



**SIEMENS**

**Design Examples  
of Semiconductor  
Circuits**

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**1966 Edition**

# Design Examples of Semiconductor Circuits

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**The circuit designs and descriptions collected in this booklet are to demonstrate by examples the manifold possibilities of semiconductor component applications. To offer a good survey, cases of similar applications have been grouped in chapters.**

**No guarantee is given for the circuits as far as patent licenses are concerned.**

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# **1. Audio Frequency Amplifiers**

In the field of audio frequency amplifiers the complementary stages become more and more important. These are transformerless output stages which have complementary transistors, i.e. a pnp and a similar npn transistor. Now there are three of these transistor pairs available: the pair AC 127/AC 152 for a power output up to 1 W, the pair AC 176/AC 153 for 2.5 W and the pair AD 161/AD 162 for a power output up to 5 W, eventually also 10 W (effective output power).

For higher output power two pnp output stage transistors driven by a complementary pair have been successful. The required phase shift already takes place in the driver, and the final stage will be driven by two antiphase voltages.

Because of low cut-off frequencies, the conventional germanium alloy junction transistors often are insufficient for Hi-Fi output stages. A phase shift causes a higher load of the transistors at frequencies in the upper third of the audible range. This limits the attainable output power at higher frequencies and also increases the distortion factor.

An essential improvement can be obtained by diffused transistors, e.g., the new types AD 166 and AD 167. As a consequence of the higher cut-off frequency of these transistors, the above mentioned difficulties occur at frequencies beyond the audible range. In order to avoid an overload of these transistors, the transmitted power also has to be limited at higher frequencies, e.g. 30 kHz.

## **1.1 Increase of the Power Dissipation at Higher Frequencies**

At higher frequencies the rise of the power dissipation described above depends on the kind of load, the base-emitter capacitance of the transistor and the circuit itself.



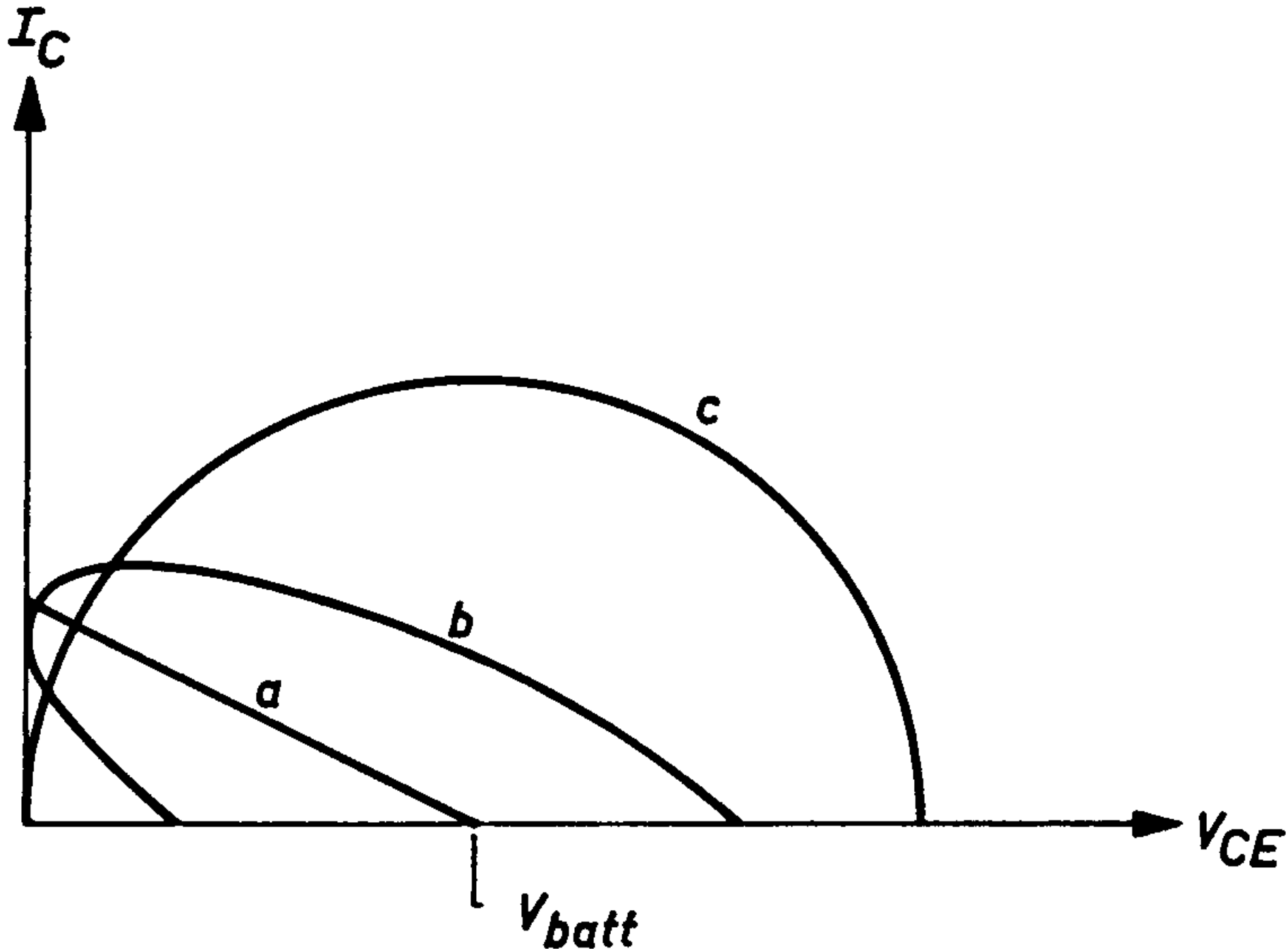


Fig. 1.1

Figure 1.1 shows the load line of an AF amplifier with an ohmic load (a), with a complex load (b) and with an imaginary load (c). Case (a) will probably never occur in practice, while case (b) happens often in AF amplifiers because the load is represented by a loudspeaker (inductivity). In the operating case, the quiescent point moves along an ellipsis whereby high current and high voltage will occur simultaneously in the transistor, i.e. the dissipation power periodically will attain very high values.

The influence of the emitter-base capacitance, which also is harmful, may be described as follows: the capacity is charged in the ON condition of the transistor. In order to turn off the transistor it has to be discharged. The time of discharge depends on the capacitance and the resistance which carries the discharge current. It is possible that during the discharge a high collector current flows and the voltage increases simultaneously. This rise in voltage is enforced in circuits with transformers by transformation in the transformer coils and in transformerless push-pull stages by the turn-on action of the other transistor.

If higher frequencies, e.g. frequencies of more than 10 kHz, are to be transmitted with a high power output one has to use transistors with a small emitter-base capacitance. The transistors AD 166 and AD 167 meet this condition. In any case one can also improve the frequency characteristics by a low-resistance drive.

## 1.2 Phono-Amplifier for Earphone Operation

The signal supplied by crystal pick-ups can be sufficiently amplified by simple circuits in order to drive dynamic earphones. Since stereophonic systems have been introduced, the reproduction with earphones is not limited any longer to ordinary listening devices. This kind of reproduction has also become important in home applications because a stereophonic effect can be experienced with earphones.

Fig. 1.2 shows a two-stage amplifier which perfectly suits this case. A frequency-correcting filter section and the volume control are incorporated in the input as well as the balance control for stereophonic operation — for the latter this amplifier has to be installed twice. The frequency-dependent negative feedback results in an emphasis on bass tones up to 12 db (at 100 Hz); it may be regulated by the potentiometer  $P_1$ .

The output stage operates in a common collector circuit the small output impedance of which warrants a good matching with the internal resistance of only 50—400  $\Omega$ , usually in dynamic earphones. Thus a low distortion factor is obtained.

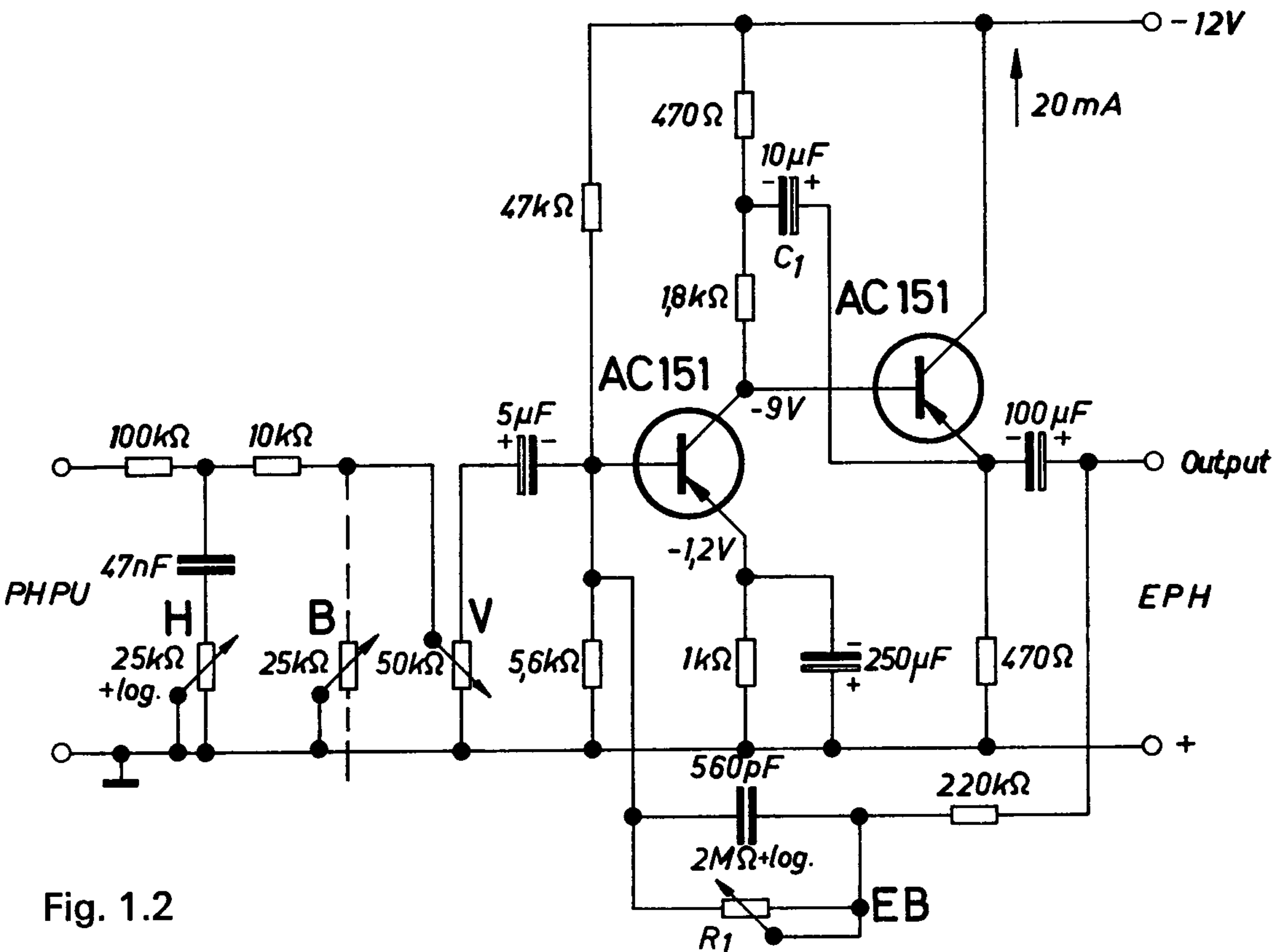


Fig. 1.2

The high input impedance of the output stage in a common collector circuit loads the pre-amplifier stage only to a small extent so that a high voltage gain is obtained. The decoupling of the load resistance by the condenser  $C_1$  even increases this gain because the load resistance of the pre-stage is connected in parallel only to the base-emitter diode of the output stage and, thus, the input impedance does not decrease. In order to obtain a high input impedance necessary for the matching to the crystal pick-up, series resistances are preceding the amplifier which, of course, reduce the voltage gain of the entire layout. For that reason the amplifier mainly has the purpose of an impedance transformer. By using transistors with a high gain one attains a voltage amplification of 1 to 2.

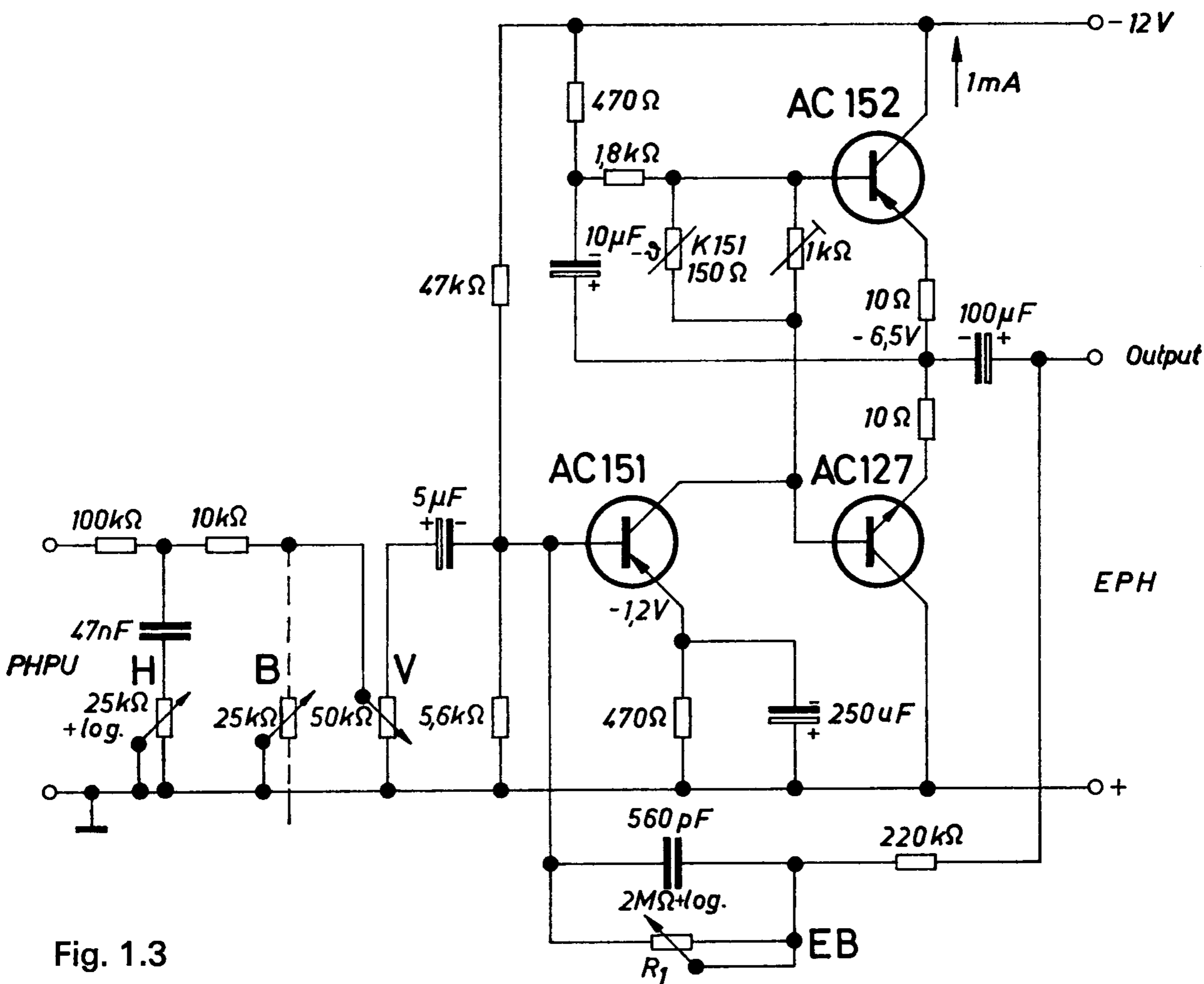


Fig. 1.3

In Fig. 1.3 the output stage is in a push-pull arrangement. Because of the small quiescent current this layout is especially suitable for portable equipment.

Technical Data	Fig. 1.2	Fig. 1.3	
Battery voltage	12	12	V
Quiescent current	20	3	mA
Max. load current	20	30	mA
Min. load resistance	50	50	$\Omega$
Input impedance	100	100	k $\Omega$
Max. power output	6 to 10	150	mW
Distortion factor	2	2	%
Voltage gain for open-circuit	1.2 to 2	1 to 1.5	
Range of tone control 100 Hz	12	12	db
10 kHz	26	26	db
Max. ambient temperature	50	50	$^{\circ}\text{C}$

### 1.3 Supplementary Amplifier for Dictaphones

In dictaphones reproductions usually are performed by earphones. Fig. 1.4 shows, however, that without any difficulty one can build a supplementary amplifier which makes the connection of a loudspeaker possible.

The voltage stabilized power supply provides a voltage of 9 V, and a current of 100 mA will be drawn at the maximum amplifier power output of 400 mW. The necessary input voltage for the amplifier driven to full output is 300 mV; the signal to noise ratio has at the output the value of 72 db. The circuit has no negative feedback for alternating currents.

Technical Data	
Supply voltage	220 V, 50 Hz
Power output	400 mW
Load resistance	10 $\Omega$
Input voltage for the amplifier driven to full output	300 mV

Transformer Tr 1 :

M 42/15 Dyn. Sheet IV/0.35, 0.5 L\*, alternate stacking

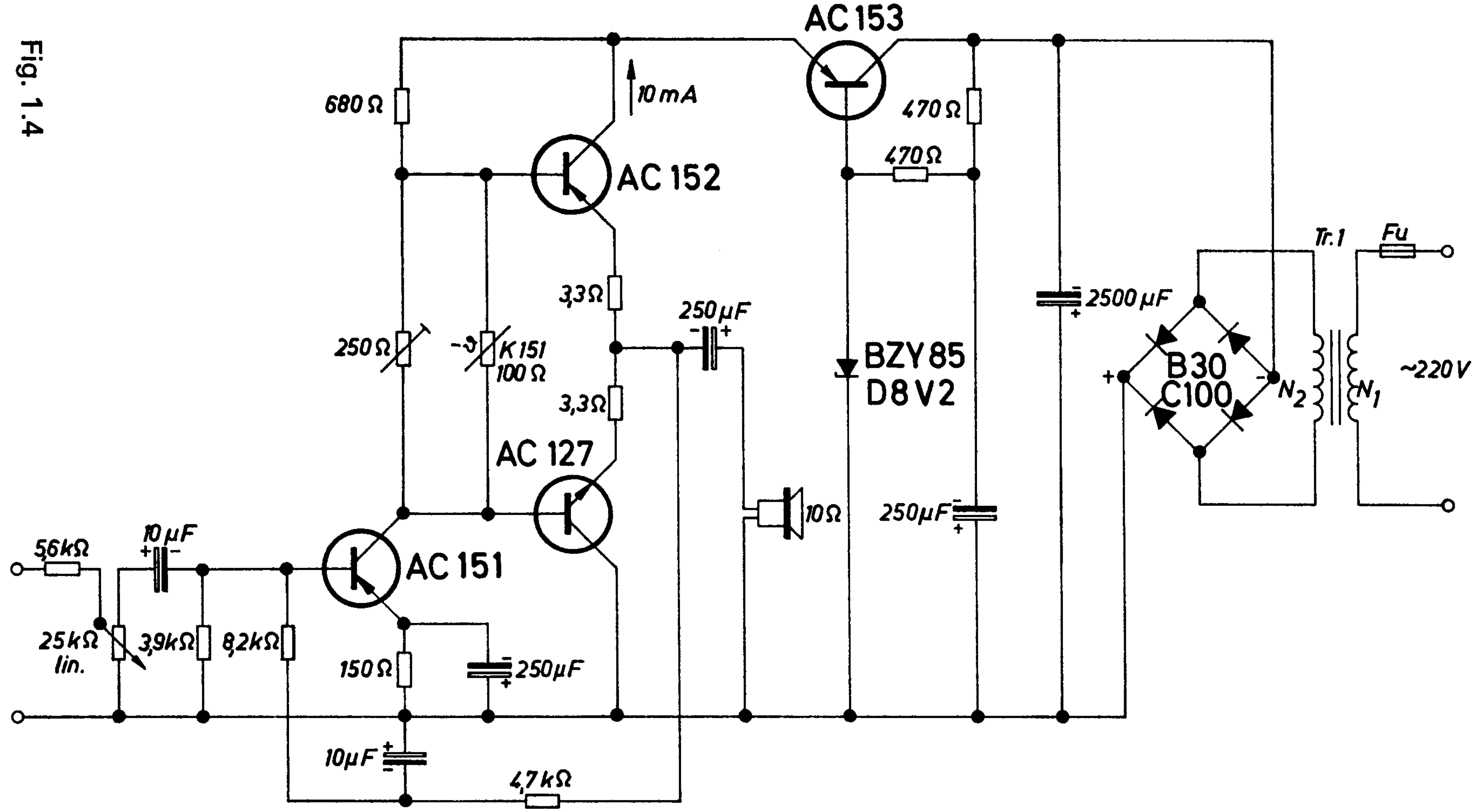
$N_1 = 4600$  turns of 0.1 CuL\*

$N_2 = 250$  turns of 0.4 CuL

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\* see page 106

Fig. 1.4



## 1.4 Audio Frequency Amplifier with a Silicon Transistor in the Input Stage

By using a npn transistor in the pre-stage and a pnp transistor in the driver stage of an AF amplifier one can save a coupling capacitor because galvanic coupling is possible. The difference between the circuit in Fig. 1.5 and the layout of the Siemens transistor quartet Q 610 is that instead of the pre-stage transistor AC 127 a silicon planar transistor is employed. The high gain of this transistor causes a higher input impedance and the low cut-off current warrants a high temperature stability of the AF amplifier which is DC coupled in 3 stages. Therefore one can design a highly resistive biasing circuitry that does not considerably damp the input impedance of the pre-stage.

### Technical Data

Supply voltage	9 V	Input voltage for	
Power output	1 W	a power output of 125 mW	5.6 mV
Load resistance	8 $\Omega$	Input impedance	35 k $\Omega$

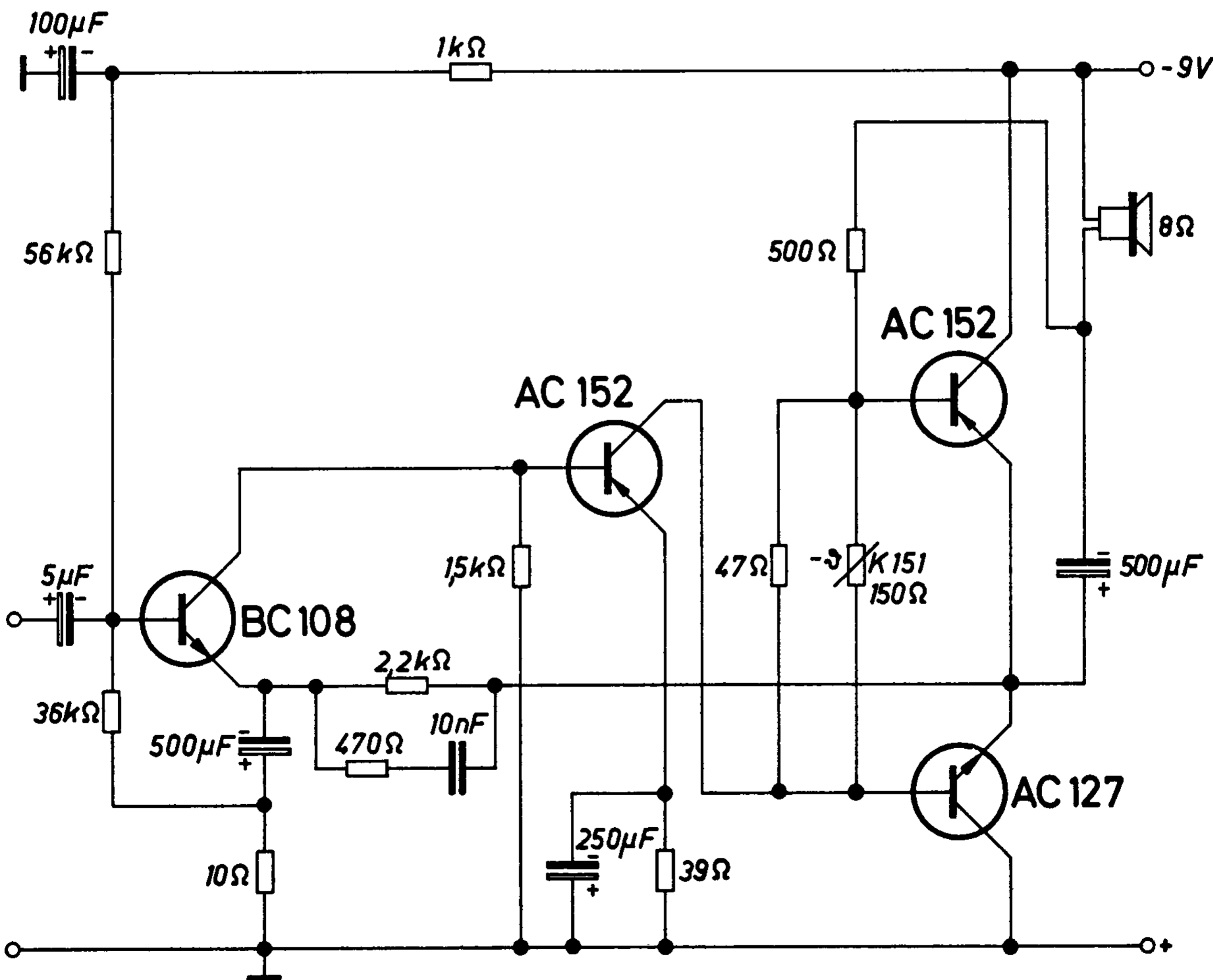


Fig. 1.5

## 1.5 Transformerless Power Stage for Portables and Car Radios

Transformerless power stages are more and more in use because of their low requirements of material. The only handicap for a general application is the relationship between supply voltage, power output and load resistance which is fixed because of the absence of a matching pad. Only 2 of the 3 mentioned magnitudes can be freely selected, and, in addition, the load resistance is limited by using loudspeakers with established internal resistances.

For the operation with car batteries, which mostly have a voltage of only 6 V, a high output power in the car radio plug-in units is desirable. Used as portable radios with a dry-cell battery of mostly 9 V, the power output should be lower in order to save battery power. Therefore, it is required that at smaller operating voltages a higher power output has to be provided. The relationship between operating voltage  $V_{CC}$ , load resistance  $R_L$  and power output  $P_o$  can be determined as follows:

In transformerless power stages the transistors are connected in series. For that reason only half of the operating voltage is available for each of the two transistors. The RMS output power is equal to the product of the RMS values of the AC voltage by the AC current or equal to half the product of the two peak values.

$$P_o \approx \frac{\frac{V_{CC}}{2}}{2} I_c = V_{CC} \frac{I_c}{4}$$

The load resistance is

$$R_L \approx \frac{\frac{V_{CC}}{2}}{I_c} = \frac{V_{CC}}{2 I_c}$$

The result is a peak current of

$$I_c \approx \frac{V_{CC}}{2 R_L}$$

By substituting this equation in the above function for the power output  $P_o$ , one obtains the relationship:

$$P_o \approx \frac{V_{cc}^2}{8 \cdot R_L}$$

In this equation the influence of the transistor cut-off voltage and further losses are not taken into consideration.

If, for example, in the 6 V operation, a power output is required that is twice as much as in the 9 V operation, the load resistance has to be decreased to a quarter of its value at 9 V. Starting from the usual value of 8  $\Omega$  one obtains a loudspeaker resistance of only 2  $\Omega$  which can hardly be realized. In order to avoid this difficulty, a bridge circuit with a 6 V supply voltage is suggested. In this case, the total operating voltage is available for a drive to full output and the computation executed above for usual transformer output stages changes as follows:

$$P_o \approx \frac{V_{cc} \cdot I_c}{2}$$

$$R_L \approx \frac{V_{cc}}{I_c} \text{ or } I_c \approx \frac{V_{cc}}{R_L},$$

$$\text{then } P_o \approx \frac{V_{cc}^2}{2 R_L}$$

Under usually equal conditions ( $V_{cc}$  and  $R_L$  equal) the quadruple power output  $P_o$  is achieved by a switch over to the bridge operation. If, as in this case, the operating voltage is simultaneously changed from 9 V to 6 V, one obtains about twice the output power.

Fig. 1.6 shows a circuit suitable for this application. At 6 V operation a peak current of approximately 1 A is achieved in the output stage; therefore, one employs the complementary pair AC 153/AC 176. The driverstage of this pair requires an operating current of about 20 mA. This operating current is too high for portable radios (saving of battery power). Therefore the bridge branch used in the battery operation is preceded by a smaller complementary pair consisting of the transistors AC 127/AC 152.



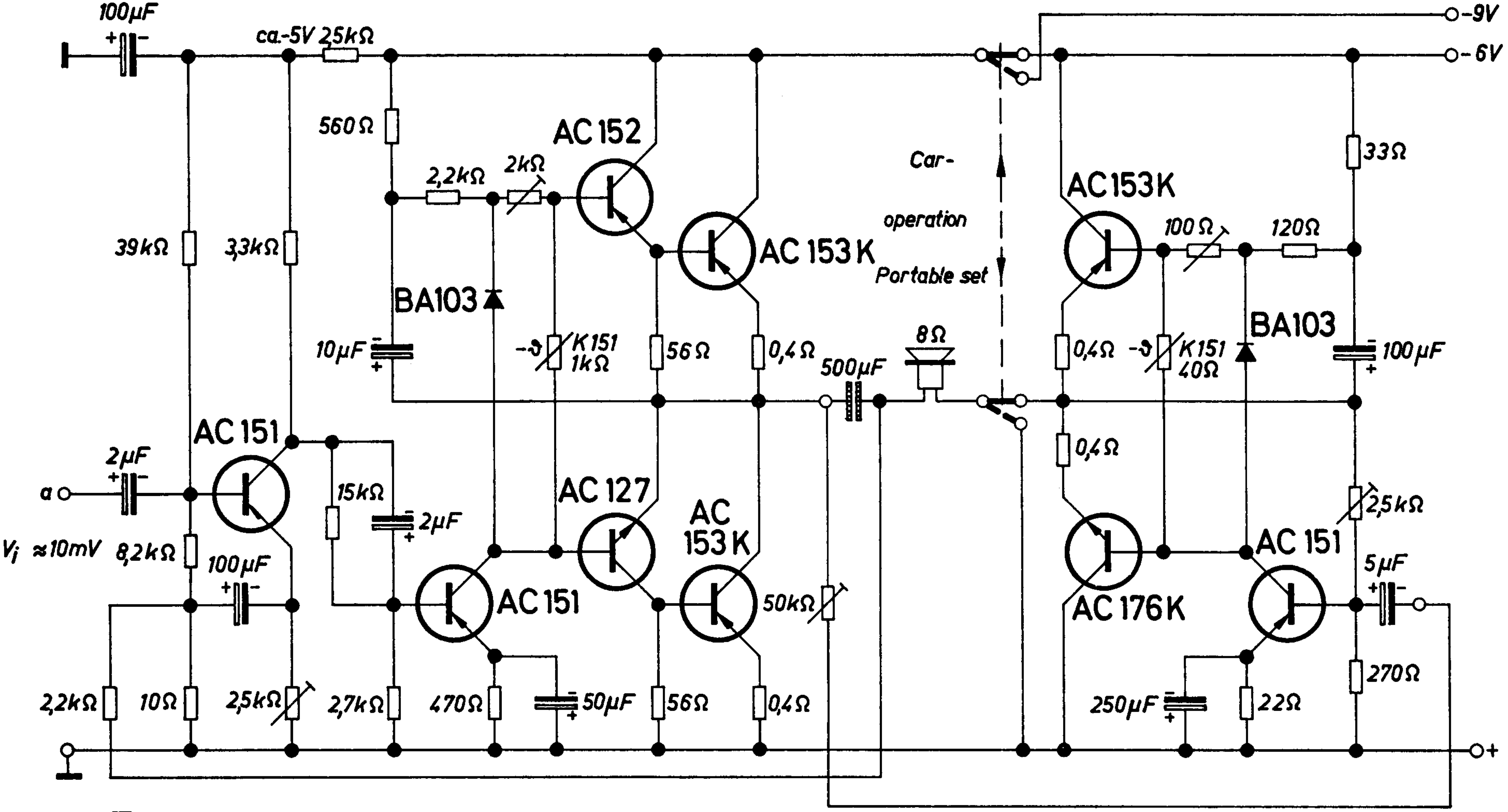


Fig. 1.6

This pair operates in a push-pull arrangement and requires a quiescent current of only 1 to 1.5 mA. As in this case the phase reversal already occurs in the driver stage, two pnp transistors AC 153 in the power stage can be used.

The quiescent currents of the output stages have to be temperature-stabilized. This can be achieved by two negative temperature coefficient thermistors K 151. Since the battery voltage may vary to a high extent, a stabilization of the quiescent point with silicon diodes BA 103 operating in forward direction is provided.

Also in the bridge operation the output condenser of 500  $\mu\text{F}$  necessary for portable sets has to be connected in series to the load resistance if the output potentials of the two bridge branches are not equal. This may occur, for example, if one branch is installed in the portable radio and the other branch in the car plug-in. Then a bipolar type has to be selected for the condenser.

This circuit seems to require a relatively high number of components; it presents, however, a possible solution in transformerless layouts.

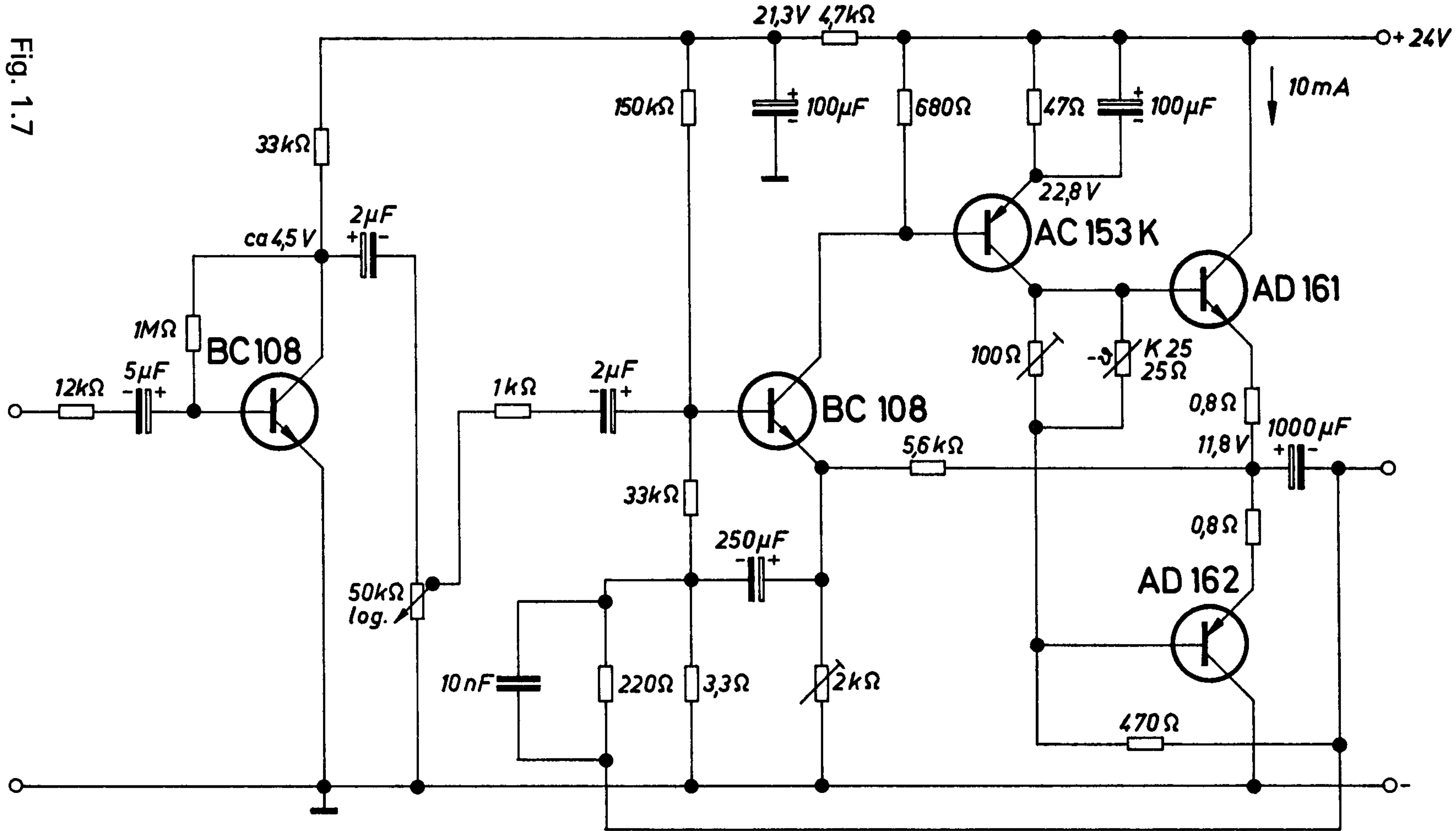
#### Technical data

Supply voltage	6 V	9 V
Power output	2 W	1 W
Load resistance	8 $\Omega$	8 $\Omega$
Input voltage for the amplifier driven to full output	10 mV	10 mV

### 1.6 Audio Frequency Amplifier 24 V / 7 W

The circuit in Fig. 1.7 is suitable for all those applications in which the minuspole has to be connected to chassis. If npn transistors are used in the pre-stage, design advantages will result. In the output stage one applies the high power complementary pair AD 161/AD 162.

Fig. 1.7



## Technical Data

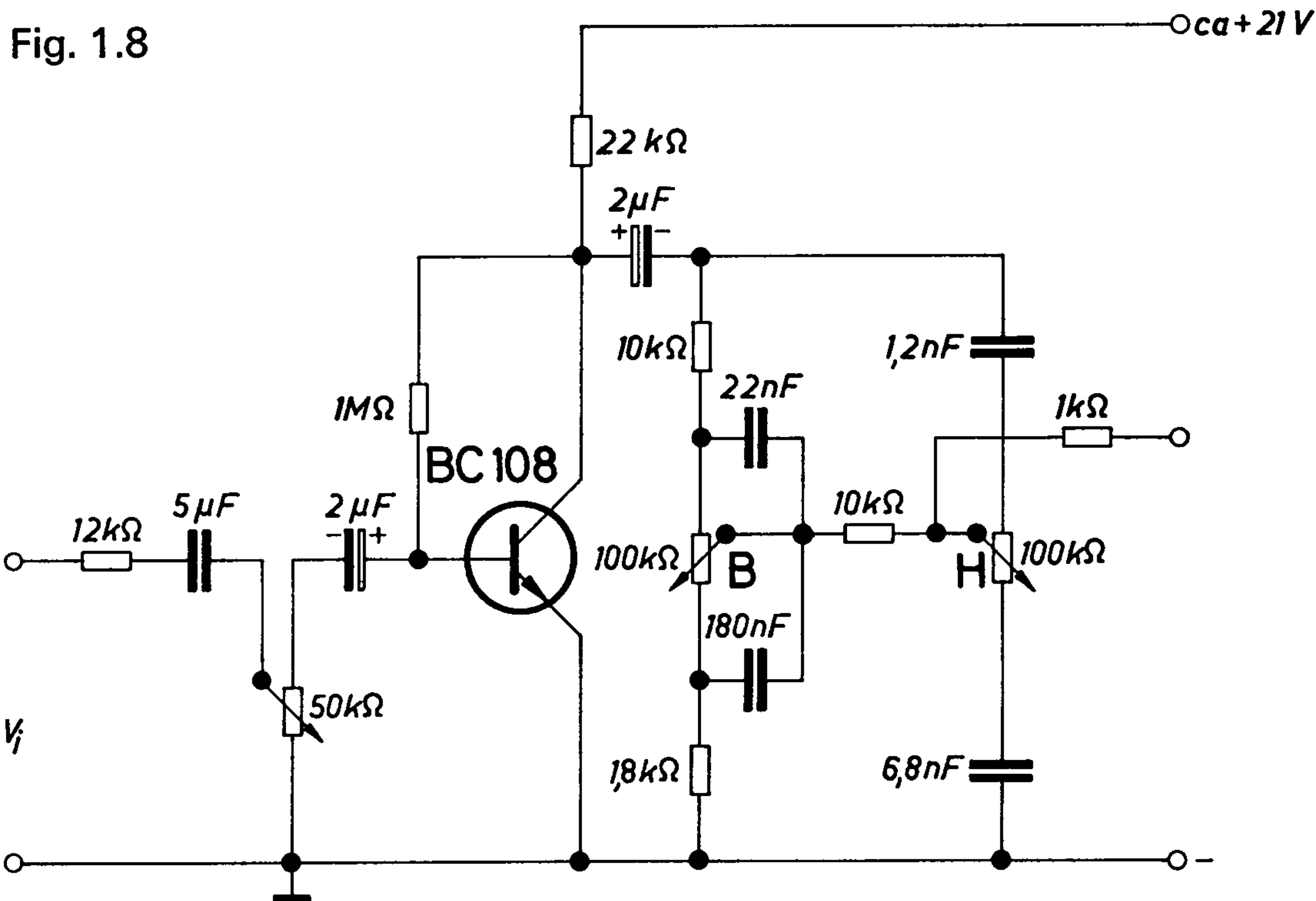
Supply voltage	24 V
Operating current	40 to 470 mA
Power output	7 W
Load Resistance	8 $\Omega$
Input voltage for maximum power output	3 mV
Input impedance	15 k $\Omega$
Power band width ( $P_o=5.6$ W; $K=2\%$ )	25 Hz to 20 kHz
Noise voltage at the output for fully advanced volume control	10 mV

## 1.7 Tone Control

The tone controller in a bridge circuit, according to Fig. 1.8, is applicable for the AF amplifier shown in Fig. 1.7. It has to precede the second prestage. Hereby the input sensitivity of the amplifier decreases slightly (input voltage approximately 24 mV for a drive to full output); therefore, a supplementary prestage probably has to be applied.

The diagram in Fig. 1.9 indicates the tone correction which is possible with this arrangement.

