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Design Examples of Semiconductor Circuits

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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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Contents

1. Audio Frequency Amplifiers

Because of their advantages for low frequency applications, such as low LF noise, high current gain and small cut-off currents, silicon planar transistors are used more and more in AF pre-stages. In the following examples of AF amplifiers, this trend is taken into consideration and, therefore, in pre-stages there is the almost exclusive use of the universal types BC 107, BC 108, and BC 109.

The electrically equivalent epoxy types BC 147, BC 148 and BC 149 may, of course, be applied in the same manner as the transistors in the smaller epoxy case BC 167, BC 168 and BC 169. The last figure of the three digit series number always indicates the correspondence of types, e. g., the epoxy types BC 147 and BC 167 correspond to the metal type BC 107 etc.

Because of the high current gain and cut-off frequency of these transistors, the latter being very high for AF applications, special circuit designs are sometimes necessary in order to prevent undesired oscillations. More details will be given in the following examples.

Transformerless stages with complementary transistors have made considerable progress in AF power stage applications (complementary transistors being pnp and npn transistors with inverse electrical characteristics.). So far, these transistors are available for medium and higher power applications and are manufactured only of germanium. For that reason, further on alloyed germanium transistors are used in the AF power stages. These transistors have low saturation voltages and, therefore, can be used with high efficiency even at lower operating voltages. The germanium alloy technology allows the production of transistors with very good linear gain characteristics, i. e. transistors showing high current gain even at high collector currents.

High current gain of the power stage transistors is of special importance for transformerless power stages because the driver stages can be operated at lower currents. For that reason the pair AC 187 K/188 K with a current gain of $B > 100$ has been developed in addition to the complementary pair AC 153 K/AC 176 K.

With the larger complementary pair AD 161/AD 162 the number of high gain pairs has been increased.

As AF pre-amplifiers, our integrated AF amplifiers TAA 111, TAA 121 and TAA 131 are most useful. Besides their uncomplicated circuit design, these amplifiers offer an almost temperature-independent gain. For a temperature range from —30 to 125° C the voltage gain of the amplifier TAA 121 changes, in the average, only between 64 and 68 dB. This low temperature dependence of the gain results from the compensation of the temperature caused variation in collector resistance by the change of the emitter-base voltage in the succeeding transistor.

As shown in chapter 1.3., with such an amplifier a power stage for an output power of approximately 1.5 W may be driven.

1.1. Stereo Pre-Amplifier

The pre-amplifier shown in Fig. 1.1. is suitable as driver of the 15 W HiFi amplifier described in chapter 1.9. of the booklet "Design Examples of Semiconductor Circuits 1966". In the pre-amplifier silicon planar transistors are used, and the amplifier operates perfectly up to an ambient temperature of 70 °C. The input stage is in common collector configuration so that a very high input impedance of approximately 750 kΩ is attained. The input stage is designed in such a way that it can handle even high input voltages without being overdriven. The volume and balance controls are connected to the input stage via a series resistor so that the setting of the potentiometer does not substantially influence the input impedance of the amplifier.

With the following tone control network the emphasis or deemphasis on bass and treble can be adjusted up to 20 dB.

The quiescent point of the following amplifier stage is set via a resistor by the emitter potential of transistor T_4. This guarantees a perfect temperature stability of the circuit. If, for instance, the current in transistor T_3 increases due to a temperature rise, the bias voltage at the base of transistor T_4 tends towards negative values. This causes a decrease of the voltage drop at the emitter resistor of this stage, and transistor T_3 obtains a lower bias voltage via the resistor of 4 MΩ. A similar compensation effect occurs with variations in the operating voltage. The power amplifier already mentioned can be connected to the end stage of the amplifier via a bias resistor of 1 kΩ.

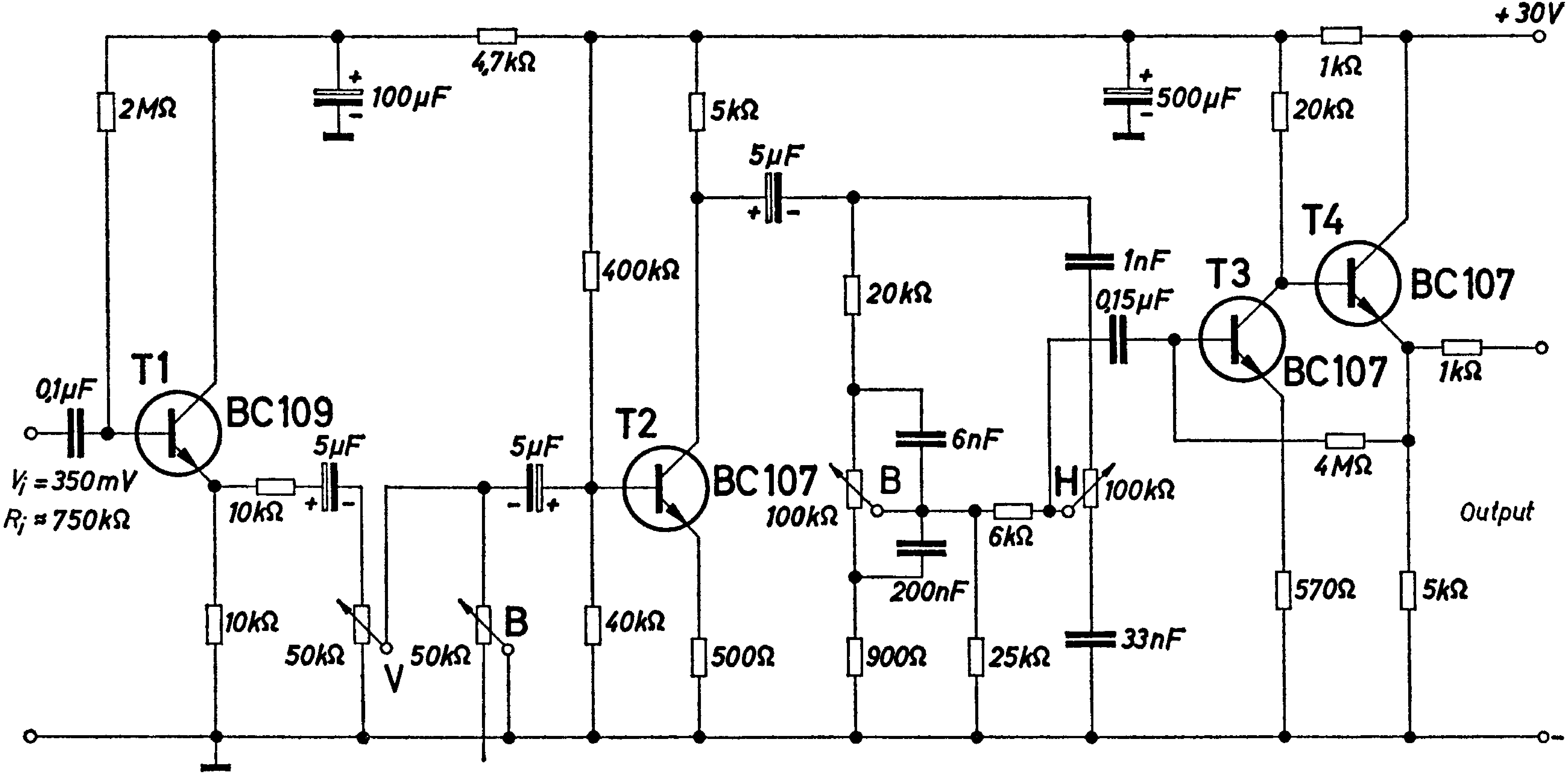

+30V
2MΩ
100μF
4,7kΩ
5kΩ
500μF
1kΩ
20kΩ
5μF
T4
1nF
400kΩ
20kΩ
T3
BC107
0,15μF
T1
0,1μF
BC109
BC107
1kΩ
5μF
T2
5μF
6nF
4MΩ
V_i = 350mV
BC107
B
H
100kΩ
R_i ≈ 750kΩ
10kΩ
100kΩ
6kΩ
Output
200nF
570Ω
5kΩ
10kΩ
B
40kΩ
33nF
50kΩ
50kΩ
500Ω
900Ω
25kΩ
V

Fig. 1.1.

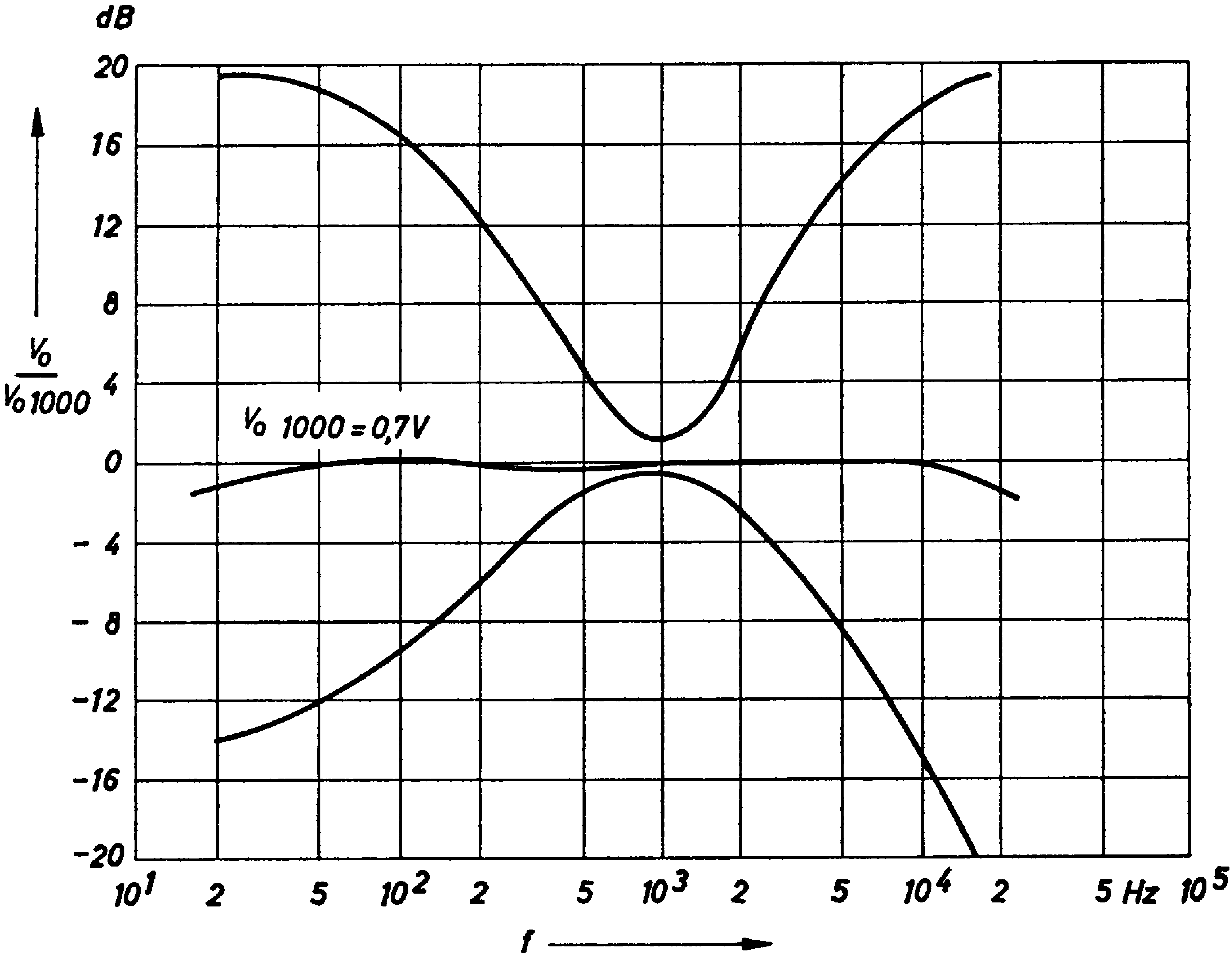


Fig. 1.2.

The following technical data apply to the whole layout, i. e., for the pre-amplifier here described plus the power amplifier in chapter 1.9. of our booklet "Design Examples of Semiconductor Circuits 1966". The same applies to the tone control curves.

Technical data of the complete amplifier

Operating voltage	30	V
Operating current	50 to 1000	mA
Output power	15	W
Load	4	Ω
Input voltage for the amplifier driven to full output ($f = 1$ kHz)	350	mV
Unweighted signal-to-noise ratio	>50	db
Max. input voltage without limiting ($f = 1$ kHz)	5.3	V
Input impedance ($f = 1$ kHz)	≈750	kΩ

1.2. Switchable Pre-Amplifier for Microphones and Magnetic Pick-Ups

The amplifier, the circuit shown in Fig. 1.3., can be connected in series to the amplifier described in the previous chapter. The frequency response of the amplifier is adjustable. A linear frequency characteristic can be turned on, if the amplifier is used as microphone amplifier and the required voltage frequency characteristic can be switched on, if it is used as pre-amplifier for a magnetic phono pick-up.

In the first case a frequency independent negative feedback, and in the second case a frequency dependent feedback section is switched into the feedback loop of the second stage.

In comparison to a frequency dependent attenuation this method of obtaining the desired frequency response has the advantage of a lower distortion factor. In order to prevent that the operating point of the transistor will be shifted by switching, the first transistor has to be decoupled for DC currents from the feedback section. This is achieved by a capacitor of 250 μF. The quiescent point is set as already described in the previous chapter.

The input impedance of the amplifier is substantially determined by the base-resistor of 47 kΩ, because the input impedance of the first transistor itself always exceeds this value.

The negative feedback capacitor of 68 pF between collector and base of the second transistor reduces the tendency towards oscillations which exists as a result of the high gain and the high cut-off frequency of transistor BC 109.

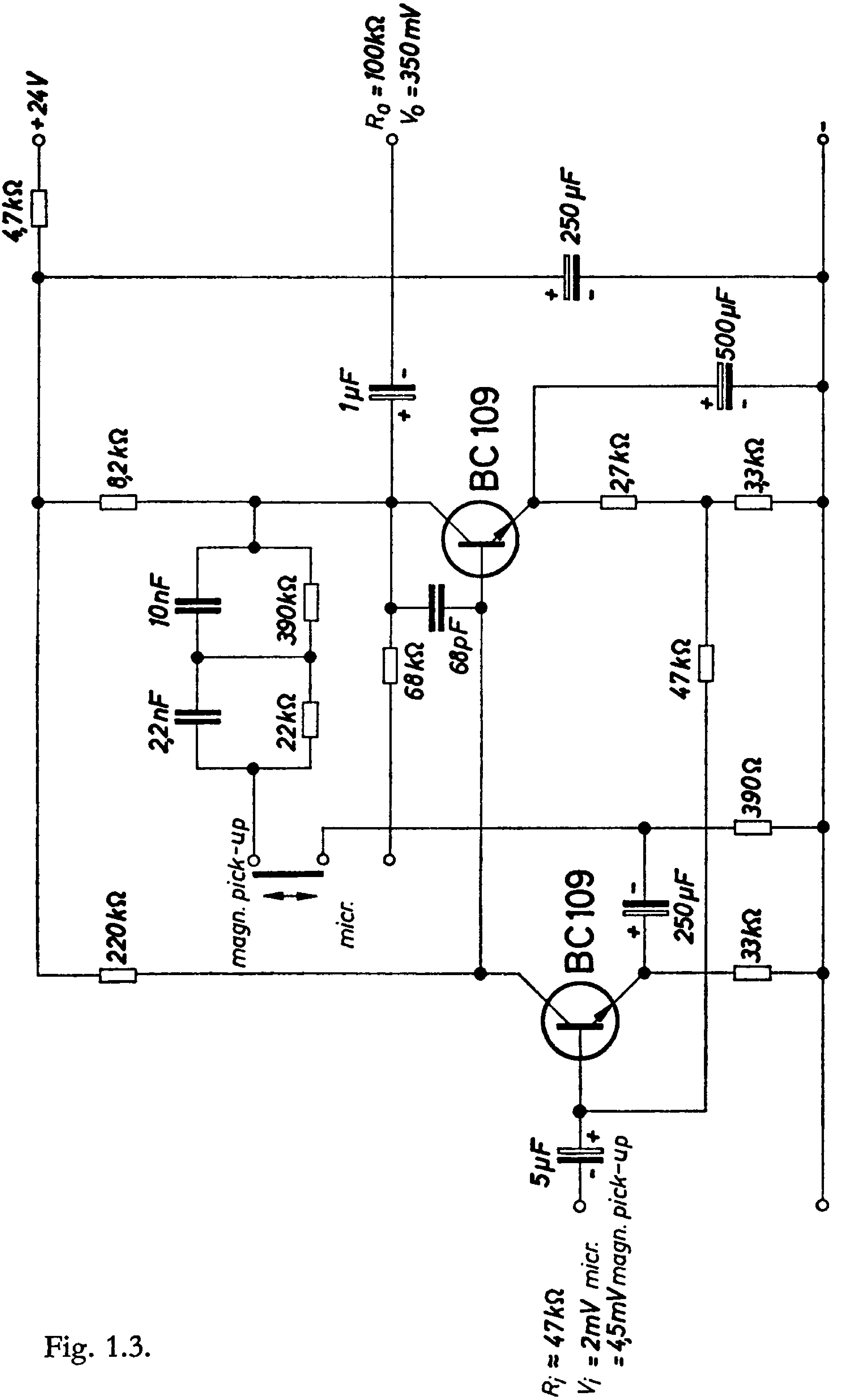


Fig. 1.3.

In order to obtain the extraordinarily good noise characteristic of the first stage, a small quiescent current of approximately 100 μA has been adjusted.

The frequency characteristic for both cases of operation is shown in Fig. 1.4.

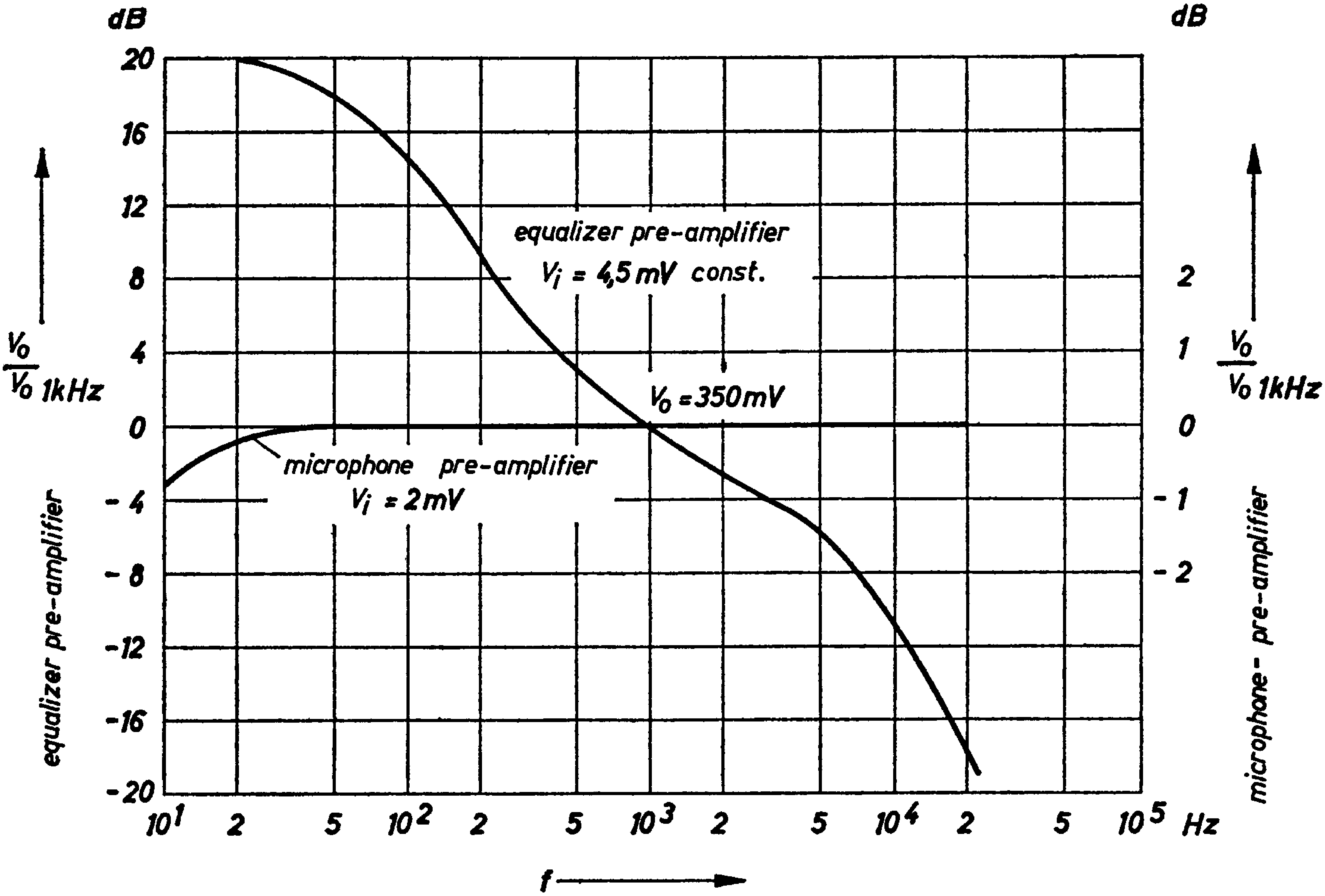


Fig. 1.4.

Technical data	Microphone pre-amplifier	Equalization pre-amplifier	
Operating voltage	24	24	V
Operating current	0.85	0.85	mA
Output voltage ($f = 1$ kHz, $R_L = 100$ k Ω)	350	350	mV
Input voltage ($f = 1$ kHz)	2	4.5	mV
Max. input voltage without limiting ($f = 1$ kHz)	20	43	mV
Input impedance	47	47	kΩ
Distortion factor (Output voltage 350 mV)			
$f = 100$ Hz	0.3	0.2	%
$f = 1$ kHz	0.3	0.1	%
$f = 20$ kHz	—	0.2	%
Unweighted signal-to-noise ratio	>50	>50	db

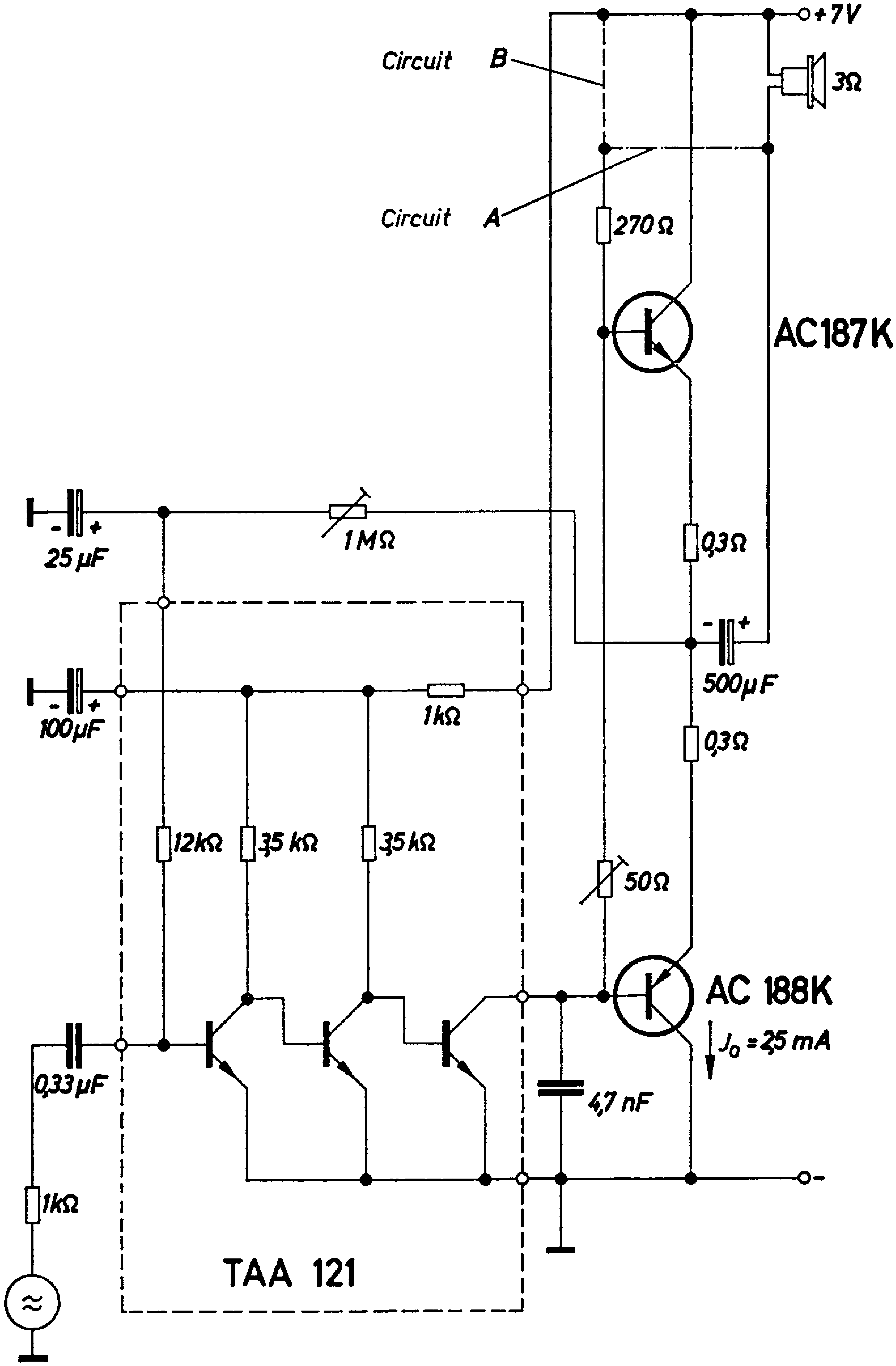

+7V
Circuit B
3Ω
Circuit A
270Ω
AC187K
25 µF
1MΩ
0,3Ω
500µF
100µF
1kΩ
0,3Ω
12kΩ
3,5 kΩ
3,5 kΩ
50Ω
AC 188K
J_0 = 2,5 mA
0,33µF
4,7 nF
1kΩ
TAA 121

Fig. 1.5.

1.3. AF Amplifier with an Integrated Circuit

The AF amplifier TAA 121 in integrated technology has a maximum permissible operating voltage of 7 V. In transformerless amplifiers the output stage can be driven only to a voltage of approximately the magnitude of the supply voltage of the driver stage, i. e., max. 7 V for the TAA 121. At the smallest loudspeaker resistance of 3 Ω the maximum output power is:

$$P_o = \frac{V_{cc}^2}{8\ R_o} = 2.04\ \text{W}$$

This theoretic value is not practically obtainable. In the circuit design shown in Fig. 1.5. a maximum output power of 1.4 W can be obtained. In this case, however, the supply voltage for the driver stage has to be tapped off at the load resistor; this causes a slight increase in power (circuit design A). If the driver is connected directly to the operating voltage, an output power of only 1.1 W is obtained. The total distortion factor for circuit B, however, is smaller. It amounts in the average to 2.5% in circuit B compared to 5% in circuit A.

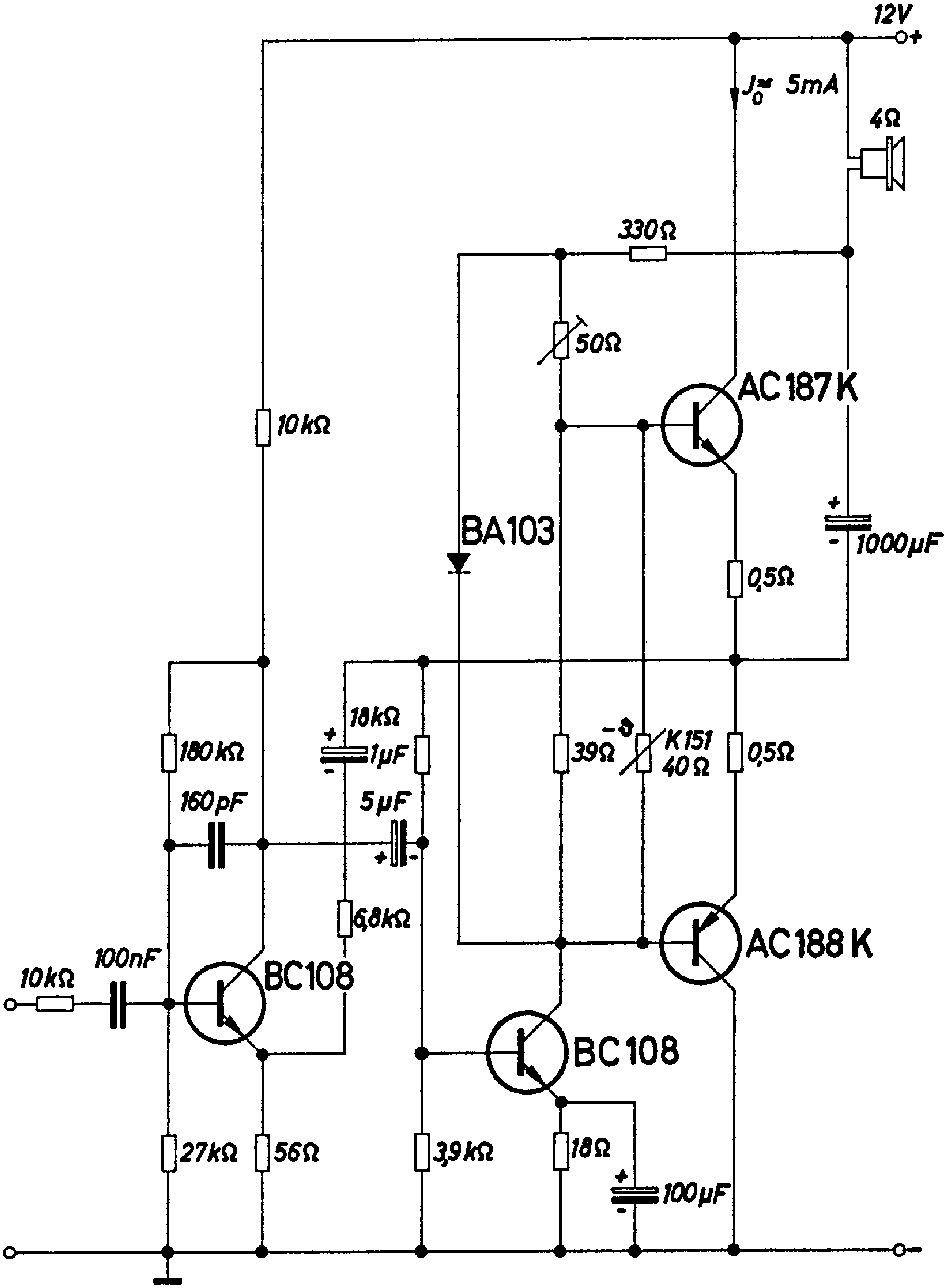
Technical data

Operating voltage	7	V
Load resistor	3	Ω
Quiescent current of the output stage	2.5	mA
Quiescent current of the driver stage	13	mA
Total quiescent current	20	mA
Maximum output power		
Circuit A	1.4	W
Circuit B	1.1	W

1.4. Transformerless AF Amplifier 12 V/3,6 W

With the complementary pair AC 187 K/AC 188 K an output power of 3.6 W can be achieved at an operating voltage of 12 V. The corresponding circuit is shown in Fig. 1.6.

In the pre-stage the feedback capacitor between collector and base, already mentioned above, has been inserted.



12V
+
$J_0 \approx 5mA$
4Ω
330Ω
50Ω
AC 187 K
10kΩ
BA103
+
-
1000µF
0,5Ω
18kΩ
+
1µF
-
180kΩ
39Ω
-ϑ
K 151
40Ω
0,5Ω
160pF
5µF
+
-
6,8kΩ
AC188 K
10kΩ
100nF
BC108
BC108
27kΩ
56Ω
3,9kΩ
18Ω
+
-
100µF
-

Fig. 1.6.

The quiescent current of the power stage is stabilized for temperature variations by a negative temperature coefficient thermistor and for supply voltage variations by a diode biased in forward direction. This compensation of the supply voltage variation is required particularly, if a voltage supply with a high internal resistance is used for the operation of the amplifier.

A negative feedback from the center tap of the power stage to the input stage of the amplifier is achieved by a RC section.

Technical data

Operating voltage	12	V
Operating current	23 to 440	mA
Output power	3.6	W
Load	4	Ω
Input voltage for maximum output power	150	mV
Input impedance	35	kΩ
Frequency range (3 db)	50 Hz to 20	kHz
Distortion factor ($f = 1$ kHz, $P_o = 1$ W)	1	%

1.5. Transformerless AF Amplifier 20 V/10 W

The complementary pair AD 161/AD 162 allows the design of amplifiers with an output power of up to 10 W.

Fig. 1.7. and 1.8. show two different types of circuit designs; the first one uses a germanium transistor as driver, whilst in the second circuit design silicon transistors are used both in the pre-stage and in the driver stage. In the circuit shown in Fig. 1.7. there is a DC and AC feedback from the center tap of the output stage to the pre-stage. By that means the voltage at the center tap of the output stage is kept constant. A RC section between the emitter of the pre-stage and ground reduces the effect of the negative AC feedback. In both circuit designs the quiescent point of the output stage is stabilized for voltage and temperature variations by a silicon diode and a NTC thermistor, respectively. The magnitude of the collector resistor in the driver stage depends on the current gain of the output stage transistor. For the lower gain groups of the transistors AD 161 and AD 162 the resistor should have a value of 82 Ω and for the higher gain groups a value of 180 Ω.

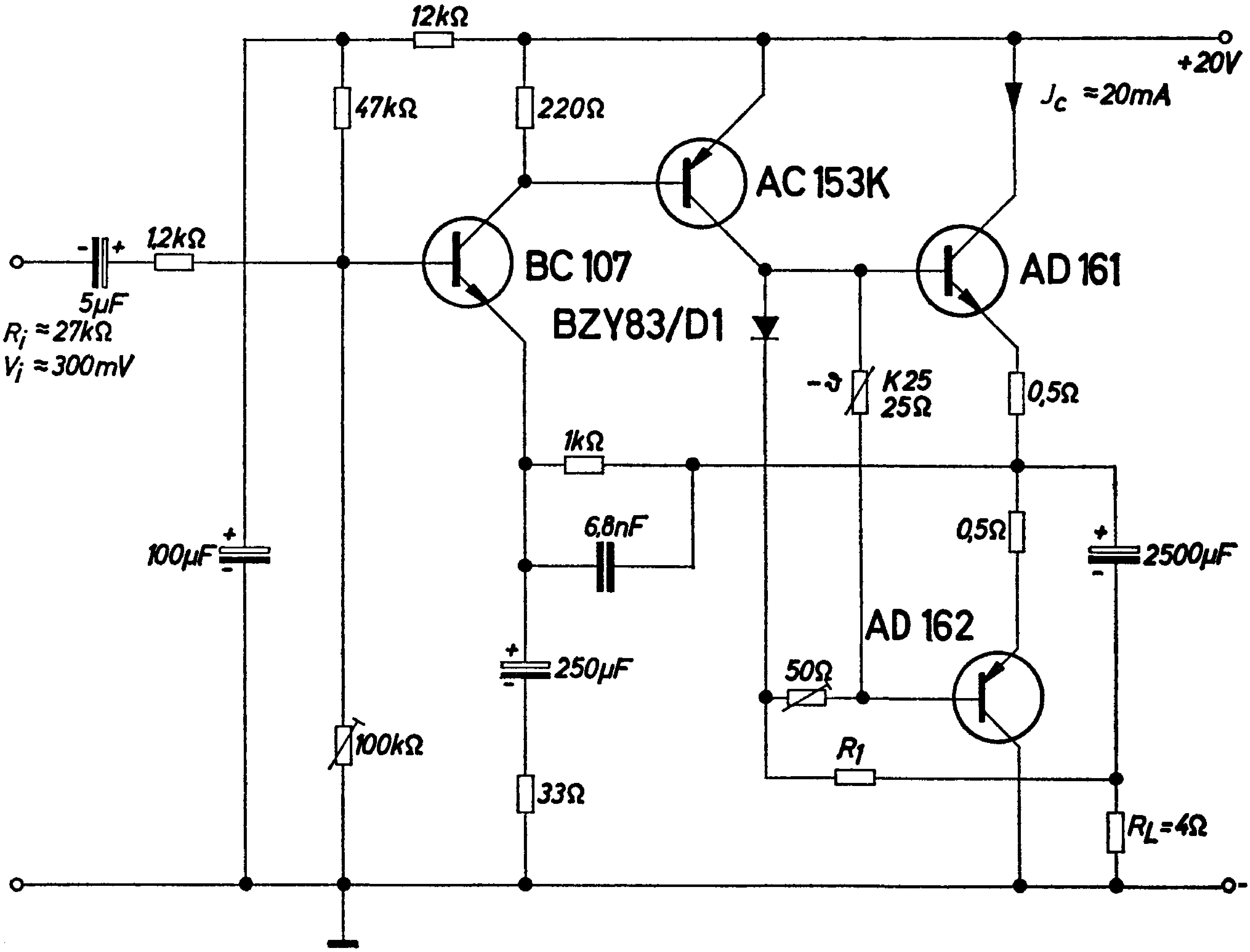
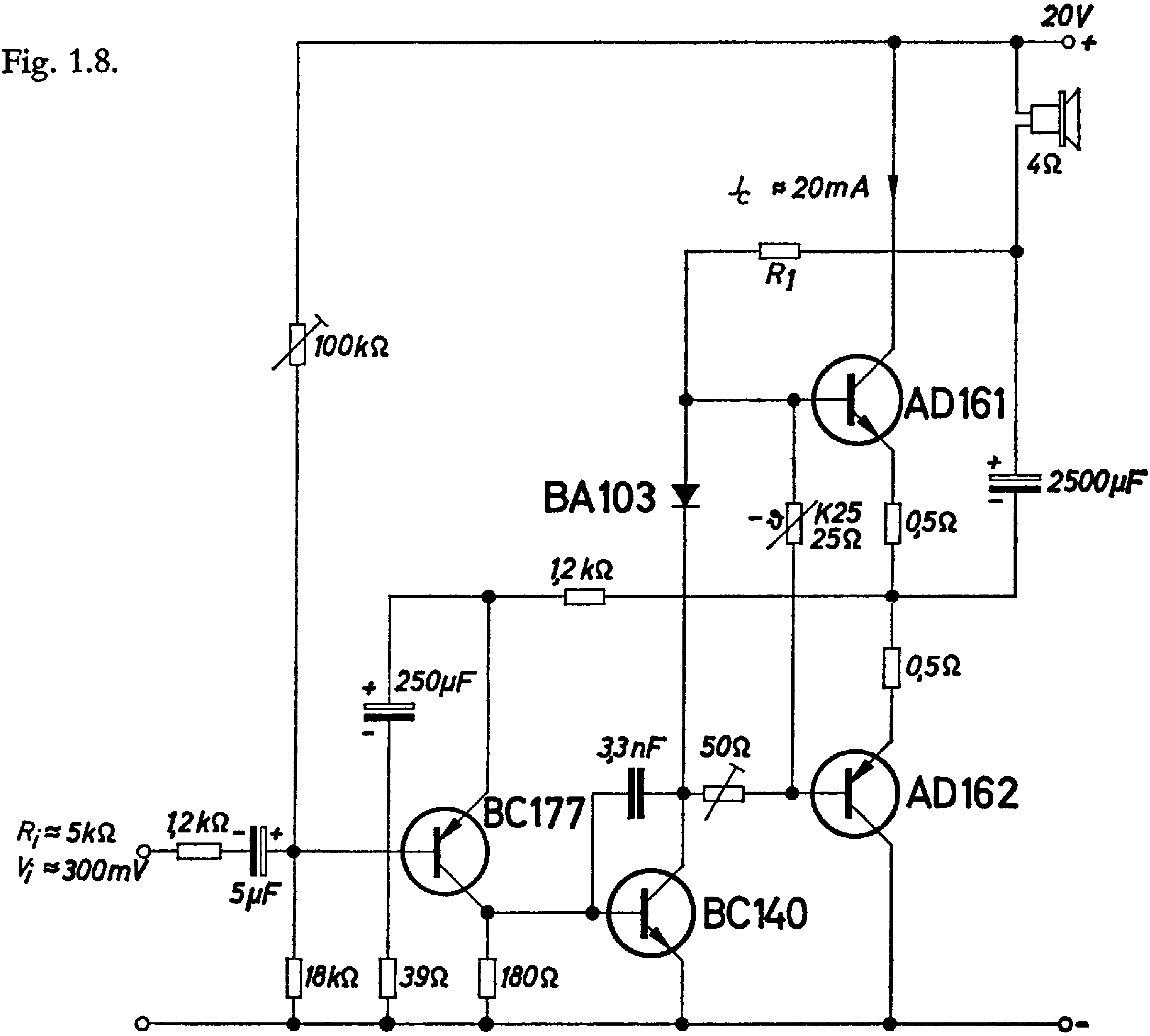


Fig. 1.7.

Technical data	Fig. 1.7.	Fig. 1.8.	
Operating voltage	20	20	V
Max. operating current	850	850	mA
Output power	10	10	W
Load	4	4	Ω
Frequency range (3 db)	23 Hz to 23 kHz	23 to 23	kHz
Input impedance	27	12	kΩ
Input voltage for amplifier driven to full output ($f = 1$ kHz)	300	300	mV
Distortion factor ($f = 100$ Hz to 10 kHz, $P_o = 5$ W)	<1	<1	%

Fig. 1.8.



1.6. Transformerless AF Amplifier 24 V/10 W

In the output stage of the circuit shown in Fig. 1.9. power transistors are used. These transistors have a higher cut-off frequency than the normal alloyed power transistors but still have the advantage of low saturation resistance and good current gain linearity. For that reason power amplifiers with high tone quality can be designed with these transistors. Since no complementary pairs of these transistors are available, the required phase reversal for the transformerless drive of a push-pull power stage has to be achieved in the driver stage. In this example a complementary pair AC 127/AC 152 is used for that purpose. The stabilization of the quiescent current in the output stage is accomplished for temperature changes by a NTC thermistor

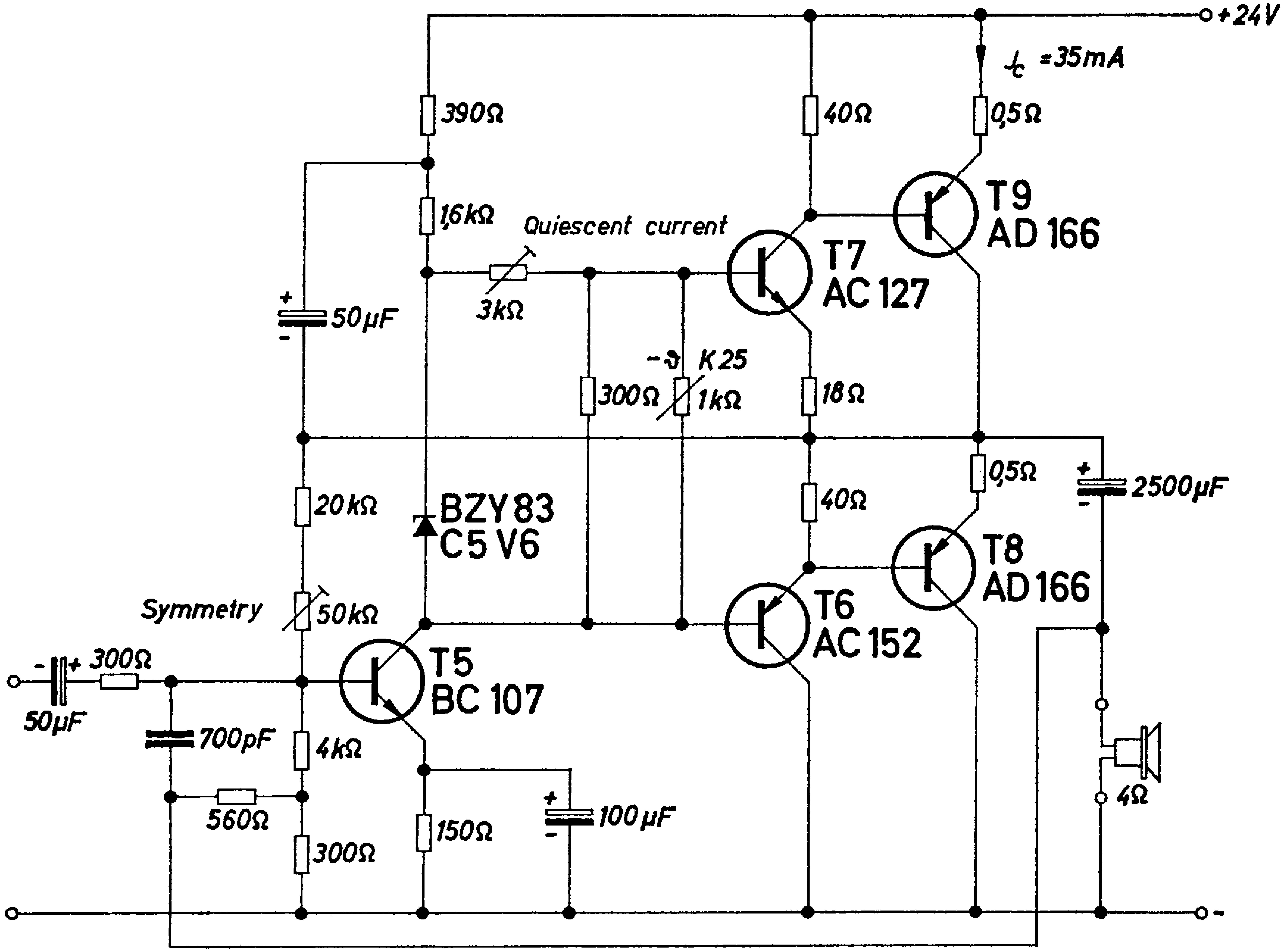


Fig. 1.9.

K 25 and for variations of the supply voltage by a Zener diode BZY 83. The NTC thermistor is to be mounted on the same heat sink as the power stage transistors. For the design of the amplifier a goal has been a small distortion factor in addition to the good frequency characteristic. This has been made possible by the transistors used. The distortion factor is below 0.4% in the entire frequency range. For this power stage the pre-amplifier described in chapter 1.1. can be used, as well as the switchable pre-amplifier for microphones and magnetic pick-ups discussed in chapter 1.2.

Technical data

Operating voltage	24	V
Operating current	50 to 100	mA
Output power	10	W
Load	4	Ω
Input voltage for the amplifier driven to full output	500	mV
Input impedance	300	Ω
Signal-to-hum ratio $P_o = 100$ mW, mains hum voltage $= 0{,}27$ V, $R_s = 200\ \Omega$, $R_L = 4\ \Omega$	60	db
Unweighted signal-to-noise ratio ($P_o = 100$ mW)	85	db

1.7. Transformerless AF Amplifier 50 V/45 W

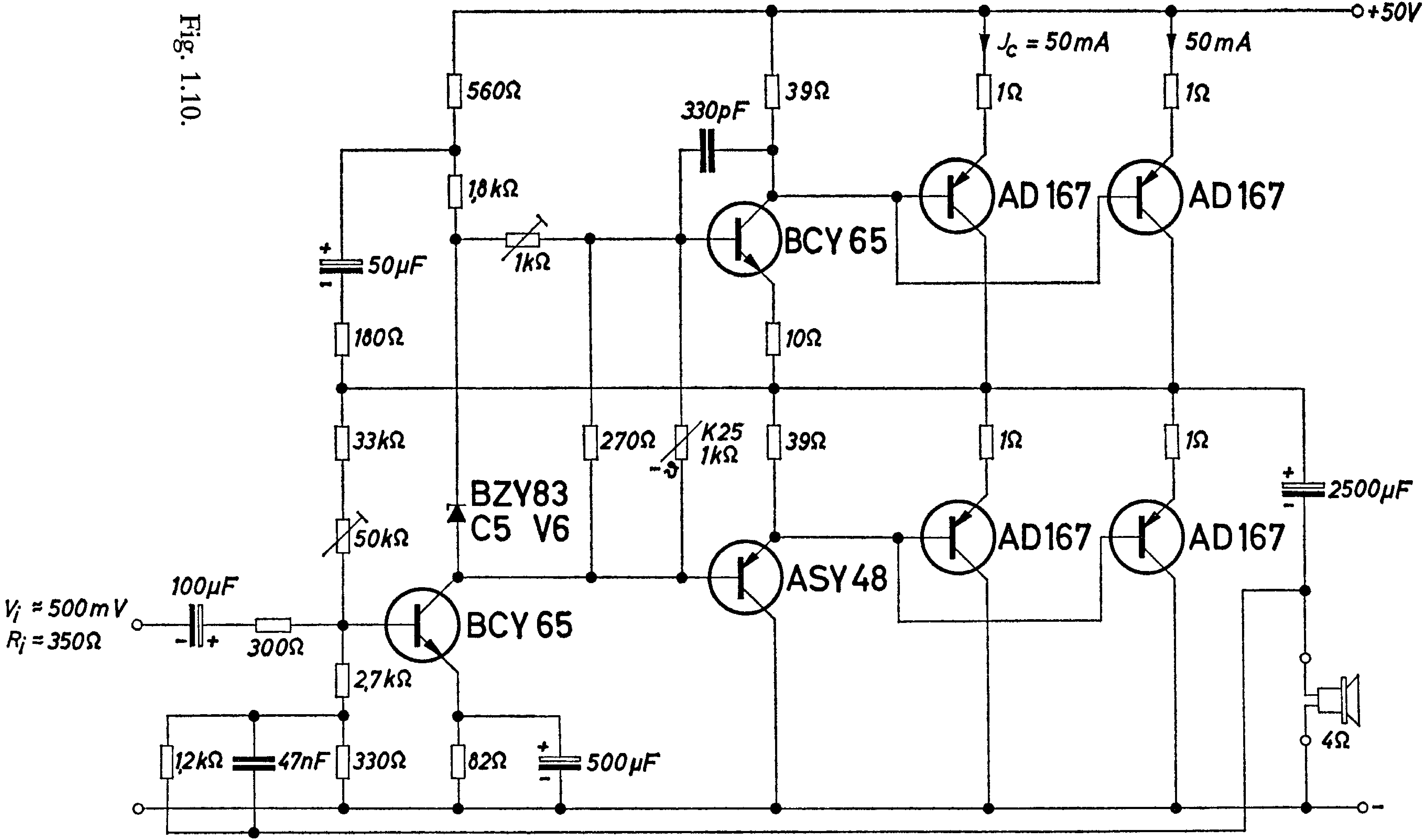
In push-pull operation an output power of 45 W at an operating voltage of 50 V can be obtained by two power drift-transistors AD 167 each connected in parallel. The reproduction achieved in the design shown in Fig. 1.10. fulfills the requirements of the HiFi sets.

