

# 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.

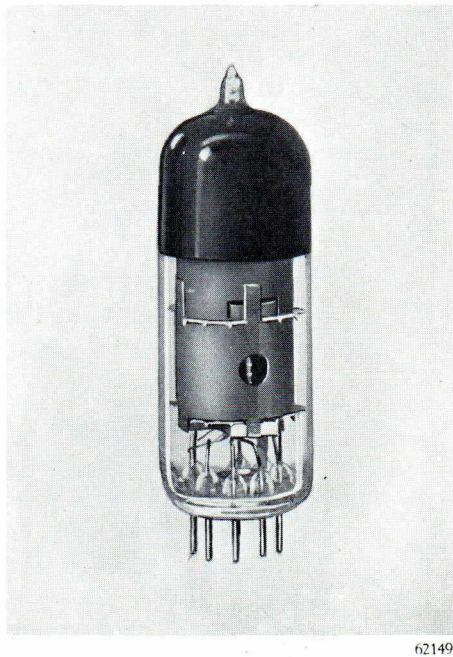


Fig. 16. The R.F. pentode EF 80 (actual size).

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.

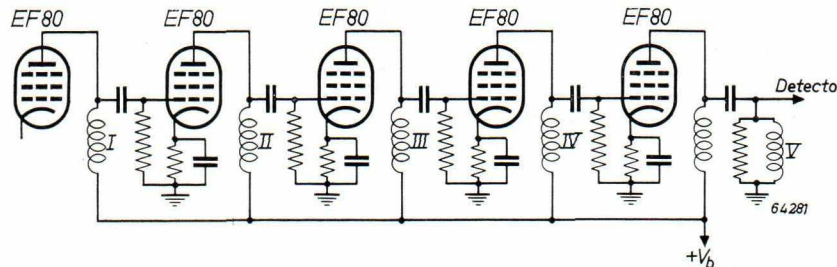


Fig. 17. Basic diagram of a four-stage stagger-tuned I.F. amplifier.

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 staggering<sup>1)</sup>. 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  |

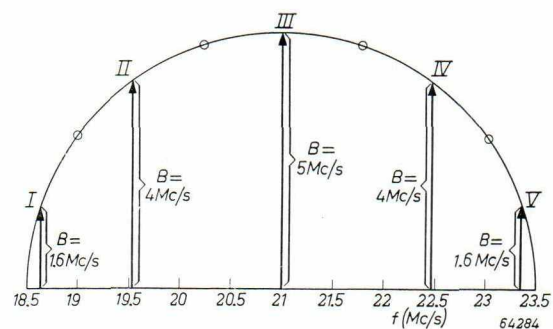


Fig. 18. Stagger diagram for a four-stage I.F. amplifier centred at 21 Mc/s.

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  |

<sup>1)</sup> 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

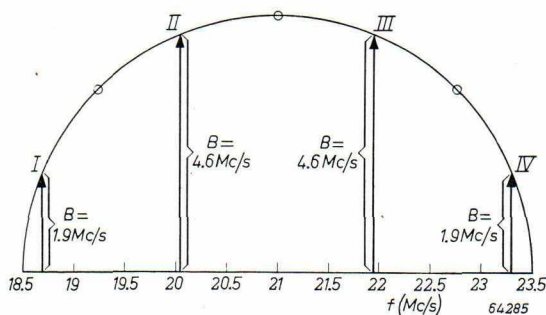


Fig. 19. Stagger diagram for a three-stage I.F. amplifier centred at 21 Mc/s.

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  $\Omega$ , a grid

leak of 8400  $\Omega$  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  $\Delta C_i$  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  $\Delta C_i$  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.

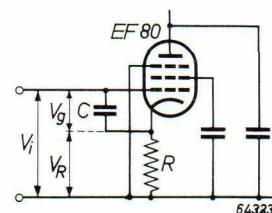


Fig. 20. Basic circuit for the compensation of input capacitance variation.

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 CHANGER<sup>1)</sup>

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.

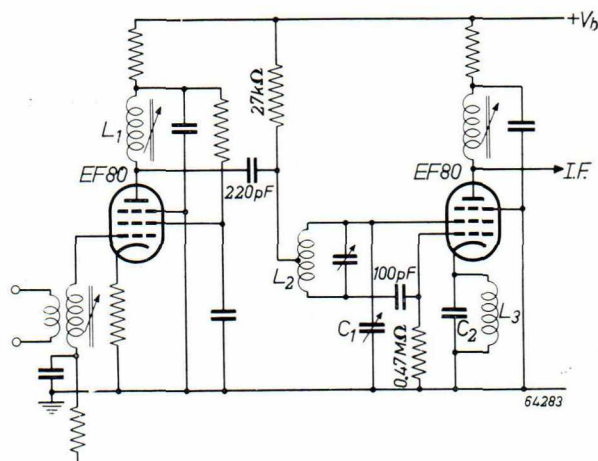


Fig. 21. Front end for a television receiver or an F.M. receiver with one EF 80 as R.F. amplifier and the other as self-oscillating frequency changer.