Fig. 1.8



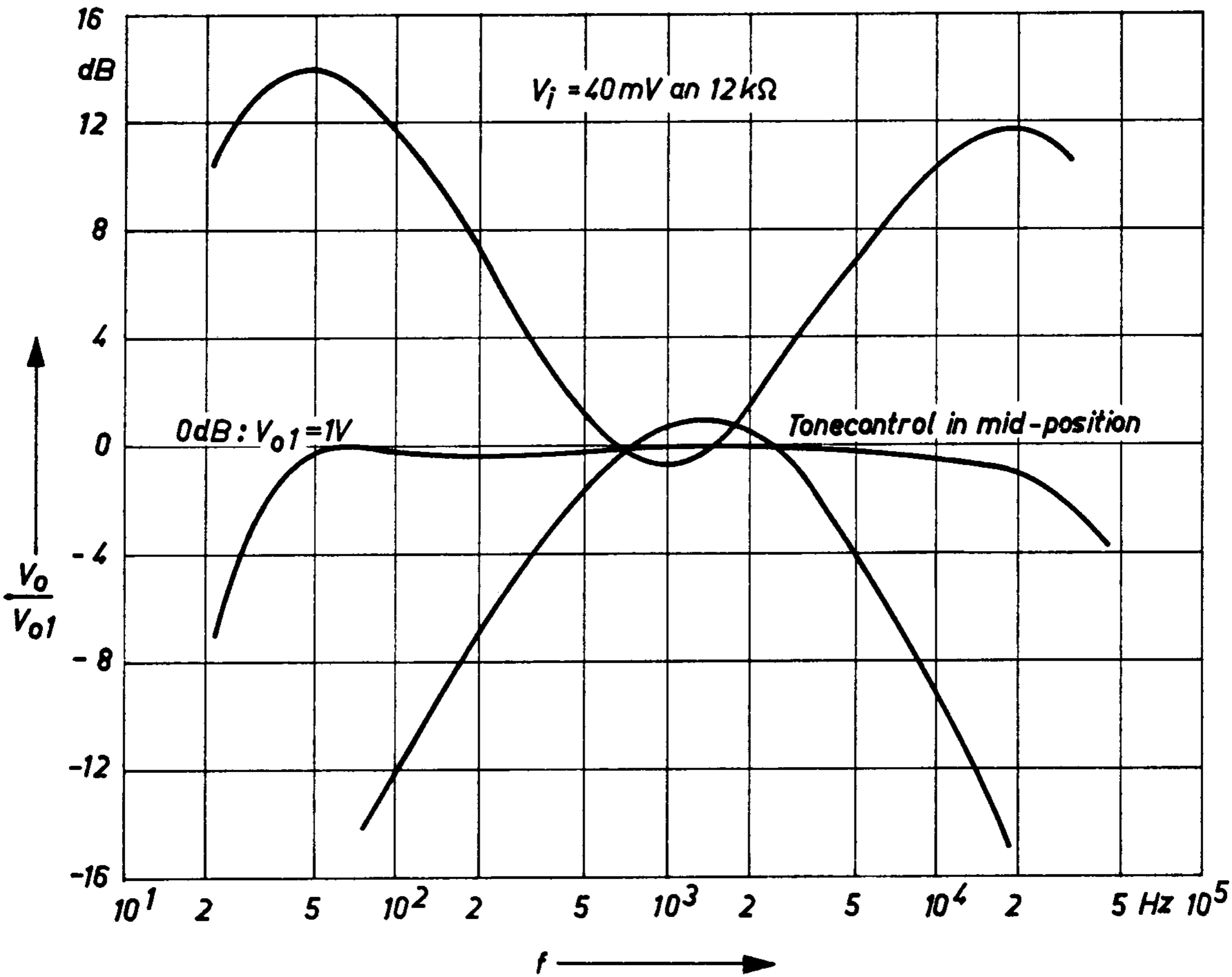


Fig. 1.9

## 1.8 Audio Frequency Amplifier 32 V / 26 W

For an output power of more than 10 W there are still no suitable complementary transistors available. Therefore it has become more and more successful for transformerless power amplifiers to drive two pnp transistors with a complementary pair.

Fig. 1.10 shows an appropriate circuit. In order to obtain a good frequency characteristic, the output stage is driven by a source which is as low-resistive as possible. A balance controller is provided at the input for stereo-operation. The amplifier has very small distortion factors as indicated in the diagrams of the Fig. 1.11 and 1.12.

Fig. 1.10

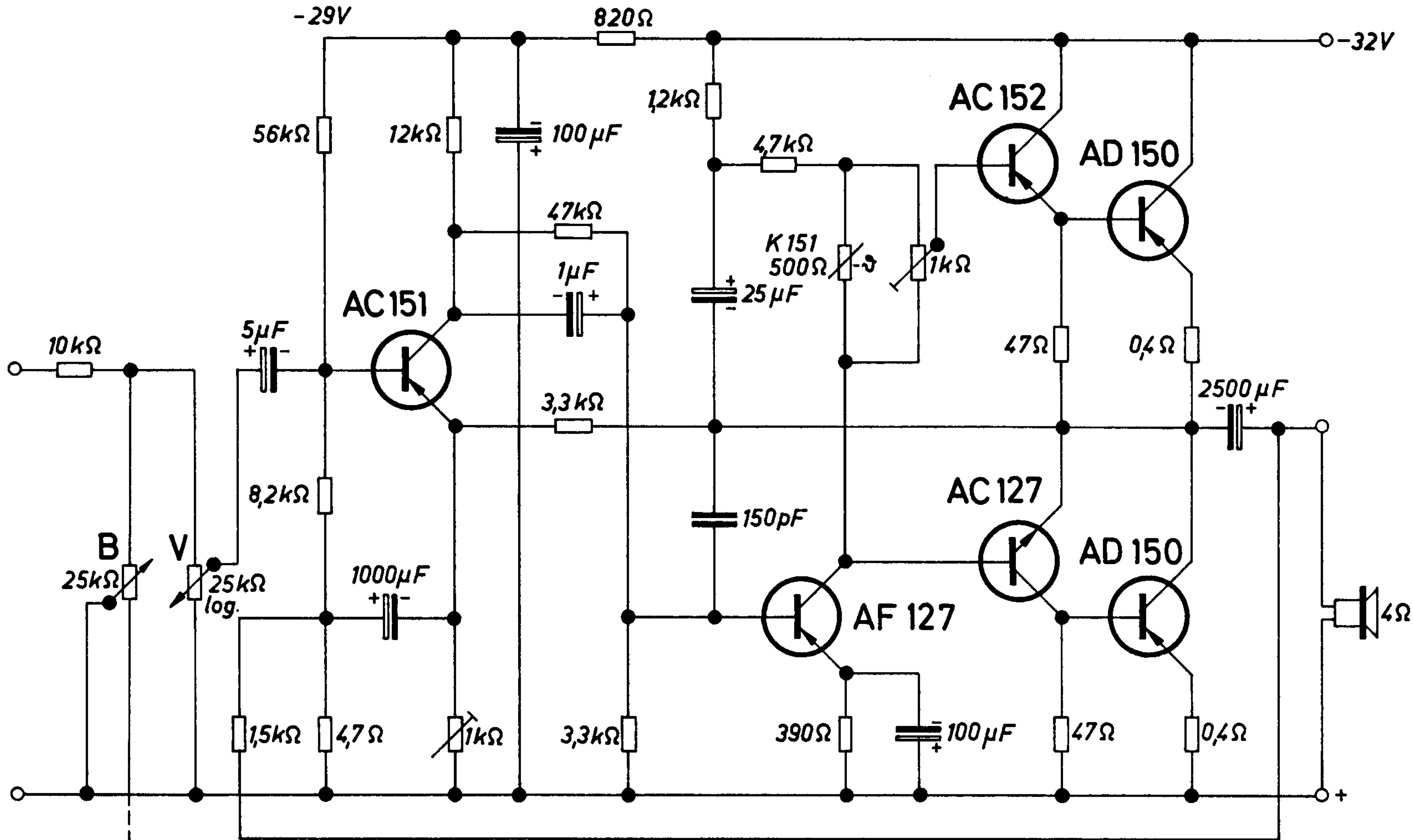


Fig. 1.11

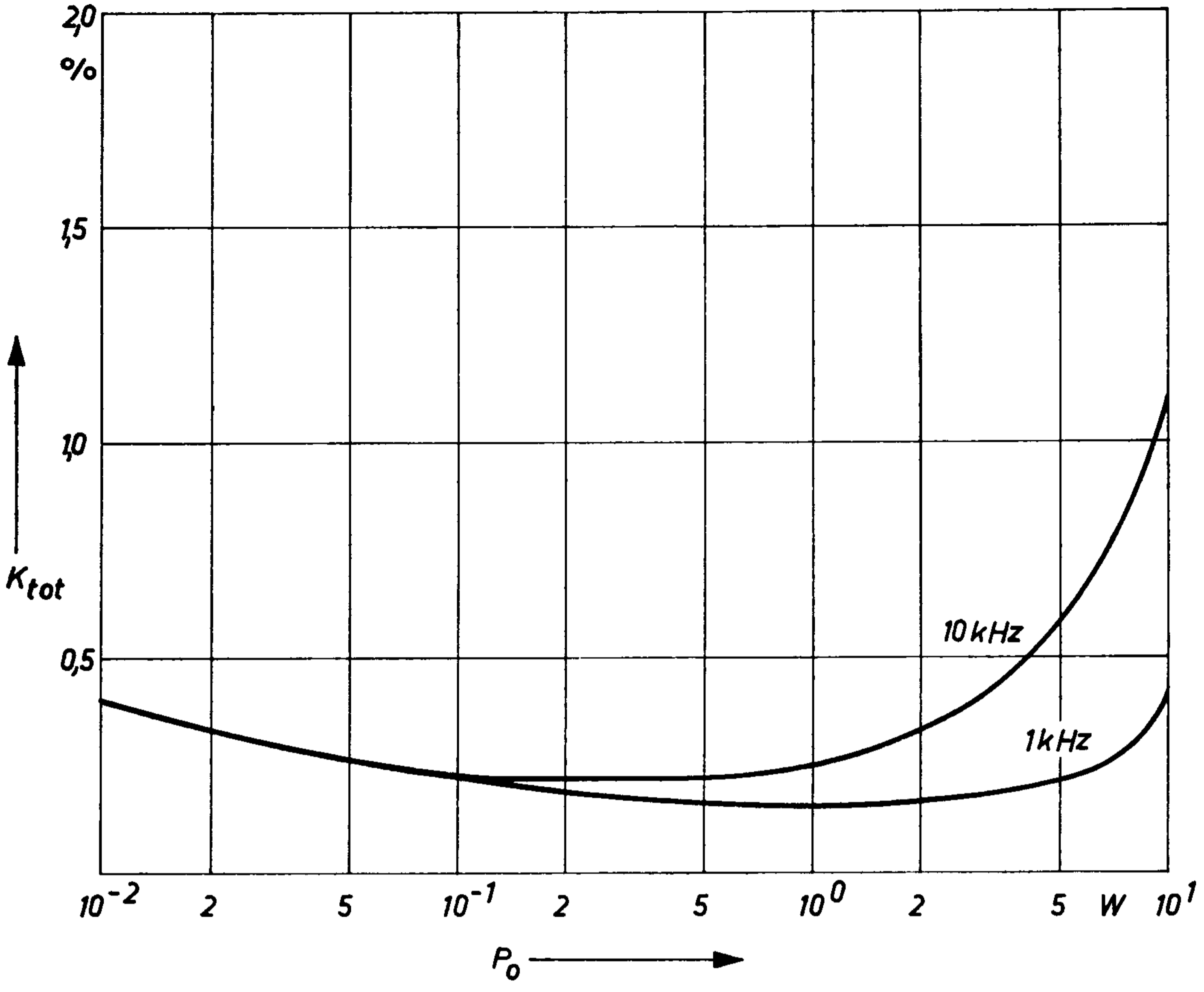
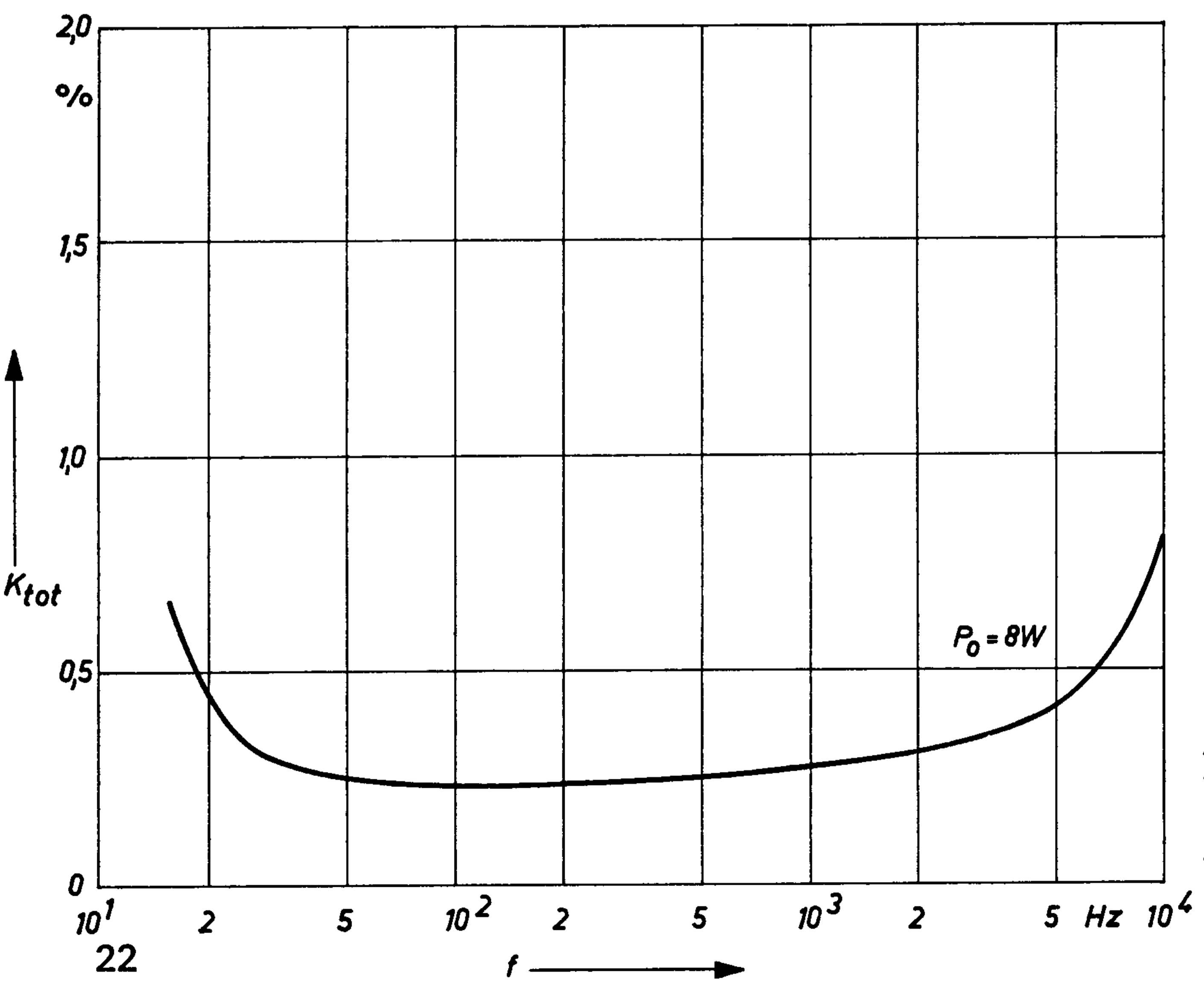


Fig. 1.12



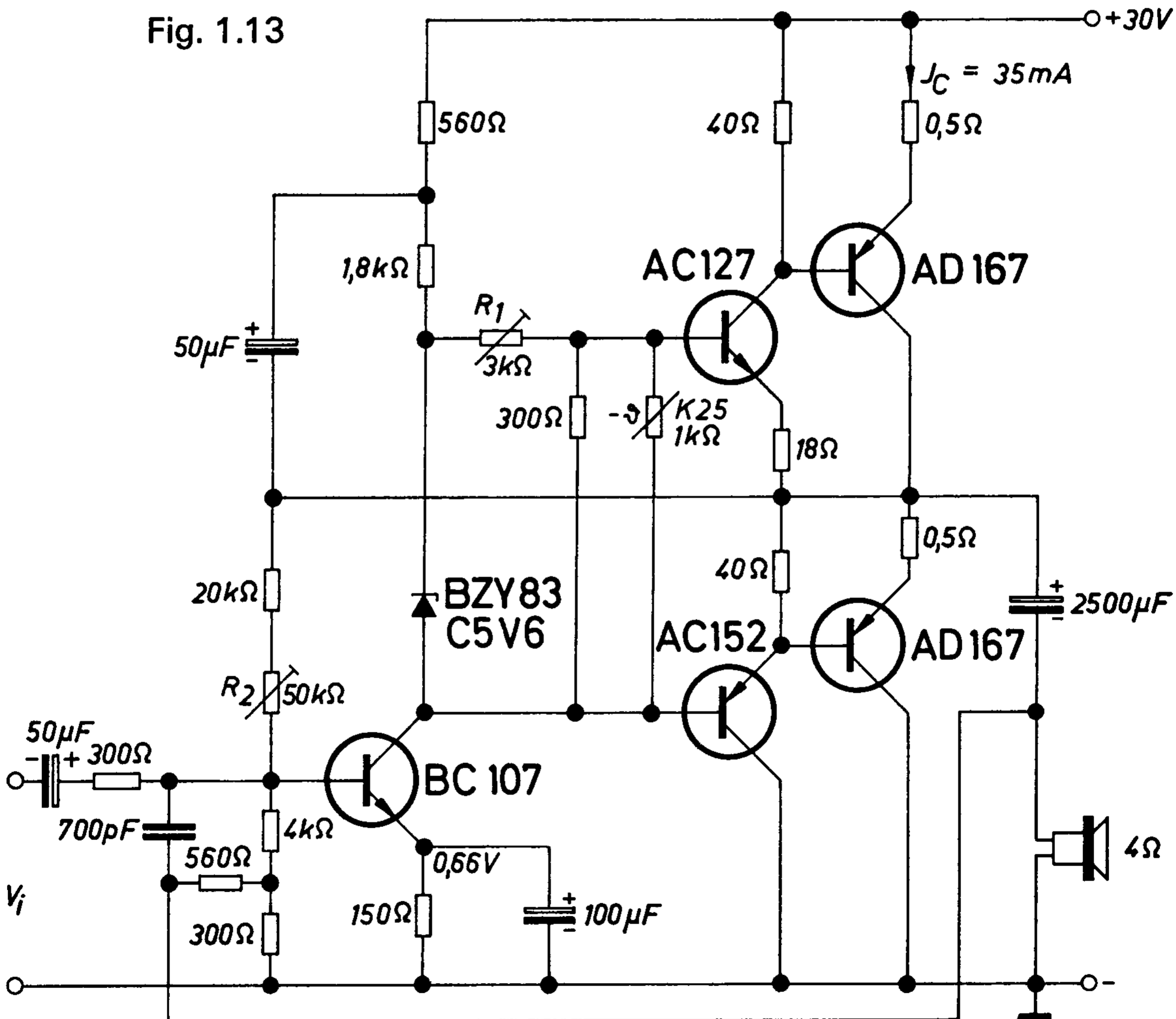
$K_{tot}$  = total distortion factor

## 1.9 Hi Fi Amplifier 30 V/15 W

The powerstage transistors of Hi Fi amplifiers have to meet several requirements. Above all they should have a linear characteristic up to high signal levels and a high cut-off frequency. The cut-off frequency of alloy-junction power transistors is in common emitter circuits which are usually used in AF output stages, approximately 10 to 15 kHz — consequently slightly below the threshold of audibility; therefore, a good reproduction of the total audible range only will be made possible by a negative feedback and a low-resistance drive. Lately, power-drift transistors with a considerably higher cut-off frequency are available by which the above mentioned difficulties can be avoided. These are the types AD 166 and AD 167.

Fig. 1.13 shows the circuit of an AF output stage with the transistors AD 167 for a power output of 15 W.

Fig. 1.13





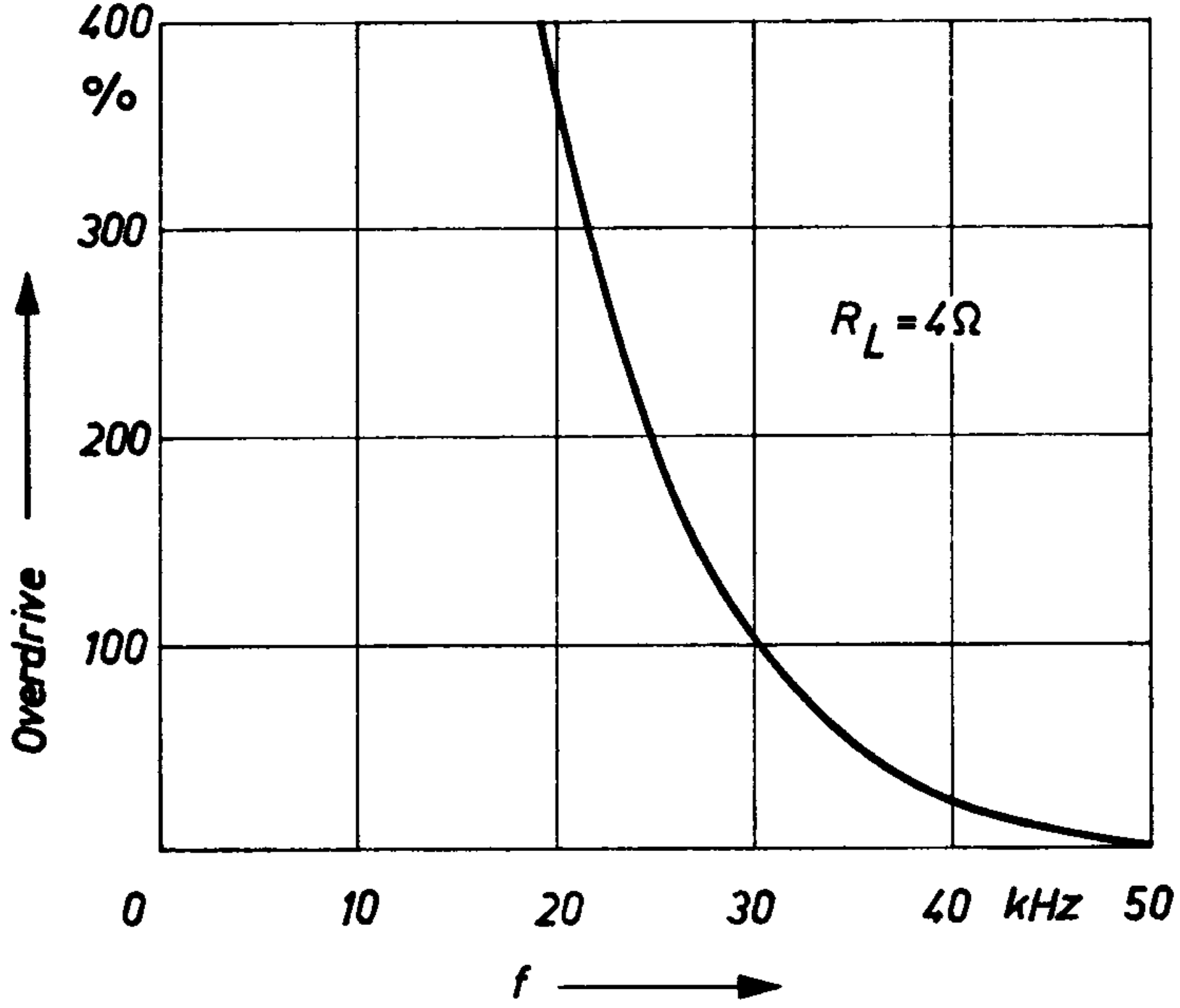
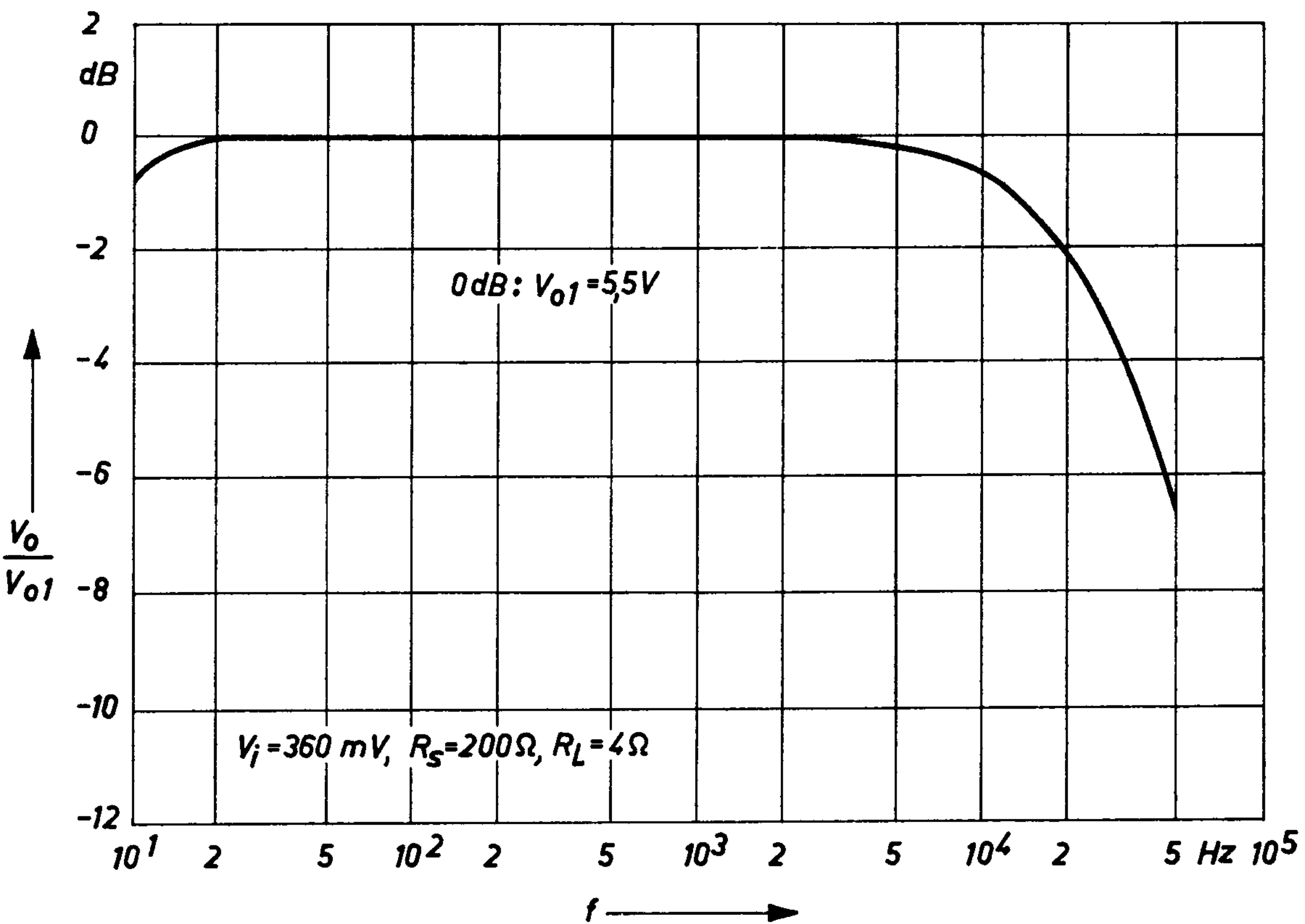


Fig. 1.14

Fig. 1.15

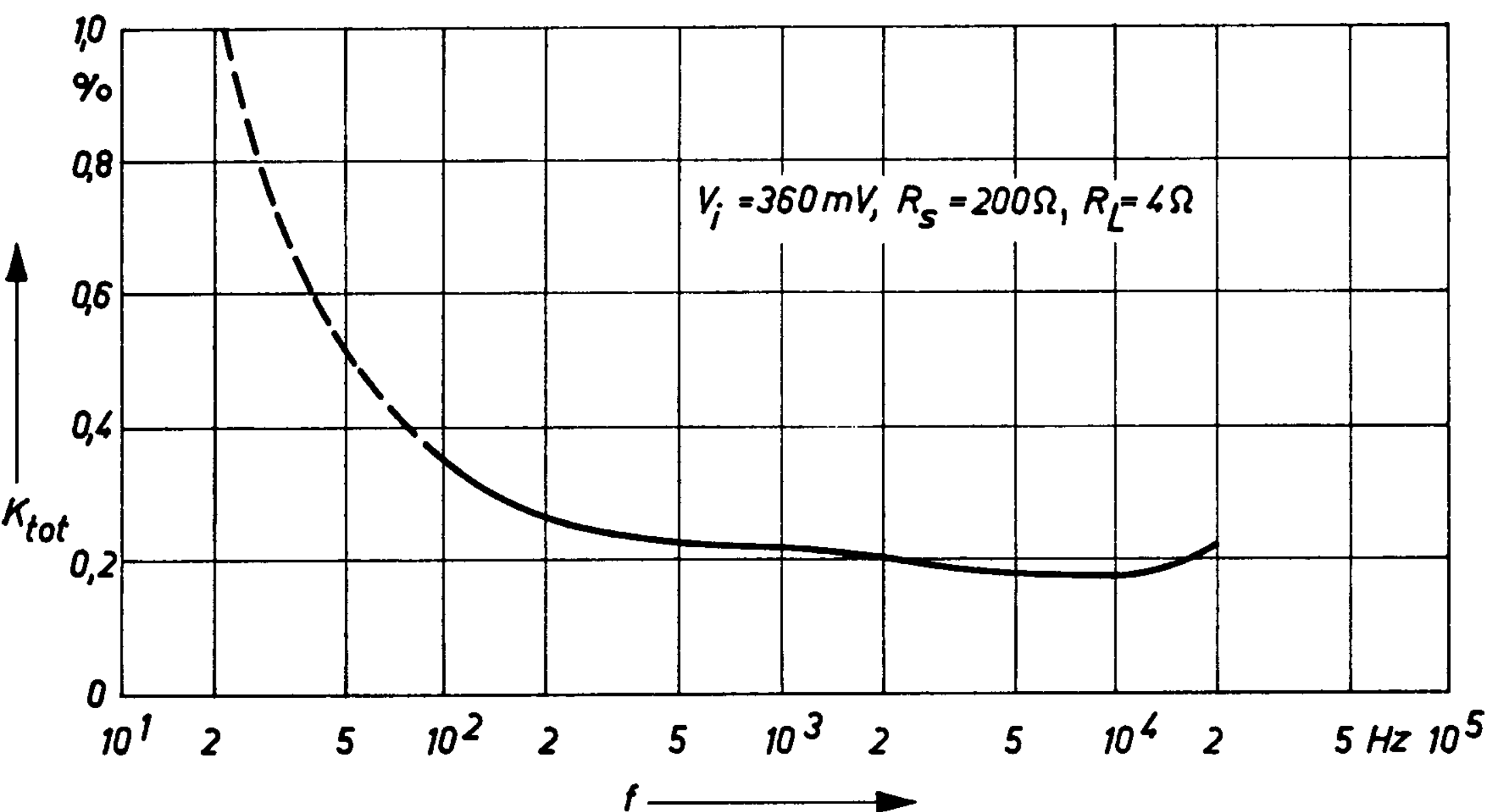


The phase reversal which is necessary for the drive of the power stage is accomplished in the push-pull driver stage with the complementary pair AC 127/AC 152. The average loss in the output stage transistors amounts to 5.5 W at a frequency of 1 kHz. This loss considerably increases, however, at frequencies beyond 15 kHz, especially in the case of overdriving. Therefore, the pre-amplifier has to be designed in such a way that overdriving at higher frequencies will be avoided. Fig. 1.14 shows the allowable overdrive vs. the transmitted frequency.

The stabilization of the quiescent current of the output stages is obtained by a negative temperature coefficient thermistor K 25 for change of temperature and by a Zener diode BZY 83 for a variation in the supply voltage. The negative temperature coefficient thermistor is to be mounted to the same heatsink as the output transistors.

In the design of the amplifier one tried to achieve above all a smaller distortion factor in addition to the good frequency characteristic made possible by the applied transistors. Fig. 1.16 shows how the distortion factor depends on the frequency. In fact, the distortion factor is below 0.3% in the whole transmission range.

Fig. 1.16



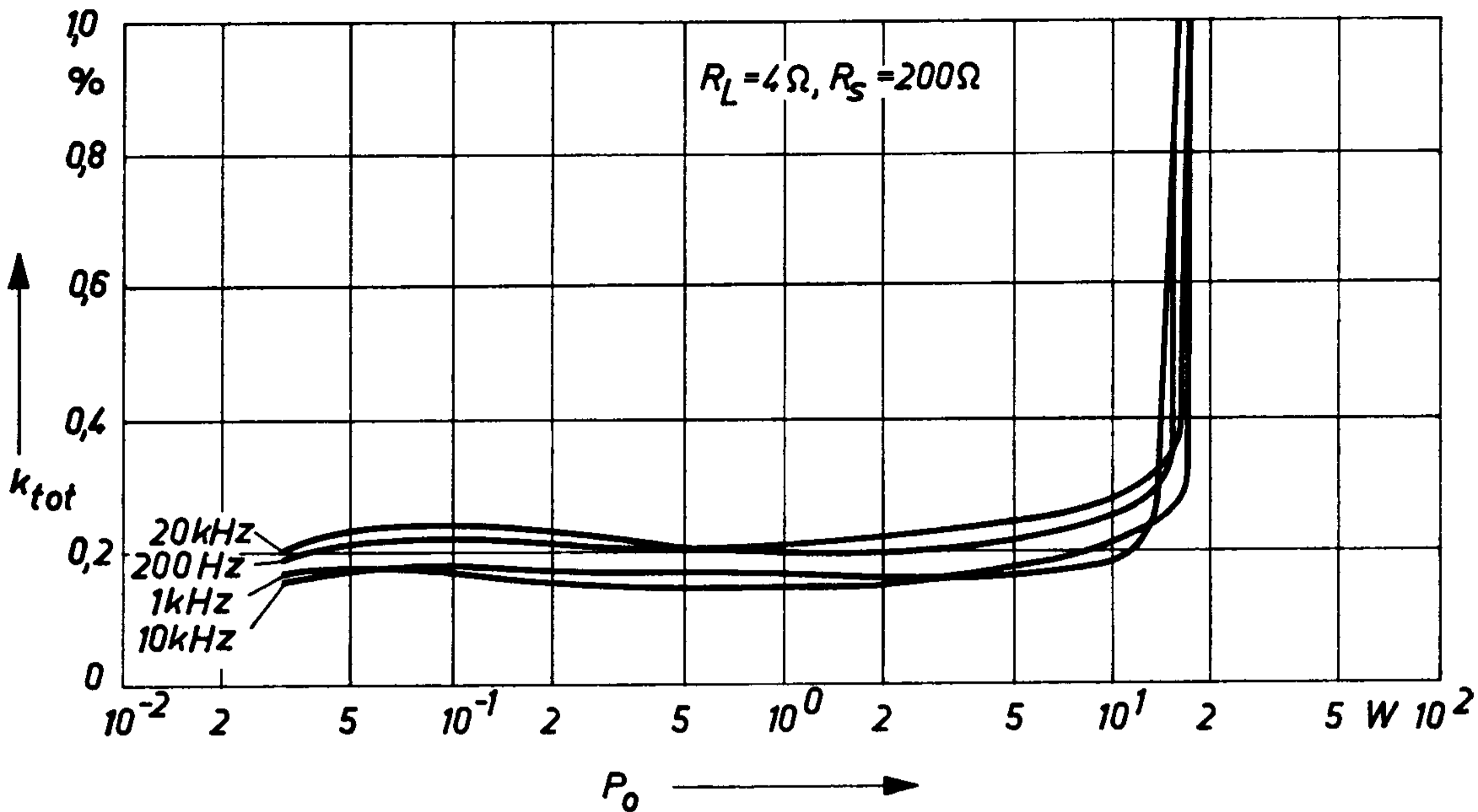


Fig. 1.17

Fig. 1.17 finally demonstrates the function of the distortion factor vs. the output power for four different frequencies. The distortion factor remains below 0.3% at all frequencies until the limit is reached where the amplifier is driven to full output.

#### Technical Data

Supply voltage	30 V
Supply current	50 to 1000 mA
Power output	15 W
Load resistance	4 $\Omega$
Input voltage for the amplifier driven to full output	500 mV
Input impedance	> 300 $\Omega$
Hum signal ratio ( $P_o = 100$ mW, hum voltage of the mains supply = 0.27 V $R_S = 200 \Omega$ ; $R_L = 4 \Omega$ )	60 db
Unweighted signal-to-noise ratio ( $P_o = 100$ mW)	> 85 db

### 1.10 Transformerless Audio Frequency Amplifier 40 W

Usually only such pairs are called complementary pairs where the characteristics of the pnp and the npn transistors correspond to each other to a large extent. The circuit design for transformerless

output stages, which is very simple for such complementary pairs, may also be applied under certain conditions for transistor pairs consisting of transistors with different characteristics. It is only required to incorporate a high negative feedback which is accomplished most expediently by operating the transistor in a common collector circuit. In the circuit indicated in Fig. 1.18 the transistors BUY 13 and AU 105 are employed for the output stages. The BUY 13 is a npn silicon power transistor and the AU 105 a pnp germanium power transistor. In spite of these unequal transistor types the distortion factor remains below 1 % of the whole audible frequency range without an overdrive of the amplifier. The output stage is preceded by a push-pull driver stage which also is provided both with a npn silicon transistor and a pnp germanium transistor, but with lower power. This stage also operates in a common collector circuit. Since the prestage is in a common collector configuration, the whole power part has no voltage gain, that means, it only operates as impedance transformer. The operating point of the three stages is adjusted in the prestage as they are galvanically coupled. The DC-output voltage is set at the output by the potentiometer  $P_1$ . The potentiometer  $P_2$  adjusts the operating point of the output stages. The quiescent current of these stages should not be set too low with regard to the distortion factor. A value of  $I_c$  (quiescent) = 25 mA is recommended.

In order to warrant the temperature stability a compensation circuit with a negative temperature coefficient thermistor K 151 is provided. This negative temperature coefficient thermistor has to be mounted with a good heat conduction to the transistor BUY 13. It prevents an increase of the collector current at a temperature rise of the output stage transistors by reducing the base voltage of the driver stage.

The capacity of the condensers  $C_1$  and  $C_3$  depends on the required lower cut-off frequency. In Fig. 1.18 the value of  $f_1 = 40$  Hz will be attainable. The condenser  $C_3$  has to be reliable with regard to the switching operation because it is constantly recharged with high currents at a drive of the amplifier to full output. The condenser  $C_2$  warrants equal alternating current potentials at the inputs of the two driver transistors. For providing the required current a power supply of 60 V, 1 A is necessary. A suitable power supply is described in chapter 7.3. It must have an electronic safety-device because the output stage transistors of the AF amplifier are endangered by load peaks and short circuits.

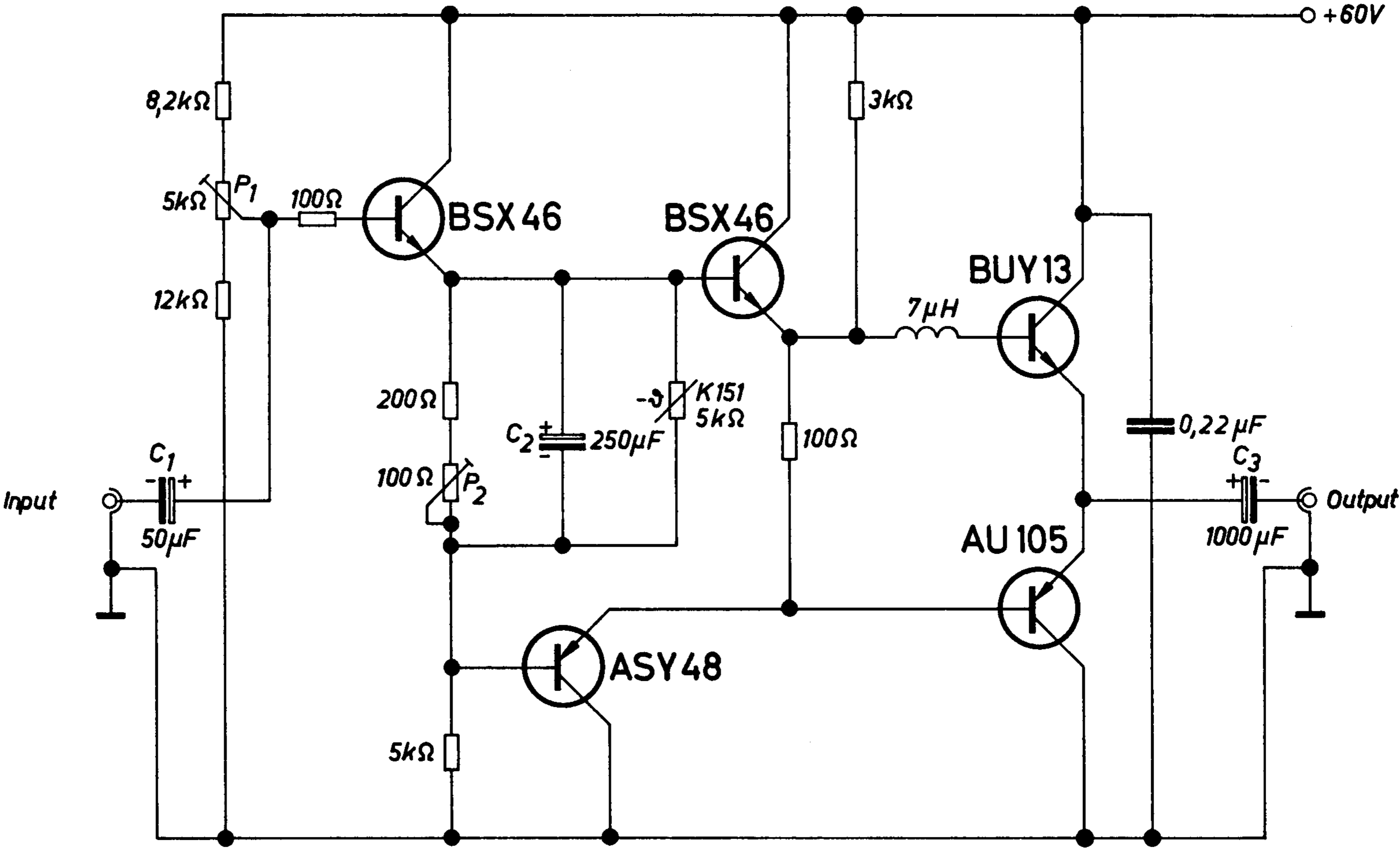


Fig. 1.18

## Technical Data

Supply voltage	60 V
Supply current	1 A
Power output	40 W
Load resistance	8 $\Omega$
Input voltage for the amplifier driven to full output	25 V
Input impedance	approx. 5 k $\Omega$
Frequency range (— 3 db)	20 Hz to 25 kHz
Distortion factor (power output 30 W)	$\leq 1 \%$

### 1.11 Three Stage Low Frequency Amplifier in Monolithic Technology

When planar technology is applied, diodes as well as transistors, condensers and resistors can be incorporated in a silicon chip. This fact makes it possible to build complete circuits in a uniform substrate. Semiconductor circuits become more and more important not only because of their small size but also on account of their reliability of operation and of the simple layout of equipment.

All the components produced in the planar technology result from diffusion of certain materials into silicon. An oxide masking bounds the areas of these zones. One starts with a silicon wafer which is covered by a coating of silicon dioxide ( $\text{SiO}_2$ ). This oxide is stable (quartz) and prevents the diffusion of most of the doping materials used for diffusion processes. By photolithographical methods "windows" through which the doping materials can penetrate will be etched into this oxide coating.

