SIEMENS

Design Examples of Semiconductor Circuits

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SIEMENS AKTIENGESELLSCHAFT

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Similar applications have been grouped in chapters to offer a good survey.

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Contents

1. RF- and AF-Circuits

1.1. Broadband Aerial-Amplifier using 2 × BFT 66

In the frequency range up to 1 GHz the broadband-transistor BFT 66 is especially applicable in low-noise IF-stages, preamplifiers and broadband amplifiers with low output power. Due to the fact that the transistor has a hermetically sealed metal-case TO-72 and on account of the multi-layer metalisation using platinum and gold it can also be recommended for professional applications.

The BFT 66 offers the following features:

- a high punch-through stability, and by that a high current gain as well as a low current dependent part of the noise figure,
- a high transition frequency and therefore a high power gain as well as a low noise figure to upper end of the UHF-range.

The BFT 66 has a noise figure of 1 dB e.g., and the maximum power gain is 12 dB at 800 MHz.

Noise Behaviour

The noise of bipolar transistors is particularly determined in the range of white noise by the thermal noise of the base resistor, by the recombination noise of the basic-emitter diode and the shot noise of the collector diode. The latter part is responsible in general for the current dependence of the noise figure. Therefore the transistor should be operated at low collector currents. But this is not feasible, since the achievable transmittance and linearity are not sufficient for broadband applications. Thus reasonable values for the collector current are in the range of 2 and 5 mA.
With the transistor BFT 66 a minimum noise figure not higher than 1 dB is achievable at $f = 10$ MHz and collector currents of 3 mA. This is equivalent to a temperature difference of a noisy resistor of only 77 K. The user of the BFT 66 will appreciate the fact that the noise figure has its minimum at 75 Ω and it increases only slightly if a noise mismatching occurs, or if the collector current is higher than the above mentioned. Due to the high transition frequency the excellent noise characteristics are maintained in the total UHF-range.
Fig. 1.1.1. shows a one-stage circuit. It is operated at a collector current of 4 mA and the power gain is 22 dB in a frequency range of 1 to 300 MHz (−3 dB). The noise figure is 1 dB at $R_G = 60\,\Omega$, $f = 25$ MHz and 1.6 dB at $f = 300$ MHz.
The high power gain values are obtained as a result of the higher transition frequency, which is already 2 GHz at a collector current of 3 mA. The flat maximum with $f_T = 4$ dB typical, is achieved at $I_C = 25$ mA.
The collector capacitance being proportional to the base area negatively influences the amplification characteristics as well as the resultant internal feedback. By the minimum distances between the stripes of the collector structure a small capacitance is obtained, i.e. the feedback has a low value.
In **Fig. 1.1.2.** a 2-stage amplifier is shown. Its frequency range is 25 to 1000 MHz, the VSWR at input and output is not greater than 2.
In the broadband operation of the transistor its dynamic range is extended below the signal limitation created by non-linear distortions and the specified noise-to-signal ratio. If a circuit with more than one stage is used, by this feature, the distortion part of the prestage will be extremely low.

The range of the greatest linearity begins at collector currents of 8 mA. An output voltage of 180 mV is achieved at an intermodulation attenuation of 60 dB.

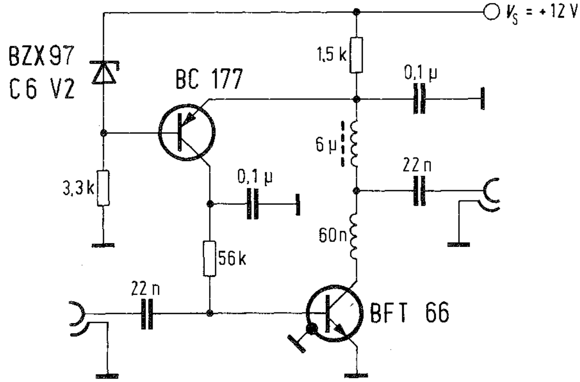


Fig. 1.1.1.

List of Capacitors used in the Circuit 1.1.1.

Quantity	Components	Ordering codes
2	0.1 µF/63 V ceramic cap.	B 37449–A 6104–5001
2	0.022 µF/16 V ceramic cap.	B 37305–A 1223–7001

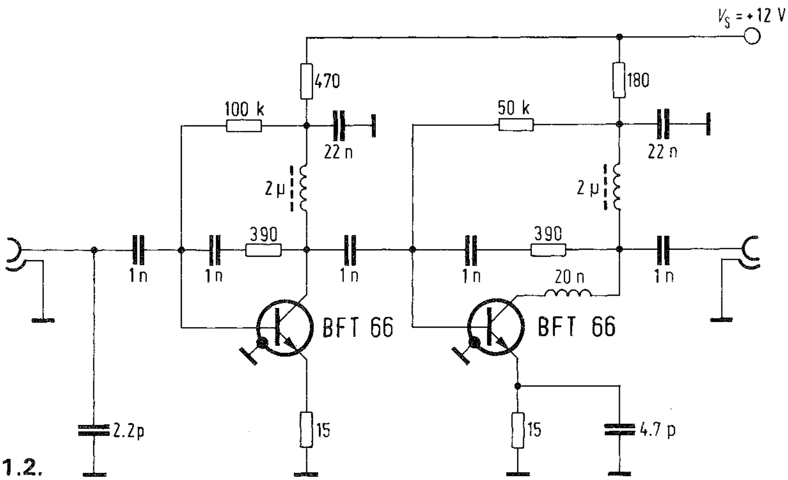


Fig. 1.1.2.

List of Capacitors used in the Circuit 1.1.2.

Quantity	Components	Ordering codes
2	0.022 µF/63 V ceramic cap.	B 37449–A 6223–S 001
5	1000 pF/500 V ceramic cap.	B 37235–J 5102–S 001
1	2.2 pF (2 pF)	B 38112–J 5020–C 000
1	4.7 pF (5 pF)	B 38112–J 0550–D 000

The great dynamic range of the BFT 66 is demonstrated by the following test results which were achieved with the broadband amplifier circuit according to **Fig. 1.1.2.**: Single-tone-control: $f = 200$ MHz, $R_G = 50\,\Omega$, attenuation of harmonics 30 dB, output voltage $V_0 = 500$ mV.
Two-tone-control: $f_1 = 798$ MHz, $f_2 = 802$ MHz, $R_G = R_L = 50\,\Omega$, intermodulation attenuation 60 dB, output voltage $V_0 = 170$ mV.

1.2. Medium Wave Receiver using TCA 440

The TCA 440 is a monolithic IC characterized by a high integration. It is especially applicable with regard to AM-receivers, battery or mains operated. This IC contains a symmetrical prestage with AGC, a balanced mixer, an oscillator and an IF amplifier with AGC. The controlled prestage assures an excellent large signal behaviour, e.g. a voltage of 2.6 V_{pp} at the input of the IC is processed without any distortions. Since the push-pull mixer operates multiplicatively, only a few harmonic mixing products are generated. The oscillator is separated from the mixer and oscillates up to the short wave range. Both prestage and IF amplifier are controlled independently. Thus an AGC-range of about 100 dB is achievable.

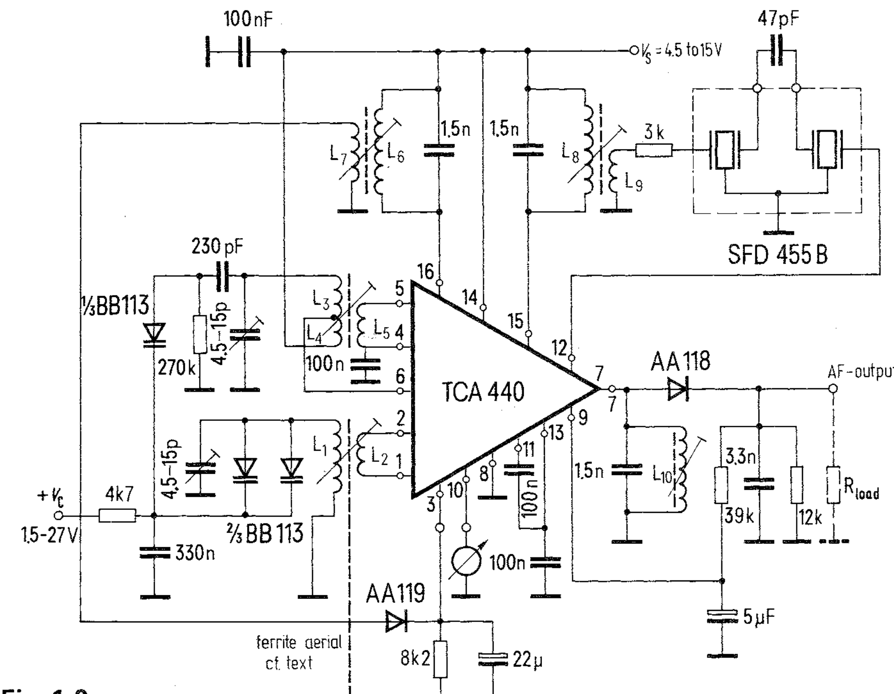


Fig. 1.2.

The symmetric composition of the IC has many advantages. The direct influence of the IF signal is avoided and a high stability against wild oscillations is obtained.
The AGC-amplifier of the IF-stage supplies a tuning voltage of max. 600 mV. It is available at pin 10 of the IC, having an internal impedance of about 400 Ω and can be indicated by a moving-coil instrument (500 µA, $R_i = 800\,\Omega$ or 300 µA, $R_i = 1500\,\Omega$).

Because of the internal stabilization, the supply voltage may change within a wide range (4.5 to 15 V), but the electrical characteristics are guaranteed, even if the IC is used in portable receivers. Over the total tuning range excellent tracking is achieved by the triple-capacitance diode BB 113 which consists of three equal systems being on the same chip. To realise the required capacitance ratio of 1 to 10 two diode-systems, connected in parallel, are used in the prestage because of the existing parallel capacitances of the circuit.

Characteristics

Supply voltage		4.5 to 15 V	
Supply current	at	4.5 V	7 mA
	at	9 V	10.5 mA
	at	15 V	12 mA
Frequency	at V_c =	1.5 V	520 kHz
	at V_c =	27 V	1640 kHz

Signal-to-noise ratio*) $\frac{S+N}{N}$	required field intensity	V_{AF} at $R_L = 10\ k\Omega$
10 dB	$2.1 \times 10^2\ \mu V/m$	24 mV_{rms}
26 dB	$1.4 \times 10^3\ \mu V/m$	35 mV_{rms}
40 dB	$7.8 \times 10^3\ \mu V/m$	43 mV_{rms}

*) Signal-to-noise ratio is measured according to
DIN 43500 appendix 1 at $f = 600$ kHz   m = 0.3 $f_{mod} = 1$ kHz

Coil data

Ferrite aerial

L_1 =	43 turns	12 × 0.04 mm ∅	enamelled copper wire, silk covered, one layer
L_2 =	6 turns	0.1 mm ∅	double-enamelled copper wire (over L_1, at the end being connected to earth)

on Siemens aerial rod B 61 610–J 1017–X 025

Oscillator coil

L_3	110 turns	0.1 mm ∅	double-enamelled copper wire with Vogt filter kit D 41-2519
L_4	15 turns	0.1 mm ∅	
L_5	22 turns	0.1 mm ∅	

The Q-factor of L_3/L_4 is to be less than 100 at 975 kHz

L_3	beginning	hot end of the circuit
	end	pin 6 of TCA 440
L_4	beginning	pin 6 of TCA 440
	end	V_s
L_5	beginning	pin 5 of TCA 440
	end	pin 4 of TCA 440

IF-Circuit

Tuner-circuit coil	L_8	70 turns	12 × 0.04 mm ∅, stranded, enamelled copper wire, silk covered
		L = 80 μH, Q = 118	
Coupling coil	L_9	10 turns	12 × 0.04 mm ∅, stranded, enamelled copper wire, silk covered

Demodulator circuit

Tuner-circuit coil	L_{10}	70 turns	12 × 0.04 mm ∅, stranded, enamelled copper wire, silk covered
		L = 80 μH, Q = 118	

IF-control circuit

Tuner-circuit coil	L_6	70 turns	12 × 0.04 mm ∅, stranded, enamelled copper wire, silk covered
		L = 80 μH, Q = 118	
Coupling coil	L_7	50 turns	12 × 0.04 mm ∅, stranded, enamelled copper wire, silk covered

With the exceptions of L_1 and L_2 all coils have been wound on Vogt filter kit D 41-2519.

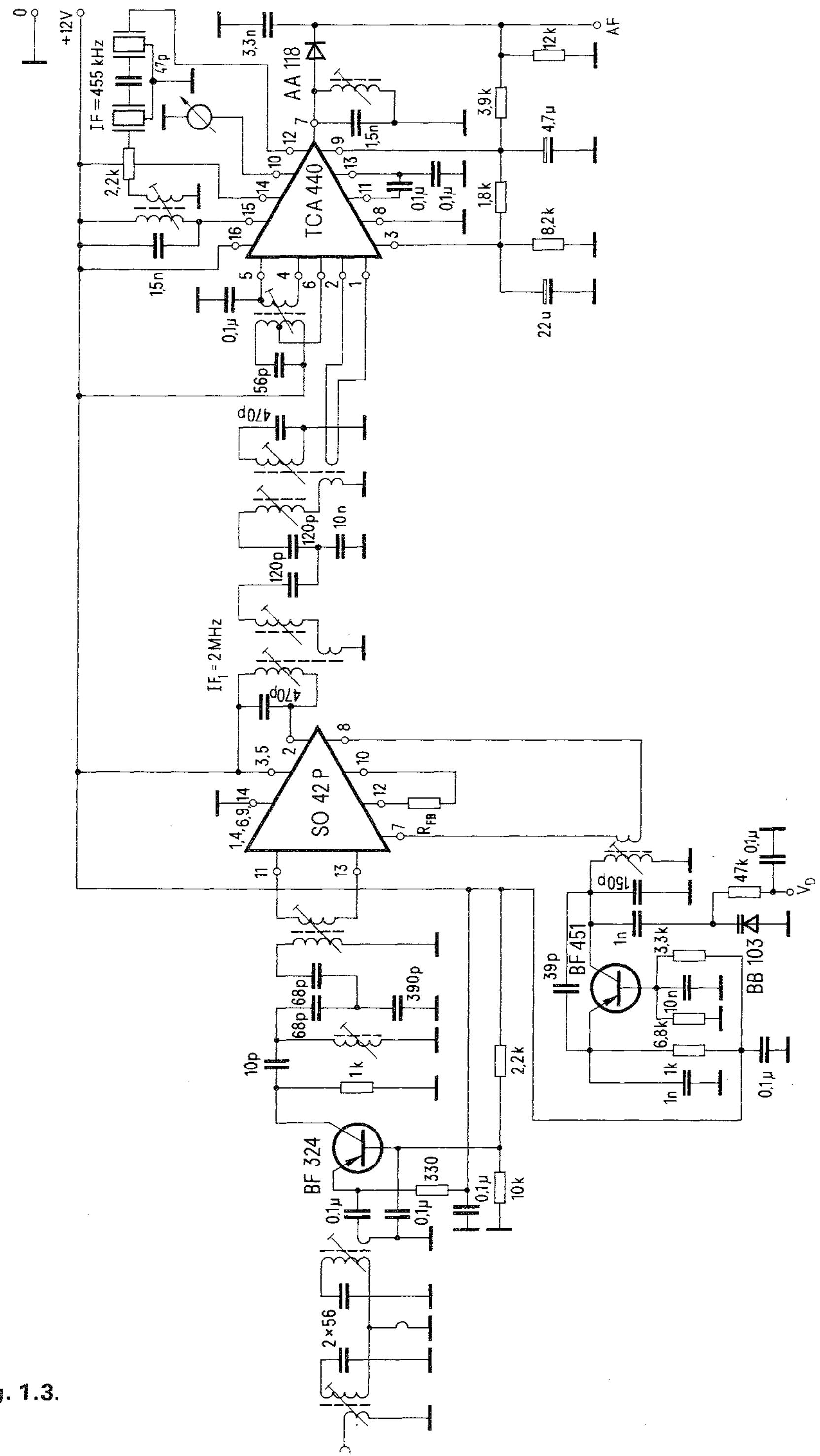
The ceramic filter SFD 455 B is produced by Stettner Corp.

List of Capacitors used in the Circuit 1.2.

Quantity	Components	Ordering codes
4	0.1 μF/63 V ceramic cap.	B 37449–A 6104–S 001
1	0.33 μF/100 V MKM-stacked film cap.	B 32541–A 1334–J 000
1	3.3 μF/25 V styro.	B 31310–A 3332–H 000
3	1.5 nF/25 V styro.	B 31310–A 3152–H 000
1	4.7 pF (5 pF) cer.	B 38112–J 5050–D 000
1	220 pF (230 pF) styro.	B 31310–A 3221–H 000
1	22 μF/40 V electrolytic cap.	B 41286–A 7226–J
1	4.7 μF (5 μF) 63 V electrolytic	B 41315–A 8475–Z
2	trimmer capacitor 4.5 to 15 pF	—

1.3. Double Heterodyne Receiver for the Short-Wave Band

Receivers available on the market today with short-wave bandswitching operate with tuned RF circuits and an AGC before the first mixer. The selection achieved by two tuned r.f. circuits is limited compared with the bandwidth of the short-wave bands. Therefore, strong interfering stations of the total band are not suppressed sufficiently and crossmodulations are generated in the first mixing stage. An AGC of the RF will be without any effect regardless of where the control information is tapped down. If the bandwidth before the first mixer is controlled the AGC will react when signals of both a weak and strong station are received. However, the weak signal will be faided out by the noise. If the AGC is derived from a stage which is behind the IF-selection circuit, no control voltage will be generated in this



+12V
0
AF
IF = 455 kHz
AA 118
TCA 440
IF1 = 2 MHz
SO 42 P
RFB
BF 324
BF 451
BB 103
VD

Fig. 1.3.

case, i.e. the RF stage as well as the mixer will be overdriven and crossmodulations are created. By increasing the selectivity the difficulties will be diminished, but this measure is limited by the achievable coil quality. The only alternative is to improve the dynamic range of the RF and first mixing stage. This is demonstrated at the following circuit shown in **Fig. 1.3.**

The band selection before the first mixing stage is arranged by two band-pass filters. They are coupled by a non-controlled transistor BF 324, which compensates the filter loss. It operates in a common-base circuit to guarantee a wide dynamic range. For the first mixing stage the IC S 042 P is used. It is linearized by a negative feedback resistor. The RF-signal is supplied to the bases of those transistors responsible for the current impression. Their emitters are connected by the negative feedback resistor, R_{FB}, determining the large signal behaviour of the mixing stage. This resistor should not have too high a value since it increases the noise figure and by that a higher gain of the prestage is required.

The signal generated in the separate oscillator is supplied to the four transistors, which are arranged as cross-coupled differential amplifier. They operate as switches and thus only a few distortions are produced.

Behind the mixing stage a four-section-filter is connected. It is responsible for the first IF-selection. The IC TCA 440 is provided for the second mixing stage and the second IF-amplifier.

List of Capacitors used in Circuit 1.3

Quantity	Components	Ordering codes
8	0.1 μF/63 V cer.	B 37449–A 6104–S 001
2	1000 pF/500 V cer.	B 37235–J 5102–S 001
2	10 nF/250 V cer.	B 37452–A 2103–S 008
1	39 pF/250 V cer.	B 38612–J 2390–J 008
1	150 pF styro.	B 31310–A 5151–H 000
2	68 pF styro.	B 31310–A 5680–H 000
2	470 pF styro.	B 31310–A 5471–H 000
2	120 pF styro.	B 31063–A 5121–H 000
3	56 pF/63 V styro.	B 31310–A 5560–H 000
1	47 pF/250 V cer.	B 38612–J 2470–J 008
2	1500 pF/250 V styro.	B 31310–A 3152–H 000
1	4.7 μF/63 V electrol.	B 41315–A 8475–Z 000
1	22 μF/40 V electrol.	B 41286–A 7226–F 000
1	10 pF/500 V cer.	B 38110–J 5100–D 000
1	390 pF cer.	—

1.4. Receiver using TCA 440–27 MHz

Fig. 1.4. shows the RF-circuit of a 27 MHz-receiver. By using the IC TCA 440 a very compact and a well arranged device can be attained. Because of the excellent large-signal features of this prestage a distortion-free operation is possible, even if more than one receiver is running at the same time.

The aerial is loosely coupled to the front end circuit to achieve a good selectivity as well as to diminish the detuning created by different aerial lengths. The main selection of the IF section is achieved by a cheap ceramic filter. Whereas the far-off selection is obtained by an additional LC-circuit. Another tuned circuit connected to the output of the IC suppresses the broad-band noise of the IF-amplifier. The AGC-voltage is generated by a germanium diode. The control information for the RF-stage is derived from the IF-control. An emitter follower is connected to the positive end of the tuned output circuit. It provides a low impedance to drive the following modulator transistor well, the base of which is clamped by an Si-diode in such a way that without signal no collector current will flow. If an RF-signal is applied, a voltage drop will occur at the collector resistor during the positive half-wave.

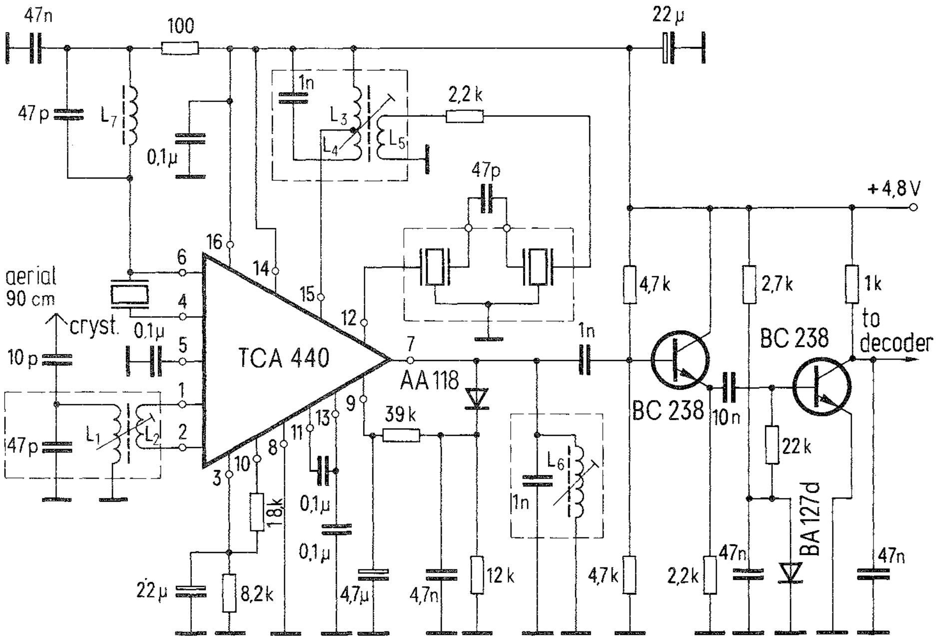


Fig. 1.4.

The information of the transmitter signal is determined by its pulse spaces. During the pulse separation the level at the collector rises to the value of the supply voltage. Thus the demodulated signal is available as a train of needle pulses at the collector. It can be decoded by the following decoder.

Characteristics

Supply voltage	4.8 V
Supply current	11 mA
Input voltage required for a stable output signal at the decoder output	3 µV

Crystal: Receiver crystal for the 27 MHz-band.

Coil data

L_1	11 turns	0.15 enamelled copper wire
L_2	5 turns	0.15 enamelled copper wire
L_1, L_2	with Vogt coil kit D 71–2499.1 without tub core	
L_3	27 turns	3 × 0.05 enamelled copper wire, silk covered
L_4	68 turns	3 × 0.05 enamelled copper wire, silk covered
L_5	28 turns	3 × 0.05 enamelled copper wire, silk covered
L_6	95 turns	3 × 0.05 enamelled copper wire, silk covered
L_3 to L_6	with Vogt coil kit D 71–2498.1	
L_7	6 turns	0.2 enamelled copper wire wound on tubular core B 62110–K 123, 5 × 1.6 × 6

List of Capacitors used in Circuit 1.4.

Quantity	Components	Ordering codes
4	0.1 µF/63 V ceramic cap.	B 37449–A 6104–S 001
2	22 µF/40 V electrolytic cap.	B 41286–A 7226–T 000
1	4.7 µF/63 V electrolytic cap.	B 41315–A 8475–Z 000
1	10 pF/500 V ceramic cap.	B 38110–J 5100–D 000
3	47 pF/63 V styroflex cap.	B 31310–A 5470–H 000
3	47 nF/63 V ceramic cap.	B 37449–A 6473–S 001
1	1 nF/500 V ceramic cap.	B 37235–J 5102–S 001
1	4.7 nF (5 nF)/500 V ceramic cap.	B 37232–J 5472–S 001

1.5. FM-Tuner with BF 324 and BF 451

The **Fig. 1.5.** shows a circuit of an FM-tuner with self-oscillating mixing stage and capacitance diodes for tuning. It was designed to enable a simple and economic FM-concept.

Both symmetrical and asymmetrical aerials can be connected to the input, since it is a broad-band one. BF 324 serves as a prestage transistor. It operates at an emitter current of 4.5 mA, so that a good compromise between noise figure and large signal stability is achieved. Oscillations in the UHF-range are avoided by a ferrite bead, at the output of the transistor.

The self-oscillating mixing stage operates with a transistor, type BF 451. It is coupled to the RF-circuit by a capacitor with a low capacitance, whereat performances for a power matching are not considered to be a benefit to good sensitivity. The oscillator circuit and the IF-output stage are well decoupled by means of a capacitor with a small capacitance, respectively a choke. This has an advantage in so far as the oscillator coil has not to be tapped.

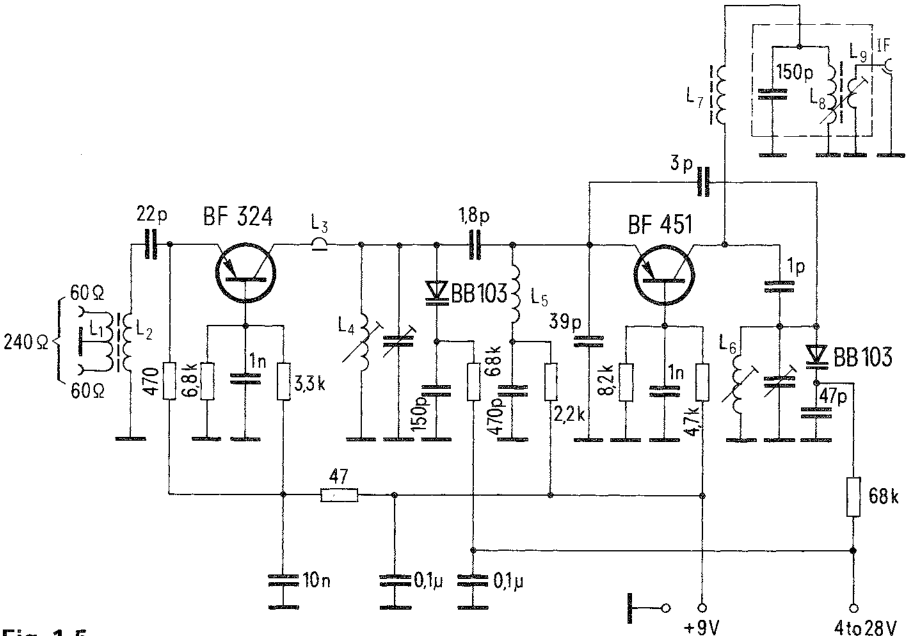


Fig. 1.5.

Characteristics

Supply voltage	9 V
Supply current	7 mA
Tuning voltage	4 to 28 V
Input impedance	60 Ω
Output impedance	60 Ω
Power gain	27 dB
RF-band width	2 MHz
IF-band width	500 kHz
Noise figure	5 dB

Coil data

L_1	6 turns	0.15 mm ∅ enamelled copper wire, centre tap
L_2	5 turns	0.25 mm ∅ enamelled copper wire
L_1, L_2	on cylinder core, B 61110–U 17 2 × 6	
L_3	ferrite bead B 62110–M 11 3.5 × 1.2 × 5.2	
L_4, L_6	6 turns	0.8 mm ∅ enamelled copper wire on core 5 mm ∅
L_5	8 turns	0.5 mm ∅ enamelled copper wire on core 5 mm ∅
L_7		6.8 µH
L_8	12 turns	0.2 mm ∅ enamelled copper wire
L_9	2 turns	0.2 mm ∅ enamelled copper wire
L_8, L_9	with Vogt coil kit D 41–2520	

List of Capacitors used in Circuit 1.5.

Quantity	Components	Ordering codes
1	22 pF/500 V ceramic cap.	B 38116–J 5220–J 000
2	1 nF/500 V ceramic cap.	B 37235–J 5102–S 001
2	0.1 μF/65 V ceramic cap.	B 37449–A 6104–S 001
1	10 nF/250 V ceramic cap.	B 37452–A 2103–S 008
2	150 pF/63 V styroflex cap.	B 31310–A 5151–H 000
1	470 pF/63 V styroflex cap.	B 31310–A 5471–H 000
1	1.8 pF (2 pF)/500 V ceramic cap.	B 38112–J 5020–C 000
1	1 pF/500 V ceramic cap.	B 38110–A 5010–C 000
1	47 pF/63 V styroflex cap.	B 31310–A 5470–H 000
1	3 pF/500 V styroflex cap.	B 38112–J 5030–D 000

1.6. UHF-Tuner with 2 × AF 379 and AF 280 offering an excellent Large Signal Characteristic

Crossmodulation problems of TV-tuners have been solved by the introduction of the well known principle of using a high-current transistor and a PIN-diode control network. However, this does not apply to interfering transmitters, which are in the neighbouring frequency of the tuned station. Their signal, amplified by the prestage and attenuated by the selective band-pass filter, is supplied to the mixing stage, where the crossmodulation products are generated. Since an essential improvement of the selectivity cannot be expected by the use of varicap diodes for tuning, the large-signal features of the mixing stage have to be bettered.
For this application the transistor AF 379 is especially favoured. It offers a very good mixing linearity, if it is operated at collector currents between 4 and 5 mA. The oscillator power required for the maximum drive of the mixing transistor can be realized by an AF 280, whereat a sufficient decoupling between oscillator and mixing stage is guaranteed.
Fig. 1.6. shows the circuit of a designed and tested tuner.
The prestage is coupled to the band-pass filter by a capacitor with a small capacitance. A collector choke improves the amplification at the lower end of the frequency range. The mixing stage is coupled capacitively to the band-pass filter and the oscillator coupling is also a capacitive one. The oscillator circuit has high impedance in order to achieve a good stability against temperature influences. The amplitude of the oscillation keeps its optimum value even at lower frequencies.

Characteristics

Frequency range	470 to 860 MHz
Tuning voltage	1 to 28 V
Supply voltage	12 V
Supply current	20 mA
Power gain	25 to 30 dB
RF band width	12 to 15 MHz
IF band width	7.5 MHz
Noise factor	5 to 6.5 dB
Reflexion coefficient	<0.6
Temperature drift of the oscillator at $\Delta T = 15^{\circ}$	$+100/-300$ kHz
Admissible interfering voltage for 1 % cross-modulation with a distance of 2 channels	60 mV (e.m.f./2 of a TV signal at 60 Ω)

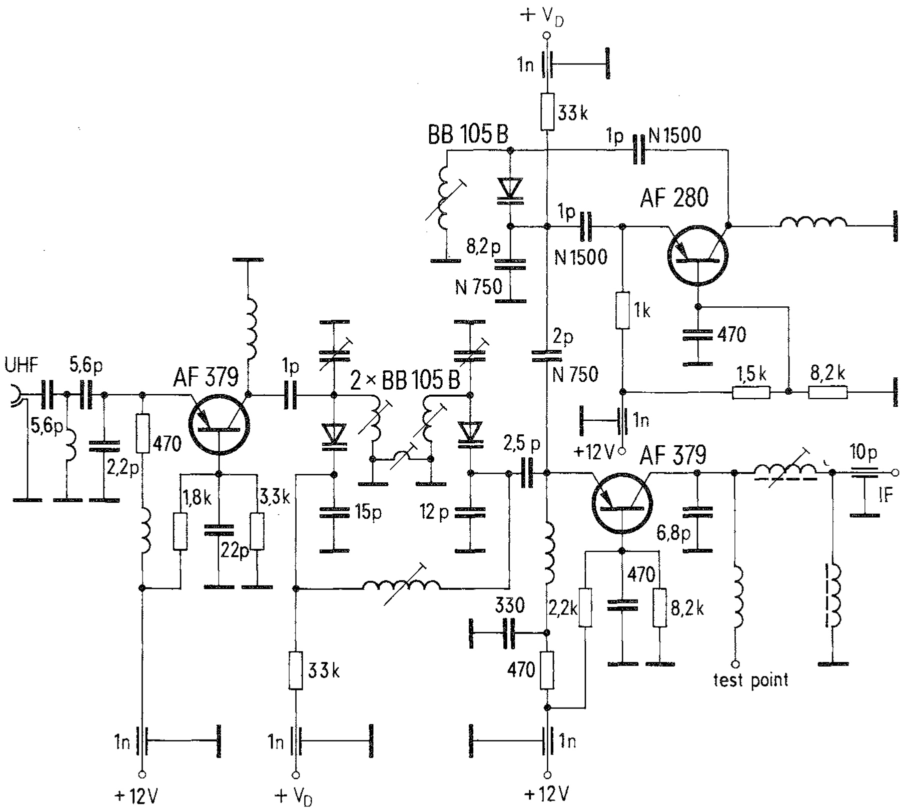


Fig. 1.6.

List of Capacitors used in Circuit 1.6.

Quantity	Components	Ordering codes
2	5.6 pF (V > 30 V)	B 38312–J 4050–C 600
1	2.2 pF (V > 30 V)	B 38265–J 5020–C 200
1	22 pF (V > 30 V)	B 38286–J 5220–J 005
1	1 pF (V > 30 V)	B 38262–J 5010–B 000
1	15 pF (V > 30 V)	B 38285–J 5150–J 005
1	12 pF (V > 30 V)	B 38285–J 5120–J 005
1	2.5 pF (V > 30 V)	B 38265–J 5020–C 500
1	330 pF (V > 30 V)	B 37205–A 5331–M 001
1	8.2 pF N 750 (V > 30 V)	B 38386–J 5080–D 205
1	2 pF N 750 (V > 30 V)	B 38266–J 5020–C 000
2	1 pF N 1500 (V > 30 V)	Stettner
2	470 pF (V > 30 V)	B 37291–B 5471–S 005
1	6.8 pF (V > 30 V)	B 38282–A 5060–F 805
1	10 pF (V > 30 V)	B 37103–A 5100–K 6
5	1 nF (V > 30 V)	B 37109–A 5102–S 6

1.7. AF-Amplifier, Class-A, 3 W at 4 Ω, V_s = 115 V

A class-A-amplifier being suitable for a power supply voltage of 155 V can be designed by using the AF-section of the IC TBA 460 and the transistor BU 111.
The voltage of the input divider consisting of resistors R_1 and R_2 and the voltage across the emitter resistor R_5 are compared at the differential input of the IC (cf. **Fig. 1.7.**) and the latter is adjusted to the value of the divider voltage. Therefore the operation of the output stage is very stable. The required gain is determined by the feedback resistors R_3 and R_4. By an RC-circuit connected in parallel to R_3 the frequency response can be corrected accordingly.
With the described application a supply voltage of 18 V is provided for the IC type TBA 460. A range of 5 to 18 V is also applicable, if the voltage divider consisting of R_1 and R_2 and the resistor R_6 are proportioned accordingly.

Characteristics

Supply voltage for the output transistor V_{S1}	115 V
Supply current I_{S1}	95 mA
Supply voltage for the IC, type TBA 460 V_{S2}	18 V
Supply current	25 mA
Maximum output power (k = 1 %, f = 1 kHz) $P_{nom.}$	3 W
Load R_L	4 Ω
Frequency range (−3 dB)	63 Hz to 15 kHz
Input voltage V_i	20 mV
Input impedance R_i	36 kΩ
Thermal resistance R_{thCASE} of the heatsink for the transistor BU 111	≦4 K/W

Transformer data

Core EI 60, dyn. sheet IV/0.35 mm
Laminated in the same direction, total air gap: 1 mm

Winding data n_1 = 80 turns, 0.5 mm, enamelled copper wire 2 × 0.1 mm paper insulated
n_2 = 2000 turns, 0.18 mm, enamelled copper wire
n_3 = 80 turns, 0.5 mm, enamelled copper wire 2 × 0.1 mm paper insulated

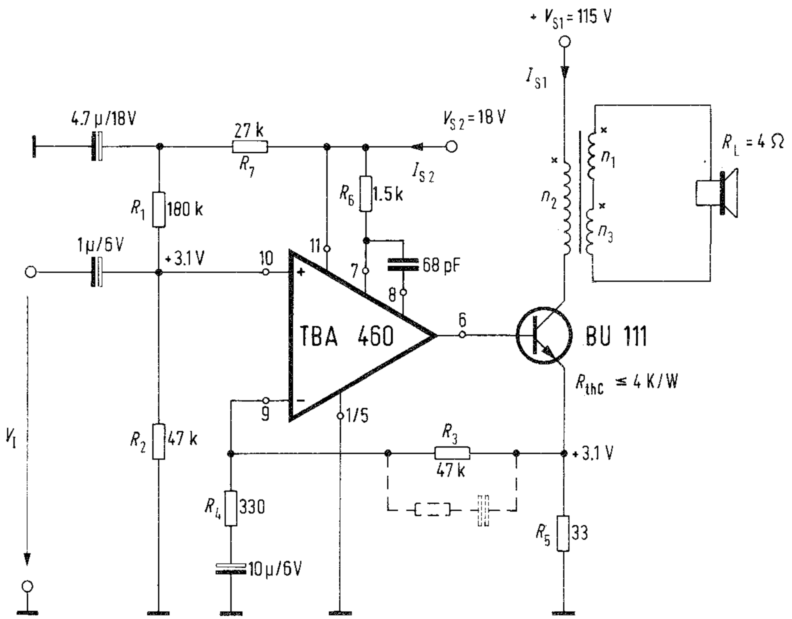


Fig. 1.7.

List of Capacitors used in Circuit 1.7.

Quantity	Components		Ordering codes
1	Electrolytic Capacitor	1 µF/63 V	B 41315–B 9105–Z 000
1	Electrolytic Capacitor	4.7 µF/63 V	B 41315–A 8475–Z 000
1	Electrolytic Capacitor	10 µF/63 V	B 41286–A 8106–T 000
1	Styroflex Capacitor	68 pF/63 V	B 31310–A 5680–H 000

2. Circuits for Monochrome TV-Receivers

2.1. Pulse Separation, Phase Comparison, Horizontal-Deflection-Generator and Driver

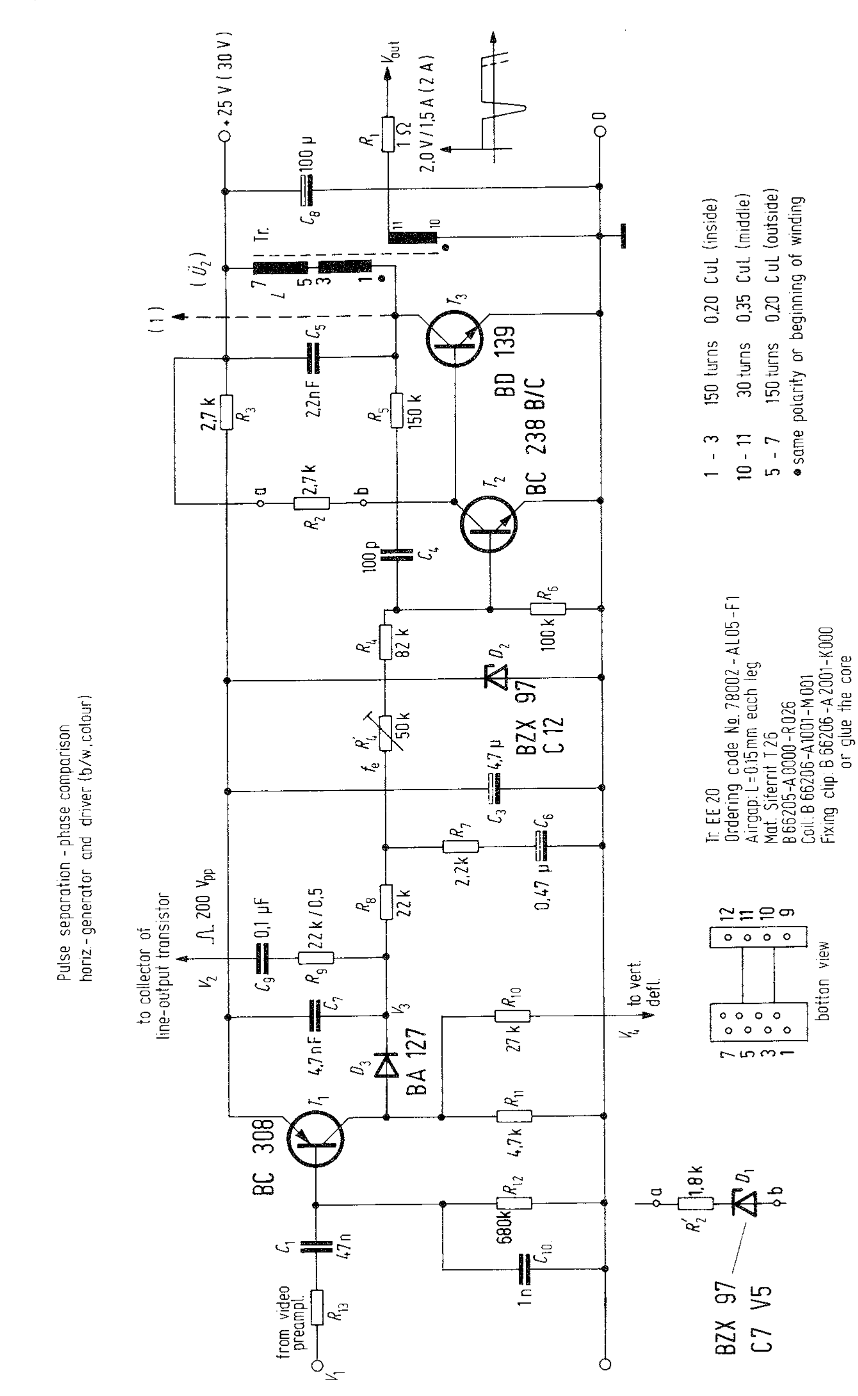
Every engineer developing new circuits endeavours to increase the efficiency as well as to improve the functions. In addition, the amount of components used should be reduced to a minimum. This applies not only to circuits with IC's but also to those using discrete semiconductors. In considering these conditions a simple circuit for processing the horizontal deflection pulses has been developed. It offers a variety of functions (cf. **Fig. 2.1.1.**).

The transistor T_1 effects the pulse separation of negatively directed line pulses of the video signal coming from the video preamplifier. The amplitude of this signal can range between 1 and 5 V_{pp}. It is supplied via the capacitor C_1 and the resistor R_{13} to the base of the transistor T_1 negatively biased. When the signal is missing practically the total positive supply voltage is applied to the capacitor C_7 via the switch-through transistor T_1, and the conductive diode D_3. Thus the horizontal oscillator operates at nominal frequency f_0.

If there is a video signal, a positive bias voltage arises at C_1 as a result of the greater negative synchronizing pulses. This voltage shifts the video signal into the cut-off region, so that only the negative synchronizing pulses drive the transistor, (but in this case intensively). Negative pulses with an amplitude of 12 V and a duration of about 4 µs are available at the collector.

A reference pulse coming from the horizontal line output-transistor via R_9 is integrated by the capacitor C_7. The result is a saw-tooth voltage, which is superimposed upon the supply voltage of 12 V. The synchronizing pulse passes the transistor T_1 and at a momentary stance (phase comparison) the delta voltage is added to the supply voltage. Thus it is possible to shift the delta voltage up and down, whereby its average value, both as a negative and positive voltage, is added to the 12 V-level. The resistors R_8 and R_7 in conjunction with the capacitor C_6 realize the filtering for pulses going direction h-generator. The capacitor C_6 integrates rapid voltage changes which are caused by interference pulses, so that the horizontal generator can constantly maintain its frequency. In any case a stabilization of the 12 V power supply voltage is necessary. In the described application the diode D_2 is responsible for the operation. The achievement is such that the picture phase position is only slightly affected by the reactions of the sound circuit and power supply, caused by load changes, especially those of the beam current.

Fig. 2.1.1. shows the h-generator circuit consisting of T_2 and T_3. Principally it is applicable for all kinds of TV-receivers. Its oscillating frequency can be retuned by a dc voltage, whereby a low control current is required. In this application a z-diode type BZX 97 C 12 stabilizes the voltage to a level of 12 V and thus also the frequency of the oscillator. The oscillation starts with a positive sine half-wave when the transistor BD 139 is not conductive. The transistor BC 238 is switched on as long as the voltage at capacitor C_4 turns it off again. This happens when the sine voltage at the transformer decreases. Thereafter the transistor T_3 is turned on and energy is stored by the inductor L. When the transistor T_3 is non-conductive the sine-oscillation begins again with an increasing amplitude. The following is achieved by capacitor C_5. The flyback voltage increases to a value between 50 and 120 V only



Pulse separation - phase comparison
horiz.-generator and driver (b/w, colour)
to collector of
line-output transistor
200 Vpp
from video
preampl.
BC 308
BA 127
BZX 97
C 12
BC 238 B/C
BD 139
+25 V (30 V)
Vout
1 Ω
2,0 V / 1,5 A (2 A)
to vert.
defl.
BZX 97
C7 V5
1 - 3 150 turns 0,20 CuL (inside)
10 - 11 30 turns 0,35 CuL (middle)
5 - 7 150 turns 0,20 CuL (outside)
• same polarity or beginning of winding
Tr. EE 20
Ordering code No. 78002-AL05-F1
Airgap: L = 0.15 mm each leg
Mat. Siferrit T 26
B 66205-A0000-R026
Coil: B 66206-A1001-M001
Fixing clip: B 66206-A2001-K000
or glue the core
bottom view

Fig. 2.1.

at the rated frequency. If supply voltages other than indicated are chosen in this application sample, a suitable capacitor with an accordingly higher dc-voltage characteristic has to be used.
This h-generator can be named as a combined sine-blocking-oscillator. It can be synchronised directly by the described phase comparison circuit. By the components R_2 and D_1 a delayed, but then an intensive start of the oscillation is attained. Generally a value of 2.7 kΩ is sufficient for the resistor R_2.

Characteristics

Supply voltage	25 V (12 to 30 V)
Supply current	80 mA
Control signal PBS (T_1)	1 to 5 V
Control current (T_2)	80 µA
Output voltage	2 V
Output current max.	1.5 A
Admissible load	1.3 Ω
Frequency	15.625 kHz ± 1 kHz (from $V_s \geqq 10$ V)

Table

V_S V	I_B at $R_B = 0.67\,\Omega$ A	I_B at $R_B = 1\,\Omega$ A	$1/f$ µsec	V_{CE} (T_3) V
5	0	0	78	30
10	0.3	0.2	67	45
15	0.7	0.5	62	55
20	1.2	1	63	75
25	1.7	1.3	65	95
30	2	1.7	67	120

The above table indicates the essential behaviour of the described h-generator (V_S, f, V_{CE}, I_B for the output stage). This h-driver is only loaded during the conductive period. During the blocking interval a half-sine voltage V_{CE} is generated at the collector capacitor of the driver transistor. By using an additional diode BA 104 (1) the flyback voltage can be clamped to a value of e.g. 100 V, if driver transistors with a sufficiently high reverse voltage are not available.

List of Capacitors used in Circuit 2.1.

Items		Components	Ordering codes
C_1	1	MKH 0.047 μF/250 V	B 32234–B 3473–M 000
C_2			
C_3	1	4.7 μF/100 V electrolyt.	B 41283–A 9475–T 000
C_4	1	100 pF/160 V styrofl. 2.5%	B 31110–A 1101–H 000
C_5	1	2.2 nF/160 V styrofl. 5%	B 31310–A 1222–H 000
C_6	1	0.47 μF/100 V MKM	B 32540–A 1474–S 000
C_7	1	4.7 nF/250 V MKM	B 32540–A 3472–K 000
C_8	1	100 μF/40 V electrolyt.	B 41316–B 7107–Z 000
C_9	1	MKH 0.1 μF/400 V	B 32234–B 6104–M 000
Others:			
Tr		Driver transformer (see also fig. 2.1.)	B 78002–AL 05–F 1

2.2. A Simple, Regulated Pulse Mode Power Supply for Portable TV-Sets without Mains Isolation (for Picture Tubes 110°/30 cm)

Fig. 2.2.1. shows a circuit for a horizontal deflection on the principle of pulse control. This circuit has no mains isolation.
This part of the deflection circuit corresponds to the already known low-voltage circuit. It consists of the line switching transistor T_2, the parallel diode D_8, the horizontal deflection unit L_j, the charging electrolytic capacitor C_{11}, the flyback capacitor C_{10}, the linearity coil Lin and the S-capacitor C_9. The primary winding 5–7′ of the line transformer is part of an autotransformer. By means of the tap (No. 7) flyback voltage across the capacitor C_{10} can be matched to the operating voltage $V_1 \approx 300$ V. The transistor T_1 operates as a current switch line-frequency synchronised. It is driven by the transformed flyback voltage via the windings 3 and 4 of the line transformer. By means of the z-diode D_3 the base voltage of the transistor T_1 is controlled, and thus its collector current. The circuit can be applied to 220 V-mains or battery operation. The filtering after the rectifier D_1 is achieved by a circuit consisting of C_2, R_3 and C_3. The collector of the pulse mode switching transistor is connected to the capacitor C_3 via the diode D_2 and its emitter is joined to the primary winding 5–7′. The diode D_6 enables the boosting in conjunction with the horizontal deflection circuit and converts the battery voltage from 12 V to 17 V. The circuit itself operates with the boosted voltage of 12 V and supplies the current for the TV-set. The beginning of the oscillation is achieved by the surge voltage at the capacitor C_4 when the set is switched on. The horizontal deflection generator (not shown in the schematics) immediately starts to oscillate and drives via the transformer Tr 2, the deflection transistor T_2 intensively. The latter controls via the h-transformer Tr 1, the pulse mode switching transistor T_1 and accordingly the increasing supply voltage. Thus the beginning of the oscillation is safely assured for the transistors T_1 and T_2.

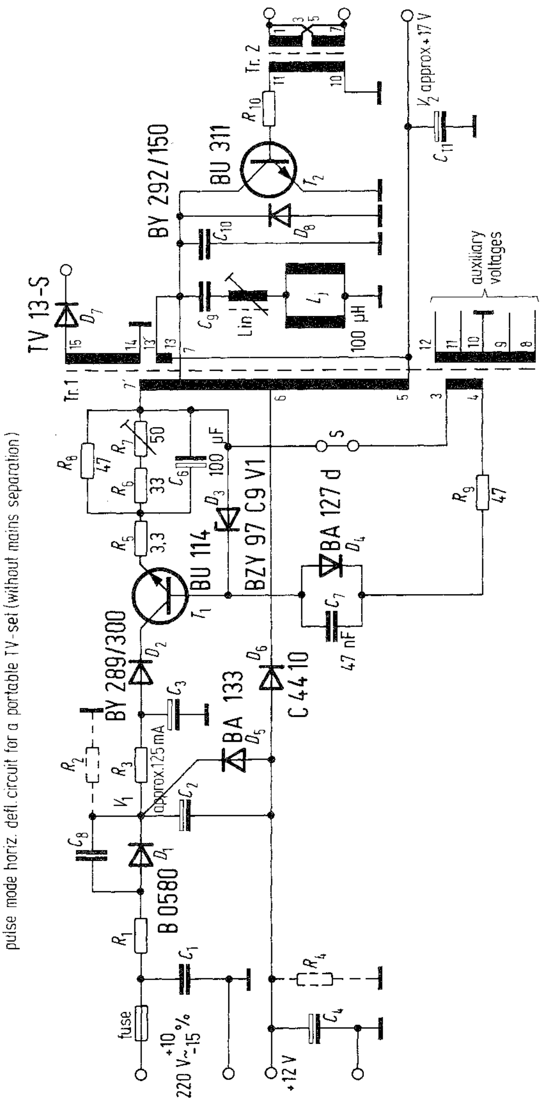

pulse mode horiz. defl. circuit for a portable TV-set (without mains separation)
TV 13-S
BY 292/150
BU 311
Tr. 2
Tr. 1
V_2 approx. +17 V
auxiliary voltages
100 µH
Lin
BU 114
BZY 97 C9 V1
BA 127 d
100 µF
47 nF
C 4410
BA 133
BY 289/300
B 0580
approx. 125 mA
fuse
220 V~ +10 −15 %
+12 V

Fig. 2.2.1.

The hum voltage across the electrolytic capacitor C_4 is isolated from the secondary supply voltage by the diode D_6. After the circuit is turned off, switch on can be repeated instantly. The picture width change affected by beam current variations is about 3mm. A picture width change depending on supply voltage fluctuations is not existent within the control range of 190 to 250 V ac.
For the deflection transistor T_2 the type BU 311 is used. It offers a V_{CBO}-voltage of 200 V. The type BU 114 having a V_{CBO} of 250 V serves as the pulse mode switching transistor.
The pulse generated in the winding 3–4 and controlling the BU 114 is differentiated by an RC-circuit. By that it becomes narrower and thus the power dissipation of the transistor is reduced. The gain tolerances of the transistor as well as the voltage tolerances of the z-diode D_3 are compensated by the potentiometer R_7, which also adjusts the picture size. It is very important that the described circuit operates extremely well, not only under normal conditions but also during the switching on and off periods. The circuit consisting of R_1 R_3, C_1, C_2, C_3 and C_8 and being connected in front of the switching transistor BU 114, is sufficient to achieve only a small degree of interference to the mains. Besides that an essential interference elimination is obtained by directly connecting primary and secondary windings. This measure shows the effect of a short circuit for interferences.
Since the oscillation of the h-generator and the control of the transistor T_1 are immediately interrupted when a short circuit of V_2 occurs, the transistors T_1 and T_2 are extremely well protected.
When the circuit is battery operated a current flows via the booster diode D_6 into the winding 6–5 after switch on. Therefore the capacitor C_{11} is charged and a secondary supply voltage of 17 V is generated. The diode D_2 is used for isolation purposes at booster-operation. It prevents any voltage influences from the 220 V mains-part, which will affect the flyback pulse at T_2. It is recommended to open the switch S, so that the control circuit for T_1, which is no longer required, is not consuming energy (1 to 2 W). This switch should operate directly in conjunction with the one responsible for selecting mains or battery operation. The diode D_5 prevents the capacitor C_2 from being connected to the wrong polarity. The data of the h-transformer is in principle the same as that used in the circuit 2.3.1. with the following exception. The winding 1–2 of circuit 2.3.1. is not separated in the circuit 2.2.1. but a part of the autotransformer winding 5–6–7–7'. The described circuit has been tested in conjunction with a picture tube type A 31–120 W.

List of Components used in the Circuit 2.2.1.

Semiconductors	Ordering codes
T_1 = NPN-power-switching transistor BU 114	• Q 62702–U 118
T_2 = NPN-power-switching transistor BU 311	• Q 62702–U 210
D_1 = Silicon-rectifier BO 580	• C 66047–A 1005–A 5
D_2 = Silicon-rectifier BY 289/300	• C 66047–A 1028–A 9
D_3 = Silicon z-diode BZY 97 C 9 V 1	• Q 68000–A 951–F 82
D_4 = Silicon diode BA 127 d	• Q 60201–X 127
D_5 = Silicon diode BA 133	• Q 60201–X 133
D_6 = Booster diode C 4410	• BY 293/150
D_7 = Selenium HV-rectifier stick TV 13 S	• Q 70–H 0922
D_8 = Silicon rectifier BY 292/150	• C 66047–A 1045–A 3

Capacitors

C_1	= RFI suppression cap. 0.1 µF ± 20%, 630 V–	● B 81121–C–F 12
C_2	= Electrolytic cap. 100 µF + 50 – 10%, 350 V–	● B 43050–B 4107–T
C_3	= Electrolytic cap. 100 µF + 50 – 10%, 350 V–	● B 43050–B 4107–T
C_4	= Electrolytic cap. 1000 µF + 50 – 10%, 16 V–	● B 41010–A 4108–T
C_6	= Electrolytic cap. 100 µF + 50 – 10%, 10 V–	● B 41283–B 3107–T
C_7	= MKH-cap. 0.047 µF ± 20%, 250 V–	● B 32234–B 3473–M
C_8	= Polypropylen cap. 2200 pF ± 10%, 630 V–	● B 33061–A 6222–K
C_9	= MKH-cap. 4.7 µF ± 20%, 100 V–	● B 32234–A 1475–M
C_{10}	= Polypropylen cap. 82 nF ± 5%, 630 V–	● B 33065–A 6823–J
C_{11}	= Electrolytic cap. 470 µF + 50 – 10%, 250 V–	B 43306–A 2477–T

Resistors

R_1	= Wire-wound resistor 4.7 Ω, 7 W	B 52136–A 040–J 7
R_2	= Film resistor 150 Ω ± 10%, 0.5 W	B 51263–A 4154–K
R_3	= Wire-wound resistor 150 Ω, 4 W	B 52135–A 151–J
R_4	= Film resistor 2,2 kΩ ± 10%, 0,25 W	B 51262–A 4222–K
R_5	= Wire-wound resistor 3.3 Ω, 1 W	
R_6	= Film resistor 33 Ω ± 10%, 0.5 W	B 51263–A 4330–K
R_7	= Potentiometer 50 Ω, 0,5 W	
R_8	= Film resistor 47 Ω ± 10%, 0.5 W	B 51263–A 4470–K
R_9	= Film resistor 47 Ω ± 10%, 1 W	B 51266–A 4470–K
R_{10}	= Wire-wound resistor 1 Ω, 0.5 W	

Others

Tr_1 = Horizontal-deflection transformer, type 241–158079, manufactured by Plessey-Arco, Italy, Firenze, Via San Pietro-A Quarachi 250
Tr_2 = Horizontal-deflection driver transformer, technical data see Fig. 2.1.1.
L_j = Deflection unit 100 µH for picture tube A 31–120 W
Lin = Linearity control, suitable with the deflection unit L_j
Fu = Medium slow-acting fuse 0.8 A

The black dots indicate those components, that are available ex stock from our Central Components Depot in: 8510 Fürth-Bislohe, Postbox 146.

2.3. Regulated Pulse Mode Power Supply with Mains Isolation for 30 cm-Picture Tube

The 220 V-power-supply shown in the following circuit **(fig. 2.3.1.)** uses a one-way rectifier B 0580. The pulse mode part, being able to operate with mains isolation has already been described in the previous chapter.
By the switching transistor T_1 the rectified mains voltage ranging between 230 and 300 V is connected at line frequency to the primary winding of the h-transformer, whereby the transistor T_1 operates in the active, highly resistive area of the family of characteristics, i.e. not in the area of the collector-emitter-saturation voltage.

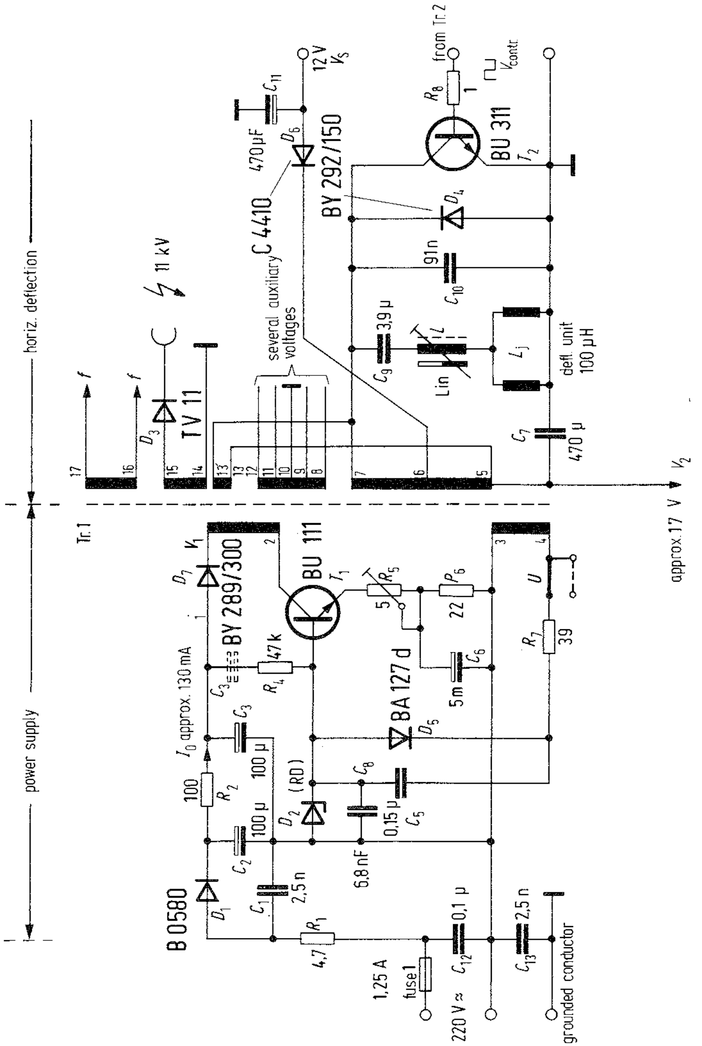

power supply
horiz. deflection
Tr.1
B 0580
BY 289/300
BU 111
BA 127 d
TV 11
11 kV
several auxiliary voltages
C 4410
BY 292/150
BU 311
470µF
12 V
V_S
from Tr.2
$V_{contr.}$
91n
3,9 µ
Lin
defl. unit
100 µH
470 µ
approx.17 V
V_2
I_0 approx. 130 mA
100
100 µ
2,5 n
4,7
6,8 nF
0,15 µ
(RD)
47 k
5m
22
39
1,25 A
fuse1
220 V ≈
0,1 µ
grounded conductor

Fig. 2.3.1.

The transistor is controlled via the resistor R_7 and a differentiation capacitor C_5 by the negative feedback voltage. The z-diode D_2 clips the control voltage and thus in conjunction with the emitter resistors an excellent current stabilization is achieved. The pulses at the collector have a duration of about 9 µs. Therefore the power dissipation of the transistor T_1 is reduced (see also **Fig. 2.4.2.**). The current control occurs at relatively low collector voltages.

This circuit offers the advantage in that the picture size will be nearly independent if the mains voltage varies in a range of $\pm 10\%$ to $\pm 15\%$. Between 200 V ac and 250 V ac the picture size stability is better than 1 %. The voltage V_s is adjusted by the potentiometer R_5 at rated mains voltage and dark picture. To receive a good power dissipation feature heat sinks have to be used with the following thermal resistance: for T_1 approx. 3.5 K/W and for T_2 approx. 5 K/W.

By using resistor R_4 a reliable oscillator start is assured. If a 2.2 µF-capacitor is added an electronic switch-off is obtained when secondary short-circuits occur. Unfortunately the feature of a continuous oscillation start is given up by this measure. It is important that the capacitance of the secondary capacitors is rated as low as possible in order to achieve a low-loss beginning of the oscillation, i.e. a safe one.

Battery operation is possible by using the booster diode C 4410. In this case the diode D_7 interrupts the current flowing via the primary winding between the transistor T_1 and the capacitors C_2 and C_3. By the switch U the base-circuit of the transistor can be turned off and thus a low-loss booster operation is possible (about 2 to 3 W). Besides that a wrong polarity being applied to capacitors C_2 and C_3 (-20 V) is avoided. If picture size variations, depending on change of battery voltage, are allowed but only by an extremely small amount, then a simple series control circuit can be added. It consists of the transistor T_4, type BD 433, which is loaded with only 2.5 W/2 A. The resistor connected in parallel to the transistor T_4 dissipates about 10 W at a high battery voltage of 16 V(see **Fig. 2.3.2.**). At a low battery voltage of 11 V the transistor current is 2 A and the transistor is practically conductive. The parallel transistor T_3 which is always used because of the sound and the prestage control in this case operates as a prestage for T_4. The bias for the transistor T_4 has to be supplied from a higher voltage (e.g. 17 V), otherwise it cannot be controlled sufficiently at an 11 V-operation, i.e. the voltage drop across this transistor is too high. The described circuit has been tested in conjunction with a picture tube A 31–120 W.

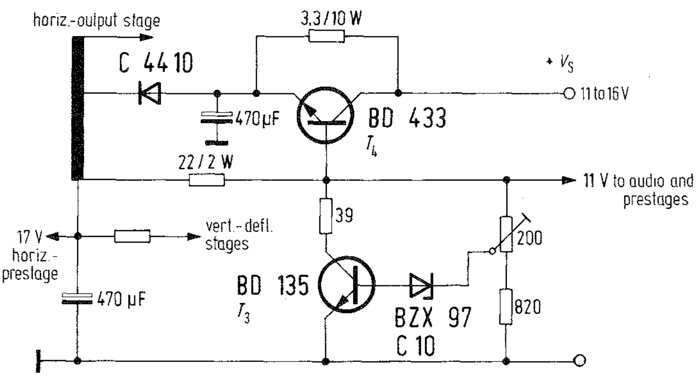


Fig. 2.3.2.

List of Components used in the Circuit 2.3.1.

Semiconductors	Ordering codes
T_1 = NPN-power switching transistor BU 111	● Q 62702–U 84
T_2 = NPN-power switching transistor BU 311	● Q 62702–U 210
D_1 = Silicon rectifier B 0580	● C 66047–A 1005–A 5
D_2 = Silicon z-diode BZY 97 C 9 V 1	● Q 68000–A 952–F 82
D_3 = Selenium HV-rectifier stick TV 11 S	● Q 70–H 0920
D_4 = Silicon rectifier BY 292/150	● C 66047–A 1045–A 3
D_5 = Silicon diode BA 127 d	● Q 60201–X 127
D_6 = Booster diode C 4410 d	● BY 293/150
D_7 = Silicon rectifier BY 289/300	● C 66047–A 1028–A 9

Capacitors	
C_1 = Polypropylen cap. 2200 pF ± 10%, 630 V	● B 33061–A 6222–K
C_2 = Electrolytic cap. 100 µF + 50 – 10%, 350 V–	● B 43050–B 4107–T
C_3 = Electrolytic cap. 100 µF + 50 – 10%, 350 V–	● B 43050–B 4107–T
C_5 = MKH-cap. 0.15 µF ± 20%, 100 V–	● B 32234–A 1154–M
C_6 = Electrolytic cap. 4700 µF + 50 – 10%, 10 V–	● B 41010–B 3478–T
C_7 = Electrolytic cap. 470 µF + 50 – 10%, 25 V–	● B 41012–A 5477–T
C_8 = MKM-cap. 6800 pF ± 10%, 250 V–	● B 32540–A 3682–K
C_9 = MKH-cap. 3.3 µF ± 10%, 100 V–	● B 32234–S 1335–K
C_{10} = Polypropylen cap. 82 nF ± 5%, 630 V–	● B 33065–A 6823–J
C_{11} = Electrolytic cap. 470 µF + 50 – 10%, 16 V–	● B 41283–A 4477–T
$C_{12} + C_{13}$ = RFI suppr. cap. 0.1 µF ± 20%, 250 Vdc/ac + 2 × 2500 pF ± 10%	● B 81321–A–E 14

Resistors	
R_1 = Wire-wound resistor 4.7 Ω, 7 W	B 52136–A 040–J 7
R_2 = Wire-wound resistor 100 Ω, 3 W	B 52135–A 101–J
R_4 = Film resistor 47 kΩ ± 10%, 1 W	B 51266–A 4473–K
R_5 = Potentiometer 5 Ω, 3 W	
R_6 = Film resistor 22 Ω ± 10%, 1 W	B 51266–A 4220–K
R_7 = Film resistor 33 Ω ± 10%, 1 W	B 51266–A 4330–K
R_8 = Wire-wound resistor 1 Ω, 0.5 W	B 52190–A 9010–K

Others

Tr_1 = Horiz.-deflection transformer, see list of components for circuit 2.2.1 (similar)
Tr_2 = Horiz.-deflection driver transformer, see list of components for circuit 2.2.1.
L_j = Deflection unit 100 µH, applicable for picture tube A 31–120 W
Lin = Linearity control, suitable for the deflection unit L_j
KT_1 = Heat sink ≧3.5 K/W
KT_2 = Heat sink ≧5 K/W
Fu = Medium slow-acting fuse 1 A

The black dots indicate those components that are available ex stock from our Central Components Depot in: 8511 Fürth-Bislohe, Postbox 146.

