The new Inter AC adapter and converter Handbook of Dipl.-Ing Jörg Rehrmann

Preface, general and important information
1 50-Hz transformers and chokes
2 50 Hz rectifier and filter circuits
2.1 The half-wave rectifier
2.2 The Brckengleichrichter
2.3 The midpoint rectifier
2.4 rectifier with voltage multiplication
2.5 Three-phase rectifier
2.6 filter circuits for rectifier
2.7 inrush current for rectifier

3 Linear DC-DC converter

4 Phase Angle

5 Voltage converter with switched Capacities
5.1 Rechteckgeneratioren for voltage converter
5.2 voltage multiplier connected with capacities
5.3 Power inverter with switched capacities
5.4 voltage divider connected with capacities

6 Voltage transformer with chokes
6.1 Abwrtswandler with chokes
6.2 Aufwrtswandler with chokes
6.3 Inverse converter with chokes
6.4 SEPIC converter

7 The flyback converter

8 Forward converter
8.1 Unregulated single-ended forward converter
8.2 Regulated single-ended forward converter
8.3 Unregulated push-pull forward converter
8.4 Regulated push-pull forward converter
8.4.1 phase shift PWM converter

9 Stray field disposal at blocking
9.1 Thermal stray field Disposal
9.2 Regenerative stray field Disposal

and flow transducers
9.1 Thermal stray field Disposal
9.2 Regenerative stray field Disposal

10th Resonant converter

11th Sinusoidal and trapezoidal converter
11.1 single-ended sine wave converter
11.2 push-pull sine wave converter

12th Passive and active power filter / PFC
12.1 Passive Entstrfilter
12.2 Active Netztfilter / power factor correction

13th Voltage converters for Special Applications
13.1 Auxiliary power generators in mains powered Whips
13.1.1 Auxiliary power generation in 50-Hz technology
13.1.2 Auxiliary power generation with high operating frequency

13.2 ballasts and transformers for lamps
13.2.1 converter for LED lamps
13.2.2 converter for gas discharge lamps
13.2.3 Electronic halogen transformer
13.3 Hochspannungsgereratoren
13.3.1 High voltage transformer for continuous operation
13.3.2 High voltage transformer for pulsed operation
13.4 Inverter

14th Formulary
Preface, general and important information
1 50-Hz transformers and chokes
1.1 The gebruchlichen core designs
1.2 The dimensions of 50-Hz transformers
1.2.1 The required Kerngre a transformer
1.2.2 The calculation of the number of turns
1.2.3 Calculation of Drahtstrken a transformer
1.3 The leakage transformer
1.4 The three-phase transformer
1.5 The autotransformer
1.6 The Einschaltproblem of Power Transformers
1.7 50-Hz chokes

2 50 Hz rectifier and filter circuits
2.1 The half-wave rectifier
2.2 The Brckengleichrichter
2.3 The midpoint rectifier
2.4 rectifier with voltage multiplication
2.5 Three-phase rectifier
2.6 filter circuits for rectifier
2.7 inrush current for rectifier

3 Linear DC-DC converter
3.1 The shunt regulator
3.2 The Lngsregler
3.2.1 shunt element than Lngsregler
3.2.2 The unregulated "Lngsregler"
3.2.3 The regulated Lngsregler
3.2.4 Low-dropout regulator
3.2.5 Voltage regulation with integrated Lngsreglern
3.2.6 variable integrated Lngsregler with adjustable current limit
3.2.7 High-voltage regulator with standard components
3.3 Constant current sources
3.3.1 Simple power sources for small-signal applications
3.3.2 power sources with high accuracy

4 Phase Angle
4.1 Power control and dimmer with phase control
4.2 rectifier with phase cut control

5 Voltage converter with switched
5.1 Rechteckgeneratioren for voltage converter
5.2 voltage multiplier connected with capacities
5.3 Power inverter with switched capacities
5.4 voltage divider connected with capacities

Capacities
5.1 Rechteckgeneratioren for voltage converter
5.2 voltage multiplier connected with capacities
5.3 Power inverter with switched capacities
5.4 voltage divider connected with capacities
6 Voltage transformer with chokes

6.1 Abwrtswandler with chokes
6.2 Aufwrtswandler with chokes
6.3 Inverse converter with chokes
6.4 SEPIC converter

7 The flyback converter
7.1 Simple flyback converter for low input voltages
7.2 Simple regulated flyback converter
7.3 Regulated flyback converter with direct Stromberwachung
7.4 Regulated flyback converters with indirect Stromberwachung
7.5 PWM flyback converter

8 Forward converter
8.1 Unregulated single-ended forward converter
8.2 Regulated single-ended forward converter
8.3 Unregulated push-pull forward converter
8.4 Regulated push-pull forward converter
8.4.1 phase shift PWM converter

9 Stray field disposal at blocking
9.1 Thermal stray field Disposal
9.2 Regenerative stray field Disposal

and flow transducers
9.1 Thermal stray field Disposal
9.2 Regenerative stray field Disposal

10th Resonant converter

11th Sinusoidal and trapezoidal converter
11.1 single-ended sine wave converter
11.2 push-pull sine wave converter

12th Passive and active power filter /
12.1 Passive Entstrfilter
12.2 Active Netztfilter / power factor correction

PFC
12.1 Passive Entstrfilter
12.2 Active Netztfilter / power factor correction

13th Voltage converters for
13.1 Auxiliary power generators in mains powered Whips
13.1.1 Auxiliary power generation in 50-Hz technology
13.1.2 Auxiliary power generation with high operating frequency
13.2 ballasts and transformers for lamps
13.2.1 converter for LED lamps
13.2.2 converter for gas discharge lamps
13.2.3 Electronic halogen transformer
13.3 Hochspannungsgereratoren
13.3.1 High voltage transformer for continuous operation
13.3.2 High voltage transformer for pulsed operation
13.4 Inverter

Special Applications
13.1 Auxiliary power generators in mains powered Whips
13.1.1 Auxiliary power generation in 50-Hz technology
13.1.2 Auxiliary power generation with high operating frequency
13.2 ballasts and transformers for lamps
13.2.1 converter for LED lamps
13.2.2 converter for gas discharge lamps
13.2.3 Electronic halogen transformer
13.3 Hochspannungsgereratoren

13.3.1 High voltage transformer for continuous operation
13.3.2 High voltage transformer for pulsed operation
13.4 Inverter

14th Formulary

Caution! Important Notes

Many of the circuits described working with dangerous voltages. This mainly concerns mains powered circuits, and high-voltage generators of higher performance.Even voltages above 60 volts are in principle already to be regarded as dangerous. High voltage generators that provide short-circuit currents of 5 mA, however, are relatively harmless for healthy people. Note, however, that charged filter capacitors can also temporarily give much larger currents in high-voltage low-power generators. As a warning, I have inserted in all circuits with a high probability result in hazardous voltages a small skull icon. Of course, dangerous voltages can occur at many other circuits with appropriate dimensioning. These circuits may be edited and composed only by skilled technicians who have sufficient knowledge of safety regulations and dealing with hazardous voltages.
The circuits shown are no blueprints for complete equipment, but provide only partial solutions for circuits dar. Who uses these circuits in its devices, you must ensure that the power supplies are adequately protected, so no danger of any type can assume. This concerns in particular the protection of circuits with suitable fuses, which are not always shown in the various circuits, and the Protection from contact with high voltages. Furthermore, the user must provide the shown clocked power supplies suitable for noise protection. These are usually not included in the circuits. The author accepts no liability for any damages caused by using one of the circuits shown.

General notes on the diagrams shown

The circuits described are basically tested and functional. Nevertheless, it is not excluded that still marking or drawing errors have crept. Furthermore, it is also possible that the circuits due to adverse effects of component tolerances or not function properly. I will strive to correct all known errors after the book and improvements in the online edition.

In order not to clutter up the diagrams with many unnecessary information which would impede the clarity, I have omitted some labels. This mainly refers to information of components, which are obviously due to their function. When resistors are, unless otherwise stated, to normal ¼-watt types. Resistors should be used with significantly lower maximum load capacity is to make this in individual cases. Resistors, where can permanently applied voltages above 150 volts should be divided into several individual resistors connected in series, although this is not always explicitly stated in the diagram. Alternatively, the use of higher-duty metal oxide resistors is possible. When the value of information I left out the unit. Pure numerical values are in ohms. A trailing small k ohms means and a large M is Mohm.
The capacitors units are usually labeled. For pure numeric values, the value is specified in pF. Voltage data I have often omitted in control circuits, as they usually the smallest market values are stress arising out of the application and sufficient. When used in timing circuits, where demand low tolerance and high temperature stability, should always be used film capacitors. For very small capacitance values and high-quality ceramic capacitors are suitable. In resonant circuits with high power and high frequencies usually the type MKP capacitors or even higher duty FKP types are used. Since it is absolutely necessary for proper function, I have marked those capacitors in the circuit diagrams accordingly.
The capacitors used in switching power supplies of all kinds are exposed to a particularly high ripple current. This concerns in particular the electrolytic capacitors that are in front of or

behind the power switches (diodes and transistors). In order to avoid unnecessary losses and undue heating of electrolytic capacitors, it is recommended that in such cases, so-called low-ESR electrolytic capacitors use (ESR = equivalent series resistance). These are electrolytic capacitors with very low internal resistance, which can often be recognized by their ultra slim design. Alternatively, the internal resistance by paralleling several smaller capacitors reduced, which is usually cheaper. Problems with too high internal resistance occur predominantly at Niedervoltelkos, since the current load is very high and significantly reduce already low loss stresses the efficiency of the converter.
When the diode very often the lack of details. Simple diode in low-power range are, unless otherwise specified, always universal diode type 1N 4148th For power diodes in switching converters should always be used with reverse recovery times of less than 100 ns ultrafast diodes. Blocking voltage and maximum current carrying capacity resulting from the specific application.
For reactors I have often given only the inductance. With coils that are charged primarily with high-frequency alternating currents, a ferrite core must be used with an air gap. This mainly affects the resonator coil of resonant converter, the simple auxiliary power generators and chokes the electronic lamp ballasts. Alternatively, air coil or coils can be used on ferrite rod or roll cores here. The smallest size for given requirements but can always reach for a core with an air gap, not just the device into saturation.
At higher power RF chokes significant losses by the proximity effect in solid copper wires can occur with large cross-section. In such cases, it is necessary to wind the coil with HF stranded wire. The abbreviation "Cul" stands for copper wire and the number presented is the wire diameter in mm. In HF wire number and diameter of individual strands is indicated.
For "normal" range most affordable storage chokes powder toroidal core chokes. These are suitable for the buck converter in section 6, the forward converter and as a power supply ballast of the push-pull sine wave converter. When the transformers also used the concrete core and coils are partially specified data. But these are intended only as an example and can be optimized in a particular case for the application in a converter. The most important formulas that I have summarized again in the formulary in the Appendix.
For the construction of individual pieces will not have been avoided to make the required transformer itself. When building a converter transformer a few things to consider. With coils or transformers to be provided with insulation between the coils or sheets, it must be ensured that the insulation sheet is 2-3 mm wider than the coil body and the side edges of the film at a distance of a few mm 1-2 mm are cut. Thus, the edge of the film to bulge upwards and prevents the coil wire between the spool and the film edge by slipping into the lower layers and there possibly causing a shorted coil Especially critical is lost with power transformers, since lost possibly the protective separation between power and low-voltage side. For power transformers, whether 50 Hz or high frequency transformers, isolation between mains and low voltage side must be at least 4 kV test voltage tolerated. Accordingly, the insulation film between the coil of each of these pages must be more particularly thick.Depending on the thickness of these should then be wrapped at least three times. Overall, such coils should be separated by about 0.5 mm insulation. Furthermore, the layers should be wound at least in the area of the disconnecting device not quite up to the edge of the bobbin. Here could otherwise very easily flashovers occur between the coils. The film itself needs to min. 200 ° C to be heat resistant, so that even at excessive operating temperature, the protective insulation is not compromised. Very practical is an additional layer of paper that retains its strength at high temperatures.

