Bipolar Power Transistors in Electronic Ballasts

Selection Criteria and System Requirements

Issue: 11. 95

Table of Contents

Introduction

Better light efficiency and longer lifetime are becoming the main reasons for the substitution of incandescent bulbs with fluorescent lamps. Fluorescent lamps are supplied from the line voltage and must be driven via a ballast unit. There are three types of ballast units; the almost conventional type, the conventional type with lower power dissipation and the electronic ballast unit. By using the electronic ballast unit, the greatest economic profit is possible. A principal distinction can be made between the electronic ballasts for compact lamps – where the lamp and the ballast are mechanical units which can not be separated without destruction – and the electronic ballast for industrial lamps where lamps are independent and can be exchanged individually.

TEMIC has been offering a type program of high-voltage bipolar transistors which is specially developed and optimized for the requirements in the field of electronic ballast units for many years. This paper describes the various requirements for the bipolar power transistors. Rough mathematical calculations will give an overview with regard to the device performance of the ballast unit. Application-specific data of bipolar power transistors and their benefits for the customer will be explained in order to make the devices suitable to the requirements in electronic lighting systems. In addition, some measures to increase the power factor will be discussed. Finally, some circuit design examples of electronic ballast units will be provided.

Lighting Systems

Comparisons given in table 1 show the advantage of electronic ballast units for fluorescent lamps over **all** other lighting systems. This is explained through the better conversion of electrical to light power – a result of the lower power loss in the ballast as well as of the longer life time of the lamps. The column „Fluorescent Lamp/Electronic ballast“ includes compact and industrial lamps.

Compact and Industrial Ballasts

In contrast to compact lamps, it is not acceptable that electronic ballasts for industrial lamps are destroyed when the fluorescent lamp fails. In principle, an emergency switch-off is needed for this type of ballast. This, however, increases the circuitry.

In case of lamp failure, there are two possible operation modes which may occur:

- No ignition because the glass of the lamp is broken or the tube is oxygen-contaminated while the lamp filament is intact. This leads to a permanent ignition mode, the transistors become overheated and, if the transistors are not switched off, are destroyed.
- If the lamp filament is burned down, there is normally no danger for the electronic ballast unit, because the series resonant circuit is open as shown in figure 1.

An additional essential feature of industrial ballasts is the possibility of lamp exchange during operation. This is practicable because the lamp interrupts the switching operation of the ballast. After re-installation, lamp ignition is started. Several possible configurations are shown in figure 1.

Suitable protection measures in case of lamp failure are discussed in the chapter „Measures for Protection“.

Table 1. Comparison of typical operating and system characteristics of the most common lighting systems

Characteristics	Incandescent Lamp	Fluorescent Lamp	Fluorescent Lamp	Fluorescent Lamp
		Conventional ballast	Low-power converter ballast	Electronic ballast
W/lm-ratio	100%	40%	< 40%	20%
Life time in h	1000	5360	5360	8000
Light	Flicker-free	Cathode flickering	Cathode flickering	Flicker-free
Stroboscopic effect	No	Yes	Yes	No
Humming noise	No	Yes	Yes	No
Starting behavior	Immediately	Flickering when started	Flickering when started	Without flickering
Light efficiency	5.60%	22.50%	22.50%	27.80%
Alternation numbers	–	15000 (60 s/ 150 s)	15000 (60 s/ 150 s)	500000 (60 s/ 150 s)
Frequency	50 Hz	50 Hz	50 Hz	30 - 40 kHz
Series choke	No	Big/ heavy	Big/ heavy	Small/ light
Choke loss	No	18.30%	< 18.30%	9.90%
Factor of phase angle	1	Inductive	Inductive	0.95 (cap.)
Power factor	1	Inductive	Inductive	1 with PFC*

Note: * PFC = **P**ower **F**actor **C**orrection

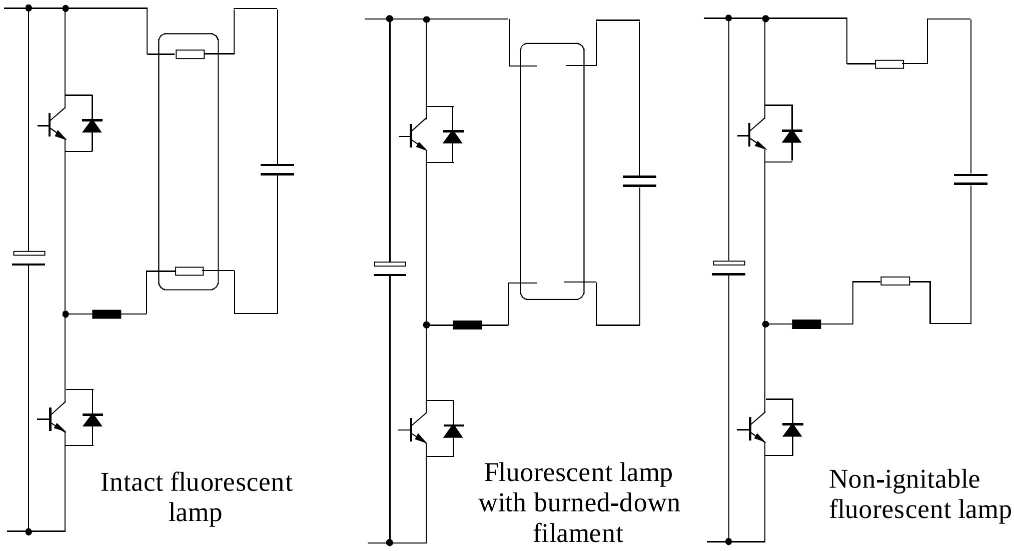


Figure 1. Operation modes of electronic ballast units

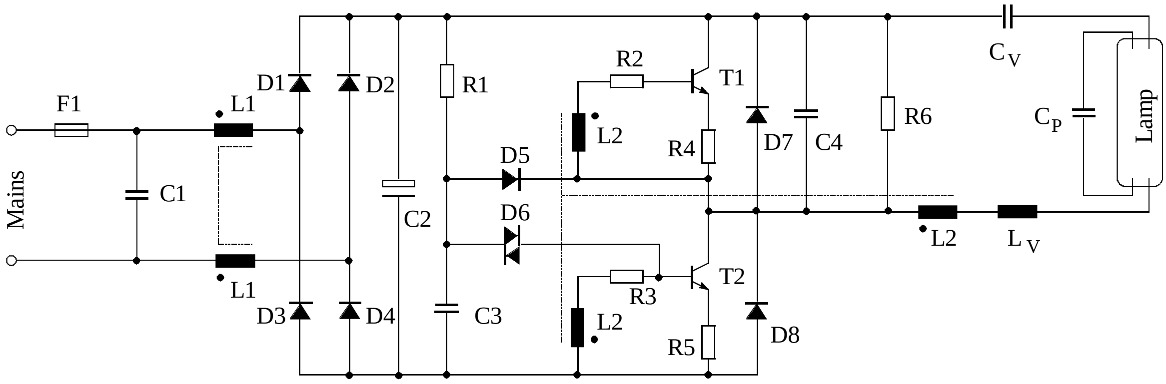


Figure 2. Typical circuitry of an electronic ballast unit

Design Parameters

When designing an electronic ballast for fluorescent lamps, the parameters of the lamp, such as power, ignition voltage, operating voltage and the resistance values of the filament, have to be known. Most of the European electronic ballasts work with a configuration that has a ring core as a saturation transformer (see figure 2). The converter is therefore self-oscillating. It can be assumed that the oscillating frequency is nearly constant in a given operating range. Most of the circuits work in the range of 30 - 40 kHz because the light efficiency of the fluorescent lamp is at its maximum value above 35 kHz. Higher frequencies, however, cause higher switching losses and higher expenditure of the line filter respectively of electromagnetic compatibility.

