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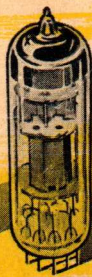
**WORLD-
SERIES**



EB 91



ECC 81



ECL 80



EF 80



PL 82



PL 83



PY 80 - PY 82



PL 81



EY 51

RECEIVING TUBES FOR

Television

PHILIPS ELECTRONIC TUBE DIVISION

WORLD SERIES
RECEIVING TUBES
for
TELEVISION

ECC 81

EF 80

EB 91

PL 83

PL 82

ECL 80

PL 81

PY 80

EY 51

PY 82

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not imply a licence under any patent.*

PREFACE

Within the space of a mere fourteen years, television has developed from a scientific toy to a public service which daily provides entertainment and instruction to hundreds of thousands of people on both sides of the Atlantic.

The scepticism which originally prevailed concerning the possibility of exploiting the new art has given place to complete confidence in both technical and commercial circles, and television now supports a thriving industry which is assured, by plans already well advanced, of rapid growth and development not only in Great Britain and in the U.S.A. but also in countries which, so far, have not instituted a television service.

Many problems had to be solved before television could be brought to its present state of perfection at which, from the technical point of view, mass popularity can be assured. But mass popularity depends not only upon technical perfection but also upon the availability of receivers of good performance at low cost. The production of such equipment is, of course, the task of the setmaker.

In the design of high-quality television receivers, the tubes employed play an important part. Tubes originally produced for normal broadcast reception, however efficient they may be for that purpose, are not necessarily the most suitable for television reception. Indeed, such have been recent developments that it is safe to say that unless tubes which have been specially developed for television service are employed, receivers of the highest performance combined with low cost cannot be realized.

Setmakers will find in this publication details of a number of new tubes, all of which meet the highest requirements. In their design, electrical properties consistent with the best performance of their various functions has been the prime consideration, and the tubes are therefore suitable for use in equipment even in the highest price class.

At the same time adaptability to receivers of low cost has not been overlooked. For example, adoption of the so-called "DC/AC" technique makes possible substantial economies in the dimensions, weight and cost of the equipment, since a heavy and expensive mains transformer is rendered unnecessary.

Based upon the latest practice in tube design, and produced in the Noval execution, the new tubes combine the highest electrical performance with very compact size — even in the case of the power types such as the line output tube.

On all counts, therefore, the tubes represent an important contribution to modern television receiver technique.

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INTRODUCTION

The special Bulletin (E.Z. 3307 E 3/50), dealing with the characteristics and the application of television receiving tubes, published in the spring of 1950, has aroused considerable interest. Since then the spectacular development of television receiver technique has continued and it has become necessary to edit a new and much revised Bulletin in which several new types of receiving tubes are included. This Bulletin, therefore, gives characteristics and application notes on the following range of tubes from which a suitable type can be chosen for every possible function in a modern television receiver.¹⁾

PL 81 (21A6) Line-output pentode; Noval base.
 PY 80 Booster diode; Noval base.
 PY 82 Mains rectifier; Noval base.
 EY 51 (6X2) High-voltage half-wave rectifier; wire connections.

Before describing in detail each of the new tubes it will be advantageous to review them as a whole and to classify them according to their functions in a television receiver. For this purpose a block diagram showing the main parts of a television receiver is represented in fig. 1. In this diagram the appropriate types are indicated for each func-

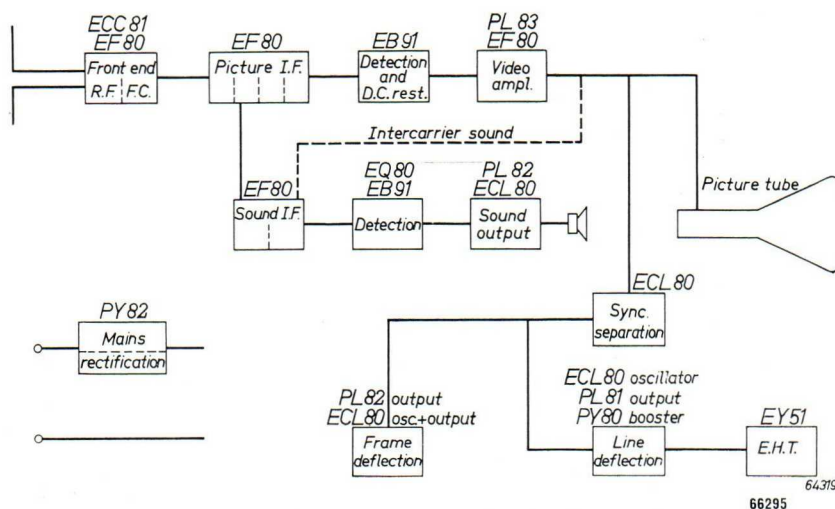


Fig. 1. Block diagram of a typical television receiver.

ECC 81 (12AT7) Double triode for R.F. amplification or frequency changing; Noval base.
 EF 80 (6BX6) R.F. pentode; Noval base.
 EB 91 (6AL5) Double diode; miniature 7-pin base.
 PL 83 (15A6) Video output pentode; Noval base.
 PL 82 Output pentode, frame or sound; Noval base.
 ECL 80 (6AB8) Multi-purpose triode-pentode; Noval base.

¹⁾ The deflection circuits described in this Bulletin are based upon the use of a narrow-angle picture tube. A Bulletin giving circuits for wide-angle deflection will be available shortly.

tion and it is seen that often more than one type is available for one specific application. This is further elucidated in the following notes.

Front end

In the so-called low band from 40 to 70 Mc/s the EF 80 can be used for R.F. amplification and for frequency changing. In the high band from 174 to 216 Mc/s the application of the double triode ECC 81 has advantages owing to the lower noise level that can be obtained with this tube, but the EF 80 is also suitable at these frequencies, especially in more simple receivers where the maximum of sensitivity is not required.

Picture I.F.

According to the sensitivity aimed at in the design, four or three stages of picture I.F. amplification preceding the video detector will normally be used. The tube for this application is the EF 80.

Video detection and D.C. restoration

For this application the double diode EB 91 is available. The cathodes of this tube are separate, so that one section may be used as the video detector and the other as D.C. restorer.

Video amplification

Two types, the PL 83 and the EF 80, are indicated for this application. In receivers with a single video stage the EF 80 can be used but the PL 83 is capable of providing a higher undistorted output voltage at a given bandwidth or, alternatively, a larger bandwidth at a given maximum output voltage. For systems with the higher definition the PL 83 is therefore to be preferred. When two stages of video amplification are used the EF 80 may serve as the pre-amplifier followed by a PL 83. It should be borne in mind, however, that, in view of the possibility of microphony, the amplification between the input of the EF 80 and the control electrode of the picture tube should be kept below 25.

Sound I.F.

In the conventional arrangement where the sound I.F. is tapped off from the picture I.F. amplifier, two stages of amplification equipped with EF 80 will normally be required preceding the F.M. detector. In a receiver with intercarrier sound one stage of sound I.F. with the EF 80 will be sufficient, whilst, depending upon the detector system employed, it will sometimes be possible to dispense with separate sound I.F. amplification.

Sound detection

For this application the special F.M. detector and limiter tube, type EQ 80¹⁾, is available. In other detector systems operating with a double diode, the EB 91 may be used.

Sound output

Two different types are available for this application. The pentode section of the ECL 80 can provide an output of 1.75 W at a supply voltage of 200 V. With certain types of F.M. detector giving

a low A.F. output some pre-amplification preceding the output stage is required. The use of the ECL 80 has the advantage that its triode section can be used for this purpose. If a higher output is required the pentode PL 82 can be used, which is capable of giving an output of about 4 W. With the EQ 80 as F.M. detector, giving a high output voltage, pre-amplification between the detector and the PL 82 is not required.

Sync. separation and amplification

Owing to its short grid base, when the pentode section of the ECL 80 is operated with a low screen-grid voltage, this tube is eminently suitable as a sync. separator. The triode section may then be used for further clipping and amplification of the sync. signal.

Frame deflection

In this section of the receiver the ECL 80 can serve for two functions, viz. the triode as blocking oscillator and the pentode as frame output tube. The pentode section can provide sufficient peak anode current in all cases where a narrow-angle picture tube is used. With double-D scanning, which involves a higher E.H.T. voltage for the picture tube, it will be necessary to feed the anode of the pentode section from the boosted high tension in the line output circuit. For all other applications where a higher peak anode current than the ECL 80 can deliver is required, in particular with wide-angle picture tubes, the output pentode PL 82 can be used.

Line deflection

In the oscillator section of this part the ECL 80 can again be used, either with the pentode or the triode section as blocking oscillator or with both sections in a multivibrator arrangement.

A special line output pentode having a high anode insulation and which is capable of delivering a high peak anode current is the PL 81. This pentode will normally be used in combination with the booster diode PY 80, in a circuit for increasing the efficiency of the output stage and for improving the linearity of the scan. Details of this circuit are given in the application notes on the PL 81.

E.H.T. rectifier

In modern television receivers the E.H.T. is normally obtained by rectifying the flyback pulses in the line circuit. A special diode of the wire-in type for this application is the EY 51. The heater

¹⁾ This tube is described in another Bulletin in this series, No. 20/62/3500 E 11/50, dealing with F.M. reception.

wattage of the EY 51 is only small, so that it is possible to feed the heater from a separate winding on the line output transformer.

Mains rectifier

For the power supply section of the receiver the PY 82 is available. This half-wave rectifier can deliver a rectified output current of 180 mA, so that in a simple receiver one tube is sufficient, whilst in a more elaborate design two tubes are required.

From the above notes it may be seen that the range of television tubes presented in this Bulletin covers all possible functions likely to be encountered in a receiver. For several functions more than one type is available, so that a suitable tube lay-out can always be found for any television standard now in use, whilst in the tube characteristics there is ample reserve to cover possible future developments.

In recent years the setmaking industries have contributed much towards reducing the cost, size and weight of receivers, for instance by the introduction of the transformerless technique. In such a receiver the heaters of the tubes must be connected in series. For this reason the range of receiving tubes presented here has been designed for series operation of the heaters, the heater current being 0.3 A. Of the PL 81, the PL 83 and the PY 80, having a heater voltage greater than 6.3 V, equivalent E-types are also available.

The adoption of a heater current of 0.3 A is a deviation from the conventional practice with broadcasting tubes, where a heater current of 0.1 A is used. There are, however, some good reasons for this. In the first place, in a television receiver one heater chain would be impossible with 0.1 A heaters, so that several chains would be needed in parallel, each with its own series resistor, and this, of course, would be a more expensive arrangement. With tubes having a heater current of 0.3 A, in the majority of cases one single heater chain suffices and consequently this is the cheapest solution. Secondly, the adoption of a heater current of 0.3 A has made it possible to design some of the tubes in this range for a heater voltage of 6.3 V, so that these types are suitable for both series and parallel operation.

There are, of course, many possible arrangements of the supply system of a receiver. The arrangement in a receiver with parallel operation of the heaters needs no special comment. In a receiver designed

for series operation of the heaters, at the higher mains voltages one heater chain will be employed, whilst the required high-tension supply is obtained by direct rectification from the mains. At the lower mains voltages two heater chains will normally be required, whilst the high tension may be obtained either by direct rectification from the mains or by the inclusion of a comparatively small auto-transformer between rectifier and mains. Some special points of interest relating to series operation of heaters in a television receiver remain to be discussed.

The tubes described in this Bulletin have been so designed as to have a very favourable heater-cathode insulation. There is, however, always inevitable capacitance between the heater and the other electrodes, particularly the cathode, and this may result in objectionable hum interference if the position in the heater chain of the particular tube in which the interference might be caused is not judiciously chosen. One end of the heater chain is normally connected to the receiver chassis, to which most of the cathodes are also connected. It is therefore advisable to choose the position in the heater chain of, for example, the frequency changer and the picture tube in such a way that the potential between heater and cathode is a minimum. The heaters of other tubes not likely to cause hum interference, such as the mains rectifier, should then be included in the other end of the heater chain. It is also possible that undesirable coupling effects are encountered between the various stages of the receiver via the heater chain. It is then necessary to use by-pass capacitors of about 1500 pF between the heater terminal on the socket and the chassis, this applying especially to the picture channel of the receiver. Finally, the following safety measure should always be included to prevent temporary overheating of the cathodes of some of the tubes: When the set is switched on the heater voltage of those tubes having the lowest heater wattage — which implies a low thermal inertia — will temporarily exceed the rated value, and this tends to reduce the useful life of these tubes. This can be avoided by using a series resistor having a high negative temperature coefficient (NTC resistor), so that the heater current is initially low and the warming up is governed by the thermal inertia of the NTC resistor.

R.F. DOUBLE TRIODE ECC 81

DESCRIPTION

The ECC 81 is a double triode on Noval base, intended primarily for use as oscillator-mixer and as R.F. amplifier in television receivers. A centre-tapped heater is employed, giving alternative ratings of 6.3 V, 0.3 A and 12.6 V, 0.15 A.

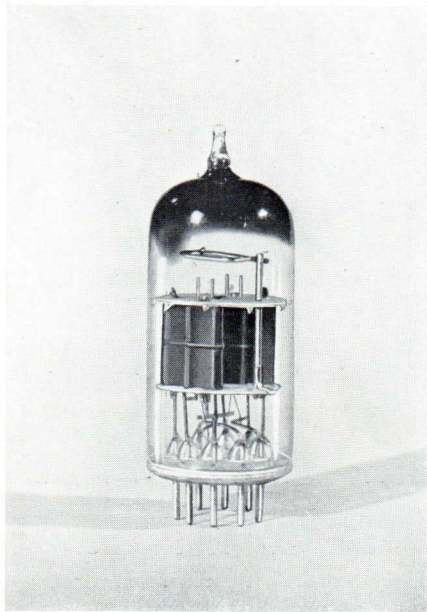


Fig. 2. The double triode ECC 81.

The higher performance of a triode compared with a pentode in the frequency range from 100 Mc/s to 300 Mc/s is well known and the provision of two high-quality triodes with separate cathodes in a single envelope ensures maximum circuit economy and flexibility. This double triode is suitable for all the radio-frequency stages in a receiver, i.e. R.F. amplifiers, mixers and local oscillators.

The electrical characteristics of the tube have been designed to meet the combined requirements of an efficient amplifier and a general-purpose radio-frequency triode. The interelectrode capacitances have therefore been kept low and the slope and the amplification factor fairly high.

APPLICATION

1. THE ECC 81 AS R.F. AMPLIFIER

In the 100 Mc/s to 300 Mc/s frequency bands the noise level due to static and cosmic noise is very low, but full advantage of these favourable conditions can only be gained when the receiver noise level is reduced to a minimum. This can be done by using triodes in the R.F. amplifier and mixer stages.

A triode can be connected in three ways — grounded grid, grounded cathode and grounded anode. The theoretical noise factors for the three circuits are very similar but the terminal impedances are different and in practice it has been found that the grounded-grid and grounded-cathode circuits give the best results.

The grounded-grid circuit has been widely used as a single-stage amplifier, since the grid forms an effective screen between input and output circuits and reduces regenerative feedback to a tolerable level.

A single grounded-cathode stage can be used but neutralization of the feedback impedance between anode and grid is difficult, especially when the circuit has to operate over a range of frequencies. However, when the grounded-cathode stage is followed by a grounded-grid amplifier the voltage gain of the first stage is approximately unity and neutralization can be accomplished in a simple way. This cascade combination of two triodes has an overall stability and voltage gain as good as those of an equivalent pentode but with the lower noise factor of a triode. Using an ECC 81, this circuit can be constructed with one tube.

In some receivers it may be advantageous to use a balanced push-pull circuit and then again the ECC 81 is eminently suitable. The receiver designer is therefore able to choose any one of a number of circuits according to individual requirements.

Grounded grid

A typical grounded-grid circuit is shown in fig. 3. This circuit is essentially a feedback amplifier, as the output current flows through the input circuit. The input impedance due to feedback is $(R_i + R_a)/(\mu + 1)$, where R_a is the resonant anode load

resistance. For values of R_a which are small compared with R_i , this expression reduces to $1/S$, which is about $200\ \Omega$ for the ECC 81. The input impedance is therefore very low and the voltage gain in the aerial circuit L_1 is small, but on the other hand the bandwidth is very large and over a very wide range of frequencies no tuning is necessary. The cathode coil L_1 can be used to match directly to an unbalanced feeder by means of a suitable tap, or it may be replaced by a balance-to-unbalance transformer to match a balanced feeder.

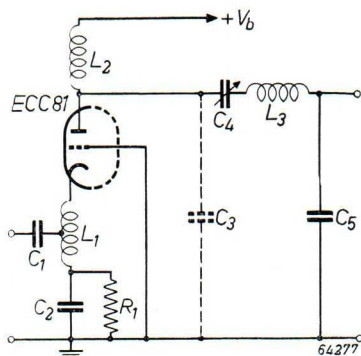


Fig. 3. One section of the ECC 81 as a grounded-grid amplifier.

An auto-transformer may be used for the output coupling but at the higher frequencies a π -filter is more convenient. The input capacitance of this filter C_3 is the output capacitance of the tube. The capacitance C_5 is adjusted to give impedance matching to the subsequent stage and C_4 is the tuning capacitor. The input impedance of the filter is approximately $(C_5/C_3)^2 R_{in}$, where R_{in} is the input impedance of the next stage, and this has to be matched to the output impedance of the tube for optimum power transfer. This output impedance is $R_i + \mu R_g$, where R_g is the total effective dynamic resistance in the grid circuit. R_1 , C_2 and L_2 are the bias components and the choke for the anode supply.

Typical figures for the performance of an ECC 81 in this circuit at a frequency of 200 Mc/s are: gain 12 db and noise factor 7.5 at a bandwidth of 6 Mc/s. In a practical circuit arrangement the two halves of the ECC 81 may be used separately for different frequency bands or, alternatively, may be connected in parallel as a single triode with double the slope and half the anode impedance.

Grounded cathode

In contrast with a pentode a single grounded-cathode triode amplifier is not suitable as a radio-

frequency amplifier in a receiver since the necessary neutralization of the anode-grid feedback capacitance is difficult to maintain, especially over a wide range of frequencies. Without neutralization stability can be achieved by employing a very low voltage gain, with a resultant loss in voltage sensitivity and an increase in noise level due to the second amplifier or mixer.

These disadvantages can be eliminated by using a cascade combination of two triodes, i.e. the two sections of an ECC 81. A non-critical neutralizing circuit can then be used purely to obtain the minimum noise factor and without affecting the stability. A typical circuit is shown in fig. 4. In this circuit L_2 is the coil for compensating the anode-grid capacitance of the first triode. Either L_3 or L_5 may be tuned to the signal frequency, L_4 and L_3 or L_5 being R.F. chokes.

The anode of the grounded-cathode triode is coupled to the cathode of the second triode, which is a grounded-grid amplifier. The input impedance of the grounded-grid stage is approximately $1/S$, so that the voltage gain of the first triode is unity. The output current of the first stage flows through the cathode circuit of the second stage, with the result that the overall gain of the system is SR_a , i.e. that of a pentode with equivalent mutual conductance. The combined circuit is stable, the overall

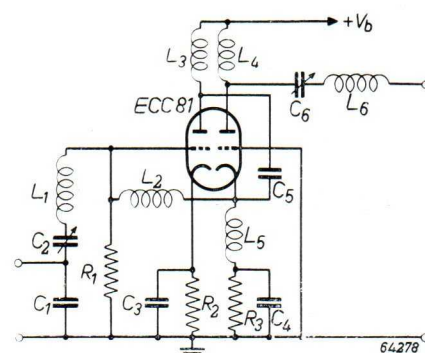


Fig. 4. Cascade combination of the two triodes of the ECC 81. One section is operating as a grounded-cathode and the other as a grounded-grid amplifier.

gain is high, and owing to the power gain of the first stage and the negative feedback of the grounded-grid amplifier the noise factor is that of the first triode.

Typical performance figures for a circuit of the type illustrated at a frequency of 200 Mc/s are: gain 13 db and noise factor 6.5 at a bandwidth of 11.5 Mc/s.

This cascade combination of the ECC 81 is very

suitable for use in a tuner in conjunction with another ECC 81 operating as mixer and local oscillator. For some television applications it may be necessary to provide two distinct frequency bands with switched channels in each band. Channel selection can be accomplished by tuning all the R.F. amplifier and oscillator circuits, but the relatively complex tuning mechanism tends to lower the performance of the receiver compared with a fixed-tuned circuit. An alternative is to use a high intermediate frequency and to increase the bandwidth of the R.F. amplifier circuits, so that channel selection can be achieved by tuning the local

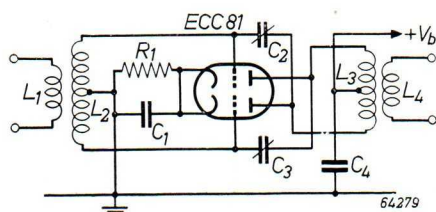


Fig. 5. The ECC 81 as a push-pull R.F. amplifier.

oscillator only. The image frequency is thereby placed outside the television bands and, although the performance of the wide-band circuits is not so good, the deterioration due to the necessity for tuning is eliminated. Owing to its high input impedance the EF 80 pentode is very suitable as a high-frequency I.F. amplifier. The reduced gain to be obtained with a wide-band R.F. amplifier increases the contribution of the mixer to the total noise and a triode mixer is therefore essential.

Push-pull amplifier

A push-pull amplifier may be conveniently constructed around an ECC 81. This circuit has a high L -to- C ratio, high gain and low noise factor. A typical circuit for operation at 200 Mc/s is shown in fig. 5.

A balanced input transformer L_1L_2 couples the balanced feeder to the control grids, and neutralizing capacitors (C_2, C_3) of the order of 1 pF are cross-connected between the respective anodes and grids. The output may be unbalanced and fed into a single-ended mixer and local oscillator, or balanced and fed into a push-pull mixer with push-pull local oscillator using two additional ECC 81 tubes. Typical performance figures for this R.F. amplifier are: gain 21 db, bandwidth 2.5 Mc/s, noise factor 5.

2. THE ECC 81 AS FREQUENCY CHANGER

As a frequency changer in high-frequency receivers the ECC 81 possesses two distinct advantages: the separate mixer and oscillator sections are contained in a single envelope, and the noise level of a triode mixer is sufficiently low to allow its use with a relatively low-gain, wide-band R.F. amplifier.

The conversion characteristics are given in figs 13, 14 and 15, p. 15-16, and it will be seen that the maximum conversion conductance is 2 mA/V at $V_b = 170$ V. Curves are also given for the anode impedance R_i and for the effective mutual conductance S , which is the value obtained when the tube is driven by both an oscillator voltage and a signal voltage of 100 mV at intermediate frequency.

A typical frequency changer circuit with the ECC 81 is shown in fig. 6. The oscillator voltage for the mixer is derived from a small capacitor of the order of 1–2 pF connected directly between the oscillator circuit and the mixer grid. This capacitor should be adjusted until the optimum oscillator drive is obtained. The oscillator voltage at the mixer grid should be 2–2.5 V according to the anode voltage used, and the drive should be adjusted so that when tuning over a frequency range its minimum value is equal to the optimum value. This is necessary to avoid an excessive spread in performance from tube to tube due to the rapid decrease in conversion conductance at low oscillator voltages.

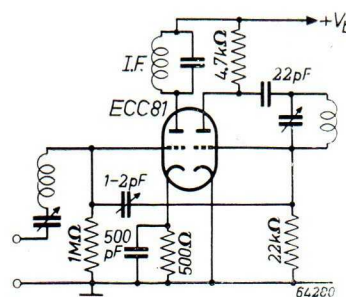


Fig. 6. The ECC 81 as a frequency changer.

The curves were obtained with a grid leak of 1 MΩ and no cathode bias was used. At $V_a = 170$ V the grid current for optimum drive is 3.3 μA, which represents a bias of 3.3 V.

In the event of a complete or partial failure of the oscillator circuit the cathode current of the mixer would become excessive. It is therefore desirable to operate the valve with cathode bias. The cathode current at optimum drive is 6.0 mA, so that a 500 Ω cathode resistor is required. Under these

conditions any value of grid leak less than 1 M Ω may be used but a high value is desirable to limit grid current. The grid current will be zero for an oscillator voltage less than that required to overcome the grid-current starting point and to drive the valve into grid current on the oscillator peaks.

The high mutual conductance of the ECC 81 makes it an efficient local oscillator at high frequencies and this tube may be used with any of the conventional oscillator circuits, of which fig. 6 gives an example. Typical performance figures for a frequency changer of this type operating in con-

junction with a push-pull R.F. amplifier as illustrated in fig. 5 are: total gain (between 50 Ω aerial and 2.2 k Ω anode load of the mixer) 31 db, bandwidth 7 Mc/s, noise factor 9. When this circuit is used in a television receiver operating without inter-carrier sound or in an F.M. receiver, vibrations in the air surrounding the tube or of the chassis may give rise to a modulation of the oscillator frequency. This can be avoided by using a screening can around the tube and, if necessary in a particular case, a resilient tube holder.

TECHNICAL DATA

HEATER DATA

Heating: indirect by A.C. or D.C.;
series or parallel supply

Heater voltage	V_f	6.3	12.6 V
Heater current	I_f	0.3	0.15 A

BASE CONNECTIONS AND DIMENSIONS (in mm)

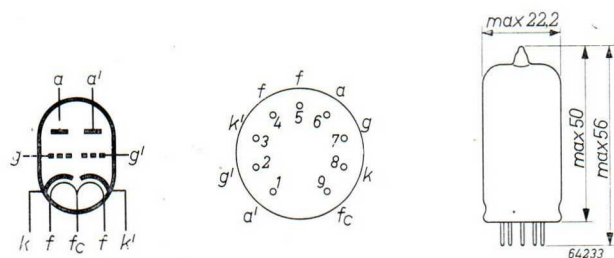


Fig. 7.

Mounting position: any

CAPACITANCES

	Section 1	Section 2
C_{ag}	1.45	1.45 pF
C_g	2.5	2.5 pF
C_a	0.45	0.35 pF
C_{ka}	0.15	0.15 pF
C_{kf}	2.5	2.5 pF
$C_{k(g+f)}$	5	5 pF
$C_{a(g+f)}$	1.6	1.5 pF
$C_{gg'}$	< 0.005 pF	
$C_{au'}$	< 0.4 pF	

TYPICAL CHARACTERISTICS

Anode voltage	V_a	170	200	250 V
Grid bias	V_g	-1 *)	-1.5	-2.35 V
Anode current	I_a	10	10	10 mA
Mutual conductance	S	6	5.5	4.9 mA/V
Amplification factor	μ	62	57	43

*) At this adjustment grid current may occur. In case where this is inadmissible it is advisable to adjust the bias at -1.5 V.

LIMITING VALUES (each section)

Anode voltage at zero anode current	V_{ao}	max. 550 V
Anode voltage	V_a	max. 300 V
Anode dissipation	W_a	max. 2.5 W
Cathode current	I_k	max. 15 mA
Heater voltage during warming up period for both sections together		max. 9.5 V *)
(for V_f normal = 6.3 V)		max. 19 V
(for V_f normal = 12.6 V)		
Voltage between heater and cathode	V_{kf}	max. 90 V
External resistance between control grid and cathode with automatic bias	R_g	max. 1 M Ω
External resistance between heater and cathode	R_{kf}	max. 20 k Ω

*) When the ECC 81 is used in a receiver or any other circuit with series supply of the heaters it is necessary to use a device limiting the initial current after switching on. This may be an NTC resistor in series with the heater chain.

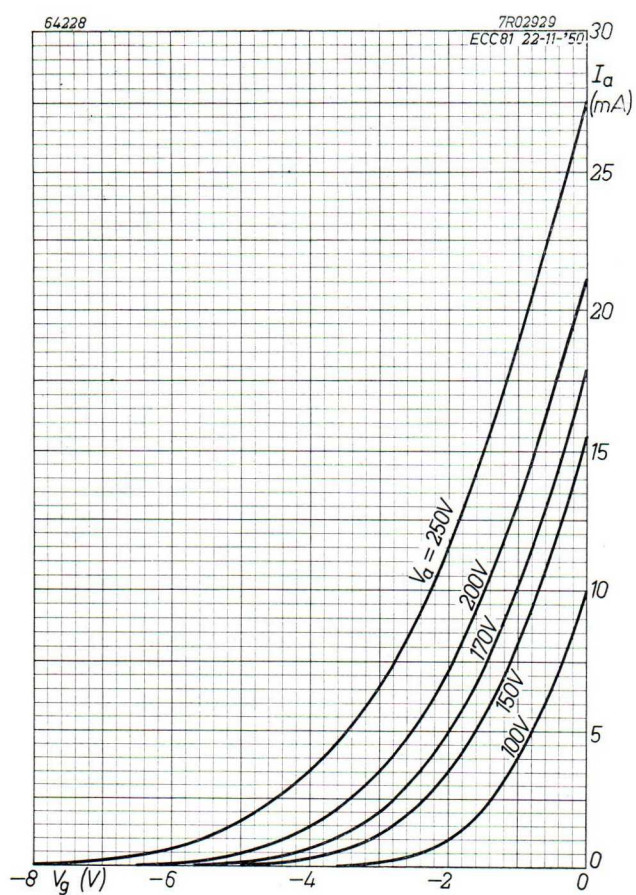
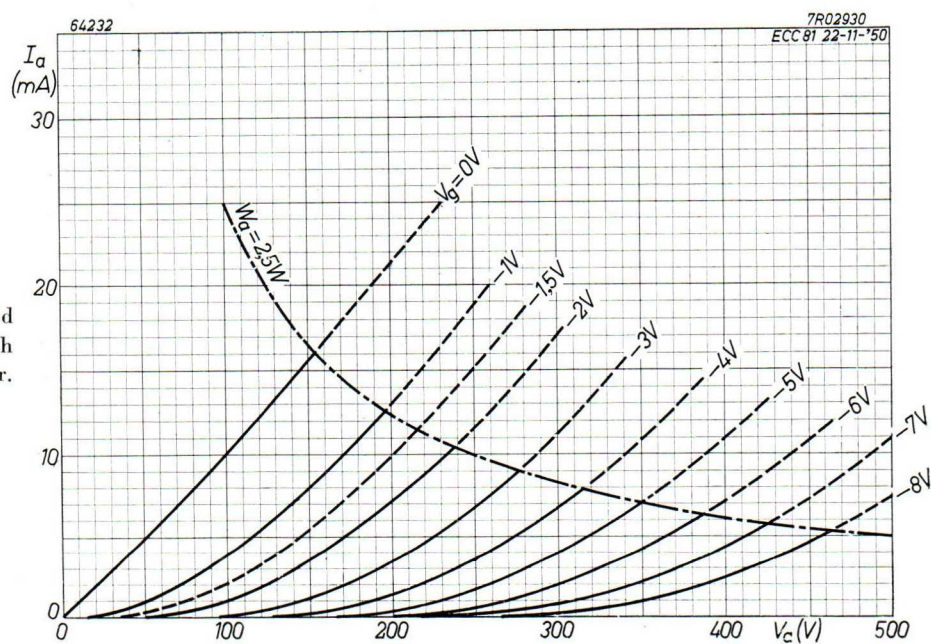


Fig. 8.
Anode current plotted
against grid voltage, with
anode voltage as para-
meter.

Fig. 9.
Anode current plotted
against anode voltage, with
grid voltage as parameter.



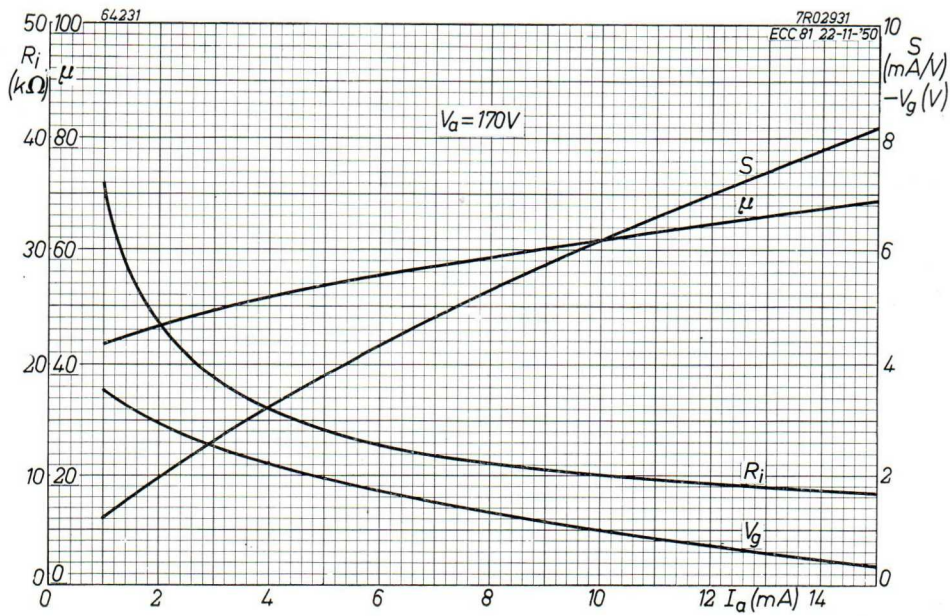


Fig. 10.
Mutual conductance, amplification factor, internal resistance and grid voltage plotted against anode current, for anode voltage of 170 V.

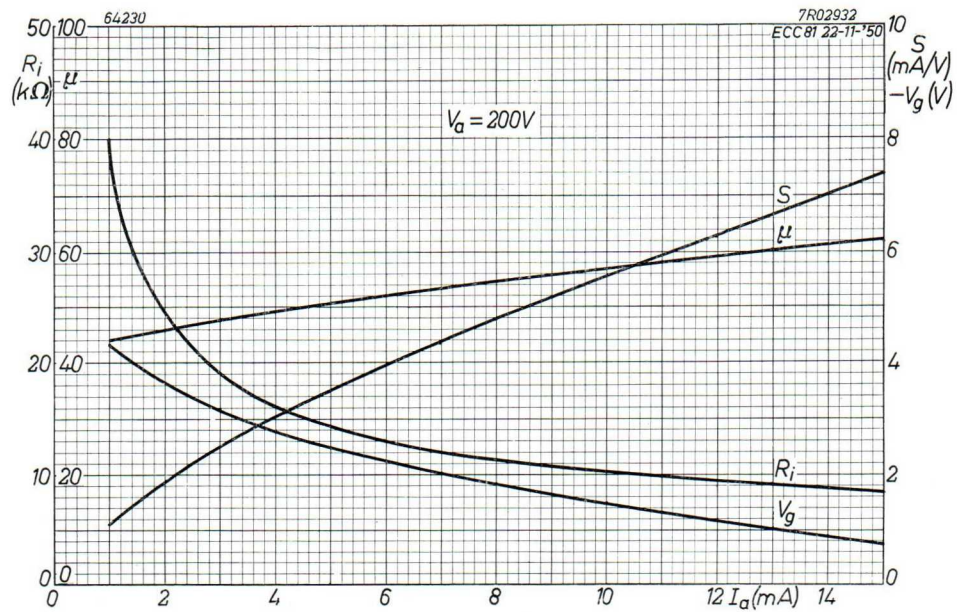


Fig. 11.
Mutual conductance, amplification factor, internal resistance and grid voltage plotted against anode current, for anode voltage of 200 V.

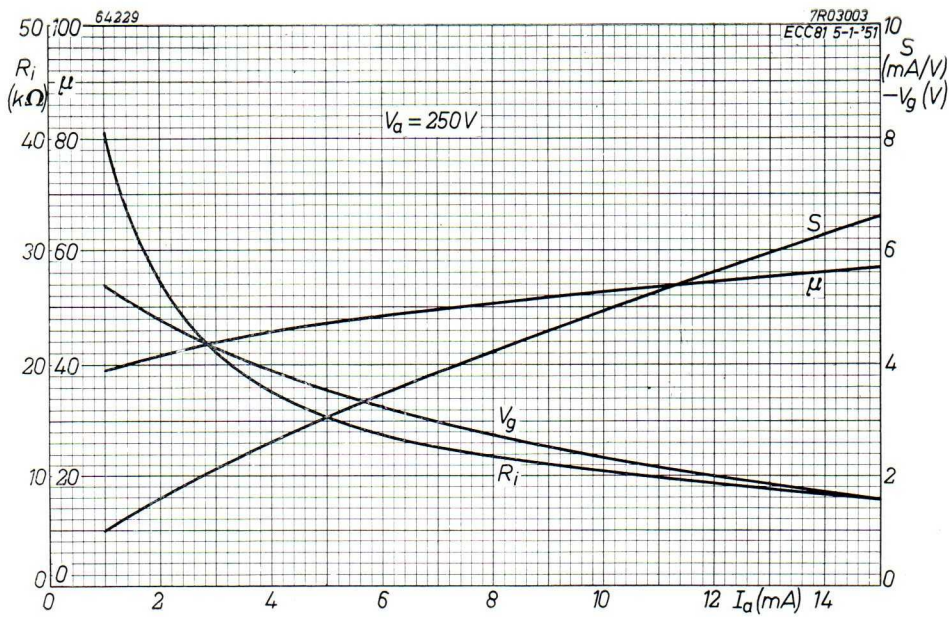


Fig. 12.
Mutual conductance, amplification factor, internal resistance and grid voltage plotted against anode current, for anode voltage of 250 V.