For larger high-frequency transformers, the coils are usually made self-supporting without a full bobbin. Because of the low number of turns and the large wire cross sections there is a normal bobbin anyway not very helpful. Essentially, there are three problems occur in the production of:

1 The coil support
This must first be established. For this purpose are very good plastic pipes. If no matching tubes or get the core has a rectangular cross section, one can get the support of several layers also produce a stable film. It is important to pay attention to generous match between core and coil former. When winding the carrier narrowed slightly and may not otherwise fit on the core.

2 Fixing the wire ends of a coil
Thus, the coil connection does not pull from the reel with mechanical tensile load, it should be provided with a strain relief. For this you can create a narrow, approximately 5 cm long strip of fabric tape with the sticky side up to about the middle of the first and last turns of each coil. The tape is then folded behind the first, or last turn, and then touch the adhesive surface of the tape and glue. With thin wires <0.3 mm, I recommend to take the wire ends and twist several times to form a strand. This increases the mechanical stability of the lead significantly for self-supporting coils.

3 Fixation of the documents
In the upper layers cantilevered coils tend to decrease in diameter towards the edges. This results in that the outer turns easily slip, and finally out to the side of the coil. This can be remedied using double-sided tape as insulation between the layers. The wire is then fixed from above and below by the adhesive tape.

When high-frequency power transformer is greater at lower voltages can happen that the coil is required only one or two turns. Here the use of copper foil or thin sheet rather than one thick wire is recommended. A thin sheet has the advantage that a few turns can be distributed uniformly over the entire width of the available winding space, which causes a low leakage inductance. The proximity effect also occurs not so much because of the large surface in thin films in appearance. For very large currents, it is advisable to place a thick plate to use multiple stacked with a layer of varnish insulated from each other films.
With a bolt-cores, eg UI cores, only integer turns are possible. A winding is uniquely defined by the fact that the wire is exactly one closed magnetic circuit defined by the (core). As the wire passes accurate and is completely irrelevant and affects most of the leakage inductance of.
For nuclei with multiple limbs, for example, EI cores, one can apply theoretically half turns. One revolution always runs through both legs, only a half between the middle and outer thighs. Coils for power transmission may also isosceles with more cores consist only of integer Windungzahlen. At half-turns, eg for a forward converter with a core without air gap, it is not guaranteed that compensate for the problems caused by the load current magnetic fields of secondary and primary coil exactly. This can result in partial saturation effects in the core and a substantial increase in the leakage inductance. Due to the required transmission ratio are half turns absolutely necessary, you can make do with the fact that one whole turn attaches to the two outer legs of an EI core and this parallel.
With increasing air gap length, but this limitation is becoming less important. Of web and rod cores, the exact number of turns can then anyway not so easily determined. In principle, any fraction of a turn is possible and allowed. Contributes especially to the coil part of the wire which is tightly wound on the core.

1 50-Hz chokes and transformers

To (Power) to convert AC voltages are 50 Hz transformers still the easiest and usually cheapest voltage converter. Even in the age of fast switching power supplies and power semiconductors, the 50-Hz transformers are an essential part of electronics. Finally, with these components is a no problems with Oberwelleneinstreuung in the 230-volt AC or high frequency radiations. In terms of reliability and immunity to power surges such transformers are likely to still be second to none.
Although the operation of a transformer is in almost every book, but I would still like to summarize again: The primary coil is connected to an AC voltage, eg 230 V mains voltage down. The coil current produces a magnetic field which in turn induces a voltage in the primary coil. Thus, the coil current not to skyrocket, the voltage

induced in the primary coil voltage as the applied (power) must be approximately equal tension. Thus, the applied voltage forces the magnetic field inducing it.Succeeds in a second coil, the secondary coil so placed that they will be carrying the same magnetic flux, the same voltage per turn is induced in the primary coil as well as in it. Is charged to the secondary coil, it creates an opposing magnetic field that the primary magnetic field weakens. Since the primary magnetic field is imposed, but in its strength due to the applied primary voltage, it can be maintained only by the primary current corresponding additional secondary load. Thus, it is then transferred to the primary output of the secondary coil. During the construction of transformers, there are still practical problems: first, relatively high magnetic field amplitudes are required at 50 Hz to induce an appreciable voltage in the coil and on the other it is very difficult to ensure that through the two coils of the same magnetic River flows. Both problems can be largely solved by using a closed core of soft iron. The high Pemeabilität iron (permeability or conductivity for magnetic fields) for setting up a certain strength is only about a zehntausenstel magnetic flux of the current required would be required for an air coil. Thus, the construction of 50-Hz transformers is at all possible. The high permeability of soft iron for magnetic fields also also ensures that there is hardly a field line takes the shortcut through the air and so practically all of the magnetic flux must pass through the iron core. So that all coils are located on the core carrying the same flow automatically. The conditions are unfortunately not so ideal if the secondary coil of the transformer is loaded with a current. The opposing magnetic field generated by the secondary coil reduces the effective magnetic permeability of the iron and causes the one or the other field line to, but to take a shortcut through the air and on the secondary coil over. This undesirable, called scattering effect is even stronger as the coils are spatially separated from each other. The practical effect of dispersion is that already existing with the ohmic resistance of the copper wire nor an inductive component, the so-called leakage inductance added. Thus, the scattering increases the internal resistance of the secondary voltage and thus their load dependence.

Another problem are the voltages induced in the iron core and the eddy currents flowing therethrough. If you were to use a solid iron core of ordinary soft iron, the efficiency of the transformer would not only deteriorate significantly, but the iron core would be hot and cause severe cooling problems. Since the induced voltages are inevitable in the core, the eddy currents can devalue only by reducing the electrical conductivity of the iron. The most effective reduction of conductivity is achieved by dividing the core into as many individual sheets. The plates lie in the direction of the magnetic field lines, so that the conductivity of the core for the magnetic field is not impaired. However, the eddy currents flowing perpendicular to the magnetic field lines, the limits between the sheets that are insulated from each other, not overcome. It can significantly smaller eddy currents flowing within the individual plates then only. These residual eddy currents can be reduced again by the electrical conductivity of the iron is reduced significantly by the addition of a few percent silicon.

1.1 The common core designs

The core designs are highly standardized and are called according to plate size and shape. The following types are commonly used:

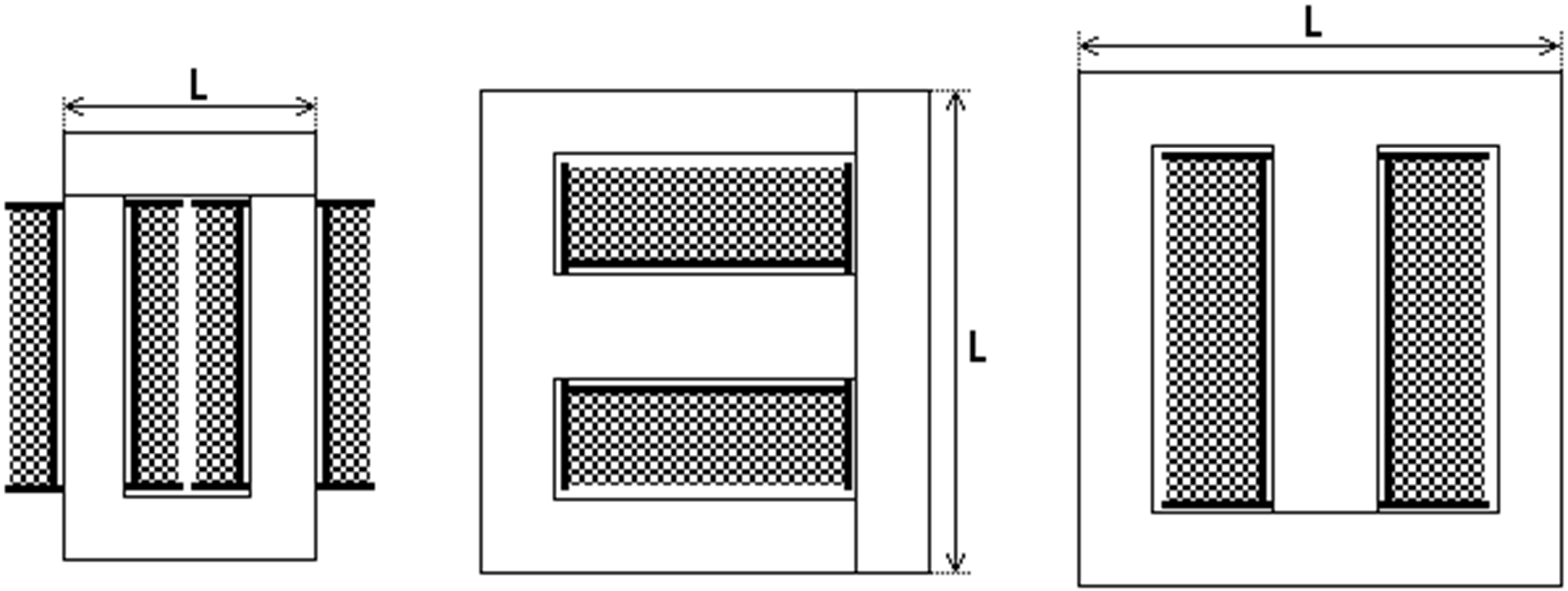


Figure 1.1 A UI core Figure 1.1 B 1.1 C EI core image M-core

Figure 1.1 A shows the simple UI core, which, as the name suggests, consists of individual plates in the U and I shape. The UI core both legs can be provided with a coil bobbin. The easiest way to accommodate a spool on the primary and on the other the secondary coil would be. Unfortunately you bought with this simplification, a significant increase in scattering, which increases the internal resistance of the output voltage accordingly. UI-core transformers are therefore almost always constructed symmetrically, so that the primary and secondary coil is one half on the one side and the other coil former. Of course, this splitting increases the manufacturing costs. The advantage is the flat construction (use for flat transformers) and good utilization of the iron core, which is almost completely wrapped.Which consist of a U-and I-shaped sheet metal layers are typically alternately pushed in opposite directions into the coil body. This increases the stability of the laminated core and mainly reduces the effective air gap of the core. In Figure 1.1 B the most common in young Transformers EI core is shown. As with the single UI core plates are arranged in opposite directions and the EI-core. The advantage of the EI core is that all the coils can be inexpensive wound on a bobbin. Depending on requirements, the windings above the other (low scatter, low internal resistance) or in two separate chambers next to each other (good protective isolation) are arranged. For ease of manufacture, the E-and I-package is often compressed to the bobbin and is welded to the outer edges. Since the welding seam runs only along the outer edge, and so does not form a closed ring around the core, no significant eddy currents can flow there. EI core transformers have with market plate dimensions, in contrast to the relatively flat UI core transformers, rather a cubic design.