2.4. Horizontal Deflection Circuit using a 185 V-Power Supply with Mains Isolation (Suitable for Picture Tubes 110°/30 to 65 cm)

The circuit of the power supply shown in **Fig. 2.4.1.** uses a controlled thyristor-rectifier. The rectified 50 Hz-sine half-wave voltage is smoothed by two electrolytic capacitors (in this application 2 × 220 µF/B 43306–B 4227–T 000). The regulation and rectification are achieved by the thyristor BSt B 0246 which is periodically triggered by a phase-shifting control circuit consisting of a voltage divider which compares a part of the output voltage with the constant one of the z-diode BZX 97 C 24. In this case only a feedback control is applied. The difference of both voltages is supplied to the base of the control transistor BC 237, operating as an amplifier. It reacts as a variable impedance consisting of a collector resistance (47 kΩ), the internal impedance of the transistor, the capacitance of 0.1 µF at the collector and an additionally phase shifting RC-circuit. The output voltage of the phase shifter is supplied to the trigger-gate of the thyristor via the diac type A 9903. The triggering occurs after the sine half-waves have reached their maximum and it is shifted on their trailing edges, according to the amount of the output voltage. A standard filtering circuit reduces the 50 Hz-hum to a value of about 2.5 V_{pp}. The output voltage of 185 V changes within a range of ±1 %, if the mains voltage varies ±10%.

The principle operation of this pulse mode power supply has already been described in chapters 2.2. and 2.3.

The mains isolation is obtained by the h-transformer. The switching transistor T_1 applies at line frequency the rectified supply voltage of 185 V to the primary winding 1–2 of the h-transformer, whereby the transistor T_1 is driven into the area of collector voltage saturation. Thus a high voltage with a low source impedance is generated. The transistor T_1 is controlled by the negative feedback voltage across the winding 3–4 via the resistors R_{15}, R_{14}, and the capacitor C_{10}. The diode D_3 supplies the starting voltage. The pulses of the collector current should have a duration of 9 µs so that the power dissipation can be extremely reduced when the high voltage coil is favourably tuned to the 3rd or the 5th harmonic.

The current is controlled at very low collector emitter voltages. Since the supply voltage is constant (185 V), the BU 114 can be used as a switching transistor and the two 22 µF-capacitors C_7 and C_8 have only to be rated to 250 V max.

The described circuit offers the advantage in that the picture size is independent of mains voltage variations. If this voltage changes between 200 V ac and 250 V ac the picture size stability will be about 1 %. The voltage V_2 is adjusted by the potentiometer R_{14} at a rated mains voltage. To achieve a power dissipation of about 5 W for T_1 a heat sink with a thermal resistance of approx. 5 K/W has to be used. A cooling block with the same thermal resistance is also recommended for T_2. The beginning of oscillations is attained by the resistor R_{13} and the capacitor C_9. The latter assures that the circuit is immediately turned off if a secondary short circuit occurs. The clamping diode D_4 protects the transistor T_1 against high-tension arc-overs of the picture tube. It is important that the capacitance of the secondary capacitors is rated as low as possible, in order to achieve a low-loss and a fast beginning of the oscillation.

It has been experienced that the h-transformer essentially affects a troublefree oscillation start. Therefore the h-transformer should oscillate with no input signal at a

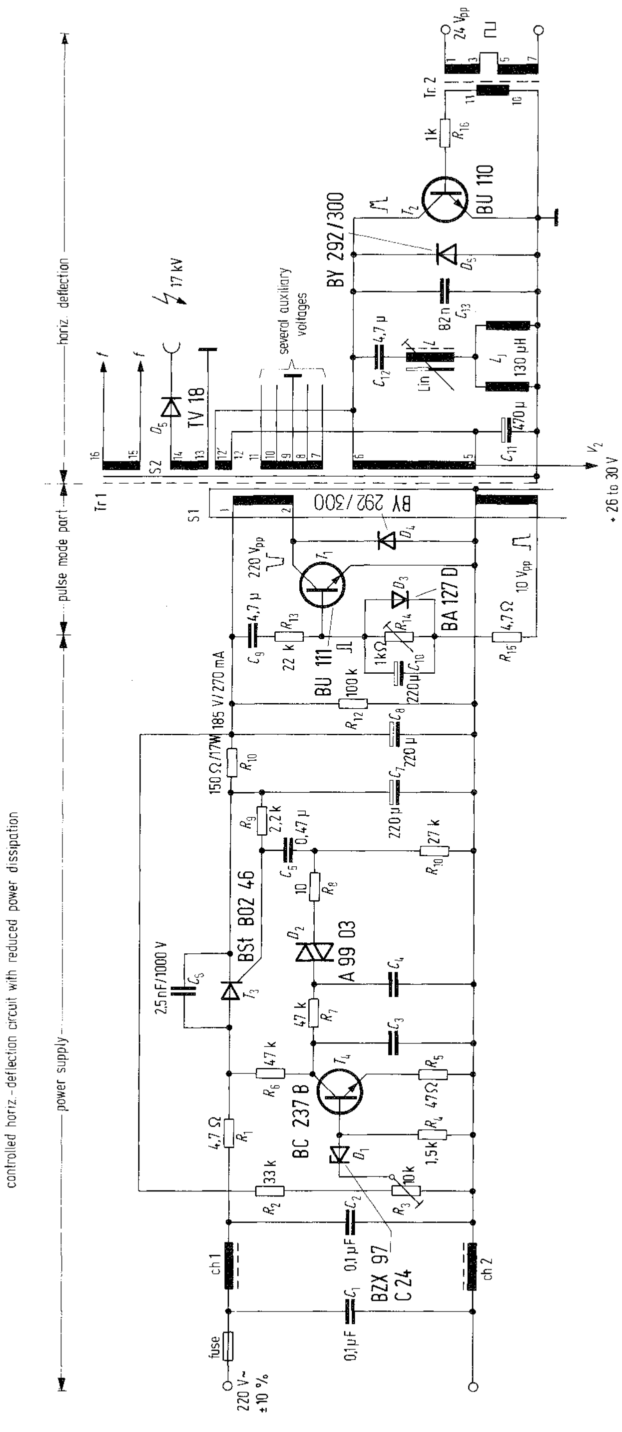

controlled horiz.-deflection circuit with reduced power dissipation
power supply
pulse mode part
horiz. deflection
220 V~ ±10 %
fuse
ch 1
ch 2
BC 237 B
BZX 97 C 24
BSt B02 46
A 99 03
BU 111
BA 127 D
BY 292/300
BU 110
TV 18
17 kV
several auxiliary voltages
Tr 1
Tr. 2
+ 26 to 30 V
24 Vpp
185 V/270 mA
150 Ω/17W
2,5 nF/1000 V
220 Vpp
10 Vpp
130 µH

Fig. 2.4.1.

slightly higher frequency than in the synchronized state. Even at low voltage (e.g. $< {}^{1}/_{3} \times V_2$) the natural frequency should not differ very much from the desired one. To avoid parasitic oscillations in the MHz-range effected by circuit capacitances, a protection capacitor of 6.8 to 10 nF has to be connected from the base of T_1 to the reference voltage terminal. The circuit described above has been tested in conjunction with a picture tube A 61–120 W.

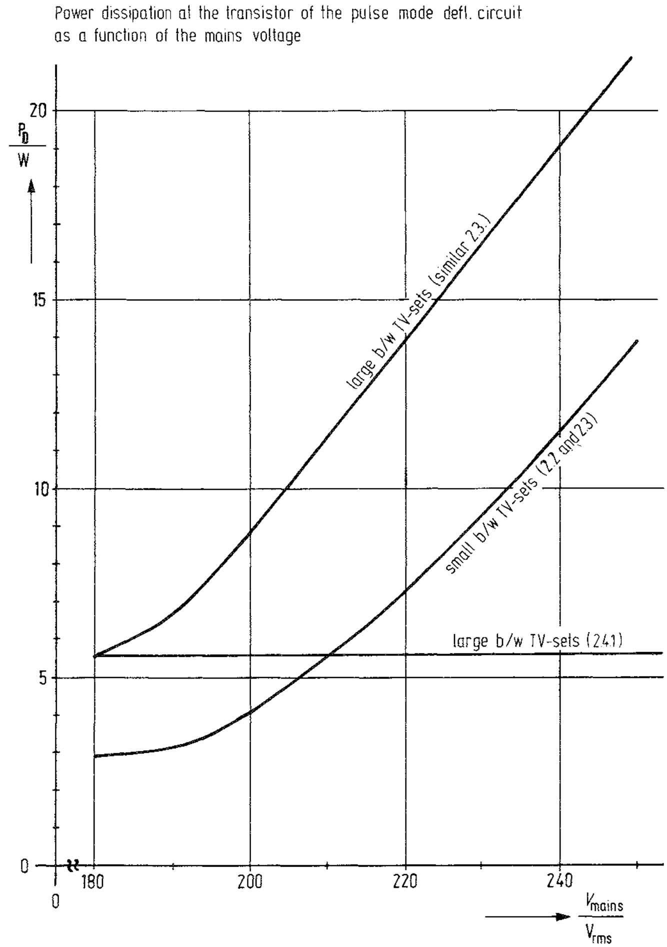


Fig. 2.4.2.

Interference Elimination Measures

Very often pulse mode power supplies create interferences which are transmitted via the lines of a TV-set to its aerial (open dipole). To eliminate these disadvantages the following measures are recommended.

1. The set is operated without mains isolation. Thus the source of interference is nearly short-circuited. A probable remainder can be suppressed by the chokes Ch_1, Ch_2, the capacitors C_1 and C_2 (see Fig. 2.4.1.).
2. The primary side of the transformer is separated from the secondary by means of two screenings. Thus a good interference suppression is achieved, even if the secondary side of the transformer is not connected to ground. Both primary and secondary interference generators are short-circuited separately and there is no potential difference between the two screenings. An additional suppression is obtained by connecting a capacitor with a capacitance of about 2.5 nF from one of the supply lines to the secondary chassis. It has to be mentioned that the collector of T_1 is not to be shunted by the existing capacitance between the transistor system and the heat sink.

Safety and Economy

The described circuit has been designed with the aim that the switching transistors do not run the risk of being overloaded or driven into the range of a second breakdown each time the set is switched on. The oscillation is started and stopped in such a way that the transistor does not operate in the prohibited area of the family of characteristics.

Short circuits, e.g. those applied to the capacitors C_{11} and C_{13}, do not affect transistors with good characteristics. However, the influence is stronger than under normal conditions. Besides that short-circuits on the secondary side (overload) have a stronger effect on the transistors, if no intermittent oscillation arises. Arc-overs of the picture tube are without any influence. But "scratching" short-circuits existing during several periods at the capacitor C_{13} are dangerous for T_1 and T_2, since high currents are generated in the winding 5–6 (>100 A). If the short-circuit is eliminated these currents will create high tensions and thus highly inverse currents as well as high voltage peaks are applied to the transistor T_1. Protection of T_1 and T_2 is achieved by a RCD-circuit or by a diode D_4 connected to T_1. But why should one use an expensive protection circuit, when the mentioned conditions never occur in practice. To receive a sufficient safety at tests or other unfavourable conditions additional resistors can be connected to base and collector of T_1. However this measure often affects the function of the circuit.

A safe operation is a matter of course, however, more important is the requirement that the transistors are safely protected against overloads when the set is switched on or off. This is obtained by the described circuit. If one should require more safety measures it can of course be achieved, but these might be considered as exaggerations in accordance with economy, function and simplicity of the circuit.

List of Components used in the Circuit 2.4.1.

Semiconductors	Ordering codes
T_1 = 1 PNP-power-transistor BU 111 (BU 114)	• Q 62702–U 84
T_2 = 1 PNP-power-transistor BU 110	• Q 62702–U 83
T_3 = 1 Thyristor BSt B 0246	• Q 66048–A 1402–A 6

T_4	= 1 PNP-transistor BC 237 B	● Q 62702–C 277
D_1	= 1 Silicon z-diode BZX 97 C 24	● Q 62702–Z 1244
D_2	= 1 Diac A 9903	● C 66047–Z 1304–A 1
D_3	= 1 Silicon diode BA 127 d	● Q 60201–X 127–D 9
$D_4 + D_6$	= 1 Silicon rectifier BY 292/300	● C 66047–A 1045–A 5
D_5	= 1 Selenium HV-rectifier stick TV 18 S	● Q 70–H 0892

Capacitors

$C_1 + C_2$	= 2 RFI suppression cap. 0.1 μF ± 20%, 630 V=	● B 81121–C–F 12
$C_3 + C_4$	= 2 MKM-stacked film cap. 0.1 μF	● B 32540–A 1104–J
C_5	= 1 MKH-cap. 0.47 μF/400 V	● B 32231–C 6474–M
C_6	= 1 Polypropylen cap. 2200 pF ± 10%, 630 V	● B 33061–A 6222–K
$C_7 + C_8$	= 2 Electrolytic cap. 220 μF + 50 – 10%, 350 V=	● B 43306–B 4227–T
C_9	= 1 Electrolytic cap. 4.7 μF, 350 V	● B 43283–B 4475–T
C_{10}	= 1 Electrolytic cap. 220 μF, 63 V	● B 41010–A 8227–T
C_{11}	= 1 Electrolytic cap. 470 μF + 50 – 10%, 40 V	● B 41010–A 7477–T
C_{12}	= 1 MKH-cap. 4.7 μF ± 20%, 100 V=	● B 32234–A 1475–M
C_{13}	= 1 Polypropylen cap. 82 nF ± 5%, 630 V=	● B 33065–A 6823–J

Resistors

R_1 = 1 Wire-wound resistor 4.7 Ω, 11 W
R_2 = 1 Resistor 33 K/1 W
R_3 = 1 Variable resistor 10 kΩ/0,5 W
R_4 = 1 Resistor 1.5 K, 0.1 W
R_5 = 1 Resistor 47 K, 0.1 W
R_6 = 1 Resistor 47 K, 1 W
R_7 = 1 Resistor 47 K, 0.1 W
R_8 = 1 Resistor 10 Ω, 0.3 W
R_9 = 1 Resistor 2.2 K, 0.3 W
R_{10} = 1 Resistor 22 K, 0.3 W
R_{11} = 1 Wire-wound resistor 150 Ω ± 5%, 17 W
R_{12} = 1 Resistor 100 kΩ, 0.5 W
R_{13} = 1 Filmresistor 47 kΩ ± 10%, 2 W
R_{14} = 1 Potentiometer 1 k, 2 W
R_{15} = 1 Filmresistor 4.7 Ω ± 10%, 2 W

Others

Tr_1 = Horizontal-deflection transformer
Tr_2 = Driver transformer for horiz.-deflection circuit (see chapter 2.1.)
L_j = Deflection unit 105 to 130 μH, suitable for the picture tube A 61–120 W
Lin = Linearity control, suitable for the deflection unit L_j
Fu = Medium slow-acting fuse 1.25 A
Ch_1 = Ch_2 = chokes 4 mH, 1 A

The black dots indicate those components that are available ex stock from our Central Components Depot in: 8510 Fürth-Bislohe, Postbox 146.

2.5. Monochrome Vidicon-TV-Camera

The block diagram of **Fig. 2.5.** demonstrates the different functions of a monochrome TV-camera. Fig. 2.5.4. shows a vidicon unit consisting of a deflection unit, a vidicon and a preamplifier, the latter incorporates a vidicon balancing network for test purposes operating an external video signal. The preamplifier also contains a frequency response equalization circuit, so that its video signal M shows only the resolution losses generated by the aperture failure of the pickup tube. The signal M is supplied to the mains amplifier, incorporating the total pulse processing for the camera, whereby all pulses are derived from one sync signal, which is fed to the main amplifier. Therefore an external synchronizing of the camera is easily possible. Preamplifier as well as main amplifier are dc-coupled. The latter has a clamping control circuit which assures, by means of a black-level reference band placed in front of the camera target pickup tube, the black-level is constant, independent of the dark current or interference signals. The main amplifier also contains the white level limiter, the blanking stage and a circuit which adds the sync pulses to the multiplex signal.

The deflection unit generates a blanking signal. It is amplified and blanks the cathode of the vidicon. To control the black level by a reference band in front of the target as mentioned above, the device has to operate with a shortened flyback time of about 5 µs, since the porch is used for generating the reference during the horizontal blanking interval. Therefore the blanking pulses available from the deflection unit are accordingly smaller than those used for a standard blanking of the main amplifier. If there is no deflection signal the vidicon is automatically turned off by the blanking signal A*. Fig. 2.5.1. shows a dc-converter operated by the 12 V power supply of the camera. It supplies the voltages for the vidicon, the deflection unit and the camera-preamplifier. The circuit of Fig. 2.5.2. shows the vidicon socket circuit in conjunction with the filtering components and the trigger amplifier for the blanking pulses.

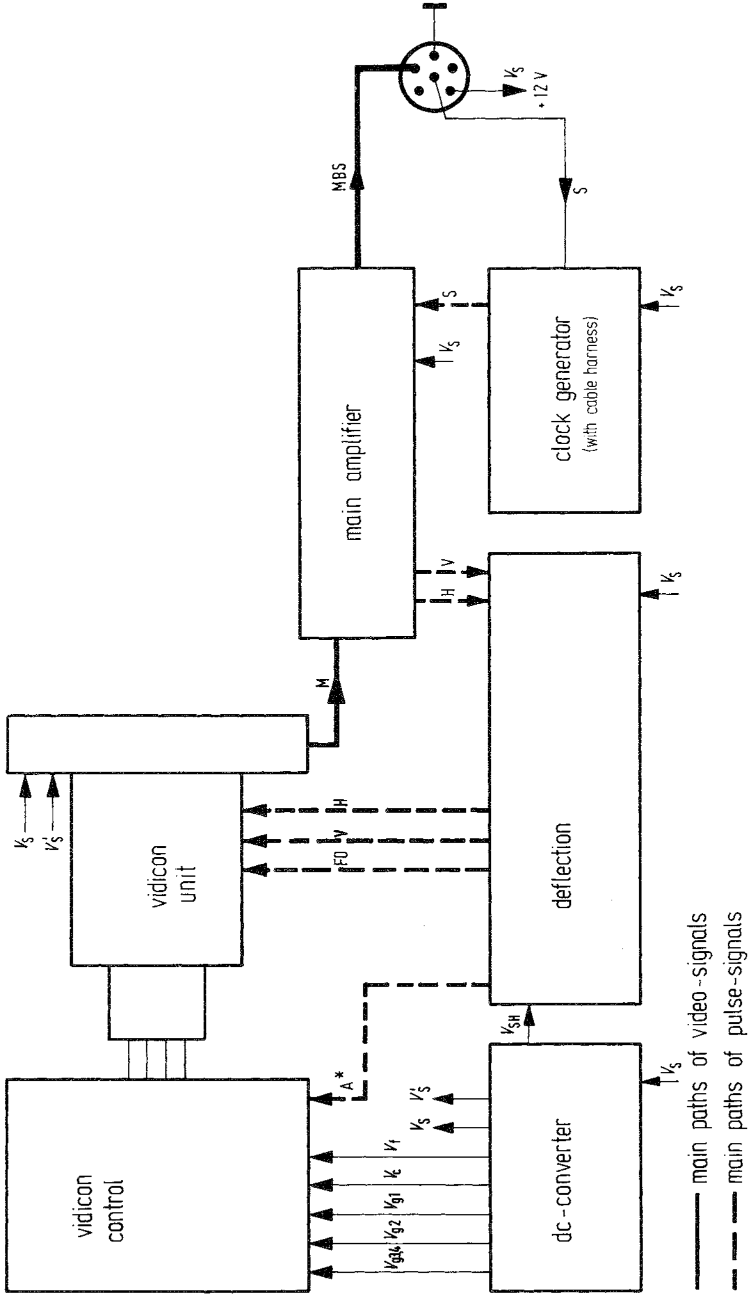
The camera contains a clock generator, also maintaining the connections to the other circuits. This generator is equipped with an automatic switch which allows the useage of external sync signals instead of the internal ones for synchronizing the camera via the main amplifier. If there is no external signal the clock generator supplies within 2 ms, internal sync pulses automatically. The external synchronization can be achieved either by sync pulses or by a video signal containing standardized sync pulses. By a switch of the clock generator the internal synchronizing can be enforced even if the external one is also existent. The sync pulse S used in the camera, the video signal M and the power supply voltage of +12 V are available at the 6 pins of the standard plug.

On account of their high resolution, their low inertia, and their good protection against overexposures and burn-in effects, multi-diodes-vidicons of the family XQ 1200 are favoured for applications in TV-cameras.

To take full advantage of these good features, it is essential that highly sophisticated circuits are designed to meet the requirements of these cameras. The circuit to be described was developed especially for operations with a multidiode vidicon. Details of the different functions can be found in the following chapters.

2.5.1. DC-Converter

This unit generates from a 12 V power supply the different supply voltages for a TV pickup camera. This circuit operates as a blocking oscillator at line frequency and it is triggered by positive line pulses applied at the terminal X_{13}.



main amplifier
clock generator
(with cable harness)
deflection
vidicon unit
vidicon control
dc-converter
MBS
S
M
H
V
FO
A*
V_S
V_S'
V_{SH}
V_f
V_c
V_{g1}
V_{g2}
$V_{g3,4}$
+12 V
main paths of video-signals
main paths of pulse-signals

Fig. 2.5.

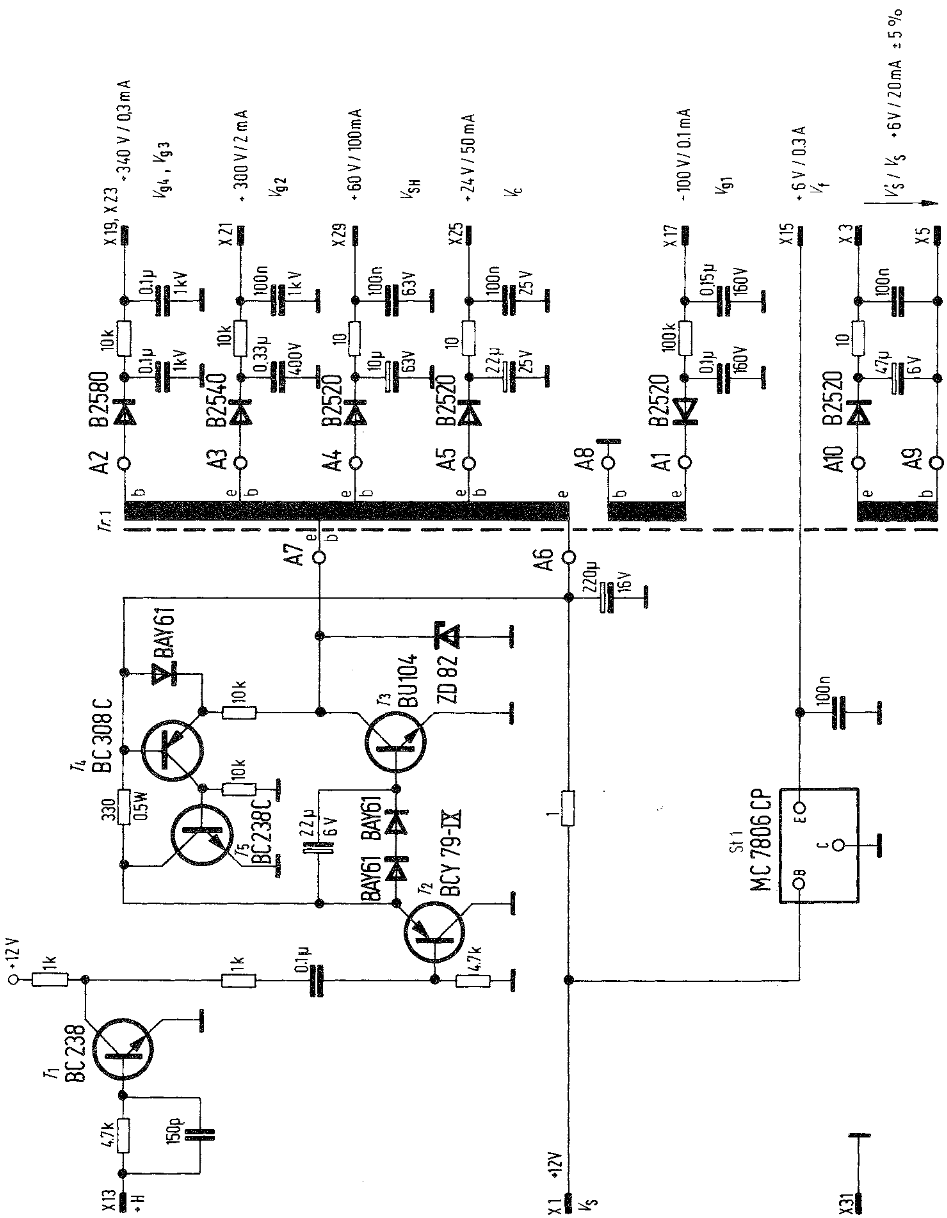

X 19, X 23 +340 V / 0,3 mA
V_{g4}, V_{g3}
X 21 +300 V / 2 mA
V_{g2}
X 29 +60 V / 100 mA
V_{SH}
X 25 +24 V / 50 mA
V_c
X 17 −100 V / 0.1 mA
V_{g1}
X 15 +6 V / 0.3 A
V_f
X 3
X 5
V_S' / V_S +6 V / 20 mA ±5 %
B2580
B2540
B2520
A1
A2
A3
A4
A5
A6
A7
A8
A9
A10
Tr. 1
BC 308 C
BC238C
BAY61
BU104
ZD 82
BCY 79-IX
BC 238
St 1
MC 7806 CP
X 13 +H
X 1 V_S +12V
X 31
+12 V

Fig. 2.5.1.

The switching transistor T_3 is controlled by an inverter T_1 via the driver stage T_2. The feedback transistors T_4 and T_5 improve the switching behaviour. The operating voltages for the multidiodes vidicon (300 and 340V), the voltages for the horizontal deflection (+60 V), the operating voltage for the blanking at the cathode of the vidicon (+24 V) as well as the negative voltage for grid one of the vidicon, are supplied by the transformer Tr_1 serving as an autotransformer. Besides that an ungrounded voltage of +6 V is available for the preamplifier of the camera. The filament voltage of +12 V for the vidicon is generated by the integrated regulator St_1. Since the dc-converter operates at line frequency, interference signals are not visible.

Transformer Tr_1 for the dc-converter

Bobbin:	B 65702–A 0000–M 001
1st Layer:	Beginning right above, 15.5 turns 0.1 enamelled copper wire tap right below; 45 turns 0.1 enamelled copper wire (= 60.5 turns total). Insulation 2 layers of Tesa-film 108/12 mm wide.
2nd Layer:	Continue to wind with 55.5 turns 0.1 enamelled copper wire (= 116 turns); end is left above (insulated sleeving). Insulation as mentioned above.
3rd Layer:	Beginning left above, 7.5 turns 0.5 enamelled copper wire tap left below, 7.5 turns 0.5 e.c.w. Insulation as mentioned above.
4th Layer:	Continue to wind 8.5 turns, 0.5 e.c.w. tap right below, 6 turns 0.5 e.c.w., end left below. Insulation as mentioned above.
5th Layer:	Beginning right above, 18 turns 0.4 e.c.w., end left above. Insulation as mentioned above.
6th Layer:	Beginning right above, 44 turns, 0.15 e.c.w., end left above, all wire ends are covered with insulated sleeving.
Core:	B 65701–L 0630–A 022 (30 ⌀ × 19).
Holder:	B 65705–A 0001–X 000.

2.5.2. Vidicon Control

This unit, serving as a bracket board for the vidicon socket, contains all filtering components for the supply voltages of the different vidicon electrodes. The switching transistor T_1 is responsible for the blanking of the signal at the vidicon cathode. It receives via the terminal X_{27} external blanking pulses which are generated in the deflection unit 2.5.3. The voltages for grid 3 and 1 (beam current) are adjustable by the potentiometers P_1 and P_2. The vidicon is turned off automatically if there are no blanking pulses. In this case the base of transistor T_1 is switched off and a positive voltage is supplied via R_2 to the cathode **(Fig. 2.5.2.)**.

2.5.3. Deflection Unit

This unit contains all circuits required to drive the deflection coils of a 1"-vidicon, i.e. focusing, horizontal and vertical deflection. Besides that there is a circuit for flyback blanking operations. The block diagram demonstrates the different functions

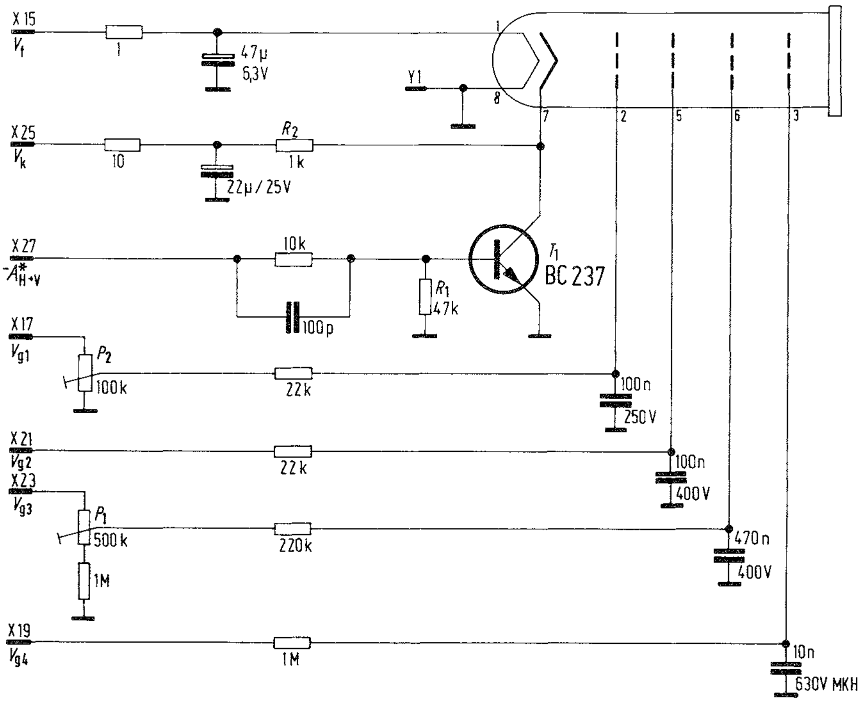


Fig. 2.5.2.

(Fig. 2.5.3.). The vertical deflection amplifier is current controlled. The deflection current is measured as an actual value by a sensing resistor and is compared with the desired value of the current applied to the input of the amplifier. The horizontal deflection operates as a voltage controlled circuit, whereby a strong feedback of the deflection amplifier 7 is also utilized.

The reference signal for the vertical deflection is achieved by charging a capacitor from a constant current source (no. 2). At the beginning of each flyback the capacitor is discharged by means of the switch 4, which is controlled by the positive vertical deflection pulses. The reference signal is saw-tooth shaped in accordance with the required deflection current. The reference signal for the horizontal deflection consists of square wave pulses with superimposed saw-tooth ones. This corresponds to the required voltage at the horizontal deflection coil. The horizontal reference voltage is achieved by a constant current source (no. 1), the current of which flows through a capacitor and a resistor connected in series. During the flyback this RC circuit is shorted by the switch no. 3 being controlled by the positive horizontal deflection pulse. Both constant-current sources (no. 1 and no. 2) can be commonly adjusted by means of potentiometer 5, whereby the picture size is also controlled. The additional potentiometer 6 only influences the source no. 2, i. e. the picture height.

Due to the voltage control and the different tolerances a linearity potentiometer is required for the horizontal deflection. It is connected in series with the deflection coil and adjusts the resistive component of the deflection circuit. A low dc current is

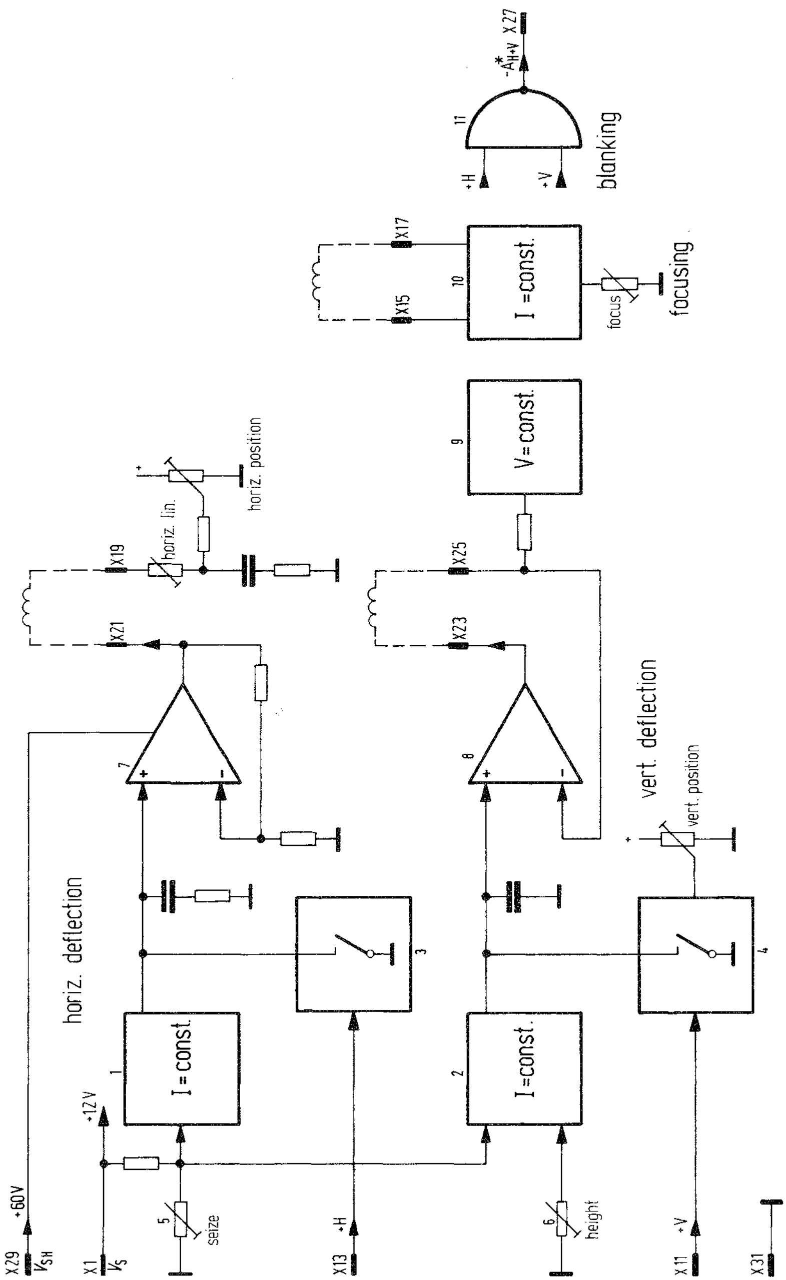

horiz. deflection
vert. deflection
I = const.
I = const.
V = const.
I = const.
blanking
focusing
focus
horiz. lin.
horiz. position
vert. position
seize
height
+60V
+12V
+H
+V
X29
X1
X13
X11
X31
X19
X21
X23
X25
X15
X17
X27

Fig. 2.5.3.

superimposed to the ac current flowing through the horizontal deflection coil to adjust the horizontal picture position. The vertical deflection is dc coupled to avoid capacitors with high capacitances. To achieve an ac current for the vertical deflection coils the low end of the deflection circuit (precision resistor) is connected to the constant current source no. 9. The focusing is obtained by the constant current source no. 10. Both positive, horizontal and vertical pulses being used for the flyback control are combined in a NOR-gate to blanking pulses which are available at terminal X_{27}. These pulses have to be amplified by a special circuit being incorporated in the vidicon control unit (2.5.2.).

The circuit of **Fig. 2.5.3.1.** shows further details. The constant currents for the reference signals are adjusted by means of the voltage divider consisting of R_1, R_2 and P_1. It is connected to the base of the emitter follower T_1 which is utilized to compensate for the temperature effects of the base emitter diodes of the following constant-current transistors T_2 and T_3. The emitter resistor of T_3 is variable and adjusts the picture height. C_1 is charged by T_3, and discharged by T_4 during the flyback interval. Thus a saw-tooth voltage is generated at C_1. The discharging voltage is determined by R_3, P_2 and R_4. The operation point of the following amplifier, as well

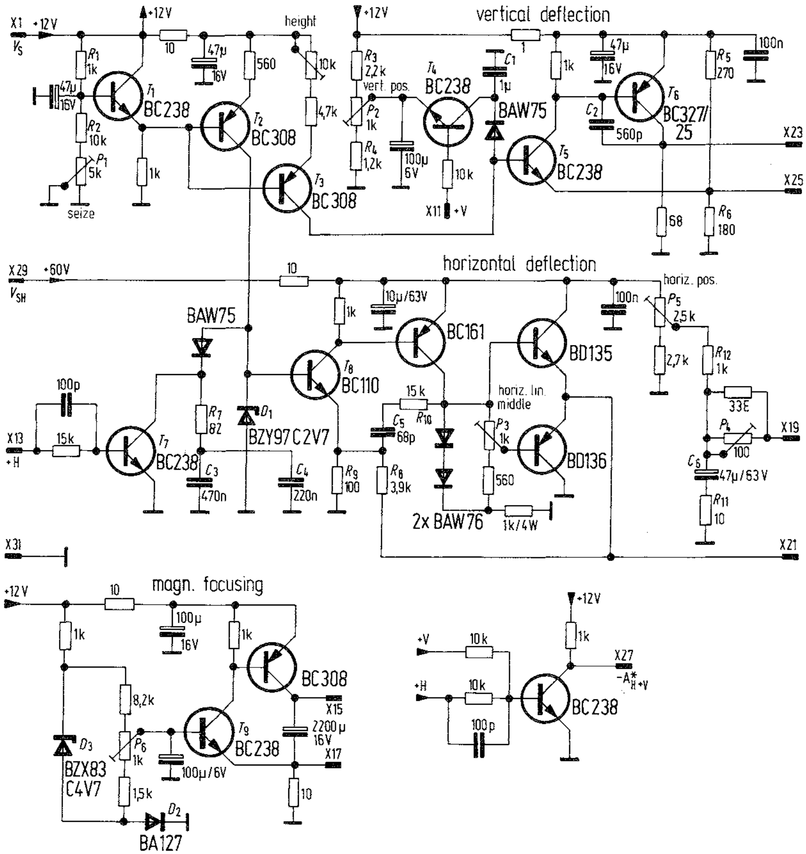


Fig. 2.5.3.1.

as the vertical picture position, is set by the potentiometer P_2. The amplifier incorporates the transistors T_5 and T_6. The reference signal (desired value) is applied to the base of T_5 whereas the actual value of the deflection current measured by the resistors R_5 and R_6 is available at the emitter of T_5. C_2 limits the bandwidth of the amplifier and assures that no undesired oscillations are generated.

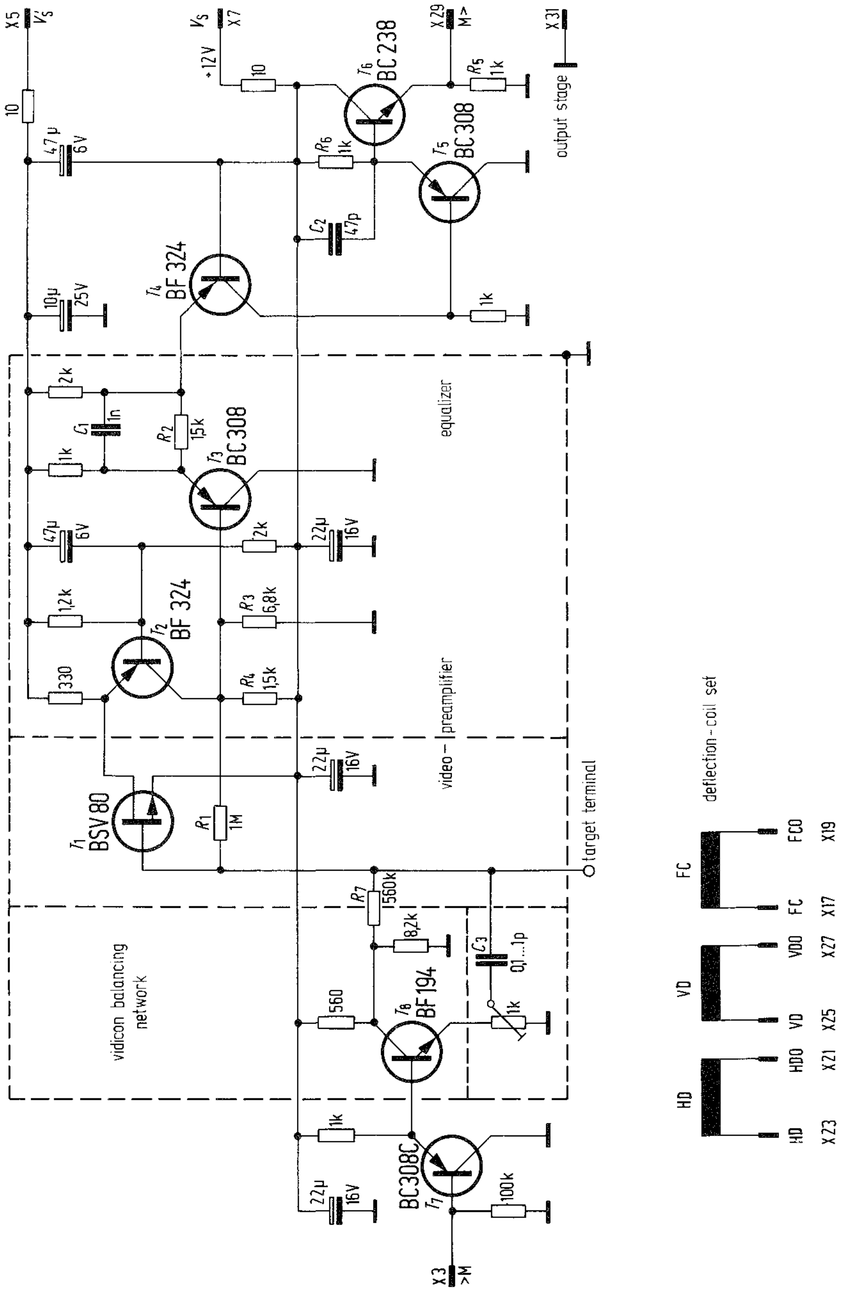
The transistor T_2 serves as a constant current source for the horizontal reference signal. At the RC-circuit consisting of R_7, C_3 and C_4 a pulse with a superimposed saw-tooth one is generated. T_7 is turned on by positive line pulses during the flyback interval. The z-diode D_1 prevents the reference signal from increasing if no driving pulse is applied. The deflection amplifier operates as a complementary push-pull type. The reference signal (desired value) is applied to the base of T_8. The actual value is supplied via the voltage divider R_8 and R_9 from the output terminal X_{21} to the inverted input of the amplifier (emitter of T_8).

The potentiometer P_3 is for adjustment so that the current supply joins in at just the middle of a line without any irregularities. The RC-circuit consisting of R_{10} and C_5 equalizes the frequency response. P_4 acts as a linearity control potentiometer, C_6 is the coupling capacitor for the deflection circuit, and the deflection current is measured across R_{11}. A dc current is supplied from P_5 via R_{12} to the horizontal picture-position potentiometer.

In the far left corner of the following schematic 2.5.3.1. the constant current source for the static magnetic focusing is shown. The diodes D_2 and D_3 compensate for the temperature effects of the base-emitter diode of T_9. The focusing current is adjusted by the potentiometer P_6, i.e. by the bias of T_9.

2.5.4. Vidicon Unit

The vidicon unit consists of a video preamplifier, a balancing network for test purposes and the deflection coils for the vidicon, whereby these circuits are connected to the harness of the camera by means of a 31-pin-plug. The video preamplifier incorporates the input stage, the equalizing circuit and the output stage. It is suitable for applications with multidiode vidicons, in accordance to a plate voltage of 10 V. Therefore a very simple dc coupled amplifier can be realized. The input stage consists of a FET T_1 and a bipolar transistor T_2 being in cascade connection. The gain is determined by the negative feedback resistor R_1. The operating voltage for the input stage is 6 V. It is added to the power supply voltage of 12 V generally used in this camera system. Due to the high internal gain the frequency is only influenced by the input capacitance and the input resistance connected in parallel. Therefore a very simple equalization can be achieved by means of an RC-circuit. It is accomplished by the differential amplifier T_3, T_4, whereby the resistor R_2 and the capacitor C_1 are connected directly to the emitter of T_3. The bias of the FET T_1 is set by the resistors R_3 and R_4. Since a gate voltage of -2 V is required the total voltage between gate and ground is $+10$ V. This corresponds to the target voltage, required by multidiode vidicons. By the differential amplifier the high dc level at the base of T_3 can easily be converted to a low dc level at the base of T_5. AC currents of opposite phase are superimposed to the dc currents of T_1 and T_2 respectively T_3 and T_4. Therefore no ac loading of the power supplies occurs, i.e. no cross-coupling or earth currents exist. This has also been achieved for the output stage T_5 and T_6 by using two complementary emitter followers, and by adequately proportioning the resistances R_5 and R_6. If longer cables are connected to the output X_{29}, this symmetrizing can also be obtained by incorporating the cable capacitance to C_2. For test purposes an active



vidicon balancing network
video – preamplifier
equalizer
output stage
target terminal
deflection - coil set
BSV 80
BF 324
BC 308
BF 324
BC 238
BC 308
BF 194
BC308C
X5
X7
X29
X31
X3
+12 V
HD
VD
FC
HD0
VD0
FC0
X23
X21
X25
X27
X17
X19

Fig. 2.5.4.

vidicon balancing circuit is added. By a suitable dimensioning and by using complementary transistors it has been achieved that no ac-loading of the power supply occurs. Thus the quantity of the filtering components, in the amplifier, can be minimised, i.e. a space saving construction with an optimum of electrical behaviour is attained. The test signal is supplied with the appropriate level to the vidicon balancing circuit via the terminal X_3. Since the harmful shunt capacitance of the resistor R_7 will result in an increase of the signal current at higher frequencies, it is compensated by an antiphase, capacitive current supplied via C_3. For its adjustment a multiburst signal is applied via the vidicon balancing circuit. Then the output signal of the video amplifier is set to a minimum at high frequencies. With a device built up as a demonstration sample the harmful shunt capacitance could be reduced to a very low value because of screening the resistor R_7 and due to a careful build-up. Therefore only a small compensation was required and thus compensation measures can be avoided up to frequencies of 10 MHz. These errors are caused by the phase shift between the signal at the collector, and the emitter of the transistor T_8. They result in unwanted effects, especially if the shunt capacitance is high.

2.5.5. Camera Amplifier

The circuit processes the signal from the preamplifier of the TV pickup-camera and forms a standardized video signal. A frequency response equalization is not necessary since this has already been realized in the preamplifier described in chapter 2.5.4. The main amplifier can be operated either dc or ac coupled. It contains a clamp control circuit for restoration or clamping of the black level.
In conjunction with a test mask placed in front of the pickup tube and by using deflection circuits with a shortened flyback interval, it is possible to create a black-level reference on the porch. By that a compensation of the dark current and elimination of scattered light effects can be achieved. This described circuit also contains a white-level limiter, a blanking stage, and a sync-pulse adder, which is connected to the video signal path.
Besides that it contains all pulse processing necessary for the operation of a TV-camera, whereby all pulses are derived from the supplied S-pulses, or from any video signal incorporating S-pulses. Therefore an external synchronization is easily possible. For external pulses there are available: clamping pulses, positive h-pulses with a duration of 5 μs for driving the horizontal deflection and positive V-pulses with a duration of 1 ms for driving the vertical deflection.
The circuit is shown in **Fig. 2.5.5.1.** The input amplifier can be selectively operated either with a direct coupling or a capacitive one using a bootstrap circuit. The black-level stabilization is achieved by a correction signal ΔW, being supplied via the resistor R_1. The white level limiter is combined with the blanking stage. The black-level limiter is blanked with sync pulses (−S) by the diode D_1. It also enables a mix-in of the sync pulses. The reference signals for black and white level are derived from the voltage divider consisting of R_2, R_3, P_1 and R_4. P_1 also supplies the actual value for the clamping circuit. Therefore a very constant black level is obtained at the camera output. All stages between the test pickup circuit for the clamping control and the camera output respectively the mix-in circuit of the black level are temperature compensated by using complementary transistors. The transistor T_1 serves as a driver for the sample and hold circuit of the clamping control. The transistor T_2 compensates for the temperature effects of T_1 which supplies the signal to the control amplifier IC_1.

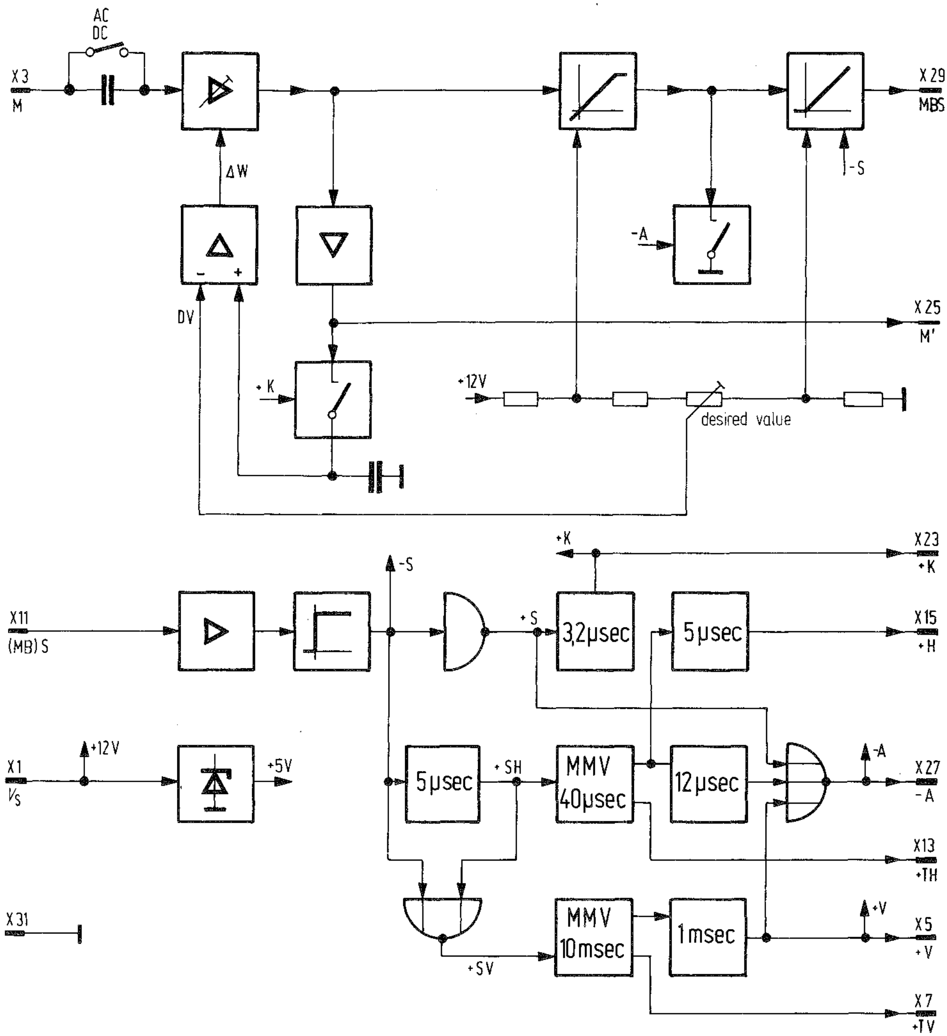


Fig. 2.5.5.

The first stage of the pulse processing circuit is a preamplifier with inverted inputs. Its bandwidth is limited to about 1 MHz to reduce interference influences. Then a pulse isolation circuit and a differential stage with a limiter T_4 follow. The latter shortens the sync pulses of the horizontal synchronization to about 5 µs, so that vertical sync pulses, also shortened, are generated. These pulses are combined in an OR-gate with regular sync pulses in such a way that all line pulses are blanked out and only five positive vertical sync pulses are available at the output of the gate IC_2, during the interval of the vertical sync pulse. To suppress the tails and the vertical sync pulses with double frequency, the horizontal sync pulses SH are applied to a monostable multivibrator (IC_3 and T_5), which has a delay time of about 40 µs. The

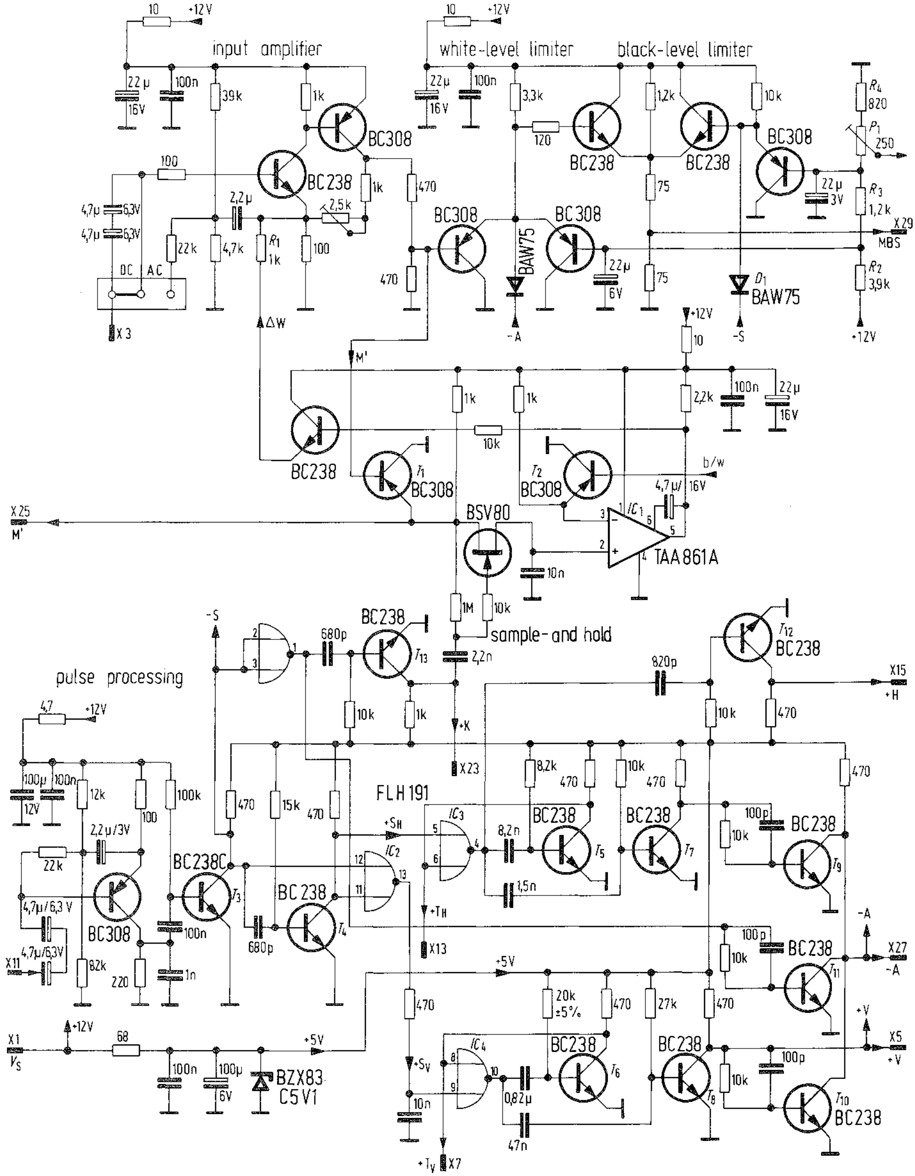


Fig. 2.5.5.1.

vertical sync pulses SV are supplied to another monostable multivibrator (IC_4 and T_6). It has a delay time of about 20 ms and suppresses the multiple pulses. The pulses TV and TH generated by these multivibrators can be externally utilized. By differentiation and limiting achieved in the stages T_7 and T_8 blanking pulses with the frequency of the horizontal deflection and those corresponding to the vertical deflection are

achieved. These are combined with a blanking signal composition A by the transistors T_9 and T_{10}. For forward shifting of the leading edge of the horizontal blanking pulses, sync pulses are added to the signal A via T_{11}. Also by differentiation and limiting in the stage T_{12}, horizontal pulses with a duration of 5 µs are generated to drive the horizontal deflection circuit. The positive V-pulses are supplied to terminal X_5 in order to drive the vertical deflection circuit. The clamping pulses K required for the clamp control are generated from the trailing edge of the positive sync pulses S by differentiation and limiting in T_{13}. They are externally available at the terminal X_{23}.
The described circuit is characterized by a very safe operation. By using a dc-coupled, temperature compensated video-amplifier, and by the clamping control circuit, a direct coupling from the vidicon to the output can be realized. Thus the minimum of signal distortion is received. Since all pulses are derived from the sync signal, an external synchronization of the TV camera is easily possible. In this case only a switch-over from the S signal, internally produced by the camera clock generator, to the external S signal is required. This can be automatically obtained by a simple circuit incorporating a clock generator.
Therefore an externally synchronized clock generator is not required, i.e. the elaborations are minimized.
Besides that the described circuit can also be used as a stabilizing amplifier. In this case a switch over from dc to ac coupling has to be considered and a composed multiplex blanking sync signal (MBS) can be applied, e.g. to terminal X_3. If the same signal is also supplied to X_{11} it achieves the synchronization of this total device.

2.5.6. Clock Generator and Harness

This unit is responsible for the synchronization of the complete TV-camera. It also incorporates all connections to the other units, previously described. The clock generator is to be used in systems, where all vertical and horizontal pulses, the blanking and clamping signals of the source being synchronized are derived from one S-pulse. Therefore, it only generates S-pulses, although a production of h, v, a- and k-pulses is practically possible because of the chosen circuit concept and by using additional gates. By driving all auxiliary pulses from the S signal a very simple concept is also attained for an external synchronization of the video-signal sources. It can be realized by any video-signal containing standardized sync pulses. For this purpose an external automatic-synchronization is considered for the clock generator. It automatically switches from the operation with an internal S signal, to the one with an external S signal, derived from a video signal. If no external sync signal exists the internal S signal is automatically switched over, i.e. an operation with internal S signal is forced.
The function of the clock generator shown in the block diagram **Fig. 2.5.6.** is as follows:
All signals are digitally processed. The clock pulses are supplied by a crystal generator (1), oscillating with 20 times the line frequency (312.4 kHz). The double line frequency is obtained by a decade divider (2). The decoder (10) supplies the different line pulse trains for generating the sync pulses. The divider no. 3 divides the double line frequency by 5 and the divider no. 11 by 2 in order to achieve the line frequency. The divider 3 indicates 2.5-line-intervals for the pre and post-tails as

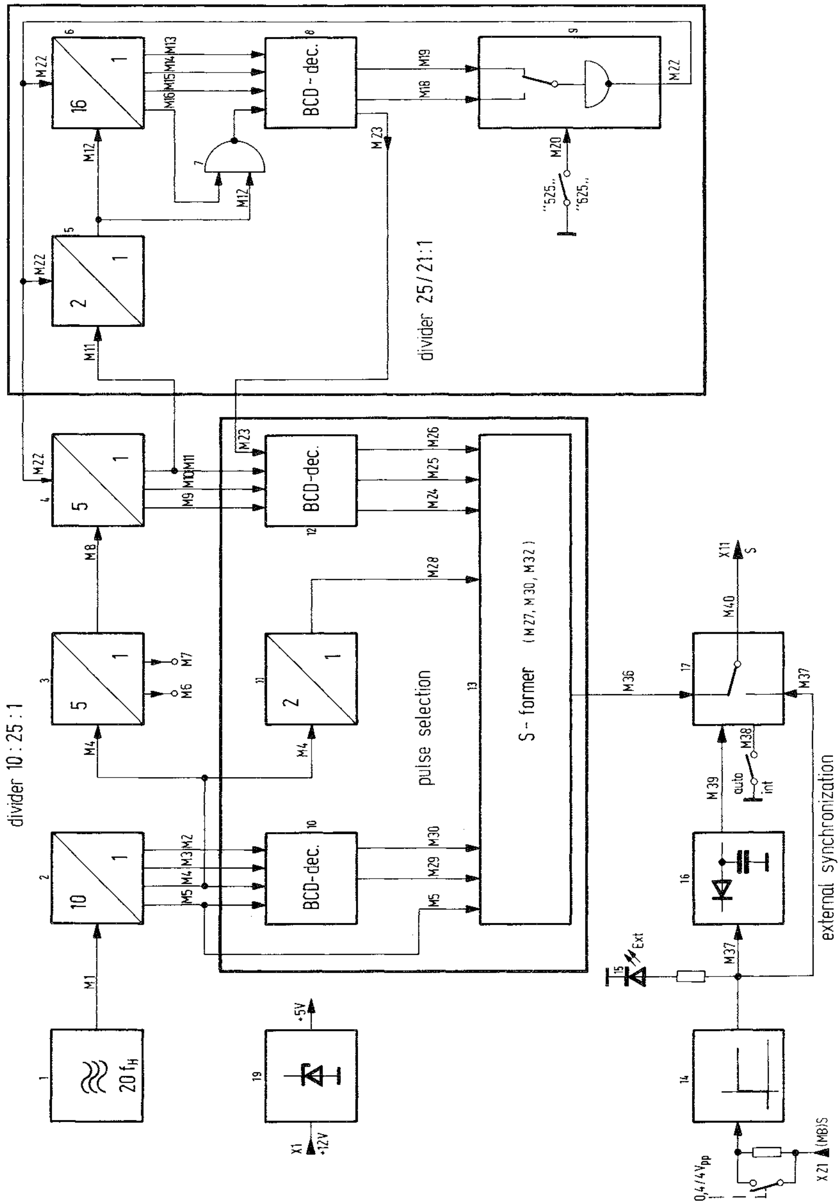

divider 25/21:1
BCD-dec.
divider 10 : 25 : 1
pulse selection
S - former (M27, M30, M32)
external synchronization
20 f_H
+5V
+12V
"525.."
"625.."
auto
int
Ext

Fig. 2.5.6.

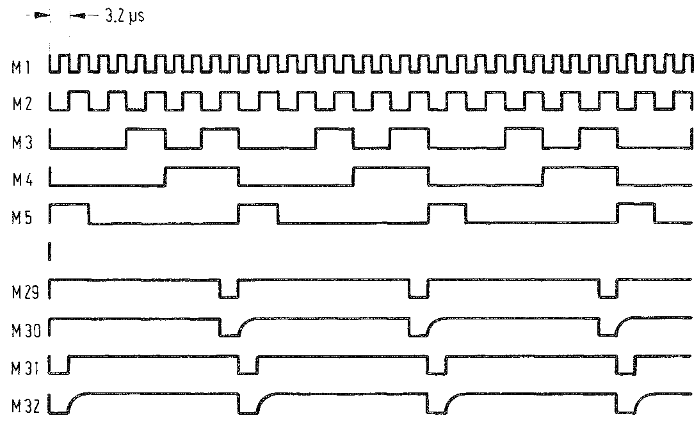


Fig. 2.5.6.1.

well as the wide v-sync pulses. The following chain of dividers (4, 5, 6) generate the field frequency. The double line frequency has to be divided by 625 according to the CCIR standards and by 525 according to the FCC standards. Since a division by 25 is already achieved by the stages 3 and 4 a counter with a selectable division of 25 or 21 in addition is required. Gate 7 and decoder 8 supply the suitable reset pulses, which are applied to the reset inputs of the dividers 5 and 6 by the reset selection circuit 9. The reset input of the divider 4 is also involved for special reasons, resulting from the concept of the used IC's. The gate pulses for the tails and the v-pulses are produced by the decoder 12. The sync signal composition S is composed by a combination of gates (13).
For external synchronization any video signal can be applied to the terminal X_{21}. Its sync pulses are separated by stage 14 and rectified by stage 16. They generate a control voltage by which switch 17 can be operated. This switch is responsible for selecting the operation of an externally or internally supplied S signal, which is applied to the output X_{11}. By means of an additional switch an operation with an internal sync signal can be selected instead of the automatic operation.
Besides that the amplitude of the externally supplied sync signal can be chosen as being 4 V_{pp} or 0.4 V_{pp}, corresponding to the video signal. The presence of an external video signal, with sync pulses, is indicated by an LED (15) regardless of which operation mode has been chosen. An operation according to the standards of 625 or 525 lines can also be selected by this switch. The operation voltage is 5 V. If the supply voltage of a device should be 12 V a regulating circuit (19) has to be considered.
The circuit shown in **Fig. 2.5.6.3.** is described as follows: The crystal generator oscillates at 20 times the line frequency. The IC_1 serves as the first decade divider, whereas the IC_3 is responsible for the division by 2 in order to produce the line frequency. The latter also contains the first 5:1 divider of the chain producing the field frequency. The other dividers of the chain (5:1 and 2:1) are incorporated in the IC_4, and one (16:1) in the IC_6. The decoder IC_7 and the gate G_1 decode suitable reset pulses, whereby M 18 is used for the 525-line-standard and M 19 for the 625-line-standard.

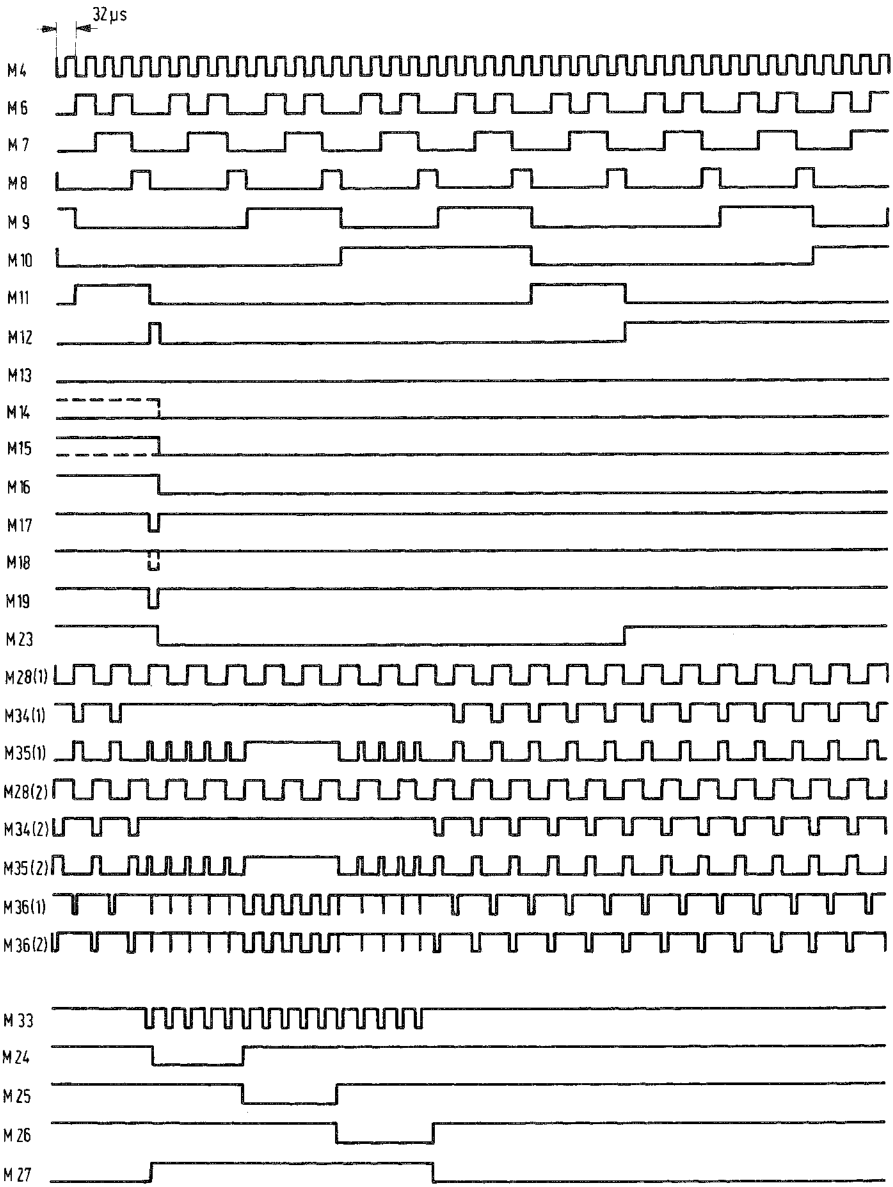

32µs
M4
M6
M7
M8
M9
M10
M11
M12
M13
M14
M15
M16
M17
M18
M19
M23
M28(1)
M34(1)
M35(1)
M28(2)
M34(2)
M35(2)
M36(1)
M36(2)
M33
M24
M25
M26
M27

Fig. 2.5.6.2.