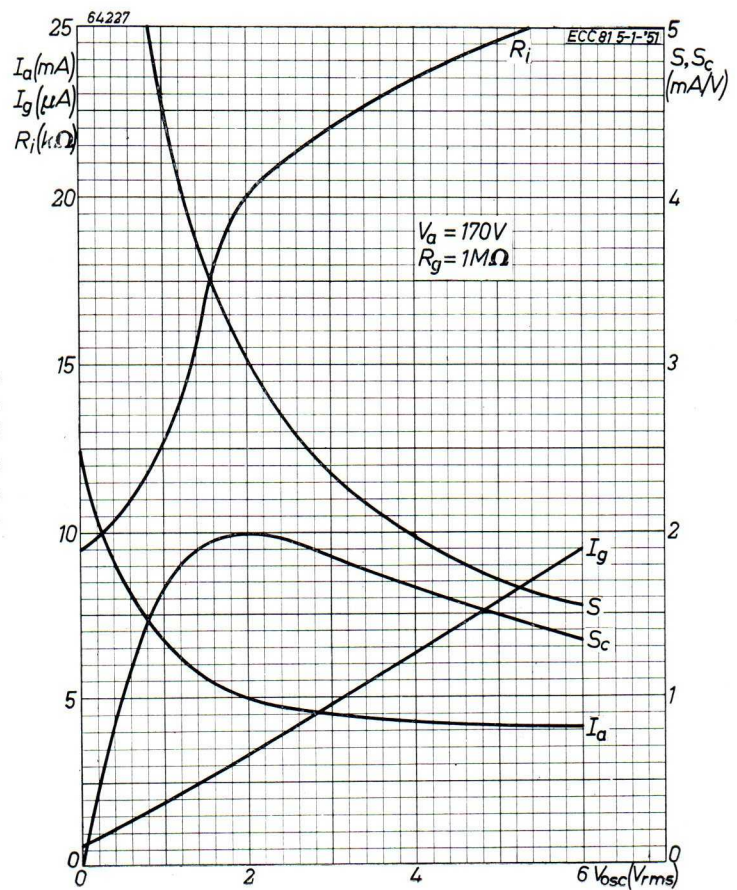


Fig. 13.
Performance curves of the ECC 81 as frequency changer at anode voltage of 170 V. S is the effective mutual conductance for a signal of 100 mV at intermediate frequency.

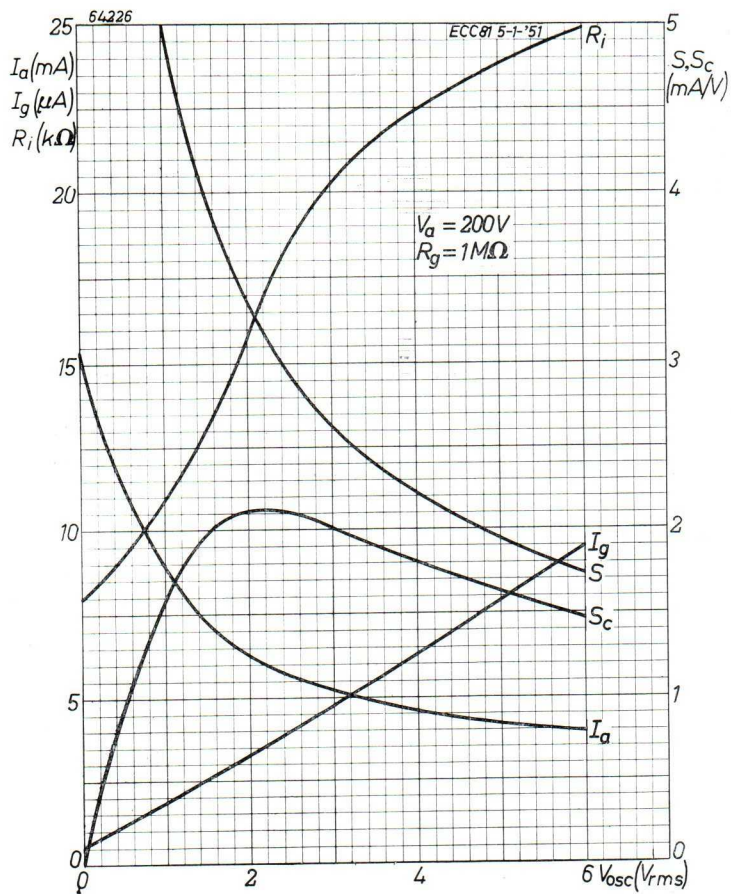
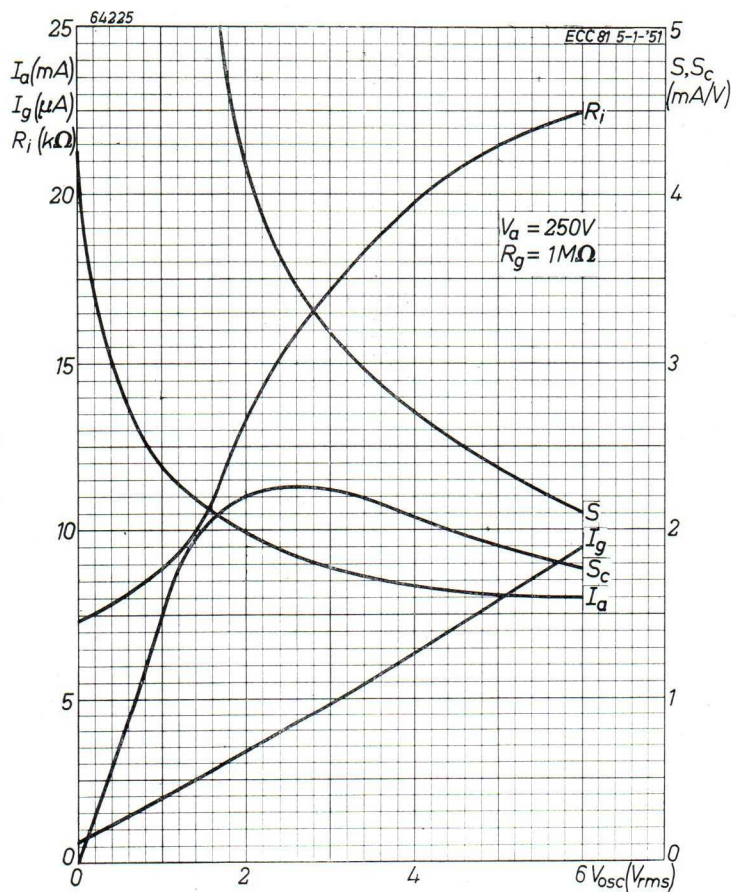


Fig. 14.
Performance curves of the ECC 81 as frequency changer at anode voltage of 200 V. S is the effective mutual conductance for a signal of 100 mV at intermediate frequency.

Fig. 15.
Performance curves of the ECC 81 as frequency changer at anode voltage of 250 V. S is the effective mutual conductance for a signal of 100 mV at intermediate frequency.



R.F. PENTODE EF 80

DESCRIPTION

The EF 80 is a high-slope R.F. pentode on the Noval base, designed primarily for wide-band R.F. and I.F. amplification in television receivers. Its characteristics render the tube suitable for use also as a video amplifier in receivers of a simple type, as a frequency changer, and in some forms of synchronizing-pulse separator circuits.

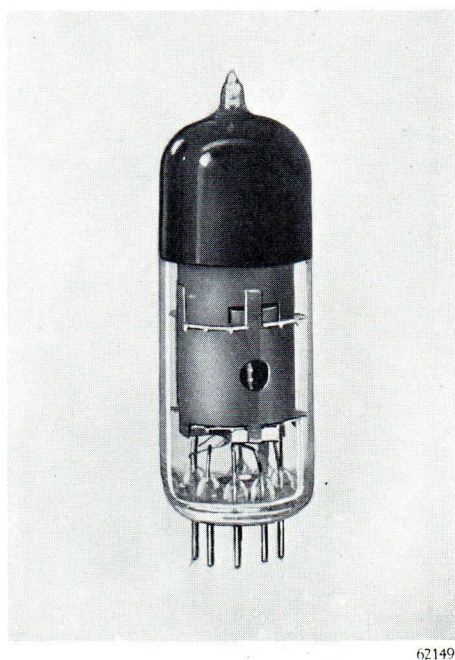


Fig. 16. The R.F. pentode EF 80 (actual size).

As with other tubes in the range described in this documentation, the EF 80 can be used with an H.T. line voltage of 170 V without sacrificing performance. A line voltage of about 170 V is commonly encountered in television receivers with a transformerless supply system when the mains voltage is 220 V. Due to the provision of two cathode leads, the input conductance has been reduced, and this is of particular importance at the higher frequencies in the television bands.

APPLICATION

1. THE EF 80 AS I.F. AMPLIFIER

The carrier frequencies of the low television band for channels I to IV of the 625 line system lie

between about 40 and 70 Mc/s. Two possible circuit arrangements are available for the amplification of these signals to a level suitable for video detection. In receivers intended for the reception of one channel only a T.R.F. amplifier can be used but, owing to the high carrier frequencies, efficient screening is required, whilst it is also difficult to adapt such a receiver for the reception of more than one channel, if this should become necessary. A superhet arrangement is therefore often preferred and most of the gain is then provided by the I.F. amplifier, the preceding stages mainly serving for frequency conversion and some pre-selection. Moreover, the frequency response of the R.F. part of the receiver is usually very flat, so that the shape of the response curve of the receiver is almost entirely determined by the I.F. circuits.

A normal figure for the required I.F. signal at the video detector for average picture contrast with a direct-viewing picture tube and one stage of video amplification is 2 V. Assuming that one stage of R.F. amplification and a frequency changer are employed, giving a total gain of 20 between the aerial terminals and the grid of the first I.F. tube, a total I.F. gain of 10,000 is required when the sensitivity for the centre frequency of the pass band has to be $10 \mu\text{V}$, and one of 1000 when a sensitivity of $100 \mu\text{V}$ is considered sufficient. A convenient figure for the 3 db bandwidth with the 625 line system is 5 Mc/s and, in order to keep image frequencies outside the television band and to minimize feedback of harmonics of the detected I.F. signal into the first stages of the receiver, the centre of the I.F. band should be chosen at about 21 Mc/s.

A basic diagram for a four-stage stagger-tuned I.F. amplifier is given in fig. 17. A very useful characteristic quantity of a tube is the product of gain and bandwidth, GB , which is equal to $S/2\pi(C_i + C_o)$, in which S is the static mutual conductance and C_i and C_o are the input and output capacitances respectively. With the EF 80 this figure is 110 and this would imply that a voltage gain of 22 could be achieved when the 3 db bandwidth of the anode circuit is 5 Mc/s. However, the GB figure is a typical tube characteristic and there-

fore does not take into account the unavoidable increase in total capacitance of the anode circuit caused by the wiring and the distributed capacitance of the coil. Moreover, the GB figure is given for the cold capacitance of the tube, and for a tube

the centre of the I.F. band and the horizontal distances of the other perpendiculars represent the detuning of the other circuits with respect to this centre. Moreover, the length of each perpendicular gives half the required bandwidth of each circuit.

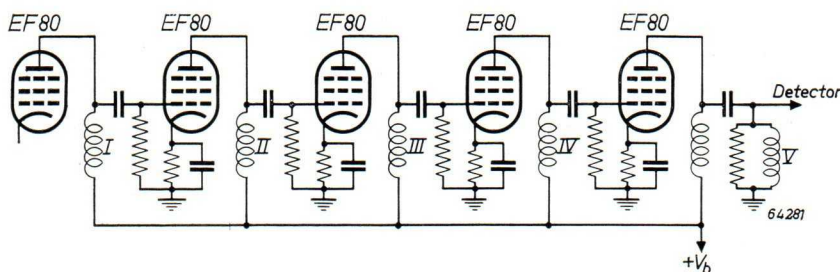


Fig. 17. Basic diagram of a four-stage stagger-tuned I.F. amplifier.

in operation the actual capacitances are somewhat greater. A practical figure for the product $G \cdot B$ of the EF 80 is 55 when the capacitance in the anode circuit is not intentionally increased, which implies tuning of the circuits by means of slugs of powdered iron. A stage gain of 11 can therefore be obtained when the bandwidth is 5 Mc/s. This method of calculation may also be applied to a stagger-tuned amplifier when the circuits are mutually detuned and damped according to the method of flat staggering¹⁾. A four-stage amplifier with EF 80 will therefore provide a total I.F. gain of 10,000 and one with three stages a total gain of 1000. With these figures allowance is made for some decrease in gain due to the inclusion of trap circuits for the suppression of sound, adjacent sound and adjacent picture signals. Also the inclusion of small non-bypassed cathode resistors, for compensating the variation of the input capacitance of the tubes caused by the contrast control, will result in a somewhat lower gain. A simple method for determining the tuning frequencies of each of the five tuned circuits of the four-stage amplifier of fig. 17 and the required bandwidths of each circuit will now be given.

From a semi-circle, subdivided into 5 equal parts, a perpendicular is dropped to the diameter from the centre of each part (see fig. 18). The scale of this figure should be so chosen that the diameter represents the required 3 db bandwidth of the whole amplifier, which in our case is 5 Mc/s. The base of the central perpendicular now represents

The tuning frequencies and bandwidths of the circuits of fig. 17 may now be tabulated as follows:

Circuit No.	Tuning frequency	Bandwidth
I	18.6 Mc/s	1.6 Mc/s
II	19.5 Mc/s	4 Mc/s
III	21 Mc/s	5 Mc/s
IV	22.5 Mc/s	4 Mc/s
V	23.4 Mc/s	1.6 Mc/s

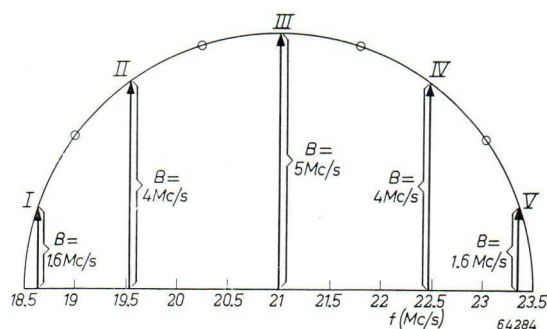


Fig. 18. Stagger diagram for a four-stage I.F. amplifier centred at 21 Mc/s.

The staggering scheme for a three-stage I.F. amplifier also with a bandwidth of 5 Mc/s and a centre frequency of 21 Mc/s is represented in fig. 19. Such an amplifier with three EF 80 tubes provides a gain of 1000 and the stagger frequencies and bandwidths are:

Circuit No.	Tuning frequency	Bandwidth
I	18.7 Mc/s	1.9 Mc/s
II	20.05 Mc/s	4.6 Mc/s
III	21.95 Mc/s	4.6 Mc/s
IV	23.8 Mc/s	1.9 Mc/s

¹⁾ See A. G. W. Uitjens, Fundamental Problems of H.F. and I.F. Amplifiers for T.V. Reception, Part I: Gain and Bandwidth, Electr. Appl. Bull. 11, p. 205, 1950 (No. 11).

From this stagger diagram it appears that it is not always necessary to have a circuit tuned to the centre frequency. In fact such a circuit will only occur in an amplifier having an uneven number of tuned circuits.

The required bandwidth of each individual circuit is usually obtained in the following way. All coils of the amplifier are given equal Q 's and the bandwidth of each circuit is adjusted to the desired value by choosing a suitable value for the grid leak resistor (fig. 17). In determining the value of this resistor it is not necessary to take the input damping of the EF 80 into account, this damping having a

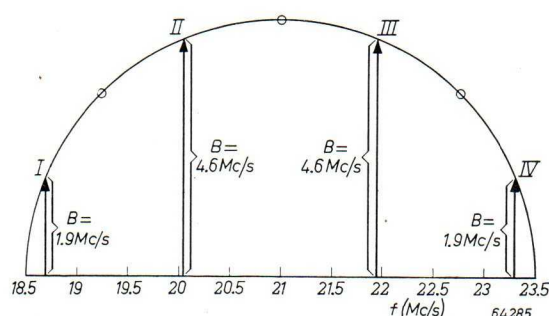


Fig. 19. Stagger diagram for a three-stage I.F. amplifier centred at 21 Mc/s.

very favourable value. The damping on the last I.F. circuit caused by the detector diode, however, may be quite considerable, its value being approximately equal to the diode load. When the smallest bandwidth of the circuits is obtained from the stagger diagram the minimum Q of the coils may be calculated with the aid of the following formula:

$$Q = \frac{f_0}{B},$$

in which f_0 is the tuning frequency of the circuit in question and B the bandwidth. For instance, the minimum Q of circuit No. I in the stagger diagram of fig. 18 is $Q = 18.6/1.6 = 11.6$. In the examples given above the total tuning capacitances have been assumed to be 22 pF, which is about twice the sum of the input and output capacitances of the tube. The total parallel resistance of this circuit is now:

$$R = \frac{Q}{2\pi f_0 C} = \frac{11.6}{2\pi \times 18.6 \times 10^6 \times 2.2 \times 10^{-11}} = 4500 \Omega.$$

Assuming the actual coil Q to be 25, this being a value which can easily be reached in practice, the parallel resistance of the circuit would be $25/11.6 \times 4500 = 9700 \Omega$. In order to arrive at the required total parallel resistance of 4500 Ω , a grid

leak of 8400 Ω should be chosen in this case. The grid leaks of the other stages can now be calculated in a similar way.

Referring once more to fig. 18, it will be clear that in the case of large numbers of mutually detuned circuits the bandwidths of the outermost circuits in the stagger diagram may become inconveniently small. This difficulty can be avoided by using two or more identical staggered groups in cascade, but this involves some loss of total gain. According to this method a five-stage amplifier, thus with six tuned circuits, can be composed of 2 identical groups of 3 circuits or of 3 groups of 2 circuits. If, however, the 3 db bandwidth of each group is again chosen equal to the required total bandwidth of the amplifier, the drop at the limits of the total pass band will be 6 db when two identical groups are employed and 9 db in the case of three groups in cascade. It is therefore necessary to give each group a bandwidth greater than the total bandwidth of the amplifier. The factors with which the required total bandwidth of the whole amplifier must be multiplied to obtain the bandwidth of each group can be found in the following table.

Number of staggered circuits in each group	Number of identical groups				
	1	2	3	4	5
1	1	1.56	1.96	2.27	2.56
2	1	1.25	1.41	1.54	1.61
3	1	1.16	1.25	1.32	1.37
4	1	1.11	1.19	1.23	
5	1	1.09	1.15		
6	1	1.075	1.11		
7	1	1.05			

From this table it appears that the required bandwidth of each group increases with the number of groups employed. In an amplifier with six tuned circuits, for example, the factor is 1 in the case of one staggered group, 1.16 for two identical groups of three circuits and 1.41 for three groups of two staggered circuits. In the latter case when the total bandwidth of the amplifier has to be 5 Mc/s each group must have a bandwidth of 7 Mc/s and, since the product of gain and bandwidth is constant for a given type of tube, the result will be a smaller gain.

It is interesting to note the difference in perform-

ance with regard to gain between an amplifier employing circuits all tuned to the same frequency and a stagger-tuned amplifier. The four-stage stagger-tuned amplifier of fig. 17 gives a gain of 10,000. In another four-stage amplifier, however, in which the five circuits are all tuned to the same frequency, the bandwidth of each individual circuit would have to be 2.56 times the required total bandwidth. Whilst the stage gain of the stagger-tuned amplifier was 10, in the latter case this would be reduced to about 4, giving a total gain of $4^4 = 256$.

The inclusion of traps for the suppression of the sound carrier and for adjacent sound and picture signals will influence the shape of the response curve of the I.F. amplifier. This will usually necessitate small corrections of the tuning frequencies and dampings of the I.F. circuits when these are determined according to the method represented in figs 18 and 19. Also the sequence of the circuits will often be chosen differently, for various reasons, although this does not influence the shape of the response curve. The first stages of the picture I.F. amplifier are normally also used to amplify the I.F. sound signal, which is usually tapped off after the second I.F. stage. Economies may then be effected in the design of the sound I.F. amplifier by choosing the most favourable stagger frequencies with regard to sound I.F. amplification for the first stages of the picture I.F. amplifier. This, however, often conflicts with the requirements of picture contrast control and a compromise must then be sought. The contrast of the television picture is normally adjusted by varying the bias of the tubes in the first stages of the receiver. Owing to unavoidable oscillator-frequency shift the bias of the mixer stage may not be varied and contrast control must therefore be achieved by varying the bias of the R.F. stage and of the first two I.F. stages. Variation of the bias, however, results in a variation ΔC_i of the input capacitances of the tubes and this may upset the total response curve of the receiver. Therefore, those I.F. circuits from the stagger diagram having the largest bandwidths should be used to precede those tubes of which the grid bias is varied, as this keeps the influence of the contrast control upon the total response curve as small as possible. As may be seen from the curves in the data section, when the slope of the tube is varied between 7.4 and 0.74 mA/V, the variation in input capacitance of the EF 80 is 1.95 pF. This variation may be inadmissible and special methods are available

to reduce this variation. A simple and efficient form of ΔC_i compensation will now be briefly described.

Compensation of input capacitance and resistance variation

To give a complete explanation of this method it would be necessary to consider many effects, including those of lead inductances, stray capacitances and transit time. The practical value of such an analysis however is limited, as lead inductances and stray capacitances occurring in production cannot be accurately predicted. The most favourable values of the components for compensating the variation in input capacitance and resistance are therefore normally determined by experiment, but the main principles should, of course, be understood.

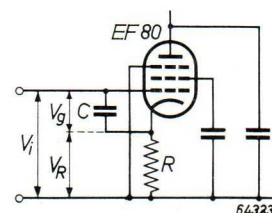


Fig. 20. Basic circuit for the compensation of input capacitance variation.

In the basic circuit of fig. 20 a non-bypassed resistor is included in the cathode lead, C representing the capacitance between grid and cathode of the tube. Due to electronic effects this capacitance varies with the bias. In the cut-off condition C has a value of 7.2 pF, this gradually increasing to about 9.4 pF when the grid bias is reduced to -2 V. In the first case the effective capacitance measured across the input terminals is practically 7.2 pF, R having a low value compared with the reactance of C . In normal operation, however, a voltage V_R occurs across the cathode resistor, so that the voltage across C is reduced by the factor $1/(1 + SR)$, in which S is the static mutual conductance of the tube. The object now is to keep the charging current of C constant, so that the effective input capacitance of C is also kept constant. In the above example the factor $1 + SR$ should therefore have a value $9.4/7.2 = 1.3$, so that, since S is 7.4 mA/V, the cathode resistor must have a value of $0.3/S = 41 \Omega$. This resistor would indeed be required for optimum compensation of the input capacitance variation, but the result would be that the effective mutual conductance of the tube is reduced by a factor

$1/1.3 = 0.77$. A lower value of R still giving adequate compensation is therefore normally used, a practical value for the EF 80 being $27\ \Omega$.

The variation in effective input resistance is also reduced by the inclusion of a non-bypassed cathode resistor. At cut-off, when the input resistance of the tube is very high, the input damping caused by the current flowing through C and R prevents the effective input resistance from reaching a very high value. However, this additional damping is also present in normal operation, so that, although the effective input resistance of the tube is somewhat increased by the effect of the negative feedback, the total effective input resistance measured across the input terminals will usually be slightly lower than that which would be obtained without a non-bypassed cathode resistor. This may be improved upon by shunting the cathode resistor by a small capacitor having a value of 5 to 10 pF.

2. THE EF 80 AS R.F. AMPLIFIER

As in all R.F. amplifiers, the connecting leads should be kept as short as possible in order to minimize stray capacitance and self-inductance. The lay-out should be such as to reduce the risk of feedback, a centre screen on the tube holder connected to the chassis being an advantage.

Normally, common decoupling may be employed for anode and screen. If trouble is experienced due to feedback through the anode-to-grid capacitance, it may be profitable to use separate decoupling for anode and screen, as this results in a lower effective feedback capacitance. In order to obtain the highest value of input resistance the cathode pins 1 and 3 should be interconnected. In the 100 Mc/s to 300 Mc/s frequency bands the noise level due to static and cosmic noise is very low. The use of a triode may therefore be preferred at these frequencies.

3. THE EF 80 AS FREQUENCY CHANGER¹⁾

Fig. 21 shows a typical circuit for operation at the lower television carrier frequencies, in which the EF 80 is employed as a self-oscillating frequency changer. The arrangement will be recognized as a conventional Colpitts oscillator with the coil connected between control grid and screen.

By connecting the R.F. input circuit L_1 to a point on the oscillator coil at which the oscillator voltage

is a minimum, coupling between the oscillator and the R.F. input is negligible, so that there is no "pulling" of oscillator frequency when tuning the R.F. circuit.

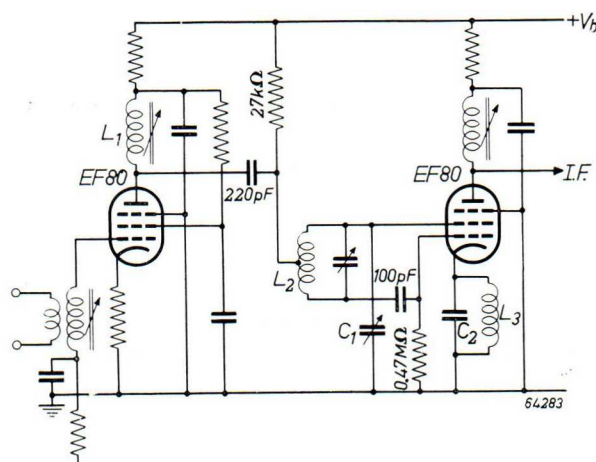


Fig. 21. Front end for a television receiver or an F.M. receiver with one EF 80 as R.F. amplifier and the other as self-oscillating frequency changer.

The oscillator coil L_2 is tapped at the centre, and the oscillator voltage at this point is made a minimum by adjusting the capacitor C_1 connected between the screen grid and earth. In practice it is possible to use a capacitor of fixed value for C_1 .

The inductance of the oscillator coil should be small compared with that of the R.F. coil L_1 , and this can be achieved by using a high Q , high C oscillator circuit, and choosing the oscillator frequency higher than the signal frequency.

The amplitude of the oscillation should be that which gives a direct voltage in the order of 2.5 V across the oscillator grid resistor. This voltage can be determined by measuring the oscillator grid current.

The input impedance of the frequency changer at carrier frequency is inherently low, mainly because the control grid is coupled to the screen. Owing to the large bandwidth employed with television this is no disadvantage. In receivers for F.M. broadcast reception, however, the required bandwidth is much smaller and the input impedance may then be increased by using a cathode capacitor C_2 , which will produce a negative component of input resistance. The value of this capacitor can best be determined experimentally by starting with, for example, 500 pF and reducing this progressively. The shunt coil L_3 , which is required for connecting the cathode to the chassis for D.C., should have a value of about one microhenry.

¹⁾ See B. G. Dammers and L. J. Cock, Comments on Frequency Changers for 30 Mc/s to 120 Mc/s for T.V. and F.M. receivers, Electr. Appl. Bull. 11, p. 105, 1950 (No. 6/7).

4. THE EF 80 AS VIDEO AMPLIFIER

The normal direct-view picture tube requires a maximum drive voltage of about 60 V peak-to-peak, including sync. signal, derived from the video stage. This signal can be delivered by an EF 80 operating with cathode compensation. In the case of a projection tube however, which requires a drive voltage of about 100 V peak-to-peak, a considerably greater anode current swing is required with the same anode load. It is then advisable to use the PL 83.

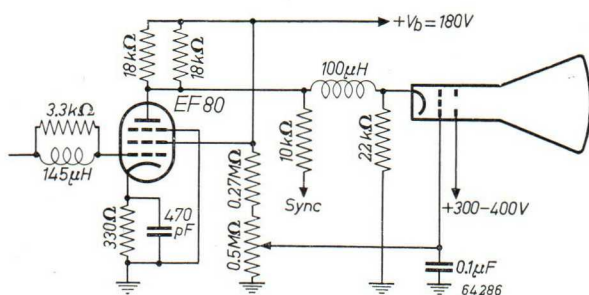


Fig. 22. The EF 80 as a video amplifier with cathode compensation.

A suitable circuit with the EF 80 for driving a picture tube of the direct-viewing type is given in fig. 22. Under quiescent conditions the tube is operated at an anode current of 6.5 mA and a screen-grid current of 1.8 mA. For 60 V peak-to-peak output voltage an anode current swing of 9.4 mA is required. The gain of the stage is 12.

A series peaking coil of 145 μ H shunted by 3300 Ω

resonates with the input capacitance of the EF 80 at about 5 Mc/s, thus improving the response at the high video frequencies. Moreover, I.F. frequencies are attenuated by this series circuit. The shunt resistor is included to prevent overshoot. It is assumed that the black level of the input signal is at chassis potential.

In this stage cathode compensation has been applied by shunting the cathode resistor with a capacitor of 470 pF. At high frequencies the negative feedback caused by the cathode resistor decreases as a result of the by-passing effect of the capacitor, thus compensating the loss in gain due to parasitic capacitances in the anode circuit. This compensation is effective up to 3.5 Mc/s.

A series peaking coil of 100 μ H between the anode of the EF 80 and the cathode of the picture tube extends the frequency characteristic to about 4.5 Mc/s. The self-capacitance of this coil should be small, and wave winding is therefore recommended.

The output signal of the video stage is applied to the sync. separator via a resistor of 10 k Ω , thus preventing an increase in the capacitance of the anode circuit. Since direct coupling is used between the EF 80 and the cathode of the picture tube, a positive variable potential can be applied to the grid for the brightness control (0.5 M Ω potentiometer). A direct voltage of 300-400 V with respect to the chassis is required for the first anode of the picture tube. In a receiver with a high-tension line voltage of about 180 V, this voltage may be taken from the booster capacitor in the line output circuit.

TECHNICAL DATA

HEATER DATA

Heating: indirect by A.C. or D.C.;
series or parallel supply

Heater voltage	V_f	6.3 V
Heater current	I_f	0.3 A

CAPACITANCES

C_{g1}	7.5 pF
C_a	3.3 pF
C_{ag1}	< 0.007 pF
C_{ak}	< 0.006 pF
C_{g2}	5.4 pF
C_{g1g2}	2.6 pF
C_{g1f}	< 0.15 pF
C_{kf}	5.0 pF

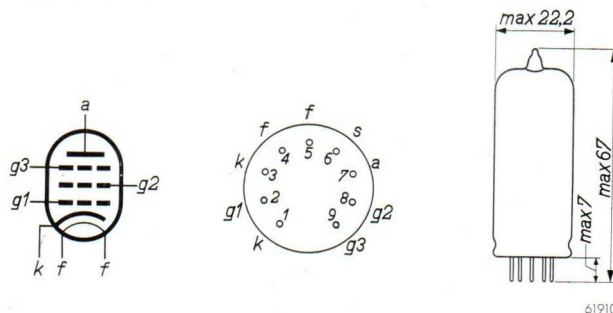
BASE CONNECTIONS AND
DIMENSIONS (in mm)

Fig. 23.

Mounting position: any

TYPICAL CHARACTERISTICS

Anode voltage	V_a	170	200	250 V
Suppressor-grid voltage	V_{g3}	0	0	0 V
Screen-grid voltage	V_{g2}	170	200	250 V
Control-grid voltage	V_{g1}	-2	-2.55	-3.5 V
Anode current	I_a	10	10	10 mA
Screen-grid current	I_{g2}	2.5	2.6	2.8 mA
Mutual conductance	S	7.4	7.1	6.8 mA/V
Internal resistance	R_i	0.5	0.55	0.65 M Ω
Amplification factor between screen grid and control grid	μ_{g2g1}	50	50	50
Equivalent noise resistance	R_{eq}	1	1.1	1.2 k Ω
Input damping at 50 Mc/s (socket pins 1 and 3 interconnected)	r_{g1}	10	12	15 k Ω

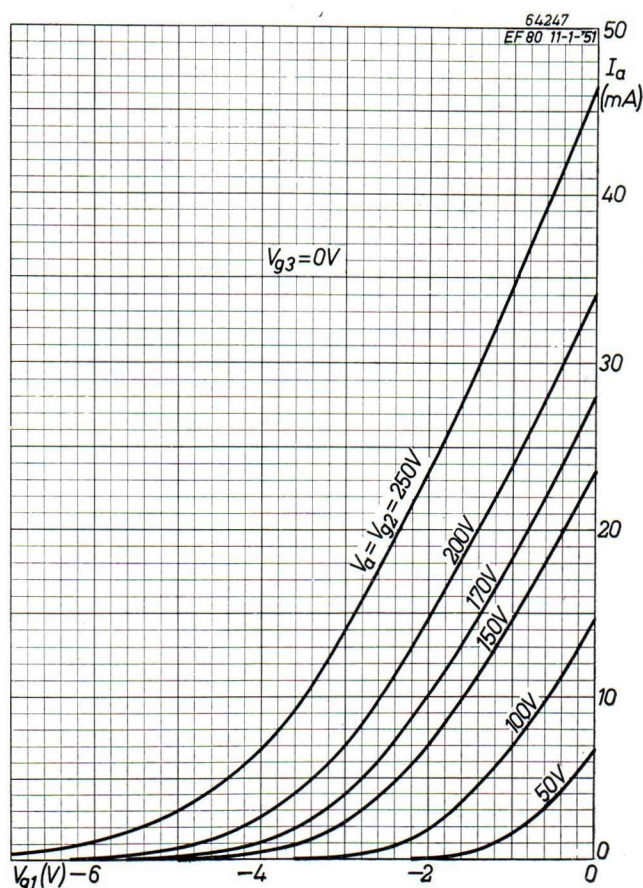


Fig. 24. Anode current plotted against control-grid voltage for anode and screen-grid voltages between 50 V and 250 V.

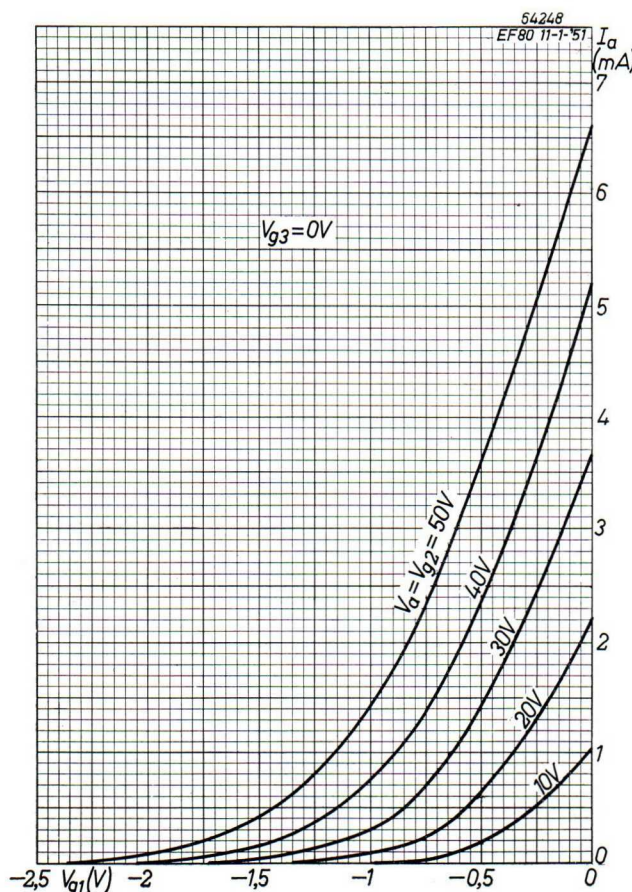


Fig. 25. Anode current plotted against control-grid voltage for anode and screen-grid voltages between 10 V and 50 V.