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.

<sup>1)</sup> 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.

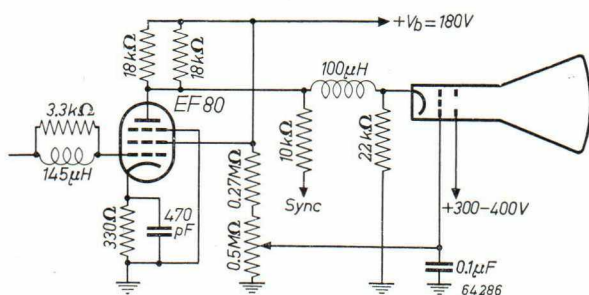


Fig. 22. The EF 80 as a video amplifier with cathode compensation.

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  $\mu$ H shunted by 3300  $\Omega$

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  $\mu$ 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 $\Omega$ , 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 $\Omega$  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     |

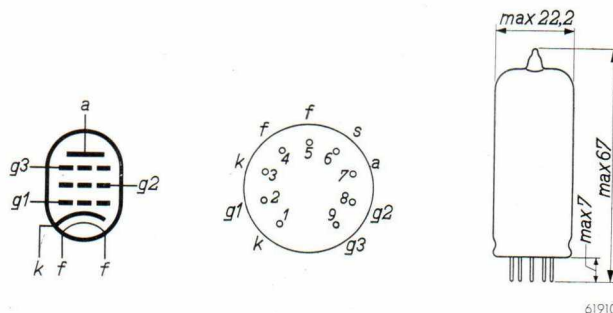
BASE CONNECTIONS AND  
DIMENSIONS (in mm)

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 $\Omega$ |
| Amplification factor<br>between screen grid<br>and control grid     | $\mu_{g2g1}$ | 50  | 50    | 50              |
| Equivalent noise<br>resistance                                      | $R_{eq}$     | 1   | 1.1   | 1.2 k $\Omega$  |
| Input damping at<br>50 Mc/s (socket pins 1<br>and 3 interconnected) | $r_{g1}$     | 10  | 12    | 15 k $\Omega$   |

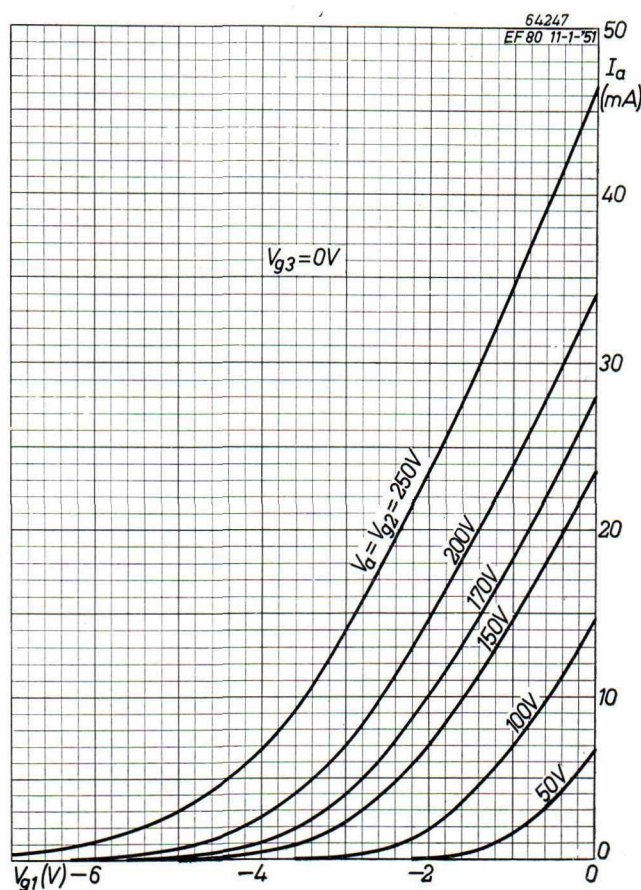


Fig. 24. Anode current plotted against control-grid voltage for anode and screen-grid voltages between 50 V and 250 V.