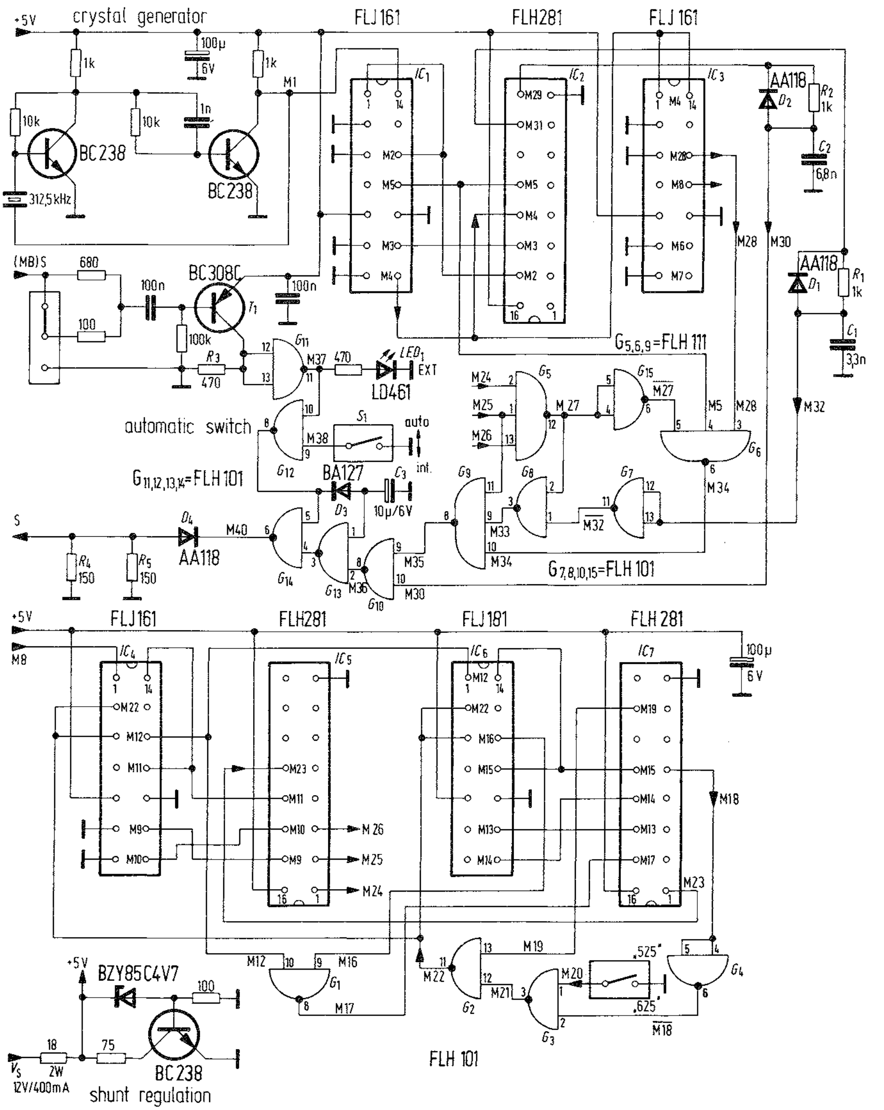

crystal generator
FLJ 161
FLH281
FLJ 161
312,5 kHz
BC238
BC308C
automatic switch
LD461
auto
int.
BA127
G11,12,13,14=FLH 101
G5,6,9=FLH 111
G7,8,10,15=FLH 101
AA118
FLJ161
FLH281
FLJ181
FLH 281
BZY85C4V7
shunt regulation
12V/400mA
FLH 101
„525"
„625"

Fig. 2.5.6.3.

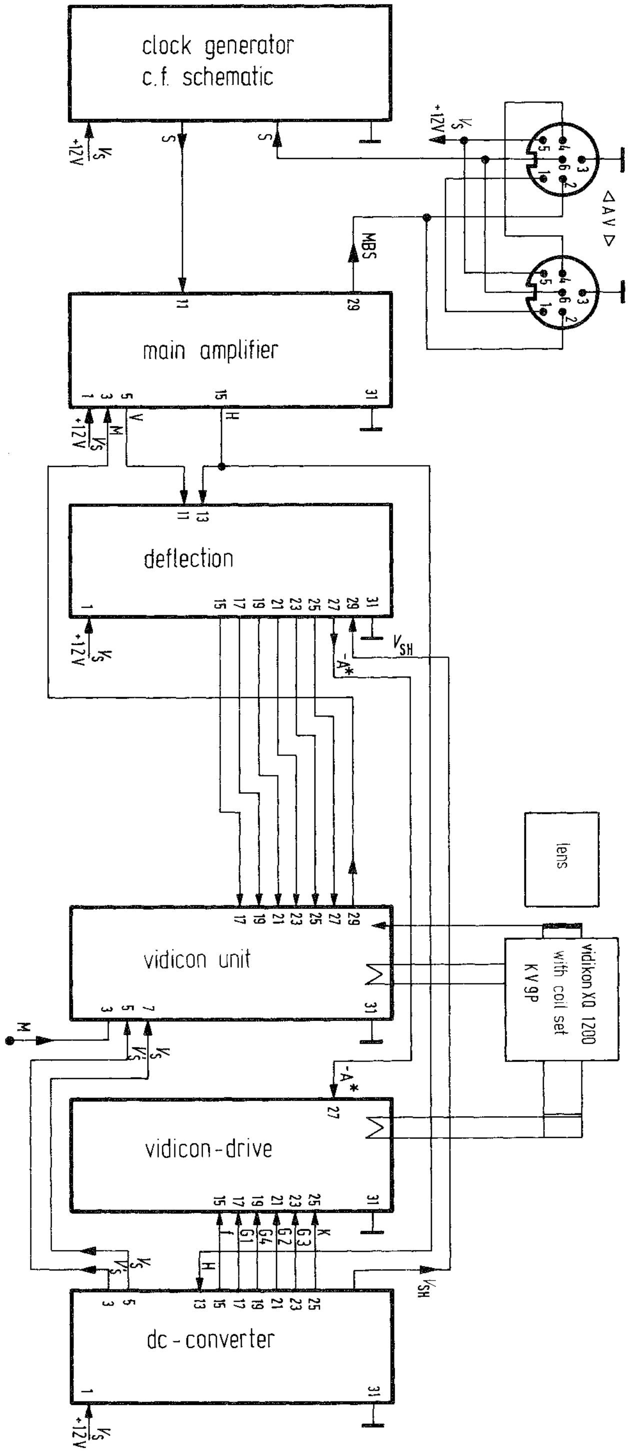


Fig. 2.5.6.4.

total block diagram

The appropriate reset pulses are set by gates G_2 to G_4. The decoder IS_5 forms the gate pulses for the tails and the vertical sync pulses, i.e. M 24 to M 26. The basic pulses for forming the tails (M 31) and the leading edge of the sync pulses (M 29) are decoded by the decoder IC_2. The duration of the tail pulses has to be extended slightly in comparison to the raster determined by the clock frequency. This is achieved by the combination of components D_1, R_1 and C_1. Besides that the leading edge of the sync pulses has to be slightly delayed, which is obtained by the circuit consisting of D_2, R_2 and C_2. A wide auxiliary vertical pulse M 27 is generated from M 24, M 25 and M 26 by means of the gate G_5. Gate G_6 combines M 28 and M 5 and produces a continuous h-sync pulse train, which is only interrupted by the auxiliary v-pulse M 27. The tail pulses are inverted by gate G_7 and supplied together with M 27 to the gate G_8. At its output these pulses are available as a pulse train (M 33) during the time of the auxiliary v-pulses. By the combination of M 33 and M 34 one part of the sync-signal composition has already been produced. It has only to be interrupted at gate G_9 by the pulse M 25 during the time of the transmitted vertical sync pulse. The output signal M 35 of the gate G_9 shows a positive pulse at same, where the subdivided vertical sync pulses should later appear in the sync signal composition. This positive pulse is periodically interrupted at gate G_{10} by the signal M 30. Since the interruption duration occurs at time intervals where no pulses are transferred, M 30 can be supplied as a continuous pulse train. Moreover by the leading edge shift of M 30 all pulses are shortened. Thus an exactly defined leading edge of the sync pulse is achieved without showing any jitter. At the output of gate G_{10} the internal sync signal composition M 38 is available. An additional switching circuit is provided for the automatic switch-over to an external sync signal. The pulse separation stage T_1 is triggered by a variable attenuator. If an external sync signal is applied, positive sync pulses are available at R_3. These are inverted by the gate G_{11} and negative sync pulses apply at its output (M 27). If there are no external sync pulses available, a positive dc voltage drops across R_3. The output voltage of gate G_{11} is zero and the LED Le_1 cannot emit light. But if sync pulses exist, the output voltage is zero only during the interval of the pulses. During the rest of the time the output level is positive, i.e. the diode emits light and indicates that external sync pulses exist. The external sync pulses respectively the dc voltage is processed by gate G_{12}. If the switch S_1 is closed, the output voltage at gate G_{12} is positive. If S_1 is open the output voltage is positive when no external sync signal is applied. However, positive sync pulses are available, when the switch S_1 is open and when external sync signal exists. In other words, either positive sync pulses or a positive dc voltage are available at the output of gate G_{12}. If positive sync pulses are supplied to the diode D_3, serving as a rectifier, a low voltage drops across the capacitor C_3, since it is nearly discharged via the diode D_3 during the short sync pulse separation.

The capacitor C_3 cannot be charged to any degree during the short duration of the positive sync pulses supplied from gate G_{13}. But when there is no external sync signal or when the switch is in the internal position a positive dc voltage is generated at the output of gate G_{12}. In this case C_2 is positively charged and enables gate G_{13} so that the internal sync pulses M 36 can pass. Thus they are supplied to the output via gate G_{14}. However, if there are external sync pulses, gate G_{13} is blocked and a positive voltage is available at its output. This voltage enables gate G_{14} so that the positive sync pulses supplied from gate G_{12} can pass. At point M 40 negatively directed sync pulses are available. They are transferred to the output X_{11} via the diode

D_4. The resistors R_4 and R_5 limit the output voltage. A simple shunt regulation circuit is considered for obtaining the power supply voltage of 5 V.

List of Capacitors used in the Circuit 2.5.1.

Quantity	Components	Value	Rated Voltage (V)	Ordering codes
3	Ceramic disc-capacitors Type 3, SDPN	100 nF	16	B 37302–A 1104–Z 001
2	Ceramic flat-capacitors Type 2, ZDHU	100 nF	63	B 37449–A 4104–S 001
1	Styroflex capacitor	150 pF	25	B 31310–A 3151–H 000
1	MKM-stacked film capacitor	100 nF	100	B 32540–A 1104–J 000
1	MKM-stacked film capacitor	150 nF	100	B 32540–A 1154–J 000
3	MKH-capacitors	100 nF	400	B 32234–B 6104–M 000
1	MKH-capacitor	330 nF	400	B 32234–B 6334–M 000
1	Electrolytic capacitor	10 μ	63	B 41316–A 8106–Z 000
1	Electrolytic capacitor Tantalum, sinter type (solid)	22 μ	10	B 41313–A 3226–Z 000
1	Electrolytic capacitor Tantalum, sinter type (solid)	22 μ	25	B 41315–A 5226–Z 000
1	Electrolytic capacitor Tantalum, sinter type (solid)	47 μ	10	B 41315–A 3476–Z 000
1	Electrolytic capacitor	220 μ	16	B 41316–A 4227–Z 000

List of Capacitors used in the Circuit 2.5.2.

Quantity	Components	Value	Rated Voltage (V)	Ordering codes
1	Styroflex capacitor	100 pF	25	B 31310–A 3101–H 000
1	MKM-stacked film capacitor	100 μF	100	B 32540–A 1104–J 000
1	MKH-capacitor	10 nF	400	B 32234–B 6103–M 000
1	MKH-capacitor	100 nF	400	B 32234–B 6104–M 000
1	Electrolytic capacitor Tantalum, sinter type (solid)	22 μ	25	B 41315–A 5226–Z 000

List of Capacitors used in the Ciruit 2.5.3.

Quantity	Components	Value	Rated Voltage (V)	Ordering codes
1	Styroflex capacitor	68 pF	63	B 31310–A 5680–H 000
2	Styroflex capacitors	100 pF	25	B 31310–A 3101–H 000
1	Styroflex capacitor	560 pF	25	B 31310–A 3561–H 000
1	MKM-stacked film capacitor	1 µF	100	B 32541–A 1105–J 000
1	MKM-stacked film capacitor	220 nF	100	B 32541–A 1224–J 000
1	MKM-stacked film capacitor	470 nF	100	B 32541–A 1474–J 000
1	Ceramic flat-capacitor Type 2, ZDHU	100 nF	63	B 37449–A 4104–S 001
1	Ceramic disc-capacitor Type 3, SDPN	100 nF	16	B 37302–A 1104–Z 001
1	Electrolytic capacitor	10 µF	63	B 41316–A 8106–Z 000
3	Electrolytic capacitors Tantalum, sinter type (solid)	47 µF	25	B 41286–B 5476–T 000
1	Electrolytic capacitor	47 µF	63	B 41316–A 8476–Z 000
2	Electrolytic capacitors Tantalum, sinter type (solid)	100 µF	10	B 41286–B 3107–T 000
1	Electrolytic capacitor Tantalum, sinter type (solid)	100 µF	16	B 41286–A 4107–T 000
1	Electrolytic capacitor	2200 µF	16	B 41012–C 4228–T 000

List of Capacitors used in the Circuit 2.5.4.

Quantity	Components	Value	Rated Voltage (V)	Ordering codes
1	Styroflex capacitor	47 pF	63	B 31310–A 5470–H 000
1	Styroflex capacitor	1 n	25	B 31310–A 3102–H 000
1	Electrolytic capacitor Tantalum, sinter type (solid)	10 µF	25	B 45181–B 3106–M 000
2	Electrolytic capacitors Tantalum, sinter type (solid)	47 µF	6.3	B 45181–A 0476–M 000
3	Electrolytic capacitors Tantalum, sinter type (solid)	22 µF	16	B 45181–A 2226–M 000

List of Capacitors used in the Circuit 2.5.5.

Quantity	Components	Value	Rated Voltage (V)	Ordering codes
2	Styroflex capacitors	680 pF	25	B 31310–A 3681–H 000
1	Styroflex capacitor	820 nF	25	B 31310–A 3821–H 000
1	Styroflex capacitor	1 nF	25	B 31310–A 3102–H 000
1	Styroflex capacitor	1.5 nF	25	B 31310–A 3152–H 000
1	MKM-stacked film capacitor	8.2 nF	250	B 32540–A 3822–HJ 000
1	MKM-stacked film capacitor	10 nF	250	B 32540–A 3103–J 000
1	MKM-stacked film capacitor	47 nF	250	B 32540–A 3473–J 000
1	MKM-stacked film capacitor	820 nF	100	B 32541–A 1824–J 000
3	Ceramic disc-capacitors Type 2, SDPN	100 pF	500	B 37215–B 5101–S 001
1	Ceramic disc-capacitor Type 2, SDPN	2.2 nF	500	B 37238–J 5222–S 001
1	Ceramic disc-capacitor Type 3, SDPN	10 nF	16	B 37305–A 1103–Z 001
7	Ceramic disc-capacitors Type 3, SDPN	100 n	16	B 37302–A 1104–Z 001
2	Electrolytic capacitors Tantalum, synter type (solid)	2.2 µF	16	B 45134–J 4225–S 001
4	Electrolytic capacitors Tantalum, sinter type (solid)	4.7 µF	6.3	B 45134–J 2475–S 001
1	Electrolytic capacitor Tantalum, sinter type (solid)	4.7 µF	16	B 41315–A 4475–Z 000
2	Electrolytic capacitors Tantalum, sinter type (solid)	22 µF	10	B 41315–A 3226–Z 000
4	Electrolytic capacitors Tantalum, sinter type (solid)	22 µF	25	B 41315–A 5226–Z 000
1	Electrolytic capacitor Tantalum, sinter type (solid)	100 µF	6.3	B 41286–B 3107–T 000
1	Electrolytic capacitor Tantalum, sinter type (solid)	100 µF	16	B 41286–A 4107–T 000

List of Capacitors used in the Circuit 2.5.6.

Quantity	Components	Value	Rated Voltage (V)	Ordering codes
1	Ceramic disc-capacitor Type 2, SDPN	1 nF	500	B 37235–J 5102–S 001
1	Styroflex capacitor	3.3 nF	25	B 31310–A 3332–H 000
1	MKM-stacked film capacitor	6.8 nF	250	B 32540–A 3682–K 000
2	Ceramic disc-capacitors Type 3, SDPN	100 nF	16	B 37302–A 1104–Z 001
1	Electrolytic capacitor Tantalum, sinter type (solid)	10 μF	6.3	B 41315–A 2106–Z 000
2	Electrolytic capacitors Tantalum, sinter type (solid)	100 μF	10	B 41286–B 3107–Z 000

2.6. Linearity Test Demodulator

Meeting the tolerance limits of the raster geometry, plays an important part in receiving a good quality when TV-pictures are picked up or played back. Many different methods for testing the linearity of a TV-picture are known. The most popular is described in the following procedure. A special test diapositive, e.g. a raster test picture, is scanned with the pickup tube and electronically compared with a raster test signal, produced by an electronic test signal generator. The corresponding method for testing playback devices is as follows. The electronically produced raster is converted to a visual picture and compared with a raster of a test diapositive projected on the screen.

Between the comparison test points of the electronically generated raster with ideal geometry and the one of the test diapositive there exists a relatively large difference. Therefore linearity errors cannot always be determined. When the requirements of a picture quality are extremely high, the linearity error has necessarily to be determined at each picture point, i.e. the so-called differential linearity error has to be found.

The circuit described in the following allows determination of this error by means of a bar pattern which has a tilt angle of 22° with reference to the vertical picture axis. If this pattern is scanned in a horizontal direction a carrier frequency of 4.43 MHz is obtained. For testing the horizontal linearity the video signal corresponding to this bar pattern is applied to a FM-demodulator. To test the vertical linearity the phase relationship of two successive bars of the pattern are compared. Then the linearity error is calculated.

With TV-camera tests the test diapositive is illuminated by using a bank of lights and projected to the light sensitive layer of the pickup tube by means of a lens.

The scanned multiplex signal M of the pickup tube is applied to a video amplifier and then supplied to the linearity test device. All geometry errors of the electro-optical and the light-optical image are included.

When monitors are used for testing they are synchronized by sync pulses and adjusted in such a way that a white scanning raster is achieved on the screen. The raster is projected upon the test diapositive by means of a lens.

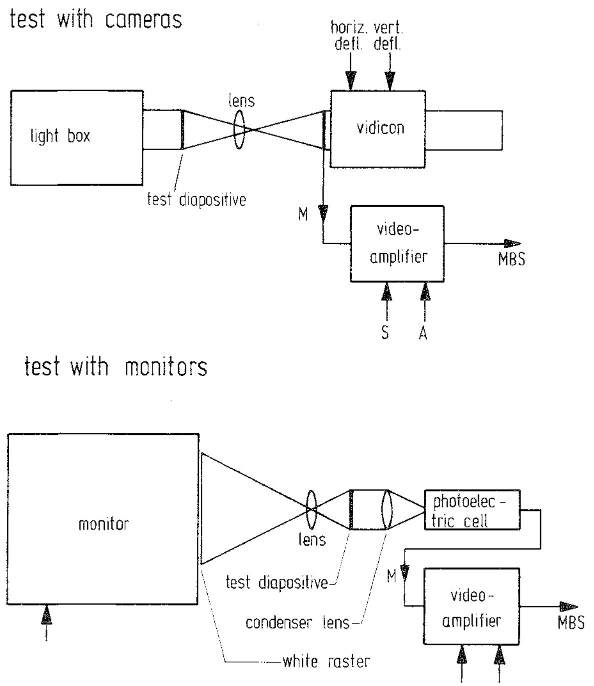


Fig. 2.6.

By using a condenser lens, a photoelectric cell and a video amplifier, a multiplex signal is produced. It is supplied to the linearity test device. In this application all electro-optical as well as light-optical defects of the image enter the test result. Only if the electro-optical image defects are of interest, as it is required e.g. at direct display units a lens of top quality has to be used. The principle of testing with monitors corresponds to the flying-spot method of scanning.

Fig. 2.6.1. shows the block diagram of the described linearity test device. Its functions are as follows. By the switch 9 two different test demodulators can be selected. For testing the horizontal linearity the multiplex signal, applied to terminal X_3, passes the band-pass filter 1 being tuned to 4.43 MHz. The following limiter 2 supplies a signal without any amplitude modulation directly to one of the inputs of the multiplier 4 and via the 90° phase shifter 3 to its second input. Stages 2 and 4 are incorporated in the IC TBA 120 operating as a FM-demodulator. For testing the vertical linearity the input signal is supplied to the multiplexer 8 via the bandpass filter 5 and the limiting amplifier 6. Besides that the signal from stage 6 is delayed by the duration of two lines in the delay-line circuit 7 and applied to the second input of the multiplexer. Thus a phase comparison of two successive lines is possible. The stages 6 and 8 are incorporated in an IC type TBA 120. The output signals of both demodulators are supplied selectively to the output amplifier 14 via the switch 9. Blanking and sync pulses are added to enable either a regular or a waveform-monitoring by applying any multiplex signal containing sync pulses to the terminal X_{11}.

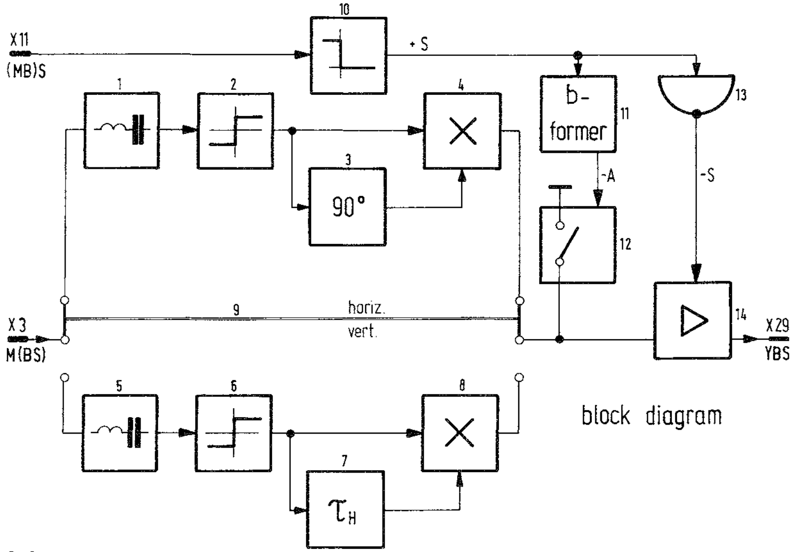
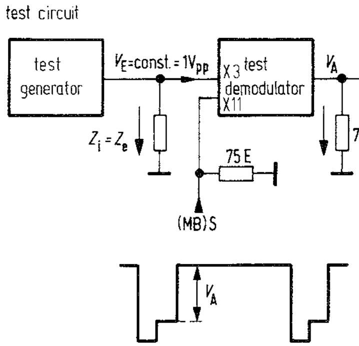


Fig. 2.6.1.

The sync pulses are separated by the amplitude filter 10 and supplied to the blanking former 11. It produces blanking pulses which short-circuit the input signal of the amplifier 14 by the blanking circuit 12. The sync pulses are inverted by gate 13 and added with negative polarity to the video signal. Thus a video blanking signal composition YBS is available at the output of the amplifier 14.

When using this test demodulator calibration curves indicating the relation between linearity error and output voltage are required. These curves can easily be obtained for testing the horizontal linearity by using a variable signal generator. An output voltage change of about 70 mV is obtained for a linearity error of 1 %. This corresponds to a frequency deviation of about 44 kHz at a carrier frequency of 4.43 MHz. The measurement range of the described device is about ± 100 kHz, which refers to an error of ± 2.5%. Since image converters of top quality require a linearity error of less than 1 %, this range is completely sufficient. The calibration curves for the vertical linearity can be derived from those obtained for the horizontal linearity, if one considers which phase relationship between two successive lines will occur, when a linearity error exists. In the case of the slanting bar pattern **(cf. Fig. 2.6.3.)** the phase difference between two successive lines is 90° if no linearity error is existent.

If the delayed and non-delayed signals are multiplied, the dc voltage component of the output signal will be zero. This refers to the centre point of the descriminator curve. With a linearity error of $+100$% the lines have twice the distance as before. This results in a phase jump of 180° from one line to another, i.e. in an additional phase shift of $\Delta\varphi = 90°$ with reference to a linearity error of zero. This corresponds to one of the descriminator curve humps. With a linearity error of -100% all lines are combined to one, since the picture height is zero. The phase shift of two successive lines is also zero, i.e. phase jumps of $\Delta\varphi = -90°$ occur in relation to the one without any linearity error. This corresponds to the second hump of the demodulator curve (see also **Fig. 2.6.3.** below). As the values of the voltages for the phase shift measurement of the horiz. linearity with a 4.43 MHz-bar-pattern



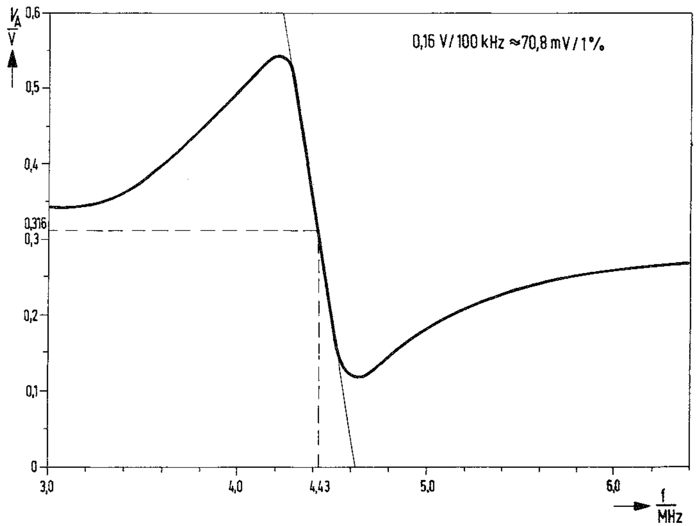


Fig. 2.6.2.

of $\pm 90°$ are already known from the calibration curve of the horizontal linearity, the descriminator sensitivity for the vertical linearity curve can be directly calculated. It is 6 mV for 1 %. In voltage reading there will be no difficulty if a better quality oscilloscope is used.

The circuit built up in praxis, is shown in **Fig. 2.6.4.** The test demodulator for the horizontal linearity consists of the input bandpass filter, the IC TBA 120 and the balanced phase shifting circuit fi_2 (90°). The capacitor C_1 suppresses interfering signals with twice the carrier frequency.

The test demodulator for the vertical linearity incorporates an input circuit (fi_3), an IC type TBA 120, a filter fi_4 and an ultrasonic delay line for 4.43 MHz, whereby its coils for the input- and output-matching are already integrated in the device. Since the delay line does not exactly delay the signal for a duration of a line, but only for an integral multiple of the carrier frequency, one line receives extra oscillations

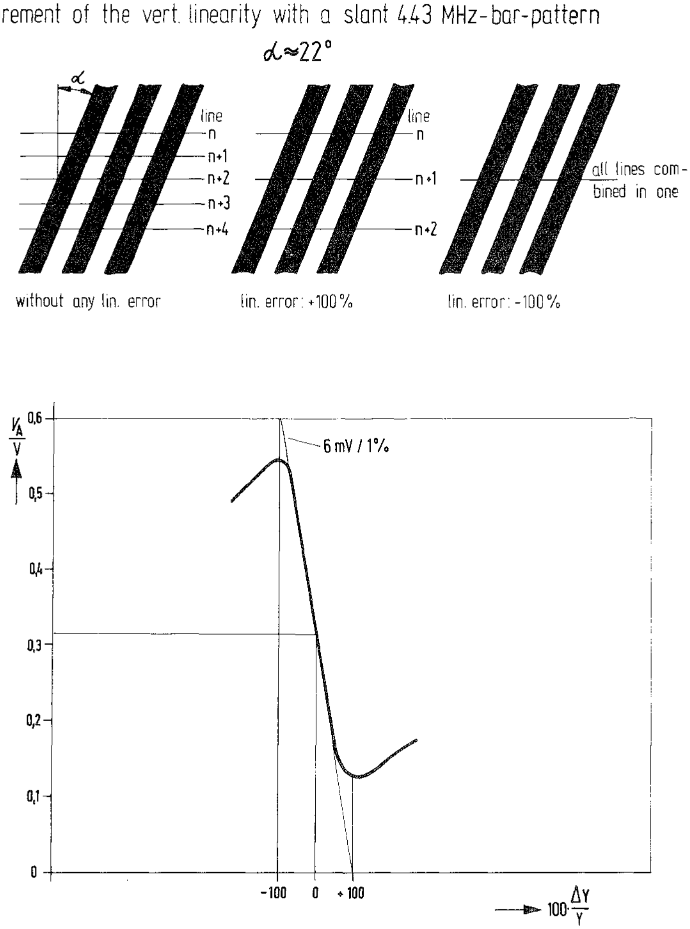
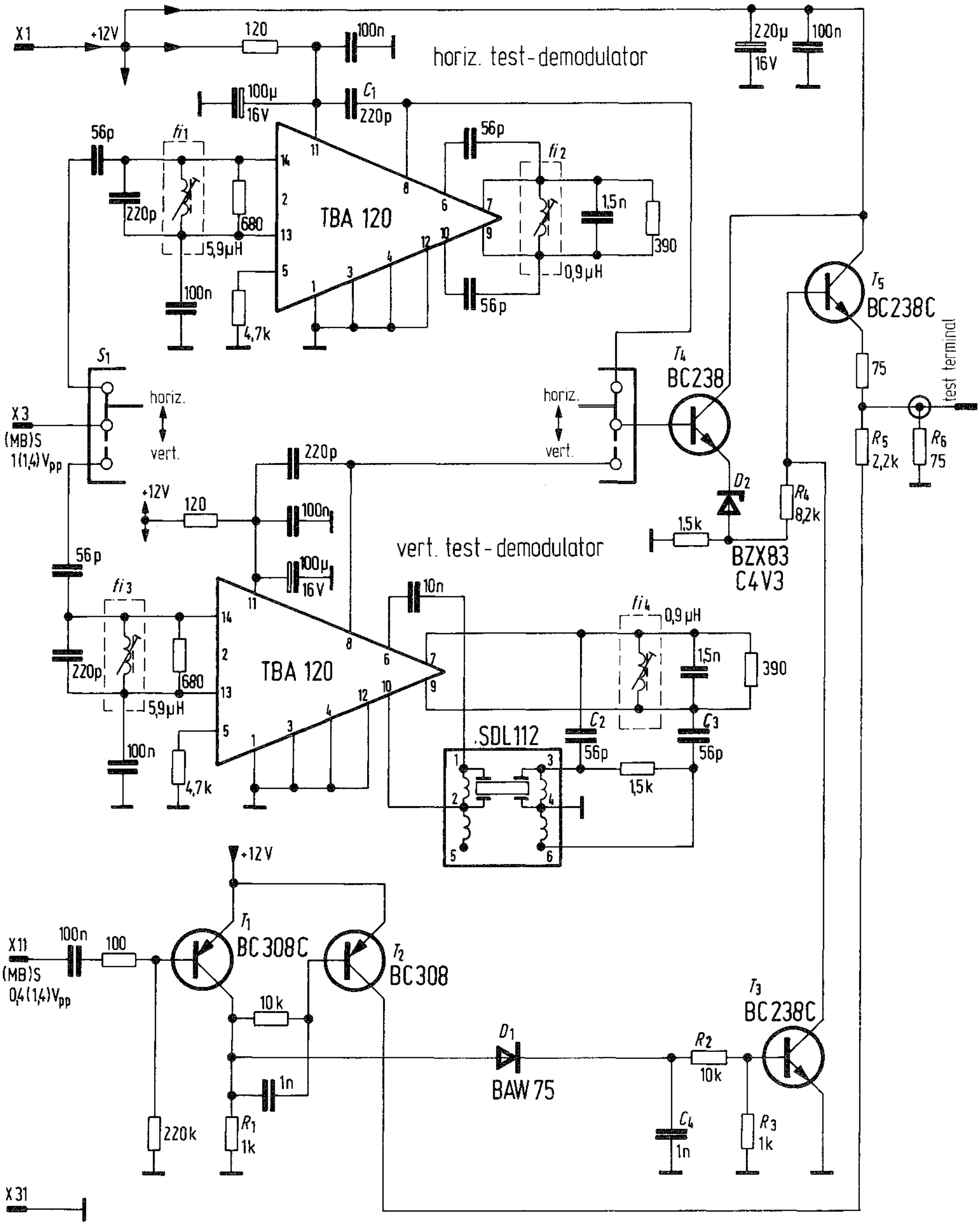


Fig. 2.6.3.

during the interval of a $\pm 1/4$ period. Therefore a balanced 90° phase shifter is additionally required to drive the demodulator circuit of the TBA 120 well. The phase shifting circuit consists of the components C_2, C_3 and fi_4. The delay line is asymmetrically driven by the limiting amplifier.
The sync and blanking pulses are isolated in stage T_1. The following stage T_3 forms the blanking pulses. The positive sync pulses available across the resistor R_1 rapidly charge the capacitor C_4 via the diode D_1, so that the transistor is immediately turned on. The discharging, i.e. the switching time of the transistor T_3 is determined by the time constant achieved by C_4, R_2 and R_3. It is about 11 µs. By the switch S_1 the signal required for the test is supplied to the emitter follower T_4. The z-diode D_2 compensates for the dc voltage offset. The blanking is achieved by short-circuiting the signal at R_4 during the blanking intervals. Therefore a blanked video signal is already applied to the base of the transistor T_5. The positive sync pulses available



X1
+12V
120
100n
horiz. test-demodulator
220µ
16V
100n
100µ
16V
C1
220p
56p
fi1
14
11
TBA 120
2
13
5
1
3
4
12
10
8
6
7
9
56p
fi2
1,5n
390
0,9µH
56p
220p
680
5,9µH
100n
4,7k
S1
horiz.
vert.
X3
(MB)S
1(1,4)Vpp
T5
BC238C
75
test terminal
R5
2,2k
R6
75
T4
BC238
D2
R4
8,2k
1,5k
BZX83
C4V3
220p
+12V
120
100n
100µ
16V
vert. test-demodulator
56p
fi3
10n
fi4
0,9µH
1,5n
390
220p
680
5,9µH
100n
4,7k
C2
56p
C3
56p
SDL112
1,5k
+12V
T1
BC308C
T2
BC308
X11
(MB)S
0,4(1,4)Vpp
100n
100
10k
1n
220k
R1
1k
D1
BAW75
R2
10k
T3
BC238C
C4
1n
R3
1k
X31

Fig. 2.6.4.

across R_1 are inverted by T_2 and via R_5 added to the video signal at the test-output. The built-in resistor R_6 is not necessary if a regular or a waveform-monitor is connected via a cable terminated by a resistance of 75 Ω.
The best way to adjust the test demodulator is to use the blanked colour subcarrier of a PAL-colour-carrier-generator (4.43 MHz). The output voltage generated during the colour subcarrier train, has to be the same as during the rest of the time for the active line period. The test frequency has been chosen for very special reasons. Cheap delay lines as they are employed in PAL-TV-receivers are proportioned to the same frequency. Therefore they can be utilized for this application as well. The described device is mainly suitable for the european standard of 625 lines. A modification can easily be realized in order to meet the american standard of 525 lines. In this case a frequency of 3.58 MHz has to be used, and a variety of cheap delay lines are also available for this frequency. As a result of an exact calculation the same test diapositive can be used by displaying a smaller section to achieve the lower test frequency. The tilt angle of 22° can be maintained.

List of Capacitors used in the Circuit 2.6.

Quantity	Components	Value	Rated Voltage (V)	Ordering codes
6	Styroflex capacitors	56 pF	63	B 31310–A 5560–H 000
2	Styroflex capacitors	220 pF	25	B 31310–A 3221–H 000
2	Styroflex capacitors	1 nF	25	B 31310–A 3102–H 000
2	Styroflex capacitors	1.5 nF	25	B 31310–A 3152–H 000
2	Ceramic disc-capacitors Type 2, SDPN	220 pF	500	B 37205–A 5221–S 001
1	Ceramic disc-capacitor Type 3, SDPN	10 nF	16	B 37305–A 1103–Z 001
6	Ceramic disc-capacitors Type 3, SDPN	100 nF	16	B 37302–A 1104–Z 001
2	Electrolytic capacitors Tantalum, sinter type (solid)	100 µF	16	B 41286–A 4107–T 000
1	Electrolytic capacitor	220 µF	16	B 41286–A 4227–T 000

3. Circuits for Colour TV-Receivers

3.1. Video-IF-Circuit using TBA 440 N/P and an AFC-Integrated Circuit TCA 890

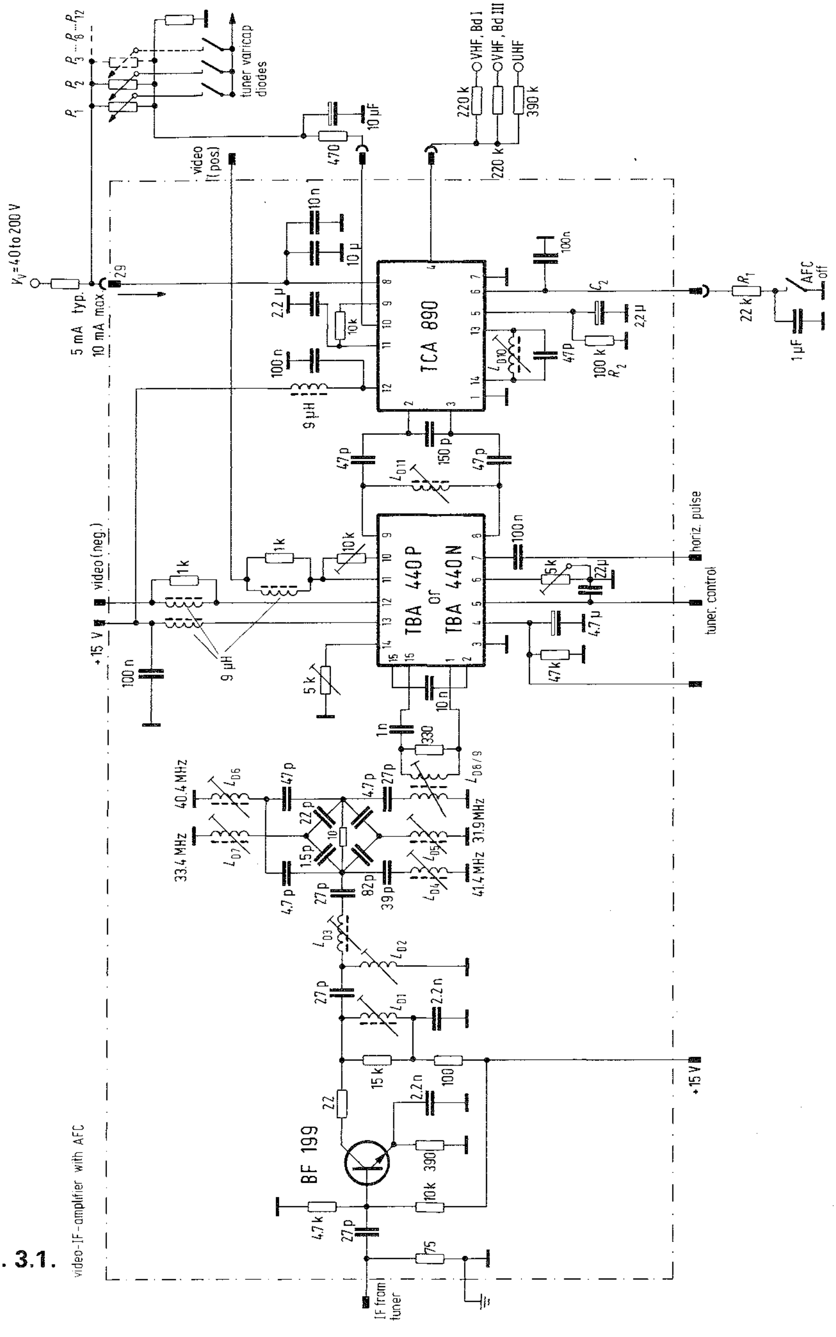


Fig. 3.1. video-IF-amplifier with AFC

Fig. 3.1. shows the total circuit consisting of four parts:

Buffer Amplifier using BF 199
Compact Filter
Video-IF-Amplifier with TBA 440 N/P
AFC-Circuit, type TCA 890.

By using the buffer amplifier, with BF 199, the achievement is as follows. Every tuner matches the video-IF-amplifier without adjustment. The input impedance is 75 Ω. The attained gain of 16.5 dB compensates not only for the loss of the compact filter, but an amplification is additionally disposable.

The compact filter incorporates seven circuits determining the transmission characteristic. Four of these are traps for the adjacent sound and picture carrier frequencies and for the attenuation of the accompanying sound. This compact filter is conventionally aligned by attenuating the demodulator circuit L_{D11}, by means of a 100 Ω-resistor. Firstly the traps have to be adjusted. Then the filter sides and its tilts are aligned by L_{D1}, L_{D3} and $L_{D8/9}$. By decreasing the inductance of L_{D2} (spreading out the coil), the left side of the filter can be correctively influenced, i.e. the sound-trap attenuation. The filter is symmetrically coupled to the IC TBA 440 N/P without any connection to earth.

The TBA 440 N or P includes a high-gain regulated video-IF-amplifier, a controlled demodulator, two low-resistance video outputs, as well as the complete key control and the delayed AGC for the tuner. Both white and black level are separately adjusted at the n as well as the p-type. The white levels of the video signal at the positive and negative output do not depend on battery level. An internal temperature stabilization guarantees an accurate operation in a temperature range of −25 to +60°C. The only difference between types TBA 440 N and TBA 440 P is the polarity of the control voltage for the tuner prestage:

TBA 440 N is suitable for prestages with npn-transistors,
TBA 440 P is favoured for those with pnp-transistors.

Both p and n-types are able to directly control pin-diode attenuators, i.e. the current available at pin 5 is sufficiently high. Therefore no additional transistors are required. The output signal of the internal limiting amplifier incorporated to the IC, is supplied via pins 8 and 9 to an auxiliary filter circuit tuned to 38.9 MHz.

Filtering chokes with an inductance of 9 µH are adapted to the power supply lines and the video outputs. To the chokes of the latter 1 kΩ-resistors are connected in parallel to improve the transient behaviour.

By means of the 5 kΩ-potentiometer connected to pin 14 the basic voltage, without any signal, i.e. the white level can be adjusted. The beginning of the keyed control and thus the black level, are determined by the 10 kΩ-potentiometer at pin 10. The delayed AGC for the tuner can be adjusted by the 5 kΩ-potentiometer connected to pin 6. For the internal keyed control a negative h-pulse with an amplitude of 2 to 5 V_{pp} has to be applied via the 100 nF-capacitor to pin 7. The RC-circuit consisting of a 47 kΩ-resistor and a 4.7 µF-capacitor, determines, at pin 4, the time constant of the keyed AGC. The 10 nF-capacitor connected to pins 2 and 15 achieves an internal dc feedback. It has to be soldered as close as possible to the terminals of the IC to avoid unwanted oscillations of the video IF amplifier.

The power supply voltage is 15 V. A pulse of 16.5 V is momentarily allowed. The output current at pins 11 and 12 should not exceed a value of 5 mA (to chassis) respectively minus 1 mA (to plus).

The input of the IC TCA 890 (pins 2 and 3) is connected to the auxiliary filter circuit L_{D11} via a capacitively balanced voltage divider. A constant level is provided for the discriminator by the internal limiting amplifier.
An additional LC-circuit also tuned to the carrier frequency of 38.9 MHz, is coupled to pins 13 and 14. Together with the integrated coupling capacitor it represents the phase shifting circuit for the coincidence demodulator, the dc voltage of which is amplified in a following amplifier. It controls not only the reference voltage source via pins 11 and 9, but also the voltage at one end of the tuning potentiometers via pin 10. Thus a constant degree of control is attained over the total tuning range.
The temperature-compensated reference voltage is produced in the TCA 890. Therefore external reference components, as previously used, are no longer necessary. It has an impedance of about 5 Ω at a current of 5 mA which is supplied via a series resistor from a voltage of 40 to 200 V (terminal 29) to pin 8 of the IC. If the current varies from 2 to 10 mA the tuning voltage only changes by 50 mV (maximum).
If pin 6 is connected via the 22 kΩ-resistor R_1 to the chassis, the AFT is turned off. The amplitude of the AFT-voltage superimposed to the reference voltage and the voltage at the tuning potentiometers end can be influenced by differently high series resistances connected to pin 4. The ratio of the currents at pin 10 and 11 is determined by the dimensioning of the coincidence circuit and it reaches already its maximum at a deviation of 50 to 100 kHz.
The filter coils L_{D1}, L_{D3} to L_{D11} are available from the company Toko Elektronik, Düsseldorf (ordering code: type 10 K). The capacitances are incorporated to the filter circuits. L_{D2} is an air-spaced coil consisting of 6 turns of enamelled copper wire 0.35 ∅; the inner diameter of the coil is 3 mm.

List of Capacitors used in the Circuit 3.1.

Quantity	Value	Ordering codes
1	27 pF	B 38116–J 5270–J
2	47 pF	B 38116–J 5470–J
1	150 pF	B 37215–B 5151–M 001
1	1 nF	B 37235–J 5102–M 001
2	2.2 nF	B 37238–J 5222–M 001
2	10 nF	B 37449–A 6103–S 1
5	100 nF	B 37449–A 6104–S 1
1	22 µF/25 V	B 41315–A 5226–Z
1	2.2 µF/25 V	B 41315–A 5225–Z
1	4.7 µF/16 V	B 41315–A 4475–Z
1	10 µF/40 V	B 41315–C 7106–Z
not on the pc board		
1	10 µF/25 V	B 41315–B 5106–Z
2	1 µF/40 V	B 41315–C 7105–Z

3.2. Sound-Section for Monochrome and Colour TV-Receivers with AF-Plug-Connection for VCR-Devices according to DIN 45482

The difference between the circuit shown in **Fig. 3.2.** to others previously described, is as follows: A connection for an AF-signal and for a switching voltage is provided. The terminal no. 2 for the AF-connection is responsible for both the AF-input and output. If a constant AF-voltage is required at this terminal for recording with a VCR-device, no switching voltage is to be applied to terminal 1. For the "play-back" from a VCR-device, however, a voltage of +12 V has to be supplied to terminal 1. For the FM-IF-amplifier with automatic volume control the IC TBA 120 U is used. The differences as opposed to the well known type TBA 120 S are as follows:

- Both additional AF signal-input and output before the electronic volume control.
- A constant AF output level if the power supply voltage changes between 10 and 18 V.
- Relatively insensitive against hum of the supply voltage.
- Very little residual IF voltages are superimposed to the AF output-signal; therefore no video-IF interference, created by the harmonics of the sound-IF are to be expected.
- A selection according to volume control characteristic is no longer required.

The 5.5 MHz sound-IF-signal is supplied via the terminal 10 to the ceramic filter type SFE 5.5 MA (Stettner Murata, Corp.), and then to pin 14 of the TBA 120 U. The FM signal is separated from any possible AM components by the extremely sensitive limiting amplifier and applied to the coincidence demodulator. The coincidence circuit consisting of L_1 and C_2 is connected to the IC via pins 7 and 9. The coupling capacitors being responsible for the phase shifting are already integrated into the TBA 120 U. The non-controlled AF signal is available at pin 12. It is supplied to the VCR-device via the emitter follower T_1, the resistor R_1, the capacitor C_2 and the terminal 2 of the pc board, when no switching voltage is applied to terminal 1. For the replay from a VCR-device a switching voltage is supplied to terminal 1 and thus the transistor 1 is turned off, since the level at its emitter is more positive than at its base. Then the AF signal is supplied from terminal 2 to the base of transistor T_2 via the capacitor C_2 and the two resistors R_1 and R_3. The amplified AF signal is picked up at the collector of T_2 and applied to pin 3 of the IC via the coupling capacitor C_3. During the playback operation the switching voltage at terminal 1 of the pc board, is applied via the diode D_1 and the resistor R_4 to pin 13 of the IC at the same time, and turns off the limiting amplifier of the sound-IF.
At pin 8 of TBA 120 U the AF signal, having already passed the automatic volume control circuit, is available. The control range covers 85 dB. This is shown in **Fig. 3.2.1.** as a function of the resistance R_x connected to the IC.
In conjunction with the internal impedance of the output stage and the capacitor C_4 connected from pin 8 to pin 11, a de-emphasis is achieved. The AF-signal is amplified by the transistors T_3 and T_4 and directly supplied to the base of transistor T_6 and via two diodes to the base of T_5. The two diodes and the resistors R_5 and R_6 compensate for the temperature effects of the output stages. By the feedback to the base of T_4 the operating point of the output stages is symmetrically adjusted.

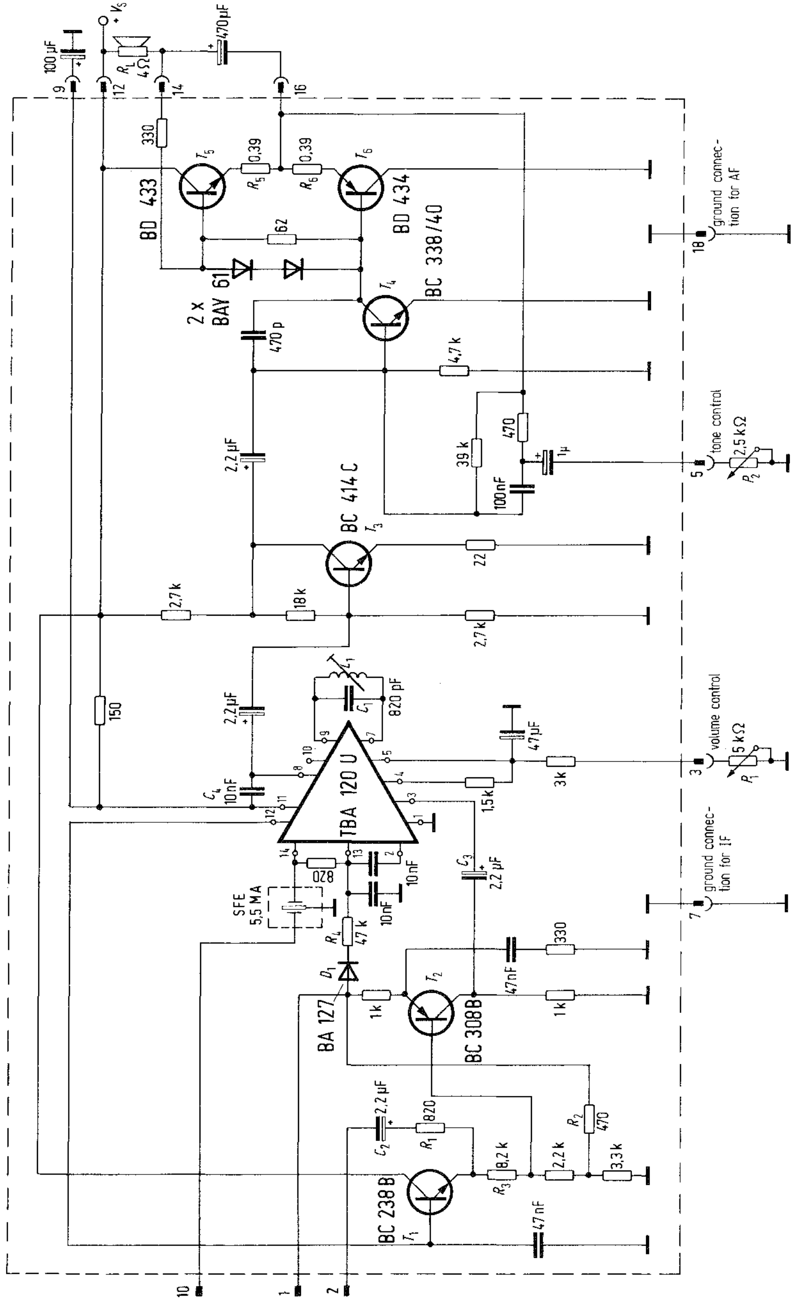

TBA 120 U
BD 433
BD 434
BC 338/40
2 x BAY 61
BC 414 C
BC 308B
BC 238 B
BA 127
SFE 5,5 MA
100 µF
470 µF
+ V_S
R_L 4 Ω
ground connection for AF
tone control
2,5 kΩ
volume control
5 kΩ
ground connection for IF
820 pF
2,2 µF
10 nF
47 µF
470 p
4,7 k
39 k
100 nF
2,7 k
18 k
150
330
0,39
62
22
3 k
1,5 k
47 k
1 k
47 nF
8,2 k
2,2 k
3,3 k
470
820

Fig. 3.2.

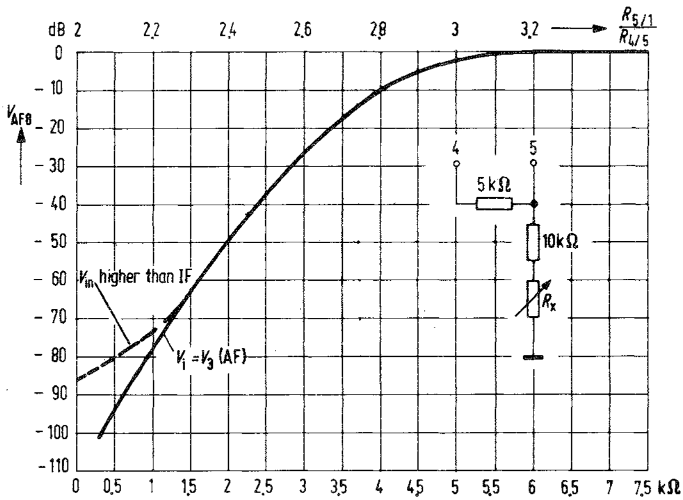


Fig. 3.2.1. with an electrolytic capacitor of 47µF connected from pin 11 to ground

Moreover, it depends on the frequency and can be varied by the tone control potentiometer P_2. **Fig. 3.2.2.** shows the frequency response of the output stages. The two different curves belong to the extreme positions of the tone control potentiometers, i.e. centre tap to the high end respectively to the grounded one.

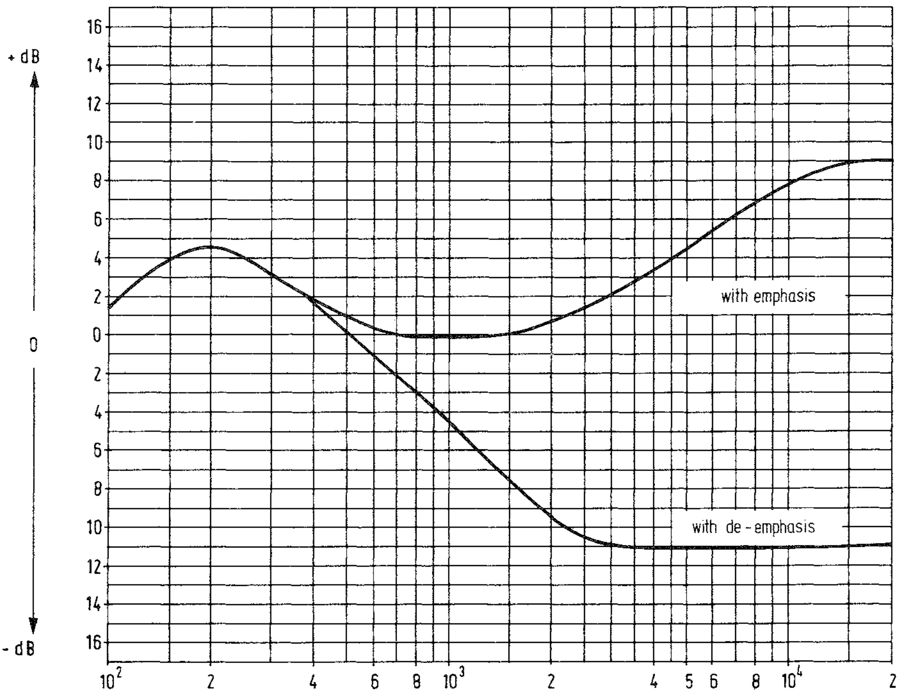


Fig. 3.2.2.

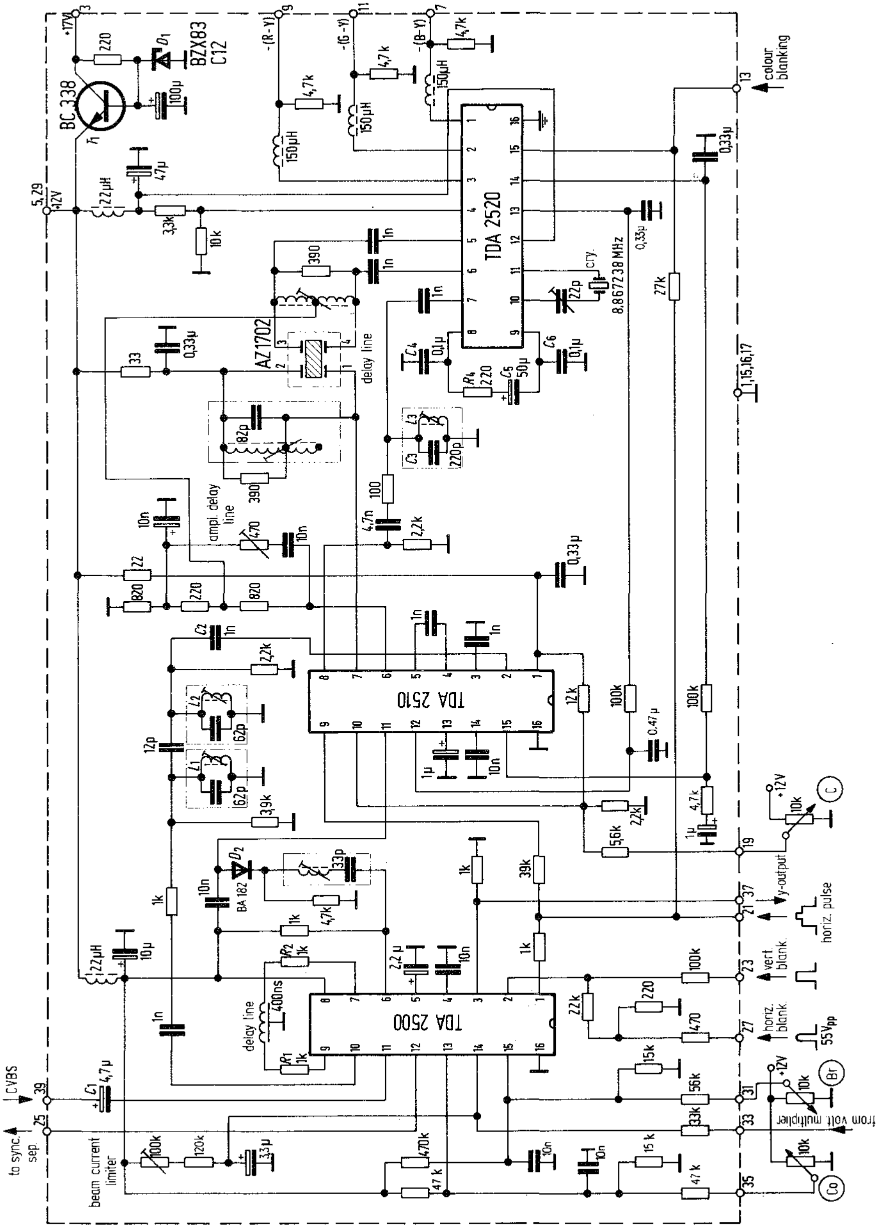
The bias of the transistor T_5 is obtained by a so-called bootstrap circuit, offering the advantage of a low residual voltage at transistor T_5 when driven to its maximum.

List of Capacitors used in the Circuit 3.2.

Quantity	Value	Ordering codes
1	470 p	B 37205–A 5471–M 001
1	820 p	B 37208–A 5821–M 001
2	47 nF	B 37449–A 6473–S 001
3	10 nF	B 37449–A 6103–S 1
1	100 nF	B 37449–A 6104–S 1
4	2.2 µF/25 V	B 41315–A 5225–Z
1	47 µF/25 V	B 41286–B 5476–T
1	1 µF/40 V	B 41315–C 7105–Z
not on the pc board		
1	100 µF/16 V	B 41286–A 4107–T
1	470 µF/25 V	B 41012–A 5477–T

3.3. Colour Decoder using TDA 2500, TDA 2510 and TDA 2520

Fig. 3.3. shows a complete colour decoding circuit incorporating the luminance pre-amplifier. It is equipped with the IC's TDA 2500, TDA 2510 and TDA 2520. The combination of the transistor T_1 and z-diode D_1 maintains a voltage of 12 V, required for the supply of the IC's.
The complete colour video signal is applied via terminal 39 and the coupling capacitor C_1 to pin 11 of the TDA 2500. The impedance matching for the short-time-delay-line between pin 7 and 9 is achieved by resistors R_1 and R_2. A trap tuned to the frequency of the colour subcarrier is connected to pin 6. It can be switched off by the diode D_2 when monochrome signals are received. The pre-amplified contrast-controlled blanked and clamped luminance signal is available at pin 3 of the IC and at pin 12 a composite video signal with negative amplitude of 3 V_{pp} is disposable to drive the amplitude separator. The colour signal is picked up at pin 10 and coupled to a two-circuit-filter using the coils L_1 and L_2. Since the contrast control for the luminance as well as the chrominance is obtained by the TDA 2500 it has to be blanked to its full capacity during the flyback time to achieve a constant "burst" amplitude. This is automatically attained by applying a positive pulse to pin 1. In parallel to the so-called electronic potentiometer an additional blanking influence is maintained to achieve a constant "burst" amplitude being independent of the contrast control. The connections to the potentiometers controlling the brightness (Br), the contrast (Co), the colour saturation (C) and the beam current can be gathered in details from Fig. 3.3. A control voltage derived from the cascade, and being proportional to beam current is applied to terminal 33. The chrominance signal filtered in the bandpass filter L_1/L_2 is supplied via the coupling capacitor C_2 to pin 2 of the TDA 2510 and separated from the burst. The latter, available at pin 8, is applied to the third IC, type TDA 2520, via the circuit L_3/C_3 being responsible for fine adjustment of the burst phase.



TDA 2520
TDA 2510
TDA 2500
BC 338
BZX83 C12
AZ1702
delay line
ampl. delay line
BA 182
8,867238 MHz
CVBS
to sync. sep.
beam current limiter
from volt multiplier
y-output
horiz. pulse
vert. blank.
horiz. blank.
55 Vpp
colour blanking
-(R-Y)
-(G-Y)
-(B-Y)
+17V
+12V
400ns

Fig. 3.3.

The controlled chrominance signal is supplied from pin 7 to the input, as well as from pin 6 directly to the output of the ultrasonic delay line (No. 2).
The adjustment of an equal amplitude of the delayed and non-delayed signal is realized by the potentiometer R_3. Both signals are added and subtracted simultaneously and supplied by two branches to pins 5 and 6 of TDA 2520 to become synchronously demodulated. This IC contains the crystal generator for the colour reference signals, the synchronous demodulators for the (B–Y) and (R–Y) chrominance signals, the PAL phase switch as well as a matrix, producing the (G–Y) signal. The crystal controlled oscillator operates at a frequency of 8.86 MHz which is twice that of the colour subcarrier. The oscillation is divided, by using two stages, to the desired frequency of 4.43 MHz. Triggering the dividers by both the leading and the trailing edge of the 8.86 MHz oscillation two reference signals with a phase shift of 90° are internally generated in the IC. Therefore the external components previously required for the phase shifting are no longer necessary. Besides that the adjustment of the 90° phase shift can be saved.
The 8.86 MHz oscillations are conventionally compared with the burst by the phase comparison and re-adjusted by means of a reactance circuit. The filtering components C_4, C_5, C_6 and R_4 are connected to the pins 8 and 9 of the IC.
The colour difference signals are available at pins 1, 2 and 3 and are supplied to the following matrixing circuit via terminals 7, 9 and 11.
At terminal 13 of the pc board a blanking signal is applied to eliminate the colour at certain picture spots. This is especially required for having a grey background when figures or letters are faded in.
The temperature compensation is sufficient for a dc coupled matrixing and colour output stage due to the one achieved by the three IC's in conjunction with the black-level clamping circuit of the TDA 2500. But great store has to be set so that a hum-free 12 V supply voltage is applied. It has not to contain any other radio interference signals.

List of Capacitors used in the Circuit 3.3.

12 pF	Ceramic capacitor	B 38116–J 5120–J
33 pF	Ceramic capacitor	B 38116–J 5330–J
220 pF	Ceramic capacitor	B 37205–A 5221–M 001
7 × 1 nF	MKM-capacitors	B 32540–A 3102–K
4.7 nF	MKM-capacitor	B 32540–A 3472–K
7 × 10 nF	MKM-capacitors	B 32540–A 3103–J
2 × 0.1 μ	MKM-capacitors	B 32540–A 1104–J
4 × 0.33 μ	MKM-capacitors	B 32540–A 1334–J
0.47 μ	MKM-capacitor	B 32540–A 1474–J
2 × 1 μF	Electrolytic capacitors	B 41315–C 7105–Z
2.2 μF	Electrolytic capacitor	B 41315–A 5225–Z
2 × 4.7 μF	Electrolytic capacitors	B 41315–A 4475–Z
10 μF	Electrolytic capacitor	B 41315–B 5106–Z
2 × 47 μF	Electrolytic capacitors	B 41286–B 5476–Z
100 μF	Electrolytic capacitor	B 41286–A 4107–T

3.4. RGB-Matrix and Output-Stages for PI-Tubes

Because of new picture-tube concepts considering devices with no separate terminal for the control grid, a clear tendency towards RGB-controlling exists. To compensate for the tolerances of the characteristics of the different beam positioning systems the black levels have to be separately adjustable at the individual picture tube cathodes. This offers the advantage in that no special temperature effects compensation of the colour decoder outputs have to be taken into consideration. Besides that the requirements for the hum-free supply voltages are essentially lower.

The inputs of the RGB-module are dimensioned for the colour difference output signals of the TAA 630 S or the TDA 2520 and for the luminance output signal of the TBA 560. The colour difference signals are supplied to the base of the preamplifier transistors T_2, T_3 and T_4, whereas the luminance signal is applied to its emitter via the emitter follower T_1. By subtraction it is achieved:

$$(R - Y) + Y = R$$
$$(G - Y) + Y = G$$
$$(B - Y) + Y = B$$

The three different primary signals drop across the collector resistors of the transistors T_2, T_3 and T_4, consisting of a fixed resistor (R_1, R_2, R_3) and an adjustable one. To reduce the influence of the undesired Miller-capacitance between base and collector a 10 pF-capacitor each (C_1, C_2, C_3) is connected to base and chassis. Thus the frequency response is also improved. These small capacitances will not influence the colour difference signals. The primary signals are coupled to the output stages by the electrolytic capacitors C_4, C_5 and C_6. The emitter-followers T_5, T_6 and T_7 drive the output transistors T_8, T_9 and T_{10} to achieve a sufficiently high input impedance for the clamping control circuit and to compensate for the current-gain tolerances of the output transistors, type BF 458. The voltage gain of the output stages is determined by the ratio of emitter-to-collector resistance.

The resistors R_{13}, R_{14} and R_{15} are connected to the collector of the output transistors and the cathodes of the picture tube to protect the transistors against arc-overs. Each emitter resistor has a 1 nF-capacitor in parallel (C_4, C_5, C_6) to equalize the frequency response. An accentuating choke of 55 µH and a 3.9 kΩ-resistor in parallel, are inserted into each cathode supply and are mounted on the pc-board of the picture tube.

To set the operating point of the output stages a positive current is applied via the resistors R_7, R_8 and R_9 to the base of the output Darlington-transistors. The black-level control consists of an unbalanced circuit using a clamping diode. During the flyback time a defined level of the picture signal is present. Therefore the capacitors C_7, C_8 and C_9 are charged by the clamp pulses. When the clamp pulse has passed, the one electrode of the capacitor has a zero-level and the other a negative voltage which is supplied as a controlled variable to the base of the output Darlington-transistors via the resistors R_{10}, R_{11} and R_{12} as well as the RC-circuits. This current following in the opposite direction and depending on the difference to the correct black-level partially eliminates the positive currents through the resistors R_7, R_8 and R_9. The black-level of each primary signal can be adjusted by the potentiometers R_{16}, R_{17} and R_{18}, which influence the amplitude of the blanking pulses.

It is convenient to connect 10 kΩ-resistors in parallel to these potentiometers so that blanking pulses are always present if the potentiometers should fail. Thus an overload of the output stages is avoided.