LIMITING VALUES

Anode voltage at zero anode current	V_{ao}	max.	550 V
Anode voltage	V_a	max.	250 V
Screen-grid voltage at zero screen-grid current	V_{g2o}	max.	550 V
Screen-grid voltage	V_{g2}	max.	250 V
Control-grid voltage (grid current +0.3 μ A)	V_{g1}	max.	-1.3 V
Heater voltage during warming-up period		max.	9.5 V
Voltage between heater and cathode	V_{kf}	max.	150 V
Cathode current	I_k	max.	15 mA
Anode dissipation	W_a	max.	2.5 W
Screen-grid dissipation	W_{g2}	max.	0.7 W
External resistance between control grid and cathode with self bias	R_{g1}	max.	1 M Ω
External resistance between control grid and cathode with fixed bias	R_{g1}	max.	0.5 M Ω
External resistance between heater and cathode	R_{kf}	max.	20 k Ω

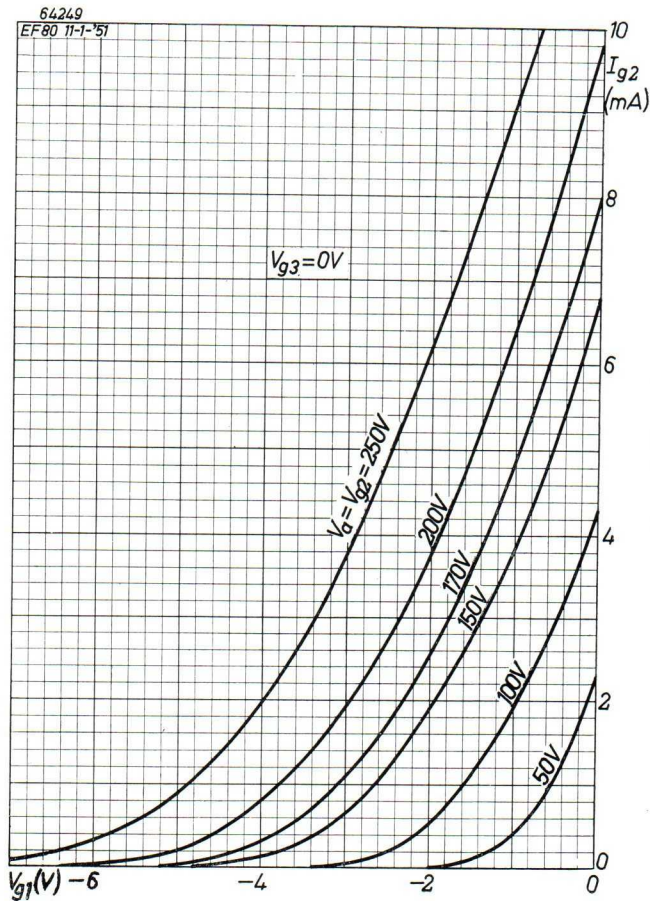
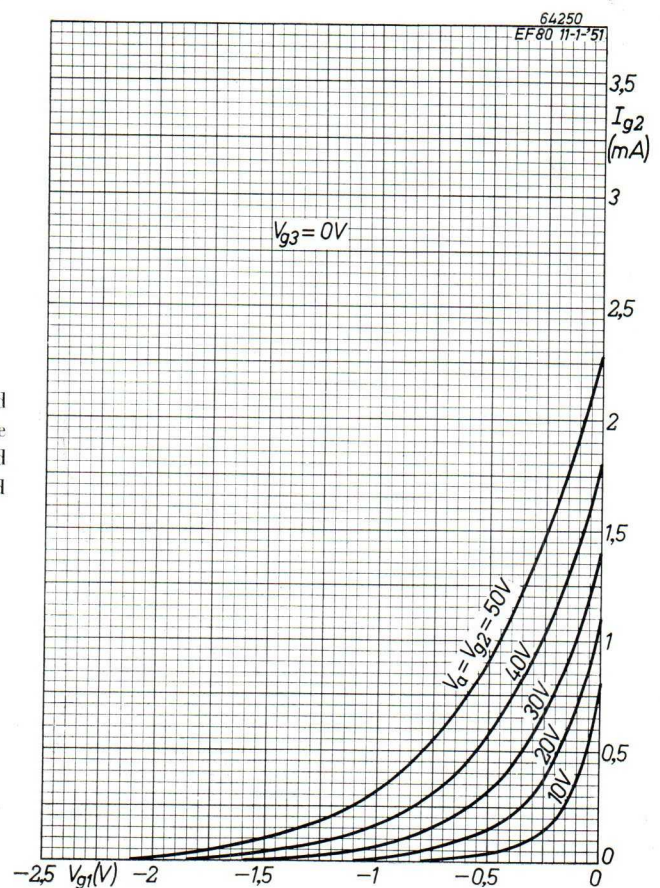


Fig. 26.
Screen-grid current plotted
against control-grid voltage
for anode and screen-grid
voltages between 50 V and
250 V.

Fig. 27.
Screen-grid current plotted
against control-grid voltage
for anode and screen-grid
voltages between 10 V and
50 V.



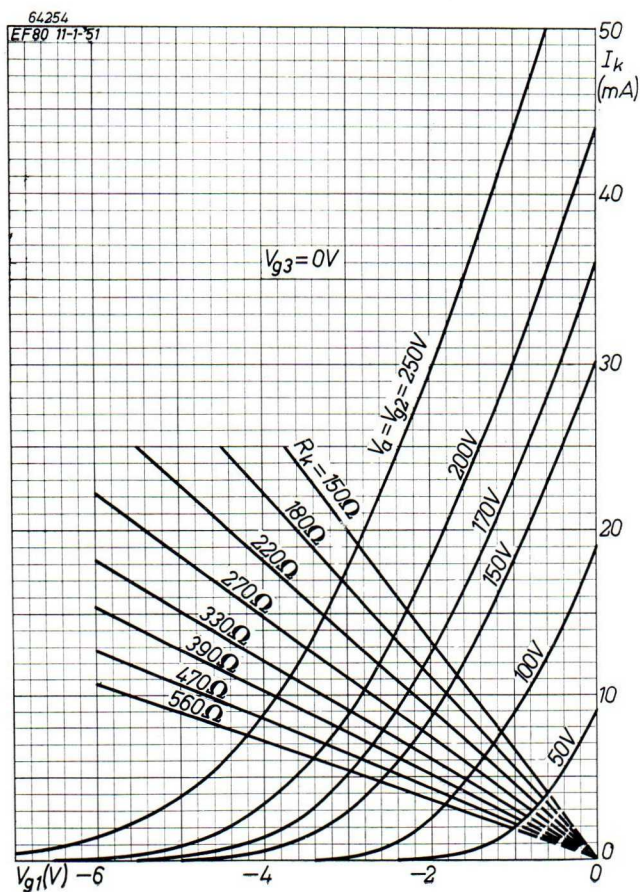
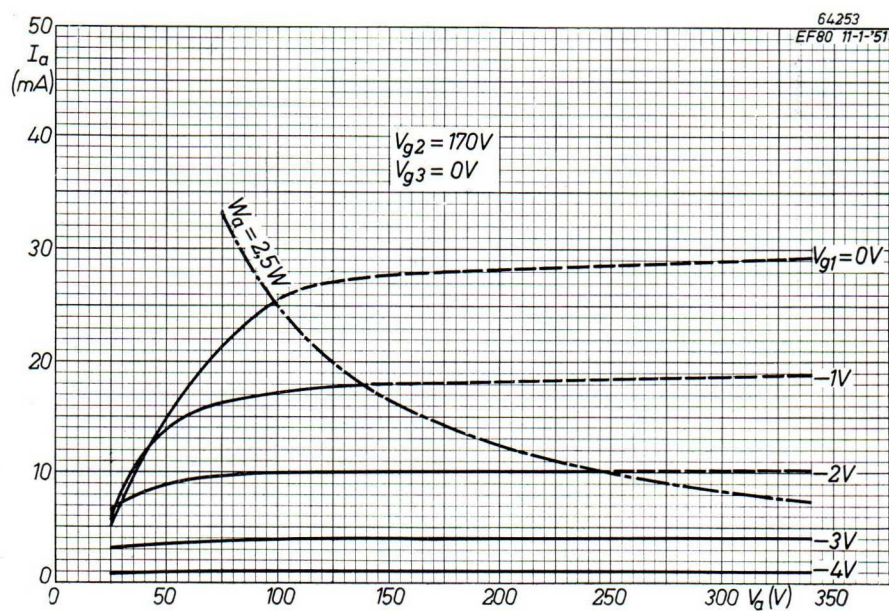


Fig. 28.
Cathode current plotted
against control-grid volt-
age, with anode and screen-
grid voltage as parameter.

Fig. 29.
Anode current plotted
against anode voltage for
screen-grid voltage of
170 V, with control-grid
voltage as parameter.



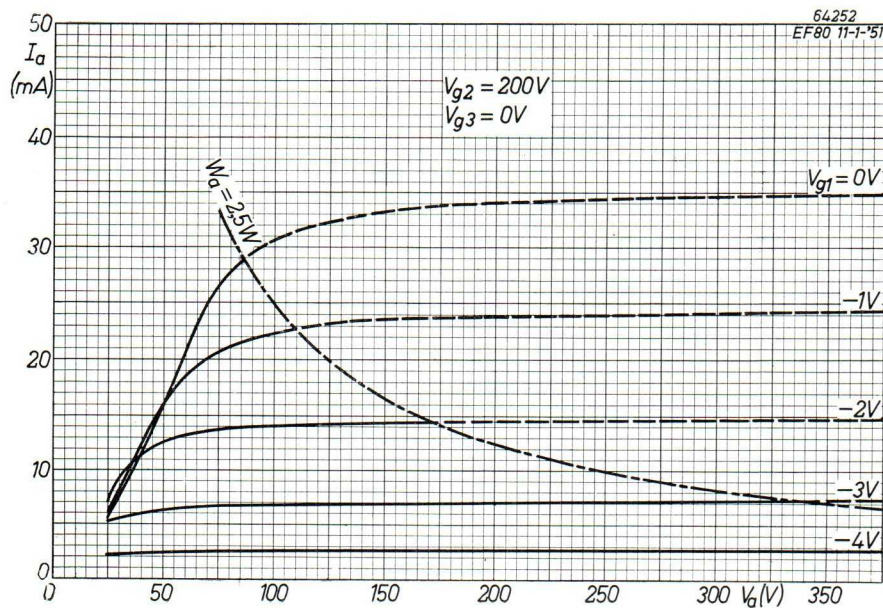
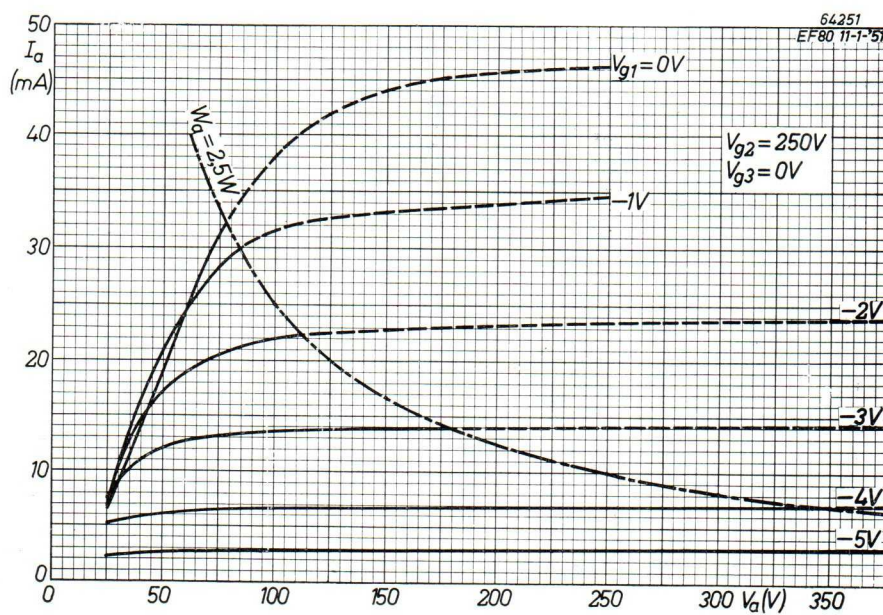


Fig. 30.
Anode current plotted against anode voltage for screen-grid voltage of 200 V, with control-grid voltage as parameter.

Fig. 31.
Anode current plotted against anode voltage for screen-grid voltage of 250 V, with control-grid voltage as parameter.



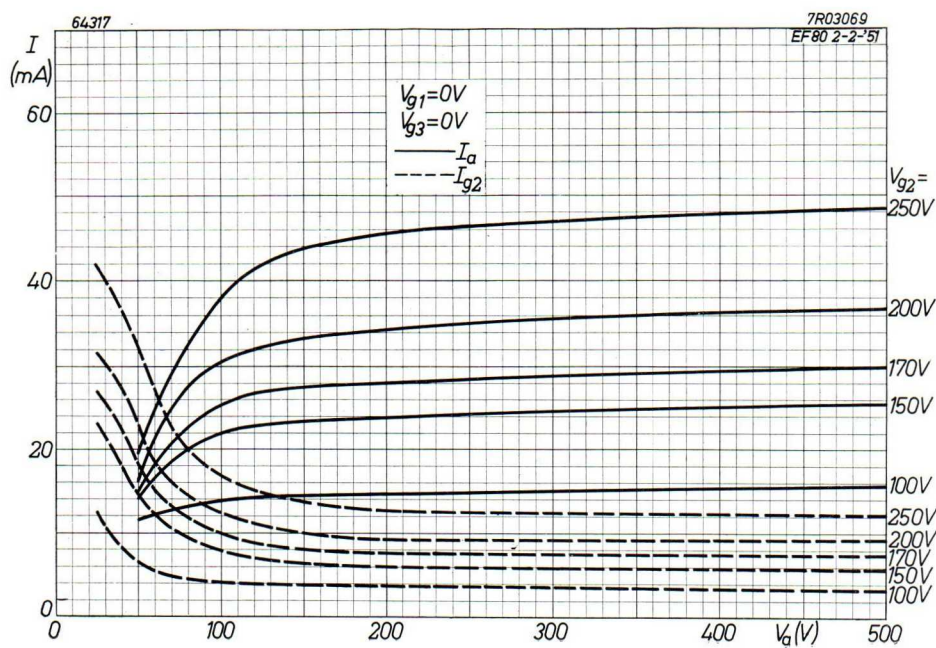


Fig. 32.
Anode current and screen-grid current plotted against anode voltage for control-grid voltage of 0 V, with screen-grid voltage as parameter.

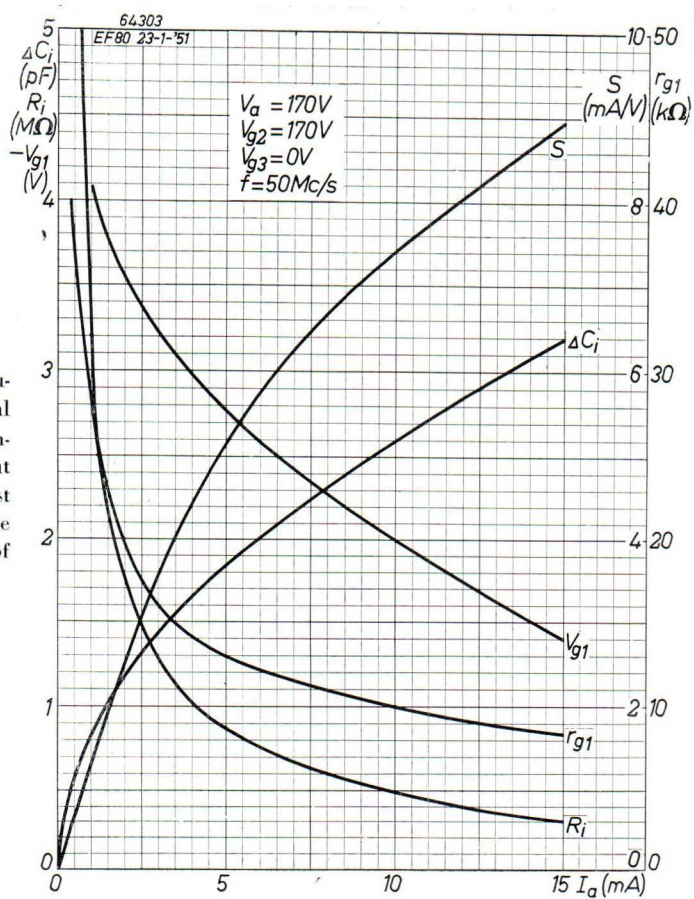


Fig. 33.
Control-grid voltage, mutual conductance, internal resistance, input capacitance variation and input damping plotted against anode current for anode and screen-grid voltage of 170 V.

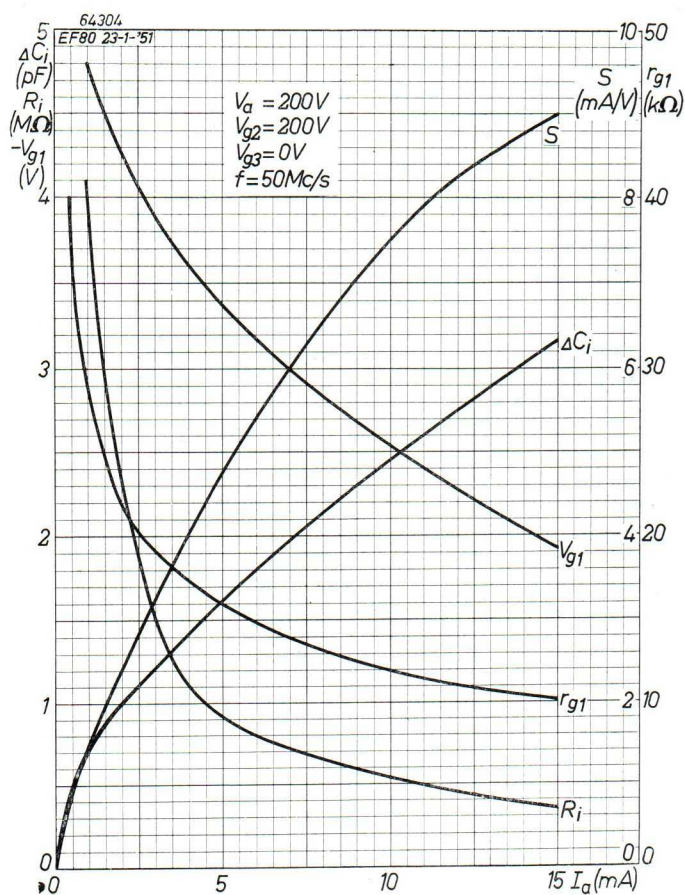


Fig. 34.
Control-grid voltage, mutual conductance, internal resistance, input capacitance variation and input damping plotted against anode current for anode and screen-grid voltage of 200 V.

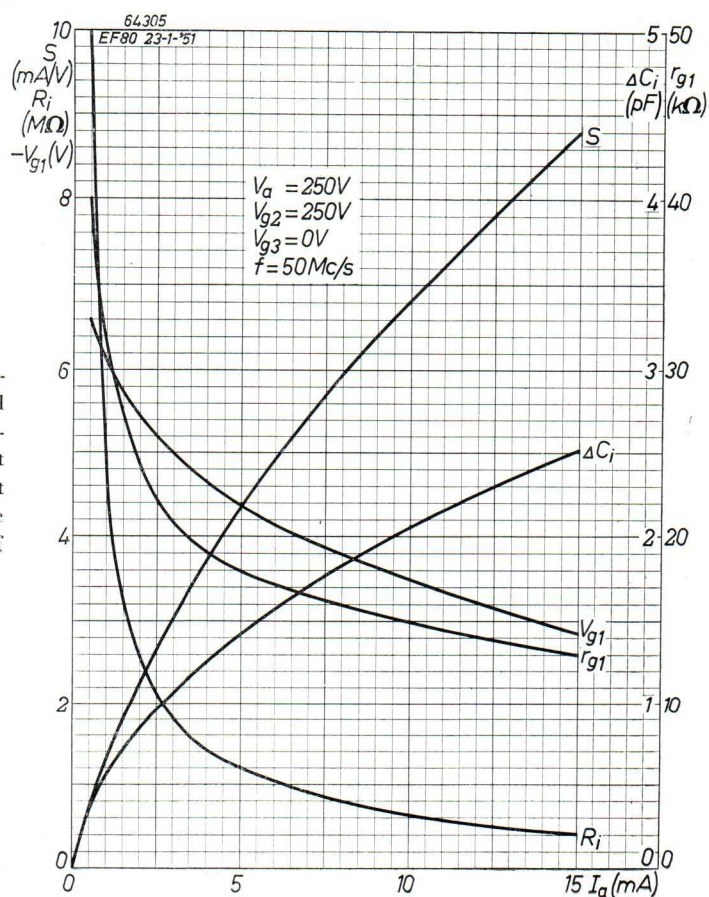


Fig. 35.
Control-grid voltage, mutual conductance, internal resistance, input capacitance variation and input damping plotted against anode current for anode and screen-grid voltage of 250 V.

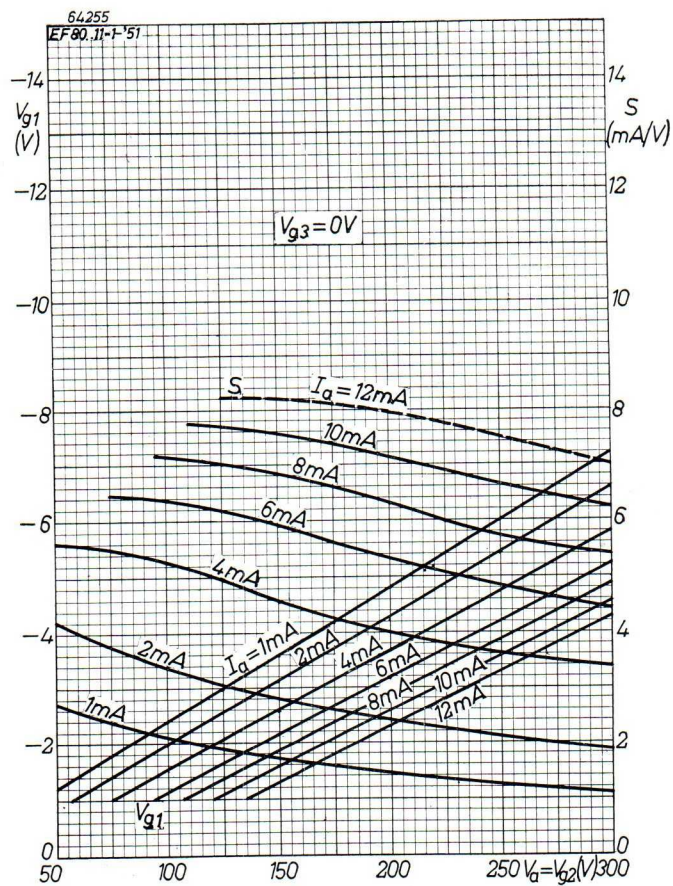


Fig. 36.

Relation between control-grid voltage and mutual conductance and anode and screen-grid voltage, with anode current as parameter.

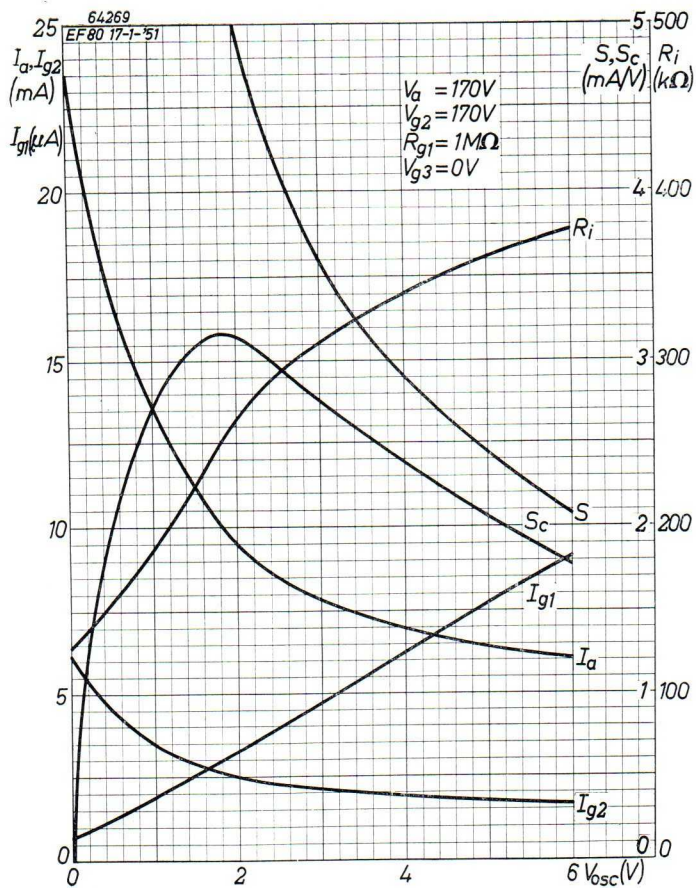


Fig. 37.

Performance curves of the EF 80 as frequency changer for anode and screen-grid voltage of 170 V.

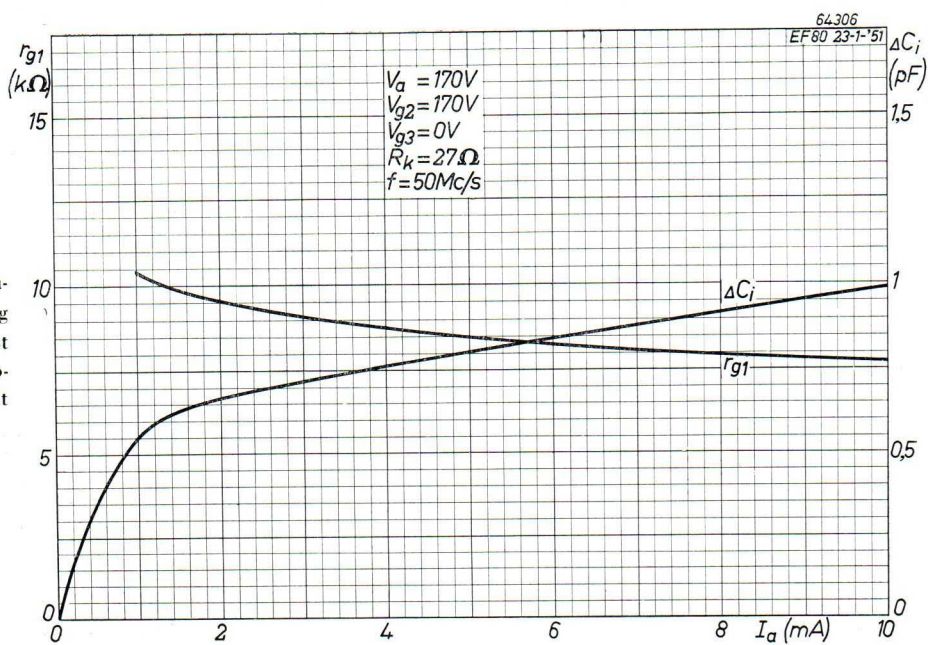


Fig. 38.

Input capacitance variation and input damping at 50 Mc/s plotted against anode current in a cathode-compensated circuit ($R_k = 27 \Omega$).

DOUBLE DIODE EB 91

DESCRIPTION

The EB 91 is a 7-pin miniature double diode. Each section has an independent cathode, and an electrostatic screen is provided between sections.

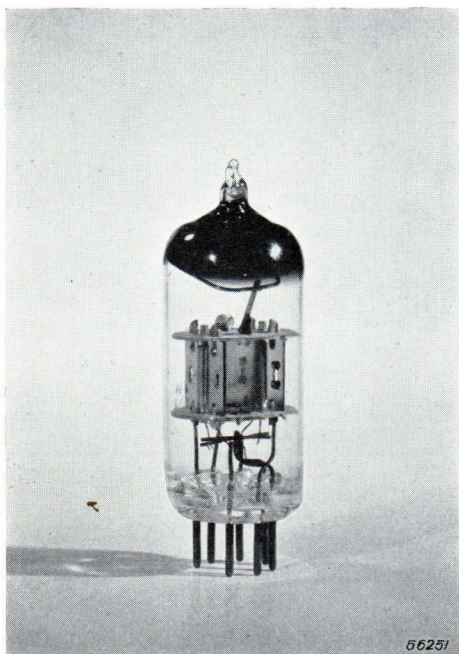


Fig. 39. The double diode EB 91 (actual size).

One diode section may be used as vision demodulator, and the other section for such functions as D.C. restoration. In the sound section of a receiver the EB 91 may also be used as an F.M. demodulator, for example in a ratio-detector circuit.

APPLICATION

The most important application of the EB 91 in a television receiver is in the vision demodulator stage. Only one diode section is required for this purpose and the other section can therefore be used as a D.C. restorer. The polarity of the diode used for demodulation depends upon a number of factors, such as the type of transmission (either positive or negative modulation may be used in the transmitter), the method of modulating the spot brightness of the picture tube (the video signal may be applied either to the cathode or to the grid of the picture tube) and the number of stages used for

video amplification. Cathode modulation of the picture tube is almost exclusively applied nowadays and, in the case of a direct-viewing tube, one stage of video amplification will usually be employed for systems operating with 405, 441, 525 and 625 lines.

In the 625-line system the modulation is negative, which means that the carrier amplitude is maximum during the sync. pulses. A signal of the polarity shown in fig. 40 is therefore required at the cathode of the picture tube, whilst the signal at the grid of the video amplifying tube should have a polarity as shown in fig. 41. It can therefore be seen that the polarity of the diode should be as illustrated in fig. 42, in which also the voltage across the load is represented. This signal would be directly suitable for the excitation of the video amplifying tube, provided the voltage at the dotted line (black level) in fig. 42 is equal to the required standing bias. This is not normally the case and since, moreover, in the absence of a signal, the video amplifying tube would operate without any bias, a separate bias voltage is applied either by the inclusion of a cathode resistor or by a fixed grid voltage. It is then necessary to apply D.C. restoration, so that the black level of the complete signal of fig. 42 coincides with the standing bias of the

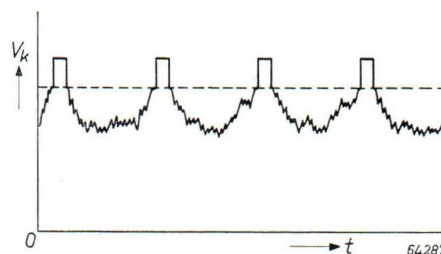


Fig. 40. Polarity of the signal required at the cathode of the picture tube.

video tube. For this purpose the remaining section of the EB 91 can be used.

Fig. 43 gives the complete circuit of the EB 91 used as vision demodulator (diode section II) and as D.C. restorer (diode section I). The potential of the whole circuit with respect to the chassis is $-V_g$, which is the fixed bias for the video amplifying tube. The bias line is normally decoupled by

an electrolytic capacitor of about $100\ \mu\text{F}$, whilst the internal resistance of the bias source is low, so that the potential of point *A* may be regarded as a fixed reference. It is therefore necessary to add to the signal appearing at $R_3 + L_1$ a voltage corresponding to the black level. This voltage is obtained by rectifying the I.F. signal with the diode section *I*. The operation of this diode needs some further explanation.

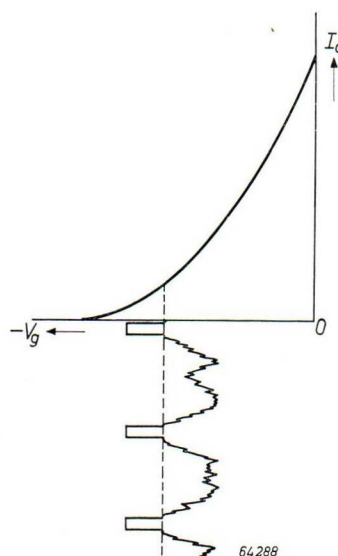


Fig. 41. Signal at the control grid of the single video stage.

It will be clear that the voltage supplied by the restorer must be free from modulation, whilst the load on the preceding I.F. circuit should be low so as to prevent undue damping. A load circuit with a high time constant and a high resistance should therefore be used. This implies that the detector efficiency of the restorer diode will be considerably greater than that of the detector diode, which has a

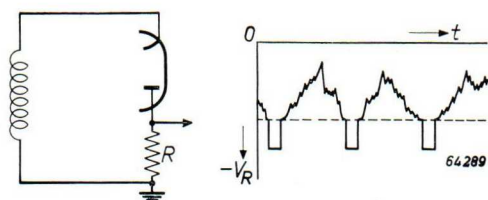


Fig. 42. Connection of the detector diode and the voltage across the load resistor.

load of only $3.9\ \text{k}\Omega$. For average picture contrast and with one stage of video amplification the I.F. signal required at the cathodes of the EB 91 is about $2\ V_{\text{rms}}$ during the sync. pulses. This I.F. signal will give rise to a voltage across the detector load R_3 of about $-1.4\ \text{V}$ also during the sync. pulses, which corresponds to a black level of $0.7 \times (-1.4) = -1\ \text{V}$. An equal voltage of op-

posite polarity must be delivered by the restorer diode. If R_1 is short-circuited a constant direct voltage of $2\ \text{V}$, corresponding to the peaks of the sync. pulses, arises across R_2 . The efficiency of the

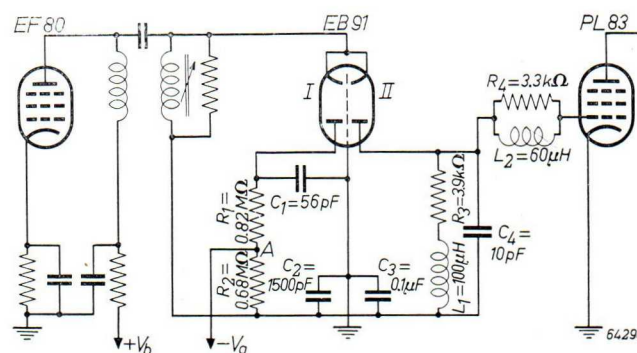


Fig. 43. Detector and D.C. restorer circuit with the double diode EB 91.

restorer circuit is now reduced with the aid of R_1 to such an extent that the voltage across R_2 becomes $1\ \text{V}$, as required.

It will be noticed that the time constants of R_2C_3 and R_1C_1 have been so chosen as to differ considerably. This has been done intentionally, so as to prevent variations in the restorer voltage across R_2 during short interference pulses, such as arise from car ignition and telephone dialling. With this circuit the voltage across R_2 does not change appreciably during a short interference pulse, the increase in rectified voltage occurring across R_1 , which together with C_1 has a low time constant. Therefore, at the end of the interference pulse the restorer circuit quickly resumes normal operation, whereas with a high time constant of the total load the increase in rectified voltage and the resultant increase in picture brightness would persist for some considerable time.

In the load of the diode section *II* (fig. 43) a shunt peaking coil L_1 of $100\ \mu\text{H}$ is included to compensate the effect that the shunt capacitor C_4 and the wiring capacitance have upon the frequency response. For the same purpose a series peaking coil L_2 of $60\ \mu\text{H}$ is connected in series with the input of the video amplifying tube. This coil resonates with the input and wiring capacitance of the PL 83 at a frequency of about $5\ \text{Mc/s}$, thus improving the response at the high end of the video frequency band whilst I.F. signals are attenuated. Damped oscillations modulating the picture brightness would, however, occur after each rapid change in signal level if this resonant circuit were not adequately damped. A damping resistor R_4 of $3.3\ \text{k}\Omega$ is therefore connected across L_2 .

TECHNICAL DATA

HEATER DATA

Heating: indirect by A.C. or D.C.;
series or parallel supply

Heater voltage	V_f	6.3 V
Heater current	I_f	0.3 A

LIMITING VALUES OF EACH SECTION FOR
THE USE AS RECTIFIER

Supply voltage	V_{tr}	max. 150 V _{rms}
Average anode current	I_d	max. 9 mA
Reservoir capacitor	C	max. 8 μ F
Internal resistance of supply	R_t	min. 300 Ω
Peak voltage between heater and cathode	V_{kfp}	max. 330 V ¹⁾

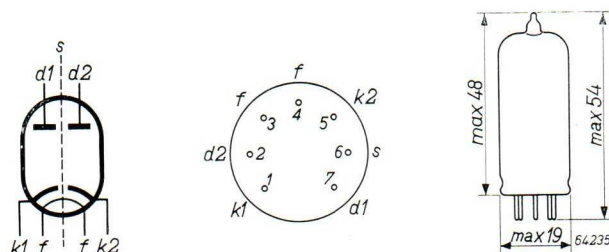
BASE CONNECTIONS AND
DIMENSIONS (in mm)

Fig. 44.

Mounting position: any

CAPACITANCES (with external screen)

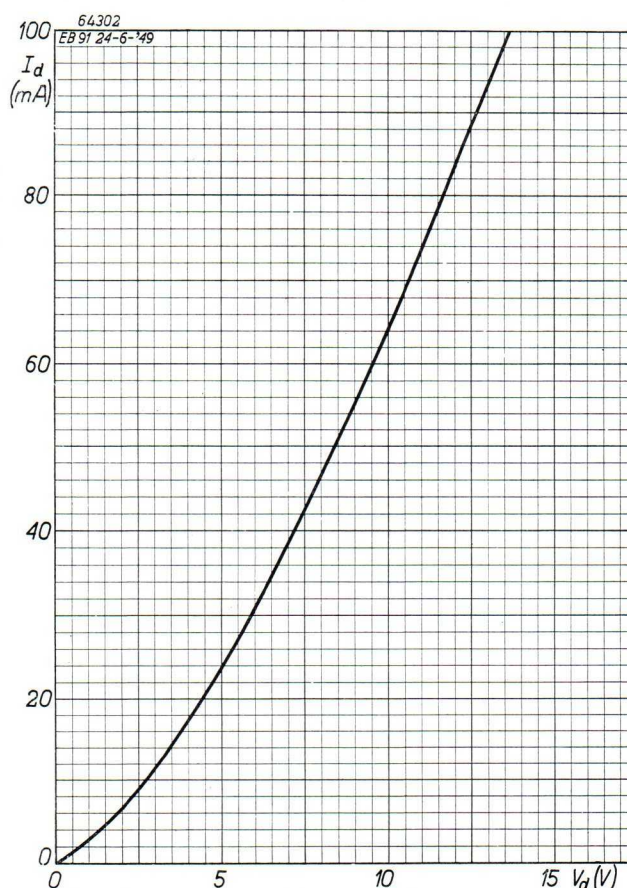
C_{d1}	3.0 pF
C_{d2}	3.0 pF
C_{d1-d2}	< 0.026 pF
C_{k1}	3.5 pF
C_{k2}	3.5 pF

LIMITING VALUES (each section)

Inverse anode voltage	$V_{d invp}$	max. 420 V
Average anode current	I_d	max. 9 mA
Peak anode current	I_{dp}	max. 54 mA
Voltage between heater and cathode	V_{kf}	max. 150 V
Peak voltage between heater and cathode	V_{kfp}	max. 330 V ¹⁾
External resistance between heater and cathode	R_{kf}	max. 20 k Ω ²⁾
Heater voltage during the warming-up period		max. 9.5 V

¹⁾ Cathode positive with respect to heater.

²⁾ If a high non-decoupled resistance between heater and cathode must be used it may be advisable, in order to avoid hum, to reduce the heater voltage to about 5 V.

Fig. 45. Anode current plotted against anode voltage
for each diode.