A npn silicon transistor BCY 65 and a pnp germanium transistor ASY 48 are used for the complementary driver stage.

The emitter resistors of 1 Ω guarantee an equal current load of the four power stage transistors. All three stages of the amplifier are galvanically coupled. The quiescent currents of all stages are set by the potentiometer of 3 kΩ. Besides the above the design is similar to that described in chapter 1.6.

A cooling surface with a thermal resistance of R_{th} <2.5 °C/W is provided for each of the 4 power stage transistors. Additional cooling is also required for the transistors of the driver stages. The pre-amplifiers shown in chapter 1.1. und 1.2. can be used also for this amplifier.

The distortion factor in dependence on the output power at different frequencies and the permissible drive beyond saturation are shown in the diagrams in Fig. 1.11. and 1.12.



+50V
J_c = 50mA
50mA
560Ω
39Ω
1Ω
1Ω
330pF
1,8kΩ
AD 167
AD 167
1kΩ
BCY 65
50μF
180Ω
10Ω
33kΩ
270Ω
K25
1kΩ
39Ω
1Ω
1Ω
2500μF
BZY83
C5 V6
50kΩ
AD167
AD 167
ASY 48
100μF
V_i ≈ 500mV
R_i = 350Ω
300Ω
BCY 65
2,7kΩ
1,2kΩ
47nF
330Ω
82Ω
500μF
4Ω

Fig. 1.10.

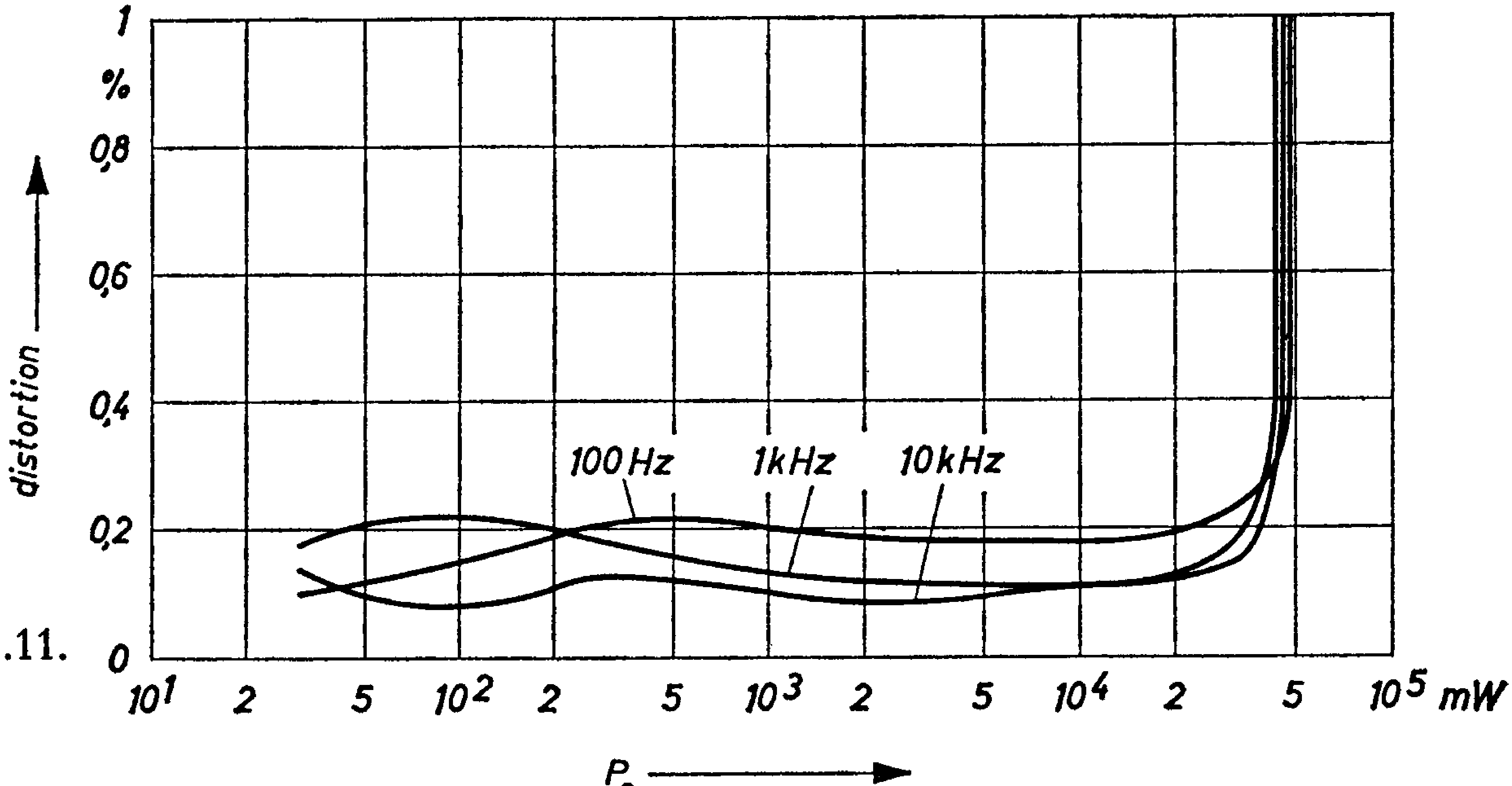


Fig. 1.11.

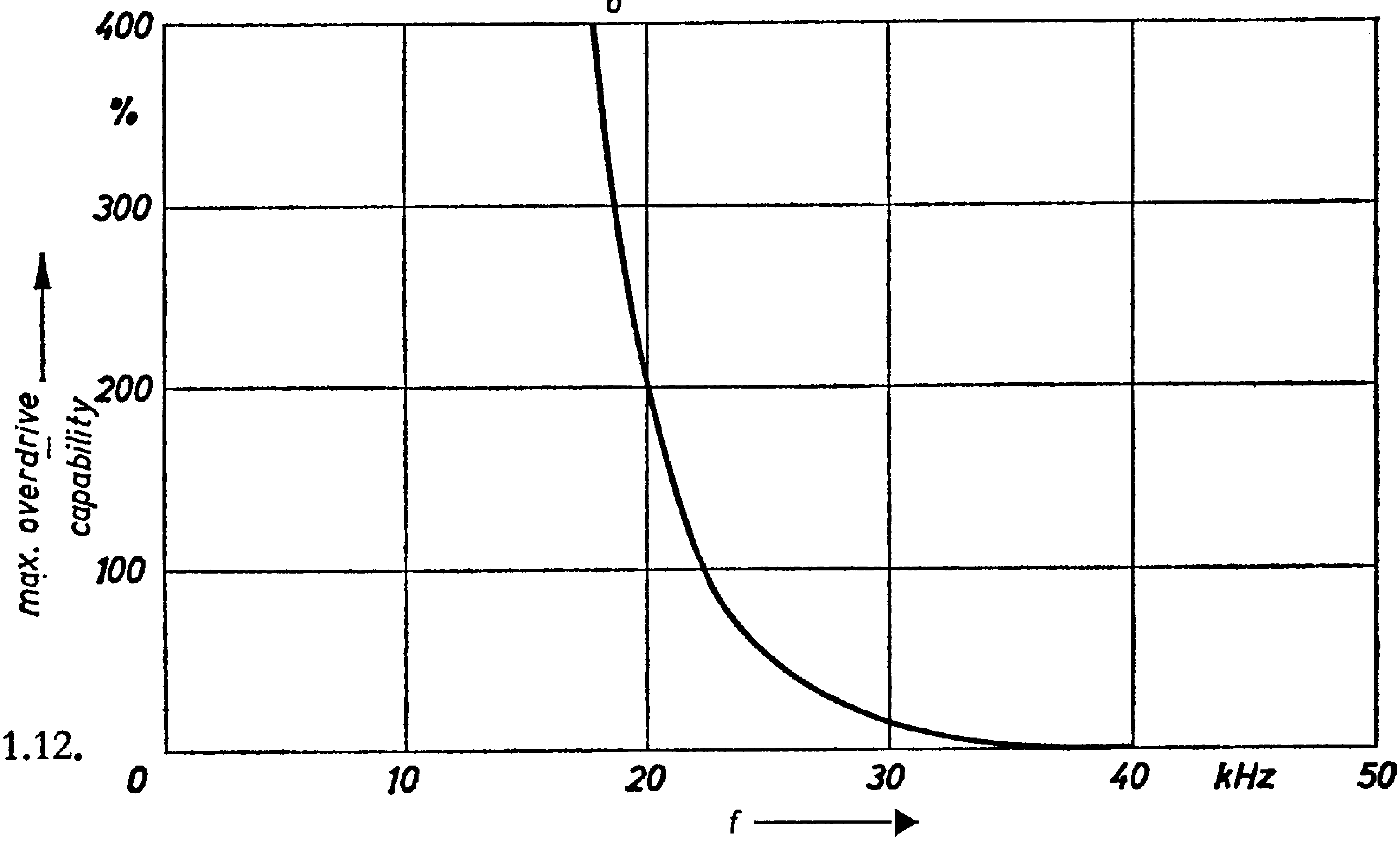


Fig. 1.12.

Technical data

Operating voltage	50	V
Operating current	0.13 to 1.65	A
Output power ($K < 1\%$)	45	W
Load	4	Ω
Distortion factor	<0.4	%
Frequency range	10 Hz to 17	kHz

Input voltage for max. output power ($f = 1$ kHz)	500 mV
Input impedance	350 Ω
Unweighted signal-to-noise ratio referred to an output power of 50 mW	80 db

2. Choppers and Oscillators

As opposed to low-frequency amplifiers for which new transistors make further designs possible, the design of transistor choppers and voltage converters has not changed within the last years. In former booklets many designs for push-pull choppers, single-base forward pass oscillator and single-base blocking oscillators have been described. In this booklet, therefore, we only show special possibilities of application for transistor choppers and transistor oscillators. A saw-tooth generator with a four-layer tetrode will be described among others.

2.1. Single-Base Blocking Oscillator for Solar Battery Operation

Silicon photo-voltaic cells are used for direct conversion of luminous energy into electric energy, i. e., they are in a position to deliver electric power depending on the illumination. The illumination falling on the cell yields a certain open-circuit voltage and a certain short-circuit current depending on the intensity.

If a battery would be charged directly by a silicon photo-voltaic cell — in this case also called solar cell — at economic power matching a charging could only occur at an extremely high illumination. The illumination would have to be so intense that the voltage supplied by one or several solar cells under load exceeds the voltage of the battery being charged. Especially for highly variable light intensity (e. g. daylight) one achieves in such a way no optimal utilization of the solar cells.

A much better ratio can be accomplished by connecting a single-base blocking oscillator to the solar battery. There is no stable ratio between the input and the output voltage of this single-base blocking oscillator, but the voltage transformer depends on the case of operation. This results from the fact that in the single-base blocking oscillator energy is stored in the oscillator transformer during the time of current flow in the transistor. This energy discharges to the load during the switch-off time of the transistor. The energy discharges to a high resistive load with high voltage and low current and to a low resistive load with low voltage and high current. The single-base blocking oscillator shown in Fig. 2.1. delivers at an operating voltage of 0.5 V to 2 V such a high output voltage that a 6 V

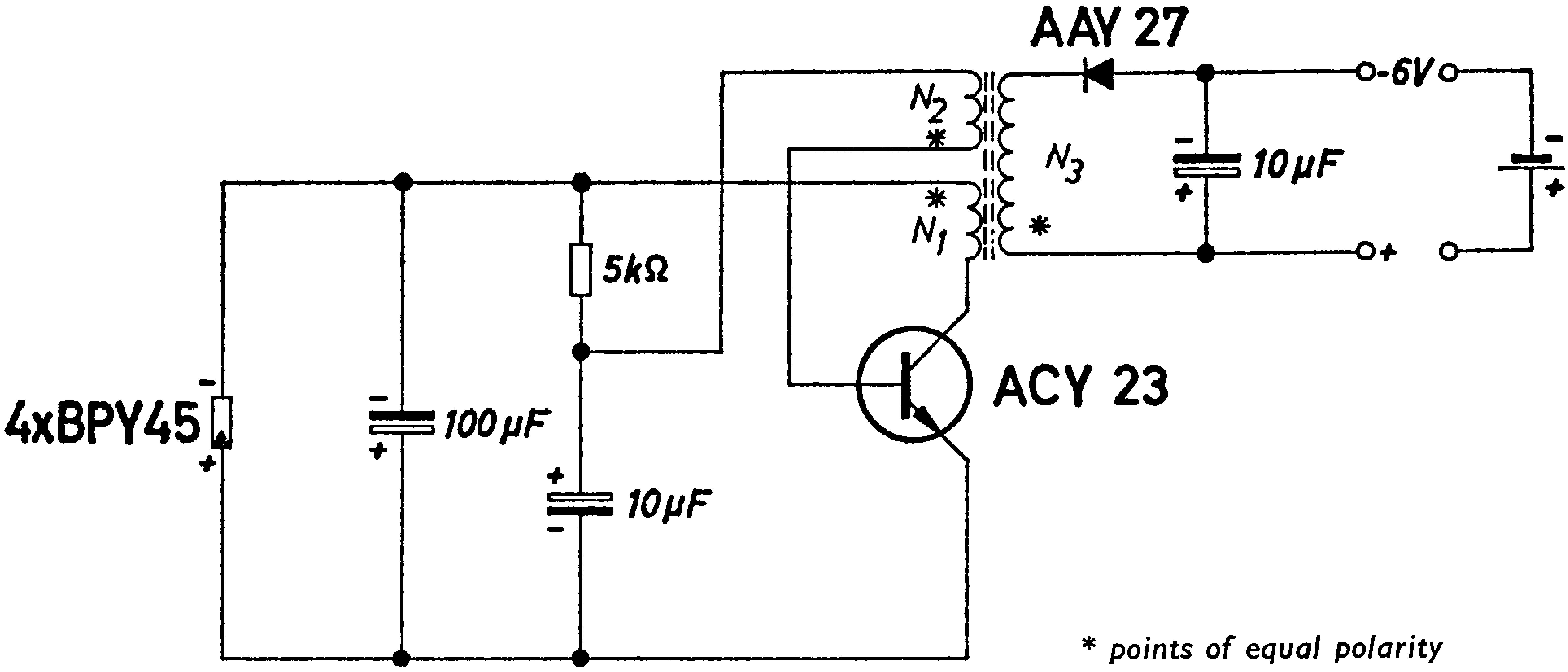


Fig. 2.1.

battery, for example, is constantly charged. In consequence a battery is constantly charged by solar batteries via such a chopper even with extremely variable illumination. Four silicon solar-elements connected in series have been used as solar battery. The chopper supplies an output power of approximately 3 mW if the photo-voltaic cells are illuminated by daylight.

Technical data

Battery voltage	0.5 to 2	V
Current consumption	2.5	mA
Oscillation frequency	ca. 10	kHz
Efficiency	ca. 60	%

Transformer: Siferrite pot core B65561-A0250-A022

$N_1 =$ 270 turns CuL 0.1
$N_2 =$ 60 turns CuL 0.08
$N_3 =$ 1000 turns CuL 0.05

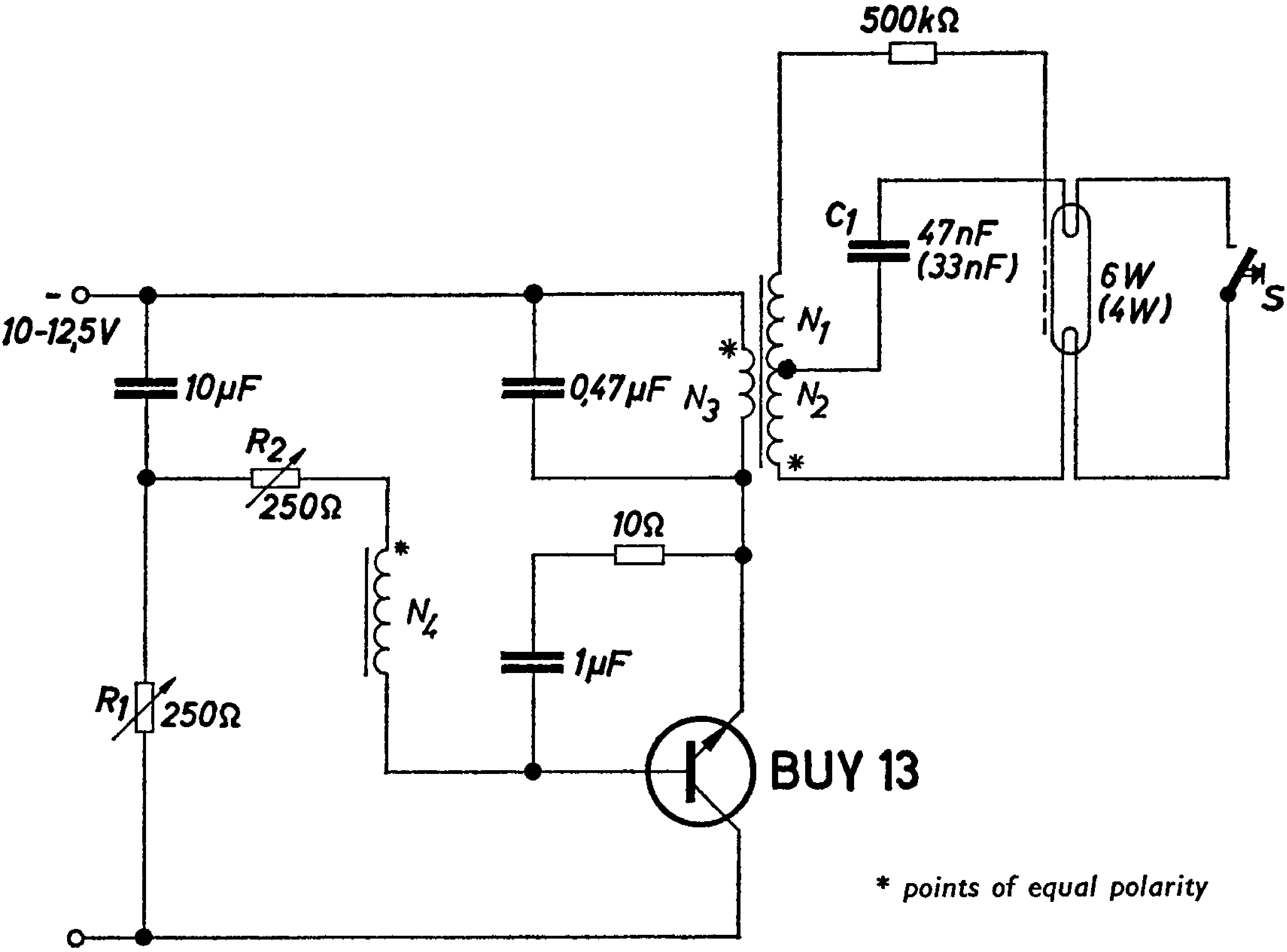


Fig. 2.2.

2.2. Sine-Wave Generator for Fluorescent Lamps

Choppers with transistors usually generate a square-wave output voltage. If higher oscillation frequencies are required, for example, beyond the audible range, the higher saturation values, which are difficult to determine, cause considerable core-losses. The total losses are increased by the transistor-switching losses. With regard to the efficiency it may be better to use a sine-wave generator instead of a chopper. The sine-wave generator shown in Fig. 2.2. has an oscillation frequency of approximately 17 kHz, i. e., it delivers no audible interfering oscillations. A 6 W or a 4 W fluorescent lamp can be connected to the output of the sine-wave generator. If a 4 W fluorescent lamp is connected, the coupling capacitor limiting the heater current and the operating current has to be reduced to a value of 33 nF.

The ignition voltage can be applied to an outer electrode, for example, to a filament at the glass bulb. The ignition coil N_1 has to be wound carefully since the voltage per turn has a value of almost 1 V.

For adjusting the circuit the switch S is closed and at the lowest operating voltage the potentiometer R_1 is turned back to such a degree that the generator safely starts oscillating. Then the resistance of the potentiometer R_2 is decreased to such a value that the tube ignites after a short closure of switch S.

Technical data

Operating voltage	10 to 12.5	V
Operating current	ca. 12	A
Efficiency	35 to 40	%

Transformer: Siferrite E-core B66231-A0200-K026

N_1 = 740 turns 0.08 CuL
N_2 = 80 turns 0.15 CuL
N_3 = 16 turns 0.4 CuL
N_4 = 23 turns 0.1 CuL

2.3. Saw-Tooth Generator with a Thyristor–Tetrode

In the saw-tooth generator illustrated in Fig. 2.3. the charging and discharging of a capacitor is controlled by a thyristor tetrode. As this semiconductor component is relatively new, we shall describe in this booklet the characteristics of the thyristor tetrode BRY 20 before any details concerning the circuit will be given. This component consists of four semiconducting layers with a doping sequence PNPN. The exterior p zone is called the anode, the exterior n zone is named the cathode. Two so-called gates are located between anode and cathode. The gate next to the anode is called G_A, the gate next to the cathode is named G_K. In the thyristor tetrode BRY 20 it is mainly the gate G_K which is used for controlling the thyristor. It is of importance that at this gate this component may not only be switched on but off as well. This is the essential difference between the thyristor-tetrode BRY 20 and the so far known four-layer diodes and also the tube thyratron which do have many characteristics in common with semiconductor thyristors.

The thyristor is switched on at the gate G_K by a current pulse of positive polarity. It remains switched on until the supply voltage is switched off or decreased to such a value that a certain minimum current, the so-called holding current, is undercut, or until a negative pulse is applied to the gate G_K. A higher energy is required at the

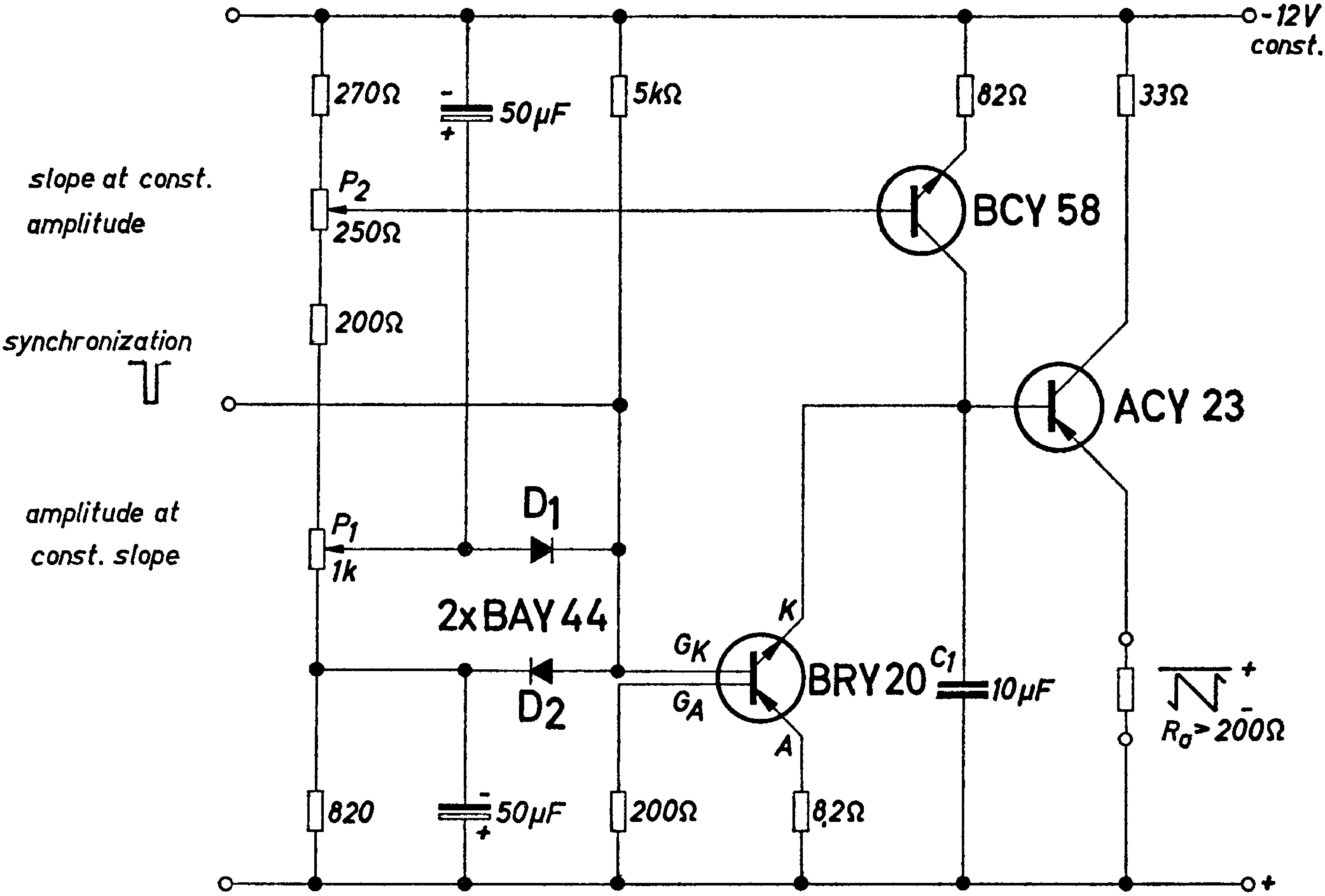


Fig. 2.3.

gate for switching off of the thyristor tetrode BRY 20 than for switching on because the so-called switch-off current gain is extremely lower than the so-called switch-on current gain. In order to prevent interference effects the gate G_A is to be connected to the positive potential. In the circuit shown in Fig. 2.3. the capacitor C_1 is charged via the transistor BCY 58. The charging current and the steepness of the slope of the generated saw-tooth pulse are set by the potentiometer P_2. The capacitor C_1 is charged until the voltage at the capacitor becomes as high as the voltage which has been set by the potentiometer P_1 at the gate G_K of the thyristor tetrode BRY 20. As soon as this occurs, the thyristor tetrode is switched on and the capacitor discharges via the protective resistor of 8.2 Ω. The potential of the thyristor cathode changes towards positive values. As soon as the negative voltage applied to the gate G_K via the diode D_2 is negative compared to the cathode potential, the negative pulse required for switching off is applied to the gate G_K, and the thyristor is switched off. The discharge of the capacitor is now finished, and the next charging process starts again.

As the voltage at which the charging of the capacitor is finished is set by the potentiometer P_1 the amplitude of the saw-tooth wave is controlled by this potentiometer P_1. For a constant steepness of the saw-tooth slope this is similar to the frequency setting of the saw-tooth voltage.

As already mentioned, the slope steepness of the saw-tooth is set by the potentiometer P_2, i. e., by the charging current of the capacitor.

Switching pulses are delivered to the gate by the charging capacitors of 50 μF, i. e., by a sufficiently low-ohmic source. Hereby very stable switching points are achieved.

In this circuit the amplitude of the output signals can be adjusted by the potentiometer P_1 between 4 and approximately 8 Vpp, whereas the steepness of the slope can be set by the potentiometer P_2 to values of 0.4 to 0.8 V/ms. For decoupling an amplifier stage is connected to the amplifier output.

A synchronization of the saw-tooth generator is possible at the gate G_K with positive pulses.

Technical data

Operating voltage	12	V const.
Operating current	ca. 55	mA
Load	>200	Ω
Output signal	4 to 8	Vpp
Slope of the signal	0.4 to 0.8	V/ms

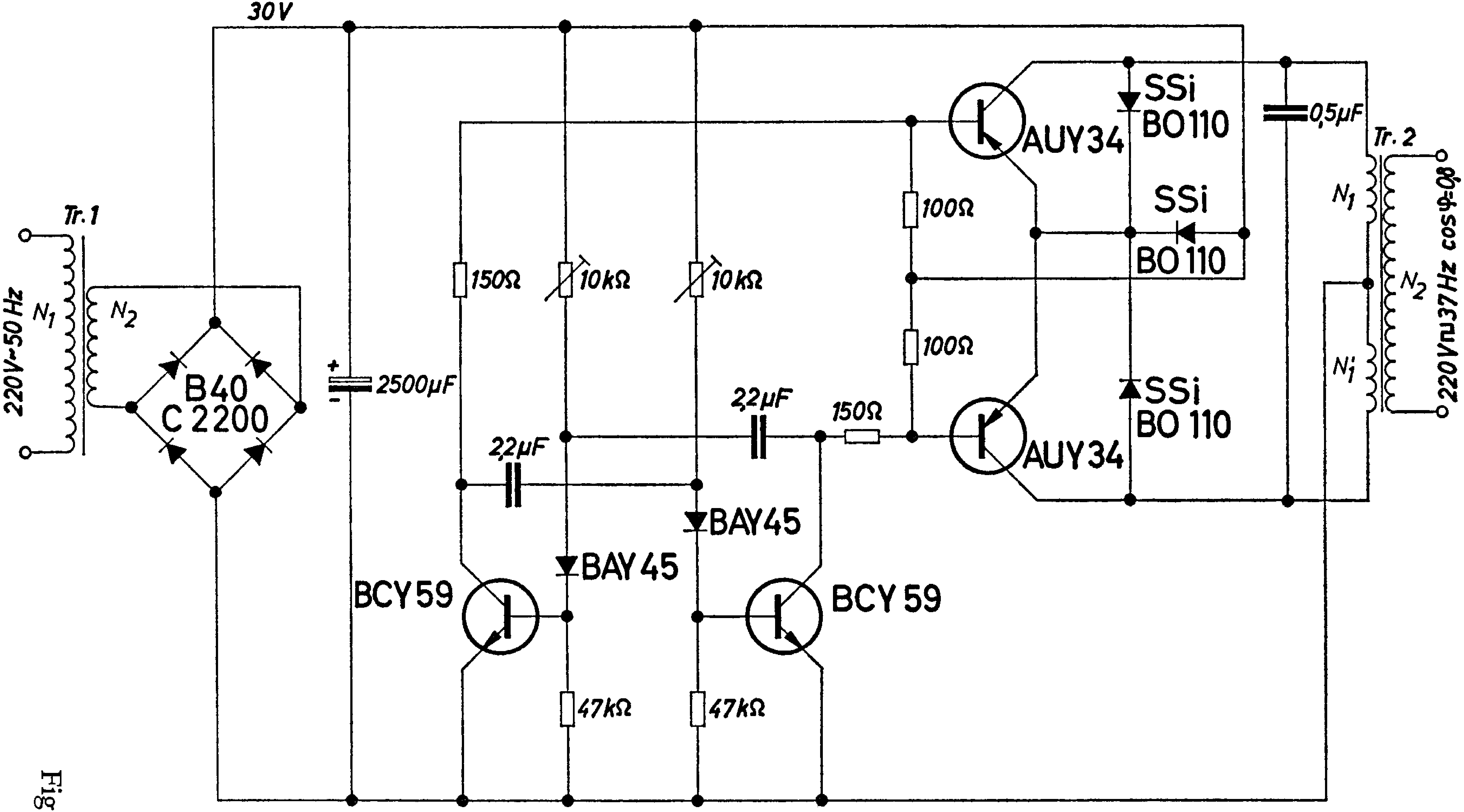

30V
Tr.1
220V~50Hz
N_1
N_2
B40
C2200
2500μF
150Ω
10kΩ
10kΩ
2,2μF
2,2μF
BAY45
BAY45
BCY59
BCY59
47kΩ
47kΩ
150Ω
100Ω
100Ω
AUY34
AUY34
SSi
BO110
SSi
BO110
SSi
BO110
0,5μF
Tr.2
N_1
N_2
N_1'
220V~37Hz cos φ=0,8

Fig. 2.4.

2.4. Frequency Converter 50 Hz/37 Hz

It is not possible to convert an AC voltage of 50 Hz to one of, say, 37 Hz by frequency division. In the example, shown in Fig. 2.4., therefore, at first the AC voltage is transformed and rectified. An astable multivibrator oscillating at the required frequency drives a power stage with the transistor AUY 34. The steepness of the slope of the multivibrator voltage is sufficiently high so that the power transistors can be directly connected without the switching losses being too high.

In order to protect the transistor from too high dissipation power at inverse operation which may occur in cases of open-circuit or inductive load, protective diodes are connected in parallel to the collector-emitter path. A third diode located in the emitter circuit of both transistors generates a low bias voltage at the emitter. Hereby an excellent switch-off of the power transistors is achieved during the pulse pause.

Technical data

Operating voltage	220 V	50 Hz
Output power	30 W	
Output voltage	220 V	37 Hz
Permissible complex load	$\cos \varphi = 0.8$	
Efficiency	ca. 70	%
Max. ambient temperature	60	°C
Total cooling surface for the power transistors	ca. 100	cm^2

Transformers

Tr 1: EI 78/26, Dyn. sheet IV/0.35 alternate stacking
N_1 = 1100 turns 0.32 CuL
N_2 = 140 turns 0.8 CuL
Tr 2: EI 84/42 Dyn. sheet IV/0.35 alternate stacking
$N_1 = N_1'$ = 192 turns 0.7 CuL joint lead wound coil
N_2 = 1600 turns 0.24 CuL

2.5. Frequency and Amplitude Stabilized Sine-Wave Generator

The RC generator shown in Fig. 2.5. operates in the principle of the Wien bridge and is appropriate for a frequency range of 5 Hz to 500 kHz. At the variation of the operating voltage of $\pm$ 10%

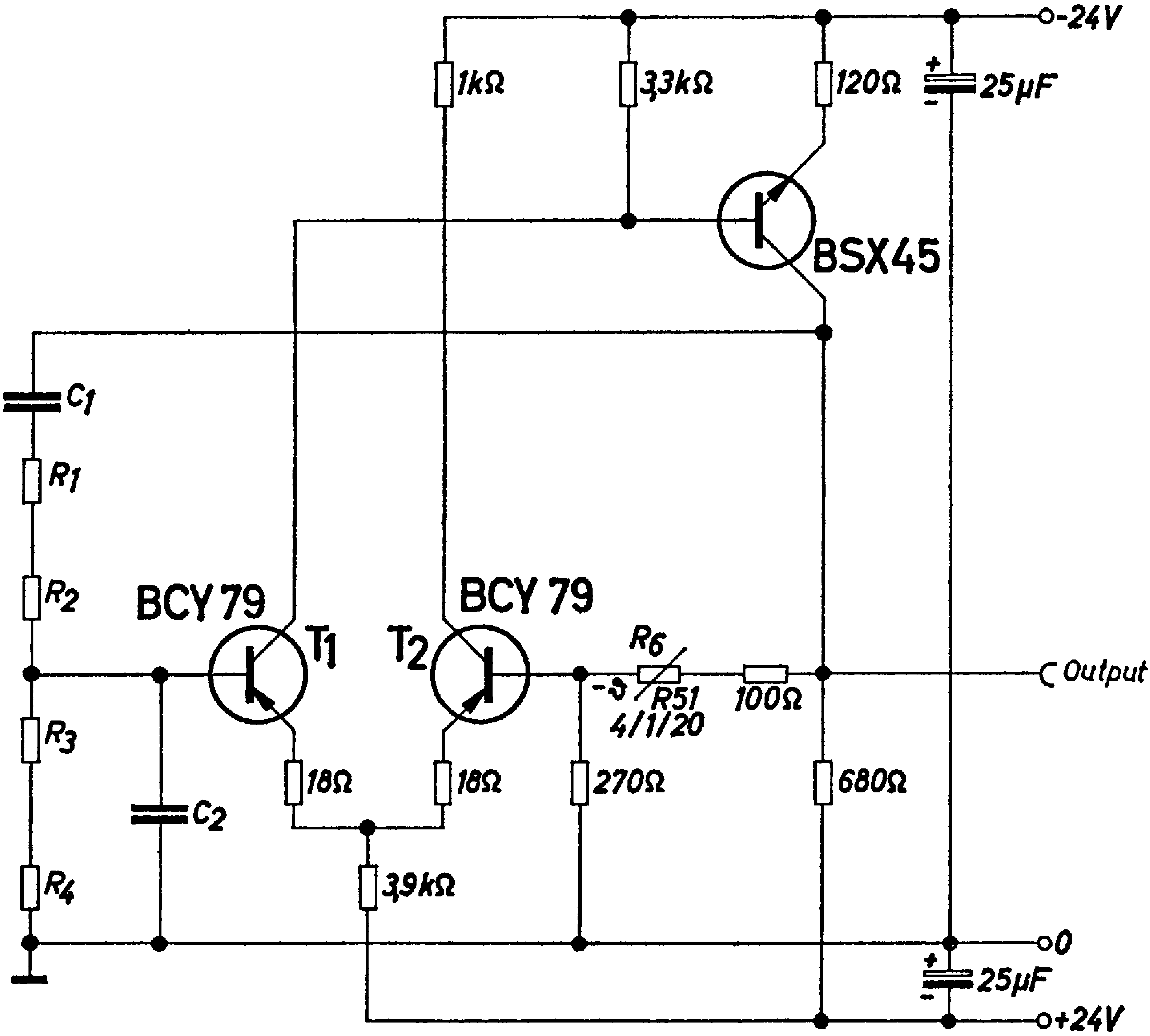


Fig. 2.5.

the frequency deviation is smaller than 10^{-4}. The constancy of the frequency vs temperature depends only on the resistors and capacitors used in the Wien bridge. A frequency deviation of $<10^{-4}/°C$ can be obtained by a suitable combination of components with different temperature coefficients. The constancy of frequency in this RC generator, therefore, is much better than in LC generators in this frequency range. The Wien bridge consists of the capacitors C_1 and

C_2 and the resistors R_1 to R_4. The magnitude of these capacitors and resistors determines the oscillation frequency

$$f_0 = \frac{1}{2\pi\, RC}$$

whereby $R = R_1 + R_2 = R_3 + R_4$ and $C = C_1 = C_2$.

For the right choice of values the sum of the resistors R has to be between 1.2 and 10 kΩ and the capacitors C_1 and C_2 must have a capacity of at least 150 pF. The output voltage of the bridge is stabilized by a negative temperature coefficient thermistor R 51. The stabilized output voltage is fed back via the Wien bridge to the input of the differential amplifier. The amplitude of the output voltage amounts to approximately 6 V. A DC voltage of maximum ± 200 mV, however, is superimposed to this AC voltage.

Technical data

Operating voltage	±24	V
AC-Output voltage	6	V
Maximum output current	10	mA
Distortion factor in the range from 50 Hz to 300 KHz		
for an open circuit	< 0.5	%
for a load of 600 Ω	< 1	%
Frequency change for a variation in operating voltage of ± 10%	$< 10^{-4}$	
Change in amplitude with temperature	ca. $-5 \cdot 10^{-3}$	1/°C

3. Multivibrators and Delay Circuits

The three basic circuits, the astable, the bistable, and the monostable multivibrator have already been described in all details in former editions of our semiconductor circuit examples. In the following chapter some interesting cases of application for these circuits will be shown. They are partly circuits with germanium transistors, which were already discussed in former booklets, and which we now show with silicon transistors. With monostable multivibrators, for example, a better constancy of delay time can be achieved because the smaller cut-off currents of silicon transistors do not load the time-determining charging capacitors to such a high extent.

Since there are now pnp silicon transistors available the multivibrators can also be built with complementary silicon transistors. An applicable example is shown in chapter 3.2.

For the design of multivibrators our extinguishable thyristor tetrode is an interesting component. It replaces in its function as a single component an entire bistable multivibrator, as it can be switched on and switched off by gate pulses. A representative example is given in chapter 3.7.

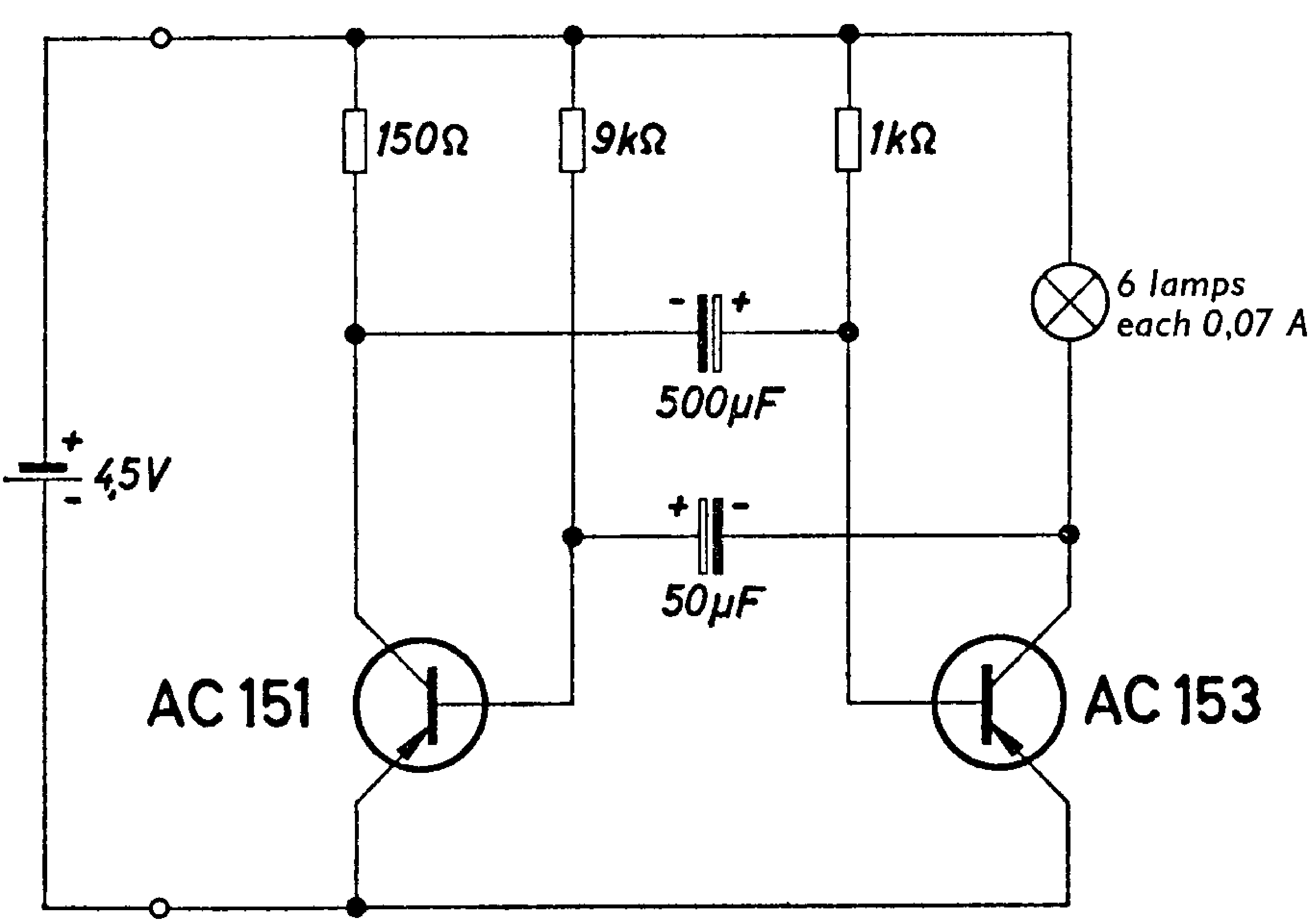


Fig. 3.1.

3.1. Blinker Circuit

Fig. 3.1. shows a blinking device with two transistors to be used in a blinking belt as it is worn, for example, by policemen or other persons who work in highway areas. The operating voltage is 4.5 V; thus a single flashlight battery may be used for the power supply. Six bulbs with a total power consumption of 1.65 W can be plugged into the blinking belt. The blinking frequency is 1.5 Hz on- and off-times are approximately equal.

Technical data

Operating voltage	4.5	V
Lamp power	1.65	W
Blinker frequency	1.5	Hz
On-time	320	msec
Off-time	320	msec

3.2. Blinker Circuit with Complementary Transistors

Astable multivibrators designed with complementary transistors do have several advantages compared to the usual circuit described in the previous chapter.

The essential advantage is the fact that a comparatively small charging capacitor can be used. Assuming an equal pulse time it can be smaller than in usual circuits by the factor of the transistor current gain. If the switch-on time of the load is much shorter than the switch-off time, a much better efficiency is obtained with this circuit, because both transistors always are simultaneously switched on or switched off. In the conventional circuit one transistor is switched on when the other is switched off. In battery operation this improvement of efficiency may be important.

In order to explain the circuit shown in Fig. 3.2. one has to start at an instantaneous state, at best in the pulse pause, e. g., when both the transistors are switched off. The transistor T_1 remains switched off as long as the voltage at the capacitor C_1, which is charged via the load and the series resistors R_1 and R_2, becomes as high as the threshold voltage of the base-to-emitter diode of transistor T_1. Then, at first a small base current will flow. The collector current of transistor T_1 flows via the base of transistor T_2. Thus the transistor T_2 is switched on and the voltage drop in the load (bulb) causes a

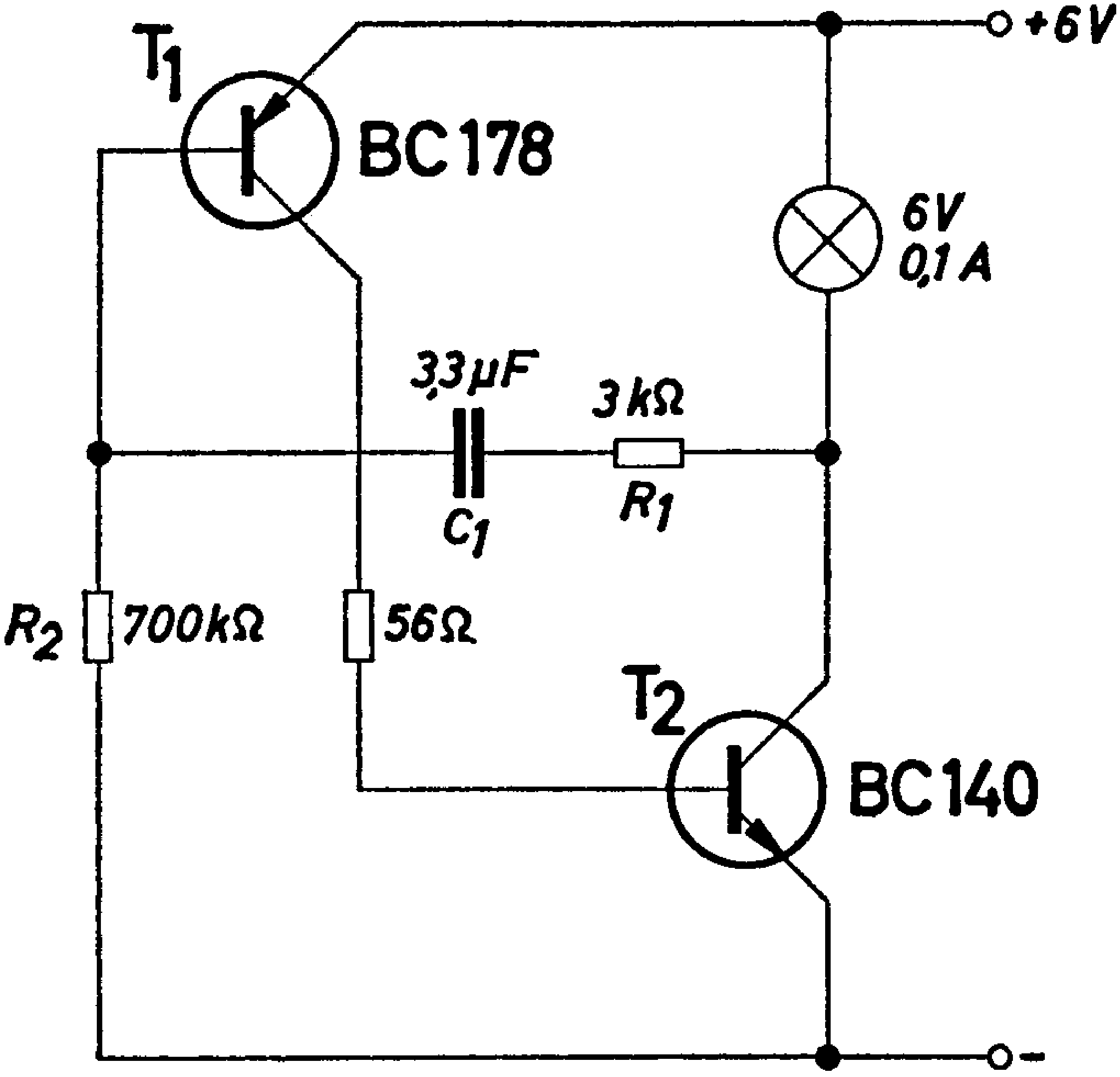


Fig. 3.2.

trend of the collector potential towards negative values. The base current of the transistor T_1 increases fast due to the feedback via the resistor R_1 and the capacitor C_1. This speeds up the switching of the circuit. Both transistors are switched on and the bulb burns. The capacitor C_1 is discharged via resistor R_1 and the base-to-emitter diode of transistor T_1. As soon as the capacitor C_1 is discharged to such a level that the base current is too low to keep the transistor T_1 switched on, the multivibrator flips back again. The duration of the pulse pause, therefore, is mainly determined by the resistor R_2 and the pulse time by the resistor R_1. The voltage increase at the collector of transistor T_2 towards positive values will be applied to the base of transistor T_1 again via the feedback. Therefore, the switching is very fast again. The generated voltage does really have an exact square wave shape, so that switching losses at the transistors remain very low.