A resistance is established, for example, by diffusion of such doping materials through a "window" of a special shape. The doping materials produce conductivity characteristics different from those of the substrate. The resistance depends on the surface and the cross section of the doped zone as well as on the doping level. The resistance is separated from the substrate by a pn layer. Several resistances can be connected in series without influencing each other.

Diodes also can be produced according to the process described above. In this case the diffused zone serves as the p or n region and the substrate as the n or p region, respectively.

The diffusion of a transistor proceeds similarly to that of a planar transistor. Two diffusions are necessary, one for the base and one for the emitter, which take place in several "windows" in the oxide coating. The substrate makes up the collector.

In this technology there exist two possibilities of making condensers. Either, one uses the capacitance of a reverse biased pn junction, i.e., a diode, or, one evaporates a metal film on the oxide coating and uses the oxide as a dielectric. For the first mentioned type of condensers the capacitance depends substantially on the voltage and a leakage current will occur. On the other hand relatively high capacitance values can be attained in a small area because the depletion layer serving as a dielectric is very thin. They are, however, non-reversible condensers. For the second type of condensers one encounters neither a voltage dependence nor a leakage current but only low capacitance values (bipolar) can be realized.

Because of the small size of the components, integrated circuits can be used for hearing aids and microphone preamplifiers. It is certain, however, that the amplifiers will not be employed only in these cases. Semiconductor circuits considerably help building the equipment in well-ordered layouts and in attaining a high reliability of operation. The danger of getting dry joints is, for example, much smaller, if large parts of the whole circuit are diffused into one substrate.

Fig. 1.19 indicates the circuit of an AF amplifier in integrated circuitry. The 3 transistors and 5 resistors located in the shaded field are diffused into a small silicon chip of approximately  $1 \text{ mm}^2$ . Together with the additional components an AF amplifier is obtained for a battery voltage of  $V_{BB} = 3 \text{ to } 7 \text{ V}$ .

The transistors are direct-coupled. The externally connected potentiometer P sets the quiescent current of the first stage and also that of all the other stages. Because of the small cut-off currents of planar transistors the temperature dependence of the operation point is insignificant. This guarantees the temperature stability of the whole circuit. With the amplifier a voltage gain of 65 db is attained. The maximum output voltage is 1 V at a load

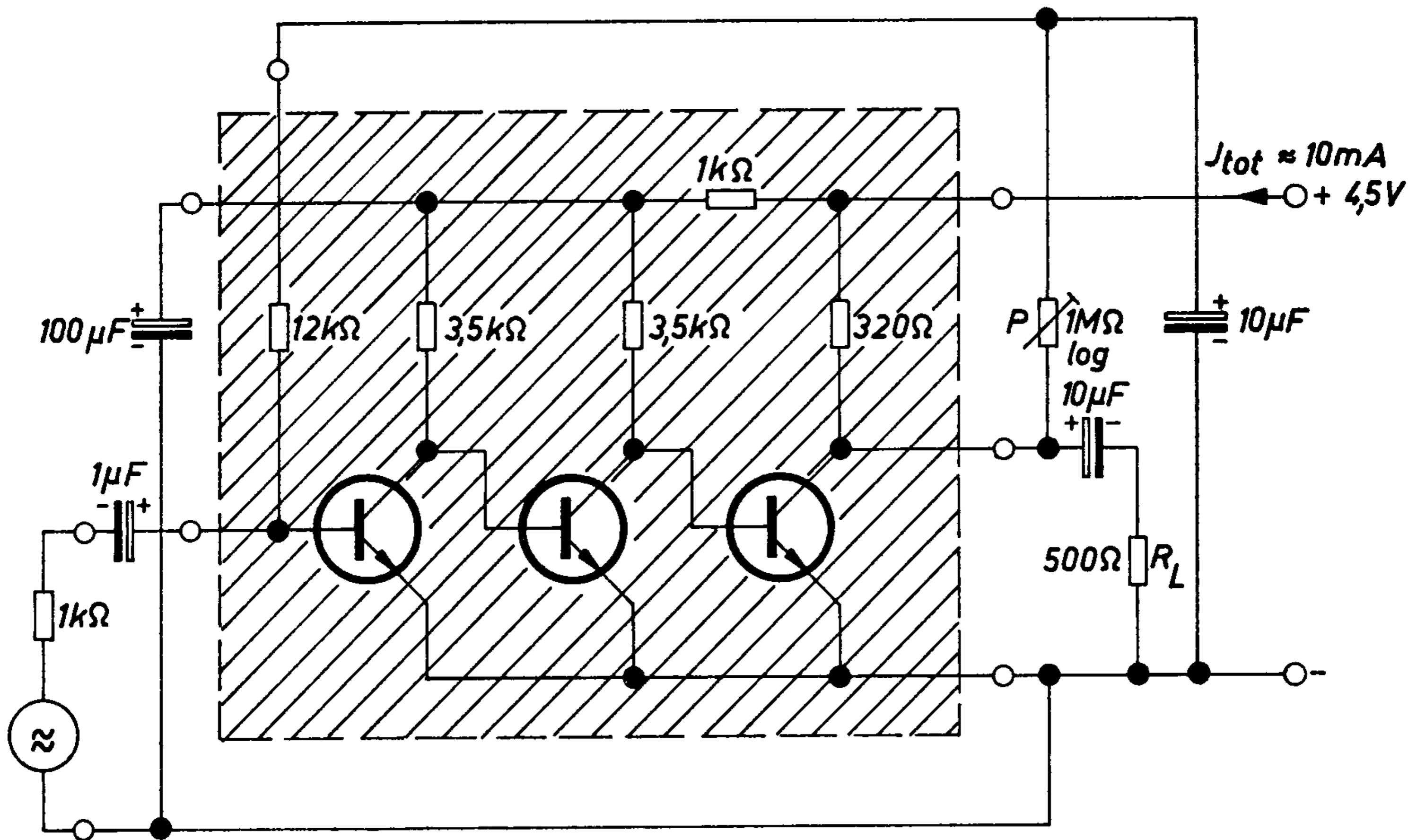


Fig. 1.19

resistance of  $500 \Omega$ . The 3 db frequency limits are at 50 Hz and 150 kHz, respectively.

The integrated circuit called TAA 111 is encapsulated in the standard case TO-5.

The attainable output voltage  $V_{o \max}$  depends on the magnitude of the load resistance  $R_L$ . The diagram in Fig. 1.20 shows this relationship. The amplifier operates correctly within a temperature range of  $-30 \text{ }^\circ\text{C}$  to approximately  $+130 \text{ }^\circ\text{C}$ . Fig. 1.21 indicates how the amplifier (voltage gain  $G_V$  and output voltage  $V_{o \max}$ ) operates in this temperature range. Lastly the diagram in Fig. 1.22 depicts the dependence of the output voltage and the voltage gain on the operation voltage.



### Technical Data

Supply voltage

4.5 V

Supply current

approx. 10 mA

Input impedance

$\geq 3$  k $\Omega$

Load resistance

500  $\Omega$

Max. output voltage

$\geq 1$  V

Voltage gain

65 (>62) db

Frequency range (—3 db)

50 Hz to 150 kHz

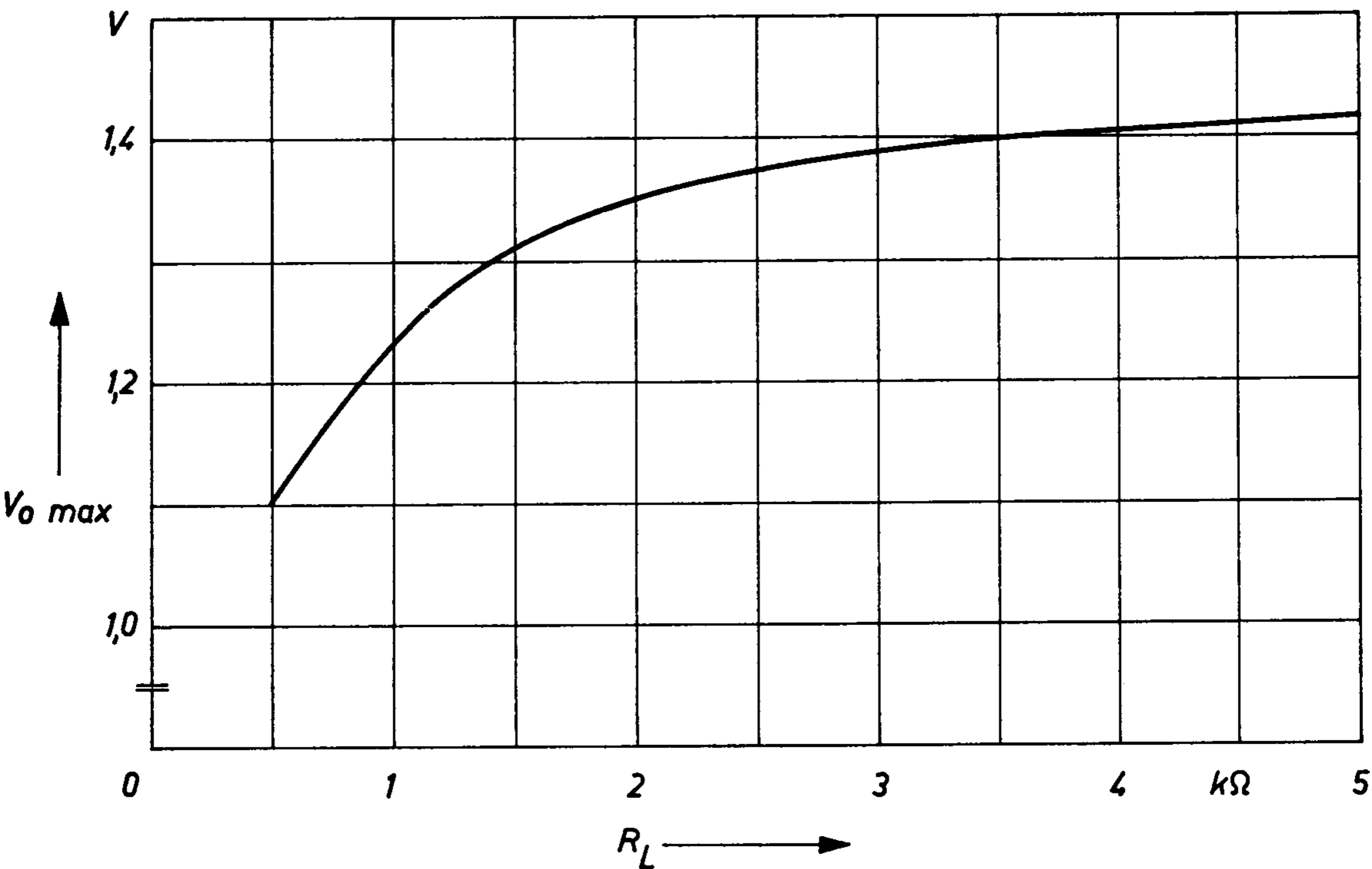


Fig. 1.20

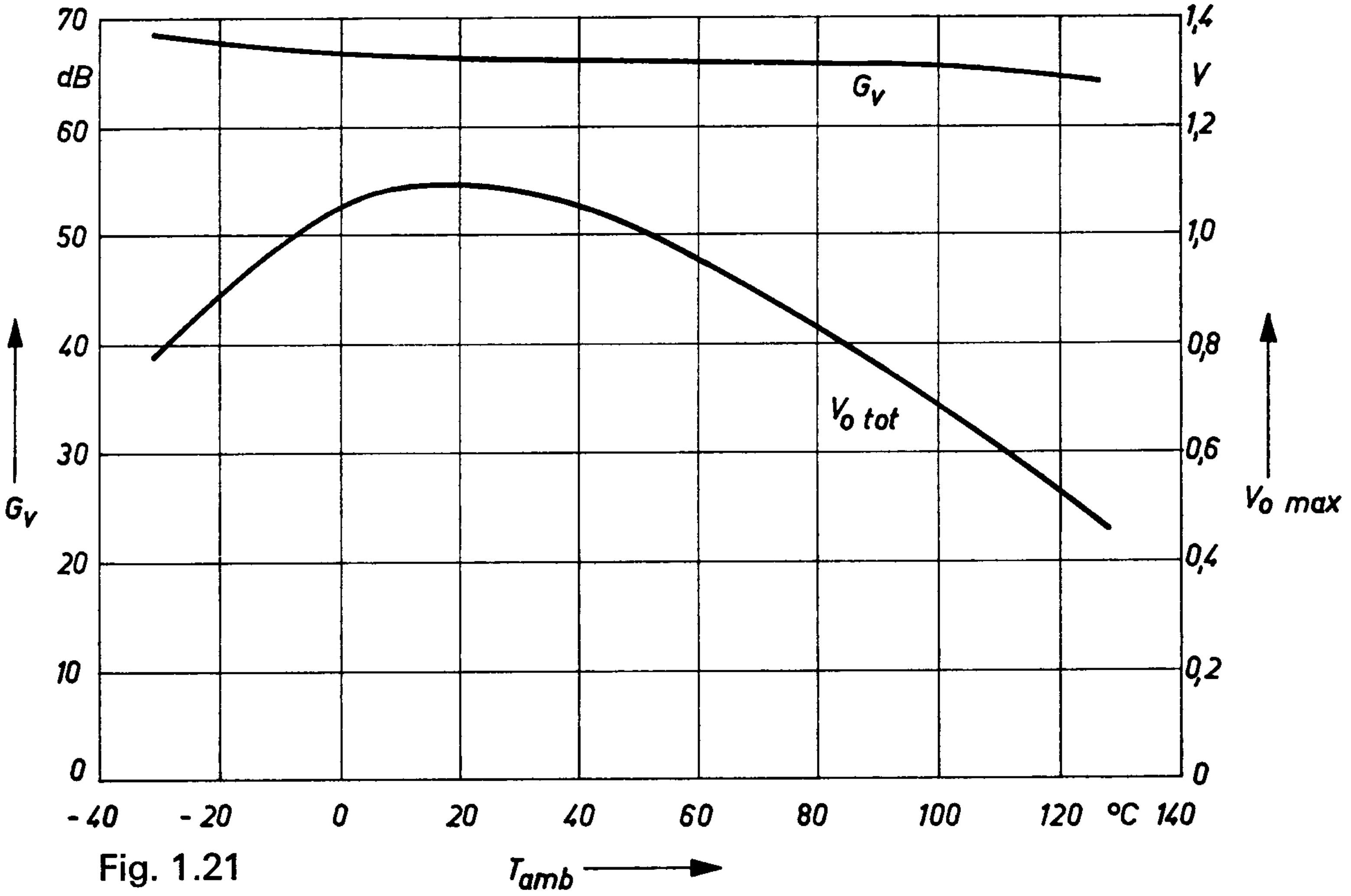


Fig. 1.21

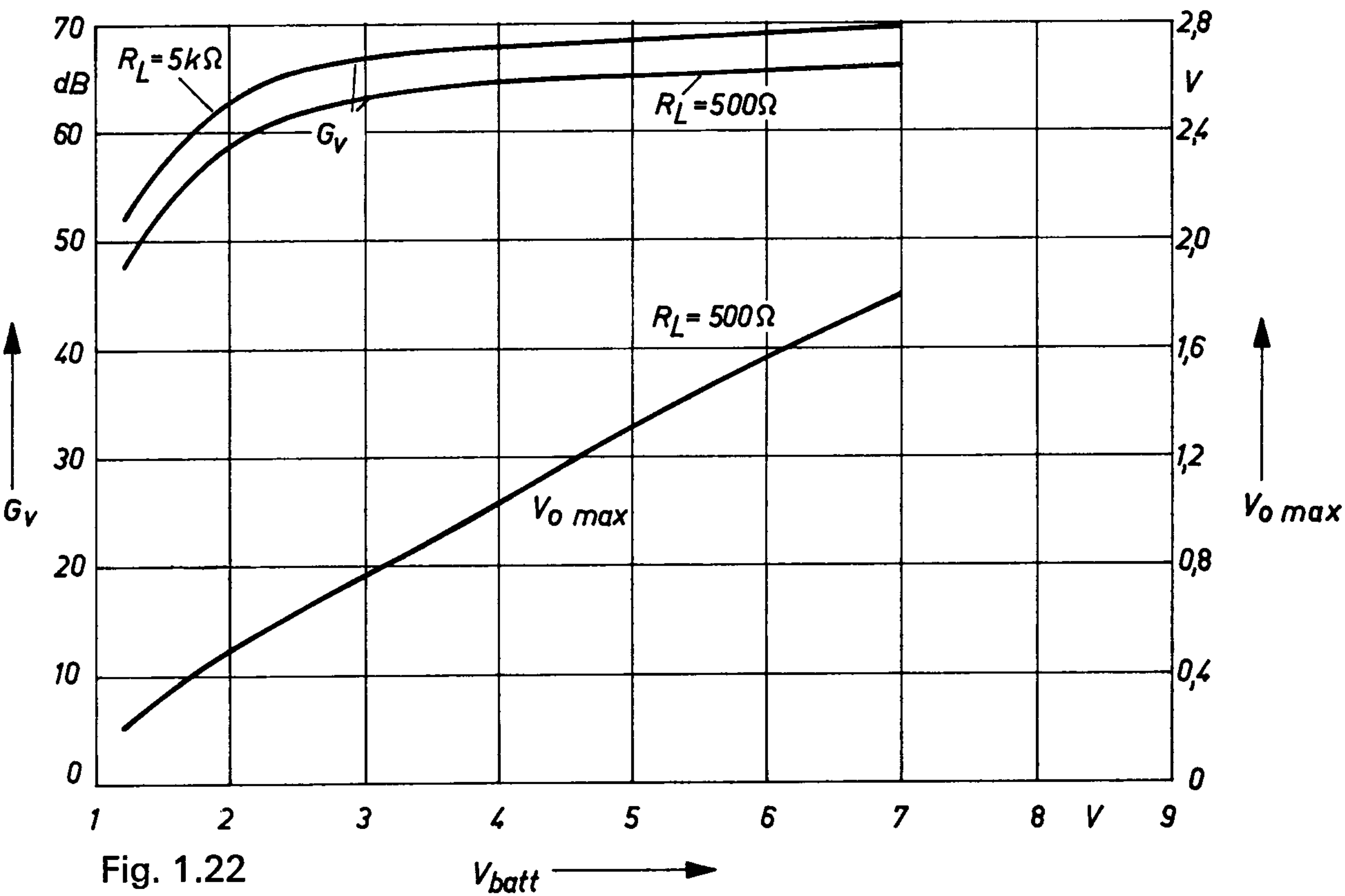


Fig. 1.22

## 2. Choppers and Oscillators

A large number of chopper circuits have already been described, partly even tabulated in former booklets. For that reason only special circuits in this field will be described.

### 2.1 Warning Blinker with a Glow-Discharge Tube

Warning blinkers should be distinguished by a low current consumption and a high efficiency of the luminous source. Therefore, one operates with "cold light", that means, with glow-discharge tubes or neon lamps. Fig. 2.1 shows a circuit supplied by a 6 V battery for generation of an operating voltage necessary for a glow-discharge tube with an ignition voltage of approximately 2500 V.

It is the combination of a chopper and a blocking oscillator. Due to the feedback winding  $N_2$  the chopper oscillates with a frequency of 2.5 kHz. During that part of each period in which the transistor is turned off, the reaction voltage rises at the output as high as the ignition voltage of the neonlamp, and the energy which is stored in the transducer during the other part of each period is discharged via the tube. The glow-discharge tube burns. It would burn constantly if a condenser  $C_1$  would not be incorporated in the feedback loop. Due to this condenser, a blocking oscillator effect is superimposed on the chopper. The chopper can oscillate only until the condenser  $C_1$  is charged by the base current of the transistor. When the charging process is finished, the base current cannot flow any longer and the chopper stops oscillating. Thereafter, the condenser  $C_1$  is discharged and the chopper starts oscillating again. The time constants are adjusted in such a way that the glow-discharge tube burns at intervals of 1 s for a period of 50 ms.

Special care has to be taken for the insulation in the transformer. It is recommended to provide a paper insulation for each layer of the coil  $N_3$  because the coil voltage per turn may — in case of a pulse — amount to 2 V.

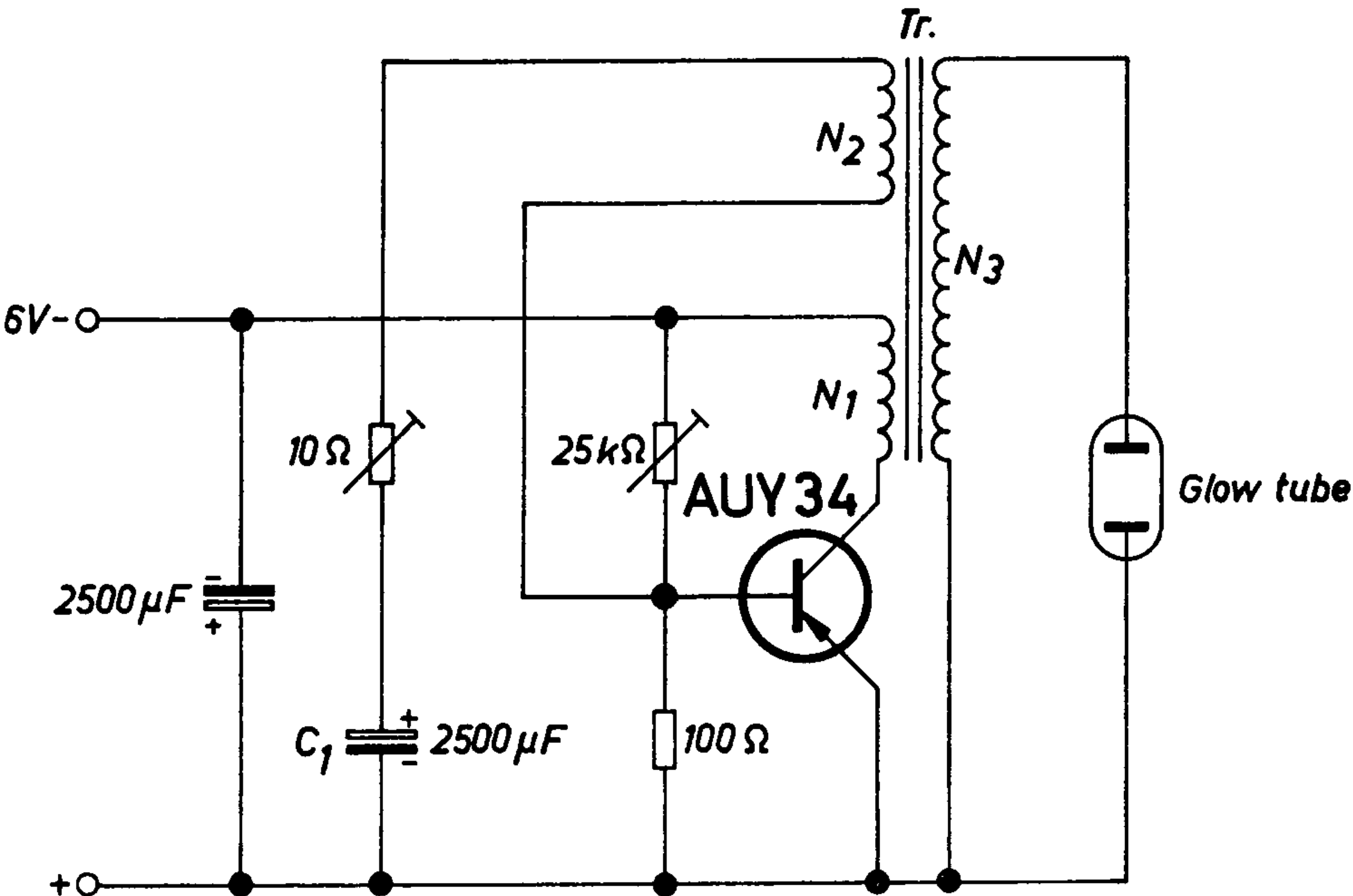


Fig. 2.1

### Technical Data

Supply voltage	6 V ± 15%
Supply current (medium value)	10 mA
Collector peak current	2.6 A
Oscillation frequency	2.5 kHz
Period of lighting of the glow-discharge tube	50 msec
Blinker frequency	1 Hz
Output peak voltage	2500 V
Maximum current of the glow-discharge tube	50 mA
Ignition voltage of the glow-discharge tube	2500 V
Conducting voltage of the glow discharge tube	800 V
Max. operating temperature	60 °C

### Transformer Tr

Siferrite E-core: B 66 241—A 0440—K 026

$N_1 = 50$  turns 0.7 CuL

$N_2 = 12$  turns 0.25 CuL

$N_3 = 1500$  turns 0.12 CuL (well insulated, voltage per turn up to 2 V)

## 2.2 Chopper 20 W / 30 Hz

The chopper shown in Fig. 2.2 has been developed to be used as a calling signal generator. Because of the low oscillator frequency and the high operating voltage a bridge circuit with four germanium transistors AU Y 34 with high maximum voltage ratings had to be used. There is no doubt that the relatively low power output of 20 W in the bridge would have permitted the substitution of two transistors by condensers. This, however, had to result in an uneconomically high capacitance because of the low oscillation frequency. In this circuit the value of the battery voltage appears at each transistor as reverse voltage ( $V_{CEO}$ ).

### Technical Data

Supply voltage	60 V $\pm$ 30%
Supply current	450 mA
Power output	20 W
Output voltage	60 V
Oscillation frequency	30 Hz
Collector peak current	800 mA
Efficiency	75 %

### Transformer Tr

M 65/27 Dyn. sheet IV/0.35, 0.5 L, alternate stacking

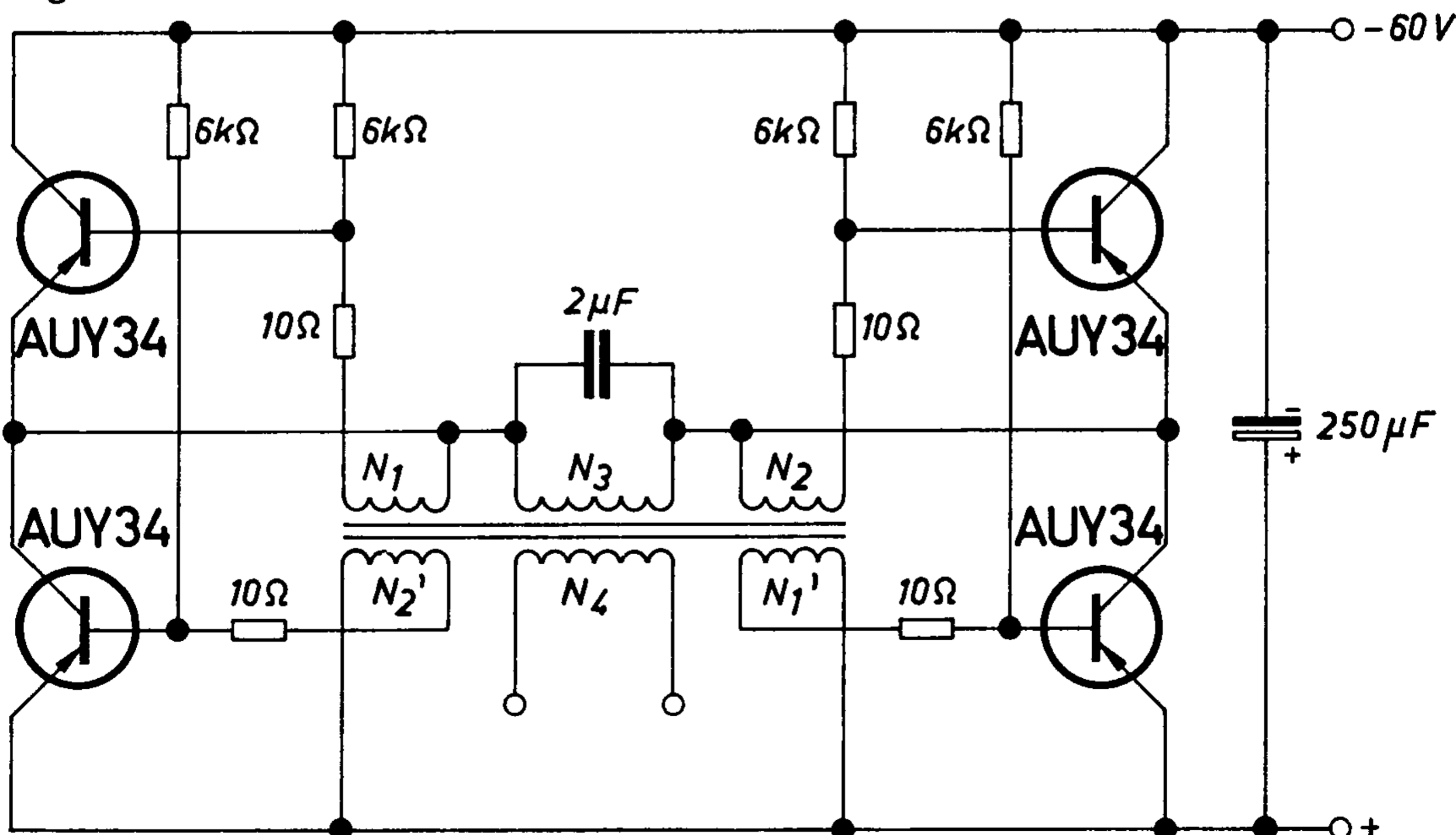
$N_1 = N'_1 = 12$  turns 0.25 CuL joint lead wound coil

$N_2 = N'_2 = 12$  turns 0.25 CuL

$N_3 = 700$  turns 0.4 CuL

$N_4 = 750$  turns 0.3 CuL

Fig. 2.2



## 2.3 Sine-Wave Generator 150 VA / 10 Hz

The power sine wave generator shown in Fig. 2.3 consists of an oscillator stage and an output stage with 2X6 transistors AU Y 22. Because of the low oscillation frequency of 10 Hz very large transformers have to be used. The sine wave generator is suitable for driving servomotors. An equal load of the output transistors is attained by an inverse feedback with one transformer coil at the emitter of each transistor.

Each of the 14 transistors AU Y 22 including also the two transistors of the oscillation stages have to be mounted separately on heatsinks with a thermal resistance of not more than 3.5 °C/W. Then a maximum ambient temperature of 45 °C is permissible.

The balance of the device can be obtained by the potentiometers  $R_1$  and  $R_2$ . They have to be set in such a way that at normal operation the sine-wave voltage will not be distorted. Because of the large size of the circuit only parts of it have been tested.

### Technical Data

Supply voltage	24 V (max. 30 V)
Supply current	10.8 A
Power output	150 VA, $\cos \varphi = 0.8$
Output voltage	60 V
Oscillation frequency	10 Hz
Total efficiency	58 %
Max. ambient temperature	45 °C

### Transformer Tr 1

M 102/35, Dyn. sheet III/0.5, 1 mm L, unit-directional stacking

$N_1 = N'_1 = 320$  turns 0.5 CuL, joint lead wound coil

$N_2 = N'_2 = 360$  turns 0.2 CuL, joint lead wound coil

$N_3 = 3360$  turns 0.18 CuL, insulated

$N_4 = N'_4 = 220$  turns 0.5 CuL, joint lead wound coil

### Transformer Tr 2

EI 170/65, Dyn. sheet III/0.5, alternate stacking

$N_1 = N'_1 = 55$  turns 2.0 CuL, joint lead wound coil

$N_2 = N'_2 = 55$  turns 0.9 CuL, " " " "

$N_3 = N'_3 = 55$  turns 0.9 CuL, " " " "

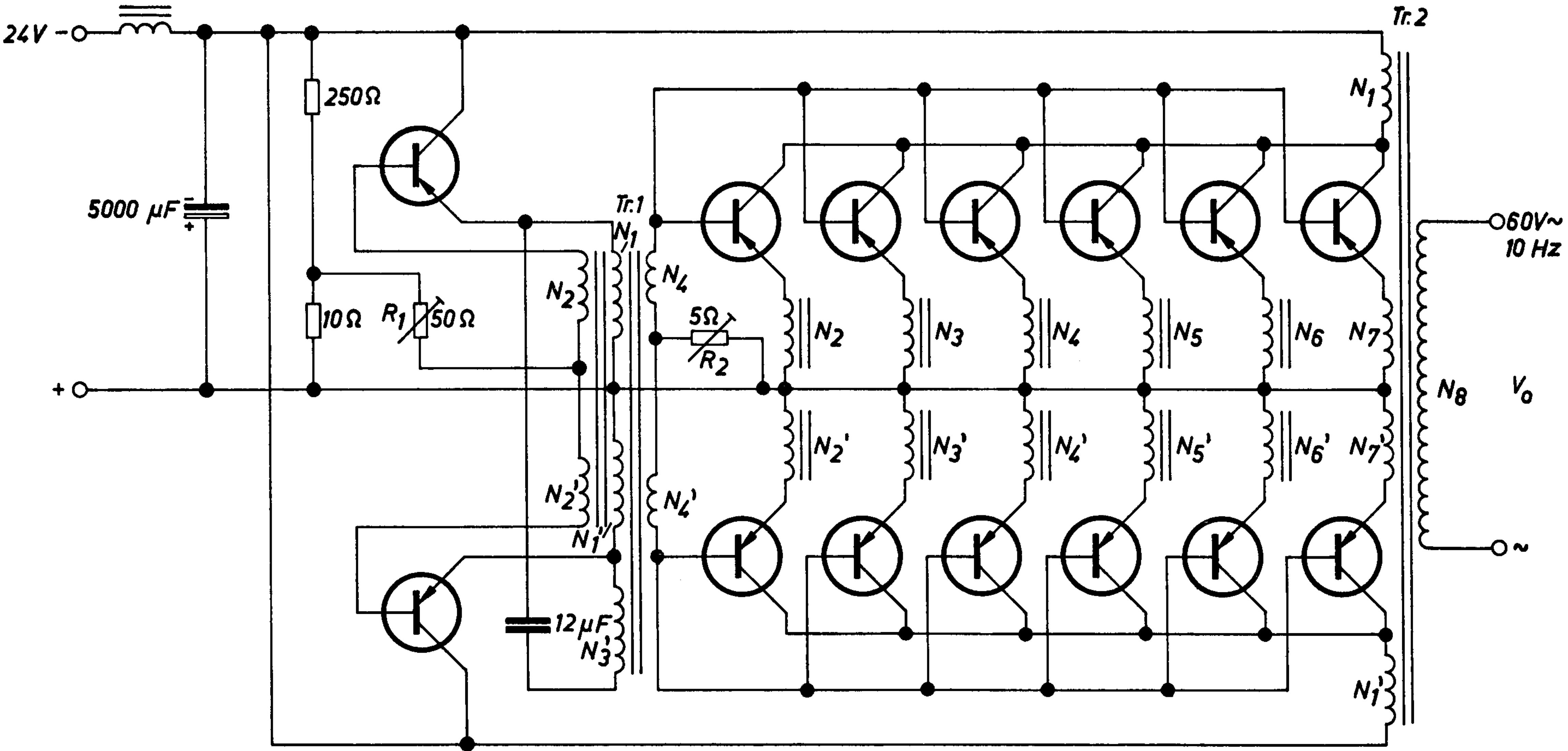
$N_4 = N'_4 = 55$  turns 0.9 CuL, " " " "

$N_5 = N'_5 = 55$  turns 0.9 CuL, " " " "

$N_6 = N'_6 = 55$  turns 0.9 CuL, " " " "

$N_7 = N'_7 = 55$  turns 0.9 CuL, " " " "

$N_8 = 470$  turns 1.0 CuL



All transistors AUY 22

Fig. 2.3

### **3. Multivibrators**

In control engineering multivibrators are generally employed for pulse generating and pulse shaping and as timing units. New components, as for instance, positive temperature coefficient thermistors offer new circuit designs and enlarge the range of applications for multivibrators.

#### **3.1 Astable Multivibrators with Positive Temperature Coefficient Thermistors**

If in the known basic circuit for bistable multivibrators the feedback resistors are replaced by PTC thermistors, the new circuit operates as an astable multivibrator. PTC thermistors are components the resistance of which increases with temperature. The PTC thermistors manufactured by Siemens show a step characteristic, i.e., they considerably change their resistance in a small temperature range, e.g. by the factor  $10^4$  in a temperature range of  $50\text{ }^\circ\text{C}$ .

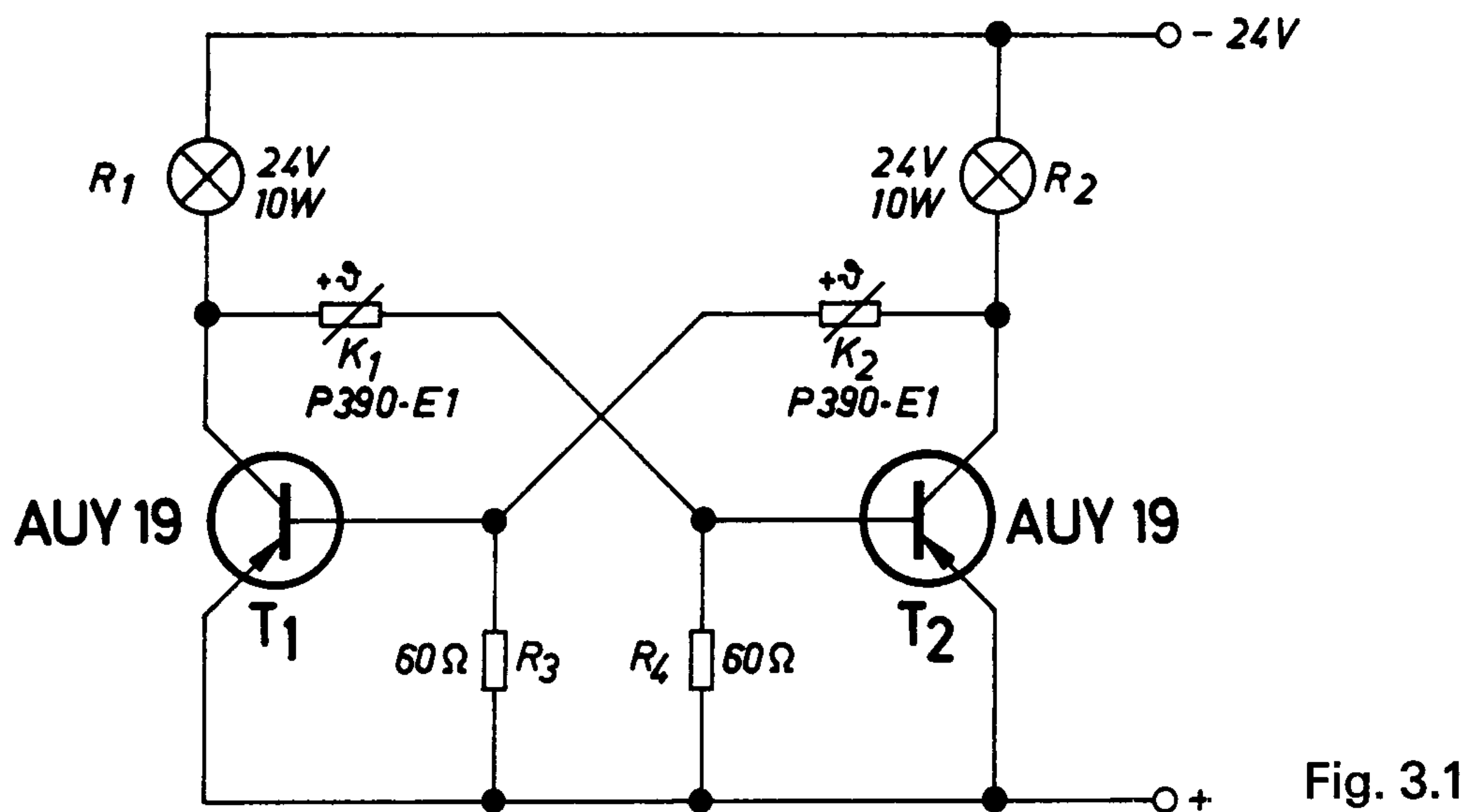
For the description of the astable multivibrator in Fig. 3.1 one has to start with an instantaneous value, for instance, with that value when the transistor  $T_1$  is turned on. Then the transistor  $T_2$  is turned off because there is only a low voltage at the collector of transistor  $T_1$  (cut-off voltage) and, therefore, no base current for transistor  $T_2$  is supplied via the PTC thermistor  $K_1$ . Because of the same reason this PTC thermistor remains cold. PTC thermistor  $K_2$ , however, will be heated up, as a large part of the operating voltage drops at thermistor  $K_2$  which carries the base current necessary for turning on the transistor  $T_1$ . As soon as the Curie temperature of the PTC thermistor is obtained (that is the temperature at which the step in resistance occurs), the transistor  $T_1$  is turned off and the transistor  $T_2$  is turned on because, then, the base current for transistor  $T_2$  is supplied via the low resistive PTC thermistor  $K_1$  which is still at a low temperature. Then, the PTC thermistor  $K_1$  heats up and the process starts again.

The current of the PTC thermistors can be adjusted by the resistors  $R_3$  and  $R_4$ . Hereby, the oscillation frequency of the multivibrator may be varied.



The higher the Curie point of the employed PTC thermistor, the better is the temperature stability of the oscillation frequency. Different from the usual astable multivibrators, the oscillation frequency depends very much on the operating voltage in the arrangement described above because the PTC thermistor current increases with the operating voltage so that the heat-up time decreases.

Therefore, the multivibrator is only suitable for applications with no strict requirements but offers the advantage of being low-resistive. In the circuit shown in Fig. 3.1 the base currents have values of about 15 mA so that high collector currents can be switched. For usual astable multivibrators one would have to use very large capacitors or additional amplifier stages in order to attain this high switching power at low frequencies.