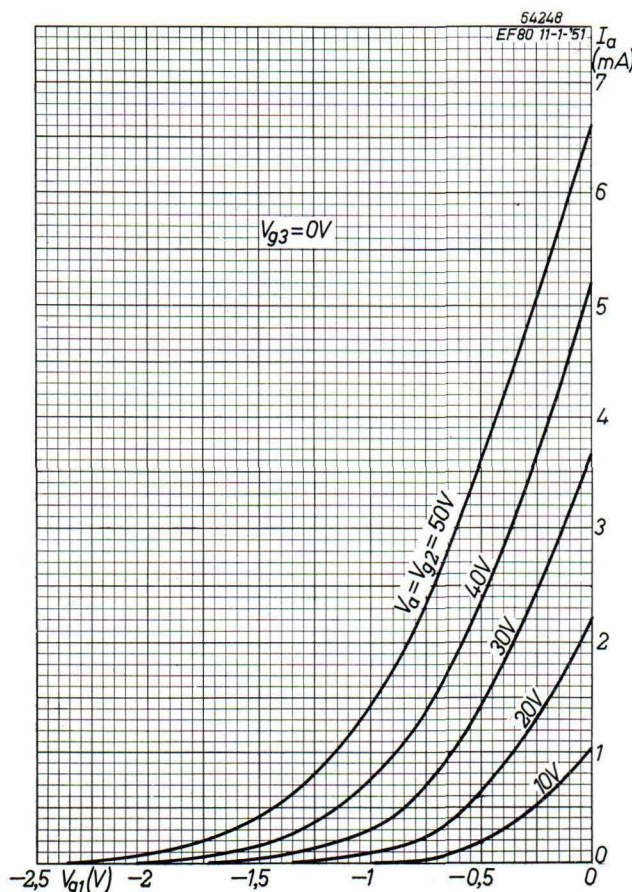


Fig. 25. Anode current plotted against control-grid voltage for anode and screen-grid voltages between 10 V and 50 V.

## 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 $\mu$ 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. | 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 $\Omega$   |
| with fixed bias                                                     | $R_{g1}$  | max. | 0.5 M $\Omega$ |
| External resistance between heater and cathode                      | $R_{kf}$  | max. | 20 k $\Omega$  |

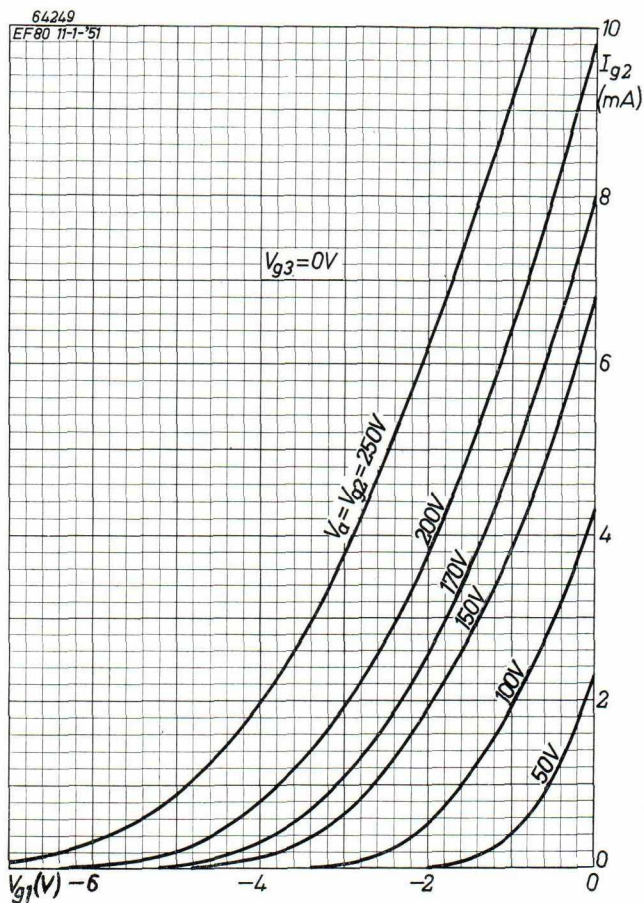
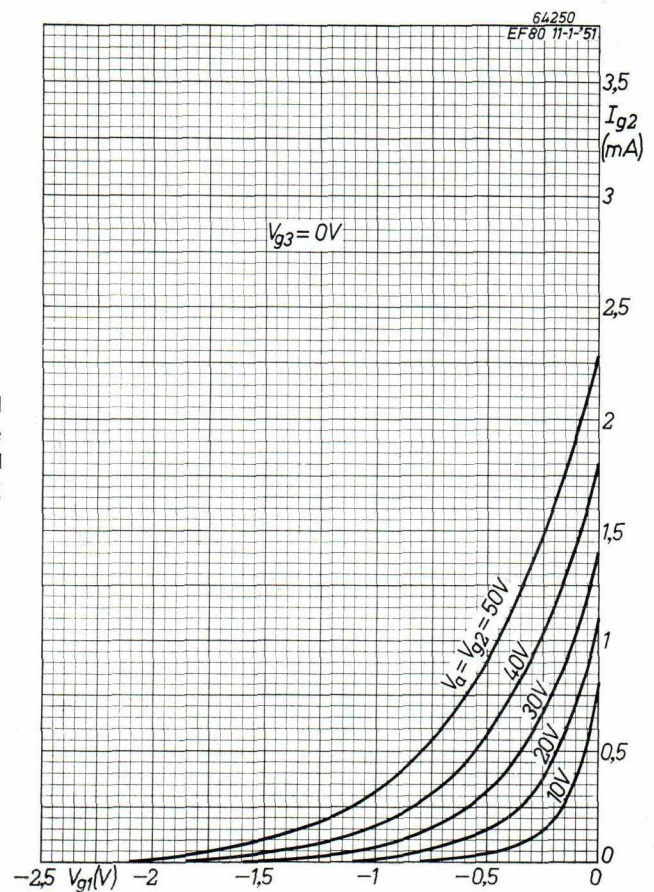