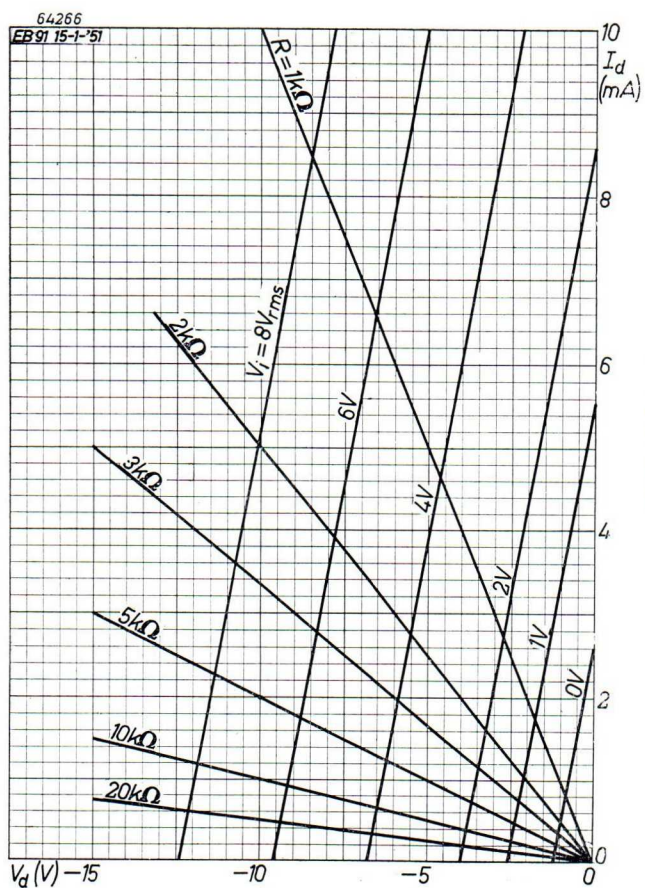


Fig. 46.
Output characteristics for each diode, for input voltages between zero and 8 V.

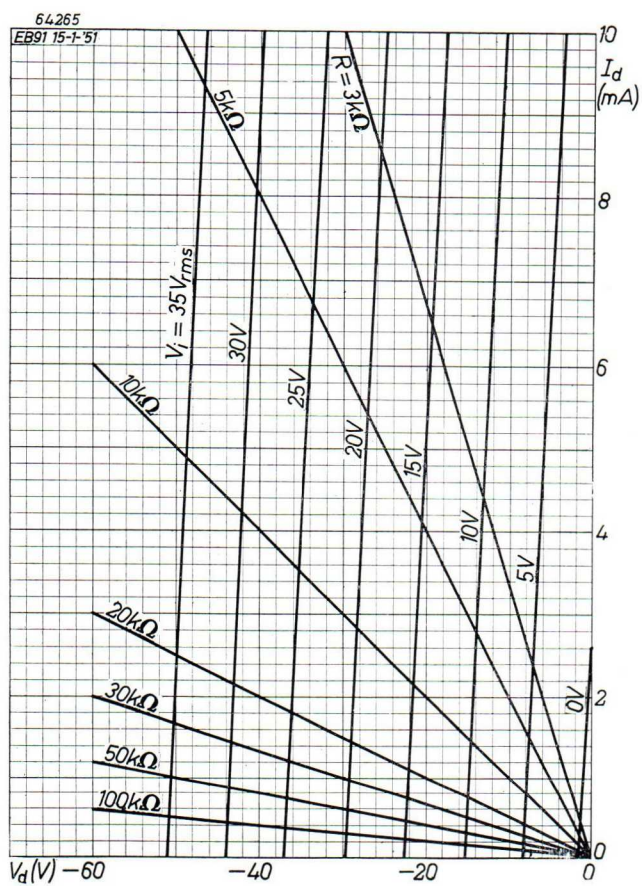
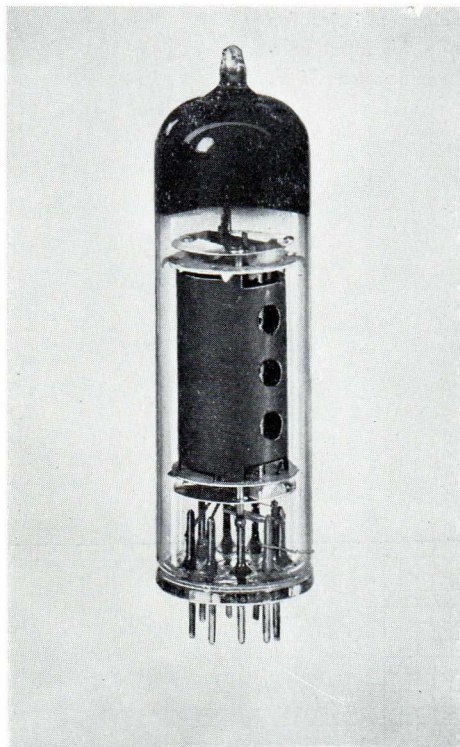


Fig. 47.
Output characteristics for each diode, for input voltages between zero and 35 V.

VIDEO OUTPUT PENTODE PL 83

DESCRIPTION

The PL 83 is a pentode on Noval base intended for use as the video output tube in television receivers. At a screen-grid voltage of only 170 V a considerable anode current swing can be obtained,



62148

Fig. 48. The video output pentode PL 83 (actual size).

so that the anode load resistance can be kept low and thus high-frequency compensation is simplified. The PL 83 can therefore be used in cases where the EF 80 would be inadequate.

Important features of the PL 83 are its high mutual conductance and small internal capacitances. Special attention has been paid to the mechanical design so as to produce a robust and rigid assembly, which minimizes the risk of microphony.

APPLICATION

1. THE PL 83 AS VIDEO AMPLIFIER WITH CATHODE COMPENSATION

A circuit of a video output stage employing the PL 83 with cathode compensation and a high-

tension line voltage of 180 V is represented in fig. 49. Compared with anode compensation by means of a shunt peaking coil, cathode compensation has the advantage that it permits of a simple and inexpensive circuit lay-out. Owing to the negative feedback, however, the gain of a cathode-compensated stage is usually lower.

In the circuit of fig. 49 it is assumed that the total capacitance in the anode circuit of the PL 83 is 35 pF. With a cathode resistor of 470 Ω and an anode load of 5 k Ω the capacitor shunting the cathode resistor should have a value of 390 pF, thus providing equal time constants in the anode and in the cathode circuit. A 3 db drop in frequency response occurs at about 5 Mc/s when the signal is applied directly to the control grid of the PL 83.

Under static conditions with $V_{g1} = -5.6$ V the anode current is 10.5 mA and the screen-grid current 1.4 mA. The gain of the stage is 7.5 and, if the black level of the input signal is at chassis potential, a peak video output voltage of 80 V, excluding sync, can be obtained at the anode. This signal is ample for driving any direct-viewing picture tube and also sufficient for the projection tube type MW 6-2.

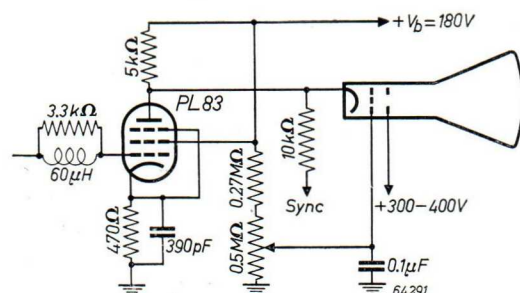


Fig. 49. The PL 83 as video amplifier with cathode compensation.

2. THE PL 83 AS VIDEO AMPLIFIER WITH ANODE COMPENSATION

Fig. 50 gives a circuit employing the PL 83 with anode compensation. The high-tension line voltage is 180 V and the circuit is suitable for driving any direct-viewing picture tube or a projection tube.

The series peaking coils in the control grid circuit

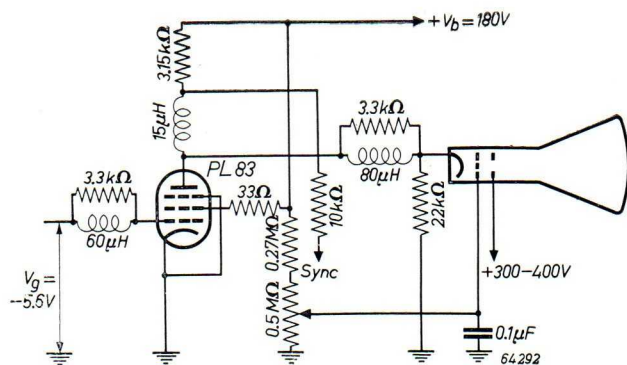


Fig. 50. The PL 83 as video amplifier with anode compensation. The stopper resistor in the screen-grid lead should not be chosen higher than 33 Ω .

TECHNICAL DATA

HEATER DATA

Heating: indirect by A.C. or D.C.
series supply

Heater voltage	V_f	15 V
Heater current	I_f	0.3 A

BASE CONNECTIONS AND DIMENSIONS (in mm)

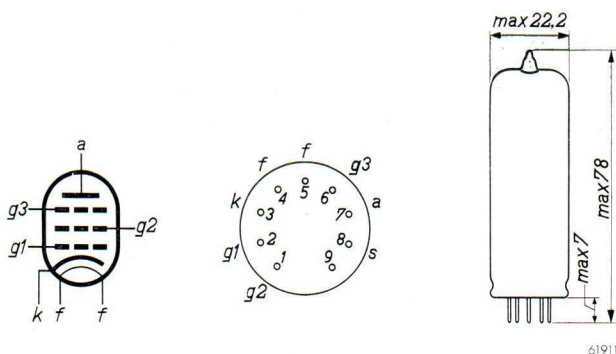


Fig. 51.

Mounting position: any

CAPACITANCES

C_{g1}	10.4 pF
C_a	6.6 pF
C_{ag1}	< 0.1 pF
C_{g1f}	< 0.15 pF

of the PL 83 and in the cathode of the picture tube are shunted by damping resistors so as to prevent overshoot.

The anode current under static conditions is 11 mA and the screen-grid current 1.5 mA when the grid bias is -5.6 V. If the black level of the input signal is at chassis potential a peak output voltage of about 100 V can be delivered, whilst the gain is 25. This gain is substantially higher than that of a stage with cathode compensation, but, on the other hand, anode compensation is more complicated.

TYPICAL CHARACTERISTICS

Anode voltage	V_a	170	200 V
Suppressor-grid voltage	V_{g3}	0	0 V
Screen-grid voltage	V_{g2}	170	200 V
Control-grid voltage	V_{g1}	-2.3	-3.5 V
Anode current	I_a	36	36 mA
Screen-grid current	I_{g2}	5	5 mA
Mutual conductance	S	10.5	10.5 mA/V
Internal resistance	R_i	0.1	0.1 M Ω
Amplification factor			
between screen grid and			
control grid	μ_{g2g1}	24	24

LIMITING VALUES

Anode voltage at zero anode			
current	V_{ao}	max.	550 V
Anode voltage	V_a	max.	250 V
Screen-grid voltage at zero			
screen-grid current	V_{g2o}	max.	550 V
Screen-grid voltage	V_{g2}	max.	250 V
Control-grid voltage			
(grid current $+0.3 \mu A$)	V_{g1}	max.	-1.3 V
Heater voltage during			
warming-up period		max.	22.5 V
Voltage between heater and			
cathode	V_{kf}	max.	150 V
Cathode current	I_k	max.	70 mA
Anode dissipation	W_a	max.	9 W
Screen-grid dissipation	W_{g2}	max.	2 W
External resistance between			
control grid and cathode	R_{g1}		
with fixed bias		max.	0.5 M Ω
with automatic bias		max.	1 M Ω
External resistance between			
heater and cathode	R_{kf}	max.	20 k Ω

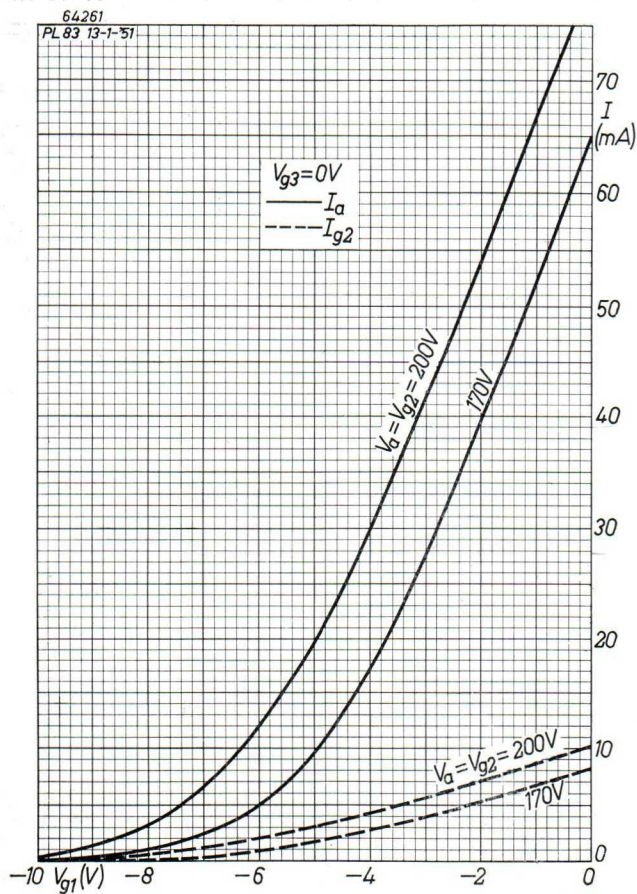
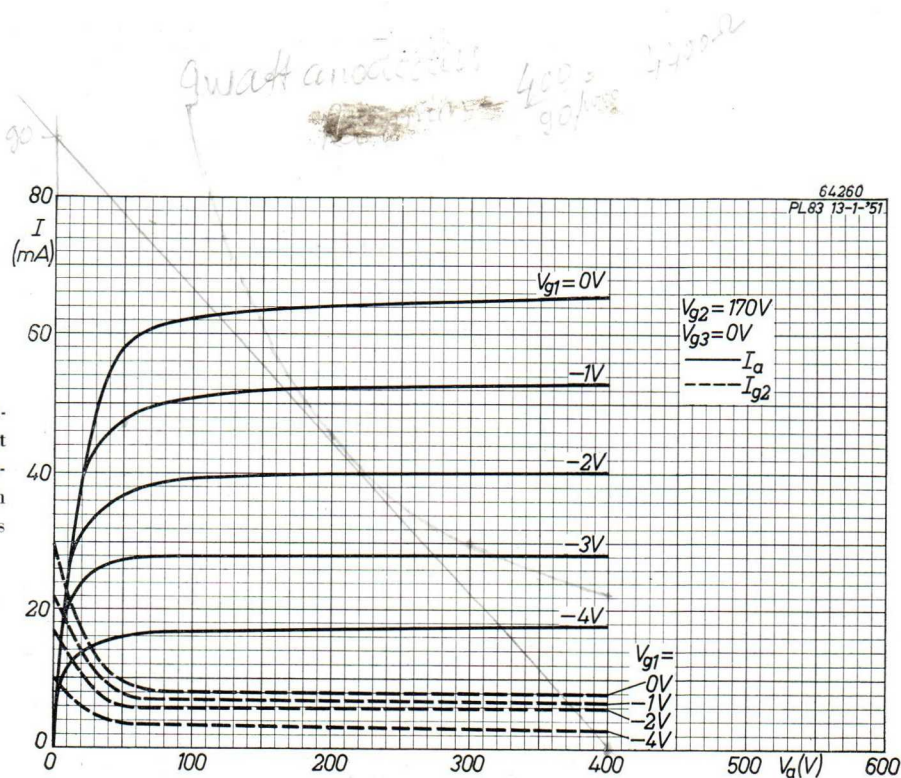


Fig. 52.
Anode current and screen-grid current plotted against control-grid voltage, with anode and screen-grid voltage as parameter.

Fig. 53.
Anode current and screen-grid current plotted against anode voltage for screen-grid voltage of 170 V, with control-grid voltage as parameter.



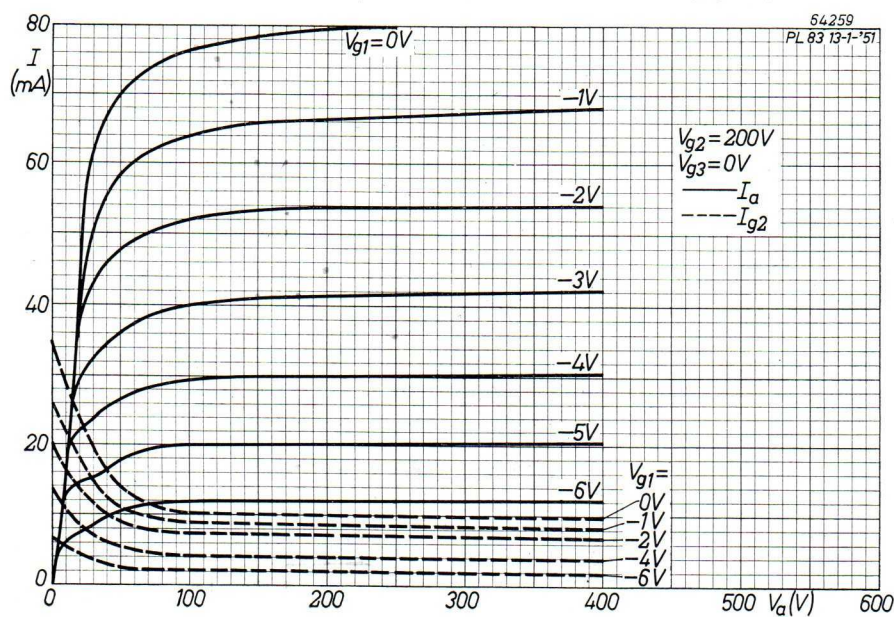
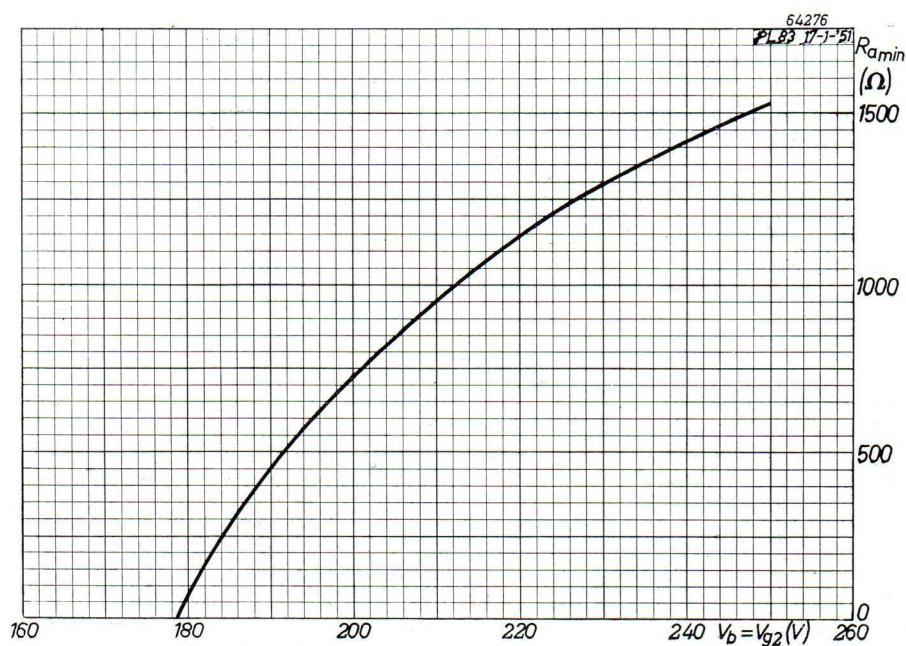


Fig. 54.
Anode current and screen-grid current plotted against anode voltage for screen-grid voltage of 200 V, with control-grid voltage as parameter.

Fig. 55.
Minimum permissible anode load resistance plotted against supply and screen-grid voltage.



OUTPUT PENTODE PL 82

DESCRIPTION

The PL 82 is an output pentode on Noval base intended for use as audio output or as frame output tube in a television receiver. Special attention has

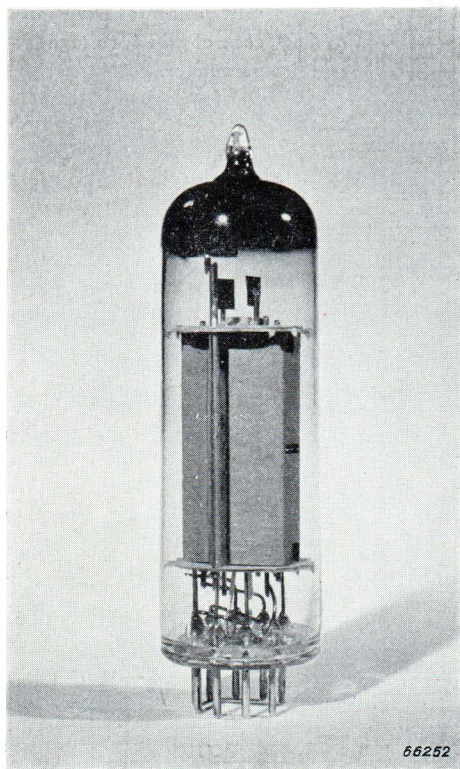


Fig. 56. The output pentode PL 82 (actual size).

been paid to the mechanical design so as to produce a robust and rigid assembly, which minimizes the risk of microphony.

There are a number of cases in which the audio output of the pentode section of an ECL 80 will be considered insufficient. The PL 82, giving an output of 4 W at 10% total distortion and at anode and screen-grid voltages of 170 V, may then be used.

For the same reason, in the frame output stage a PL 82 may be used instead of the pentode section of an ECL 80. The available peak anode current of the PL 82 is three to four times as large as that of the ECL 80, so that the former tube is particularly suitable for use in receivers employing a picture tube with wide-angle deflection.

APPLICATION

1. THE PL 82 AS SOUND OUTPUT TUBE¹⁾

As already indicated above, at a line voltage of 170 V the PL 82 is capable of giving an output of 4 W at 10% total distortion and with a matching resistance of 3 k Ω . The driving voltage required for this output is 6 V_{rms} and the bias 10.4 V. Either fixed or automatic bias may be applied. Automatic bias has the advantage that a higher value of control-grid resistor is permitted, which may increase the gain of the preceding stage. A disadvantage connected with automatic bias, however, is that, due to the voltage drop in the cathode resistor, the voltage between anode and cathode is somewhat reduced, and this also reduces the available output.

At a line voltage of 200 V the output of the PL 82 is the same as that at 170 V, although the operating conditions are somewhat different. These conditions are to be found in the tube data.

2. THE PL 82 AS FRAME OUTPUT TUBE

The PL 82 can deliver a large peak anode current at comparatively low anode and screen-grid voltages. In order to avoid incursion in the knee of the I_a/V_a characteristic, the frame output stage is normally so designed, that the minimum anode voltage, which occurs at the end of the scan, does not become lower than 50 V. From the tube characteristics it may be seen that with a screen-grid voltage of 170 V and an anode voltage of 50 V the anode current of the average tube can greatly exceed 100 mA. In order to allow for inevitable spread between individual tubes and for deterioration during life, the circuit should, however, be designed around a peak anode current not exceeding 90 mA at $V_a = 50$ V and $V_{g2} = 170$ V.

In the design of a frame output stage a choice must be made between an output transformer of small dimensions, which leads to a large peak anode current, and a larger output transformer, with which the peak anode current will be lower. Since

¹⁾ See P. D. van der Knaap and J. Jager, Matching of the Frame Output Stage to the Deflection Coils in Television Receivers, Electr. Appl. Bull. 11, p. 21, 1950 (No. 2).

there are also other important factors to be taken into account, the choice of the output transformer will be dealt with at some length.

The deflection coils used for vertical deflection have resistance and self-inductance, but the rate of change of the deflection current is so low that the voltage occurring across the self-inductance is negligible compared with that across the resistance. It may therefore be assumed that the matching transformer is loaded with a pure resistance, R_d ; see fig. 57.

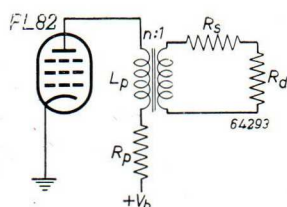


Fig. 57. Basic diagram of a frame output stage.

When it is now assumed that the operating conditions are such that a linear saw-tooth current having a peak-to-peak value I_d flows in the deflection coils, the saw-tooth current in the primary will be I_d/n , whilst a linear saw-tooth voltage occurs across the primary self-inductance L_p . Due to this saw-tooth voltage a current having a parabolic shape is added to the saw-tooth current in the primary. The relative magnitude of this parabolic component depends upon the ratio between the primary self-inductance and the load resistance transformed to the primary, and also upon the fundamental frequency of the saw-tooth current. It will be understood that a relatively low primary self-inductance, as with a transformer of small dimensions, causes a relatively large parabolic component, which considerably influences the shape of the total anode current. In the other extreme case of a very high primary self-inductance, involving a transformer of large dimensions, the parabolic component becomes very small and the total anode current has a linear shape.

The shape of the total anode current required to give a linear saw-tooth current in the deflection coils is depicted in fig. 58 for different values of T_1/T , in which $T_1 = L_p/n^2 (R_d + R_s)$ and T the duration of the frame scan.

It appears from this figure that with $T_1/T = 0.1$ the parabolic component assumes a very high value, resulting in a considerable distortion of the total anode current and a high value of the required peak

anode current. Another important disadvantage is that the minimum in anode current occurs quite a considerable time after the commencement of the scan. In fig. 58 it is assumed that the minimum value of the anode current is just zero, which is, of course, most economical in respect to current consumption. At minimum anode current the tube characteristic shows a strong curvature and it is consequently difficult to distort the driving voltage in such a way that the required shape of anode current is obtained. Non-linearity of the frame scan is therefore apt to occur at about 0.4 of the total picture height, and this is very annoying. A higher value of T_1/T should therefore be employed and this involves a higher value for the primary self-inductance

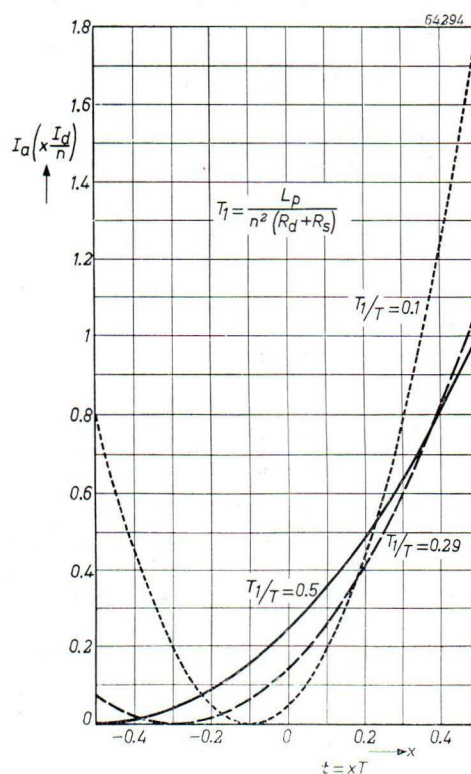


Fig. 58. Shape of the total anode current required to give a linear deflection current, for different values of T_1/T .

With $T_1/T = 0.29$ the shape of the total anode current is much more favourable but the minimum in anode current still occurs at 0.2 of the total picture height. It can be shown that with this value of T_1/T the average anode current is a minimum.

Finally, with $T_1/T = 0.5$ the minimum in anode current coincides with the commencement of the scan, which consequently is the most favourable situation. Moreover, it appears that the shape of the

anode current closely resembles that of the I_a/V_g characteristic of the tube, so that the problem of obtaining a driving voltage of the required shape is considerably simplified. With this value of T_1/T the required peak anode current reaches the minimum. A further increase of T_1/T , i.e. a higher value for L_p , does not decrease the required peak anode current and, although the total anode current assumes a more linear shape, the required average anode current becomes higher.

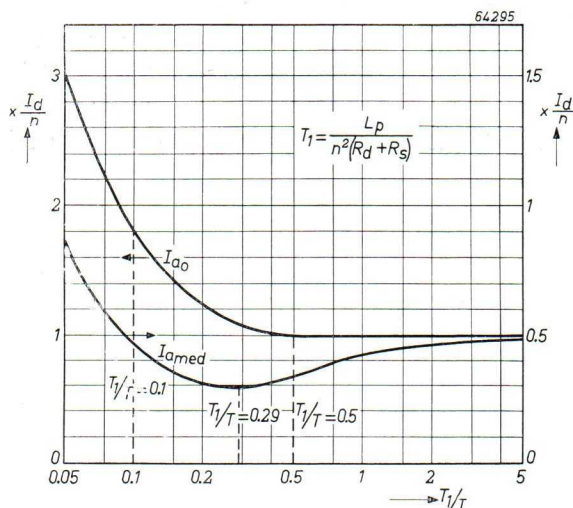


Fig. 59. Peak anode current and average anode current plotted against T_1/T .

The two important characteristics of the anode current, viz. the average anode current $I_{a \text{ med}}$ and the peak anode current I_{a0} , are plotted in fig. 59 against the values of T_1/T . It may be seen that with $T_1/T = 0.29$ the average anode current is a minimum and equal to $0.29 I_a/n$. At $T_1/T = 0.5$ the peak anode current has reached the minimum, which is I_a/n . In this case the average current is only slightly higher than that with $T_1/T = 0.29$.

By way of example an output transformer for use in conjunction with the PL 82 will now be calculated. The influence of the stray self-inductance will be disregarded, and the following practical figures have been chosen:

Supply voltage	$V_b = 170 \text{ V}$
Resistance of the deflection coils	$R_d = 50 \Omega$
Self-inductance of the deflection coils	$L_d = 50 \text{ mH}$
Deflection current peak-to-peak	$I_d = 350 \text{ mA}$
Frame frequency	$1/T = 50 \text{ c/s}$

These data refer to double-D scanning of a picture tube MW 31-16 or MW 22-16 operating with an E.H.T. of 9 kV

In view of the maximum permissible drop of the anode voltage below the H.T. line the turns ratio of the transformer must first be calculated. Assuming that the secondary has a resistance R_s of 15Ω , the drop occurring at the end of the scan across the primary self-inductance is: $\frac{1}{2} I_{dn} (R_d + R_s) = 11.4 n$. In order to avoid incursion in the knee of the I_a/V_a characteristic, and to make allowance for some additional voltage drop caused by the primary resistance and the self-inductance of the deflection coils, the above-mentioned drop must not be chosen too high, a suitable value in this case being 80 V. The turns ratio will therefore be $80/11.4 = 7$.

It must now be checked whether the additional drop, caused by the resistance of the primary and the self-inductance of the deflection coils, does not bring the momentary anode voltage below 50 V. In this example a drop of 40 V is still permissible. If the primary self-inductance is so chosen that $T_1/T = 0.5$ (see figs 58 and 59), the peak anode current occurring at the end of the scan becomes: $I_{a0} = I_d/n = 50 \text{ mA}$. If it is assumed that the primary resistance will be 500Ω , the maximum drop due to this resistance will be 25 V. Owing to the self-inductance of the deflection coils a constant voltage drop occurs during the scan across the primary. When the saw-tooth current is linear this amounts to $n \cdot L_d \cdot I_d/T$. Disregarding the duration of the fly-back, for a system operating with 50 frames per second the duration of the scan $T = 0.02 \text{ sec}$. This gives for the drop due to the self-inductance of the deflection coils, $7 \times 0.05 \times 0.350/0.02 = 6.1 \text{ V}$.

The total drop below the H.T. line now is $80 + 25 + 6 = 111 \text{ V}$, so that the minimum anode voltage at the end of the scan becomes 59 V, which is a sufficiently high value. In fact the anode voltage may be allowed to become as low as 50 V, but it should be realized that in determining $I_{a \text{ med}}$ and I_{a0} the minimum momentary anode current has been assumed to be zero. This condition can hardly be fulfilled in normal practice and the minimum value of the anode current will actually be 1 or 2 mA. This will result in a slightly higher drop in the resistance of the primary winding. As a final check the total voltage drop should again be calculated when the actual resistances of the primary and the secondary are known.

Finally, the required primary self-inductance must be calculated. It has been assumed that

$T_1/T = 0.5$, so that with $T = 0.02$ the value of T_1 becomes $T_1 = 0.01 = L_p/n^2 (R_d + R_s)$, and $L_p = 7^2 \times (50 + 15) \times 0.01 = 32 \text{ H}$.

Summarizing, the transformer data are as follows:

Primary self-inductance	32 H
Turns ratio	7 : 1
Resistance of the primary	500 Ω
Resistance of the secondary	15 Ω
Direct current in the primary	(see fig. 59) 16.5 + 2 mA

From the tube characteristics it may be seen that when the screen-grid voltage is 170 V a bias of about 15 V is required. If this bias is obtained automatically by means of a cathode resistor the

available supply voltage will be reduced by an amount equal to the cathode voltage. This might well bring the momentary anode voltage below the tolerable level but, on the other hand, in most television receivers operating with a transformerless supply system from the 220 V mains the available H.T. line voltage is somewhat higher than 170 V, for example 185 V. With fixed bias there is, of course, no reduction of the available high tension.

The driving voltage for the tube must have a shape that gives the required form of anode current. According to the shape of the available saw-tooth voltage and the figure of T_1/T , provision has to be made for obtaining the required shape of driving voltage. The means employed for this are given in the application notes of the ECL 80.

TECHNICAL DATA

HEATER DATA

Heating: indirect by A.C. or D.C.;
series supply

Heater voltage	V_f	16.5 V
Heater current	I_f	0.3 A

BASE CONNECTIONS AND

DIMENSIONS (in mm)

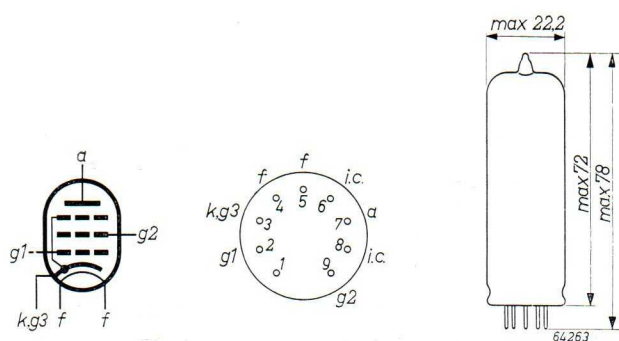


Fig. 60.

Mounting position: any

CAPACITANCES

C_{g1}	11 pF
C_a	8.3 pF
C_{ag1}	< 1 pF
C_{g1f}	< 0.15 pF

TYPICAL CHARACTERISTICS

Supply voltage	V_b	170	200 V
Anode voltage	V_a	170	200 V
Screen-grid voltage	V_{g2}	170	V
Screen-grid resistor	R_{g2}	—	680 Ω
Control-grid voltage	V_{g1}	—10.4	—13.9 V
Anode current	I_a	53	45 mA
Screen-grid current	I_{g2}	10	8.5 mA
Mutual conductance	S	9.5	8 mA/V
Internal resistance	R_i	20	24 k Ω
Amplification factor			
between screen grid and			
control grid	μ_{g2g1}	10	10

OPERATING CONDITIONS FOR SOUND

OUTPUT (one tube in class A)

Supply voltage	V_b	170	200 V
Anode voltage	V_a	170	200 V
Screen-grid voltage	V_{g2}	170	V
Screen-grid resistor	R_{g2}	—	680 Ω
Control-grid voltage	V_{g1}	—10.4	—13.9 V
Anode current	I_a	53	45 mA
Screen-grid current	I_{g2}	10	8.5 mA
Mutual conductance	S	9.5	8 mA/V
Internal resistance	R_i	20	24 k Ω
Matching resistance	R_a	3	4 k Ω
Power output			
($d_{tot} = 10\%$)	W_o	4	4 W
Input signal			
($d_{tot} = 10\%$)	V_i	6	7 V _{rms}
Input signal			
($W_o = 50 \text{ mW}$)	V_i	0.5	0.55 V _{rms}

OPERATING CONDITIONS FOR SOUND OUTPUT (two tubes in push-pull)

Anode voltage	V_a	170	200	V
Screen-grid				
voltage	V_{g2}	170	200	V
Common cathode				
resistor	R_k	100	135	Ω
Matching resist-				
ance between				
anodes	$R_{aa'}$	4	4	k Ω
Input signal	V_i	0	9.3	0
Anode current	I_a	2x46	2x50	2x45
Screen-grid				
current	I_{g2}	2x8.7	2x17	2x8.5
Power output	W_o		9	12 W
Total distortion	d_{tot}		5	5 %

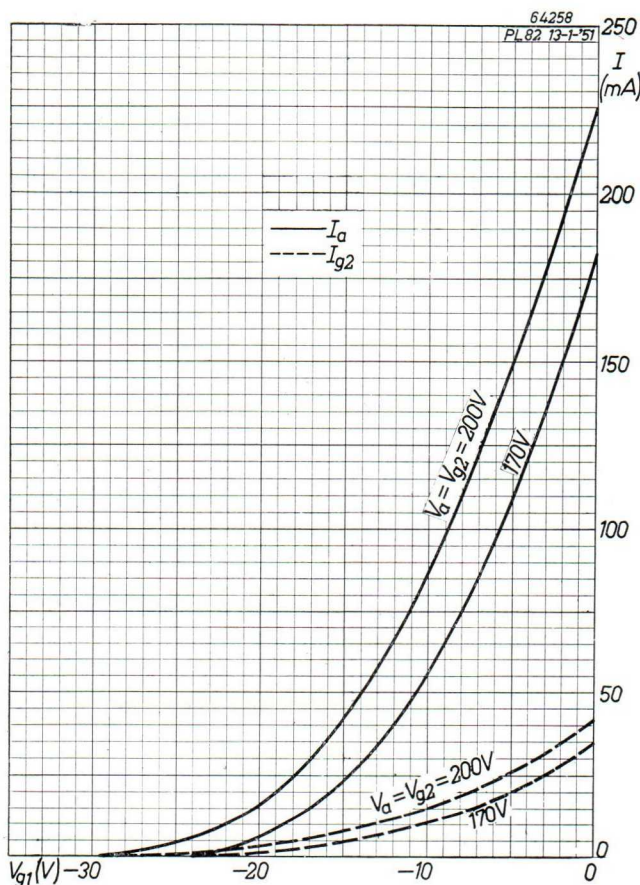


Fig. 61. Anode current and screen-grid current plotted against control-grid voltage with anode and screen-grid voltage as parameter.