An improvement of the EI core, the M core is shown in Figure 1.1 C. The sheets consist of a layer now only one piece. The magnetic flux must overcome only the nor an air gap in the center leg. This child is still minimal because the outer legs do not allow pulling the sheet. The installation of the sheets is this something difficult, maybe a reason why the M cores are somewhat out of fashion. Instead zusammenzuzetzen the core of individual sheets, you can also wrap it in a long strip. This not only simplifies the manufacture of the core, but veringert, because of the similarity of field lines and sheet metal course, even the stray fields. In Figure 1.1 D, a so-called C-core is seen. In order to move the sheet stack in a bobbin, it must first be cut open (hence the name). Thus, the air gap does not form an integral assembly, the cut surfaces of the core pieces to be ground flat and pressed together permanently with the aid of a steel clamping band around the core. Common types are both shown in Figure 1.1 D, consisting of four core pieces version with double outer leg, the Mimics the EI or M-core and the corresponding the UI core simple version with two bobbins and only two core pieces. Cut cores are found only rarely, since they were largely replaced by the cost and better toroids.

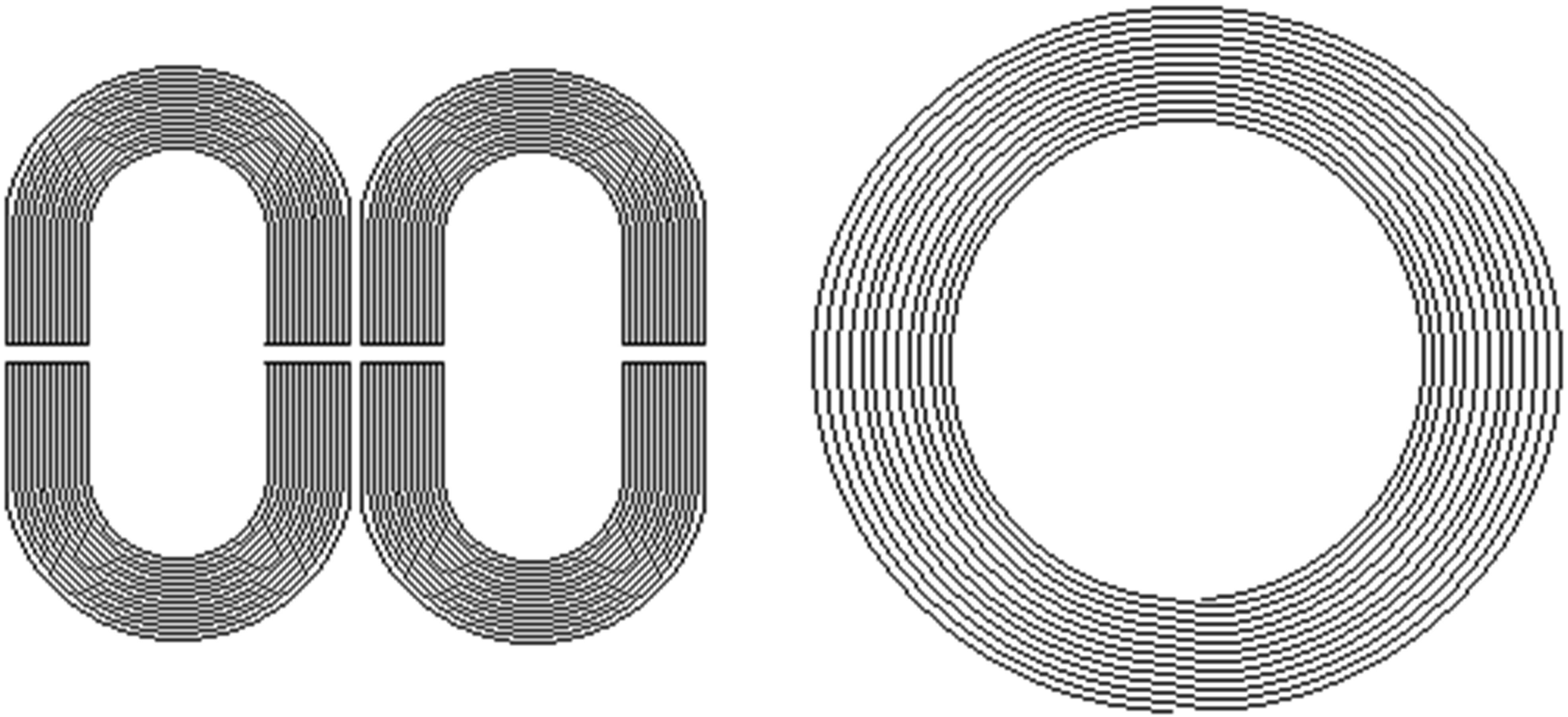


Figure 1.1 Figure 1.1 D C-core E toroidal core

1 The toroidal core shown in Figure 1.1 E is actually the ideal core. In contrast to the C-core band in the iron toroid core circular wound and not cut. Since the magnetic field lines of an annular air coil (toroid) would run there anyway, where the iron core is, they can not take any shortcuts outside of the coil. The toroidal transformer is therefore almost stray field free with uniform winding. However, this does not mean that the toroidal transformer leakage inductance does not. The double meaning of the term stray field occurs here very clearly to light. While most users are only interested in the urgent outward Störstreufeld that induces unpleasant ripple voltages, especially in audio equipment that is not urgent outward, likewise annular leakage field is responsible within and between the primary and the secondary windings for the leakage inductance. Due to the close arrangement of the primary and stray inductance of the secondary coil, however, is lower than for

all other core shapes. This is also the isolation of large and somewhat problematic, especially at high voltages. Toroidal transformers are therefore primarily used for applications up to 230 Volts AC. For higher voltages used in the main EI or UI cores. For the winding of toroids special winding machines are required. Only such machines allow for economic production of these transformers.

1.2 The dimensions of 50-Hz transformers

A transformer is then optimally exploited when the iron is just not falling into the saturation in the core. In this case, the maximum magnetic flux and thus also the maximum induction voltage is reached. The voltage required for a certain number of turns is minimal and the potential for vorgegebenenm winding cross-section wire diameter maximum. With a lower number of turns to drive the core would inevitably lead to saturation, and the greatly reduced inductance of the primary coil during saturation leads to a rapidly increasing current, leading along with the saturation losses in the iron core of the transformer an overload. Increasing the number of turns of wire on the primary and secondary coil is longer and thinner (it has to be thin to fit on the limited winding cross section). This not only increases the leakage inductance, the resistor and additional wire leads to a lower capacity of the transformer.

1.2.1 The required size of a transformer core

The size of the transformer for a required performance is an empirical value. Standardized forms for these core values are at the transformer manufacturers but roughly known. If in doubt, take the next largest core. Anyone who is not in possession of this experience, can help with the similarity laws. This one looks for a known standard with known nuclear power in the mid range, ie EI60/21 with 20 VA. Because of the inductive reactive power-related share of the total power absorbed by the transformer, the power is usually expressed as apparent power with the abbreviation VA. A significant deviation of the apparent power of the active power output actually achieved are found primarily in transformers with very low power and / or high leakage inductance (leakage transformers). Normally, the VA can be directly replaced by Watts. The core message of the laws of similarity is that in a true to scale enlargement or reduction of an original by a factor k all routes to the k-fold increase in size or smaller. This is actually natural and easy to understand, but sometimes has amazing consequences. As an immediate consequence of this fact is that all surfaces to the k^2 times, all volumes and masses zoom in and out to the $k^{3\text{-fold.}}$ If one wishes to a transformer to k times larger, the cross-sectional area of the iron core increases, thus, the maximum magnetic flux to the ^{2} k-fold. The required number of turns are then reduced to the k^2 times, the Windungsumfang increased to k times the length of the wire windings is therefore reduced to k times. At the same time is now, at a $^{2\text{-fold}}$ reduced by k turns, the $k^{2\text{-fold}}$ winding cross section is available. The wire cross-section can thus be increased by k^4 times. Because of the extension of the Windungsumfanges while reducing the number of turns around the k^2 times the wire resistance finally reduced to the $k^{5\text{-fold.}}$ The allowable power dissipation is approximately proportional to the surface, so that k is increased by $^{2\text{-fold.}}$ Since the power loss increases with the square of the load current, that is also the square of the transmission power, the reduction of the resistance wire ^{5} to the k-fold allows an increase in the transmission power at the $k^{2.5.}$ The increase in surface area and power dissipation allowed to k^2 allows to further increase the transmission power by k times. This yields the following general conversion rule: If the dimensions of a transformer at a constant multiplied by the frequency and type of construction scale k, the achievable transmission power multiplied by the factor $k^{3.5.}$

What is astonishing about this result is that an increase in the transformer output to rise more than the volume and mass. However, the iron losses are proportional to volume. Since enlargement of the transformer, the volume increases faster than the cooling serving surface, these losses fall with transformer size is more significant.Eventually, the sheet thickness must be reduced or the core for additional cooling. While the transmission power increases by a factor $k^{3.5,}$ increases the power loss, according to the cooling surface, only to $k^{2,}$ which means a considerable increase in the efficiency of large transformers: From this calculation, there are two other advantages of large transformers result. The high efficiency of large transformers with low leakage inductance goes well with a low internal resistance and copper along. Accordingly, the output voltage is smaller transformers is load dependent than the Great. Small transformers are for these reasons also unempfinlicher against overload. Very small transformers to about EI 30 are usually even permanently short-circuit proof. The open circuit voltage is less transformers (eg EI30) have the same reasons with about 1.5 times the normal load voltage are applied. The efficiency is then at normal load below 70%. Conclusion: Thus, large power transformers work more effectively than small ones.

One problem with the translation of the output classes is that the standard cores are not always the same shape (similar). Often, only the thickness of the laminated core is increased to increase performance. Ideally, the iron core in the spool should, however, have a square cross section (more preferably be circular). For square or circular cross-sections can be known, wrap with minimal wire length, the maximum cross-sectional area. For a rough calculation of the expected transmission power but the similarity analysis is a good orientation. If, for example, a transformer with EI core transmit a power of 80 VA (watts), one first takes the above-mentioned 20-VA transformer EI60/21 pattern. The power that is expected to quadruple, ie, $k = {}^{3.5}$ $k = 4$ so $4^{1/3,\,5} = 1.49$. The same shape core would then have an edge length of about 89 mm. The core would be the closest standard EI88 core.