Fig. 26.  
Screen-grid current plotted  
against control-grid voltage  
for anode and screen-grid  
voltages between 50 V and  
250 V.

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



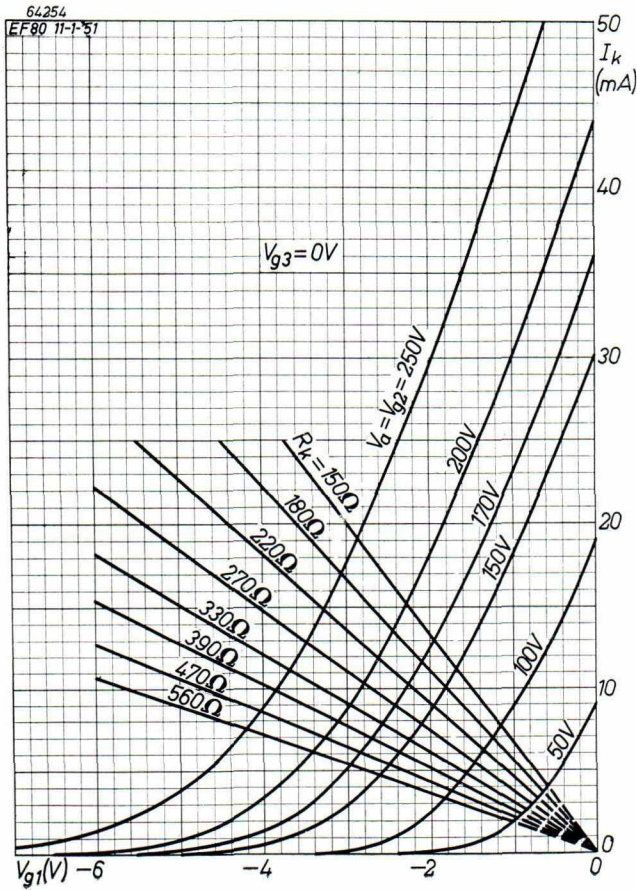


Fig. 28.  
Cathode current plotted  
against control-grid volt-  
age, with anode and screen-  
grid voltage as parameter.

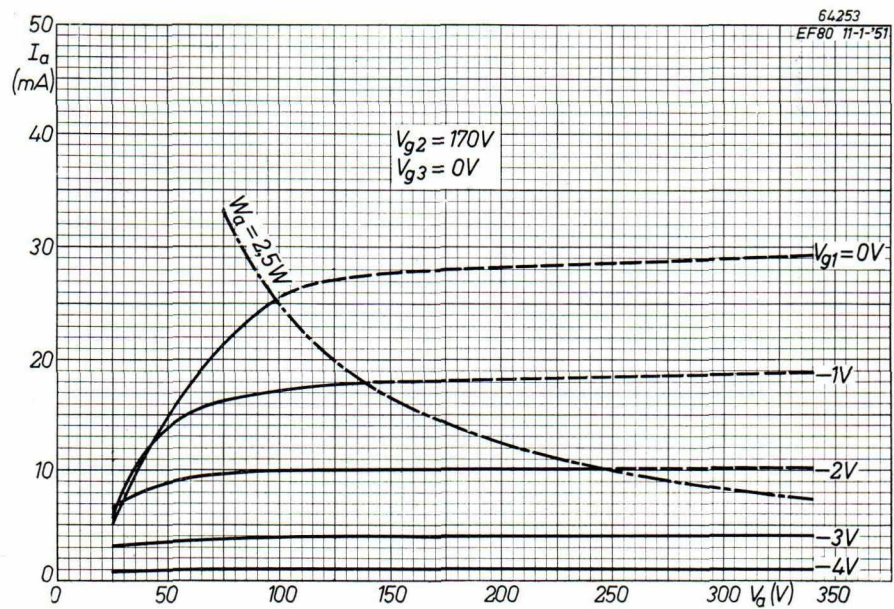


Fig. 29.  
Anode current plotted  
against anode voltage for  
screen-grid voltage of  
170 V, with control-grid  
voltage as parameter.