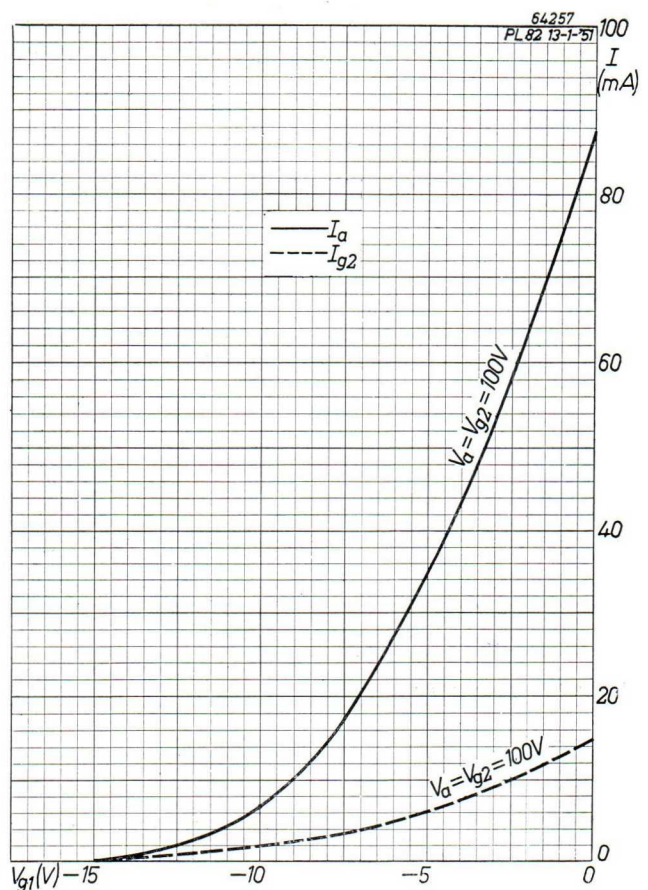


Fig. 62. Anode current and screen-grid current plotted against control-grid voltage for anode and screen-grid voltage of 100 V.

LIMITING VALUES

Anode voltage at zero anode

current V_{a0} max. 550 V

Anode voltage

V_a max. 250 V

Positive peak anode voltage

(10% of the duration of a cycle with a maximum of 2 m sec.)

max. 2.5 kV

Negative peak anode voltage

max. 500 V

Screen-grid voltage at zero

screen-grid current

V_{g20} max. 550 V

Screen-grid voltage

V_{g2} max. 250 V

Control-grid voltage

(grid current +0.3 μ A)

V_{g1} max. -1.3 V

Heater voltage during

warming-up period

max. 24.5 V

Voltage between heater and

cathode

V_{kf} max. 200 V

Cathode current

I_k max. 75 mA

Anode dissipation

W_a max. 9 W

Screen-grid dissipation

W_{g2} max. 2.5 W

External resistance between
control grid and cathode

with fixed bias R_{g1} max. 0.4 M Ω

with automatic bias R_{g1} max. 1 M Ω

External resistance between

heater and cathode R_{kf} max. 20 k Ω

To allow for inevitable spread between individual tubes and for deterioration during life in frame output applications, the circuit should be designed around a peak anode current not exceeding 90 mA at $V_a = 50$ V and $V_{g2} = 170$ V.

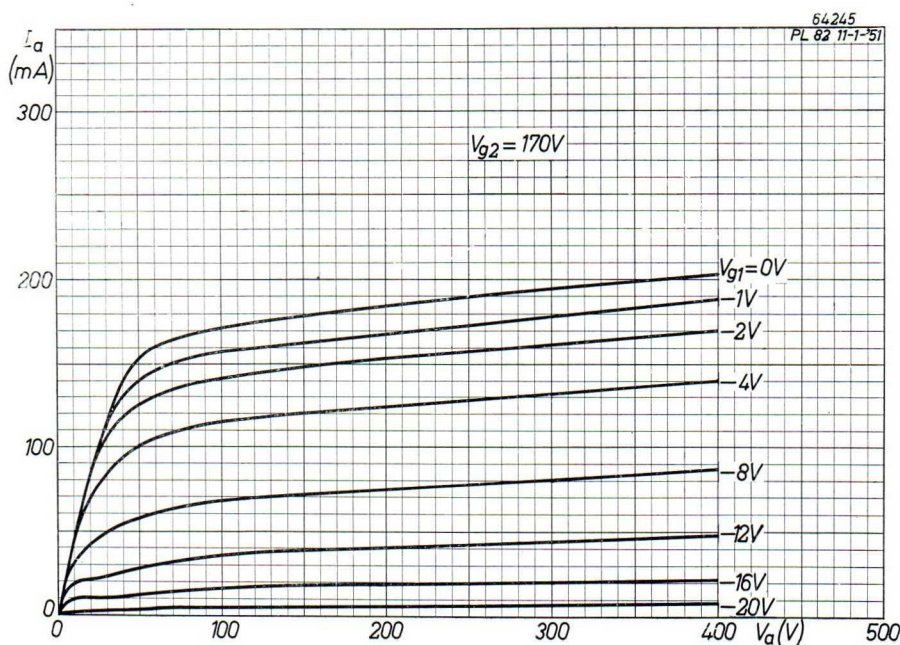


Fig. 63.
Anode current plotted
against anode voltage for
screen-grid voltage of
170 V, with control-grid
voltage as parameter.

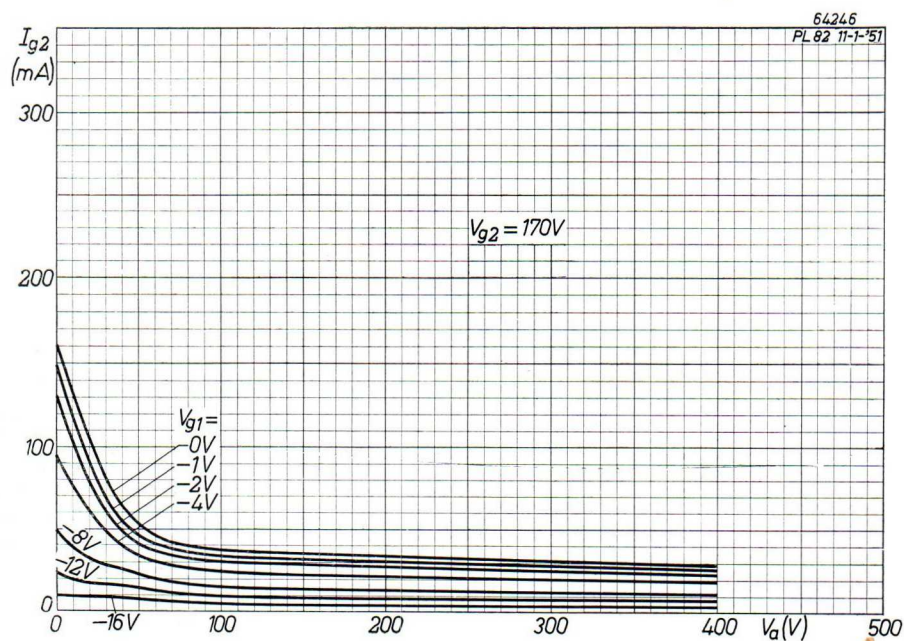


Fig. 64.
Screen-grid current plotted
against anode voltage
for screen-grid voltage of
170 V, with control-grid
voltage as parameter.

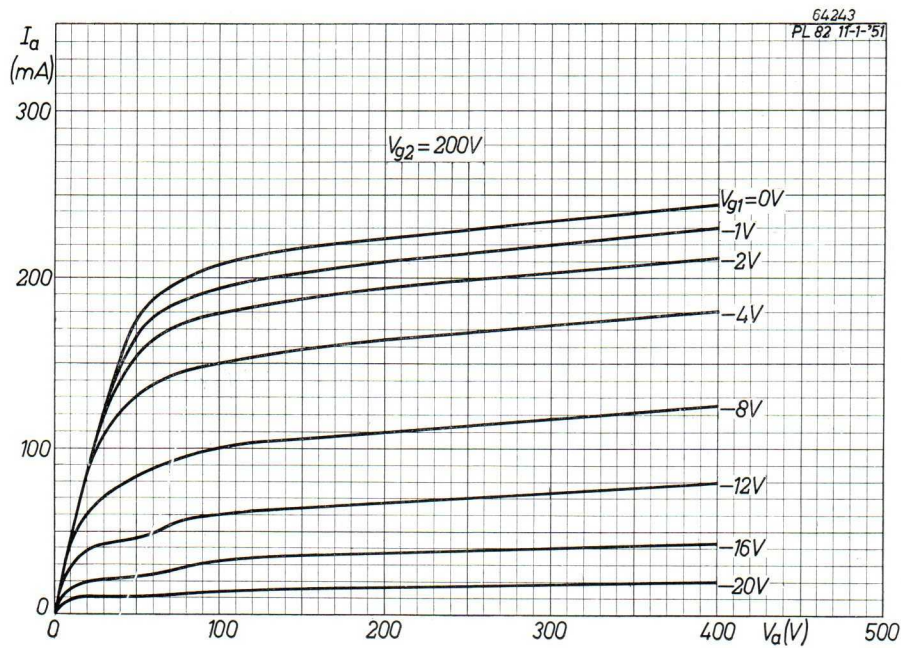
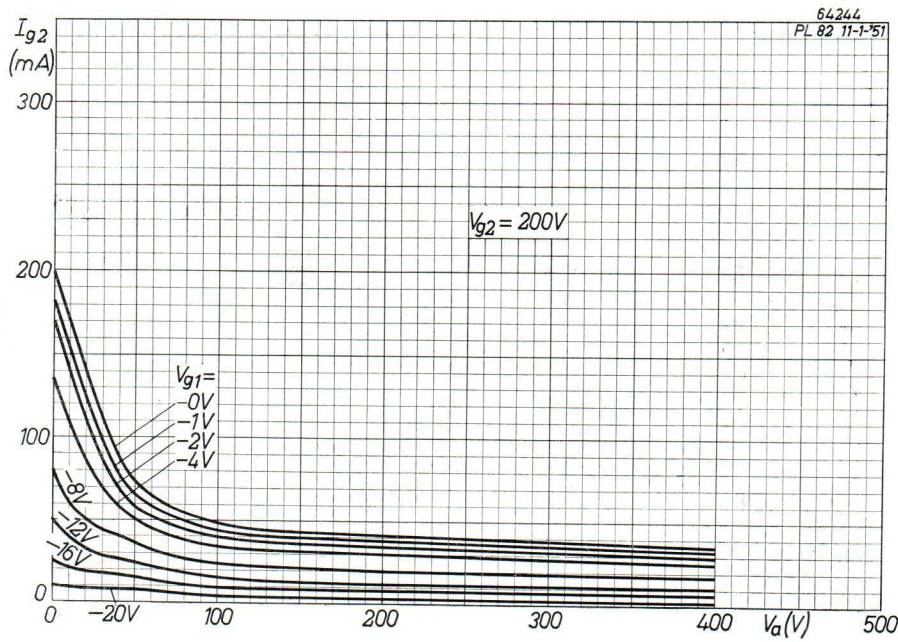


Fig. 65.
Anode current plotted
against anode voltage for
screen-grid voltage of
200 V, with control-grid
voltage as parameter.

Fig. 66.
Screen-grid current plotted
against anode voltage
for screen-grid voltage of
200 V, with control-grid
voltage as parameter.



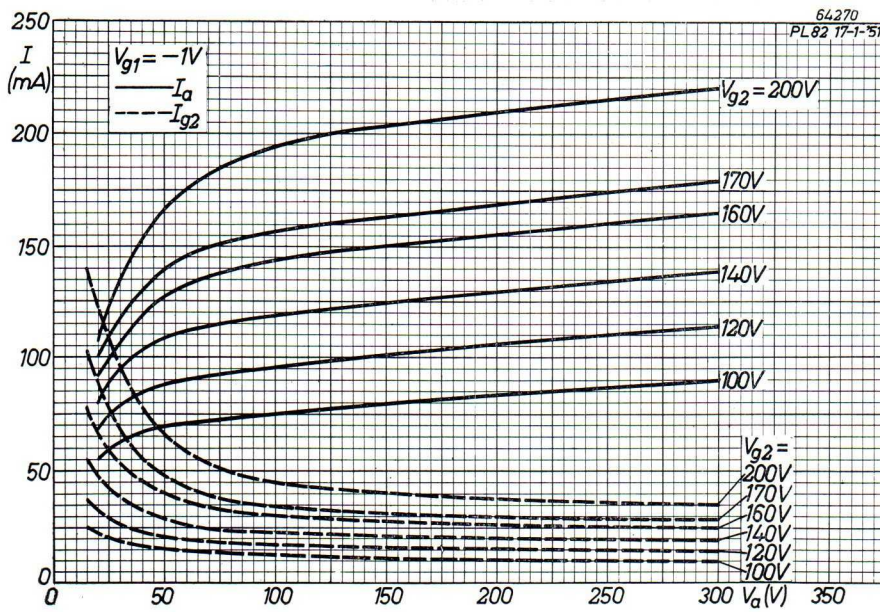


Fig. 67.
Anode current and screen-grid current plotted against anode voltage for control-grid voltage of $-1V$, with screen-grid voltage as parameter.

Fig. 68.
Anode current, screen-grid current, input voltage and total distortion plotted against output power for anode and screen-grid voltage of $170V$.

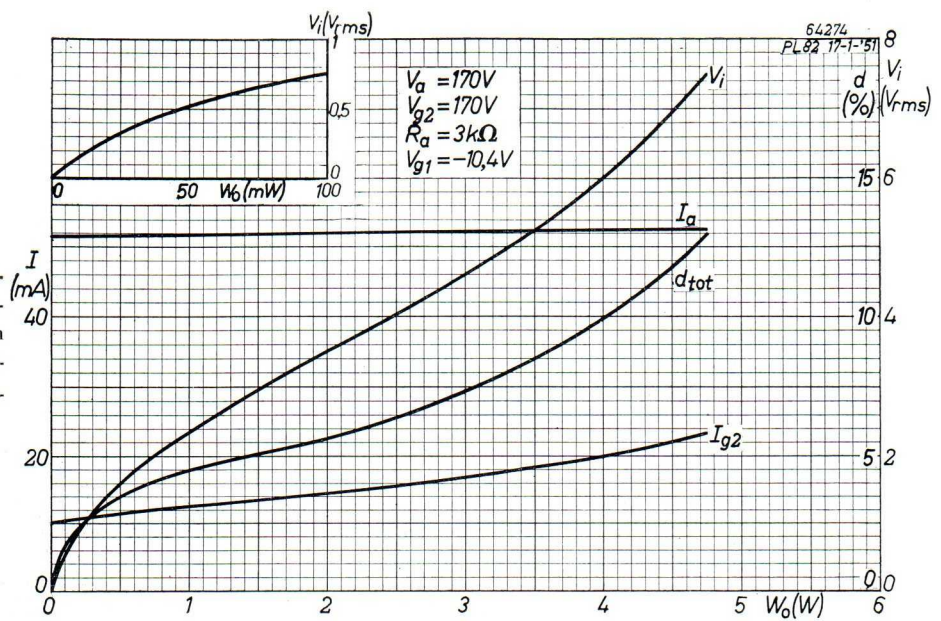


Fig. 69.

Anode current, screen-grid current, input voltage and total distortion plotted against output power for anode and screen-grid supply voltage of 200 V.

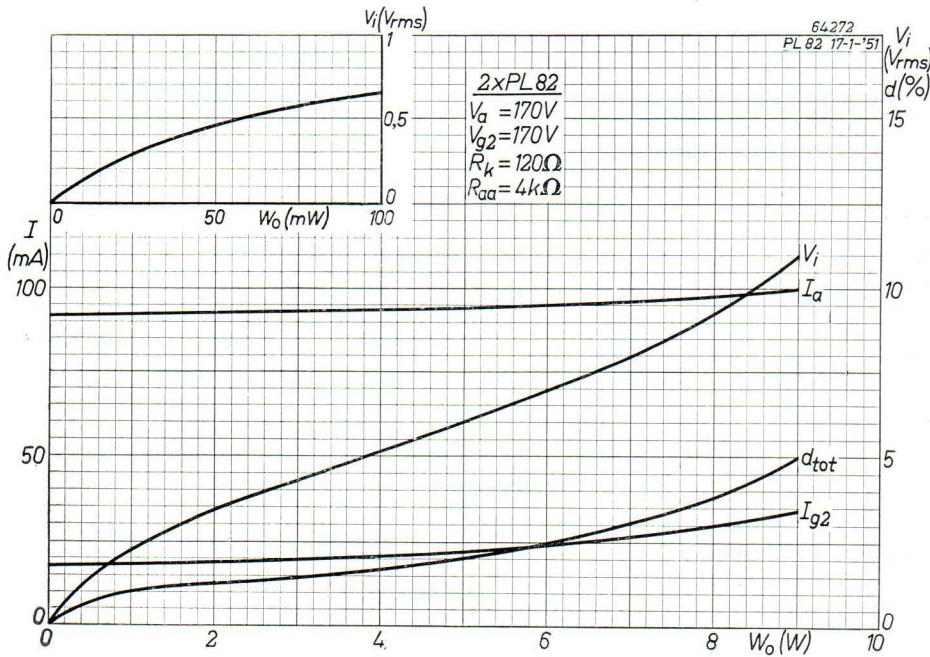
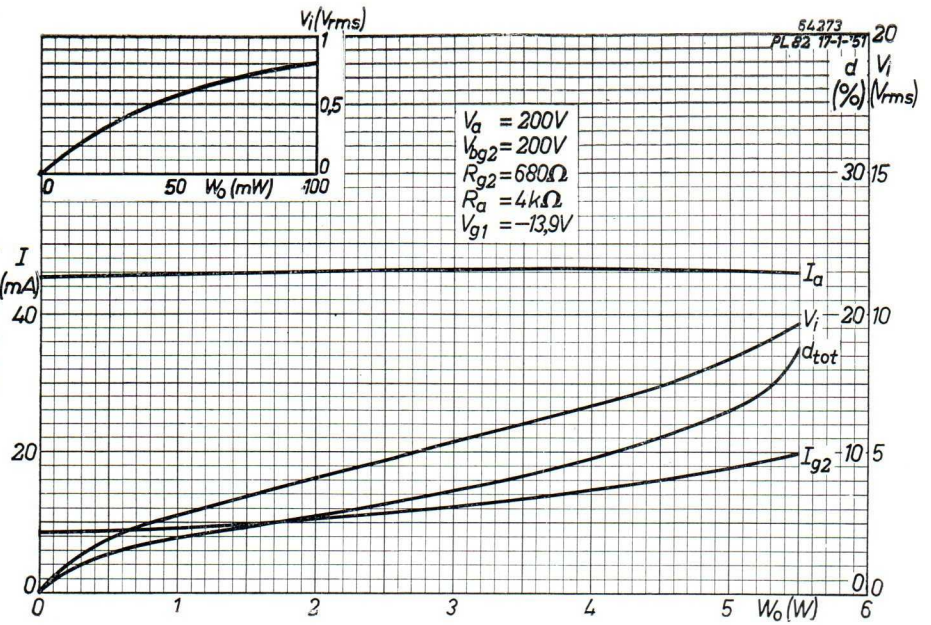
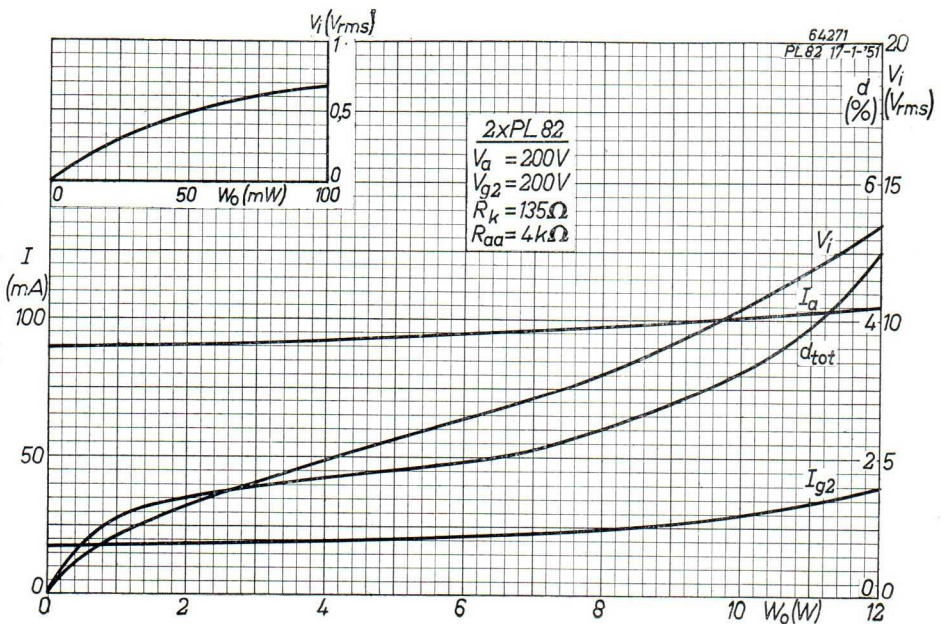


Fig. 70.

Anode current, screen-grid current, input voltage and total distortion plotted against output power of two tubes PL 82 in push-pull for anode and screen-grid voltage of 170 V.

Fig. 71.

Anode current, screen-grid current, input voltage and total distortion plotted against output power of two tubes PL 82 in push-pull for anode and screen-grid voltage of 200 V.



MULTI-PURPOSE TRIODE-PENTODE ECL 80

DESCRIPTION

In all stages of the picture channel and in the I.F. stages of the sound channel of a television receiver single tubes are used almost exclusively. In other stages, however, tubes with two electrode

supply, A.C. or D.C., the heater voltage and current being 6.3 V and 0.3 A respectively.

Some of the wide variety of functions for which the ECL 80 may be employed are listed below:

	Triode section	Pentode section
1	Frame-blocking oscillator	Frame output
2	Line-blocking oscillator	Frame output
3	A.F. voltage amplifier	Sound output
4	Frame or line oscillator	Sync. separator
5	Half of frame multivibrator	Frame output
6	Frame interlace tube	Sync. separator
7	One half of line multivibrator	Other half of line multivibrator

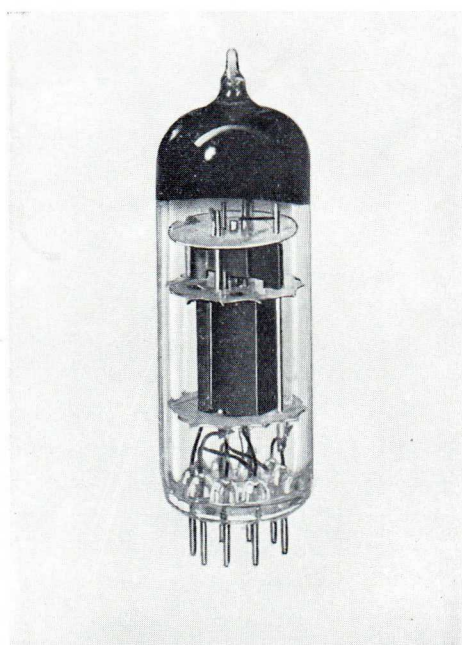


Fig. 72. The triode-pentode ECL 80 (actual size).

systems in one envelope are frequently used. This obviously reduces the total number of tubes to be employed, thus effecting economies in chassis space and also in the cost of the receiver. The most commonly used multiple tubes are double diodes and double triodes, but in some applications a pentode system would offer advantages. However, each section of a double pentode of small dimensions will normally be incapable of providing sufficient power to permit it to be used in the frame output stage or as the sound output tube. These considerations have led to the design of a triode-pentode, type ECL 80, which is not only suitable for a large number of functions but is also of small dimensions. The heater of this triode-pentode is suitable for either series or parallel

APPLICATION

1. TRIODE AS FRAME BLOCKING OSCILLATOR

PENTODE FOR FRAME OUTPUT (rectangular picture 7 kV)

Provided deflection coils of reasonable efficiency are available, the pentode section of the ECL 80 can be used as the frame output tube at a high-tension line voltage of 170 V to 200 V for a picture tube MW 31-16 or MW 22-16 operating with 7 kV E.H.T. and presenting a rectangular picture. Double-D scanning is also often used and, having regard to the larger picture area, it is advisable to employ a higher E.H.T., for example 9 kV. This increases the required deflection power quite considerably, and if the ECL 80 is again to be used as the frame output tube it is necessary to feed the anode of the pentode section from a higher supply voltage, which may well be the boosted high tension in the line output circuit. As previously shown, it is also possible in this case to use a PL 82 instead, which can be fed from the normal high-tension line.

Fig. 73 gives a circuit of the ECL 80 as frame blocking oscillator and frame output tube operating with a line voltage of 180 V. This circuit is suitable for the vertical deflection of a picture tube MW

31-16 or MW 22-16 provided the E.H.T. is not higher than 7 kV and a rectangular picture is used.

The pentode grid is supplied with a saw-tooth voltage from the grid circuit of the blocking oscillator, as a more linear saw-tooth is obtained for the same H.T. voltage than from an RC network in the

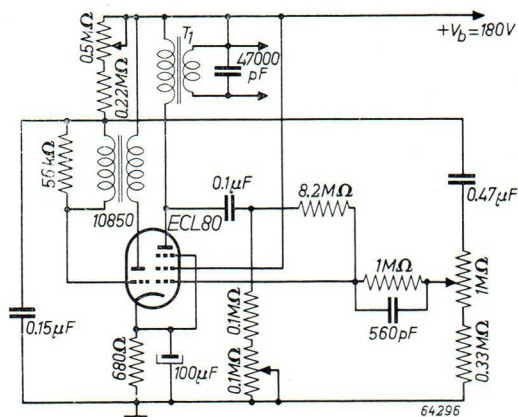


Fig. 73. Circuit for vertical scanning of a rectangular picture on a picture tube MW 22-16 or MW 13-16 operating at 7 kV second-anode voltage.

anode circuit. A blocking transformer type 10850 can be used, the turns ratio of which from the grid circuit to the anode is 2 : 1. The horizontal hold is effected with a potentiometer of 0.5 MΩ in the grid circuit. For the synchronization negative-going pulses can be applied to the anode of the triode section.

The data of the frame output transformer T_1 are not given in fig. 73 because these depend upon the characteristics of the deflection coils used. An output transformer can be calculated on the lines indicated in the application notes of the PL 82, but it should be realized that adequate primary self-inductance must be provided so that the peak anode current does not exceed about 26 mA at $V_a = 50$ V and $V_{g2} = 170$ V. As indicated in the diagram, a capacitor of 47,000 pF should be connected across the transformer secondary in order to prevent interference from the line coils.

Heavy negative feedback is applied across the pentode section of the ECL 80. This feedback prevents any tendency towards microphony, whilst, moreover, the time constant of the RC network feeding back the anode alternating voltage is so chosen that a parabolic component is added to the driving voltage. The amplitude of the parabolic component, and thus the linearity of the deflection current, can be regulated with the potentiometer of 0.1 MΩ. The component values indicated only

serve as a general guide and depend upon the primary self-inductance of the transformer used.

It will be noticed that the resistor between the pentode grid and the height control is shunted by a capacitor of 560 pF. This capacitor has practically no influence during the scan, but during the flyback it results in a negative pulse being supplied to the control grid and thus blocking the tube. Lengthening of the flyback by the flow of anode current is therefore prevented.

2. TRIODE AS FRAME BLOCKING OSCILLATOR

PENTODE FOR FRAME OUTPUT (double-D picture 9 kV)

It has already been indicated that for a double-D picture on a picture tube operating at 9 kV second-anode voltage a higher deflection power is required than in the case of a rectangular picture and 7 kV E.H.T. It is then necessary either to feed the anode of the pentode section of the ECL 80 from a higher supply voltage or to use a PL 82 instead.

Fig. 74 gives a circuit in which the ECL 80 is used. This circuit is intended to operate in conjunction with the deflection and focusing unit type 10914, data of which are given in the application notes of the PL 81, where a line output circuit is described operating with a booster diode and giving a boosted H.T. voltage of 410 V. This voltage is also used to supply the anode of the frame output pentode. The circuit of fig. 74 is capable of providing the vertical deflection on a picture tube MW 22-16 or MW 31-16 for a double-D picture and 9 kV E.H.T.

The sync. signal is applied to the blocking oscillator via a double integration circuit, which gives a better interlace than with single integration. The charge capacitor C_4 in the triode grid circuit has been kept low so as to facilitate synchronization with the integrated sync. signal. A sync. signal of about 75 V peak-to-peak is required at the terminal S, and this can be obtained with the separator circuit given in fig. 75. A standard blocking transformer type 10850 with a turns ratio of 2 : 1 between L_2 and L_1 has been used. The potentiometer R_5 is the vertical hold control.

The saw-tooth voltage for driving the output pentode is obtained in the anode circuit of the triode section, which is supplied with 410 V. R_{13} and C_9 in series with the charge capacitor C_8 serve to give a negative peak during the flyback, which

blocks the pentode. In the circuit previously described this peak was obtained by reducing the negative feedback across the pentode for high frequencies. An advantage of the arrangement given here is that a high feedback factor is maintained

Deflection current peak-to-peak	540 mA
Average anode current of the pentode	9.6 mA
Peak anode current of the pentode	21.5 mA
Screen-grid current	2.2 mA
Cathode voltage	approx. 10.5 V

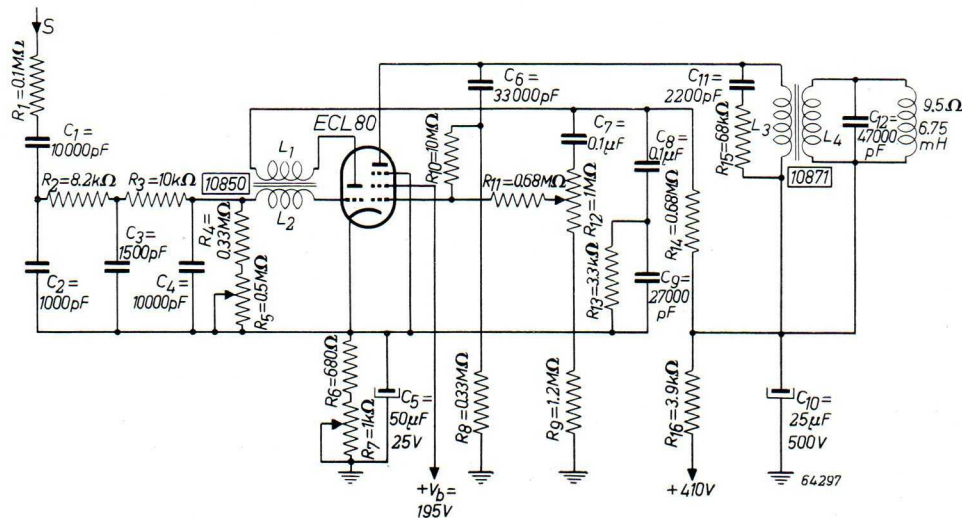


Fig. 74. Circuit for vertical scanning of a double-D picture on a picture tube MW 22—16 or MW 31—16 operating at 9 kV second-anode voltage. The anodes of the triode section and the pentode section are fed from the boosted high-tension supply of 410 V in the line output circuit (see fig. 98).

also for high frequencies, which is favourable with regard to microphony.

There are two possible arrangements for the height control. According to one arrangement the charge resistor can be varied, but this has the disadvantage that, owing to the high time constant of $R_{14}C_8$, there is a time delay in the adjustment of the picture height. The method represented in fig. 74, see R_{12} , does not possess this disadvantage.

Negative feedback is applied across the output pentode via C_6 and R_8 . The values of these components are so chosen that the required parabolic component is obtained in the anode current. It may be necessary, when large deviations in component values are to be expected, to make R_8 variable for adjusting the linearity. The linearity of the beginning of the scan is adjusted with the potentiometer R_7 in the cathode circuit.

A standard output transformer type 10871 is used in this circuit. Owing to the high supply voltage and the consequently high turns ratio of this transformer, a high positive voltage peak occurs across the primary during the flyback. This peak has been kept below 1200 V by the inclusion of the damping circuit $R_{15}C_{11}$.

The currents and voltages in the output circuit are as follows:

3. THE ECL 80 AS SYNCHRONIZATION AMPLIFIER AND CLIPPER

A circuit in which the ECL 80 is used as synchronization amplifier and clipper is given in fig. 75. The composite video signal, the sync. pulses of which are positive, is applied to the control grid of the pentode section via a normal blocking

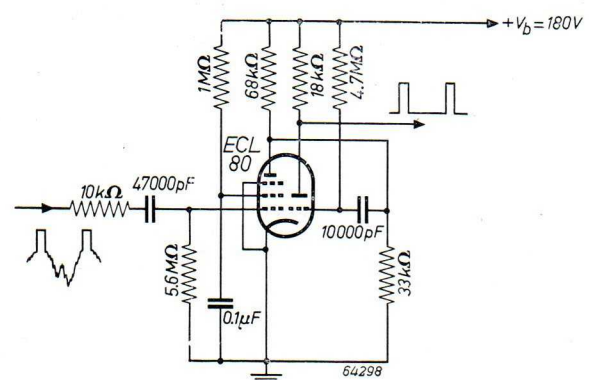


Fig. 75. Synchronization amplifier and clipper employing ECL 80.

capacitor and a resistor of 10 kΩ, the sole purpose of which is to prevent an undue increase of the capacitance in the anode circuit of the preceding video stage. The screen-grid voltage is chosen low, giving a grid base of only about 2 V, thus ensuring

effective separation of sync. and video signals. The average anode voltage is about 48 V and during the sync. pulses this drops below the knee in the I_a/V_a characteristic, so that the peaks of the sync. pulses are limited. At the anode of the pentode section negative-going sync. pulses of 45 V peak-to-peak are available, and these are applied to the triode grid for further amplification and limiting. Finally the output pulses of the triode section are positive-going and the amplitude is 75 V peak-to-peak.

4. TRIODE AS A.F. VOLTAGE AMPLIFIER PENTODE FOR SOUND OUTPUT

Extensive information referring to these uses are given under "Technical Data". The ECL 80 permits the design of an audio amplifier of higher sensitivity than that of a normal single audio stage. Precautions should be taken, however, to avoid two possible sources of oscillation.

Oscillation at a low audio frequency may occur due to feedback through the common cathode resistor circuit when a common cathode resistor, tapped for the triode, is used. This may be avoided by using a long time constant in the cathode circuit ($C_k = 100 \mu\text{F}$) and a short one for the coupling between the triode anode and the pentode grid

($C_{g1} = 5000 \text{ pF}$; $R_{g1} = 0.68 \text{ M}\Omega$). The A.C. earth return of the circuit feeding the triode grid should be connected to the cathode, as otherwise the time constant of $C_{g1}R_{g1}$ may have to be further reduced.

Oscillation at a high audio frequency may occur due to coupling through the capacitance between the pentode anode and the triode grid. The impedance in the triode grid circuit should therefore be kept as low as possible ($R_{gt} = 0.2 \text{ M}\Omega$). The output transformer primary should also be shunted with a high-frequency RC damping circuit to prevent a rise in the anode load impedance of the pentode at the higher frequencies. In extreme cases it may be necessary to use a capacitor between triode anode and triode grid, or between triode grid and earth.

The tendency towards oscillation at a high frequency as mentioned above may, however, most conveniently be reduced by applying negative feedback from the anode of the pentode section to the triode anode. A resistor of $0.68 \text{ M}\Omega$ will suffice for this purpose. The effect of this form of feedback on distortion and sensitivity is such that the sensitivity is reduced to about half that obtained without negative feedback whilst the distortion is considerably reduced, especially at low output; for example at 0.1 W without negative feedback $d_{\text{tot}} = 2.5\%$, and with negative feedback $d_{\text{tot}} = 0.5\%$.

TECHNICAL DATA

HEATER DATA

Heating: indirect by A.C. or D.C.;
series or parallel supply

Heater voltage	V_f	6.3 V
Heater current	I_f	0.3 A

BASE CONNECTIONS AND DIMENSIONS (in mm)

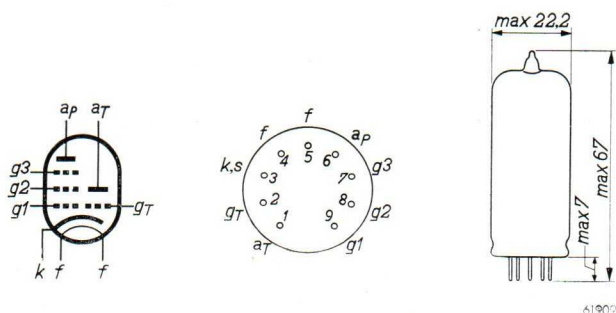


Fig. 76.