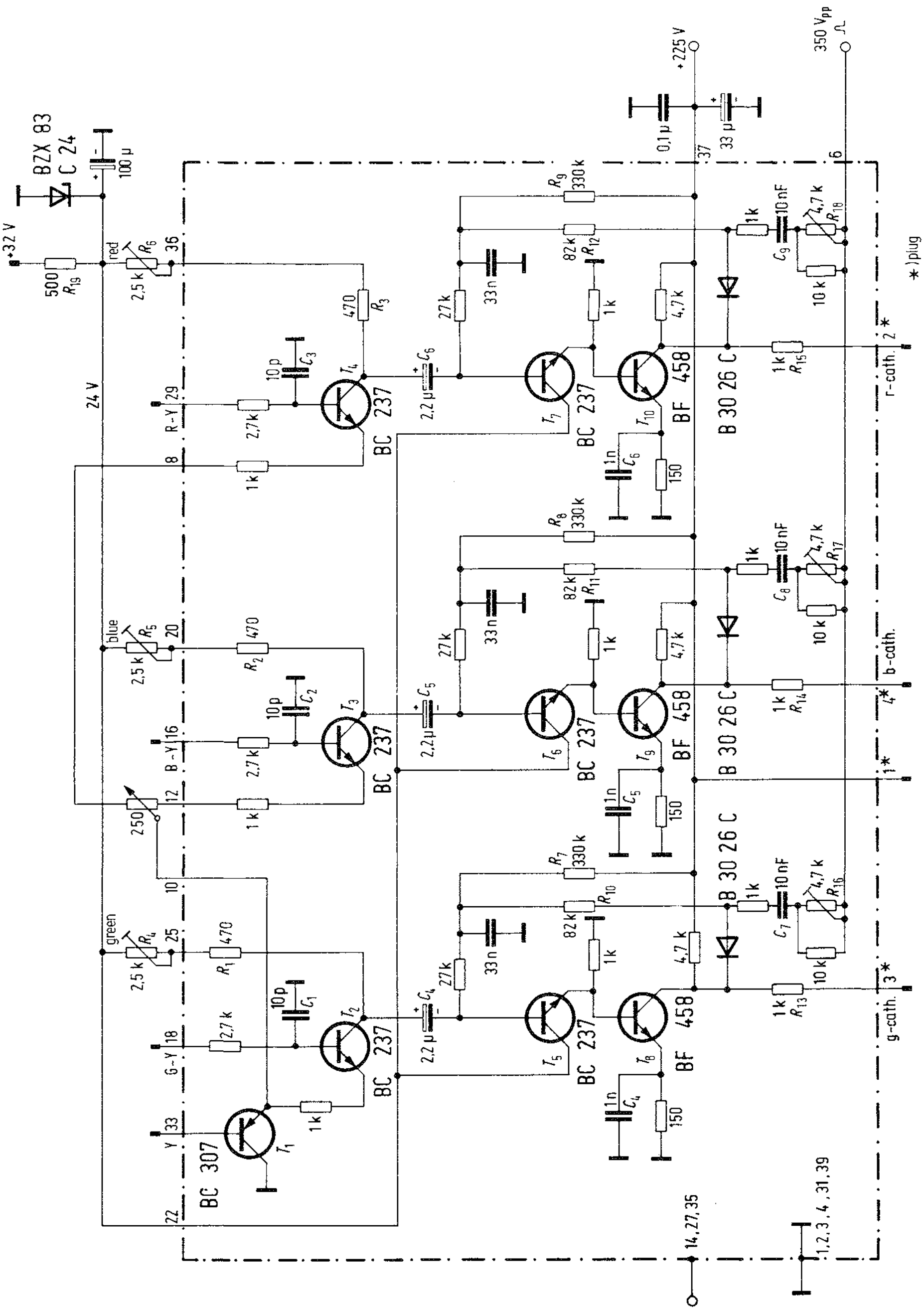

BZX 83
C 24
100 µ
+32 V
500
24 V
red
blue
green
R-Y
B-Y
G-Y
Y
BC 307
BC 237
BF 458
B 30 26 C
+225 V
0,1 µ
33 µ
350 Vpp
r-cath.
b-cath.
g-cath.
*) plug
14, 27, 35
1, 2, 3, 4, 31, 39

Fig. 3.4.

Required input levels:

Brightness signal:	V_E	$= 2\ V_{pp}$
Colour difference signal:	$-V_{(R-Y)}$	$= 4\ V_{pp}$
Colour difference signal:	$-V_{(G-Y)}$	$= 2.4\ V_{pp}$
Colour difference signal:	$-V_{(B-Y)}$	$= 5\ V_{pp}$
Output signal:	$V_{R,G,B}$	$= 100\ V_{pp}$

Supply voltage:

For prestage:	V_s	= 24 V at 30 mA (or 32 V with a series resistor R_{19})
For output stage:	V_s	= 225 V

List of Capacitors used in the Circuit 3.4.

3	MKM-stacked film capacitors	0.001 μF	B 32540–A 3102–K
3	MKM-stacked film capacitors	0.01 μF	B 32540–A 3103–J
3	MKM-stacked film capacitors	0.033 μF	B 32540–A 3333–J
1	MKM-stacked film capacitor	0.1 μF	B 32540–A 3104–J
3	Ceramic capacitors	10 pF	B 38112–J 5100–C
3	Electrolytic capacitors	2.2 μ/40 V	B 41313–A 8225–Z
1	Electrolytic capacitor	47 μ/250 V	B 43052–C 2476–T
1	Electrolytic capacitor	100 μ/25 V	B 41283–B 5107–T

3.5. Horizontal-Deflection Circuit for Different Colour Picture Tubes using BU 208

The standard circuit for the horizontal deflection using BU 208 can mostly be modified to the different types of picture tube by changing some of the resistors, capacitors or wire-wound components. Generally the pincushion correction is different in each application case. Therefore the circuit achieving optimal correction for a specific picture tube type is described on the following pages. **Fig. 3.5.** shows the basic circuit and table 1 contains the components that have to be changed, the circuit for driving the horizontal deflection coils and the corresponding pincushion correction circuits.

The BU 208 is operated in a common-base circuit. As it is driven at its emitter by T_2, an accurate switching-through and turning off is achieved.

The positive driving pulse from the TBA 920 triggers the transistor T_3 previously turned off. Up to that time the transistor T_2 was conductive, since it was positively biased by the 15 Ω-resistor connected to the auxiliary power supply of +5 V. Now it is switched off. The emitter current of T_1 charges the 0.68 μF-capacitor to its maximum. The base current of the BU 208 flows via the 2.7 Ω-resistor, the 25 μF-capacitor and the 1 Ω-resistor to chassis. This takes about 3 μ sec and afterwards the BU 208 is switched off within a time of 0.7 μs. By the 0.68 μF-capacitor connected to chassis and emitter an optimum delay time is obtained.

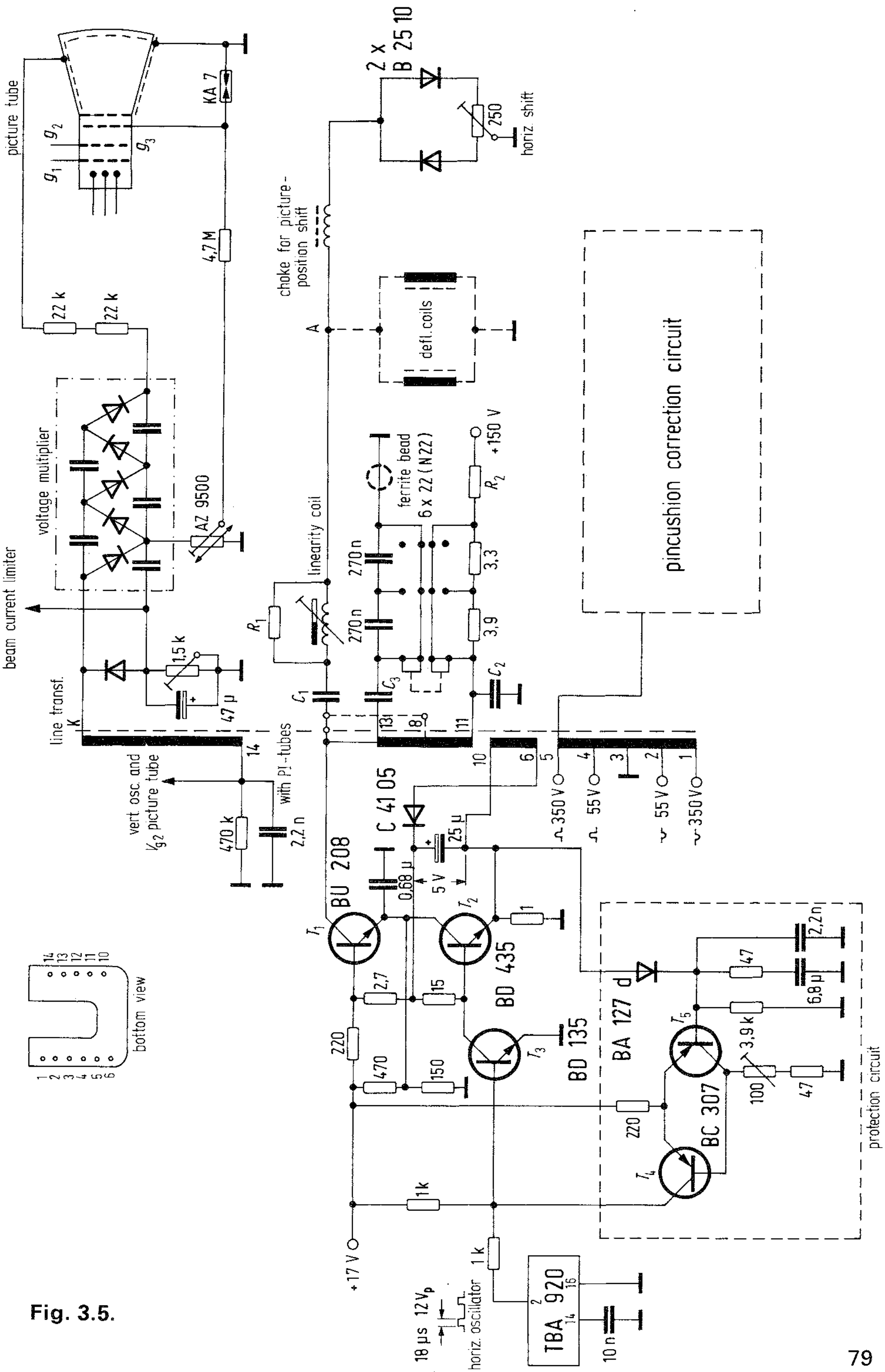

picture tube
KA 7
4,7 M
22 k
22 k
voltage multiplier
AZ 9500
beam current limiter
line transf.
K
1,5 k
47 µ
vert. osc. and
V_{g2} picture tube
470 k
2,2 n
with PL-tubes
choke for picture-position shift
A
defl. coils
2 x B 25 10
250
horiz. shift
pincushion correction circuit
linearity coil
R_1
270 n
270 n
ferrite bead
6 x 22 (N22)
+150 V
R_2
3,3
3,9
C_1
C_2
C_3
350 V
55 V
55 V
350 V
C 4105
25 µ
BU 208
0,68 µ
5 V
T_1
T_2
BD 435
2,7
15
220
470
150
T_3
BD 135
BA 127 d
47
2,2 n
6,8 µ
T_5
3,9 k
100
47
BC 307
220
T_4
protection circuit
1 k
1 k
+17 V
TBA 920
10 n
horiz. oscillator
18 µs
12 V_p
bottom view

Fig. 3.5.

When power is supplied to the circuit, a current flows from the +17 V-line via the 220 Ω-resistor to the base of the BU 208. During operation the transistor is biased via the 2.7 Ω-resistor by the auxiliary supply voltage of +5 V. The epibase transistor BD 435 encapsulated in a plastic package (SOT-32) is especially suited as emitter-load transistor, since the saturation voltage is only 1 V at 4 A. The picture width can be varied by means of a plug-in switch. In this case the high tension is practically constant.

Protection Circuit

The protection circuit consisting of the transistors T_4 and T_5 prevents damage of the line output transistors when short-circuits or picture flash-overs occur. If the voltage dropping across the 1 Ω-emitter-resistor being proportional to the collector current of the BU 208 exceeds the "protection level", adjusted through the 100 Ω-potentiometer, the transistor T_5 becomes non-conductive. Now the transistor T_4 is conductive and turns the driver transistor T_3 on, i.e. the transistor T_2 is no longer driven and T_1 is turned off. The resetting automatically occurs after a time of several milliseconds, i.e. when the 6.8 µF-capacitor is discharged via the 3.9 kΩ-resistor. An RC-circuit with two time constants (2.2 nF/47 Ω in parallel to 6.8 µF) is connected to the base of T_5 to protect the output-stage transistors against surge currents due to short-circuit.

3.5.1. 110°-Standard-Neck Picture Tubes with Delta Configuration and 110°-Uniline Picture Tubes

The components not indicated in the basic circuit can be found in Table 1. **Fig. 3.5.1.** shows the circuit for the pincushion correction and for driving the deflection coils.

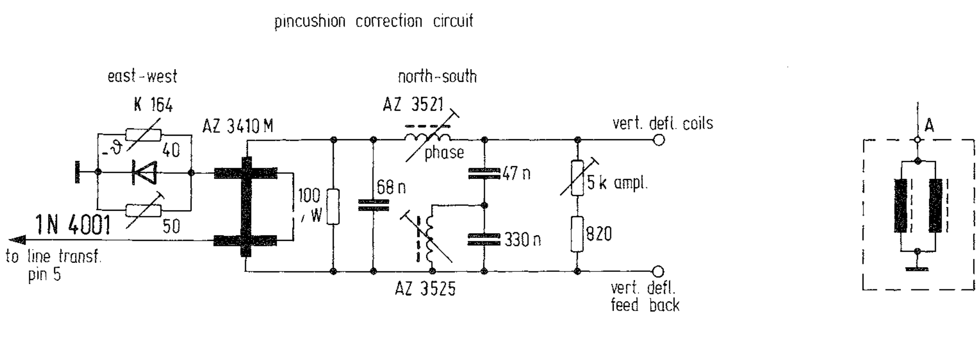


Fig. 3.5.1.

For the total north-south and east-west correction only a 30 mm-transductor type AZ 3410 M is required. Both correction circuits can be adjusted independently. A circuit, filtering harmonics and consisting of the coil AZ 3525 as well as of two capacitors (330 nF/47 nF), is added to compensate the so-called "moustache" distortions, i.e. the S-shaped ones. The phase is adjusted through the AZ 3521.

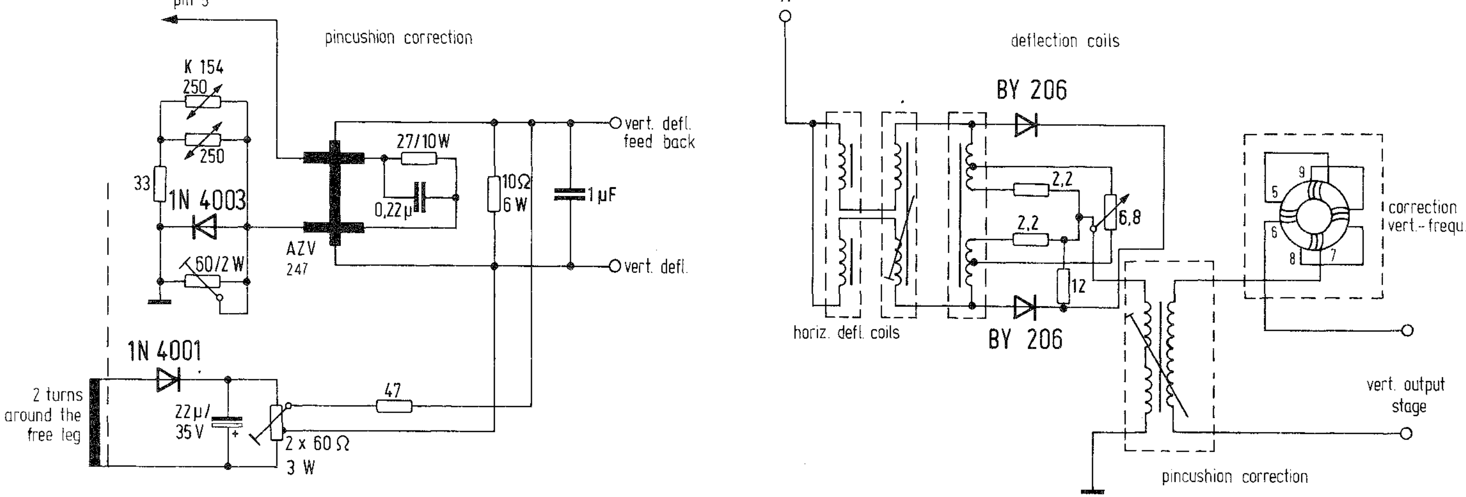


Fig. 3.5.2a.

Table 1

Solution for Picture Tube Type	Cascade Type	Line Transformer Type	Linearity Coil Type	Picture Position Choke	C_1	C_2	C_3	R_1	R_2	Pincushion correction circuit	Circuit for driving the horizontal-deflection Coil
110° Standard neck delta-configuration	TVK 52 S	AZ 3102	AZ 2302	AZ 3520	0.68 µ	1 µ	12 n	1.5 k	22	3.5.1.	3.5.1.
110° Uniline tubes	TVK 52 S	AZ 3102	AZ 2302	AZ 3520	0.6 µ	1 µ	12 n	1.5 k	22	3.5.1.	3.5.1.
20 AX-system	TVK 52 S	AZ 3102	AZ 2302	AZ 3520	0.47 µ	3.3 µ	12 n	1.5 k	33	3.5.2.	3.5.2.
RIS-Inline tubes 18″ and 22″	TVK 52 S	AZ 3102	AZ 2302	AZ 3520	0.39 µ	1 µ	12 n	1.5 k	30	3.5.3.	3.5.2.
PI-Tubes 16″, 20″ and 27″	TVK 76	AZV 320	AZV 331	AZV 335	2.7 µ	3.3 µ	14 n	120	33	3.5.4.	3.5.4.

The combination-transductor is premagnetized through the diode 1 N 4001. The east-west correction can be adjusted by the 50-Ω-resistor connected in parallel to the diode and the NTC-resistor, type K 164, which is responsible for compensation of the temperature effects. If required, the quality of the "inner-pincushion"-correction can be improved by the rotatable permanent-magnet. Although the correction circuit inductively loads the flyback transformer, neither a disadvantageous high-tension modulation nor a pincushion dependency on the beam current is evident. This fact is especially achieved by the compensating effect of the 22-Ω-resistor R_2 inserted in the power supply line. The adjustment of the pincushion correction is to be made in the following sequence:

1. all adjustors to mid-position, turn out the core of the AZ 3525 as far as possible,
2. adjust the east-west correction through the 50-Ω-potentiometer and correct it by the permanent-magnet,
3. adjust north-south phase through AZ 3521,
4. adjust north-south amplitude by the 5-kΩ-potentiometer,
5. optimize the pincushion correction by means of the harmonic filter circuit (AZ 3525); this is not necessary with RIS-tubes.

3.5.2. 20 AX-Systems

No additional correction circuit for convergence errors is required for the 20 AX-system. To compensate the production tolerances of the picture tube and the deflection unit, a four-pole toroid coil is covering the core of the deflection unit. Thus a 45/45 four-pole field with diagonal axis is generated. By changing the current distribution in the two parts of the horizontal and the vertical deflection coils a 0/90 four-pole field is achievable with horizontal as well as vertical frequency.

Fig. 3.5.2a. shows a pincushion correction circuit using a transductor. The corrections achieved are sufficient for standard operations. A correction of the keystone distortions can be obtained by a dc magnetization. To produce this the line voltage is rectified by a diode 1 N 4004 connected to two windings around the free leg of the flyback transformer.

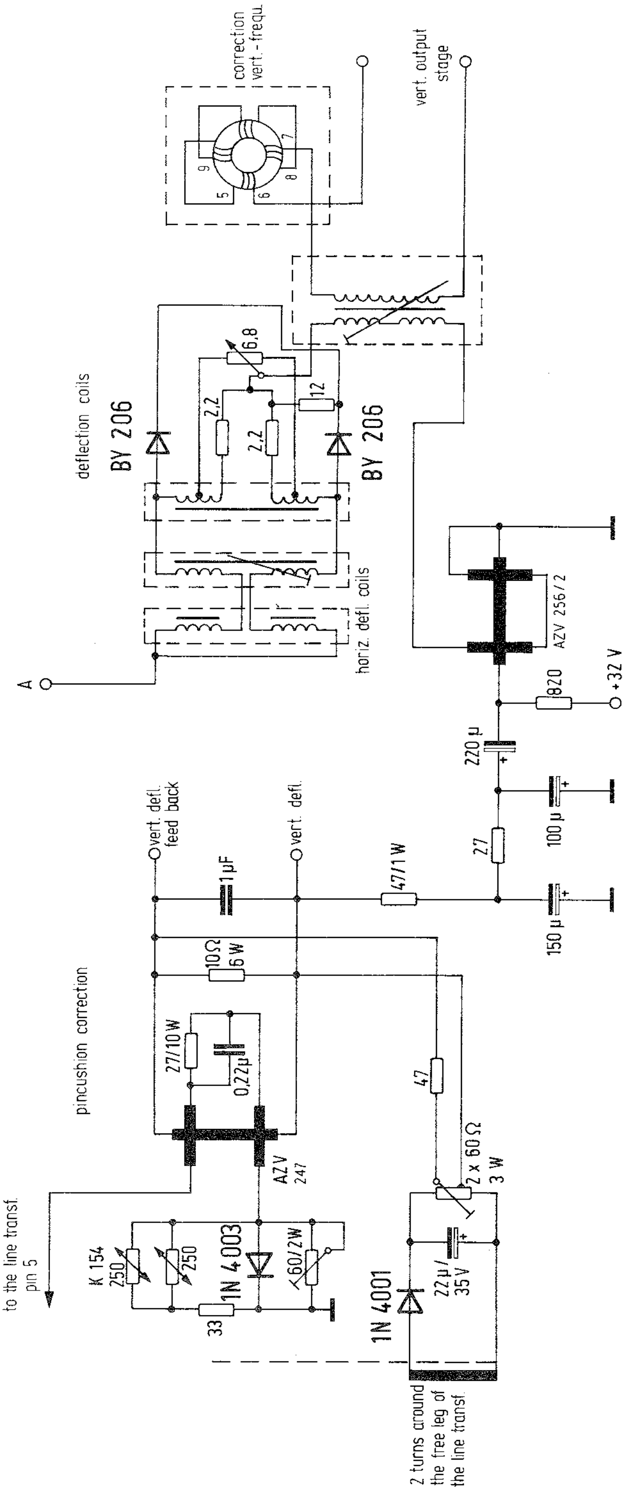
For top-quality requirements the correction circuit shown in **Fig. 3.5.2b.** is recommended. It consists of a parallel and a series-transductor. The dc magnetization of the latter is maintained via the 820 Ω-resistor from the power supply voltage of 32 V.

3.5.3. RIS-Inline Tubes 18" and 22"

The deflection coils are driven by differential coils as shown in the circuit of **Fig. 3.5.3.** Thus the red-green centre-line convergence of the left and right picture part can be symmetrized. The pincushion correction alignment is attained by three adjustments only.

3.5.4. PI-Tubes 16", 20" and 26"

Because of the low impedance of the toroid horizontal deflection coils the deflection unit is not directly connected to the collector of the transistor BU 208 but via a tap of the flyback transformer (terminal 8). For the transformer a U 59-core is used in order to avoid operation in the saturation area of the material when it is heavily loaded. For this particular picture tube a pincushion correction using a parallel transducer is completely sufficient (see **Fig. 3.5.4.**). The components have to be proportioned in such a way that the low impedance of the deflection unit is met.



correction vert.-frequ.
vert. output stage
deflection coils
BY 206
BY 206
6,8
12
2,2
2,2
horiz. defl. coils
A
AZV 256/2
+32 V
820
220 µ
100 µ
150 µ
27
47/1 W
vert. defl. feed back
vert. defl.
1 µF
10Ω 6 W
pincushion correction
27/10 W
0,22µ
47
AZV 247
2 x 60 Ω 3 W
to the line transf. pin 5
K 154 250
250
1N 4003
60/2W
33
1N 4001
22 µ/ 35 V
2 turns around the free leg of the line transf.

Fig. 3.5.2b.

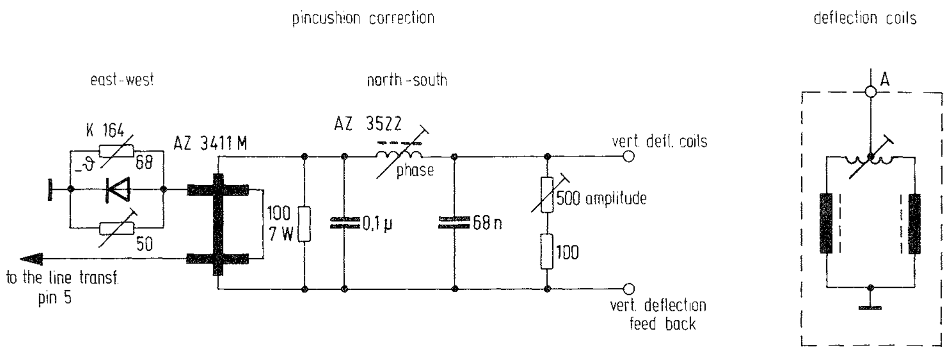


Fig. 3.5.3.

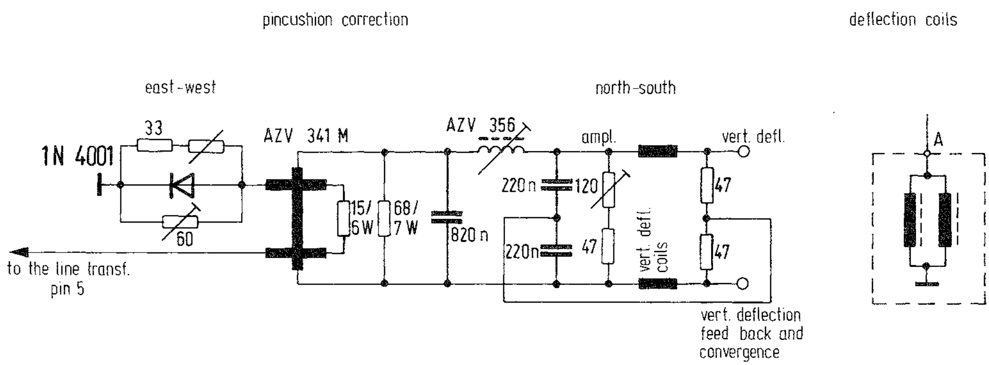


Fig. 3.5.4.

List of Capacitors used in the Circuit 3.5.

1	MKM-stacked film capacitor	2.2 nF	B 32540–A 3222–K
1	MKM-stacked film capacitor	0.68 µF	B 32540–A 1684–J
1	Electrolytic capacitor	22 µ/25 V	B 41315–A 5226–Z
1	MKL-capacitor	6.8 µ/25 V	B 42110–D 3685–M

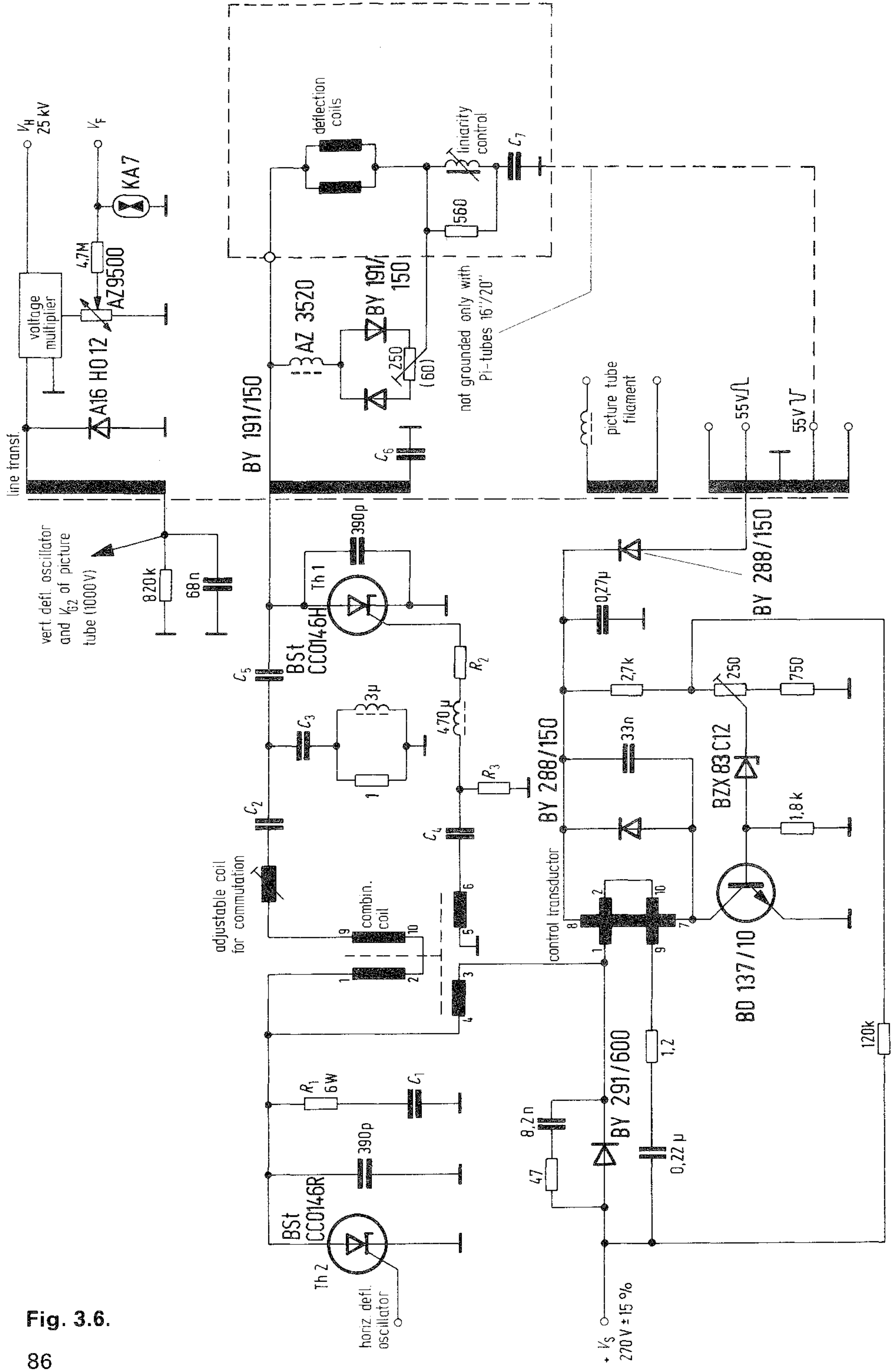
S-capacitor: see section "horiz. deflection"

3.5.1.	3.5.1.	3.5.2.	3.5.3.	3.5.4.
110°-Standard neck tube, delta configuration	110°-Uniline tube	20 AX-system	RIS-Inline 18″ and 22″	PI-tubes 16″, 20″ and 27″
0.68 μF = 3 × 0.15 μF B 32541–A 3154–J and 1 × 0.22 μF B 32541–A 3224–J	0.6 μF = 4 × 0.15 μF B 32541–A 3154–J	0.47 μF = 4 × 0.12 μF B 32541–A 3124–J	0.39 μF = 4 × 0.1 μF B 32541–A 3104–J	2.7 μF = 4 × 0.68 μF B 32541–A 1684–J
1 × 68 nF B 32541–A 3683–J 1 × 47 nF B 32541–A 3473–J 1 × 0.33 μF B 32541–A 1334–J	1 × 68 nF B 32541–A 3683–J 1 × 47 nF B 32541–A 3473–J 1 × 0.33 μF B 32541–A 1334–J	1 × 0.22 nF B 32541–A 3224–J 1 × 1 μF B 32541–A 1105–J 1 × Electrolytic capacitor 22 μF/40 V B 41283–B 7226–T	1 × 0.1 μF B 32541–A 3104–MJ 1 × 68 nF B 32541–A 3683–J	1 × 0.22 μF B 32541–A 3224–J 2 × 0.22 μF B 32541–A 3224–J 1 × 0.33 μF B 32541–A 1334–J 1 × 0.47 μF B 32541–A 1474–J
		at series transductors in addition 2 × 100 μF/40 V B 41283–B 7107–T 1 × 220 μF/40 V B 41283–A 7227–T		

3.6. Thyristor Horizontal-Deflection Circuits for Colour TV-Sets

Thyristor circuits for the horizontal deflection have been used for several years now. The fundamental circuit shown in **Fig. 3.6.** is applicable for the following types of picture tube:

110°-standard neck picture tube, delta configuration
110°-thin neck picture tube, delta configuration
PI-Tubes, 16″, 20″ and 27″
RIS-inline tubes 18″ and 12″
20 AX-system.



deflection coils
linearity control
C_7
560
V_H 25 kV
V_F
KA7
4,7M
AZ9500
voltage multiplier
A16 HO12
line transf.
BY 191/150
AZ 3520
BY 191/150
250 (60)
not grounded only with Pi-tubes 16"/20"
picture tube filament
55V
55V
C_6
vert. defl. oscillator and V_{G2} of picture tube (1000V)
820k
68n
390p
Th 1
BSt CC0146H
C_5
R_2
470μ
3μ
C_3
1
R_3
C_4
C_2
adjustable coil for commutation
combin. coil
control transductor
BY 288/150
0,27μ
27k
250
750
33n
BY 288/150
BZX 83 C12
1,8k
BD 137/10
120k
BY 291/600
1,2
8,2n
47
0,22 μ
R_1 6W
C_1
390p
BSt CC0146R
Th 2
horiz. defl. oscillator
$+V_S$ 270 V ± 15 %

Fig. 3.6.

The deflection is divided into three sections:

forward stroke,
interruption of the forward stroke,
flyback.

According to these sections three resonant-circuits are successively driven.

1. The forward-stroke circuit.
   It consists of the switched off thyristor Th_1, the series-connected linearity potentiometer, the S-capacitor and the horizontal deflection coils. The inductance of the latter can be matched to the one of the driving circuit by means of the flyback transformer in parallel to the deflection coils. The shape of the deflection current is determined during the forward stroke by the linearity potentiometer and the S-capacitor.
2. The flyback thyristor Th_2 is triggered by a pulse of the horizontal-oscillator just prior to when the forward stroke is finished. This starts the flyback. A commutating current flowing oppositely to the forward stroke one, turns off the forward stroke thyristor when the zero axis is crossed. The resonant-circuit being responsible for this triggering process consists of the switched-off flyback thyristor Th_2 the commutating coil, incorporated to the combination coil, the commutation adjustment coil, the capacitor-T-circuit and the turned-off forward-stroke thyristor Th_1.
3. Flyback.
   After the forward-stroke thyristor had been switched off during the zero-axis crossing the following flyback resonant-circuit is active: flyback thyristor Th_2, the commutating coil being a part of the combination coil, the commutating adjustment coil, the capacitor-T-circuit, the inductor resulting from deflection coil and flyback transformer, the linearity potentiometer and the S-capacitor.

The losses generated during forward and backward stroke have to be compensated to avoid a decrease of the deflection voltage amplitude. In order to achieve this, energy is transfered from the power supply V_S via the input-choke, included in the combination coil. For compensating changes of the main voltage and the load excessive energy is given back in measured quantities through the control transductor. During the commutating interval, i.e. when the flyback thyristor is switched off, a sawtooth current flows through the input choke which then stores energy ($0.5 \times L \times I^2$). After turning on the flyback thyristor (beginning of the forward stroke) the stored energy is transfered to the commutating capacitors. The amount of energy supplied to the deflection coils during the flyback depends on how highly the capacitor-T-circuit is charged. The energy transport from the capacitor-T-circuit to the charging capacitor of the power supply is determined by the control transductor which controls by that the resulting energy supply. The amplitude of the pulse produced through the flyback transformer is the control information for this transductor.
The following table indicates the components which have to be used in the fundamental circuit (**Fig. 3.6.**) to meet the requirements of the different picture tubes. The following pincushion correction circuits are accordingly recommended.

3.6.1. For 110°-Standard Neck Picture Tubes with Delta Config.

The pincushion correction is completely maintained by a transductor which is passively driven (**Fig. 3.6.1.**). The alignment is achieved by the following control possibilities: east-west amplitude, dc magnetization of the transductor, north-south amplitude, north-south phase.

Picture Tube Type	Line Transf. Type	Lin. Control Type	Combi. Coil Type	Com. Adjustm. Coil Type	Raster Parallel-Transductor Type	*N/S* Phase Coil Type	Series Trans-ductor Type	Control Trans-ductor Type	Cas-cade Type	C_1	C_2	C_3	C_4	C_5	C_6	C_7
110° standard neck picture tube, delta configuration	AZ 2110	AZ 2302	AZ 2468	AZ 2531	AZ 2412 MZ	AZ 3521	—	AZ 2422	TVK 52 S	5 n	0.18 µ	68 n	0.22 µ	0.18 µ	12 µF	0.68 µF
110° thin neck picture tube, delta configuration	AZ 2107	AZ 2300 A	AZ 2644	AZ 2531	AZ 2402 MZ	AZ 2521	AZV 2414	AZ 2422	TVK 52 S	5 n	0.15 µ	56 n	0.22 µ	0.15 µ	0.68 µ	1.8 µF
PI-tubes 16″/20″	AZ 2106	AZ 2303 A	AZ 2646 A	AZ 2531	AZ 2413 A	AZ 2521	—	AZ 2422	TVK 52 S	5 n	0.1 µ	33/ 40 n	0.22 µ	82 n	0.68 µ	1.5 µF
PI-tubes 27″	AZ 2107	AZ 2300 A	AZ 2644	AZ 2531	AZ 276 MZ	AZV 254	AZV 2414	AZ 2422	TVK 52 S	5 n	0.12 µ	47 n	0.22 µ	0.15 µ	0.68 µ	1.8 µF
RIS-Inline tube 18″ to 22″ incl.	AZV 217	AZ 2300 A	AZ 2644	AZ 2531	AZV 249 MZ/2	AZ 3522	—	AZ 2422	TVK 52 S	2.2 n	0.15 µ	56 n	0.22 µ	0.12 µ	0.68 µ	1.8 µF
20 AX-system	AZV 218	AZ 2302	AZV 272/2	AZ 2531	AZV 255 MZ	AZ 2502	AZV 256/2	AZ 2422	TVK 76	4.7 n	0.1 µ	51 n	0.22 µ	0.22 µ	12 µ	0.39 µF

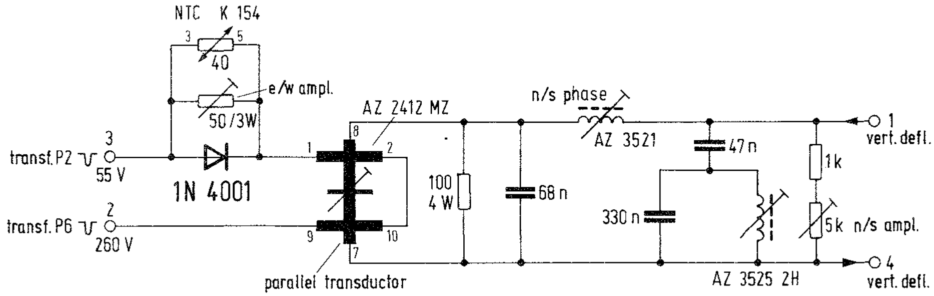


Fig. 3.6.1.

3.6.2. 100°-Thin Neck Picture Tubes with Delta Configurations

The circuit for these types of picture tube is shown in **Fig. 3.6.2.** For higher requirements a series transductor can be additionally used for compensating the east-west inner pincushion.

3.6.3. For PI-Tubes 16"/20"

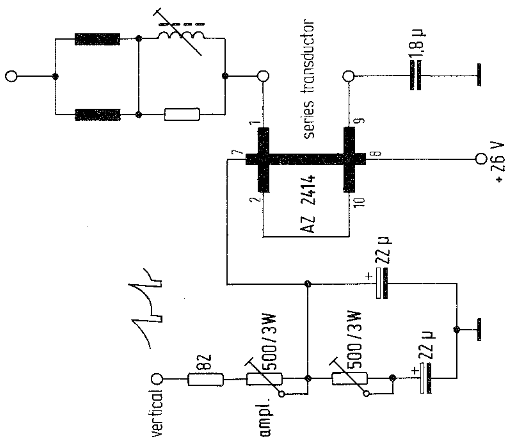
For pincushion correction the simple circuit shown in **Fig. 3.6.3.** is recommended.

3.6.4. For PI-Tubes 27"

The pincushion correction for this picture-tube type uses two passively controlled transductors (**Fig. 3.6.4.**). The parallel transductor AZV 276 MZ in conjunction with the phase coil AZV 254 compensates for both the north-south and the east-west pincushion distortions. The series transductor AZ 2414 corrects the east-west inner pincushion distortions.

3.6.5. For RIS-Inline Picture Tubes 18" and 22"

The pincushion correction circuit is shown in **Fig. 3.6.5.** The transductor AZV 249 MZ/2 is passively driven. It compensates the north-south as well as the east-west pincushion distortions. The corresponding north-south phase coil is identified by AZ 3522.


series transductor
1,8 μ
AZ 2414
+ 26 V
22 μ
vertical
82
ampl.
500/3W
500/3W
22 μ

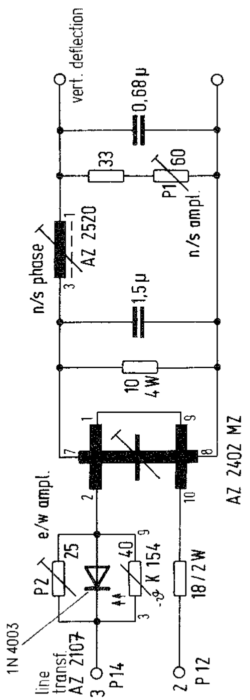

vert. deflection
0,68 μ
33
P1
60
n/s ampl.
n/s phase
AZ 2520
1,5 μ
10
4 W
AZ 2402 MZ
e/w ampl.
P2
25
40
K 154
18/2 W
1N 4003
line
transf.
AZ 2107
3
P14
2
P 12

Fig. 3.6.2.

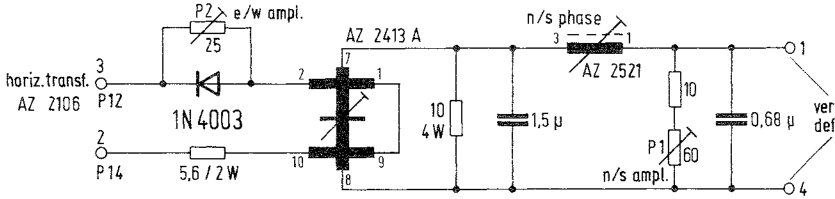


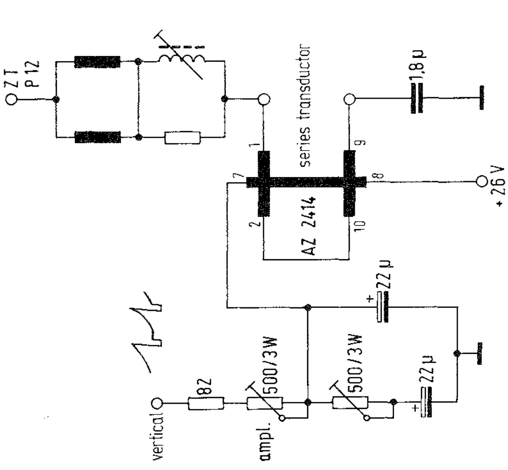
Fig. 3.6.3.

3.6.6. For AX 20-Systems

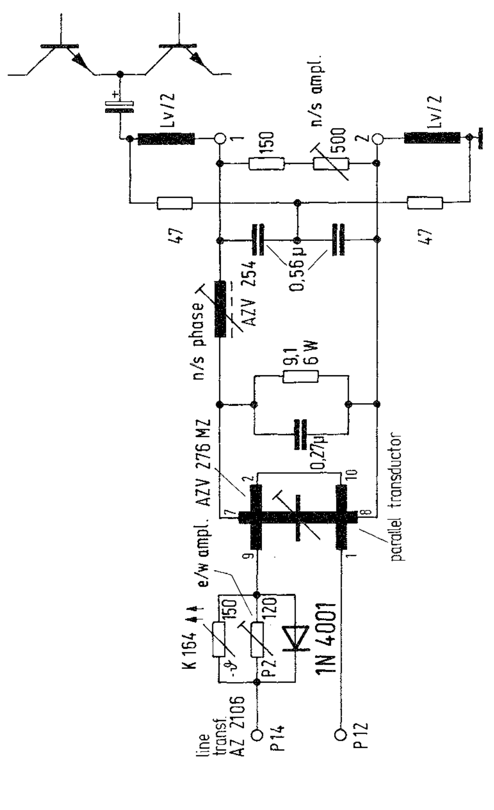
Since the north-south pincushion distortions are unimportant no correction is necessary. The compensation of the 14% east-west-pincushion distortions is achieved by two transductors of the material N 41, which allows an overcompensation of 4%. The main correction as well as the one of the keystone effect is realized by the parallel transductor, driven through the vertical saw-tooth voltage. The passively controlled series transductor AZV 256/1 achieves the inner pincushion correction and also supports the parallel transductor AZV 255 MZ. The latter is connected to the winding 10–11 of the flyback transformer because of the better cross coupling. This is more effective than triggering through an auxiliary winding supplying the adequate pulses. The keystone correction is achieved by applying a dc current to the control coil of the parallel transductor. This dc current is superimposed to saw-tooth ones, flowing in the coil. Therefore the correction maximum of the transductor is shifted, which results in a keystone correction.
The dc voltage is produced by rectifying the flyback voltage supplied from two windings around the free leg of the flyback transformer. By use of a potentiometer with a centre tap an adjustment in both directions is possible.
The parabolic control current for the series transductor AZV 256/2 is generated by integration of the deflection voltage.
The so-called "corner-winking" created by a very strong pincushion correction can be reduced to a negligible extent by suitably dimensioning the inductance of the commutating coil.

List of Capacitors used in the Circuit 3.6.

1 pc	MKM-stacked film cap.	0.33 µF/100 V	B 32540–A 1274–J
1 pc	MKM-stacked film cap.	0.033 µF/250 V	B 32540–A 3333–J
1 pc	MKM-stacked film cap.	0.22 µF/400 V	B 32892–A 4224–K
1 pc	—	(no Siemens type available, before 68 nF)	
1 pc	MKH-capacitor	100 nF/1.5 kV	B 32227–A 1104–M
1 pc	Ceramic capacitor	390 pF/1.5 kV	B 38890–A 1391–M1



Z T
P 12
series transductor
1,8 µ
1
9
7
8
2
10
AZ 2414
+ 26 V
22 µ
vertical
82
500/3W
ampl.
500/3W
22 µ



Lv/2
1
150
n/s ampl.
500
2
Lv/2
47
47
0,56 µ
AZV 254
n/s phase
9,1
6 W
0,27µ
AZV 276 MZ
2
10
7
8
e/w ampl.
9
1
parallel transductor
K 164
150
120
P2
1N 4001
line
transf.
AZ 2106
P 14
P 12

Fig. 3.6.4.

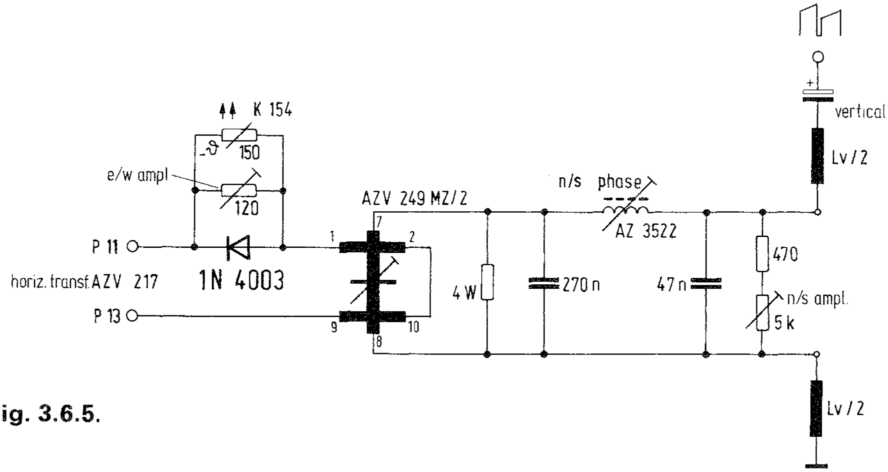


Fig. 3.6.5.

List of Capacitors used in the Circuit 3.6.1.

C_1	1 pc MKH-capacitor	5 nF/1.6 kV	B 32237–A 1502–S
$C_2/C_3/C_5$	1 pc Triple-layer cap.*	0.18 μF/68 nF/0.22 μF	
C_4	1 pc MKM-stacked film cap.	0.22 μF/250 V	B 32541–A 3224–J
C_6	1 pc —	12 μF/160 V	—
C_7	1 pc MKM-stacked film cap.	0.68 μF/400 V	B 32892–A 4684–K
	1 pc MKM-stacked film cap.	68 nF/250 V	B 32540–A 3683–J
	1 pc MKM-stacked film cap.	47 nF/250 V	B 32540–A 3473–J
	1 pc MKM-stacked film cap.	330 nF/100 V	B 32540–A 1334–J

List of Capacitors used in the Circuit 3.6.2.

C_1	1 pc MKH-capacitor	5 nF/1.6 kV	B 32237–A 1502–S
$C_2/C_3/C_5$	1 pc Triple layer capacitor*	0.15 μF/56 nF/0.15 μF	
C_4	1 pc MKM-stacked film cap.	0.22 μF/250 V	B 32541–A 3224–J
C_6	1 pc MKM-stacked film cap.	0.58 μF/250 V	B 32542–A 3684–J
C_7	1 pc —	1.8 μF/400 V	—
	1 pc MKM-stacked film cap.	0.68 μF/100 V	B 32540–A 1684–J
	1 pc MKH-capacitor	1.5 μF/250 V	B 32234–A 3155–K
	2 pcs AL-electrolytic cap.	22 μF/40 V	B 41283–B 7226–T

List of Capacitors used in the Circuit 3.6.3.

C_1	1 pc MKH-capacitor	5 nF/1.6 kV	B 32237–A 1502–S
$C_2/C_3/C_5$	1 pc Triple-layer capacitor*	0.1 μF/33 nF (40 n)/0.082 μF	
C_4	1 pc MKM-stacked film cap.	0.22 μF/250 V	B 32541–A 3224–J
C_6	1 pc MKH-capacitor	0.68 μF/250 V	B 32542–A 3684–J
C_7	1 pc —	1.5 μF/400 V	—
	1 pc MKH-capacitor	1.5 μF/250 V	B 32234–A 3155–K
	1 pc MKM-stacked film cap.	0.68 μF/100 V	B 32540–A 1684–J

* Ordering codes on request

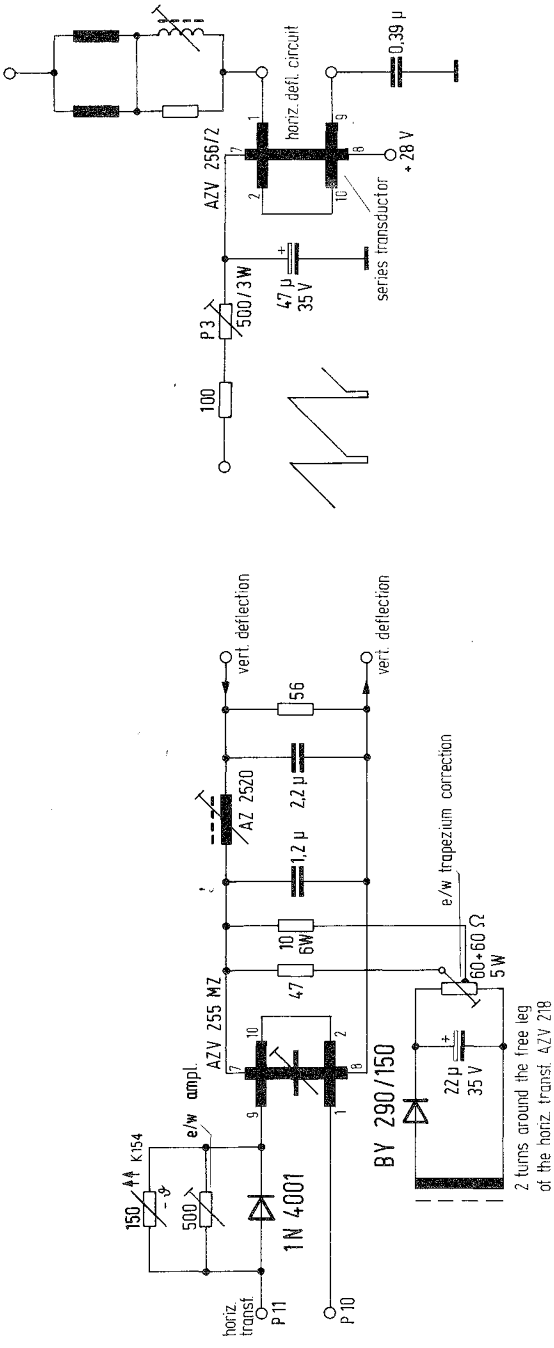

0,39 µ
horiz. defl. circuit
AZV 256/2
+ 28 V
series transductor
47 µ
35 V
500/3 W
P 3
100
vert. deflection
vert. deflection
56
2,2 µ
AZ 2520
1,2 µ
10
6W
47
AZV 255 MZ
e/w trapezium correction
60+60 Ω
5 W
22 µ
35 V
BY 290/150
2 turns around the free leg
of the horiz. transf. AZV 218
e/w ampl.
1 N 4001
150
K154
500
horiz.
transf.
P 11
P 10

Fig. 3.6.6.

List of Capacitors used in the Circuit 3.6.4.

C_1	1 pc MKH-capacitor	5 nF/1.6 kV	B 32237–A 1502–S
$C_2/C_3/C_5$	1 pc Triple-layer capacitor*	0.12 μF/47 nF/0.15 μF	
C_4	1 pc MKM-stacked film cap.	0.22 μF/250 V	B 32541–A 3224–J
C_6	1 pc MKM-stacked film cap.	0.68 μF/250 V	B 32542–A 3684–J
C_7	1 pc —	1.8 μF/400 V	—
	1 pc —	0.27 μF/100 V	—
	2 pcs —	0.56 μF/100 V	—
	2 pcs AL-electrolytic cap.	22 μF/40 V	B 41283–B 7226–T

List of Capacitors used in the Circuit 3.6.5.

C_1	1 pc —	2.2 nF/1000 V	—
$C_2/C_3/C_5$	1 pc Triple-layer capacitor*	0.15 μF/56 nF/0.12 μF	
C_4	1 pc MKM-stacked film cap.	0.22 μF/250 V	B 32541–A 3224–J
C_6	1 pc MKM-stacked film cap.	0.68 μF/250 V	B 32542–A 3684–J
C_7	1 pc —	1.8 μF/400 V	—
	1 pc MKM-stacked film cap.	47 n/250 V	B 32540–A 3473–J

List of Capacitors used in the Circuit 3.6.6.

C_1	1 pc MKH-capacitor	5 nF/1.6 kV	B 32237–A 1502–S
$C_2/C_3/C_5$	1 pc Triple-layer capacitor*	0.1 μF/51 nF/0.22 μF	
C_4	1 pc MKM-stacked film cap.	0.22 μ/250 V	B 32541–A 3224–J
C_6	1 pc —	12 μ/160 V	—
C_7	1 pc —	0.39 μ/400 V	—
	1 pc MKM-stacked film cap.	0.22 μF/100 V	B 32541–A 1224–J
	1 pc MKM-stacked film cap.	1 μF/100 V	B 32541–A 1105–J
	1 pc MKL-capacitor	2.2 μF/63 V	B 32110–F 9225–K
	1 pc AL-electrolytic cap.	22 μF/40 V	B 41283–B 7226–T
	1 pc AL-electrolytic cap.	47 μF/40 V	B 41283–C 7476–T

* Ordering codes on request

3.7. Vertical Deflection Circuit using TCA 880 for Picture Tubes of the 20 AX-System

The vertical deflection circuit shown in **Fig. 3.7.** employs as active components two epibase transistors BD 437/BD 438, and an IC type TCA 880. The latter contains the vertical frequency oscillator, an S-correction, which is externally adjustable, a symmetrical amplitude set, which also allows a beam-dependent amplitude control as well as the drivers for the external output-stage transistors. A blanking pulse selectable with positive or negative amplitude is available at pin 4. Its duration is adjustable.

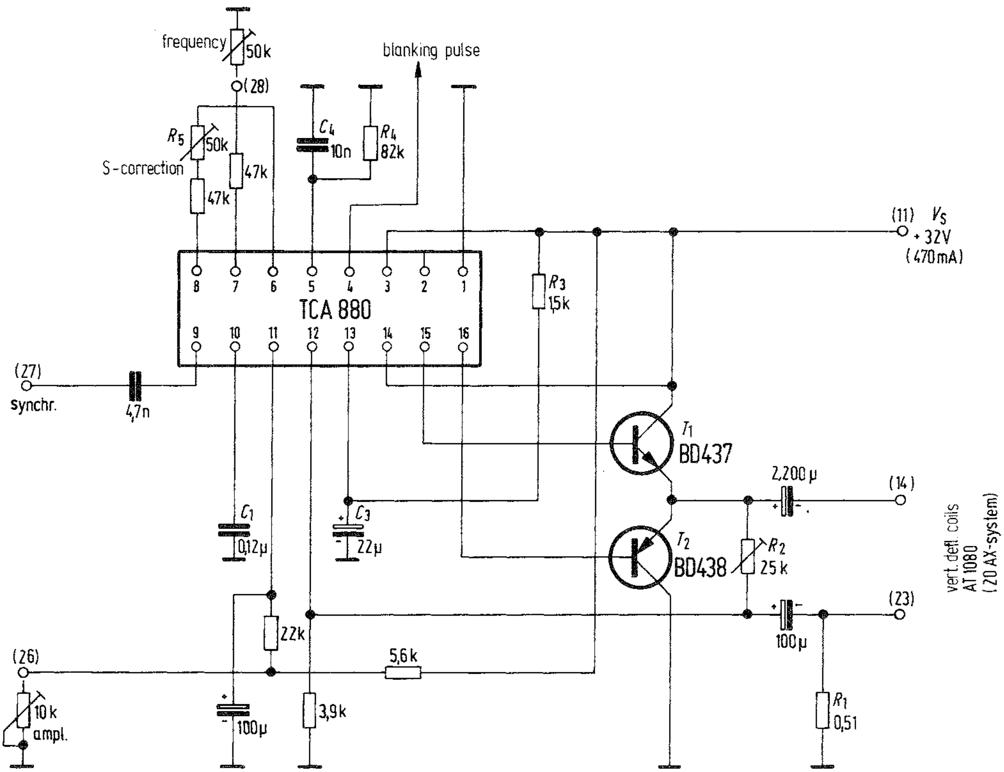


Fig. 3.7.

The vertical sync pulse is applied to pin 9. It synchronizes the vertical oscillator operating according to the principle of a threshold switch. The frequency is determined by the capacitor C_1 connected to pin 10. The fine tuning is realized by influencing the charging and discharging current sources through an external potentiometer connected to pin 7. The complementary output stage transistors BD 437 and BD 438, connected to pins 15 and 16 supply the energy required by the deflection coils.

The voltage drop across the resistor R_1 is produced by the current of the deflection coils. It is applied as a negative-feedback voltage to pin 12. Besides that a dc voltage adjustable by the potentiometer R_2 is also supplied to pin 12. It generates a symmetrical operating point of the output-stage transistors.

The vertical gain control is internally achieved through an electric potentiometer by applying a variable dc voltage to pin 11. If this dc voltage is obtained by rectification of the high voltage at the cascade, it depends on the beam current and can be utilized to compensate picture size variation generated by beam current changes.

The IC requires internally a supply voltage of 10 V. Therefore an RC-circuit consisting of R_3 and C_3 has to be connected to the 32-volt terminal (11) of the external power supply and the pin 13. The RC-circuit incorporating R_4 and C_4 determines the blanking time of the pulse available at pin 4. An S-correction is accomplished by the potentiometer R_5 connected to pins 6 and 8.

Compared to circuits using discrete semiconductors the solution shown in Fig. 3.7. is distinguished by the fact that less external components are required and that it offers a vertical AGC, depending on a dc current. Besides that it is characterized by blanked sync pulse input which is especially intensive against interferences and by the possibility of achieving a simple S-correction.

List of Capacitors used in the Circuit 3.7.

Quantity	Components	Value	Ordering codes
1	MKM-stacked film capacitor	4.7 nF	B 32540–A 3472–K
1	MKM-stacked film capacitor	0.01 µF	B 32540–A 3103–J
1	MKM-stacked film capacitor	0.15 µF	B 32540–A 1154–J
1	Electrolytic capacitor	22 µ/25 V	B 41313–A 5226–Z
2	Electrolytic capacitors	100 µ/16 V	B 41286–A 4107–T
1	Electrolytic capacitor	2200 µ/25 V	B 41010–B 5228–T

3.8. Self-Oscillating Power Supply for Colour TV-Receivers with Mains Isolation

Power supplies with mains isolation are essentially advantageous for the total concept of a colour TV-receiver. The aerial can be directly coupled to the front end and besides that sockets for video recorders, head phones and tape recorders can be incorporated to the set without any additional elaborations.
The principle of operation has already been described in the Design Examples of Semiconductors, Edition 1974/75. But some improvements with regard to the switching behaviour of the transistors have been introduced. Besides that an RFI-suppression circuit assures that the maximum of the specified interference level is not exceeded. The circuit of the power supply (**Fig. 3.8.**) operates upon the principle of a non-synchronized, self-oscillating dc converter. It oscillates with a frequency between 20 and 28 kHz. All output voltages are stabilized and thus an open-loop-proof as well as a short-circuit-proof operation is achieved. The mains supply voltage can vary between 180 and 265 V. Mains voltage fluctuations of $\pm 20\%$ are reduced to $<1\%$ at the output. The hum voltage of about 18 V_{pp} at the input capacitor is decreased to a value of 0.3 V_{pp} at the 225 V-output. As against conventional circuits without mains isolation the circuit including the isolation requires only a minimum of additional elaborateness in components. The transformer has to be moulded to meet VDE-standards.

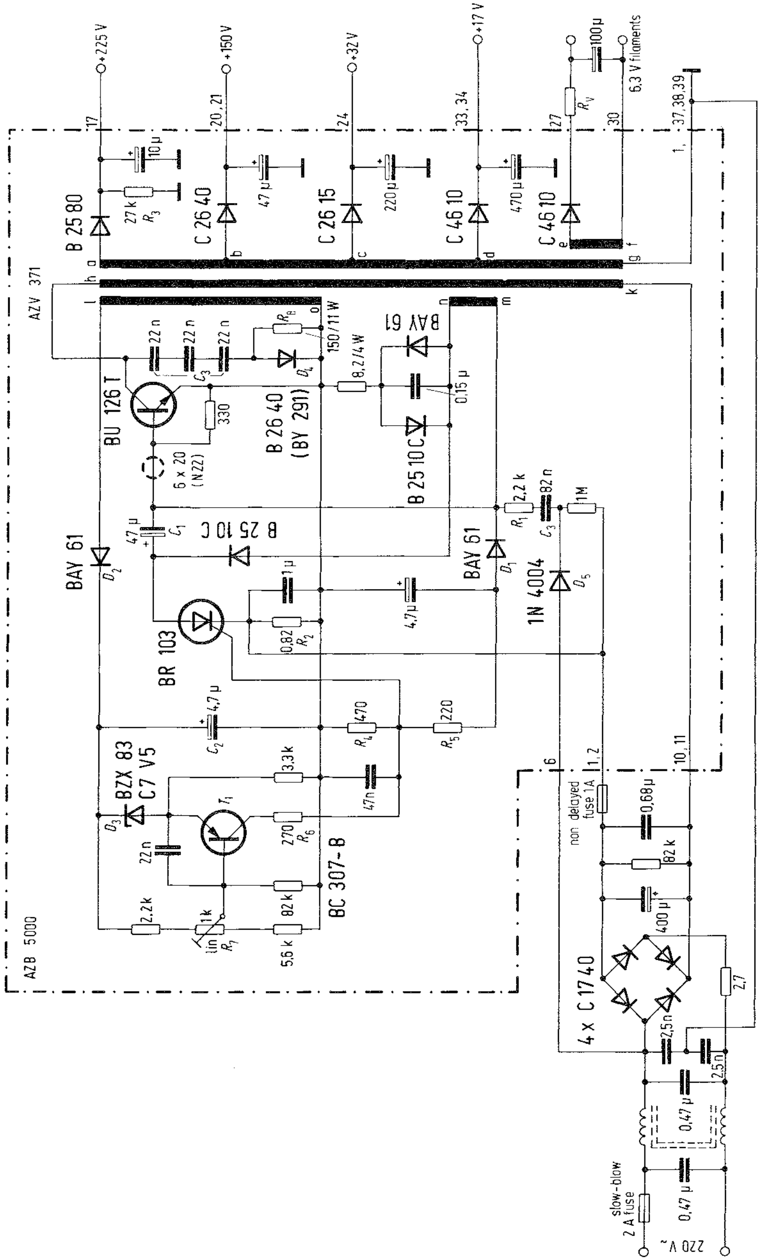

+225 V
+150 V
+32 V
+17 V
100 µ
6,3 V filaments
17
20, 21
24
33, 34
27
30
37, 38, 39
1
B 25 80
27 k
R_3
10 µ
C 26 40
47 µ
C 26 15
220 µ
C 46 10
470 µ
C 46 10
R_V
AZV 371
150/11 W
R_8
22 n
22 n
22 n
C_3
D_4
BU 126 T
330
B 26 40
(BY 291)
8,2/4 W
BAY 61
0,15 µ
B 25 10 C
6 x 20
(N22)
47 µ
C_1
B 25 10 C
2,2 k
R_1
82 n
C_3
1M
BAY 61
D_2
BAY 61
D_1
1N 4004
D_5
1 µ
4,7 µ
BR 103
0,82
R_2
4,7 µ
C_2
470
R_4
220
R_5
BZX 83
C7 V5
D_3
3,3k
47n
T_1
22 n
270
R_6
82 k
2,2k
1k
lin
R_7
5,6 k
BC 307-B
AZB 5000
6
1, 2
10, 11
non delayed
fuse 1A
0,68 µ
82 k
400 µ
4 x C 17 40
2,7
2,5 n
2,5 n
0,47 µ
0,47 µ
slow-blow
2 A fuse
220 V~

Fig. 3.8.

Description of Functions

The mains voltage is rectified in a bridge circuit (4 × C 1740) and smoothed through an electrolytic capacitor of 400 μF. The BU 126 T operates as switching transistor. The control of the output voltages is attained by affecting the energy which is stored in a transformer during the forward period. A balance is achieved between the energy stored during the forward period and the one released during the reverse period. The stored energy is dosed by controlling the collector peak-current of the switching transistor BU 126 T.
When the oscillation is started by a pulse derived from the mains voltage, it is then maintained through the inverse-feedback winding n–m. A voltage drops across the resistor R_2 which depends on the collector current of the BU 126 T. Through the voltage divider R_5/R_4 the gate of the switching-off thyristor BR 103 is biased to a level of -2 V referred to the cathode.
The negative voltage for the divider is generated through the diode D_1 from the voltage across the feedback winding during the reverse period. The voltage drop across the resistor R_2 now opposes the negative bias of the gate. As soon as the triggering level (about $+0.7$ to 1 V) is exceeded, the thyristor is triggered. When it is switched on a negative level is achieved at the base of the BU 126 T, by the capacitor C_1, and this turns off the thyristor. The thyristor remains conductive during the switching and reverse time of the BU 126 T and it is turned off at the zero-axis crossing, which is achieved by the polarity change of the reverse-feedback voltage. The max. possible collector peak-current of the BU 126 T depends on the dimensioning of the voltage divider R_4/R_5. If the divider is not loaded the maximum is about 4 A_{pp}. If it is loaded by the transistor T_1 and the resistor R_6 the negative bias is changed during the control process. If the bias is high the auxiliary thyristor only triggers at high collector peak-currents, i.e. much energy is stored in the transformer. To decrease the amount of stored energy the bias is simply reduced by loading the voltage divider. The control information is generated from the winding l–o, which is tightly coupled to the windings of the output voltages. The diode D_2 generates a dc voltage which depends on the output voltage. The control transistor is turned on when the voltage at C_2 exceeds a fixed level adjusted by the potentiometer R_7. Through this transistor the negative bias at the gate of the BR 103 is reduced. Therefore the thyristor triggers earlier and the transistor BU 126 T turns off at lower collector peak-currents.
The combination of the rectifier diode D_4, being in parallel to the resistor R_8, and of the RC-circuit consisting of R_8 and C_3 essentially reduces the switching losses of the transistor BU 126 T.

Auxiliary circuit for the oscillation start

Defined starting pulses of a 5-ms-duration are generated from the mains ac-voltage by means of the diode 1 N 4004 and the RC-circuit C_3/R_1. They are applied to the base of the BU 126 T and the transistor becomes conductive. Thus the oscillation begins.

Open-loop operation

The power supply is suitable for power outputs of about 200 W if a EE 42-core is used and up to 280 W with a EE 55-core. Between open-loop operation and a load of 40 W it runs in an intermitting 50 Hz-operation.

At loads less than 40 W the switching frequency rises. If the period time is less than 25 µs the turn-off time of the thyristor cannot be reached. The thyristor remains turned on and the oscillation is interrupted. A new beginning of the oscillation is only possible with the next starting pulse. During the open-circuit operation a pulse train with spacings of 20 ms is generated. The resistor R_3 acts as basic load to prevent a very high increase of the output voltages. It has been practically experienced that thus a high safety is achieved at open-loop operations.