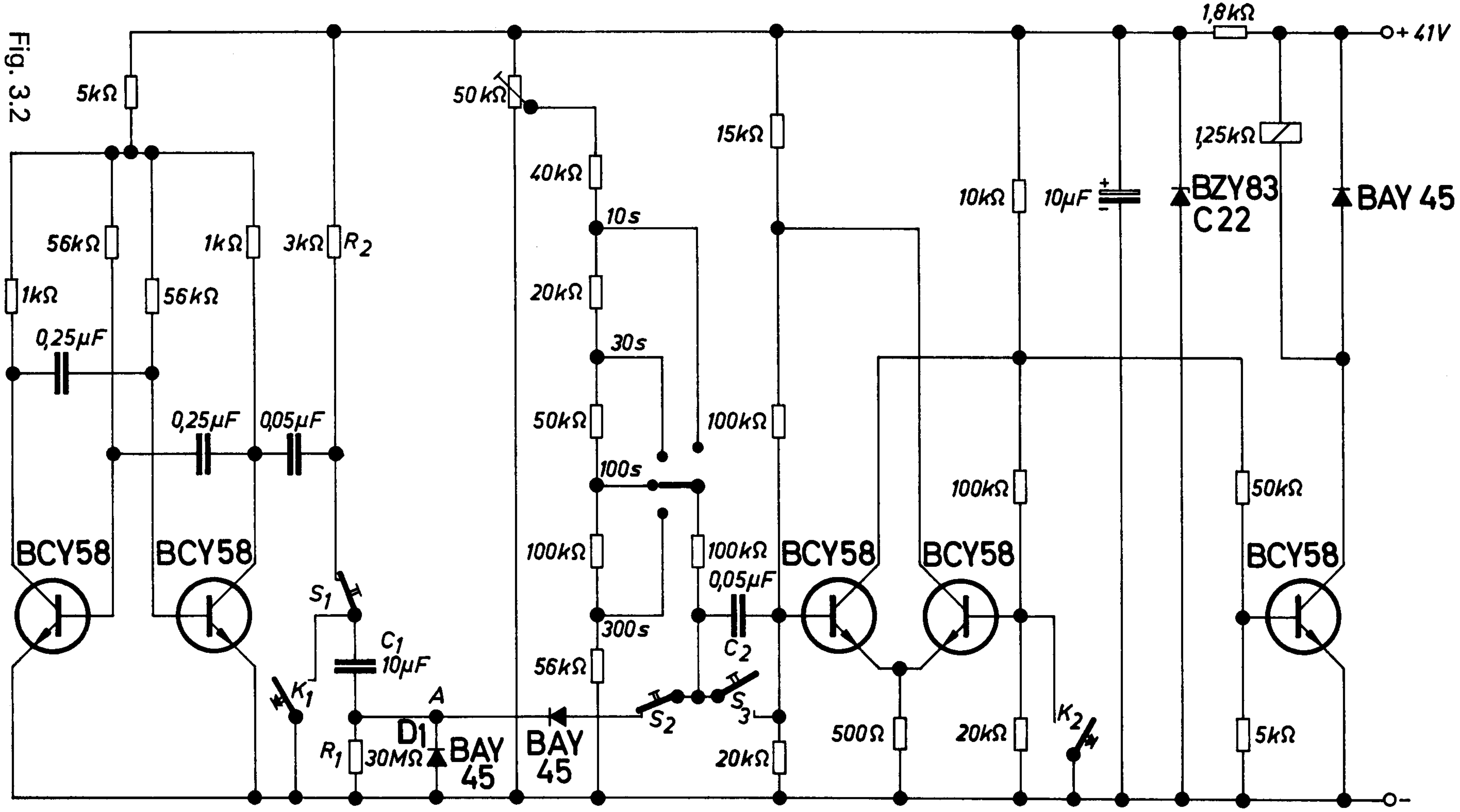


The circuit shown in Fig. 3.1, for instance, is suitable as blinker-signal operator. Two bulbs with 10 W each are alternately blinking at a frequency of 2 Hz.

### 3.2 Timing Pulse Generators 10 to 300 Seconds

Fig. 3.2 indicates a delay circuit appropriate, for example, for the control of washing machines. Time determining is a RC combination the condenser of which is charged with a DC voltage being superimposed by pulses. In this example the delay time is adjusted in stages by a potential divider.

Fig. 3.2



If the switch  $S_1$  and  $S_2$  are closed and switch  $S_3$  is opened, the delay process starts. The condenser  $C_1$  is charged via the resistors  $R_1$  and  $R_2$ . At the first moment, almost the total operating voltage drops at the resistor  $R_1$ . The voltage drop decreases according to an e-function as does the charging current of the condenser. If an amplifier with an adjustable threshold level is connected to this resistor, one obtains a delay circuit which only operates inaccurately. Since the threshold level is only slowly attained, small changes of the threshold level already cause considerable changes of the delay time. Besides, the resistance of the RC combination has to be relatively low because, otherwise, the input resistance of the transistor influences the delay time. For longer delay times a condenser with a very high capacitance is needed and, thus, electrolyt condensers have to be employed. The temperature dependence of the capacitance of these two condensers results in a proportional dependence of the delay time.

For this reason, a multivibrator is connected to the timing circuit, which modulates the charging current in such a way that voltage peaks of both polarities of approximately 5 V are superimposed upon the voltage at the resistor  $R_1$ . At the end of the charging procedure for the condenser  $C_1$ , these voltage peaks are much higher than the DC voltage drop at resistor  $R_1$ , whereby also negative voltage peaks appear at point A. After having reached a certain magnitude, they reverse a bistable multivibrator via the condenser  $C_2$ . The required magnitude of the pulse depends on the amount of the antiphase voltage which is applied to the input of the multivibrator via a resistor chain. Switching occurs as soon as the pulses of negative voltage become a little larger than the adjusted antiphase voltage. This voltage is, in the example of Fig. 3.2, switchable in stages which results in delay times of 10 to 300 seconds. Of course, a potentiometer may also continuously adjust the delay. As indicated in the circuit, in this pulse superimposing device a high resistive RC-combination may be employed so that already long delays can be attained by small capacitances. One may apply, for instance, a paper condenser.

A relay is connected to the bistable multivibrator by an amplifier stage. This relay picks up after the lapse of the delay time.

Before the circuit is put into operation again, the switches  $S_1$  and  $S_2$  have to be opened and the switch  $S_3$  has to be closed.

Coupling capacitor  $C_2$  is discharged with switch  $S_3$ . Besides, the keys  $K_1$  and  $K_2$  have to be pushed. The condenser  $C_1$  of the time determining section is discharged via key  $K_1$  and diode  $D_1$  and the bistable multivibrator is set again to the initial position by key  $K_2$ .

At an ambient temperature of 50 °C and at the maximum delay of 300 s the error is —5%, referred to normal room temperature of 22 °C. At an ambient temperature of 70 °C the error amounts to —10%.

### 3.3 Electronic Warning Device

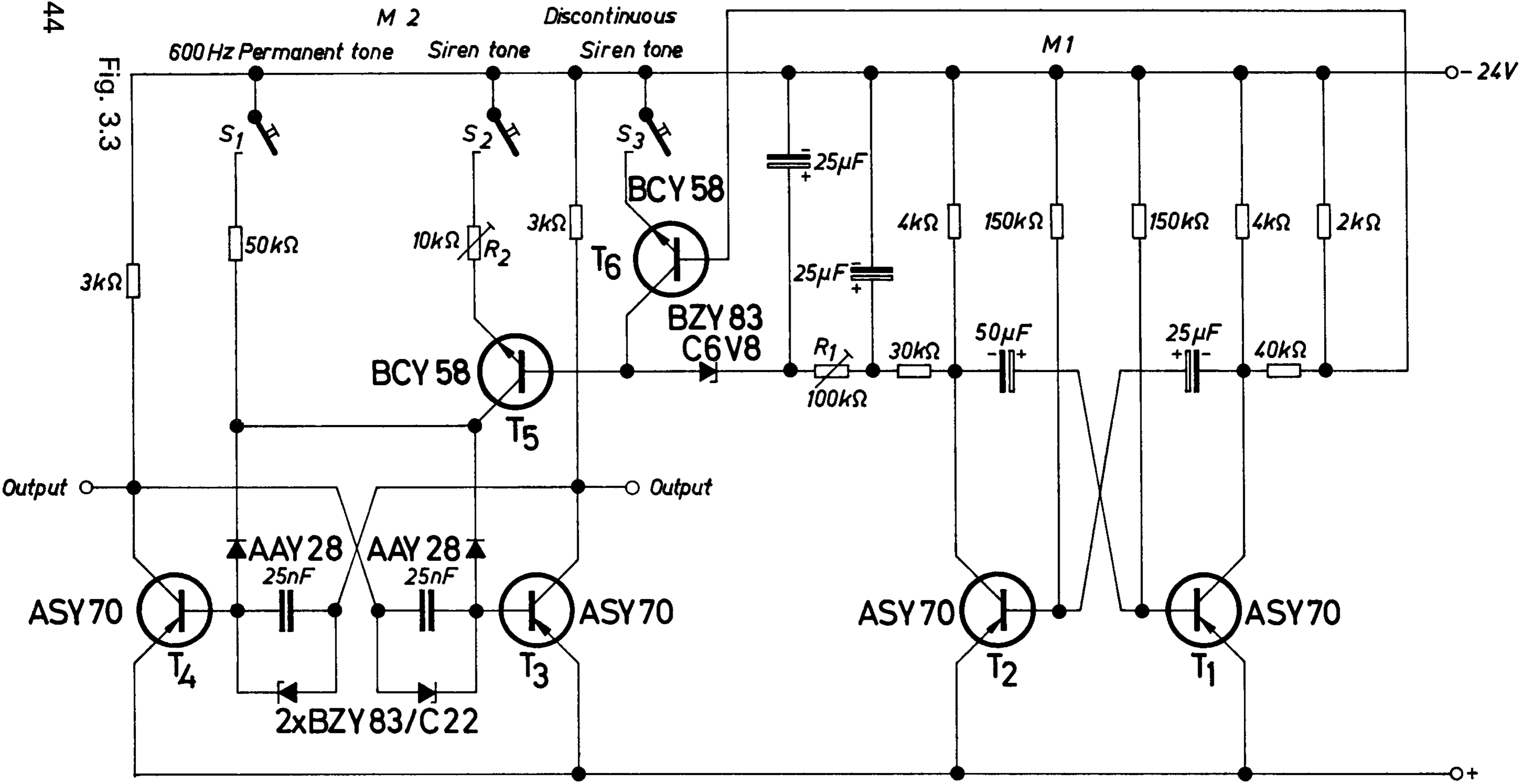
The electronic warning device the circuit of which is shown in Fig. 3.3 generates three different warning tones. They can be initiated by separate switches according to the urgency. A sirene tone is employed for cases of highest danger. At other degrees of danger one may switch on an interrupted sirene tone or a 600 Hz permanent tone. Such arrangements apply in boilerhouses and engine rooms of ships.

The sirene tone is generated by the combination of two multivibrators. Multivibrator  $M_1$  oscillates with a frequency of 1/6 Hz. Its square-wave voltage is converted by means of a filter chain into a pulsing DC voltage with an approximate sine wave form.

In the rhythm of this voltage the supply voltage for the frequency determining section of the multivibrator  $M_2$  will be varied via transistor  $T_5$ . By this means the oscillation frequency of this multivibrator varies in a range from 400 to 1000 Hz. This normal sirene tone is turned on with the switch  $S_2$  which is installed as an error indicator at especially critical points of the controlled plant. The interrupted sirene tone is generated by grounding the driving circuit of the multivibrator  $M_2$  via transistor  $T_6$  in the rhythm of 3 seconds after having closed the switch  $S_3$ . Thus, a sirene tone of 3 sec. occurs always followed by a pause of equal duration.

When switch  $S_1$  is closed, a constant bias voltage is applied to the driving circuit. The warning apparatus generates an uninterrupted tone of 600 Hz. The output signal appears at the collector of the transistors  $T_3$  and  $T_4$  and may be applied to a loudspeaker.

Fig. 3.3



The Zener diode in the coupling between the two multivibrators decreases the DC current portion of the voltage pulsing in the form of a sine-wave.

The sirene tone can be adjusted by two potentiometers. The frequency deviation is varied by potentiometer  $R_1$  and the frequency position by potentiometer  $R_2$ .

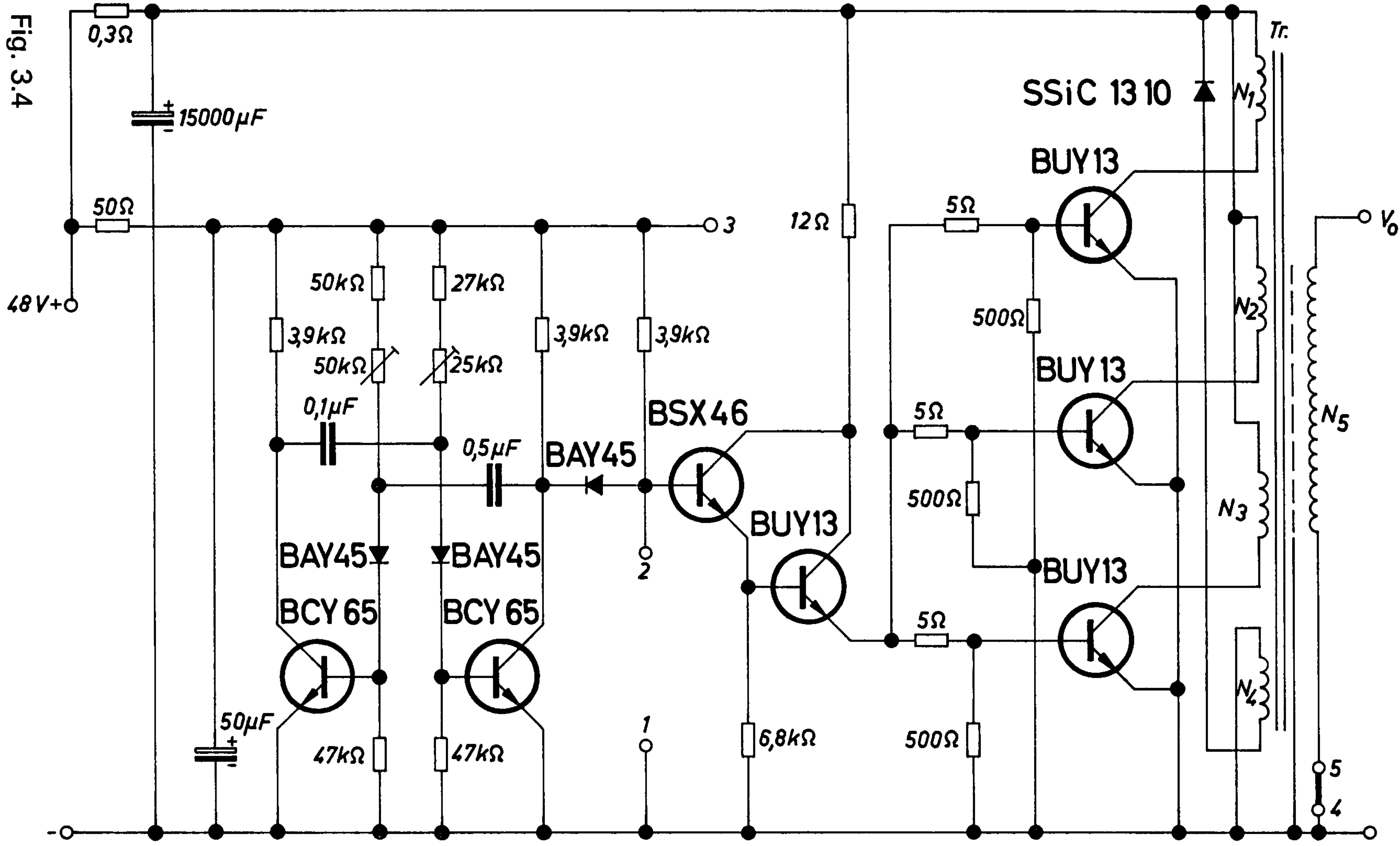
### 3.4 Pulse Generator for 1 kW

Very often pulse generators are required which supply short pulses of high power at certain intervals. Fig. 3.4 shows a circuit, which generates 1 kW output pulses of 2 ms duration at intervals of 20 ms.

An astable multivibrator serves as a timing circuit. Its pulsing time can be set by two potentiometers in the biasing circuits of the transistors. A power stage with 4 silicon power transistors BUY 13 will be driven by an amplifier. Three transistors are connected in parallel and supply the pulse energy at three separate primary coils of the output transformer. The inductive energy of the transformer, which results from turning off the transistors, is fed back to the battery by means of the winding  $N_4$  and a silicon power rectifier. This avoids damaging voltage peaks at the transistors and the efficiency of the equipment will be improved.

The power stage cannot be overloaded and short-circuited. In order to protect the output transistors, the installation of an electronic fuse is required. Wire fuses are too slow-acting for this application.

Fig. 3.5 shows the circuit of a suitable turn-off automatic. The poles 1 to 5 shown in Fig. 3.4 and 3.5 indicate how the electronic fuse has to be inserted into the equipment. Potentiometer  $P_1$  is located in the output circuit. If the load current exceeds a certain permissible value, the bistable multivibrator of the electronic fuse flips and the pulse amplifier (Fig. 3.4) will be short-circuited. Thus, the response threshold of the fuse is adjusted by the potentiometer  $P_1$ . Simultaneously the temperature dependence of the threshold voltage for the multivibrator, which has the value of 4 mV/°C, has to be observed. At low tempera-



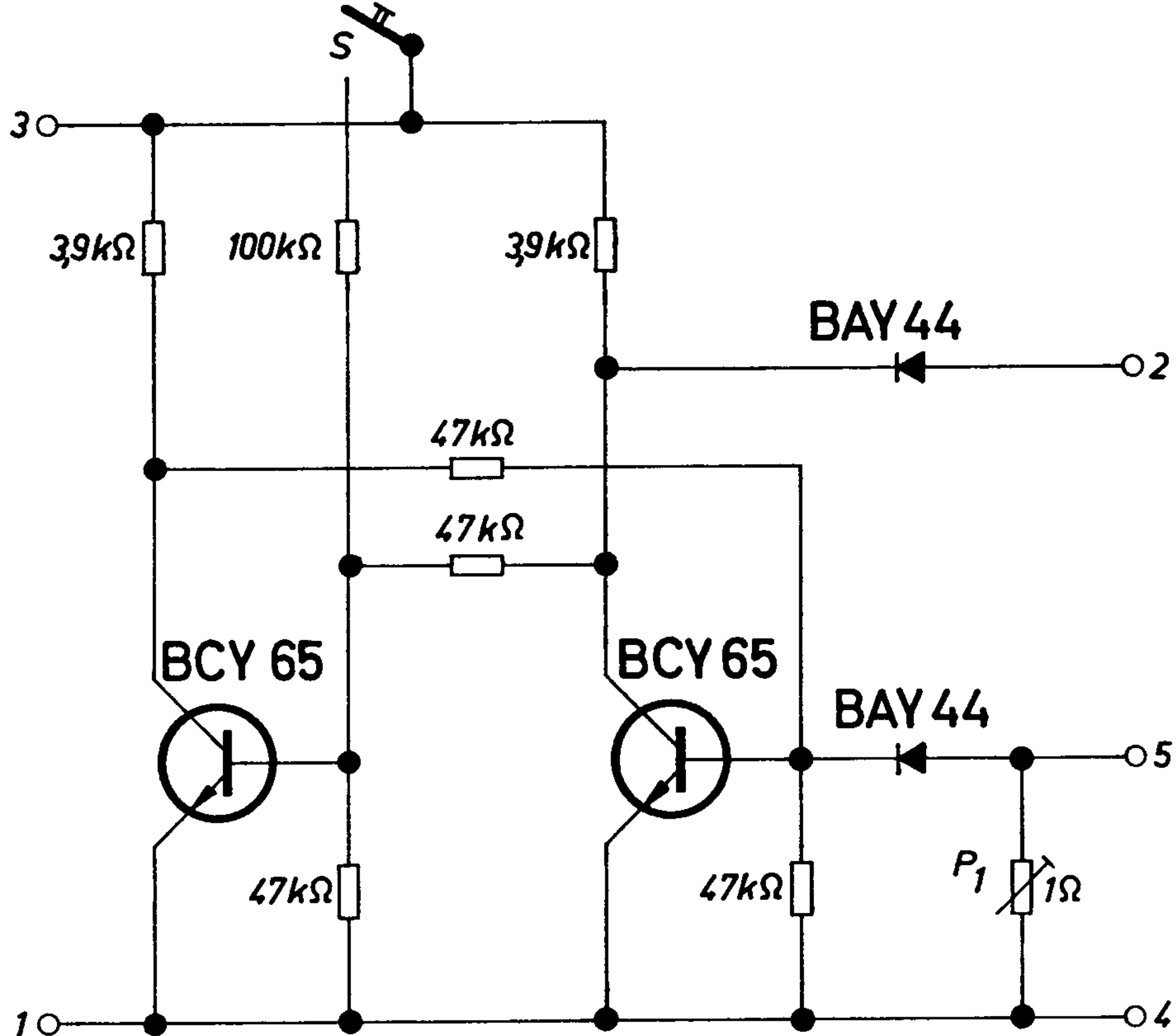


Fig. 3.5

tures, therefore, the fuse responds only at higher values of the load current. For that reason the adjustment of the response threshold has to be set for the lowest expected temperature.

In normal operation the circuit is temperature stable. The maximum permissible ambient temperature is mainly determined by the heat-up of the transformers and by the condensers.

The voltage source is pulsatingly loaded. For that reason, a charging condenser of  $15000 \mu\text{F}$  and a series resistance of  $0.3 \Omega$  serve as protection. The charging condenser and the series resistance of  $0.3 \Omega$  can only be removed if the current source is able to stand reverse currents and to supply a surge power of  $1.3 \text{ kW}$ , and the total internal resistance of the current source (terminals included) does not exceed  $0.1 \Omega$ .

Fig. 3.6 indicates the pulse shape of the output voltage  $V_o$  of the reverse voltage  $V_{\text{CE}}$  at the transistors and that of the entire collector current of the three output transistors under load and in open-circuit. It clearly can be seen how the voltage reaction is clipped by the regain of power with the winding  $N_4$  of the transformer.



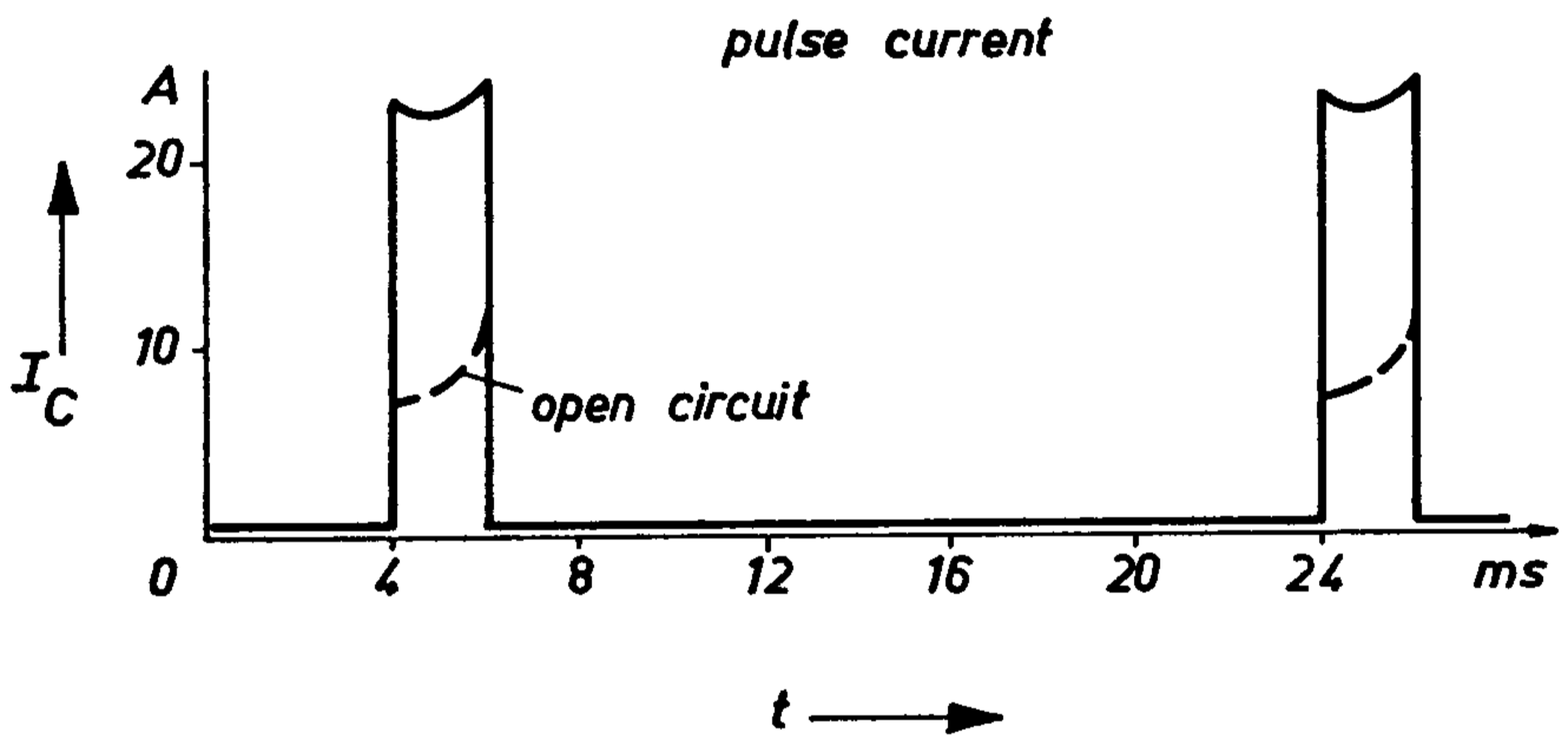
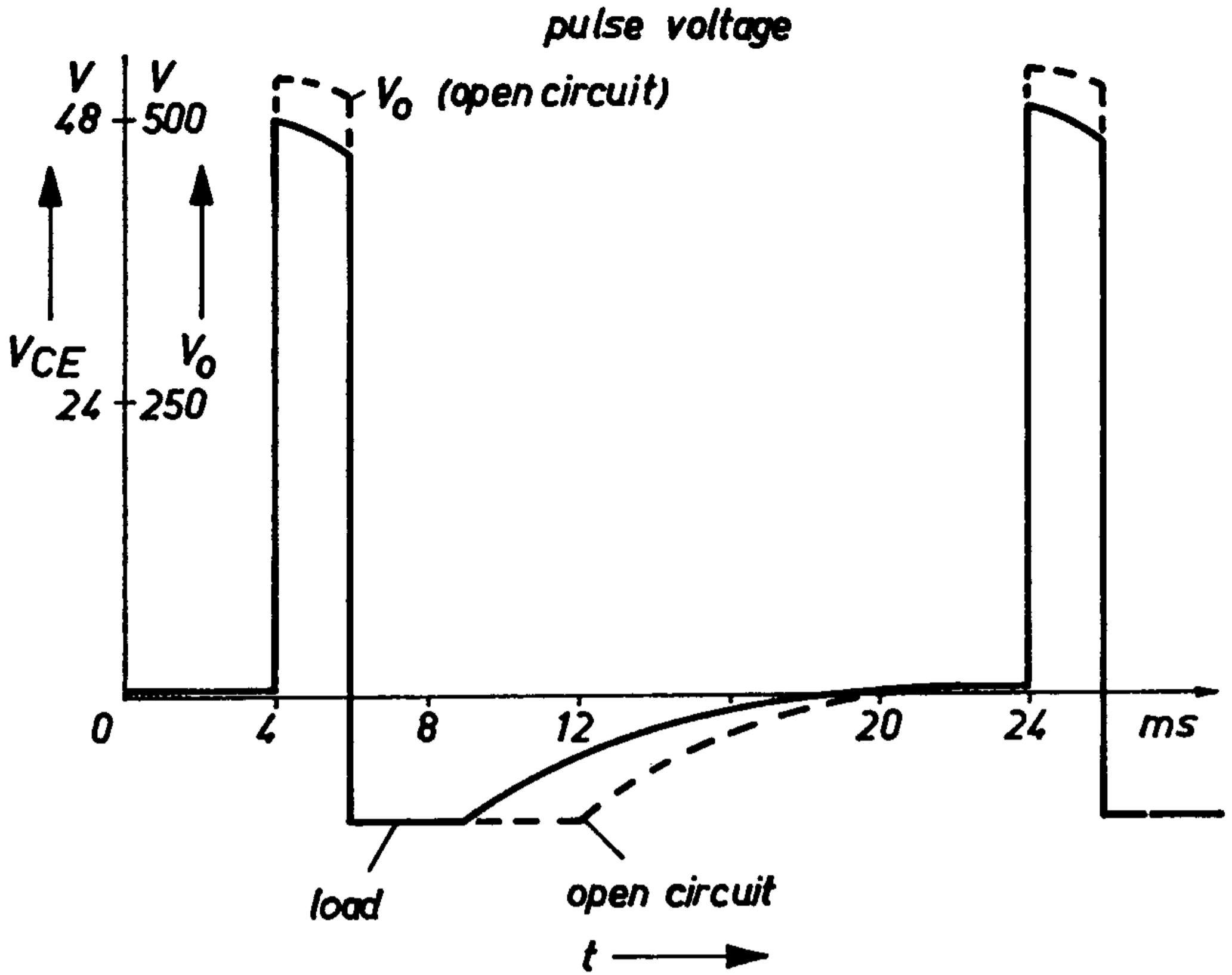


Fig. 3.6

**Technical Data**

Supply voltage	48	V
Supply current for rated load operation	2.9	A
Supply current for open-circuit operation	0.4	A
Max. load resistance	250	$\Omega$
Pulse power output	1	kW
Output pulse voltage	500	V
Pulse duration	2	ms
Pulse sequence	20	ms
Average efficiency	74	%

Total peak current of the three power-stage transistors	24	A
Rise time of the pulse	< 50	$\mu\text{s}$
Required cooling surface for the three power-stage transistors	240	$\text{cm}^2$
Max. ambient temperature	70	$^{\circ}\text{C}$

Transformer Tr: EI 130 a, Dyn. sheet IV/0.35, alternate stacking

$N_1 = N_2 = N_3 = 90$  turns 1.2 CuL  
 $N_4 = 265$  turns 0.9 CuL  
 $N_5 = 1080$  turns 0.6 CuL

## 4. Photo Amplifiers

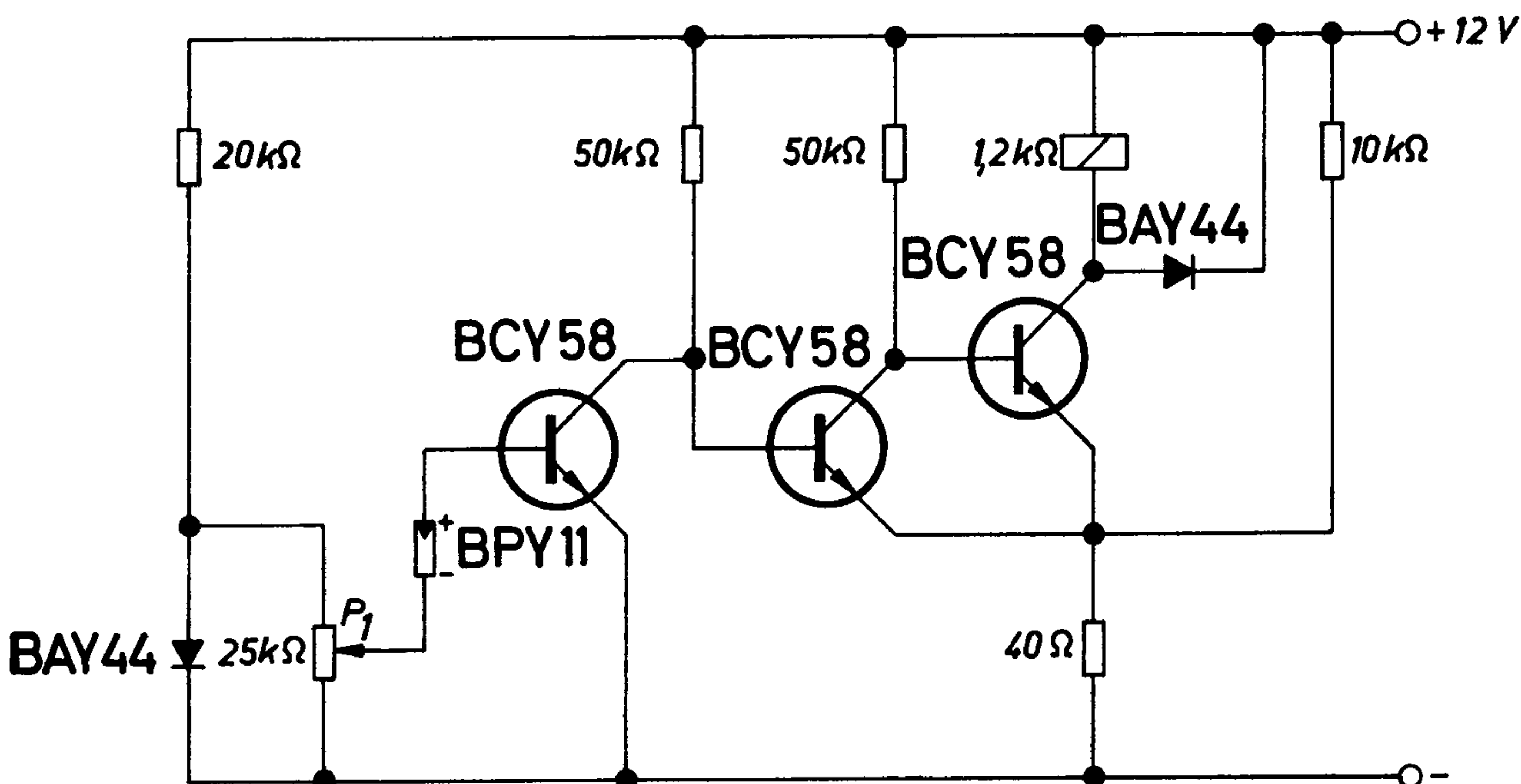
For the conversion of light signals into electrical signals two different semiconductor components are available: the photo diodes and the photo-voltaic cells. The photo diodes mostly are operated as photo resistors e.g. a voltage is applied to the photo diode and the change of the photo current (reverse current) with the intensity of illumination is evaluated. By that means high sensitivity is also attained with photo diodes of small areas. The only disadvantage is a relatively high temperature dependence of these diodes. Since the reverse current of a diode is temperature dependent, a temperature variation results in a similar effect as a change in light intensity.

Silicon photo-voltaic cells mostly have a very small allowable reverse voltage of, for example, 0.5 V, and, therefore, cannot be operated as photo resistors. The direct conversion of luminous energy in electrical energy is applied in these components. In order to obtain a sufficient sensitivity, the photo-voltaic cells are mostly designed with larger photosensitive areas. The silicon photo-voltaic cells, however, do have the advantage to facilitate the design of temperature independent control circuits because cut-off currents are without any influence.

## 4.1 Sensitive Photo-Amplifier with Silicon Transistors

The short-circuit current of silicon photo-voltaic cells is directly proportional to the light intensity. The smaller the current signal which is required for the succeeding amplifier, the more sensitive is a photo amplifier with these cells. For that reason, silicon planar transistors are suitable for input circuits of photo amplifiers. They do have high current amplification and small cut-off currents. Thus, they can be operated at low quiescent points. A very low base current, i.e. photo current, is already sufficient for driving the amplifier. One difficulty, however, still has to be surmounted. The open-circuit voltage of silicon photo-voltaic cells amounts to approx. 0.5 V while the threshold voltage at the base of planar transistors is about 0.7 V. Therefore, in order to overcome this threshold voltage, a certain voltage has to be added to the photo-voltaic cell for driving a planar transistor. As the threshold voltage of transistors is temperature dependent it is recommended to gain this bias voltage by a silicon diode operated in forward direction because the forward voltage of this diode has approximately the same temperature coefficient.

Fig. 4.1



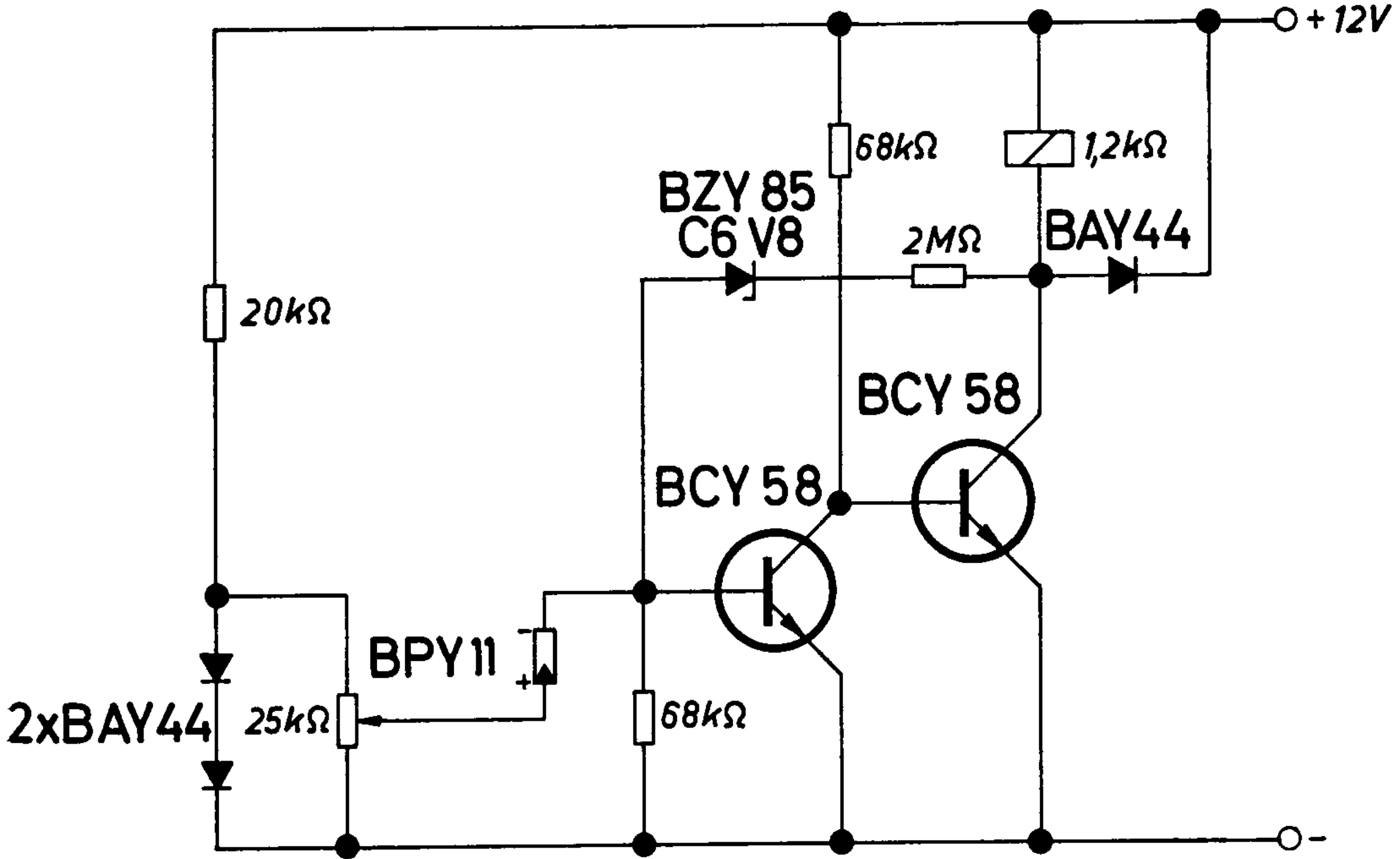


Fig. 4.2

Besides, the bias voltages become considerably independent from the change of the operating voltage because of the steep forward characteristic of these diodes.

Fig. 4.1 shows a three-stage photo amplifier for which the above mentioned method is applied. The potentiometer  $P_1$  sets the bias voltage at such a value that the output relay does not pick up if the photo-voltaic cell is darkened. If suitably adjusted, the relay already picks up at an intensity of illumination of 100 lx. A turn-over characteristic of the circuit is attained by the emitter resistor which belongs both to the second and the third stage. The same sensitivity can also be obtained by the two-stage circuit in Fig. 4.2. In order to warrant that the relay also picks up when the photo-voltaic cell is illuminated, a sufficiently high bias voltage is provided to turn on the first stage at darkness. For that reason, two silicon diodes are connected in series. The voltage of the photo-voltaic cells subtracts from the bias voltage, whereby the input transistor is turned off and the relay picks up. In this case the turn-over characteristic is obtained by a feedback with a zenerdiode and a resistor. Feedback may also be arranged with a resistor of 4 M $\Omega$  only. Then, however, the sensitivity will decrease.

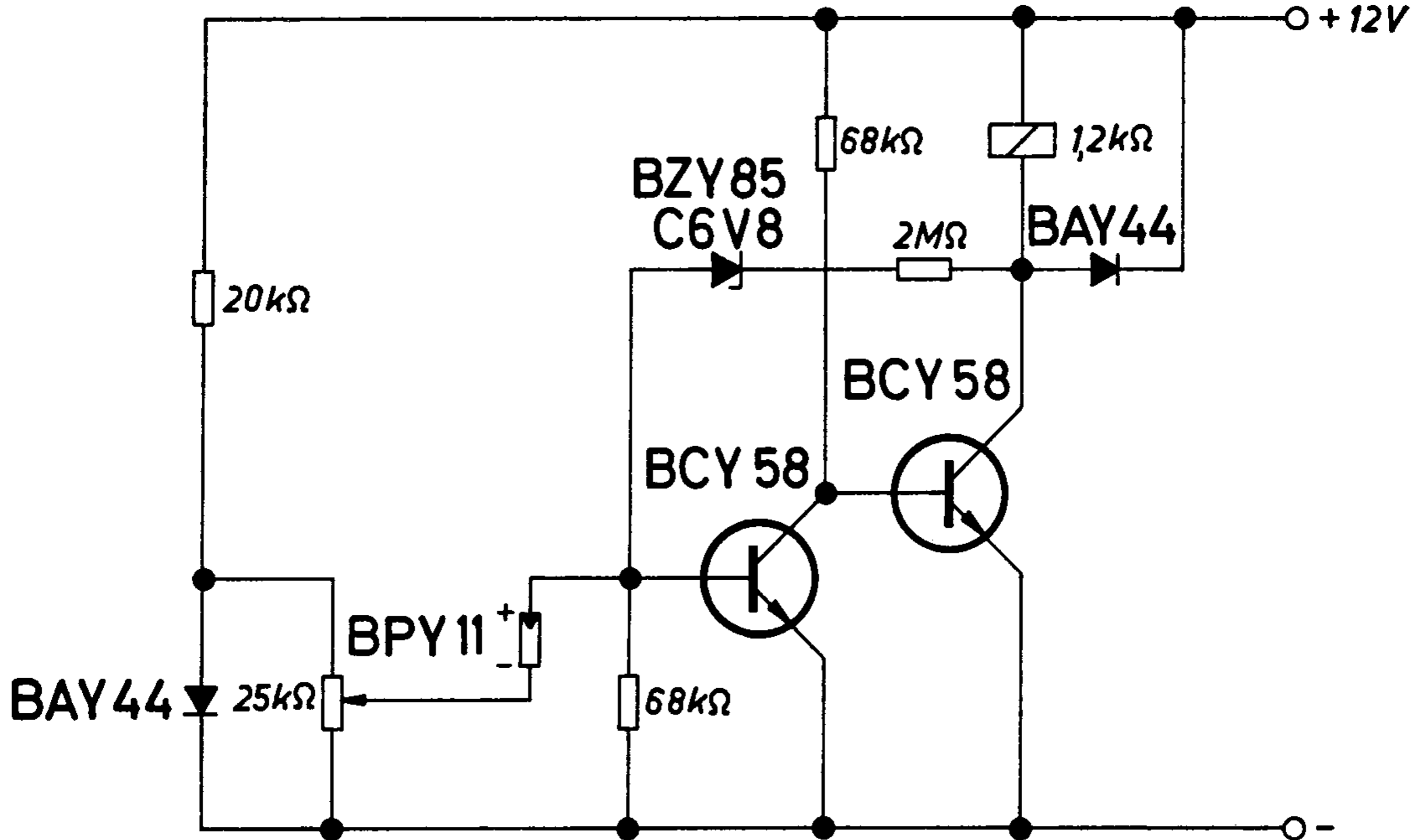


Fig. 4.3

The circuit in Fig. 4.3 differs from that in Fig. 4.2 only in the fact that the photo-voltaic cell BPY 11 gets reverse polarity, and that a smaller bias voltage is adjusted. The input circuit, therefore, corresponds again with that in Fig. 4.1. Since it is a two-stage amplifier in comparison with that circuit, a phase reversal will occur, and the output relay drops out at illumination of the photo-voltaic cell. The sensitivity is again 100 Lx.