1.2.2 The calculation of the number of turns

If one has once decided on a core, the required number of turns very well calculated. Assuming that soft iron cores can be magnetized to a magnetic field strength of 1.5 Tesla, can be achievable at 50 Hz current voltage, which is the induction voltage of a winding, calculated as follows: I would like to first of all a qualitative approach to connection between the transformer voltage and magnetic field to insert. From the alternating current theory is known that in an ideal coil, the coil current leads the voltage by 90 ° or □ / second This means inter alia that the current is zero and the maximum voltage at the zero crossing is at a maximum. The current value is irrelevant for the following discussion and in practice. However, he is always proportional, so also in phase with the magnetic field. The magnetic field in the iron core is therefore in the positive peak of the AC voltage is zero and is built up to the zero crossing to its maximum value. To the negative peak of the voltage has to be reduced again to zero. Until the next zero crossing of the voltage has reached the maximum value in the opposite direction, to then finally to the positive peak point of the voltage completely degrade again. The structure of the magnetic field of a coil thus always takes place in a quarter period, starting at the apex to the next zero crossing of the voltage. The current voltage of a coil is the same as the temporal change in the magnetic flux in the iron core (flux change per second). If one of a voltage to the coil, the flow increases in proportion to voltage and time. As the voltage varies with time, voltage, and time can not be easily multiplied, but it has a definite integral be calculated. Who is not familiar with calculus, I would like at this point so spare, and who are familiar with it, looking anyway, what comes out.Figuratively, one can imagine the magnetic flux as the area under the current-voltage curve in any case. A special feature of the unit sine function (peak 1, angle in radians plotted) is that the area under the half-wave is exactly two. The area under a quarter period is accordingly one. In order to transfer these very abstract area into reality, must (the one) will be multiplied by the peak value of the current-voltage $\hat{U}_1$ and those related to the period T 2π radians of the AC the whole. The maximum magnetic flux in the core then errechet with the formula $\hat{U} = \hat{U} = {}_1\, T/2\pi\, {}_1\, 1/2\pi f$. By rearranging the equation, then the maximum current voltage results with $\hat{U}_1 = 2\pi f \quad = 2\pi f A$. Where f is the frequency in Hz, A is the cross sectional area of the iron core in square the maximum magnetic field strength in

Tesla (1.5 T). Substituting the fixed normally sizes f = 50 Hz and = 1.5 T in the equation and suggests the units (the physicist may pardon this sacrilege), we obtain the formal wrong but practically useful formula $\hat{U} * _1$ = 470 A or for the effective voltage U_{eff} = 333 * A. If eg a power transformer are wound, the iron core has a square cross section with 4 cm edge length, we obtain a maximum current of voltage $\hat{U}_1$ = 333 x 0.04 x 0.04 = 0.53 $V_{rms.}$ At 230 Volts AC 433 then turns to be wound.

1.2.3 The calculation of the wire thickness of a transformer

As a final size of the wire cross-section is calculated. To the coils, a winding cross section is allocated first. Typically, the primary coil is about half and the secondary coil (s) the other half of the available cross-section coil. When the area ratio of said plurality of secondary coils of each coil should be approximately the same to the respective power component. Dividing the space allocated by the number of turns, one obtains the theoretical cross section of each wire winding. Because of the extra insulation and any gaps between the wires but you will not use much more than half of the allocated area. By precise winding technique and rectangular wire cross-sections of the so-called filling factor can be even more in the case of series products.

Since round enamelled copper wires always the diameter is specified, the wire cross section must still be converted into diameter $D = 2 \sqrt{} A / \pi$.

2 50 Hz rectifier and filter circuits

In this chapter I would like to treat the rectifier and filter circuits, which are located in virtually all electronic equipment that operate on 230 Volts AC. Most of these circuits consist of a bridge rectifier and an electrolytic capacitor, but even in these simple models there is to learn a few things to know.

2.1 The half-wave rectifier

The half-wave rectifier consists only of a diode and a Siebelko. The application is mainly limited to low power supply voltage, where you want to save the bridge rectifier. At times, as rectifiers were still expensive and possibly composed of many large selenium cells, we used half-wave rectifier for the supply of larger consumers. The disadvantages of wave rectifier occur but just at high power light: As the Siebelko is charged only 50 times per second, this must be twice as large as in Zweiweggleichrichtern that the electrolytic capacitor charging 100 times per second. Since only one half wave of the AC voltage is charged to obtain an asymmetric distortion of the mains voltage, which makes creating especially the power transformers because of the resulting DC component. In addition, the leads are still burdened with an increased RMS current. Although the time average of the input current amount is always identical to the output current, the half-wave rectifier, the input current can but only 50 times per second during the short time of the apex of a half-wave flow. To get to the same output stream of a full wave rectifier, the current pulse must have approximately twice the thickness. In the ohmic resistance R of the supply line, the power loss $P = RI^2$ is reacted at that time. While the frequency of the current pulses at 50 Hz cut in half, roughly quadrupled the converted during a pulse output. In the middle is finally obtained at approximately the double half-wave rectifier conduction loss as the full-wave rectifier.

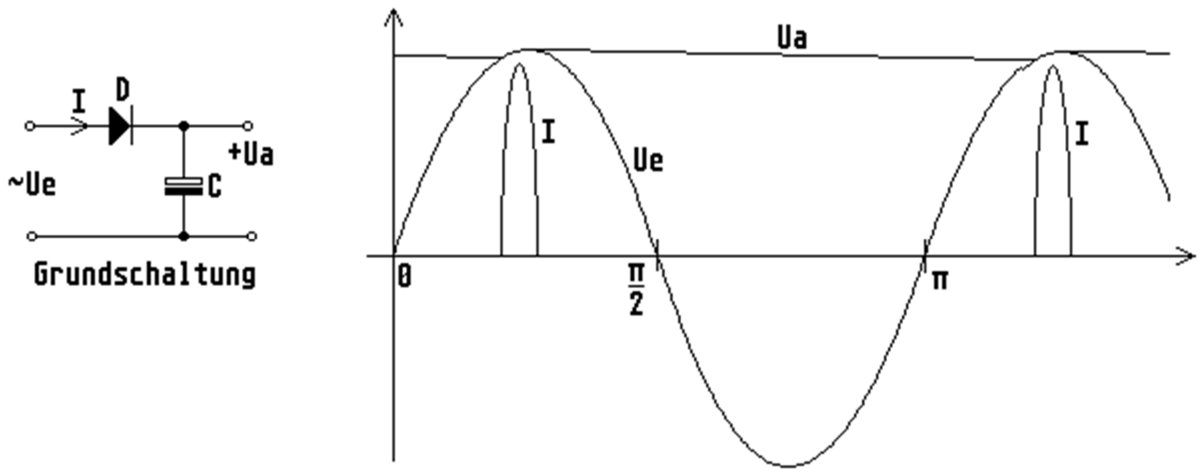


Figure 2.1 Voltage and current waveforms at the half-wave rectification

In Figure 2.1, the voltages Vin, Vout and the current I plotted against each other. Shortly before the vertex of the positive sinusoidal half wave, if the input voltage exceeds the voltage at Siebelko, diode D starts to conduct. At this moment, therefore, the current pulse the Siebelko C starts charging. Behind the apex of the input voltage drops below the capacitor voltage and the current through the diode comes to a halt. To the next vertex, where the process begins again, the electrolytic capacitor to discharge. The discharge rate of the capacitors depends on its capacity and the discharge current. To calculate the capacity, you have to consider how far the electrolytic capacitor must discharge between two current pulses, without the function of the consumer is compromised. This can be done easily remember that a 1 amp loaded with 1000 uF electrolytic capacitor to 1 volt / ms discharges. When the half-wave rectification at 50 Hz (20 ms) would mean 20 volts discharge. In the selection of the diode is to make sure that they can withstand the impact of the current pulses, especially the starting current, and that a reverse voltage is applied to it during the negative vertex up to twice the peak value of the input AC voltage. At 230 Volts AC are the example 2 x 230 x 1.41 = 649 volts. To survive mains spike, power rectifier diodes are used for one-way rectification 1000-1300 volts reverse voltage. Especially in rectifiers that operate directly on the 230-volt mains voltage, the current pulse when switching on with a discharged Elko can rise to several hundred amperes. To avoid possible damage to the switch, diode and other components, the inrush current is limited by a resistor in series with the diode. Due to the high RMS current in the diode, however, the power loss in the resistor is relatively high. In case of power rectifiers for devices with small and medium-sized power resistor values on the order of 5 ohms are common. This value represents a good compromise between the necessary limitation and loss of power in continuous operation dar. Often, disc-shaped thermistors are used for inrush current limiting.This devalue in operation because of their resistance heating and thus reduce the power dissipation in continuous operation. Another frequently used technique is the short-circuiting of the limiting resistor after switch to a relay or a thyristor or triac.

2.2 The bridge rectifier (Graetz bridge)

The main advantage of bridge rectifier compared to the half-wave rectifier is that both the positive and the negative half wave of the AC voltage to charge the filter capacitors are used. This has several advantages: The Siebelko now needs only half the time between two current pulses received the minimum supply voltage upright. Strictly speaking, the backup time is even shorter because the gap between two consecutive half sine waves is much shorter than 10 ms. The period in which no power is available, is comparatively short. For a half-wave rectifier is about 10 ms absolutely no tension. Even a tolerant because consumers can easily suffer dropouts.

The Siebelko the bridge rectifier needs therefore, with the same power, be a maximum of only half as large as the half-wave rectifier, a significant space and cost savings for high performance. The symmetric load, the voltage source is especially important in case of power transformers. Firstly, the performance potential of the transformer is much better used and on the other you risk not overload the transformer due to saturation effects in the iron core. In the half-wave rectification occur in the transformer coils resulting DC currents, which can very easily lead without an air gap for magnetic saturation of the iron core with seeds.