At an operating voltage of 6 V and at the highest possible ambient temperature, the leakage current of capacitor C_1 should not exceed 3 µA.

Technical data

Operating voltage	4.5 to 6	V
Operating temperature	—40° to +60°	C
Pulse frequency	0.58 Hz ± 10	%
Pulse time	50	ms
(Switch-on time of the bulb)		

3.3. Monostable Circuit for a Timing Element

With the circuit shown in Fig. 3.3. a starting delay or a breaking delay can be obtained depending in which collector circuit of the transistors used the relay is located.

The monostable multivibrator becomes unstable by pushing key K. At this moment capacitor C_1 is charged to the total operating voltage. It has been charged via the relay 2, the resistor R_1, the diode D_1, and the base-to-emitter diode of the transistor T_1.

By pushing key K transistor T_2 is switched on and transistor T_1 is switched off via a feedback. The capacitor C_1 now is being discharged (or, rather, charged with opposite polarity) via the collector-emitter path of transistor T_2 and via the resistors R_1, R_2, and R_3. A charging with opposite polarity, however, does not occur, as at the zero crossing of the voltage, the capacitor diode D_1 becomes conductive again and transistor T_1 is switched on. The transistor T_2 is switched off via the resistor R_4 and the stable state is reestablished.

The stable state of operation is always set immediately after the operating voltage is applied to the circuit, because capacitor C_2 at the base of transistor T_2 keeps the base potential momentarily at a low value.

The function of the delay circuit is not influenced if key K remains pushed, as the release pulse is applied to the circuit via a capacitor.

As already mentioned in this circuit a relay may first pick up and then drop again after a certain time or vice versa. If only one of these two functions is desired, e. g., the first one, a smaller charging capacitor can be used if the circuit is changed slightly. Then the circuit with transistor T_1 has to be more highly resistive. Relay 1 is replaced by a resistor of 25 kΩ. Resistor R_2 can be increased to a maximum value of 2 MΩ. Hereby the value of capacitor C_1 can be reduced to an eighth (for the same delay time).

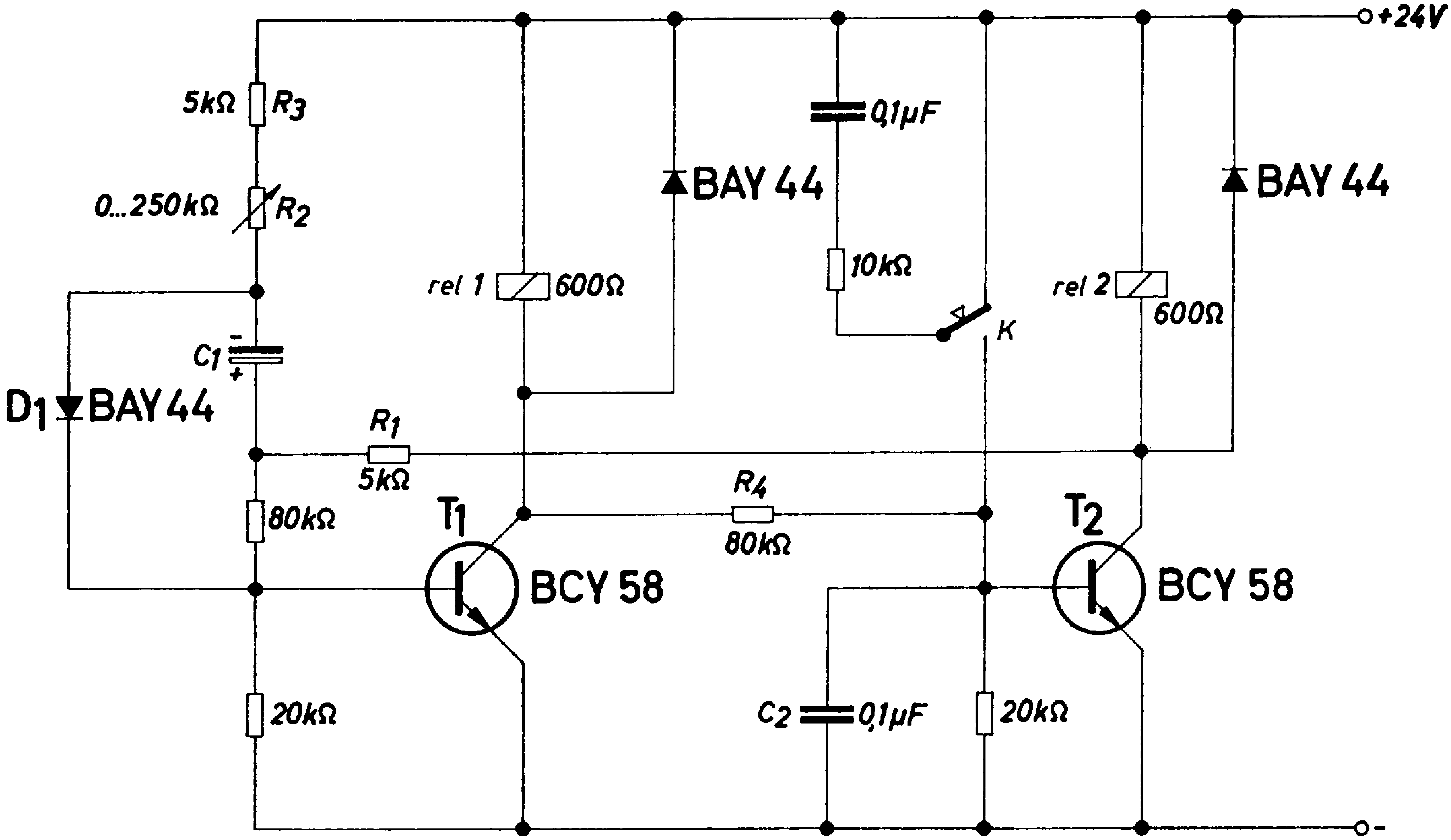


Fig. 3.3.

Fig. 3.4. indicates another possibility of decreasing the value of the capacitor C_1. (The circuit shown in Fig. 3.3. is enlarged by means of an extra pre-stage, and the charging resistor R_2 may have a maximum value of 4 MΩ). Therefore, for the same delay time, the charging capacitor C_1 needs to be only a sixteenth of the value in the circuit shown in Fig. 3.3. The charging capacitors are to have low leakage currents.

Technical data	Fig. 3.3.	Fig. 3.4.	
Operating voltage	24	24	V
Maximum pulse time $C_1 = 6\ \mu F$	1.1	17.6	s
Tolerance in switching time at a temperature of 80 °C referred to 25 °C	7	10	%

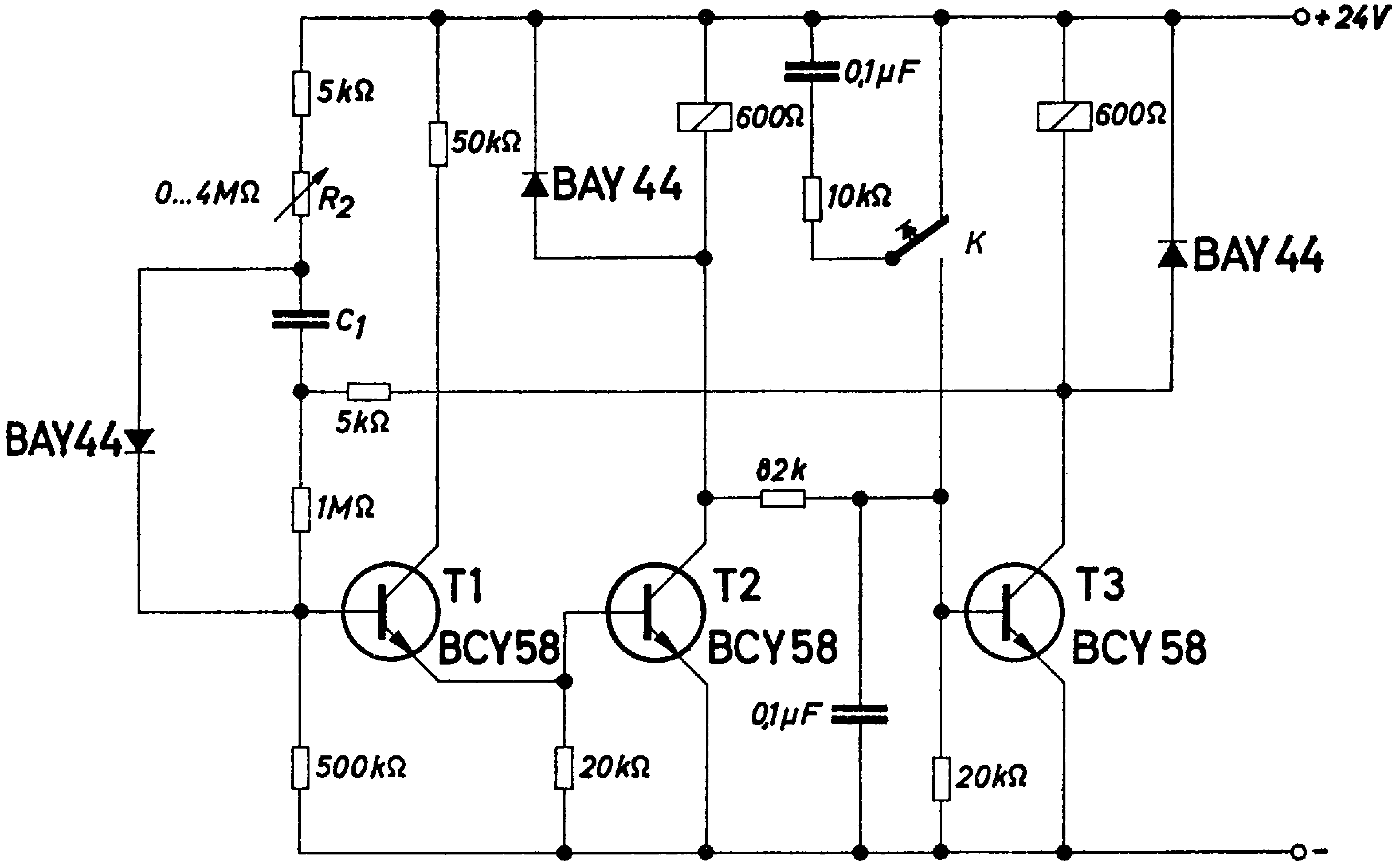


Fig. 3.4.

3.4. Delay Circuit for 3 to 60 Minute Intervals

In the delay circuit shown in Fig. 3.5. a capacitor C_1 is charged by a DC current on which periodic pulses are superimposed. The pulses are generated by an astable multivibrator.

If switch S_1 is closed, the process of delay starts. Capacitor C_1 is charged via the resistors R_1 and R_2. The superimposed pulses are led via the capacitor C_2. At the beginning of discharging almost the entire operating voltage drops at the resistor R_1. This voltage drop decreases when the capacitor is charged. As soon as the pulse peaks of the charging current and thus the voltage drop undercut a certain value, the bistable multivibrator consisting of the transistors T_3 and T_4 flips. The voltage at which the flipping occurs depends on the magnitude of the bias voltage which is set by resistor R_3 at the input of the bistable multivibrator. As soon as a voltage at point A of the circuit is more negative than at point B, the diode D_1 is forward biased and the starting pulse is applied to the bistable multivibrator.

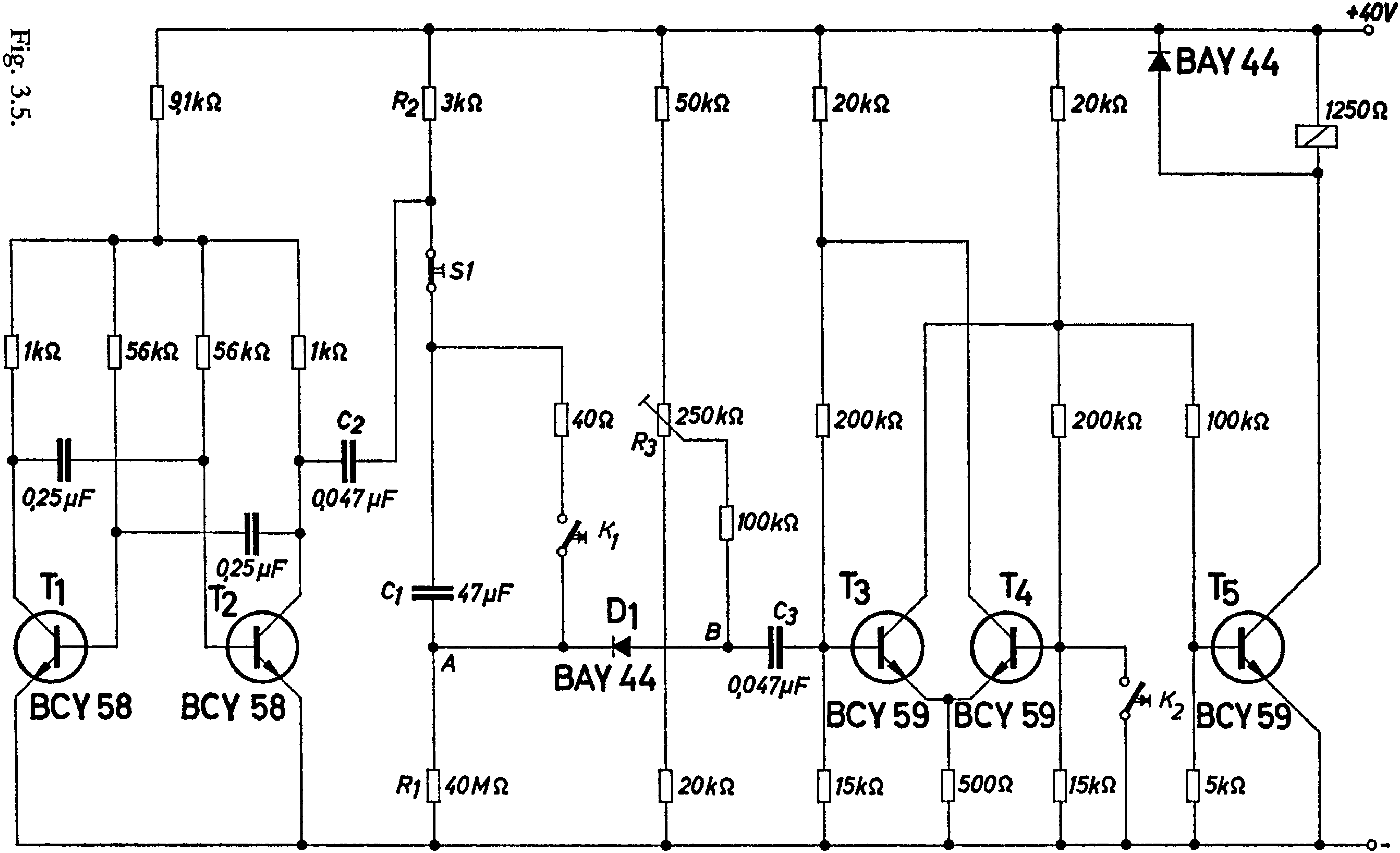

+40V
9,1kΩ
R2
3kΩ
50kΩ
20kΩ
20kΩ
BAY 44
1250Ω
S1
1kΩ
56kΩ
56kΩ
1kΩ
40Ω
250kΩ
R3
200kΩ
200kΩ
100kΩ
C2
0,25µF
0,047µF
100kΩ
K1
0,25µF
C1
47µF
T1
T2
D1
T3
T4
T5
A
B
C3
BAY 44
0,047µF
K2
BCY 58
BCY 58
BCY 59
BCY 59
BCY 59
R1
40MΩ
20kΩ
15kΩ
500Ω
15kΩ
5kΩ

Fig. 3.5.

Thus the magnitude of the delay time is set at resistor R_3. This does have the advantage that not, as usually, the highly resistive charging resistor R_1 has to be controlled, but a relatively low-resistance potentiometer can be used for time setting. The superimposition of pulses makes a decoupling for DC currents with the capacitor C_3 between driving section and amplifier possible. Hereby a very good temperature stability is achieved.

At the output a switching amplifier is connected to the bistable multivibrator. This amplifier switches the relay after lapse of the delay time.

The circuit can be set to starting position by pushing the keys K_1 and K_2.

The capacitor C_1 has to have a small leakage current (e.g. MKL capacitor).

Technical data

Operating voltage	40	V
Delay time	3 to 60	min.
Tolerance of the delay time at an ambient temperature of 50 °C referred to 20 °C	—4	%

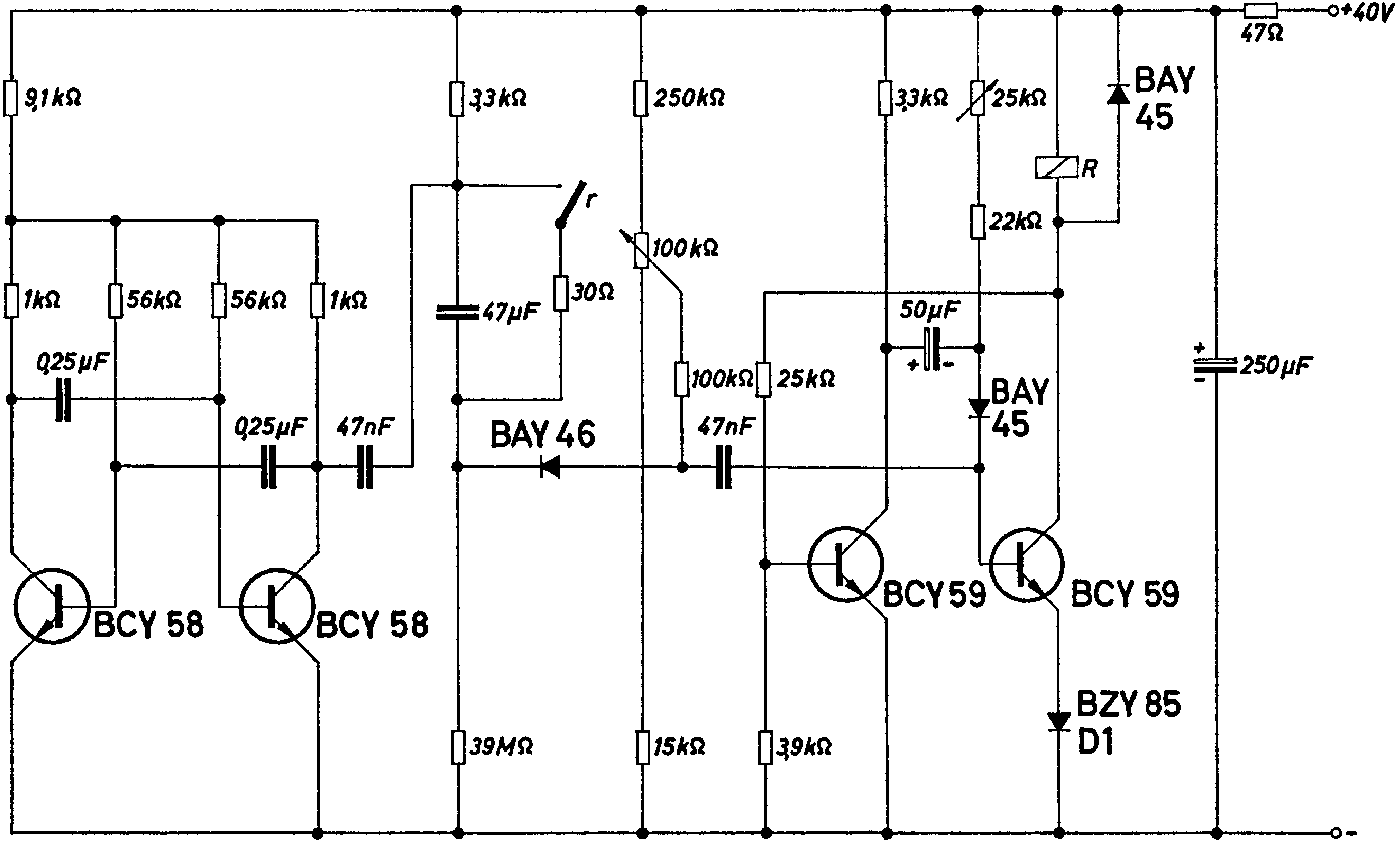

+40V
47Ω
9,1kΩ
3,3kΩ
250kΩ
3,3kΩ
25kΩ
BAY 45
R
r
22kΩ
100kΩ
1kΩ
56kΩ
56kΩ
1kΩ
47μF
30Ω
50μF
0,25μF
100kΩ
25kΩ
250μF
BAY 45
0,25μF
47nF
BAY 46
47nF
BCY 58
BCY 58
BCY 59
BCY 59
BZY 85 D1
39MΩ
15kΩ
3,9kΩ

Fig. 3.6.

3.5. Electronic Time Switch

The circuit shown in Fig. 3.6. operates in principle in the same way as the circuit described in the previous chapter. The only difference is that in this circuit a monostable multivibrator is located at the output. The multivibrator is driven at intervals of 0.3 to 1.2 hours and generates output pulses of 1 to 3 seconds duration. Such a circuit can be used for the control of short processes which have to occur periodically in relatively long intervals, e. g., greasing of movable parts.

Technical data

Operating voltage	40 V ± 10	%
Pulse pause	0.3 to 1.2	hours
Pulse time	1 to 3	s
Maximum ambient temperature	60	°C
Maximum timing error in a temperature range from 0 to 40 °C	± 5	%

3.6. Starting Delay Circuit

The circuit shown in Fig. 3.7. operates as a real starting delay circuit, i. e., a relay at the output picks up only a certain time after the operating voltage has been applied to the circuit. At the input of the circuit a bridge has been arranged consisting of the resistors R_1 and R_2, the capacitor C_1 and the resistors R_3 and R_4. In the zero branch of the bridge there is the input of transistor T_1. The diode D_1 only has to prevent that a too high reverse voltage is applied to the input of the transistor T_1. When the operating voltage is applied to the circuit, capacitor C_1 is charged via resistors R_1 and R_2. Both transistors remain switched off as long as the voltage at the capacitor C_1 is sufficiently high to drive the transistor T_1 beyond saturation. Then the transistor T_2 also is switched on and the relay picks up.

As soon as the operating voltage is switched off the relay drops immediately and the capacitor C_1 is discharged via diode D_2 and the resistors R_3 and R_4. Now the next switch-on process may follow.

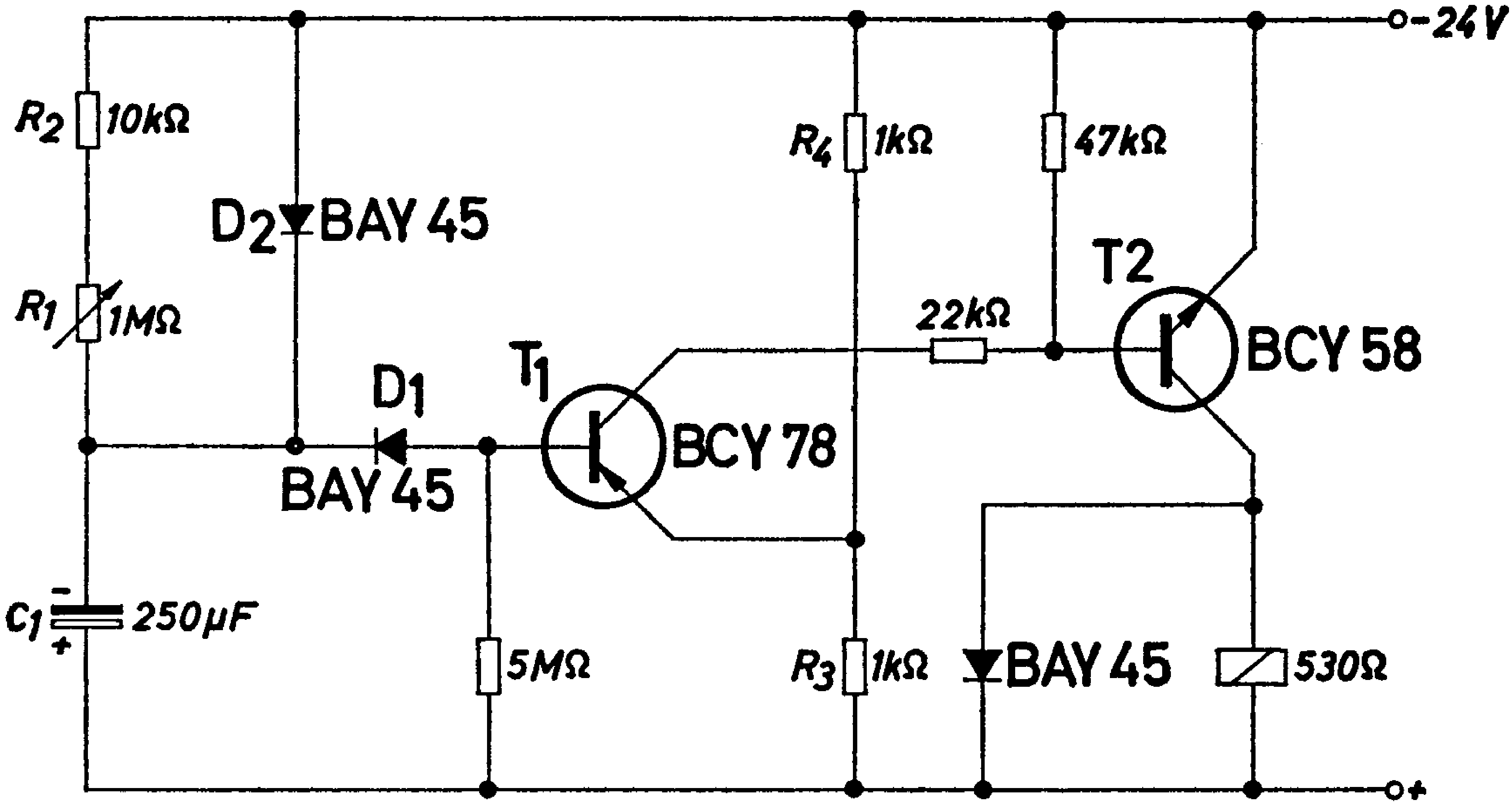


Fig. 3.7.

Technical data		
Operating voltage	24 V ± 20	%
Delay in starting	1.7 to 170	s
Repetition time	2.5	s
Maximum ambient temperature	60	°C

3.7. Delay Circuit with Thyristor-Tetrode

By using our thyristor tetrode BRY 20 a delay circuit can be built with one single active component, as indicated in Fig. 3.8. The circuit operates like a mechanical time-lag relay that picks up only a certain time after the operating voltage has been applied to the circuit.

If the constant operating voltage of 24 V is applied to the circuit shown in Fig. 3.8., the capacitor C_1 is charged via the relay and the resistors R_1 and R_2. The capacitor C_1 and the input of the thyristor tetrode BRY 20 as well as a Zener diode D_1 are connected in parallel. As long as the voltage at the capacitor is lower than the Zener voltage of the diode, the thyristor-tetrode remains switched off. The capacitor C_2, which is connected in parallel with the input, prevents the thyristor from being switched on when stray pulse peaks occur.

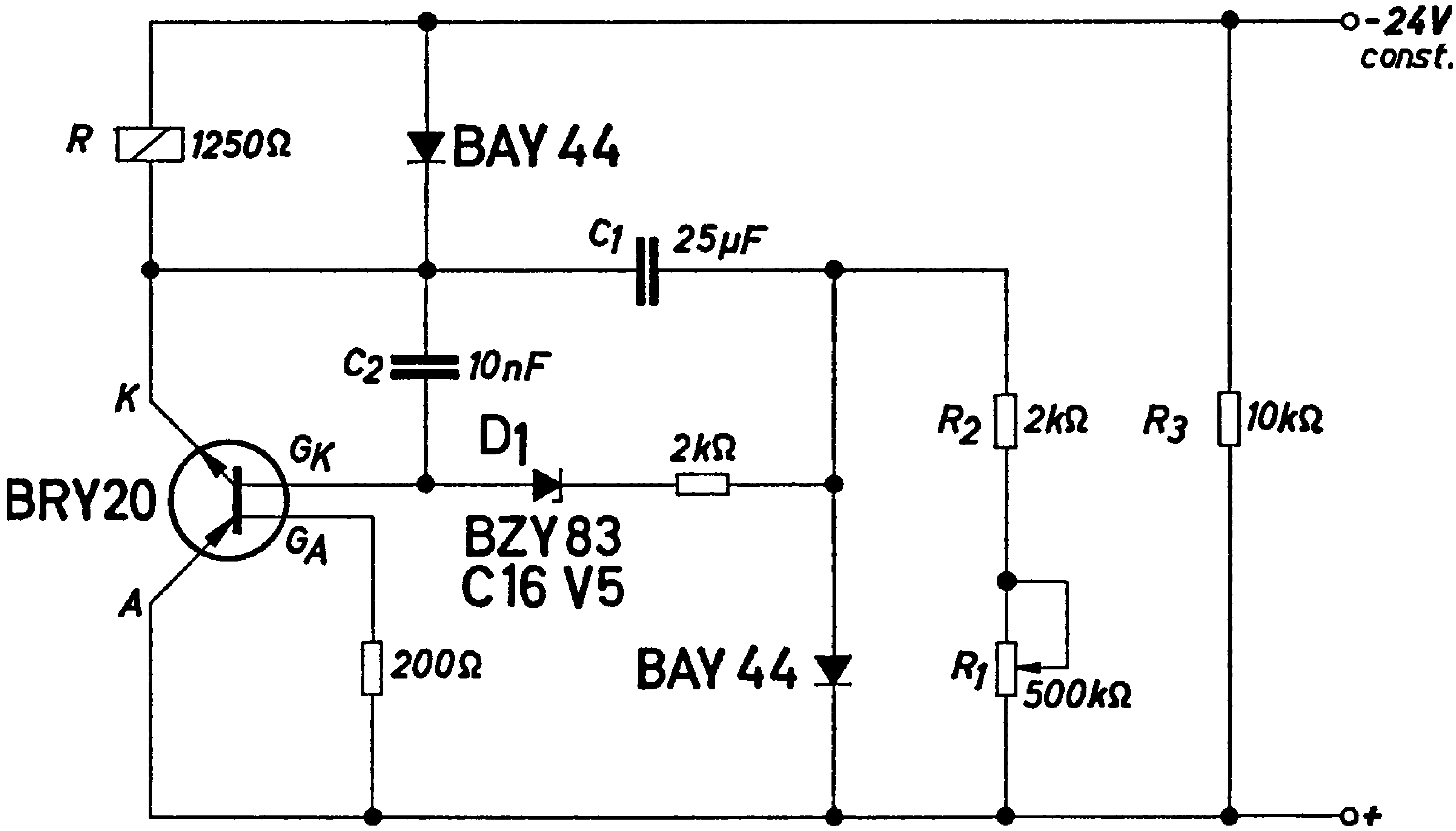


Fig. 3.8.

As soon as the voltage at the capacitor C_1 exceeds the Zener voltage of diode D_1, the thyristor tetrode is switched on and the relay picks up.

The time for attaining this voltage depends on the charging current of the capacitor C_1 which can be set by resistor R_1: So in this way the delay time is controlled.

As soon as the thyristor tetrode is switched on, the capacitor C_1 is discharged rapidly via the thyristor.

If the operating voltage is immediately turned off after the relay has picked up, the capacitor C_1 is discharged via the resistor R_3. In order to keep the delay time as constant as possible the operating voltage has to be stabilized.

Technical data

Operating voltage	24	V const.
Operating current	2 to 22	mA
Delay time	0.1 to 10	s
Repetition time	20	ms
Tolerance of the delay time at an ambient temperature of 70 °C referred to room temperature	± 2	‰

Relay: Trls 154 C according to TBv 65422/93 D

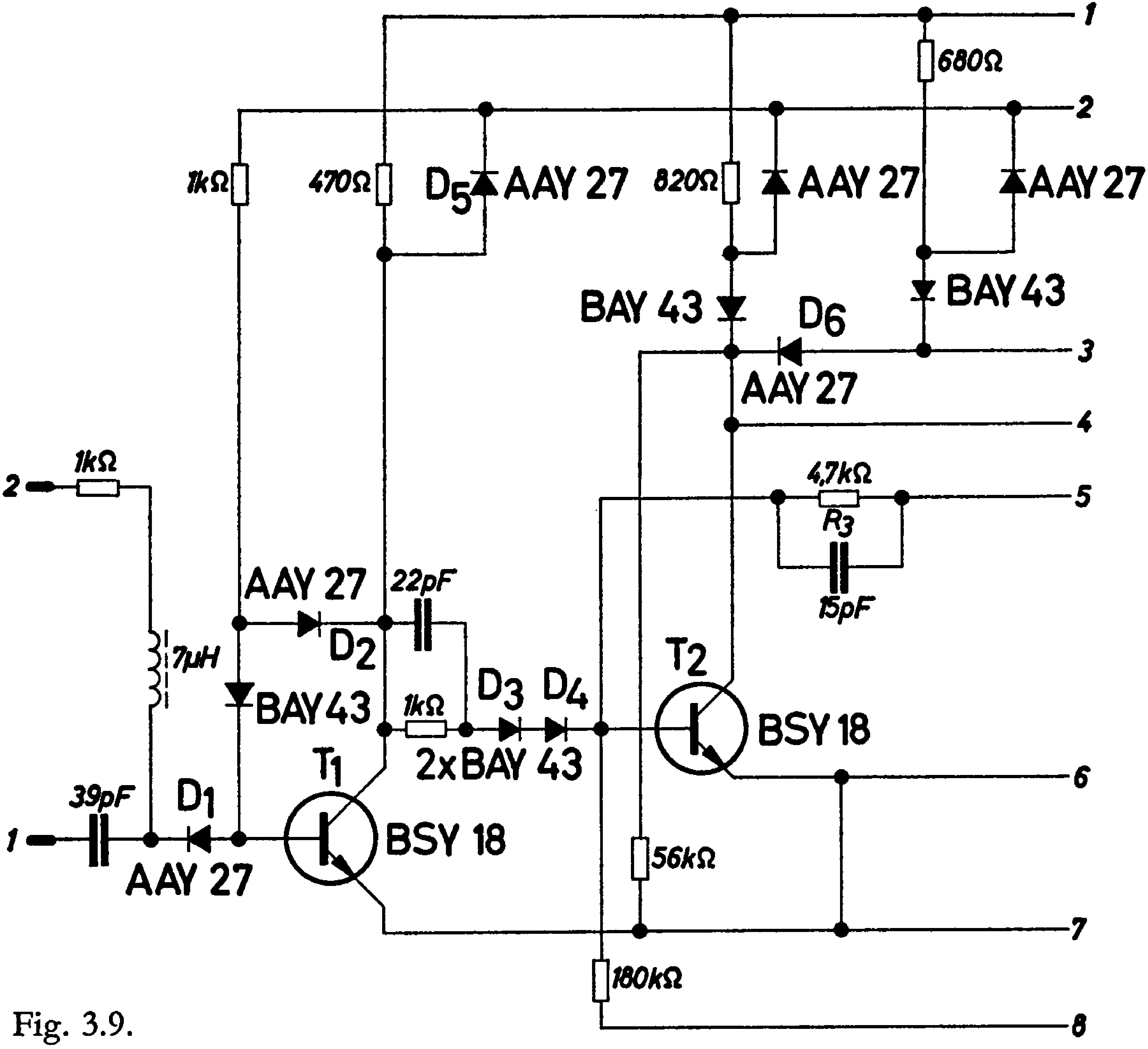
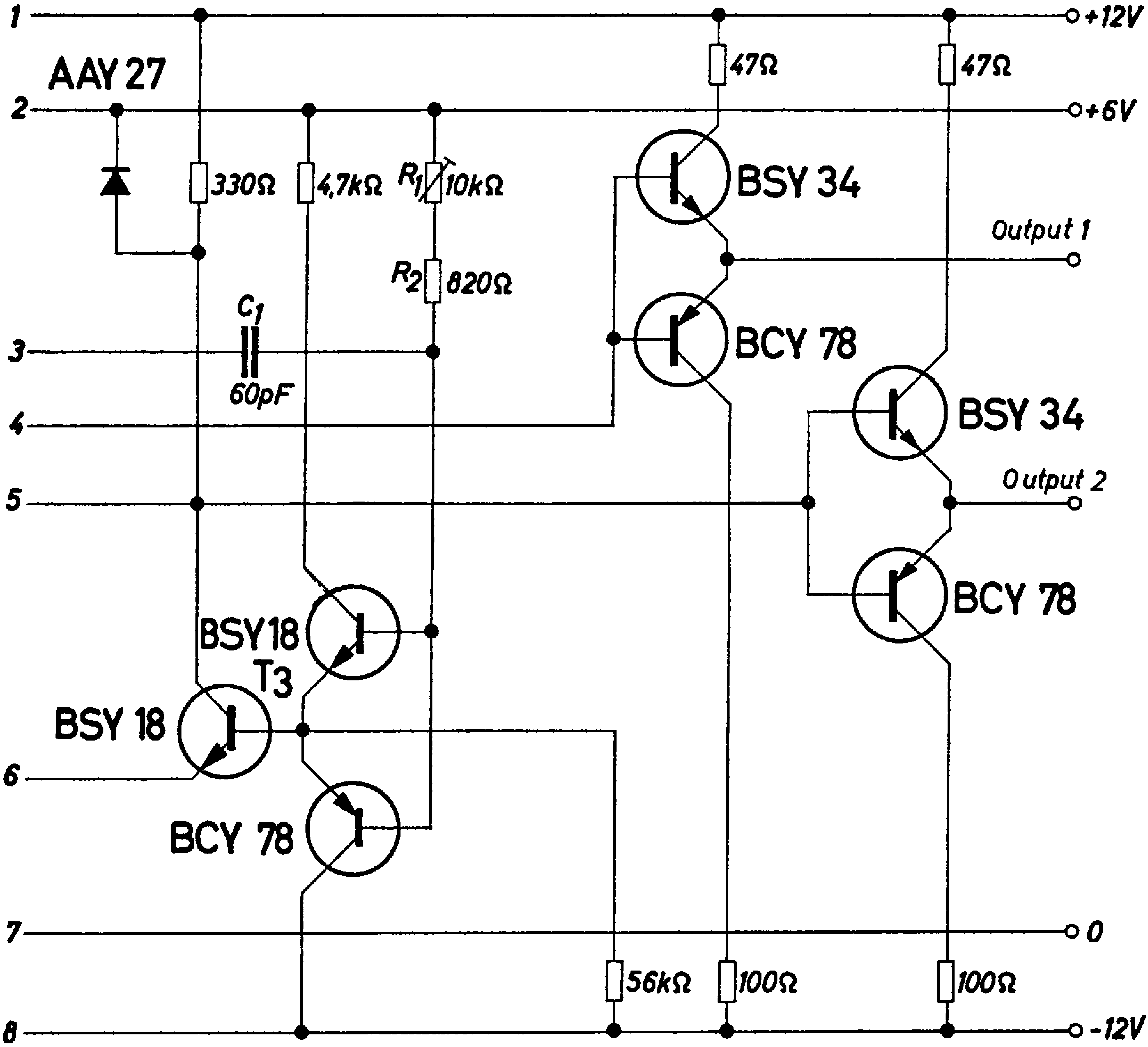


Fig. 3.9.

3.8. Monostable Multivibrator up to 10 MHz

The monostable multivibrator shown in Fig. 3.9. operates perfectly at frequencies up to 10 MHz. The rise and fall time of the output pulses is about 20 ns.

The transistor T_1 is switched on in the stable state, transistor T_2, then, is switched off. Both silicon diodes D_3 and D_4 prevent the saturation voltage of transistor T_1 from switching on transistor T_2. The multivibrator has two inputs. One input (2) prepares the switching to the unstable state while the other input (1) serves for the switching itself. At input 2 the circuit can be blocked by a positive voltage of 6 V. The diode D_1, then, is reverse-biased and blocks the switch-on pulses. A switching of the monostable multivibrator at input 1 by the negative edge of a pulse, e. g., by a pulse



step of +6 V to zero, is only possible, if there is no voltage at this input.

The transistor T_1 is switched off for a short time so that transistor T_2 is switched on. The collector potential of the switched off transistor T_1 is kept at a value of 6 V by diode D_5. This warrants a fast switch-on of transistor T_1 after the switching of the monostable multivibrator. The diode D_2 prevents a saturation of the conductive transistor and also makes a fast switch-off of transistor T_1 possible.

As soon as transistor T_2 has been switched on, the capacitor C_1 is discharged via the resistors R_1 and R_2, the diode D_6, and the transistor T_2. When the discharging is finished, transistor T_3 is switched on and the monostable multivibrator flips back in the stable state. The transistor T_2 is switched off by a feedback via the resistor R_3.

The switching of transistor T_3 occurs via an emitter-follower stage consisting of a npn and a pnp transistor. One transistor is switched on and the other transistor is switched off by a positive or negative input pulse and vice versa. The transistor being switched on operates as low-resistance load for the transistor being switched off. This results in a fast switching process. Similar stages are connected to the two outputs, i. e., to the collectors of transistors T_2 and T_3. Thus output pulses of a very steep slope are achieved. As the collector resistance of transistor T_2 has been divided by the diode D_6, the charging current of capacitor C_1, which flows after flipping into the stable state, does not influence the square-wave shape of the pulses. The width of the output pulses is set by the RC constant of the discharging circuit (resistors R_1 and R_2, capacitor C_1). At a constant value of the capacitor C_1, the pulse width can be varied by the potentiometer R_1 in a ratio of about 1:10.

Technical data

Operating voltage	± 12 and $+6$	V
Output amplitude	6	V
Minimum load resistance	50	Ω
Maximum operating frequency	10	MHz
Rise time of the output pulse	20	ns
Fall time of the output pulse	25	ns
Rise time of the input pulse	<30	ns
Voltage of the input pulse	4 to 6	V

3.9. Astable Multivibrator up to 10 MHz

The astable multivibrator shown in Fig. 3.10. is appropriate for operating frequencies up to 10 MHz. In order to describe its function one has to start from an instantaneous state, e. g., from that time of the period when transistor T_1 is switched off. The transistor T_2 is then switched on. The capacitors C_1 and C_2 are charged via this transistor and the resistor R_1. As soon as the potential at the emitter of transistor T_1 becomes negative compared to the potential at its base, transistor T_1 becomes conductive and transistor T_2 is switched off.

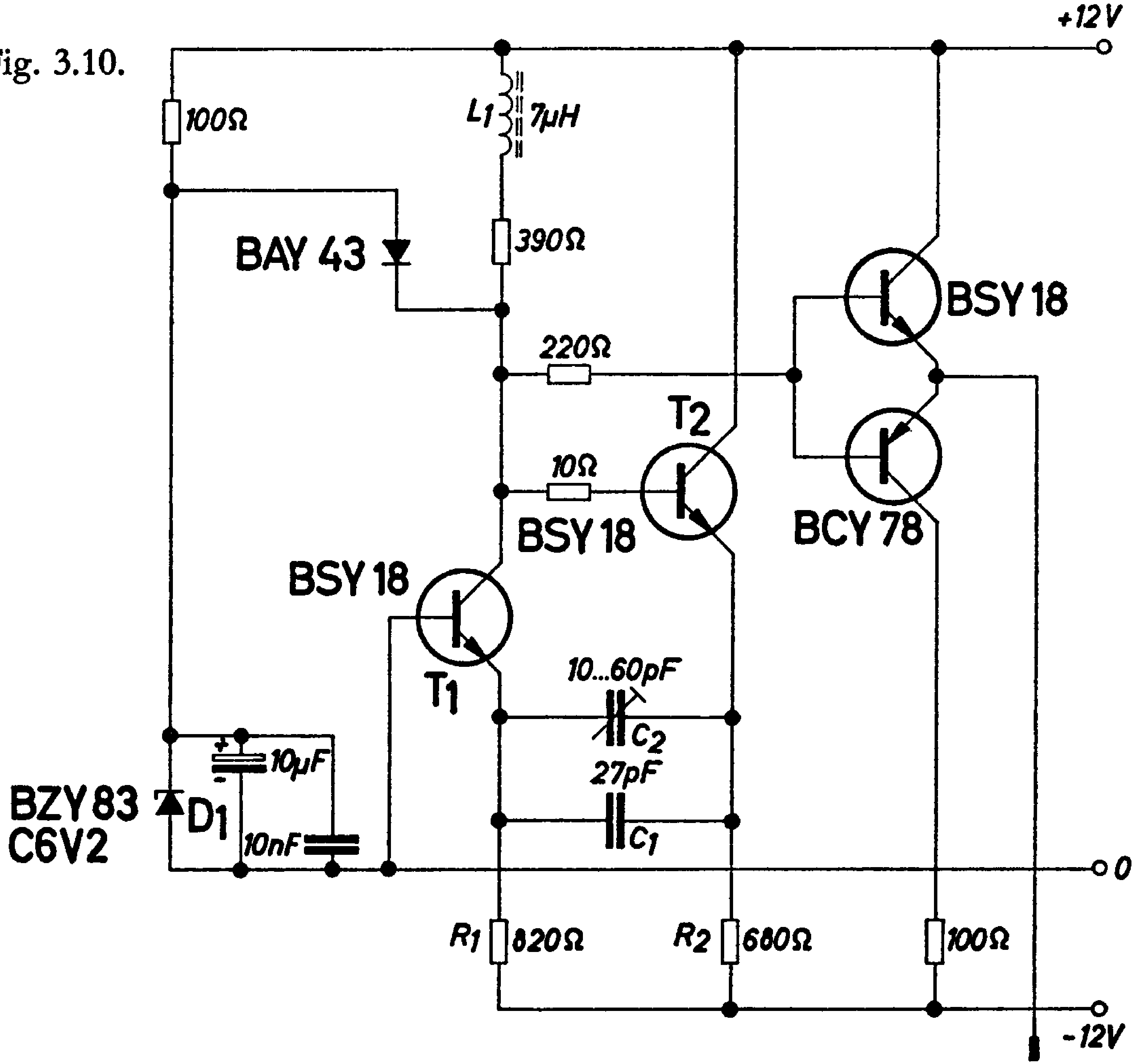
As in the previous example, the collector potential of transistor T_1 is also kept at a value of about 6 V by a Zener diode in order to effect fast switching. The capacitors C_1 and C_2 now are discharged via resistor R_2 and transistor T_1 until the potential at the emitter of transistor T_2 becomes negative with respect to the potential at its base. Then, the multivibrator flips back into that state at which our description started. The switching stage with two complementary transistors already described in the previous chapter is connected to the output of the multivibrator.

The width of the pulses is adjusted by resistor R_2, and the pulse pause is set by resistor R_1. At a constant duty cycle the frequency of the astable multivibrator can be influenced by varying capacitor C_2.

Technical data

Operating voltage	±12	V
Output amplitude	6	V
Rise time and fall time of the output pulses	<20	ns

Fig. 3.10.