The components of the series resonant circuit, ballast choke (L_V) and parallel capacitor (C_P) influence the operating performance of the ballast. The capacitor C_V is necessary to avoid a dc portion being applied to the lamp as this would reduce the life time of the lamp.

Equivalent Circuit

Three different modes can be separated under correct operation:

- Pre-heating of the filament with non-ignited lamp (figure 3)
- Ignition mode with non-ignited lamp (figure 3)
- Permanent operation with ignited lamp. The lamp characteristic is similar to a Zener diode (figure 7)

Ignition

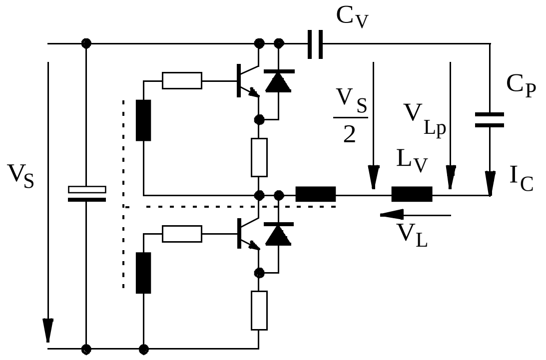


Figure 3. Equivalent circuit in the ignition mode

The filament has been neglected in all equivalent circuits. The following equation for the voltages is therefore valid:

$$\frac{\vec{V}_S}{2} = \vec{V}_{Lp} + \vec{V}_L \qquad (1)$$

and then:

$$\frac{\vec{V}_S}{2} = j \times \vec{I}_C \times \left(\omega \times L - \frac{1}{\omega \times C}\right) \qquad (2)$$

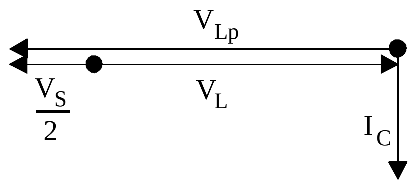


Figure 4. Example of a vector diagram with non-ignited lamp

Figure 4 shows the vector diagram in the ignition mode, assuming capacitive de-tuning of the resonant circuit. This means that the operating frequency of the converter is below the resonant frequency. The maximum attainable ignition voltage is defined as:

$$\left|\vec{V}_{C_P}\right|_{max} = \left|\frac{\vec{V}_S}{2}\right| \times \left|\frac{\omega_0{}^2}{\omega^2 - \omega_0{}^2}\right| \qquad (3)$$

The maximum ignition current is the peak collector current and is given with:

$$\hat{I}_{C\,max} = \sqrt{2} \times \left|\vec{V}_{C_P}\right|_{max} \times \omega \times C \qquad (4)$$

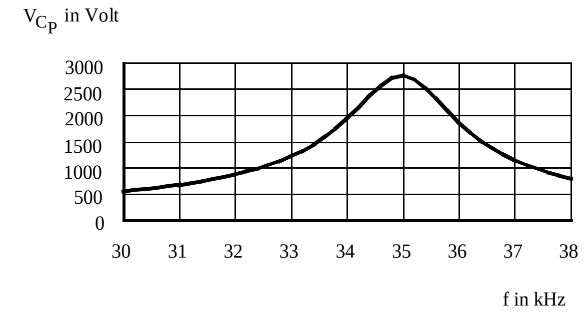


Figure 5. Voltage at parallel capacitor C_p as a function of the operating frequency (f_0 = 35 kHz)

The voltage at the parallel capacitor must be limited to defined ranges for the operating modes „pre-heating of the filament“ and „ignition“,. In the „pre-heating mode“, the voltage must be below the smallest ignition voltage of the lamp (valid at room temperature). In the „ignition mode“, the peak value of the voltage calculated in (3) has to be higher than the highest ignition voltage of the lamp. The plot shown in figure 5 is calculated with (5).

$$V_{C_{Ignit.}} = \frac{V_S}{\sqrt{2}} \times \sqrt{\left(\frac{f_0{}^2 - f^2}{f_0{}^2}\right)^2 + \left(2 \times R_{fil.} \times \omega \times C_P\right)^2} \qquad (5)$$

where:

f_0 = resonant frequency

f = operating frequency

$R_{fil.}$ = resistance of filament

There are two possibilities to change the lamp voltage before ignition:

- Change the operating frequency
- Change the resonant frequency by variation of L_V or C_P

Normal Operation

Under normal operation, calculations describing the behavior of the ballast become more complex because there is no proceeding possibility to use equations and to calculate the parameters, like the lamp current in a closed loop.

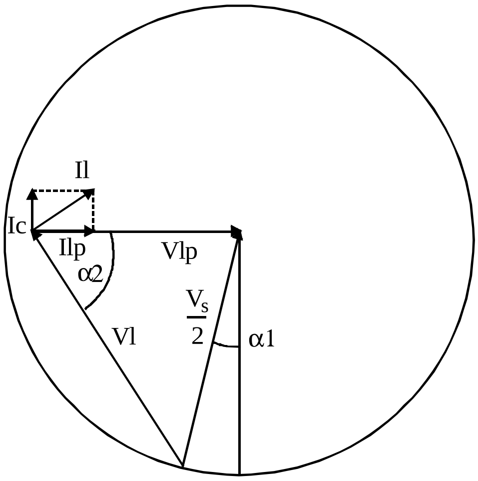


Figure 6. Example of a vector diagram during normal operation of the ballast

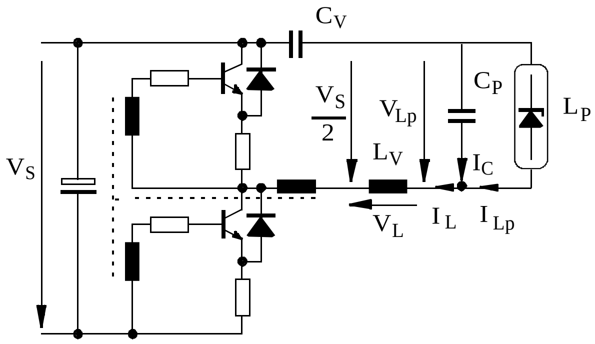


Figure 7. Equivalent circuit under normal operation

The construction of the vector diagram shown in figure 6 is based on the following considerations:

In the first step, the angle α_1 must be fixed arbitrarily, then the solution can be found with iterative calculations by variation of α_1. This iteration starts with the voltage equation in (1) and the current node equation:

$$\vec{I}_L = \vec{I}_C + \vec{I}_{Lp} \qquad (6)$$

where:

$$\left|\vec{I}_C\right| = \omega \times C_P \times \left|\vec{V}_{Lp}\right| \qquad (7)$$

The value of the current flowing through the parallel capacitor C_P (I_C) is now fixed. It is assumed that the vector of the lamp voltage (V_{Lp}) is parallel to the real axis of the vector. The vector of the current through the capacitor is therefore parallel to the imaginary axis.

$$\mathrm{Re}\left(\frac{V_S}{2}\right) = \sin\alpha1 \times \left|\frac{\vec{V}_S}{2}\right| \qquad (8)$$

$$\mathrm{Im}\left(\frac{V_S}{2}\right) = \cos\alpha1 \times \left|\frac{\vec{V}_S}{2}\right| \qquad (9)$$