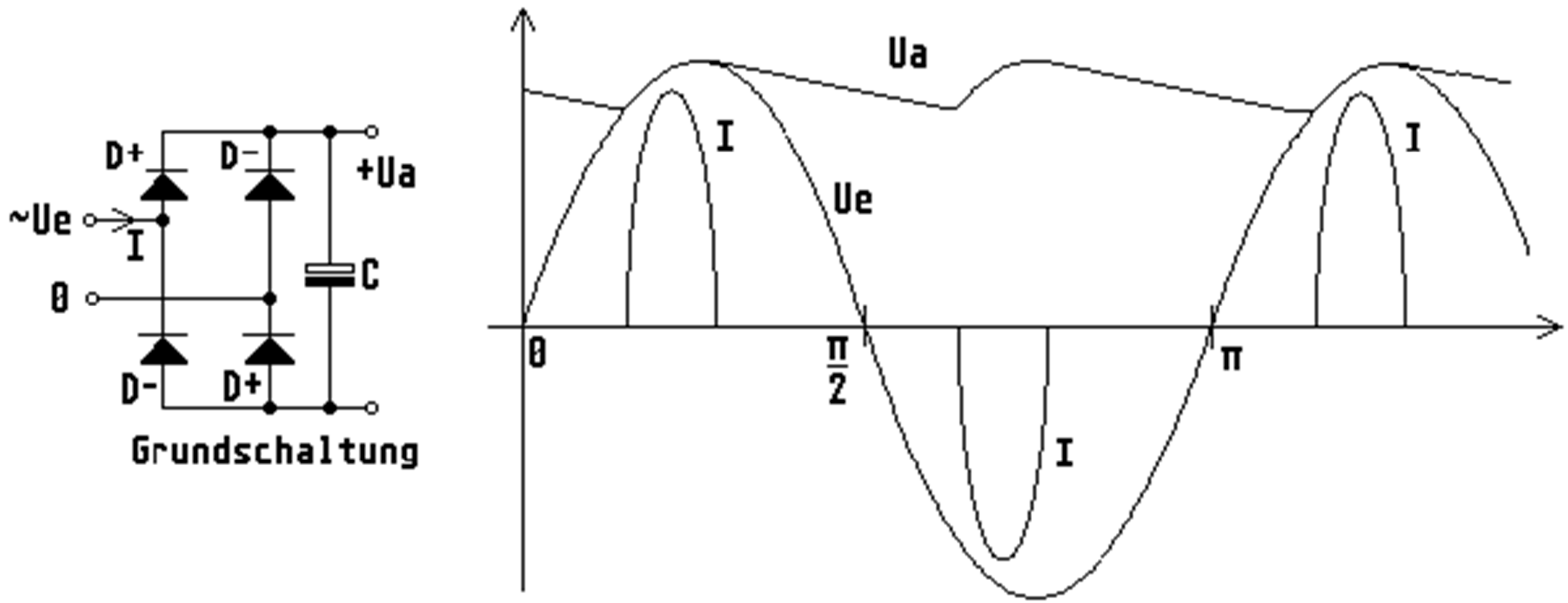


Figure 2.2 Voltage and current waveforms at the bridge rectifier

In Figure 2.2, the basic circuit consisting of four diode bridge rectifier is shown. Regardless of the polarity of the input voltage Ue Ue the negative pole of one of the lower diode is always connected to the negative pole and the positive pole of Ue on one of the upper diode to the positive terminal of the filter capacitor. In the positive peak of Ue, the diodes D + and the negative, the diodes D-switched. The maximum peak reverse voltage across the diode bridge rectifier is the only simple peak value of the AC voltage. Of course, the bridge rectifier also has disadvantages: Since the current must always flow through two diodes, one has to silicon diodes at higher load a voltage drop of about 2 volts. For small AC voltages this loss can already strong consequence. By using low-voltage Schottky diodes (up to 60V reverse voltage) can be the loss of power and of course the power dissipation in the rectifier approximately halve. At higher voltages, Schottky diodes bring no benefit, since then similarly high flow stresses reach as ordinary silicon rectifier.

Another disadvantage of rectifier bridges is that the input and output voltage have no common ground. Therefore, either the input or output voltage must be potential-free. This means for example that in measurements behind a bridge rectifier that operates directly from 230 volt mains, always an isolating transformer must be installed. Otherwise, one would inevitably cause grounded when touching the ground of the oscilloscope to the negative pole of the filter capacitors short-circuited.For security reasons, you should use the isolation transformer but anyway when working in the mains voltage range. As rectifiers were very large and expensive, the need of four diodes was a major drawback of the bridge rectifier. Since silicon diodes and bridge rectifiers are also ready to get very cheap today, this factor has become meaningless. The limitation occurs in the same manner as the half-wave rectifier.

2.3 The midpoint rectifier

The center of the rectifier allows despite saving two diodes, to take advantage of the bridge rectifier in terms Siebelko. However, the operation of the center is a network transformer with double secondary winding needed.

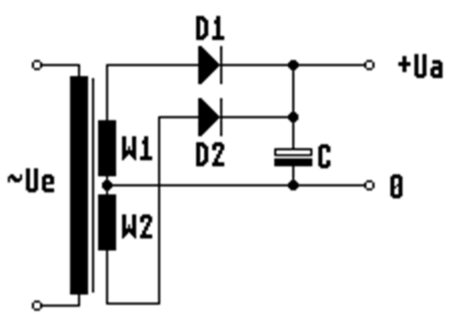


Figure 2.3 A
Simple midpoint rectifier

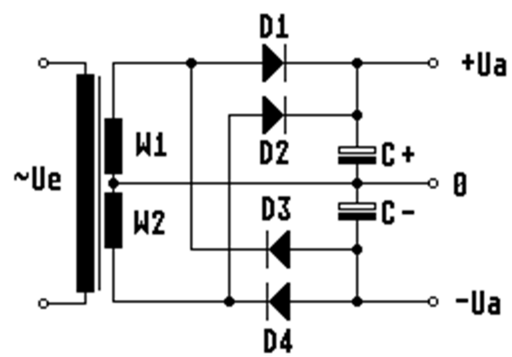


Figure 2.3 B Double midpoint rectifier

The windings W1 and W2 are connected so that they represent two equal and opposite phase AC voltages available. In Figure 2.3 A, a simple middle rectifier is shown. Each of the diodes D1 and D2 operating as a half-wave rectifier. However, since the diodes operate in push-pull, of the capacitor C is therefore charged twice per period every 10 ms. When dimensioning the filter capacitors can proceed as in the bridge rectifier. Of course, here, the half-wave rectification of the individual diodes leads to a resultant direct current to the windings W1 and W2. The harm the transformer but nothing as cancel arising in W1 and W2 opposite DC magnetic fields are mutually exclusive. The simple midpoint rectifier is mainly used at times as diodes costs still play an essential role. Today it is used only at low voltages to avoid the double loss of voltage (and power consumption) of the bridge rectifier. At higher voltages, then outweighs the disadvantage of higher cost transformer due to the additional connections and additional copper wire.

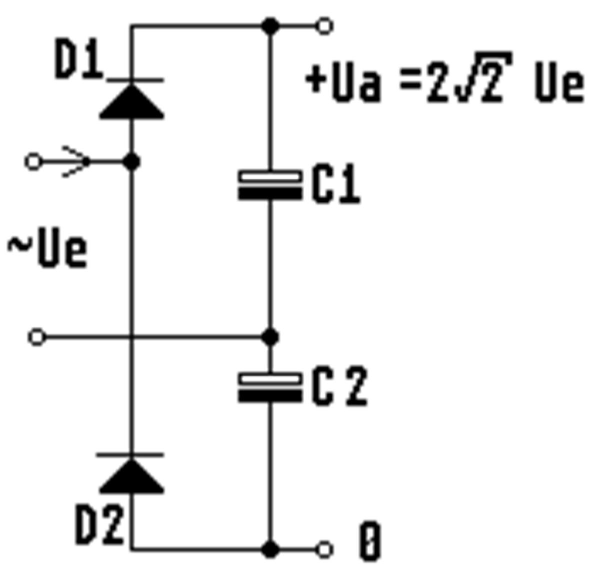
Figure 2.3 B shows that on the other hand can be used very advantageously for symmetrical supply voltages double the center tap. For this, the previously unused negative half-waves of the winding tension can be easily rectified by the diodes D3 and D4 and smoothed by the electrolytic capacitor C. Positive and negative output voltage thereby have a common ground. Who the diodes of the circuit considered in more detail, will find that it is identical with the bridge rectifier. Therefore, the dual center tap allows the use of standard rectifier bridges. Since the center of double rectifier is comprised of two independent single, unbalanced load, a strong positive and negative power supply is allowed. Another application of the double center circuit is the generation of two operating voltages of the same polarity. This is usually defined as the negative output voltage mass and obtained two positive output voltages, one of which is twice as large as the other. A practical example is the generation of a +5 volt and 12 volt power supply for TTL circuits having logic blocks and analog technology.

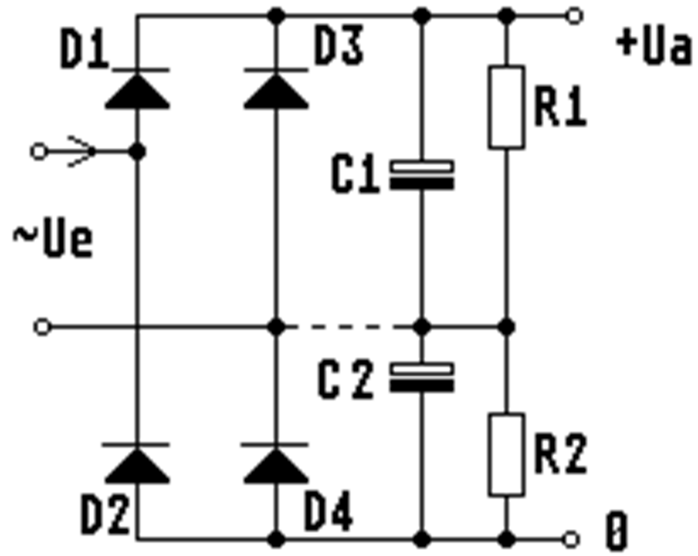
2.4 rectifier with voltage multiplication

Sometimes in a circuit higher voltages are required, as the peak value corresponding to the available AC voltage. This can have different reasons:

1. When the AC voltage is not variable voltage.
2. A transformer with an additional winding would be too expensive.
3. In a high-voltage power supply has a high voltage winding would be too expensive.

The circuits described are all "transformer-friendly" because it puts positive and negative half-waves alike and because of the capacitive coupling prinzipiel allow any resulting DC currents. The multiplier circuits are the simplest Verdopplerschaltungen. In Verdopplerschaltungen obtained as the output voltage peak-peak value of the input AC voltage. Which again corresponds to the maximum blocking voltage of the wave rectifier, voltage at 230 volts so 649 volts.





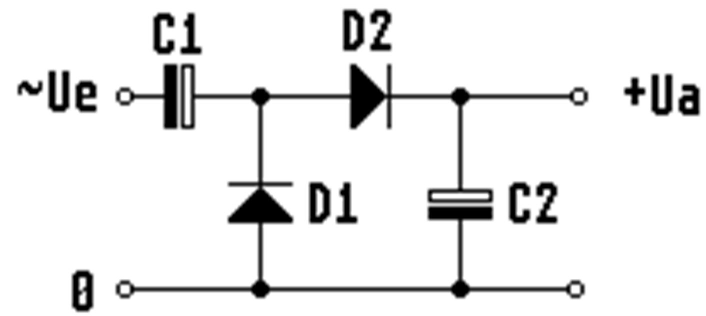


Figure 2.4 A
Doubler,
symmetrical (Delon)

Figure 2.4 B
Doubler,
Bridge rectifier or
Doubler,
symmetrical (Delon)

Figure 2.4 C
Doubler,
asymmetric (Villard)

In the Verdopplerschaltungen there are two different circuit versions with different advantages and disadvantages. In Figure 2.4 A symmetrical version is shown. The diodes D1 and D2 each operate as a half-wave rectifier. D1 aligns the positive half-wave of the same, which is sieved with C1. Accordingly, the negative half-wave is rectified by D2 and sieved. At the two capacitors is then present the total peak-to-peak value of the AC voltage. The advantage of this circuit is that it can be optionally used in conjunction with a bridge rectifier and an optional doubler or bridge as a bridge rectifier. A very popular application is the power rectifier, which can be selectively switched 115-230 Volts AC. In the "230-volt mode" diodes D1 to D4 are working as a bridge rectifier. C1 and C2 are simply connected only in series and provide a total capacity of half the value of C1 and C2. C1 and C2 also need therefore only half the DC voltage Ua endure. The resistors R1 and R2 are equal and should ensure that the total voltage is distributed equally between C1 and C2. The exact value is not critical, but should be large enough that no additional cooling problems or unnecessary losses. In the 115 volt mode the junction of diodes D3 and D4 is connected directly to that of C1 and C2 (dotted line). D3 and D4 are now directly parallel to the capacitors C1 and C2 in the reverse direction and have no function. Left therefore remains only the simple doubler from Figure 2.4 A. For the interpretation of subsequent filter circuits should be noted that the residual ripple frequency, although it is two-way rectifier with 50 Hz ripple frequency is 100 Hz. This can be explained that the 50-Hz fundamental frequency together with its odd harmonics will be wiped out by the in-phase, but a half period time-delayed addition of the individual sawtooth ripple voltage thereby. Remain the even harmonics, which again form a sawtooth wave at twice the fundamental frequency. Enter into this in more detail at this point but would lead too far. A disadvantage of this doubler is how the bridge rectifier that input or output voltage must be isolated, ie have no common ground.