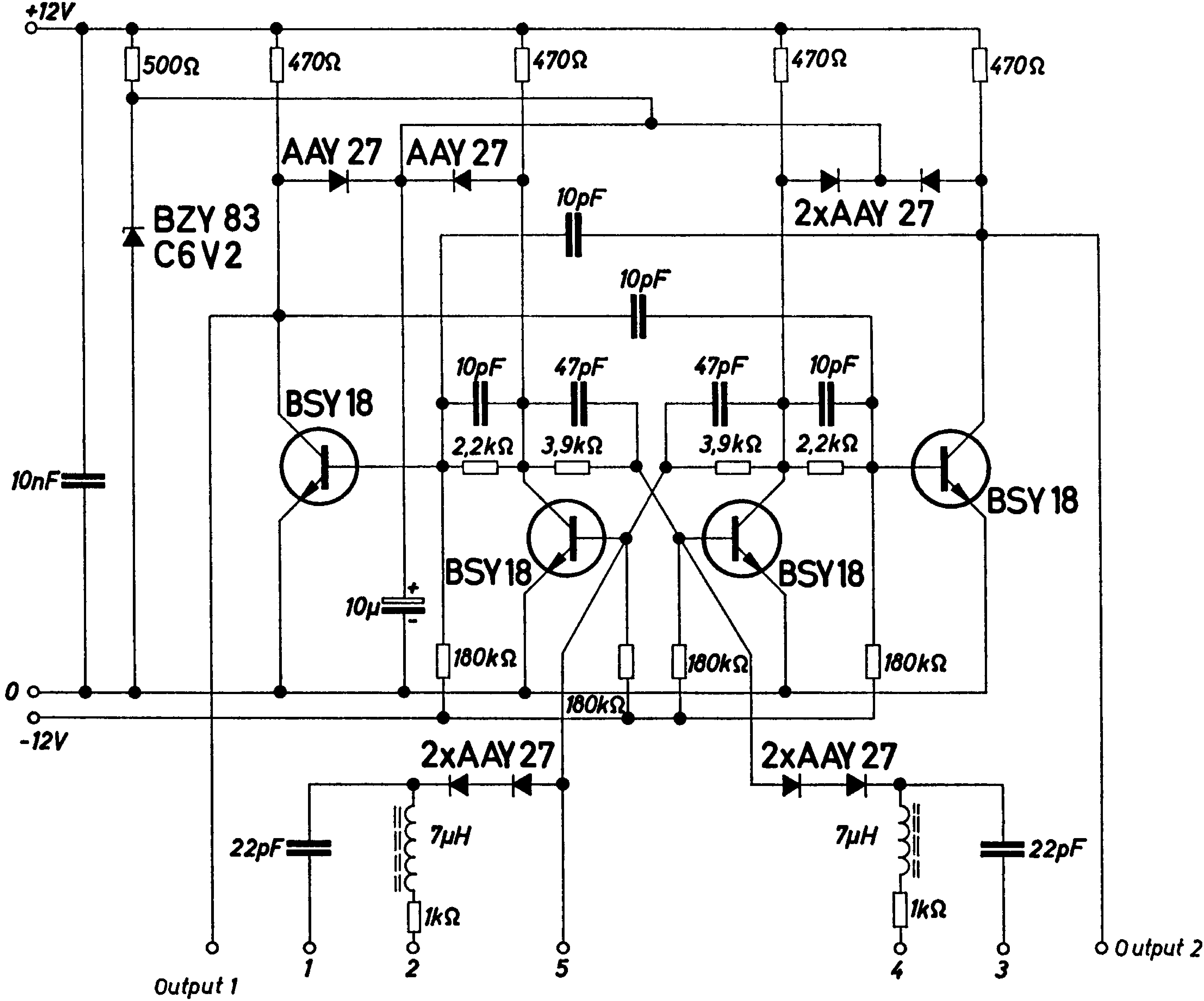

+12V
500Ω
470Ω
470Ω
470Ω
470Ω
AAY 27
AAY 27
2xAAY 27
BZY 83
C6V2
10pF
10pF
10pF
47pF
47pF
10pF
2,2kΩ
3,9kΩ
3,9kΩ
2,2kΩ
BSY 18
BSY 18
BSY 18
BSY 18
10nF
10µ
180kΩ
180kΩ
180kΩ
180kΩ
0
-12V
2xAAY 27
2xAAY 27
22pF
7µH
7µH
22pF
1kΩ
1kΩ
Output 1
1
2
5
4
3
Output 2

Fig. 3.11.

3.10. Bistable Multivibrator up to 20 MHz

In the bistable multivibrator shown in Fig. 3.11. there are again two inputs. One input prepares the switching (2 and 4) and the other input serves for the switching itself (1 and 3). In order to make it possible that the multivibrator can be switched at the inputs 1 or 3 by the negative edge of a pulse it is required that no voltage is applied to the corresponding inputs 2 or 4. With a positive voltage of 6 V at these inputs the bistable multivibrator will be blocked. The multivibrator can be reset at input 5 by a DC signal. Each branch of the multivibrator consists of two transistors. The coupling resistors are by-passed by capacitors in order to ensure fast pulse transmission. As already described in chapter 3.8. the collector potential of the transistor is set at a value of 6 V in order to obtain quick switching of the transistors.

Technical data

Operating voltage	± 12	V
Output pulse	6	V
Rise and fall time of the output pulse	<25	ns
Delay of the output pulse referred to the input pulse	<30	ns
Input pulse	3 to 6	V
Rise time of input pulse	<50	ns

4. Photo Amplifiers

In addition to the germanium photo diodes and silicon photo-voltaic cells already known there is now also a silicon photo transistor which delivers high output signals even at relatively low intensities of illumination. This results from the high current gain in the transistor chips used for these photo transistors. In this chapter several examples of photo amplifiers with germanium photo diodes, silicon photo-voltaic cells and also one example with the new photo transistor will be explained.

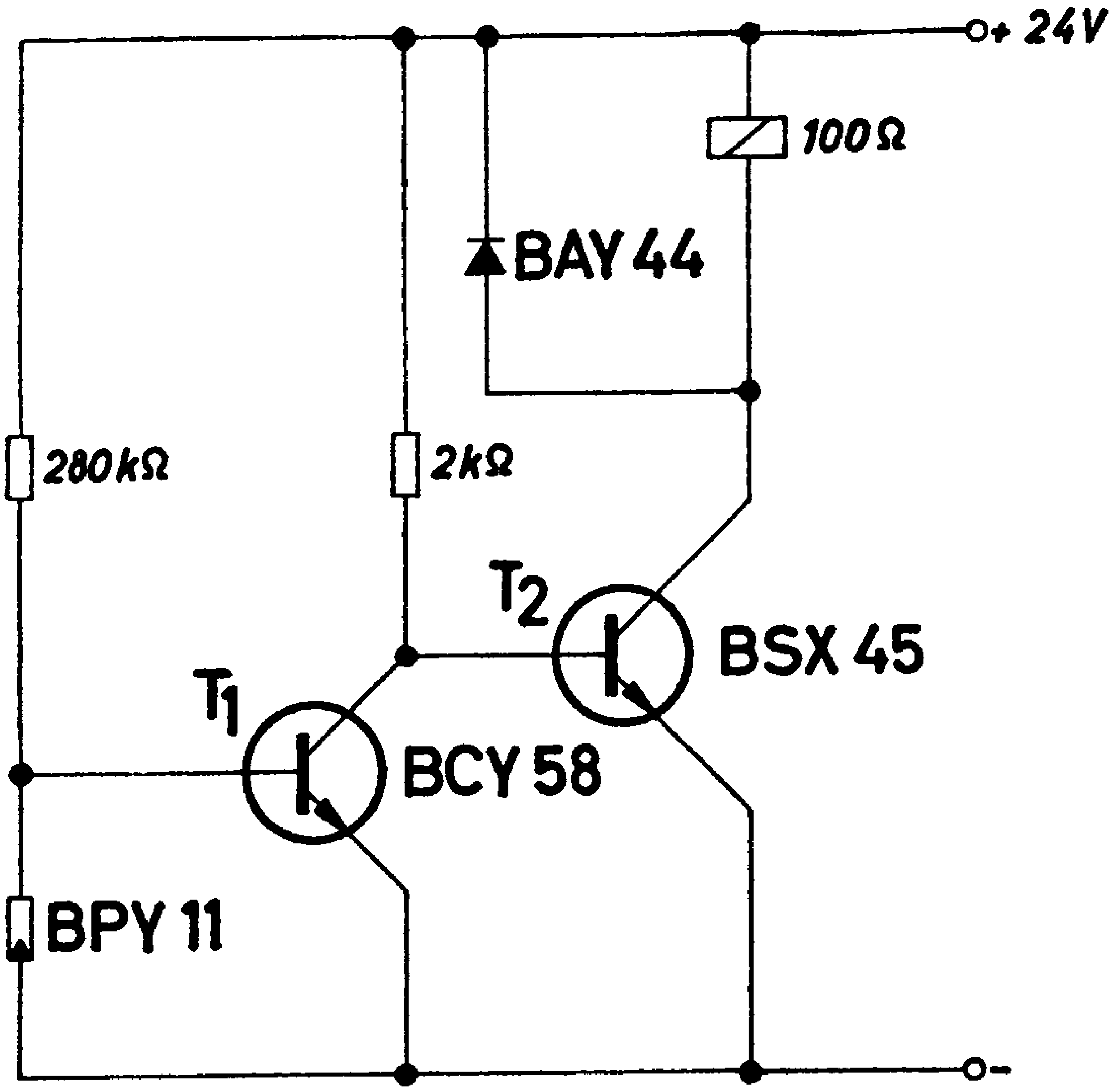


Fig. 4.1.

4.1. Switching Amplifier with a Photo-Voltaic Cell

Fig. 4.1. shows a simple switching amplifier with two transistors. With this amplifier a relay can be switched by illuminating the photo-voltaic cell. As long as the photo-voltaic cell is not illuminated, transistor T_1 is conductive and transistor T_2 is switched off. With sufficient illumination the transistor T_1 is short-circuited at the input,

transistor T_2 is switched on, and the relay picks up. This process is already started by the photo-voltaic cell for a photo current of min. 35 μA.

Technical data

Operating voltage	24	V
Operating current	12 to 200	mA
Min. photo current	35	μA
Max. ambient temperature	70	°C

4.2. Photo Amplifier for Light Pulses

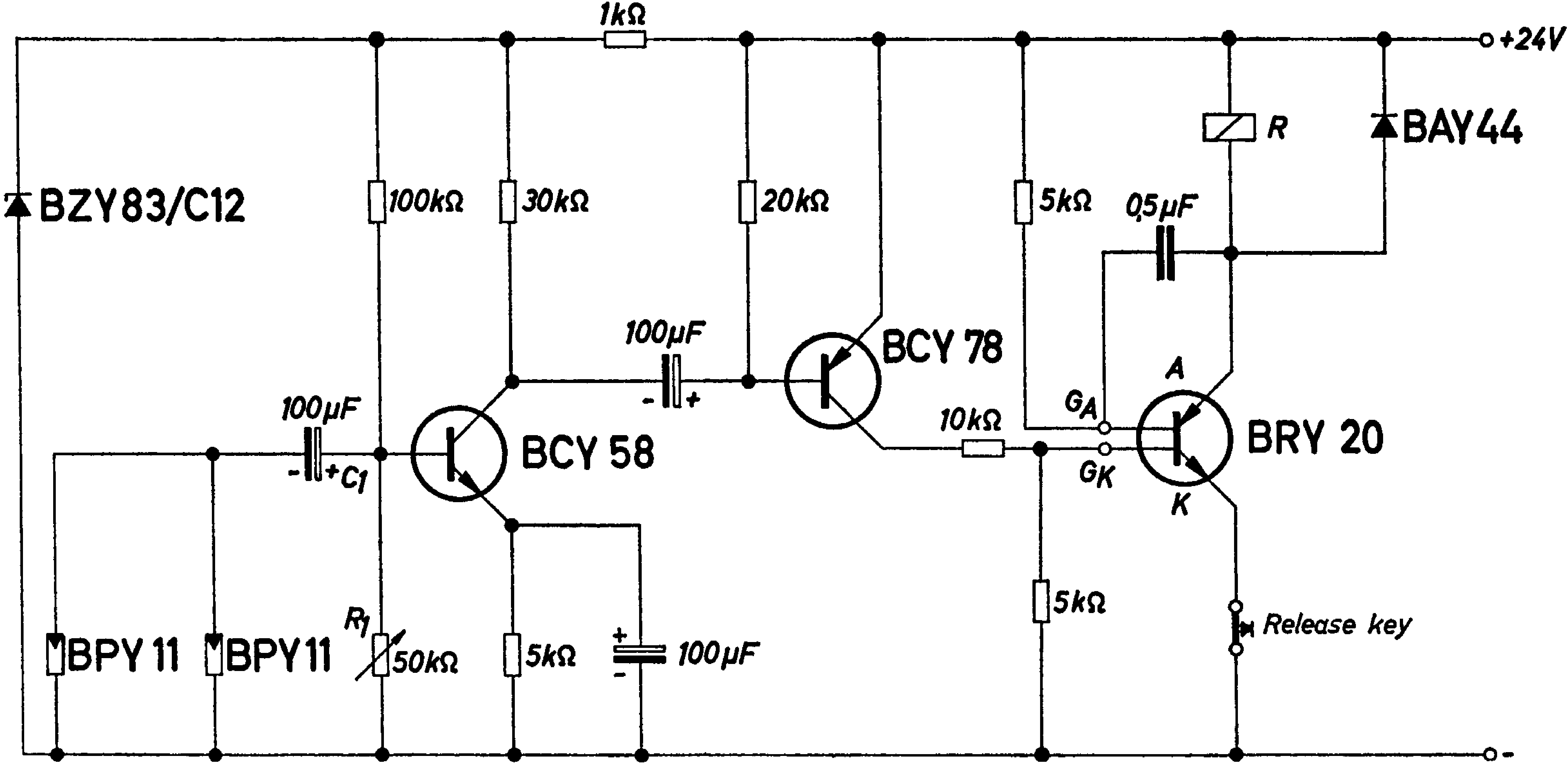
The photo amplifier shown in Fig. 4.2. is an AC or pulse amplifier. If the intensity of illumination at one of the two photo-voltaic cells at the input is changed, a short pulse is applied to the first transistor via the capacitor C_1. This pulse will be amplified and applied to the input of a thyristor-tetrode. The thyristor-tetrode then is switched on, and the relay at the output picks up. There are two photo-voltaic cells at the input to guarantee that the circuit operates perfectly even if one light barrier fails, for instance, by contamination.

The relay at the output remains picked up until the release key is pushed, as the thyristor-tetrode BRY 20 operates like a bistable multivibrator, that means, that a short input pulse is sufficient to flip it into another stable state.

Technical data

Operating voltage	24	V
Minimum width of the input pulse	5	ms
Maximum ambient temperature	60	°C

Relay R: Small round-type relay 6 V23006



1kΩ
+24V
BZY83/C12
100kΩ
30kΩ
20kΩ
5kΩ
R
BAY44
0,5µF
100µF
BCY 78
A
GA
BRY 20
10kΩ
GK
K
100µF
C1
BCY 58
5kΩ
Release key
BPY 11
BPY 11
R1
50kΩ
5kΩ
100µF

Fig. 4.2.

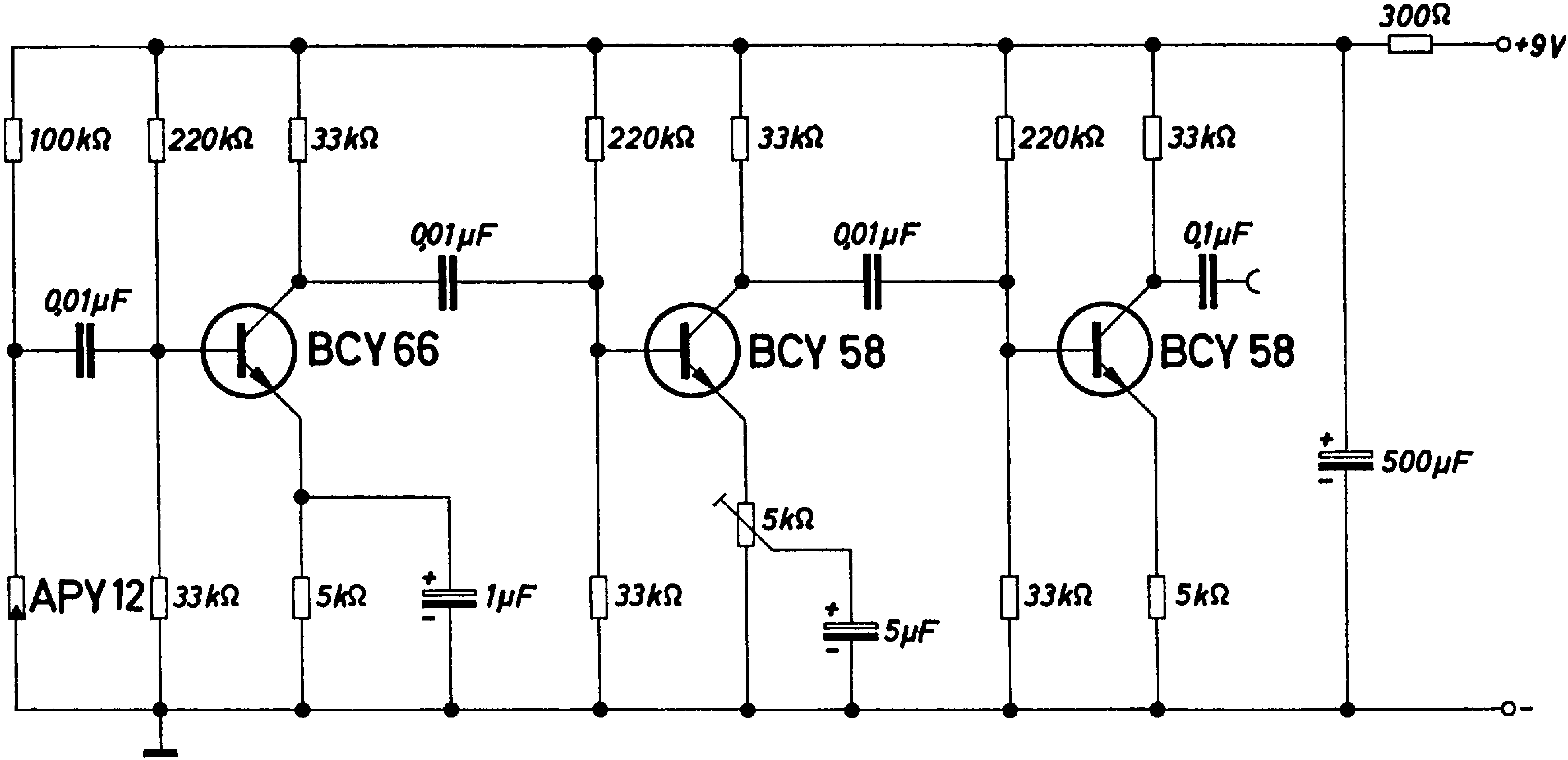

300Ω
+9V
100kΩ
220kΩ
33kΩ
220kΩ
33kΩ
220kΩ
33kΩ
0,01µF
0,01µF
0,01µF
0,1µF
BCY 66
BCY 58
BCY 58
500µF
5kΩ
APY 12
33kΩ
5kΩ
1µF
33kΩ
5µF
33kΩ
5kΩ
–

Fig. 4.3.

4.3. AC Photo Amplifier

In photo amplifiers which operate as DC amplifiers, the input sensitivity is determined by the temperature dependent cut-off currents of the input transistor. A considerable increase in input sensitivity is possible by using AC photo amplifiers, because the input sensitivity of these AC photo amplifiers is limited only by the noise of the first transistor. AC photo amplifiers do have another advantage. They do not response to continuous light, i. e., daylight. For that reason, light barriers also can operate in illuminated rooms. The circuit of such an AC photo amplifier is shown in Fig. 4.3. A normal AC amplifier with a total gain of 86 dB is connected to the photo-voltaic cell APY 12. The amplification can be controlled at the emitter resistor of the second stage. For application as a light barrier this amplifier is used as a pre-amplifier only. In the above form, however, it may be used for measuring purposes.

Technical data

Operating voltage	9	V
Operating current	0.5	mA
Frequency range (3 dB)	400 Hz to 14 kHz	

4.4. Light Barrier with Delay

The circuit of the light barrier indicated in Fig. 4.4. delivers two different output signals depending on whether the light barrier is interrupted for a shorter or a longer time. A photo transistor has been used as control device as this transistor already delivers a relatively high output pulse at low light intensities.

If the illumination of the photo transistor BPY 62 is interrupted, the potential at its collector changes towards positive values. The monostable multivibrator consisting of the transistors T_1 and T_2 is switched to the unstable state via the capacitor C_1, and the relay R_1 drops out. After lapse of the delay time of 20 to 100 ms, which can be set by the potentiometer R_3, the relay picks up again.

The transistor T_3 is switched on as long as the photo transistor is illuminated. If the light beam is interrupted the transistor T_3 is switched off. Hereby the capacitor C_2 will be charged via the resistors R_1 and R_2. As soon as the voltage at this capacitor becomes that high that transistor T_4 is switched on, the relay R_2 also picks up. Thus the relay

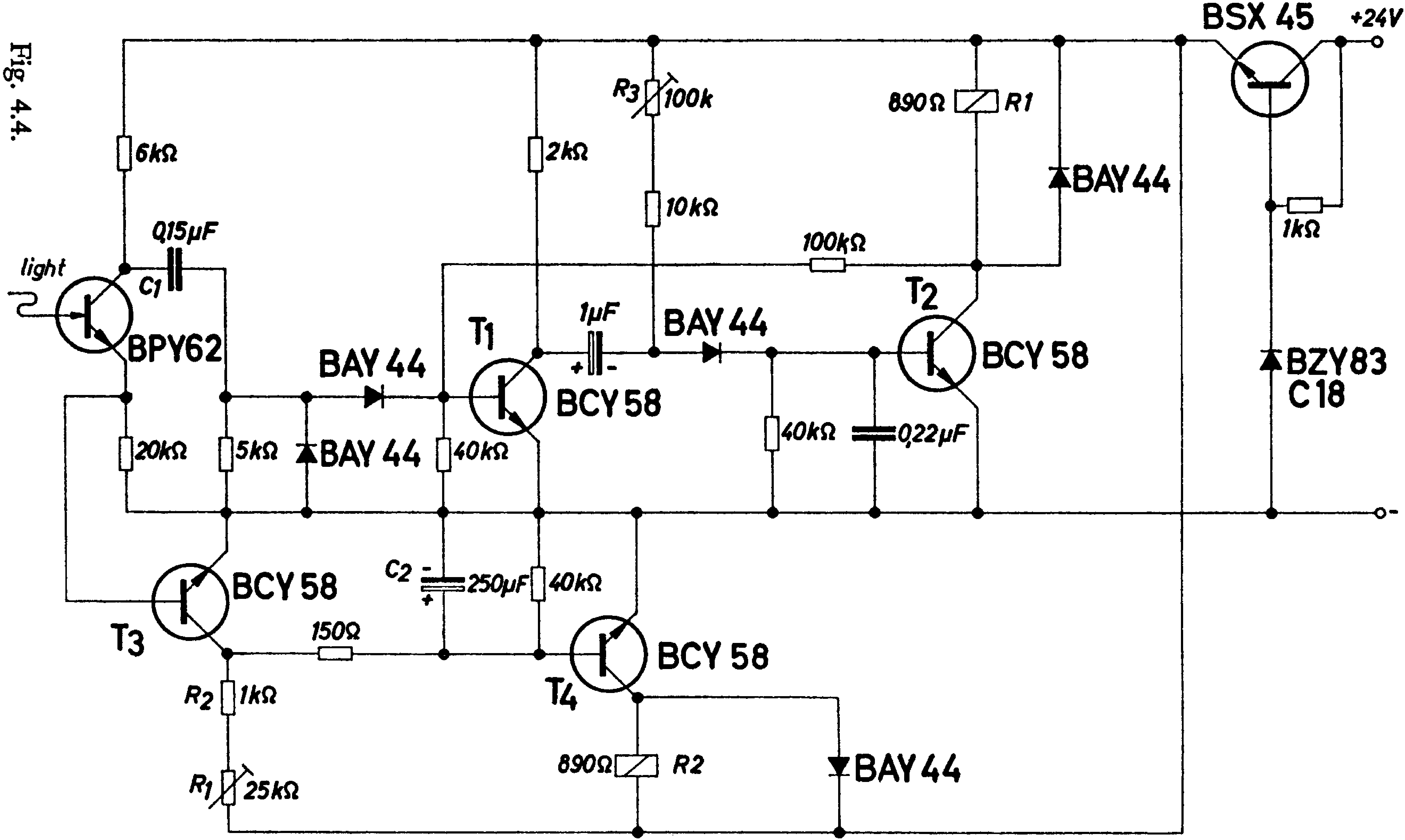

BSX 45
+24V
R3
100k
890Ω
R1
6kΩ
2kΩ
BAY 44
10kΩ
1kΩ
0,15μF
100kΩ
light
C1
T2
1μF
BAY 44
T1
BPY62
BCY 58
BZY83
C18
BAY 44
+
-
BCY 58
40kΩ
0,22μF
20kΩ
5kΩ
BAY 44
40kΩ
C2
-
250μF
40kΩ
+
BCY 58
T3
150Ω
BCY 58
R2
1kΩ
T4
890Ω
R2
BAY 44
R1
25kΩ
-

Fig. 4.4.

only attracts if the light barrier remains interrupted for a longer time as it corresponds to the delay time adjustable by resistor R_1. In order to prevent that this delay time varies too much, the operating voltage has to be stabilized by an arrangement of a transistor and a Zener diode. In order to guarantee perfect operation of the circuit, the illumination of the photo transistors has to be that intensive that a collector current of 2 mA is flowing.

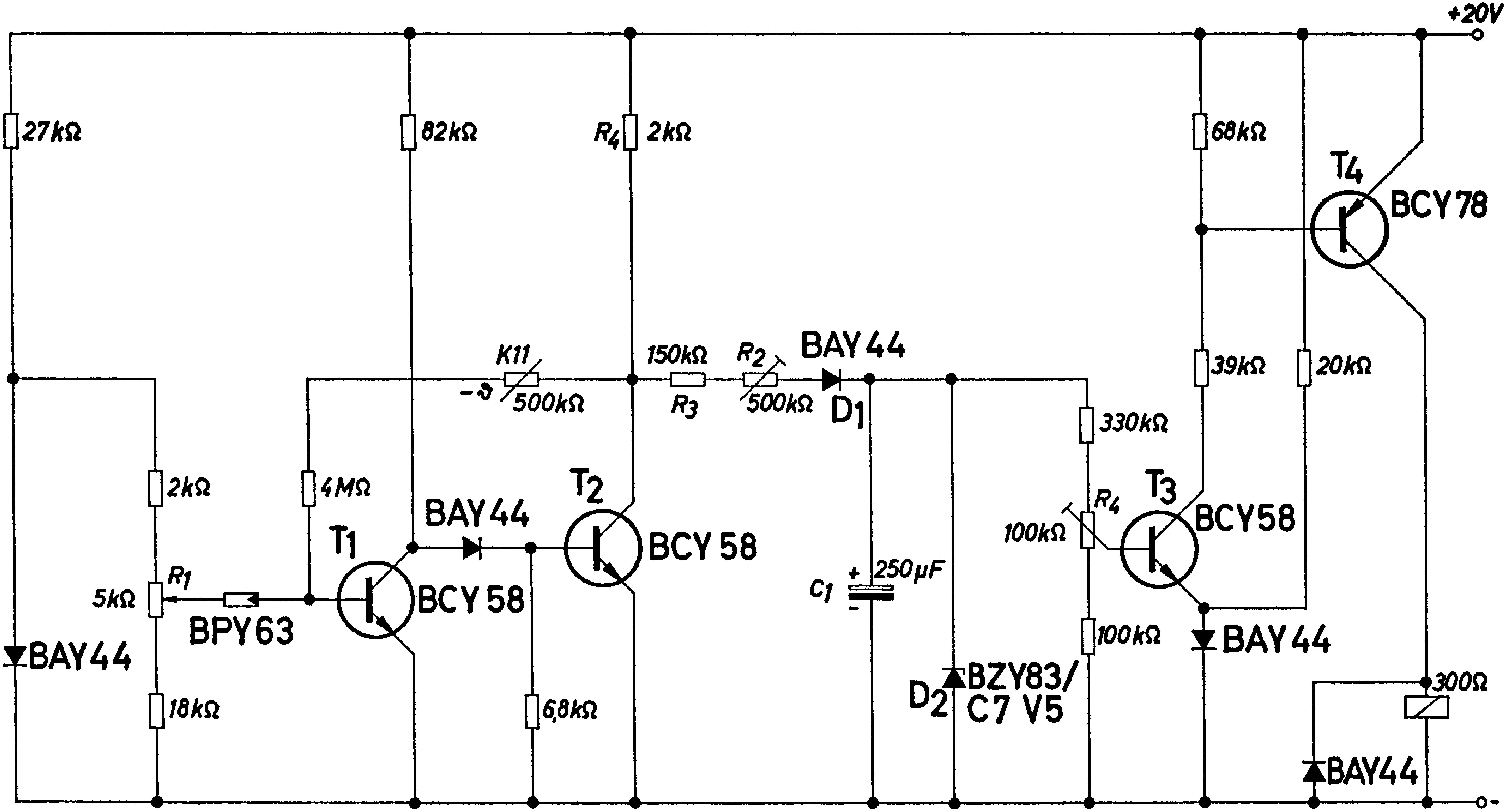
Technical data

Operating voltage	24	V
Pick-up time of the relay R_1	20 to 100	ms
Delay time for relay R_2	approx. 0.5	s

4.5. Twilight Switch with Delay

For the twilight switch a switch-on and switch-off delay is required in order to prevent that short variations in brightness (i.e. shadow of a flying bird, headlight of a car) cause a release. The circuit shown in Fig. 4.5. consists of a sensitive photo amplifier and a delay section with a switching amplifier. A silicon photo-voltaic cell, BPY 63 to which a bias voltage is applied, is connected to the input of the photo amplifier. This bias voltage is obtained by a forward biased silicon diode. For that reason its magnitude is constant, independent of variations in the operating voltage. Another advantage is the fact that the temperature dependence of the input sensitivity of transistor T_1 is compensated. Therefore, the complete photo amplifier is most temperature stable. The magnitude of the bias voltage as well as the input sensitivity can be set at the potentiometer R_1. If the photo-voltaic cell is illuminated, transistor T_1 is switched off and transistor T_2 becomes conductive. The capacitor C_1 then is charged via the resistors R_2, R_3, and R_4, and the diode D_1. A switching amplifier is connected to this capacitor via a voltage divider. As soon as the voltage at capacitor C_1 is high enough to switch on transistor T_3, the relay at the output picks up. The capacitor is charged up to the voltage of Zener diode D_2. For that reason the switching amplifier first remains switched on when the illumination of the photo-voltaic cell at the input has been interrupted, until the voltage of capacitor C_1 has dropped so low that the transistor T_3 is not kept in the conductive state anymore. Then the relay at the output drops out. The switch-off delay consequently is set by the potentiometer R_4. In this simple

Fig. 4.5.


+20V
27kΩ
82kΩ
R4 2kΩ
68kΩ
T4
BCY78
K11
-ϑ
500kΩ
150kΩ
R3
R2
500kΩ
BAY 44
D1
39kΩ
20kΩ
330kΩ
2kΩ
4MΩ
T2
BAY 44
BCY 58
T1
BCY 58
R1
5kΩ
BPY63
BAY44
18kΩ
6,8kΩ
C1
250µF
D2
BZY83/
C7 V5
T3
BCY58
R4
100kΩ
100kΩ
BAY 44
300Ω
BAY44

delay circuit the switch-on and switch-off delays depend on each other, i. e., the switch-on delay is set by the potentiometers R_2 and R_4 while the switch-off delay is determined by the potentiometer R_4. For that reason at first the switch-off delay has to be adjusted by potentiometer R_4 and then the switch-on delay has to be set by potentiometer R_2.

Technical data

Operating voltage	20	V
Sensitivity	10	Lux
Switch-on delay	50 to 70	s
Switch-off delay	30 to 60	s

5. Control Circuits

In this chapter all those circuits are described which are suitable for control techniques, e. g., temperature control circuits with thermistors as transducer and general purpose DC- and AC-amplifiers.

5.1. Control Circuit for a Stove Plate

The simplest way to control temperature of stove plates is to turn off the heating at a certain temperature and to turn it on again when it cools down. Because of the thermal inertia of the plates it is not possible to obtain a constant temperature. This disadvantage can be reduced by a control of the filament power the magnitude of which is varied. This can be achieved by series resistors, by voltage variation at the element, or by operation with current pulses, as used in the example shown in Fig. 5.1. The circuit consists of an astable multivibrator with the transistors T_1 and T_2 the duty cycle of which can be adjusted by resistor R_1. Hereby the operating time of the relay in the collector circuit of transistor T_2 can be changed between 0.6 and 12 s. The pulse pause has a constant value of 5 s. In this way the desired temperature of the stove plate will be determined in advance. The higher the desired temperature the longer will be the operating time of the relay. In addition the adjusted temperature is controlled by a NTC thermistor sensor K 273. It is arranged in a bridge with the variable resistor R_2 which sets the switching temperature. When the temperature drops below the adjusted value, transistor T_3 is switched on. The multivibrator is blocked by the conductive transistor T_3 so that the relay remains picked up. The heating remains constantly turned on via an operating contact of the relay until the desired temperature is reached .Then, the heating power will be decreased as already described by the initiation of multivibrator oscillations.

Instead of the relay at the output a power transistor may be used. The design of this stage, however, depends very much on the operated heating system and on the operating voltage.

The silicon diodes BAY 44 at the base of the transistors protect the emitter-base diode against too large inverse voltages: the permissible value for diffused transistors is rather small.

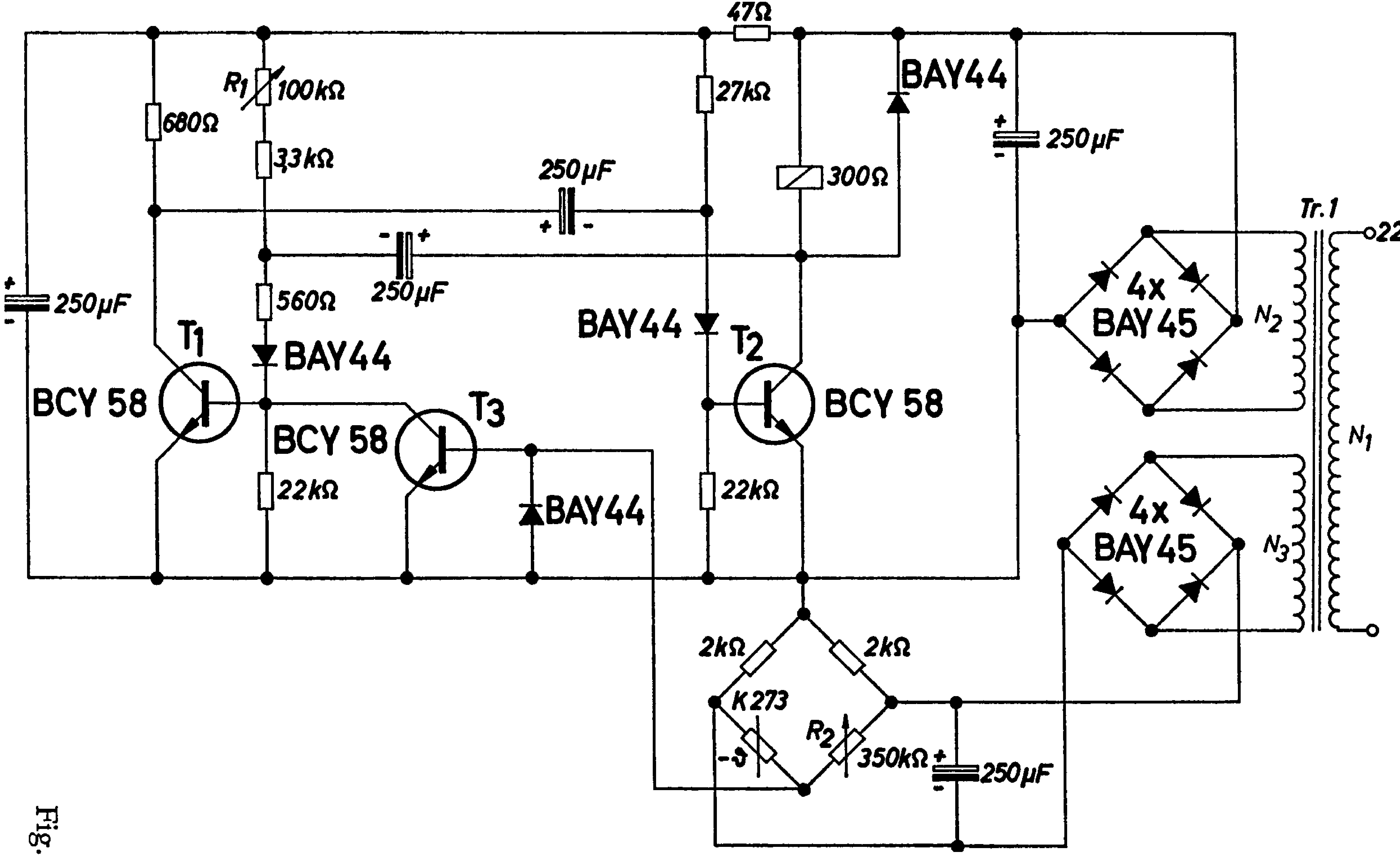

47Ω
R1
100kΩ
680Ω
33kΩ
27kΩ
BAY44
250µF
300Ω
250µF
250µF
250µF
560Ω
BAY44
T1
BAY44
T2
BCY 58
BCY 58
BCY 58
T3
22kΩ
BAY44
22kΩ
Tr.1
220V~
4x
BAY 45
N2
N1
4x
BAY 45
N3
2kΩ
2kΩ
K 273
R2
350kΩ
250µF

Fig. 5.1.

Technical data

Operating voltage	220 V 50	Hz
Adjustable temperature	25 to 95	°C
Pulse time	0.6 to 12	s
Pulse pause	5	s

Transformer Tr 1: EI42/14 Dyn. sheet IV/0.35 alternate stacking
N_1 = 4400 turns 0.06 CuL
Isolation 3×0.05 paper
N_2 = 480 turns 0.16 CuL
N_3 = 300 turns 0.16 CuL

5.2. Temperature Control Circuit with Thyristor-Tetrode

The circuit shown in Fig. 5.2. differs from the circuit usually used for temperature control only in that a thyristor-tetrode has been used in the output. At the input there is a bridge with the NTC thermistor K 22 which can be operated up to a temperature of max. 200 °C. In order to avoid that the remainder of electric power heats up the NTC thermistor too much, it is to be mounted with a thermal resistance $R_{thamb} \leq 300$ °C/W. The amplifier responds to a bridge unbalance of 130 mV. The switch hysteresis loop, that means the difference between switch-on and switch-off of the controller, therefore, amounts to 1—5 °C in the adjustable temperature range of 60—200 °C. The smallest hysteresis loop occurs at the lowest setting temperature.

The thyristor tetrode BRY 20 at the output is operated with an AC voltage of 20 V. During a half-wave the thyristor tetrode is always switched off, during the other half-wave it is switched on if the bridge is unbalanced at the input.

Technical data

Operating voltage for the bridge	20	V
Operating voltage for the switching amplifier	12	V
Operating voltage for the thyristor-tetrode	20 V 50	Hz
Adjustable temperature	60—200	°C
Control accuracy at 60 °C	<1	°C
at 150 °C	<2	°C
at 200 °C	<5	°C

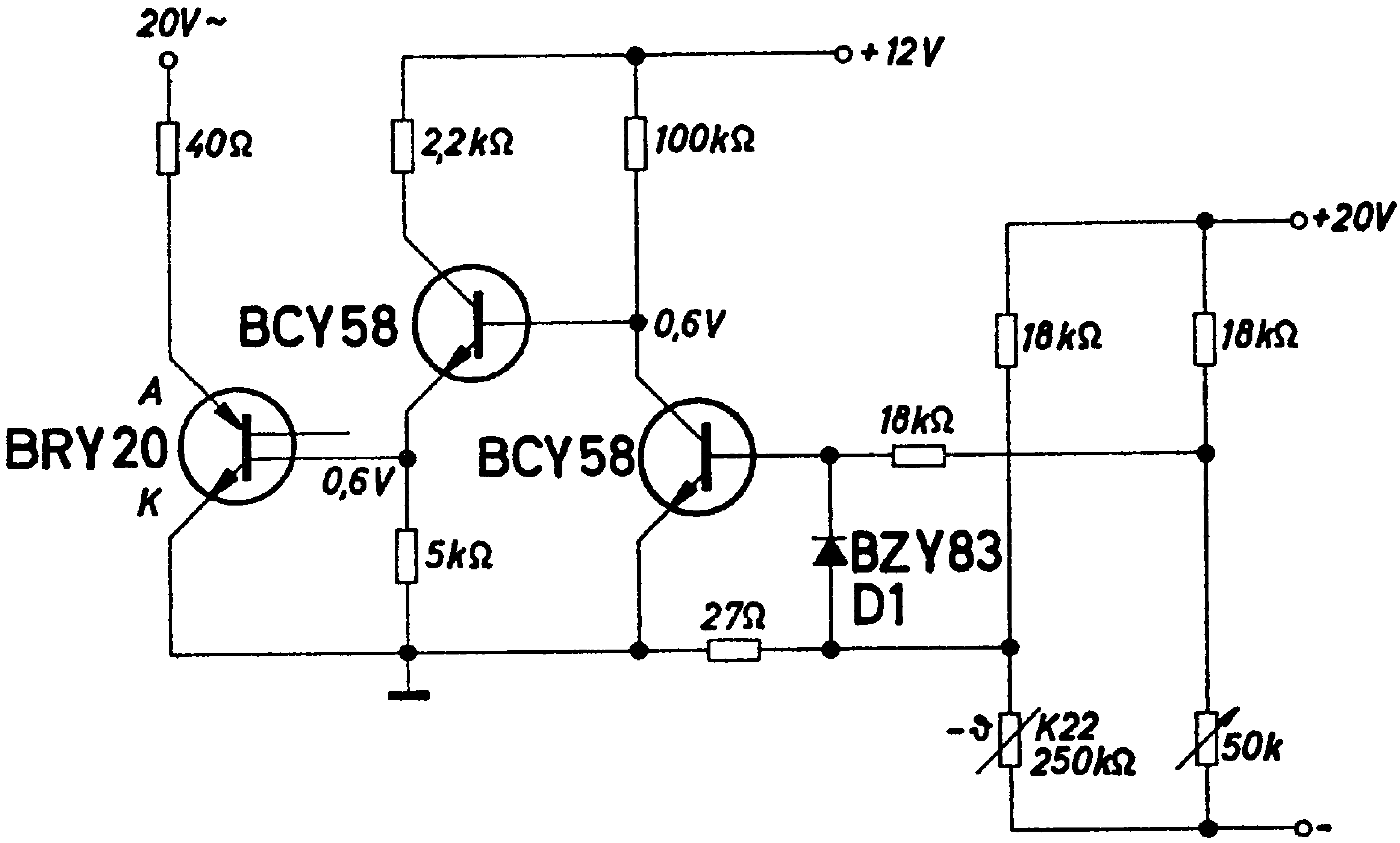


Fig. 5.2.

5.3. Temperature Control with Positive Temperature Coefficient Thermistors

Temperature control circuits are used in many diverse fields, i.e., for controlling the motor temperature, or the temperature in a molding press, etc. Very often a great number of test points has to be controlled. Therefore, the circuit has to be designed in such a way that many test points can be controlled with the lowest possible expenditure. On the other hand it must be easy to find out at which test point the temperature is too high so that the damage can be repaired immediately. The circuit shown in Fig. 5.3. meets all these requirements. As thermometer probes PTC thermistors are used which considerably increase their resistance at a certain temperature (transient characteristic). This characteristic guarantees a high reliability, as, e.g. if a feedline is broken, a high temperature is simulated, causing the alarm to release or turning off the device. An interference at the input caused by a short circuit is a seldom occurrence.

The switching temperature depends mainly on what kind of PTC thermistor is used. Within a small range it can be additionally adjusted by the potentiometers R_1 and R_2. It is recommended to adjust it in such a way that the switching amplifier S_1 connected to the voltage divider via the resistor R_1 starts operating at a lower temperature than the switching amplifier S_2, which is connected to the voltage divider via resistor R_2. Then the switching amplifier S_1 initiates an early warning and the master warning is released only if the temperature keeps rising. The additional amplifiers S_3 and S_4 by which relays are switched are connected to the switching amplifiers mentioned above. At master warning the motor can be switched off, for instance, in the case of a motor control, by the amplifiers S_3 and S_4.

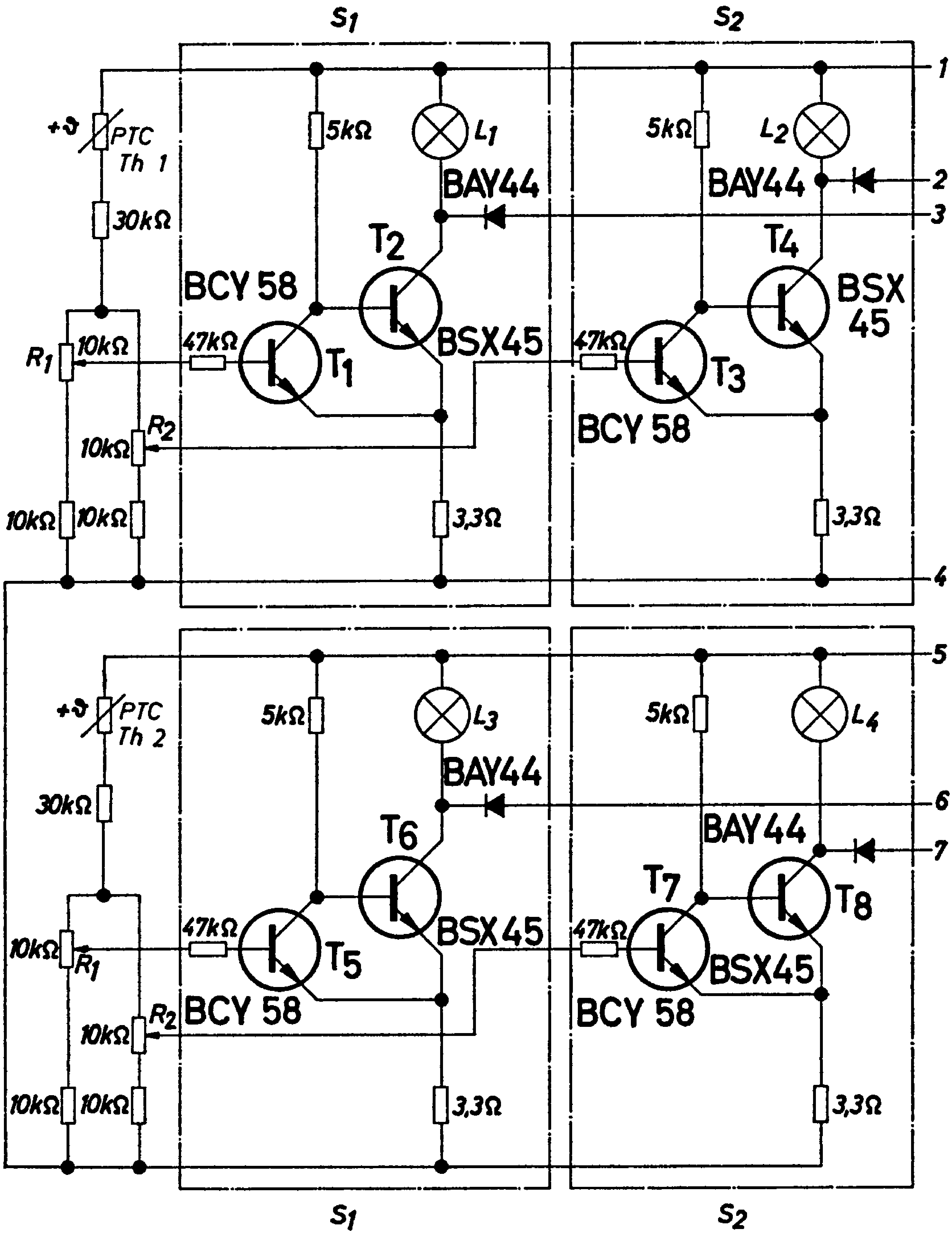
In order to multiply the test points in the simplest way, the voltage dividers are laid out in various arrangements with the PTC thermistor and the switching amplifiers S_1 and S_2. The switching amplifiers S_3 and S_4 have to be installed only once for an arbitrary number of test points. All the outputs of the switching amplifiers S_1 and S_2 are connected to the amplifiers S_3 and S_4.

Locating the failure is easily possible by the bulbs at the outputs of the amplifiers S_1 and S_2, indicating what test point has released the alarm.