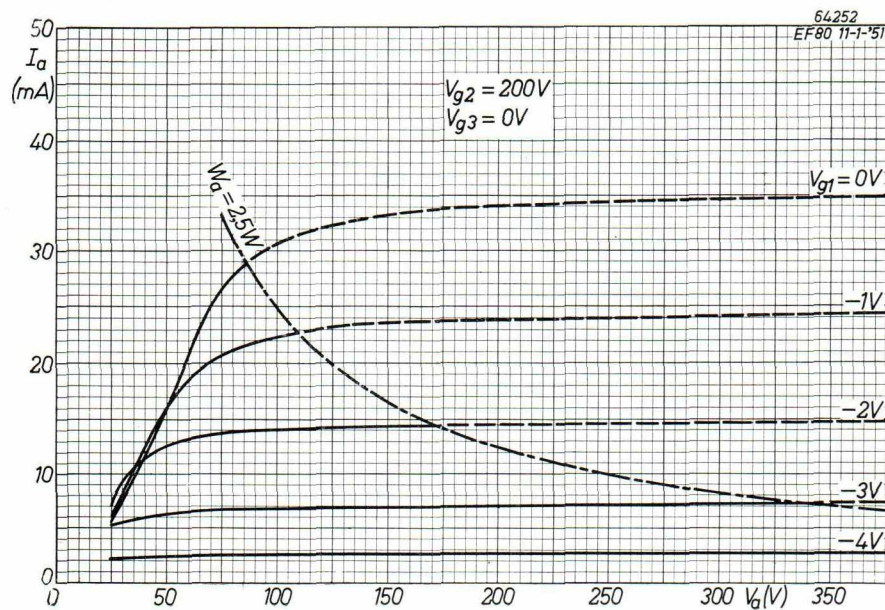


Fig. 30.

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

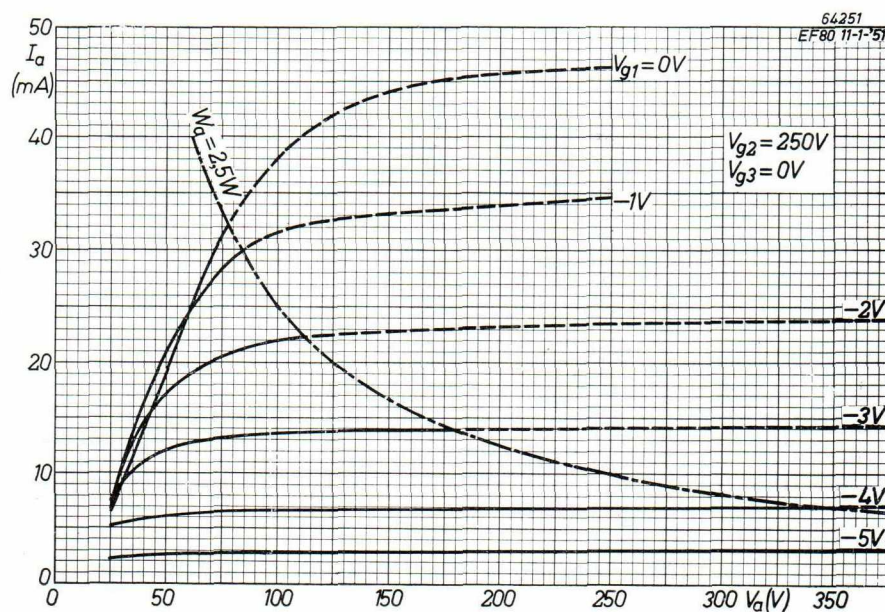


Fig. 31.