## 5. Control Circuits

New semi-conductor components offer new circuit designs in control engineering. The positive temperature coefficient thermistors with a ceramic substrate, which considerably enlarge our program of temperature dependent resistors, are of special importance.

PTC thermistors of semiconducting ceramic material show the effect that the resistance steeply rises within a relatively small temperature range. Temperature coefficients up to  $+50\%/^{\circ}\text{C}$  may occur. That means that the resistance increases by 50% if the temperature rises  $1^{\circ}\text{C}$ . At a temperature rise of  $50^{\circ}\text{C}$  the resistance increases by  $10^4$ . The semiconducting material of these PTC thermistors is a doped ferroelectric ceramic e.g. bariumtitanate ( $\text{Ba Ti O}_3$ ). The ceramic which has insulation characteristics obtains a special kind of conductivity by doping. Above the so-called Curie point the resistance, as already described, rises with a dropping dielectric constant.

The Curie temperature itself and the range of the rise in resistance can be determined by the composition of the material. For the time being, Siemens supplies PTC thermistors with Curie temperatures of 40, 60, 80, 120 and  $180^{\circ}\text{C}$ . The components have the shape of disks. A terminal wire is soldered to both sides of the disk. In addition there exist PTC thermistor pills encapsulated in a glass case similar to that which is used for diodes. This type of PTC thermistors can be used especially for liquids. Three examples will follow. The slope of the current rise in resistance decreases proportional to the voltage because the PTC thermistors have a varistor effect.

PTC thermistors always can be used in cases where a change of temperature should be converted into an electrical signal. In that case it is possible to use the entire rise of resistance (switching effect) or the continuous rise within a small temperature range.

A change of the ambient medium also causes an electric signal e.g. for transitions from air to liquids or for transitions from stationary to flowing air or liquids, respectively. For the case first mentioned, examples are given in the chapters 5.2, 5.3 and 5.4.

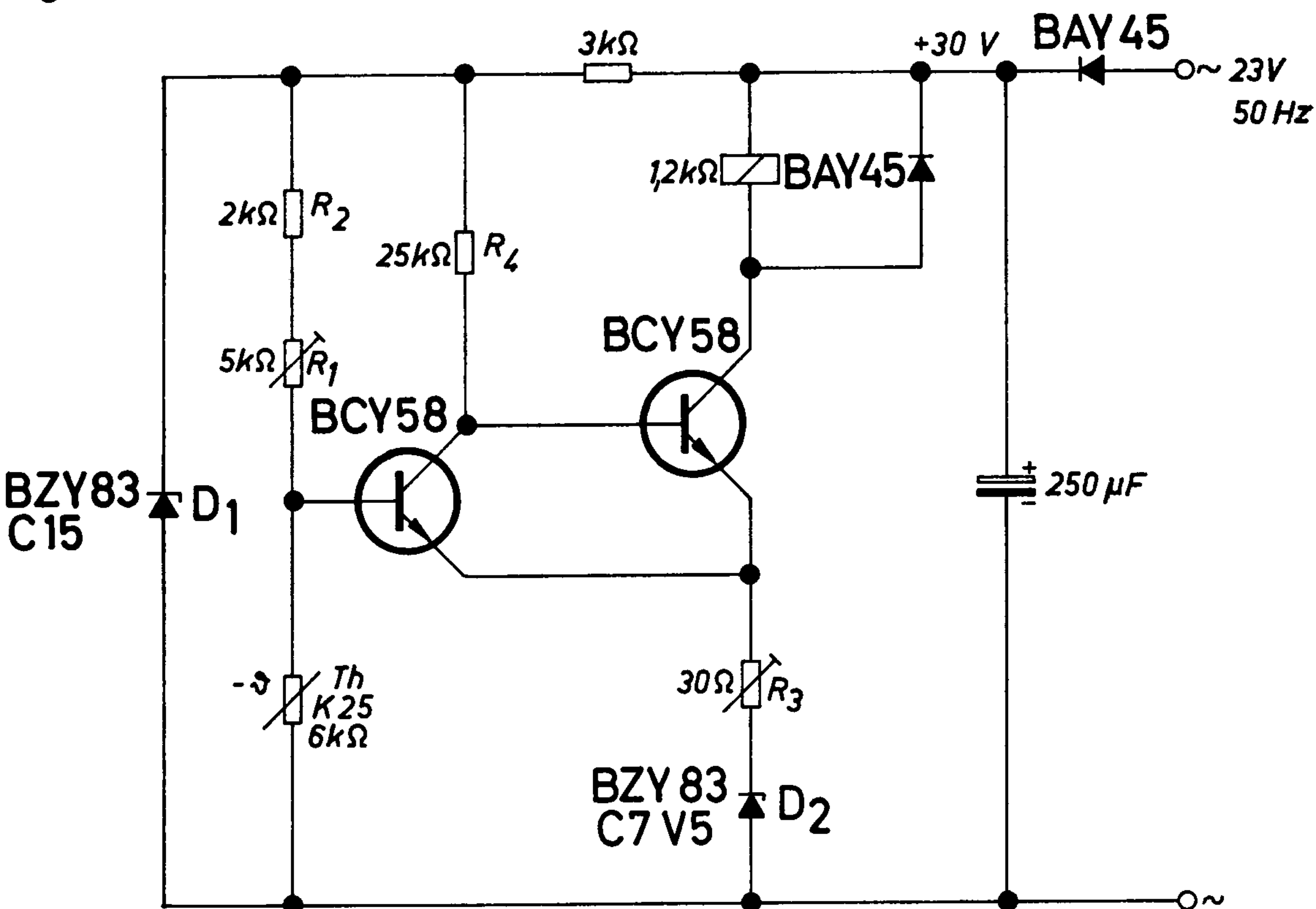
The magnetoresistor which is described in chapter 6 is another new semiconductor component for control engineering.

## 5.1 Control Circuit for Room Heaters

For the heating control in rooms, temperature control circuits are required, the response threshold of which is approximately set at normal room temperature of 22 °C within a small range. Besides the switching accuracy, the difference between the turn-on and the turn-off temperature, has to be adjustable in order to compensate for the different inertness of the various heating systems. In the circuit shown in Fig. 5.1 the threshold temperature can be set between 20 and 40 °C and the switching accuracy between  $\pm 0.6$  and  $\pm 3$  °C.

The resistors  $R_1$ ,  $R_2$  and the NTC thermistor  $Th$  form a potential divider which is fed by a voltage stabilized by a zener diode. This zener diode is to be mounted to a heatsink in order to keep its operating temperature low. As the zener voltage is dependent on temperature, the control accuracy decreases according to the operating temperature of the zener diode.

Fig. 5.1



The emitter potential of the amplifier which is connected to the potential divider is raised to the value of + 7.5 V by means of a zener diode. The transistor only turns on when a voltage higher than this threshold voltage appears at the base of the first transistor. This occurs when the resistance of the NTC thermistor increases, that means, when the ambient temperature drops below a certain value. This threshold value can be set by the potentiometer  $R_1$ . After the first transistor has been turned on, the second one turns off, and the relay drops out. The heating can be switched by a normally closed contact of the relay. Resistor  $R_3$  in the emitter terminal sets the switching accuracy. At the maximum value of 30  $\Omega$  an accuracy of  $\pm 3$  °C is attained. Without the resistor the optimum accuracy is  $\pm 0.6$  °C. The switching function of the relay can be reversed with reference to the ambient temperature if the NTC thermistor and the resistors in the potential divider are exchanged with each other. This arrangement has the advantage that the sensor can be connected to chassis. Another advantage is that the heating is turned off, if lead disconnection occurs between sensor and switching amplifier, in other words, if the terminal between potential divider and the base of the input transistor is broken. In this case, the second transistor is turned on via resistor  $R_4$ , the relay picks up, and the heating is turned off because, as already mentioned, a normally closed contact of the relay is used for this purpose.

### Technical Data

Supply voltage	30 V DC or 23 V AC, 50 Hz
Switching range	20 ° to 40 °C
Switching accuracy	$\pm 0.6$ to $\pm 3$ °C



## 5.2 Leakage Indicators

Leakage indicators are required for various liquid containers e.g. for oil transformers and oil tanks. In the first case, a certain oil level has to be controlled. In the second case, double-walled oil tanks are used; the hollow space between the two tanks is filled with water. If there is a leak in the exterior tank, water runs outward; if there is a leak in the interior tank, water flows into the oil tank which is never completely filled. Consequently, the water level drops in both cases. For that reason, an indicator can be used for control which is similar to that in the mentioned oil transformer.

PTC thermistors are suitable for this purpose. At operation in open air and in liquids the different heat conduction and the resulting variation of the PTC thermistor resistance is used for these

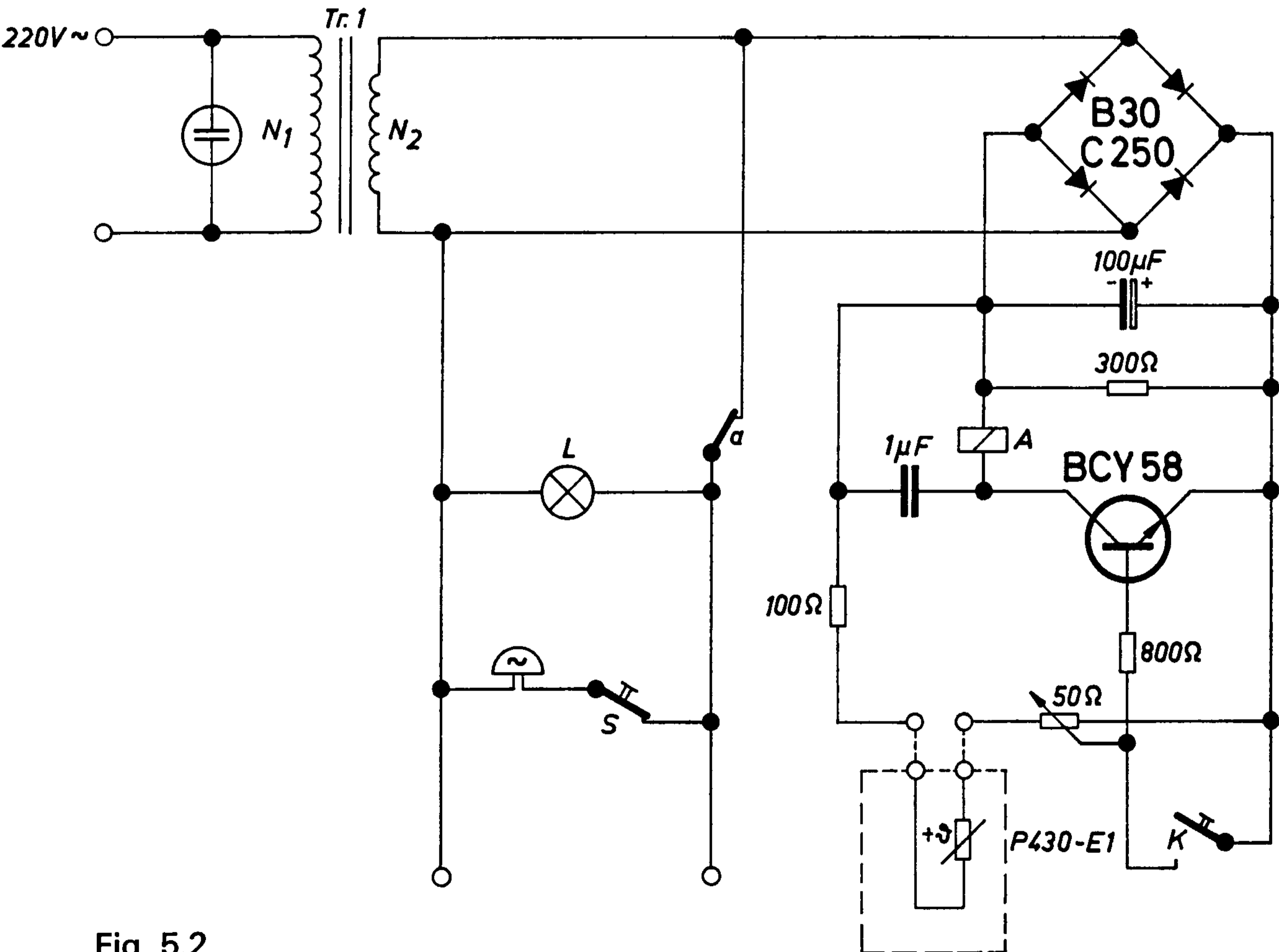


Fig. 5.2

applications. Fig. 5.2 shows the circuit. PTC thermistor P 430 E-1 is attached to a suitable spot on the tank. As long as it is surrounded by liquid, the PTC thermistor is cold i.e. low resistive. At the transistor base a positive voltage is attained which turns on the transistor. The relay is activated. As soon as the liquid-level drops and the PTC thermistor is operated in open air it heats up and the resistance increases. The voltage at the center tap of the potential divider, which the base of the transistor is connected with, shifts in such a way that the transistor will be turned off, and the relay will drop out. Thus, for example, a warning device may be released. In this example the warning device consists of an indicator lamp and an alarm clock which can be turned off by switch S. Readiness for operation can be tested by key *K* and will be indicated by a glow-discharge lamp in the power supply.

Transformer Tr 1

EI 54/18, Dyn. sheet IV/0.35, alternate stacking

$N_1 = 2810$  turns 0.13 CuL

$N_2 = 250$  turns 0.35 CuL

Relay A: Trls 154c according to TBv 65420/93a

### **5.3 Overfill Control for Tanks by means of Positive Temperature Coefficient Thermistors**

In order to avoid that oil flows out while the oil tanks are refilled, an overfill control is required which locks the inlet valve as soon as the tank is filled up.

For reasons of economy, the sensor which is installed in the tank has to be simple and inexpensive. The amplifier mounted in the tank truck, however, is allowed to be more costly. In the circuit shown in Fig. 5.3 only the PTC thermistor has to be installed in the tank. While the tank is filled, it is connected to the amplifier by a two-wire lead. Sensor and amplifier are fed by a power supply consisting of a chopper and a voltage stabilizing circuit. The PTC thermistor is located in a branch of a voltage divider which is fed by the stabilized voltage. When the device is turned on, the valve of the tank stays shut as long as the PTC thermistor first operated in air, reaches its final temperature. This temperature is higher than the so-called Curie temperature. The PTC thermistor, therefore, becomes high resistive. For that reason the transistor  $T_1$  obtains no base current so that  $T_1$  is turned off.

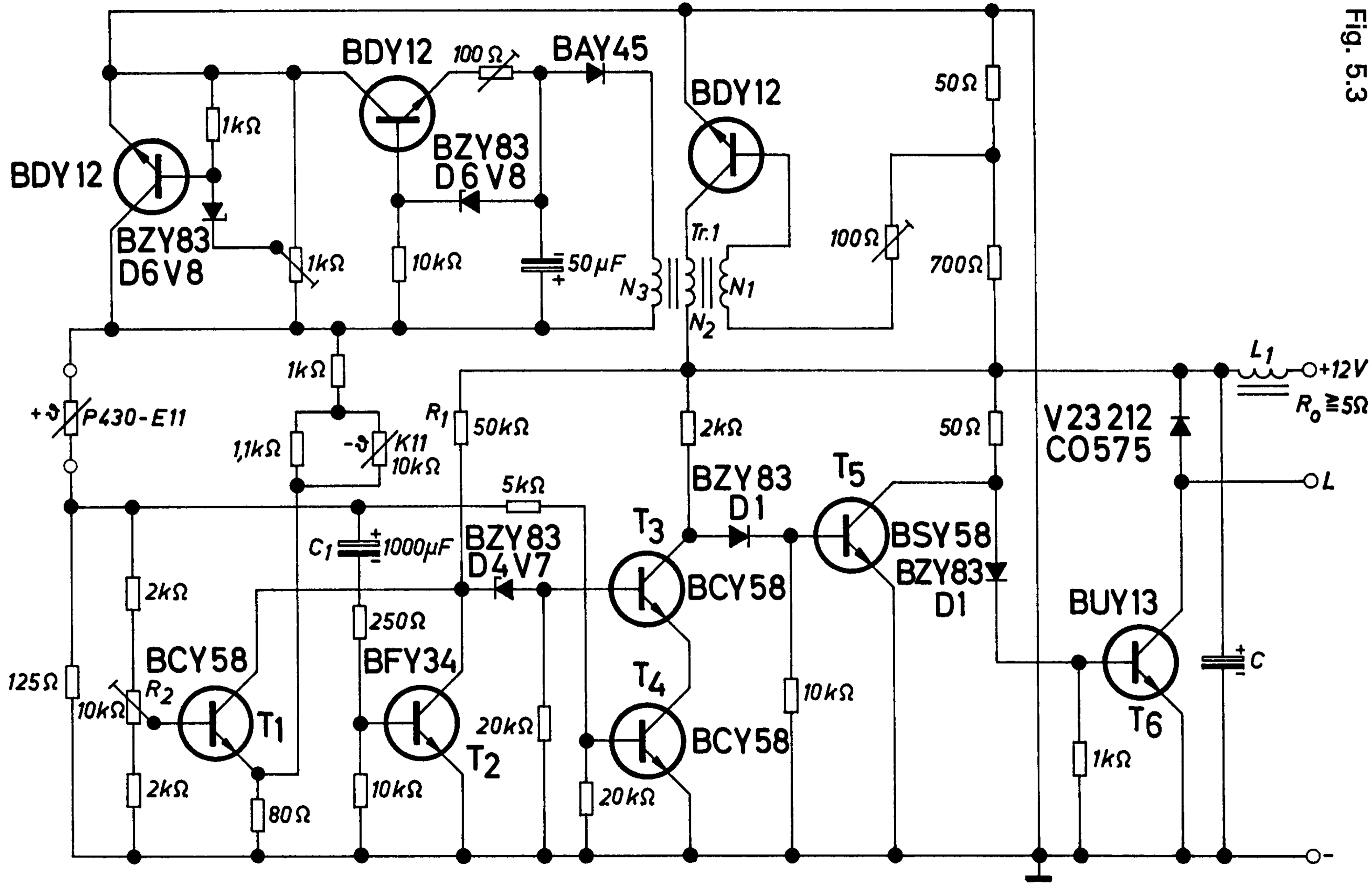


Fig. 5.3

Transistor  $T_3$  is coupled to transistor  $T_1$  by a zenerdiode via resistor  $R_1$ . Transistor  $T_3$  is turned on as long as transistor  $T_1$  is turned off. Then, transistor  $T_5$  is turned off, transistor  $T_6$  is turned on, and the valve is open.

If the liquid level rises so far that the PTC thermistor is immersed, the latter becomes cold and low-resistive. The potential at the center tap of the voltage divider shifts in such a way that transistor  $T_1$  is turned on. Hereby, the input of the transistor  $T_3$  is short-circuited and the valve is shut by the transistors  $T_5$  and  $T_6$ . The silicon diodes in the base lead of the transistors  $T_5$  and  $T_6$  increase the input threshold voltage of these transistors in order to turn them off safely when the preceding transistor is turned on.

The threshold level of the amplifier is set by the potentiometer  $R_2$ . It depends on the emitter bias voltage of the first transistor. The emitter bias voltage is tapped off at a voltage divider which also is connected to the stabilized voltage. This voltage divider also contains a NTC thermistor which stabilizes the threshold value of the amplifier for temperature changes.

In order to turn the valve off quickly when the PTC thermistor is immersed in the liquid, a supplementary stage with the transistor  $T_2$  is provided. This transistor  $T_2$  already is turned on at a slight change of the PTC thermistor current by the charging current pulse for the condenser  $C_1$ . Transistor  $T_2$ , consequently, initiates the same effect as transistor  $T_1$  after a stronger cooling of the PTC thermistor.

For the design of the circuit great care has been taken for highest possible reliability in case of a breakdown. If, for example, an interruption occurs between sensor and amplifier, transistor  $T_4$  will receive no driver current and will be turned off as well as transistor  $T_6$ . At normal operation, the small current flowing through the hot PTC thermistor is already sufficient for turning on this transistor. Consequently, transistor  $T_3$  which is connected in series to  $T_4$  only can operate if the PTC thermistor is connected to the amplifier. The circuit is protected against short-circuits in the cable between PTC thermistor and amplifier because a small resistor at the input appears as a cold PTC thermistor. This is the case of operation of the PTC thermistor immersed in the liquid when the valve is shut. At the output a high power silicon transistor is employed. In spite of that, it is necessary to clip the reaction voltage by a silicon rectifier when the valve is shut.

## Technical Data

Supply voltage	12 V $\pm$ 30 %
Output resistance (valve)	$> 5 \Omega$
Permissible oil temperature	— 25 to + 60 °C
Permissible ambient temperature	— 25 to + 75 °C
Response time	$< 1$ s

## Transformer Tr 1

Siferrite Pot Core B 65581-A 0000-R 026

$N_1 = 5$  turns 0.4 CuL

$N_2 = 30$  turns 0.5 CuL

$N_3 = 40$  turns 0.5 CuL

## 5.4 Electronic Synchronizing Circuit

In order to be able to connect two AC generators they have to have equal numbers of revolutions and equal instantaneous values of the output voltage (same phase). The well-known synchronized bright and synchronized dark method with light bulbs still can be used for 50 Hz generators because the above mentioned requirement is given for a sufficient length of time (1/10 sec). For generators with a much higher frequency, e.g. 400 Hz, an electronic synchronizing circuit is required. In the circuit in Fig. 5.4 the two output voltages of the generators are applied to each of both inputs of a differential amplifier via transformers. As long as the instantaneous values of the voltages are not in accordance with each other, the differential amplifier sends out pulses the width of which corresponds to the phase difference between the two input voltages. These pulses have the shape of square waves because the input signals will be clipped by means of zener diodes. Dependent on the phase position of the two input voltages, the output pulses of the differential amplifier may have a negative or positive potential. They are, therefore, rectified in a bridge circuit and applied to a switching amplifier. In order to avoid a response of the amplifier already after a single zero crossing of the phase difference, the filter condenser  $C_1$  is installed at the input of this switching amplifier.

Only if phase coincidence for several periods is guaranteed, that means, if in any case the same number of revolutions of the two generators is warranted, the switching amplifier does actuate.

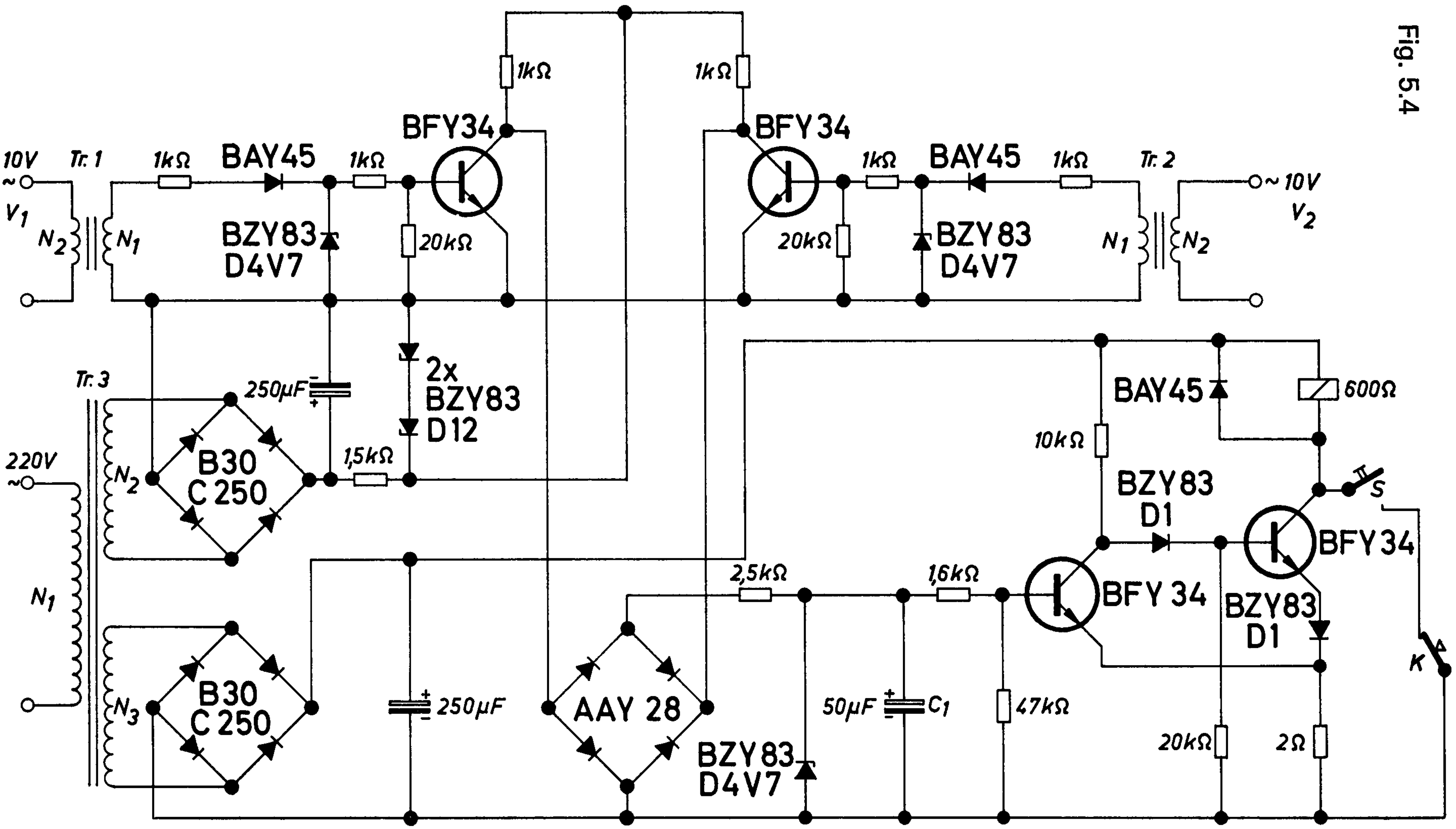


Fig. 5.4

Instead of the switching amplifier a measuring instrument can also be employed after the bridge rectifier without filter. Then the indication of this instrument is proportional to the phase angle between the two measured voltages because, as already mentioned, the width of the pulses generated by the differential amplifier corresponds to this phase angle.

### Technical Data

Supply voltage	220 V, 50 Hz
Input voltage	10 V
Input power	2.5 W
Frequency range of the test voltages	250 to 500 Hz

### Transformers

Tr 1 and Tr 2: EI 42/14, Dyn. sheet IV/0.35, alternate stacking

$N_1 = 640$  turns 0.2 CuL

$N_2 = 130$  turns 0.35 CuL

Tr 3: M 65/27, Dyn. sheet IV/0.35, 0.5 L, alternate stacking

$N_1 = 1550$  turns 0.25 CuL

$N_2 = 250$  turns 0.35 CuL

$N_3 = 200$  turns 0.5 CuL

## 5.5 Driving of a Servomotor with Photo-Voltaic Cells

If the two photo-voltaic cells of the circuit shown in Fig. 5.5 are illuminated, the servomotor rotates in a certain direction. If one of the two photo-voltaic cells is illuminated by a light intensity exceeding 200 lux, the trigger circuit which is connected to the photo-voltaic cell does actuate. The collector receives a voltage from the output stage consisting of complementary transistors, that is, one pnp and one npn transistor (pair) until the illumination of the photo cells drops below the threshold value of 200 lux. The potential of the voltage at the motor depends on which of the two photo cells is illuminated. The kind of operation of the complementary output stage may be compared to the output in transformerless AF amplifiers in which the complementary transistors also are driven by a phase coincident input signal and supply an AC signal for the loudspeaker. The sensitivity of the trigger may be set by the resistor  $R_1$ .

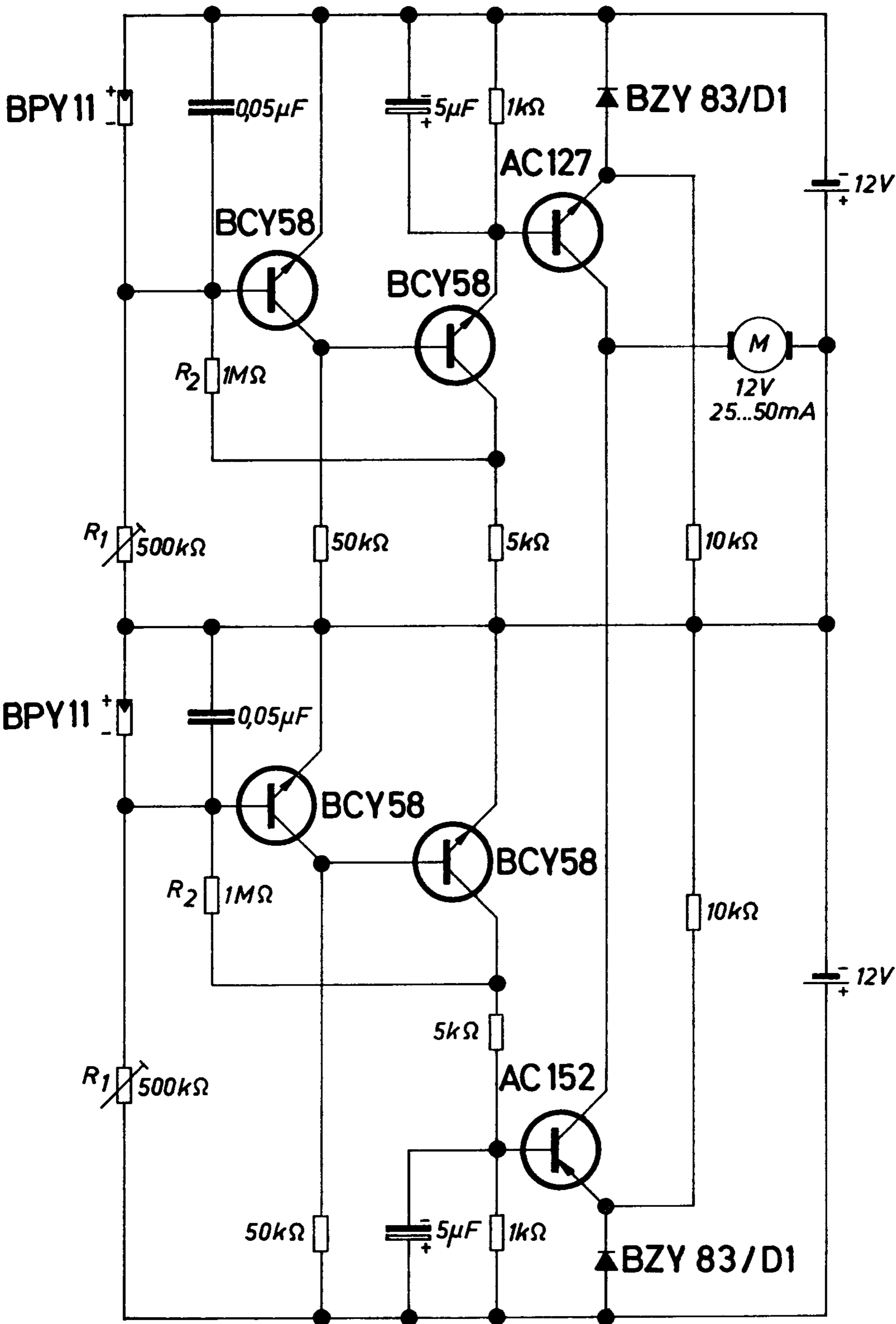


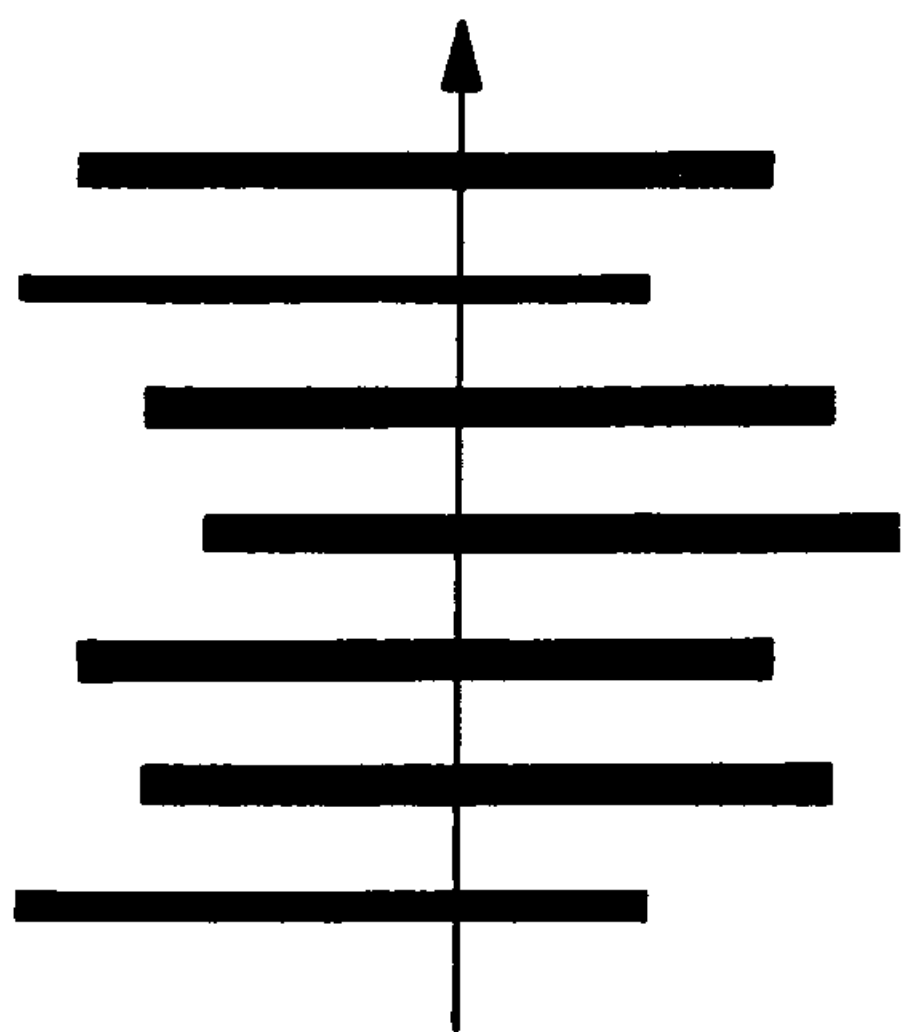
Fig. 5.5



## 6. Circuits with Hall-Effect Devices and Magnetoresistors

Besides the hall-effect devices which have been part of our program for several years, a new component depending on magnetic fields, the so-called magnetoresistor, is now available. While hall-effect devices are suitable for control and measurement in which a voltage dependent on magnetic fields is required, magnetoresistors may always be used when a resistor is to be changed in dependence on a magnetic field. These new semiconductor components indicate an increase in resistance by 6 to 20 times at a magnetic induction of 10 kG. The only component the function of which also results from the change in resistance in the magnetic field is the bismuth spiral which only shows an increase of the resistance of less than 50% for the same induction.

*without magnetic field*



*with magnetic field*

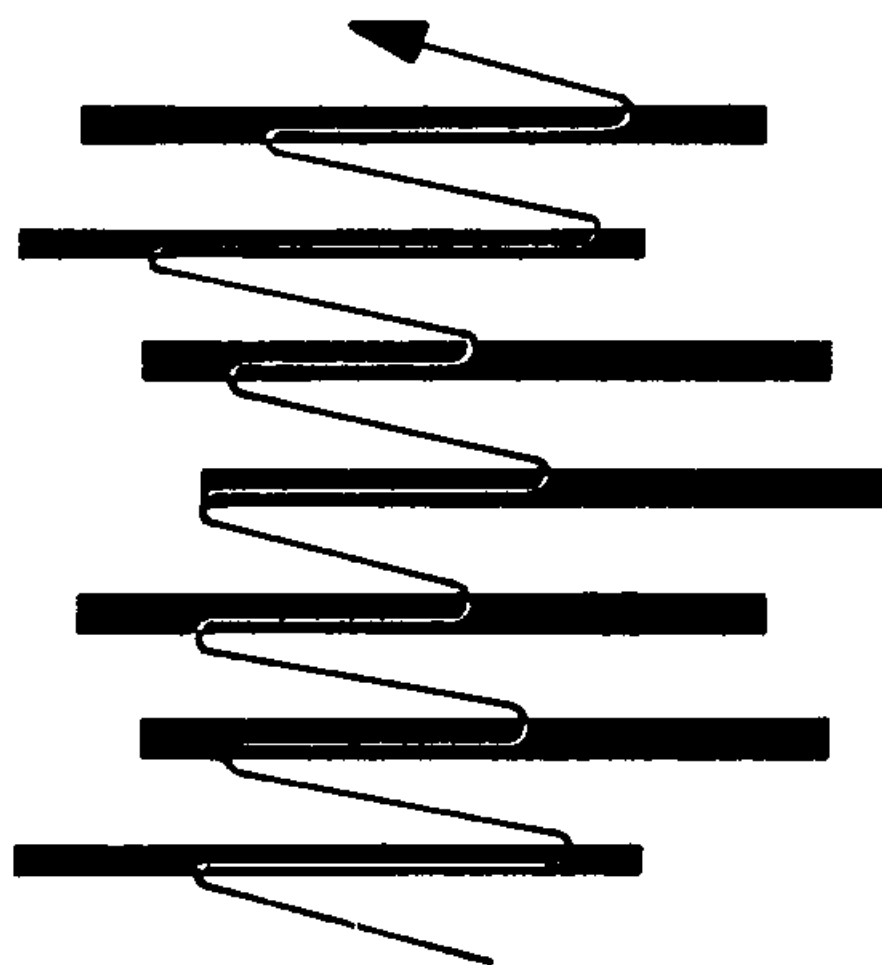


Fig. 6.1

With magnetoresistors the resistance increase results from the fact that the current paths through the magnetoresistor are deflected in the magnetic field.

In order to attain a considerable change of resistance, metallic conducting regions are created in distances of some thousandth up to some tenthousandth millimeter by enclosures of nickel antimonide needles. During deflection in the magnetic field these regions cause an increase of the total length of the current paths.

Fig. 6.1 schematically indicates this process. At both polarities of the magnetic field, the resistance of the magnetoresistor changes in the same direction. Up to a magnetic field intensity of 3 kG the resistance changes quadratically with the induction. At higher inductions, the dependence of the resistance on the induction becomes more and more linear.

The magnetoresistors are suitable for measurement and control of magnetic fields down to temperatures of  $-60^{\circ}\text{C}$ .

Above all magnetoresistors may be used as resistors which can be controlled continuously without contacts. The control may occur either by a permanent magnet the position which is changed in relation to the semiconductor or by the current of an electromagnet in the air gaps of which the semiconductor is located. As magnetoresistors are two-port networks and for the mentioned types the resistances amount to 10 to 500  $\Omega$  without magnetic field, they can be installed in conventional circuits without any difficulties. Magnetoresistors always behave like normal resistors, in other word, the instantaneous values of current and voltage relate to each other according to the occasional resistance.

## 6.1 Hall-Voltage Amplifier

In order to amplify hall voltages, amplifiers with low input impedances and best possible linearity are required. Fig. 6.2 shows the circuit of a DC amplifier with an input impedance of 5  $\Omega$ . The amplifier consists of an input stage with a high negative feedback, which has to be operated at a constant temperature. Therefore, the installation of this stage in a small thermostat is recommended. The input stage is succeeded by a driver stage with complementary transistors. This stage drives the push-pull power stage. The operation point of the prestage transistor is set by the potentiometer  $R_1$ , the collector quiescent current of the following stages by the potentiometer  $K_2$ . This amplifier also may

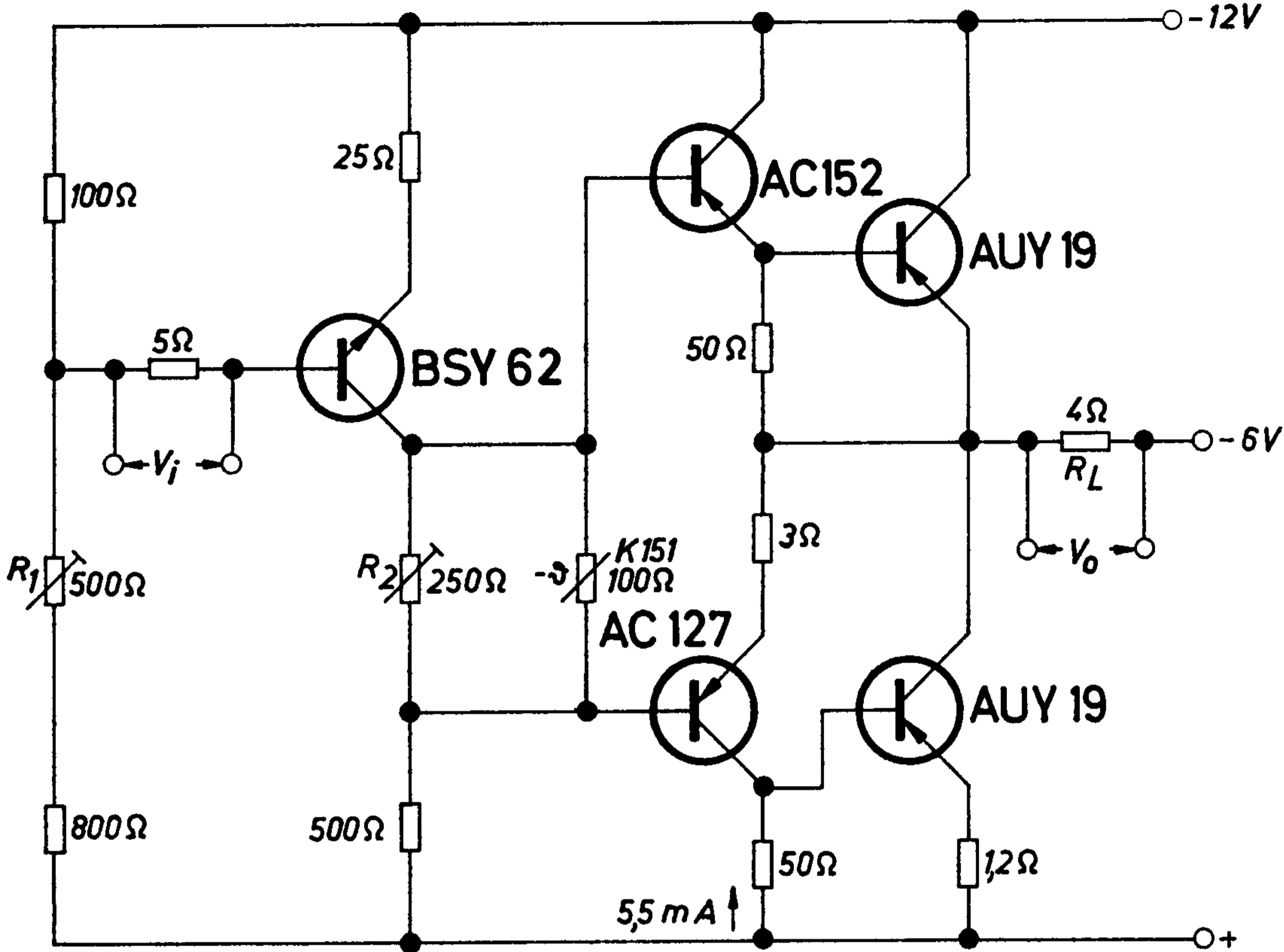


Fig. 6.2

be employed as an AC measuring amplifier. The maximum output power is 2 W at a distortion factor of 3.5%. The upper cut-off frequency has a value of 30 kHz.