In Figure 2.4 C, the asymmetric doubler can be seen. D1 and C1 work as an ordinary half-wave rectifier, where C1 is charged with the simple peak value of the AC voltage. Since the negative terminal of C1 is connected directly to the input AC voltage, the DC voltage is added to the positive terminal to the alternating voltage.Thus, a voltage is generated at the positive pole, swinging their value between zero and the peak to peak value of the AC voltage. The positive peak of this pulsating DC voltage is then connected with the electrolytic capacitor C2 to D2, then the double peak is at the same voltage as available. C1 is charged during the negative and C2 during the positive half-wave. The residual ripple frequency is in the asymmetric doubler therefore only 50 Hz dimensions of C2 is therefore also like to make the half-wave rectifier. During the positive vertex, if C2 is charged, C1 is discharged and there is except through the ripple amplitude, an additional voltage loss. A sensible sizing it would be safe if the power loss through the discharge of C1 and C2 is equal relative to each Kondersatorspannung. As C2 to double the voltage across C1 is applied as this is precisely the case when the capacitance of C1 is twice that of C2. The great advantage of asymmetric doubler circuit is that the input and output voltage have a common ground and that they can be cascaded. With two capacitors and two diodes per phase voltages to each of the peak to peak value per stage can be added. In Figure 2.4 a three Vervielfacherkaskade D is shown. The output voltage is then ideally, almost 8.5 times the RMS value of the AC voltage, AC voltage at 230 volts that would be nearly 2000 volts.

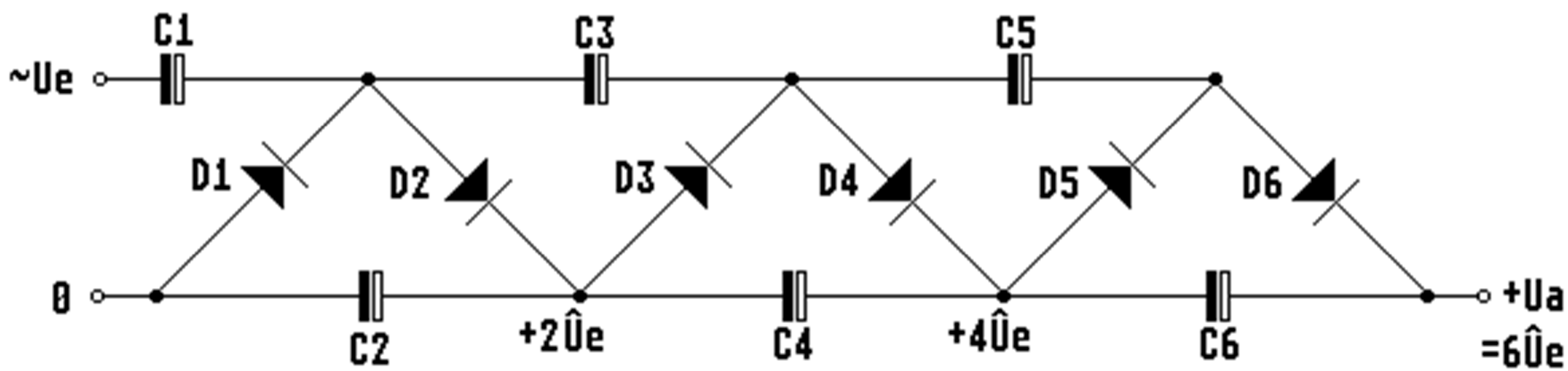


Figure 2.4 Three-D Vervielfacherkaskade for positive output voltages

High voltages are used mainly for the production Vervielfacherkaskaden. This way you can avoid the problems due to isolation rather expensive high voltage windings. This applies particularly to smaller power transformers that do not allow high output voltages due to their mechanical dimensions that way. The efficiency of the cascade is very high, since at higher voltages, the forward voltages of the diodes are not insignificant and cause the capacitors with good sizing hardly any losses.The optimal sizing of the capacitors is somewhat more complicated than for the doubler. Note, first, that except C1 applied the double peak voltage capacitors at all.Clearly, it should be clear that, for a voltage multiplier the input current must be substantially greater than the load current Vout the output voltage. If, again requires the condition that the same relative ripple voltage to be built in all capacitors, the capacitors have to be designed accordingly in the first stage is greater than that of the last stage. This is especially important if cost of expensive high-voltage capacitors are to be saved. Are the cost of the capacitors no major role, are used because of the ease of procurement, mostly capacitors with the same capacity. It is relatively complicated and difficult to see through, to consider the charge transfer in the individual capacitors at each half cycle. With a simple consideration, however, it is faster to the destination: a load current at the output, which naturally can not flow through the capacitors, and therefore may occur through the diodes D1-D6 as the resulting DC current flows only. Since only capacitors are connected to the connection points of the diode, this DC branch, and, therefore, may not need to be the same in all diodes. Since each diode turns on only once per period, the transmitted during the current flow phase in charge must be equal to all the diodes. For the optimization of capacity should we start in the last stage. Here we consider only the positive peak of the alternating voltage when a portion of the charge of C1, C3 and C5 is transmitted to filter capacitors C2, C4 and C6. At C5 and C6 are the same voltages. Therefore, their capacity must be equal. Harder it is already at C3 and C4. C4 in the same size and charge currents of D4 (on C6), D6 of flow.C4 on a total of twice the charge enters as on C6. At the same ripple voltage so C4 must be twice as large as C6. The same will also apply to C3, C5 be twice as large as the must, because it transfers the charge of C5 through C6 and D6 and the equal charge of C4 through D4. Then finally get the charges of D2, D4 and D6 C2. C2 must therefore be three times greater than C6. Since C1 is applied to only the simple peak voltage, it must be twice as large as C2, that is six times as large as C6 again. The whole thing can of course continue as desired, however, the relative Kapazitätzänderung between the first stages with increasing number of stages from, so it is not always worthwhile to use different capacitors. In the previous discussion, I have only to each other determines the ratio of the capacitance values. To specify an absolute dimensioning, take back the last stage of which is so loaded with exactly the output filter capacitor current. In an n-stage cascade-optimized, wherein the first filter capacitor is n times as large as the last, a capacitor is formed on the last n-th of the total ripple. The last condenser is therefore dimensioned so that when a maximum load current of the n-th maximum allowable ripple voltage can not be achieved. The calculation of the ripple voltage is identical with that of wave rectifier. The stress of diodes is basically the same in all stages, but are subject to D1 and D2 increased Einschaltstrombelastung. When using different sized capacitors should also be noted that in the event of a short circuit of the output voltage discharge the capacitors of the first stages of the diodes in the final stages and can destroy them. A protective resistor for current limiting should be mounted in the output line.

About the properties of the multiplier circuits are various rumors floating around that I want to correct it again at this point.

1. The efficiency is very high in principle. If the forward voltage of the diode is low relative to the input voltage, the losses are very low.
2. The internal resistance of the output voltage can be very low when the capacitors are sufficiently large.
3. All diodes are loaded equally in the middle to the output stream. The exposure of the diode is otherwise independent of the input current and the number of stages.

2.5 Three-phase rectifier

Stands in a three-phase device available, it is useful to provide greater DC load with a bridge rectifier for AC. The three-phase rectifier is almost the same way as the normal bridge rectifier, but has a third pair of diodes to connect the three phases.

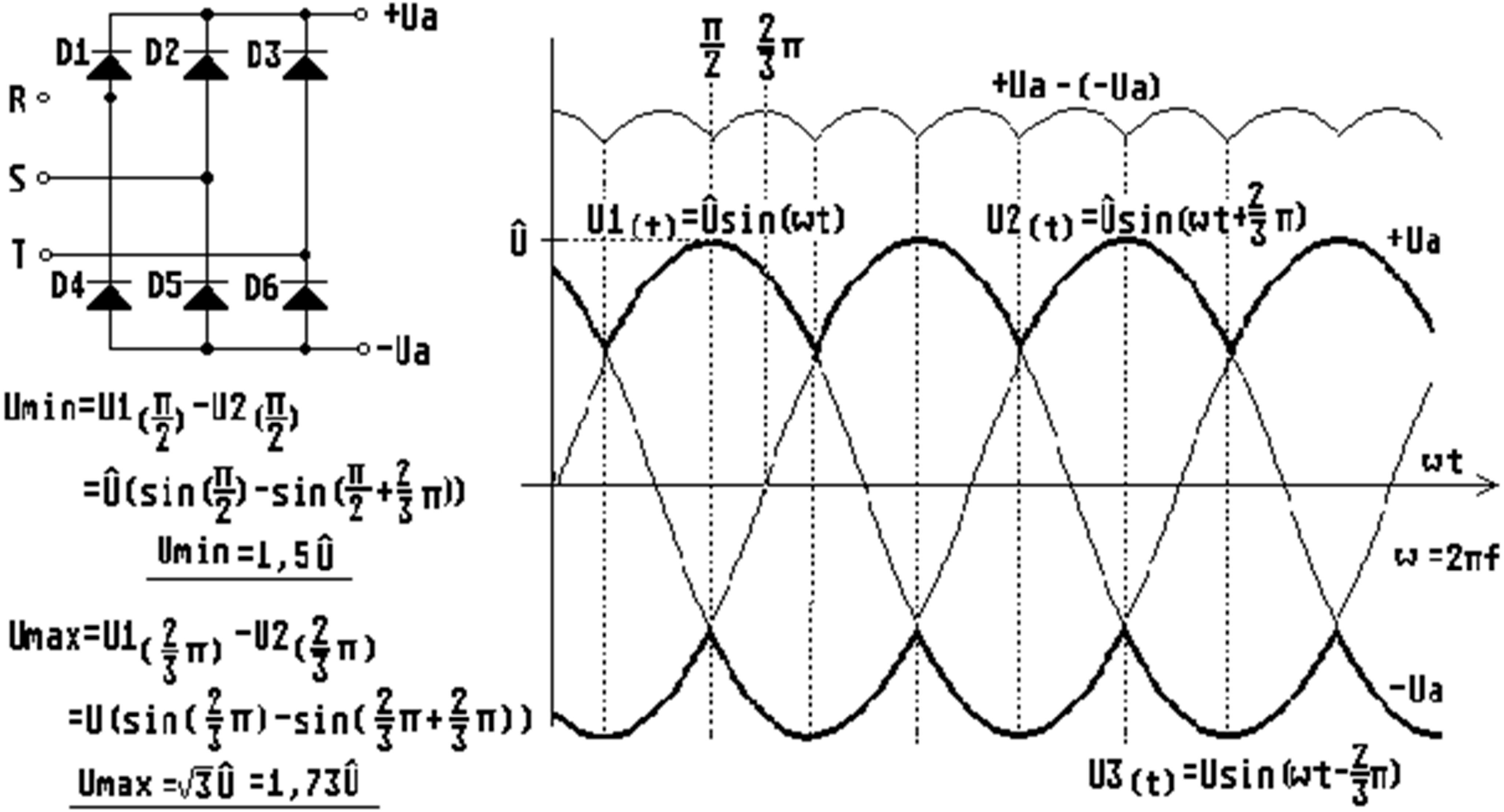
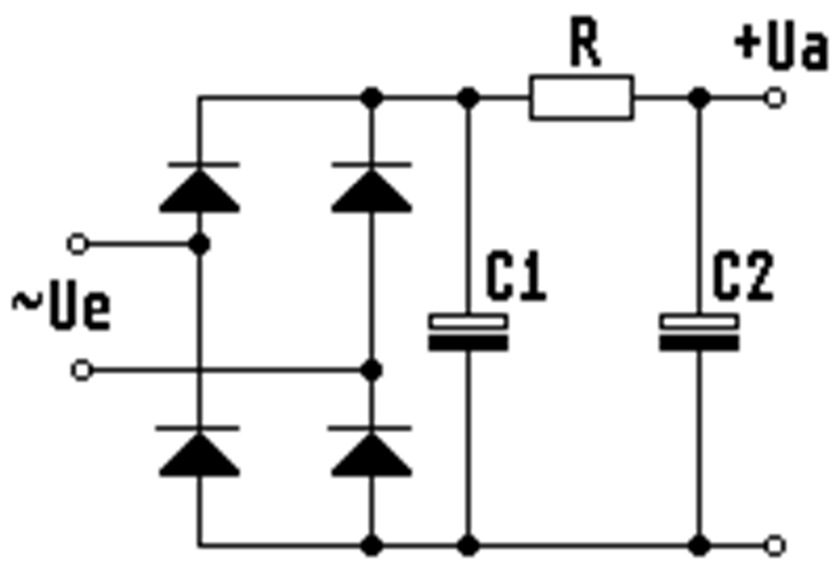