While as already mentioned, the switching temperature depends mainly on the kind of PTC thermistor used, because of its steep rise in resistance at a certain temperature, the switching points depend only to a small extent on changes in the operating voltage. Thus it is not necessary to stabilize the operating voltage as required, for instance, for the operation of NTC thermistor sensors in voltage dividers.

Technical data

Operating voltage	12	V
Operating current	11 to 460	mA
Setting temperature, dependent on the applied PTC-thermistor (P310-C11 to P450-C11)	ca. 50 to 220	°C
Ambient temperature	0—60	°C



S1
S2
1
+ϑ PTC Th 1
30kΩ
5kΩ
L1
BAY44
T2
BCY 58
BSX45
47kΩ
T1
R1
10kΩ
R2
10kΩ
10kΩ
10kΩ
3,3Ω
5kΩ
L2
BAY44
2
3
T4
BSX 45
47kΩ
T3
BCY 58
3,3Ω
4
5
+ϑ PTC Th 2
30kΩ
5kΩ
L3
BAY44
6
T6
BSX45
47kΩ
T5
BCY 58
10kΩ
R1
R2
10kΩ
10kΩ
10kΩ
3,3Ω
5kΩ
L4
BAY44
7
T7
T8
47kΩ
BSX45
BCY 58
3,3Ω
S1
S2

Fig. 5.3.

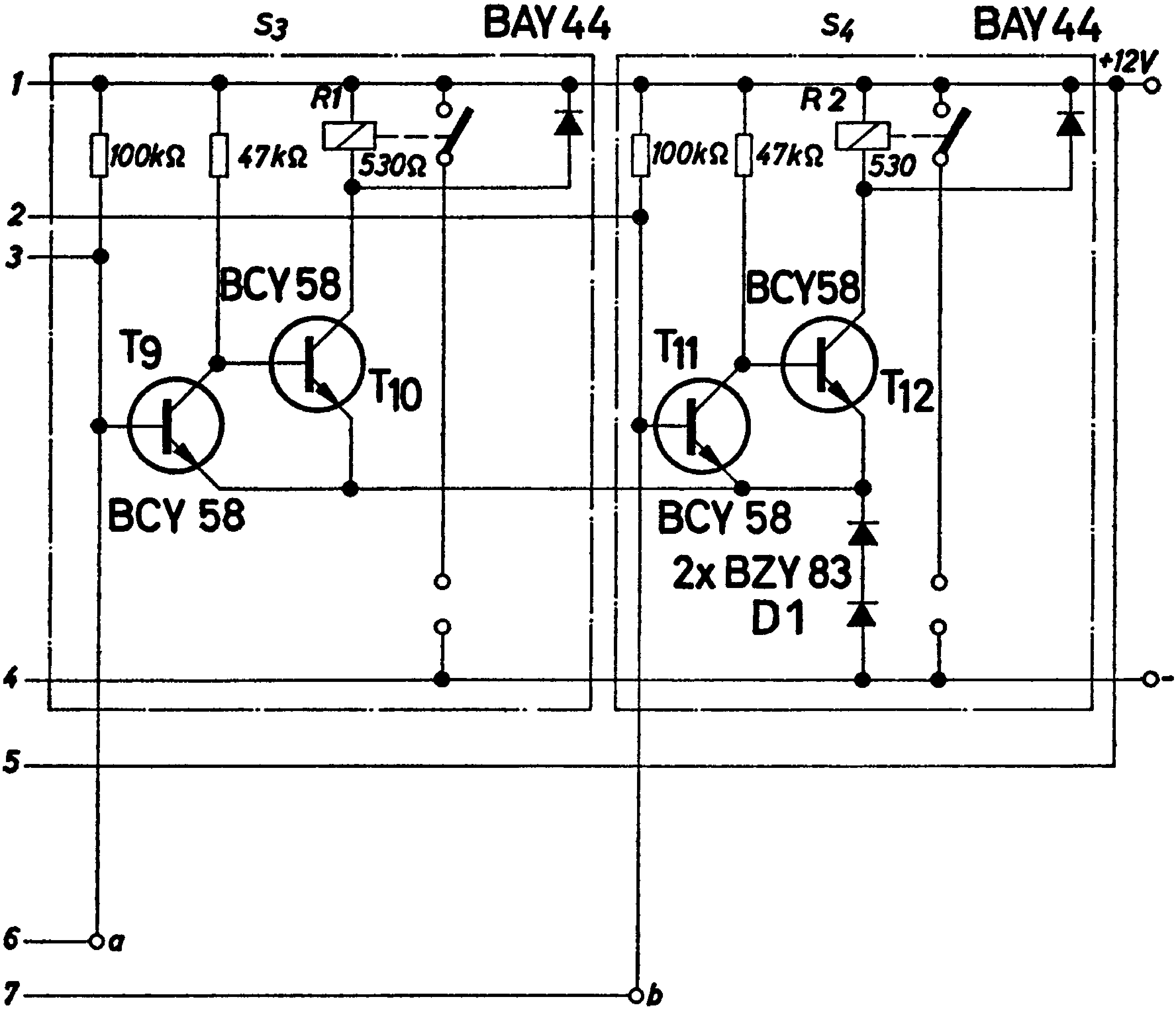

S3
BAY 44
S4
BAY 44
+12V
1
2
3
4
5
6
7
a
b
R1
530Ω
100kΩ
47kΩ
R 2
530
100kΩ
47kΩ
BCY 58
T9
T10
BCY 58
BCY58
T11
T12
BCY 58
2x BZY 83
D1

Fig. 5.3.

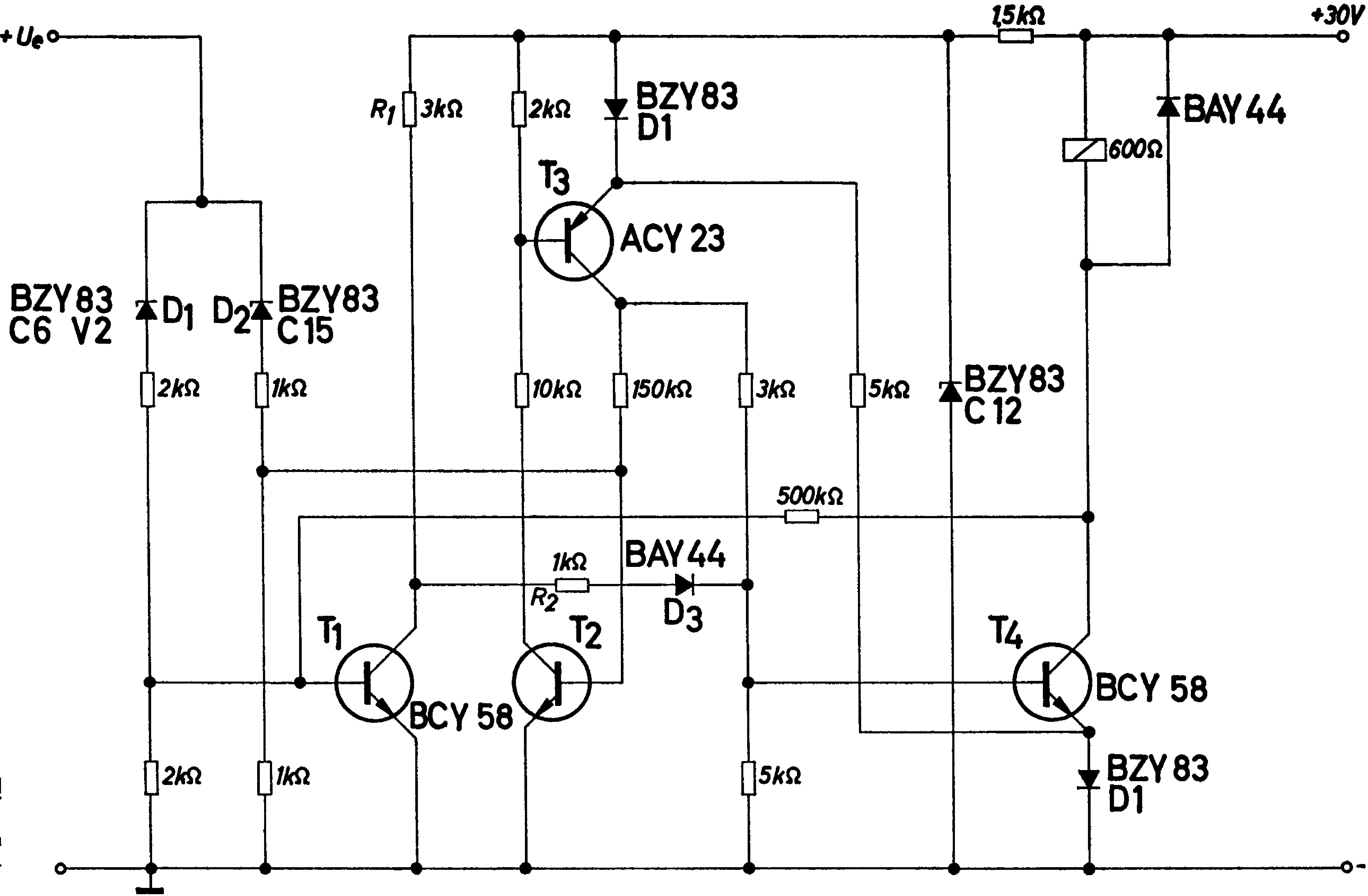

+Ue
1,5kΩ
+30V
R1
3kΩ
2kΩ
BZY83
D1
BAY44
600Ω
T3
ACY 23
BZY83
C6 V2
D1
D2
BZY83
C15
2kΩ
1kΩ
10kΩ
150kΩ
3kΩ
5kΩ
BZY83
C 12
500kΩ
1kΩ
BAY44
R2
D3
T1
T2
T4
BCY 58
BCY 58
2kΩ
1kΩ
5kΩ
BZY83
D1

Fig. 5.4.

5.4. Voltage Setting Switch

The circuit shown in Fig. 5.4. is appropriate for controlling a voltage setting which may vary within a certain range. If it exceeds or falls below this limit, a relay picks up at the output.

If the input voltage V_I drops below the lowest limit value, no current flows via the two Zener diodes D_1 and D_2, and the transistors T_1, T_2 and T_3 are switched off. Positive potential is applied to the input of transistor T_4 via the resistors R_1, R_2 and the diode D_3. Thus, transistor T_4 becomes conductive, and the relay picks up.

As soon as the input voltage has risen to the lower limit value, the Zener diode D_1 becomes conductive, and transistor T_1 is switched on. Hereby diode D_3 becomes backward biased and the base current of transistor T_4 is cut off. The transistor T_4 is switched off, and the relay drops out. When the voltage at the input keeps increasing and reaches the upper limit value, a current flows via the Zener diode D_2, and the transistors T_2 and T_3 become conductive. The transistor T_3 receives the necessary driving voltage via the transistor T_2. The transistor T_4 is switched on again via transistor T_3 and the relay picks up.

Consequently the relay is not energized as long as the voltage V_I is within a certain range. As soon as the voltage drops below the lower limit or exceeds the upper limit, the relay picks up.

The voltage limits can be made adjustable by a slight change of the input circuit. The input voltage V_I has to be applied to the Zener diodes D_1 and D_2 via voltage dividers.

In the input circuit shown in Fig. 5.4. the voltage limits depend on the Zener voltage of the diodes used which spreads within a certain range.

The temperature influences the voltage limits only to a small extent. Within a range of 25 to 60 °C, it only changes by a few percent.

Technical data

Operating voltage	30	V
Operating current (relay dropped out)	ca. 15	mA
Maximum input voltage V_I	24	V
Maximum ambient temperature	60	°C
Lower voltage limit	ca. 6.2	V
Upper voltage limit	ca. 15	V

(both dependent on the actual Zener voltage of the diodes D_1 or D_2)

5.5. DC Voltage Measuring Amplifier with Transistor Chopper

The input sensitivity of DC amplifiers with transistors is limited above all by the temperature dependent transistor parameters. In DC amplifiers of especially high sensitivity one takes advantage of the fact that AC amplifiers can be built for much higher sensitivity because they are limited only by transistor noise. The DC signal is then chopped and the resulting AC signal amplified.

The circuit indicated in Fig. 5.5. shows such an amplifier. The chopper with the transistor T_5 is driven by an astable multivibrator (transistors T_1 and T_2) via an amplifier stage with the transistor T_3. The amplifier stage has to increase the steepness of the edges of the square wave pulses generated by the multivibrator. The driving current for the chopper transistor T_5 is kept constant by transistor T_4 which is arranged as a current source.

The chopper transistor is inversely operated as transistors at this kind of operation do have the lowest saturation voltage. The constant base current has been set to a very high value. It is 300 μA at a collector current of transistor T_5 of 1 to 5 μA only. This current results from the input signal of 100 to 500 μV and the input resistance of 100 Ω.

As the saturation voltage at the transistor even at inverse operation and at full drive beyond saturation is about 500 μV, it had to be compensated by special designs. This is achieved by the voltage drop at resistor R_1. The temperature compensation is accomplished by a combination of a NTC thermistor K11 and a resistor R_2.

The driving frequency of the chopper amounts to 5 kHz. If the frequency were too high, disturbing pulses are superimposed on the desired signal when the chopper transistor is switched on.

The square-wave AC voltage generated in the chopper is amplified by a three-stage AC amplifier. At the output of this amplifier the signal is rectified and led to a high impedance measuring instrument. In order to minimize the influence of the rectifier diodes upon the signal magnitude, the output signal is transformed.

The gain of the AC amplifier is set at the potentiometer R_4. The desired deflection at the measuring instrument can be set by applying the calibration voltage to the input. The driving current for the chopper transistor is adjusted by potentiometer R_3. One chooses a value which results in the lowest interference voltage at the output.

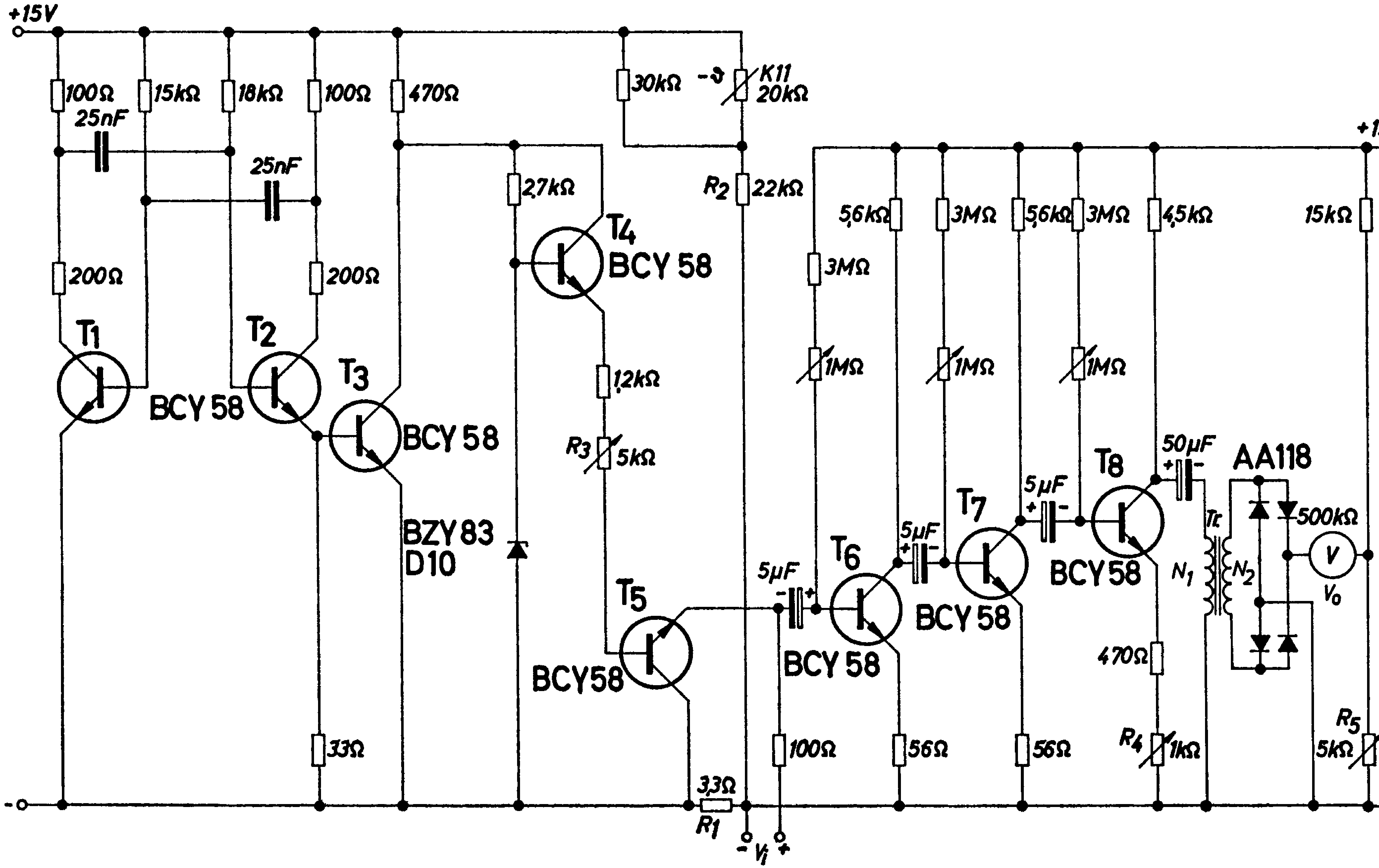

+15V
100Ω
15kΩ
18kΩ
100Ω
470Ω
30kΩ
K11
20kΩ
25nF
25nF
200Ω
200Ω
2,7kΩ
R2
22kΩ
+15V
5,6kΩ
3MΩ
5,6kΩ
3MΩ
4,5kΩ
15kΩ
3MΩ
T4
BCY 58
T1
T2
T3
BCY 58
BCY 58
1MΩ
1MΩ
1MΩ
1,2kΩ
R3
5kΩ
50μF
AA118
T8
T7
5μF
5μF
5μF
500kΩ
BZY 83
D10
T6
T5
Tr
N1
N2
V
V0
BCY 58
BCY 58
BCY 58
BCY58
BCY 58
470Ω
33Ω
100Ω
56Ω
56Ω
R4
1kΩ
R5
5kΩ
3,3Ω
R1
Vi

Fig. 5.5.

In a test circuit the base current had the value of 300 μA as already mentioned. The remainder of the interference voltage can be compensated at the output by a reverse voltage (resistor R_5). The diagram shown in Fig. 5.6. explains the relation between the input and output voltage with and without compensation of the interference voltage at the output. It can be seen that by compensation input voltages even below 100 μV can be amplified with a relatively small inaccuracy.

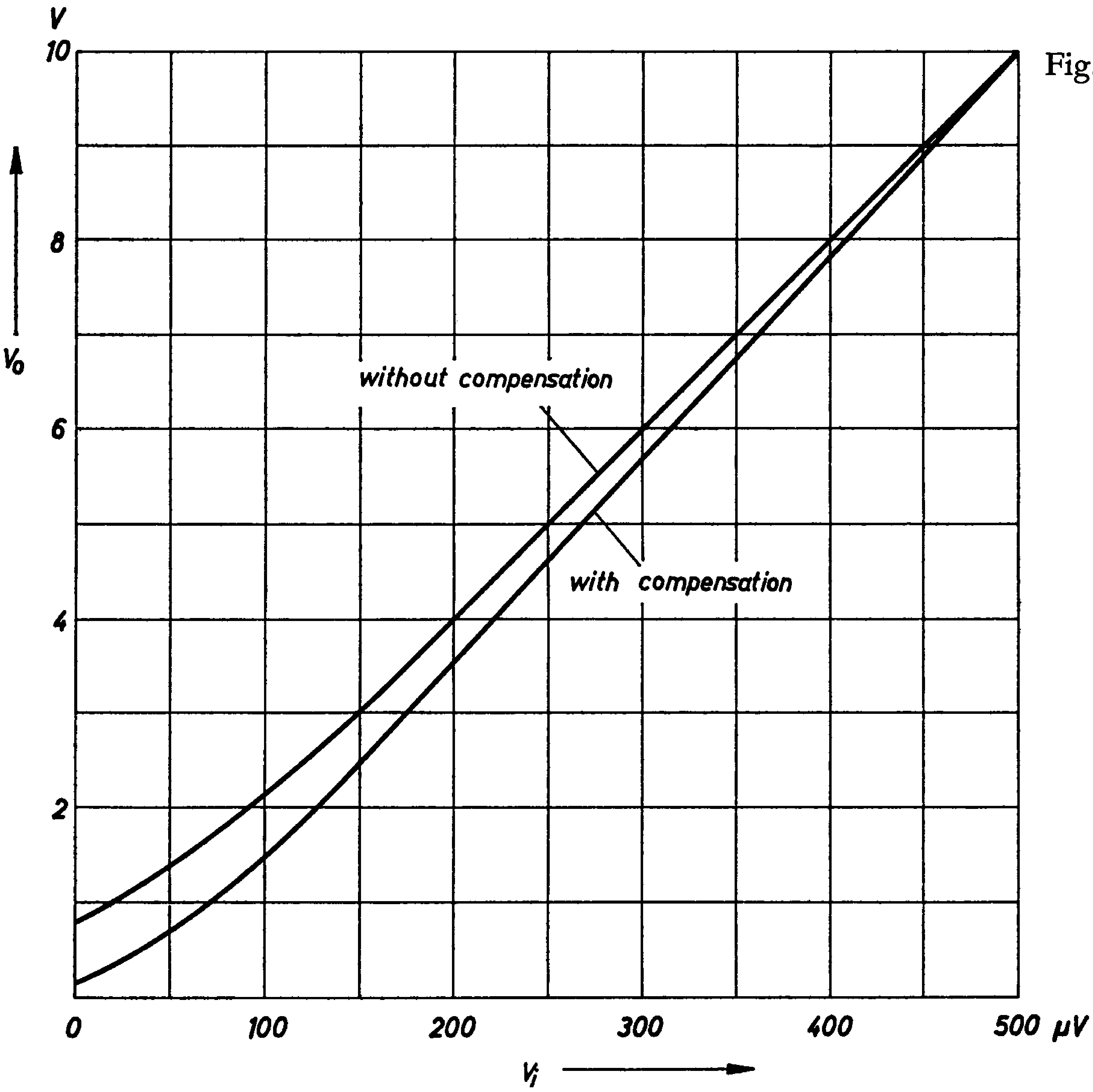


Fig. 5.6.

Technical data

Operating voltage (for chopper and amplifier separated)	2×15	V
Input sensitivity	100 to 500	µV
Output voltage	10	V

Transformer T: Siferrite pot-cores B65571-A0000-R026
N_1 = 267 turns 0.09 CuL
N_2 = 1330 turns 0.09 CuL

5.6. Direction Control of Rotation for Small Type Motors

The circuit shown in Fig. 5.7. makes it possible to change the direction of rotation of a small-type motor by using only one operating-voltage source. It can be used for motors with an operating current up to 100 mA.

The circuit consists of a bridge with 4 transistors. By applying negative pulses to the input A or the input B either the transistors T_1 and T_3 or the transistors T_2 and T_4 are switched on.

If, e.g., a negative pulse of min. 1.5 mA reaches the base of transistor T_1, it is switched on, and the potential at its collector becomes positive. Hereby transistor T_4 is switched off because the potential at its emitter is increased by the voltage drop at resistor R_1. The potential at the collector of transistor T_2 changes towards negative values. Because of the coupling via the resistor of 10 Ω a blocking signal is applied to the base of transistor T_4. In a similar way, a positive signal gets to the base of transistor T_3 from the collector of transistor T_1, so that the npn transistor T_3 becomes conductive. As a result of this coupling it is possible to switch the transistors very fast.

If a negative signal is applied to input B, the transistors T_2 and T_4 are switched on in the same way already described. Hereby the current direction is changed via the load resistor R_1, i.e., consequently via the motor. Thus the direction of rotation is inversed.

Technical data

Operating voltage	12	V
Operating current	100	mA
Driving current	>1.5	mA
Switching time	ca. 25	µs

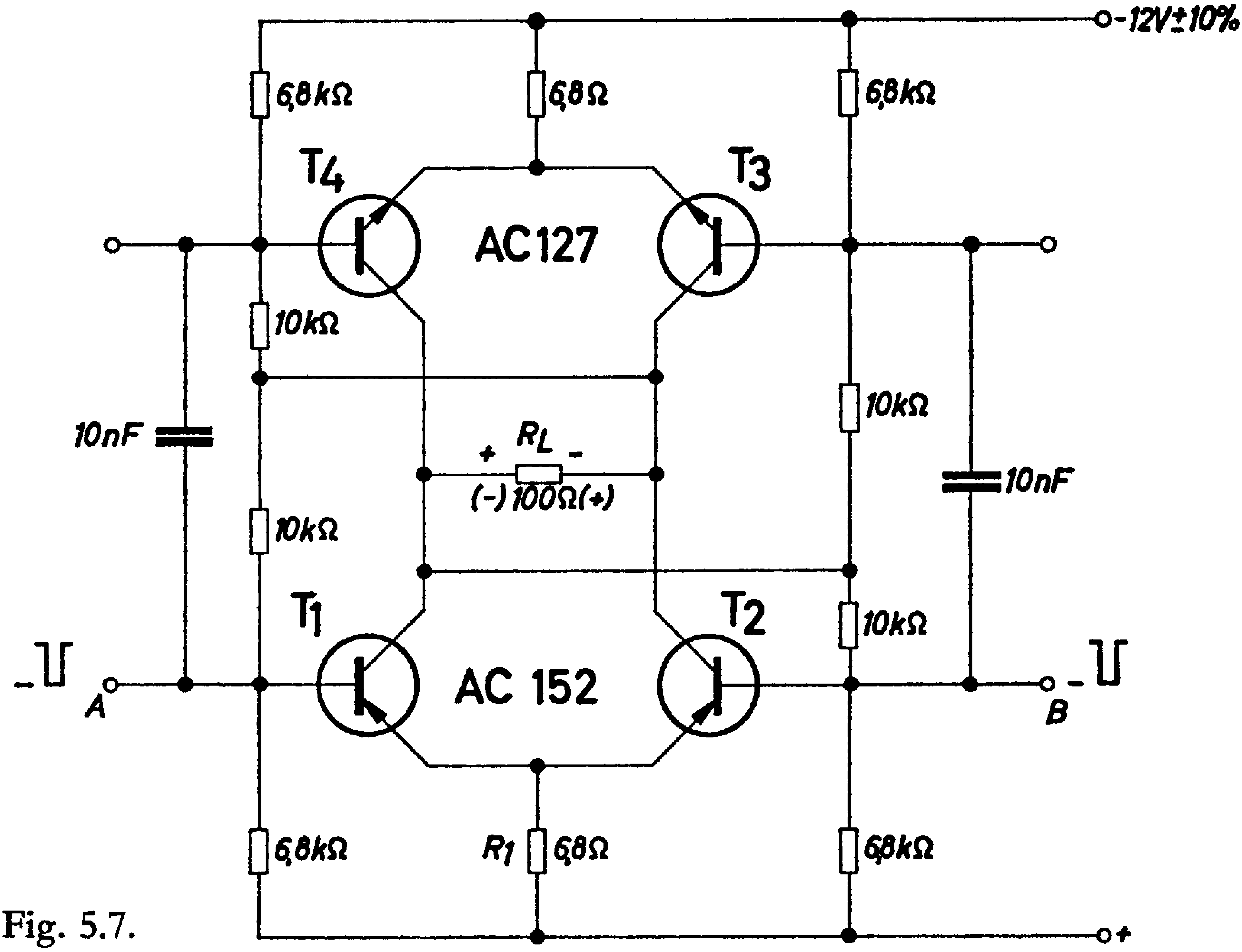
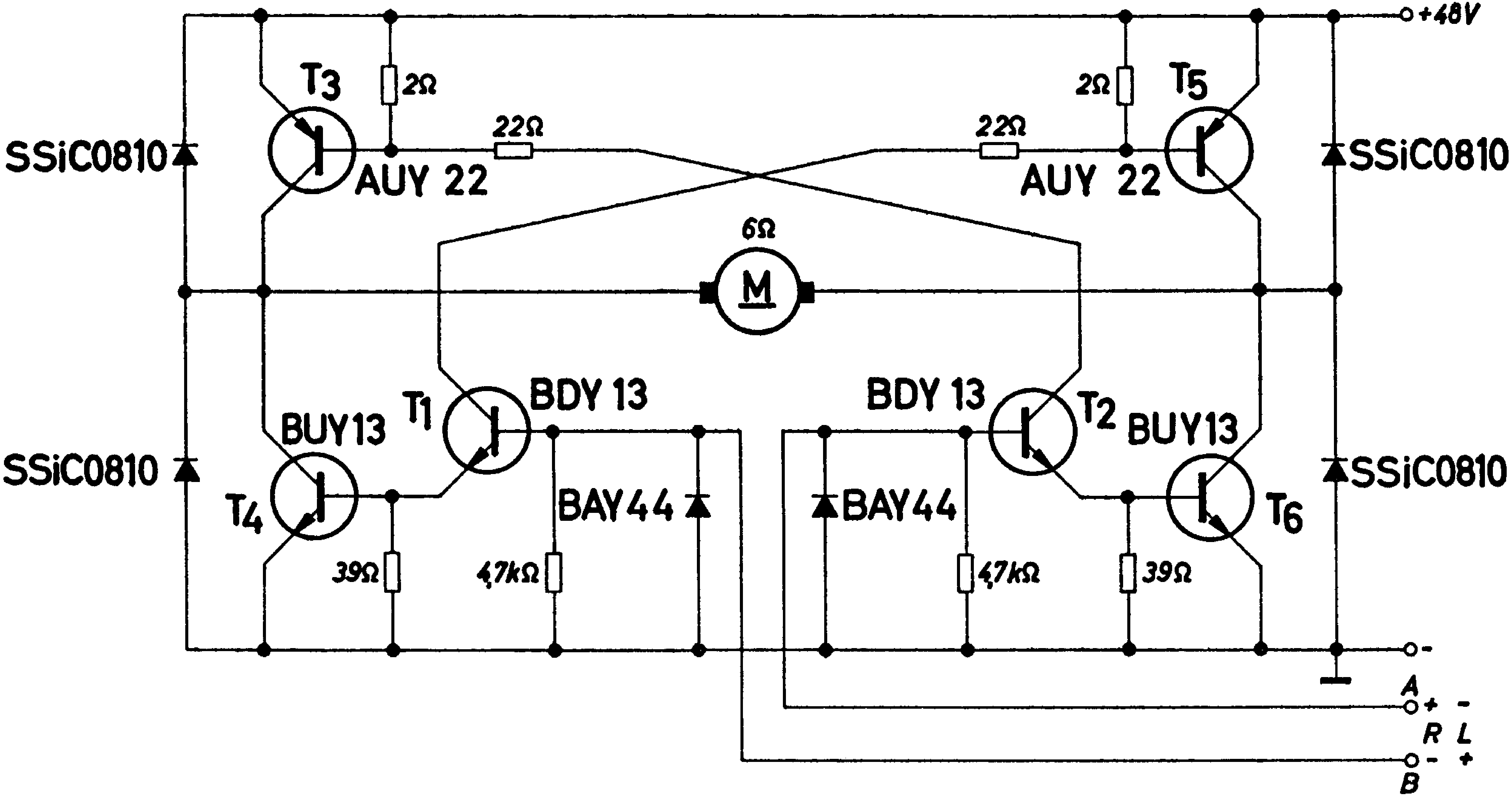

-12V±10%
6,8kΩ
6,8Ω
6,8kΩ
T4
T3
AC127
10kΩ
10kΩ
10nF
+
RL
-
(-) 100Ω(+)
10nF
10kΩ
T1
T2
10kΩ
AC 152
A
B
6,8kΩ
R1
6,8Ω
6,8kΩ
+

Fig. 5.7.



+48V
T3
2Ω
22Ω
AUY 22
SSiC0810
T5
2Ω
22Ω
AUY 22
SSiC0810
6Ω
M
T1
BDY 13
BUY13
T4
SSiC0810
39Ω
4,7kΩ
BAY44
BDY 13
T2
BUY13
T6
SSiC0810
BAY44
4,7kΩ
39Ω
-
A
+ -
R L
- +
B

Fig. 5.8.

5.7. Switching of a Power Motor

Fig. 5.8. shows a circuit for changing the direction of rotation in a motor of medium power. As in the example described above the motor again is located in a diagonal of a bridge circuit consisting of 4 transistors. Always two diagonally opposite transistors of the bridge are driven by one pre-stage. Dependent on the polarity of the input voltage at the poles A and B one of the two pre-stage transistors becomes conductive. If, e.g., pole A is positive, transistor T_2 is switched on. Then transistors T_3 and T_6 are switched on by T_2 and a current flows from the positive pole of the voltage supply of 48 V via the transistor T_3, the motor and transistor T_6. As soon as the driving voltage changes its polarity at the input, transistor T_3 is switched on and the motor obtains a current in reverse direction via the transistors T_5 and T_4. This changes the direction of rotation. The silicon diodes are connected in parallel to the power transistors in order to draw off the inductive energy resulting from switching the motor.

The power transistors are to be mounted on heat sinks, the thermal resistance of which should be $R_{th} \leqq 2.5$ °C/W for each transistor.

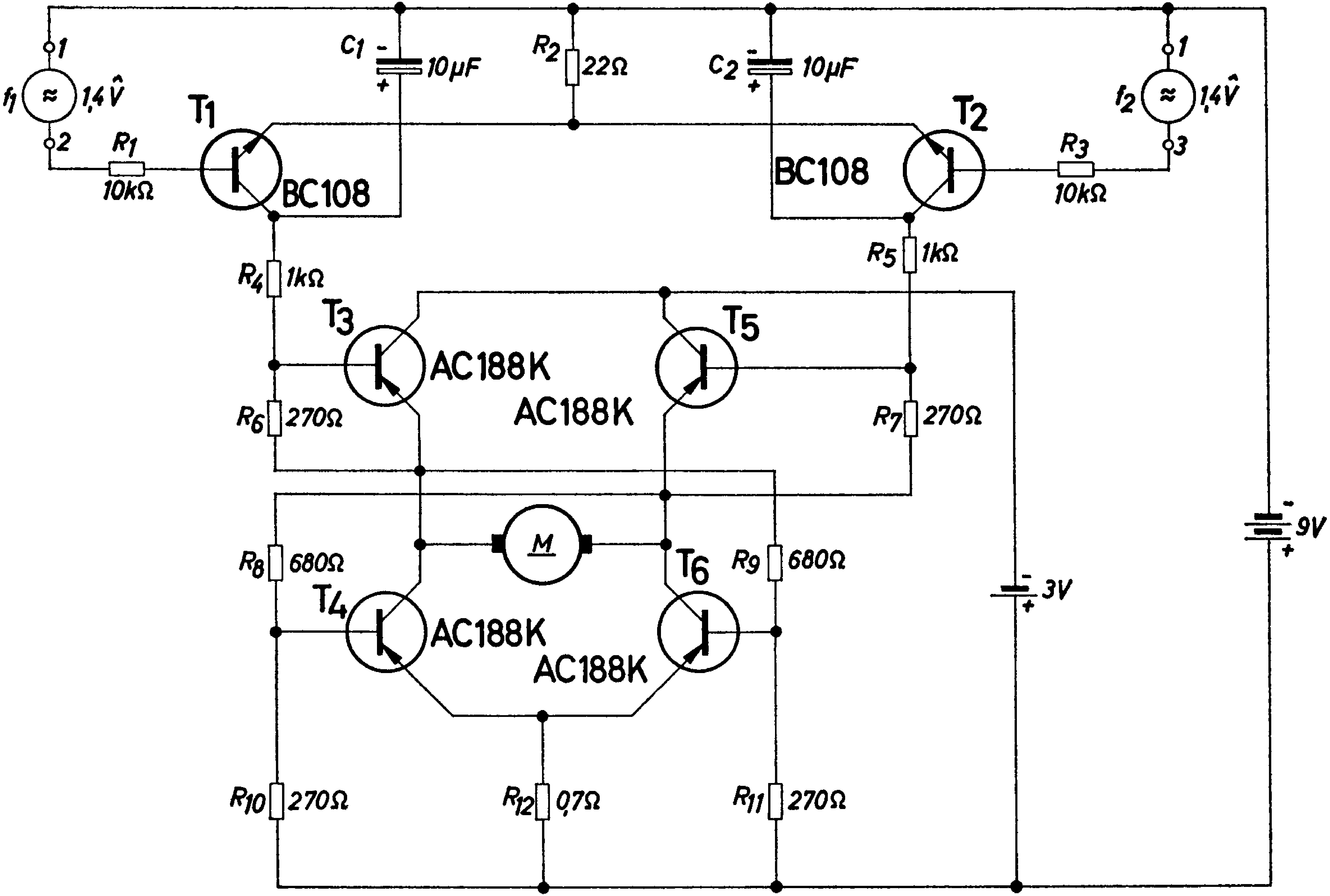
Technical data

Operating voltage	48 + 15 %	V
Maximum operating current	8	A
Minimum load resistor	6	Ω
Driving voltage	3	V
Driving current	60	mA
Maximum ambient temperature	60	°C

5.8. Switching of Small Type Motors with LF Signals

For simple remote control circuits LF-voltages are often used for signalling. The circuit indicated in Fig. 5.9. shows how the direction of rotation of small type motors can be reversed by such LF signals. As in the example mentioned above the motor is again arranged in a bridge circuit consisting of 4 transistors. The bridge is driven by two pre-stages. Low-frequency signals with the frequency f_1 and f_2 serve as driving voltage. The base-to-emitter diode of the pre-stage transistors rectify the AC signal and the capacitors C_1 and C_2 which are located at the collector filter this DC voltage. The base resistors R_1

Fig. 5.9.



T1
BC108
T2
BC108
T3
AC188K
T4
AC188K
T5
AC188K
T6
AC188K
f1
f2
1,4 V̂
C1 10µF
C2 10µF
R1 10kΩ
R2 22Ω
R3 10kΩ
R4 1kΩ
R5 1kΩ
R6 270Ω
R7 270Ω
R8 680Ω
R9 680Ω
R10 270Ω
R11 270Ω
R12 0,7Ω
M
3V
9V

and R_3 limit the magnitude of the input signal and increase the angle of current flow of the rectifier. Due to the emitter resistor R_2, which both pre-stages have in common, the second pre-stage transistor is switched off when the first one is conductive.

If transistor T_1, for example, is switched on by an input signal of the frequency f_1, the transistors T_3 and T_6 are provided with a well filtered DC driving signal via the resistors R_4, R_6, R_9 and R_{11}. The motor rotates in a certain direction.

If a driving signal of the frequency f_2 is applied to the input of transistor T_2, the transistors T_4 and T_5 are switched on and the current through the motor flows in reverse direction. This changes the direction of rotation. The advantages of this circuit are the high input resistance (approx. 15 kΩ) and the high total efficiency resulting from the fact that no power is consumed if there are no input signals.

Technical data

Operating voltage	9 and 3	V
Maximum motor current	1	A
Minimum driving voltage	1.4	A
Input resistance	15	kΩ
Maximum ambient temperature	70	°C
Input frequencies	0.5 to 50	kHz

5.9. Indicator for Resistance Variations

The circuit shown in Fig. 5.10. responses to a slight change of the resistor between two sensor pins at the input. The circuit, therefore, may be used in order to measure a liquid level or to indicate humidity in porous materials.

Depending on the current gain of the input transistors used, a current of 10 μA is sufficient between the two electrodes of the sensor for the response of the circuit.

Technical data

Operating voltage	20	V
Input sensitivity	ca. 10	μA

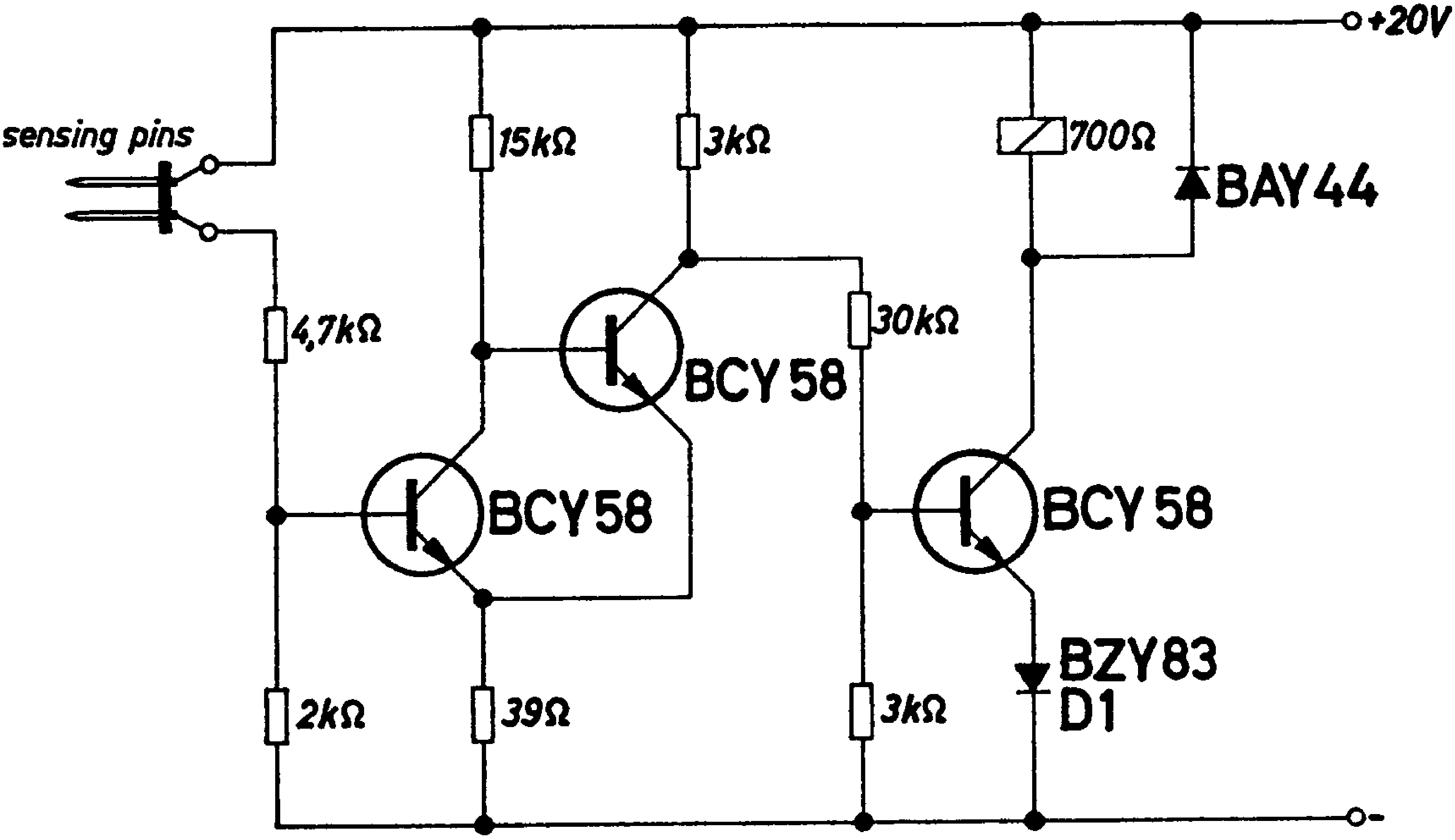


Fig. 5.10.

5.10. Highly Sensitive Bridge Amplifier

As already described in several examples, a considerably higher input sensitivity can be obtained with AC transistor amplifiers than with DC amplifiers. The sensitivity of bridge amplifiers, therefore, can be substantially increased by using an AC voltage instead of a DC voltage for the supply of the bridge. For the design of the amplifier one has to take into consideration that at unbalance an AC bridge delivers an output signal in which only the phase position of the signal voltage indicates the sense of the bridge unbalance.

In the example shown in Fig. 5.11. this problem is solved by supplying the bridge with a square-wave AC voltage with unequal pulse-pause time. The direction of the bridge unbalance is indicated in the driving signal for the amplifier in such a way that the shorter square-wave pulses are in one case positive, in the other case negative. This driving signal is amplified in two stages and is applied to the demodulator stages via a transformer. Each of the two demodulator stages has a push-pull coil. The voltages appearing in these push-pull coils are separately rectified and filtered. Because of the small load at the output of the demodulator circuit a peak value rectification is

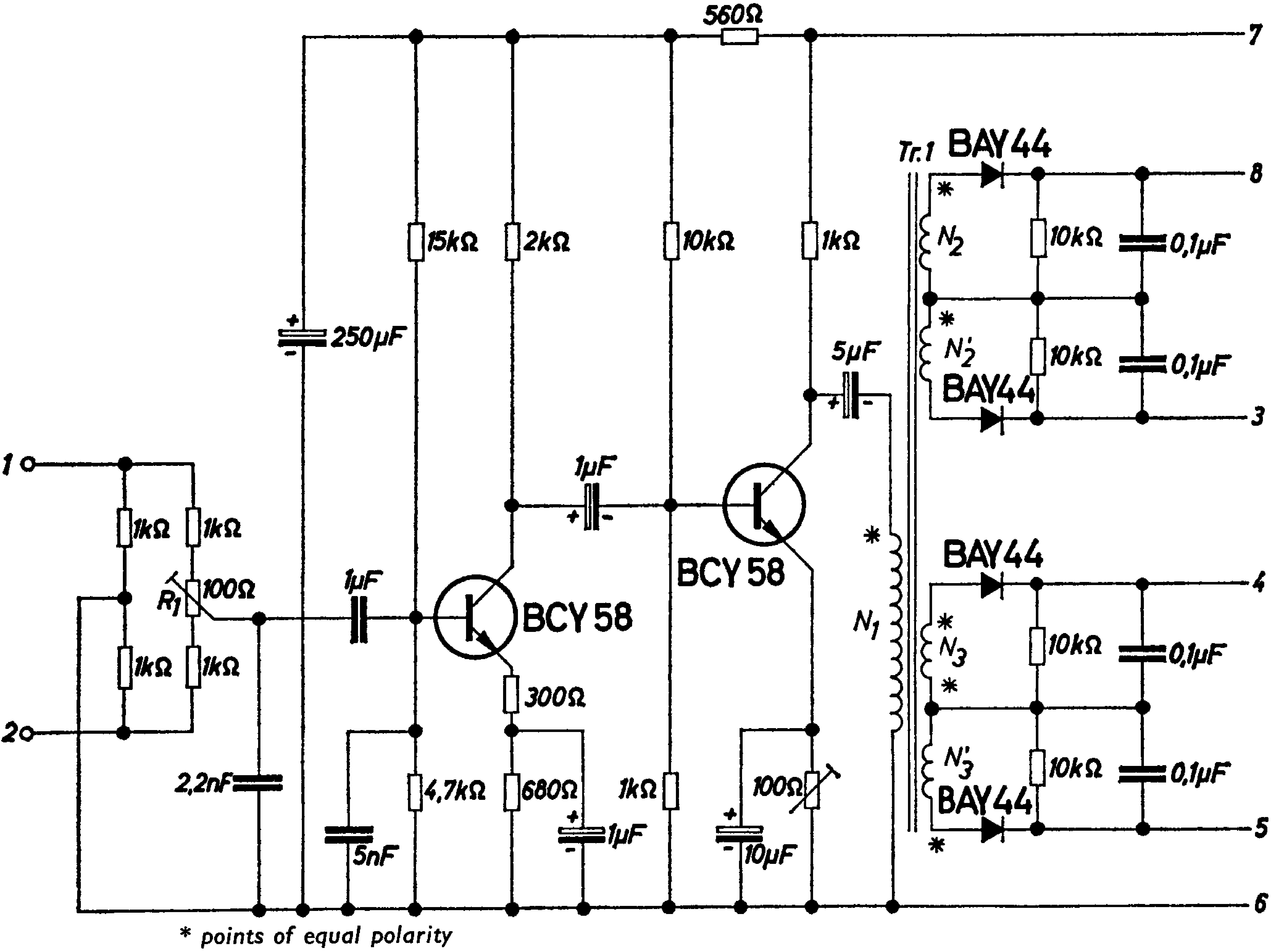


Fig. 5.11.

obtained, that means, the capacitors at the output are charged up to the respective peak values of the voltages in the push-pull coils. Since the capacitors are connected in series, one obtains as output voltage the difference between both the voltage peak values. If the difference is of positive polarity, the connected switching amplifier responds and the relay picks up.