Fig. 6.3 shows the linearity of the design if employed as a DC amplifier. Input voltages of both polarities can be used with the same accuracy.

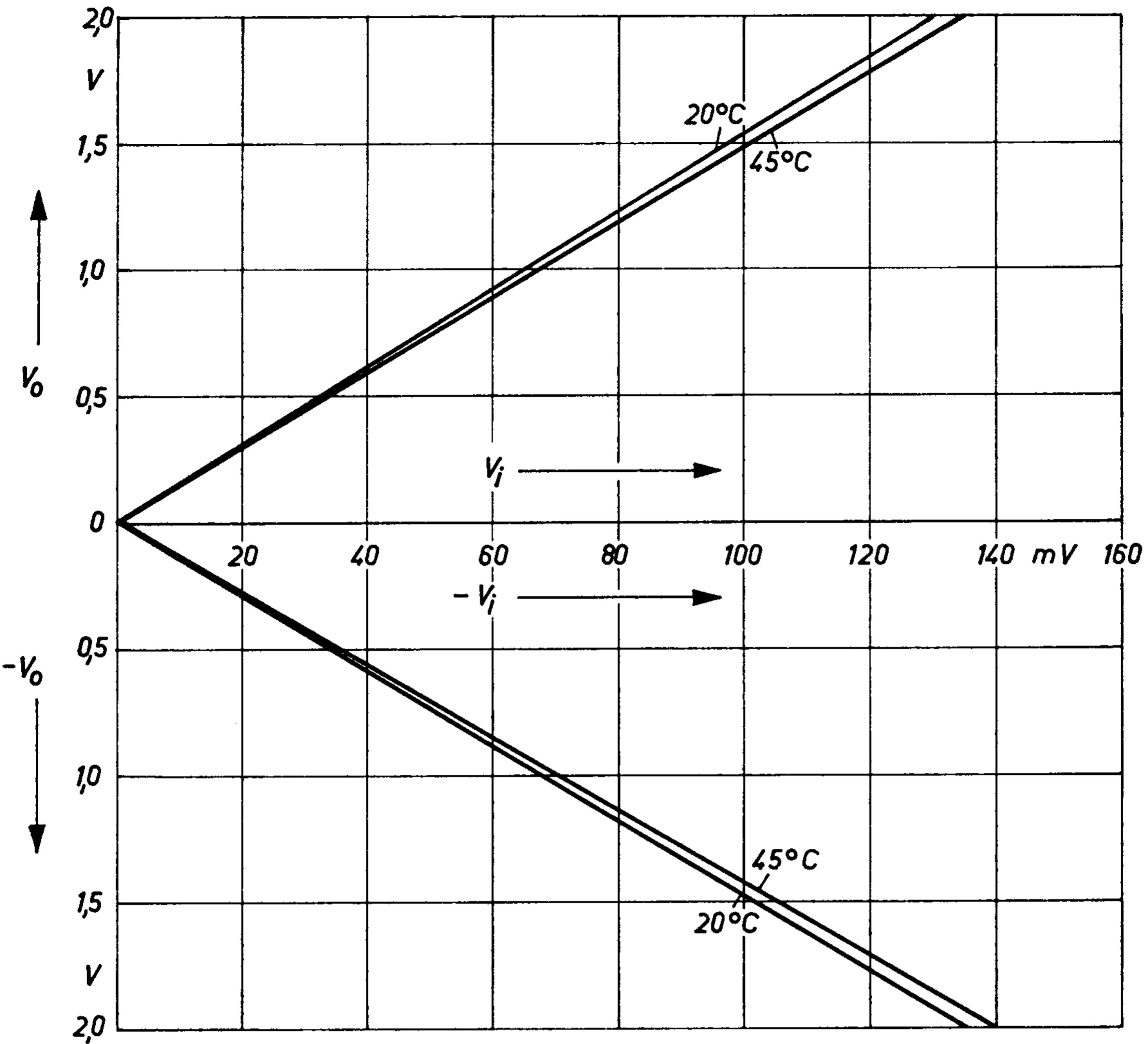


Fig. 6.3

### Technical Data

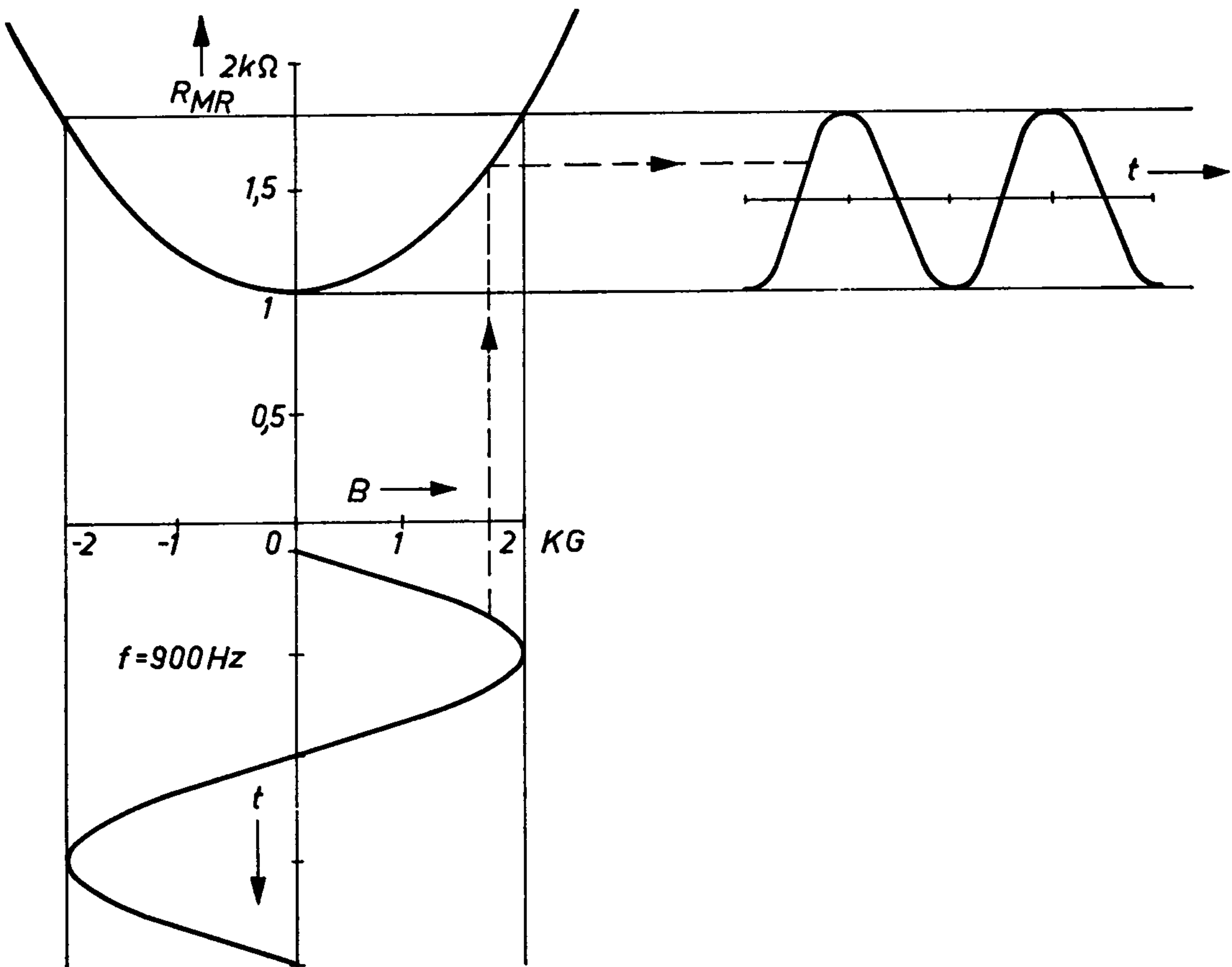
Supply voltage	$2 \times 6 \text{ V}$
Input voltage	150 mV
Input impedance	$5 \Omega$
Output voltage	2 V
Output impedance	$4 \Omega$
Max. ambient temperature	$45^\circ\text{C}$

## 6.2 Chopper with Magnetoresistors

Choppers are more and more employed for highly sensitive DC amplifiers because considerably higher sensitivity can be obtained with AC transistor amplifiers than with DC amplifiers. In the first case, the very low noise limits the input sensitivity, while in the second case, the cut-off current of the transistor influences the sensitivity. As the cut-off current is temperature dependent, the possible input sensitivity of DC transistor amplifiers decreases to the same extent as the temperature range in which the amplifier is to operate.

For that reason, the DC signal is chopped in sensitive amplifiers, in other words, this signal is converted to a proportional AC signal which will be amplified by an AC amplifier and then will be rectified again. Most important is that such a chopper does have

Fig. 6.4



a low zero drift so that small input signals also will be converted into proportional AC signals. In transistor choppers the accuracy at small input signals is mainly determined by the saturation voltage of the transistors.

The magnetoresistors described at the beginning of this chapter are suitable for the design of accurate choppers. Fig. 6.4 shows that the resistance of magnetoresistors, which are located in the air gap of a magnet, changes with double the frequency of the magnetic induction. Fig. 6.5 schematically demonstrates such a chopper. The magnetoresistors are inserted in the air gap of a transformer to which a constant AC voltage is applied. This AC voltage generates a constant induction by which the resistance of the magnetoresistors always will be changed with twice the frequency by a certain amount. For that reason, an AC voltage which can be tapped off at the resistor  $R_1$  is directly proportional to the input voltage. Then this AC voltage can be amplified linearly and rectified again. Two magnetoresistors must be employed because otherwise a zero drift would occur. The magnetoresistor and its terminals form a loop in which a certain voltage is induced. Both magnetoresistors are connected in such a way that these induced voltages will compensate each other.

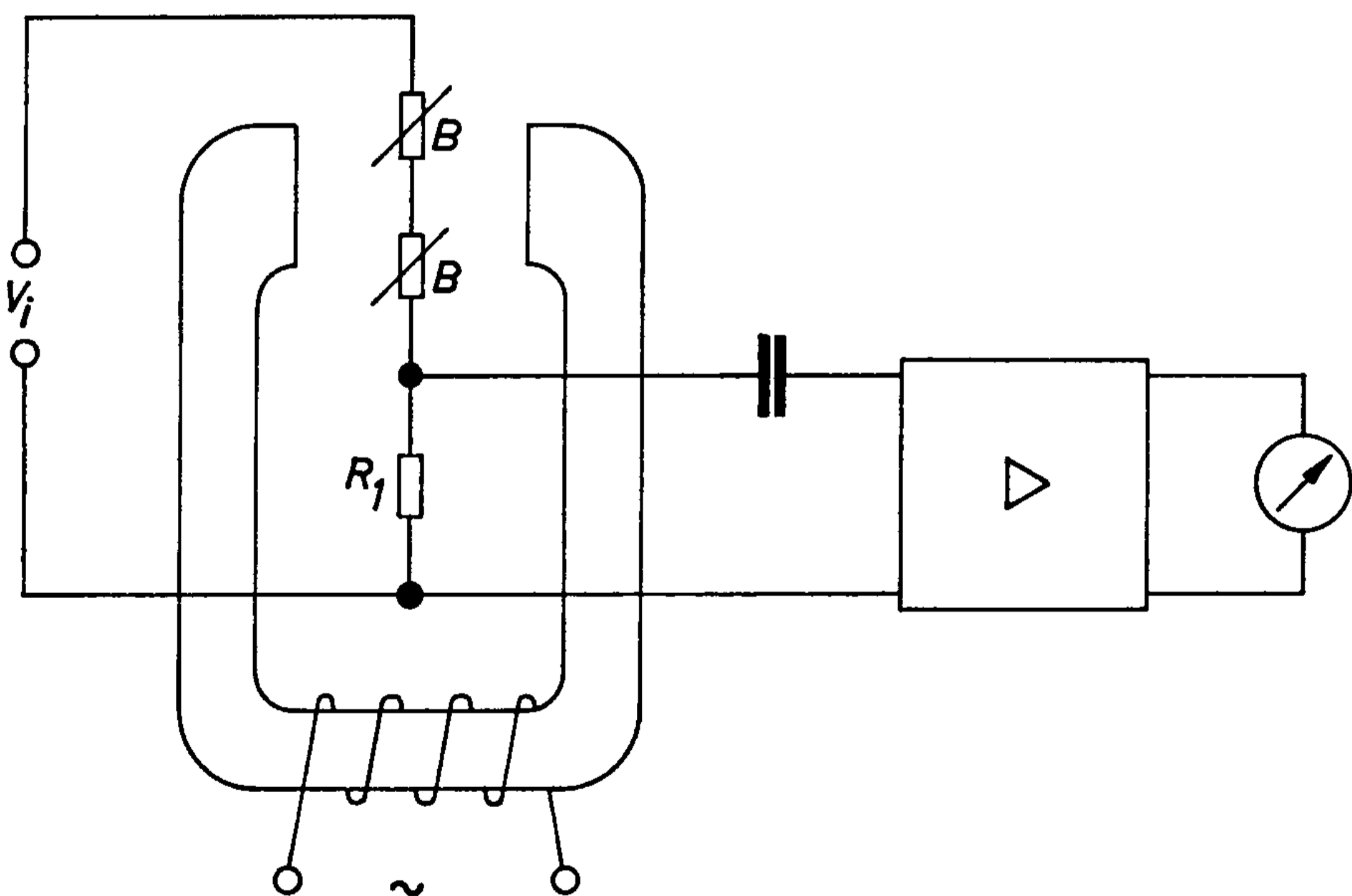


Fig. 6.5

## **7. Controlled Power Supplies**

Low operating voltages are required for transistor equipments. For higher power requirements high operating currents are obtained which can be sufficiently filtered in a conventional way only with great requirements of material. A variation of the operating voltage often is a handicap so that controlled power supplies have to be employed. Thus, controlled power supplies have to keep the operating voltage constant and have to decrease the filtering requirements. Nowadays, these power supplies are always equipped with transistors. Different types of power supplies are used for this purpose. In the simplest type, a transistor is driven as a variable series resistor. These power supplies, however, are only short circuit-proof to a certain extent. Therefore, they mostly contain a supplementary electronic fuse. The parallel control offers a second possibility. In this case, the transistor is again driven as a variable resistor but this time parallel to the load. This type of circuit is automatically short circuit-proof. This circuit, however, has one disadvantage. The total energy is constantly absorbed by the power supply even when no load is connected.

Voltage stabilized power supplies for two position controls are suitable for high power requirements where a filter condenser is charged by pulses. In this case, the power transistors only operate as switches. Therefore, they dissipate a small power only. In this type of circuit, however, the filtering requirement is slightly higher than for the types of power supplies which have been described above.

### **7.1 Voltage Stabilized Power Supplies for Two Position Control 31 V, 3.3 A**

High output power can be attained by a pulse-scanned, voltage-stabilized power supply with relatively small dissipation power in the controlling section. A condenser is charged by pulses, whose widths or sequence frequencies depend on the load.

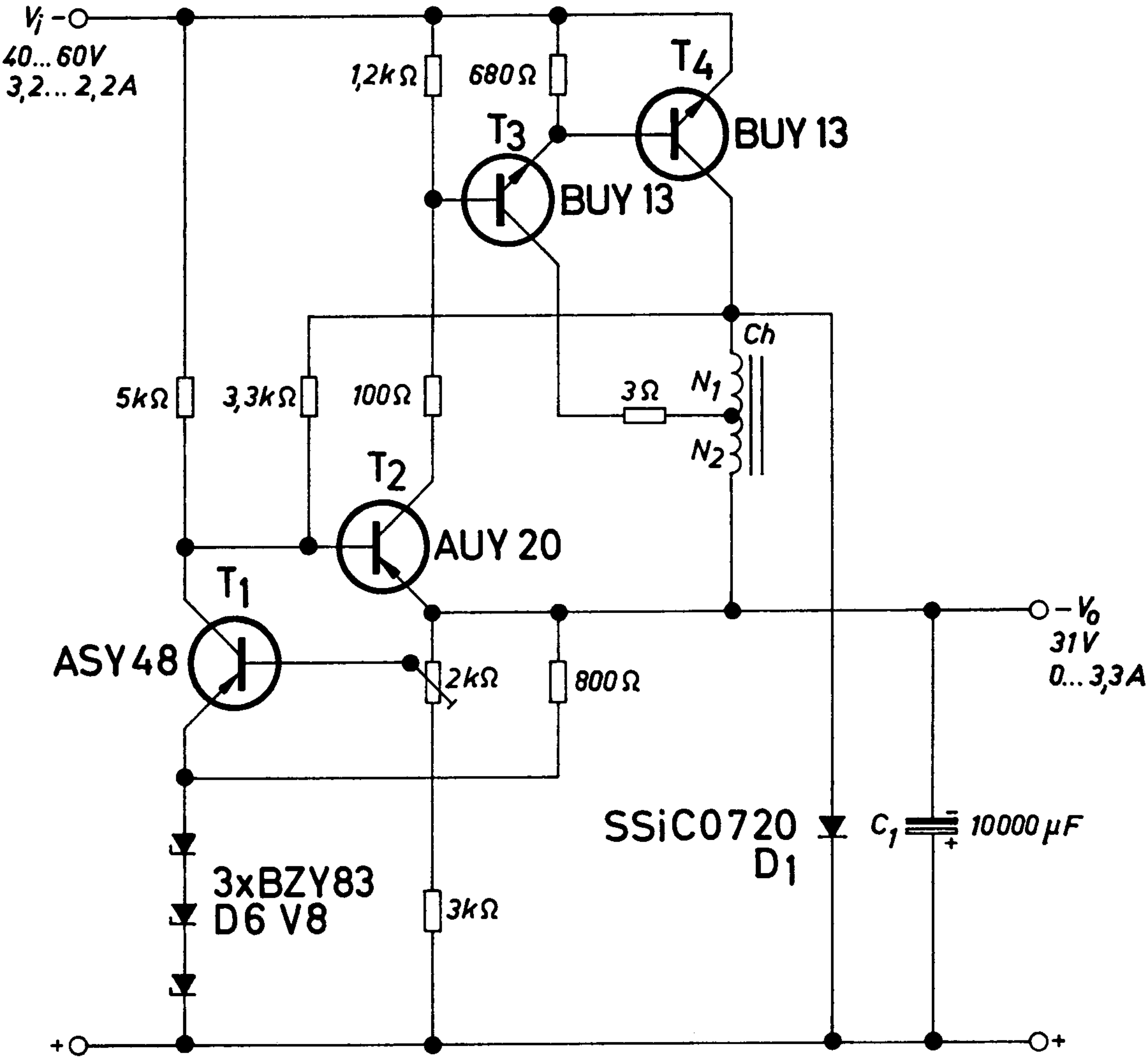


Fig. 7.1

Hereby, material requirements may be saved for the powerstage and a higher efficiency will be attained. But, as already mentioned in the beginning of this chapter, more filters are required at the output in comparison to the continuously controlled power supplies.

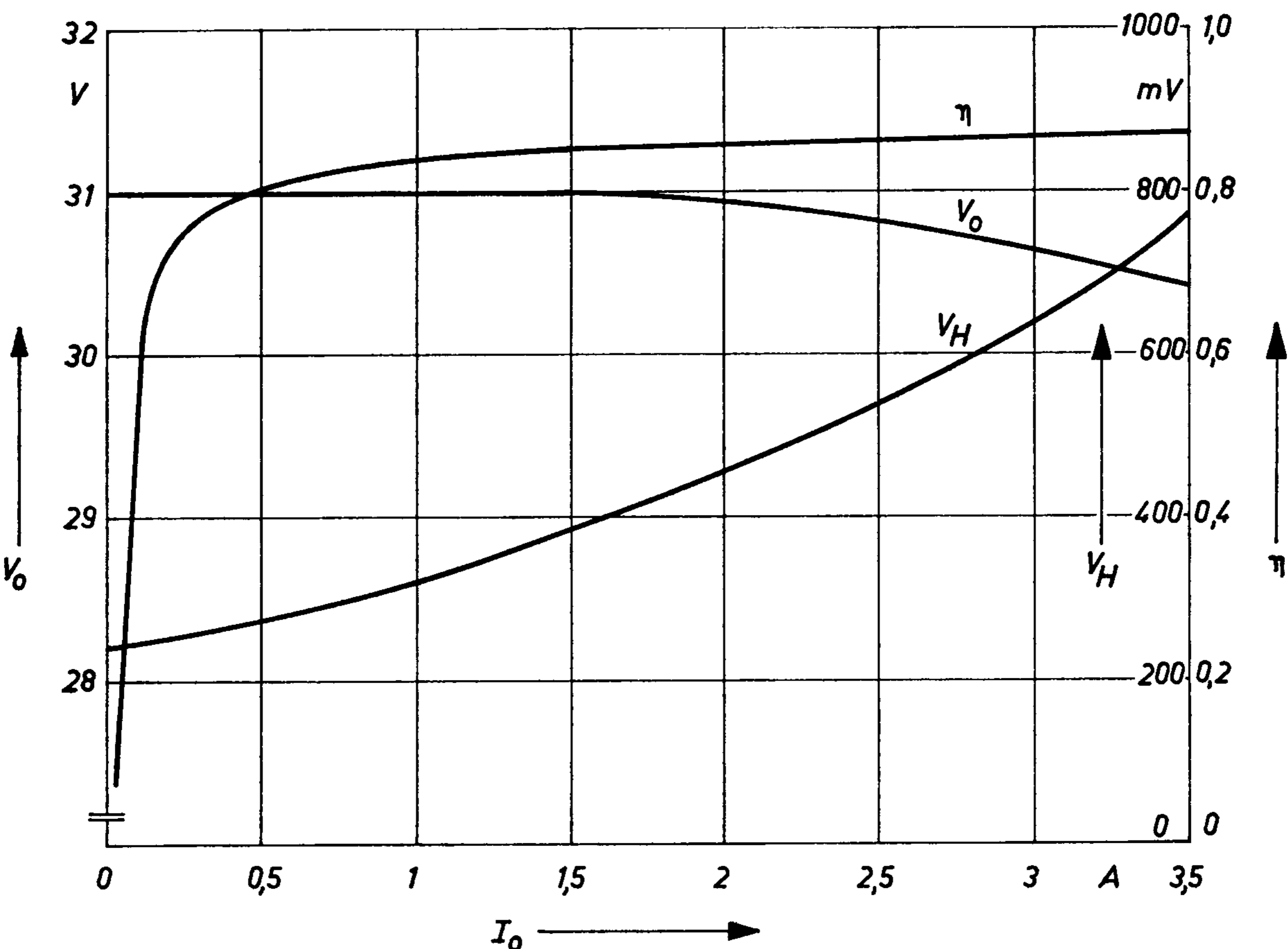
Fig. 7.1 indicates the circuit of a pulse-controlled, voltage-stabilized power supply for a output voltage of 31 V and with a maximum load current of 3.3 A. One immediately can recognize the low requirements in relation to the maximum output power of approximately 100 W. If the output voltage at the filter condenser  $C_1$  exceeds a certain value, transistor  $T_1$  is turned on.



This occurs because the voltage at the base of this transistor obtains a higher negative value than the emitter potential, which is kept constant by the zener diodes. The threshold value can be set by the potentiometer. As soon as transistor  $T_1$  is turned on, the transistors  $T_2$ ,  $T_3$  and  $T_4$  are turned off. Hereby the charging of the condenser will be stopped. Only the inductive energy which is stored in the choke Ch is discharged via the diode  $D_1$  to the condenser  $C_1$ . This improves the efficiency. Now the condenser  $C_1$  discharges via the load resistance. As soon as the output voltage drops below the adjusted value, transistor  $T_1$  turns off, and the other three transistors turn on. Then, condenser  $C_1$  is recharged via transistor  $T_4$  and via the choke. In this circuit the recharging of the condenser occurs at a frequency of approximate 200 Hz.

At a variation of the input voltage of  $\pm 10$  V the output voltage changes in the case of a load variation between open circuit and full load only by a value of  $\pm 1$  V. Output voltages between 19 and 34 V can be adjusted by the potentiometer. The ripple of the

Fig. 7.2



output voltage has for open circuits and for small loads (load current  $\leq 1$  A) a value of approximately 250 mV (peak value); in case of full load, the ripple attains a value of approximately 700 mV.

Fig. 7.2 indicates, how efficiency  $\eta$ , output voltage  $V_o$  and the hum voltage  $V_h$  depend on load current  $I_o$ .

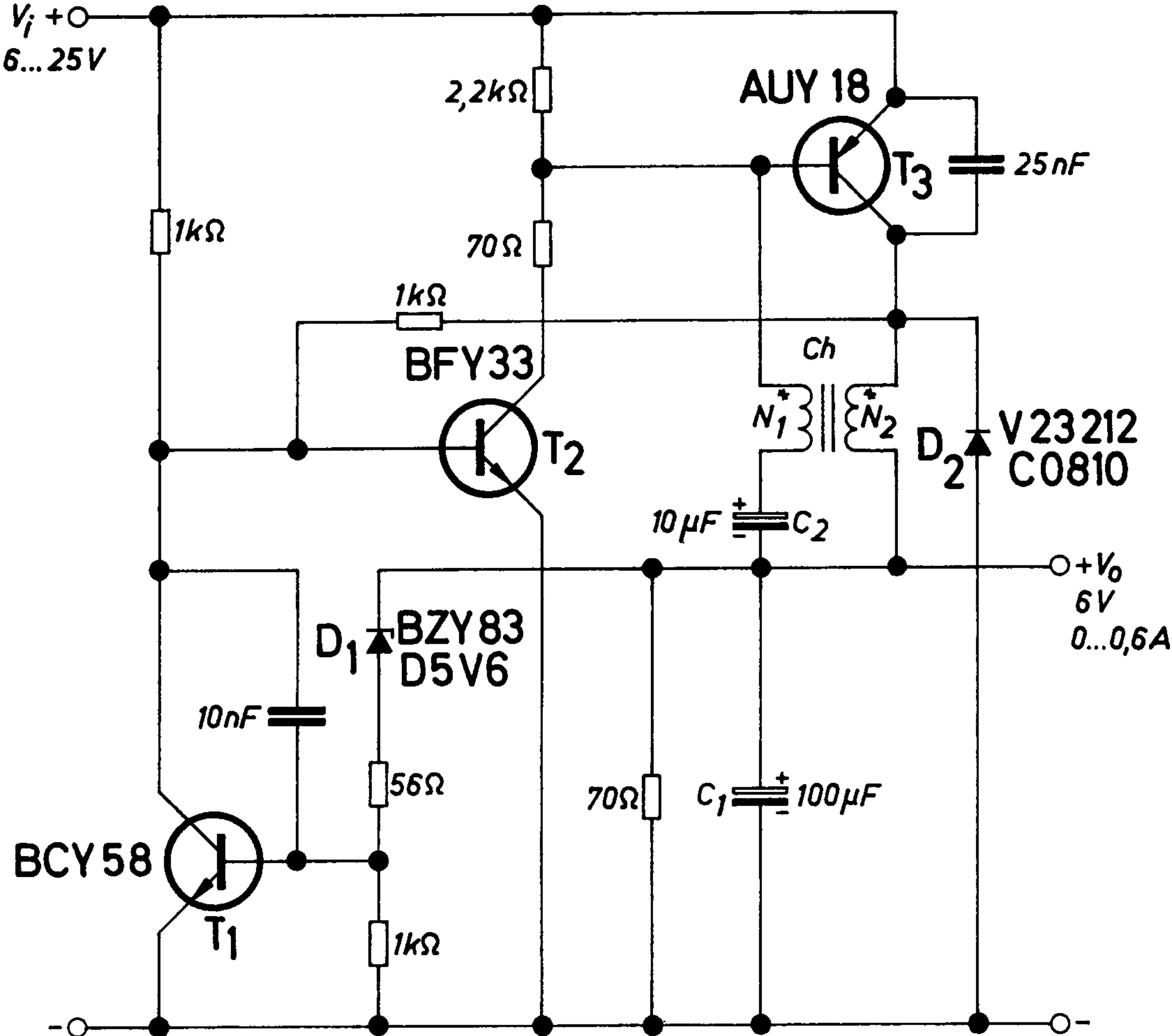
#### Technical Data

Input voltage	40 to 60 V
Output voltage (adjustable)	19 to 34 V
Load current	0 to 3.3 A
Choke Ch: M 74/32, Dyn. sheet IV/0.5, 0.5 mmL, unit-directional	
$N_1 = 9$ turns 2.0 CuL	stacking
$N_2 = 55$ turns 2.0 CuL	

## 7.2 Voltage Stabilized Power Supply for Two Position Control 6 V, 0.6 A

The circuit shown in Fig. 7.3 operates in the same way as the circuit in Fig. 7.1 does. If the voltage at the filter condenser  $C_1$  exceeds the value 6 V, transistor  $T_1$  turns on because the zener diode  $D_1$  becomes conductive by exceeding the zener voltage. By that means, the input of transistor  $T_2$  is short-circuited so that this transistor and also transistor  $T_3$  turn off. Then, the energy stored in the choke Ch during the current flow in transistor  $T_3$  is discharged via the diode  $D_2$  to the condenser  $C_1$ . As soon as the voltage at the condenser  $C_1$  drops by discharge via the load so far that the zener diode  $D_1$  is in reverse condition, the transistor  $T_1$  also will be turned off and the transistors  $T_2$  and  $T_3$  are turned on. Now, condenser  $C_1$  is recharged via transistor  $T_3$  and the choke Ch until the zener voltage at the diode  $D_1$  will be attained again. Then the described process starts over again.

In order to attain the smallest possible circuit size, a small inductivity and a small filter condenser have been employed for this circuit. A high switching frequency and a larger ripple wave of the constant voltage result from this fact. In order to improve the efficiency of the circuit, care has been taken that the transistor  $T_3$  is quickly turned on. For that reason the voltage which is generated in a supplementary coil  $N_1$  of the choke Ch is applied to the base of the transistor  $T_3$  and, thus, the switching



\* points of equal polarity

Fig. 7.3

of the device is accelerated. This is effective not only for the turn on but also for the turn off of the transistor. As soon as the transistor  $T_3$  is turned off, the polarity of the voltage changes at the choke. Thus, a reverse voltage instantaneously is applied to the input of transistor  $T_3$  via the coil  $N_1$ . The condenser  $C_2$  decreases the current peaks which occur when the transistor is turned off.

This power supply always applies in these cases where a constant output voltage is to be gained from a highly varying input voltage.

#### Technical Data

Supply voltage            6 to 25 V

Output voltage            6 V

Load current              0 to 0.6 A

Choke Ch: Siferrite-pot core B 65561 -A 0063-A 025

$N_1 = 40$  turns 0.1 CuL

$N_2 = 120$  turns 0.35 CuL

### 7.3 Controlled Power Supply 60 V, 1 A

The power supply shown in Fig. 7.4 is suitable for the current supply of the transformerless AF amplifier described in chapter 1.10. This is a power supply with series control i.e. the transistor  $T_1$  operates as a variable series resistor.

In the power supply an electronic fuse is incorporated (Fig. 7.5) which turns off the power supply when overload or short circuit occur. A voltage peak of 90 V appears at transistor  $T_1$ ; therefore, a high voltage germanium power transistor has been employed. This transistor is driven by two prestage transistors connected in series because the high reverse voltage also appears at these transistors. Due to a voltage divider in the base circuit the reverse voltage drops at the two transistors. Terminals which are equally marked indicate how the electronic fuse in Fig. 7.5 has to be connected to the power supply in Fig. 7.4. The power supply is turned off by the transistor  $T_5$  which is driven into saturation when the fuse is actuated and, thus, the transistor  $T_1$  is turned off.

Transformer Tr: M 102a, Dyn. sheet IV/0.35, 0.5 L, alternate

$N_1 = 760$  turns CuL 0.55

stacking

$N_2 = 221$  turns CuL 0.8

Fig. 7.4

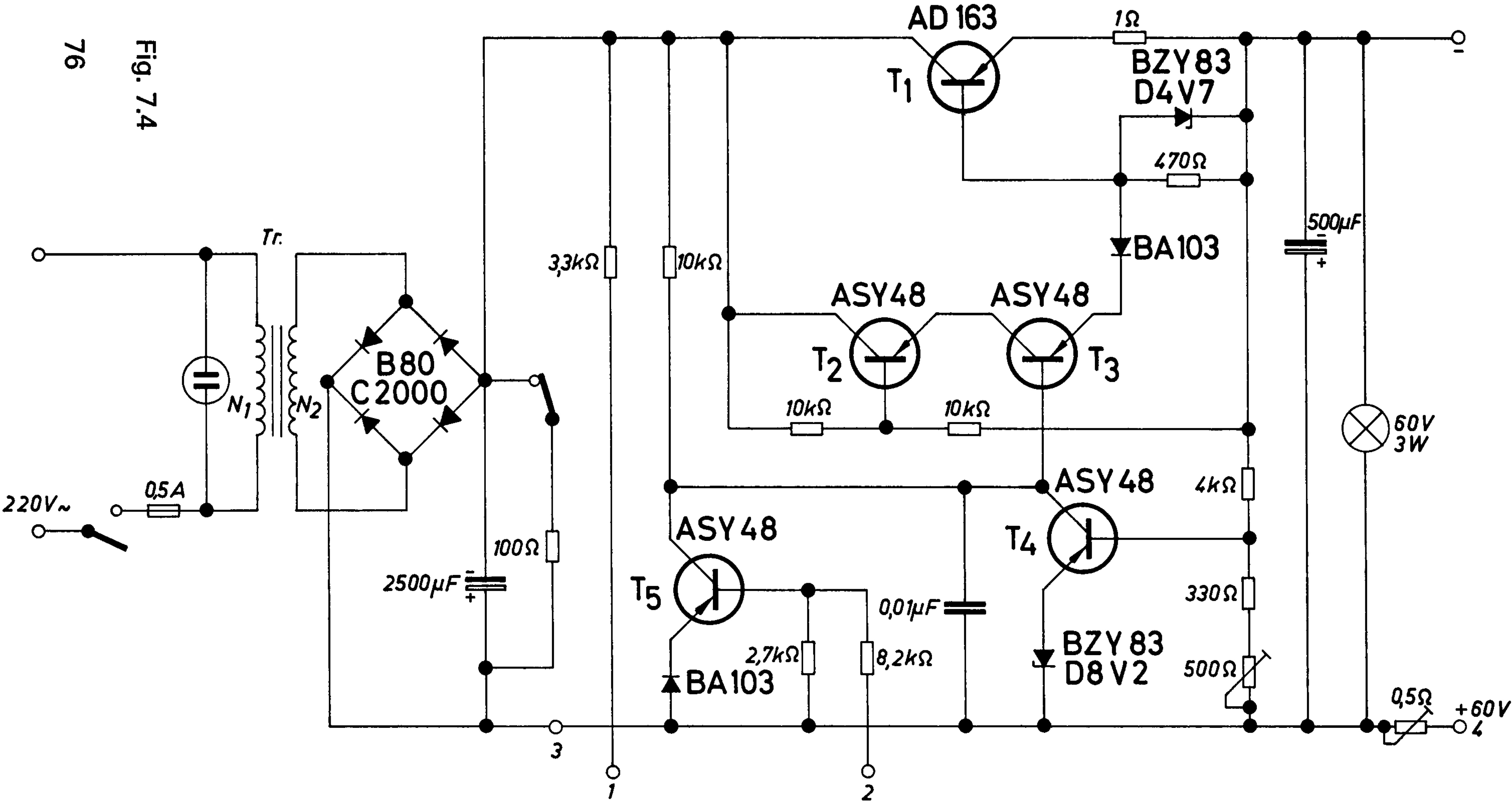
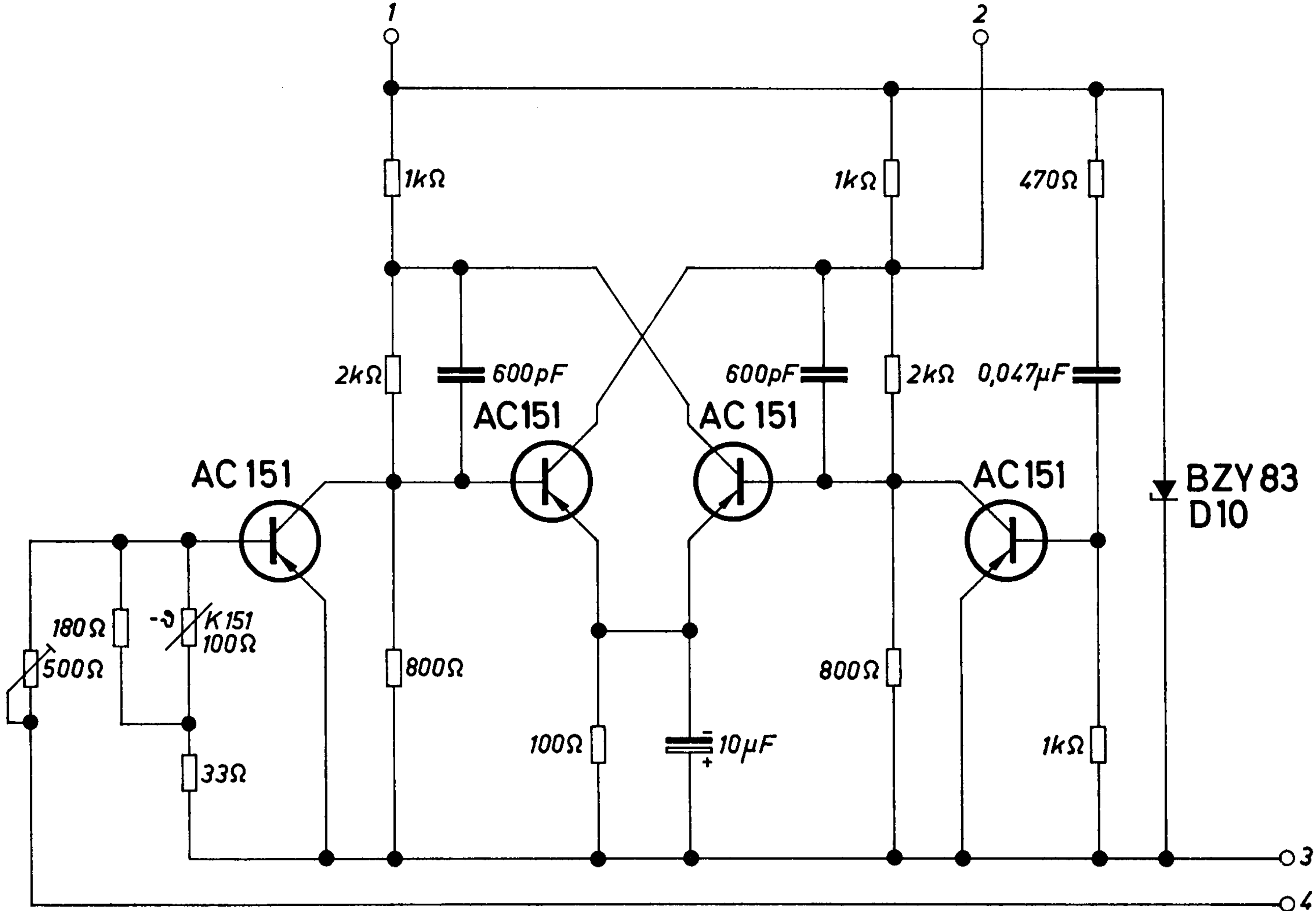


Fig. 7.5



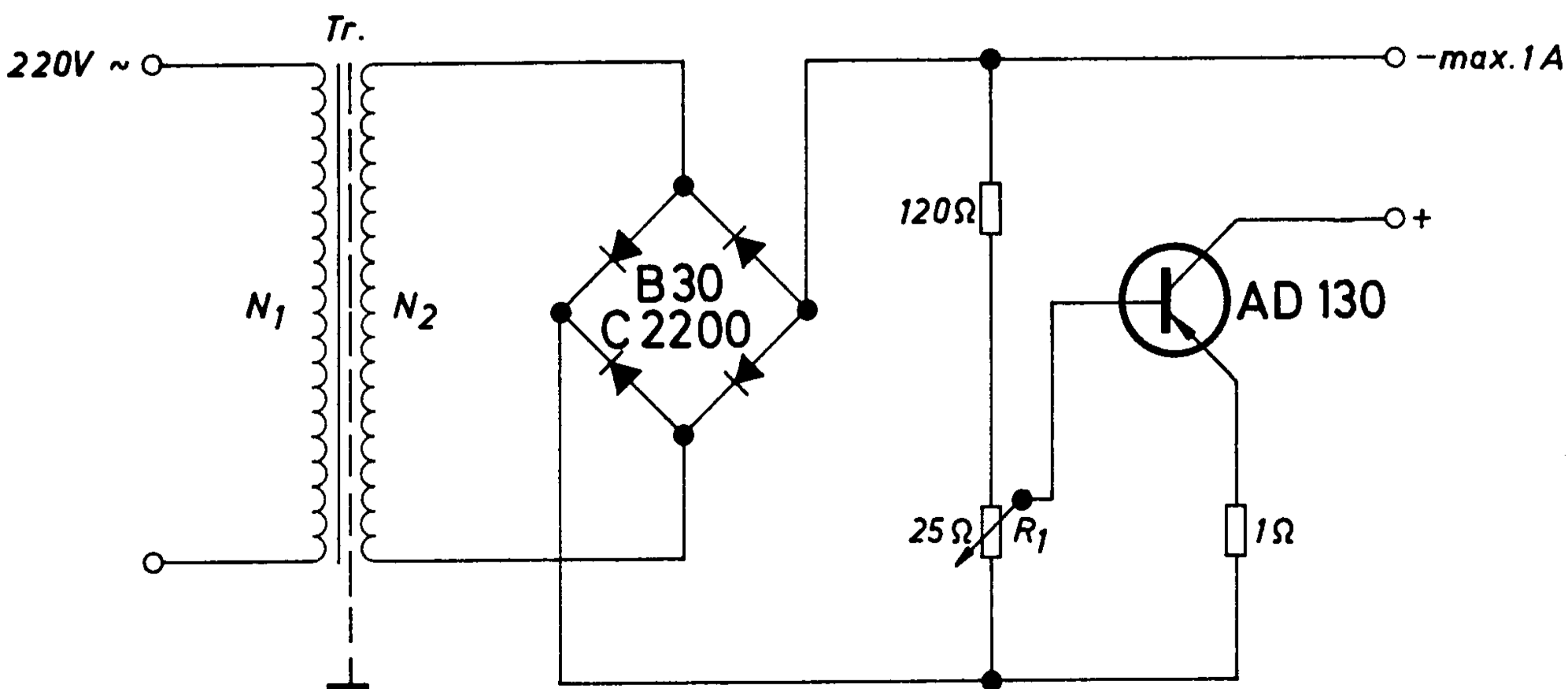
## 7.4 Simple Current Stabilized Power Supply

Control transformers are required for toys such as model railways. They have to be reliable and in any case short-circuit-proof. As it is basically unimportant whether the output voltage or the output current is controlled, a current control power supply is suitable because it is automatically short-circuit proof. The output current amounts to 1 A max. and is adjusted by the potentiometer  $R_1$ . The power transistor is to be mounted to a heatsink with a thermal resistance of 3 °C/W max. Thus it corresponds to an efficient cooling surface of approximately 250 cm<sup>2</sup>.