The real and the imaginary part of the inverter output voltage V_S can be calculated with the actual angle α_1. The real part, the imaginary part and the modulus of the vector of the voltage at the ballast choke L_V (V_L) can be calculated with these results.

$$\mathrm{Re}(V_L) = \mathrm{Re}\left(\frac{V_S}{2}\right) + \left|\vec{V}_{Lp}\right| \qquad (10)$$

$$\mathrm{Im}(V_L) = \mathrm{Im}\left(\frac{V_S}{2}\right) \qquad (11)$$

$$|V_L| = \sqrt{\mathrm{Re}(V_L)^2 + \mathrm{Im}(V_L)^2} \qquad (12)$$

The angle between the lamp voltage vector and ballast choke voltage vector can be calculated by using equations (9) and (12).

$$\alpha2 = \arcsin\left(\frac{\mathrm{Im}\left(\frac{V_S}{2}\right)}{|V_L|}\right) \qquad (13)$$

The modulus of the current flowing through the ballast choke L_V can be calculated by using equation (12).

$$\left|\vec{I}_L\right| = \frac{\left|\vec{V}_L\right|}{\omega \times L} \qquad (14)$$

$$\mathrm{Im}(I_L) = \cos\alpha2 \times \left|\vec{I}_L\right| \overset{!}{=} \left|\vec{I}_C\right| \qquad (15)$$

The iteration can be stopped if the condition in (15) is valid. Only if the imaginary part of I_L is equal to the modulus of I_C can a possible solution be achieved.

$$\mathrm{Re}(I_L) = \sin\alpha2 \times \left|\vec{I}_L\right| = I_{Lp} \qquad (16)$$

Equation (16) helps to calculate the value of the lamp current, equation (17) to calculate the lamp power.

$$P_{Lp} = U_{Lp} \times I_{Lp} \qquad (17)$$

The possible lamp power can therefore be calculated with the knowledge of the four parameters L_V, C_P, ω and V_s. Otherwise, this is only possible if the operating frequency is the resonant frequency. In this case, the equations (1) and (2) from the application note 002 (November 1991) can be used:

$$L_V = \frac{V_S \times V_{Lp}}{2 \times \omega_0 \times P_{Lp}} \qquad (18)$$

$$C_P = \frac{1}{\omega_0{}^2 \times L_V} \qquad (19)$$

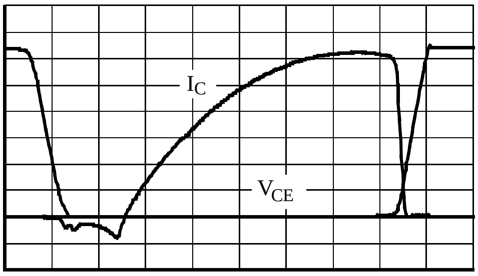
Requirements Regarding Power Transistors

TEMIC's aim is to optimize and to develop all of its bipolar power transistors for customer-defined requirements. It is therefore necessary to know the typical conditions wherein the components are able to work well. The most important system variables for electronic ballast units for fluorescent lamps are the following:

- Current load in the case of ignition and normal operation with maximum expected value of line voltage
- RBSOA, voltage requirement
- Switching behavior, base-drive conditions, switching frequency
- Cooling conditions, power dissipation, ambient temperature
- Dependence of the case temperature through other parts of the circuit

Standard electronic ballast units are „low-cost applications", where the base drive conditions are in principle not optimum. The transistor is activated by a base current when the anti-parallel free-wheeling diode is active and the collector current is negative. This leads to an immense overcharge of the collector-base diode. Transistors which are not optimized react in such a case with long storage times and extreme fall times.

Lowest values of the power dissipation are necessary to operate with such transistors under ambient temperatures in the range of 100°C - 110°C without any heatsink.



V_{CE} : 50 V/div. I_C : 0.2 A/div. T : 2 µs/div.

Figure 8. Current and voltage course

Power Capability

TEMIC specifies the power transistors of the SWOT (**S**imple s**W**itch-**O**ff **T**ransistor) type range especially for the requirements of electronic lighting application to give the user the best possible support for selection. The maximum collector current a TEMIC bipolar power transistor can operate in application is therefore the propagated collector current given in the data sheet.

TEMIC defines the typical collector current as a dc-current gain of 4 and a collector-emitter saturation voltage of 2 V. A factor for the overcurrent for electronic ballast units can be fixed to be the quotient of the maximum ignition current and the normal operating current. This leads to a typical value of 6 and furthermore, the definition given below of the „system-collector current":

$$I_{Csys} = \frac{I_C}{6} \qquad (20)$$

TEMIC SWOTs, driven with a dc-current gain of 4, have a typical collector-emitter saturation voltage of 0.1 V for this collector current. An example of the BUF630 for this definition is shown in figure 9.

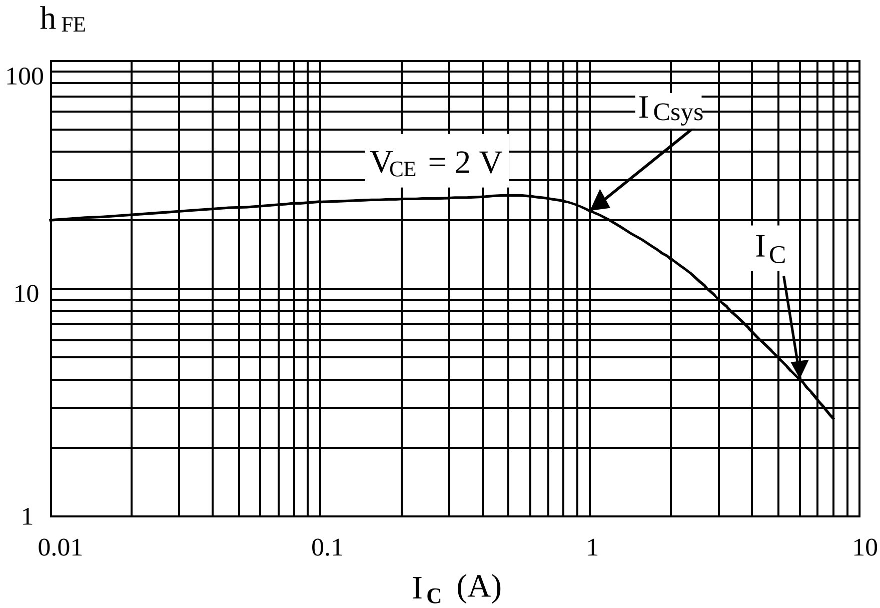


Figure 9. Definition of collector- and system current

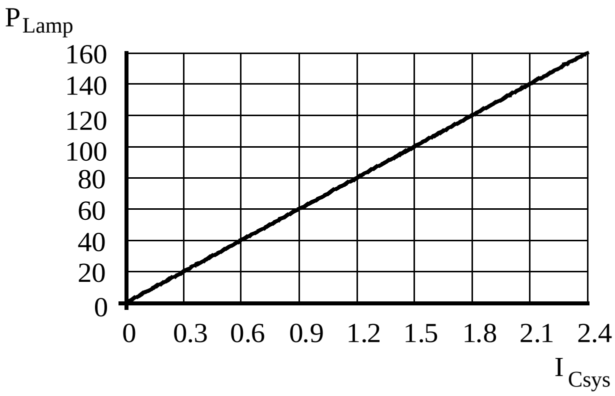


Figure 10. Relationship of system current / lamp power

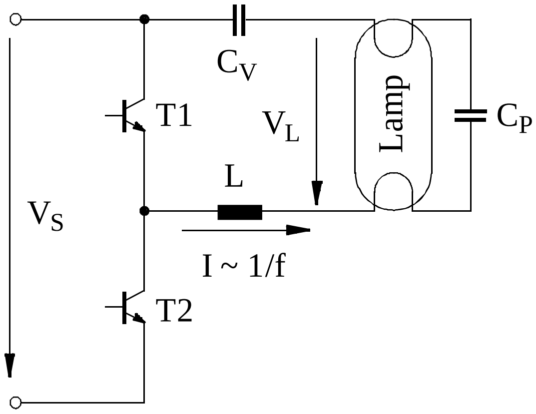


Figure 11. Output power

In general, a system current of 15 mA represents a lamp power of 1 W for electronic ballast units at 230 V mains. This relation is shown in figure 10.