Standard operation

The standard-operation ranges between 40 W and about 180 W-loads.

Short-circuit operation

If a short-circuit occurs at one of the outputs the self-oscillating dc converter changes to an intermitting operation. The continuous sequence of collector peak-current pulses is replaced by a pulse train with a spacing of 20 ms. This behaviour results from the 50-Hz-starting-circuit. At the same time the spacing of the individual current pulses is increased to about 2 ms. Besides that the collector voltage of the BU 126 T is reduced from about 600 V_{pp} to max. 380 V. The maximum ratings of the BU 126 T are not exceeded during short-circuit operation, which is realized by a special starting circuit. The short-circuit current is determined by the energy stored in the transformer. By reducing the amount of collector current pulses per time unit, the stored energy can be decreased.
First the collector peak current of the BU 126 T will increase in the case of a short-circuit. This results in triggering the thyristor and in switching off the BU 126 T. At the same time the voltage across the winding n–m and thus also the bias at the divider consisting of R_4 and R_5 are reduced. Therefore the thyristor triggers at lower collector currents.
The transistor can be switched only, when the thyristor is turned off and when a new starting pulse is available. Collector-current pulses are only possible during the defined starting period of 5 ms, which follows with a distance of 20 ms (50 Hz-mains frequency). As only single charging pulses are produced the short-current flowing through the output stage diodes is very low, i.e. fuses are only required in exceptional cases.
The described power supply is available as a complete module under the ordering code number AZB 5000. However the bridge-rectifiers, the charging capacitors and the RFI-filtering circuit are not included.

Technical data:

Mains voltage range	180 to 265 V (150 to 265 V)
Nominal output voltages	225 V=/0.1 A
	150 V=/0.7 A
	32 V=/0.6 A
	17 V=/1.2 A
	6.3 V=/0.9 A for the picture tube heating a series resistor is required.

List of Capacitors used in the Circuit 3.8.

Quantity	Components	Value	Ordering codes
4	MKM-stacked film capacitors	0.022 μF/250 V	B 32540–A 3223–J
1	MKM-stacked film capacitor	0.047 μF/250 V	B 32540–A 3473–J
1	MKM-stacked film capacitor	0.15 μF/10 V	B 32540–A 1154–J
1	MKM-stacked film capacitor	1 μF/100 V	B 32541–A 1105–J
1	MKM-stacked film capacitor	0.33 μF/250 V	B 32540–A 3333–J
1	MKM-stacked film capacitor	0.47 μF/250 V	B 32540–A 3473–J
1	Electrolytic capacitor	10 μF/350 V	B 43052–A 4106–T
1	Electrolytic capacitor	47 μF/250 V	B 43052–C 2476–T
1	Electrolytic capacitor	220 μF/40 V	B 41286–A 7227–T
1	Electrolytic capacitor	470 μF/25 V	B 41012–A 5477–T
1	Electrolytic capacitor	4.7 μF/16 V	B 41315–A 4475–Z
1	Electrolytic capacitor	4.7 μF/40 V	B 41315–B 7475–Z
1	Electrolytic capacitor	47 μF/40 V	B 41286–B 7476–T

3.9. Tint Control for TV-Receivers According to NTSC-Standards

In TV-receivers applying to the PAL-standards the received burst signal and the one of the reference oscillator have a fixed phase relation. Contrary to this, TV sets operating according to the NTSC-standards contain a correction circuit to compensate colour purity errors generated by differential phaseshifts during the transmission. At PAL-receivers this correction is automatically achieved by line-by-line keying of the (R–Y) component with a following interpretation of the results obtained from two successive lines.

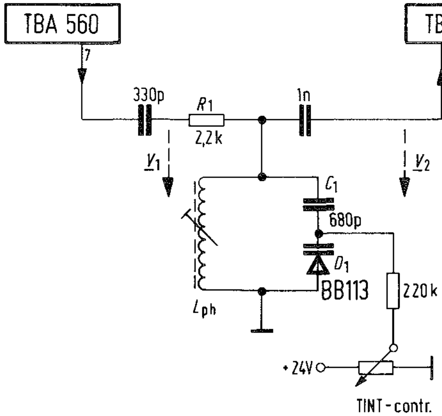
In principle the phase shift can be realized at the chrominance signal as well as at the reference signal. But the engineering effort is smaller for the latter, since the total bandwidth of the modulation spectrum has not to be transmitted. The solution shown in **Fig. 3.9.** recommends the use of a resonant-circuit with an AM-tuning-diode BB 113, the capacitance of which is varied by a dc voltage.

When the resonant frequency of the circuit coincides with that of the chrominance signal no phase shift occurs. However, when the resonant signal is detuned towards lower frequencies the output signal lags behind and when it is detuned towards higher frequencies the voltage leads in phase (as shown in Fig. 3.9. on the right side). The reference oscillator circuit synchronizes upon the phase-shifted burst which results in a phase shift between the transmitted burst and the reference signal, being supplied to the demodulator. The phase shift produces the desired change of the tint. The resonant circuit has to be tuned in such a manner that its frequency coincides with that of the respective colour subcarrier (at NTSC-standards e.g. 3.58 MHz), when the tint-control potentiometer is in its centre position.

Data of the coils L_{ph}

5 filter 10 K (Toko)

40 turns 0.12 enamelled copper wire (without screening can)



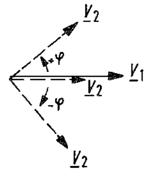


Fig. 3.9.

List of Capacitors used in the Circuit 3.9.

1 × 330 pF	B 37205–A 5331–M 1
1 × 680 pF	B 37208–A 5681–M 1
1 × 1 µF	B 37235–J 5102–M 001

3.10. Appropriate Connection for VCR-Devices

The DIN-standard 45482 not only considers the high-frequency connection of a VCR-device to a TV-receiver but also one, utilizing the video signal. All levels and conditions are standardized for these operations.
In section 3.2. a connection for an AF recorder/replayer has already been described. Now the completion to a VCR-device using the video signal is shown in **Fig. 3.10.** It consists of an emitter follower for matching the input impedance of the VCR-device and of a circuit which switches off the video-IF-amplifier by means of a 12 V-voltage during the replay operation.

Mode of record operation

The video signal available at pin 11 of the TBA 440 N/P passes the sound trap and by that the sound carrier with a frequency of 5.5 MHz is eliminated. The matching to the input impedance of the VCR-device is achieved by the emitter follower in conjunction with the resistors R_2 and R_3.
The circuit is proportioned for an input impedance of 75 Ω. When the connecting plug is pulled an equivalent resistor of $R_1 = 75\,\Omega$ is added via the contacts of the socket and the resistors R_2 and R_3 maintain, in conjunction with the input impedance of the emitter follower, the desired source impedance of 75 Ω. In the position "record" no switching voltage is applied to terminal 1. The diodes D_1 and D_2 are turned off and the video-IF-amplifier regularly operates.

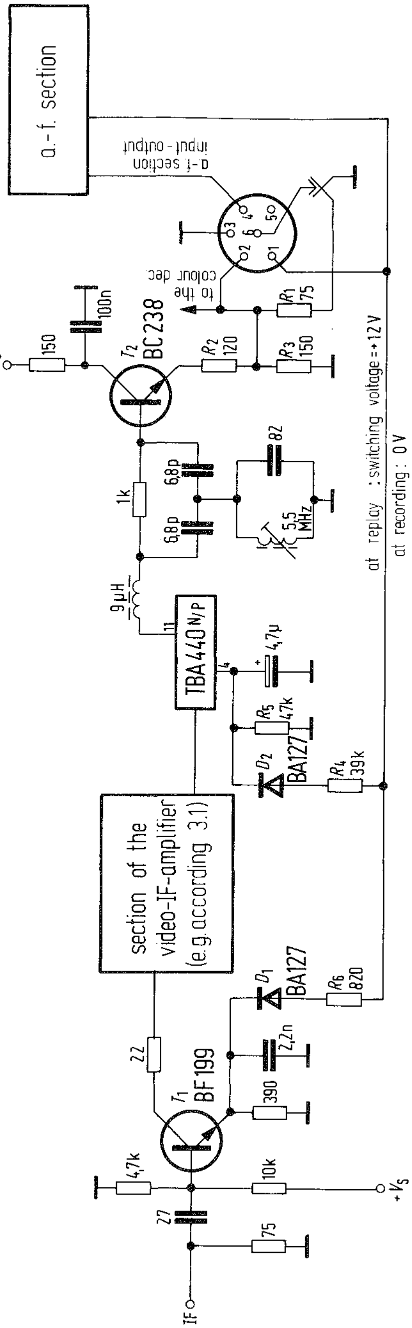

a.-f. section
a-f section
input - output
to the
colour dec.
section of the
video-IF-amplifier
(e.g. according 3.1)
TBA440 N/P
BC238
BF199
BA127
at replay : switching voltage = +12 V
at recording : 0 V
IF
+VS
5,5 MHz
9 µH

Fig. 3.10.

Mode of replay operation

At the operation "replay" the VCR-device supplies a switching voltage of 12 V to terminal 1. It passes the resistor R_4, the diode D_2 and it is applied to pin 4 of the IC TBA 440 N/P. The voltage drop at R_5 reduces the gain of the IC by a minimum of 52 dB (value according data sheet). Besides that the gain of the tuner is also diminished.
A positive voltage is applied to the emitter of transistor T_1 via the resistor R_6 and the diode D_1 in order to suppress the received composite picture signal, if large signals are picked up at the aerial and if the tuner control should be misadjusted. The transistor is switched off and the gain of the video-IF amplifier is additionally reduced by about 40 dB.
The video-signal of the VCR-device is supplied via the connection-terminal 2 to the input of the colour decoder. Since the video-IF-amplifier is controlled in such a way that it offers no gain, the emitter follower T_2 only receives a dc voltage, which is available at terminal 2 and the level of which is within the tolerances indicated by the DIN-Standards. In this case the matching to 75 Ω is also achieved by the internal impedance of the emitter follower and the two resistors R_2 and R_3.

List of Capacitors used in the Circuit 3.10.

2 pcs	6.8 pF	B 38112–J 5060–C 800
1 pc	100 nF	B 37449–A 6104–S 1

3.11. Correction-Matrix for a Composite-Picture-Signal Generator

With colour broadcasting special, relative response curves of phosphors being utilized in the picture tube for the picture build-up are standardized. They are principally shown in **Fig. 3.11.2.** These curves also contain negative components due to the fact that fully saturated colours cannot be obtained by the phosphors available today. These negative portions are practically useless for the replay of saturated colours with a colour TV-receiver, since in this case the picture tube is operated below the cut-off voltage. With a colour TV-camera having only three pick-up tubes it is also impossible to realize the response curves by only using a light divider with negative components of the characteristics. Therefore the chromaticity diagram is compromized, i. e. not only fully saturated colours cannot be reproduced, but also the saturization of the other colours is additionally reduced. Besides that the hue is also negatively influenced. An improvement is possible when the negative components of the curves are realized at least at the signal generation in the colour TV-camera and if only one limitation of the picture tube control is applied. By this measure fully saturated colours cannot, of course, be reproduced, but the colour saturation within the chromaticity diagram can be generally improved.
It is principally possible to produce through a linear transformation the response curves with negative components as of **Fig. 3.11.2.** from those shown in **Fig. 3.11.3.** and having only positive portions. Besides that the luminance signal can be additionally generated. The logic operation is achieved by the matrix according to equation 1. Electrical circuits realizing this transformation are presently used in all types of studio TV-cameras.

$$\begin{vmatrix} B_2 \\ G_2 \\ R_2 \\ Y_2 \end{vmatrix} = \begin{vmatrix} a_{11} & a_{12} & a_{13} \\ a_{21} & a_{22} & a_{23} \\ a_{31} & a_{32} & a_{33} \\ a_{41} & a_{42} & a_{43} \end{vmatrix} \bullet \begin{vmatrix} B_1 \\ G_1 \\ R_1 \end{vmatrix} \tag{1}$$

$$\begin{vmatrix} R_2 - Y_2 \\ B_2 - Y_2 \end{vmatrix} = \begin{vmatrix} k_R & 0 \\ 0 & k_B \end{vmatrix} \bullet \begin{vmatrix} R_1 - Y_2 \\ B_1 - Y_2 \end{vmatrix} \tag{2}$$

Unfortunately the realization of these circuits is joint with a great engineering effort, as a large quantity of positive as well as negative coefficients is required.

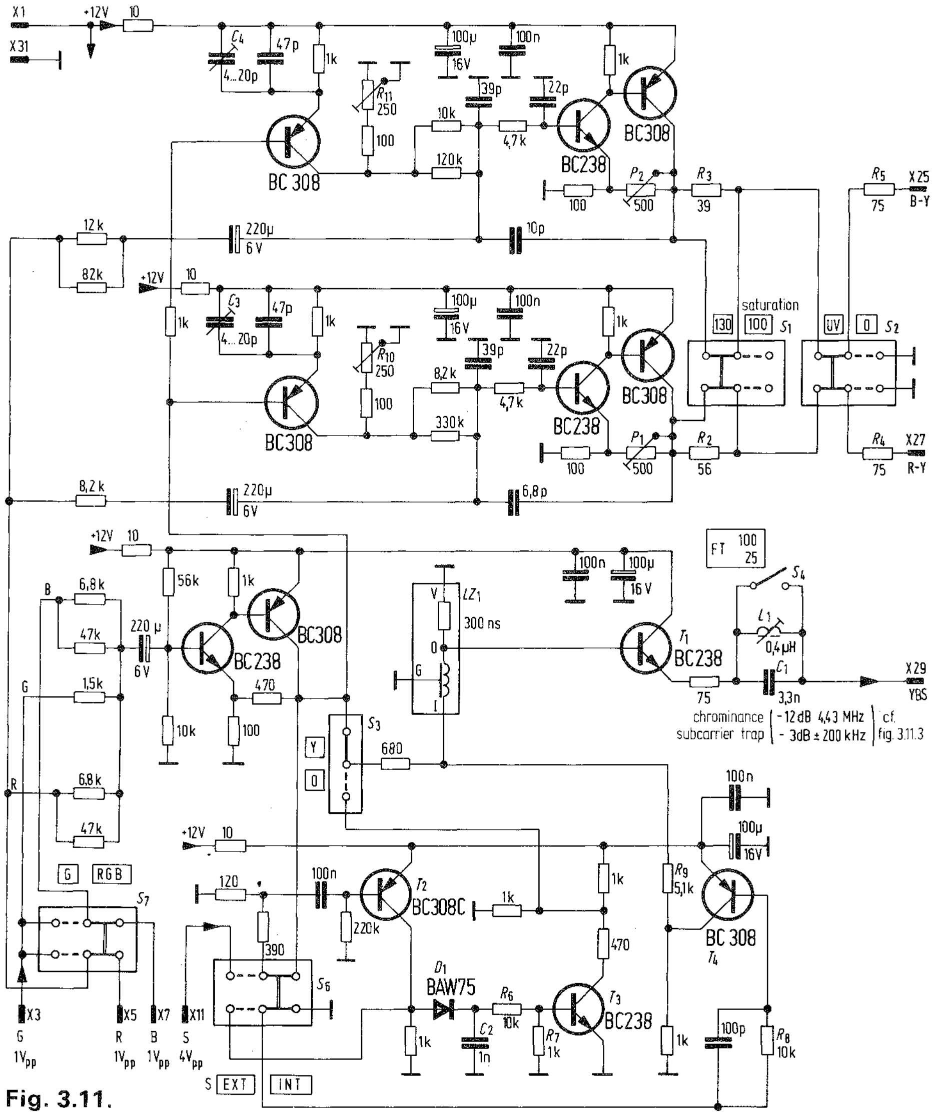


Fig. 3.11.

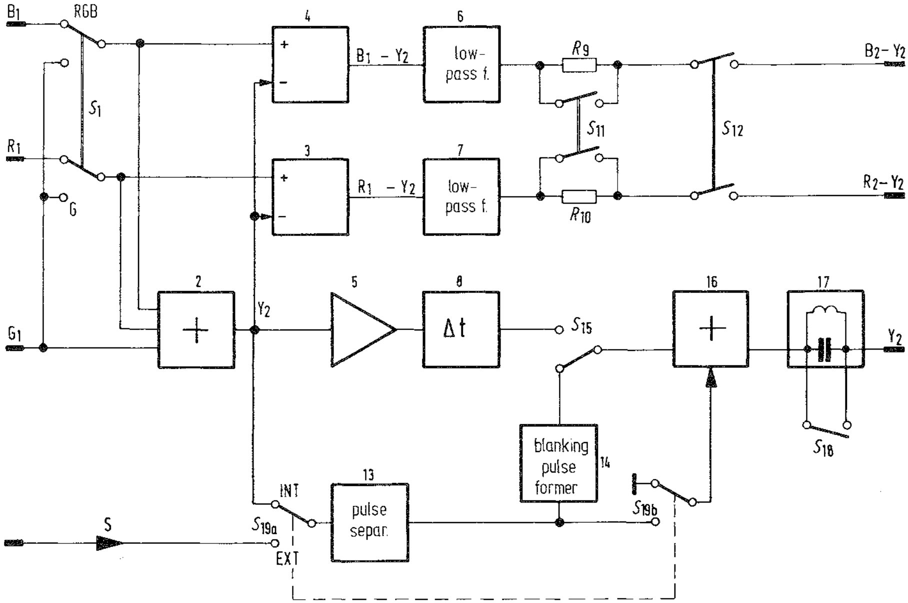


Fig. 3.11.1.

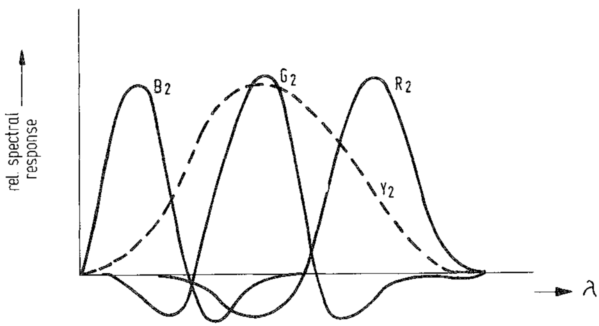


Fig. 3.11.2.

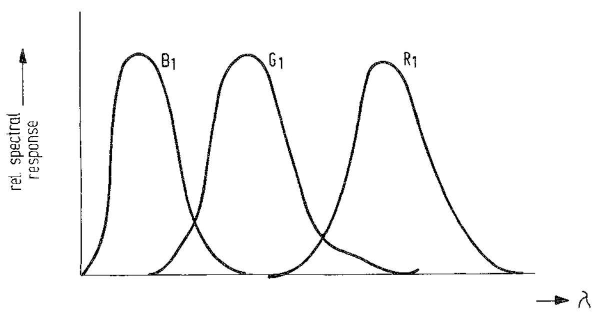


Fig. 3.11.3.

chrominance subcarrier trap

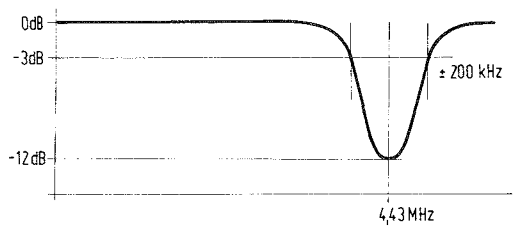


Fig. 3.11.4.

If only the luminance signal Y_2 is obtained by a matrix containing B_1, G_1 and R_1 (this only requires positive components), then the transformation of the other signals can be combined in a matrix of the colour difference signals, as it has been experienced. This is described in equation 2.

As it can be generally derived, only a gain increase for the colour difference signals is necessary to achieve the desired colour correction. Thus a very easy circuit can be realized and it also offers the advantage in that one luminance signal and two difference signals are supplied which can be directly applied to a colour decoder for transmission of the composite picture signal. By using such a circuit the colour reproduction of older devices and of simple industrial TV-cameras can be essentially improved.

With the aid of the block diagram shown in **Fig. 3.11.1.** the functions of the mentioned circuit are described as follows. In the RGB-position of the switch 1 the three primary signals R_1, G_1 and B_1 are applied to the luminance-signal matrix 2, which is a simple counter stage. In the two differential amplifiers 3 and 4 the luminance signal is subtracted from the red and blue signal to produce the difference signals $(B_1 - Y_2)$ and $(R_1 - Y_2)$. The colour information being incorporated to these difference signals can be transmitted via circuits with a reduced band width, which is achieved by the low-pass filters 6 and 7. These active filters amplify simultaneously the colour difference signals so that they are available at the output as the difference signals $(B_2 - Y_2)$ and $(R_2 - Y_2)$ with a higher amplitude if the switches S_{11} and S_{12} are closed. The latter signals contain the desired corrections for the colour saturation. By means of the switch S_{12} the colour difference signals can be switched off, so that only a monochrome operation is maintained. The switch S_{11} adds the series resistors R_9 and R_{10}, which eliminate the influence of the colour saturation correction for test purposes.

The luminance signal Y_2 is supplied to the throw-over switch S_{15} via the amplifier stage 5 and the delay line which compensates for the propagation time of the low-pass filters. By means of this switch an operation with the luminance signal or with blanking pulses generating a grey area can be chosen. In the latter case only the

picture corresponding to the colour-difference signals is seen on the screen. This can be utilized for the white level adjustment of a camera, since the colour difference signals have to read zero, i.e. a uniform grey area has to be produced.
In the stage 16 the sync pulses are added to the luminance signal. A trap (17) being tuned to the subcarrier frequency and suppressing cross-colour-effects of the coded PAL-signal, is connected to its output. The trap can be short-circuited by the switch S_{18}.
The blanking pulses are generated from the sync pulses in the former 14. The sync pulses can be derived either from supplied RGB-signals, if they already contain sync pulses, or from external signals applied by means of the switch S_{19}. When the sync pulses are internally produced their addition to the luminance signal is switched off, since corresponding sync pulses are already generated from the input signals R, G, B. The stage 13 separates the sync pulses from any video signal containing S-pulses.
Further details of the colour correction matrix can be had from the circuit shown in **Fig. 3.11**. It consists of a luminance signal matrix, two subtracting stages, where the $(R-Y)$ and the $(B-Y)$ signal are produced, two active low-pass filters with output amplifiers for the $(R-Y)$ signal respectively the $(B-Y)$ signal. The saturation can be reduced by means of the switch S_1 adding the resistors R_2 and R_3. The switch S_7 turns off the colour difference signals. By the resistors R_4 and R_5 the amplifier input impedance of 75 Ω is achieved.
The switch S_3 allows the choose between an operation with luminance signal and the grey level produced by the blanking pulses. An emitter follower serving as an output stage for the luminance signal is connected to the delay line LZ_1 compensating for the propagation time. The trap for the colour subcarrier is tuned to 4.43 MHz and has a 3 dB-band width of ±200 kHz. It contains the components L_1 and C_1 and can be short-circuited by the switch S_4. The switching from internal to external synchronization is obtained by means of the switch S_6. The blanking pulses are produced by extending the positive sync pulses available at the collector of the transistor T_2. The latter operates as a pulse separator. The capacitor C_2 is rapidly charged via the diode D_1 and slowly discharged via the resistors R_6 and R_7. The transistor T_3 is turned on by the positive pulse and switched off when C_2 is discharged via R_6 and R_7. Thus a blanking pulse with a duration of about 12 µs is generated at the collector of T_3. It is supplied to the luminance signal amplifier via switch S_3. At external synchronization positive sync pulses are applied via S_6 and R_8 to the base of transistor T_4, operating as an inverter for the sync pulses, which are added to the luminance signal via R_9 before they enter the delay line.
By means of the switch S_7 the signal supplied to the green-signal input can also be applied to the red-signal and blue-signal input of the matrix. Thus a control of the white level adjustment is possible. At this operation each of the two colour difference signals of the output are adjusted to zero through the resistors R_{10} and R_{11}. The adjustment is made firstly for lower signal frequencies, whereas C_3 and C_4 enable also an alignment for higher signal frequencies. Therefore colour contours are safely avoided. At the RGB-position of the switch S_7 the gain for the colour difference signals is adjusted through the potentiometers P_1 and P_2 and by using a colour bar generator. The subcarrier trap, consisting of L_1 and C_1 is tuned to an attenuation maximum by using a subcarrier signal with a frequency of 4.43 MHz when the switch S_4 is open. At all tests the outputs have to be terminated with an impedance of 75 Ω.

List of Capacitors used in the Circuit 3.11.

Quantity	Components	Value	Rated voltage (V)	Ordering codes
1	Styroflex capacitor	10 pF	63	B 31310–A 5100–F 000
2	Styroflex capacitors	22 pF	63	B 31310–A 5220–F 000
2	Styroflex capacitors	39 pF	160	B 31861–A 1390–F 000
2	Styroflex capacitors	47 pF	63	B 31310–A 5470–H 000
1	Styroflex capacitor	1 nF	25	B 31310–A 3102–H 000
1	Styroflex capacitor	3.3 nF	25	B 31310–A 3332–H 000
1	Styroflex capacitor	100 pF	25	B 31310–A 3101–H 000
1	Ceramic disc-capacitor type 1 B, SDPM	7 pF	500	B 38112–J 5070–D 000
5	Ceramic disc-capacitors, type 3, SDPN	100 nF	16	B 37302–A 1104–Z 001
4	Electrolytic capacitors Tantalum, sinter type	100 µF	16	B 41286–A 4107–T 000
3	Electrolytic capacitors Tantalum, sinter type	220 µF	10	B 45181–A 1227–M 000
2	Ceramic disc trimmers	4.5 to 20 pF	160	STETTNER 7 S-Triko 02

4. Optoelectronic Circuits

4.1. Table Model of a Battery Operated Quartz Clock using a Field-Effect LCD-Device

By using the C-MOS IC's 5437 and 5442 as well as the Siemens field-effect LCD-device type FAN 41320 R (character height = 13 mm) a table model of a 12 hour-quartz-clock running more than 2 years with a miniature 1.5 V-battery (e.g. Varta 7244) can be realized. Besides hours and minutes both seconds and the 2-digit date can be displayed by pushing a button.
The circuit is shown in **Fig. 4.1**. The IC 1, type 5437, contains a crystal-controlled 32.768 kHz-oscillator and 9 dividers, which reduce the frequency to 64 Hz. The operating voltage is 1.5 V typ. and 1.3 V min.

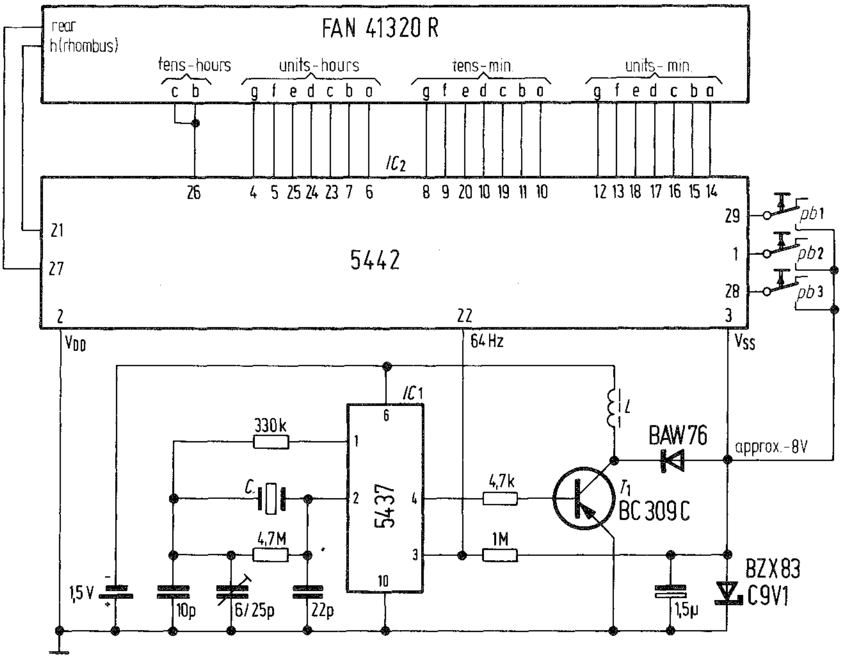


Fig. 4.1.

The hours, minutes, seconds and the date are generated from the 64 Hz pulses by the IC 2, type 5442, operating as a driver. Its 7-segment-coded output information is supplied to the LCD-device FAN 41320 R. The function of the clock can easily be checked due to the fact that a colon or a rhombus sign is always displayed on the LCD-device with a clock frequency of 1 Hz. The polarity of all display outputs is reversed 32 times per second to assure that the LCD-device operates with a dc voltage which guarantees a long life time. The supply voltage of the driver ranges between 7 and 9 V. It is produced from the 1.5 V-battery by means of a blocking oscillator which connects, during a time of about 15 μs a small inductance to the IC and the

1.5 V-battery terminal, via the transistor T_1. The current of the coil increases like a saw-tooth to about 4 mA. After blocking the transistor the flyback pulse is utilized to produce the negative supply voltage of about 8 V.
A z-diode is provided to assure that the adjustable maximum voltage of 10 V is not exceeded during the no-load operation. The trigger pulses for the base of transistor T_1 are supplied from the oscillator IC (2) type 5437. The repetition frequency is 1024 Hz.

Setting Procedures

By the three push buttons (pb 1 to pb 3) either the required information is selected (hours, minutes, seconds or the date), or the time setting is achieved. The different display statuses are indicated by the table shown in **Fig. 4.1.1.**

	1	2	:	3	6
normal position	hours tens	units	rhombus clocking	minutes tens	units
push button pb_2 pushed	off	A or P	off	date tens	units
push button pb_2 pushed	off	off	rhombus clocking	seconds tens	units

Fig. 4.1.1.

The following rules apply:

Pushing of pb 1

The minute display (last two characters) steps on by one unit with the seconds-clock.

Pushing of pb 3

The hour display (first two characters) steps on by one unit within seconds-clock frequency.

Non-recurring push of pb 2

On the two right sections the date is displayed (only day of the month) and on the second section from the left the letter A or P appears. This means before (a.m) or after noon (p.m). The correct setting of the date is achieved as previously described in the procedure: pushing pb 1 or pb 3.

Repeated pushing of pb 2

On the two right sections the seconds are displayed. When the pb 2 is momentarily pushed again the standard display operation is obtained. If the pb 2 is not pushed the device automatically switches to a standard displaying at the following minute change.

Technical Data:

Battery voltage	1.5 V (min. 1.30 V)
Power consumption	
Total	1.5 V/40 μA
Display FAN 41320 R	8 V/13 μA
Oscillator/Divider 5437*)	1.5 V/ 5 μA
Driver 5442*)	8 V/ 1 μA
Efficiency of the converter	70%
Operating time with a 1.5 V-battery type Varta Mignon 7244	>20,000 hours.

Crystal
f = 32.768 kHz, type Motorola C-L 17

Coil Data
L = 6 mH

Siferrite pot core	⌀ 9 × 5, N 30 without air gap, type B 65517–A 0000–R 030, n = 50 turns with 0,1 enamelled copper wire
or	⌀ 5.8 × 3.3 N 30 without air gap, type B 65501–J 0001–Y 030, n = 63 turns with 0.1 enamelled copper wire

*) Both IC's are available from Solid State Scientific Corp., Pasadena, California, USA.

List of Capacitors used in the Circuit 4.1.

1 pc	Styroflex capacitor	10 pF	B 31310–A 5100–F 000
1 pc	Styroflex capacitor	22 pF	B 31310–A 5320–F 000
1 pc	Electrolytic capacitor	1.5 μF/25 V	B 45181–B 3155–M 000

4.2. Miniature Light Barrier for a Shaft Position Encoder or a Revolution Counter

Miniature light barriers are required for shaft position encoders, since light transmitter and receiver are closely facing each other by a distance of a few millimeters. For this application a practical combination is achieved by using the light emitting diode LD 261 and the phototransistor BPX 81. Both components have the same epoxy case with an edge length of 2.2 mm. The LED operates in the infrared range at about 950 nm, since the efficiency is essentially higher than that of the visible radiation. The circuit described in the following converts interruptions of a light beam into electrical pulses for counting.

The construction of a shaft position encoder is shown in **Fig. 4.2.** The distance between the transmitting and the receiving components is about 3 to 5 mm. Both are inserted in a hole with a diameter of 3 mm, whereby the opening is diminished to 1.4 mm at its front ends. A plastic disc carrying a line pattern at its circumference as shown in **Fig. 4.2.1.** is rotating between transmitter and receiver. A pervious section follows a non-pervious one and the angle position of the disc is determined by counting the quantity of sections having passed.
Assuming that the rotating disc with a diameter of about 50 mm has a pattern of 600 lines, the distance between two lines is about 0.25 mm. To increase the light-to-dark ratio at the receivers side a plate with the same grid structure is mounted in front of the transmitter-hole as shown in **Fig. 4.2.2.** If the position of the grid on the rotating disc coincides with the one of the plate, the phototransistor receives a maximum of light. If both grid pattern are displaced with half the distance of two lines, the received light becomes a minimum. As the transmitter is rotable and adjustable in its position an efficiency maximum can be achieved.

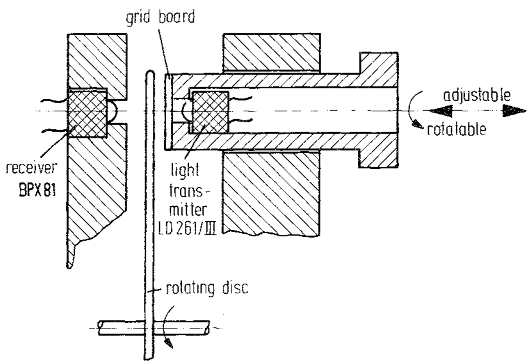


Fig. 4.2.

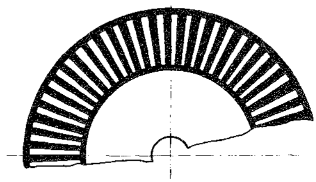


Fig. 4.2.1.

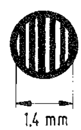


Fig. 4.2.2.

The circuit is shown in **Fig. 4.2.3.** The emitting diode LD 261 is operated at a current of about 20 mA.
The collector current of the potentiometer varies between about 3 µA (minimum) and about 12 µA (maximum) when the disc is rotating. Since the minimum value is to be kept constant strong ambient light influences have to be eliminated.
The current variation is sufficient to safely trigger the op amp TAA 861, which serves as a Schmitt-trigger. The following NAND-gates (FLH 101) operating as monostable multivibrator produce a definite square pulse with a duration of about 10 µs, for each line passing the light barrier. The circuit operates up to a frequency of 40 kHz, which corresponds to about 4000 r.p.m. of the disc.

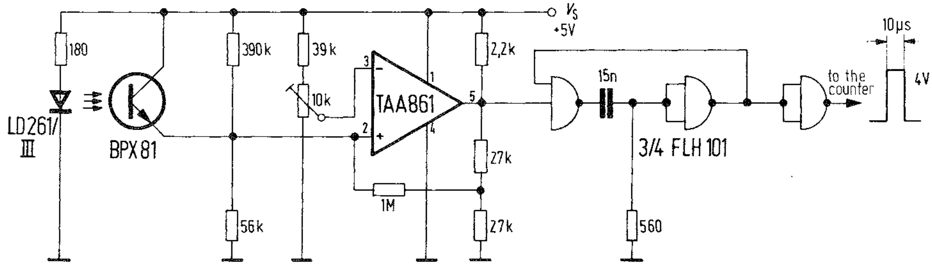


Fig. 4.2.3.

Technical Data:

Supply voltage V_s	5 V
Supply current (total) I_s	35 mA
Wave-length of the transmitted light	950 nm
Maximum counting frequency	40 kHz
Duration of the output pulses	10 µs
Amplitude of the output pulses	4 V

List of Capacitors used in Circuit 4.2.

1 pc	MKM-stacked film capacitor	15 nF	B 32540–A 3153–J

4.3. Light Barrier using TCA 105

The light barrier shown in **Fig. 4.3.1.** consists of the GaAs light-emitting diode LD 261, the phototransistor BPX 81 and the integrated threshold switch TCA 105. The LED is operated at a constant current to meet the total range of the power supply voltage being between 4.5 V and 27 V. The IC itself is specified for a wider range. The constant current source is realized by the transistor T_1, the diodes D_1 and D_2 as well as the two resistors R_1 and R_2. By the two diodes an independent, nearly constant voltage is achieved at the base of T_1. The constant current of the transistor can be adjusted by the potentiometer R_2.

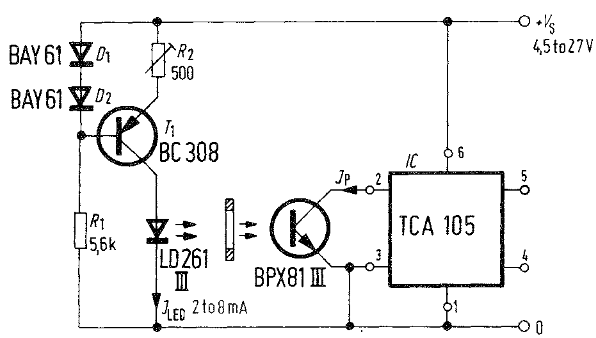


Fig. 4.3.1.

Parameter changes of the components created by temperature and ageing effects are compensated for if the photocurrent of the phototransistor is chosen four times higher than the required input threshold current of the TCA 105, i.e. about 200 µA. The output signal is available at the two anti-valent outputs of the IC (pins 4 and 5).

Adjustment

The light barrier is adjusted by setting the LED-current. If the IC is operated in the test circuit as shown in **Fig. 4.3.2.** the current of the LED has to be set in such a way that a voltage of 400 mV is available between pins 1 and 2 of the TCA 105.

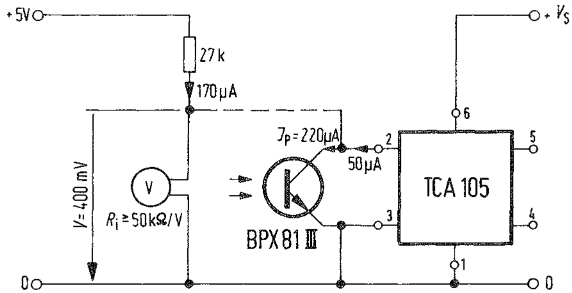


Fig. 4.3.2.

Technical Data

Supply voltage	4.5 to 27 V
Supply current	3.5 to 11.3 mA
LED current	2 to 8 mA
Supply current of the IC	3.3 mA
Ambient temperature range	−25 to +70 °C.

4.4. Optical Weight-Quantizer for Large Scales

The optoelectronic circuit described in **Fig. 4.4.** facilitates the weight quantization of large scales, whereby a 3-stage LED-display indicates the difference of the adjustment.

The incandescent lamp Gl_1 illuminates the two photodiodes PD_1 and PD_2. The first is covered by a slot diaphragm, which is moved up and down by the balance arm of the scale with a stroke of 4.5 mm, corresponding to the balance difference. A voltage, being proportional to the balance difference, drops across the resistor R_1 and is supplied to the three op amps TCA 335 operating as threshold switches. The reference voltages V_1, V_2 and V_3 are produced by the photocurrent of the photodiode PD_2 and drop across the resistors R_2, R_3 and R_4. They are supplied to the non-inverted inputs of the TCA 335. If the voltage across the resistor R_1 exceeds the reference value then the corresponding LED's LD_1, LD_2, LD_3 are switched on. An inverse function can be achieved by interchanging inputs 2 and 3 of the op amps. Since both photodiodes are illuminated by the same incandescent lamp, brightness changes created by aging or supply voltage variations are ineffective.

The common mode voltage, necessary for operating the op amps drops across the diodes D_1, D_2 and D_3.

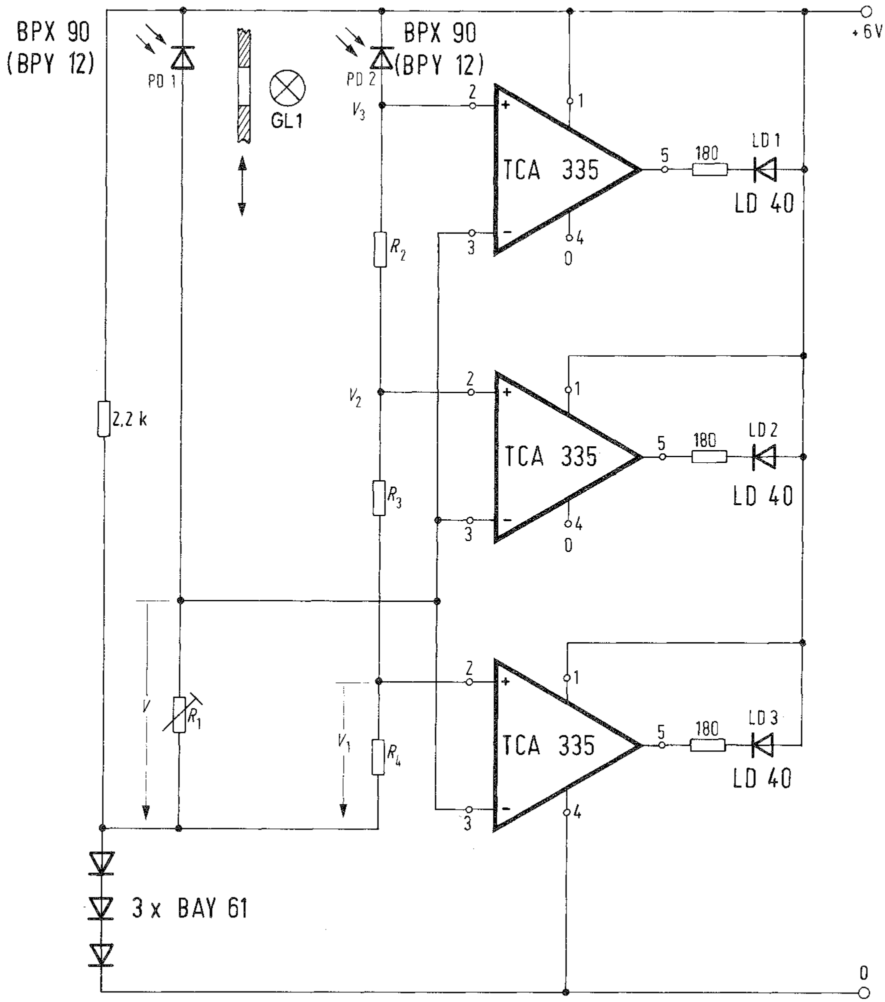


Fig. 4.4.

4.5. Optically Code Reading Regardless as to whether Different Kinds of Papers have Different Reflexion Coefficients

When identifying stroke markings placed on different kinds of papers the uncertainty exists that the code is erroneously read due to different reflexion coefficients.
The circuit described in the following and shown in **Fig. 4.5.** avoids this difficulty by means of an additional compensation track. The two phototransistors FT_1 and FT_2 being connected in series serve as a voltage divider, the centre tap of which is joint to the inverted input of the amplifier OP. To each phototransistor belongs a LED.

Both are connected in parallel, whereby the pair consisting of Le_1 and FT_1 serves for the compensation track and the one incorporating Le_2 and FT_2 functions for the reading track. Therefore the influence of a reflexion coefficient of the paper is eliminated and the reading result is determined only by the different reflexion of the strokes.

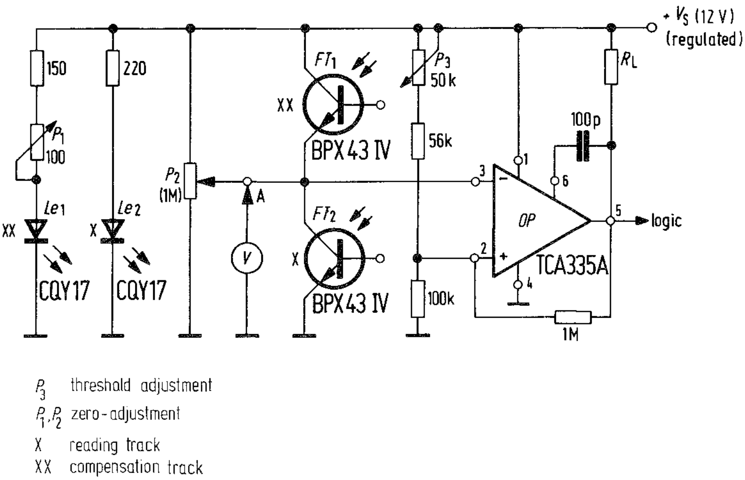


Fig. 4.5.

Adjustment procedure

Firstly the potentiometer P_2 is adjusted that a level of $0.5 \times V_s$ is measured at point A. During this procedure the phototransistors have to be completely covered. Then a paper of any kind without stroke markings is inserted into to the readchannel and P_1 is adjusted in such a way that point A has a level of $0.5 \times V_s$. The threshold for the stroke markings is determined by the potentiometer P_3.

List of Capacitors used in the Circuit 4.5.

1 pc styroflex capacitor 100 pF/160 V B 31310–A 1101–H 000.

4.6. Highly Sensitive Threshold Switch for Optoelectronic Applications

With the aid of the TPV 63-circuit, manufactured by a multichip procedure, switching operations can easily be achieved even if the currents are in the nA-range. Application examples are as follows, light barriers, twilight switches, automatic exposure timers and devices for optical character reading.
The TPV 63 incorporates a Darlington amplifier with an unusually low input current consumption of 20 pA. Its output can be relatively high loaded (e.g. currents of 70 mA at 10 V). In conjunction with a photodiode it is favoured for general purpose applications especially for a low-cost optoelectronic threshold switch.

4.6.1. Highly Sensitive Circuit

Fig. 4.6.1. shows the circuit of a simple but highly sensitive twilight switch. If the photodiode D_1 is exposed to light, a photo current flows and resistor R_1 produces a voltage drop. As soon as this voltage drop is sufficiently high (approx. 0.7 V) the voltage at the non-inverting input (a) of the operational amplifier becomes more positive than the voltage at the other input (b) and the output voltage is then at maximum. The output current of the operational amplifier drops to zero. The transition range between the two conditions of the operational amplifier is approx. 0.5 mV as a function of the input.

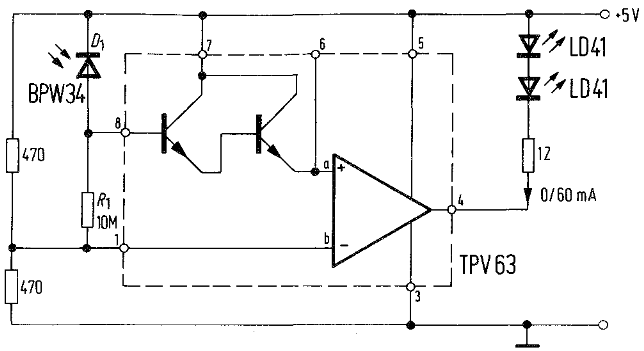


Fig. 4.6.1.

The threshold is adjustable by the resistor R_1. If a photodiode type BPW 34, offering a sensitivity of about 50 nA/Lux and a resistance of $R_1 = 10\ M\Omega$ are chosen, the circuit will already react at illuminations between 1 and 2 Lux. The maximum switching frequency being determined by the resistor R_1 and the parallel self-capacitance of the photodiode (approx. 40 pF) is about 800 Hz.
Small incandescent lamps (e.g. Osram type 3641 with 3.8 V and 70 mA) or LED's as shown in **Fig. 4.6.1.** can be driven by the opamp if its maximum output current of $I_q = 70$ mA is not exceeded. For larger loads a relay or an amplifier have to be used in addition.

Technical data

Supply voltage	5 V
Output current (without any illumination)	70 mA max.
Output current (acting operation)	$<7\ \mu A$
Supply current (acting operation)	6 mA
Threshold of the illumination at a colour temperature of 2850 K	1 to 2 Lux at $R_1 = 10\ M\Omega$
Switching frequency	800 Hz max.
Residual voltage at the output (pin 4) if the op amp is switched through ($I_q = 70$ mA)	<1 V

4.6.2. Circuit with Frequency Response Compensation

In case a higher switching frequency is required, the self-capacitance of the photodiode D_1 has to be compensated by a positive feedback. For this application a circuit as shown in **Fig. 4.6.2.** will suit.

In this layout the emitter output 6 of the darlington stage is connected to the photodiode D_1 by a coupling capacitor C_1. As the darlington amplifier does not reverse the signal, and the voltage again is about 1, the voltage at the photodiode D_1 is practically unaffected by a signal current variation. The photodiode capacitance is ineffective because of no-load current in the network and the cutoff frequency is increased by a factor of 10.

Though this condition is a feedback, there is no danger of self excitation due to the signal being taken from the emitter terminal of the darlington amplifier, its voltage gain is somewhat less than 1. The positive end of diode D_1 is connected to the supply voltage terminal via a resistor to take effect of the positive feedback via C_1.

A switching frequency of about 50 KHz is achieved at a threshold of about 10 Lux and at $R_1 = 1\ \mathrm{M\Omega}$.

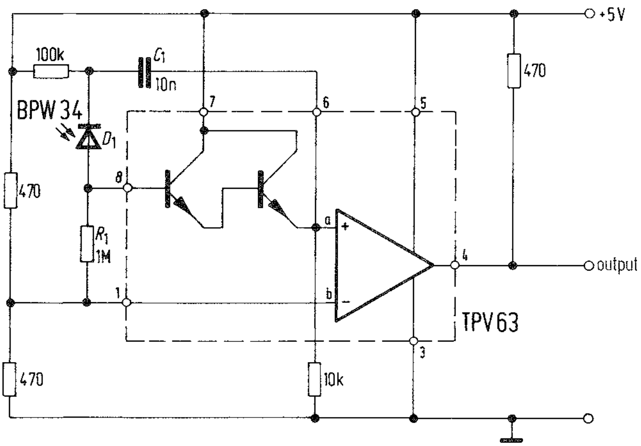


Fig. 4.6.2.

List of Capacitors used in the Circuit 4.6.2.

1 pc	MKM-stacked film capacitor	10 nF	B 32504–A 3103–J

4.7. Optical Combustion Control and Fire Protection Circuit

In conjunction with the phototransistor BPY 62 the op amp TCA 335 is used for designing two intrinsically safe circuits for optical combustion control of oil and gas burners. The circuits react accordingly to the characteristic flame flicker and the constant component of the ambient light does not result in an alarm signal release. Applying slight modifications both circuits can be used for fire protection devices.

The phototransistor BPY 62 (**cf. Fig. 4.7.1.**) detects the characteristic brightness changes of a flame (f = 1 to 20 Hz) and converts them into voltage signals via the resistor R_1. The function of this circuit is achieved by suppressing dc components as well as the influence of alternating light with a higher frequency, probably generated by fluorescent lamps (100 Hz). Therefore a band-pass filter is necessarily required.

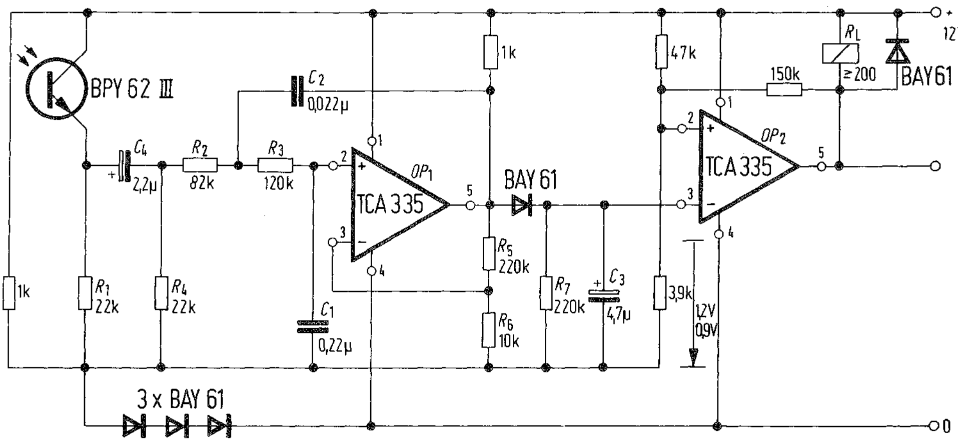


Fig. 4.7.1.

The active band-pass filter of 2nd-order in conjunction with the OP 1 (TCA 335) amplifies the signal by a factor of 20 and reduces the voltage by 12 dB per octave at frequencies higher than 20 Hz. Slow going dc voltage changes are suppressed by a simple high-pass filter (C_4 and R_4) connected to the emitter of the BPY 62.

To obtain a definite output signal a rectifier circuit and a Schmitt-trigger follow after the filter. An additional selection is achieved by applying the reference voltage to the non-inverting input of the OP 2. The employed hysteresis voltage avoids a continuous switching of the op amp which could be produced by a dc voltage superimposed by a strong ripple with a frequency of less than 10 Hz. No great elaborateness is applied to the smoothing to keep the dead time of the total system low. When the flame extinguishes the relay drops out and the supply of gas or oil is interrupted. The circuit operates intrinsically safe, i.e. the relay is switched off, when either the sensor circuit is short-circuited or interrupted or a breakdown happens with the power-supply.

If no dangerous influence of external alternating light is expected, the active low-pass filter 2nd-order can be replaced by one of 1st-order (R_2 and C_2) as shown in **Fig. 4.7.2.** Thus an essentially higher sensitivity of the circuit can be attained. The gain of the ac voltage amplifier is adjusted by the resistor R_3 and corresponds to the following voltage divider ratio

$$\frac{R_2 + R_3}{R_3}$$

The rectification and the signal processing are the same as described for circuit 4.7.1. The threshold switch reacts at a level of 1.6 V and turns off at 1.25 V.

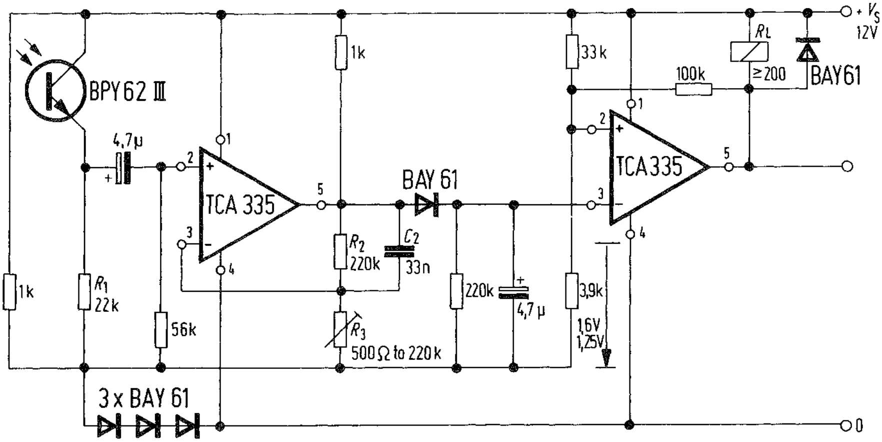


Fig. 4.7.2.

Technical Data:

Circuit	Fig. 4.7.1.	Fig. 4.7.2.
Supply voltage	12 V	12 V
Supply current (without load)	20 mA	20 mA
Max. output current	70 mA	70 mA
Alternating light selectivity	1 to 28 Hz	1 to 25 Hz
AC voltage gain	23	$\frac{R_2 + R_3}{R_3}$
Change of light intensity	5 Lx	0,25 to 50 Lx

An intrinsically-safe fire protection device is responsible for turning off the relay when a flame exists.
In this case the function of the described threshold switch is inverted. The modified circuit is shown in **Fig. 4.7.3.**
The hysteresis is produced by the transistor BCX 78.
The same circuits as shown in Fig. 4.7.1. or 4.7.2. can also serve as drivers.

List of Capacitors used in the Circuit 4.7.1.

C_4	1	Electrolytic 2.2 µF/63 V	B 41316–A 8225–Z 000
C_3	1	Electrolytic 4.7 µF/63 V	B 41316–A 8475–Z 000
C_1	1	MKM 0.22 µ/100 V	B 32540–A 1224–J 000
C_2	1	MKM 0.022 µ/250 V	B 32540–A 3223–J 000

List of Capacitors used in the Circuit 4.7.2.

	2	Electrolytics 4.7 µF/63 V	B 41316–A 8475–Z 000
	1	MKM 0.033 µ/250 V	B 32540–A 1334–J 000

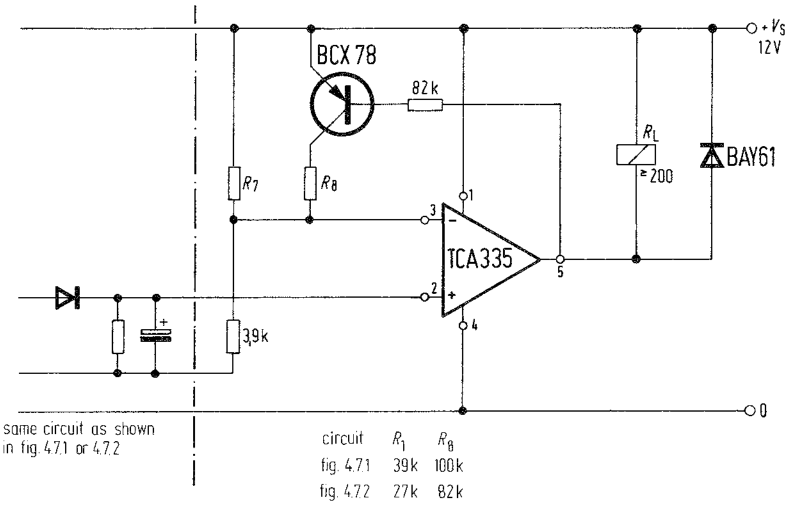


Fig. 4.7.3.

4.8. Optoelectronic Coupler CNY 17 used as a Photothyristor

As shown in **Fig. 4.8.1.** a photothyristor can be simulated by using a combination of the transistor BCX 78 and the phototransistor system of the CNY 17. To avoid an "over-head"-firing an RC-circuit has to be connected in parallel to the base-emitter junction of the BCY 78 and of the phototransistor. When the LED of the CNY 17 has reached a certain light intensity the thyristor cir,cuit triggers. It is switched on as long as the supply voltage V_s is not turned off.

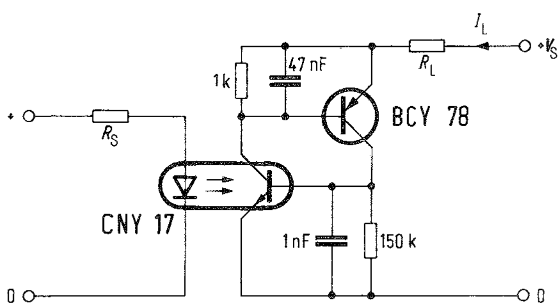


Fig. 4.8.1.

The light intensity of the LED incorporated in the CNY 17 depends on the current flowing through the diode. Therefore the circuit reacts at a certain current, which can be determined by different shunt resistors.

Technical Data:

Max. load current $I_{L\,max.}$		50 mA
Residual voltage at $I_L = 50$ mA		1.2 V
Max. supply voltage		30 V
Starting current of the LED	≈	10 mA
Turn-on time	≈	300 µs
Turn-off time	<	80 µs

List of Capacitors used in the Circuit 4.8.1.

1 pc	MKM-stacked-film capacitor	1 µF/250 V	B 32540–A 3102–K 000
1 pc	MKM-stacked-film capacitor	0.047 µF/250 V	B 32540–A 3473–J 000

4.9. Suppression of DC Component in Photocurrent of Phototransistors

In many applications phototransistors are intended to transmit only intensity-modulated light signals. Non-modulated light intensity interferes; the dc component caused by it must be suppressed.
Two circuits are described here in which the dc component remains ineffective. In the first circuit the direct current is kept constant through an automatic control system, in the second an active, frequency-dependent external resistance is used which is much smaller at low frequencies than at high ones.
Phototransistors are particularly suitable as light detectors for many applications since they are economical and, due to their amplification, offer a larger output signal than photo-diodes. Thus they are less sensitive to external interferences.
In optoelectronics a number of applications are used in which an intensity-modulated signal is superimposed upon a non-modulated one, e.g. in optical flame control, in light barriers involving moving objects, and in computerized flashlight equipment as well as slave flashlight equipment in which the primary illumination can cause interference. In many instances the suppression of the dc component is required because of the danger of overdriving through unmodulated light intensity.
Using phototransistors the dc component of the photocurrent cannot be suppressed by a coupling capacitor.

Circuit for phototransistors with base terminal

In **Fig. 4.9.1.** phototransistor T_1 and transistor T_2 form an automatic control system which regulates the voltage drop at resistor R_1, maintaining it at a constant value, independent of the unmodulated light intensity at phototransistor T_1. When the light intensity rises, a larger photocurrent I_p flows through T_1, and the voltage drop at resistor R_1 becomes greater. As a result a larger current flows to the base of T_2. The rising collector current of T_2 keeps reducing the primary photocurrent of T_2 until the voltage drop at resistor R_1 reaches its original value.
Due to the by-passing of the base-emitter junction of T_2 by capacitor C_1, this control mechanism is ineffective during rapid changes. The cut-off frequency above which the control becomes ineffective is determined by capacitor C_1 and resistor R_2.

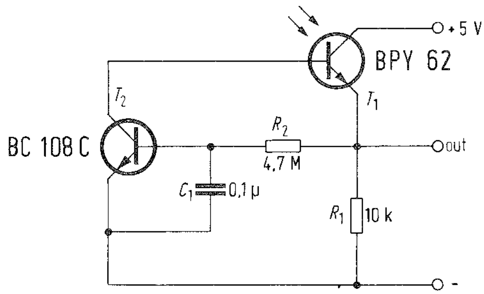


Fig. 4.9.1.

Resistor R_1 determines the quiescent current. R_2 should be as large as possible to permit small values for C_1. However, when resistance of R_2 becomes too large, the drive of T_2 is too weak. As a result the maximum light intensity at which the control still works is reduced. The maximum light intensity is also limited by the power supply voltage, because the voltage drop at R_1 must not exceed a fixed maximum value.

For the dimensioning given in **Fig. 4.9.1.** the maximum light intensity can be 25,000 lx; the voltage drop at R_1 must not exceed the value $V_{R1} = 4$ V. The photosensitivity of phototransistor BPY 62 is 2 mA/1000 lx. The dark current of the circuit is smaller than the dark current I_{CEO} of the simple phototransistor, because part of the dark current is split as residual current from T_2. The lower cut-off frequency of the circuit in the above dimensioning is $f_{gu} = 16$ Hz, the upper frequency $f_{go} = 2.5$ kHz. If an increase in the upper cut-off frequency f_{go} is required, resistance of R_1 must become smaller.

To exclude interference signals, the connection between the collector of T_2 and the base of phototransistor T_1 must be held as short as possible.

Circuit for phototransistors without base connection

The circuit shown in **Fig. 4.9.2.** is intended for phototransistors without base connection. At low frequencies the base voltage of transistor T_2 remains constant, and is determined by the voltage divider of resistors R_1 and R_2. The collector resistance of phototransistor T_1 is determined by the relatively low diffusion resistance of the base-emitter junction of transistor T_2. A large collector current can flow without resulting in a substantial decrease of the collector voltage of phototransistor T_1. For the diffusion resistance it applies that

$$R_D = \frac{k \times T}{e \times I},$$

k standing for Boltzmann constant (1.38×10^{-23} WsK^{-1});
T for absolute temperature of phototransistor T_1, in Kelvin;
e for elementary charge (1.6×10^{-19} As); and I for emitter current of transistor T_2 in Ampere.

At high frequencies the base-emitter junction is short-circuited by capacitor C_1. As a result the considerably larger differential resistance of the emitter-collector junction of transistors T_2 functions as external resistance. Parallel to it there is the series circuit consisting of capacitor C_1 and the resistors R_1 and R_2, parallel-connected through the power supply. In the circuit presented in **Fig. 4.9.2.**, the maximum light intensity for the given dimensions can amount to 20,000 lx.

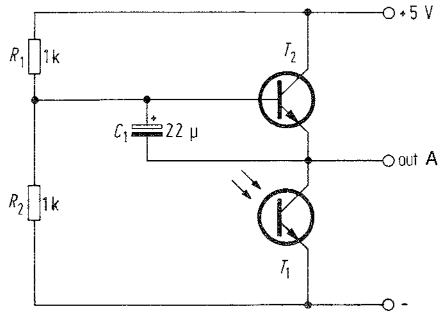


Fig. 4.9.2.

The sensitivity of photo-transistor BPX 81, used in the experimental circuit, is 2.5 mA/1000 lx. The lower cut-off frequency is $f_{gu} = 80$ Hz, the upper frequency is $f_{go} = 40$ kHz. The ac voltage at point A can be raised by increasing the resistance of R_1 and R_2. For a maximum light intensity of 20,000 lx, resistances of up to 10 kΩ are permissible.

List of Capacitors used in the Circuit 4.9.1.

1 pc	Ceramic Capacitor	0.1 µF/63 V	B 37449–A 6104–S 001

List of Capacitors used in the Circuit 4.9.2.

1 pc	Electrolytic Capacitor	22 µF/40 V	B 41286–A 7226–T 000

4.10. Phototransistor used in a Computerized Photoflash Unit

A new circuit has been designed for the receiving part of the computerized photoflash unit. It offers the advantage in that it essentially compensates all the undesired influences produced by exposure time errors, ambient light, temperature, and tolerances of the photosensitivity. A phototransistor in conjunction with an integrating capacitor connected to the emitter serves as photodetector.

A computerized photoflash unit differs from a standard one in that the duration of the photoflash is determined by a photodetector. Therefore the exposure time for a camera film is constant and does not depend on the intensity of the reflected light, i.e. the flash is interrupted sooner or later in dependence on the quantity of reflected light. **Fig. 4.10.1.** shows on principle the control circuit of a computerized photoflash unit. The photocurrent of the phototransistor charges the capacitor C_1 and thus the turn-off thyristor shown in the figure with broken lines is triggered.

A trial was conducted to find out how far exposure time errors of photoflash devices using the circuit of Fig. 4.10.1. depend on the sensitivity of the phototransistor. It has been experienced that the sensitivity changes by about 25% in a distance between 0.9 m to 4.0 m. This variation is generated through the change of the current gain depending on the collector current.

The compensation of the linearity error of a phototransistor is only partially possible because of its unavoidable characteristic tolerance. Therefore it is more convenient to use a circuit in which the value of the current gain does not essentially influence the exposure time of a computerized photoflash unit.

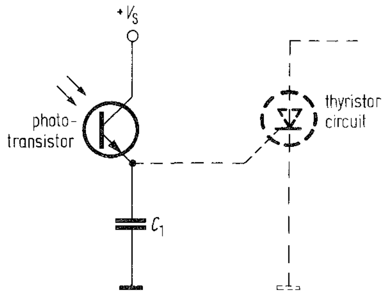


Fig. 4.10.1.

The base collector current dependence on the luminous intensity is completely linear whereas this is contrary to the one of the emitter collector current. This is founded in the fact that the base-collector-junction serves as a photodiode. Therefore a special circuit has been designed. The current generated through the light is integrated by a capacitance not being connected to the emitter of the phototransistor but to its base as shown in **Fig. 4.10.2.** At the beginning of the exposure the capacitor is not charged, i.e. the base-emitter-junction is not conductive. If the phototransistor is illuminated charge carriers are generated. A hole moves to the base terminal and positively charges the capacitor C_1 with reference to ground potential. When the capacitor is charged so that the base-collector-junction becomes conductive, the phototransistor starts to amplify, i.e. the emitter current increases. The amplified photocurrent produces a voltage drop across the load resistor R_2 and thus the following turn-off thyristor is triggered.

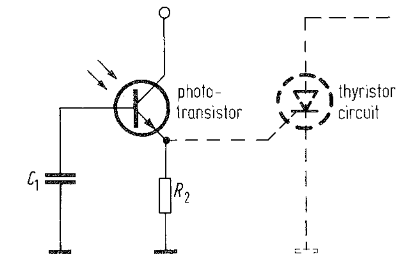


Fig. 4.10.2.