### Technical Data

Supply voltage	220 V, 50 Hz
Output current	0 to 1 A
Minimum load resistance	14 $\Omega$
Internal resistance	1.5 $\Omega$
Transformer Tr: M 65/27, Dyn. sheet III/0.5, 0.5 L, alternate stacking	
$N_1 = 1600$ turns 0.25 CuL	
$N_2 = 150$ turns 0.8 CuL	

Fig. 7.6



## **8. Radio Frequency Circuits**

Due to modern technologies it is possible to design transistors with more and more improved high-frequency characteristics. A variation of the approved germanium mesa technology, the so-called mesa II technology produces high frequency transistors with a noise figure of 3 and a power gain of 14 dB at 800 MHz.

The silicon planar technique is applied for high frequency power transistors. Considerable improvements recently have been achieved also in this field.

### **8.1 Low Noise UHF-Antenna Amplifier**

By using the amplifier with the new low noise mesa transistor AFY 42, the TV reception within the UHF range can be improved. The noise figures of  $F < 3$ , which are attainable within the whole range, improve the quality of reception even in band IV.

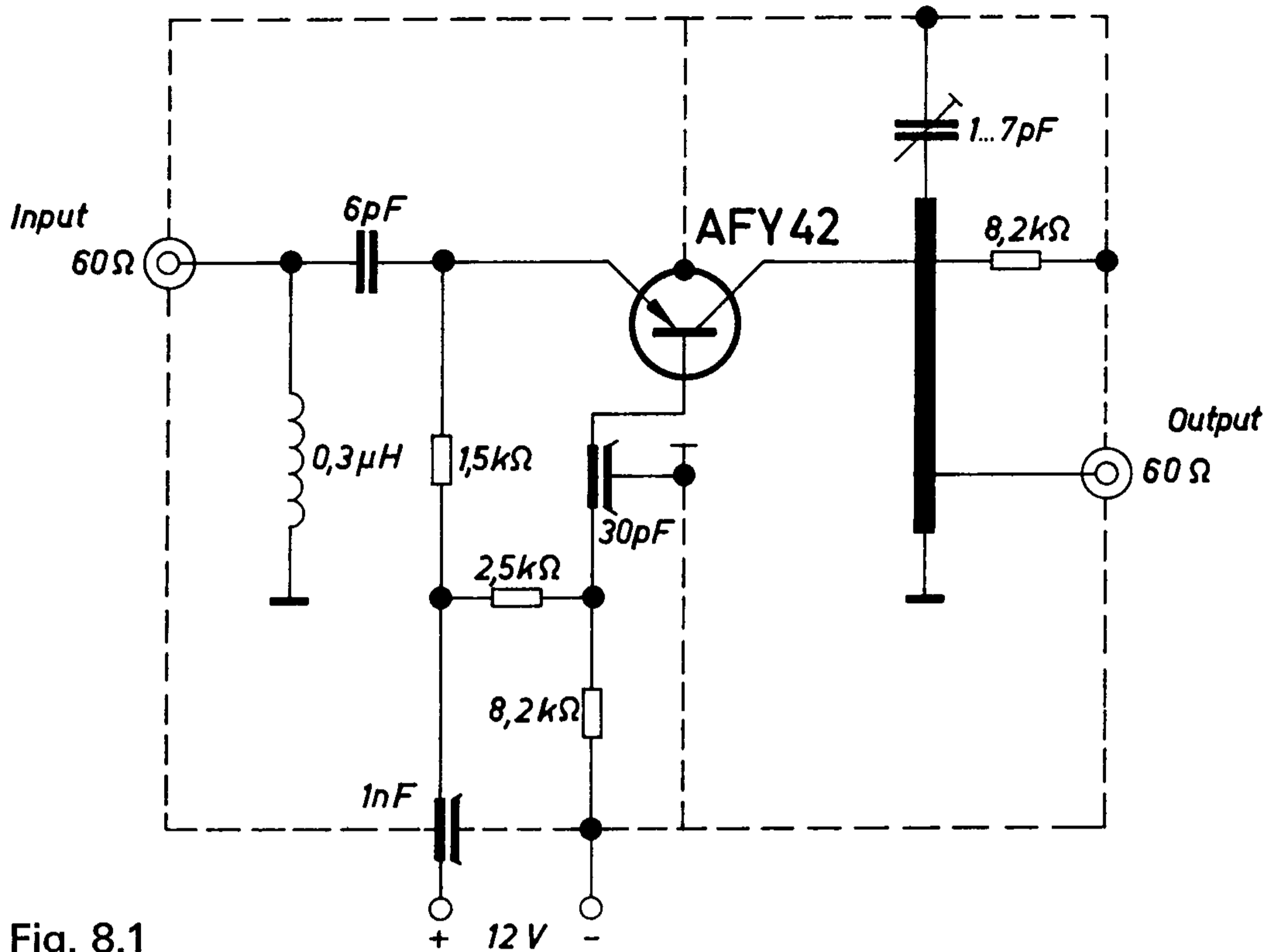
The described antenna amplifier is designed in a very simple way. Because of its small dimensions it can be installed in the dipole socket of an antenna.

The circuit is shown in Fig. 8.1. The input has a wide band characteristic because the quality of reception is of such a level that no cross modulation has to be expected. The small coupling capacitance sufficiently protects the transistor from dangerous input voltages e.g. during a thunderstorm. Because of the uniform amplification within the entire frequency range the transistor is operated in common base configuration.

A  $\lambda/4$ -circuit which can be tuned to any channel between 470 and 860 MHz is connected to the collector.

Because of the small interior reaction of the transistor, the amplifier is stable even without load at the input or the output. At maximum the amplifier may be operated at an ambient temperature up to 83 °C as the self-heating is only 7 °C.





### Technical Data

measured with an average sample AFY 42

Supply voltage	12	V
Current consumption	2.7	mA
Frequency range	470 to 860	MHz
Power gain	11 to 13	db
Noise figure	2.2 to 2.5	
Band-width	20 to 50	MHz
Input reflection coefficient	0.25 to 0.5	
Output reflection coefficient	< 0.2	

### Dimensions of

the Amplifier	45 mm × 20 mm × 15 mm
Dimensions of the resonant-line circuit	Width 10 mm Depth 15 mm Length 45 mm
Interior conductor	Length 35 mm ∅ 1 mm (silver-plated copper)

Tapping for the collector	9 mm from the trimmer side terminal
Tapping for the output	5 mm from the ground terminal

## 8.2 Non-Neutralized IF Amplifier 36.4 MHz with 2 x AFY 16 and AFY 37

The television IF amplifier shown in Fig. 8.2 contains the transistors AFY 16 and AFY 37. Because of the favorable reverse capacitance of these transistors, it is not difficult to design the amplifier without any neutralisation. Thus, at least one condenser in each stage and the neutralizing coil in the band-pass filter can be omitted. The amplifier shows perfect stability of the transmission curve even during an experimental detuning of the resonant circuits. For the control of the first stage (gain decrease control) a control range of 60 db is possible. It is recommended to set the operation point for maximum gain in the first and in the second IF stage at 3 mA. The gain of the control stage is completely regulated down at 7.7 mA. The operation point of the output stage is located at 15 V/7 mA. The band-pass filter is inductively coupled to the input of the succeeding transistor because, hereby, it is easier to attain a higher stability of the gain during temperature change and at an excessive voltage.

The power gain of the three-stage amplifier has a value of 87 db. At a synchron-pulse-compression of 30% the attainable output voltage after rectification is 6 V at a load resistor of 2.7 k $\Omega$ . The small signal and the large signal stabilization as well as the curve constance are perfect in the entire control and driving range of the output. The cut-off currents of the transistors are of no influence even at higher temperatures as the bias circuit is sufficiently low resistive.

### Technical Data (Fig. 8.2)

Supply voltage	18 V	Range of control	60 db
Power gain	87 db	Output voltage at 2.7 k $\Omega$	6 V

### Coil Data

Bobbin 5 mm  $\varnothing$ , cores siferrite U 17/4 mm  $\varnothing$

$L_T = 18$  turns 0.3 CuLL\*

$L_1 = 7$  turns 0.3 CuLL

$L_2 = 15$  turns 0.3 CuLL

$L_3 = 15$  turns 0.3 CuLL, tap at 5 turns

$L_4 = 15$  turns 0.3 CuLL

$L_5 = 8$  turns 0.3 CuLL, tap at 2 turns

$L_6 = 15$  turns 0.3 CuLL

$L_7 = 18$  turns 0.3 CuLL

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\* see page 106

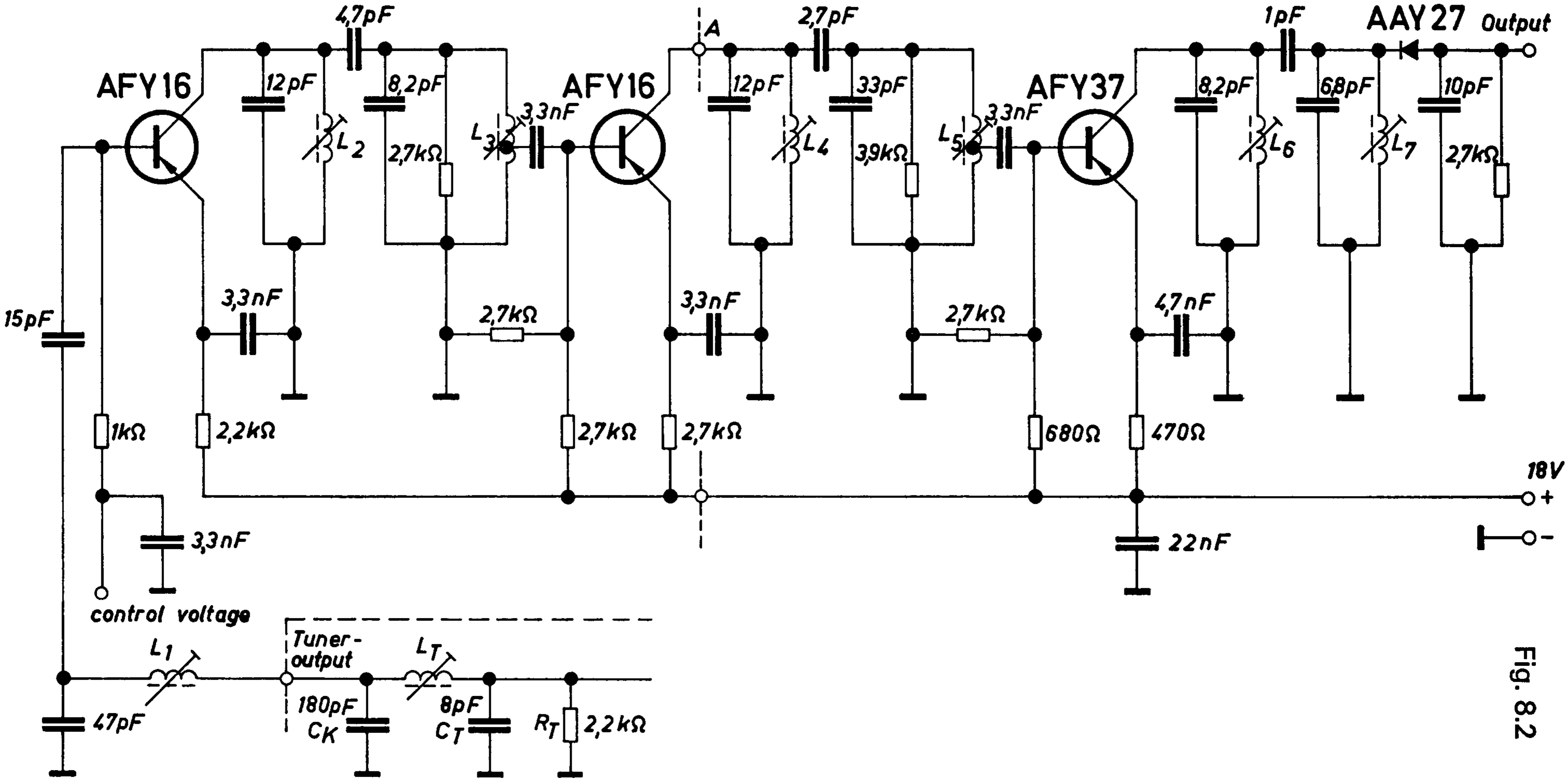


Fig. 8.2

Fig. 8.3 shows another solution for the IF output. In this case, the transistor AFY 39 is employed whereby a neutralisation is omitted again. The circuit has to be designed in such a way that the amplifier still remains stable. The amplification is for some db smaller compared to an output stage with neutralization. It has been proved that in non-neutralized output stages the spread of the transistor parameters influences the gain spread to a smaller extent than in stages with neutralization. The power amplification amounts to 20 db including the demodulation.

#### Technical Data (Fig. 8.3)

Supply voltage	18 V
Power gain	20 db
Output Voltage at 2.7 k $\Omega$	6 V

#### Coil Data

- $L_4 = 15$  turns 0.3 CuLL
- $L_5 = 7$  turns 0.3 CuLL, tap at  $1\frac{1}{2}$  turns
- $L_6 = 12$  turns 0.3 CuLL
- $L_7 = 15$  turns 0.3 CuLL

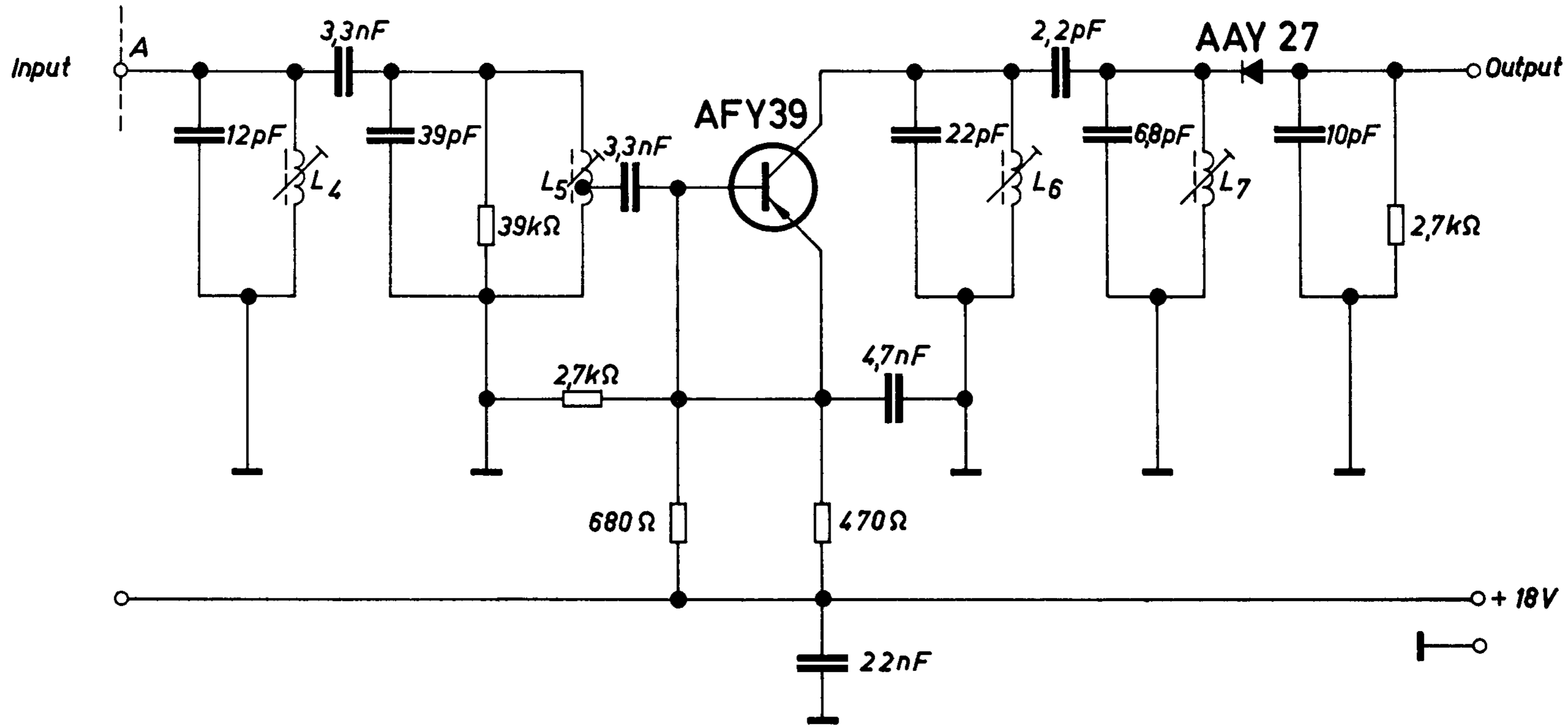


Fig. 8.3

## **9. Radio Circuits**

An increasing number of TV sets is being transistorized. New transistors and new circuit designs improve the efficiency of these transistor circuits. By employing the new mesa-transistor AF 239 in the input stage of UHF-tuners, noise figures of 5 db are attained in the entire UHF range.

With the new input transistor AF 109 R also a considerable noise reduction in VHF tuners is possible. Of special interest is the fact that mechanical tuning devices are being replaced by voltage variable capacitance diodes. With the diode BA 138 the tuning of band I and III at a voltage deviation from 3 to 30 V is made possible.

The favorable high frequency characteristics of the video transistor BF 110 are advantageous in one-stage and two-stage video amplifiers. Therefore, one example for each solution will be presented in this chapter. By using our germanium mesa transistors a high gain per stage can be obtained in radio IF-amplifiers. Suitable circuits will be given in the following chapter.

### **9.1 UHF-Tuner with Automatic Retuning**

Under unfavorable conditions, UHF tuners show a high frequency drift when the ambient temperature changes. This is not mainly due to the transistor but the frequency drift is caused by a combined effect of material expansion and condenser-and transistor drift.

Great difficulties arise from the material expansion of the cavity resonators and conductor paths. Furthermore trouble may eventually occur by the material expansion of the condenser spindle which may result in positive or negative capacitance deviations and, thus, frequency shifts dependent on the symmetry of the rotor plate and the stator plate. Resonant circuit condensers, and trimming- and ground condensers also may cause temperature dependent frequency deviations. Mostly the frequency shift will be compensated by selecting different temperature coefficients of the frequency determining circuit and coupling capacitances.

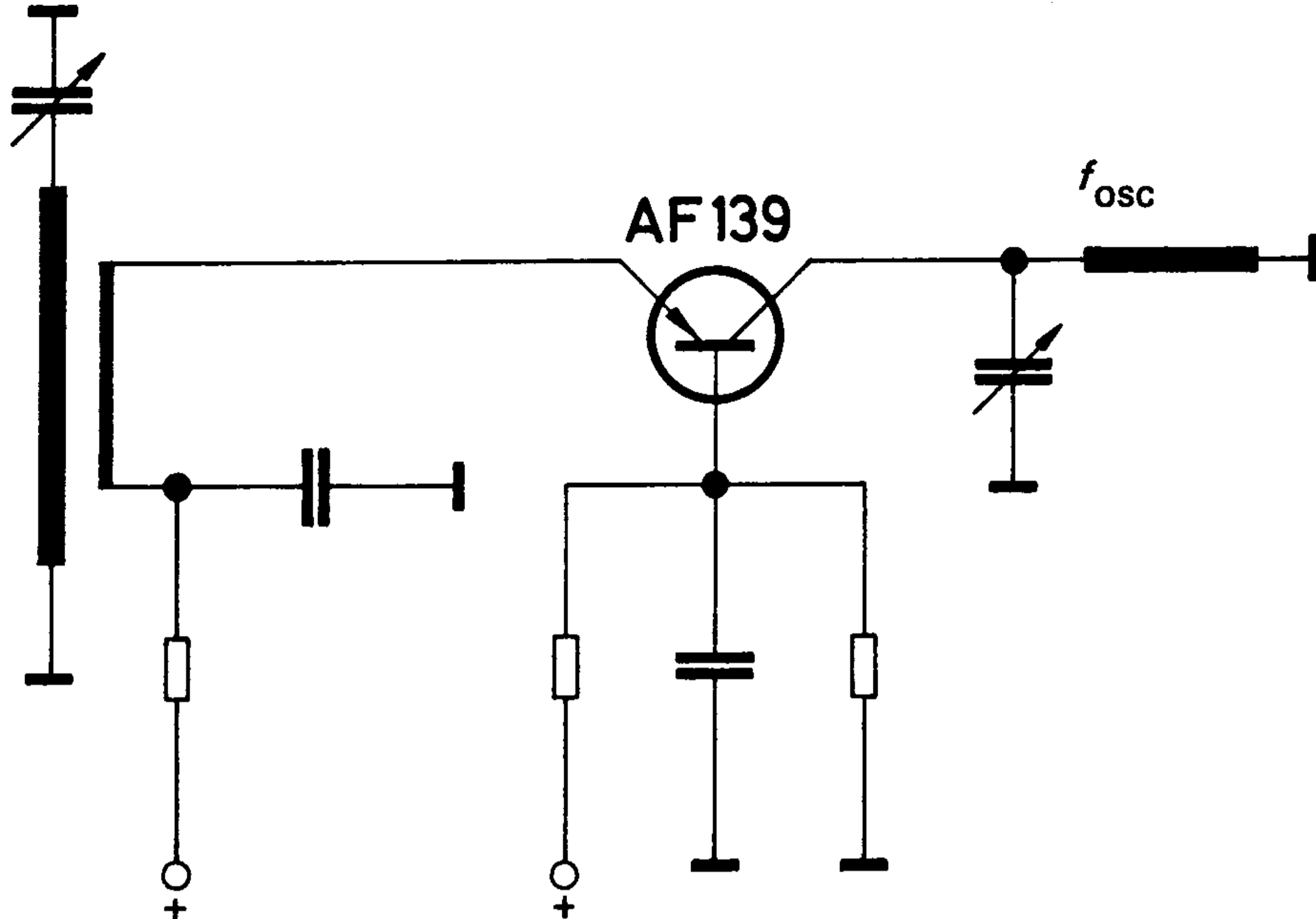


Fig. 9.1

As, however, a non linear relation between frequency and temperature often exists the compensation is not efficient to the same extent for the entire tuning range. Tolerance calculations have proved that, for instance, with two compensating condensers of 0.6 pF and 0.8 pF a shift of 2 MHz is possible due to the spreads of the capacitance and the temperature coefficient values at a temperature difference of 30 °C in the upper range of the tuner. Simultaneous extreme capacity- and temperature coefficient spreads, however, seldom occur so that the resulting shift will mostly remain under 1 MHz. Nevertheless, such shifts, even small ones, may be disturbing.

The circuit in Fig. 9.1 makes it possible to compensate the oscillator frequency drift of a UHF-tuner. If at 500 or 800 MHz the operating point in the indicated circuit will be shifted in such a way that either the current or the voltage will be kept constant, and the voltage or the current, respectively, will be varied, the result will be a frequency shift of the oscillator wave as demonstrated in Fig. 9.2. This change in frequency is caused by the output impedance of the transistor depending on the operation point. Thus, a voltage-variable capacitance diode for the retuning can be omitted. The resulting frequency shifts are sufficient in order to realize an automatic retuning. It is advantageous to operate with a constant battery voltage and a varying

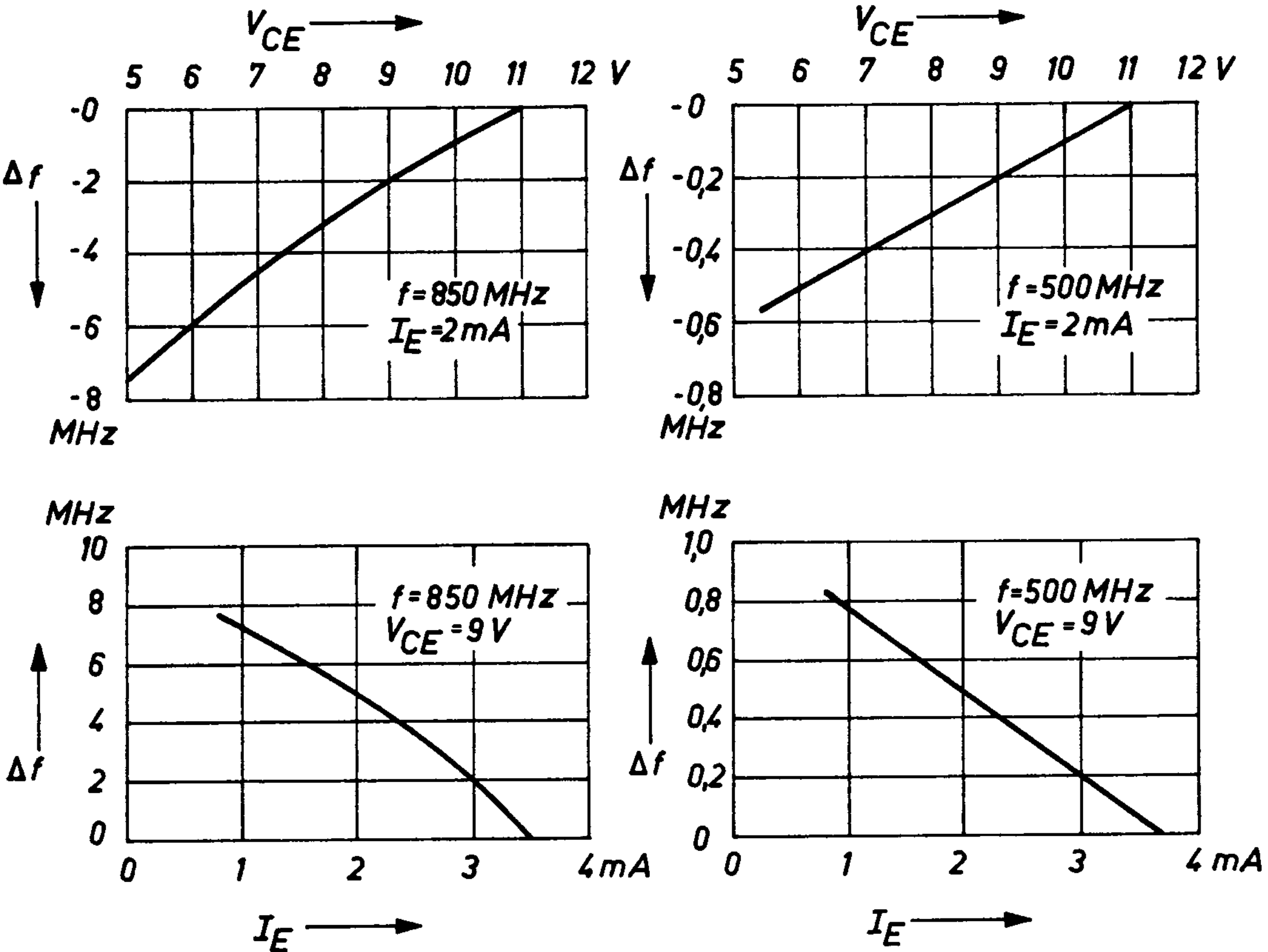


Fig. 9.2

collector current. For the mid-frequency the setting of a current of 2 mA is preferable. This value results in a nearly maximum power gain so that for the current changes no considerable loss in gain will occur.

The detuning demonstrated in Fig. 9.2 has been measured in a tuner with two transistors AF 139 in the pre-stage and in the mixer-oscillator stage. For this purpose, a frequency stable signal has been applied to the input and the change of frequency has been measured by the IF. As shown in Fig. 9.2, a change in current of 1 mA causes a change of frequency of 2 to 3 MHz. Practical experience shows that such a large frequency shift almost never occurs. A wide lock-in range, however, is very useful for an easy tuning.

Fig. 9.3 shows an experimental layout with a push-pull discriminator fed by the third stage of a three-stage TV-IF-amplifier. The discriminator coils  $L_4$ ,  $L_5$ ,  $L_6$  are coupled with the primary circuit in such a way that the passband curve will not be deformed too much. The discriminator is tuned to the video carrier frequency 38.9 MHz.



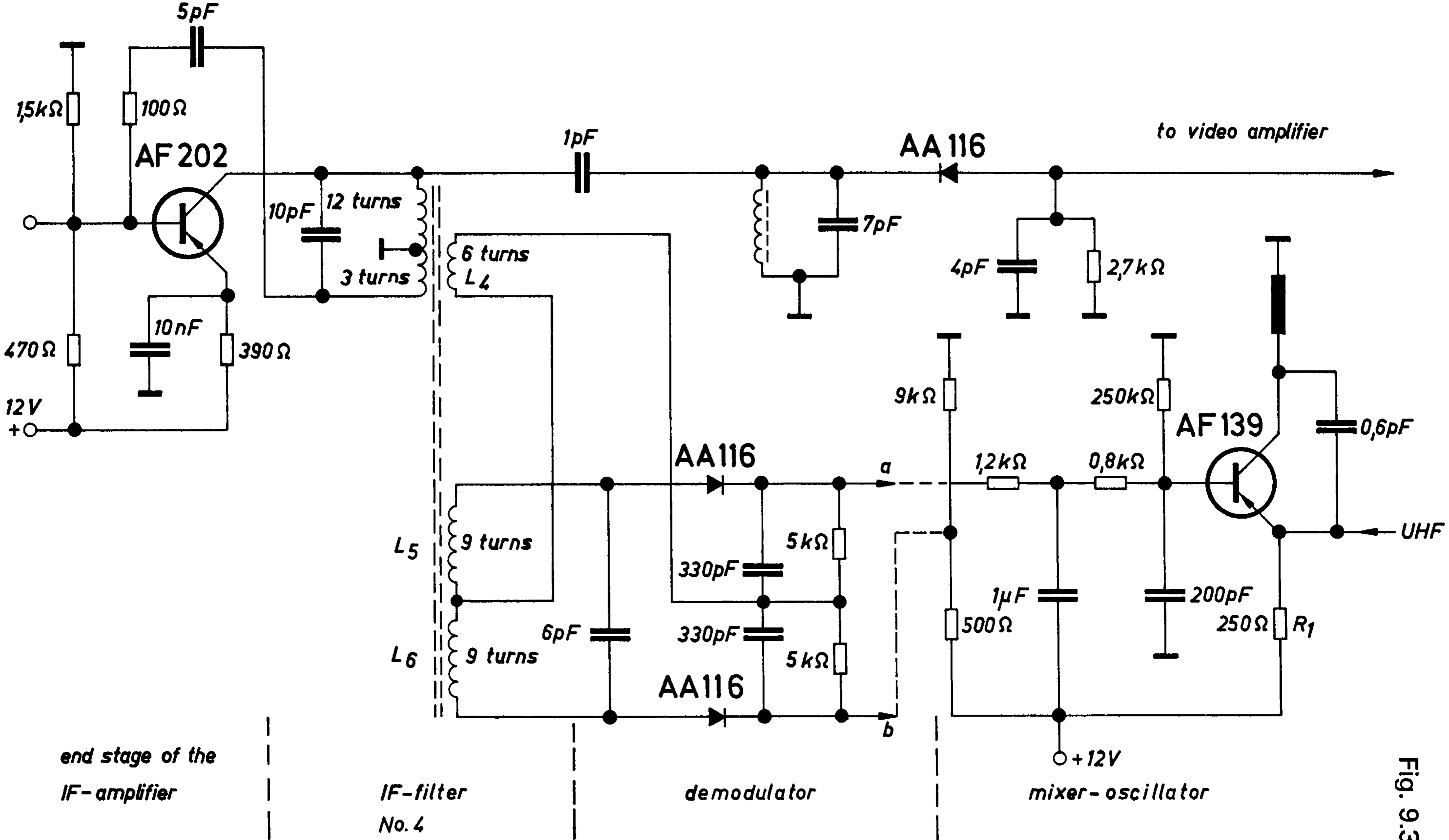


Fig. 9.3

In the experiment no additional amplifier has been connected between the terminals a and b. The discriminator directly influenced the mixer-oscillator transistor AF 139. From this fact results a relatively high base resistance. For the design it is recommended to take in account a sufficient thermal stability inspite of the fact that the discriminator also compensates the temperature shift of the transistor base current and even makes a current drive possible.

For that reason, load resistances of 5 to 10 k $\Omega$  are possible at the discriminator. The load resistances and the transistor-base-current determine the power which is drawn from the IF primary circuit in order to warrant a sufficient retuning. The required change of the collector current in the oscillator-transistor remains below 1 mA. The transistor requires a base current of  $I_B = 50 \mu\text{A}$  for a collector current of 1 mA at a minimum current gain of  $B = 20$ . As the discriminator also can compensate the increase of cut-off current due to the rise in temperature, the need of control current is augmented by the change of cut-off current. However, it probably will not exceed 100  $\mu\text{A}$ . The associated change of the control voltage remains low if a small emitter resistance for the oscillator transistor will be chosen. Due to the resistor  $R_1$  of 250  $\Omega$ , the required control voltage remains below 0.5 V. The retuning capability of the layout has been tested by an unmodulated signal at the input. The test was made at the upper range limit. When the input frequency of 6 MHz was changed corresponding to the change of the oscillator frequency, a deviation of the IF frequency of only 0.6 MHz was measured. Consequently, the control factor is 1 : 10.

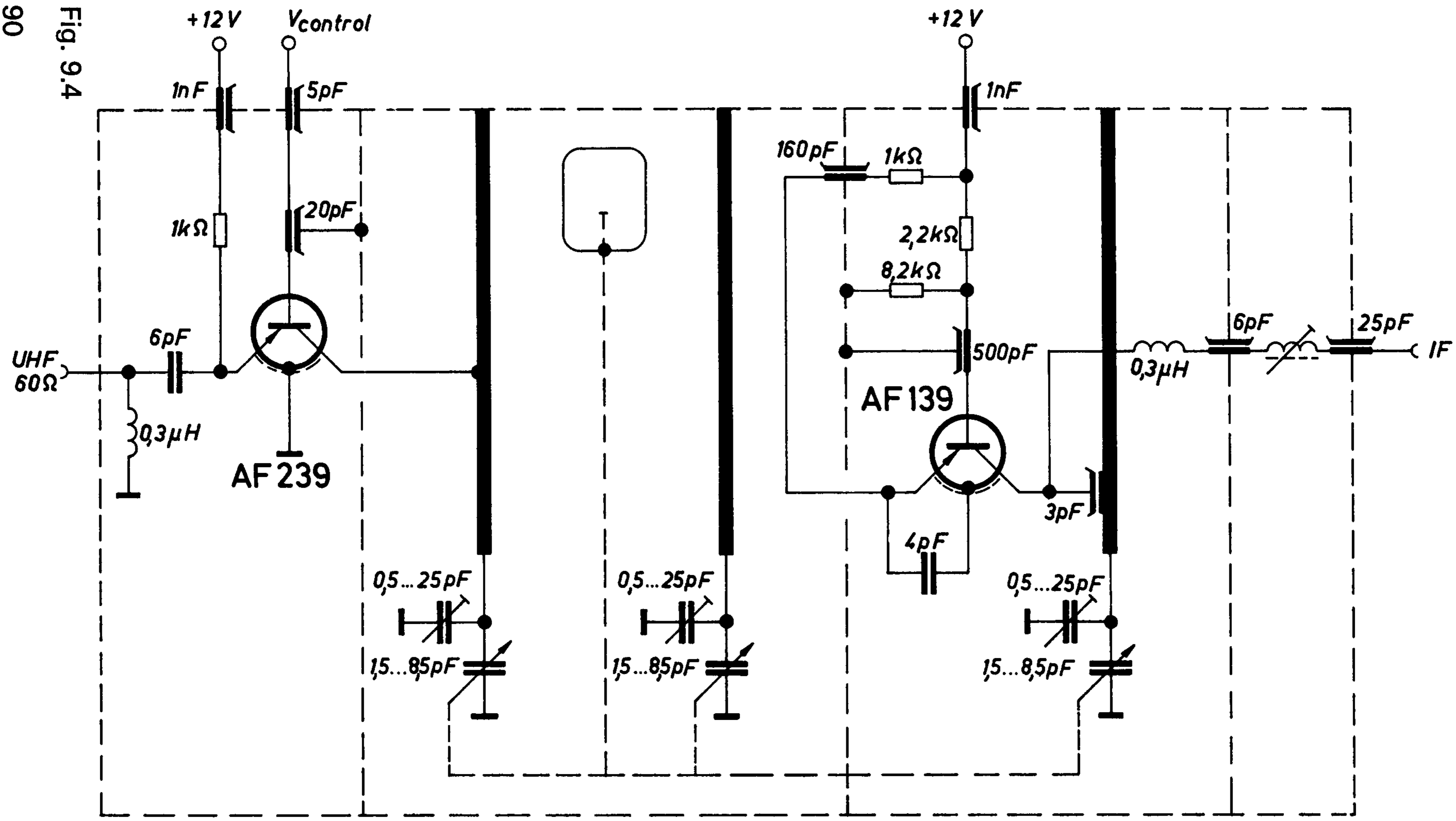
No further tests have been made so far in a complete TV set. Some problems certainly will arise by the combination of the components e.g. control oscillations are possible. Picture disturbances as a consequence of the loading by the discriminator may be avoided by its good matching. The total power gain will not considerably be decreased by the retuning.

Employed coil slug: Siferrite thread core B 63310 U 17 A 12,3

## 9.2 Low Noise UHF-tuner with AF 239

The noise qualities of a tuner are determined substantially by the characteristics of the prestage. With the transistors available today the average noise figures of UHF-tuners were at 470 MHz approximately  $F = 5$  and at 860 MHz approximately  $F = 11$ .

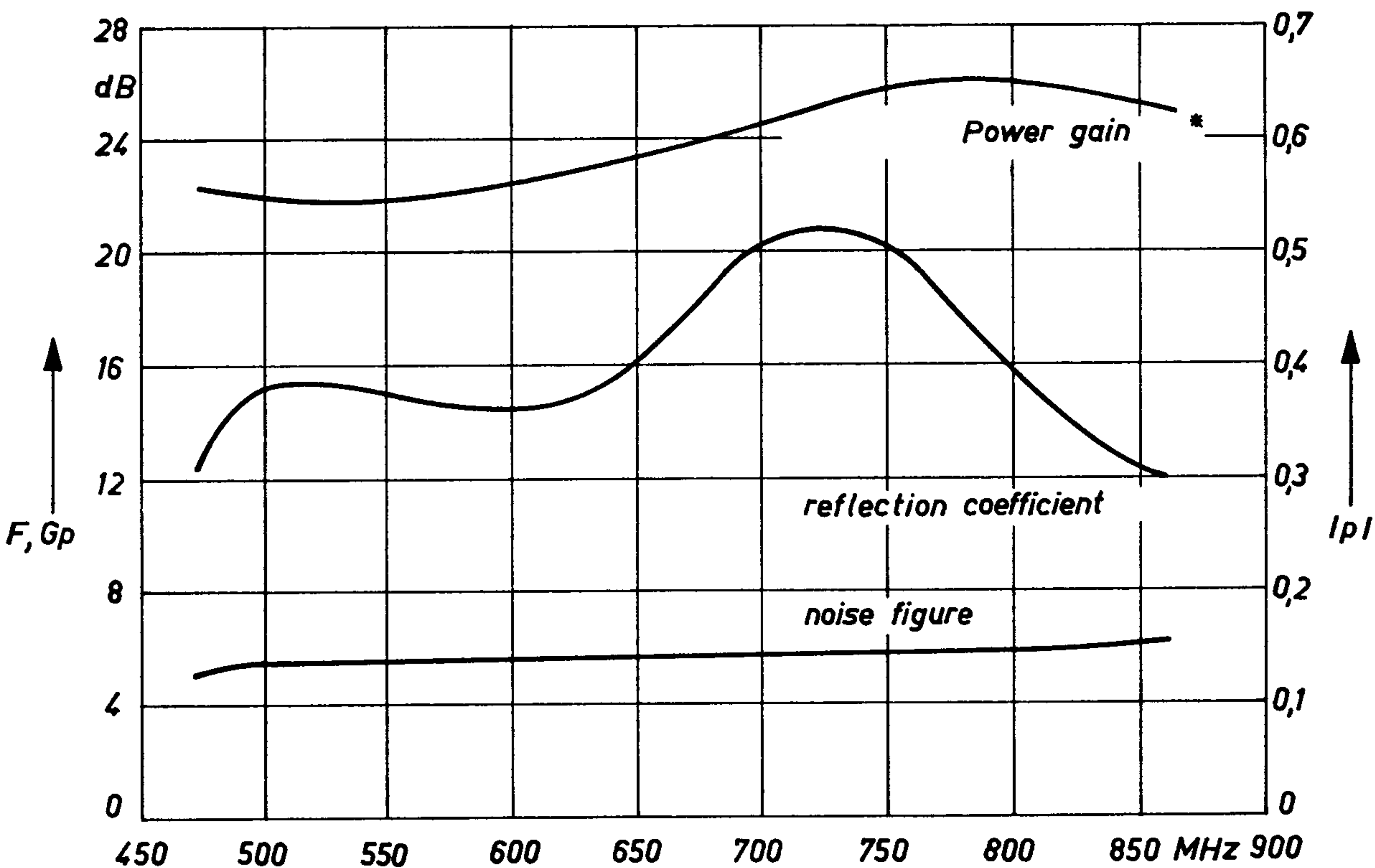
Fig. 9.4



The new mesa transistor AF 239 perfectly qualifies for use in the UHF prestages because of its excellent noise and gain characteristics. In connection with an oscillating mixer-stage, equipped with AF 139, UHF-tuners can now be built the average noise figures of which are approximately 3.5 at 470 MHz and 4 at approximately 860 MHz. Fig. 9.4 indicates such a transistorized circuit. The layout of the prestage is similar to a stage with the AF 139. The higher gain of the AF 239 requires only a careful selection of the ground points, above all that of the transistor grounding and the base capacitance. In the laboratory layout the lowest reflection-coefficients were always obtained if the connection between antenna-input and emitter terminal was very short. The maximum prestage gain is achieved at a collector current of  $I_c = 2.6$  mA. With regard to the total characteristics of the mixerstage, the mixer is operated at a collector quiescent current of  $I_c = 2$  mA. Fig. 9.5 indicates values of gain, noise figures and a reflection coefficient value which depend on the incoming frequency.