Both the demodulator circuits are laid out in such a way that they operate in phase-opposition (see indicated polarity of the secondary transformer winding). This assures that the response of a certain switching amplifier corresponds to a respective sense of unbalancing. The sensitivity of the switching amplifier will be increased by a

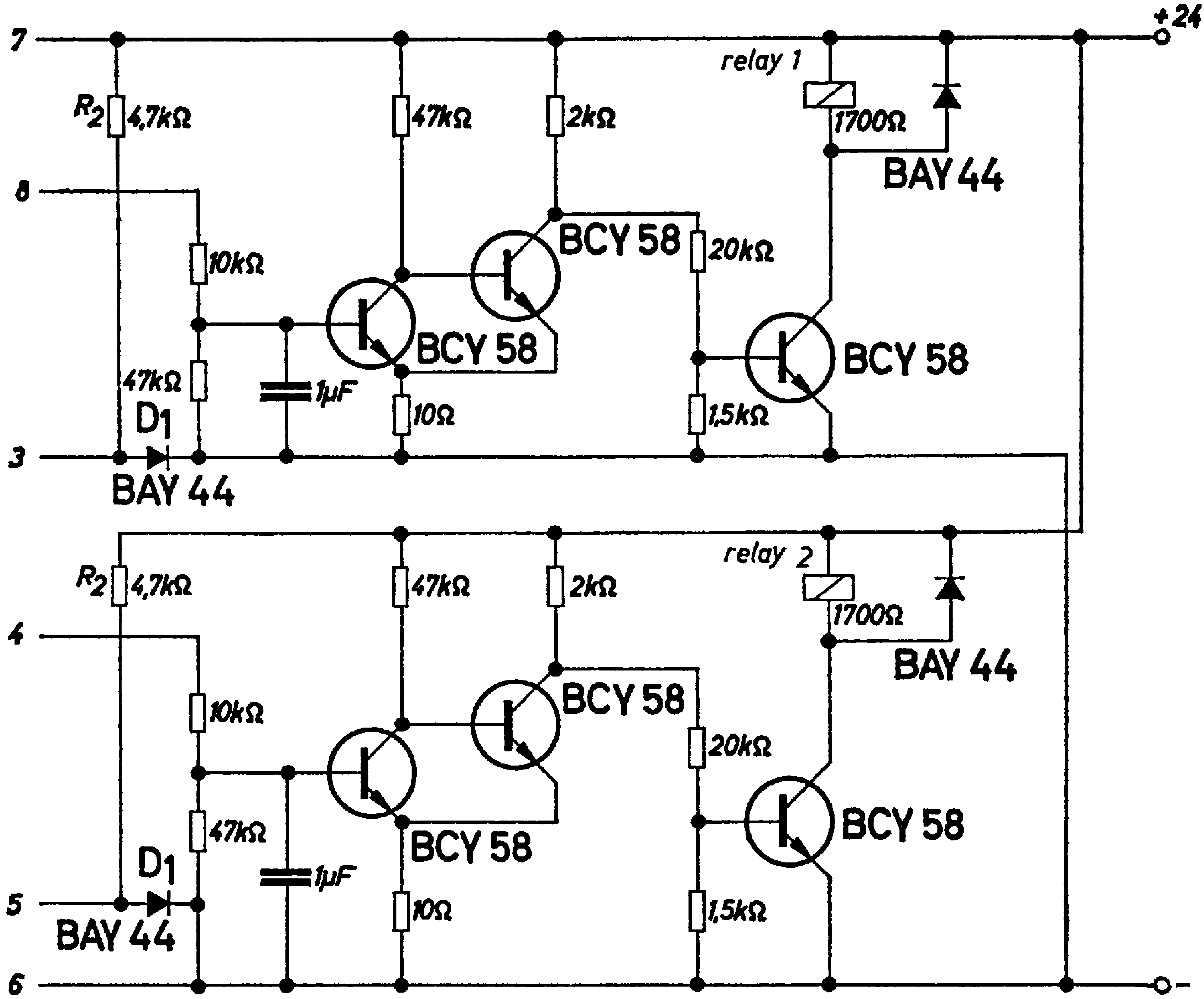


Fig. 5.11.

bias voltage of approx. 0.6 V via the resistor R_2 and the forward biased silicon diode D_1. By that means the threshold voltage of the input transistor is compensated. In the circuit of Fig. 5.11. the relay at the output picks up at an unbalancing of the bridge of only 0.15%.

Fig. 5.12. shows a circuit of the square-wave generator for the generation of the bridge supply voltage. The unsymmetrical square-wave voltage has an oscillation frequency of 8 kHz. The coupling of the bridge to the pulser is achieved by a transformer, so that the same supply voltage can be used both for the amplifier and the square-wave generator.

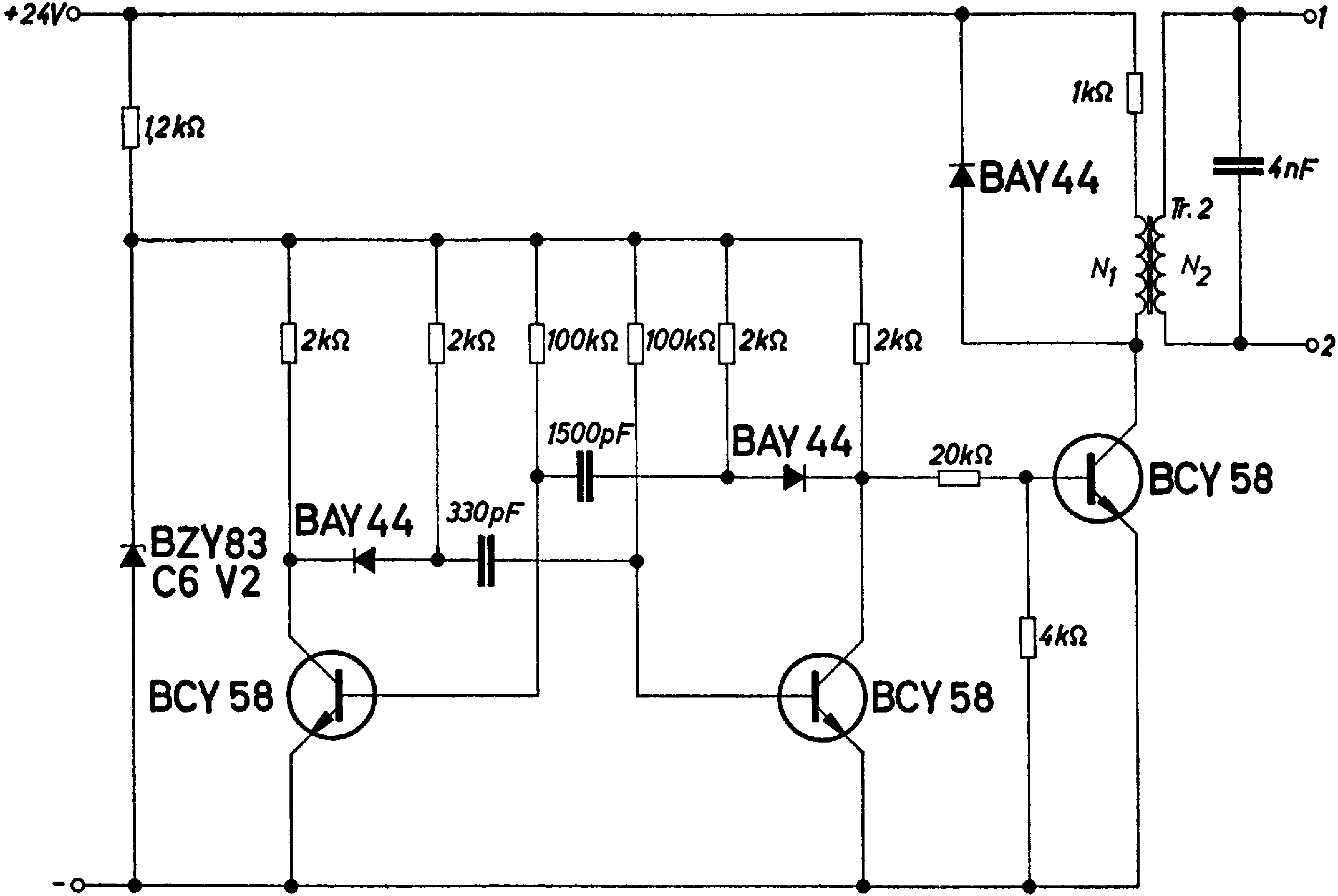

+24V
1,2kΩ
BZY83
C6 V2
2kΩ
2kΩ
100kΩ
100kΩ
2kΩ
2kΩ
1500pF
330pF
BAY 44
BAY 44
BAY 44
BCY 58
BCY 58
BCY 58
20kΩ
4kΩ
1kΩ
4nF
Tr. 2
N_1
N_2
1
2
-

Fig. 5.12.

Technical data

Operating voltage	24	V
Required bridge unbalancing for the relay pick-up	0.15	%
Frequency of the bridge voltage	8	kHz

Transformer

Tr 1 (Fig. 5.11.) Siferrite pot core B65571-A0000-R026

$N_1 = 220$ turns 0.12 CuL

$N_2 = N'_2 = N_3 = N'_3 = 200$ turns 0.08 CuL

Tr 2 (Fig. 5.12.) Siferrite pot core B65561-A0000-R022

$N_1 = 190$ turns 0.15 CuL

$N_2 = 100$ turns 0.18 CuL

5.11. AC Bridge Amplifier

In the design example of Fig. 5.13. the problem of the preceding chapter is solved in another way. The supply of the bridge is a sine wave voltage of 160 kHz. The bridge itself consists of 2 resistors and 2 capacitors, one of which is variable. For a capacitance variation of this capacitor by a value of $\pm\Delta C$ to the input of the succeeding amplifier an AC signal is applied which has a phase shift of 0 or 180° with reference to the bridge voltage, dependent on the sense of the capacitance variation. Via an impedance converter the signal is applied to a phase bridge by which the phase shift in the whole amplifier is compensated. The compensated signal is applied to the main amplifier via a further stage in common collector configuration, i. e., an impedance converter. A resonant circuit in series filters harmonics deriving from non-linearities in the amplifier and in the rectifier. Via an impedance converter at the output of the main amplifier the signal is applied to the phase-selective rectifier circuit. This impedance converter is required that the low input impedance of this rectifier stage does not too much load the amplifier. The impedance converter succeeding the balancing bridge and the phase bridge do have the same pupose. The high input impedance of these stages warrants an open-circuit operation of the two bridges.

In the rectifier circuit the phase position of the supply voltage for the balancing bridge is compared with the phase position of the amplified bridge signal. In both the branches of the rectifier circuits two

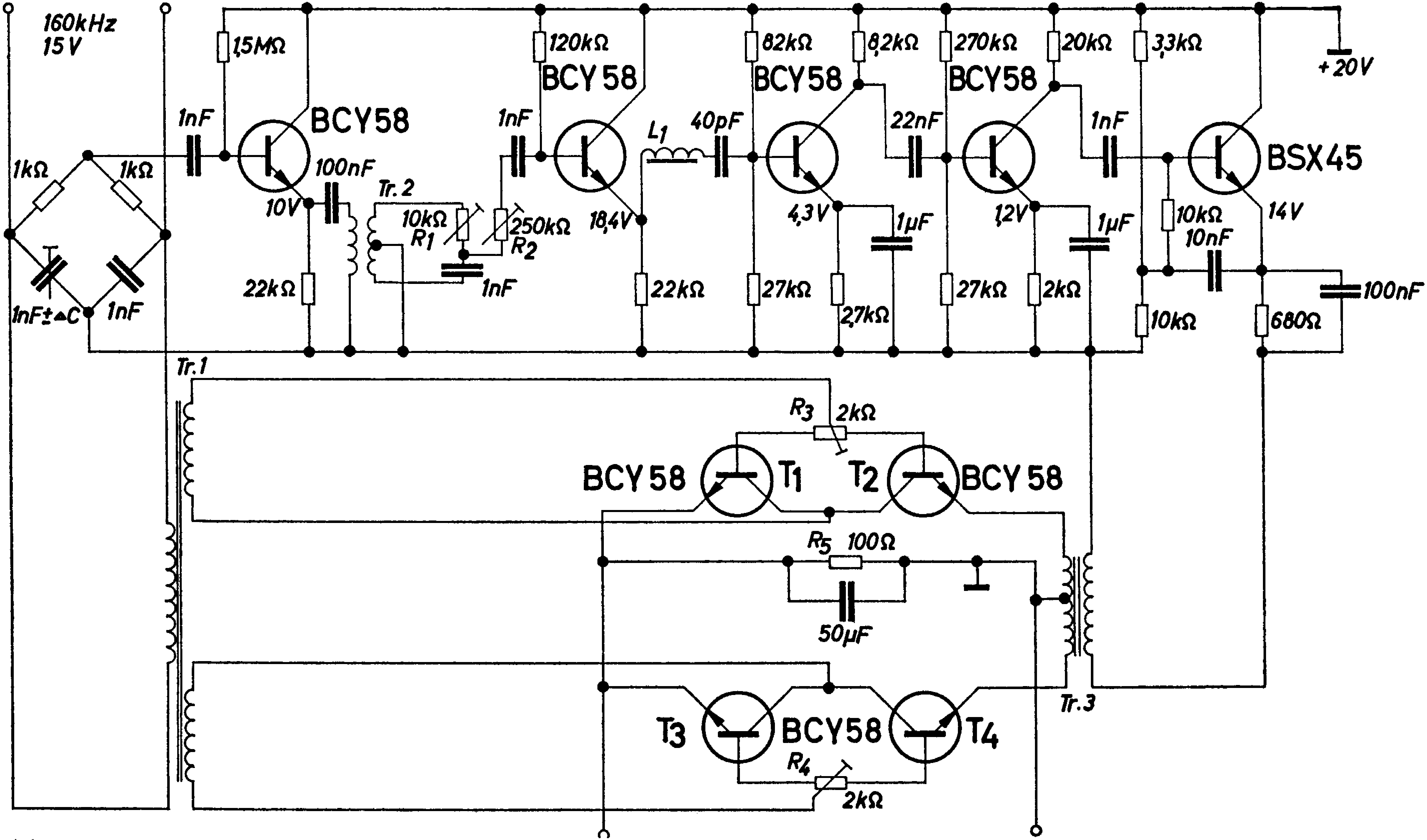

160kHz
15V
1kΩ
1kΩ
1nF±ΔC
1nF
1nF
1,5MΩ
BCY58
10V
100nF
22kΩ
Tr. 2
10kΩ
R1
250kΩ
R2
1nF
1nF
120kΩ
BCY 58
18,4V
22kΩ
L1
40pF
82kΩ
8,2kΩ
BCY58
4,3V
1µF
27kΩ
2,7kΩ
22nF
270kΩ
20kΩ
BCY58
1,2V
1µF
27kΩ
2kΩ
1nF
3,3kΩ
BSX45
10kΩ
10nF
14V
10kΩ
680Ω
100nF
+20V
Tr.1
R3
2kΩ
BCY 58
T1
T2
BCY 58
R5
100Ω
50µF
Tr.3
T3
BCY58
T4
R4
2kΩ

Fig. 5.13.

transistors are used, one of which is always operated normally, the other one inversely. By that means a good symmetry of the signal for both directions of the current is attained.

If the supply voltage has, for instance, such an instantaneous value that a positive signal is applied to the transistors T_1 and T_2 via the resistor R_3, the transistors are conductive. Due to the push-pull winding of transformer Tr 1 in the same instant the transistors T_3 and T_4 are switched off by a negative signal. The polarity at the secondary winding of the transformer Tr 3 determines in which direction the current flows via the transistors T_1 and T_2 and the output resistor R_5. During the next half-cycle the transistors T_3 and T_4 become conductive. Since in the same time interval also the polarity of the voltage at transformer Tr 3 has changed, it is guaranteed that also during the second half-cycle the current in resistor R_5 flows in the same direction. A reversal of the direction of current only occurs if there is a phase shift of 180° between the bridge supply voltage and the bridge signal, i.e., between the voltage at transformer Tr 1 and transformer Tr 3. Thus the sense of current via resistor R_5 indicates the sense of the bridge unbalance and the magnitude of the current shows the absolute value of the unbalance. The output voltage at resistor R_5 is filtered by a capacitor.

The adjustment of the circuit is achieved in the following way: At first the resonant frequency of the series resonant circuit is set. After that the phase bridge has to be adjusted. For this purpose the output voltage at resistor R_5 has to be observed on an oscilloscope and such a phase shift adjusted by resistor R_1 that a pulsing AC voltage is seen at the oscilloscope. The filter capacitor has to be disconnected for these measurements. After that the desired sensitivity of the amplifier can be set by resistor R_2. For the adjustment of the phase shift mentioned above it is recommended to set at first the highest possible sensitivity.

Equal peak values of the pulsing AC voltage at resistor R_5 are set by the resistors R_3 and R_4.

For unbalancing of the bridge in the other sense the oscilloscope has to show a curve symmetrical to the zero line. If that does not occur the adjustment of the amplifier has to be repeated.

Technical data

Operating voltage	20	V
Operating current	approx. 30	mA
Bridge Voltage	15	V
Frequency of the bridge voltage	160	kHz
Power consumption of the bridge	1.5	W
Bridge unbalancing for the output voltage of 200 mV	1	‰
Maximum bridge unbalancing	1	%
Maximum ambient temperature	60	°C

Transformer

Tr 1: Siferrite pot core B65531-K0000-R022
N_1 = 60 turns 0.1 CuL
N_2 = 2 × 60 turns 0.1 CuL, joint lead wound coil
Tr 2: Siferrite pot core B65521-J0000-R022
N_1 = 70 turns 0.12 CuL
N_2 = 2 × 10 turns 0.12 CuL, joint lead wound coil
Tr 3: Siferrite pot core B65541-K0400-K026
N_1 = 70 turns 0.18 CuL
N_2 = 2 × 35 turns 0.18 CuL, joint lead wound coil
L_1 = Siferrite pot core B65541-K0400-K026 with tuning slug
N = 230 turns 0.13 CuL

5.12. Pulse Control Circuit

In one of our former booklets the circuit of an electronic counter with presetting has been indicated. This circuit delivers an output signal after an adjusted number of input signals has occured. This circuit has the disadvantage that for a series connection of several decades not any arbitrary number can be set.

The circuit in Fig. 5.14. extends the arrangement mentioned above in such a way that, for example, for three decades connected in series all numbers from 1 to 999 can be preset.

The circuit consists of an astable multivibrator as pulser, the multivibrator consisting of three transistors. The transistors T_6 and T_7 are incorporated in the multivibrator itself and transistor T_8 decouples capacitor C_2, so that the output signal has the exact shape of a square wave. This circuit has already been described in former chapters.

The pulses are applied to the input of three counter decades, connected in series, with a diode matrix. For these counter decades the circuit mentioned above can be used; only the reset stage and the switching amplifier at the output have to be omitted. The counter decades deliver a positive pulse at the output whenever the preset number of pulses is attained. The outputs of the three decades are connected in parallel and thus form an and-gate. Only if a positive signal occurs at all outputs, the switching amplifier with the transistors T_2 and T_3 responds, and the relay is switched via the thyristor tetrode BRY 20. In this way each number from 1 to 999 can be set. Decade 1 is the unit decade, decade 2 the ten decade and decade 3 the hundred decade. If, for instance, the number 325 is set, all three decades are operating until the third decade has received three input pulses. The second decade will remain in position 2. This occurs after 320 input pulses. After further five input pulses another positive pulse occurs at output 5 of the decade 1 (switch S 1). Since there is already positive potential at the switches S 2 and S 3, the switching amplifier responds. Via a diode a second switching amplifier is driven by which the pulser is blocked. Thus no further input pulses are applied to the counter decades.

The circuit falls into the operating condition on closing switch S 4. Before that the output stage of the switching amplifier has to be put in rest position by pushing key K.

Such a pulse circuit can be used, e.g., for controlling stepping motors.

Technical data

Operating voltage	12	V
Minimum load resistance	5	kΩ

The capacitance of the capacitors C_1 and C_2 is dependent on the desired pulse frequency.

5.13. Pulse Coupling Circuit

There are difficulties, to decouple DC wise longer square-wave pulses without distortion. Normally the decoupling is achieved by a capacitor, which, however, is charged by the pulse current, so that the amplitude of the transmitted pulse decreases with increasing pulse time.

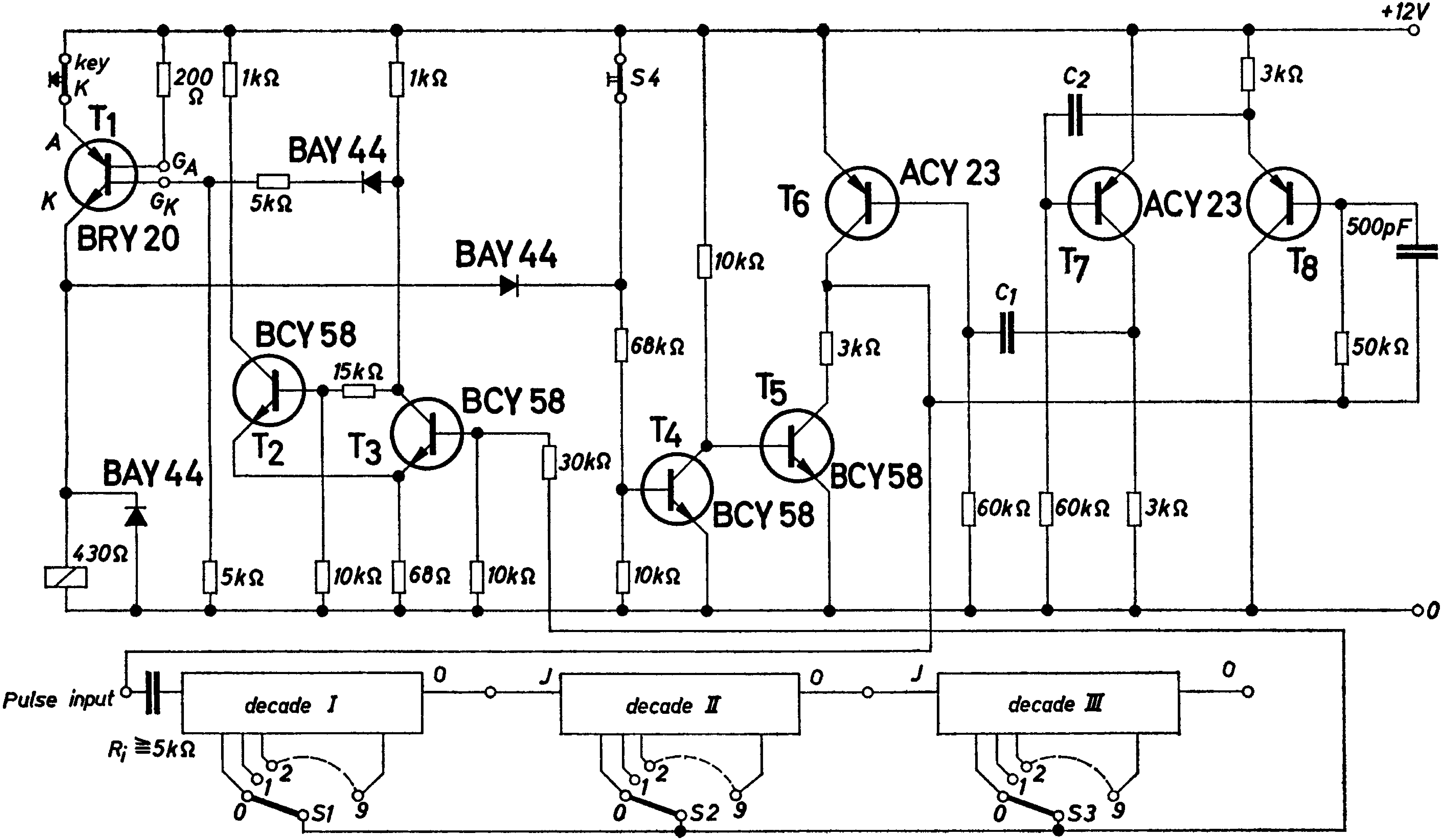

+12V
key
K
T1
A
K
G_A
G_K
BRY 20
200 Ω
1kΩ
1kΩ
BAY 44
5kΩ
BAY 44
S4
10kΩ
T6
ACY 23
C2
3kΩ
ACY 23
T7
T8
500pF
C1
BCY 58
15kΩ
BCY 58
T2
T3
68kΩ
3kΩ
50kΩ
T5
T4
BCY58
BCY 58
30kΩ
BAY 44
430Ω
5kΩ
10kΩ
68Ω
10kΩ
10kΩ
60kΩ
60kΩ
3kΩ
0
Pulse input
R_i ≧ 5kΩ
decade I
decade II
decade III
0
J
0
J
0
0
1
2
9
S1
S2
S3

Fig. 5.14.

Such a sinking of the square-wave pulse amplitude is avoided by the coupling circuit of Fig. 5.15. A tunnel diode is connected in parallel to the input of the transistor. The transistor is switched on by a positive input pulse via the capacitor C_1 and at the output a jump in potential occurs. Simultaneously the tunnel diode will be switched on if the input pulse is higher than the current peak of the tunnel diode. The tunnel diode remains traversed by current even if the pulse current decreases by charging the capacitor C_1. The voltage drop at the tunnel diode and at the germanium diode connected in series is sufficient to keep the transistor permanently conductive. Only if the input pulse is switched off, a negative pulse is applied to the input of the switching amplifier via the capacitor C_1, so that the tunnel diode is switched off again. Simultaneously the transistor will be switched off because the voltage drop at the tunnel diode has decreased to a very low value. At the output again a jump in potential occurs. The output pulse has an exact shape of a square wave, the length of which corresponds accurately to the length of the input pulse, because the transistor is always either switched on or switched off. The operating voltage of the circuit is 12 V. The collector voltage of the transistor is maintained, however, by a diode at a voltage of 6 V in the off-condition of the transistor so that fast switching of the transistor is attained.

Technical data

Operating voltage	6 and 12	V
Input voltage	3 to 20	V

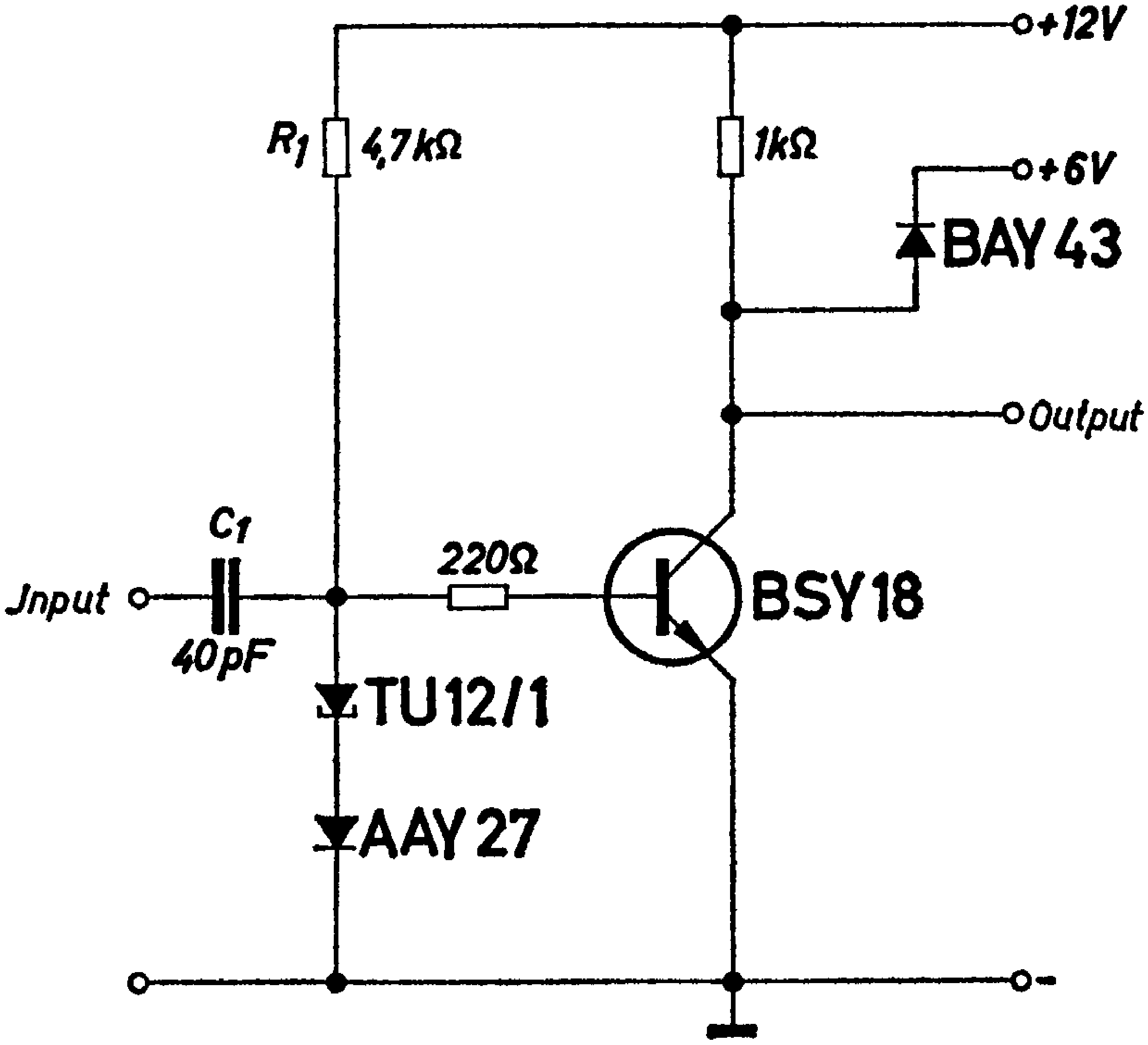

+12V
R1
4,7kΩ
1kΩ
+6V
BAY 43
Output
C1
Jnput
40pF
220Ω
BSY18
TU12/1
AAY 27
-

Fig. 5.15.

6. Controlled Power Supplies

Transistor circuits normally are operated at relatively low voltages. If transistor circuits have a higher power consumption, due to the low voltages high currents are needed. In order to minimize the change of the output voltage at load variations, power supplies with a low internal resistance are required, which can be obtained in the easiest way by a voltage control. Simultaneously also variations in the line voltage can be compensated. Furthermore by the controlled power supply filtering can be achieved which is otherwise (with conventional means, as chokes, resistors and capacitors) very expensive because of the high load currents.

6.1. Reference Voltage Source 10 V/100 mA

For an adequate design of a controlled power supply the constancy of the controlled output voltage corresponds rather accurately to the stability of the reference voltage, which normally is given by a Zener diode.

In the example of Fig. 6.1. the reference voltage is obtained by the Zener diode BZY 83/C5 V6. The temperature coefficient of the Zener voltage of the diode BZY 83/C5 V6 has a value of approx. -10^{-4}/°C. The voltage at this Zener diode will be compared with the voltage at the load in the differential amplifier consisting of transistors T_1 and T_2. For the supply line to the control circuit own conductors are provided, so that also the voltage drop which occurs in longer wires is compensated. The magnitude of the controlled output voltage can be set by a potentiometer of 200 Ω.

The signal of the differential amplifier is applied to the series transistor BDY 12 via the amplifier stages.

For overload protection a current limiting device has been provided. As soon as the voltage drop at the series resistor of 12 Ω exceeds a certain value, transistor T_3 is switched on, so that another transistor (BCY 59) becomes conductive. The collector-emitter path of the transistor is connected in parallel to the output of the differential amplifier. By that means a drive of the series transistor far beyond saturation via the pre-stages is avoided and the short-circuit current is limited approx. to 230 mA. In comparison to the conventional

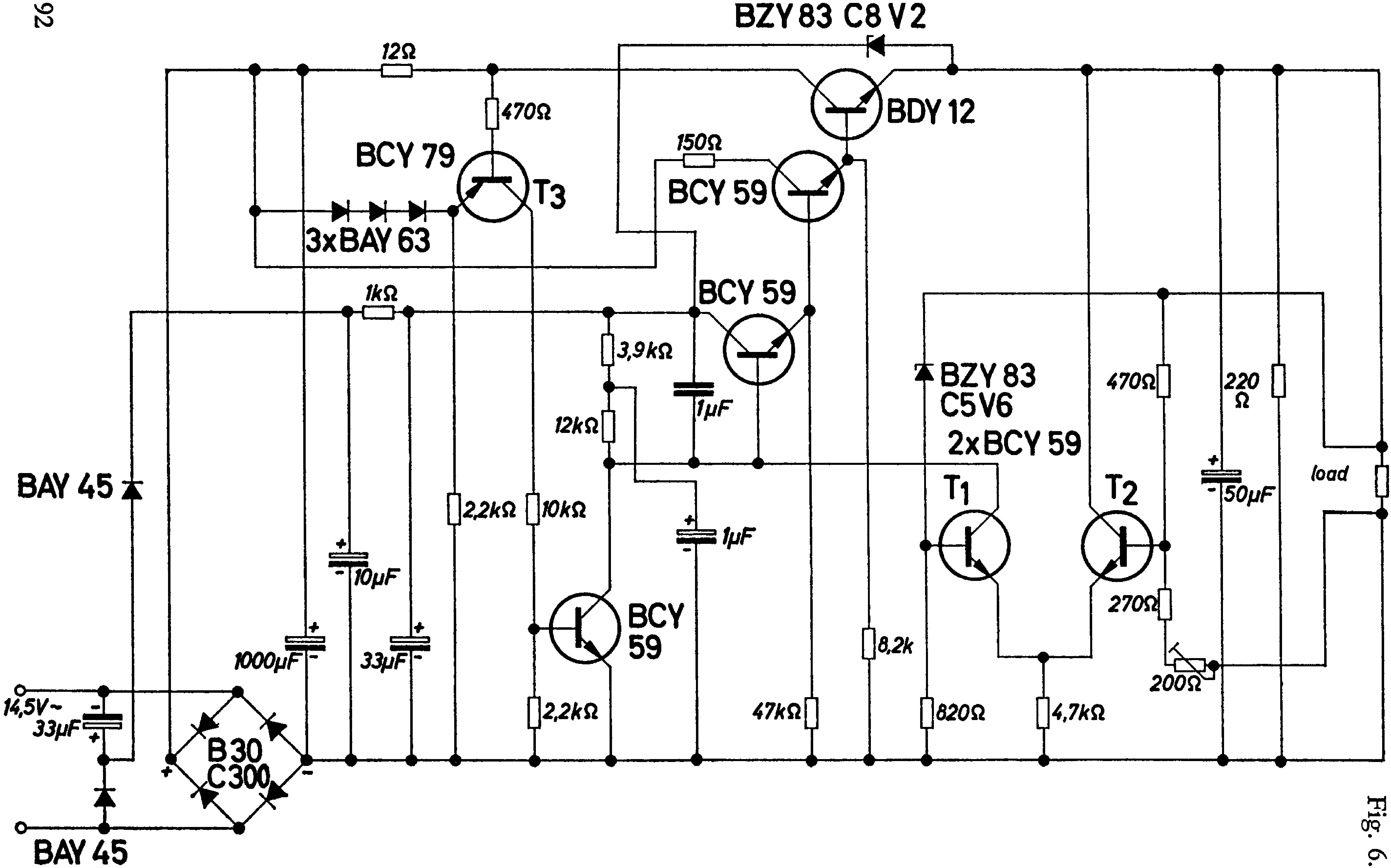

BZY 83 C8 V2
12Ω
470Ω
BDY 12
BCY 79
T3
150Ω
BCY 59
3xBAY 63
BCY 59
1kΩ
3,9kΩ
1µF
12kΩ
BZY 83
C5V6
2xBCY 59
470Ω
220
Ω
T1
T2
load
BAY 45
2,2kΩ
10kΩ
1µF
50µF
10µF
BCY
59
270Ω
1000µF
33µF
8,2k
200Ω
14,5V~
33µF
2,2kΩ
47kΩ
820Ω
4,7kΩ
B30
C300
BAY 45

Fig. 6.1.

fuses the short-circuit fuse has the advantage that after elimination of the short circuit the power supply can immediately operate again. For the cut-off fuses usually a key has to be pushed to switch on the power supply again.

Technical data

Operating current	14.5	V
Output voltage	10	V
Maximum output current	100	mA
Variation of the output voltage between open circuit and full load operation	< 1	mV
Variation of the output voltage for a change in line voltage of ±10%	< 1	mV
Hum voltage	< 1	mV
Temperature dependence of the output voltage	approx. -10^{-4}	°C
Maximum ambient temperature	60	°C

6.2. Power Supply for Small Type Motores

Small type motors can be operated at 6 V and 12 V batteries without switching by the power supply shown in Fig. 6.2. Also other loads, of course, can be fed by this power supply. For the 6 V operation the Zener diode D_1 is blocked so that transistor T_3 remains switched off. Transistor T_1 is switched on via transistor T_2 and the motor is always connected to the full operating voltage. The capacitor C_2 is discharged slowly via the high resistance of the blocked Zener diode D_1, so that the astable multivibrator, consisting of transistors T_2 and T_3, always flips for a short time after longer time intervals.

At an operating voltage of 12 V the Zener diode D_1 is traversed by current and the multivibrator operates at a duty cycle 1:4, i.e., only during a fourth of a cycle the motor is traversed by double current at double operating voltage. This results in the average in the same power consumption as for the 6 V operation.

An experiment with a small type motor showed that for this pulse operation the same power consumption is not sufficient in order to attain the total torque. By changing the capacitors C_1 and C_2, the

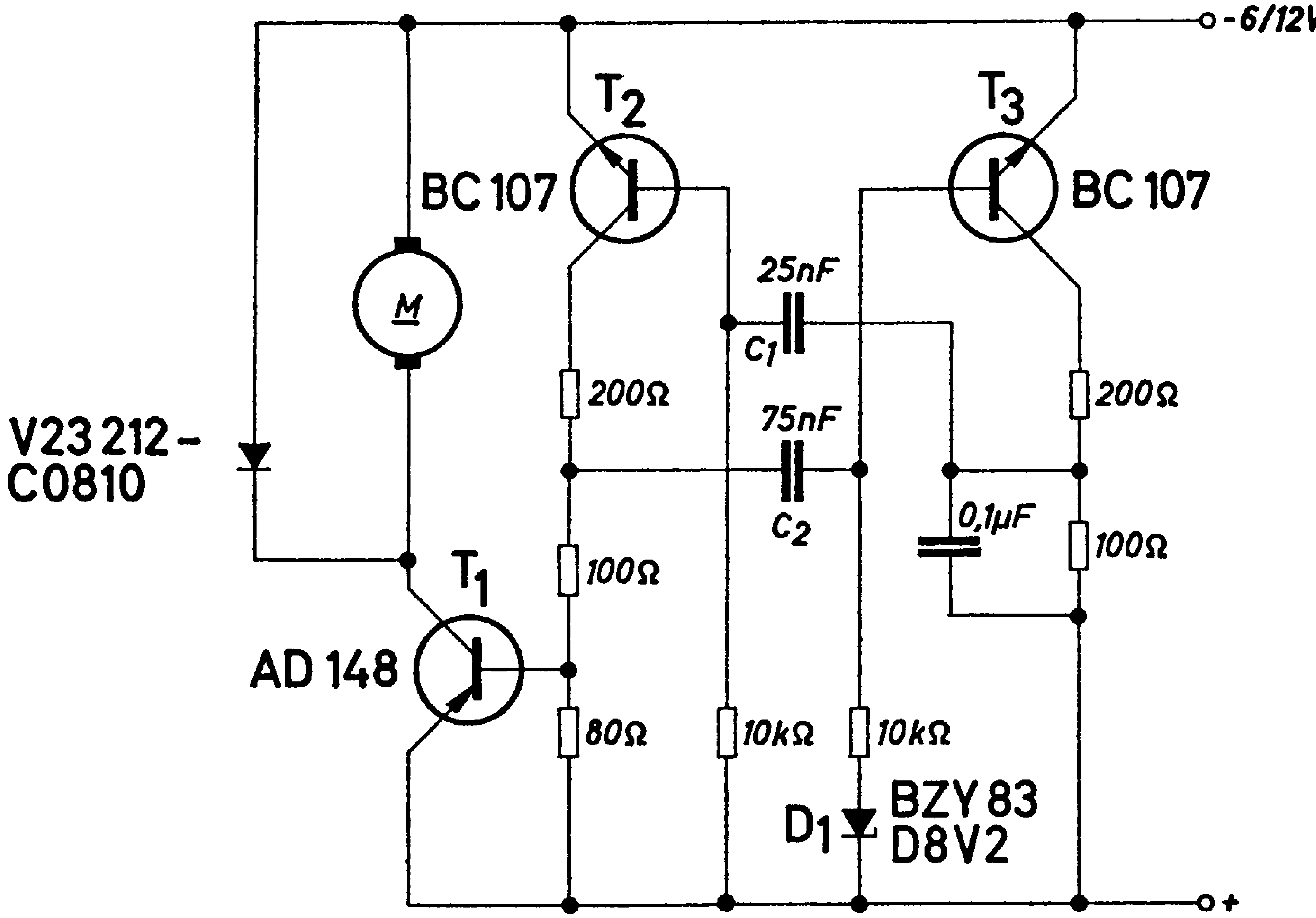


Fig. 6.2.

duty cycle can be reduced, for instance, from 1:4 to 1:2 so that in any case the total torque can be attained also for a 12 V operation. The silicon diode parallel to the motor prevents the destruction of the output transistor by breaking-voltage peaks.

Technical data

Operating voltage	6 or 12	V
Minimum load	10	Ω

7. Radio Frequency Circuits

In the RF range silicon transistors are used particularly for high power applications. The planar technology applied for silicon transistors makes the realization of required complicated system geometries for high-frequency power transistors possible. For pre-stage applications germanium transistors are superior due to the better noise characteristics and smaller feedback capacitances.

7.1. Conductor, Line- and Trouble Searching Instrument

The flat-type leads used more and more in house installations and installed directly in the brickwork require instruments by which the course of the leads or the position of line failures can be discovered. In this chapter a simple equipment will be described which fulfils at best the requirements. It consists of a transmitter (Fig. 7.1.) and a receiver (Fig. 7.2.). The transmitter or generator consists of two astable multivibrators. One multivibrator oscillates at 100 kHz (M 2) and another multivibrator at 1.5 kHz (M 1). The multivibrators are connected with each other in such a way that the 100 kHz signal is 100% modulated with 1.5 kHz, i.e., the multivibrator M 2 is switched on and off by multivibrator M 1 in a 1.5 kHz cycle. The generator will be connected to the currentless line, e.g., to the plug socket. The connection can be symmetrical (poles 1 and 3) or unsymmetrical (poles 1 and 2 or 2 and 3). Due to the limiting resistors in the output of the multivibrator M 2 the generator is short-circuit proof.

The receiver is tuned to 100 kHz. The RF signal received by a dipole or ferrite rod antenna is amplified by the integrated amplifier TAA 111 and after that demodulated. The signal is applied to a measuring instrument or to an earphone via an amplifier stage. The acoustic indication is possible because the 100 kHz signal is modulated with a low frequency. For simple applications an acoustic indication is certainly preferred. If, however, the acoustic indication can be avoided, the modulation of the 100 kHz signal may be omitted and multivibrator M 2 is sufficient as generator (Fig. 7.1.).

For the search of a line break the end of the line has to be short-circuited. For very long lines the short circuit may be omitted because the capacitance between both leads already is sufficient for

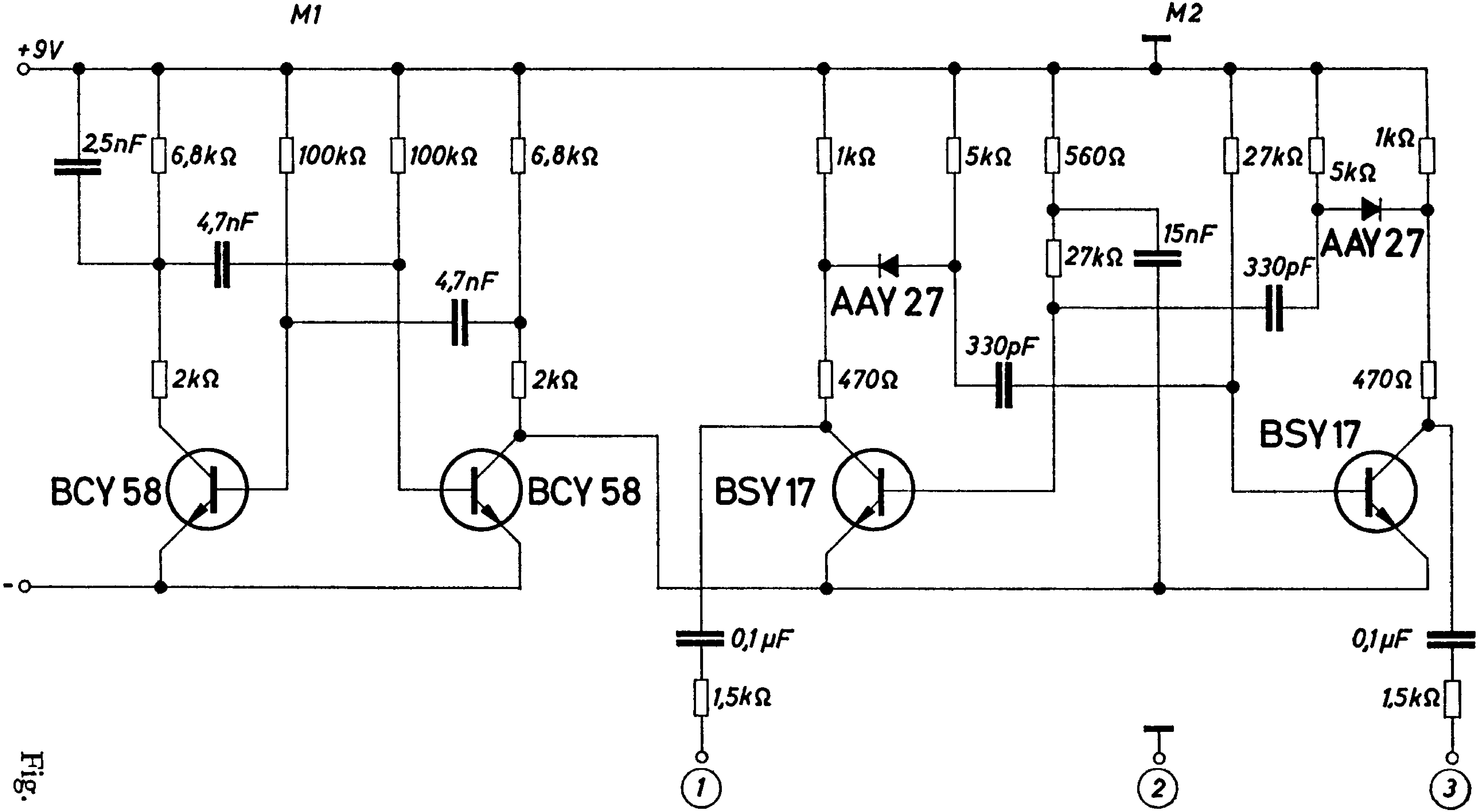
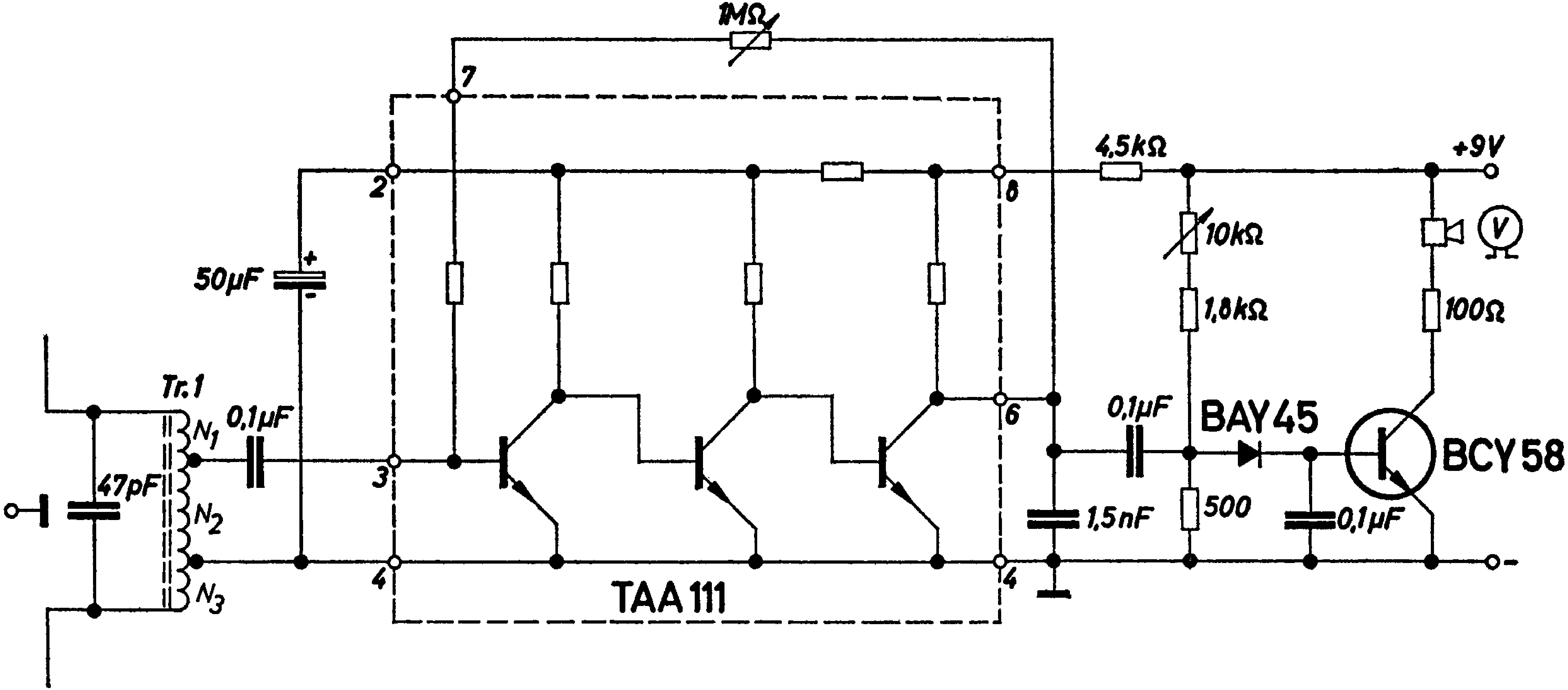

M1
M2
+9V
-
2,5nF
6,8kΩ
100kΩ
100kΩ
6,8kΩ
4,7nF
4,7nF
2kΩ
2kΩ
BCY 58
BCY 58
1kΩ
5kΩ
560Ω
27kΩ
5kΩ
1kΩ
AAY 27
AAY 27
15nF
27kΩ
330pF
330pF
470Ω
470Ω
BSY 17
BSY 17
0,1μF
0,1μF
1,5kΩ
1,5kΩ
1
2
3

Fig. 7.1.