Grouping System t_x

Most of the circuits for electronic ballast units using the half-bridge configuration have a strong dependence on output power and oscillating frequency:

$$P_L \approx V_L \times I \sim \frac{V_L}{f} \qquad (21)$$

A change of storage time in bipolar power transistors leads to a change of the oscillating frequency and, due to the series inductance, to a change of output power. Unfortunately, bipolar power transistors in electronic ballast units are driven in many different ways, depending on the special circuit dimensioning. The typical switching characteristics in the data sheets do not give any relevant information about the real dynamic behavior in an actual electronic ballast circuit. The dynamic characteristics are influenced by the dc-current gain, blocking voltage, the technology used and by the chip size. These parameters are given by the transistor design of the manufacturer (TEMIC). Other parameters like the working point, type of load and the switch-off conditions depend on the application which is given by the manufacturer of the electronic ballast unit.

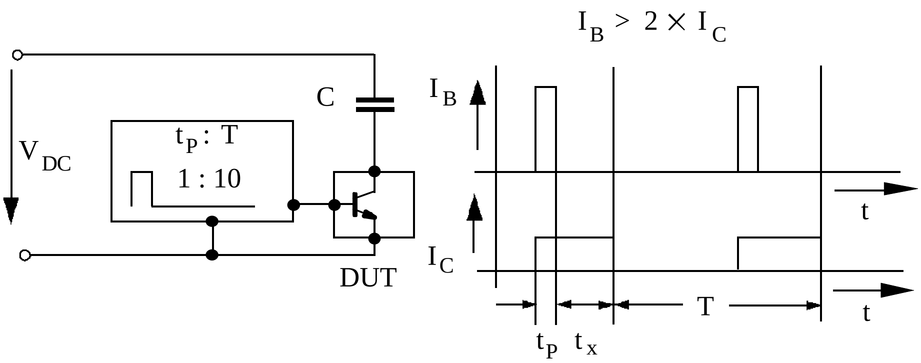


Figure 12. Typical storage time t_x for bipolar power transistors in the field of electronic ballast units

As a conclusion, there are mainly two problems to be solved:

- The spread of the storage time of the bipolar power transistor in the application must be as small as possible.
- The absolute value of the storage time of the bipolar power transistor in the application must be available to fit in the application.

TEMIC's solution:

Common to all configurations of electronic ballast units is the fact that transistors are heavily overdriven and that the reverse base current during switch-off is very low.

TEMIC therefore offers the most relevant switching parameter t_X for their bipolar power transistors. This parameter is controlled and measured in production for 100%. The basic test circuit is shown in figure 12.

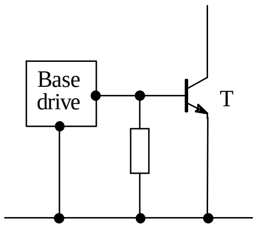


Figure 13. Base emitter resistance

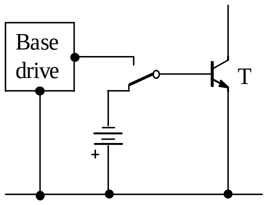


Figure 14. Negative base emitter voltage

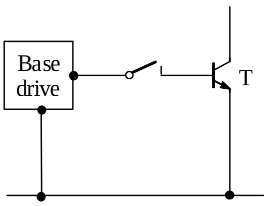


Figure 15. Base emitter open circuit

Working Voltage V_{CEW}

In over 90 % of the switching applications, the bipolar power transistors are blocked either by short circuit respectively resistance < 100 Ω between base and emitter (see figure 13) or negative base emitter voltage, as shown in figure 14, but not with a base emitter open circuit as shown in figure 15. Such switching applications are:

- Electronic ballast units for fluorescent lamps
- Electronic transformers for halogen lamps
- Switch-mode power supplies

So the value of V_{CEO} is not the parameter to determine the switching capability of a bipolar power transistor.

TEMIC defines the switching capability of a bipolar power transistor with the working voltage, as the so-called V_{CEW}.

A transistor switched off under the conditions shown in figures 13 and 14 can be switched on and off, depending on the base drive condition and of course on the collector current, up to V_{CES}.

Therefore, TEMIC defines V_{CEW} as the maximum voltage at which a TEMIC bipolar power transistor can be switched on and off without any risk at a defined collector current (I_{CW}) and base drive condition. This results in a collector current versus working voltage area which is also known as the RBSOA- or FBSOA diagram of a bipolar transistor. TEMIC defines these as „Safe Working Area". The maximum collector current in the diagram in figure 16, I_{CW}, is the collector current propagated in the data sheet. The highest working voltage is achieved if the reverse base current (I_{B2}) is greater than 10% and less than 50% of the collector current. It is assumed that the transistor should have a saturation voltage less than 2 V for the forward base current.

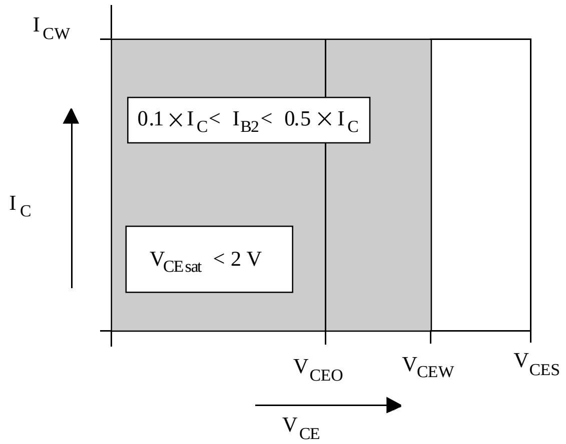


Figure 16. Principle diagram of „Safe Working Area“

The maximum working voltage V_{CEW} of the SWOT types (BUF..) is 100 V above V_{CEO} as shown in table 2:

Table 2. Propagated values of the working voltage

TO 220 Package					
Type	V_{CEO} V	V_{CEW} V	V_{CES} V	I_C A	I_{CW} A
BUF620	400	500	700	4	4
BUF630	400	500	700	6	6
BUF644	400	500	700	8	8
BUF650	400	500	700	10	10
BUF654	400	500	700	12	12

TO 220 Package (Continued)					
Type	V_{CEO} V	V_{CEW} V	V_{CES} V	I_C A	I_{CW} A
BUF642	400	500	900	6	6
BUF672	450	550	900	11	11
BUF636A	450	550	1000	5	5
BUF640A	450	550	1000	6	6
BUF646A	450	550	1000	7	7
TO 251 (DPAK) Package					
Type	V_{CEO} V	V_{CEW} V	V_{CES} V	I_C A	I_{CW} A
BUD600	250	300	600	2	2
BUD620	400	500	700	4	4
BUD630	400	500	700	6	6
BUD616A	450	550	1000	1.6	1.6
BUD636A	450	550	1000	5	5
TO 252 (DPAK SMD) Package					
Type	V_{CEO} V	V_{CEW} V	V_{CES} V	I_C A	I_{CW} A
BUD600SMD	250	300	600	2	2
BUD620SMD	400	500	700	4	4
BUD630SMD	400	500	700	6	6
BUD616ASMD	450	550	1000	1.6	1.6
BUD636ASMD	450	550	1000	5	5

Power Factor Correction

Nearly all mains-operated electronic devices without power factor correction (Fig. 17) have typical line currents like the one shown in figure 18.