Mounting position: any

CAPACITANCES

Pentode section	C_a	5.0 pF
	C_{g1}	4.5 pF
	C_{ag1}	< 0.2 pF
	C_{g1f}	< 0.25 pF
	C_{kf}	3.7 pF
Triode section	C_a	0.8 pF
	C_g	2.1 pF
	C_{ag}	0.9 pF
	C_{gf}	< 0.05 pF
	C_{kf}	3.7 pF
Between sections	C_{gt-ap}	< 0.12 pF
	C_{at-ap}	< 1.2 pF
	C_{gt-g1p}	< 0.2 pF
	C_{at-g1p}	< 0.2 pF

TYPICAL CHARACTERISTICS

Pentode section

Supply voltage	V_b	170	200	250 V
Anode voltage	V_a	170	200	250 V
Suppressor-grid voltage	V_{g3}	0	0	0 V
Screen-grid voltage	V_{g2}	170	200	V
Screen-grid resistor	R_{g2}	—	—	4.7 k Ω
Control-grid voltage	V_{g1}	-6.7	-8	-12.2 V
Anode current	I_a	15	17.5	14 mA
Screen-grid current	I_{g2}	2.8	3.3	2.6 mA
Mutual conductance	S	3.2	3.3	2.6 mA/V
Internal resistance	R_i	0.15	0.15	0.2 M Ω
Amplification factor between screen grid and control grid	μ_{g2g1}	14	14	14

Triode section

Anode voltage	V_a	100 V
Grid voltage	V_g	0 V
Anode current	I_a	8 mA
Mutual conductance	S	1.9 mA/V
Amplification factor	μ	20

OPERATING CONDITIONS AS SOUND

OUTPUT TUBE, pentode section

Supply voltage	V_b	170	200	250 V
Anode voltage	V_a	170	200	250 V
Suppressor-grid voltage	V_{g3}	0	0	0 V
Screen-grid voltage	V_{g2}	170	200	V
Screen-grid resistor	R_{g2}	—	—	4.7 k Ω
Control-grid voltage	V_{g1}	-6.7	-8	-12.2 V
Anode current	I_a	15	17.5	14 mA
Screen-grid current	I_{g2}	2.8	3.3	2.6 mA
Mutual conductance	S	3.2	3.3	2.6 mA/V
Internal resistance	R_i	0.15	0.15	0.2 M Ω
Matching resistance	R_a	11	11	17.5 k Ω
Power output ($d_{tot} = 10\%$)		1	1.4	1.55 W
Input signal ($d_{tot} = 10\%$)		3.7	4.1	5.3 V _{rms}
Power output ($\eta = 50\%$)		1.27	1.75	1.75 W
Input signal ($\eta = 50\%$)		4.4	5.1	5.9 V _{rms}
Input signal ($W_o = 50$ mW)		0.7	0.7	0.75 V _{rms}

OPERATING CONDITIONS AS SYNCHRONIZATION SEPARATOR, pentode section

Anode voltage	V_a	20	V
Suppressor-grid voltage	V_{g3}	0	V
Screen-grid voltage	V_{g2}	12	V
Control-grid voltage	V_{g1}	0	-1.45 V
Anode current	I_a	2	0.1 mA

OPERATING CONDITIONS AS A.F. VOLTAGE

AMPLIFIER, triode section

V_b (V)	R_a (k Ω)	$R_g^{(1)}$ (k Ω)	$-V_g$ (V)	I_a (mA)	$V_o^{(2)}$ (V _{rms})	V_o/V_i	d_{tot} (%)
170	47	150	3.5	1.8	22	9.5	8.7
170	100	330	3.5	1.0	24	10.5	7.6
170	220	680	3.5	0.5	24.5	11	6.5
200	47	150	4.2	2.2	27	9.5	9.0
200	100	330	4.2	1.2	29	10.5	8.0
200	220	680	4.2	0.6	30	11	6.5
250	47	150	5.5	2.8	36	9.5	9.2
250	100	330	5.5	1.5	39	10.5	8.3
250	220	680	5.5	0.75	40	11	7.0

¹⁾ Grid resistor of next tube.²⁾ Output voltage at start of I_o . At lower output voltages the distortion is proportionately reduced.

LIMITING VALUES

Pentode section

Anode voltage at zero anode current	V_{a0}	max.	550 V
Anode voltage	V_a	max.	400 V
Positive peak anode voltage (10% of the duration of a cycle with a maximum of 2 m sec.)		max.	1200 V
Negative peak anode voltage		max.	500 V
Screen-grid voltage at zero screen-grid current	V_{g20}	max.	550 V
Screen-grid voltage	V_{g2}	max.	250 V
Control-grid voltage (grid current +0.3 μ A)	V_{g1}	max.	-1.3 V
Heater voltage during the warming-up period		max.	9.5 V
Voltage between heater and cathode	V_{kf}	max.	150 V
Cathode current	I_k	max.	25 mA

Peak cathode current (10% of the duration of a cycle with a maximum of 2 m sec.)

I_{kp} max. 350 mA

Anode dissipation W_a max. 3.5 W

Screen-grid dissipation W_{g2} max. 1.2 W

External resistance between control grid and cathode

with fixed bias R_{g1} max. 1 M Ω

with automatic bias max. 2 M Ω

External resistance between heater and cathode

R_{kf} max. 20 k Ω

To allow for inevitable spread between individual tubes and for deterioration during life in frame output applications, the circuit should be designed around a peak anode current not exceeding 26 mA at $V_a = 50$, $V_{g2} = 170$ V. At $V_a = 60$ V, $V_{g2} = 200$ V the corresponding value for the peak current is 31 mA and at $V_a = 70$ V, $V_{g2} = 250$ V it is 42 mA.

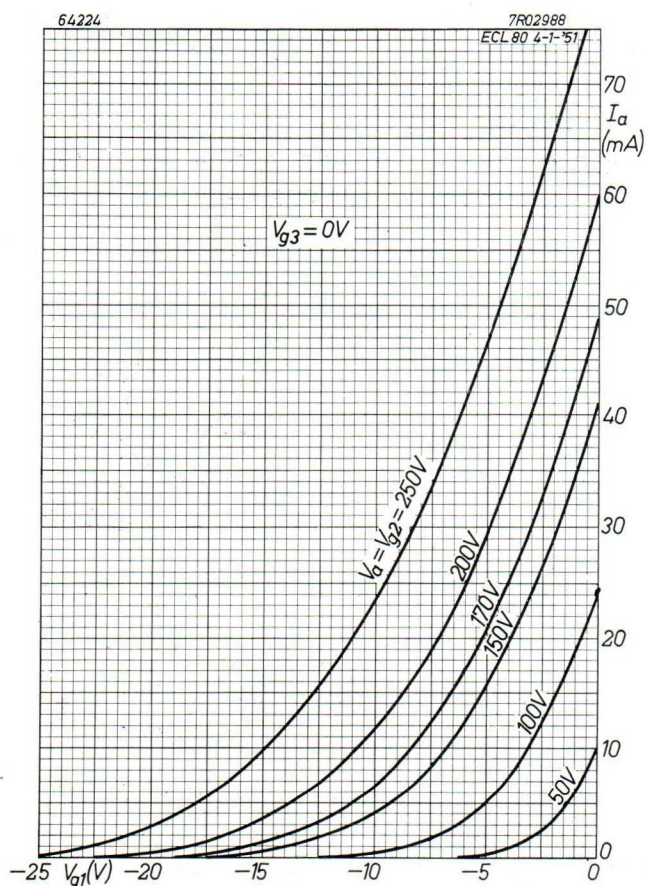


Fig. 77 Anode current of pentode section plotted against control-grid voltage for anode and screen-grid voltages between 50 V and 250 V.

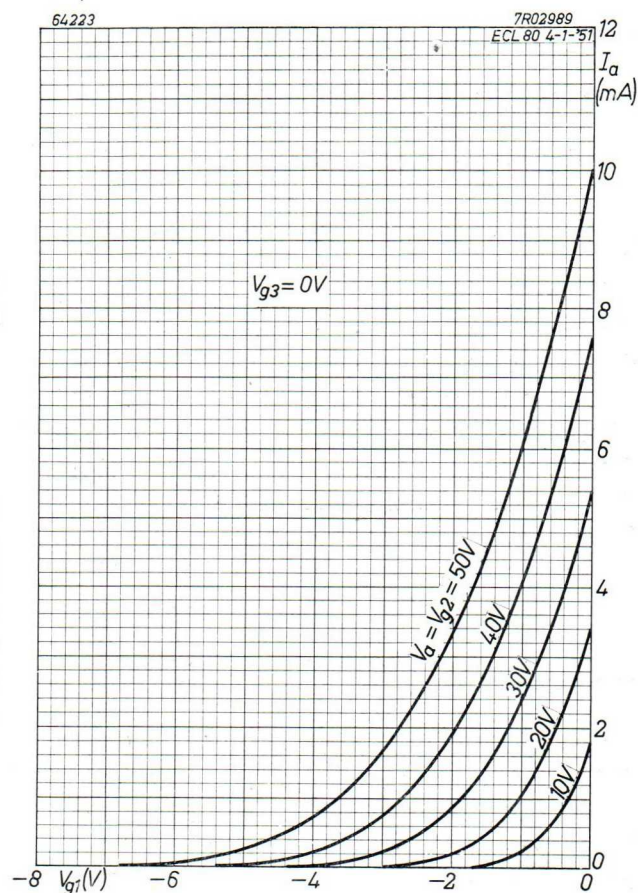


Fig. 78 Anode current of pentode section plotted against control-grid voltage for anode and screen-grid voltages between 10 V and 50 V.

Triode section

Anode voltage at zero anode

current V_{ao} max. 550 V

Anode voltage V_a max. 200 V

Control-grid voltage

(grid current $+0.3 \mu A$) V_g max. -1.3 V

Heater voltage during

warming-up period max. 9.5 V

Voltage between heater and

cathode V_{kf} max. 150 V

Cathode current I_k max. 8 mA

Peak cathode current (10% of the duration of a cycle with a maximum of 2 m sec.)

I_{kp} max. 200 mA

Anode dissipation W_a max. 1 W

External resistance between control grid and cathode

R_g max. 1 M Ω

with fixed bias max. 3 M Ω

with automatic bias

External resistance between

heater and cathode R_{kf} max. 20 k Ω

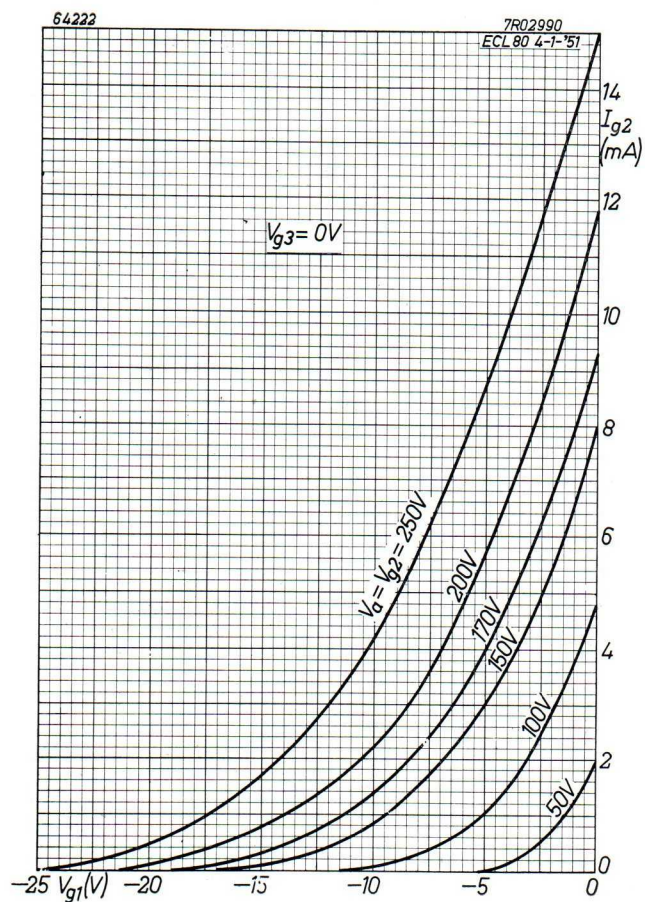


Fig. 79.

Screen-grid current of **pentode section** plotted against control-grid voltage for anode and screen-grid voltages between 50 V and 250 V.

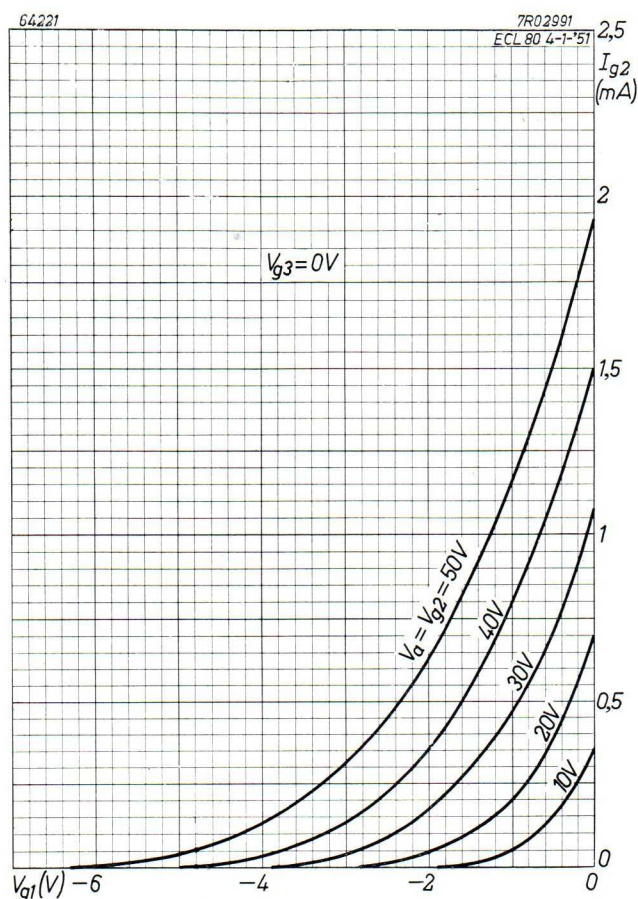


Fig. 80.

Screen-grid current of **pentode section** plotted against control-grid voltage for anode and screen-grid voltages between 10 V and 50 V.

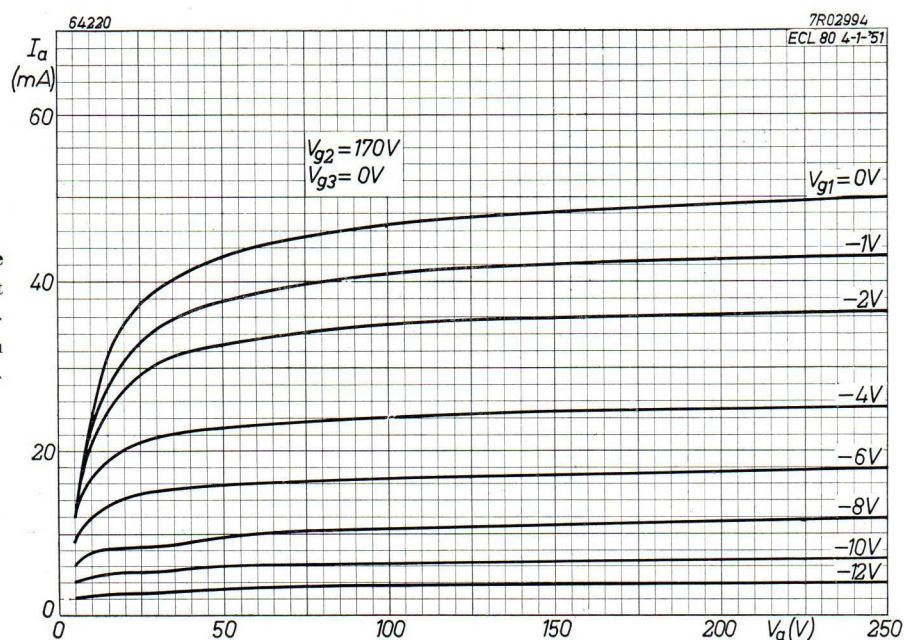


Fig. 81.

Anode current of **pentode section** plotted against anode voltage for screen-grid voltage of 170 V, with control-grid voltage as parameter.

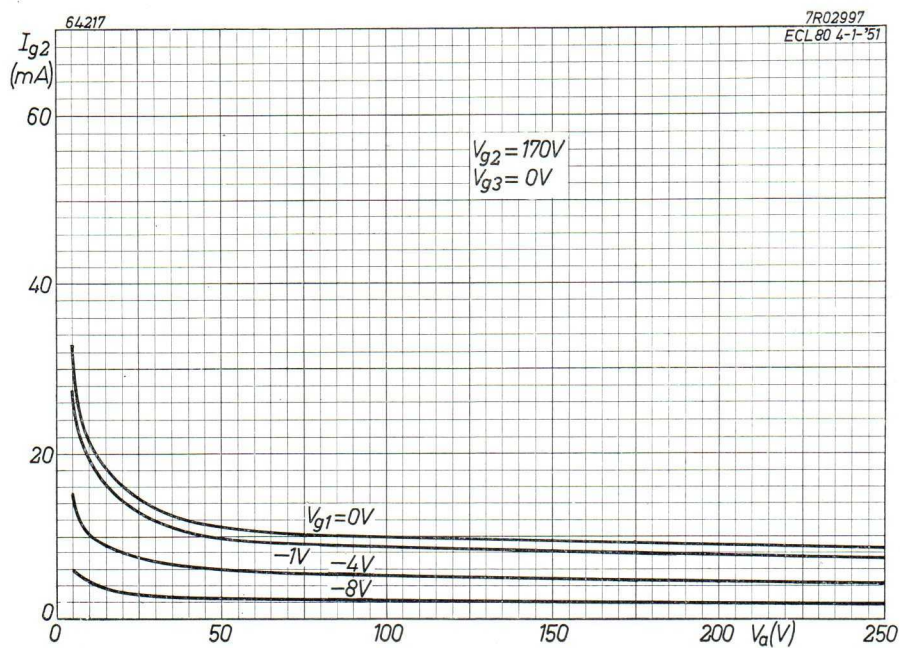


Fig. 82.
Screen-grid current of
pentode section plotted
against anode voltage for
screen-grid voltage of
170 V, with control-grid
voltage as parameter.

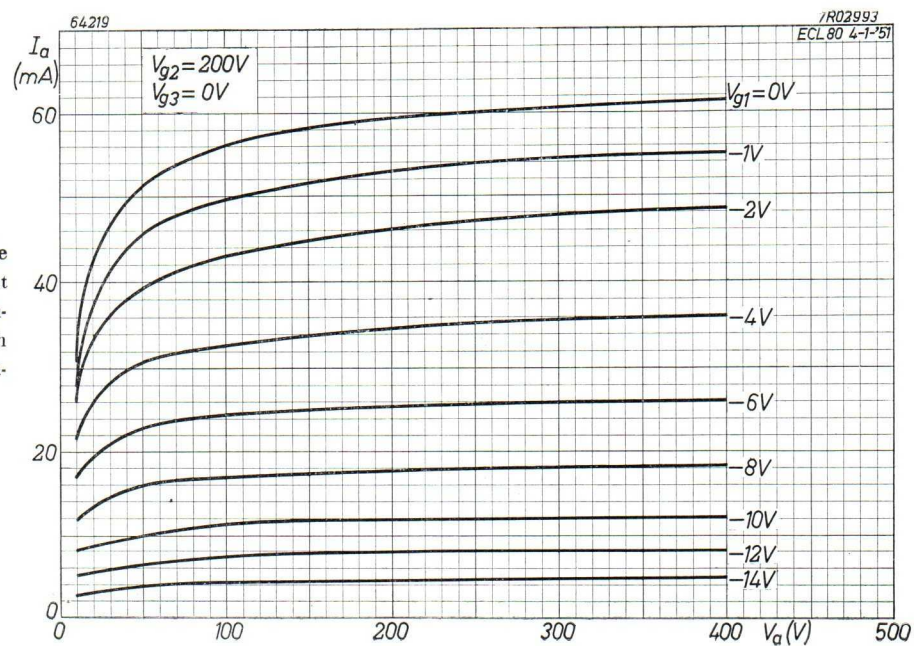


Fig. 83.
Anode current of pentode
section plotted against
anode voltage for screen-
grid voltage of 200 V, with
control-grid voltage as pa-
rameter.

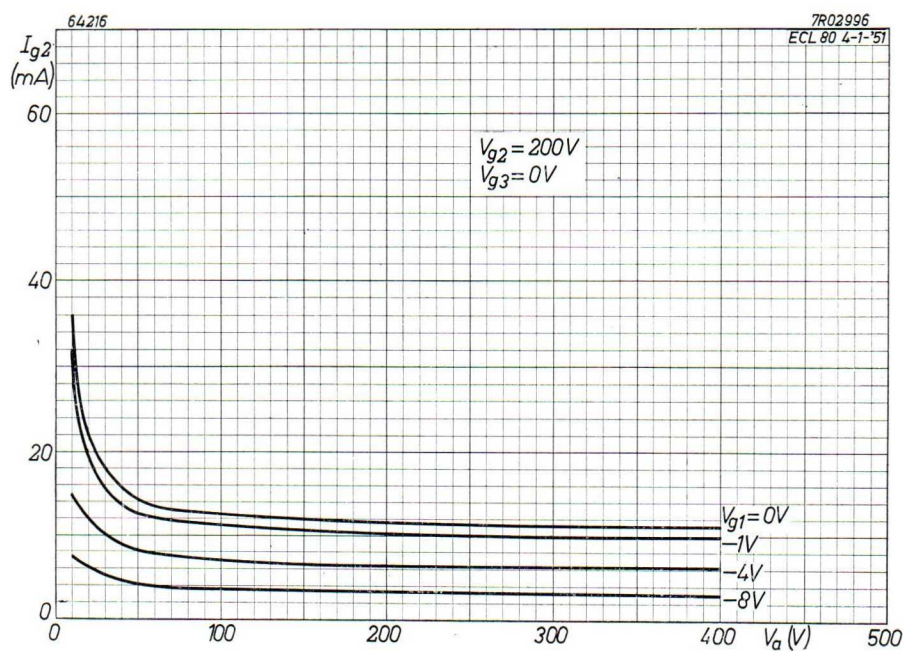


Fig. 84.
Screen-grid current of pentode section plotted against anode voltage for screen-grid voltage of 200 V, with control-grid voltage as parameter.

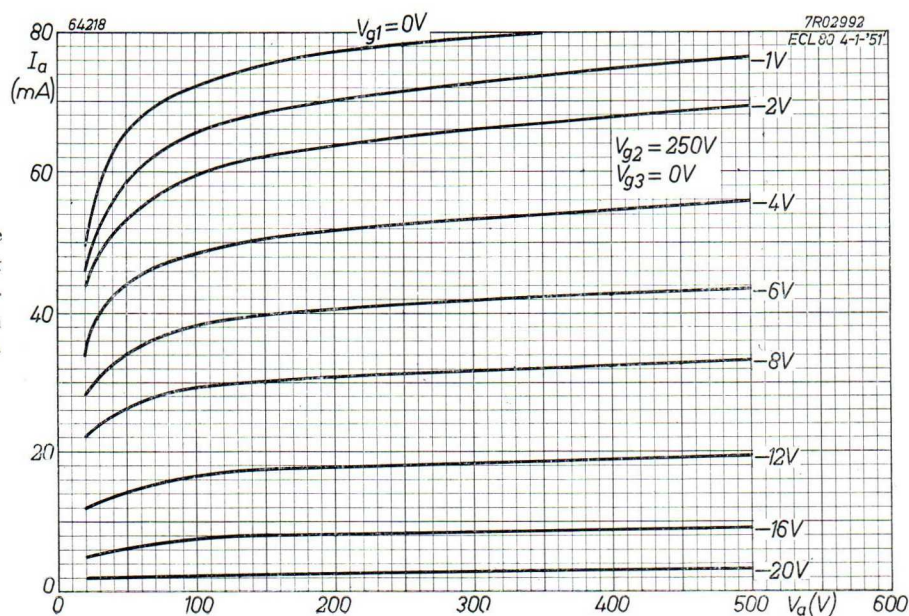


Fig. 85.
Anode current of pentode section plotted against anode voltage for screen-grid voltage of 250 V, with control-grid voltage as parameter.

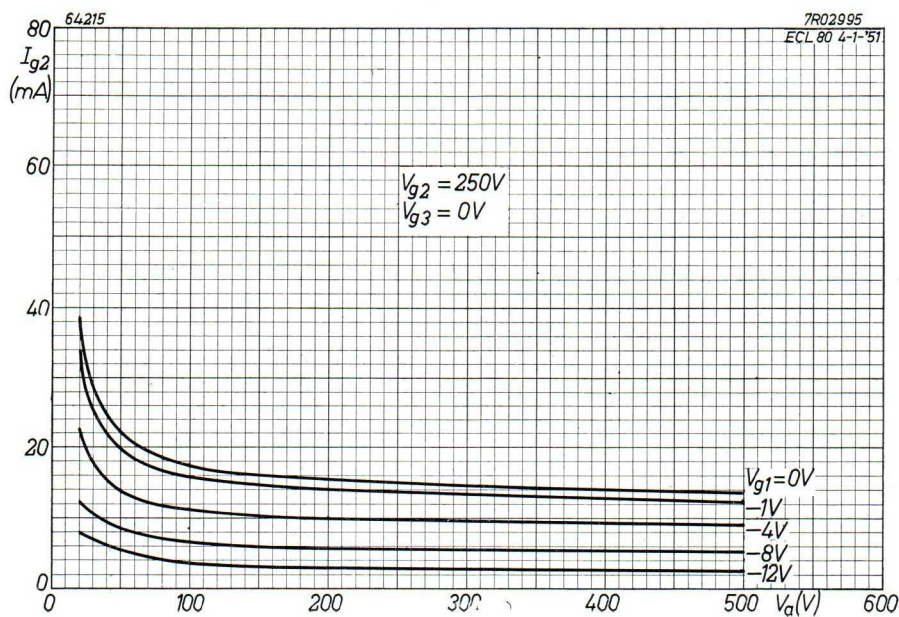


Fig. 86.

Screen-grid current of **pentode section** plotted against anode voltage for screen-grid voltage of 250 V, with control-grid voltage as parameter.

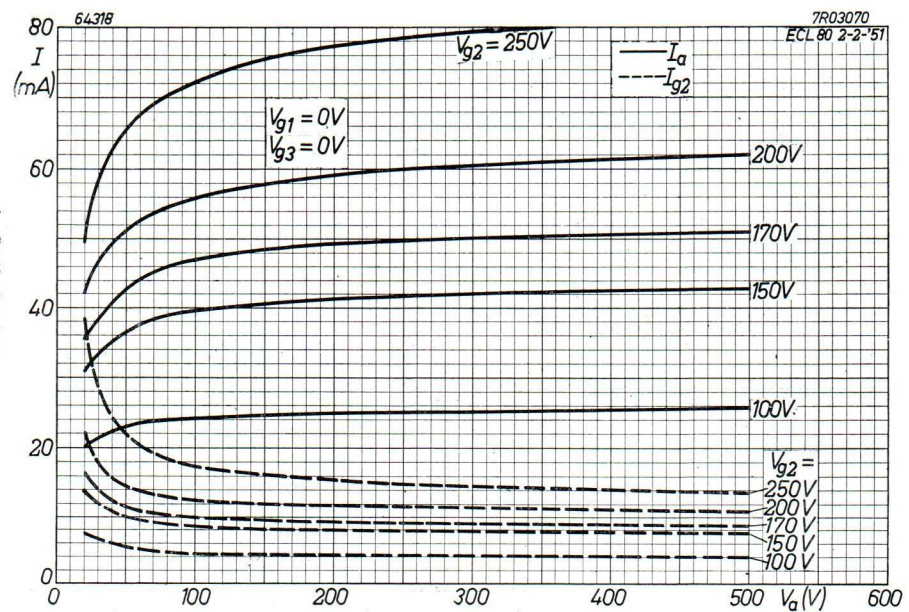


Fig. 87.

Anode current and screen-grid current of **pentode section** plotted against anode voltage for control-grid voltage of 0 V, with screen-grid voltage as parameter.

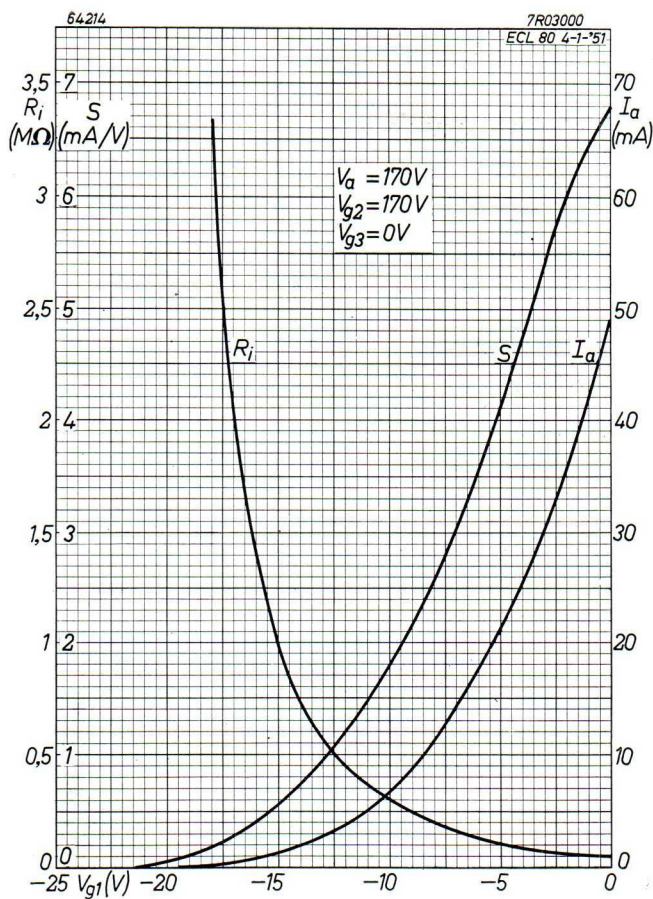


Fig. 88.
Anode current, mutual conductance and internal resistance of pentode section plotted against control-grid voltage for anode and screen-grid voltage of 170 V.

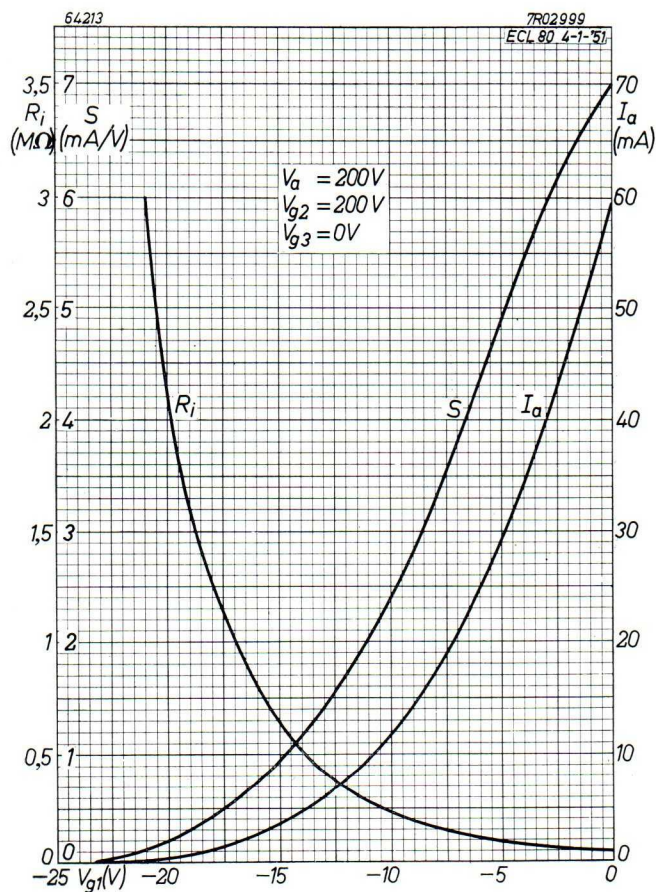


Fig. 89.
Anode current, mutual conductance and internal resistance of pentode section plotted against control-grid voltage for anode and screen-grid voltage of 200 V.

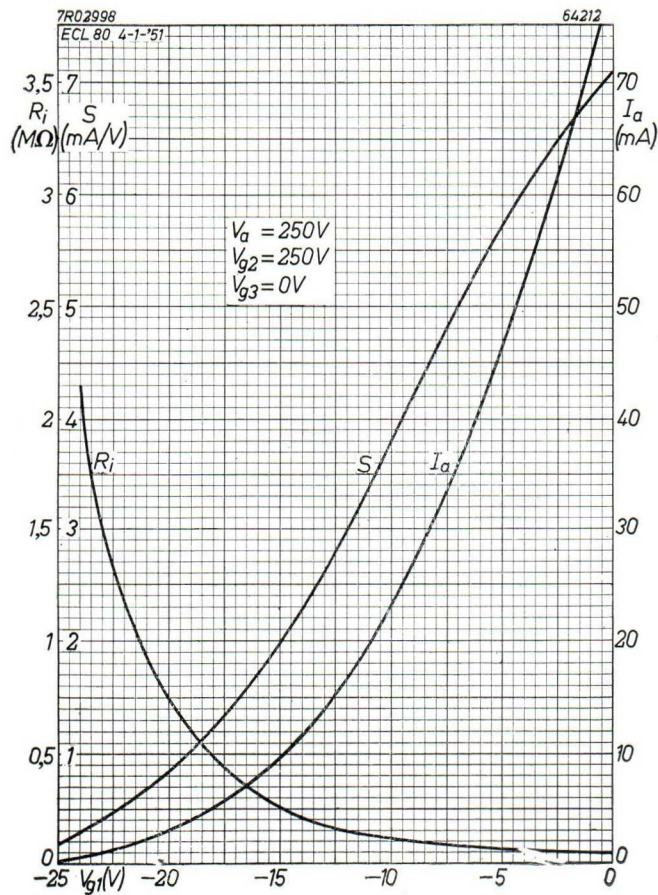
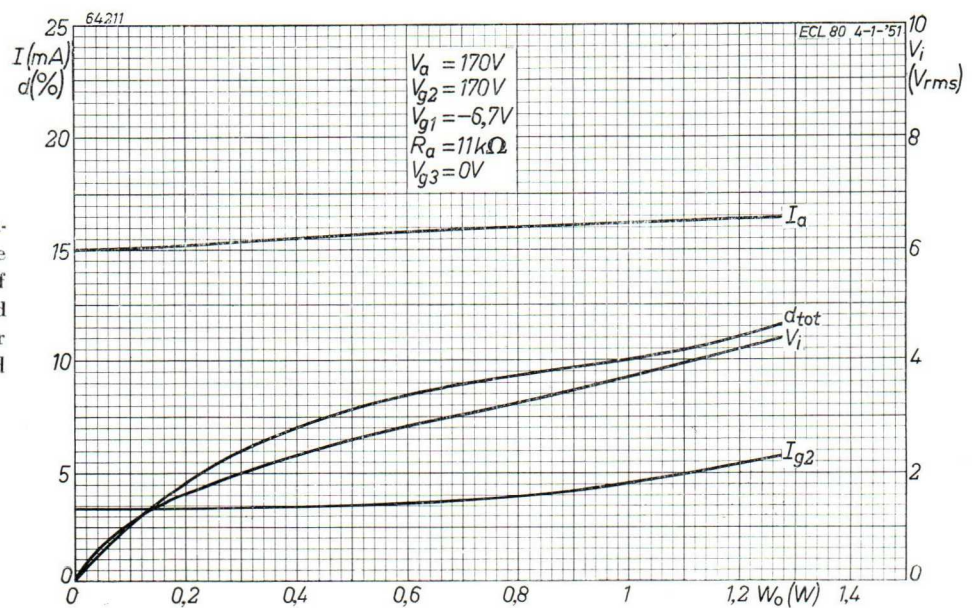


Fig. 90.
Anode current, mutual conductance and internal resistance of pentode section plotted against control-grid voltage for anode and screen-grid voltage of 250 V.

Fig. 91.
Anode current, screen-grid current, input voltage and total distortion of pentode section plotted against output power for anode and screen-grid voltage of 170 V.



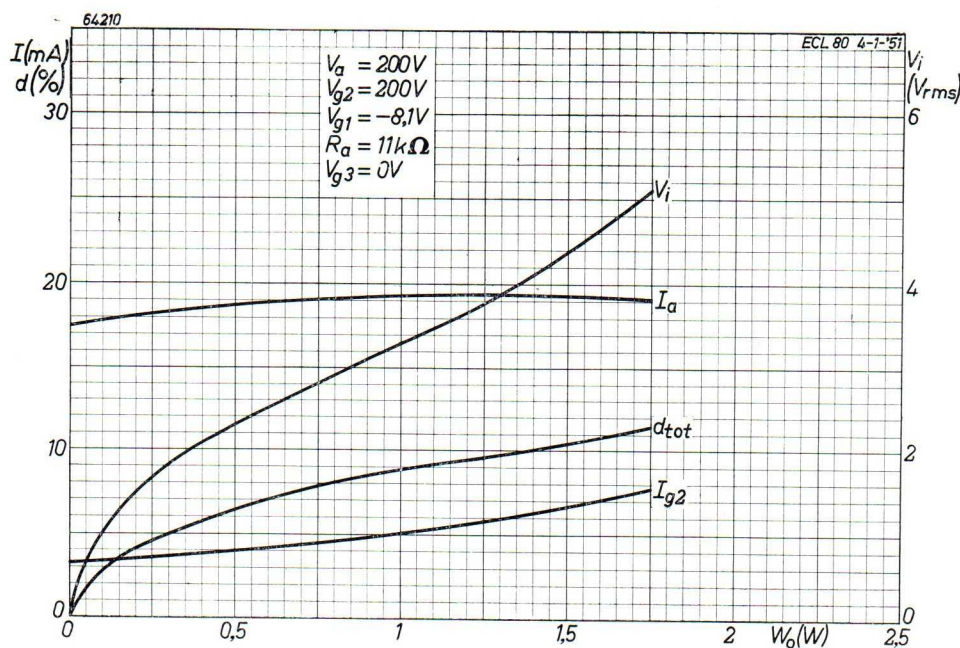


Fig. 92.