Figure 2.5 circuit and voltage conditions at the three-phase bridge rectifier

In Figure 2.5 you can see the three at 120 ° or 2/3π phase offset phases. The three diodes D1-D3 pick out each of the positive half cycle with the highest instantaneous value and form the positive output voltage Vout +. Similarly, the diodes D4-D6 of the negative half-wave form the negative output voltage Vout-. The voltages Vout + and Vout-are marked in bold in the diagram. The available DC voltage is also plotted above the differential voltage Ua + (-Vo). Particularly interesting in the AC rectifier is that the DC voltage never goes to zero and the corresponding residual ripple frequency six times the mains frequency. In order to calculate the minimum and maximum output voltage of the minimum point is π / 2 and the maximum point 2/3π calculated. At two points, the phases U1 and U2 make the output voltage. Substituting the values in the time function equations, and each of these phases is the difference between U1 and U2, we obtain the minimum and maximum value of the DC output voltage. Thus, the output voltage swings (as measured against the neutral) between 1.5 times and 1.73 times the peak value of the individual phase. At 230 Volts AC or 400 volts AC, a rectifier bridge would provide a voltage 488-563 volts (minus twice the diode forward voltage). Under these circumstances, it is then also possible to omit the Siebelko. If a switching regulator is to be operated, a noise filter must be connected in any case.

2.6 filter circuits for rectifier

When the rectified voltage without electronic voltage regulator to operate a consumer, it may be that the residual ripple still interferes with the function. For 50 Hz applications, this filter circuits have hardly any significance today. The operating voltages of today are either stabilized or electronic circuit is insensitive to ripple voltages as is common for example in audio amplifiers. To reduce the ripple voltage by increasing the filter capacitors is very inefficient, since it would require very large and expensive electrolytic capacitors. Effective are several small capacitors that are connected in an RC filter network. Better, but also more expensive and larger chokes are for filtering out the ripple voltage.



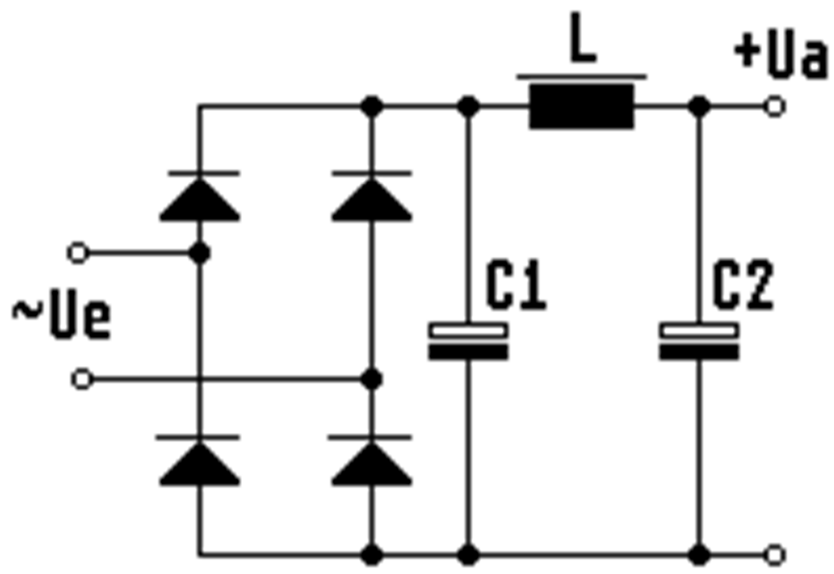


Figure 2.6 A RC filter network

Figure 2.6 B LC ladder filter

The RC filter network in Figure 2.6 A is the simplest and most common. To reduce the 100 Hz ripple voltage significantly, the cutoff frequency $f_t = 1 / (2\pi RC_2)$ should the low pass filter R, C_2 are well below 100 Hz. The resistance R must not be too large, of course, so that the power dissipation $P = RI^2$ is not too large.Much more effective the LC ladder filter is shown in Figure 2.6 B. Again, should the cutoff frequency $f_g = 1 / (2\pi \sqrt{} \sqrt{} LC_2)$ well below 100 Hz. The advantage of the choke consists in that, with the same cut-off frequency, the resistance and the loss is considerably less than the Siebwiderstand. In addition, the permeability of the screening above the cutoff frequency falls about twice as steep as the RC-filter section. The choke must be dimensioned so that the iron core by the operating current can not be caught in the magnetic saturation. Iron cores of chokes therefore always possess an air gap.

2.7 inrush current for rectifier circuits

For simple power supplies with 50 Hz power transformer, the internal resistance of the transformer is sufficient to limit the inrush current to a safe level. Are more critical rectifier circuits that operate directly on the mains voltage. The internal resistance of the 230-volt mains voltage is so low that large filter capacitors at switch currents can flow in the kA range. Although rectifier diodes tolerate very high peak currents, they are completely overloaded with such values. Quite apart from bringing such current spikes massive wear on the switch contacts with them. Furthermore, the fuses would be far too large or very sluggish, so that they do not respond by the inrush current. For these reasons at least all power rectifier must be provided with a current limit. It does not matter which is used, the rectifier circuits described. The most common design of power rectifiers is the bridge rectifier described in Figure 2.2. It has the advantage that it needs only one Hochvoltelko and a bipolar voltage source and still utilizes both half-waves. Since large Hochvoltelkos are relatively expensive, is usually chosen for cost reasons, this variant. As an alternative to the bridge rectifier would anyway only a doubler in question. Then you would have to handle with DC voltages of 600 volts, which could be problematic in semiconductors for small to medium loads. This Verdopplerschaltungen have the disadvantage that the electrolytic capacitors at 50 Hz will only download 50 times per second. The capacitors in Verdopplerschaltungen therefore need to store more total energy than bridge rectifier filter capacitors for the same performance. This would also exacerbate the extra cost and the greater the volume of construction Einschaltproblem.

In rectifier circuits for low power, the problem can be solved relatively easily. This simply inserts a small wire a resistor in series with the rectifier. A wire resistance it should be, because resistance wire has a higher heat capacity than thin metal or coal layers and therefore the inrush withstand better. Common values are around 5 ohms. That's enough to protect the diodes against overload. Much greater the resistance but may also not be, since otherwise expect a significant power loss.Because of the high proportion of the input current harmonic rms current through the resistor is significantly higher than would be expected for sinusoidal currents with the same power. To improve efficiency, it is usual to use an NTC resistor, instead of a normal resistor. After the electrolytic capacitor is charged, the thermistor heats up and becomes low, so that reduces power dissipation to a minimum. This solution is very simple, but has the disadvantage that it no longer works for a short power interruption. If the Elko discharged by the load connected to a power interruption, quickly, and a few seconds later charged, the thermistor is hot and low impedance.This can lead to a large surge that threatens the components. This could be characterized to prevent the load is switched off automatically when the line voltage is interrupted. The Elko remains largely charged until you turn the power supply. In an extended power interruption the thermistor to cool and also limit the charging current of an uncharged capacitors. Suitable thermistor for inrush current limiting are produced in disc form for this purpose and usually have values from 4.7 to 22 ohms.

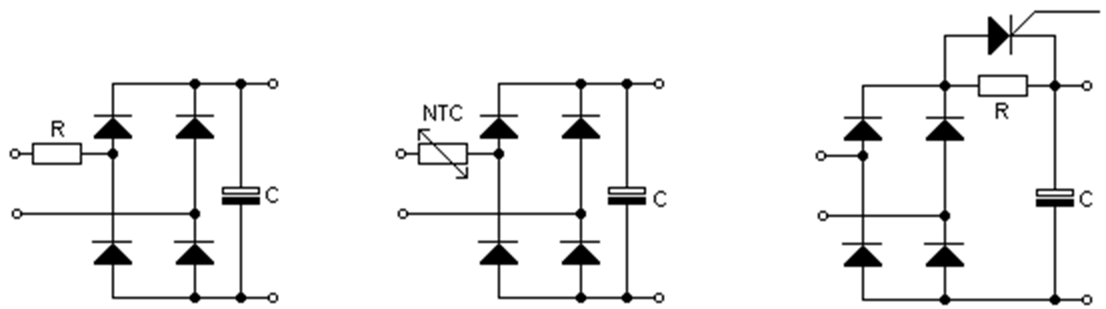


Figure 2.7 Various standard methods for limiting

A somewhat more elaborate method is that it uses a normal wire resistor for current limiting and short-circuits this by charging the capacitors. As a switch, for example, is a thyristor. This has the disadvantage that the activation of the gate at a potential of about + 300 volts may not be easy. Alternatively, the resistor can be inserted into the negative power rail and short-circuited by a triac. A triac may be triggered with a small control voltage positive with respect to the negative power rail. A fully floating bridge of resistance is possible with a relay or opto-triac.

Another electronic limiter, Which is Intended for use in switching power supplies, higher power, can be seen in Figure 11.2 D

3 Linear DC-DC converter

In this chapter I would like to summarize all forms of DC voltage regulator or converter that deliver more or less without the use of AC voltages stable DC voltages or currents. For most standard applications, there are of course already priced ICs that must be connected in the simplest case with one or two external capacitors.For specific applications it is sometimes useful to construct a discrete controller or with the appropriate wiring of a standard ICs. It is also very useful in the repair of older devices to know the basic circuits used. This category is common to all transducers, in that the output current and voltage are always less than or at most equal to the input values. Voltage or current differences are thereby converted into heat.