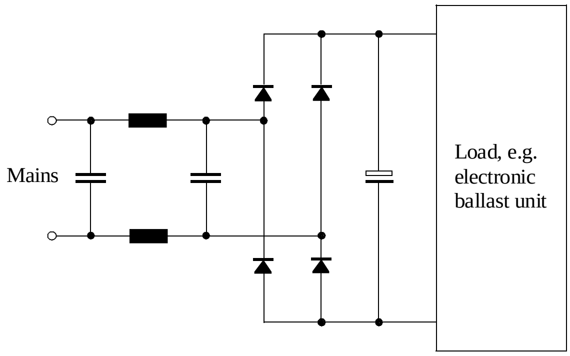


Figure 17. Typical peak value rectification of the mains supply voltage

Without considering a phase angle between voltage and current, the power factor can be defined as follows:

$$k = \frac{P}{I_{RMS} \times V_{RMS}} \qquad (22)$$

If the current and voltage are sine waves, the power factor will be $k = 1$. Assuming an impressed mains voltage (which is not always the case), the power factor is determined by the wave form of the current. As shown in figure 18, the power factor is approximately 0.5. In other words: the RMS value of the mains current is, for this example, two times higher than it could be if the power factor had been 1, while the power consumption in both cases would be the same. This is a great disadvantage in the field of energy distribution for power cables, transformers and open wire lines. The reason for this is that the line-cross section must be dimensioned for double the value without power factor correction.

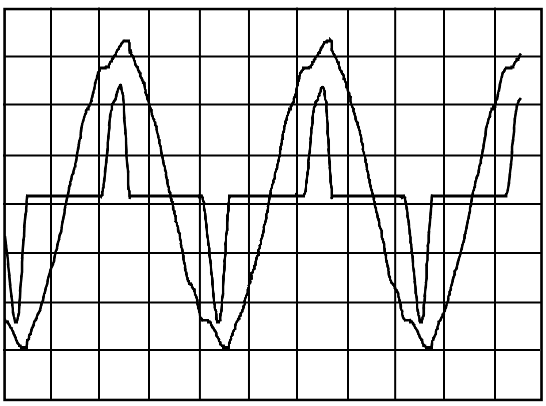


Figure 18. Typical wave forms of mains current and supply voltage without power factor correction

Usually, the power factor will be described with harmonic analysis (Fourier analysis). An unity power factor can be achieved if the current wave contains only the fundamental wave. Harmonics are included if the power factor is less than 1. The procedures can be distinguished as active and passive ones.

Passive Power Factor Correction

At the moment, there are three different known principles:

- Choke with large inductance (100 Hz)
  Disadvantages: high weight and enormous volume
- Special charging circuit for the smoothing capacitor where 2 capacitors were loaded in series and discharged by the load in parallel
  Disadvantage: a large ripple on the dc voltage
- Rectification of the high frequent oscillation of the converter output using the boot-strap principle
  Disadvantage: dependence of the dc-output level on the RF pulse current

The circuits 2. and 3. are protected by patent law and are therefore not described in detail here.

Active Power Factor Correction

Most of the circuits for active power factor correction use the boost converter configuration as shown in figure 19. The circuit is similar to a fly-back converter, but has no galvanic isolation between input and output.

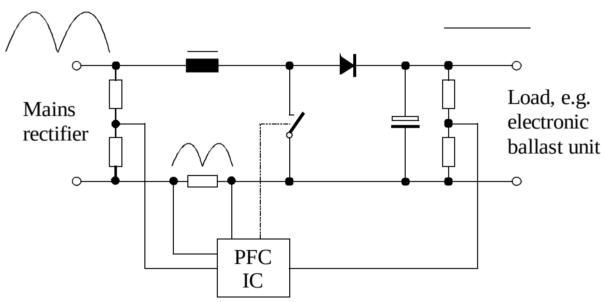


Figure 19. Typical configuration of active power factor correction circuits

The circuit regulates the mains supply current to sine-wave form and the dc-output voltage to a constant value. In order to enable the regulation, the dc-output voltage must be higher than the peak value of the rectified mains voltage. Two different operation modes for active power factor correction are known:

- The so-called discontinuous mode, where the average value – not the momentary value of the mains current – is regulated to sine wave form. This procedure is suitable for loads with nearly constant output power, e.g., electronic ballast units.
- The so-called continuous mode is where the momentary value of the mains supply current is regulated to sine wave form. This procedure is suitable for loads with greater changes in output power. A disadvantage compared to the discontinuous mode is the necessity for a higher inductance of the boost. This mode is therefore more expensive.

The regulated dc-output voltage is an enormous advantage of the active power factor correction which also simplifies the design and dimensioning of the ballast. On the other hand, more additional components are needed for the active power factor correction. Only a passive power-factor correction therefore makes sense in the field of low-cost compact lamps.

Measures for Protection

An immediate switch-off within a few seconds is necessary in case of operating problems such as lamp failure or short circuit at the output terminals of the ballast. It is not easy, however, to switch off a self-oscillating converter. In addition, a switch-off may be dangerous for the power transistor. In the following sub-chapters, some examples for switch-off circuits are described and discussed taking the semiconductor manufacturer's point of view into account.

Upper Transistor

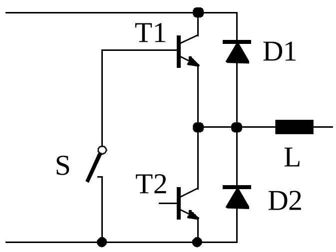
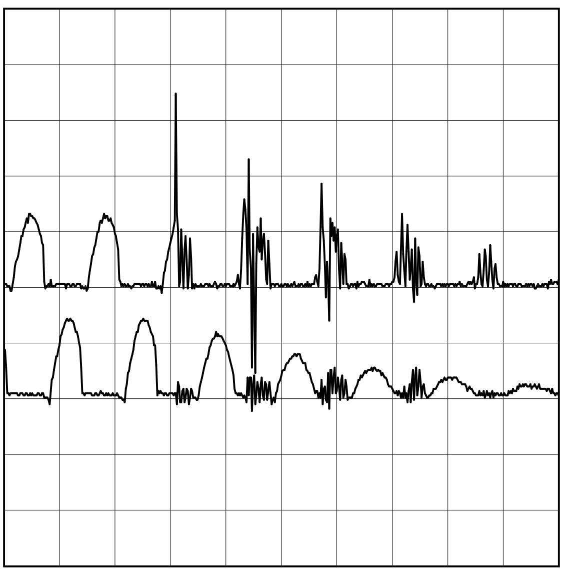


Figure 20. Switch-off T1



(above) I_{C1} : 1 A/div, (below) I_{C2} : 1 A/div, T : 20 µs/div

Figure 21. Moment of switch-off

Although the switch off is initialized (start of RF oscillations) as shown in figure 21, the converter is active for some periods in a decay process. This is much more critical because the upper transistor is not saturated during this operation when a collector current is flowing. The collector-emitter voltage of T1 is approximately the dc-supply voltage. Transistors with $V_{CEO} > V_{DC}$ are therefore necessary to ensure a safe switch-off operation The advantage of the working voltage V_{CEW} (see chapter „Working Voltage V_{CEW}") which is typically 100 V above V_{CEO} for the SWOT types **can not be used with this kind of switch-off procedure**.