Anode current, screen-grid current, input voltage and total distortion of pentode section plotted against output power for anode and screen-grid voltage of 200 V.

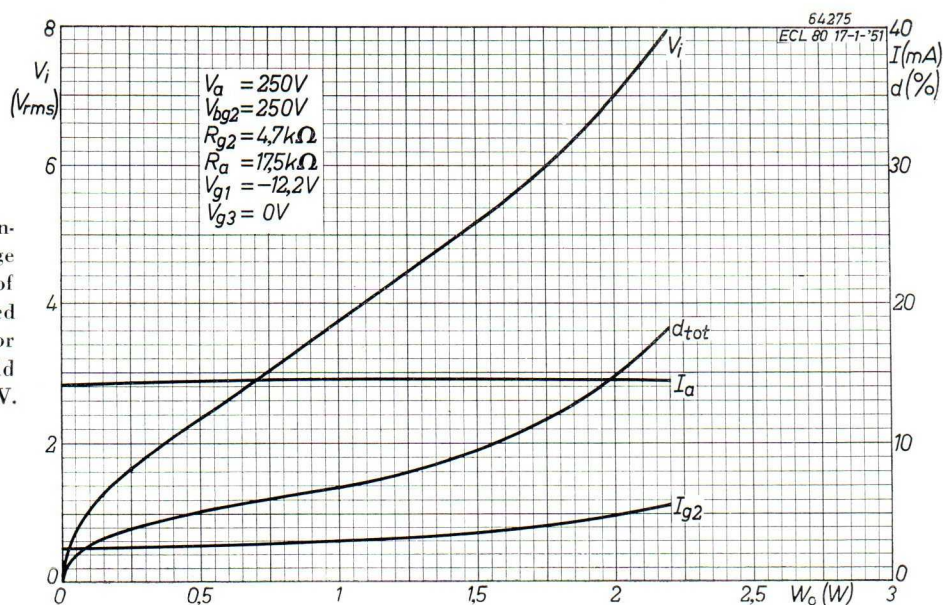


Fig. 93.

Anode current, screen-grid current, input voltage and total distortion of pentode section plotted against output power for anode and screen-grid supply voltage of 250 V.

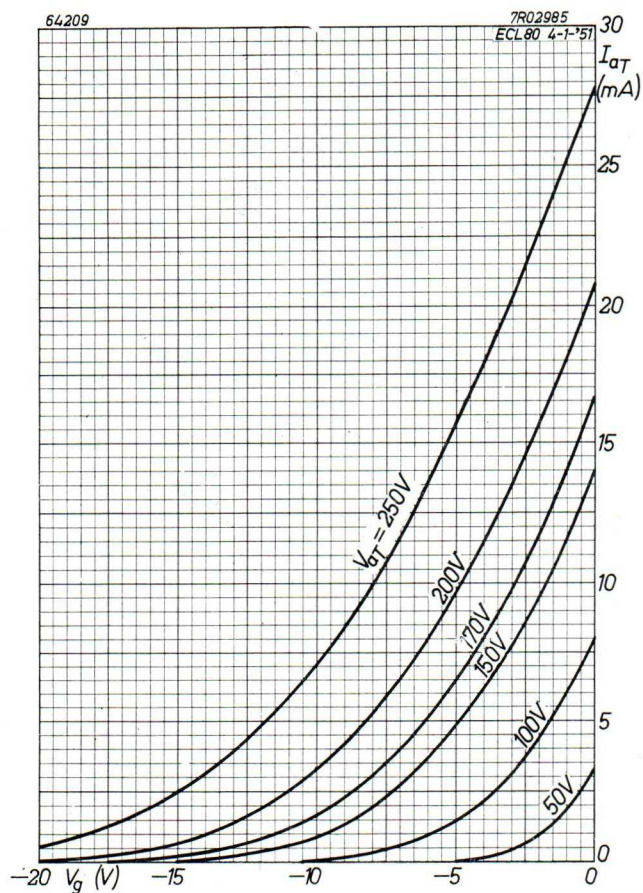


Fig. 94.

Anode current of **triode section** plotted against control-grid voltage for anode voltages between 50 V and 250 V.

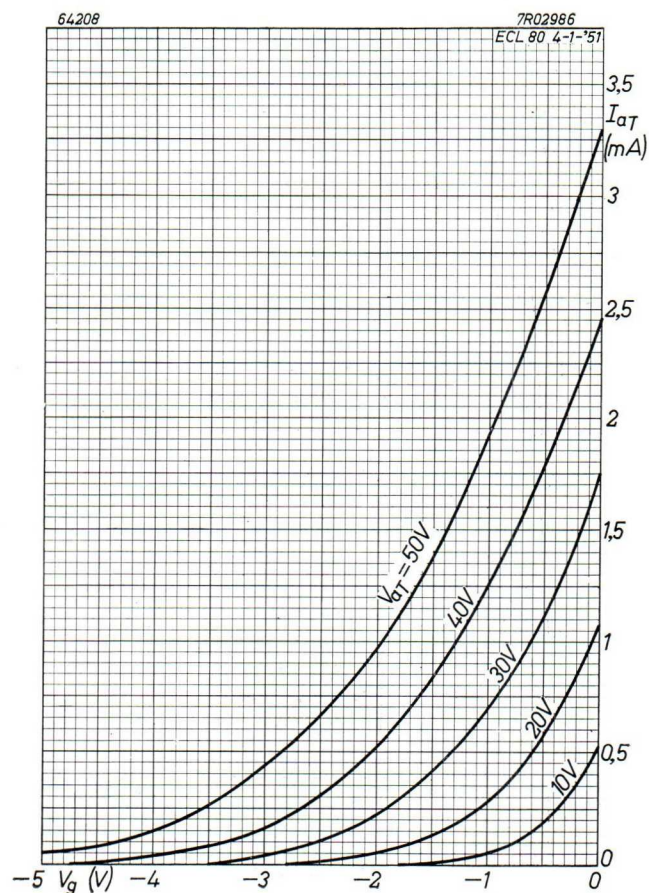
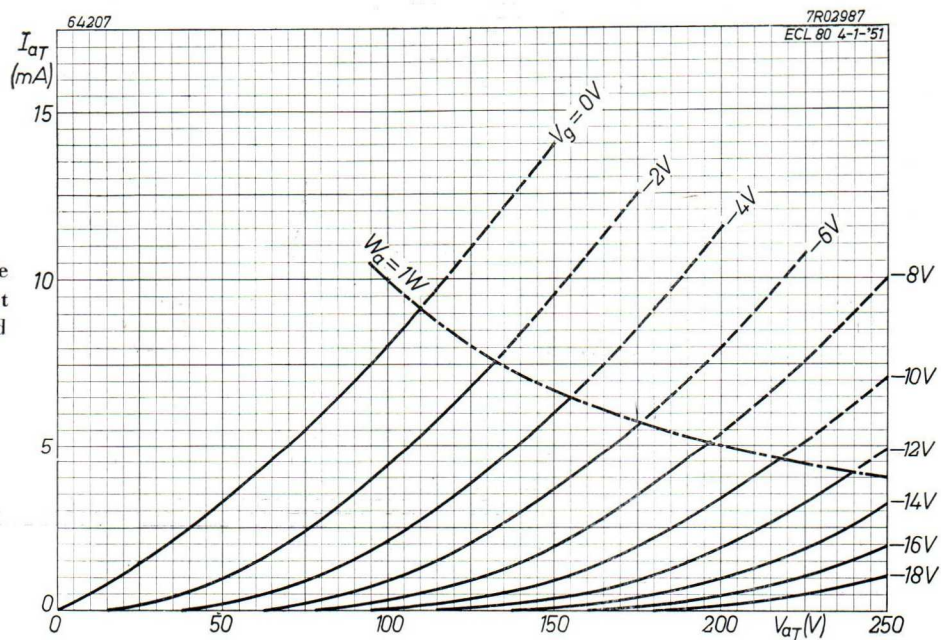


Fig. 95.

Anode current of **triode section** plotted against control-grid voltage for anode voltages between 10 V and 50 V.

Fig. 96.

Anode current of **triode section** plotted against anode voltage, with grid voltage as parameter.



LINE OUTPUT PENTODE PL 81

DESCRIPTION

The PL 81 is an all-glass power pentode on Noval base and with a top anode connection, specifically designed for operation as a line output tube. In spite of its small dimensions the average tube will

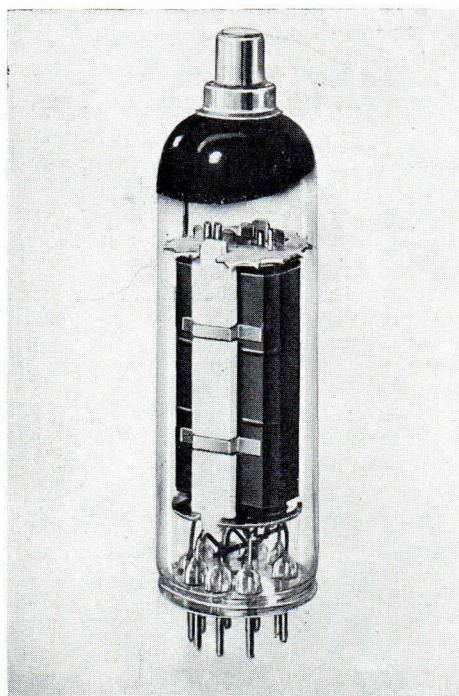


Fig. 97. The line output pentode PL 81 (actual size).

deliver a peak anode current of 380 mA at an anode voltage of 70 V and a screen-grid voltage of 170 V. The anode is adequately insulated to withstand positive and negative voltage peaks of 7 kV.

Quite apart from the use of the PL 81 in a television receiver, for which it is primarily intended, two of these pentodes can also be used in the final stage of an A.F. power amplifier. At a supply voltage of only 170 V two tubes PL 81 in push-pull B will deliver a power output of 13.5 W and at a V_b of 200 V the output is 20 W.

APPLICATION

In the line output circuit of a modern television receiver the line output pentode PL 81 will normally be used in conjunction with a booster diode type PY 80. The booster diode circuit is a device whereby the energy in the line output circuit (transformer and deflection coils), which would otherwise be dissipated uselessly in damping resistors during the flyback period, is recovered and used for boosting the H.T. supply voltage. This is particularly advantageous in equipment employing the DC/AC (transformerless) technique where the available H.T. supply voltage rarely exceeds 170 to 200 V.

The H.T. boost therefore permits operation of the line output tube at a higher anode supply voltage with corresponding reduction in anode current. In addition, it makes available a higher H.T. supply voltage for such purposes as the first-anode voltage of the picture tube and the anode voltage of the frame output stage. Another advantage connected with the use of a booster circuit consists in the fact that the voltage across the line output transformer is kept constant during the scan by the booster diode, which results in a more linear scan.

The E.H.T. voltage for the second anode of the picture tube can be obtained in many different ways, but the most convenient and efficient method used nowadays is to rectify the flyback pulses occurring at the anode of the line output tube by means of a special diode, the EY 51.

It is thus seen that the applications of the PL 81 line output pentode, the PY 80 booster diode and the EY 51 E.H.T. rectifier are closely interrelated. For this reason, in the following pages the application notes relating to these tubes have been combined. A line output circuit is given using standard components and suitable for horizontal double-D deflection with a picture tube type MW 22-16 or MW 31-16 operating at 9 kV second-anode voltage. Since the operation of a line output stage greatly depends upon the amplitude and shape of the driving voltage used, also the circuiting of the line multivibrator and the synchronizing stage is given.

1. A CIRCUIT FOR HORIZONTAL DEFLECTION FOR THE PICTURE TUBES MW 22-16 OR MW 31-16 OPERATING AT 9 kV SECOND-ANODE VOLTAGE

A complete circuit diagram is given in fig. 98. In this circuit the functions of the tubes are as follows:

ECH 42	synchronizer
ECL 80	line multivibrator
PL 81	line output
PY 80	booster
EY 51	E.H.T. rectifier

The operation of the line output circuit can best be understood by assuming a linear saw-tooth current to be present in the deflection coils; see fig. 99. For double-D deflection a current of about 900 mA peak-to-peak is required in the line coils incorporated in the deflection and focusing unit type 10914. The self-inductance of the line coils is 3.77 mH and the resistance 4.7 Ω , so that at a line frequency of 15,625 c/s (625 lines) a practically constant voltage v_s of about 60 V is developed

The voltage v_s across the secondary L_7 is stepped up by the booster winding L_8 to about 210 V during the scanning period. During this period the voltage between the points 4 and 6 is such that the anode of the booster diode PY 80 is positive with respect to the H.T. supply line V_{b1} . During the scan, therefore, the diode conducts, with the result that this voltage is effectively added to the H.T. supply. The voltage between the cathode of the PY 80 and earth is 410 V and this voltage is available for the anode supply of the PL 81, whilst it can also be used to feed the first anode of the picture tube and the anode of the frame output stage.

It should be realized that as a result of the large step-up the PY 80 is subjected to a very large peak inverse voltage. The tube, however, is adequately insulated to withstand this condition. It is, however, essential that a suitable insulation is provided in the tube holder, the cylindrical centre shield of which should be removed, whilst it is also advisable to remove the contacts 1, 6 and 7, thus further reducing the risk of flashover and minimizing the leakage through the holder. In addition, certain

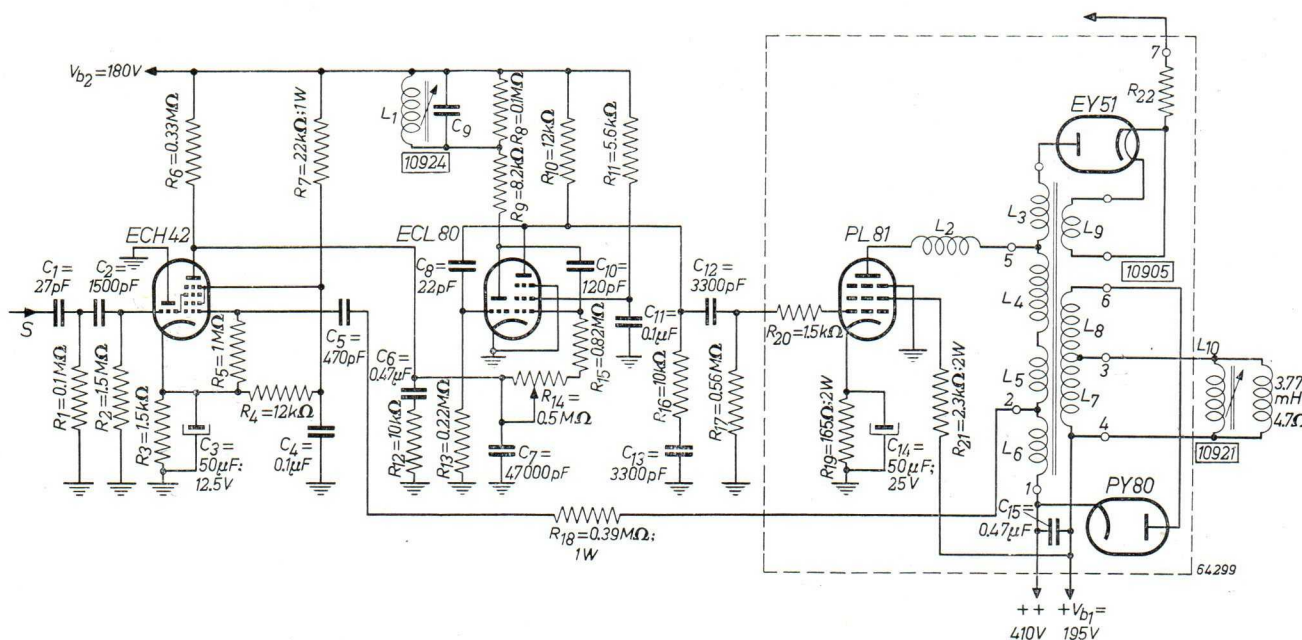


Fig. 98. Circuit for horizontal scanning of a double-D picture tube MW 22-16 or MW 31-16 operating at 9 kV second-anode voltage. The heater current of the ECH 42 is 0.23 A, so that a shunt resistor of 90 Ω is required across this heater.

during the scan. During the flyback a large sinusoidal voltage peak v_{pk} occurs. The line coils are shunted by a variable inductor type 10921, the self-inductance of which can be varied between 8.8 and 32 mH, for adjusting the current in the deflection coils.

types of holder may have to be mounted in a plate of insulating material.

The boosted H.T. supply allows of the turns ratio between the primary and the secondary of the transformer, that is between $L_4 L_5 L_6$ and L_7 , being much greater than is possible without H.T. boost,

and this results in a proportional reduction in the current consumption of the line output tube. With the transformer type 10905 the turns ratio is 6.35 to 1, and this means that the anode voltage of the PL 81 is about $410 - 6.35 \times 60 = 29 \text{ V}$

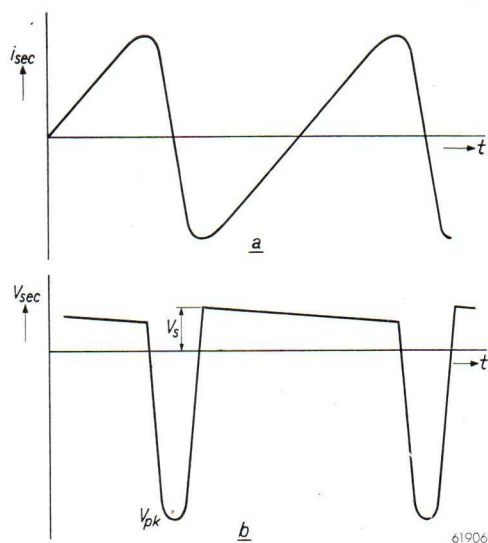


Fig. 99. The deflection current *a* and the voltage across the secondary of the line-output transformer *b*.

during the scanning period. Owing to resistance in the transformer windings and in the deflection coils the voltage drop across the primary is not entirely constant; a voltage varying between 23 and 43 V during the scan has actually been measured between anode and cathode of the PL 81. This low anode voltage might result in excessive screen-grid dissipation and for this reason a screen-grid resistor, R_{21} , has been included, whilst a cathode resistor has been used to prevent an overload of the anode upon failure of the driving voltage. The coil L_2 is incorporated in the anode connector cable for the PL 81 and serves to prevent the penetration of V.H.F. oscillations, which might occur at the beginning of the scan, into the line output circuit.

From fig. 99 it may be seen that when the anode of the PL 81 is negative with respect to the H.T. line during the scan, a very high positive voltage peak will occur during the flyback period. This can be rectified for obtaining the E.H.T. for the picture tube. For this purpose an overwind, L_3 , is provided on the transformer, and a rectifier type EY 51, the heater of which is fed from a separate winding L_9 , is used for the rectification. The EY 51 and the protecting resistor R_{22} are also incorporated in the line output transformer assembly. It will be noticed that no separate reservoir capacitor is used and it is therefore necessary to use a picture tube with

external conductive coating, such as the MW 22-16 or the MW 31-16, the capacitance of which between the internal and the external coating serves as reservoir capacitor.

In order to prevent interference with other circuits of the receiver — or with other television or broadcast receivers in the neighbourhood — it is advisable to enclose the line output circuit in a perforated screening cage. This screening is indicated in fig. 98 by the broken line.

The currents and voltages of the line output circuit are as follows:

Average values:

Supply voltage V_{b1}	195 V
Current from supply	78 mA
Anode current PL 81	65 mA
Screen-grid current PL 81	21 mA
Current drawn from 410 V terminal	13 mA
Voltage across R_{19}	15.6 V
Screen-grid voltage	143 V
E.H.T. (terminal 7)	

at 100 μA	9.8 kV
at 200 μA	9.2 kV

Peak values:

Anode current PL 81	140 mA
Screen-grid current PL 81	65 mA
Ripple at C_{15}	0.5 V
Driving voltage PL 81 peak-to-peak	130 V

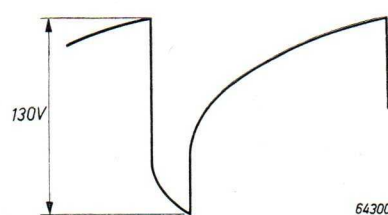


Fig. 100. Shape of the driving voltage for the PL 81.

The driving voltage of the PL 81 has a shape as represented in fig. 100, and this voltage is obtained with the multivibrator circuit equipped with ECL 80. During the flyback the line output pentode is cut off by the negative pulse, thereby preventing any additional damping, which otherwise would reduce the available peak voltage for rectification of the E.H.T.

A multivibrator with equal time constants in the grid circuits gives an output which is symmetrical, that is to say, the rectangular positive and negative blocks are of equal duration. This circuit is of the asymmetrical type, so that the duration of the

negative pulse is slightly longer than the flyback period of the line-output circuit, which is approximately 16% of the cycle.

In order to improve the frequency stability of the multivibrator a tuned circuit, L_1C_9 , has been taken up in the anode circuit of the triode section. This circuit is tuned to 15,625 c/s with the aid of an adjustable "Ferroxcube" core. The free-running frequency of the multivibrator can further be adjusted with R_{14} of 0.5 M Ω . L_1 and C_9 are available as one unit under the type number 10924. The nominal self-inductance of L_1 is 10 mH and the regulating range is from 6 to 14 mH. C_9 has a capacitance of 10,000 pF.

A novel circuit, which is very insensitive for interference, has been used for synchronizing the multivibrator. A sync. signal of 75 V peak-to-peak is applied to the grid of the triode section of the ECH 42, which is internally connected with the third grid of the hexode. This signal can be obtained with the sync. separator described in the application notes of the ECL 80. Grid rectification takes place in the triode section of the ECH 42, with the result that the third grid of the hexode section is at cathode potential only during the sync. pulses, whilst it is heavily negative during the intervals between the pulses. There is, therefore, a possibility of anode current flowing in the hexode, provided this section is not blocked by a negative voltage on the first grid. This grid is supplied with a part of the alternating voltage occurring across the primary of the line-output transformer; see terminal 2. The flyback pulses of the latter voltage are positively directed, so that grid rectification again takes place and the hexode section is consequently blocked during the intervals between the flyback pulses. Anode current can therefore only flow in the hexode section when both the sync. pulses and the line-flyback pulses occur simultaneously, whilst the average value of the anode current is determined by the degree of overlapping of these pulses. This is illustrated in fig. 101, in which the shape of the pulses is idealized.

As a result of the anode current pulses a voltage drop occurs in the anode resistor R_6 and the pulsatory character of this drop is smoothed out by C_6 and C_7 . The voltage at C_7 is applied via R_{14} and R_{15} to the control grid of the pentode section of the multivibrator and thus influences the frequency. In order to prevent hunting, it is necessary to include a resistor R_{12} in series with C_6 .

If it is assumed that the relative position of the

pulses represented in fig. 101 applies to the steady state condition, it will be clear that when the frequency of the multivibrator increases, that is when the pulses V_{g1H} move to the left relative to V_{gT} , the average anode current of the hexode section of the ECH 42 will also increase. This results in a

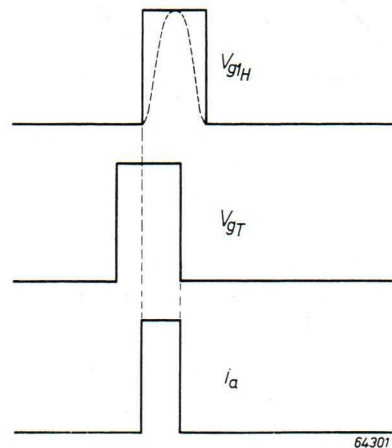


Fig. 101. The anode current i_a in the hexode section of the ECH 42 varies with the relative position of the pulses V_{gT} and V_{g1H} on the triode and hexode grids respectively.

higher voltage drop in R_6 , as a consequence of which the frequency deviation of the multivibrator is reduced. The same reasoning can be applied for the case where the frequency of the multivibrator decreases. In this case the anode current of the ECH 42 will decrease, resulting in a lower voltage drop in R_6 , which also tends to reduce the frequency deviation. It is thus seen that the multivibrator frequency is kept constant and "in phase" with the sync. pulses. In the absence of a signal, or if, as a result of an interference burst, the sync. signal fails for some time, it is necessary that the frequency of the multivibrator is maintained. For this reason the resistors R_3 , R_4 and R_7 have been so chosen that the anode currents with and without sync. signal are about equal.

On closer examination of fig. 101 it appears that the sync. pulse starts before the flyback pulse, so that the time available for the flyback of the line output circuit is reduced. This is not so serious as it might seem at first sight. Actually, the shape of the flyback pulse V_{g1H} is approximately sinusoidal, see dotted line, whilst the amplitude is very large. Therefore the hexode section of the ECH 42 can only draw current during the peak of the sinusoidal pulse, so that it is quite possible to adjust R_{14} in such a way that the beginnings of the flyback pulses and the sync. pulses coincide.

Finally, in the actual circuit differentiation has been applied for the sync. pulse, see C_1 and R_1 in fig. 98, as otherwise during the frame pulses the average anode current of the hexode would change, this resulting in curved sides at the top of the picture. It is therefore seen that fig. 101 only serves to explain the principal behaviour of the synchronizing system.

COMPONENTS FOR THE LINE OUTPUT CIRCUIT

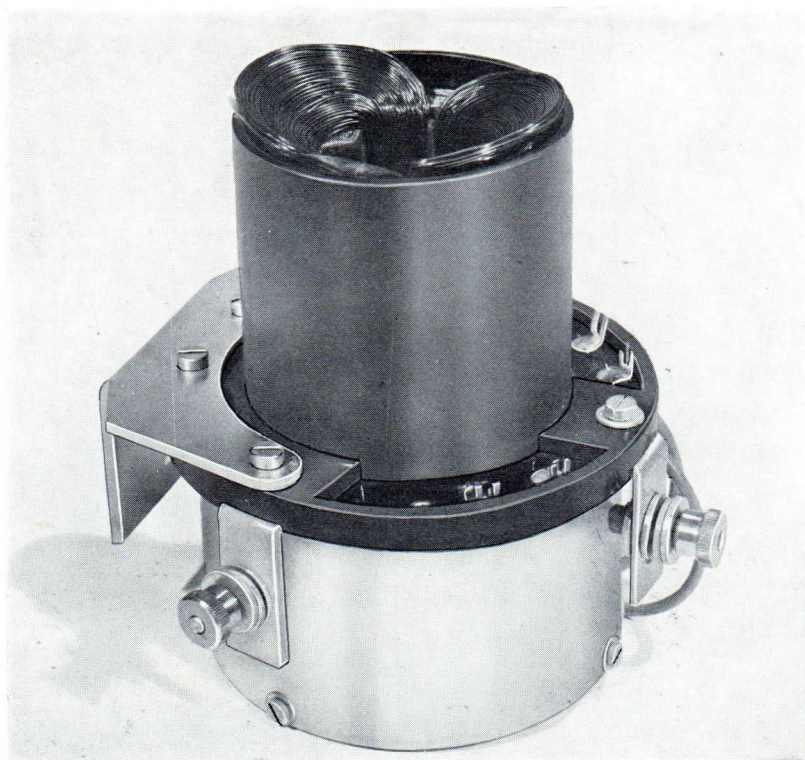
a. DEFLECTION AND FOCUSING UNIT TYPE 10914

It has already been indicated that the line output circuit is matched to the line coils incorporated in the deflection and focusing unit type 10914. This unit is intended for use in conjunction with the picture tubes MW 22-16 or MW 31-16. The data are

given below, whilst the connections and dimensions are represented in fig. 103.

Technical data

Line coils:	Inductance	3.77 mH
	Resistance	4.7 Ω
	Sensitivity (MW 22-16)	36.5 mA/cm at 9 kV
Frame coils:	Resistance	9.5 Ω
	Inductance	6.75 mH
	Sensitivity (MW 22-16)	31.5 mA/cm at 9 kV
Focus coil:	Resistance	4676 Ω
	Current	25 mA approx.
Maximum operating temperature 85 °C		
Terminals (see fig. 103)	2 — 6 line coils	
	1 — 5 frame coils	
	3 — 4 focus coil	
	5 screen.	



65378

Fig. 102. The deflection and focusing unit type 10914.

*AW focus 155-745 15.5 kV
665-815 66.5 kV*

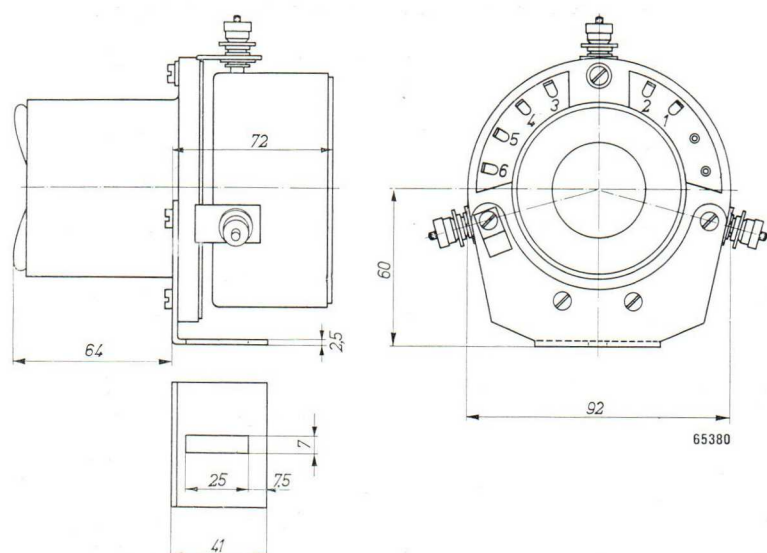


Fig. 103. Dimensions and connections of the deflection and focusing unit type 10914.

b. LINE OUTPUT TRANSFORMER TYPE 10905

This transformer has already been discussed in the preceding text in connection with the line output circuit of fig. 98. The data are given below and a dimensional drawing is represented in fig. 105.

Technical data

Type number 10905
 For use with line output pentode PL 81 and booster diode PY 80
 Line repetition rate 15,625 lines per second
 Inductance of primary

(terminals 1 and 5) 156 mH at 500 c/s with 30 mA D.C.

Turns ratio primary-secondary

(terminals 1 — 5 and 3 — 4)

6.35 : 1

Maximum operating temperature

85 °C

Terminals (see fig. 105) 1 cathode booster

2 tap on primary for sync.

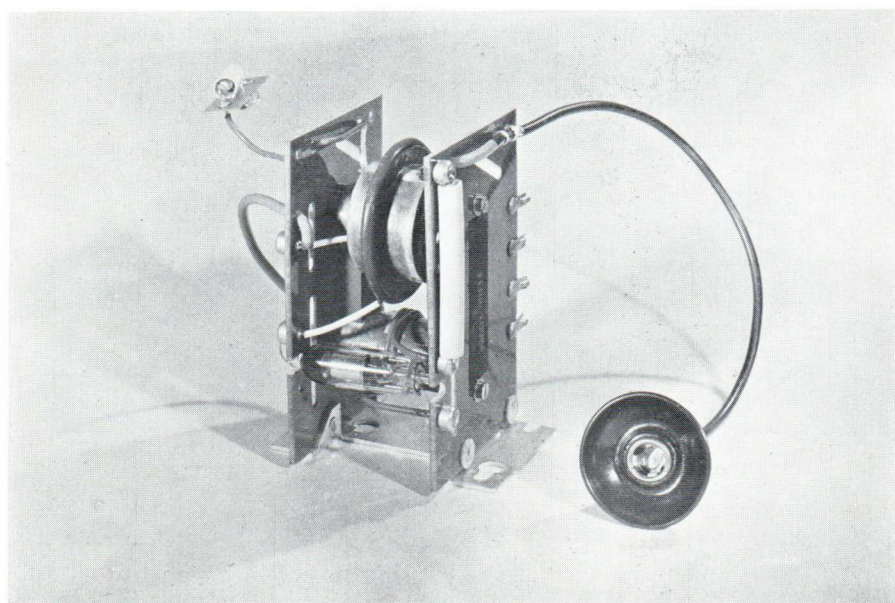
3 line coil

4 line coil and H.T. supply

5 anode output pentode

6 anode booster

7 + E.H.T.



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Fig. 104. The line-output transformer type 10905.

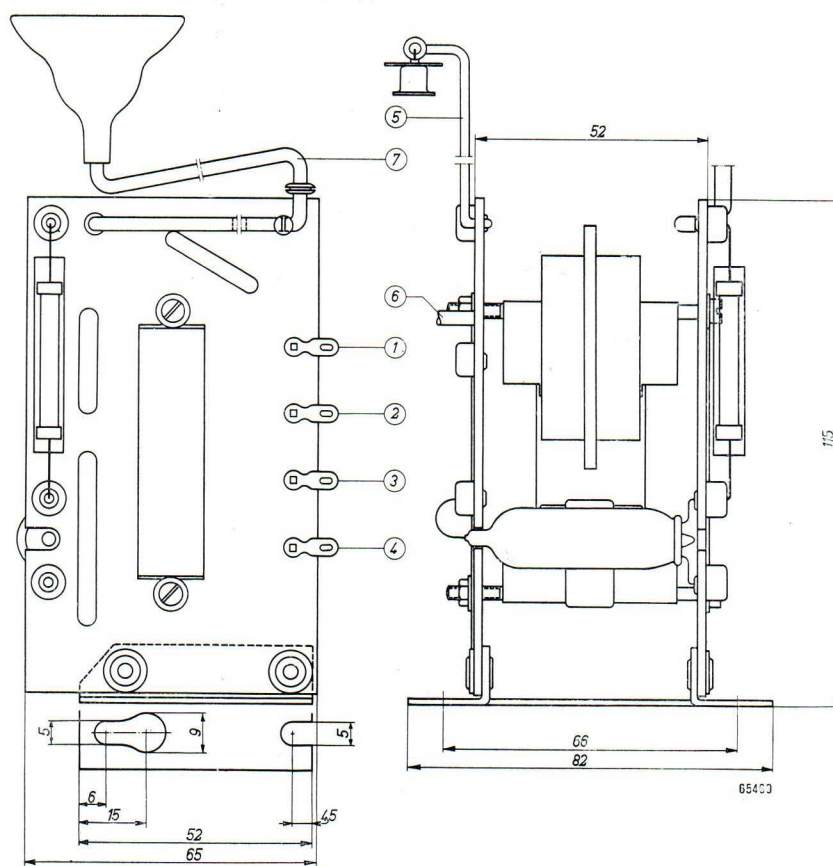


Fig. 105. Dimensions and connections of the line-output transformer type 10905.

*c. LINE AMPLITUDE CONTROL UNIT
TYPE 10921*

The line amplitude control unit type 10921 is a variable inductor with "Ferroxcube" core serving to adjust the width of the image on the screen of

the picture tube. It has been designed for use with the line-output transformer type 10905 and the line coils of the deflection and focusing unit type 10914. The data are given below and a dimensional drawing is represented in fig. 106.

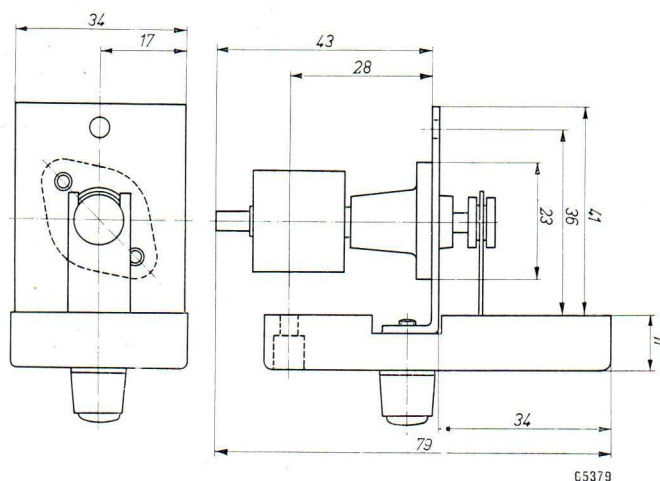


Fig. 106. Dimensional drawing of the line amplitude control unit type 10921.

Technical data

Type number	10921
Minimum inductance	8.8 mH
Maximum inductance	32 mH
Maximum operating temperature	85 °C

*d. TUNED CIRCUIT FOR LINE
MULTIVIBRATOR TYPE 10924*

This unit comprises an inductor with a nominal self-inductance of 10 mH and an adjustable

"*Ferroxcube*" core. A fixed mica tuning capacitor of 10,000 pF has been built in. The tuned circuit serves to improve the frequency stability of a multivibrator driving the line-output stage.