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

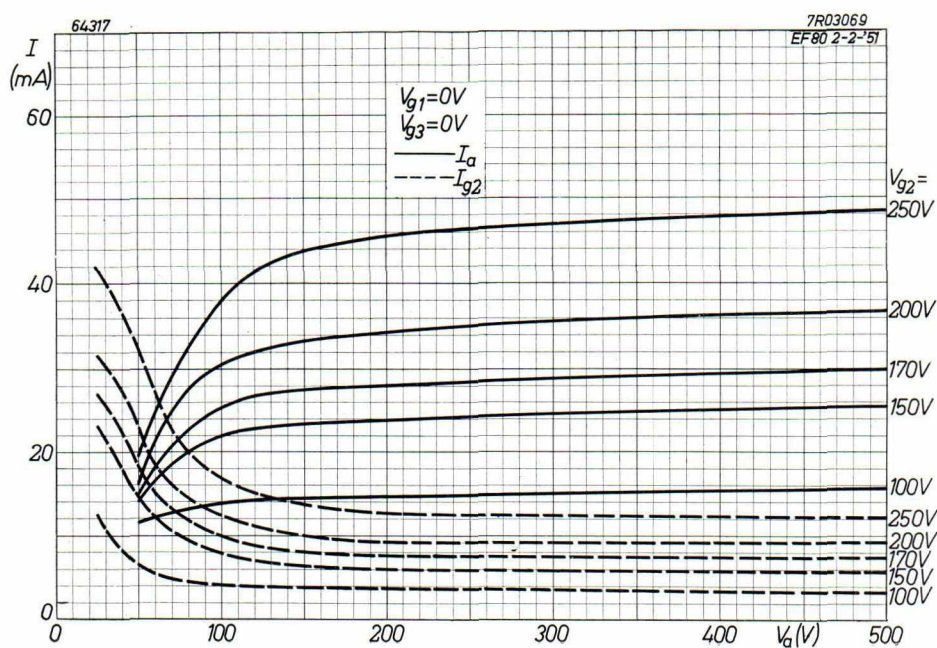


Fig. 32.  
Anode current and screen-grid current plotted against anode voltage for control-grid voltage of 0 V, with screen-grid voltage as parameter.

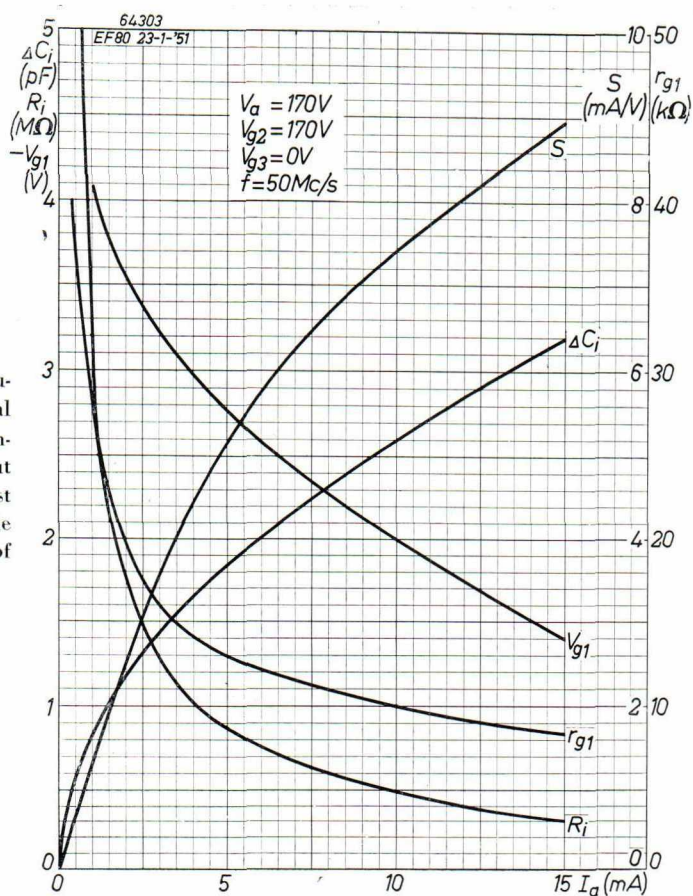


Fig. 33.  
Control-grid voltage, mutual conductance, internal resistance, input capacitance variation and input damping plotted against anode current for anode and screen-grid voltage of 170 V.