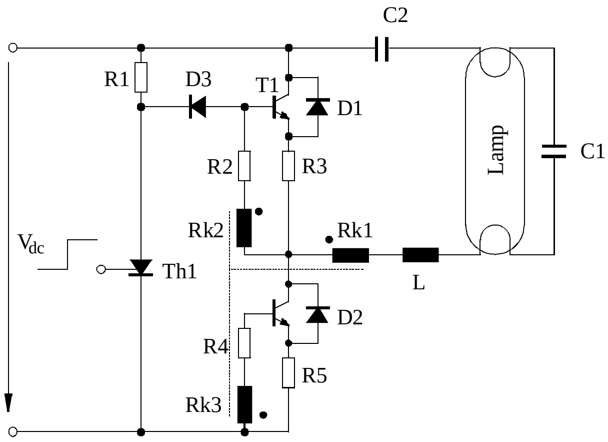


Figure 22. Example of switch-off circuit

The use of cheap power transistors with good dynamic characteristics and high dc-current gain is impossible under the described circumstances. The RF oscillation represents a high SOA load for the transistor.

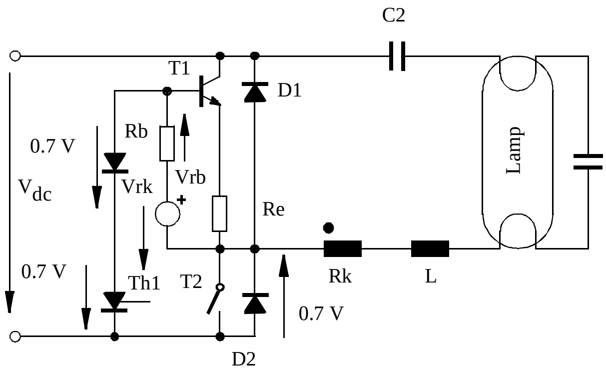


Figure 23. Equivalent circuit during switch-off under the conditions of figure 20

The equivalent circuit shown in figure 23 shows that T1 can not be switched off via Th1. The base emitter voltage of T1 and the voltage drop at R_e can be limited only to approximately 2.1 V. This does not prevent a positive base drive current of T1. If T1 is driven, V_{CB} cannot reach values smaller than V_{DC} – 1.4 V which represents a high power dissipation and a tremendous FBSOA (**F**orward-**B**iased **S**afe **O**perating **A**rea) – load.

Lower Transistor with b-e Short Circuits

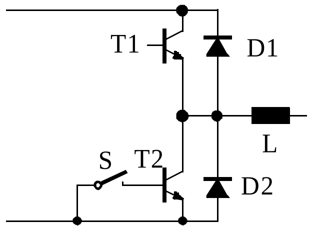


Figure 24. Switch-off T2

This switch-off procedure is particularly suitable for ballast configurations as shown in figure 22 where the resonant circuit is connected with the plus wire of the dc-supply voltage. Transistor T2 cuts off the energy supply for the resonant circuit, which means the decay process will be as short as possible.

The Switch S has to carry on the one hand the reverse base current of T2 and the base drive current which will be supplied from the driving transformer. The saturation voltage must be below 0.3 V and this could therefore cause problems. The short circuit with S should be as good a fit as possible.

Four different kinds of switches are possible:

- A „short circuit“ with an SCR that includes a saturation voltage of minimum 0.7 V (disadvantage: no absolute security, even at high temperatures)
- Due to the finite dc-current gain, a bipolar transistor switch may be a solution depending on the design of the electronic ballast unit
- A power MOSFET is a suitable switch-off device, provided that R_{DSon} is low enough (disadvantage: the limitation of the base emitter reverse voltage may have a negative influence on the dynamic behavior)
- The lowest saturation is achieved with a relay contact (disadvantage: contact bouncing or operate delay might be possible)

Lower Transistor with Interrupted Base Current

A configuration as shown in figure 25 can be used to avoid the strict requirements for the saturation voltage of the switch-off device. A base emitter resistance for T1 and T2 is necessary to ensure fast storage and fall times (t_S, t_F).

Under normal operating conditions, T1 is driven via R1. The transistor must not limit the base current of T3. D3 is necessary for the reverse base current,. R3 and R4 should have different values of the resistance to ensure symmetrical switching behavior of T2 and T3, Th1 has to be switched on in case of lamp failure. T1 will therefore be switched off and the base current is then interrupted. In addition, ignitions of the DIAC D2 are also interrupted. A power MOSFET can be used the same way instead of a bipolar transistor.

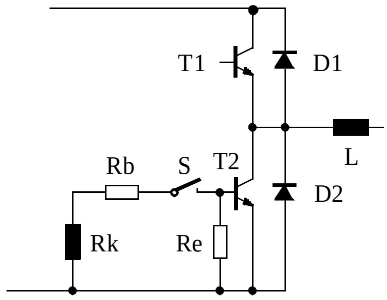


Figure 25. Interruption of base current

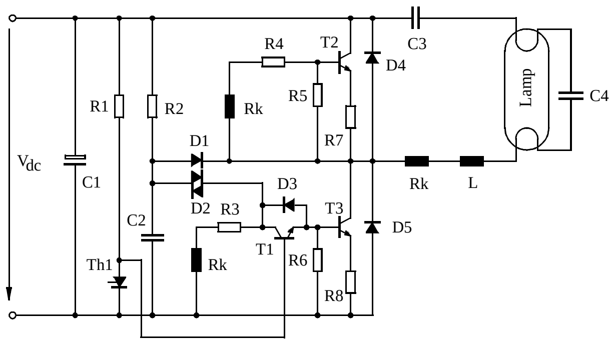


Figure 26. Interruption with a bipolar transistor

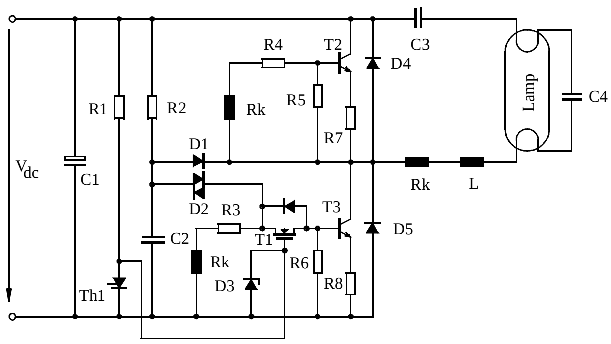


Figure 27. Interruption with a MOSFET

The functional principle of the circuit shown in figure 26 is almost similar to the one shown in figure 27. D3 is necessary to limit the gate-source voltage of T1. The maximum value of the negative voltage amplitude at R_k plus the Zener voltage must be smaller than V_{GSS} of T1 for the selection of the Zener voltage.