Technical data

Type number	10924
Minimum inductance	6 mH
Maximum inductance	14 mH
Fixed tuning capacitor	10,000 pF
Maximum operating temperature	85 °C

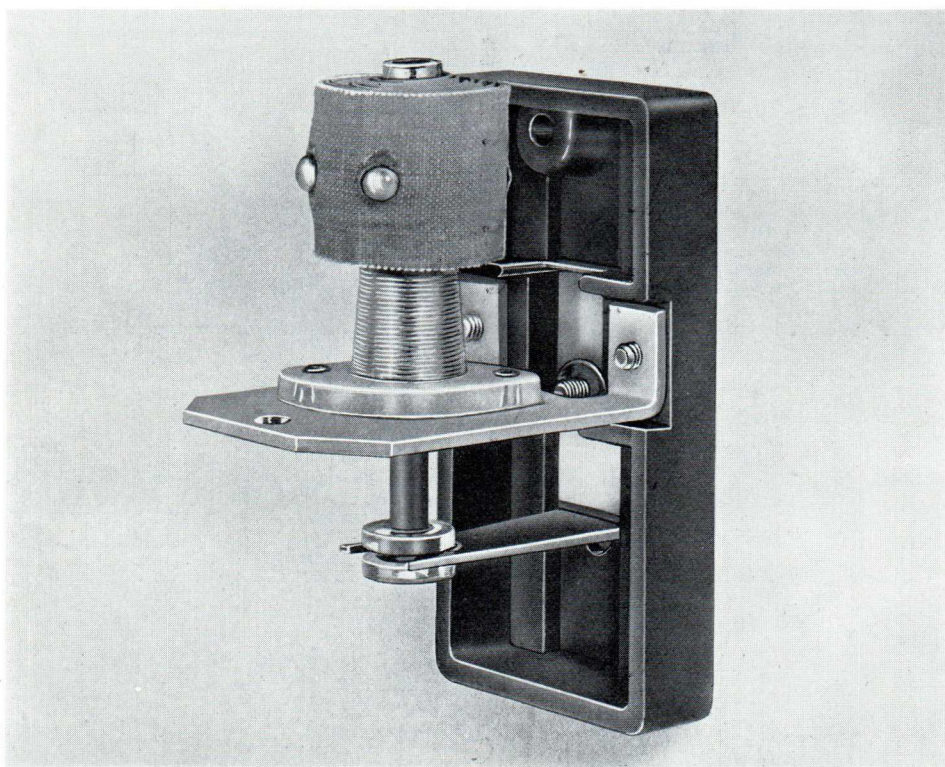


Fig. 107. The line amplitude control unit type 10921.

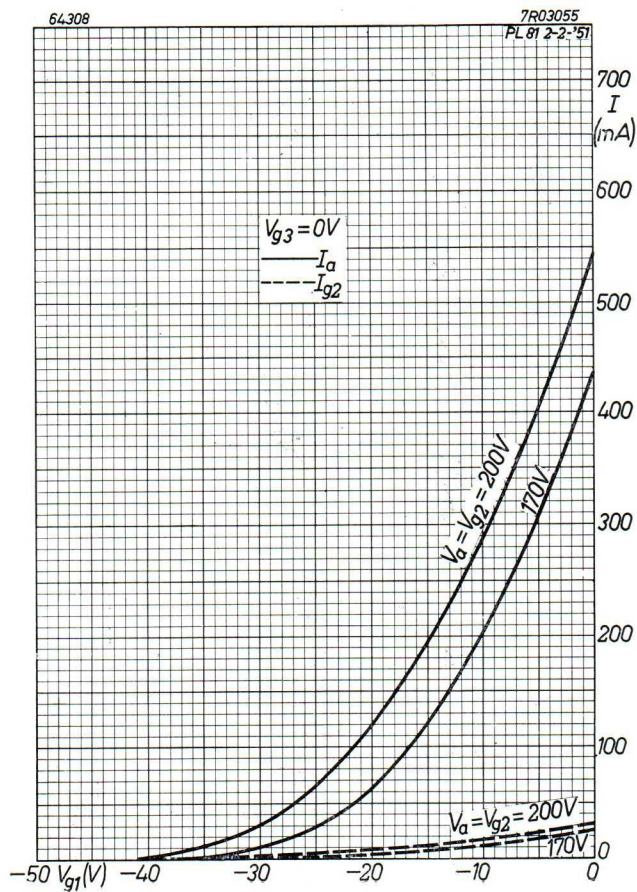


Fig. 109.
Anode current and screen-grid current plotted against control-grid voltage, with anode and screen-grid voltage as parameter.

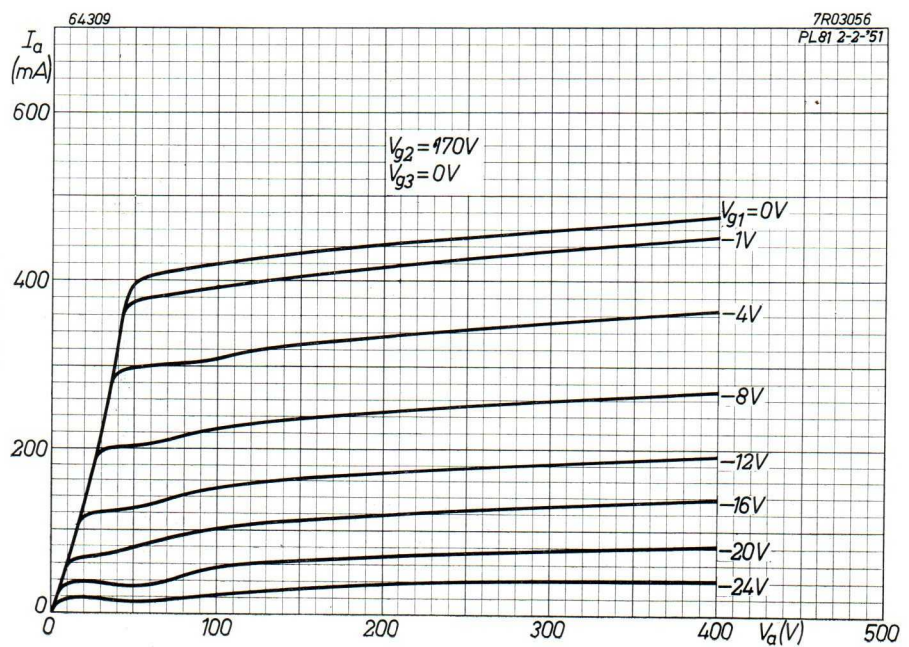


Fig. 110.
Anode current plotted against anode voltage for screen-grid voltage of 170 V, with control-grid voltage as parameter.

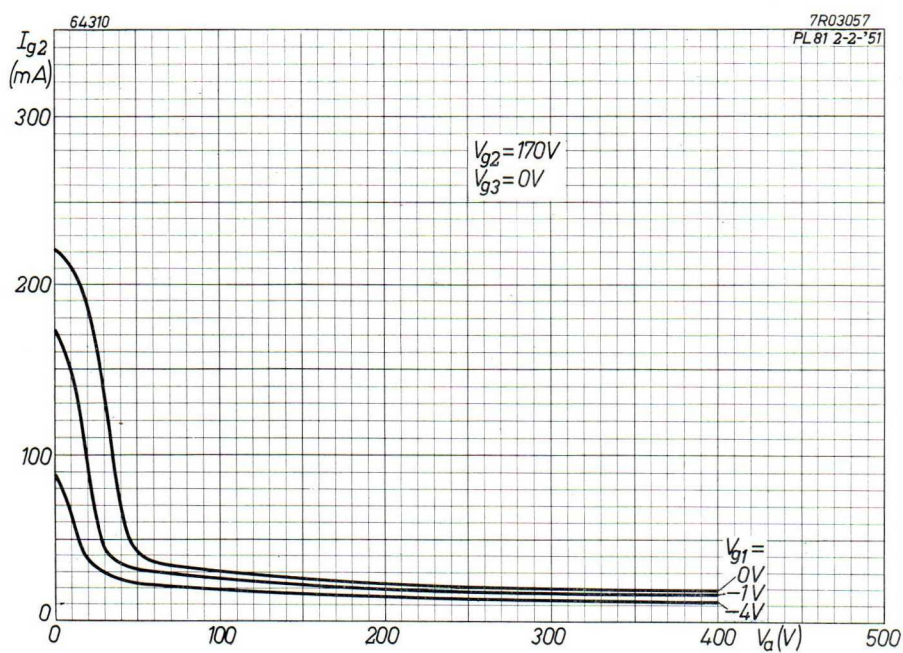


Fig. 111.

Screen-grid current plotted against anode voltage for screen-grid voltage of 170 V, with control-grid voltage as parameter.

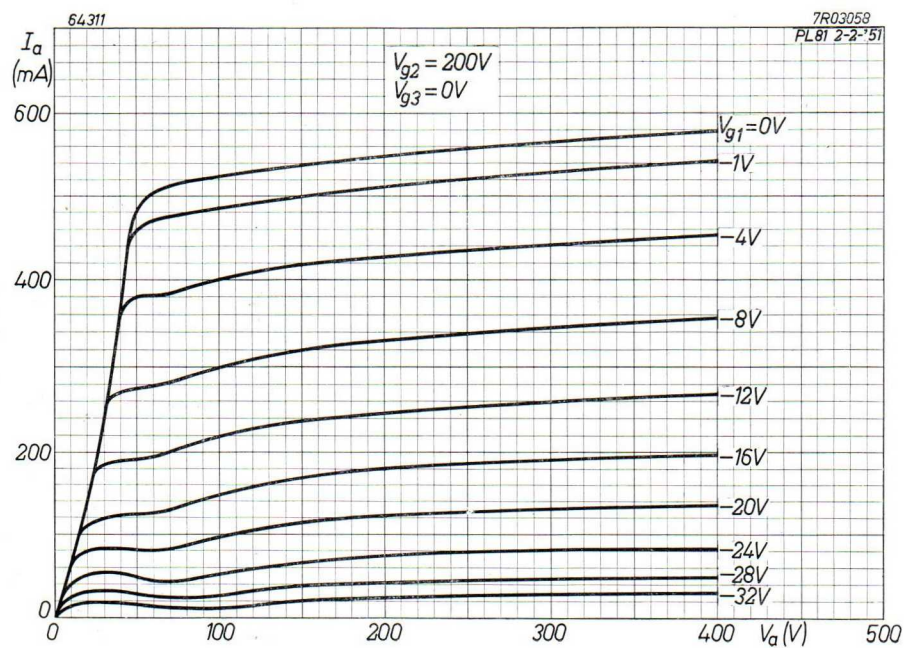


Fig. 112.

Anode current plotted against anode voltage for screen-grid voltage of 200 V, with control-grid voltage as parameter.

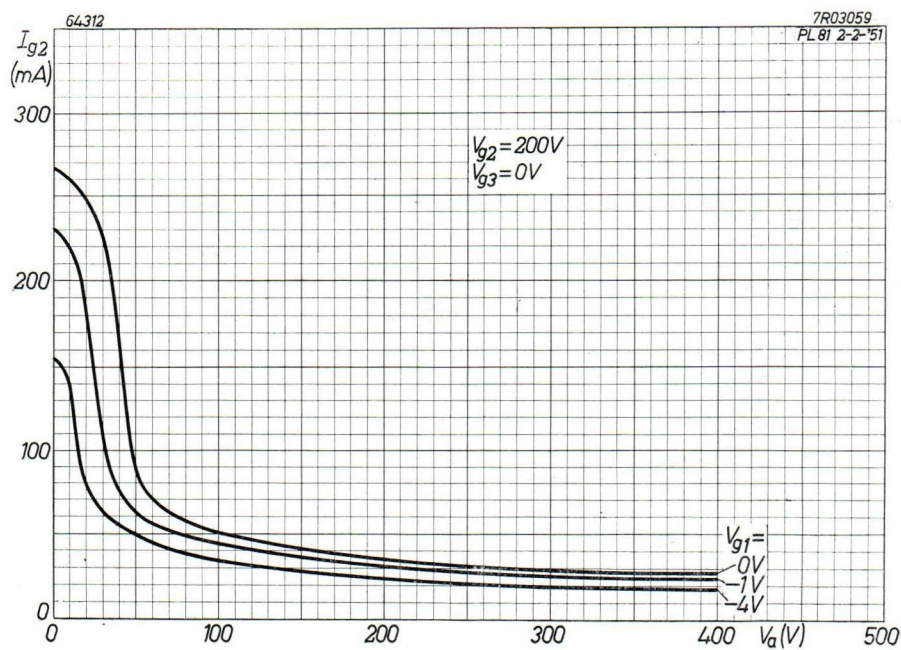
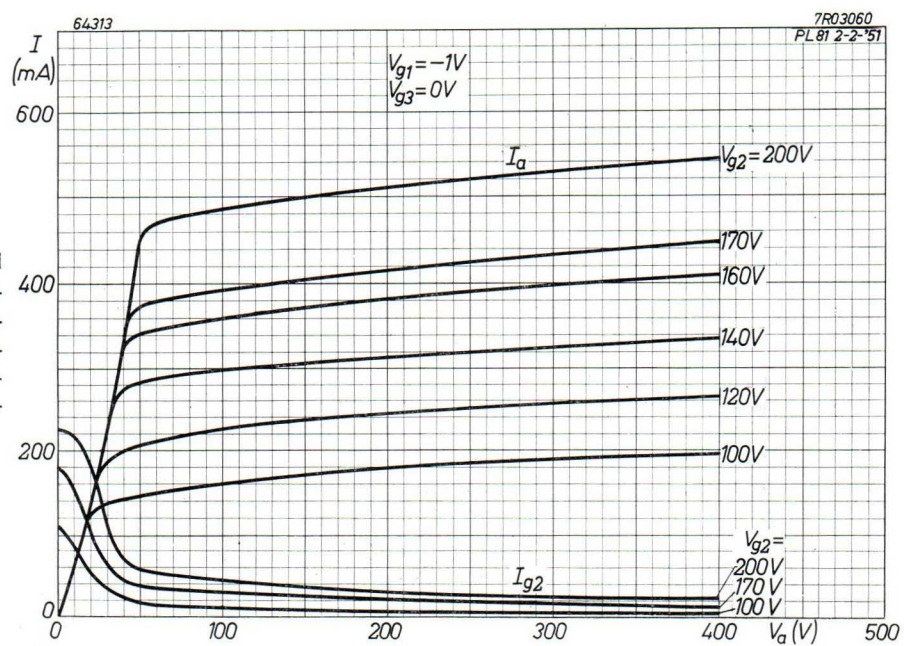


Fig. 113.
Screen-grid current plotted against anode voltage for screen-grid voltage of 200 V, with control-grid voltage as parameter.

Fig. 114.
Anode current and screen-grid current plotted against anode voltage for control-grid voltage of $-1V$, with screen-grid voltage as parameter.



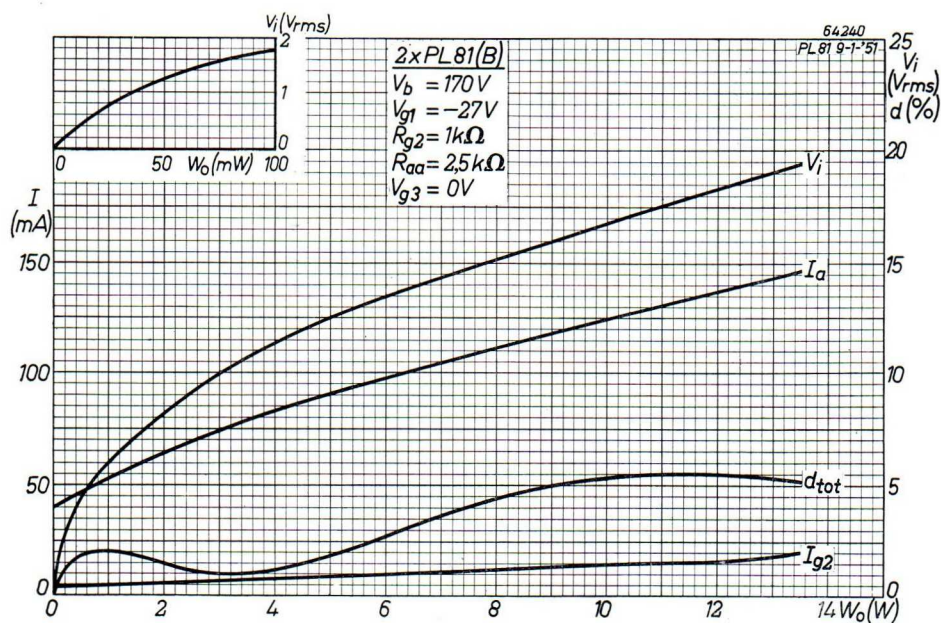
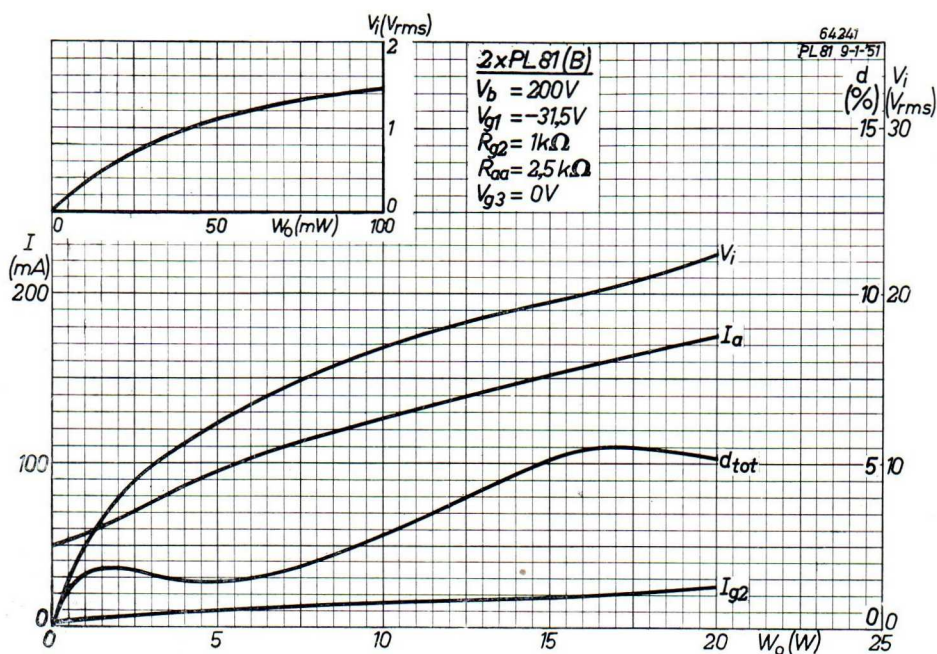


Fig. 115.

Anode current, screen-grid current, input voltage and total distortion plotted against output power of two tubes PL 81 in push-pull B for anode and screen-grid supply voltage of 170 V.

Fig. 116.

Anode current, screen-grid current, input voltage and total distortion plotted against output power of two tubes PL 81 in push-pull B for anode and screen-grid supply voltage of 200 V.



BOOSTER DIODE PY 80

DESCRIPTION

The PY 80 is an all-glass diode on Noval base, intended for the use as a booster diode. In a circuit designed to recover energy stored in the line output transformer and the deflection coils, which would otherwise be uselessly dissipated in damping resistors, the PY 80 can be employed for boosting the H.T. voltage for the line output tube, thereby considerably enhancing the efficiency of that stage.

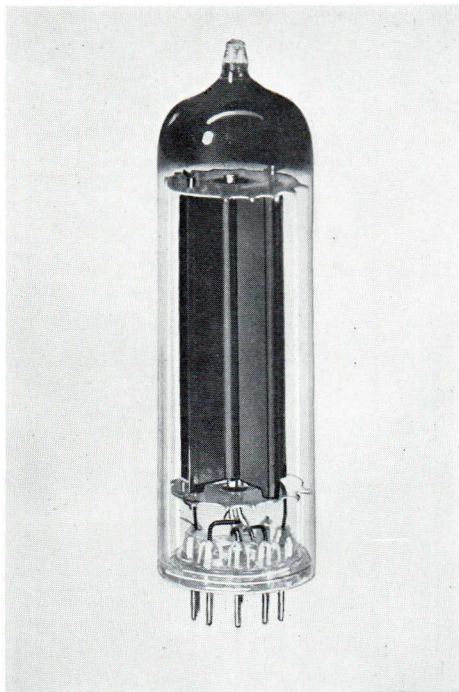


Fig. 117. The booster diode PY 80 (actual size).

TECHNICAL DATA

HEATER DATA

Heating: indirect by A.C. or D.C.;
series supply

Heater voltage V_f 19 V

Heater current I_f 0.3 A

BASE CONNECTIONS AND DIMENSIONS (in mm)

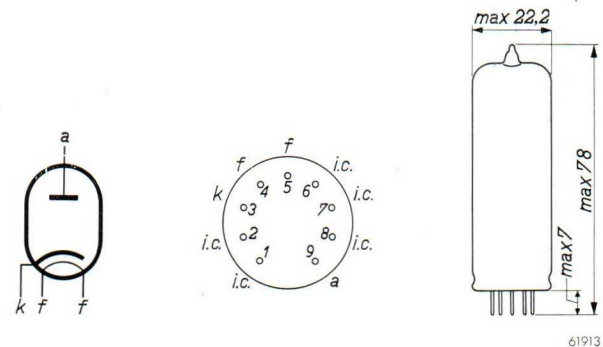


Fig. 118.

Mounting position: any

Capacitance

between anode and cathode C_a 5.5 pF

LIMITING VALUES

Peak inverse anode voltage $V_{a \text{ invp}}$ max. 4 kV¹⁾

Peak anode current I_{ap} max. 400 mA

Average anode current I_a max. 180 mA

Voltage between heater
and cathode V_{kfp} max. 650 V²⁾

Heater voltage during the
warming-up period max. 28.5 V

¹⁾ Maximum pulse duration 18% of one cycle, with a maximum of 18 μ sec.

The contacts 1—6 and 8 of the tube holder may not be used to support the wiring. When the tube holder 5908/03 is used no special precautions need be taken provided the peak inverse voltage does not exceed 3 kV. Between 3 and 4 kV, however, the cylindrical centre shield and contact 6 should be removed from the holder, this then being mounted in a plate of insulating material of about 4 cm diameter.

²⁾ Maximum 160 V_{rms} alternating voltage + maximum 450 V direct voltage. Cathode positive with respect to the heater.

In the booster application the cathode of the PY 80 is normally at a high potential with respect to the heater. The heater-cathode insulation has therefore been made adequate to withstand a potential difference of 650 V. Furthermore, the anode insulation will withstand a peak inverse voltage of 4 kV, such as may occur during the flyback.

APPLICATION

See Application Notes PL 81.

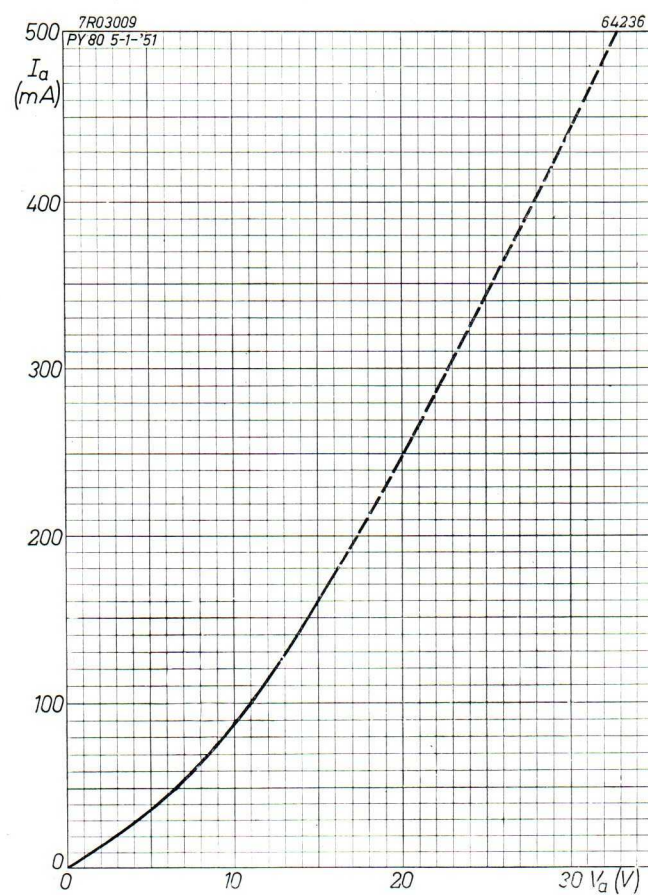


Fig. 119. Anode current plotted against anode voltage.

E.H.T. RECTIFIER EY 51

DESCRIPTION

The EY 51 is a miniature all-glass half-wave rectifier of the wire-in type. Its low heater wattage renders it particularly suitable for use in E.H.T. supply units deriving their input from the line flyback, the heater current being obtained from a subsidiary winding on the line output transformer.

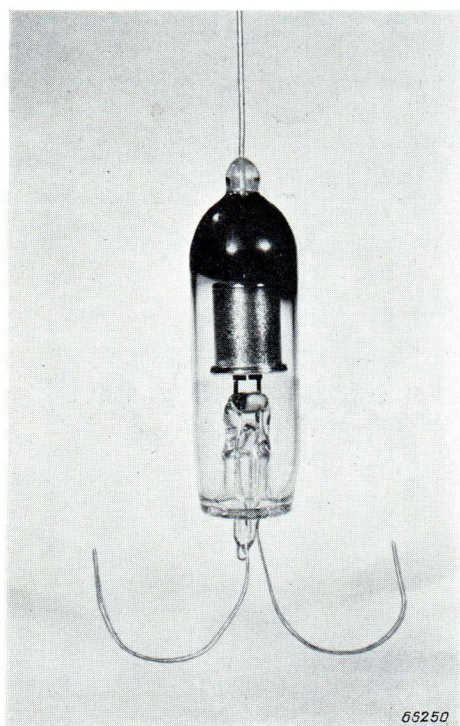


Fig. 120. The E.H.T. rectifier EY 51 (actual size).

APPLICATION

See Application Notes PL 81.

TECHNICAL DATA

HEATER DATA

Heating: indirect by A.C. or D.C.;
parallel supply

Heater voltage	V_f	6.3 V
Heater current	I_f	90 mA

CONNECTIONS AND DIMENSIONS (in mm)

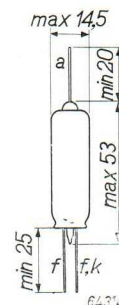


Fig. 121.

Mounting position any

Capacitance

between anode and cathode C_{ak} 0.8 pF

LIMITING VALUES

a. With sinusoidal input (mains frequency):

Input voltage	V_{tr}	max.	5 kV _{rms}
Average output current	I_o	max.	3 mA
Reservoir capacitance	C_{filt}	max.	0.1 μ F
Limiting resistance	R_t	min.	0.1 M Ω

b. With sinusoidal input (10—500 kc/s):

Peak inverse voltage	$V_{a invp}$	max.	17 kV
Average output current	I_o	max.	0.5 mA
Reservoir capacitance	C_{filt}	max.	0.01 μ F
Limiting resistance	R_t	min.	0.1 M Ω

c. With pulse input

Peak inverse voltage	$V_{a invp}$	max.	17 kV
Average output current	I_o	max.	0.2 mA
Peak cathode current	I_{kp}	max.	80 mA ¹⁾
Reservoir capacitance	C_{filt}	max.	5000 pF

¹⁾ For maximum pulse duration 1/2% of the time between two pulses, with a maximum of 5 μ sec.

Fig. 122.
Anode current plotted
against anode voltage.

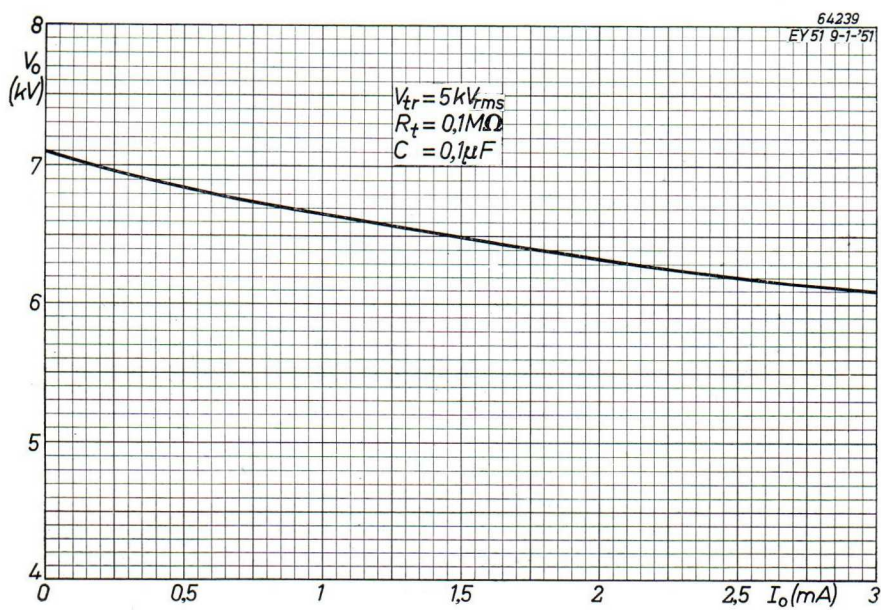
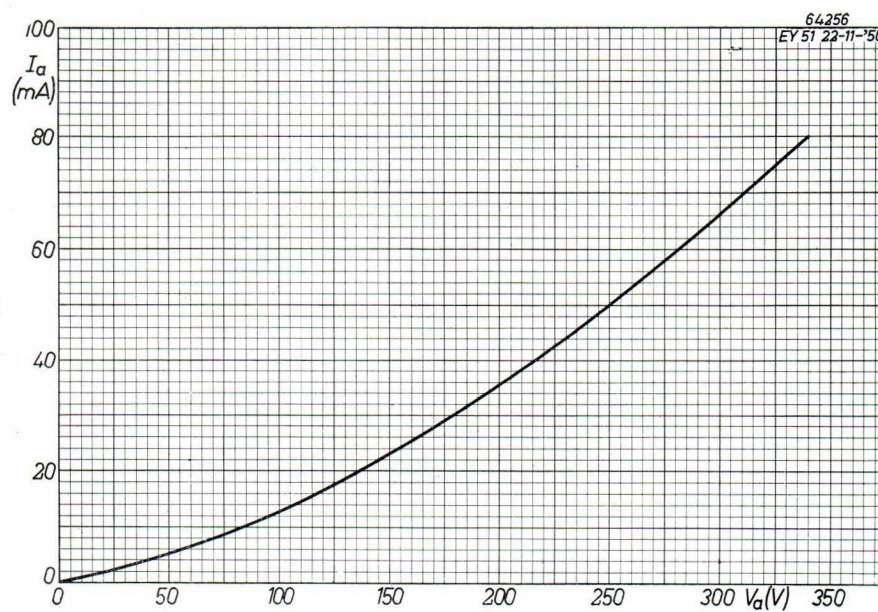


Fig. 123.
Regulation curve for si-
nusoidal input at 50 c/s.

MAINS RECTIFIER PY 82

DESCRIPTION

The PY 82 is an all-glass diode on Noval base, intended for use as a mains rectifier in receivers with series-connected heaters. Used as a half-wave

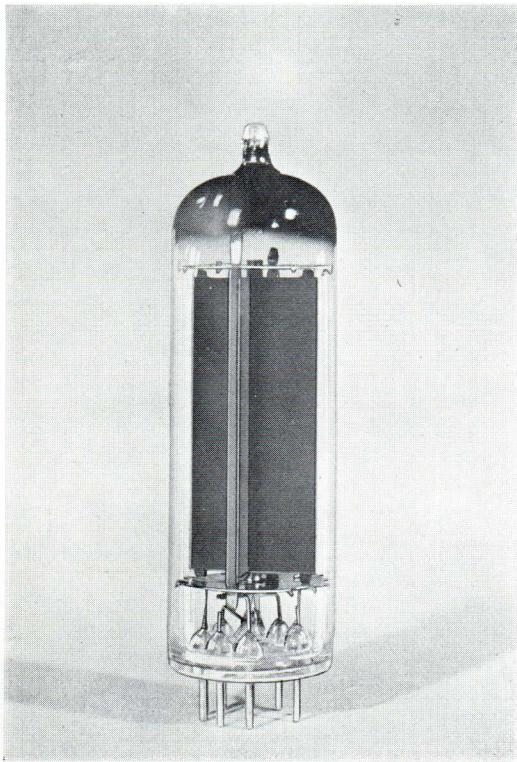


Fig. 124. The mains rectifier PY 82 (actual size).

rectifier the PY 82 will provide a rectified output of 180 mA at an input voltage of 250 V_{rms}, thus satisfying the requirements of receivers of a simple type. To supply the heavier H.T. drain for more elaborate television receivers it is necessary to employ two PY 82 rectifiers in parallel.

TECHNICAL DATA

HEATER DATA

Heating: indirect by A.C. or D.C.
series supply

Heater voltage	V_f	19 V
Heater current	I_f	0.3 A

TYPICAL OPERATING CONDITIONS

Input voltage	V_{tr}	250	240	220	200	127 V _{rms}
Reservoir	capacitance C_{filt}	60	60	60	60	60 μ F
Limiting resistor	R_t	125	105	65	30	0 Ω
Output current	I_o	180	180	180	180	180 mA
Output voltage	V_o	195	195	195	195	127 V

LIMITING VALUES

Input voltage	V_{tr}	max. 250 V _{rms}
Peak inverse voltage	$V_{a invp}$	max. 700 V
Output current	I_o	max. 180 mA
Voltage between heater and cathode	$V_{k/p}$	max. 550 V ¹⁾
Reservoir capacitance	C_{filt}	max. 60 μ F ²⁾
Heater voltage during the warming-up period		max. 28.5 V

Minimum permissible values
for the limiting resistor

V_{tr}	250	240	220	200	127 V _{rms}
R_t	100	80	40	30	0 Ω

BASE CONNECTIONS AND DIMENSIONS (in mm)

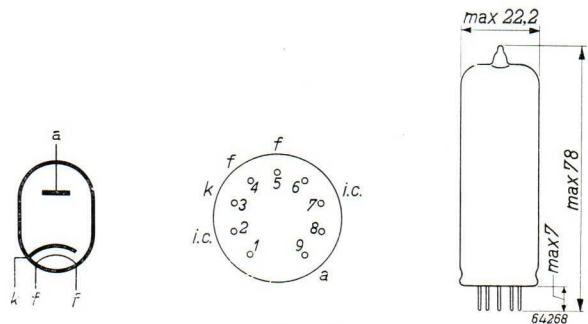


Fig. 125.

Mounting position: any

¹⁾ Maximum 220 V_{rms} alternating voltage + maximum 250 V direct voltage. Cathode positive with respect to heater.

²⁾ When two tubes PY 82 are connected in parallel the maximum reservoir capacitance is 100 μ F. In this case each anode must have the minimum limiting resistor specified above.

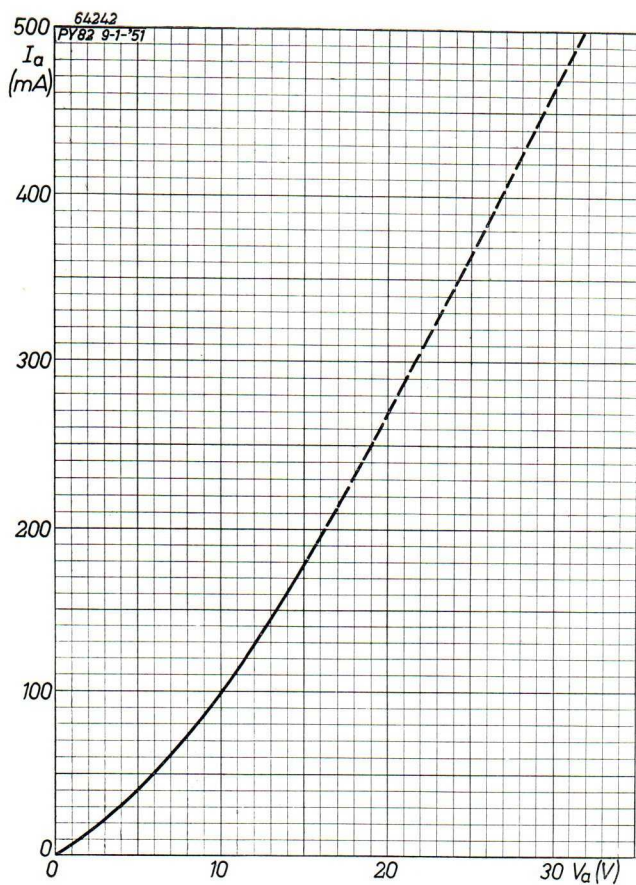
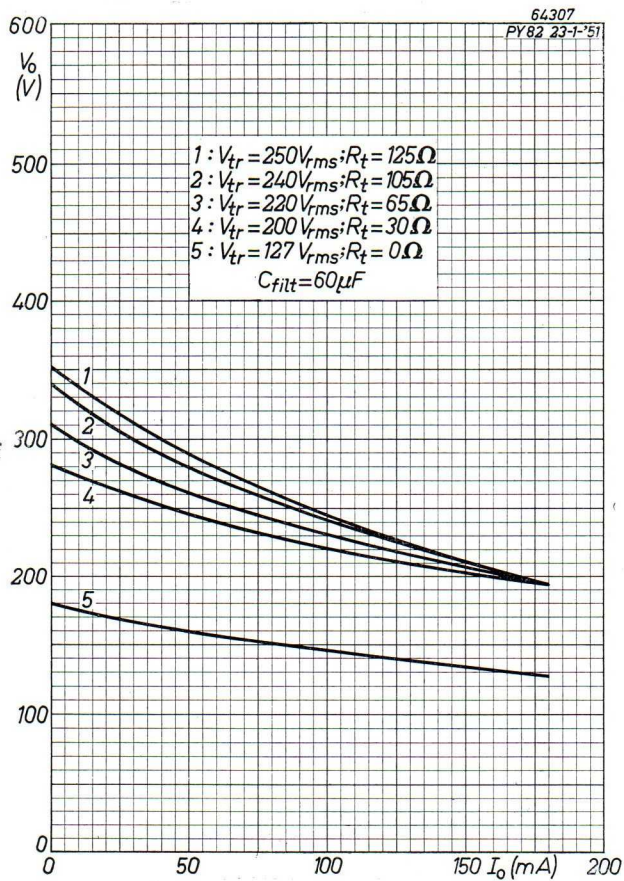


Fig. 126.
Anode current plotted
against anode voltage.

Fig. 127.
Regulation curves of
PY 82.



5000

1A