The disadvantage of the circuit shown in **Fig. 4.10.2.** is that the signal slewing rate is not fast enough, because the capacitance of the integrating capacitor C_1 is increased by the gain of the phototransistor at that instant when the base-emitter-junction becomes conductive, i.e. when there is an amplification effect. In order to improve the signal slewing rate the circuit shown in **Fig. 4.10.3.** is recommended. Here the capacitor C_1 is connected to base and emitter. If the voltage across the load resistor R_4 increases the level at the capacitors low end also rises with nearly the same amount as at the high end of C_1 connected to the base. Therefore, the capacitor C_1 usually requires no charge. The circuit according to Fig. 4.10.3. assures that at the beginning of each photoflash the capacitor C_1 always has the same charge impedance of the illumination which previously occurred. The resistors R_2 and R_3 serve as voltage divider, at which a positive voltage of 1 V referred to the level of the phototransistor

emitter is disposible before the photoflash is started. The diode D_1 is turned off. Its voltage difference effects that a current flows via the resistor R_1 into the base of the phototransistor. At its base-emitter-junctions a voltage drop, not being essentially increased by the external illumination is produced. At the beginning of the photoflash a negative pulse is applied via terminal B to the resistor R_2. By the current flowing through R_2 the diode D_1 becomes conductive and its level changes from $+1$ V to -0.7 V. This potential difference is fully transmitted via the integrating capacitor C_1 to the base of the phototransistor, which is therefore reversely biased by this voltage. Thereafter this bias is compensated by the photocurrent. The negative voltage pulse required at the beginning of the photoflash can be derived from the same voltage source, which generates the collector-emitter-voltage at the beginning of the photoflashing. The voltage at terminal A is taken from a divider being in parallel to the photoflash capacitor, i.e. it is also available before the photoflashing occurs.
The advantageous features of the circuit according to Fig. 4.10.3. compared to the one of a conventionally computerized photoflash unit are as follows:

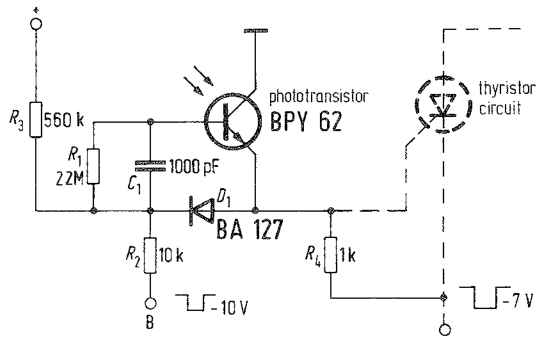


Fig. 4.10.3.

a) Exposure time failures are nearly not detectable – presuming an objective lux meter (<5%).
b) The phototransistors must not be selected according to their photosensitivity since their base-collector-junction is utilized and there is no difference in sensitivity amongst the phototransistors.
c) No neutral absorber is required, since the internal base-collector-diode of the phototransistor operates linearly. Therefore the photodetector is able to receive more light, i.e. signals with a higher amplitude are produced and the operation is trouble-free. The gate current of the thyristor does not influence the exposure time control. The total temperature coefficient is low (about 0.3% K^{-1}). If necessary the TC can be additionally decreased by applying at terminal B a pulse with a higher amplitude.
   The charging of the integrating capacitor is extremely low when the supply voltage is suddenly applied to the phototransistor.

4.11. Thyristorized Computer Photoflash Unit

Computerized photoflash units offer the feature that the duration of the photoflash is controlled by the quantity of the reflected light. Adequate devices have already been described in Design Examples of Semiconductor Circuits, edition 1974/75.

Fig. 4.11.1. shows a circuit, which, however, differs from the previously mentioned ones.
The energy required for the photoflash electrolytic capacitor of 1000 μF at 350 V is produced by a dc voltage converter using a transistor AD 136 V. A control circuit is employed avoiding that the voltage across the photoflash electrolytic capacitor does not exceed a level of 360 V. The transistor AC 152 VI is driven by the control amplifier transistor BC 238 B, being connected in parallel to the base of the converter transistor AD 136 V. If an operating voltage of 340 V has reached a glow lamp indicates that the unit is ready for operation. This pilot lamp is adjustable by the 100 kΩ-potentiometer. The current flowing through the glow lamp produces a voltage drop at the 10 kΩ-potentiometer. A part of this drop being adjustable drives the control circuit, which turns on the transistor AC 152 VI and very much reduces the amplitude of the converter oscillation. Therefore only the energy required to compensate for the losses is generated. In this case the operating voltage is constant at a level of about 350 V ± 10 V.
The voltage drop produced by the current through the glow lamp also triggers the photoflash. Before the voltage across the photoflash capacitor has reached a level of 340 V the glow lamp does not fire and it is impossible to release a photoflash.
Thus erroneous exposures are avoided at longer photoflash durations. At very short photoflashes the automatic control can still react.
The small thyristor BRY 55/300 connected to the triggering push button and respectively to the X-contact guarantees a chatter-free pulse being supplied to the triggering transformer and the switching thyristor. Besides that the X-contacts of the camera are protected since the BRY 55/300 requires a very low gate triggering current.
The 22 μF electrolytic capacitor connected in parallel to the 27 kΩ-load-resistor of the turn-off capacitor (6.8 μF) is responsible for reducing the current peak flowing through the flash tube after the turn off. In some cases this peak creates exposure time errors at photoflashes with extremely short durations. The thyristor Th_4 type BRY 55/200 supplies a sufficiently steep pulse to the gate of the turn-off thyristor. By means of the potentiometer connected to the transistor BCY 78 VIII the turn-off time can be adjusted depending on the aperture and the film material used. Therefore it is not necessary to select the thyristor Th_4 to get a definite turn-off beginning.
The 0.1 Ω-resistor connected to the cathode of the switching thyristor Th_2 is responsible for generating the supply voltage for the computer device. The negative gate voltage required essentially for turning off the switching thyristor is produced by the series circuit of a 0.15 μF-capacitor and a 470 Ω-resistor. This negative pulse generated at the anode of the turn-off thyristor Th_3 at the turn-off beginning is applied to the control gate of the switching thyristor Th_2.

List of important Components used in the above Circuit

a) Switching thyristor	BSt E 0333 T
b) Turn-off thyristor	BSt C 0733 T
c) Phototransistor	BPY 12 (can also be inversely poled)
d) Photoflash electrolytic capacitor	1000 μF/1360 V (B 43405–S 0108–Q 54)
e) Turn-off capacitor	6.8 μF
f) Fast recovery diodes	C 2605, BAY 44, B 2580 C
g) Small thyristors	BRY 55/200 and BRY 55/300

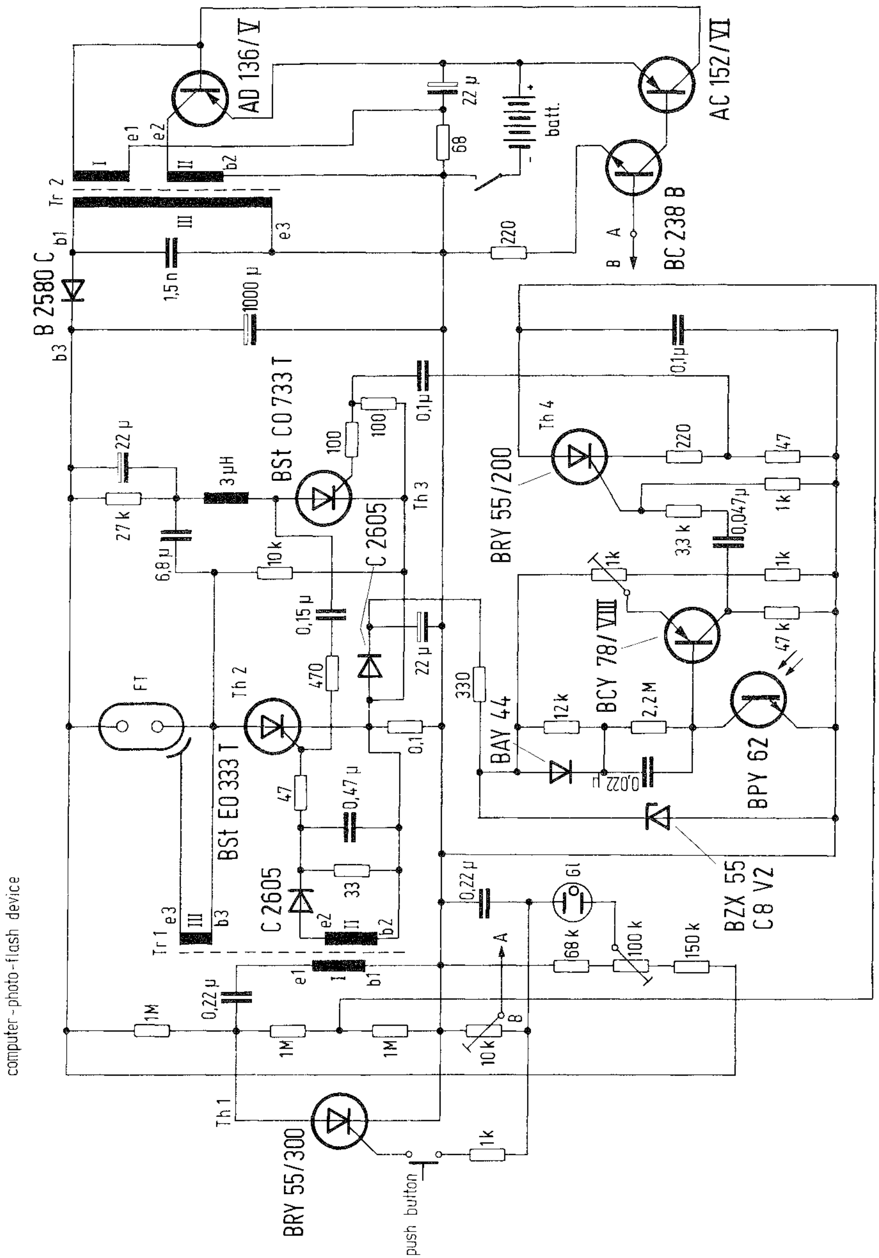

computer-photo-flash device
AD 136/V
AC 152/VII
BC 238 B
B 2580 C
Tr 2
batt.
BSt CO 733 T
BSt EO 333 T
BRY 55/200
BRY 55/300
BCY 78/VIII
BAY 44
BPY 62
BZX 55
C 8 V2
C 2605
Tr 1
Th 1
Th 2
Th 3
Th 4
FT
Gl
push button

Fig. 4.11.1.

List of Capacitors used in the Circuit 4.11.

Quantity	Component	Ordering codes
1	0.22 μF/400 V MKH	B 32231–C 6224–M
1	0.47 μF/100 V MKM	B 32540–A 1474–J
1	0.22 μF/100 V MKM	B 32540–A 1224–J
1	22 μF/25 V Electrolytic	B 41315–A 5226–Z
1	6.8 μF/350 V MKH Special	B 32231–A 3685
1	0.15 μF/250 V MKM	B 32541–A 3154–J
1	22 μF/350 V Electrolytic	B 43052–B 4226–T
1	1000 μF/350 V Electrolytic	B 43405–S 0108–Q 54
1	0.1 μF/100 V MKM	B 32540–A 1104–J
1	0.022 μF/250 V MKM	B 32540–A 3223–J
1	0.047 μF/250 V MKM	B 32540–A 3473–J
1	0.1 μF/250 V MKM	B 32540–A 3104–J
1	330 pF/1000 V Ceramic	B 37370–A 1330–S 3
1	22 μF/10 V Electrolytic	B 41315–A 3226–Z

4.12. Elverson Oscilloscope for Universal Applications

In general:

Flashlight stroboscopes, so-called Elverson oscilloscopes, can be driven either by AF-voltage peaks representing a music rhythm or by free running pulses. In the latter case the timing can be varied within wide ranges. Besides that an external synchronization is possible which is important for motion studies. By using several stroboscopes any light condition can be obtained when a sequence of motions has to be photographed.

The stroboscope can be triggered by either a break or a make contact, e.g. x-contact of a camera. The shock-proof trigger circuit described in the following is especially suited for physical test applications.

Finally the use of a flashlight stroboscope for automotive spark setting has to be mentioned.

Power Supply for the Flash Tube

Conventional flash tubes used in flash light stroboscopes require an anode supply voltage between 400 and 650 V, which can be produced without any transformer by applying a Villard rectifier circuit (voltage doubler). The generated voltage has to be less than the maximum admissible anode voltage. The capacitance of the capacitors can be calculated from the anode voltage V_L, the max. permissible power dissipation of the flash tube $P_{L\,max}$ and the maximum flashlight frequency f_{max} as follows:

$$C_2 \leqq \frac{2 \times P_{L\,max}}{V_L^2 \times f_{max}}$$

It is recommended to make $C_1 \approx C_2$, whereby attention has to be paid to the fact that pulse-proof capacitors with a sufficiently highly rated voltage are used (e.g. MP- or MKH-types).

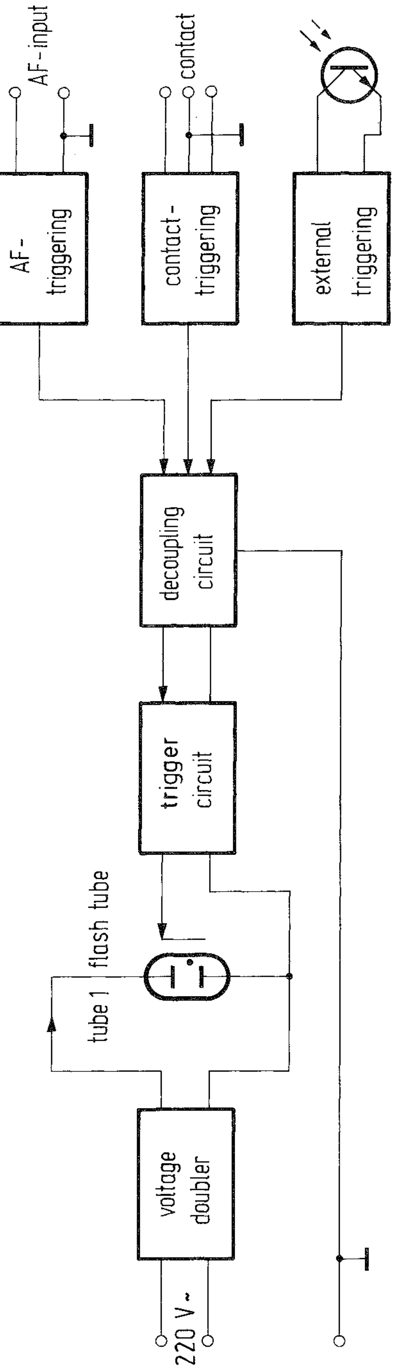

AF-input
contact
AF-triggering
contact-triggering
external triggering
decoupling circuit
trigger circuit
flash tube
tube 1
voltage doubler
220 V~

Fig. 4.12.1.

Triggering

Fig. 4.12.2. shows a suitable trigger circuit for a flash tube. The capacitor C_{3abc} is charged via the resistor R_3. When the thyristor triggers, C_3 is discharged via the primary winding of the triggering transformer Tr_1. By that a voltage pulse firing the flash tube is produced in the secondary coil. The turns ratio of the transformer is determined by the lowest required trigger voltage of the tube (some kV) and by the voltage available at C_3.
The thyristor Th_1 is triggered by the glow lamp (tube 2) at the level between 50 and 100 V when the voltage across C_3 has reached the trigger level of tube 2. Therefore the flash tube (1) is automatically triggered. The space between the pulses is determined by the following time constant:

$$T_3 = R_3 \times C_3$$

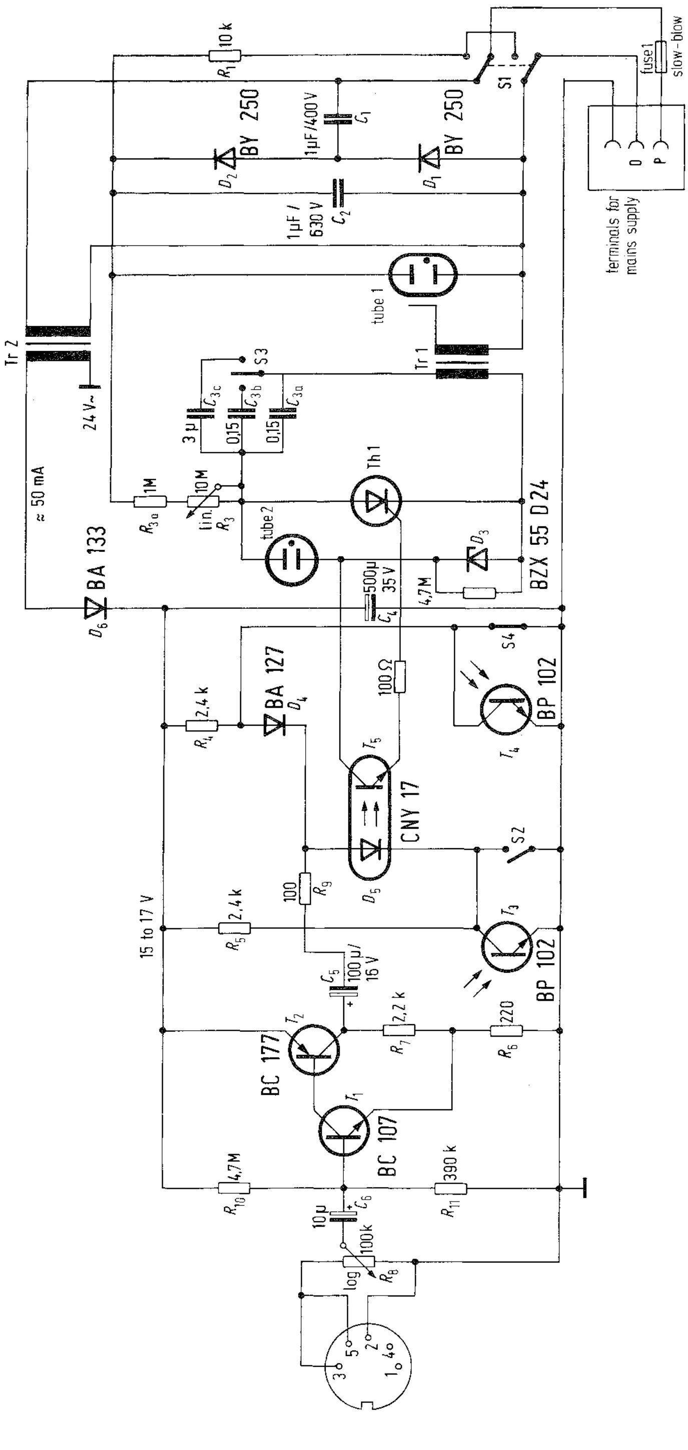
It can be varied step by step by means of the switch S_3 and continuously set by the potentiometer R_3.

Decoupling and Contact Control

As already mentioned a shock-proof start-stop-circuit is required for many applications. A solution for this coupling problem is offered by the optoelectronic coupler CNY 17, which contains an LED (D_5) and a phototransistor (T_5) in a 6-pin dual-in-line case. By that the gate circuit of the triggering thyristor Th_1 is isolated. If T_5 becomes conductive the tube 2 is switched on and the first photoflash is produced without delay. The sequence of photoflashes is immediately interrupted when the transistor T_5 is turned off. The z-diode D_3 guarantees that a maximum reverse voltage of T_5 is not exceeded.
The operation status of T_5 depends on the current flowing through D_5. It is controlled by two contacts, the making (S_2) and the breaking one (S_4). If S_2 is closed and S_4 is open a current flows through R_4 and T_5 is turned off, i.e. the photoflash is triggered. If, however, S_2 is open and S_4 is closed the diode D_5 is non-conductive and the sequence of photoflashes is interrupted. The contacts S_2 and S_4 are generally realized by a pair of jacks.
The mains isolated power supply voltage for the shock-proof start-stop circuit is produced by the transformer Tr_2 and the half-wave rectifier circuit consisting of D_6 and C_4.

External Triggering

For the operation "external triggering" the following conditions have to be guaranteed.
R_3 has to have a minimum resistance
S_3 has to be in its zero position, so that only C_{3a} determines the time. The capacitance of C_{3a} has to be lower than the one of C_3, selected at the "master" stroboscope to avoid suppression of some photoflashes. Additionally it has to be considered that an external operation with C_{3a} is not allowed in certain circumstances.
S_2 and S_4 have to be open.
R_8 has to be in its zero-position.



10 k
R1
S1
fuse1
slow-blow
terminals for mains supply
0
P
BY 250
D2
D1
1μF/400V
C1
1μF / 630 V
C2
tube 1
Tr 2
Tr 1
24 V~
S3
3 μ
C3c
0,15
C3b
C3a
≈ 50 mA
1M
R3a
10M
lin.
R3
Th1
tube 2
BZX 55 D24
D3
4,7M
500μ 35 V
C4
BA 133
D6
S4
100 Ω
BA 127
D4
2,4 k
R4
BP 102
T4
T5
CNY 17
D5
S2
100
R9
15 to 17 V
2,4 k
R5
T3
BP 102
C5
100 μ/ 16 V
2,2 k
R7
220
R6
BC 177
T2
T1
BC 107
4,7M
R10
390 k
R11
10μ
C6
100k
log
R8

Fig. 4.12.2.

AF-Triggering

If the stroboscope has to operate according to a music rhythm an AF-amplifier has to be additionally used (T_1 and T_2) as shown in **Fig. 4.12.2.** By that it is possible to drive the device at a low level of about $0.7V_{rms}$ which is available from e.g. head phones or from a monitoring output of a tape-recorder amplifier. The amplified AF-signal is disposable at the collector of T_2. R_9 protects the diode D_5 against too high peak current produced by the capacitor C_5. The sensitivity of the circuit is adjustable by the potentiometer R_8.
In this operation it has to be guaranteed that S_2 and S_4 are closed. The inertia of the stroboscope's reaction to the music rhythm can be adjusted by the potentiometer R_3. The diode D_4 being in anti-parallel connection to D_5 balances the load of the AF-amplifier.

List of Capacitors used in the Circuit 4.12.

Item	Quantity	Components	Ordering codes
1	1	1 μ/400 V MKH	B 32229–A 4105–M
2	1	1 μ/630 V MKH	B 32229–A 6105–M
3a	1	0.15 μ/100 V MKH	B 32110–E 0154–M
3b	1	0.15 μ/100 V MKH	B 32110–E 0154–M
3c	1	3.3 μ/250 V MKH	B 32110–E 2335–M
4	1	470 μ/40 V Electrolytic	B 41012–A 7477–T
5	1	100 μ/16 V Electrolytic	B 41286–A 4107–T
6	1	10 μ/63 V Electrolytic	B 41286–A 8106–T

4.13. Control Circuit for several LED-Arrays

The IC UAA 170 offers the feature that the first diode of a connected LED array emits light when the control voltage falls below the control range, and the last diode indicates that the control range is exceeded. If many LED-arrays are connected together it has to be guaranteed that only one diode emits light when a transition from one array to another occurs. This can easily be achieved by covering the first and last LED of each array.
Fig. 4.13.1. shows a circuit incorporating two UAA 170's. The light point moves from one IC to the other. Pin 9 of the UAA 170 is not connected to a LED but to a combination of a silicon diode BAY 61 and an npn-transistor BC 258 A. Both components are responsible for switching on the following IC and by that also the LED current of the following array is turned on. Every LED of the array should show a forward voltage drop, not exceeding a tolerance range of 500 mV. This also applies to the diode-transistor combination. In **Fig. 4.13.2.** a circuit being able to drive 100 LED's is shown. There are seven UAA 170's required. Two voltage dividers have to be used to achieve optimum conditions for the control and reference voltages of the IC ($\Delta V_{12/13} \approx 1.2$ V). The left divider belongs to the first four UAA 170's whereas the right one is for the last three IC's.

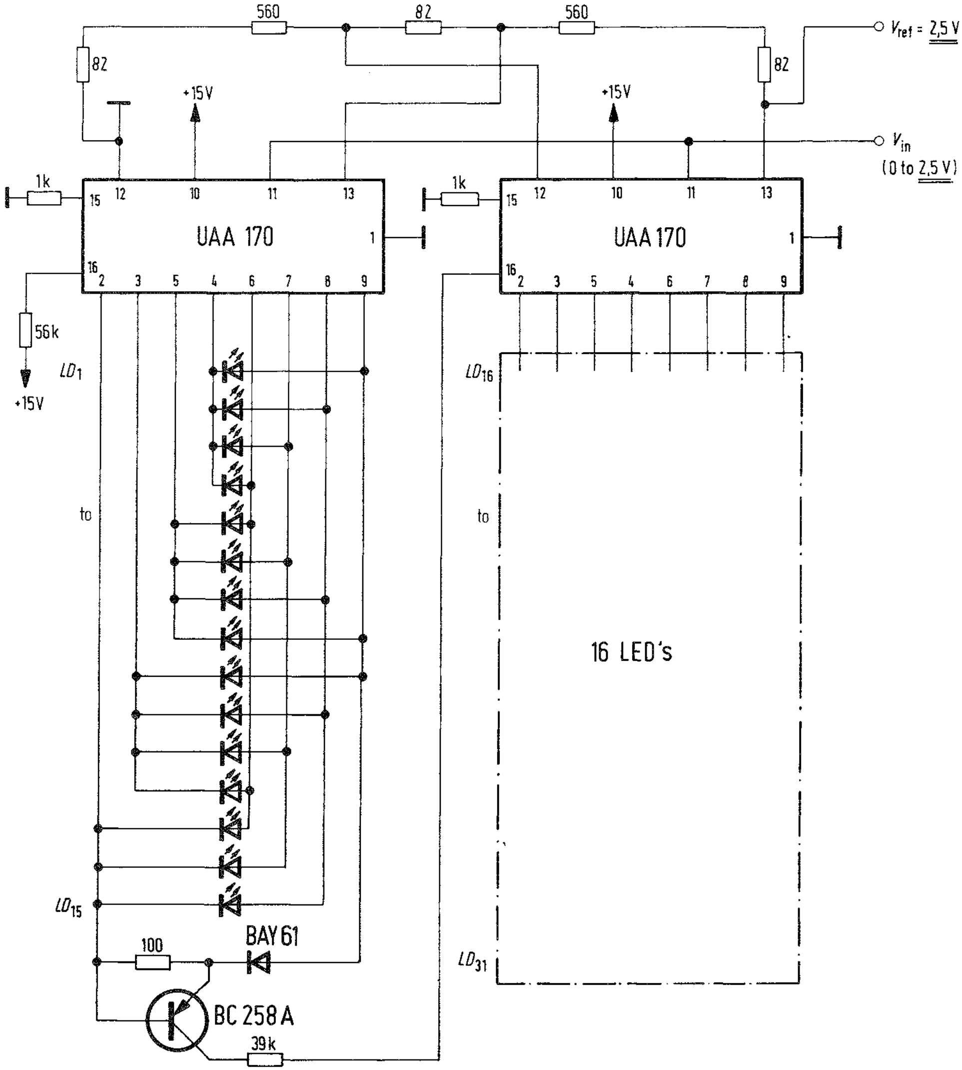
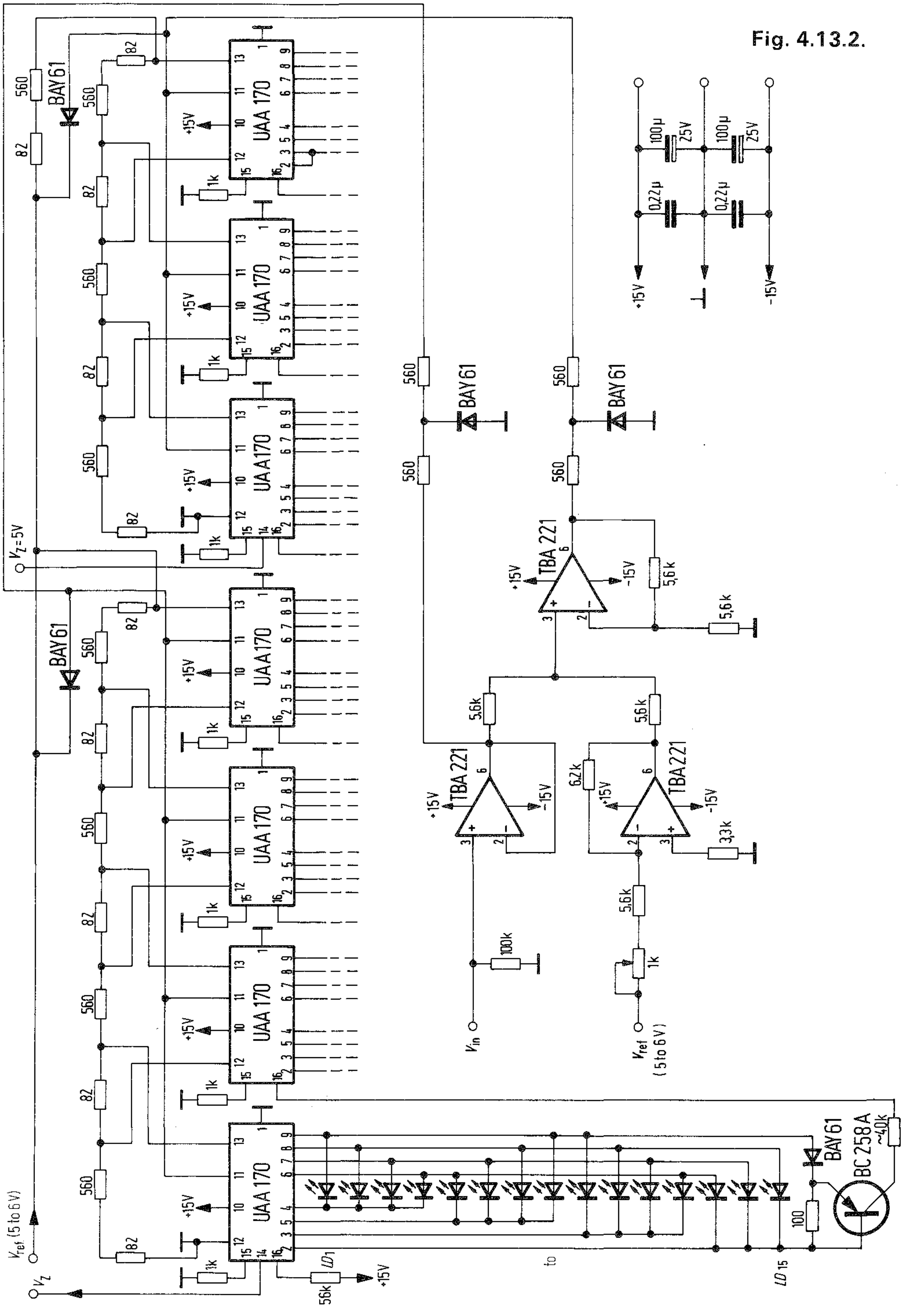


Fig. 4.13.1.

Since there are so many UAA 170's connected in cascade the input voltage range of 0 to 6 V is not sufficient. Therefore it is extended to 10 V by the aid of three op amps TBA 221 and applied to the two groups of IC's. Thus the admissible voltage of 6 V between V_{contr} and V_{ref} is not exceeded. The reference voltage is the same for each group of UAA 170 (5.87 V).
Through the application of the op amp follows:

$$V_{contr\,1} = V_{in}$$
$$V_{contr\,2} = V_{in} - V_{ref}$$

Fig. 4.13.2.



UAA 170
TBA 221
BAY61
BC 258A
~40k
+15V
-15V
$V_Z = 5V$
V_{ref} (5 to 6 V)
V_Z
V_{in}
LD_1
to
LD_{15}
560
82
1k
56k
100
100k
5,6k
6,2k
3,3k
100µ 25V
0,22µ

Only a temperature dependence of the described circuit is observable. The two diodes BAY 61 protect the inputs of the UAA 170 against negative voltages. Since a low-resistive voltage divider (82 resp. 560 Ω) has been chosen, the different currents flowing between the IC's are negligible.

List of Capacitors used in the Circuit 4.13.

Quantity	Component		Ordering Codes
2	MKM-stacked film capacitors	0.22 μF/100 V	B 32540–A 1224–J
2	Electrolytic capacitors	100 μF/25 V	B 41286–B 5107–T

5. Control-, Regulation- and Switching-Amplifier Circuits

5.1. Speed Regulation Using the IC TCA 955

The integrated circuit TCA 955 is especially favoured to regulate the speed of dc motors in movie cameras, casette and tape recorders or record players as well as of more powerful motors in control engineering.
The following features have to be especially accentuated: high regulation precision, wide operating voltage range and achievable current saving.
Fig. 5.1.1. shows the block diagram of the TCA 955 in conjunction with the external connections for the application of a few-pole tacho-generator.
The principle of regulation corresponds to a keyed control. Concerning the dc regulation, the power consumption will considerably reduce in the case the sequence of the switching frequency will be equal to or shorter than the electric motor time constant and the supply voltage higher than the required motor voltage.
At standard operation the switching frequency corresponds to twice the frequency of the speed indicator **(cf. Fig. 5.1.2.)**. Therefore the use of motors with a large number of poles is advantageous, since regulating accuracy and dynamic control action are improved. The capacitor C_3 serves as a smoothing component. A resistor is connected to pins 6, 7 and 8.
In cases, where higher indicator frequencies are not feasible and where the motor time constant is low, the switching frequency oscillator located in the TCA 955 and oscillating at higher frequencies can be used as shown in this application example.
Fig. 5.1.3. indicates both the comparison of desired and actual value measured at the comparator and the appropriate output pulses at an operation using the switching frequency oscillator.
The circuit has a sensitive frequency-dc-converter as an indicator for the actual value. It incorporates an input amplifier and a frequency doubler and can be triggered by different sensors like magneto resistors, photovoltaic cells, inductive pick ups, hall-effect devices as well as simple tacho-generators.
Therefore the regulation accuracy is no longer determined by the tacho-generator. It depends mainly on the external RC-circuit consisting of R_1 and C_2 and on the centre loop gain.
At supply voltages $\geqq 4.6$ and $\leqq 16$ V a voltage regulator incorporated in the TCA 955 becomes active and maintaines the control circuit. In case the regulation requirements are not tough, the device will also work at supply ratings between 2.2 and 6 V (pins 11 and 15 connected).
Besides that the IC features a battery level indicator (battery gauge), which drives an external LED as long as the voltage has not fallen below its minimum level.

Technical Data:

Motor No. 116225/1 with a 6-pole tacho-generator, Company Weiß; Nürnberg, motor time constant: 105 µs.

Supply voltage V_{S1}	+2 to +12 V	Torque ($V_{S1} \geqq 3$ V)	≈25 cmp
Supply voltage V_{S2}	+6 V	Switching frequency	16 kHz
Nominal speed range	1800 r.p.m.	Supply current and speed error	see Fig. 5.4.1.

The current saving is 50% at $V_{S1} = 12$ V compared to a dc control.

circuit diagram of an electronic speed control device with TCA 955

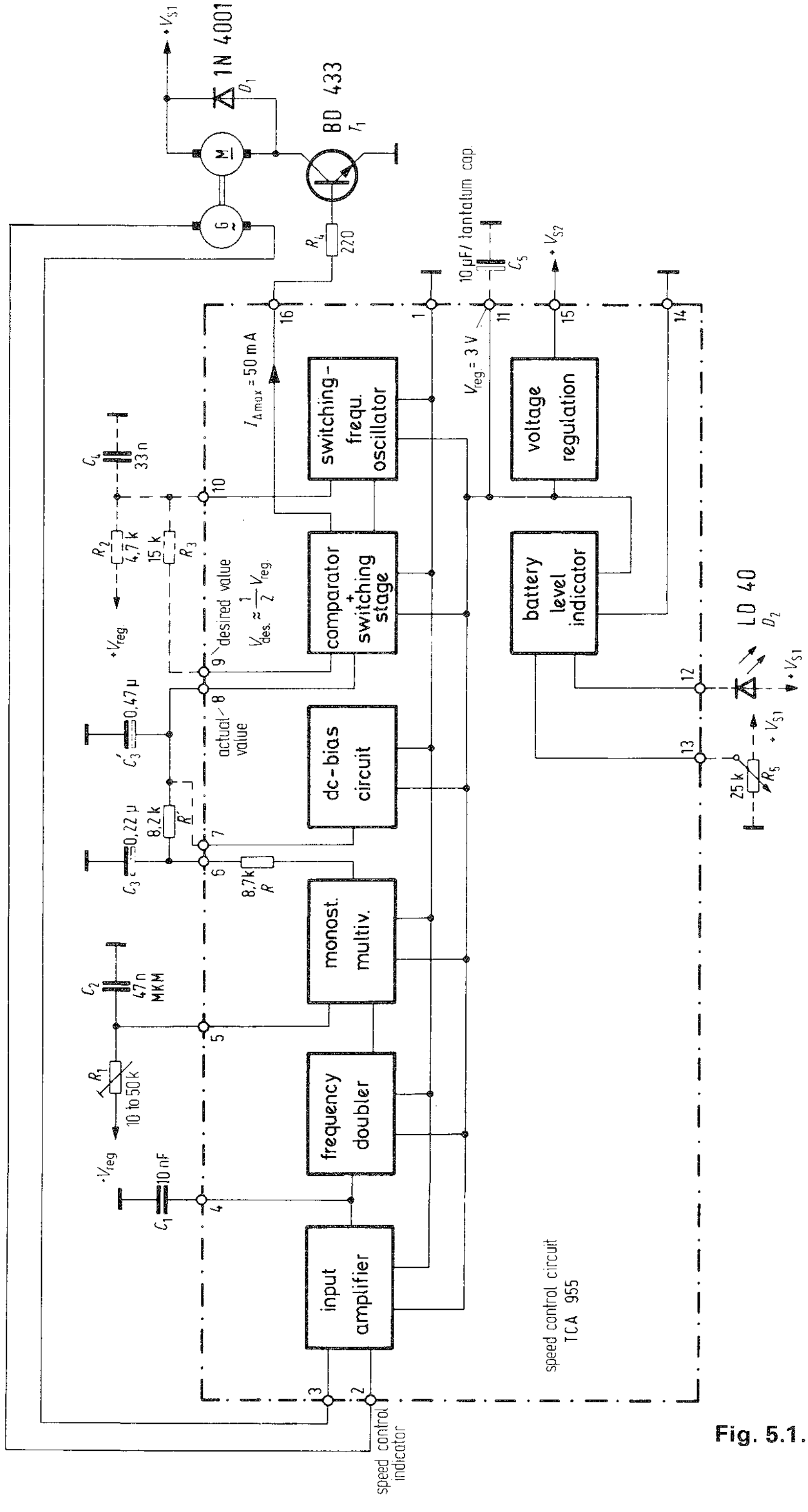


Fig. 5.1.1.

Comparison of desired value and actual one, measured at the comparator without switching-oscillator

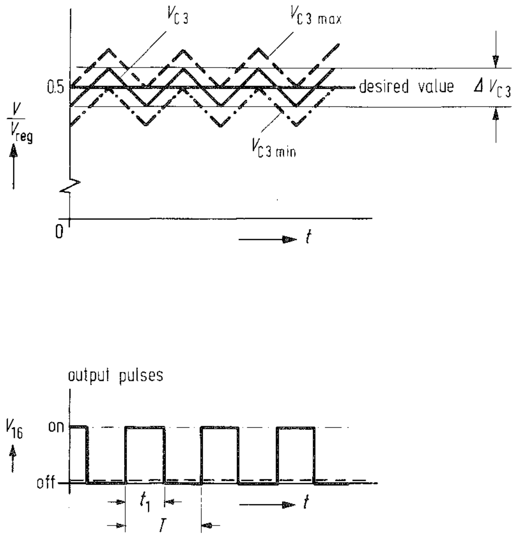


Fig. 5.1.2.

Comparison of desired value and actual one, measured at the comparator with switching oscillator

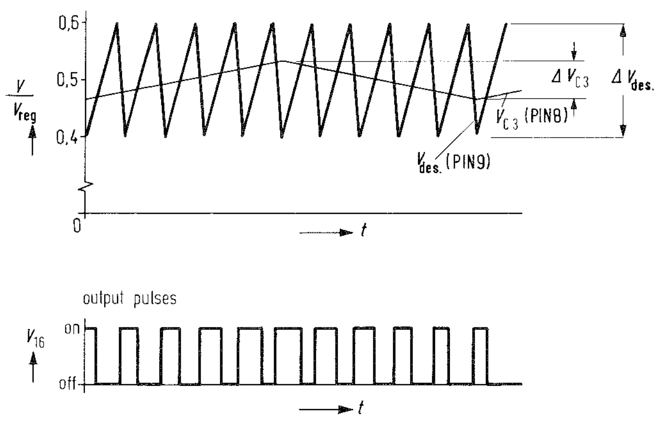


Fig. 5.1.3.

RPM-error and current consumption of the motor as a function of the supply voltage

$n_{nom.} = 1800$ r.p.m., M = const. at $V_{S1} \geq 3$ V

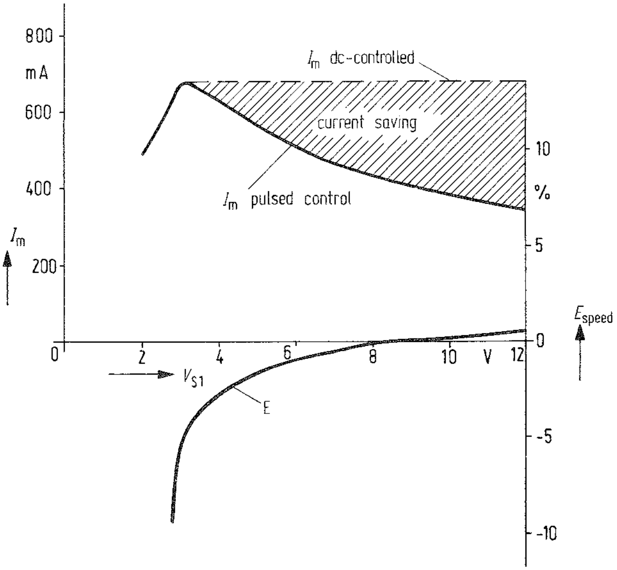


Fig. 5.1.4.

5.2. Clock Generator Using the Threshold Switch TCA 345 A

Fig. 5.2.1. shows the circuit of a clock generator operating as an astable multivibrator. It comprises the threshold switch IC TCA 345A. On account of its extremely low input current of typical 10 nA, long time intervals are achievable. The described circuit is proportioned in a way that a maximum clock pulse interval of 1 min is attained.

clock generator with threshold switch IC TCA 345 A

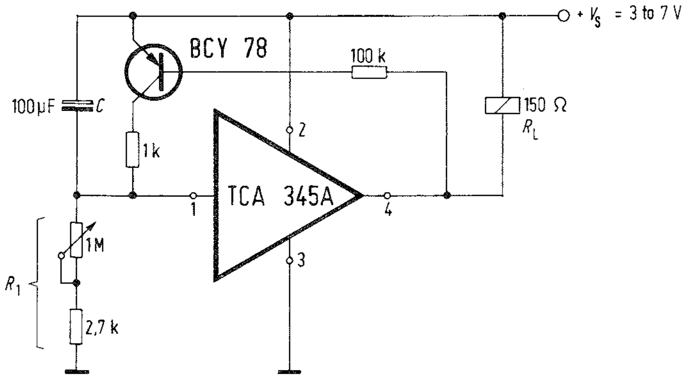


Fig. 5.2.1.

The TCA 345 A offers the advantage in that its package has only 4 pins and no protecting diode is required for switching inductive loads. The threshold levels of the TCA 345 A are proportional to the supply voltage V_S and range between 0.45 $\times V_S$ and 0.66 $\times V_S$.
Because of the non-inverting input of the IC a transistor is additionally required. The capacitor is charged via R_1 and discharged via the 1 kΩ-resistor.
The accuracy of the time intervals depends on the employed capacitor. For larger interval the use of tantalum electrolytic capacitors is necessary to achieve reasonable values ±10%.
However, this application is limited by the capacitor leakage current, i.e. long time intervals (more than 1 min) are only achievable if top-quality capacitors are used or if an adequate loss of accuracy is accepted.

Technical Data:

Supply voltage V_S	3 to 7 V
Pulse interval (adjustable)	0.16 to 60 s
Duration of the produced pulse	30 ms
Output current	30 mA typ., 70 mA max.
Supply current (at V_S = 3 V)	3 mA typ.
Ambient Temperature	−15°C to +70°C
Input current of the TCA 345 A	10 nA typ.

List of Components used in the Circuit 5.2.1.

Quantity	Components	
	Resistors	
1	1 MΩ Potentiometer 0.2 W	
1	2.7 kΩ 0.1 W	
1	1 kΩ 0.1 W	
1	100 kΩ 0.1 W	
1	Load resistor min. 150 Ω	
	Capacitor	
1	Tantalum electrolytic cap. 100 µF/10 V	B 45170–A 1107–M 000
	Semiconductors	
1	BCY 78	
1	TCA 345 A	

5.3. Pulse Duty-Factor Converter for DC-Current-Motor-Operation

For speed setting of e.g. a machine tool motor an adjustable pulse duty factor converter featuring a constant clock frequency is recommended. The pulsed operation guarantees a strong starting torque even at low speeds.
The pulse duty factor converter shown in **Fig. 5.3.1.** consists of a clock generator, which periodically charges and discharges a capacitor C from that of 0.36 to $0.72 \times V_k$ or vice versa. The pulse duration is measured in seconds according to the equation: $t_p = 8000 \times C$, whereas C has to be inserted as a quantity of Farads. A comparator compares the level at the capacitor and the adjustable potentiometer voltage. The duration of the pulse which determines the operation of the motor is achieved by the time difference between the beginning of the capacitor charging and the operation start of the comparator. The output switching transistor T_E is rated according to the required motor peak-current. The suitable clock frequency is calculated from the size and the speed of the motor.

List of Components used in the Circuit 5.3.1.

Quantity	Components
	Resistors
2	1 kΩ 0.3 W
2	10 kΩ 0.1 W
1	2.7 kΩ 0.1 W
1	5 kΩ 0.25 W Poti
2	3.9 kΩ 0.1 W
1	47 kΩ 0.1 W
1	220 Ω 0.3 W
1	560 Ω 0.3 W
1	100 kΩ 0.1 W
1	1.5 kΩ 0.3 W
1	22 kΩ 0.1 W
	Capacitors
1	Capacitor (corresponding to pulse interval) >8 V
	Semiconductors
1	BSX 45
2	BCY 58
2	TCA 315
1	BAY 61
1	BO 110
1	BZX 97 C 8 V 2
1	2 N 3045 (B $>$ 70) or
1	2 N 3055 (B $>$ 70) or
1	BD 433 (B $>$ 70)

Fig. 5.3.1.

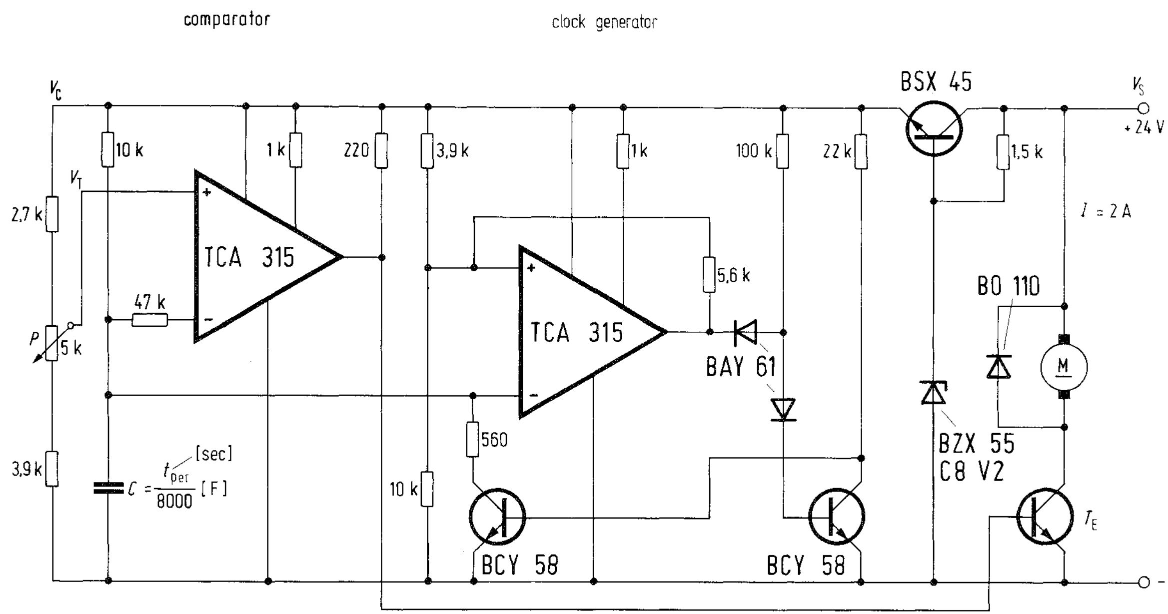
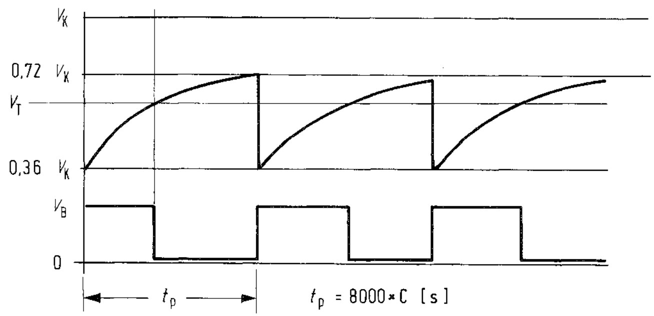


Fig. 5.3.2.



5.4. Operational Amplifier Used as an Analogue-Digital-Converter

The newly-developed circuit of an analogue-digital-converter (**Fig. 5.4.1.**) converts a small dc input voltage to a high squarewave voltage, the duty cycle and frequency of which is proportional to the fed voltage difference. The circuit is perfectly therefore suited to evaluate bridge signals. An analogue signal originating from a bridge circuit composed of PTC or NTC resistors, photodiodes, magneto resistors or other components can, for instance, be converted in pulses without great effort. The analogue-digital converter only answers within a clearly defined functional range. The two output steady states lying above the functional range may be treated as additional information. The circuit is thus particularly suited for monitoring applications.

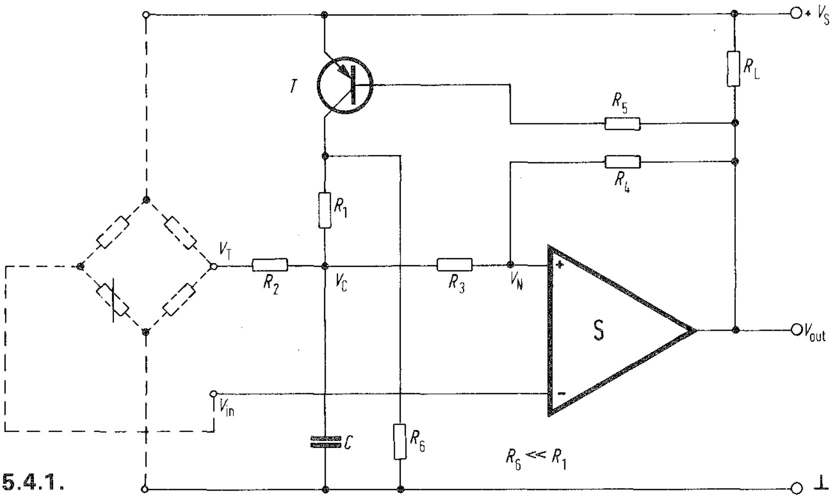


Fig. 5.4.1.

Function

The function of the analogue-digital converter consists of guiding the positive and negative feedback necessary for the generation of oscillations onto the non-inverting input (+) of the operational amplifier *S*. At the switching moment (negative slope) of the amplifier, the potential V_N is reduced at the non-inverting input via the positive feedback circuit R_4, and is increased at the same time, at point V_C via the negative feedback circuit R_5, the transistor *T* and the resistor R_1. Whilst the potential increase at point V_C is delayed for a certain period by the capacitor a switch-on time for the amplifier *S* is given. As soon as the amplifier returns to its steady bias, the potential at point V_C is sufficiently elevated. The capacitor is preferably discharged via the resistors R_1 and R_6 until the amplifier reacts and the operation is repeated.
The diagram (**Fig. 5.4.2.**) shows the duty cycle varying proportionally with the input voltage within the selectable limits $V_{C\,min}$ and $V_{C\,max}$. This proves that the effective voltage variation V_C always corresponds, independently of the input voltage level, to the continuous voltage variation ΔV_N caused by the positive feedback. As the

capacitor is always charged relative to the voltage $V_{C\,max}$ independently of the level V_E and the discharge following $V_{C\,min}$, different charge and discharge times are produced. They are exactly proportional to the input voltage V_{in}. At the output of the amplifier, a square wave voltage V_{out} is available.

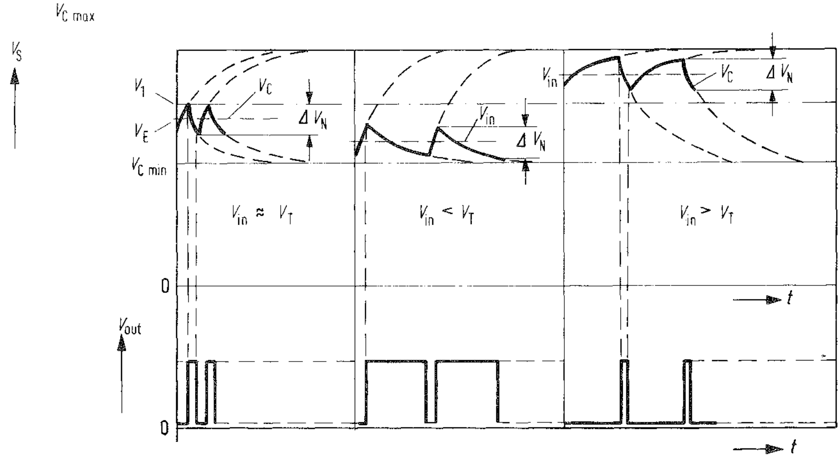


Fig. 5.4.2.

To maintain the oscillations, one has to consider certain regulations when dimensioning the circuit. The voltage variation ΔV_N caused by the positive feedback at the positive amplifier input (+) should be low with respect to the calculated statiscal voltage variation $V_{C\,max} - V_{C\,min}$ of the negative feedback (R_5, T, R_1). As long as the input voltage V_E is within $V_{C\,min}$ and $V_{C\,max}$ the amplifier is able to oscillate. If the input voltage V_E is lower than $V_{C\,min}$ the amplifier S will remain blocked. When V_E is greater than $V_{C\,max}$ amplifier S remains conductive (limit indication). The quiescent value of voltage V_C is determined by the constant partial voltage V_T.

5.5. Ice-Warning Device

In an ice-warning device (**Fig. 5.5.1.**) temperature is measured by the NTC-resistor K 22. At the temperature value to be observed +2°C for instance, the analogue-digital converter will respond (refer to section 5.4) and first activates a signal lamp intermittently and for short periods. The lumination time gets longer as the temperature sinks and, finally, stays on continuously. Instead of the lamp an LED may be used.

List of Components used in the Circuit 5.5.1.

Quantity	Component	
	Resistors	
1	100 kΩ variable 0.2 W	
1	40 kΩ K 22-NTC-resistor	
1	680 Ω 0,3 W	
1	390 Ω 0.3 W	
1	3.9 kΩ 0.1 W	
1	4.7 kΩ 0.1 W	
1	220 kΩ 0.1 W	
1	15 kΩ 0.1 W	
1	4.7 kΩ 0.1 W	
1	2.2 kΩ 0.1 W	
1	12 Ω 0.3 W	
1	680 Ω 0.3 W	
1	560 Ω 0.3 W	
	Capacitor	
1	100 μF Tantalum-electrolytic cap. 10 V	B 45170–A 1227–M 000
	Semiconductors	
1	LED LD 461	
1	BC 327	
1	TAA 865 A	
	Others	
1	Incandescent lamp 12 V/2 W	

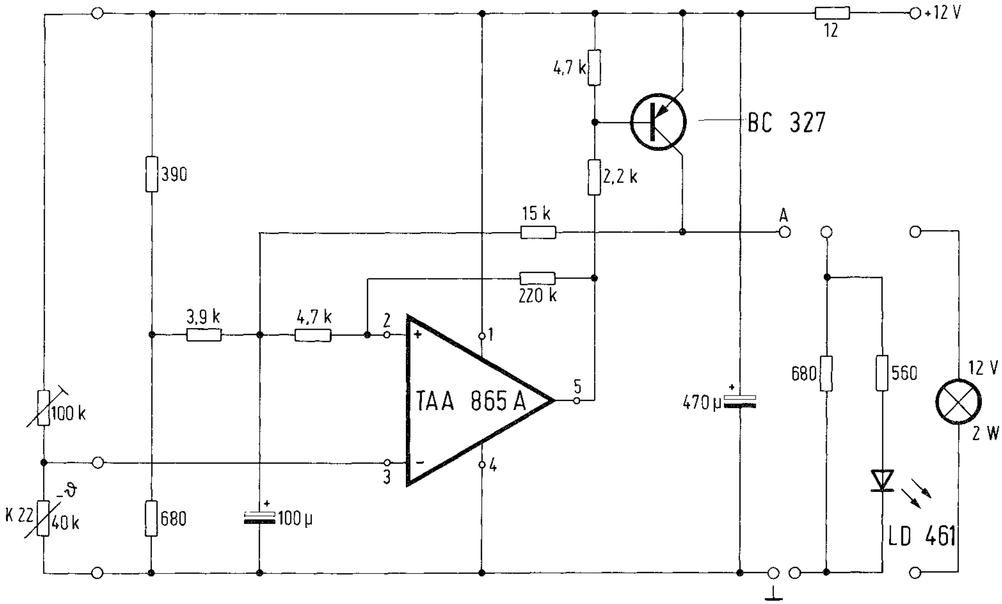


Fig. 5.5.1.

5.6. Thermometer using UAA 170

A thermometer indicating room temperature has been designed to demonstrate the application of the IC UAA 170 in conjunction with a certain number of LED's. The temperature range lies between +15°C to +30°C and the accuracy is ±0.2°C. 16 LED's (one LED for each degree) serves as an indicator. There are four green LED's type LD 57 and twelve reds, type LD 40. The NTC-resistor K 11 (100 kΩ) acts as a temperature sensor. The brightness of the LED's is controlled by the phototransistor BP 102 III depending on the ambient light.

The circuit of the thermometer is shown in **Fig. 5.6.1.** To linearize the characteristics of the NTC-resistor in the stated temperature range, a resistor with R_p = 100 kΩ is connected in parallel to the thermistor (in the circuit it is: $R_p = R_1 + R_2$).

Driving the parallel circuit with a constant current is not necessary, since the temperature range to be linearized is relatively small (+15°C to +30°C). Therefore it is sufficient to produce the current via a series resistor of R_v = 100 kΩ connected from the supply voltage of the IC, being +18 V.

The voltage corresponding to +15°C is 6.4 V. However, the admissible input voltage V_{contr} may not exceed a maximum of 6 V. Therefore the divider consisting of R_1 and R_2 is proportioned in such a way that V_{contr} is 5.4 V at +15°C.

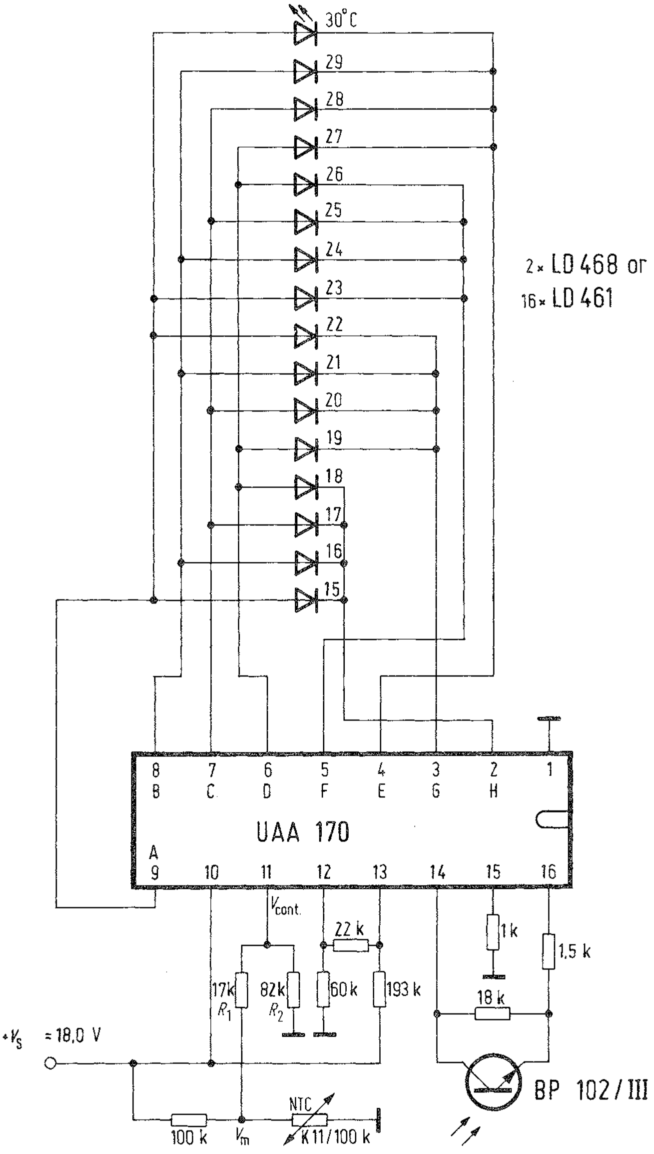

30°C
29
28
27
26
25
24
23
22
21
20
19
18
17
16
15
2× LD 468 or
16× LD 461
8 B
7 C
6 D
5 F
4 E
3 G
2 H
1
UAA 170
A 9
10
11
12
13
14
15
16
$V_{cont.}$
22 k
1 k
1,5 k
17k R_1
82k R_2
60 k
193 k
18 k
$+V_S$ = 18,0 V
BP 102/III
NTC
100 k
V_m
K 11/100 k

Fig. 5.6.1.

List of Components used in the Circuit 5.6.1.

Quantity	Components
	Semiconductors
12	LED's LD 40
4	LED's LD 57
1	IC, type UAA 170
1	BP 102/III phototransistor
	Resistors
1	NTC-resistor K11/100 kΩ
1	Resistor 100 kΩ/0.1 W
1	Resistor 17 kΩ/0.1 W
1	Resistor 82 kΩ/0.1 W
1	Resistor 60 kΩ/0.1 W
1	Resistor 22 kΩ/0.1 W
1	Resistor 193 kΩ/0.1 W
1	Resistor 18 kΩ/0.1 W
1	Resistor 1 kΩ/0.3 W
1	Resistor 1.5 kΩ/0.3 W

5.7. Applications Using the Programmable Unijunction-Transistor (PUT) BRY 56

This section describes the features of the PUT BRY 56 and how to utilize them in applications of multivibrators, saw-tooth generators, pulse generators, long-period timers as well as pulse circuits like triggers for thyristors and triacs.
In conjunction with adequate resistors, externally connected, a PUT shows the same behavior as a regular unijunction transistor. If the voltage between anode and cathode has exceeded the programmed trigger level, the so-called "peak-point voltage" V_P, the PUT is switched through. It is only turned off, if the anode-cathode voltage falls below the holding level, the so-called "valley point voltage" V_V, being about 1 V. The internal structure of a PUT is, however, quite different to the one of a conventional unijunction transistor. The PUT is more like a thyristor-tetrode consisting of four differently doped semiconductor layers (pnpn).
Fig. 5.7.1. shows the composition and the equivalent circuit diagram. The connections required to set the peak-point voltage are indicated in **Fig. 5.7.2.**

The features of the PUT BRY 56 and its advantages compared to conventional unijunction transistors (UJT) are as follows:
- The trigger voltage of the PUT (peak-point voltage V_P) can be freely chosen contrary to the one of a UJT, i.e. it is programmable and does not depend on the supply voltage.
- The trigger current I_P (peak-point current) and the holding current I_V (valley current) can be set in wide ranges by external resistors.
- Relatively low forward voltages drop of less than 1.4 V.

– Extremely low residual currents flowing from gate to anode ($I_{GAO} < 10$ nA) and to cathode, especially when high voltages are applied.
– Applicable to relatively high voltages.
– Due to the last two mentioned features of the BRY 56 and because of the low trigger current requirements the PUT is especially favoured for pulse generators oscillating with low frequencies and for long-period timers.

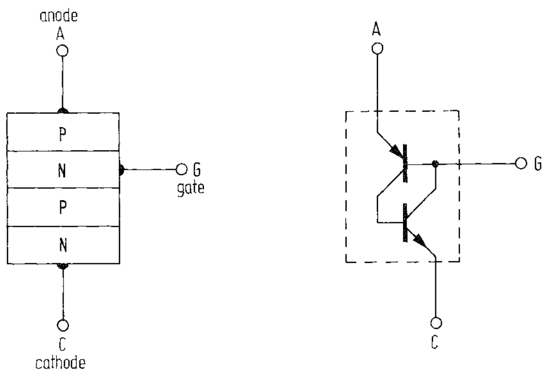
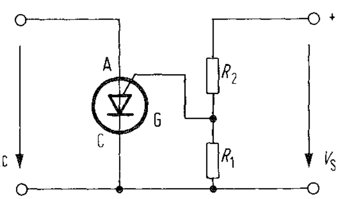


Fig. 5.7.1.



$$V_{AC} = V_P = \frac{R_1}{R_1 + R_2} V_S + V_T$$

V_T = offset-voltage of the PUT

Fig. 5.7.2.

5.7.1. Pulse Generator

In most of the applications the PUT is used in a relaxation oscillator. **Fig. 5.7.3.** shows the fundamental circuit and the achievable oscillation modes, like: sawtooth voltage and pulses with positive as well as negative polarity.

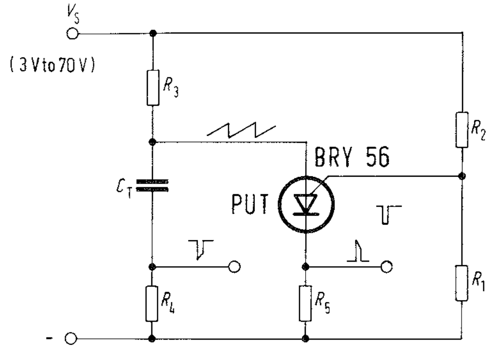


Fig. 5.7.3.

The peak-point voltage is adjusted by resistors R_1 and R_2 as follows:

$$V_P \approx \frac{R_1}{R_1 + R_2} \times V_S$$

The minimum-to-maximum rating of the current (I_P) : (I_V) is determined by the resistance of R_g, which is the parallel connection of R_1 and R_2. Also refer to the data sheet on BRY 56.

When the voltage at the capacitor exceeds the peak point level the PUT is triggered and the capacitor is discharged. As soon as the discharging current falls below the valey current, the PUT is turned off and the capacitor charges again. In this way a relaxation oscillation is generated.

Its frequency is mainly determined by the time constant being a product of the capacitance C_T and the resistance R_3 $(\tau_1 = R_3 \times C_T)$.

The resistors R_4 and R_5 limit the discharging current of the capacitor. Their resistance can be rated adequately low. The time constant $\tau_2 = (R_4 + R_5) \times C_T$ determines the duration of the pulses which are available at resistors R_4 and R_5 with positive respectively negative amplitude.

If pulses having only one polarity are required one of the resistors R_4 or R_5 can be eliminated. To achieve a safe beginning of the oscillation the resistance of R_3 has to be rated in the way that the current flowing through R_3 is on the one hand higher than the peak-point current I_p but lower than the valey current I_V.

Oscillation Conditions

It applies:

$$\frac{V_S - V_V}{I_V} < R_3 < \frac{V_S - V_P}{I_P}$$

whereas V_V is the valey voltage of about 1 V

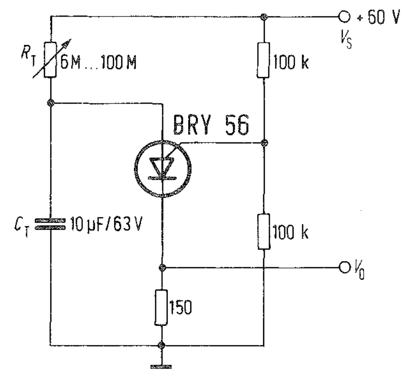
V_P is the peak-point voltage. It is adjustable by R_1 and R_2 according to the following equation

$$V_P \approx \frac{R_1}{R_1 + R_2} \times V_S .$$

The currents I_P and I_V are determined by the resistance of R_g which is the parallel connection of R_1 and R_2 (also refer to data sheet). The pulse generator can be operated at a voltage range of 3 to 70 V, whereas changes of the supply voltage will not essentially influence the frequency of the oscillations. On the contrary, the amplitude is practically proportional to the supply voltage.
A certain temperature dependence results from the influence of the off-set voltage V_T, which can be reduced, however, by choosing a higher peak-point voltage. For $R_g < 100\ \text{k}\Omega$ a partial compensation of the temperature effects are possible by adding a diode to the voltage divider consisting of R_1 and R_2 (compare also Fig. 5.7.9.). Since the resistor of the time circuit has a high resistance the described pulse generator is especially suited for producing oscillations with low frequencies.
However, a limitation for the time constant is set by the leakage current of the employed capacitor. In order not to influence the charging current of the capacitor by the leakage current (insulation current) the former should be chosen essentially higher. **Fig. 5.7.4.** shows a circuit comprising a MKL-capacitor (10 µF) and in the circuit of **Fig. 5.7.5.** an electrolytic capacitor is used (100 µF, leakage current $\ll 1\ \mu\text{A}$).

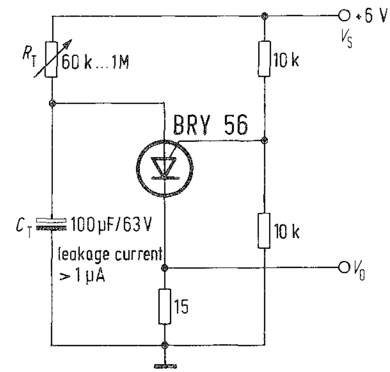
5.7.2. Metronome

The circuit of a metronome as demonstrated in **Fig. 5.7.6.** is a simple application example for a pulse generator. Its clock frequency can be adjusted by the 1 MΩ-potentiometer in a range of about 30 to 240 cycles per minute.