Lower Transistor with Negative b-e Voltage

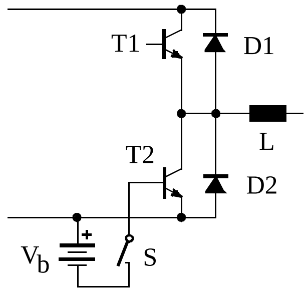


Figure 28. Switch-off T2

An emergency switch-off circuit, as shown in figure 28, is the safest procedure with regard to the bipolar power transistor. An auxiliary voltage, realized with an auxiliary winding on the ballast choke, for instance, is not accepted by most of the ballast manufacturers because of the high ratio of the number of turns to the main winding. A better solution is shown in figure 29.

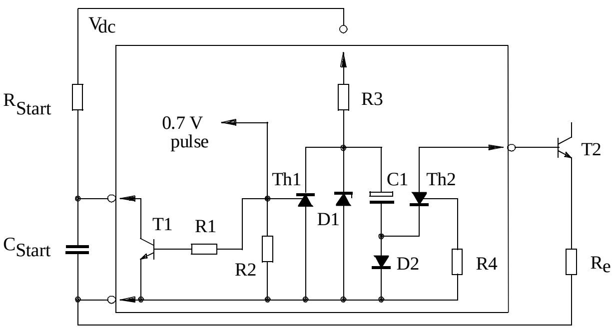


Figure 29. Example for a switch-off circuit comparable to the one shown in figure 25

The auxiliary voltage to switch off T2 is generated with C1 loaded via R3. The Z-diode D1 limits the voltage at C1 to values smaller than V_{EBO} of the power transistor T2. The electronic ballast unit can be switched off immediately if a 0.7 V pulse is applied to the gate of Th1. If Th1 is triggered, the plus terminal of C1 will have approximately ground potential and D2 is blocked. T2 is triggered via R4 and reverse base current is flowing through the circuit of base emitter of T2, Th2, Re, Th1 and C1 simultaneously. The gate-cathode voltage of Th1 is used to keep the starting capacitor discharged to avoid new trigger pulses from the DIAC (not shown in figure 29).

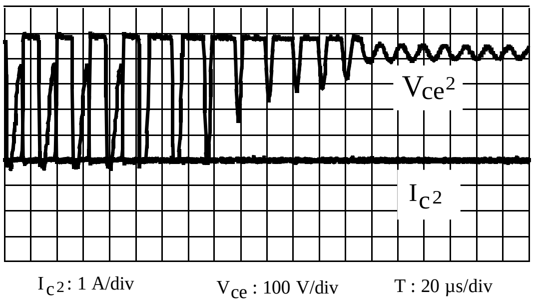


Figure 30. Moment of switch-off

If the upper power transistor is switched off as shown in figure 20, the collector current of T2 is zero. After the ballast unit has been switched off, the trace of the collector-emitter voltage may give the impression that T2 is switched on again. Nevertheless, only the parallel free-wheeling diode is switched on. The decay process takes more than 8 cycles of the switching frequency. In contrast to figure 21, it is not worthwhile mentioning power dissipation at T2 after switch-off. Under the described conditions, the maximum working voltage V_{CEW} can be used. Under the conditions of figure 20, voltages up to only V_{CEO} can be applied.

Circuit Examples

The following examples do not intend to give the impression that they are dimensioned for production requirements. There is neither a power factor correction circuit nor a line filter included. They should be seen only as proposals in order to give the possibility for a first critical look at the application of electronic ballast units.

Ballast Unit for 8 W at 120-V Mains

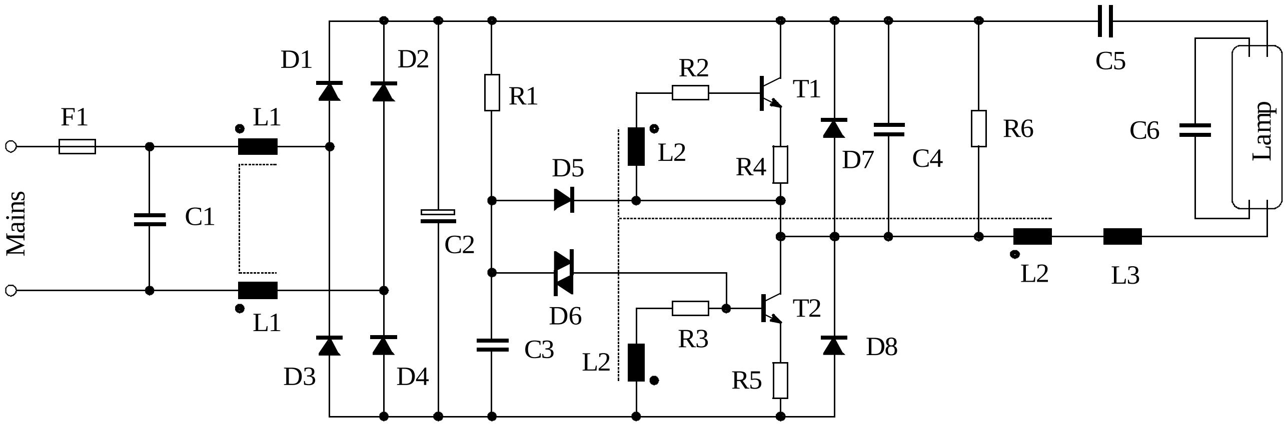


Figure 31. Circuit example 8 W / 120 V

Parts list:

T1	BUD620
T2	BUD620
D1	BYT51G
D2	BYT51G
D3	BYT51G
D4	BYT51G
D5	BYT51G
D6	G-ST2 (DIAC)
D7	BYT52G
D8	BYT52G
R1	330 kΩ
R2	68 Ω
R3	68 Ω
R4	3.3 Ω
R5	3.3 Ω
R6	330 kΩ

C1	100 nF	250 V	
C2	10 µF	250 V	
C3	22 nF	100 V	
C4	1 nF	250 V	
C5	100 nF	250 V	
C6	3.3 nF	400 V	
F1	250 mA		
L1	5 mH (2 ×)		
core	E13/4N27		d = 0 mm
winding	2 × 80 turns		∅ = 0.2 mm
L2	toroidal core		
winding	3/4/4 turns		∅ = 0.2 mm
L3	0.9 mH		
core	E16/5 N27		d = 1 mm
winding	140 turns		∅ = 0.3 mm
lamp	8 W		

Ballast Unit for 8 W at 230-V Mains

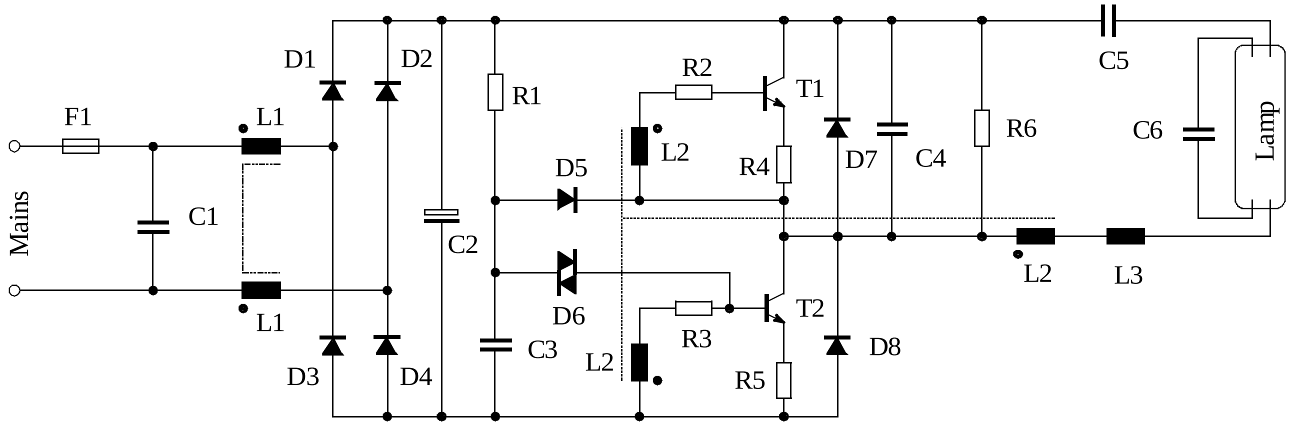


Figure 32. Circuit example 8 W / 230 V

Parts list:

T1	BUD620
T2	BUD620
D1	BYT51K
D2	BYT51K
D3	BYT51K
D4	BYT51K
D5	BYT51K
D6	G-ST2 (DIAC)
D7	BYT52K
D8	BYT52K
R1	330 kΩ
R2	68 Ω
R3	68 Ω
R4	3.3 Ω
R5	3.3 Ω
R6	330 kΩ

C1	100 nF	250 V	
C2	4.7 µF	250 V	
C3	22 nF	100 V	
C4	1 nF	250 V	
C5	100 nF	250 V	
C6	3.3 nF	400 V	
F1	250 mA		
L1	11.5 mH	(2 ×)	
core	E13/4	N27	d = 0 mm
winding	2 × 120 turns		∅ = 0.2 mm
L2	toroidal core		
winding	3/4/4 turns		∅ = 0.3 mm
L 3	2.8 mH		
core	E16/5	N27	d = 1 mm
winding	250 turns		∅ = 0.3 mm
lamp	8 W		

Ballast Unit for 40 W at 230-V Mains

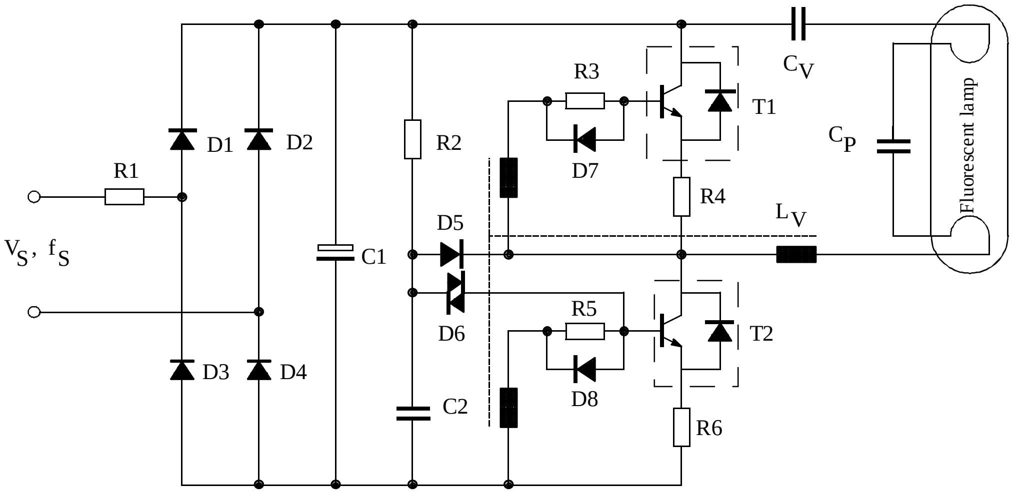


Figure 33. Circuit example 40 W / 230 V

Parts list:

T1	TE13005D
T2	TE13005D
D1	BYT51K
D2	BYT51K
D3	BYT51K
D4	BYT51K
D5	BYT51K
D6	G-ST2 (DIAC)
D7	1N4148
D8	1N 4148
R1	1 Ω
R2	470 kΩ
R3	51 Ω

R4	1 Ω		
R5	51 Ω		
R6	1 kΩ		
C1	15 µF	350 V	
C2	150 nF	250 V	
C3	100 nF	100 V	
C4	10 nF	250 V	
L1	1.94 mH		
core	E25/7	N27	d = 0.5 mm
prim. wdgs	146 turns		∅ = 0.5 mm
sec. wdgs.	2 × 3 turns		∅ = 0.5 mm
lamp	40 W	L40W/250 OSRAM	

Ballast Unit for 100 W at 230-V Mains

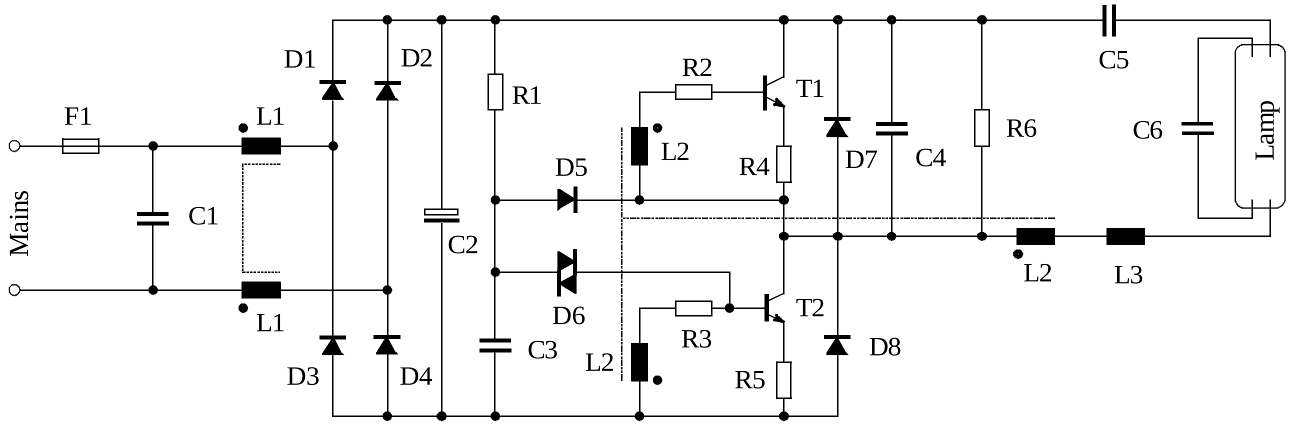


Figure 34. Circuit example 100 W / 230 V

Parts list:

T1	BUF654
T2	BUF654
D1	BYT51K
D2	BYT51K
D3	BYT51K
D4	BYT51K
D5	BYT51K
D6	G-ST2 (DIAC)
D7	BYT52K
D8	BYT52K
R1	330 kΩ
R2	15 Ω
R3	15 Ω
R4	0.22 Ω
R5	0.22 Ω
R6	330 kΩ

C1	100 nF	400 V	
C2	47 µF	350 V	
C3	22 nF	100 V	
C4	4.7 nF	400 V	
C5	220 nF	400 V	
C6	22 nF	400 V	
F1	2 A		
L1	11.5 mH	(2 ×)	
core	E25/7	N27	d = 0 mm
winding	2 × 81 turns		∅ = 0.45 mm
L2	toroidal core		
winding	4/4/4 turns		∅ = 0.65 mm
L3	0.5 mH		
core	E42/15	N27	d = 1 mm
winding	43 turns		∅ = 0.8 mm
lamp	100 W	UVA lamp	