Fig. 7.2.



1MΩ
7
2
8
4,5kΩ
+9V
10kΩ
50μF
1,8kΩ
100Ω
V
Tr.1
N1
0,1μF
6
0,1μF
BAY45
BCY58
3
47pF
N2
1,5nF
500
0,1μF
4
4
N3
TAA111

a short circuit. After the generator is connected the line is tracked by the receiver, i. e., this course is taken where the output signal is the largest. As soon as the position of the line break is reached, the high field density occurring at the line break results in an increase of the signal. Behind the line break there is a drop in field intensity. For the test a dipole or an unsymmetrical antenna is suitable. According to the description of the whole procedure the course of the line also can be located with this method.

For short circuits the receiver again will be led along the line. At the location of the short circuit there is a sudden drop of the signal. For this measurement procedure a ferrite antenna is more suitable because in this test method a magnetic field has to be indicated.

Transformer Tr 1 : (Fig. 7.2.)
Siferrite pot core B65531-KO160-AO28

N_1 = 250 turns 0.06 CuL
N_2 = 70 turns 0.06 CuL
N_3 = 275 turns 0.06 CuL

7.2. Quartz Oscillators with Transistors

Fig. 7.3. shows a circuit with a one-stage quartz oscillator of an output power of approx. 400 mW at a frequency of 27.12 MHz. The transistor BSY 34 is operated in common base configuration. The feedback is represented by a capacitor between collector and emitter. The quartz is in the transistor base circuit. For all frequencies, except the resonant frequency, the quartz causes such a strong negative feedback, so that the feedback via the capacitor is not sufficient in order to maintain an oscillation. Therefore the oscillator only oscillates in the series resonant frequency of the used quartz. The resistance R_1 will be adjusted in such a way that there is an optimal condition of oscillation. In the collector of the oscillator transistor there is a parallel resonant circuit. The output power will be coupled out via a series resonant circuit and a transforming low-pass filter. All used coils are air-core coils. By bending the air-core coils the oscillator will be adjusted to maximum output power. The transistor additionally has to be cooled, e. g., by attaching a cooling clamp to the metal encapsulation.

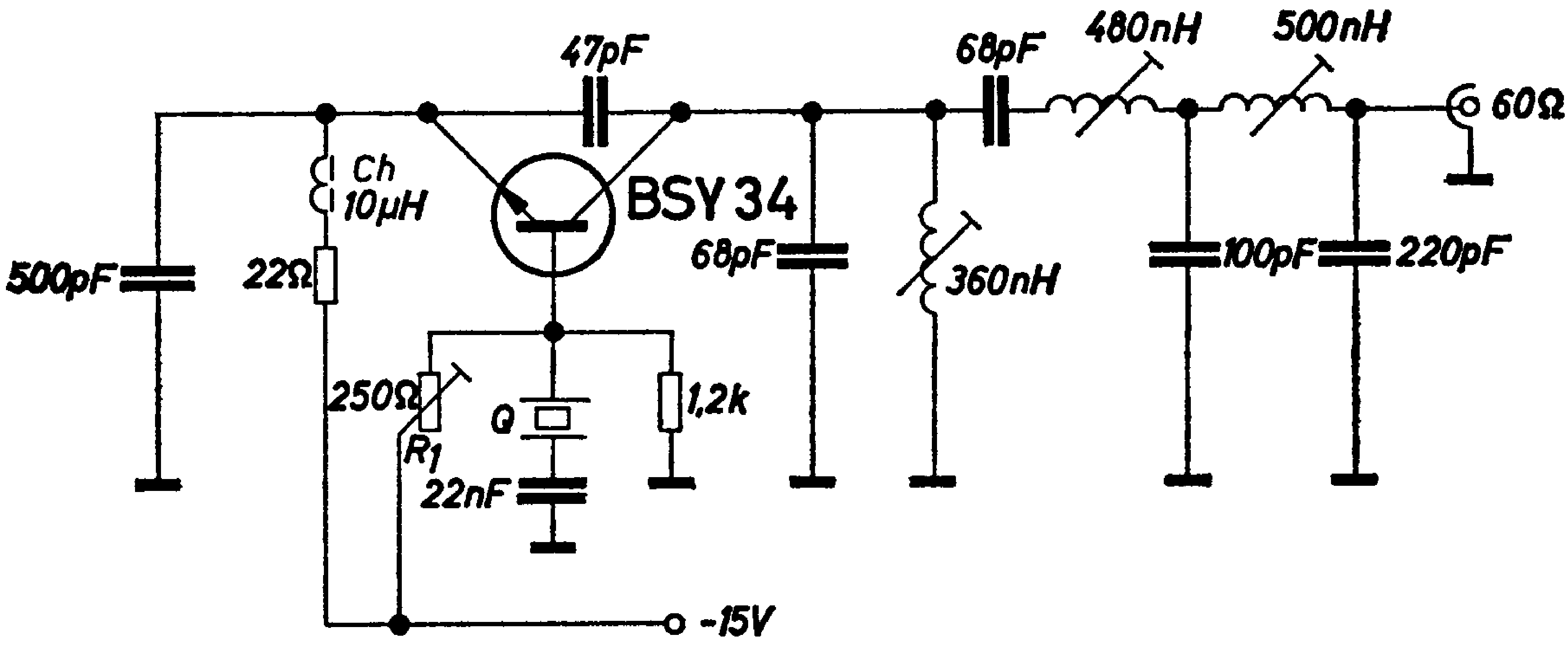


Fig. 7.3.

Technical data (Fig. 7.3.)

Operating voltage	15	V
Operating current	95	mA
Output power	400	mW
Oscillation frequency	27.12	MHz

In Fig. 7.4. a two-stage version of the oscillator is shown with which an output power of 2 W can be attained. The oscillator corresponds actually to that of Fig. 7.3. The efficiency of the whole circuit is approx. 45%. The class C end stage is coupled inductively to the oscillator so that a good matching is achieved. A special circuit for coupling out the power is not necessary. The output power can be directly tapped off at the collector of the end stage. The mechanical construction of the circuit is unsymmetrical. For both the transistors an additional cooling is provided.

Technical data (Fig. 7.4.)

Operating voltage	15	V
Operating current	300	mA
Output power	2	W
Oscillation frequency	27.12	MHz
Efficiency	approx. 45	%

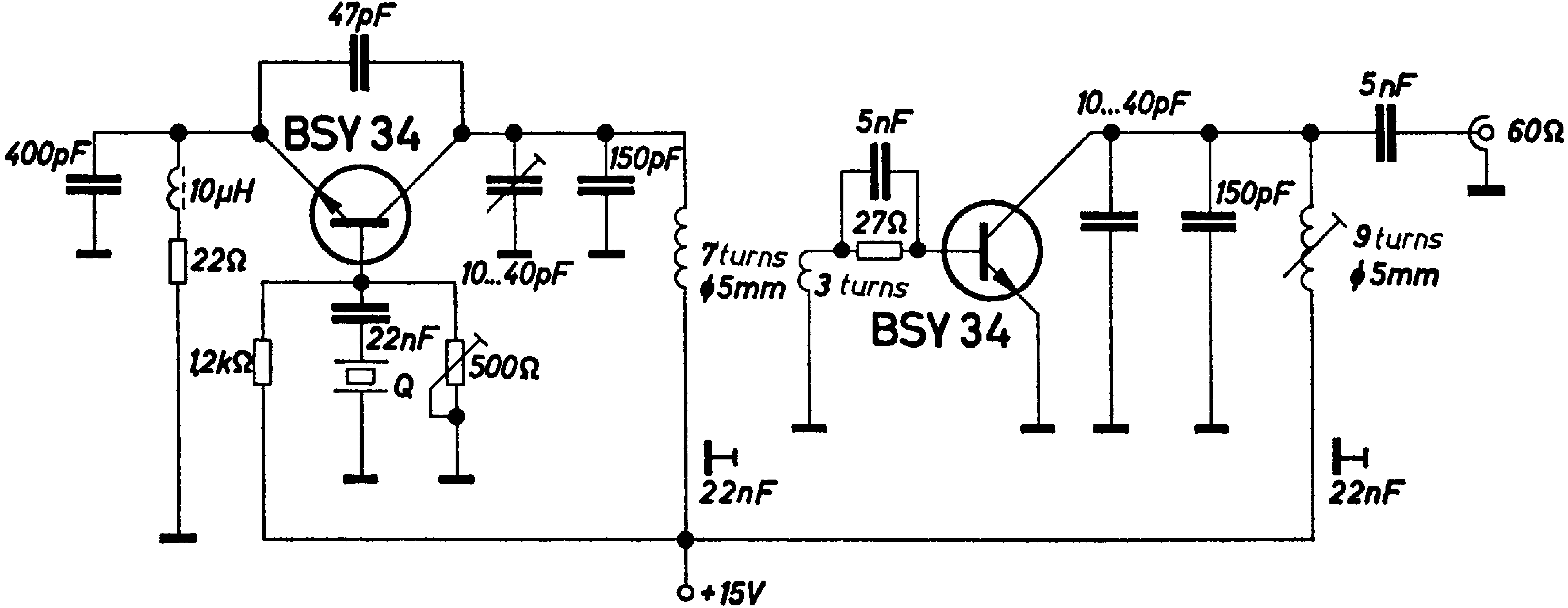

47pF
BSY 34
400pF
10µH
22Ω
1,2kΩ
22nF
Q
500Ω
10...40pF
150pF
7 turns
ϕ5mm
22nF
+15V
5nF
27Ω
3 turns
BSY 34
10...40pF
150pF
9 turns
ϕ5mm
5nF
60Ω
22nF

Fig. 7.4.

7.3. Wireless Microphone

The so-called wireless microphones are mainly used for indoor applications. The distance between the microphone and the connected transmitter on the one hand and receiver on the other hand usually amounts to some hundred meters, so that a small transmitter power is sufficient.

Fig. 7.5. shows a suitable circuit. The RF portion consists of a totally screened oscillator which is operated by a stabilized voltage and a loosely coupled buffer stage. With this two-stage lay-out a good frequency stability and low harmonic interference radiation is achieved. The frequency of the oscillator can be adjusted in a small range around 150 MHz by a trimmer capacitor C_1 to a value allowed for this operation.

The three-stage modulation amplifier has a high input impedance, it is especially designed for the connection to a crystal microphone.

The FM-modulation of the transmitter is achieved by a varicap BA 138. The frequency ratio can be adjusted by the capacitor C_2.

For enlarging the range the output circuit which is tuned with capacitor C_3 can be terminated by a $\lambda/4$ antenna.

Technical data

Operating voltage	9	V
Operating current	8	mA
Frequency (adjustable)	approx. 150	MHz

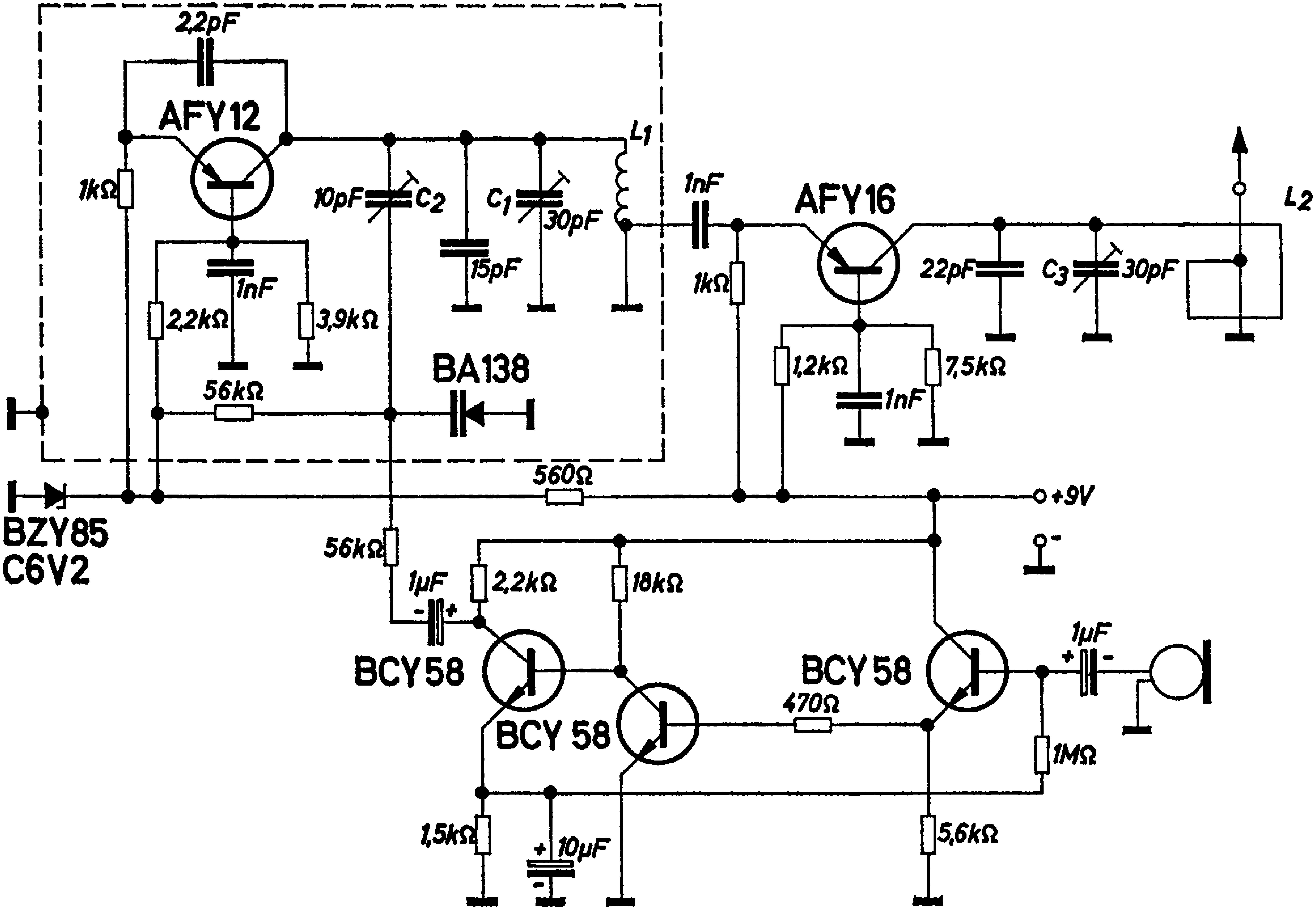
Inductances:

L_1 : 2.5 turns, 5 mm coil diameter, 1.0 CuAg with a tap-off at a wire length of 12 mm, measured from ground terminal

L_2 : Stirrup, 18 mm $\times$ 15 mm, 1.5 CuAg

7.4. Transistor Relay with Galvanic Decoupling of Input and Output Circuit

Today transistors are often used as switches. In some applications, however, the coupling between input and output of the transistor is a certain disadvantage. The circuit of Fig. 7.6. shows an example how a galvanic decoupling in transistor circuits between the input and output comparable to relay circuits is achieved.



2,2pF
AFY12
1kΩ
1nF
2,2kΩ
3,9kΩ
10pF
C2
C1
30pF
15pF
L1
56kΩ
BA138
1nF
AFY16
1kΩ
22pF
C3
30pF
L2
1,2kΩ
7,5kΩ
1nF
560Ω
+9V
BZY85
C6V2
56kΩ
1μF
2,2kΩ
18kΩ
BCY58
BCY 58
BCY 58
470Ω
1μF
1MΩ
1,5kΩ
10μF
5,6kΩ

Fig. 7.5.

The decoupling is accomplished by the application of an inductive transmission path in which the RF modulated switching pulse is transmitted.

The transistor relay consists of an oscillator and a receiver stage. The oscillator frequency has been chosen very high so that for the modulation of the oscillator a wide transmission band-width is obtained. It is at approx. 20 MHz so that a band width from 0 to about 500 kHz eventually up to 1 MHz is possible.

The quiescent point of the oscillator is stabilized by a resistor in the emitter circuit. Since it also acts as negative AC feedback it increases simultaneously the frequency stability.

The output of the oscillator is transformer-coupled to the receiver. The best coupling is accomplished by a commonly wound primary and secondary coil of the transformer. The isolation between both windings determines the permissible potential difference between input and output of the transistor relay. A value of 1000 V is obtained without any difficulties.

The RF signal rectified by a diode and filtered is applied to a switching stage. The maximum switching current amounts to approx. 50 mA. This current flows as long as the oscillator oscillates in the primary section. In the simplest case its supply voltage may be disconnected in order to initiate a switching process in the secondary section. Between input and output there is a delay of about 1 μs. For small input signals instead of a mechanical switch a transistor switching stage can be used for switching the operating voltage. (Fig. 7.6.). The required input voltage amounts to 2 to 3 V. For switching frequencies up to 100 kHz in this stage the transistor BCY 58 is most suitable, for higher frequencies the transistor BSY 17 is more preferable because of its shorter switching times.

The high frequency voltage is only necessary for the transmission of a signal. To avoid interference of neighbouring circuit parts, this voltage should be kept at a distance from the input and output. Therefore, the whole lay-out has to be installed in a shielded case. At the output additionally a choke has to be connected. A reduction of the interference voltage at the output also can be accomplished by a symmetrical lay-out of the secondary winding of the transformer and the full-wave rectifier connected. Thus, for the high frequency a bridge circuit is obtained and the filtering is facilitated because of the double frequency generated at the rectification.

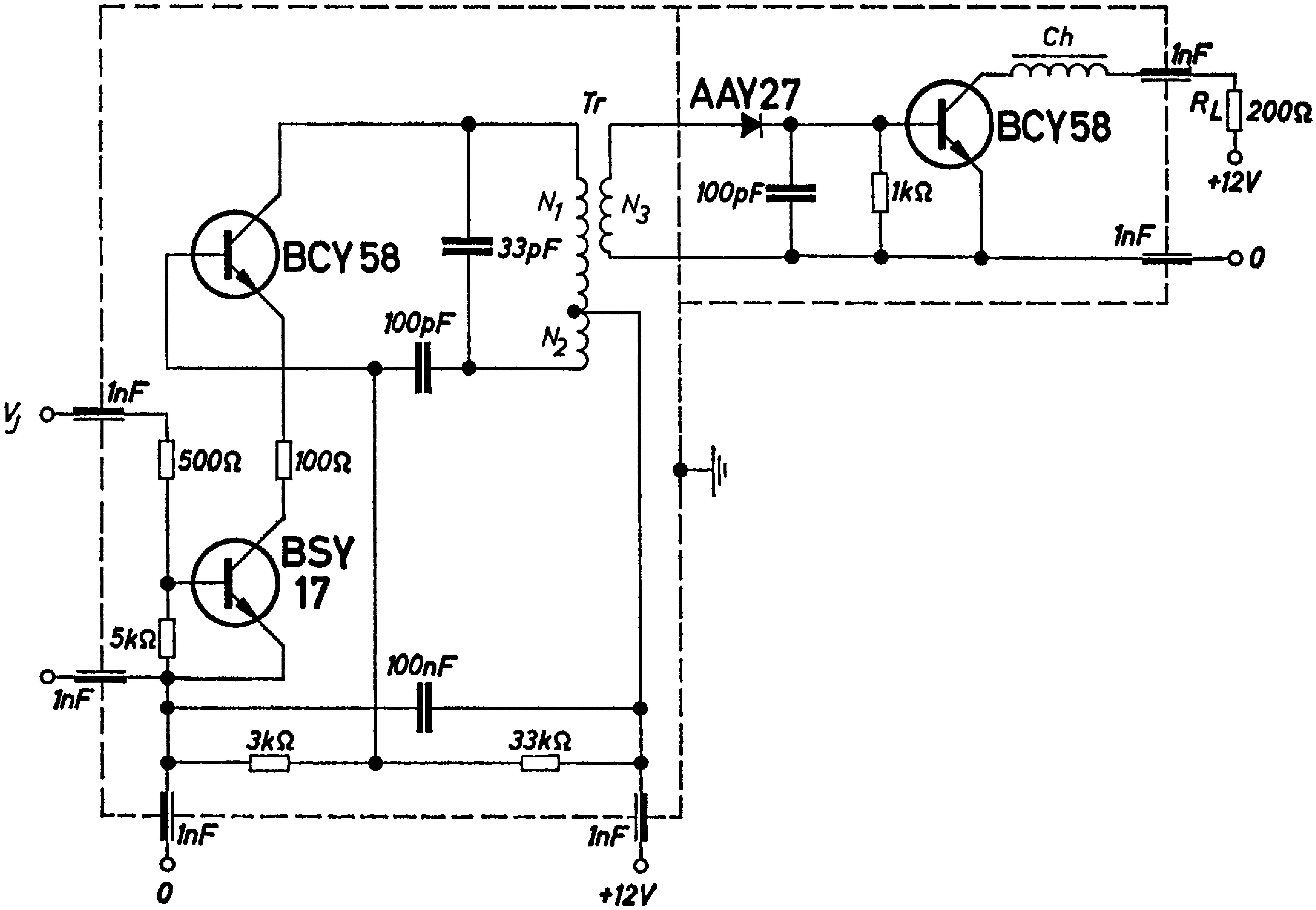

Ch
1nF
AAY27
Tr
BCY58
R_L
200Ω
+12V
100pF
1kΩ
N_1
N_3
1nF
0
BCY 58
33pF
100pF
N_2
V_j
1nF
500Ω
100Ω
BSY 17
5kΩ
1nF
100nF
3kΩ
33kΩ
1nF
1nF
0
+12V

Fig. 7.6.

Technical data

Operating voltage	2×12	V
Operating current of the oscillator	ca. 5	mA
Oscillator frequency	ca. 20	MHz
Maximum pulse frequency	500	kHz
Input voltage	2 to 3	V (or 12 V without transistor BSY 17)
Minimum input pulse width	1	μs
Pulse delay	1	μs
Switching current	50	mA

Transformer Tr. : Air core coil, bobbin 6 mm ⌀

$N_1 = 8$ turns 0.6 CuL

$N_2 = 4$ turns 0.6 CuL

$N_3 = 7$ turns 0.4 CuL (joint lead wound coil for N_1 and N_3)

Choke Ch : $L = 6\ \mu H$

8. Radio Circuits

Also this year new semiconductor components offering noticeable advantages for radio and TV circuits have been introduced. There is, for instance, the VHF switching diode BA 136, which can be used for the switch-over from the lower to the upper VHF band. The high frequency switching characteristic of this diode is so good that fully electronic VHF tuners can be designed, the electric data of which practically are as good as those of tuners with mechanical switching. In any case the omission of mechanical contacts, however, offers several advantages, as for instance, the possibility of a simple remote control.

The introduction of coloured TV requires new semiconductor components for this application. In this chapter the colour video-circuit according to the RGB concept is illustrated.

For the electronic tuning of FM tuners new diodes BB 103 and BB 104 have been developed. In one of the succeeding chapters one design example with the diode BB 103 is explained. Since the mechanical layout of the tuner substantially influences its quality, also the printed board, on which the components have been mounted, is shown.

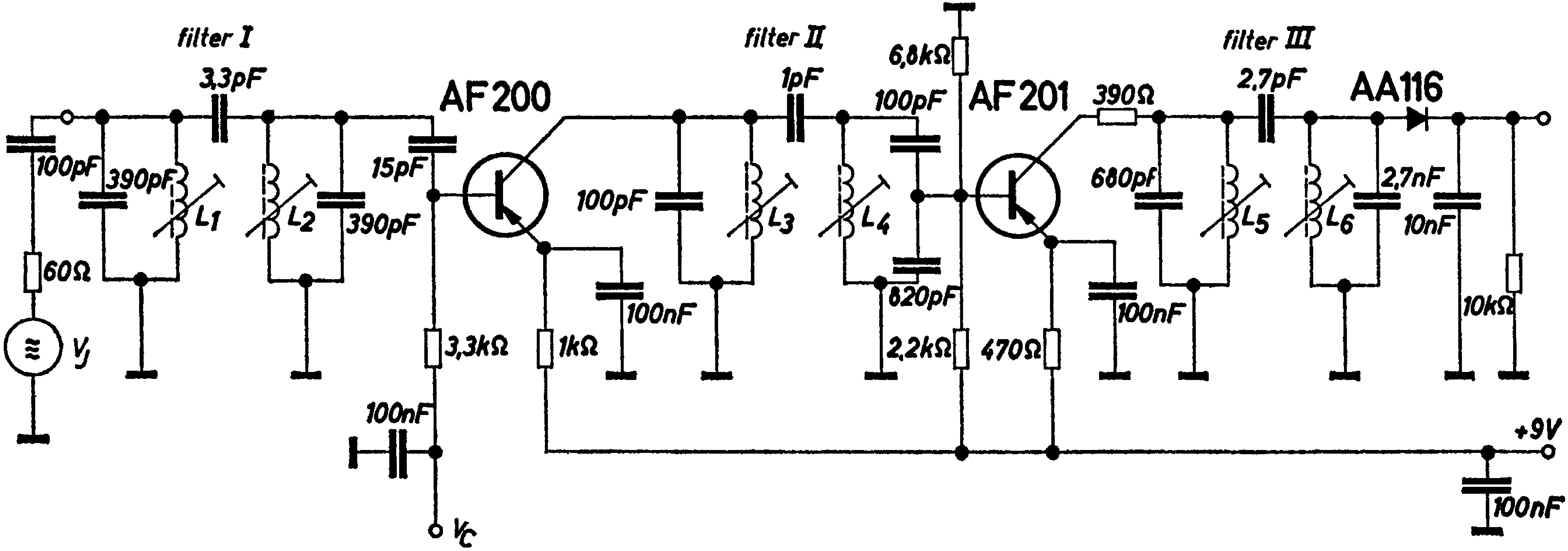
Investigations of various circuits result in further applications for germanium mesa transistors. Thus it is possible to accomplish a current upward control in AM-IF amplifiers with these transistors in a suitable circuit.

8.1. Upward Controlled AM-IF Amplifier

Because of the better characteristics for large input signals usually found in controlled transistor circuits the current upward control is preferred to the current downward control. In AM-IF amplifiers, however, difficulties arise due to the high variations of the input and output admittances occuring at current-upward controls.

Fig. 8.1. shows the circuit of a two-stage AM-IF amplifier for which in the first stage the current upward control is used. An input voltage up to 1 V at 60 Ω can be handled. The overdrive characteristics of the IF amplifier are practically independent on the

Fig. 8.1.


filter I
3,3pF
100pF
390pF
L1
L2
390pF
60Ω
Vi
15pF
AF200
3,3kΩ
100nF
VC
1kΩ
100nF
100pF
filter II
1pF
100pF
L3
L4
820pF
6,8kΩ
2,2kΩ
AF201
470Ω
100nF
390Ω
filter III
2,7pF
680pF
L5
L6
2,7nF
AA116
10nF
10kΩ
+9V
100nF

degree of the control. The IF amplifier uses the germanium mesa transistors AF 200 (control stage) and AF 201. The transistors will be operated in a non-neutralized common emitter configuration.

Only the layout of the filters I and II connected to the control transistor differ from the usual design of current downward controlled IF amplifiers. The input filter I is loaded in the primary section by the generator impedance and in the secondary section by the input impedance of the transistor. The transistor coupling has been accomplished by the principle of transformation with variable transformation ratio. Hereby the change in the transformation ratio takes place in such a way that inspite of the transistor input conductance varying during the control the attenuation of the filter remains constant over the whole control range. The change of the transformation ratio is caused by the active and reactive component of the transistor input admittance which is always dependent on the instantaneous state of control. By that means one achieves a wide range of control and a compensation of the deformation of the filter curve without any loss of power. Also for the design of output filter II of the control stage a special principle has been used which already stood the test for TV-IF amplifiers. The performance is as follows: In the uncontrolled state the filter operates as a double-tuned circuit band-pass filter with a total band-width B and a shape of the curve corresponding approximately to that of a single tuned circuit. The design has been executed in such a way that the total band-width B equals the band-width B_2 of the secondary circuit. Under control the primary circuit will be attenuated to a high extent and the total filter takes the shape of the curve and the band-width of the secondary circuit.

The range of control of this circuit is 70 db. That corresponds for an attainable total gain of 60 db to a control from +60 db to —10 db. The collector current of the transistor AF 200 varies simultaneously from approx. 5 mA to about 8 mA.

Fig. 8.2. shows the permissible input voltage vs. the downward control. For the purpose of comparison the same curve is shown for current downward control. As a measuring definition for both cases the following is valid: permissible input voltage at 60 Ω for a distortion factor of 10%. One derives from the diagram in Fig. 8.2. that down to a downward control of about 30 db both control methods are approximately equal. For a stronger downward control,

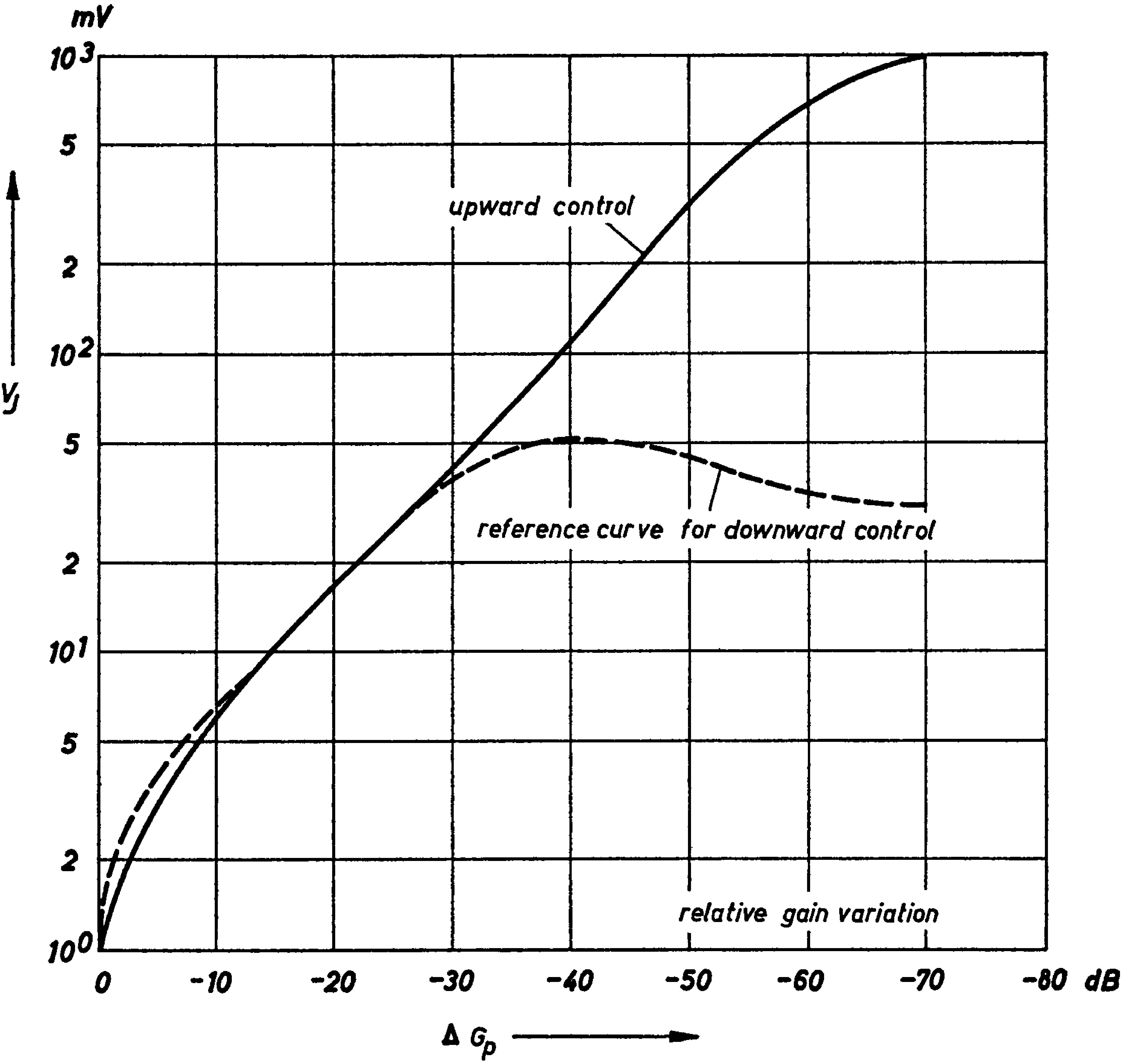


Fig. 8.2.

however, the current upward control is superior to the current downward control with regard to the permissible input voltage. Whilst in the first case a permissible input voltage of 1 V can be obtained for a downward control of 70 db, only a value of 30 mV is attainable in the second case.

The current upward control, therefore, makes a good reception of strong transmitters possible, as it is especially necessary in car radios and domestic sets.

Technical data

Operating voltage	9	V
Total gain	60	db
Range of control	70	db
Frequency	450	kHz

Inductances

$L_1 = 300\ \mu H, Q_0 \approx 120$	$L_5 = 180\ \mu H, Q_0 \approx 120$
$L_2 = 300\ \mu H, Q_0 \approx 120$	$L_6 = 40\ \mu H, Q_0 \approx 120$
$L_3 = L_4 = 1\ mH, Q_0 \approx 120$	(Q_0 means figure of merit for open circuit operation)

8.2. FM Tuner with Diode Tuning

The capacitance diodes BB 103 have been developed particularly for application in FM tuners. These are diodes with a relatively high capacitance ratio. The spread of the capacitance variation vs the voltage, however, still remains small. By that means a good tracking is achieved, when using these diodes, without the requirement of special pairing.

Besides the concept with single diodes mentioned in this chapter, circuits are known which use in each tuning branch two diodes connected in series with opposite polarity. For these concepts we offer the tuning diodes BB 104, in which both diodes are incorporated in one silicon chip and encapsulated in a plastic case. Because of the joint diffusion of both diodes one obtains a good symmetry. The FM tuner the circuit of which is shown in Fig. 8.3. consists of a pre-stage, a mixer stage and a separate oscillator. Because of its good noise characteristics the transistor AF 109 R is used in the pre-stage. A wide-band input circuit of the pre-stage guarantees better noise figures and to obtain a better filtering an inductively coupled tunable bandfilter is used between pre-stage and mixer stage, thus reducing interference originating from external sources and the harmonics of the oscillator frequency. The inductive coupling to the mixer ensures a good rejection at higher frequencies, for instance, of the image frequency.

For the tuning of the entire FM range from 87 to 108 MHz a change in voltage at the tuning diodes from 6 V to 28 V is sufficient. The small spread of initial diode capacitances and the capacitance variation difference for the bandfilter and the oscillator is adjusted by trimmer capacitors. The oscillator is loosely coupled to the mixer, in order to avoid a change of the oscillator frequency at high input signals. This is achieved by an inductance in the shape of a small loop at the emitter of the mixer transistor, which is connected in series to the inductance in the oscillator circuit. The supply voltage for the oscillator and the mixer is stabilized by a Zener diode in order to keep the oscillator frequency constant also for a varying supply voltage. For reduction of the interference radiation the pre-stage is decoupled by a shield from the other part of the tuner. The location of the shield is shown in Fig. 8.4.

Since the quality of the tuner is influenced substantially by the mechanical layout, the printed circuit with a component scheme for the board is shown in Fig. 8.4.

Technical data

Frequency	87	MHz	108	MHz
Operating voltage	12		12	V
Power gain	27		27	db
Noise figure	4.5		4	db
Tuning voltage	6		28	V
RF band-width	1.8		2	MHz
IF band-width	380		380	kHz
Oscillator interference voltage at 60 Ω	<0.15		<0.15	mV
Interference suppression				
Image frequency $f_r = f_{osc} + f_i$	62		58	db
Frequency $f_r = f_{osc} + \frac{f_i}{2}$	80		77	db

Inductances:

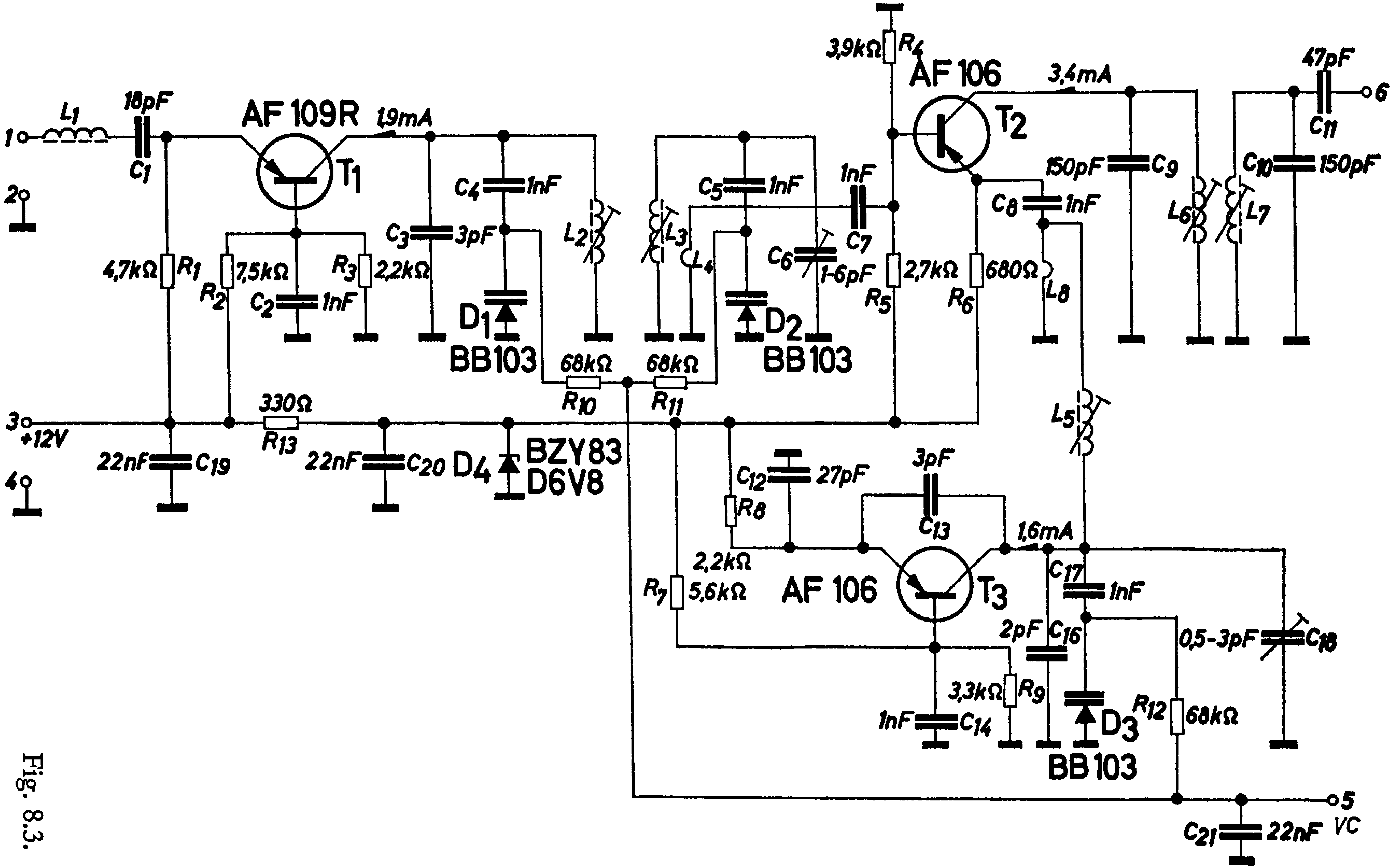
L_1 = Siferrite thread core B63310-U17-A12.3 4 turns 0.5 CuL

$L_2 = L_3$: Siferrite thread core B63310-U17-A12.3 4 turns 0.65 CuL

L_4 = 1 turn 0.65 CuL, wound on the same core as L_3

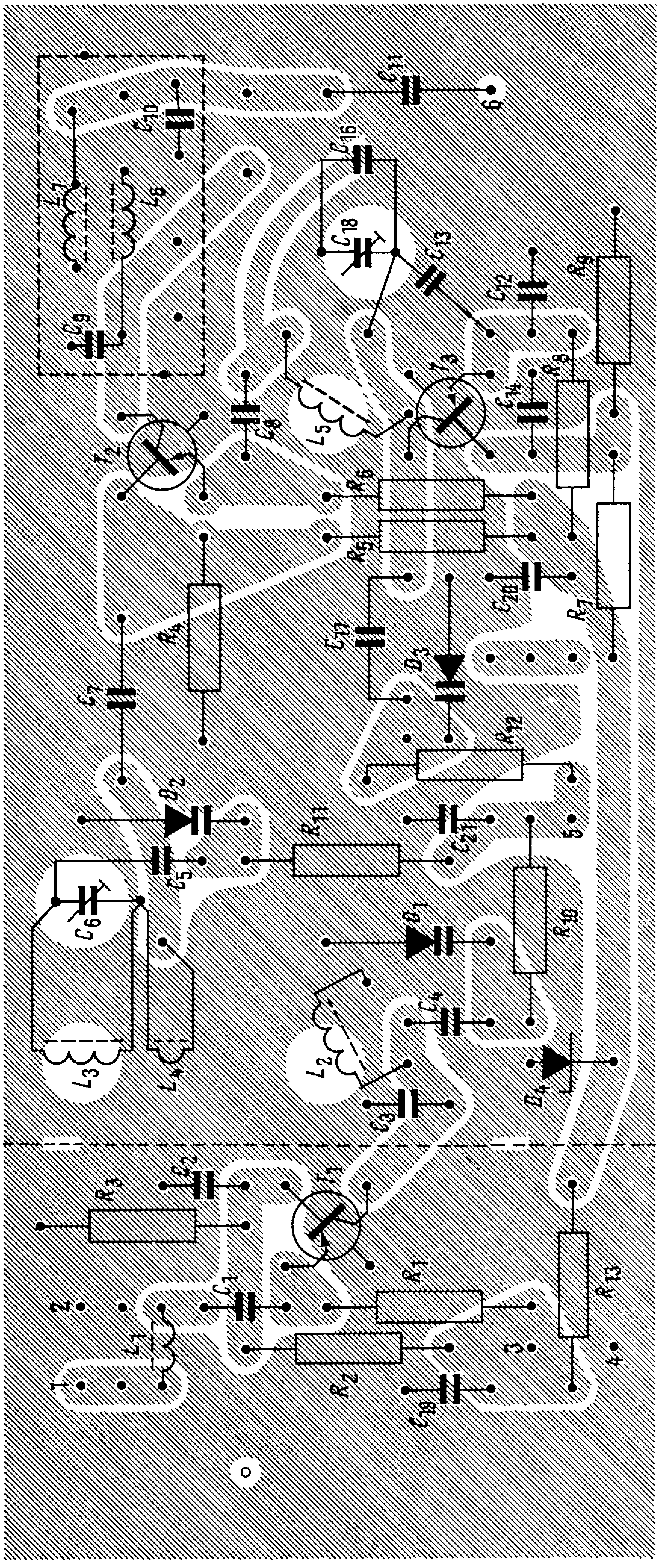
L_5 = Siferrite thread core B63310-U17-A12.3 3 turns 0.65 CuL

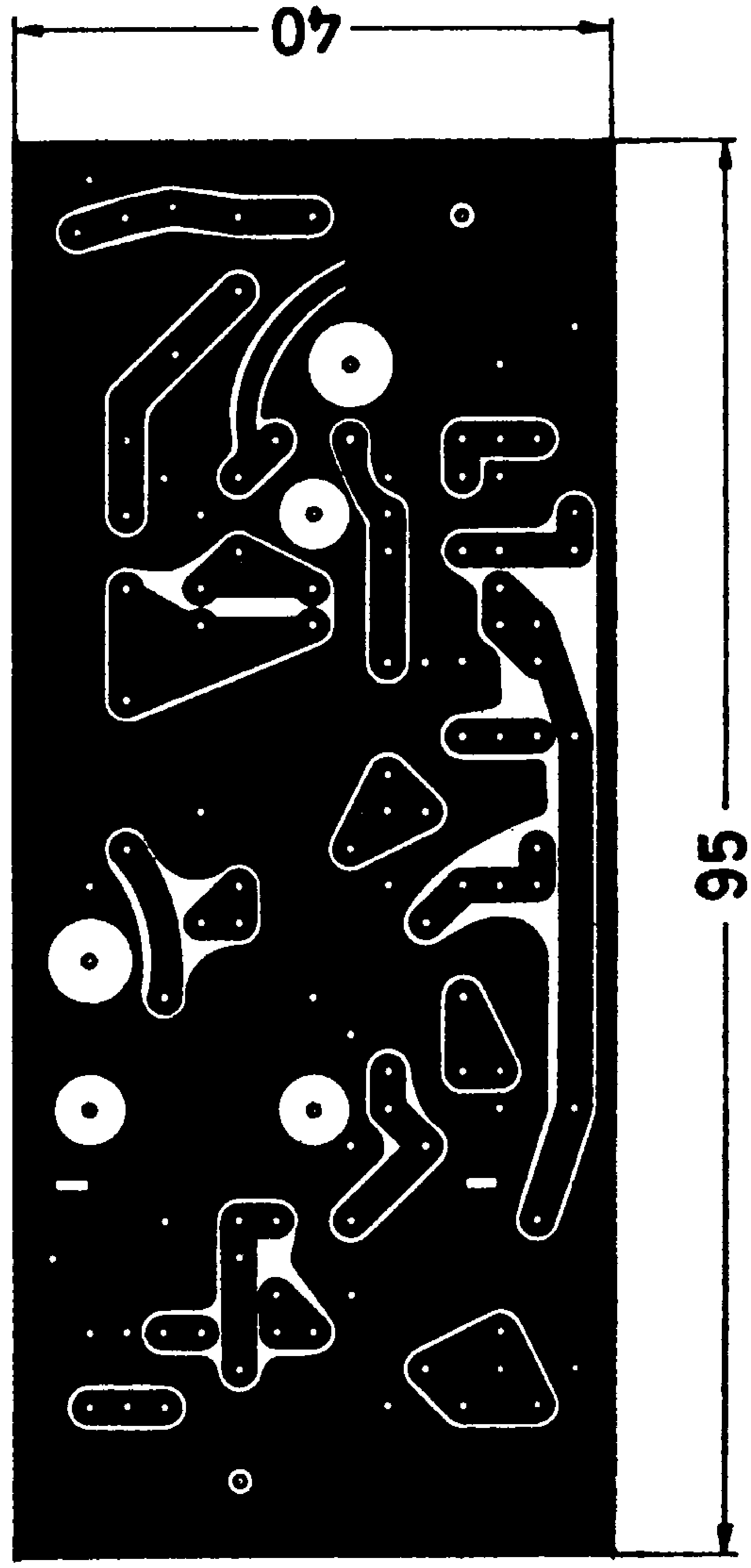
$L_6 = L_7$ = Siferrite thread core B63310-U17-A12.3 15 turns 0.12 CuL



1
L1
18pF
C1
AF 109R
T1
1,9mA
2
C4
1nF
C3
3pF
L2
L3
C5
1nF
C7
1nF
L4
C6
1-6pF
4,7kΩ
R1
R2
7,5kΩ
R3
2,2kΩ
C2
1nF
D1
BB103
68kΩ
R10
68kΩ
R11
D2
BB103
3,9kΩ
R4
AF 106
T2
3,4mA
47pF
6
C11
150pF
C9
C10
150pF
C8
1nF
L6
L7
2,7kΩ
R5
680Ω
R6
L8
L5
3
+12V
330Ω
R13
22nF
C19
22nF
C20
D4
BZY83
D6V8
4
C12
27pF
R8
3pF
C13
1,6mA
2,2kΩ
AF 106
T3
R7
5,6kΩ
C17
1nF
0,5-3pF
C18
2pF
C16
3,3kΩ
R9
1nF
C14
D3
BB 103
R12
68kΩ
5
VC
C21
22nF

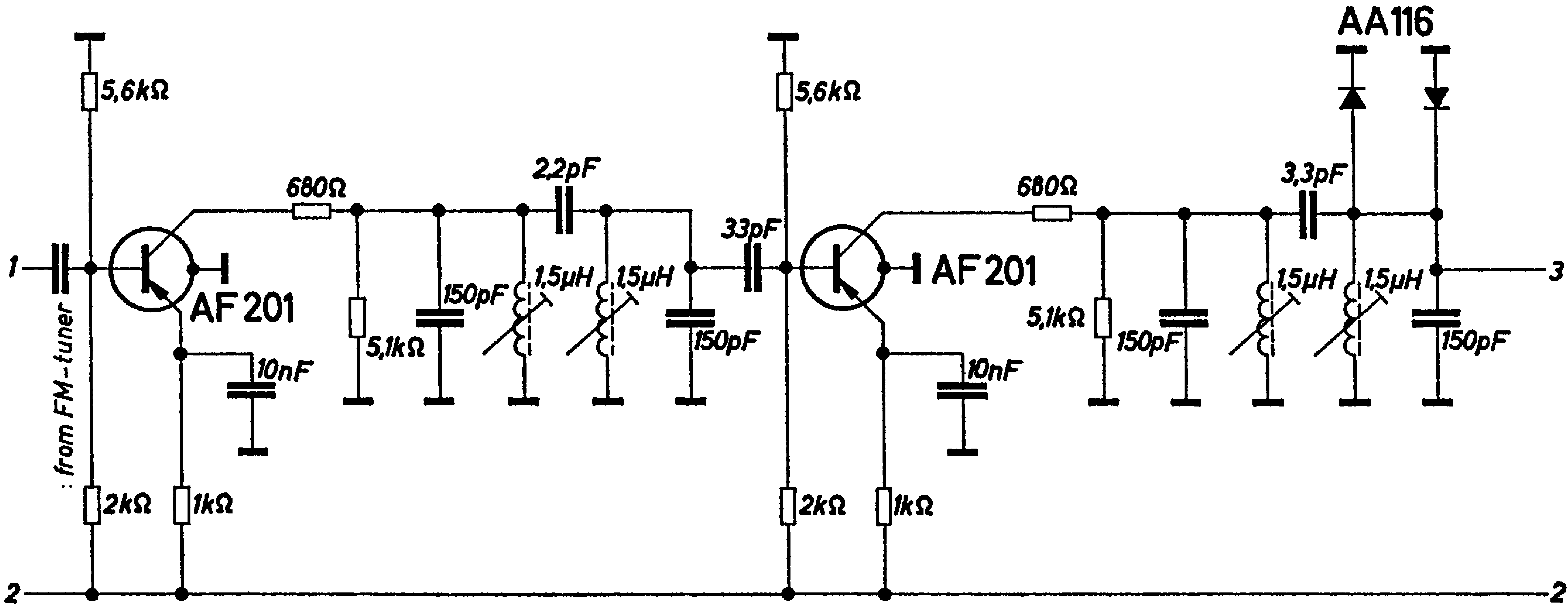
Fig. 8.3.





printed circuit, scale 1:1

Fig. 8.4.