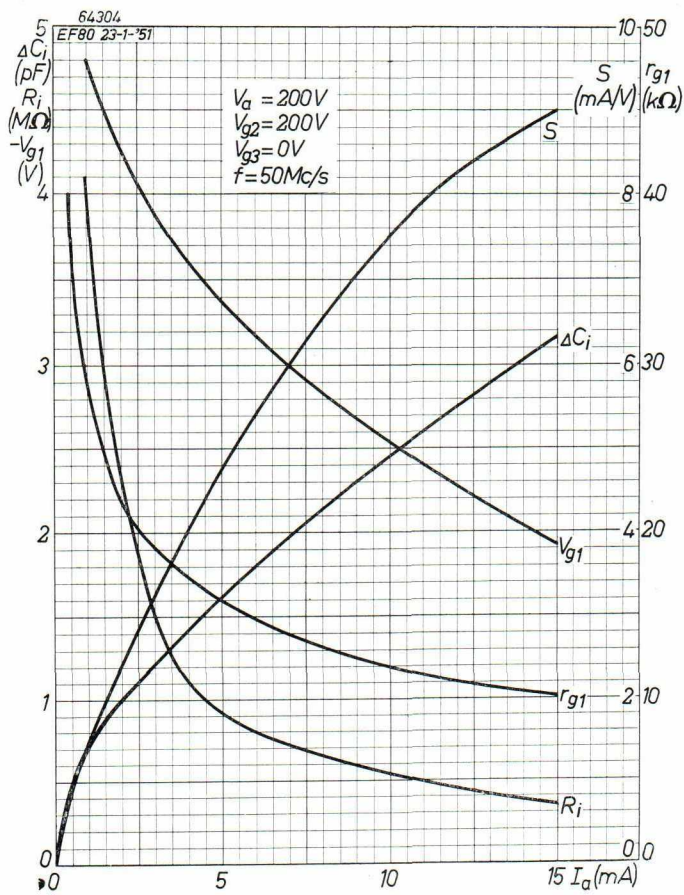
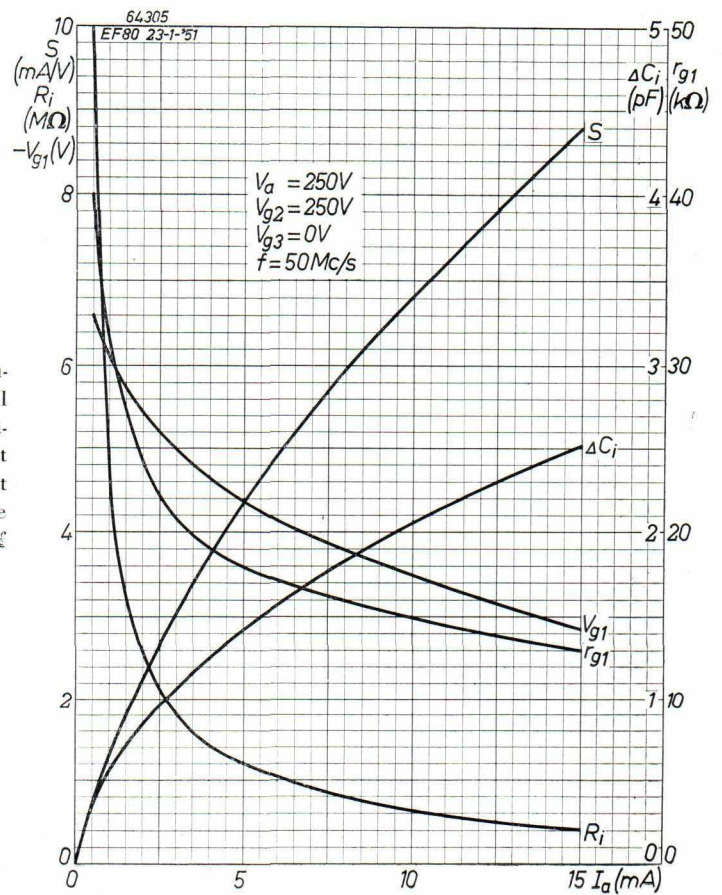


Fig. 34.  
Control-grid voltage, mutual conductance, internal resistance, input capacitance variation and input damping plotted against anode current for anode and screen-grid voltage of 200 V.

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.



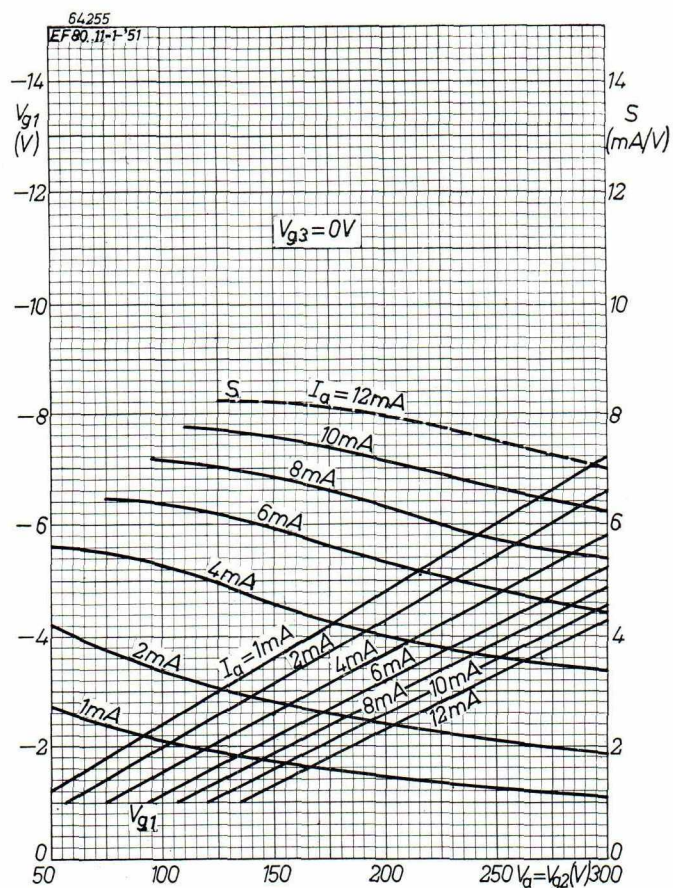


Fig. 36.