$R_T \times C_T$ = 60 to 1000 sec

Fig. 5.7.4.



$R_T \times C_T$ = 6 to 100 sec

Fig. 5.7.5.

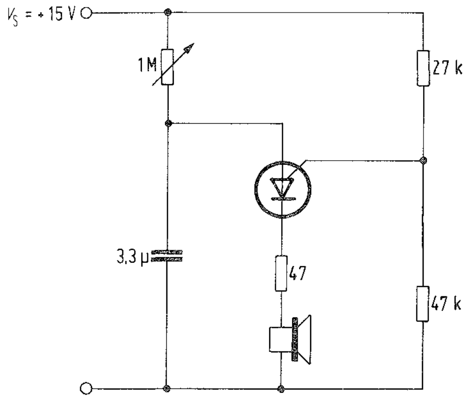


Fig. 5.7.6.

5.7.3. Trigger Circuit for a Thyristor

Fig. 5.7.7. shows a simple circuit for triggering a thyristor by directly applying the discharging pulse to it.
The adjusted resistance of the 1 MΩ-potentiometer determines the charging current and thus the charging time of the capacitor. When the voltage at the capacitor exceeds the trigger level the PUT is turned on and the discharging current of the capacitor triggers the thyristor.
The z-diode limits the voltage of the control circuit during the time of the positive half-wave and prevents a charging of the capacitor during the negative half-wave.
A triac can also be triggered in a similar way. In this case the control circuit has to receive a voltage, which is produced by a full-wave rectifier and the triggering pulses have to be transmitted by a small transformer.

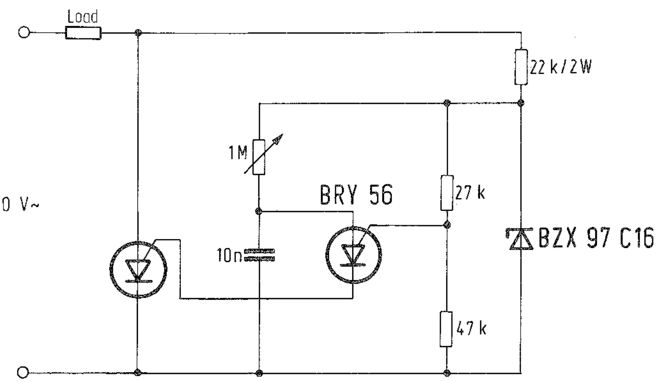


Fig. 5.7.7.

5.7.4. Long-Period Timer

If the circuit described in **Fig. 5.7.7.** (thyristor trigger) is connected to a dc instead of an ac voltage, a non-repeated time delay for the load switching-on occurs. By appropriate dimensioning of the time circuit long delay times can be achieved, e.g. 1000 sec and by that a long-period timer is realized.

5.7.5. Monostable Multivibrator

The monostable multivibrator shown in **Fig. 5.7.8.** consists of two different circuits, one is a pulse generator comprising a PUT the other is an astable multivibrator using two transistors.

In the steady state the transistor T_1 is turned on, i.e. its collector level is low. The voltage for the time circuit ($R_T \times C_T$) is picked up at the collector of T_1 and also has a low value during the quiescent state. Therefore the capacitor C_T remains uncharged. Transistor T_2 is turned on by applying a positive pulse to the input and transistor T_1 is simultaneously switched off. The collector of the latter now has a high level and the capacitor C_T can be charged via R_T and the collector resistor of T_1.

If the voltage across C_T exceeds the trigger level the PUT is switched on and by that a voltage pulse, also turning on the transistor T_1, is produced at the cathode resistor R_C. The collector level is decreased to a low level and thus a low voltage is also supplied to the R_TC_T-timing-circuit. The capacitor can only be charged if the collector levels of the bistable multivibrator have again been inverted after applying a new trigger pulse to the input.

To guarantee that the output has a defined level when the supply voltage is applied, the base of T_1 is connected to the positive power supply terminal via a series connection of resistor R_1 and capacitor C_1. Before the supply voltage is supplied C_1 is not charged and a current flows via the resistor R_1 to the base of T_1 when the supply voltage is connected. Thus the transistor T_1 is switched through.

This monostable multivibrator is especially suited for generating long pulse periods respectively long pulse durations. The latter is relatively independent of supply voltage variations, because both the control voltage at the gate of the PUT and the capacitor voltage at its anode changes in the same direction.

The recovery time is mainly determined by the time constant $\tau = C_T \times R_C$. It can be rated very low. The lower limit for the capacitance of the time-circuit capacitor is fixed by the required peak value of the pulse at R_C (cf. data sheet). The upper level is set by the leakage currents of the employed capacitors.

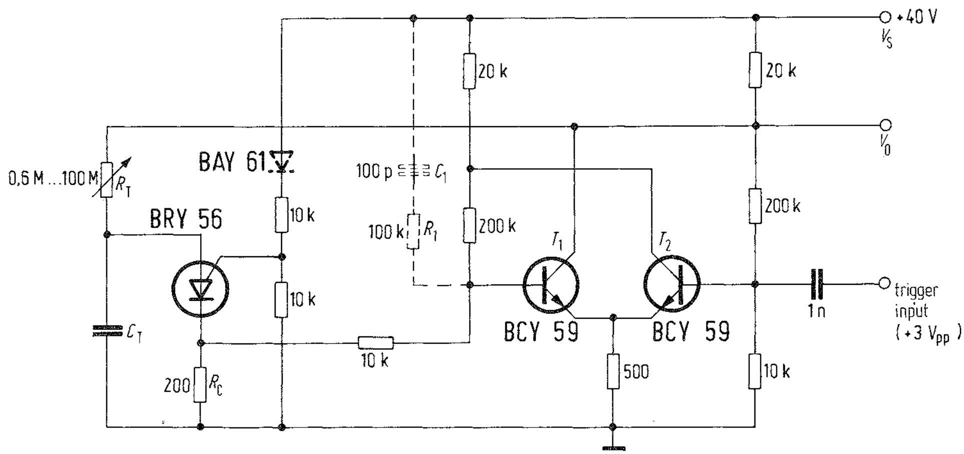


Fig. 5.7.8.

5.7.6. Saw-Tooth Generator

The saw-tooth voltage linearity of the oscillator shown in **Fig. 5.7.9.** can be improved if the capacitor C_T is charged with a constant current, produced by means of a transistor.
The duration of the saw-tooth is adjustable by the potentiometer R_3 after the amplitude has been set within a range of about 1 V to 10 V through the potentiometer R_1. A switch for selecting different capacitances of C_T is recommended for covering a wider frequency range.
The temperature drift of the base-emitter-voltage of the transistor is compensated by the diode D_1. The diode D_2 partly eliminates the temperature drift of the PUT offset voltage.
The available output current I_0 is relatively small. It is essentially determined by the resistance of the emitter resistor R_3. For $R_3 = 20\ \text{M}\Omega$ follows: $I_0 \ll 1\ \mu\text{A}$. Therefore, a following amplifier is necessary according to the kind of application.

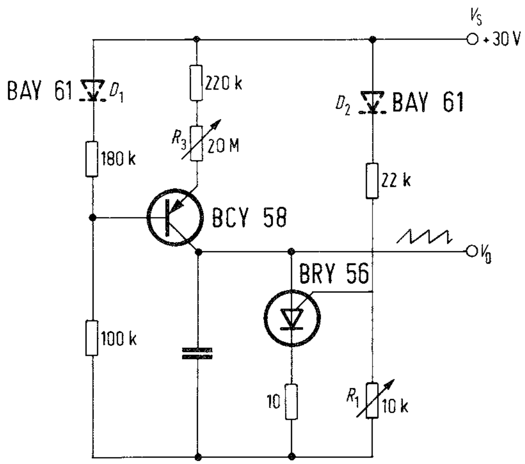


Fig. 5.7.9.

5.8. Speed-Dependent Interlocking Switch System

Speed-dependent switching operations can easily be realized by utilizing the circuit shown in **Fig. 5.8.1.** A door interlocking can be designed for instance, for an electric washing machine, i.e. the door cannot be opened when the washing drum has reached a certain speed.
In conjunction with a small permanent magnet the Hall-IC SAS 211 functions as a tacho-generator. The magnet ($12.5 \times 8 \times 6\ \text{mm}^3$) can be mounted e.g. on a belt pulley as shown in **Fig. 5.8.2.** Per a rotation each a pulse with a duty factor of about 1 : 1000 is generated. The circuit is so dimensioned that the 1 μF-capacitor can be charged via the 150 Ω-resistor during the pulse duration, independently of the speed. During the pulse space the 1 μF-capacitor is discharged via the 150 Ω and the 1 MΩ-resistors. The discharging time constant $\tau = 1\ \mu\text{F} \times 1\ \text{M}\Omega$ has been chosen in such a way that the capacitor is discharged, more or less, depending on the

speed. If the voltage at the 1 µF-capacitor is integrated, a speed-dependent dc voltage is achieved. The latter in conjunction with a threshold switch is able to realize a switching operation. The speed threshold switch is adjustable by the 5 kΩ-potentiometer.

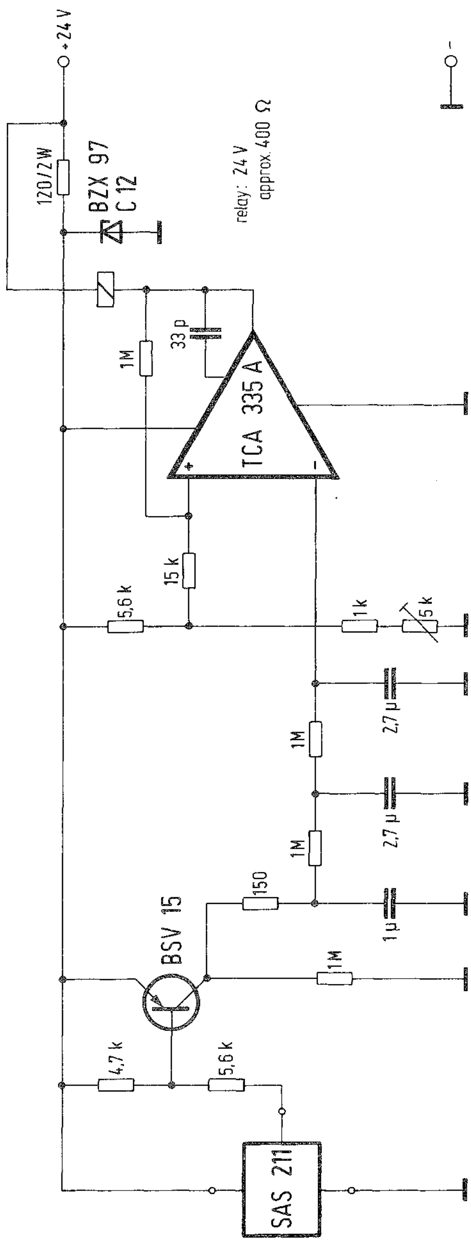


Fig. 5.8.1.

Technical Data:

Supply voltage	24 V
Temperature range	0 °C to 70 °C
Speed range for the indicated dimensioning	$0 < n \leqq 1000$ r.p.m.
Switching point for e.g. 60 r.p.m. (Adjustable by the 5 kΩ-potentiometer)	about 4 V
Dimensions of the magnet (Oxyd permanent magnet DS1)	$12.5 \times 8 \times 6\ mm^3$

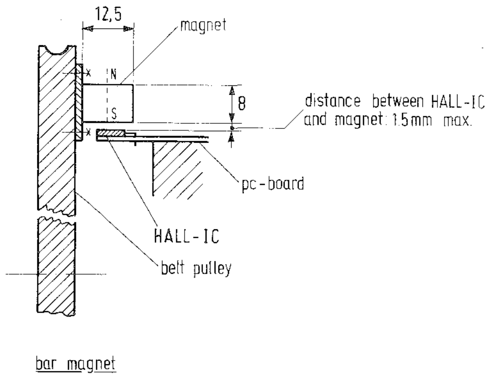


Fig. 5.8.2.

List of Capacitors used in the Circuit 5.8.1.

Quantity	Components	Ordering codes
1	1 µF/63 V	B 32120–E 9105–M 000
2	2.7 µF/63 V (3.3 µF)	B 32120–E 9335–M 000
1	33 pF/63 V	B 31063–A 5330–M 000

5.9. Amplifier for the Differential Magneto Resistor FP 211 D 155

The differential magneto-resistor-sensor FP 211 D 155 is especially suited to adapt speeds and angle positions. Contrary to other types of MR's from our programme the soft iron bars which is to be detected can be very small and mounted with extremely small distances from each other.

By using a tooth wheel, for instance, a high number of pulses per one rotation can be achieved even if the wheel diameter is relatively small. Two circuits suited for application of the FP 211 D 155 are described in **Fig. 5.9.1.** and **Fig. 5.9.2.**

Because of the small dimensions of the MR-sensor FP 211 D 155 the supply voltage has to be low; i.e. the power dissipation has to be kept low. Special considerations have to be paid to this fact if the sensor is used at high ambient temperatures of e.g. +100 °C. At temperatures between +80 °C and +100 °C a supply voltage of only 2 V is permitted.

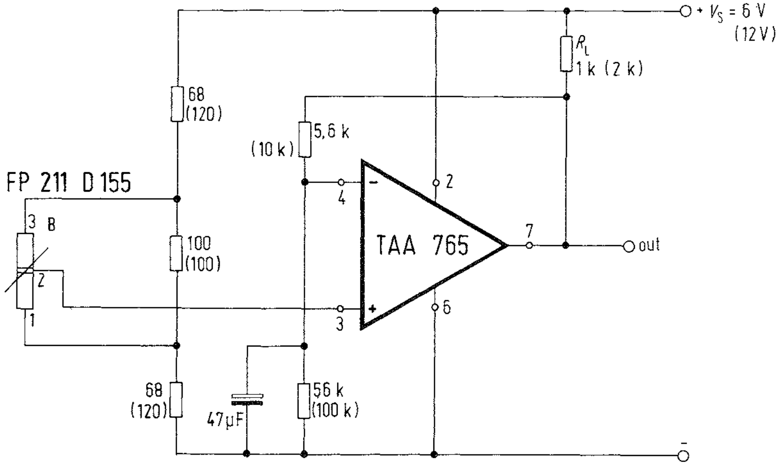


Fig. 5.9.1. Preamplifier for the differential MR: FP 211 D 155

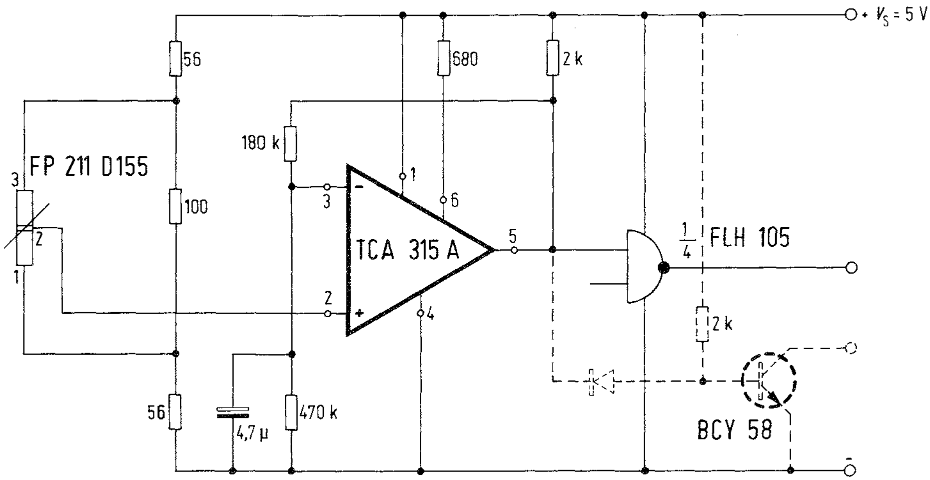


Fig. 5.9.2. Preamplifier for the differential MR: FP 211 D 155

If besides the high ambient temperature, the air gap between MR-sensor and soft iron bar is great, the signals supplied from the MR-bridge are very small. Therefore a dynamical amplifier should be used.
Due to a strong dc current inverse feedback, an individual adjustment is not necessary. In the following, the previously mentioned circuits are described in detail.
The supply voltage for the circuit of Fig. 5.9.1. is 6 V. The values in brackets correspond to a 12 V-supply.

Only a voltage of 2 V may be applied to the MR-sensor type FP 211 D 155 at a maximum operating temperature of +100 °C. Therefore the supply voltage is reduced by a divider. Its symmetrical arrangement is chosen with respect to the following op amp. The inverse feedback of the amplifier is strong (5.6 kΩ-resistor). Thus a 10% higher voltage than the level at the centre tap of the MR-sensor is available at its output, if the magneto resistors are not magnetically influenced. The voltage at the centre tap is half the supply voltage, whereas tolerances of the MR-sensor, the regular resistors and the input off set voltage may vary this value slightly. If the MR-sensor is influenced by a passing soft iron tooth, as shown in **Fig. 5.9.3.** above, its centre voltage is rapidly changed. This variation is directly applied to the non-inverting input of the op amp. It is amplified with the full gain and available as a square pulse at the output, since the amplifer operates in its saturation state. The dc current inverse feedback is not effective for the produced pulses since the inverting input is blocked by a capacitor, which also determines the lower cut-off frequency f_u respectively the maximum pulse duration. F_u is about 50 Hz at a capacitance of 47 µF and at $V_s = +6$ V.

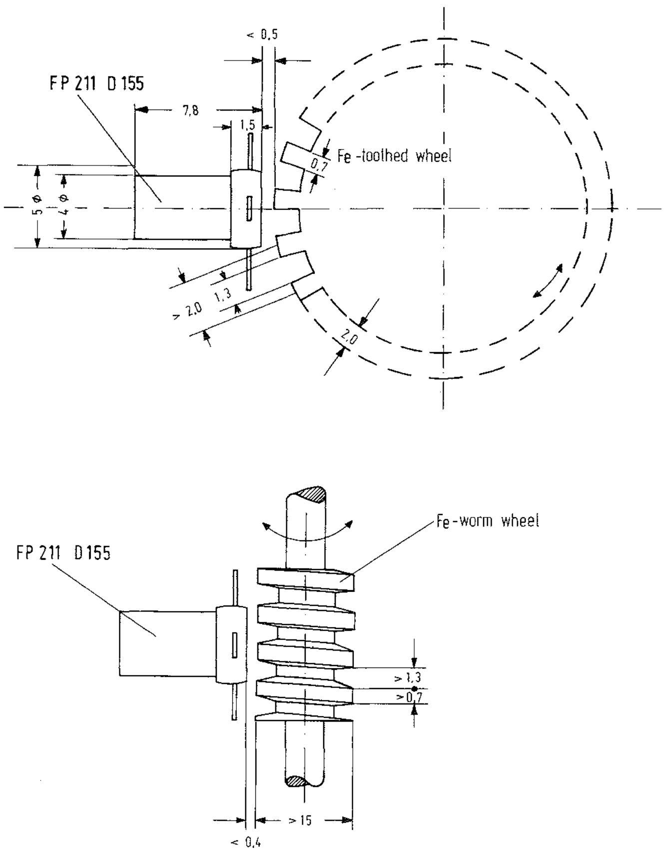


Fig. 5.9.3.

The circuit of **Fig. 5.9.2.** operates with a supply voltage of 5 V. As op amp the type TCA 315 A has to be used. Its output is connected to a TTL NAND-gate, which can also be replaced by a circuit using a transistor BCY 58 as shown in the figure by dashed lines.

Technical Data:

Total resistance	(400 ± 120) Ω
Centre symmetry	≦5%
Operating temperature	−25 to +80°C
Power dissipation	
at case temperature $T_C = 75°C$	50 mW
at ambient temperature $T_{amb} = 75°C$	25 mW
Thermal conductivity G_{thC}	5 mW/K
Cut-off frequency	20 Hz
Required resistance variation of the MR sensor at an air gap of 0.5 mm	$\frac{R_{1-2}}{R_{2-3}} > 3\%$

List of Capacitors used in the Circuits 5.9.1. and 5.9.2.

Quantity	Components		Ordering codes
1	Electrolytic capacitor	47 µF/40 V	B 41286–B 7476–T 000
1	Electrolytic capacitor	4.7 µF/63 V	B 41315–A 8475–Z 000

5.10. Regulating Unit for 24 V and 3 A max.

The **Fig. 5.10.1.** shows a transistor circuit which is able to replace a variable high-power resistor. However, the power dissipation cannot be reduced by this control circuit, it is released by the power transistor, which has to be cooled adequately.

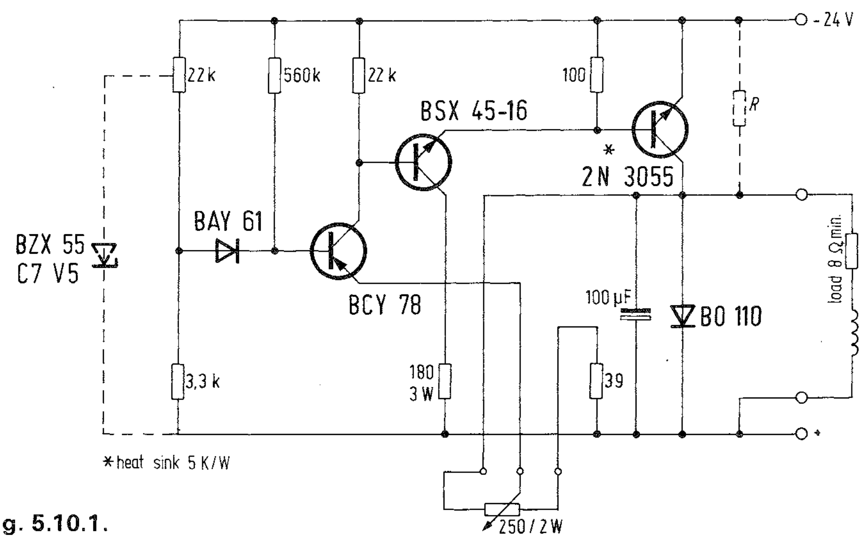


Fig. 5.10.1.

The power dissipation range can be totally utilized. The voltage across the load is regulated by adding a z-diode. The circuit is short-circuit-proof with a few exceptions. If the total adjustment range is not practically used the power dissipation can be partially taken over by an adequate resistor.

List of Capacitors used in the Circuit 5.10.1.

1 pc	Electrolytic capacitor	100 μF/40 V	B 41286–A 7107–T 000

5.11. Motor Control for Clockwise and Counterclockwise Rotation

For a servomotor 12 V/15 W or 24 V/30 W the following control circuit for counterclockwise and clockwise rotation was designed.
The input of the circuit shown in **Fig. 5.11.1.** is also TTL-compatible. The input level "L" means a clockwise rotation whereas a counterclockwise rotation is achieved by the input signal "H". If there is no input signal, all output transistors are switched-off for sure, i.e. it is assured that the motor stands still.
For the output stage the pnp/npn epibase power Darlington transistors BD 644/643 are recommended. They contain already an inverse diode each in antiparallel connection. These diodes are essentially necessary for the feedback of the motor induction energy.

Operating Data:

Supply voltage	12 V (24 V) ± 10%
Motor output	15 W (30 W)
Input signal	TTL-compatible

5.12. AC Current Switch using Thyristors

Electrical loads requiring ac currents are practically controlled by triacs, but there exist some exceptions, making an operation with thyristors connected in antiparallel more useful, e. g. for the case of a high supply voltage or of a high load current. In the following circuit shown in **Fig. 5.12.1.** the operating voltage is 380 V. Therefore two thyristors antiparallelly connected are used.
The thyristors BSt BO 146 are controlled via a small transformer, the secondary voltage of which is rectified. This dc-voltage control offers the advantage that also highly resistive loads can be switched as shown in this design example.

Technical Data:

Supply voltage	380 V_{ac}
Load	500 Ω to 80 kΩ
Control current	about 6.5 mA_{ac}
Thyristors	2 × BSt B 0146
Rectifier	2 × B 1210 B 60 C 1000/700
Capacitor	0.1 μF/1 KV B 32227–A 0104
Transformer	M 42 dynamo sheet 4

n_1 = 7400 turns enamelled copper wire
n_2 = 230 turns enamelled copper wire
n_3 = 230 turns enamelled copper wire

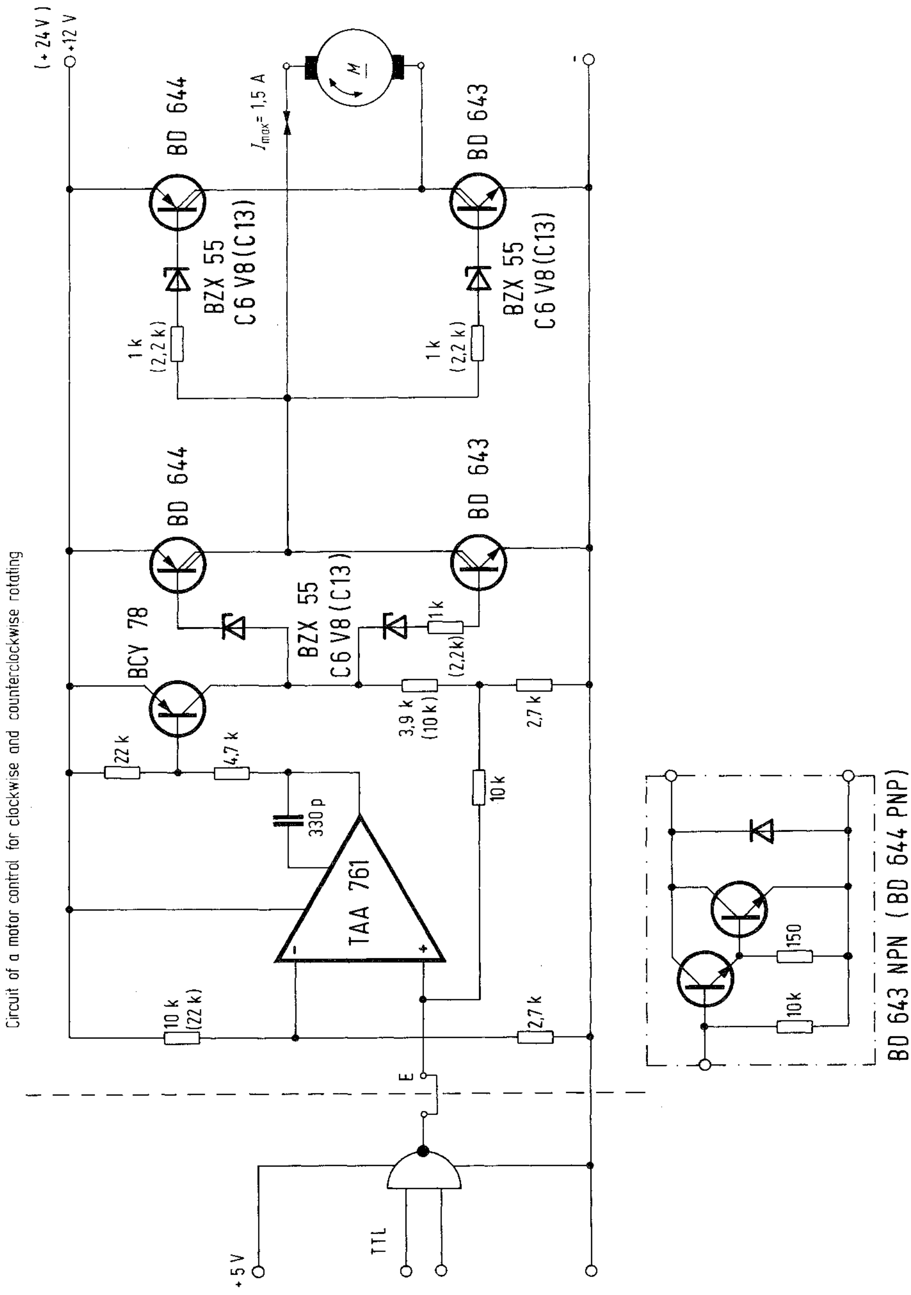

Circuit of a motor control for clockwise and counterclockwise rotating
(+24 V)
+12 V
BD 644
BD 643
BZX 55
C 6 V8 (C13)
1 k
(2,2 k)
I_{max} = 1,5 A
M
BCY 78
22 k
4,7 k
330 p
TAA 761
10 k
(22 k)
2,7 k
3,9 k
(10 k)
E
TTL
+5 V
150
BD 643 NPN (BD 644 PNP)

Fig. 5.11.1.

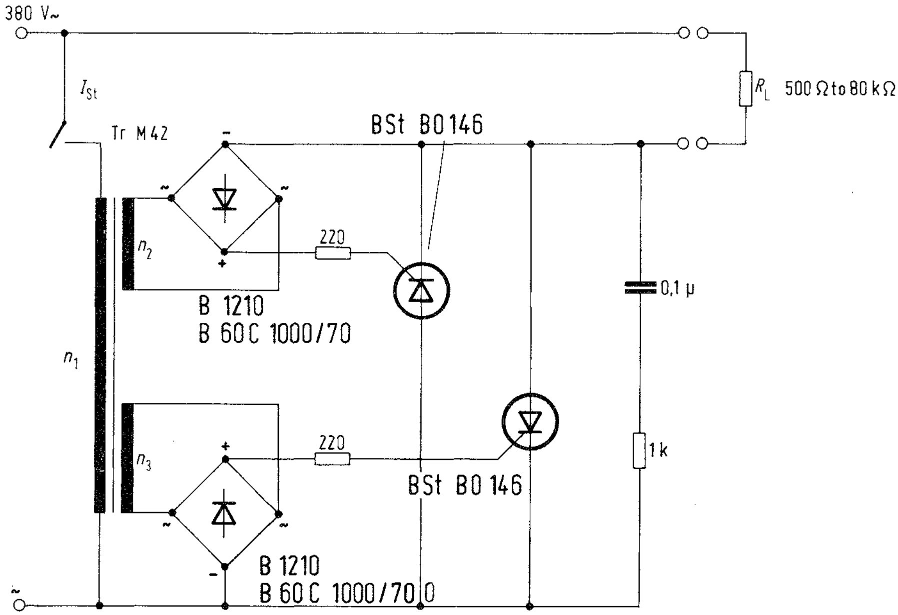


Fig. 5.12.1.

Capacitor used in the Circuit 5.12.1.

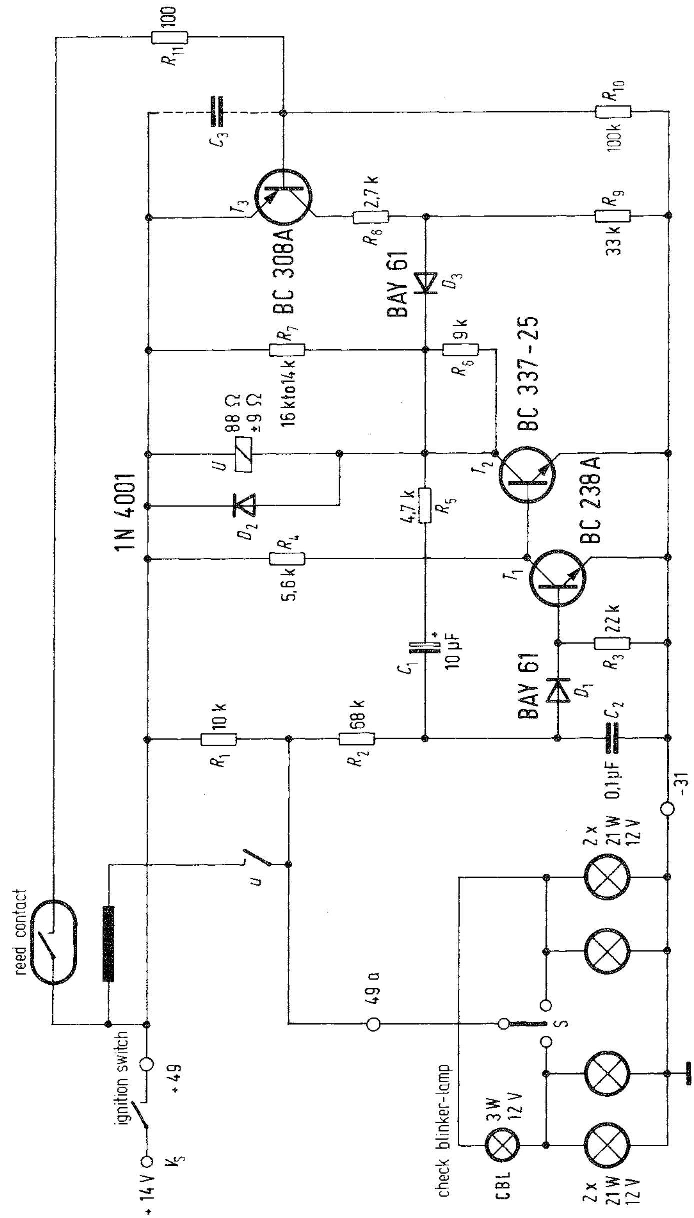
1 pc	MKH-capacitor	0.1 µF/1 KV	B 32227–A 0104

5.13. Electronic Direction and Emergency Flasher using Reed Contacts for Automobiles

According to official regulations, flashing-light direction indicators for cars have to be provided with a checking device which informs the driver, by means of a pilot lamp, when one of the flashing-light is not operating. The failure can be indicated either directly or indirectly.
Due to its specific characteristics the Siemens reed contact SK 260 can be advantageously utilized for realizing a flashing-light breakdown indicator.

Direct Indication

The warning unit for the flashing-light indicator comprises the electronical clock generator, the flashing-light switching relay and the checking device using the reed contact SK 260 (cf. **Fig. 5.13.1.**). The operation of the flashing-lights is monitored by checking their lamp current. A breakdown is indicated by a pilot lamp flashing with twice the frequency than before, i.e. the numbers of pulses increase from 90 to 170 pulses per minute. The current for the flashing-lights flows via the winding of the reed contact (11 turns). Therefore the contact synchronously operates at the flashing frequency and does not influence the one of the multivibrator being 1.5 Hz.



R_{11} 100
C_3
R_{10} 100k
T_3 BC 308A
R_8 2,7 k
R_9 33 k
BAY 61 D_3
R_7 16 k to 14 k
R_6 9 k
BC 337-25
U 88 Ω ±9 Ω
T_2
1N 4001
D_2
R_4 5,6 k
4,7 k R_5
BC 238 A
T_1
C_1 10 μF
R_3 22 k
BAY 61 D_1
C_2 0,1 μF
R_1 10 k
R_2 68 k
-31
2 x 21 W 12 V
reed contact
u
49 a
S
ignition switch
+49
+14 V
K_5
check blinker-lamp
CBL
3 W 12 V
2 x 21 W 12 V

Fig. 5.13.1.

When a breakdown for one of the flashing-lights occurs the winding current decreases by the factor 2. The exitation is also reduced by 50% and the reed contact does no longer operate. Therefore the transistor T_3 becomes conductive and the resistor R_8 is connected in parallel to R_7. By that the time constant being responsible for the turn on- and off-times is reduced. This results in an increase of the flashing frequency to about 2.8 Hz.

Technical Data:

Operating voltage range	10 to 16 V
Nominal flashing frequency at $V_S = 14$ V, $T_{amb} = 25$°C	90 pulses per min
Upper limit for the flashing frequency at $V_S = 10$ V, $T_{amb} = -40$°C	102 pulses per min
Lower limit for the flashing frequency at $V_S = 16$ V, $T_{amb} = +80$°C	81 pulses per min
Flashing frequency when one of the lamps has a breakdown	~170 pulses per min
Permissible ambient temperature range	-40°C to $+80$°C
Load of the flashing lamps indicating the direction	2×21 W + 3 W
Load of the flashing warning lamps	4×21 W
Switching relay	V 23033–C 1001–A 402

Indirect Indication

Fig. 5.13.2. shows a direction flasher employing less elaborateness. In this circuit the winding for the reed contact is directly responsible for checking the lamp current

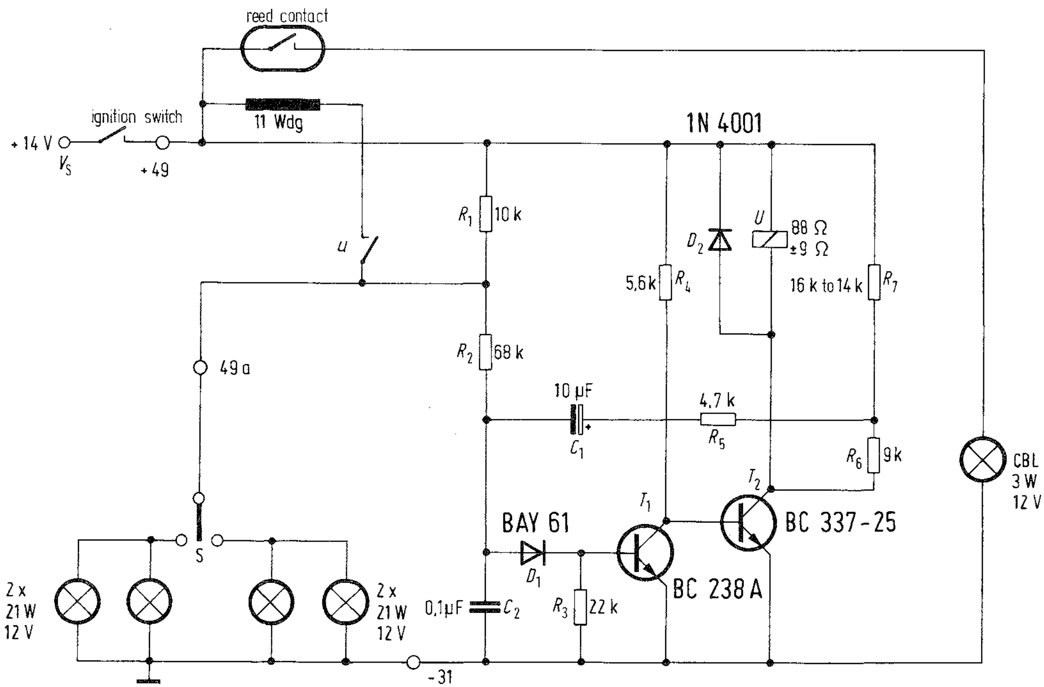


Fig. 5.13.2.

of the direction flasher. If the current of two flashing lamps flows through the winding the reed contact is closed, when the lamps are turned on, and the pilot lamp (CBL) operates at the flashing frequency. When one of the flashing direction indicator lamps have a break-down the current is not sufficiently high to sustain a pick-up of the reed contact SK 260. The pilot lamp (CBL) is no longer connected to the supply voltage and thus a failure is indirectly indicated.

List of Capacitors used in the Circuits 5.13.1. and 5.13.2.

Quantity	Components	Ordering codes
1	10 µF/63 V Electrolytic cap.	B 41283–A 8106–T
1	0.1 µF/100 V MKM-stacked film cap.	B 32540–A 1104–J 000

5.14. Electronic Regulator for Three-Phase Automotive Generators

Modern cars are equipped with three-phase generators which are very reliable in performance, require little maintenance and weigh less than dc generators while having the same capacity.

The generator voltage is controlled by a two-step action regulator which determines the duty cycle of the inductive exciter current of, e.g. 4.0 A in a 490 W-generator. Because of their low control accuracy and their rapid deterioration, conventional mechanical regulators still in use today are replaced more and more by electronic regulators.

Their dimensions are smaller than those of mechanical regulators, and they can therefore be directly attached to the generator or even installed in it without affecting their reliability.

Fig. 5.14.1. shows a circuit of a regulator, the generator and the control indicator. The three-phase generator has a three-phase current winding in the stator, and an exciter winding in the rotor. The stator winding is connected to the ac side of a three-phase rectifier bridge, consisting of silicon diodes (D_7 to D_{12}), with a battery attached to its dc current clamp. A second three-phase rectifier bridge (D_4 to D_6, and D_{10} to D_{12}) supplies the exciter current which is fed via the regulator, the two brushes and the slip rings to the exciter coil (EW).

In case of fluctuating rotational speed and load, the regulator serves to adapt the generator voltage to the battery voltage. Thereby the discharged battery must be recharged as rapidly as possible, yet without exceeding the permissible maximum voltage for the load. The charging current in a fully charged battery must be low in order to avoid overloading or damage of the battery.

The regulator for dc generators has three functions, i.e. that the voltage controller, current limiter, and directional circuit breaker. The regulator for three-phase current generators, on the other hand, needs only to respond to the voltage. The function of the directional circuit breaker is taken over by the built-in diode-bridge and a current limitation is not necessary.

The controller maintains the generator voltage at battery charging voltage level (in a 12 V battery it would be 14.3 V), by switching the current in the field coil EW on and off. During this process the range of the switch-on/switch-off ratio may lie between 0 and ∞, and the oscillation frequency may reach 3 kHz.

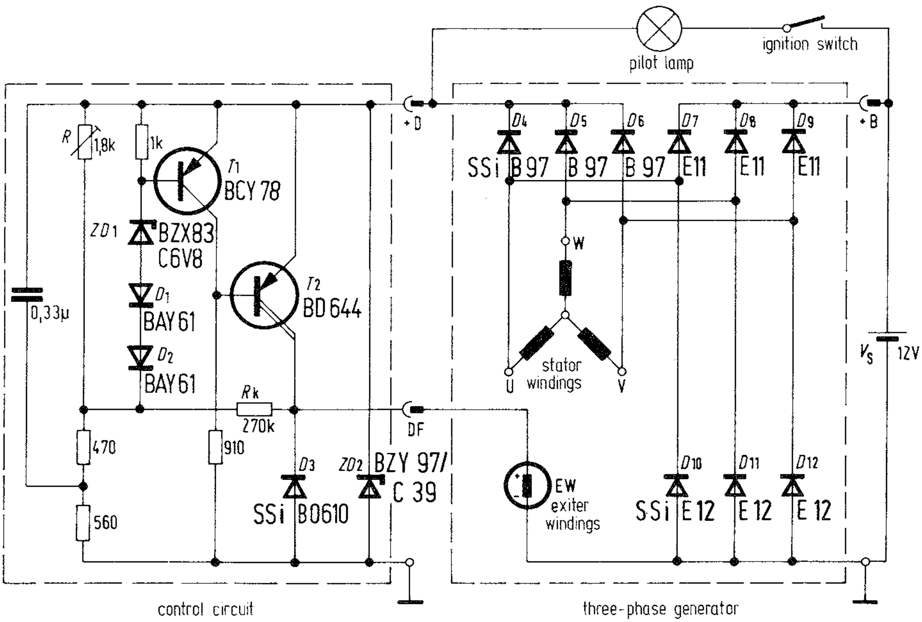


Fig. 5.14.1.

The ripple (± 0.15 V) is exclusively determined by the behavior of the generator during switching; the switching frequency, on the other hand, depends mainly on the momentary load and rotational speed of the generator. The superimposed hum voltage is produced by the bridge rectifier for the three-phase voltage.

Capacitor used in the Circuit 5.14.1.

1 pc	MKM-stacked film capacitor	0.33 μF/100 V	B 32540–A 1331–J 000

5.14.1. Mode of Operation

The mode of operation of a regulator **(Fig. 5.14.1.)** becomes evident when it is observed during rising and falling generator voltage. As soon as the voltage at the resistor R of the voltage divider exceeds the z-diode voltage, the transistor T_1 becomes conductive, and the Darlington output stage, switched on until then, is turned off. Since the field coil of the generator lies in the collector circuit of the output stage, the excitation is removed from the field coil, and the exciter current dies away across the free-wheel diode in accordance with the time constant of the exciter circuit. When the generator terminal voltage falls below the nominal value given by the z-diode, T_1 cuts off, the Darlington stage BD 644 becomes conductive, and as a result the generator is again fully excited.
The generator voltage rises rapidly more or less, depending on the load of the switched-on equipment. A defined hysteresis, produced by the feedback resistance R_K, is necessary for the regulator to maintain its two-step character. The capacitor connected in parallel to the input divider suppresses the hum of the generator voltage.

Temperature Behaviour

The effect of the temperature on the generator voltage should be adapted to the needs of the battery charging voltage. Therefore the regulator is rated for an average temperature coefficient of the generator voltage of −8 mV/K. The negative effect of temperature results from the addition of the negative temperature coefficient of the base-emitter-diode of transistor T_1, the positive coefficient of the z-diode, and the negative effect of temperature on the two series diodes BAY 61 especially provided for this purpose.

Protection of the Regulator

A power z-diode, type BZY 97 C 39 (ZD 2), limiting the voltage at the transistor to about 39 V, serves as surge voltage protection for the regulator, which may not be operated without any battery connected.
The technical data are detailed in the following table.

Technical Data on the Three-Phase Generator Regulator

Rated regulator voltage	14.3 V
Maximum current	4.5 A
Operating temperature range	−40 to +100°C
Regulator thresholds:	
Upper threshold	14.4 V
Lower threshold	14.2 V
Ripple of the control voltage	0.3 V
Temperature coefficient of the total circuit with reference to the control voltage	−6.9 to 8.5 mV/K

5.15. 24 V-Regulator for Three-Phase Automotive Generators

Fig. 5.15.1. shows the circuit for an automotive regulator rated for a 30 A-three-phase dynamo. The output stage is equipped with the npn Darlington transistor BD 647. The features and the principle of operation are the same as already described in chapter 5.14. The regulator device has 4 terminals, as the control voltage is to be picked up at generator terminal +B.
The controlled voltage of this terminal is 27.4 ± 0.3 V at an ambient temperature of 20°C, a load current of 12 A and a rotating speed of 2500 r.p.m. The exciting current for the generator rotor amounts 3 A and the maximum generator speed is 8000 r.p.m. The circuit has a temperature coefficient of −13 mV/K ± 3 mV/K referred to the controlled voltage in a temperature range of −40°C to +60°C.

List of Capacitors used in the Circuit 5.15.1.

Quantity	Components		Ordering codes
1	MKM-capacitor	0.33 μF/100 V	B 32540–A 1334–J 000
1	MKM-capacitor	0.22 μF/100 V	B 32540–A 1224–J 000

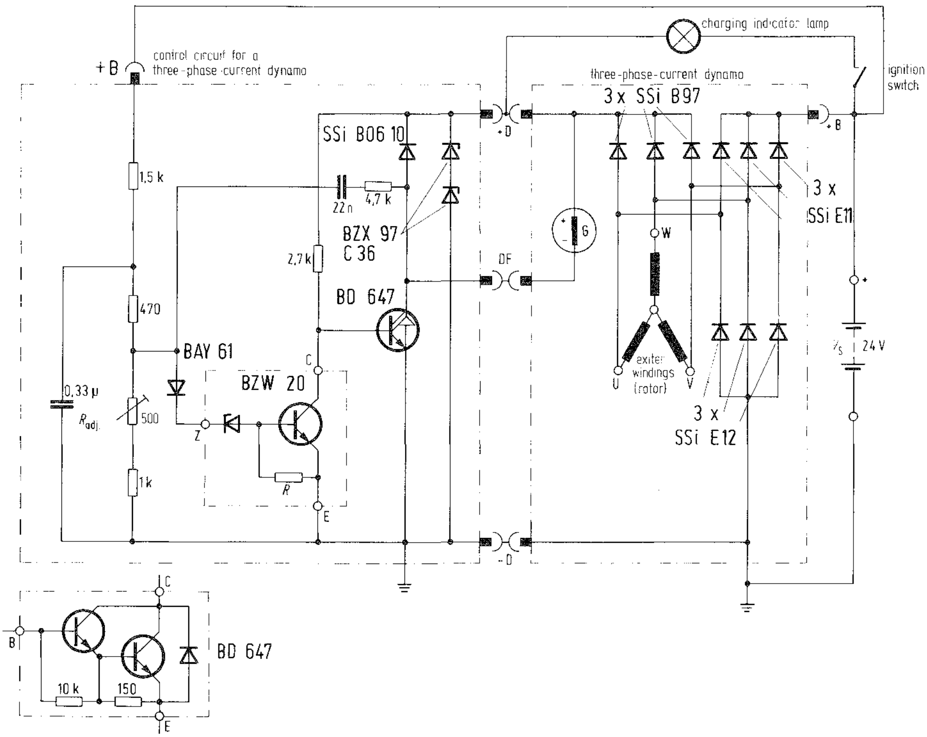


Fig. 5.15.1.

5.16. Temperature-Effect Compensation of Differential Magneto Resistor-Sensors

The so-called "field-plates" are magnetically dependent semiconductor resistors (MR's) consisting of indium antimonide and nickel antimonide. Their temperature coefficient essentially depends on the doping of the basic material. However, by using differential circuits as shown in **Fig. 5.16.1.** a suitable compensation of the temperature drift can be achieved only in a temperature interval of $\Delta T < 10\,°C$. For ranges $\Delta T > 10\,°C$ the compensation obtained by the bridge circuit is not sufficient. If however, the same temperature response of the output voltage V_0 and the internal resistance of a MR-sensor consisting of L-material is utilized an improvement can be attained **(cf. Fig. 5.16.2.)**. If the MR-sensor FP 210 L 100 is connected to a load having an input impedance $Z_i = 0$ the output current will depend on the quotient of V_0 and R_i, assuming the temperature response of both quantities is the same. In the circuit according to **Fig. 5.16.3.** the load having an impedance of zero consists of an operational amplifier the input of which remains on the same level. The current is supplied from the output of the op amp to the input via a feedback resistor R. For the output voltage it applies $V = i \times R$. The resistance of the resistor r depends on the premagnetization of the MR's. As wider the temperature range, at which the circuit has to operate, as smaller the resistance of r has to be chosen ($r \rightarrow 0$).

By the circuit shown in Fig. 5.16.3. the required supply current and the heating of the bridge resistors are essentially reduced.
An output signal temperature dependency of less than 5% is achievable in a temperature range of −40 °C to +130 °C, although the internal resistance of the MR-sensor varies more than 50% in this temperature interval.

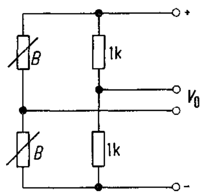


Fig. 5.16.1.

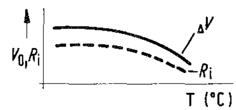


Fig. 5.16.2.

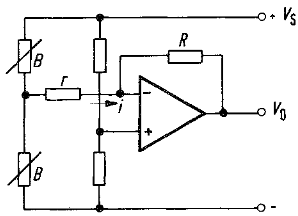


Fig. 5.16.3.

5.17. Position Indicator using a Differential MR-Sensor

In conjunction with a switching circuit showing an hysteresis effect a magneto resistor sensor, e.g. type FP 210 D 250, can be utilized to build up a position indicator. The sensor is influenced by a soft iron sheet having a thickness of 2 mm. The air gap between sensor and iron sheet should be about 0.2 mm **(cf. Fig. 5.17.1.)**. When the soft iron sheet is moved in front of the MR-sensor an output voltage V_{MR}, the shape of which is shown in **Fig. 5.17.2.** is generated.
When the iron sheet is displaced to the right a maximum and thereafter a minimum of the output voltage is received, assuming that terminal 1 is connected to the minus pole of the power supply, and terminal 3 to the plus. For the following signal processing a switching stage with a hysteresis is required. It remains in the state of the last extreme value having past **(see Fig. 5.17.4.)**.
The position of the soft iron sheet can be concluded directly from the output voltage of the switching stage if a linear displacement sensing should be required.

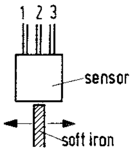


Fig. 5.17.1.

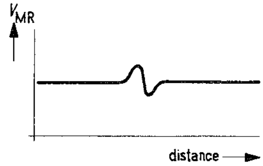


Fig. 5.17.2.

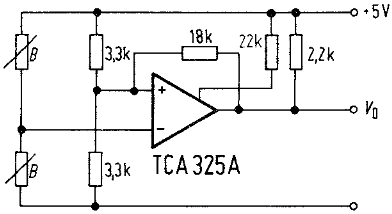


Fig. 5.17.3.

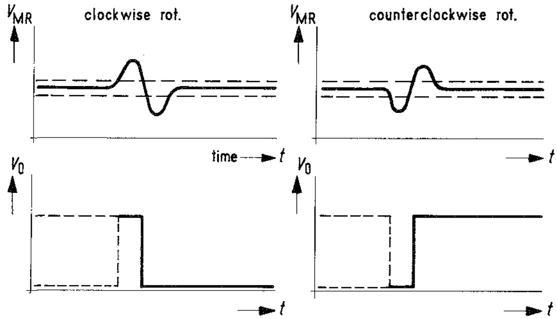


Fig. 5.17.4.

5.18. Sensing the Direction of Rotation by using a Differential MR-Sensor

If a magneto resistor sensor, e. g. type FP 210 D 250, is used for sensing the direction of a rotating toothed wheel an asymmetrical relation of tooth width to tooth space is necessary. The following dimensions are recommended for the toothed wheel **(Fig. 5.18.1.)**.

tooth height	h: $\geqq$ 2 mm	tooth width	b: 2 mm
tooth thickness	d: $\geqq$ 3 mm	tooth space	l: 6 mm

The clearance between tooth wheel and sensor should be less than 0.5 mm.
The hysteresis of the switching amplifier is rated in such a way that the output voltage range of the sensor not being magnetically influenced, is fully covered.

The centre symmetry defined as $M = \frac{R_1 - R_2}{R_1}$ (whereas $R_1 > R_2$) is less than 10% for sensors of the FP 210-family. Therefore the output voltage of a sensor not being influenced ranges between 47.3% and 52.7% of the supply voltage. Assuming an operating voltage of $V_S = 5$ V the hysteresis range of the amplifier will be 2.36 to 2.64 V. If a range of 2.32 and 26.8 is choosen for safety purposes, the sensing circuit has to be dimensioned as indicated in **Fig. 5.18.2.**

Conclusions for the rotating direction of the tooth wheel can be drawn from the average value of the output voltage, as demonstrated in **Fig. 5.18.3.**
Since the relation of tooth width to space is asymmetrical an essential difference for the average values of the output voltage is received when the wheel rotates clock or anti-clockwise. But the repetition frequency is reduced. An optimum is achieved at a ratio of 1 : 3, whereby half the highest possible frequency is obtained. The voltage depending on the rotation direction ranges between 25% and 75% of the output voltage variation of the switching stage.

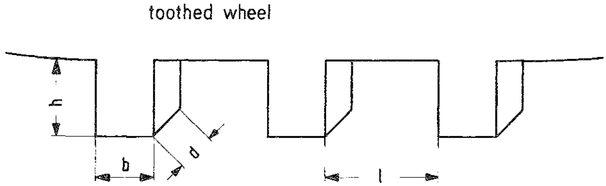


Fig. 5.18.1.

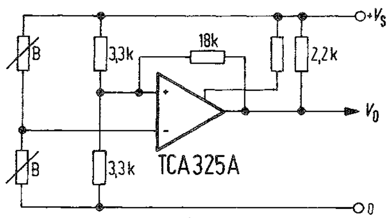


Fig. 5.18.2.

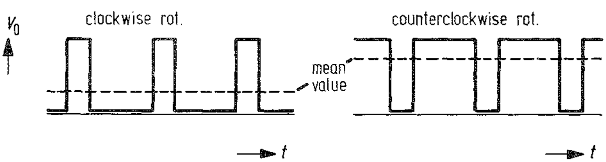


Fig. 5.18.3.

6. Power Supply Circuits

6.1. Push-Pull Chopper with an Adjustable Output Voltage

Fig. 6.1.1. and **Fig. 6.1.2.** show the basic circuits for series-driven push-pull choppers operating with a half-wave rectifier instead of the commonly used push-pull or bridge rectifiers. The latter mode of rectification offers the advantage in, at a duty cycle of 1 : 2, that the elaborateness for the smoothing can be kept extremely low. At duty cycles of $\gtrless 1:2$ only one diode is required. If the duty cycle is changed from e.g. 1 : 2 to 1 : 20, the output voltage can be varied in a ratio of 1 : 10, whereby it remains constant with the exception of the internal (low) voltage drop regardless whether the output is loaded or not. By combining a push-pull chopper and a half-wave rectifier adjustable voltage regulators can be realized and their power requirements are low, if semiconductors are employed. A circuit connected from the output to the generator can very constantly control the adjusted output voltage, whereby the internal impedance becomes very low.
Complementary pnp/npn transistors or types in the quasi-complementary npn/npn configuration can be used. There are no levels generated higher than the supply voltage, even if no load is connected to the output. The advantages of this concept are as follows:

1) Small transformers, e.g. EE 42 for about 100 W output
2) Small heat sinks for the transistors
3) The oscillating frequency is beyond the range of audibility
4) Adjustable output voltage at a load range: full to zero
5) The output voltage is very constantly controlled, $R_1 \rightarrow 0$
6) The reverse voltage of the transistors corresponds to the supply voltage

The circuit shown in **Fig. 6.1.1.** has been practically tested for $V_1 = 60$ V and $V_2 = 1$ to 13 V at 1.5 A. The circuit according to **Fig. 6.1.2.** is a suggestion for higher supply voltages V_1. However, certain measures are required in order to realize a large variation of the duty cycle.

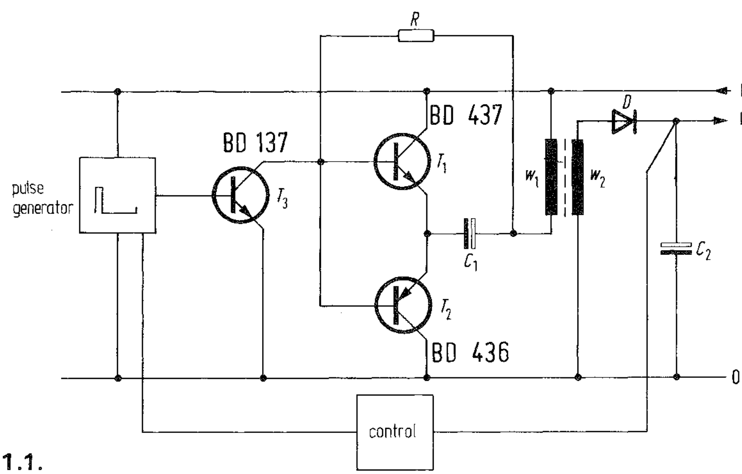


Fig. 6.1.1.

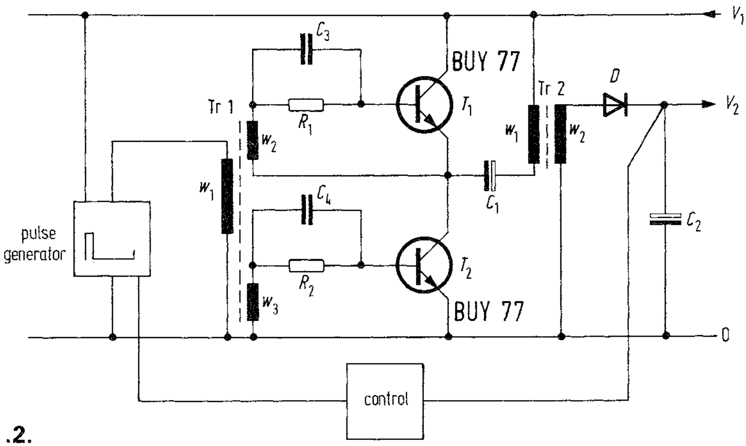


Fig. 6.1.2.

List of Capacitors used in the Circuit 6.1.2.

Item	Quantity	Components	Ordering codes
C1	1	100 μF/350 V	B 43306–B 4107–T 000
C2	1	100 μF/160 V	B 43050–E 1107–T 000

6.2. Low-Loss Power Supply 48 to 60 V/5 V, 300 mA for Digital IC-Operation and LED Displays

LED-displays requiring low power can advantageously be used in power supply systems with higher dc voltages, e.g. 60 V at telecommunication equipments. But the relatively high voltage has to be converted to the low level required by a LED-display without essentially dissipating power. If in the mentioned example a required output current of 300 mA at 5 V has to be produced from the supply voltage of 60 V by a circuit consisting of a resistor and a z-diode, the minimum power dissipation would be 19 W. If, however, the self-oscillating power supply described in **Fig. 6.2.1.** is used, the power dissipation will be reduced to about 1 W. The capacitor C_1 being connected in parallel to the load is charged via the switching transistor BSV 17 and the coil L by short, triangular current pulses. Their duration essentially depending on C_1 and L is finished when the upper threshold is reached and when the transistor BSV 17 is turned off via the op amp serving as a voltage comparator. In the course of the pulse duration most of the energy coming from the power supply is stored in the coil and it is immediately transferred to the capacitor C_1 via the diode BAY 45 when the pulse space begins. Because of the principle of operation a hum voltage is imposed upon the output voltage, being adjustable through the potentiometer P, since the following pulse is going to be generated only if the capacitor C_1 is discharged via the load to the lower threshold. As the capacitor C_1 is a part of the oscillator, its capacitance cannot be indefinitely increased to reduce the hum voltage. This can only be obtained by an additional filter circuit (C_2 and 1.5 Ω-resistor) as shown in the Fig. 6.2.1.

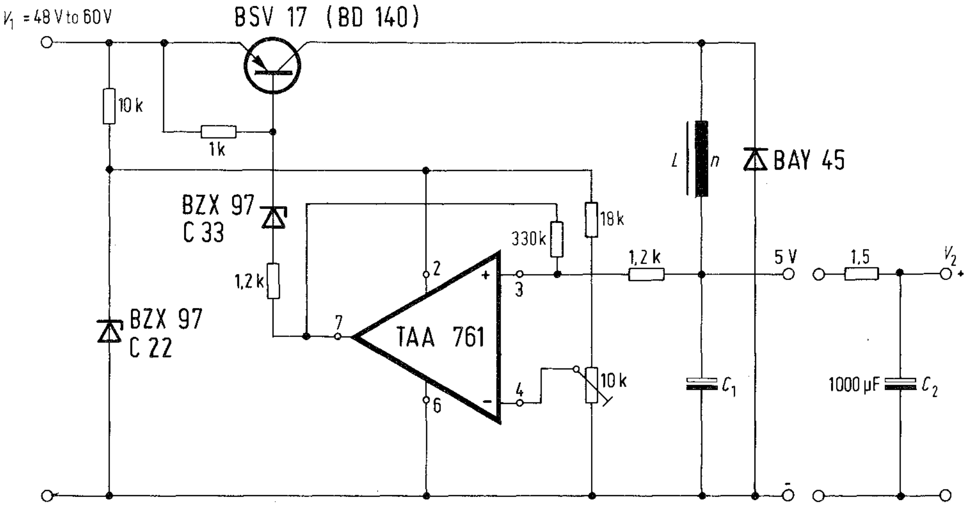


Fig. 6.2.1.

Technical Data:

Supply voltage	48 to 60 V
Maximum input current	42 mA (at 60 V) 50 mA (at 80 V)
Oscillation frequency	300 Hz to 1 kHz
Pulse duration	<120 µs
Output voltage	5 V (3 to 7 V, adjustable)
Output current	0.3 A maximum
Efficiency	60% approx.
Transistor peak current	0.7 A

Coil: Pot Core 26∅ ± 16
B 65671–L 0250–A 22
n = 230 turns, 0.3 enamelled copper wire

List of Capacitors used in the Circuit 6.2.1.

Item	Quantity	Components	Ordering codes
C1	1	470 µF/25 V	B 41010–B 5477–T 000
C2	1	1000 µF/25 V	B 41010–B 5108–T 000

6.3. Constant-Current Two-Pole for Driving Discrete LED's in a wide Voltage Range

A simple constant current source, as shown in **Fig. 6.3.1.** is recommended for operating discrete LED's when a constant brightness is required even within a wide supply voltage range. The LED-current is regulated by a circuit consisting of discrete

components. The transistor BC 168 allows an operation at a maximum voltage of 30 V. The required current is calculated as follows: $I_d = \frac{V_{BE}}{R} \approx \frac{0.7\ V}{R}$. Assuming $R = 33\,\Omega$, a current of 20 mA ± 10% is realized in a range of 4 to 20 V.
The brightness variation is negligible and practically invisible. It does not increase exactly proportional to I_d at static operations. Since the brightness decreases, however, because of increased heat, a compensation effect is attained.
The power dissipation of operations using a series resistor, raises according to a 2nd-order function, but contrary to that the power dissipation of the described circuit increases only linear.

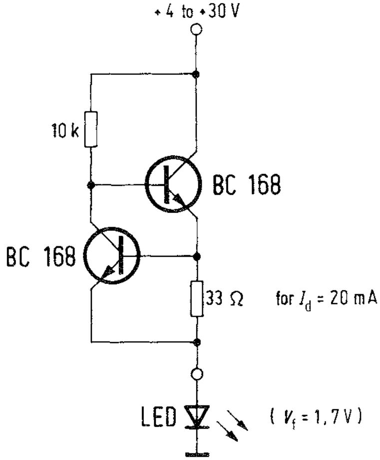


Fig. 6.3.1.

6.4. Voltage Converter for the Operation of LCD's

Summary:

Fig. 6.4.1. shows the circuit of a self-oscillating blocking oscillator which is especially favoured for generating the supply voltages required for the operation of a quartz clock with LCD-devices (e.g. AN 4132). The voltage converter supplies a maximum output of 750 µW at the required output voltage of 15 V. The efficiency ranges between 70 and 75% depending on the output power.

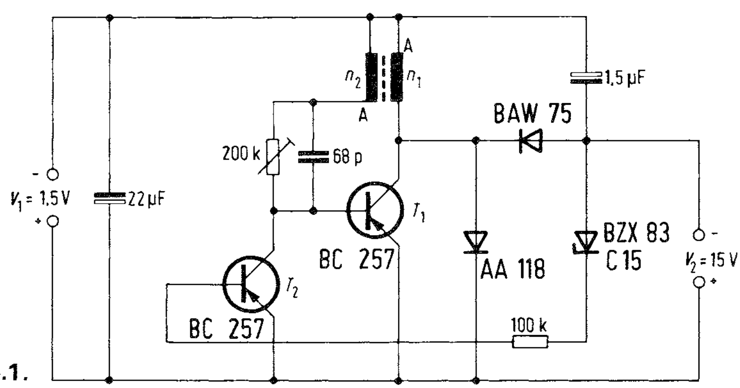


Fig. 6.4.1.

The clock-IC's manufactured in C-MOS technique continuously used today require a supply voltage of 1.5 V for the frequency divider and 15 V for the LCD-driver. Whilst for the 1.5 V-supply conventional tubular cells are available, the voltage for the driver-IC has to be produced by means of a dc voltage converter, which has to meet the following requirements:

1) High efficiency
2) Load-independent output voltage between full- and no-load operation
3) Small dimensions.

Today externally triggered blocking oscillators are frequently used in clock circuits. In this case the trigger pulses are supplied from the frequency-divider-IC. Due to the very short pulse duration of about 10 to 13 μs and the relatively long pulse separation of 1 ms a peak current of ≧200 mA is required for transmitting the energy to the blocking inductor. This high current produces a voltage drop of about 0.3 V across the switching transistor and therefore the efficiency is relatively slack (about 30 to 35%). Better results are achieved by using self-oscillating blocking oscillators, by which an efficiency of ≧70% is attainable at 15 V and at a current of about 40 μA usually required for operating a Siemens 4-digit LCD-devices. The cost for both kinds of blocking oscillators will presumably be the same.
A peak current maximum of 2 mA is allowed if an efficiency optimum is forcibly required. According to the data sheet the collector-emitter-saturation voltage of the transistor BC 257 is below 0.1 V at this current. A pot core with the dimensions 9∅ × 5 mm, N 30 without airgap and an A_L-value of 2500 nH/n^2 has been chosen as a compromise between achievable inductance and mechanical dimensions. An inductance of about 50 mH can be realized by 130 turns (0.1 enamelled copper wire), which are easily placed on the bobbin. Thus a peak current of about 1.8 mA is realized at a pulse interval of about 84 μs.
A 68 pF-capacitor is connected in parallel to the 200 kΩ-potentiometer, limiting the base current, in order to turn off sufficiently fast the switching transistor T_1. The output voltage is regulated by the z-diode BZX 97 C 15 and the transistor T_2. When the z-voltage is reached, the latter is turned on and interrupts the energy transmission to the coil.
The blocking oscillator being operated with a 1.5 cell and shown in **Fig. 6.4.1.** supplies a load-independent output voltage of 15 V. Its efficiency is 73% at an output of 750 μW (corresponding to 50 μA).

Coil Data

Siferrit pot core:	9∅ × 5 m, N 30 without air gap, B 65517–A 0000–R 030
Windings:	n_1 = 130 turns, 0.1 enamelled copper wire
	n_2 = 20 turns, 0.1 enamelled copper wire

Adjustment procedure for full load

Load the output with a resistance of 300 kΩ and adjust the load current to a value of 50 μA by the 200 kΩ-potentiometer.

List of Capacitors used in the Circuit 6.4.