### Technical Data

Supply voltage	12 V
Power gain	22 to 26 db
Noise figure	3.5 to 4 (5.5 to 6 db)
Reflection coefficient	0.3 to 0.5



\* IF-bandwidth 7.5 MHz  
dip 0.5 db

Fig. 9.5

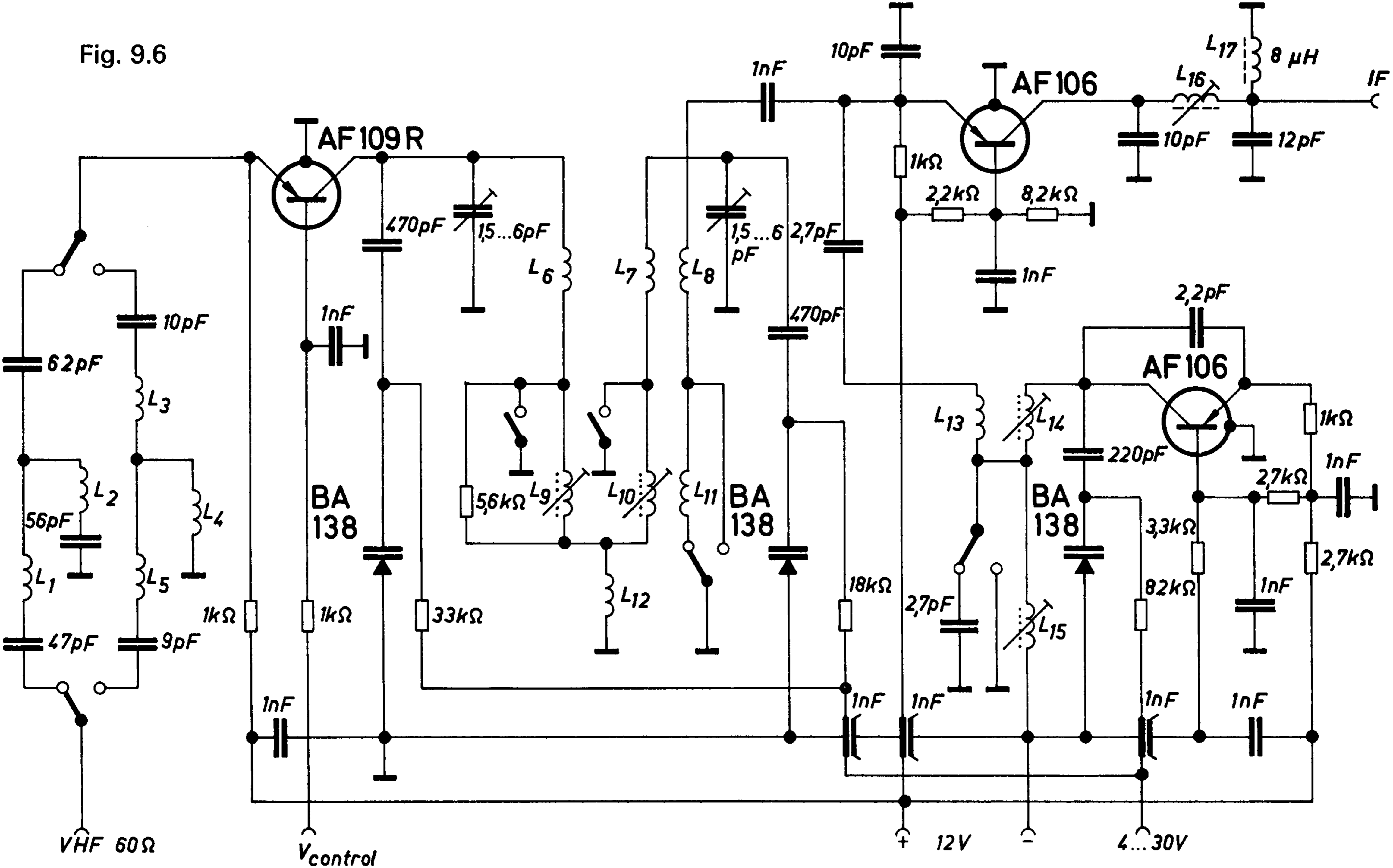
### 9.3 Electronically Tunable VHF-Tuners

An increasing number of varicaps are applied for the tuning of band I and band III, and the bands are still switched mechanically. The main advantage of this tuning principle is the fact that no movable RF-contacts are any longer required for the tuning. This essentially increases the reliability. Without the tuning mechanic a very compact layout and a simplified pushbutton control is possible.

Three epitaxial Si-planar-diodes are used for tuning, two of them in the VHF bandfilter between prestage and mixer, and one in the oscillator. The diodes which have a spread in their initial capacitance are divided into two capacitance groups. Diodes with green rings do have an initial capacitance of 3.8 to 4.9 pF ( $U_R = 30 \text{ V}$ ); diodes with blue rings do have an initial capacitance of 4.4 to 5.5 pF. The capacitance ratio is between 2.4 and 2.7 for a DC voltage change of 3 to 30 V.

Fig. 9.6 shows a tuner circuit which is comparable with former layouts with the exception of its tuning elements and its wideband input filters. The new control transistor AF 109 R is employed in the prestage; one AF 106 is used in the mixer and one in the oscillator stage. All 3 transistors are operated in common base configuration because of equal gain in band I and III. The operating point of the prestage is approximately 3.7 mA in the uncontrolled state; that of the mixer and the oscillator is about 2 mA. A wideband filter, which can be switchselected for band I and III, is at the input of the prestage. For economical reasons, this filter is generally preferred to a selectively tunable RF-circuit. Herewith a small prestage noise and a good far-off selectivity are obtained. The disadvantage is the input matching which is relatively difficult to attain and the higher cross modulation sensitivity which, however, only becomes effective if a second and stronger TV transmitter is received in the same band. The inductivity  $L_2$  which is located in the filter branch for band I and the capacitance (56 pF) connected in series increase the input attenuation in the IF range. The tunable, inductively coupled VHF bandfilter is connected to the collector circuit of the prestage transistor. The coils of band I are connected in series to the coils of band III. Tracking deviations of this filter which result from the frequency dependent input impedance of the mixer can partly be compensated by using one of the above mentioned capacitance groups in the primary section of the filter and in the secondary section. The mixer is connected by the

Fig. 9.6



coupling coils  $L_8$  and  $L_{11}$ . The primary circuit of the IF filter is located in the output. The supply voltage is applied to the collector via a choke of  $8 \mu\text{H}$ .

The oscillator has a capacitive feedback. The RF voltage is applied to the emitter of the mixer via a condenser of  $2.7 \text{ pF}$ . The RF voltage at the oscillator circuit and, thus, at the varicap is very low ( $< 2.5 \text{ V}$ ) in order to avoid synchronization troubles for the small tuning voltages.

Technical Data	Band I	Band III
Power gain <sup>1)</sup>	25 to 26	25 to 27 db
Noise figure	3 to 3.6	3.4 to 4.2
Reflection coefficient	0.4 to 0.5	0.4 to 0.5
Range of control	$>35$	$>35 \text{ db}$
VHF band width	10.5 to 12.5	10 to 12 MHz
IF rejection	$>50$	$>60 \text{ db}$
Interfering voltage for 1 % cross modulation at $60 \Omega^2$ ) ( $f_{\text{inter}} = f_{\text{signal}} + 14 \text{ MHz}$ )	$>20$	$>14 \text{ mV}$

Coil Data	Coil Diameter	Core Material
$L_1 = 10$ turns 0.35 CuL	3 mm	—
$L_2 = 17$ „ 0.35 CuL	3 mm	—
$L_3 = 3$ „ 0.35 CuL	3 mm	—
$L_4 = 3$ „ 0.35 CuL	3 mm	—
$L_5 = 3$ „ 0.35 CuL	3 mm	—
$L_6 = 2.5$ „ 0.75 CuL	4.3 mm	—
$L_7 = 2.5$ „ 0.5 CuL	4.3 mm	—
$L_8 = 1.5$ „ 0.35 CuL	4.3 mm	—
$L_9 = 13$ „ 0.35 CuL	4.3 mm	brass
$L_{10} = 13$ „ 0.35 CuL	4.3 mm	brass
$L_{11} = 3.5$ „ 0.35 CuL	4.3 mm	—
$L_{12} = 8$ „ 0.35 CuL	3 mm	—
$L_{13} = 1$ „ 0.5 CuL	4.3 mm	—
$L_{14} = 2$ „ 0.75 CuL	4.3 mm	brass
$L_{15} = 5$ „ 0.35 CuL	4.3 mm	brass
$L_{16} = 12$ „ 0.35 CuL	5 mm	B 63310 U 17 A 12.3
$L_{17} = 30$ „ 0.15 CuL	3.5 mm	B 61935 Si 31 SD

<sup>1)</sup> band width of the IF filter 7.5 MHz; dip 0.5 db

<sup>2)</sup> half of the EMF of a 100% sine modulated TV-transmitter is effective as interfering voltage.

## 9.4 TV IF Amplifier with Silicon Transistors

As already known, it is impossible to design silicon RF-transistors which with similar geometrical shape show the same RF-characteristics as germanium transistors. One has to choose much smaller structures and certain arrangements have to be made in order to decrease the feedback capacitance. One succeeded in obtaining a sufficiently small feedback capacitance in the silicon transistors BF 167 and BF 173 by the screening method. Herewith TV IF amplifiers can be built which do not require any neutralization. The control type BF 167 makes a range of control of 60 db possible. The maximum amplification is approximately 30 dB. The power gain of the total three-stage amplifier is 84 db according to Fig. 9.7. The output voltage amounts to 6 V referred to 30% synchronous pulse compression. The operation points show values as follows:

1st stage: 10 V/4 mA (unregulated)  
2nd stage: 10 V/3 mA  
3rd stage: 13 V/7.3 mA

The output stage must have a resistance of 220  $\Omega$  between collector and band filter in order to obtain at large signals a sufficient stability and constance of the transmission characteristics. Compared to a concept with pnp transistors the requirements of material are a little larger if the minuspole of the supply voltage is connected to chassis, and if the voltage supply is derived from the tube anode voltage. Then, three additional RC-filters are required in the collector circuits.

### Technical Data

Operation voltage	20 V
Power gain	84 db
Range of control	60 db
Output voltage after rectification at 2.7 k $\Omega$	6 V

### Coil Data

Bobbin 5 mm  $\varnothing$

Coil Cores B 63310 U 17 A 12.3

$L_T = 18$  turns 0.3 CuLL

$L_1 = 5$  „ 0.7 Silverplated copper wire,  
tap at 1 $\frac{1}{4}$  turns

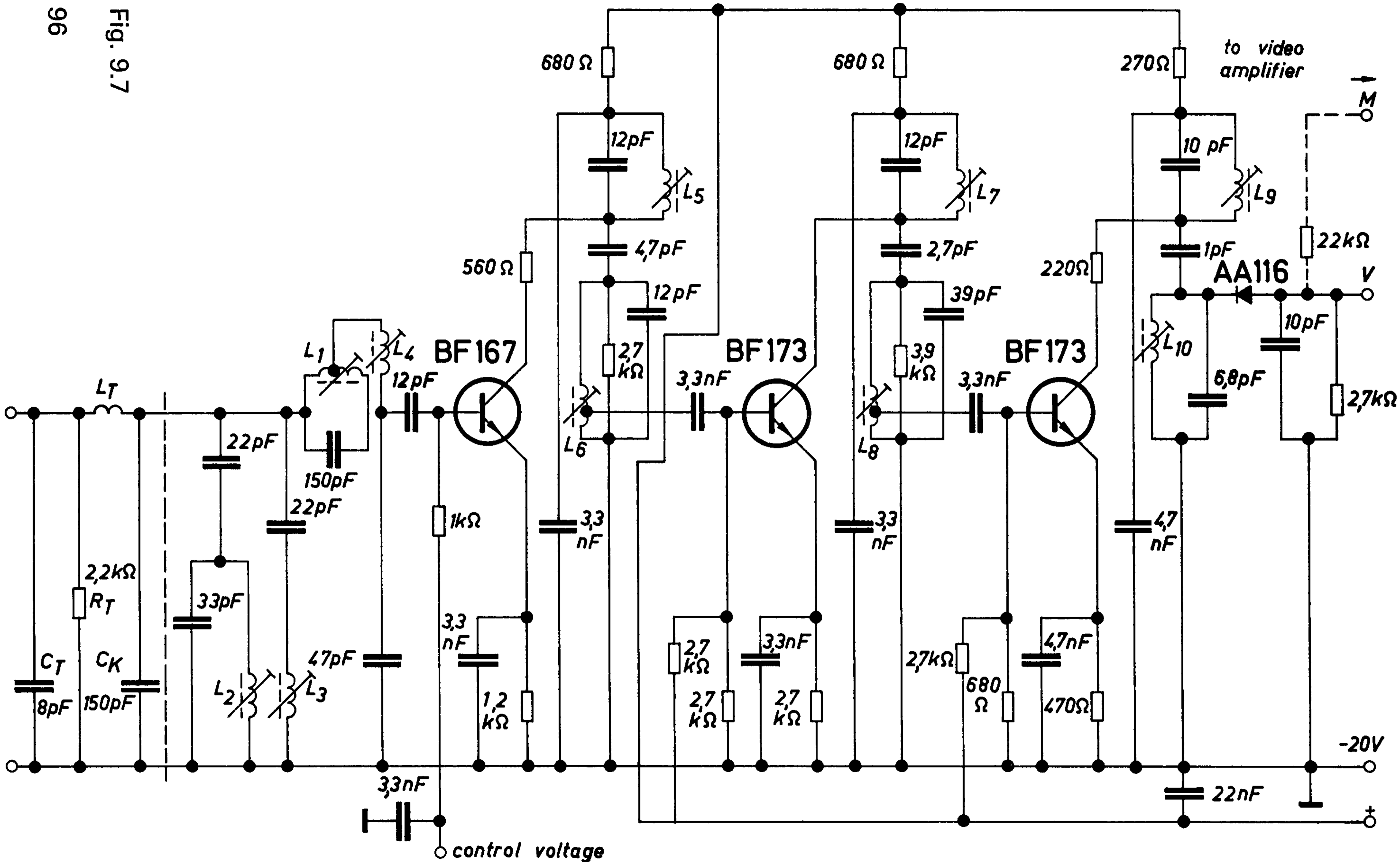
$L_2 = 9$  „ 0.7 Silverplated copper wire

$L_3 = 15$  „ 0.4 CuLL

$L_4 = 7$  „ 0.3 CuLL



Fig. 9.7



$L_5 = 14$	„	0.3 CuLL
$L_6 = 14$	„	0.3 CuLL tap at 5 turns
$L_7 = 14$	„	0.3 CuLL
$L_8 = 7$	„	0.3 CuLL tap at 5 turns
$L_9 = 14$	„	0.3 CuLL
$L_{10} = 18$	„	0.3 CuLL

## 9.5 One-Stage Video Amplifier

In Fig. 9.8 the circuit of a one-stage video amplifier is shown with the silicon planar transistor BF 110. The contrast control has to take place at the output of the amplifier with regard to the adjacent subassemblies the driving signal of which is to be as constant as possible. This fact makes contrast controlling in an one-stage video amplifier difficult. The control network at the output permits a control range of approximately 4 : 1. The black level is kept constant at the contrast control by a DC bridge. The total network acts as operating resistor, its value being 4.8 k $\Omega$ . Thus, at an effective battery voltage of 125 V the maximum dissipation power is approximately 800 mW. With a cooling clamp of 8 cm<sup>2</sup> and low-resistive biasing circuit, the collector current change which is caused by the self-heating of the transistors, can be kept at a value of  $\leq 1$  mA.

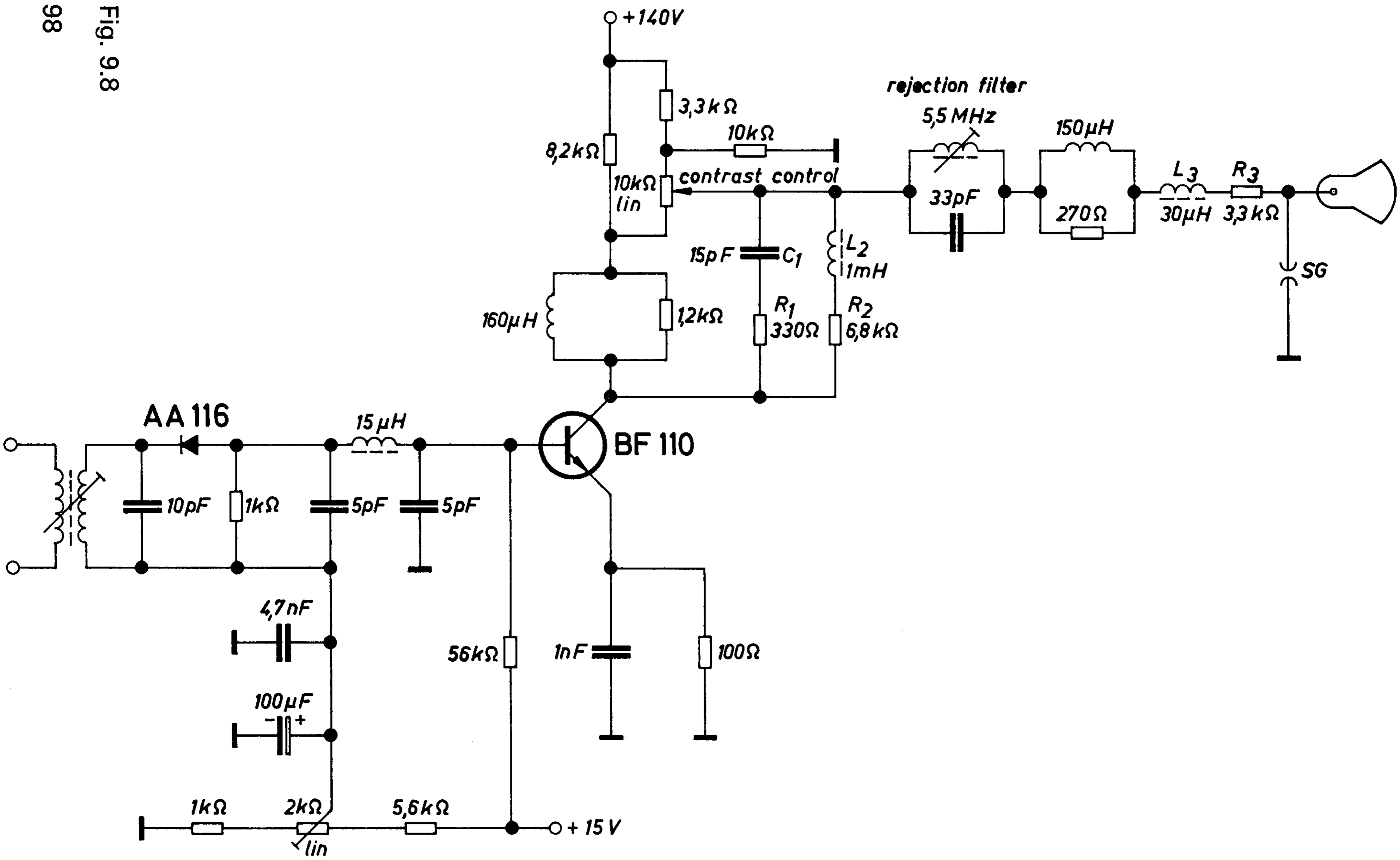
This value corresponds approximately to a shade of grey and can be accepted.

The frequency characteristic of the contrast controller is compensated by two correction filter sections ( $R_1, C_1, R_2, L_2$ ) (Fig. 9.9). The protective circuit in the cathode feedline is of special importance. It contains a resistor  $R_3$ , the inductivity  $L_3$  and a sparking gap  $F$  which avoids damages of the video transistor at high-voltage flash-overs in the cathode ray tube. It is recommended to connect this combination directly to the tube base. The driving signals for sound IF amplifiers, amplitude filter and control can be tapped off at the emitter and the collector of the transistor BF 110 with the level which is independent from the contrast control.

### Technical Data

Battery voltage	140 V
Operating voltage at the transistor	125 V
Voltage gain factor	30
Output voltage	$> 100 V_{pp}$

Fig. 9.8



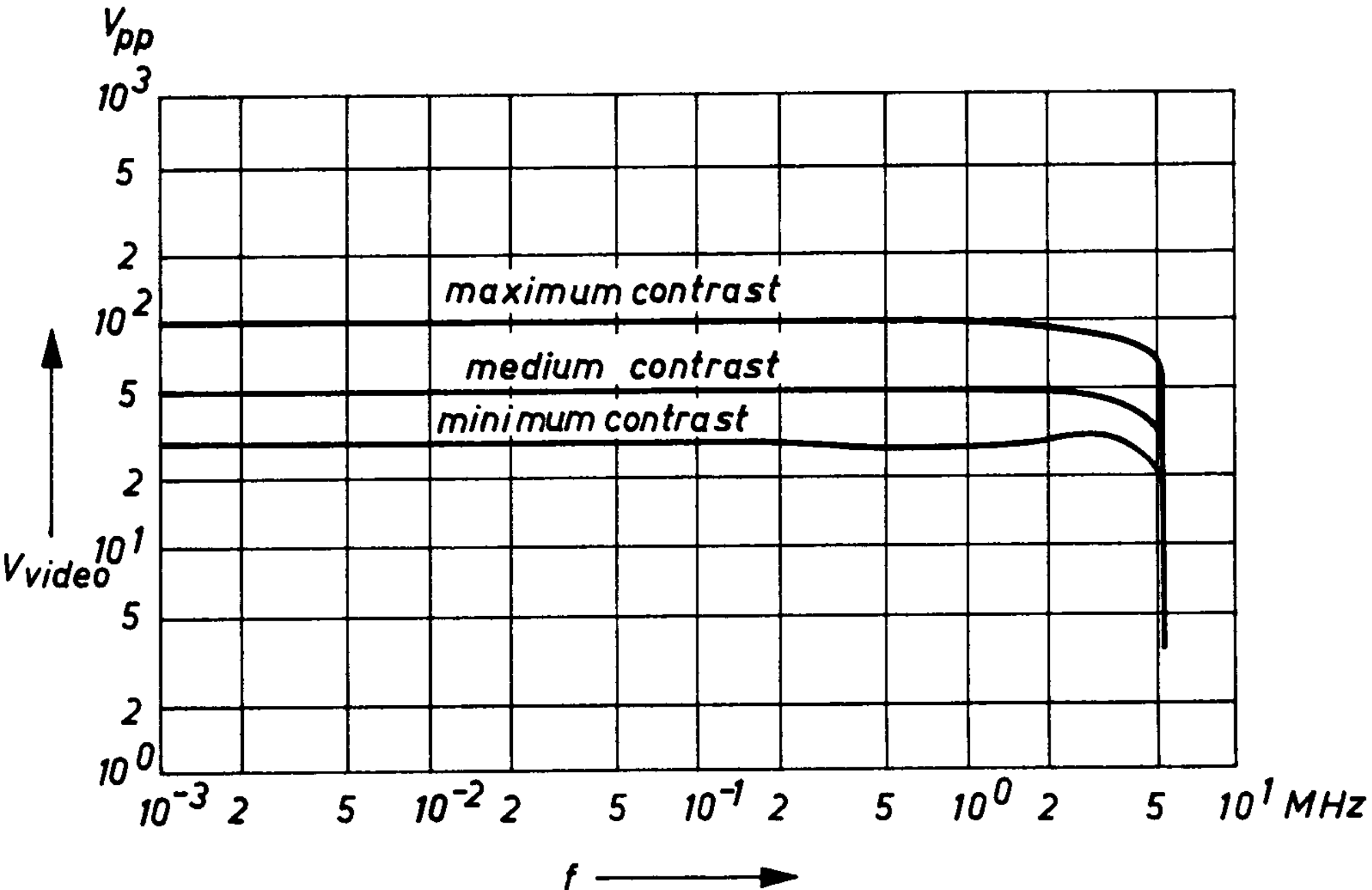


Fig. 9.9

## 9.6 Two-Stage Video Amplifier

Fig. 9.10 shows the circuit of a two-stage video amplifier with the silicon planar transistor BF 110 in the output and a BC 108 in a common base circuit as driver. The driver transistor makes it possible to build a control branch which is not sensitive as far as circuit capacitances and layout are concerned. Fig. 9.11 shows the frequency characteristics of the output voltage at various contrast settings. In this example the contrast is set at the base of the output transistor. The control network is arranged as DC bridge by which the black level can be kept constant during the control. The output stage is suitable for connection to a supply voltage of 240 V. The parallel connection of the two resistors of the potential divider at the collector serves as load resistor. The effective voltage at the transistor is 125 V if the nominal voltage is 240 V. This allows an output voltage of  $\geq 100 V_{pp}$ . In this case at excessive voltage of the mains supply or also during the heat-up time of the equipment, the permitted reverse voltage  $V_{CER}$  of the transistor will not be exceeded. The maximum dissipation power of the transistor in this design amounts to approximately 800 mW. The change of the collector current which is caused by the self-heating of the transistor is kept on a value of  $< 1$  mA by a cooling clamp of  $8 \text{ cm}^2$ .

The signals for the sound-IF amplifier, amplitude filter and control can be tapped off at the emitter or collector of the driver

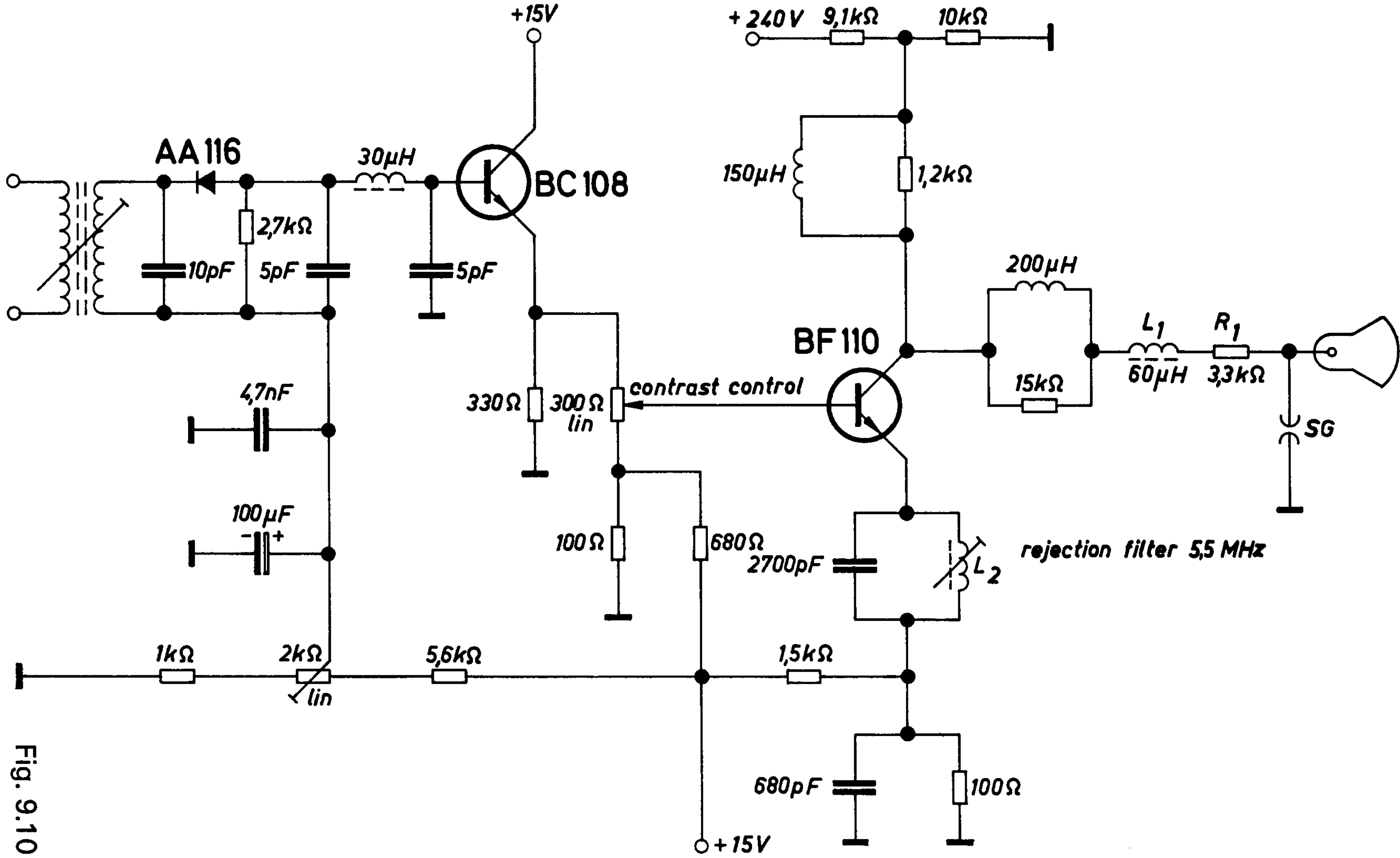


Fig. 9.10

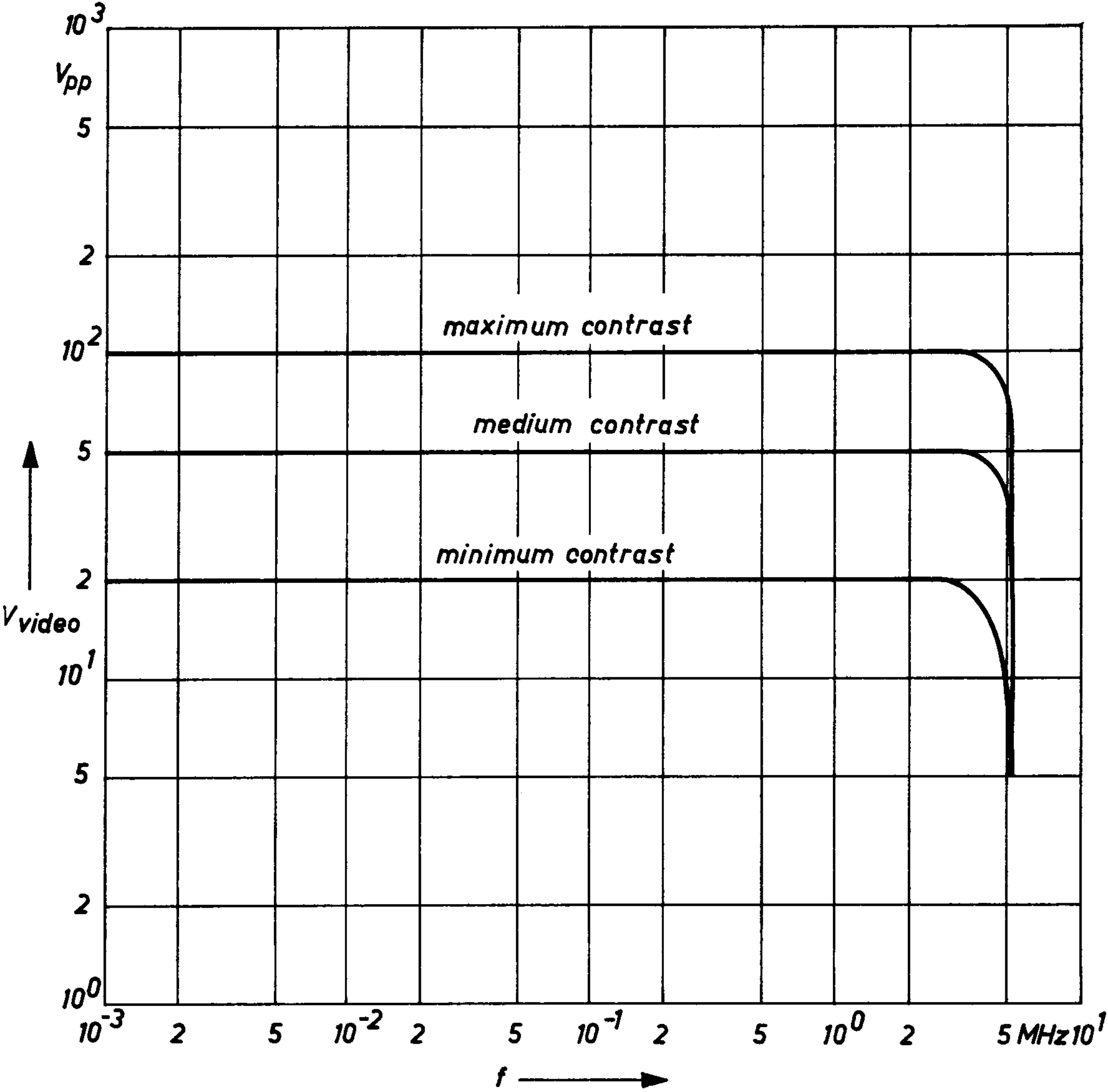


Fig. 9.11

transistor. Of special interest is the protection circuit in the cathode feedline with the resistor  $R_1$ , the inductivity  $L_1$  and the sparking gap  $F$ . The latter avoids that the output transistor is damaged by high voltage flash-overs within the tube.

#### Technical Data

Operation voltage	240 V
Operation voltage at the transistor	125 V
Voltage gain factor	30
Output voltage	$> 100 V_{pp}$

Inductivity  $L_1$ : 10 turns 0.8 silverplated copper wire, pitch 1.6 mm/turn

## 9.7 AM-FM IF Amplifier without Neutralization

In radio sets the IF-amplifiers are mostly neutralized. The reverse capacitance of the germanium mesa transistors AF 200 and AF 201 is so small that AM-FM IF amplifiers can be designed which warrant high gains without neutralization and show excellent stability. Fig. 9.12 indicates the circuit of such an amplifier, the first stage of which operates as a self-oscillating mixer in the AM-range and which is succeeded by a two-stage IF-amplifier.

The first IF stage can be controlled and a range of control of 50 db is attained. Since in most cases transistorized AM-FM IF amplifiers usually are employed in battery supplied sets, a low current consumption is required. Therefore, it is recommended to apply the current decrease control instead of the current increase-control usually employed in transistor designs inspite of slightly more unfavorable results. One of these disadvantages is the lower overdrive capability because the first IF stage has to be operated at low operating points in the case of high input signals. For this reason a damping diode is employed in the first IF circuit. The damping diode is connected to the first IF circuit which is low resistive in order to minimize modulation distortions.

The operating currents, which are indicated in the circuit, may be reduced if a particularly low current consumption is required. In this case, however, lower gain values and slightly larger bandwidths may result as depicted in the table below:

Collector current	1	2	4	mA
Output voltage (60 $\Omega$ )	13	23.5	42	mV
Input voltage (60 $\Omega$ )	0.57	0.53	0.5	mV
Band width	6.5	6	5	kHz
Power gain	28	33	38.6	db

A further possibility for the protection of the IF amplifier against excessively high input voltages is a control of the antenna input by a diode which is indicated in Fig. 9.12 by an interrupted line. In this layout also the self-oscillating mixer will be protected against excessively large signals. The attainable total gain including the demodulation is approx. 80 db. In the

FM-range (10.7 MHz) all three stages operate as an IF amplifier. In this case also the attainable gain again depends on the adjusted operating point.

Collector current	1	2	4	mA
Power gain	26*	26	31	db

\* secondary coil of the filter 3 turns instead of 2 turns at a higher current.

In order to obtain a good stability of the resonant curve for large signals a collector-resistor of 250 or 470, respectively, has to be applied.

### Technical Data

Supply voltage	9 to 12 V
Gain AM-range	approx. 80 db
FM-range	approx. 80 db
(including tuner-gain of 30 db)	
Range of control (AM-range)	50 db

### Inductivities

FM: TU. 1 = TU. 2: primary  $1.1 \mu\text{H}$ ; transformer ratio = 5.5 : 1  
 TU. 3: primary  $1.1 \mu\text{H}$ ; secondary  $2.2 \mu\text{H}$   
 AM: TM. 1: primary  $45 \mu\text{H}$ ; secondary  $320 \mu\text{H}$   
 TM. 2:  $320 \mu\text{H}$   
 TM. 3: primary  $120 \mu\text{H}$ , transformer ratio = 2 : 1



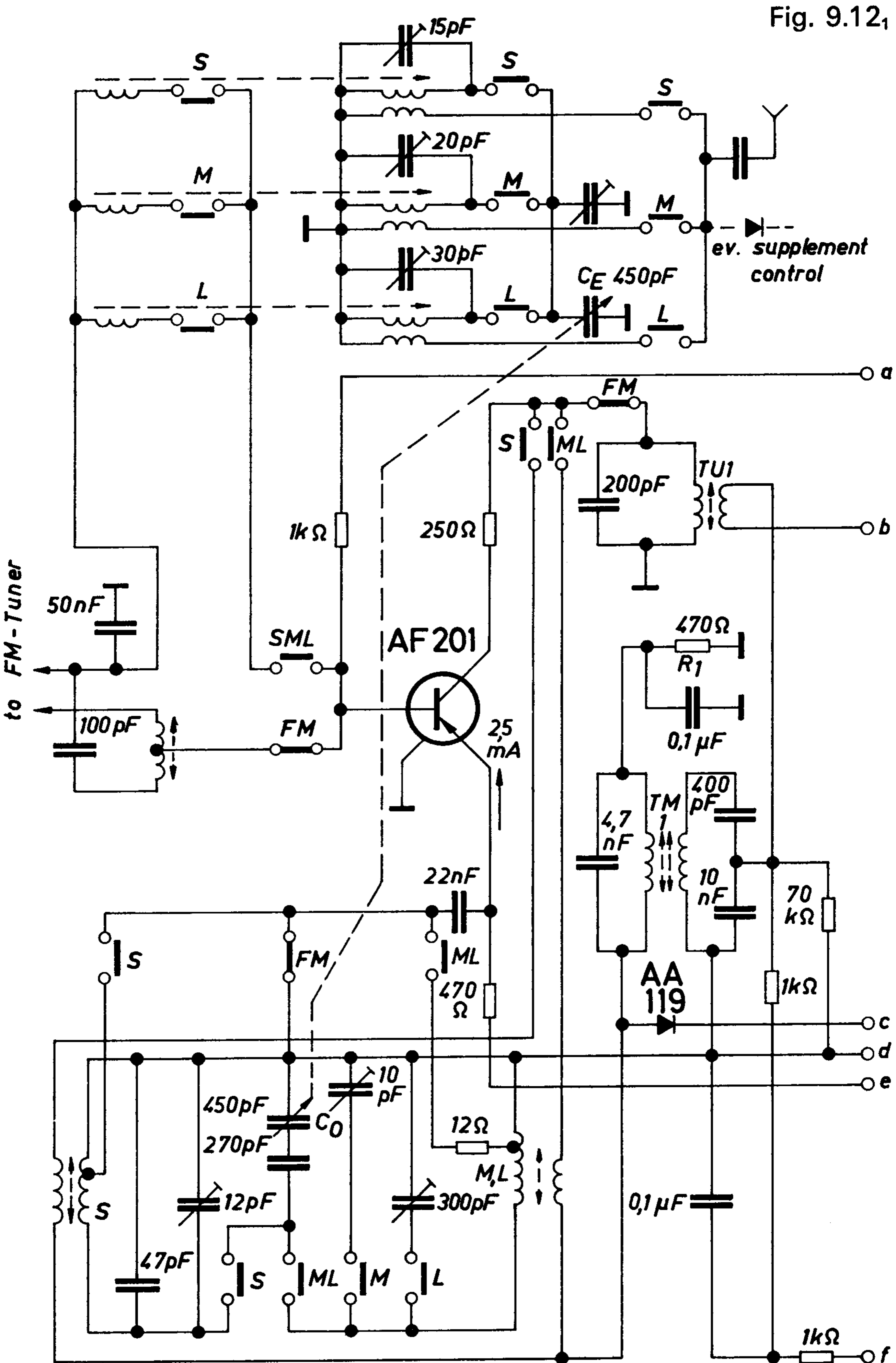
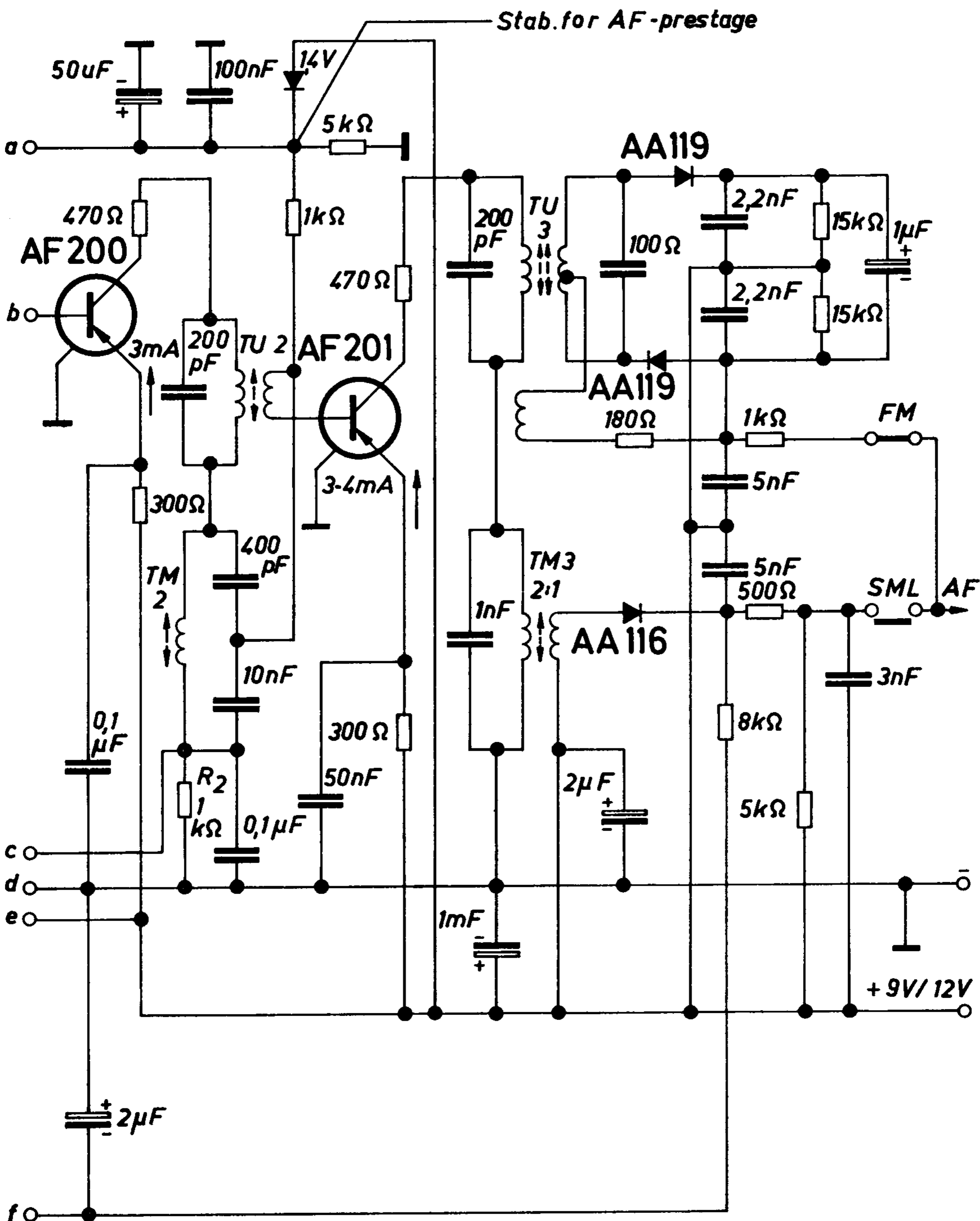


Fig. 9.12<sub>2</sub>



## **10. Abbreviations**

**CuL**      enameled copper wire

**CuLL**    double enameled copper wire

**K**        distortion factor

**L**        air gap