AA116
5,6kΩ
5,6kΩ
680Ω
2,2pF
680Ω
3,3pF
33pF
1
AF 201
AF 201
3
150pF
1,5μH
1,5μH
150pF
5,1kΩ
150pF
1,5μH
1,5μH
150pF
5,1kΩ
10nF
10nF
: from FM-tuner
2kΩ
1kΩ
2kΩ
1kΩ
2
2

Fig. 8.5.

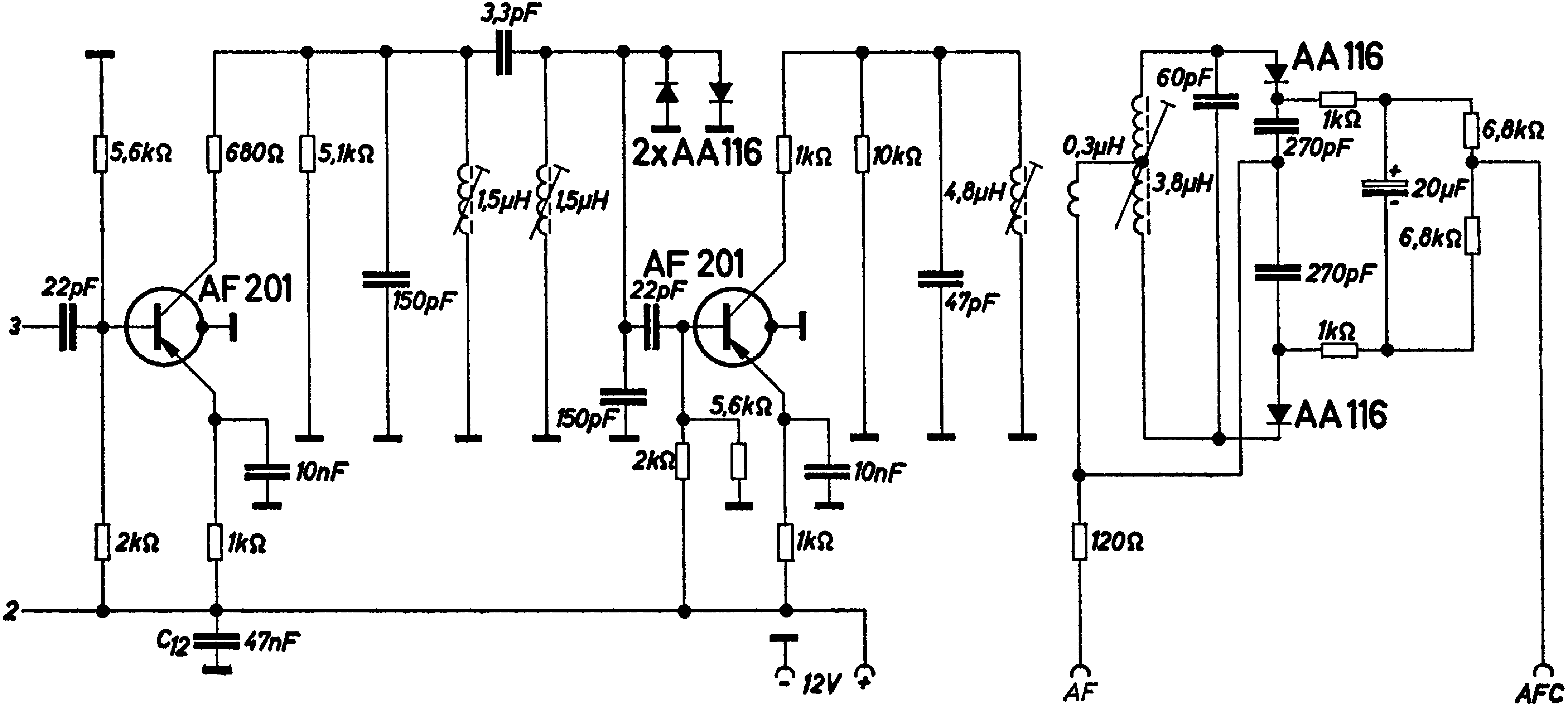

3,3pF
5,6kΩ
680Ω
5,1kΩ
1,5µH
1,5µH
2xAA116
1kΩ
10kΩ
4,8µH
AF 201
22pF
3
150pF
AF 201
22pF
150pF
47pF
10nF
5,6kΩ
2kΩ
10nF
2kΩ
1kΩ
1kΩ
2
C12
47nF
12V
AA116
60pF
0,3µH
3,8µH
1kΩ
270pF
6,8kΩ
20µF
6,8kΩ
270pF
1kΩ
AA116
120Ω
AF
AFC

Fig. 8.5.

8.3. High Quality FM-IF Amplifier

Fig. 8.5. shows a four-stage FM-IF amplifier for 10.7 MHz in which the transistors AF 201 are operated in a common-emitter circuit without neutralization. The signals are limited by four anti-parallel connected diodes at the input of the third and the fourth stage. For demodulation a ratio detector is used with a hump distance of 900 kHz. Since for that reason even at full frequency deviation only a small part of the demodulator characteristic is covered, a small distortion factor is obtained. The four-stage amplifier has been designed with regard to the special stereo requirements. The IF band-width, for example, is 270 kHz in order to obtain small distortion at the LF band-width of 53 kHz required for stereo reception. The resulting unfavorable adjacent-channel selectivity is compensated by a perfect capture ratio due to the good limiting characteristics of the circuit. To achieve a good cross-channel attenuation in the stereo decoder connected to the IF amplifier, it is necessary that the phase delay of the entire RF part (tuner and IF amplifier) is as linear as possible within the pass band. This requirement is fulfilled best by a bell-shaped pass-band curve.

Technical data

Operating voltage		12	V
Operating current		35	mA
Band-width		270	kHz
Total gain		85	db
Limiting sensitivity at 60 Ω for 30 db signal-to-noise ratio		25	μV
Signal-to-noise ratio for an input voltage of 50 μV	>	48	db
AF output voltage at 10 kΩ		40	mV
AF distortion factor at 1 kHz	<	0.5	%

8.4. Stereo Decoder with Silicon Transistors

The stereo decoder has to transform the stereo multiplex signal, which comes from the FM-demodulator, into the AF signals which correspond to both stereo channels. Both channels should have only a small cross-talk and the transmissions are not to be disturbed by audible interference voltages. The circuit of such a decoder is shown in Fig. 8.6.

An impedance transformer stage at the input causes an input resistance of approximately 200 kΩ which does not considerably load the usual FM demodulators. The stereo multiplex signal (MPX signal) coming from the emitter of the transistor of this stage is applied directly to the ring modulator at the decoder output via the secondary coil of the transformer T 3. As to the pilot tone the input stage acts as an emitter stage with a strong feedback. The amplification of this stage results approx. from the relation between the resonant resistance of the collector circuit and the emitter resistance and is, therefore, only slightly influenced by the spread of the transistor parameters. The input resistance decreases only to a small extent for the pilot tone. A biased diode (D_1) attenuates the resonant circuit to a constant amplitude. For that reason variations of the input voltage and also the pilot tone will be eliminated. In spite of the fact that the common base circuit within this frequency range has a smaller amplification than a common emitter circuit, the former has been chosen for the second pilot tone amplifier stage (T 2). This common base circuit has the advantage that the amplification is almost independent on spreads of transistor parameters, for the amplification is only determined by the mutual conductance dependent on the emitter current. The mutual conductance is approx. the same for all usual transistors within this frequency range. The emitter of this stage is coupled via an extremely low-resistive secondary coil of the preceding resonant circuit. This phase of the pilot tone is shifted by the required 90° with the capacitor connected in parallel to the emitter. The base voltage of the transistor is tapped off at the emitter of the first-stage transistor via a resistor. A doubling circuit with two diodes (D_2 and D_3) transforms the pilot tone frequency of 19 kHz at the collector circuit of the second stage to the sub-carrier frequency of 38 kHz. A succeeding RC section increases the efficiency of the doubling circuit. The sub-carrier coming from this RC section reaches the ring modulator via the third transformer (T 3), the secondary coil of which is tuned to resonance.

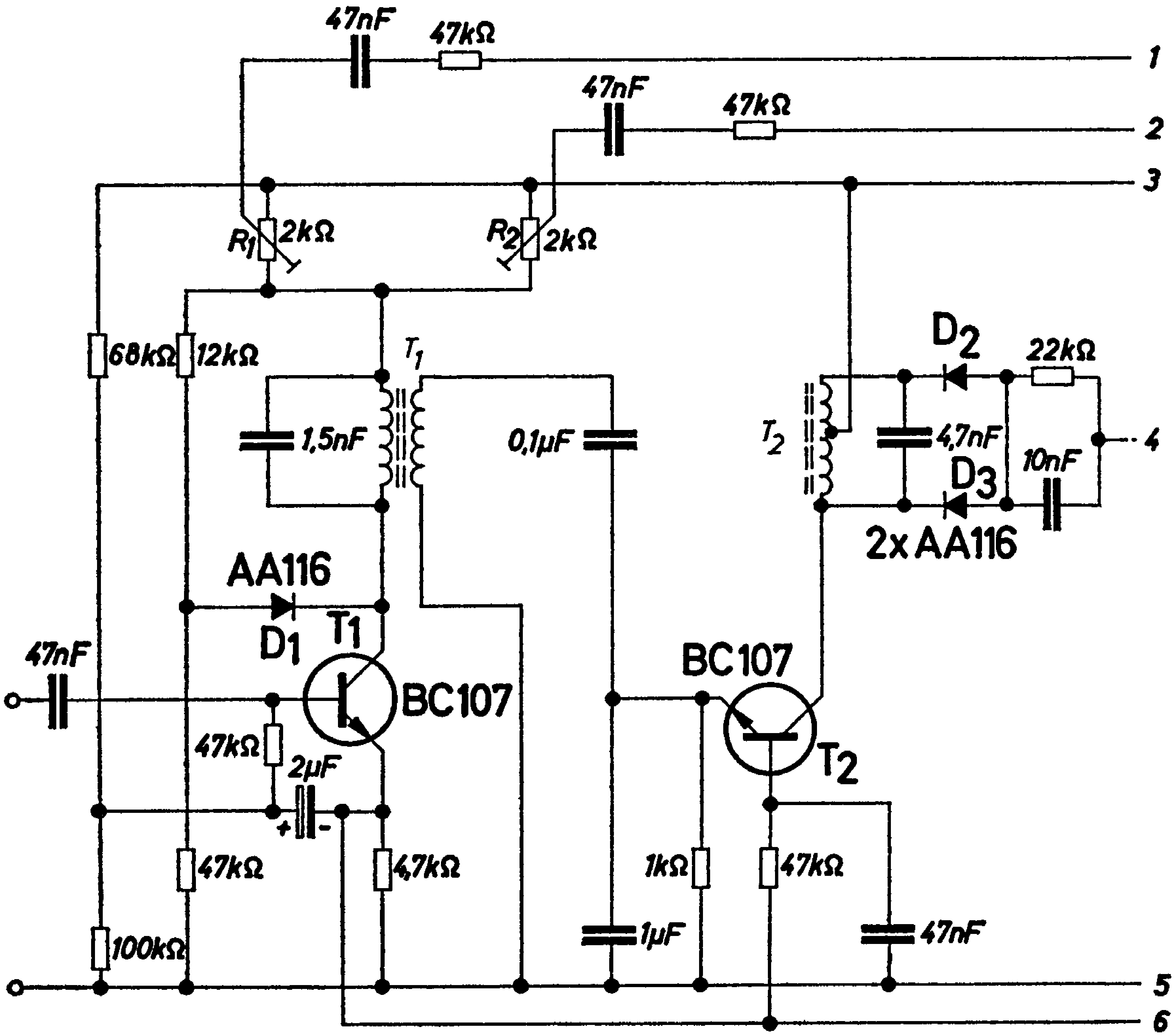


Fig. 8.6.

The ring modulator operates as point-contact rectifier; the sub-carrier voltage is only that high that at the largest permissible MPX signal a distortionless demodulation is still guaranteed. Hereby the noise dependent on the sub-carrier amplitude is kept at an optimal small value. A DC voltage in forward direction increases the efficiency especially at small signals and transfers the RF signal at monophone transmissions via the diodes. For that reason a special mono-stereo switch becomes unnecessary.

At the demodulation of the side-band information its content is decreased to a value corresponding to the demodulation efficiency. Since, however, the relation between the sum signal and the difference signal in the side band has to be preserved for achieving a

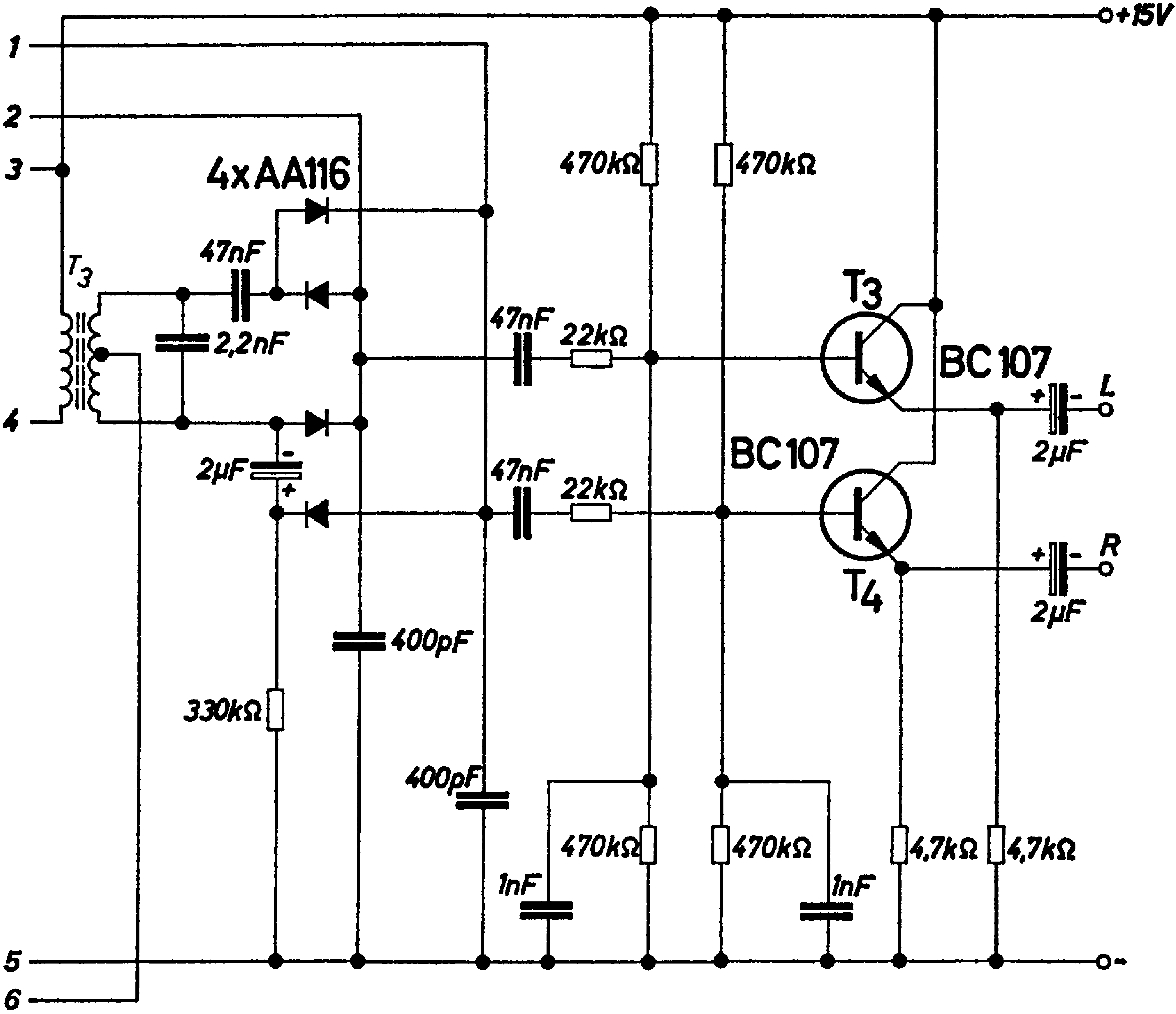


Fig. 8.6.

maximum cross-talk attenuation, the sum signal has to be decreased to a certain extent. Therefore, an antiphase voltage of the sum signal is tapped off at the two potentiometers R_1 and R_2 in the collector circuit of the first stage and led — with adequate magnitude — to the output of the ring modulator. Within the range of a constant demodulation efficiency there can be always achieved a maximum of cross-talk attenuation. The magnitude of the sub-carrier voltage has to be kept as stable as possible because the demodulation efficiency depends on it.

As already mentioned the sub-carrier voltage is kept sufficiently independent from the pilot tone voltage at the input of the decoders by the limiting effect of the diode D_1 parallel to the collector circuit of the first stage. Therefore, the MPX voltage which has a fixed relation to the pilot tone voltage, may vary at the factor 10 without changing the cross-talk attenuation considerably.

Technical data

Operating voltage	15 ± 5	V
Current consumption	approx. 10	mA
Input voltage range (MPX-signal)	approx. 50 to 500	mV
Minimum values of cross-talk attenuation in the range of		
100 Hz to 1 kHz	>30	db
1 kHz to 10 kHz	>40	db
10 kHz to 15 kHz	>35	db
Distortion factor at an input voltage of 500 mV		
Stereo	< 0.6	%
Mono	< 0.5	%
Interference	>45	db
Sub-carrier residual voltage (38 kHz)	< 5	mV
Transmission Attenuation		
Stereo	2	db
Mono	3.5	db
Input resistance	200 kΩ, parallel to 15 pF	
Output resistance with impedance converter	100 Ω, in series to 2 μF	
Smallest permissible load without impedance converter	200	kΩ

8.5. Full Electronically Tuned VHF Tuner

In our booklet 1966 a VHF tuner with diode tuning has been described in details. In that tuner the switching from the lower band to the upper band was still mechanical. Fig. 8.7. shows the circuit of a VHF tuner in which the varicaps BA 138 again have been used for tuning but which is switched electronically. The switching diodes BA 136 which have been especially developed for VHF switching are used for this electronically tunable VHF tuner. The switching diodes BA 136 show a low inductance, required at high frequencies.

A low resistance in forward direction is of special importance for these diodes. If the diode BA 136 is located in the RF input circuit, the RF losses at this resistance influence the noise-to-signal ratio. The diode BA 136 has a forward resistance of about 1 Ω only. Therefore it is perfectly suitable for the switching of high-impedance RF circuits in TV tuners with tuning diodes.

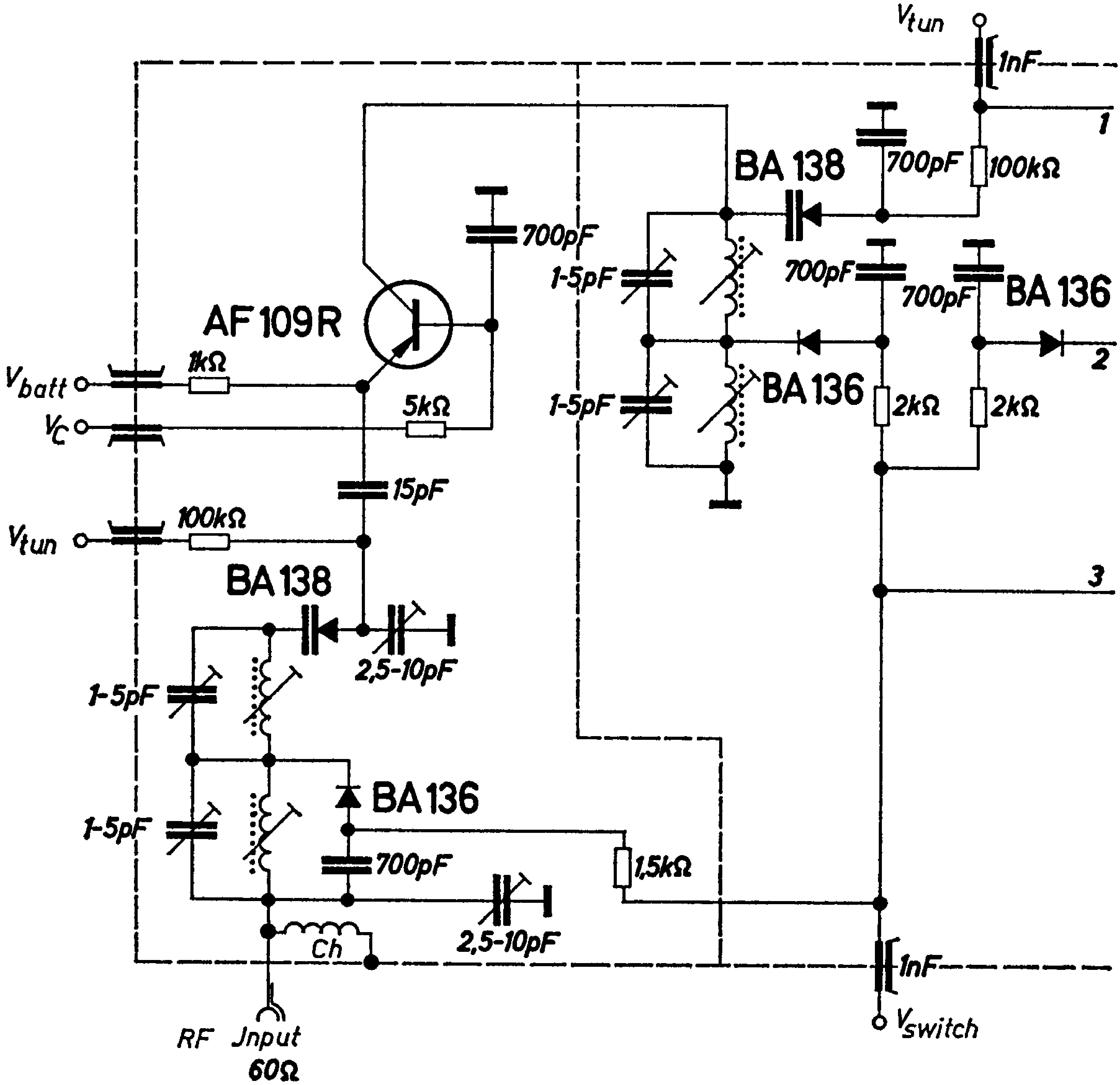
With these diodes, therefore, an electronically tunable VHF tuner can be built which shows no electronic disadvantages in comparison to a mechanically switched tuner. The electronic VHF tuner, however, offers many advantages as, e.g., free choice of the position of the switching units, independence from control levers and lever systems, easy remote control, and high switching reliability.

As in line-operated equipments the current consumption is of small importance, the operating current for the switching diodes can be set according to the requirements of the circuit. It has to be taken into consideration, however, that the noise of the diodes becomes higher with increasing currents. In this circuit a current of 6 mA is recommended for the diodes used in the intermediate pass-band filter and oscillator circuits. For the input circuit, however, a current of 8 mA has been better. For the design of the circuits for the lower band it is very important that these diodes also show a highly voltage-dependent capacitance. In order to keep this capacitance low in the reverse state, the diode has to be operated with a possibly high reverse voltage. At a reverse voltage of 30 V the capacitance has a value lower than 2 pF, which is transformed into the circuit. This bias voltage for the diodes can be tapped off from the stabilized supply voltage for the tuning diodes via a high resistance of, e.g., 100 Ω.

Because of the high series resistor for the switching diodes the load is small. The switching to the upper band occurs with a low-resistive voltage of +12 V. As the reverse voltage is connected via a high series resistance, it is not necessary to disconnect this voltage.

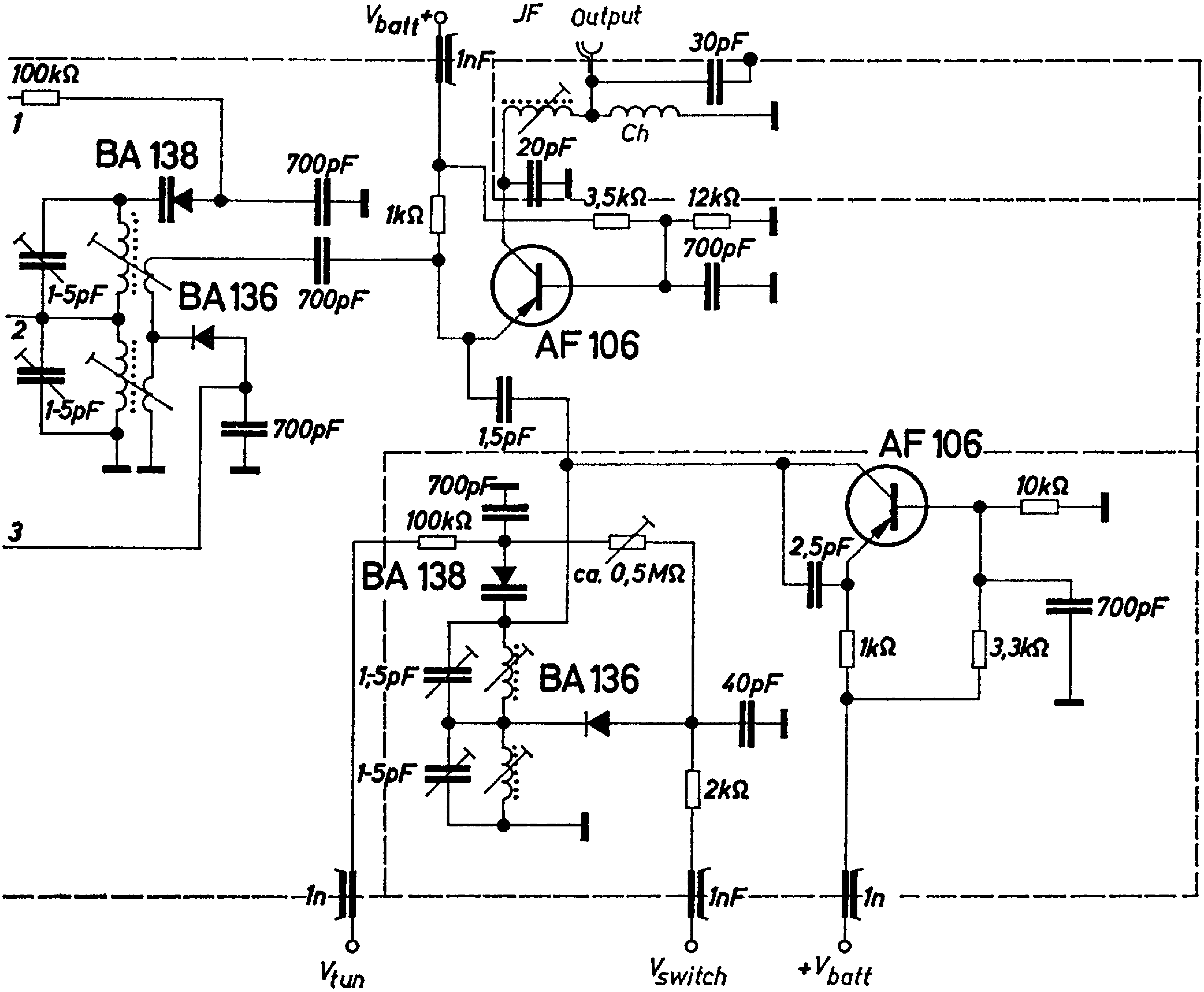
At the input of the VHF tuner a tunable RF circuit is used. A π-circuit has been chosen as this circuit corresponds at best to the requirements resulting from the switching diodes. Besides, the range switching can be achieved by one single switching diode at the input. The application of this simple circuit, however, makes a compromise necessary with regard to selectivity and band-width.

The π-section is coupled to the pre-stage transistor AF 109 R via a capacitor of 15 pF. The control voltage which is applied to the base of this transistor has to be determined in such a way that in the downward controlled state an emitter current of 2.3 mA will flow.



Vtun
1nF
1
BA 138
700pF
100kΩ
700pF
AF 109R
1-5pF
700pF
700pF
BA 136
700pF
2
Vbatt
1kΩ
1-5pF
BA 136
2kΩ
2kΩ
VC
5kΩ
15pF
Vtun
100kΩ
3
BA 138
2,5-10pF
1-5pF
BA 136
1-5pF
700pF
1,5kΩ
2,5-10pF
Ch
1nF
Vswitch
RF Jnput
60Ω

Fig. 8.7.



Vbatt+
JF
Output
1nF
30pF
100kΩ
1
Ch
BA 138
20pF
700pF
3,5kΩ
12kΩ
1kΩ
700pF
1-5pF
BA 136
700pF
2
AF 106
1-5pF
700pF
1,5pF
AF 106
700pF
10kΩ
100kΩ
3
2,5pF
BA 138
ca. 0,5MΩ
700pF
1kΩ
3,3kΩ
1;5pF
BA 136
40pF
1-5pF
2kΩ
1n
1nF
1n
Vtun
Vswitch
+Vbatt

Fig. 8.7.

The pre-stage of the tuner is decoupled from the mixer stage by the wall of a cavity resonator in order to prevent the interference of the oscillator voltage.

The band filter switched to the respective range by two switching diodes is located in the cavity resonator for the mixer stage. At reception in the upper band a third switching diode short circuits a part of the coupling coil to the mixer and hereby avoids losses of the oscillator- and the received voltage. A mesa transistor AF 106 in common-base circuit is used as mixer transistor. Its operating current is to be approximately 2 mA.

A transistor AF 106 in common-base configuration is also used in the oscillator circuit. The only difference of the circuit compared to the usual circuit is an electronic reduction of the oscillator variation in the lower band. This reduction is achieved by applying once more the tuning voltage of —30 V, which in the lower band operation also is applied to the reverse biased switching diode via 0.5 Ω to the tuning diode of the oscillator. For the switching to the upper band a small positive voltage is applied to the BA 136 diode operated in forward direction. This voltage does not considerably influence the real tuning voltage via 0.5 Ω.

Fig. 8.8. and 8.9. show the power gain, the input reflection and the noise figure in the lower band and in the upper band.

Technical data

Operating voltage	12	V
Tuning voltage	-3 to 30	V
Switching voltage	-30 and 12	V
Power gain	20 to 26	db
Noise figure	4 to 6	

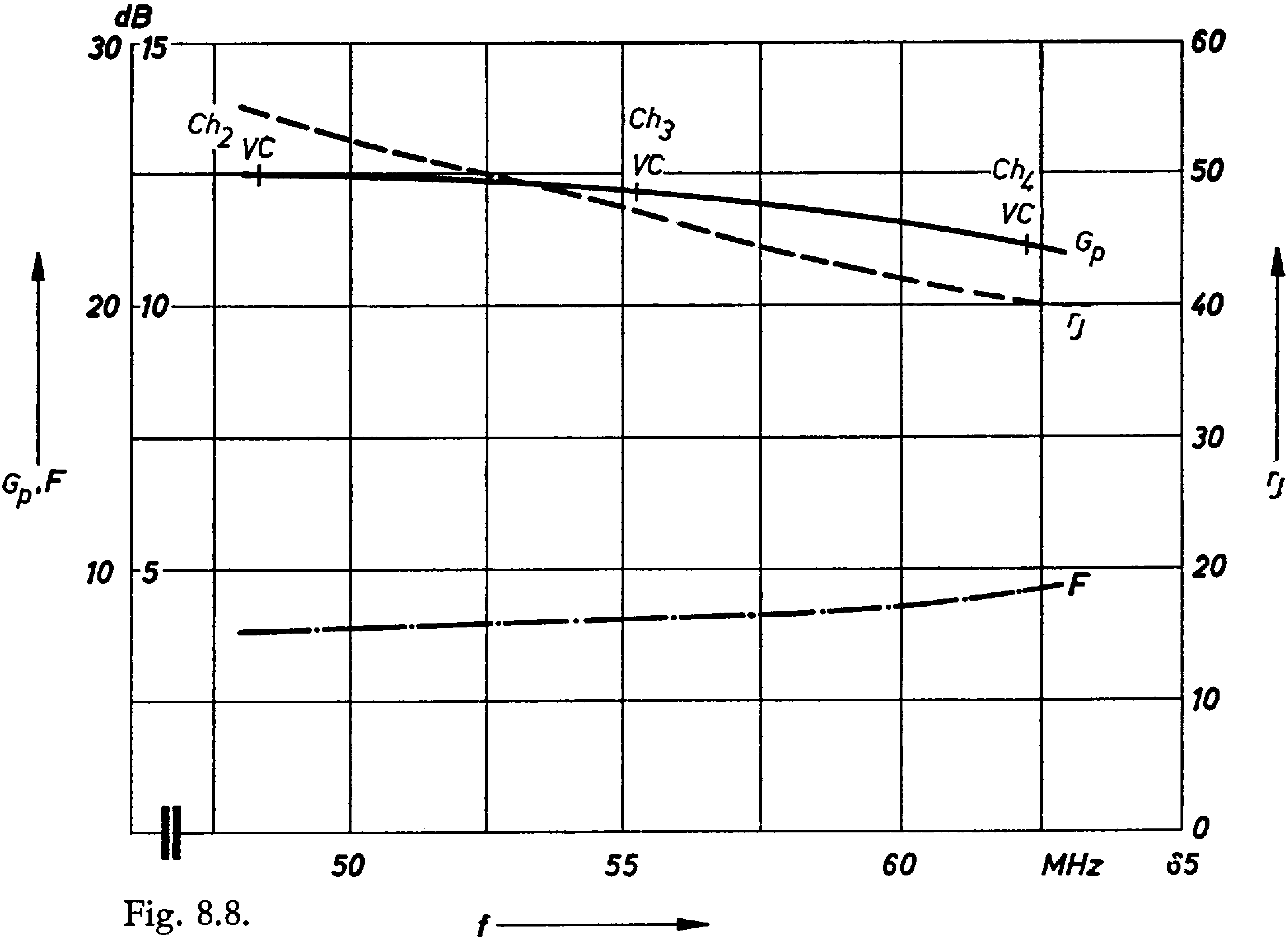

dB
30
15
20
10
10
5
Gp,F
60
50
40
30
20
10
0
rJ
Ch2
VC
Ch3
VC
Ch4
VC
Gp
rJ
F
50
55
60
MHz
65
f

Fig. 8.8.

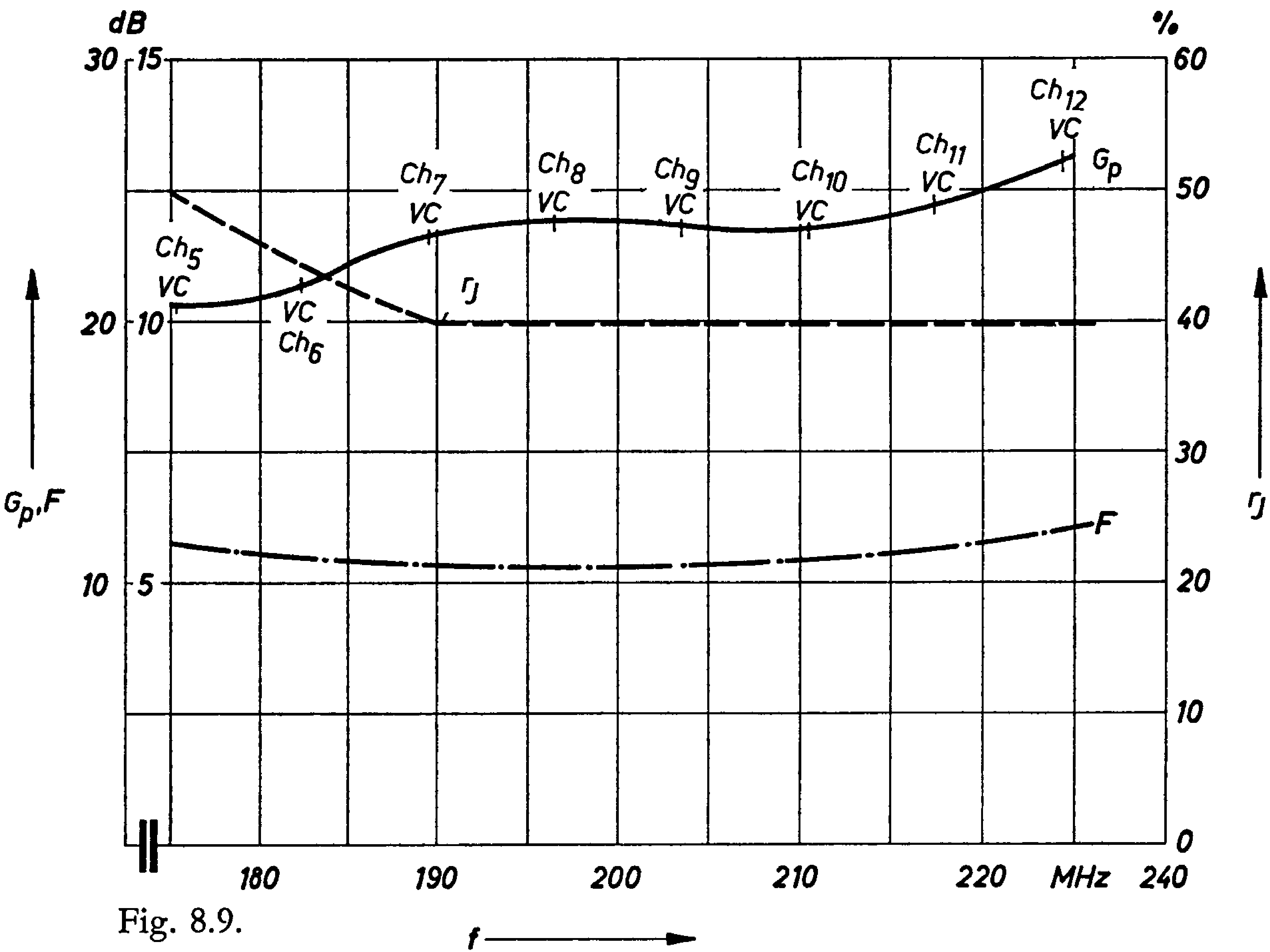

dB
%
30
15
20
10
10
5
Gp,F
60
50
40
30
20
10
0
rJ
Ch5
VC
VC
Ch6
Ch7
VC
Ch8
VC
Ch9
VC
Ch10
VC
Ch11
VC
Ch12
VC
Gp
rJ
F
180
190
200
210
220
MHz
240
f

Fig. 8.9.

8.6. Color Video Circuit in the RGB Concept

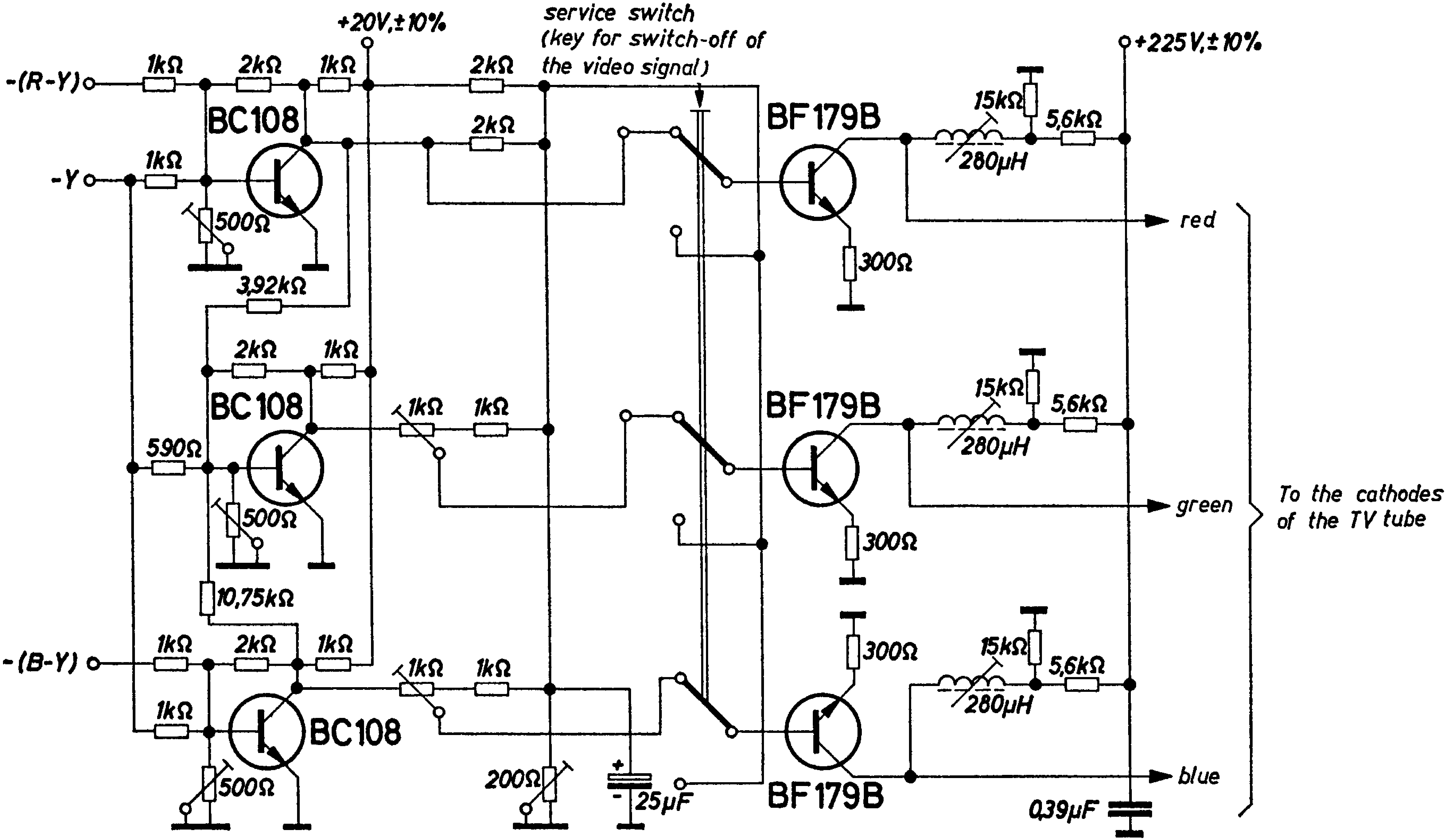
While in color video circuits according to the color difference concept the color information (R-Y), (G-Y), and (B-Y) is amplified separately from the brightness information Y, the color information according to the so-called RGB concept is amplified together with the brightness information. The drive of the color video tube is achieved at the respective cathodes of the three-beam control systems.

For that purpose the original color components R, G and B (red, green and blue color signals) have to be regained from the transmitter informations Y, (R-Y) and (B-Y) in a matrix circuit. Principally this matrix circuit could be passive, in other words it could consist of an adequately dimensioned resistance network. A precisely operating passive matrix circuit free of coupling effects causes a decrease of the signal amplitude, so that in Fig. 8.10. an active matrix circuit has been used. The active matrix circuit has good stability characteristics, extremely low mutual coupling, an uncomplicated adjustment and a low output impedance. It operates according to the principle of the adding amplifier already known in the analog computer technique and is designed with 3 transistors BC 108. The high current amplification of these transistors is of special advantage in the circuit because hereby the magnitude of the output DC voltage becomes practically independent from the operating temperature.

The signals R, G, and B obtained in the matrix circuit are amplified in three video output stages with the transistors BF 179 B. These transistors show in addition to a high permissible reverse voltage a special resistance towards tube flashing. The effective video driving voltage (picture blanking signal) is 105 V. With this voltage the desired total beam peak current of the color tube can be attained under all circumstances. A picture blanking signal of 105 V corresponds to a peak-to-peak video signal including the sync pulse of $105 \cdot 1.38 = 145V$ (picture blanking signal with sync pulse). Then the sync pulse is 40 V. Since the tube at the grid 2 or grid 1 is blocked during the fly-back retrace period by blanking pulses, it is not necessary that the sync pulse in the video amplifier is absolutely linearly transmitted. At a permissible sync pulse suppression of 50%, a picture blanking signal with sync pulse of 125 V is attained the amplifier of which is shown in Fig. 8.10.

Adjustment of the illuminant: In order to adjust the video amplifier together with the color tube for the correct reproduction of a black-

Fig. 8.10.



+20V,±10%
service switch
(key for switch-off of
the video signal)
+225V,±10%
-(R-Y)
-Y
-(B-Y)
1kΩ
2kΩ
1kΩ
2kΩ
2kΩ
1kΩ
BC108
500Ω
3,92kΩ
2kΩ
1kΩ
BC108
590Ω
1kΩ
1kΩ
500Ω
10,75kΩ
1kΩ
2kΩ
1kΩ
1kΩ
1kΩ
1kΩ
BC108
500Ω
200Ω
25μF
BF179B
BF179B
BF179B
300Ω
300Ω
300Ω
15kΩ
5,6kΩ
280μH
15kΩ
5,6kΩ
280μH
15kΩ
5,6kΩ
280μH
0,39μF
red
green
blue
To the cathodes
of the TV tube

white signal through the total reproduced grey scale, video control circuits have been mounted between the matrix circuit and the video output amplifiers. A potential is adjusted at the common point of the 3 cross-branches by means of a voltage divider; this potential corresponds to the black-level potential at the collector side of the three matrix amplifiers. From this results the fact that the three cross-branches for the black level potential are without current and that a change of the potentiometer adjustment or the video signal does not cause a variation of the black level. Three further switches mutually coupled have been arranged. They make it possible that the video signals can be switched off without changing the black level for alignment of the rejection point.

The color tube can only be adjusted in such a way that at first for switched-off video signals the three tube characteristics are set to the same rejection point by the U_{g_2} voltages at the tube (compensation of reverse voltage spreads) and after switching by adjusting the video signals G and B with regard to R (at a constant black level) the spreads of the fluorescent material efficiency are compensated.

The signal amplitude of the R signal has been determined to obtain at any time the highest possible tube beam current.

This kind of adjustment of the illuminant is very simple. The output transistors have to be operated with a mounted heat sink of $R_{th} \leqq 15\,°C/W$.

Technical data

Operating voltage	225 V and 20 V (± 10%)
Picture blanking signal with sync pulse	105 V