Quantity	Components	Ordering codes
1	1 µF/63 V Electrolytic capacitor	B 41316–C 7105–Z 000
1	22 µF/63 V Electrolytic capacitor	B 41286–A 7226–T 000
1	68 pF/250 V Ceramic capacitor	B 38612–J 2680–J 008

6.5. Power Supply 24 V/6.5 V, 17 W

The following solution demonstrates how a 6 V-lightning installation can be operated by a supply voltage of 24 V. The switch mode-circuit described in **Fig. 6.5.1.** features a high efficiency. Its clock frequency varies in a wide range according to the load- or the input voltage-changes. If the output voltage should be too low the two transistors are switched on by the comparison amplifier and the smoothing capacitor is charged via the choke as long as the required level at the output is again reached. Then the comparator flops into its blocking state due to the given hysteresis (2.2MΩ/5.6 kΩ). Thus the transistors previously conductive are also turned off.
The energy stored in the inductor is additionally transmitted via the diode C 2605 to the smoothing capacitor being more highly charged. However, the capacitor is also discharged by the amount of the hysteresis voltage of the amplifier due to the influence of the connected load. Then the switching process starts again.

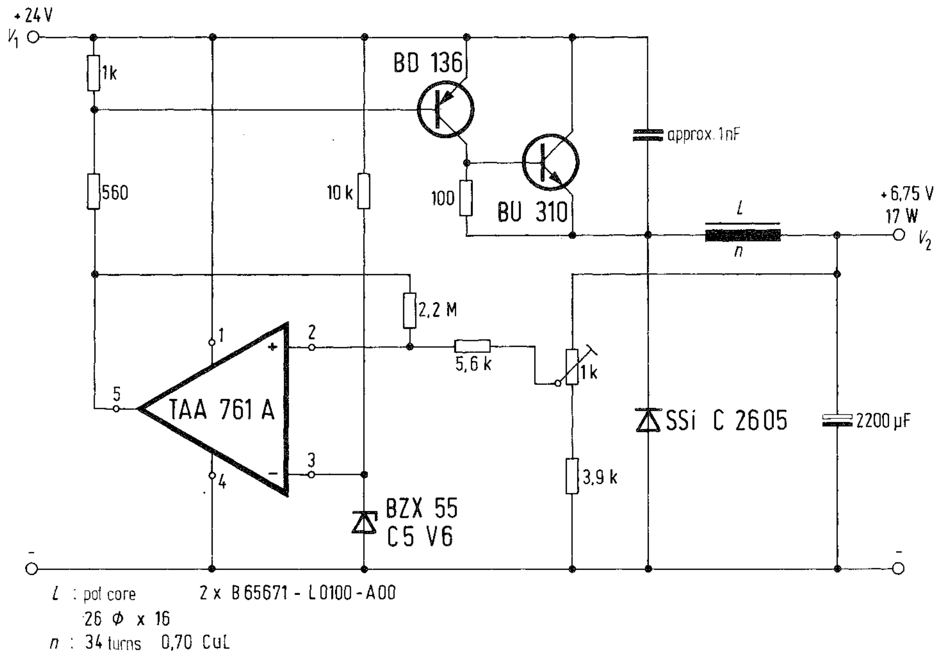


Fig. 6.5.1.

The smoothing capacitance has to be rated according to the given hysteresis voltage of the switching amplifier. The choke L is not to be operated in its saturation state, if a high efficiency is required. This is indicated by the near triangular shape of the charging and discharging choke current.
If a higher power output is desired, a smaller inductance is required, because of the charging peak current. Since this current, however, results in a stronger saturation, a greater pot core is necessary.

List of Capacitors used in the Circuit 6.5.1.

Quantity	Components	Ordering codes
1	2200 µF/16 V Electrolytic capacitor	B 41012–C 4228–T 000
1	1 nF/250 V Ceramic capacitor	B 37235–J 5102–S 001

6.6. Regulated Power Supply for a Projector Lamp 12 V/50 W

For the operation of halogen projector lamps it is very essential that their supply voltage does not depend on mains voltage variations, because they are very sensitive against excess voltage. Their life time is disproportionately reduced if only a few percent of an overvoltage exist. The circuit shown in **Fig. 6.6.1.** comprises an oscillator operating at about 20 kHz. The 12 V-halogen lamp is directly driven by the 20 kHz-sine-wave voltage, i.e. mains-isolated and without any rectifier. A sufficient regulation of 3 to 4% is attained in a voltage range of 200 to 220 Vac, whereas only a 1%-variation of the lamp voltage occurs between 220 and 240 Vac. The regulation is achieved by a constant collector current of the transistor T_1 ($R_i \rightarrow \infty$) being controlled by a z-diode BZY 97 C 9 V 1 in conjunction with the 8.2 Ω-emitter resistor.
A relatively high feedback is realized to decrease the power dissipation of the transistor and to guarantee a safe switch-on of the oscillation. The bias for the beginning operation is clamped by the diode BA 127. The diode B 2510 C allows a control with sine-peaks at the so-called class C-operation.

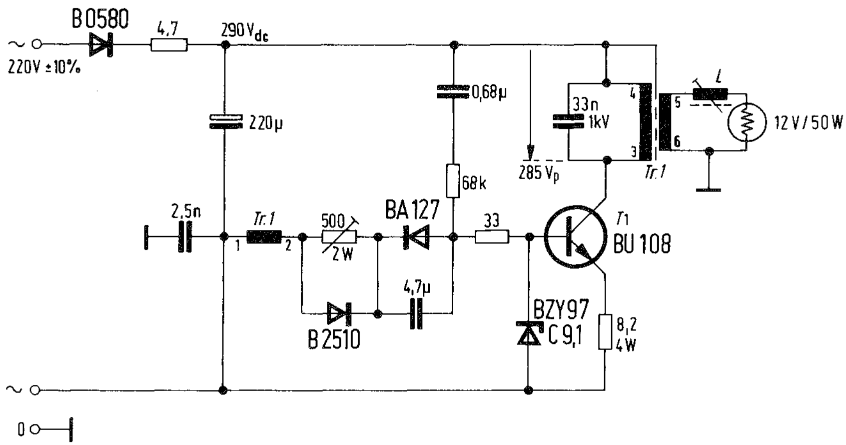


Fig. 6.6.1.

The beginning of the control at the transistor is adjusted by the 500 Ω-potentiometer at 220 Vac, in order to avoid that the power dissipation of T_1 increases too much at 250 Vac (thermal resistance of the required heat sink = 3 K/W). If the lamp is short-circuited the oscillation is immediately interrupted. At an operation with no load, however, the voltage at T_1 raises to about 950 V, whereby the power dissipation at T_1 is essentially reduced. For the matching of the lamp either a small inductance can be connected in series or the air gap has to be varied a little bit.
Always when the lamp is switched on a total short-circuit occurs. Since the transformer has a magnetic leakage being considerably high, the leakage inductance is practically connected in parallel to the primary one. The circuit only oscillates at a higher frequency, e.g. 25 kHz instead of 20 kHz, as long as the lamp filament becomes sufficiently hot, i.e. becomes a satisfactorily high resistance.

Transformer Tr_1

Siferrite U 57/28/16 (Pair) B 67334
Air gap: 1 mm each leg.

Winding 1 to 2:	10 turns,	1 × 0.45 mm ∅ enamelled copper wire	leg 1
Winding 3 to 4:	100 turns,	2 × 0.6 mm ∅ enamelled copper wire	leg 1
Winding 5 to 6:	10 turns,	6 × 0.6 mm ∅ enamelled copper wire	leg 2

List of Components used in the Circuit 6.6.1.

Quantity	Components	Ordering codes
	Resistors	
1	4.7 Ω/7 W	
1	68 kΩ/0.3 W	
1	33 Ω/2 W	
1	8.2 Ω/4 W	
1	500 Ω var. 2 W	
	Capacitors	
1	Electrolytic 220 µF/350 V	B 43306–B 4227–T 000
1	MKH 0.68 µF/400 V	B 32231–J 6684–K 000
1	33 µF/1000 V Polypropylen, low-loss	B 33066
1	MKL 4.7 µF/63 V	B 32120–E 9475–M 000
1	2.5 µF/2.5 kV	B 32237–J 2252–S 000
	Diodes, transistors	
1	B 0580	
1	BZY 97 C 9 V 1	
1	BA 127 d	
1	B 2510 C	
1	BU 108	
	Others	
1	Transformer	refer to transformer data
1	var. Inductance 5 µH/4 A	
1	Projector lamp 12 V/50 W Osram	

6.7. Regulated Power Supply 220 Vac/200 Vdc, 250 mA

The circuit shown in **Fig. 6.7.1.** supplies a constant output voltage of 200 Vdc at a constant current of 250 mA. If components with slightly other values are used this device is also suitable for a constant current of 0.5 A. If a dc/ac converter is connected to its output, for instance, a power supply is realized that is not influenced by mains voltage variations.

The series controlled transistor BU 114 (250 V/6 A or BU 111 with 400 V/6 A) is driven by a control-amplifier transistor BF 459 (300 V/100 mA) or BF 458 (250 V/100 mA).

The output voltage, being reduced by the arc drop of the one or two glow lamps, is divided and applied to the base of the BF 459 via a 12 V-z-diode. Thus a control voltage, exactly following the variations of the output level, is available at the base. If the output voltage raises, the level at the base of the BF 459 becomes more positive, i.e. the transistor becomes conductive. By that the base of the series transistor BU 114 is influenced and it becomes less conductive. Therefore the output voltage decreases.

The diode, connected to the base of BU 114 and the emitter resistor, clamps the negative base-voltage to a level of about 1 V. The two other series diodes connected with the anode to the base of BU 114 in conjunction with the 3 Ω-emitter-resistor limit the collector current of the BU 114 during the switch on. This emitter resistor sufficiently compensates gain tolerances of the series transistor as well.

With a modification of the discribed power supply the circuit shown in **Fig. 6.7.2.** comprising the emitter follower BU 110, is connected to the terminals 1, 2, and 3. As only a very low collector current is supplied via the glow lamps, an essentially improved regulation characteristic is realized.

In the circuit shown in Fig. 6.7.1. the regulation is less than ±2% when the mains voltage varies by ±10%, but with the circuit of Fig. 6.7.2. it is less than 1‰.

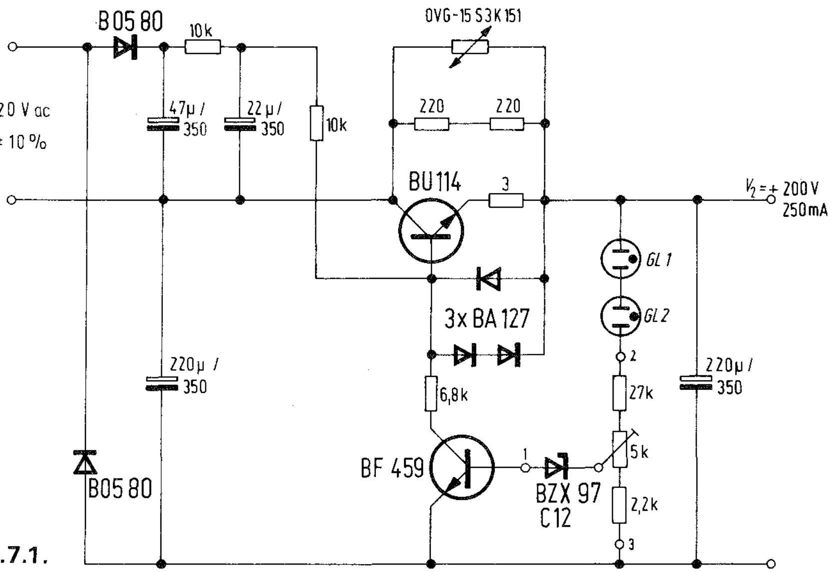


Fig. 6.7.1.

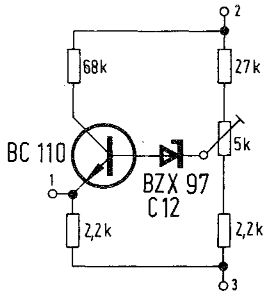


Fig. 6.7.2.

List of Capacitors used in the Circuit 6.7.1.

Quantity	Component	Ordering codes
1	Electrolytic capacitor 22 µF/350 V	B 43052–B 4226–T 000
1	Electrolytic capacitor 47 µF/350 V	B 43052–C 4476–T 000
	Electrolytic capacitors 220 µF/350 V	B 41306–B 4227–T 000

7. Digital Circuits

7.1. Input Register

Fig. 7.1. shows a 16-stage input register with the shift register FZJ 161. As soon as the start key is activated, the monostable multivibrator FZK 101 is triggered. The resulting output pulse resets via $\bar{R}$ the control counter FZJ 151, and the control flip-flop FZJ 101 to Q = L. The set inputs S of the shift registers FZJ 161 are enabled after a short delay caused by the capacitively delayed NAND-gate FZH 111, whereby the different levels of the selectors S_1 to S_{16} are considered. This is necessary as a proper information storage is assured only if set and reset inputs are simultaneously supplied with an L signal for t = 1 µs and afterwards the reset input is returned to an H-signal at least 1 µs before the set input reacts. When the start key is in its rest position again, the input gate FZH 191 is activated and clock pulses are supplied to the clock inputs of the register FZJ 161 and the counter FZJ 151. After the 16th clock pulse the control flip-flop FZJ 101 reacts and the clock gate FZH 191 is blocked by the counter output. The end of the sequence is indicated by a light emitting diode LD 461. The information stored in the register is retained, as the serial output SQ is connected to the serial input SI. The control circuit can be disabled by the repetition switch S. The register circulates the information as long as the switch S is in the position ground (L-signal). The LED's LD 468 connected to the Q-outputs of the register serve as status indicators.

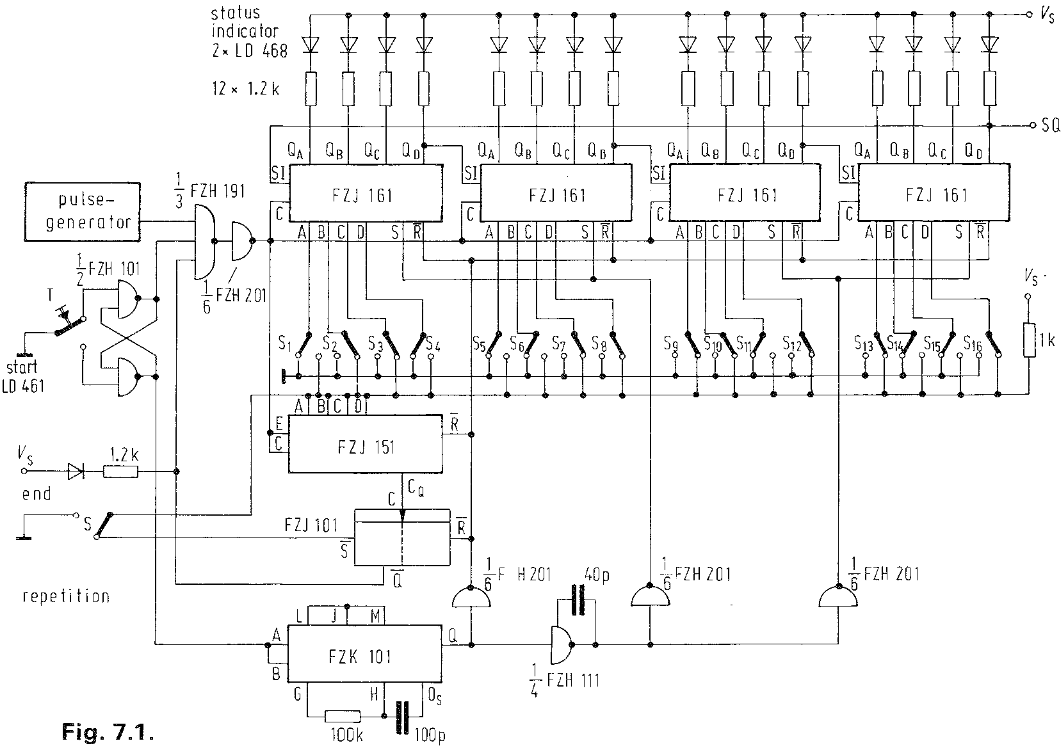


Fig. 7.1.

7.2. Arithmetic Circuits

7.2.1. Halfadder

Fig. 7.2.1. shows a halfadder which is simply realized by an exclusiv-OR-gate FZH 271.
The truth table is as follows:

inputs		outputs
A	*B*	*Q*
L	L	L
L	H	H
H	L	H
H	H	L

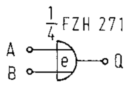


Fig. 7.2.1.

7.2.2. Halfadder with Carry Output

If the circuit shown in Fig. 7.2.1. is extended by an AND-gate FZH 251 a halfadder is attained (see **Fig. 7.2.2.**). At both the outputs the sum signal and the carry signal are available.
The truth table is as follows:

inputs		sum output	carry output
A	*A*	Σ	C_Q
L	L	L	L
L	H	H	L
H	L	H	L
H	H	L	H

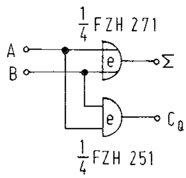


Fig. 7.2.2.

7.2.3. Fulladder

A fulladder can be realized by means of two halfadders according to **Fig. 7.2.3.** An additional gate is required for the generation of the final carry signal. The circuit is simplified if NAND-gates are used instead of AND-gates for this purpose.

inputs			sum output	carry output
A	*B*	C_I	Σ	C_Q
L	L	L	L	L
L	H	L	H	L
H	L	L	H	L
H	H	L	L	H
L	L	H	H	L
L	H	H	L	H
H	L	H	L	H
H	H	H	H	H

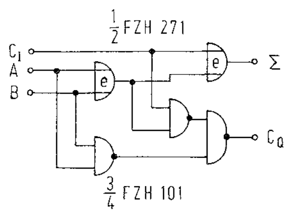


Fig. 7.2.3.

An adder network of *n* bits is formed by cascading the carry inputs and outputs as shown in **Fig. 7.2.3a.**

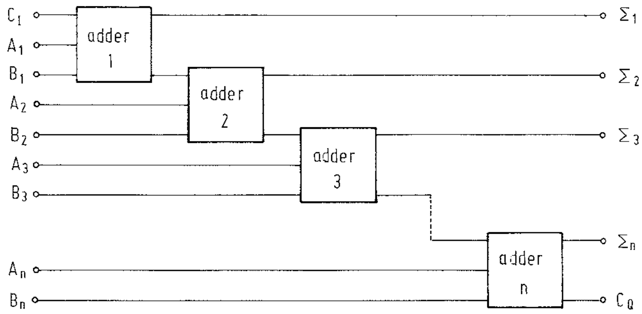


Fig. 7.2.3a.

7.3. Comparator for n Bits

A comparator which supplies the information $A = B$ and $A \neq B$ is shown in **Fig. 7.3.** No indication is given as to whether A is greater or smaller than B. The actual comparison is performed by the exclusive-OR-gate FZH 271. After having being inverted, the information is combined by the NAND-gate FZH 171. Thus output Q indicates the relation between A and B as follows:

$Q = \text{L}, \quad A = B$
$Q = \text{H}, \quad A < B \text{ or } A > B$

The circuit can be extended to n bits by means of additional diodes, type BAW 76, exclusive-OR-gates, FZH 271 and inverters FZH 201 at the extended input N_1 of the FZH 171.

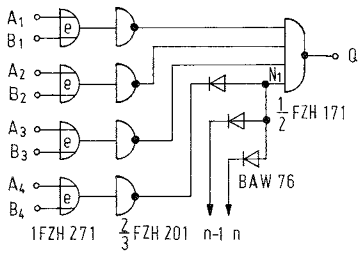


Fig. 7.3.

7.4. Pulse Circuits

7.4.1. Delay Circuits

The delay capability of LSL-gates can be used to design simple delay times up to approximately 1 s. Two basic variations are possible. **Fig. 7.4.1.** shows a delay circuit with the AND-gate FZH 251. The length of the delay is determined by the capacitor with an arbitrary capacitance added to the first gate.

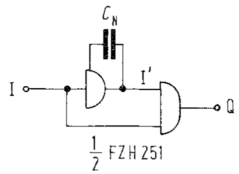


Fig. 7.4.1.

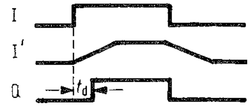


Fig. 7.4.1a.

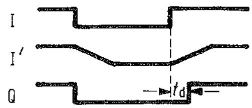


Fig. 7.4.1b.

The action on a positive pulse at the input *I* is shown in **Fig. 7.4.1a.** It can be seen that the leading edge of the output pulse is delayed whilst the trailing edge coincides with the input pulse. In the case of a negative pulse, as shown in **Fig. 7.4.1b.**, the leading edges of input and output pulses coincide whilst the trailing edge of the output pulse lags.

A delay circuit with the OR-gate FZH 291 is shown in **Fig. 7.4.1.c.** The addition of this circuit is the exact opposite of that shown above. The trailing edge of a positive pulse lags whereas the leading edge of a negative pulse is delayed. The corresponding pulse diagrams are given in **Fig. 7.4.1d.** and **e.**

Circuits which produce a lag of both the leading and trailing edges can be realized by a serial connection of the circuits shown above.

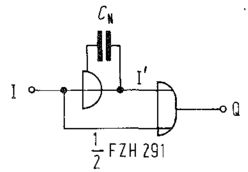


Fig. 7.4.1c.

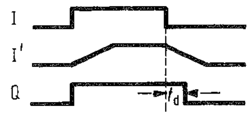


Fig. 7.4.1d.

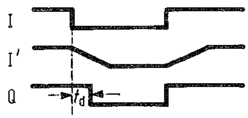


Fig. 7.4.1e.

7.4.2. Differentiating Circuits

The delay capability of LSL gates can also be used to differentiate pulses. Circuits operating this technique have advantages over those using RC-circuits, since the slope of the trailing edges is constant. Therefore dynamic inputs can be driven directly. There are two fundamental circuits possible. The first modification with the NAND-gate FZH 101 and the AND-gate FZH 251 is shown in **Fig. 7.4.2.** The time of the delay is determined by the capacitor with an arbitrary capacitance added to the first gate. A pulse is generated only at output *Q*, if the signal at input I changes from L- to H-level.

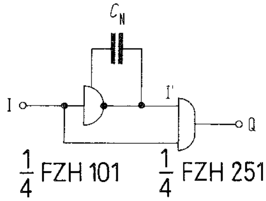


Fig. 7.4.2.

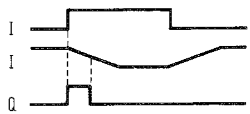


Fig. 7.4.2a.

The behaviour of the second modification with a NOR-gate FZH 281, shown in **Fig. 7.4.2b.**, is the opposite. An output signal is only generated if the signal at the input *I* changes from H- to L-level. The corresponding pulse diagrams are shown in **Fig. 7.4.2a.** and **c.**

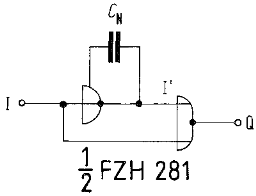


Fig. 7.4.2b.

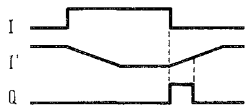


Fig. 7.4.2c.

Fig. 7.4.2d. shows a circuit using the exclusive-OR-gate FZH 271 and the delayed AND-gate FZH 251. Each change in the signal level at the input *I* generates a pulse at the output *Q*. Therefore this circuit is suitable for pulse-doubling application.

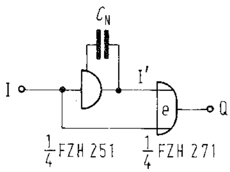


Fig. 7.4.2d.

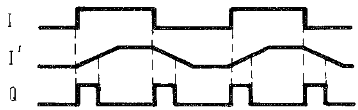


Fig. 7.4.2e.

7.5. Pulse Generator

An inexpensive self-starting oscillator can be realized with the Schmitt-Trigger FZH 241 according to **Fig. 7.5**. A feedback resistor *R* and a capacitor *C* connected to ground determine roughly the pulse duration as follows:

$$t = k \times R \times Cs$$

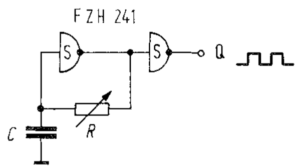


Fig. 7.5.

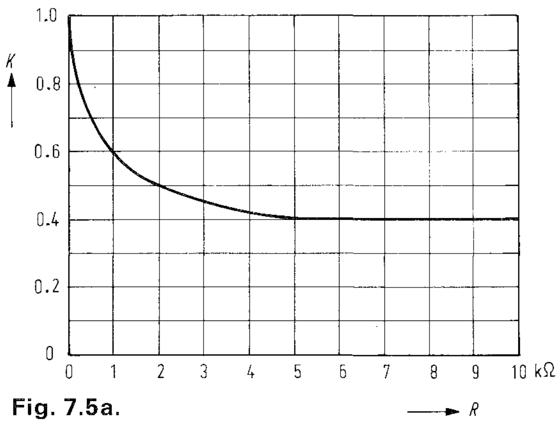


Fig. 7.5a.

The factor k depends on the resistance of R and can be derived from the typical curve of **Fig. 7.5a.** The recommended resistance variation is 0.2 k to 10 kΩ. Values below 0.2 kΩ exceed the output load factor of the FZH 241. Values above 10 kΩ reduce the charging and discharging currents to such an extent that frequency stability of the oscillator is no longer guaranteed.
Any capacitor of an arbitrary value may be used. The lowest possible pulse duration is approxymately t = 1 µs due to the propagation delay of the FZH 241.
A supply voltage variation from 11 V to 17 V will increase the pulse duration by approximately 10%. A temperature fluctuation from 20°C to −30°C will reduce the pulse duration by approximately 5% and a change from 20°C to 70°C will increase the pulse duration by 10%.
A 1 µF-capacitor should be connected directly to the supply voltage terminals to suppress possible overshooting. The capacitor is to be connected as close to the terminals of V_s and O_s as possible.
The second Schmitt-Trigger is used as a pulse shaper. Generally it is required only for feedback resistors below R = 4 kΩ.

7.6. Motor Control

Fig. 7.6. shows a control circuit of a d.c. motor operating on the principle of a windscreen wiper. When the contacts C_1 and C_2 are closing, the flip-flop circuit FZJ 111 flops. Chattering probably created is suppressed by the delaying capacitors of 68 nF and 2.2 nF. The pnp–npn output stages with BD 644 and BD 643 are controlled by the complementary outputs Q and $\bar{Q}$. The thresholds of the output stages are determined by z-diodes and are set between L- and H-level. There is an overlapping of 0.9 V between both thresholds since z-diodes with a z-voltage of 4.7 V and 5.6 V are used. Thus it is achieved that the conductive transistors are switched off before the non-conductive ones are turned on. The output currents of the gate are limited to typical 2 mA by the 1.8 kΩ-series resistors and by that loads requiring a current of 5 A max. can be driven. A sufficient cooling however is necessary.

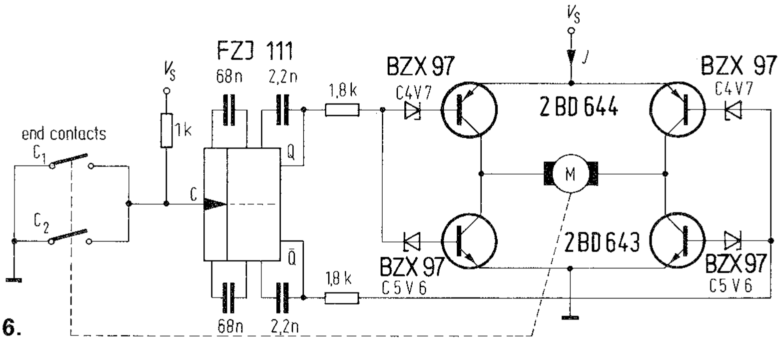
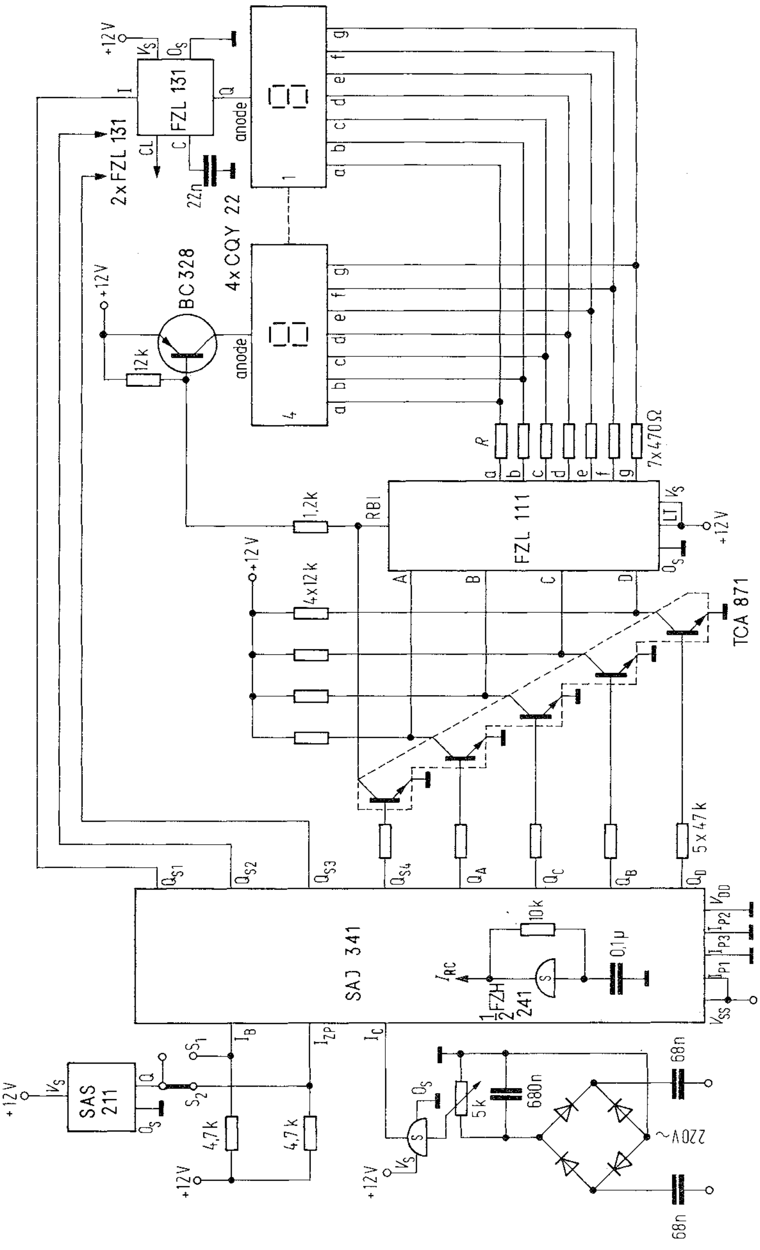


Fig. 7.6.

7.7. Digital Clock Circuits

Fig. 7.7. shows a mains-controlled digital-clock circuit using the MOS-counter SAJ 341, which has already been described in detail in the "Design Examples of Semiconductor Circuits", Edition 74/75, Section 7. Since gates of the LSL-series are used to drive the 4-digit-display device operating in a time-division-multiplex-mode only a supply voltage of 12 V is required for the total circuit. Besides that the



SAJ 341
FZL 111
FZL 131
2 x FZL 131
4 x CQY 22
BC 328
TCA 871
SAS 211
7x470Ω
4x12 k
5 x 47 k
220 V ~

Fig. 7.7.

intensity of each display segment can be individually modified by changing the resistors, if several LED's per segment are used.
The BCD-information is available at the outputs Q_A to Q_D of the SAJ 341. Since the output current of the MOS-circuit is relatively low, an op amp TCA 871 is required as an amplifier to drive the following LSL-gates. The set information is supplied by the MOS-outputs Q_{S1} to Q_{S4}. As the gates FZL 131 which control the decade displays need an input current at H-levels of only 0.1 to 0.2 mA, they can be operated directly by the outputs Q_{S1} to Q_{S3}.
The drivers FZL 131 incorporate a short-circuit protection. The IC's will only operate if a minimum capacitance of 20 nF is connected to the C-terminal. For more than one IC only one capacitor is sufficient, if the CL-terminals of all drivers are combined and the free C-terminal is connected to the V_S-terminal. The output Q_{S4} controls the driver transistor BC 328 for the 4th-digit display and the ripple-blanking-input RBI of the decoder FZL 111 suppressing the decimal 0-signal at its outputs. The decoder outputs a to g remain turned off as long as the condition $A \wedge B \wedge C \wedge D \wedge RBI = \mathrm{L}$ is satisfied. In that the 4th-digit display of the clock remains dark until ten o'clock. The segment current of the LED-display CQY 22 is determined by the output resistors R. The adjusted current flows within one fourth of the time due to the time-division-multiplex-operation. The Schmitt-trigger FZH 241 in conjunction with the RC-circuit (10 kΩ, 0.1 µF) serves as oscillator, operating at a frequency of about 10 kHz. The internal oscillator of the SAJ 341 is not used, since its frequency is too high and therefore a lightning of adjacent segments may occur if only a simple circuit for driving the display is used.
The clock is set by means of the magnetically controlled switch SAS 211, being chosen since it operates chatter-free. Hours are set in the position S_1 and the minutes in position S_2.
The SAJ 341 is programmed with $I_{P1,2,3}$ = HLL according to an operation at 50 Hz. The pulses are capacitively derived from the mains and a rectifier bridge achieves the required symmetry. The following Schmitt-trigger FZH 241 improves the noise immunity and simultaneously guarantees a definite shape of the pulse at the clock input I_C.
Fig. 7.7a. shows a simple power supply which can be operated either with a series capacitor or a mains transformer. The voltage is regulated by means of a z-diode BZY 97 C 12. The capacitance of the series capacitor is determined by the current consumption. Calculation of the maximum of the supply current without the z-diode and when the clock is in the position 10 : 28 p.m. is recommended in order to avoid an overloading of the z-diode. The following values have been experienced for the capacitance of the series capacitor:

Current mA	Capacitance µF	Ordering codes
95	1.5	B 25834–A 4155–KA or KB
100	1.6	B 25833–A 4165–KA or KB
125	2.0	B 25833–A 4205–KA or KB
140	2.2	B 25834–A 4225–KA or KB
160	2.5	B 25833–A 4255–KA or KB
190	3.0	B 25833–A 4305–KA or KB

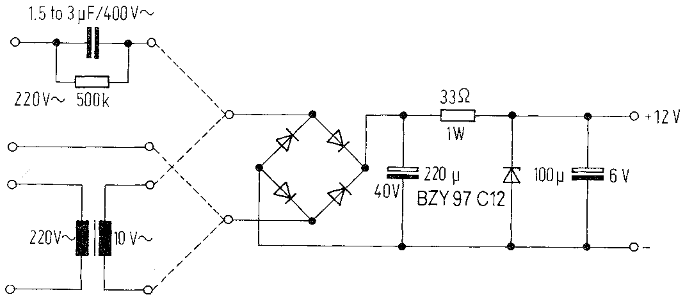


Fig. 7.7.a.

Fig. 7.7b. shows a pattern for the pc board seen from the top side, i.e. where the components are mounted.

Attention: When testing the circuit, it has to be absolutely isolated from the mains and an operation of the clock by using an isolating transformer is recommended in this case.

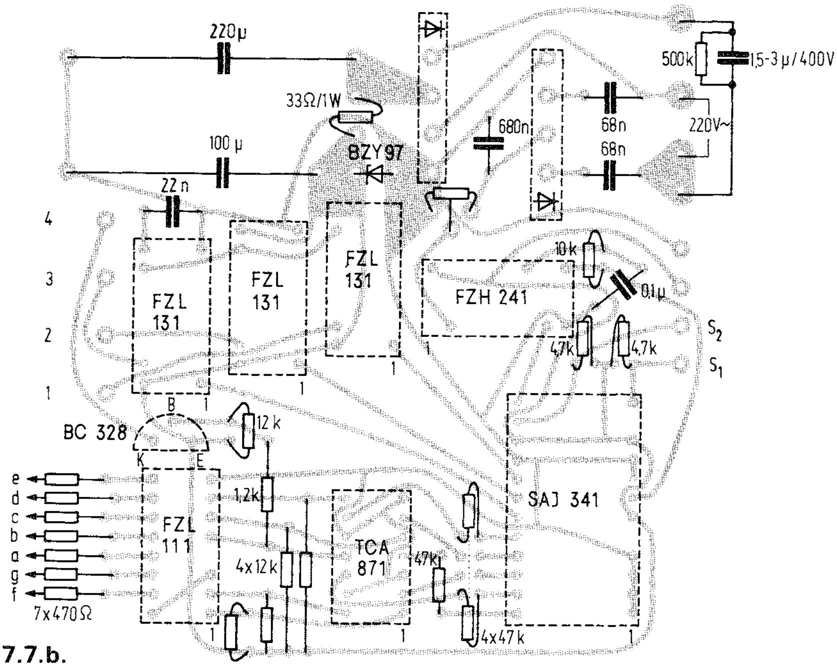


Fig. 7.7.b.

List of Components and Ordering codes

1	SAJ 341	Q 67000–J 640
1	TCA 871	Q 67000–T 2
1	FZH 241	Q 67000–H 645
1	FZL 111	Q 67000–L 156

List of Components used in the Circuit 7.7. (continued)

3	FZL 131	Q 67000–L 169
1	SAS 211	Q 67000–S 22
4	CQY 22	Q 68000–A 636
1	BC 328	Q 62702–C 312
1	BZX 97	Q 62702–Z 1237
2	B 1240	Q 66067–A 1706–A 5

Resistors	**Capacitors**	
1 33 Ω, 1 W	1 22 , 63 V	B 37449–A 6223–S 1
7 470 Ω, 1/8 W	2 68 n, 400 V	B 32220–K 3104–K 6683
1 1.2 K, 1/8 W	1 0.1 μ, 63 V	B 37449–A 6104–S 1
2 4.7 K, 1/8 W	1 0.68 μ, 100 V	B 32541–A 1684–J
1 10 K, 1/8 W	1 1.5 to 3 μ, 400 V~	see previous page
5 12 K, 1/8 W	1 100 μ, 16 V	B 41283–A 4107–T
5 47 K, 1/8 W	1 220 μ, 40 V	B 41283–A 7227–T
1 560 K, 1/4 W		

7.8. Programmable Control Circuit

The application of read-only-memories enables the user to design programmable control circuits. If a change is required in the programme only the read-only-memory has to be reprogrammed, and new development of the total circuit is not necessary. A circuit by which up to 15 devices can be controlled is shown in **Fig. 7.8.**

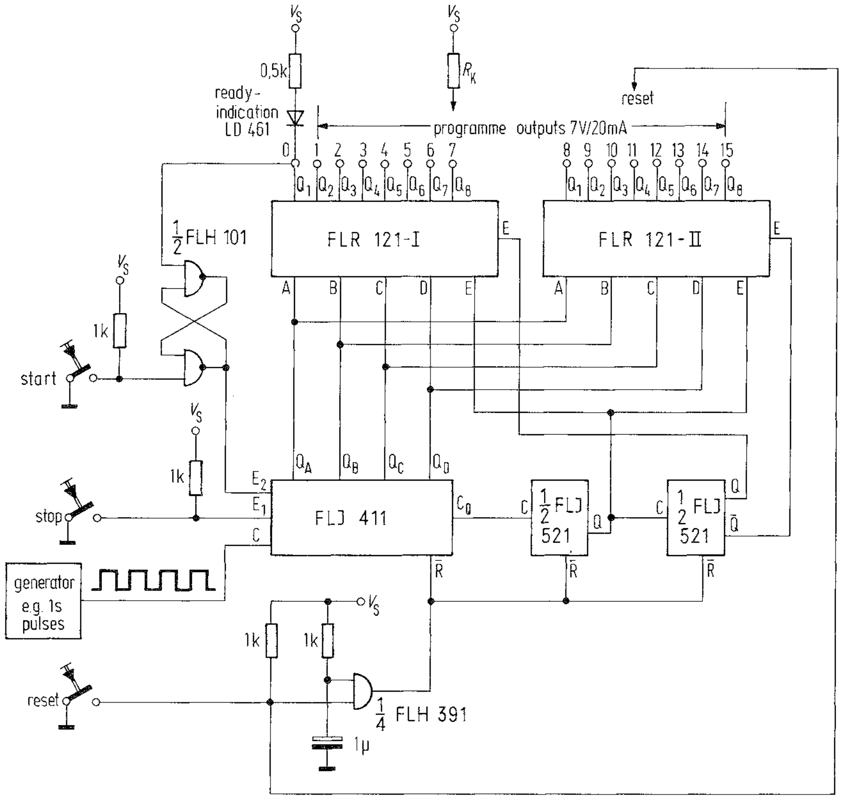
When the supply voltage is applied or when the reset push button is operated the control circuit is reset to zero-position. All programme outputs are blocked and the ready-indicator LD 461 emits light. The programme counter consisting of FLJ 411 and FLJ 521 has the state Q = L. The circuit is blocked by means of the control flip-flop connected to the ROM-output Q_1 and the counter enabling input E_2 as long as the start push button is again operated.

The programme counter is driven by pulses with a duration of 1 s. This means that the minimum operation time of the control circuits is 1 s or a multiple of same. Other times are achievable by an adequate frequency matching of the clock generator. The clock pulse has to be derived from an end position-switch or an initiator if irregular intervals are applied. The counter has 6 bit, i.e. the programme lasts for a maximum of $2^6 = 64$ s. A shorter programme can be programmed to the ROM. In this case a connection of the adequate ROM-output to the reset-circuit FLH 391 is additionally required.

The stop-push button is responsible for stopping the control in any position. When it is released the process continues.

The control circuit comprises two ROM's, type FLR 121, having open collector outputs. The minimum resistance connected to the collector may be $R_K = 250\,\Omega$ per each output. Non-programmed storage locations produce a H-signal at the output of the FLR 121.

The storage locations are binary selected at the address inputs A to E. The enabling input E serves for the selection of the storage locations. The memory I operates from 0 to the 30st second and the memory II from the 31st to the 62nd.



V_S
0,5k
ready-
indication
LD 461
R_K
reset
programme outputs 7V/20mA
0 1 2 3 4 5 6 7
8 9 10 11 12 13 14 15
FLR 121-I
FLR 121-II
1/2 FLH 101
start
stop
1k
FLJ 411
1/2 FLJ 521
generator
e.g. 1s
pulses
reset
1/4 FLH 391
1μ

Fig. 7.8.

The following list contains an example of a control programme.

Time	Counter position = address 2E	1E	E	D	C	B	A	Deci. Equivalent	Outputs L-level	H-level	Function
0	H	L	L	L	L	L	L	0	0	1 to 15	display ready →
0	H	L	L	L	L	L	H	1	1	2 to 15	
1	H	L	L	L	L	H	L	2	1	2 to 15	Control element 1
2	H	L	L	L	H	H	H	3	1	2 to 15	
3	H	L	L	L	H	L	L	4	2, 3, 4	1, 5 to 15	Control elem. 2, 3, 4
4	H	L	L	L	H	L	H	5	3, 4	1, 2, 5 to 15	Control element 3, 4
5	H	L	L	L	H	H	L	6	4	1, 2, 3, 5 to 15	Control element 4
.				.				.	.	.	.
.				.				.	.	.	.
.				.				.	.	.	.
.				.				.	.	.	.
29	H	L	H	H	H	H	L	30	7	1 to 6, 8 to 15	Control element 7
30	H	L	H	H	H	H	H	31	7	1 to 6, 8 to 15	
31	L	H	L	L	L	L	L	32	8	1 to 7, 9 to 15	Control element 8
32	L	H	L	L	L	L	H	33	8	1 to 7, 9 to 15	
.				.				.	.	.	.
.				.				.	.	.	.
.				.				.	.	.	.
.				.				.	.	.	.
54	L	H	H	L	H	H	H	55	13	1 to 12, 14, 15	reset

The ROM is programmed by the manufacturer, if a programme has been added to the order. Otherwise, the circuits will be supplied less programme.
A suitable device that applies exactly to the programming conditions indicated by the manufacturer is prerequisitely required for the programming. However, it has to be considered that other manufacturers have different conditions.

7.9. Extention for a ROM

The capacity of a ROM can be extended through a series operation of several ROM's by means of the enabling inputs *E*. **Fig. 7.9a.** shows the appropriate circuit for an extention of up to 8 FLR 121. The two binary counters serve as programme counter. The address inputs *A*, *B*, *C*, *D*, *E*, connected in parallel for all memories are controlled by the first 5 bit of the programme counter. The binary decoder FLY 151 decodes the last 3 bit of the counter and controls the enabling inputs *E*. This means, that the enabling is transmitted to the next stage after $2^5 = 32$ steps.
Fig. 7.9b. shows a circuit for the extention of up to 16 FLR 121, operating on the same principle as previously mentioned. The programme counter is, however, extended for 1 bit to 9 bit by means of the Flip-Flop FLJ 351. The circuit FLY 151 is replaced by the binary decoder FLY 141.

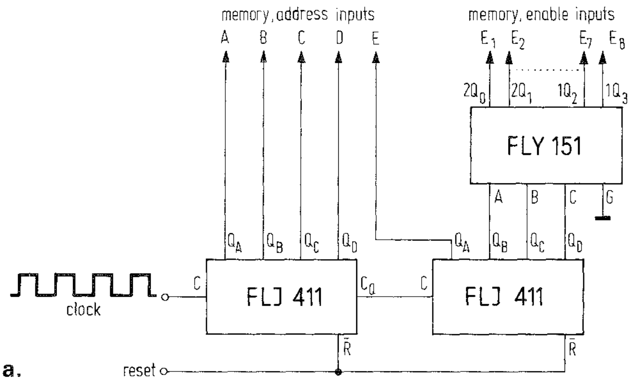


Fig. 7.9a.

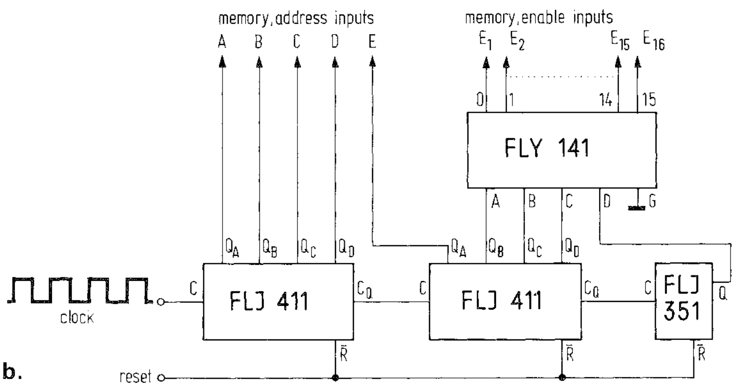


Fig. 7.9b.

The extention can be applied for more complex memory systems by using the same method. The following stage, however, requires a decoding achieved in two groups. If e.g. a programme counter with 3 FLJ 411, i.e. 12 bit, is used, the following scheme applies.

1st to 5th bit = all address inputs *A* to *E*
6th to 9th bit = all inputs *A* to *D* of the second group,
total eight 4-bit-binary decoder.
10th to 12th bit = input *A*, *B*, *C* of the first group,
one 3-bit-binary decoder.

The decoder outputs of the first group are connected to the strobe inputs of the decoders belonging to the second group. Total $2^7 = 128$ memories can be serially enabled by this method from the 6th to the 12th bit.

7.10. Actual-to-Desired Value Comparator

Control processes often require several adjustments of desired values, as with determining different end positions or for variable countings. Programmable ROM's in this case offer the advantage in that any desired values can be realized if it is within the specified capacity of the PROM. **Fig. 7.10a.** shows a circuit comprising a 256-bit-PROM FLR 121. The circuit has 8 outputs so that $2^7 = 128$ values are possible. Total $256 : 8 = 32$ memory addresses are available, i.e. 32 desired values can be programmed.

The preselection of the desired values is realized by the 2-digit BCD-selector switch. The binary BCD-converter FLH 561 converts the BCD-information to the equivalent address code of the PROM. The comparison of the desired value A and the actual one B is maintained by the 4-bit-comparator FLH 431. When both values are equal the output $A = B$ changes from L- to H-level. It has to be considered that a signal at the output $A = B$ can be shortly simulated by pulse overlapping, produced through the simultaneous change of several variables of the actual value.

If fast counters or memories are connected to the output $A = B$ an information may erroneously be produced under certain circumstances. An improvement is achieved by changing only one variable or by an information enabling in the case of B being constant. In **Fig. 7.10a.** an enabling is realized by using a NOR-gate FLH 191 connected to the input $A = B$. An L-signal applied to the enable input allows an equality indication at the output $A = B$. An H-signal produces a L-signal at the output $A = B$ independent of all the other input conditions at the FLH 431. The enabling circuit also serves for the extraction of the decimal preselection from 32 to 39. As soon as the position 31 has passed a H-signal is available at the output Q_5 of the FLH 561 and the comparator is blocked.

An extention of the memory capacity to any quantity of digits is possible when the PROM's, type FLR 121, are additionally used. In this case all memory address inputs A to E are connected in parallel.

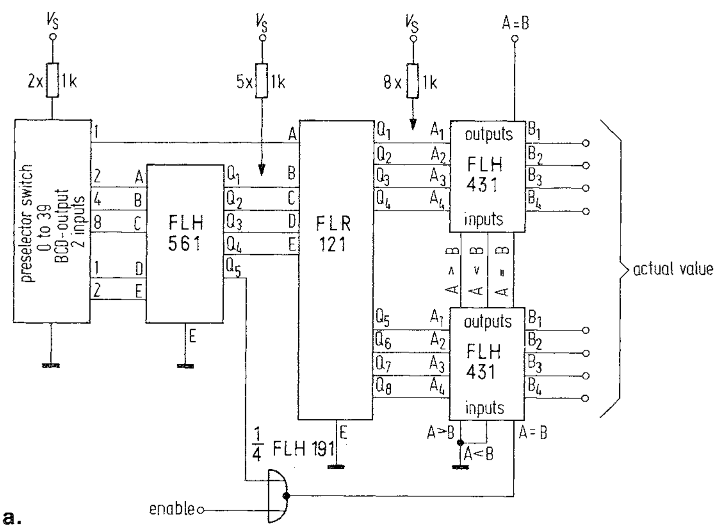


Fig. 7.10a.

Fig. 7.10b. shows a desired-to-actual value comparator incorporating the 1024-bit PROM FLR 131. The addresses are directly preselected at the inputs *A* to H by means of the 2-digit selector switch. This is economical, but limits the possible number of the desired values from 128 to 99. It has to be especially considered that the storage locations are programmed in the correct sequence. It is necessitated by driving the binary address inputs through a BCD-coded switch that the following values of the PROM can only be used:

Preselector-switch position	Corresponding memory words	Not used memory words
0 to 9	0 to 9	10 to 15
10 to 19	16 to 25	26 to 31
20 to 29	32 to 41	42 to 47
30 to 39	48 to 57	58 to 63
40 to 49	64 to 73	74 to 79
50 to 59	80 to 89	90 to 95
60 to 69	96 to 105	106 to 111
70 to 79	112 to 121	122 to 127
80 to 89	128 to 137	138 to 143
90 to 99	144 to 153	154 to 255

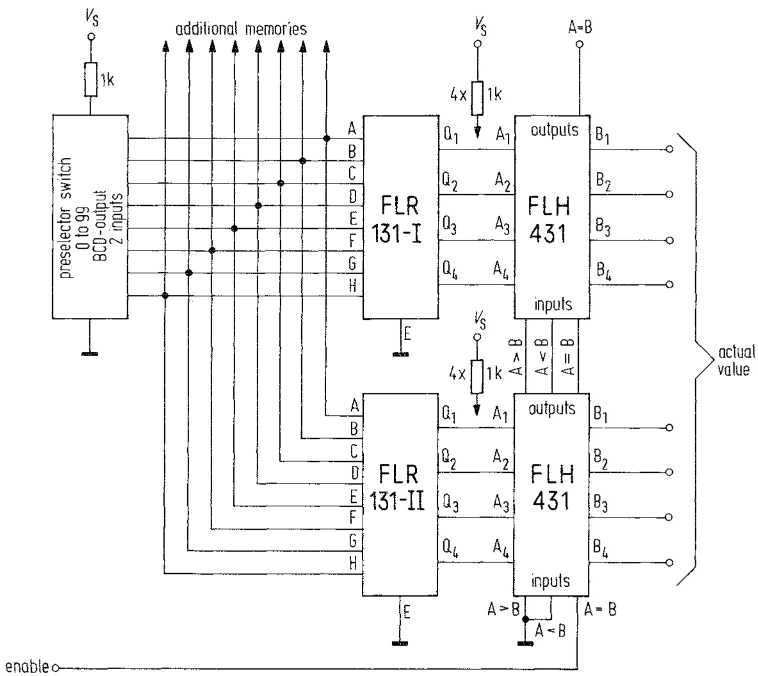


Fig. 7.10b.

The carry input $A = B$ is suited for additional enabling, as already described in Fig. 7.10a. In this case a H-level results in an indication and an L-signal in a blocking. Additional memories can be connected to the address inputs in parallel. Thus any capacity extention is possible.
The ROM is programmed by the manufacturer, if a programme has been added to the order. Otherwise, the circuits will be supplied less programme.
A suitable device that applies exactly to the programming conditions indicated by the manufacturer is prerequisitely required for the programming. However, it has to be considered that other manufacturers have different conditions.

7.11. Adjustable Pulse Generator

ROM's and RAM's are going to be more and more used for digital applications. The fast and safe reading of a special address or the shifting of an address block requires a generator, that produces a preselectable quantity of pulses. This can principally be obtained by the following solutions:

1) With a counter, the end position of which is controlled by a decoder or comparator.
2) With a preselectable reversible counter which blocks when Q = L is realized.

Circuit 1 offers the advantage against circuit 2 in that the address inputs can be directly driven by the counter outputs, if required. But circuit 2, shown in **Fig. 7.11.**, is more economical, as less components are required. It includes the reversible counter FLJ 201. The operation-mode input is not connected, which corresponds to a H-signal for reversible counting. The counters operate synchroneously and have a serial enabling by the connection from E_Q to E_I.
The carry outputs C_Q of the counters are combined by the NAND-gate FLH 121 and block the control flip-flop connected to the enable input E_I of the first stage, when it applies: Q = L and C_Q = H.
When the start push button is released the set inputs S of the counters receive L-signals and the information of the preselector switches is transmitted to the counters. The control flip-flop also being connected to the start flip-flop is blocked as long as the start button is engaged.
The Schmitt-trigger FLH 601 with the inverse-feedback 330 Ω-resistor and the 500 pF-capacitor serves as generator, oscillating at a frequency of about 5 MHz, which is adjustable by the capacitor C in a range of 1 Hz to 10 MHz. The corresponding limit capacitances are 1 mF and 100 pF. A Schmitt-trigger as a pulse former follows to the generator. Two additional Schmitt-triggers, in conjunction with an RC-circuit, consisting of a 10 kΩ-resistor and a 500 pF-capacitor produce a needle pulse at every signal change from L- to H-level of the generator. The control flip-flop is enabled when the start push button is released. But the flipping only occurs with the following needle pulse and thus a defined start is guaranteed for the counting. At the outputs Q and $\overline{Q}$ additional pulses do not appear.
The maximum operating frequency of the circuit is determined by the propagation delay of the counter during the enabling and by the additional propagation delays of the control circuits. It is 10 MHz typical and 5 MHz minimum.

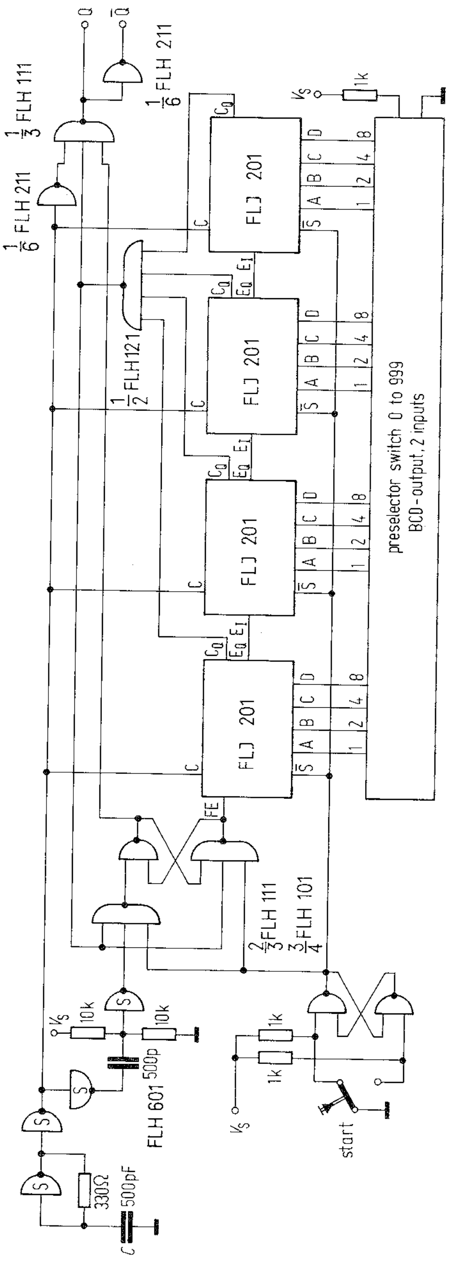

Q
$\bar{Q}$
$\frac{1}{3}$ FLH 111
$\frac{1}{6}$ FLH 211
$\frac{1}{6}$ FLH 211
$\frac{1}{2}$ FLH121
FLJ 201
FLJ 201
FLJ 201
FLJ 201
preselector switch 0 to 999
BCD-output, 2 inputs
$\frac{2}{3}$ FLH 111
$\frac{3}{4}$ FLH 101
FLH 601
500p
10k
10k
1k
1k
1k
330Ω
500pF
V_S
start

Fig. 7.11.

7.12. Electronic Burglar Alarm

Electronic warning devices are very efficiently used as burglar alarms in cars, but they are seldomly installed. The following requirements have to be met by such a protection circuit.

1) An interrupted signal is to be supplied, not a continuous one, since it is more remarcable and the surprise-effect is greater. Besides that a special signal can be chosen to be easily identified by the user of a burglar alarm. The life time of a battery is also extended by using an interrupted signal.
2) The signal has to be switched off after a certain preselected time, since a constantly emitted signal is not allowed by official regulations. Besides that battery power has to be saved and the burglar alarm device has to return to its ready-state as soon as possible.

Fig. 7.12. shows a circuit that applies to these requirements. It contains an oscillator, which rapidly switches on and off the following signal generator via a switching circuit. Thus an AM-signal is produced.

The oscillator is released by a monostable multivibrator when the operation contact is closed, i.e. it starts to oscillate. The oscillator is automatically turned off after a certain time t_M, regardless as to whether the position of the operating contact is.

The circuit is activated by a key-switch, which can be installed outside of the car. It is, however, disadvantageous because the device can be inactivated by destroying the lock, but it offers the advantage in that the driver has not to enter the car in a hurry as it is required at automatically activated burglar alarms. LSL-gates offering a high reliability and requiring no external voltage regulation circuits due to their wide operation range, can be advantageously used from 12 V and up, i.e. a minimum of elaborateness is necessary.

The main part of the described device is the circuit FZH 205, incorporating six inverters specified for an extended temperature range.

Oscillator

The oscillator shown in **Fig. 7.12.** comprises of the gates G_4 to G_6. Its period is determined by the circuit consisting of the capacitor C_2 and the series connection of resistors R_2 and R_2'.

It is: $t_{osc} = \dfrac{1}{f_{osc}} \approx 2 \times (R_2 + R_2') \times C_2$

The oscillation is switched on by bridging the RC-circuit through a negator (G_6). For the relatively low required frequency only a non-polarized electrolytic capacitor can be used as time-determining component (e.g. type B 42190–A 4107–T), since the polarity of the voltage is constantly changed. The output signal is available at the test point X.

The oscillator is enabled via the input of the gate G_5. If it receives an L-level the output of G_5 will remain at H-level. However, if an H-level is applied to the input, the output of G_5 depends on the input signal, e.g. the oscillator is switched on.

Monostable multivibrator

The last three gates of the IC FZH 205 are used to realize the monostable multivibrator. The circuit is set by a negative pulse edge applied to terminal 6. The series 10 µF-electrolytic capacitor guarantees that the monovibrator safely flops back

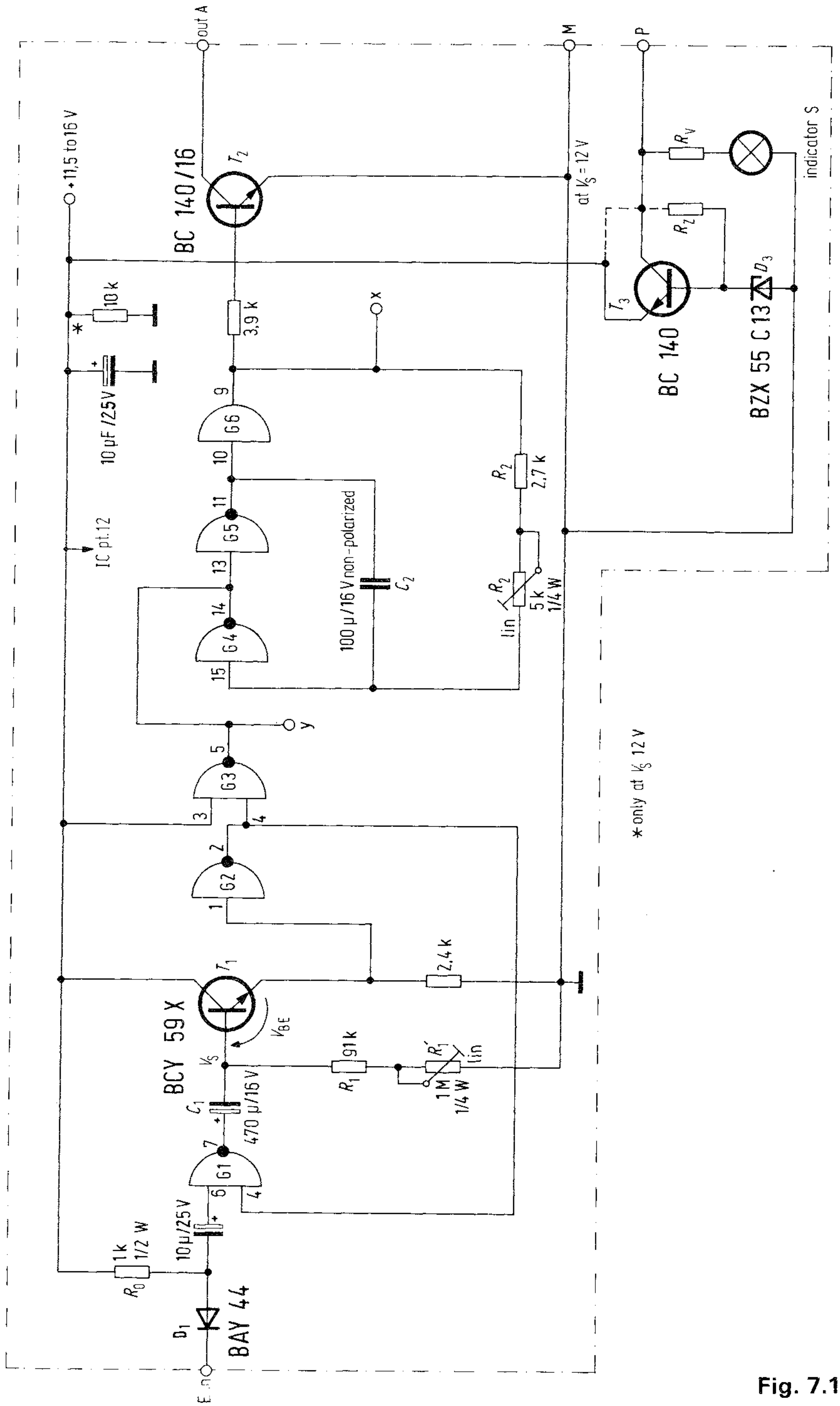

out A
M
P
+11,5 to 16 V
BC 140/16
T_2
at V_S = 12 V
R_V
indicator S
R_Z
BC 140
T_3
D_3
BZX 55 C 13
10 k
*
3,9 k
x
10 µF/25 V
9
G6
10
11
G5
13
14
G4
15
R_2
2,7 k
100 µ/16 V non-polarized
C_2
IC pt. 12
R_2
lin
5 k
1/4 W
y
5
G3
3
4
2
G2
1
*only at V_S 12 V
T_1
2,4 k
BCY 59 X
V_{BE}
V_S
91 k
R_1
R_1'
lin
1 M
1/4 W
C_1
470 µ/16 V
7
G1
6
4
10 µ/25 V
1 k
1/2 W
R_0
D_1
BAY 44

Fig. 7.12.

into its off position, even if the operating contact is closed or a longer time than t_M occurs, i.e. E is connected to chassis. A short L-pulse is supplied via the capacitor to the gate G_1, and T_1 as well as G_2 are turned on. The flyback from the output of the gate G_2 to the non-connected input of G_1 effects that the unstable state is continued and that output of G_2 does not change its H-signal. The capacitor C_1 is slowly charged in the meantime to a H-level via the resistors R_1 and R_1'. The emitter follower transistor T_1 continues to be conductive because of the voltage drop produced across the resistors $(R_1 + R_1')$ by the charging current. The voltage V_{S2} drops across the emitter resistor.
As soon as V_{S2} becomes lower than the switching voltage V_{JL} the gate G_2 flops and the current returns to its initial position. In order to get the levels required for the oscillator control (on $\triangleq$ H-signal, off $\triangleq$ L-signal) an inverter G_3, the output of which is connected to the oscillator gate G_5, follows (cf. Fig. 7.12.).
A quick recharging of the electrolytic capacitor is obtained by the resistor R_0. Therefore a charge being sufficient to set the monovibrator, is stored just after opening the operating contact. This requires however the protection diode D_1, type BAY 44, since the terminal E is usually connected to the supply voltage via the lamps of the courtesy lights and by that the burglar alarm device cannot be safely turned off.

Voltage regulation and operation indication

The voltage regulation using T_3 is only required at a supply voltage of 24 V. At a 12 V-supply collector and emitter of the transistor T_3 can be bridged by a short wire, since the LSL-gates operate in a very wide range. The operation of the device is indicated by a pilot lamp, which is connected to the supply voltage by a series resistor R_v.

Power switch

A relay has to be used if a horn is to be operated by the described circuit. This can be realized by the well-known horn-relay, which is connected to the positive pole of the battery and to terminal A. The relay pulls up when the transistor T_2 being protected against conductive voltage peaks by the diode D_2, type BAY 44, is turned on.

Installation of the burglar alarm device

The installation of the burglar alarm device is demonstrated in **Fig. 7.12.1.** The following has to be considered:

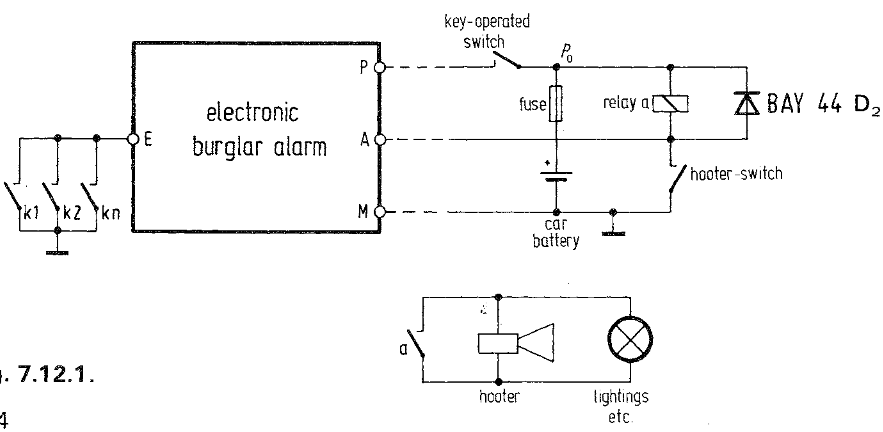


Fig. 7.12.1.

1) The relay should be mounted very close to the horn to avoid connecting lines having large cross sections. The original connection to the horn contact has not to be changed but it has to be considered that in some cars the line feeding the horn is interrupted by the ignition switch. Therefore it has to be assured that the terminal P_0 (cf. Fig. 7.12.1.) is directly or via a fuse connected to the battery, since the burglar alarm device has to also operate when the ignition is switched off. The same applies to the feeding of the horn itself.
2) The device is activated by the key switch. It should be mounted at a discrete but easily accessible place.
3) The operating contacts have to always obtain a connection to the minus pole of the battery when the contacts are closed. As many contacts as possible can be used for protecting a car. Usually the door contacts are utilized, since one of its lines can be interrupted near the courtesy lights. Besides that contacts for the bonnet or roof rack etc. can be installed in parallel as shown in Fig. 7.12.1.

List of Capacitors used in the Circuit 7.12.1.

1 pc	C_2	100 µF/16 V	B 42190–A 4107–T
2 pcs	Electrolytics	10 µF/25 V	B 41313–B 5106–Z
1 pc	Electrolytic	470 µF/16 V	B 41283–A 4477–T