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

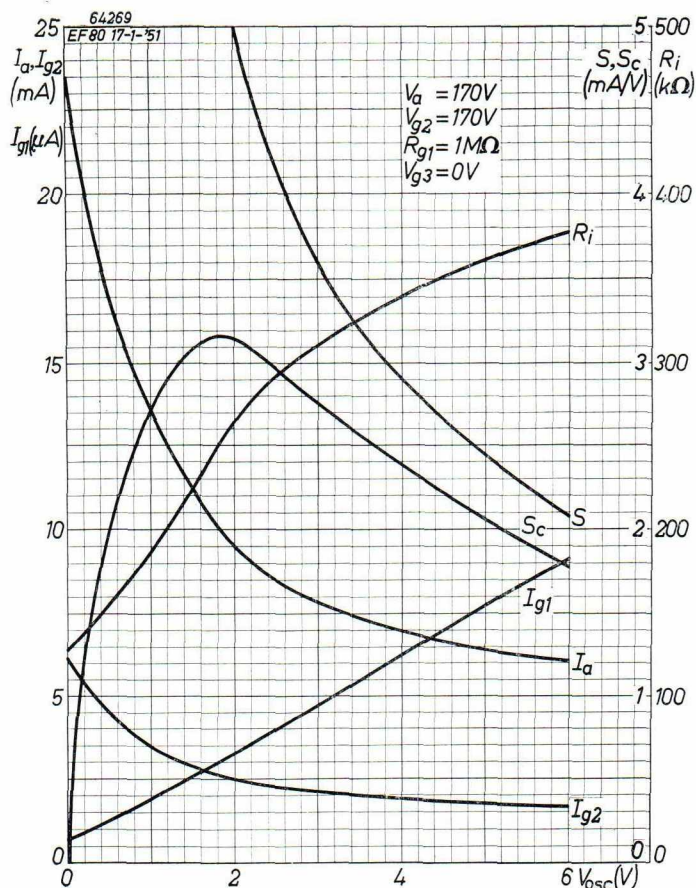


Fig. 37.

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

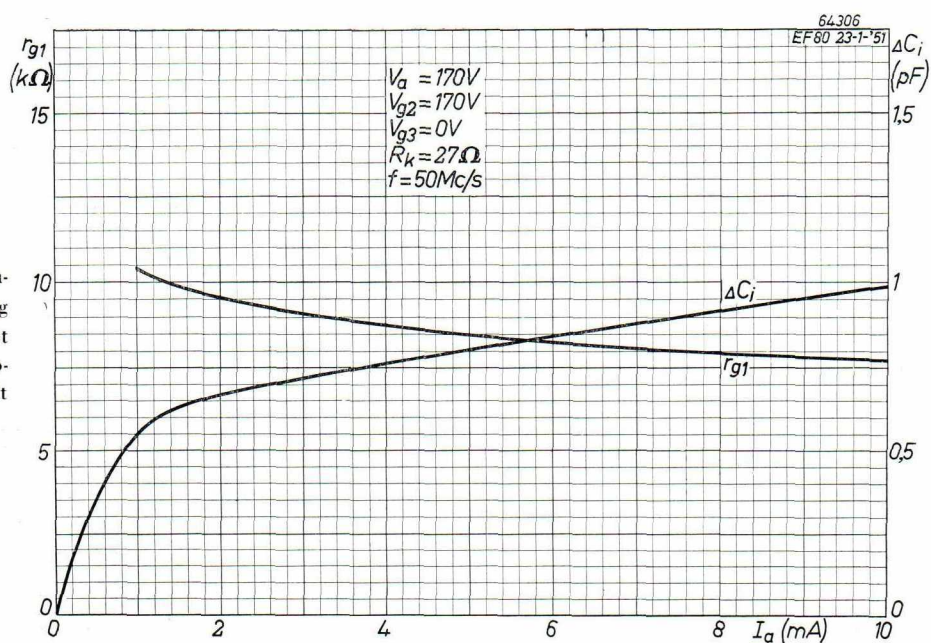


Fig. 38.

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