3.1 The shunt regulator

The shunt regulator is the simplest but also unwirtschaftlichste voltage regulator. The name comes from English and means about shunt regulator. The basic principle is that an existing voltage is charged to the extent that the desired voltage is stable. If the voltage source is connected upstream to low, its internal resistance is increased by an additional resistor. The best known shunt element is probably the zener diode is their second main application of the surge. Zener diodes can be, at least at voltages above 2 volts, and operates in the reverse direction using a controlled breakdown behavior of the barrier layer. Before Halbleiterära neon lamps were used with an operating voltage of about 70-140 volts (depending on the structure and gas filling) to generate relatively stable reference voltages mainly. Found in the higher voltage range and can be found now and then VDR resistors. However, the characteristics of these devices allow only a rough stabilization. Today, they are used only as surge. Since shunt regulator work very uneconomical, they are now used only for the generation of auxiliary or reference voltages with very little stress. The internal resistance of the voltage source to the load voltage Ub is increased with the first resistor Rv. Rv should be such that at maximum output current and minimum input voltage (1 mA) flowing through this still required for the function of Shuntelementes minimum current.

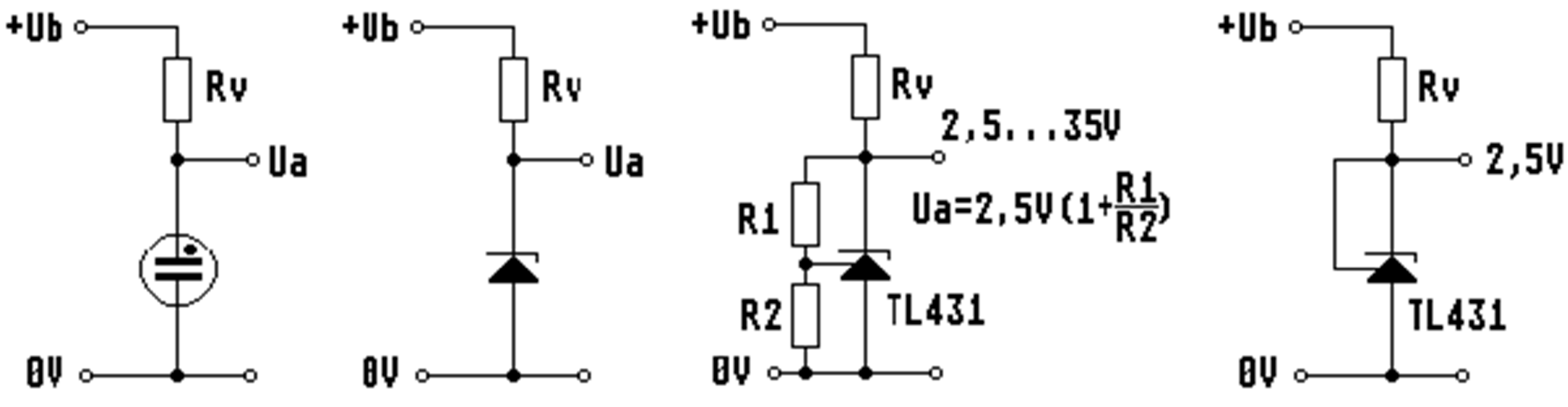


Figure 3.1 A
Neon lamp as
Shunt element

Figure 3.1 B
Zener diode as
Shunt element

Figure 3.1 C
Adjustable shunt
controller with IC TL431

Figure 3.1 D
Special case of
Figure 3.1 c: R1 = 0

In the four sketches the connections of the common Shuntelemente are shown. Previously made to the property of gas discharge lamps advantage to keep at a relatively large discharge current range is a reasonably stable operating voltage. Figure 3.1 A shows such Glimmlampenstabilisierung. Image 3.1B shows the most common form of the shunt regulator with a Zener diode. Zener diodes, such as resistors, with voltage values in the E24 series, 5% Tolerance and various performance classes. The temperature stability of normal Zener diodes is only mediocre and unsuitable for measurement purposes. If the shunt regulator, a reference voltage, for example, produce an A / D converter, specific reference diodes are used. With reference diodes newer design, it is actually ICs with two connection pins. These ICs

have low values as a shunt element in relation to stress tolerance, temperature drift and dynamic internal resistance.
An especially interesting is the IC TL431 shunt element that was brought by many manufacturers under similar names on the market. With this IC, the voltage measuring input and the current sink has been performed separately. Figure 3.1 C shows a simple wiring of the TL431. The voltage measuring input has a temperature-stabilized threshold of 2.5 volts. Using the voltage divider R1/R2, the output voltage is divided down to 2.5 volts. If the output voltage, the output current IC increases so far that the voltage remains stable. By the ratio of R1 and R2 can be set as any output voltage from 2.5 to 35 volts. However, please note that the input current of the IC is about 2 uA temperature dependent and can vary by up to about 1 uA. Depending on the required accuracy should be up to 1000 times greater than the input power of the cross-flow fluctuations in the voltage divider to the 100th It then follows that R2 is an order of magnitude from 2.2 to 22 ohms. In Figure 3.1d, the circuit is simplified somewhat. The TL431 operates as normal reference diode with 2.5 volt supply voltage. A special feature in the diagram of the TL431 is the identification of the connections. Since the component was derived from a zener diode and the circuit symbol and the name of the terminals, cathode and anode, was acquired. As with the Zener diode, the anode negative and the cathode connected to the positive terminal of the TL431 with the power source. The control connection is commonly referred to as "Reference".
The TL431 is so versatile that it is used not only in linear regulators but also in many switching regulators. The most common design is the 92-plastic housing. Other variants are the DIP8 and SO8 (SMD) package. Meanwhile, they are also available in 3-pin SOT23 package.

3.2 The linear regulator

The main difference shunt regulator is that between Spannungein and output a setting element, usually a transistor is, the conductivity of which is adjusted so that the output voltage maintains its desired value. The big advantage is that the source must deliver only as much power as is really needed at the output. When linear regulator must therefore only the voltage drop across the control element can be burned as power loss multiplied by the load current. With a linear regulator can therefore higher power supply voltage also be realized. She especially likes to be used in EMC sensitive areas because they produce no high-frequency interference, and its output voltage is very low noise with good controls. Since there already are a lot cheaper standard ICs for linear regulators, they can be used anywhere where it does not depend primarily on high efficiency.

3.2.1 shunt element than the longitudinal control

In principle, a shunt element is suitable as a series regulator. The shunt member is then placed between the input and output voltage. This is the simplest form of the series regulator, but the back has some disadvantages: The shunt element reduces the input voltage to a fixed value. The output voltage is just as stable or unstable as the input voltage. Moreover, such arrangements without special precautions are generally not short circuit proof. The main application area is mainly limited to the production of smaller auxiliary voltages from an existing stable operating voltage or the reduction of operating voltage in order not to exceed the permissible limits of an IC. In Figure 3.2.1 A simple example is shown in which one or more CMOS logic blocks are powered by a stable 24 V supply voltage. As CMOS devices from the CD40 ... Series may only be operated with up to 15 volts, the 24 V supply voltage is reduced with a 10-V Zener to 10 volts

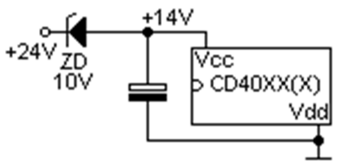
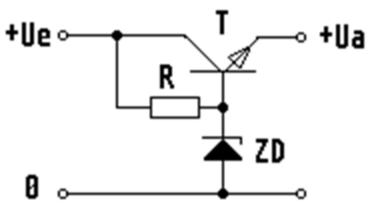
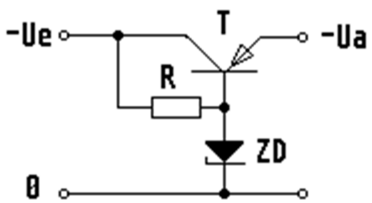


Figure 3.2.1 shunt element (Zener diode) as a linear regulator

3.2.2 The unregulated "linear regulator"
In this form it is basically a combination of shunt and series regulator, where the advantages of the simple structure of the shunt regulator is connected to the better efficiency of the series regulator.





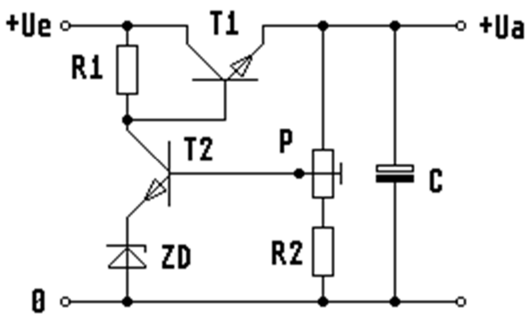
3.2.2 A Picture
Regulator for positive
Output voltages

Picture 3.2.2 B
Negative regulator for
Output voltages

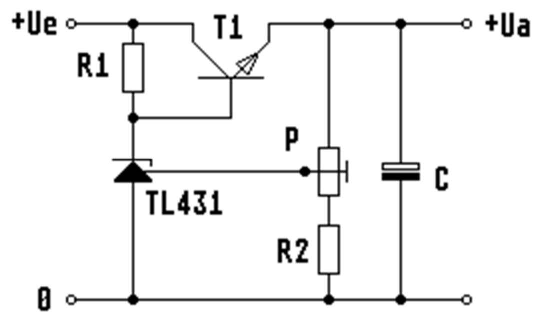
The shunt regulator consists of a resistor R and zener diode ZD. The series transistor T operates as an emitter follower and only serves to increase the capacity of the shunt regulator. Depending on the current gain of the transistor used in the output voltage Vout may be charged to the voltage of several hundred times to ZD, while only a relatively low no-load current flows through R.
Of course this has Einfachstregler again some disadvantages: In addition to the lack of short-circuit current limit and the output voltage stability is mediocre. Even if it succeeds by using high quality reference diodes to keep the voltage at ZD stable, the output voltage is lower than the threshold voltage of the transistor, the Zener voltage. The threshold voltage can also, depending on the temperature and strain, are between about 0.4 and 0.8 volts. The output is therefore an uncertainty range of about 0.4 volts.
This type of controller works equally well for both positive and negative voltages. For negative voltages is simply the NPN replaced by a PNP transistor and zener diode polarity (Figure 3.2.2 B).

3.2.3 The controlled linear regulator

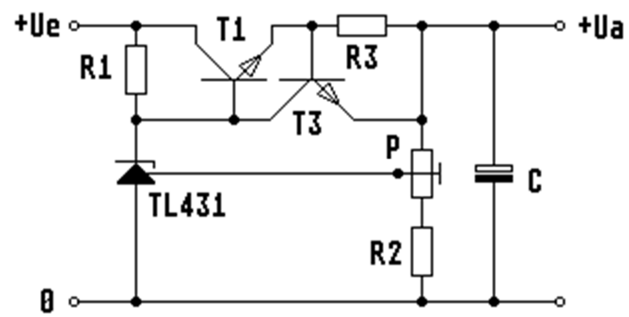
In contrast to unregulated linear regulator, which is not actually a control, the controlled series regulator includes a control circuit that measures the output voltage and holds the more or less precisely constant. From circuit complexity ago it hardly plays a role in whether the output voltage is fixed or variable.



3.2.3 A Picture
Original form of the voltage regulator



Picture 3.2.3 B
High stability control with TL431



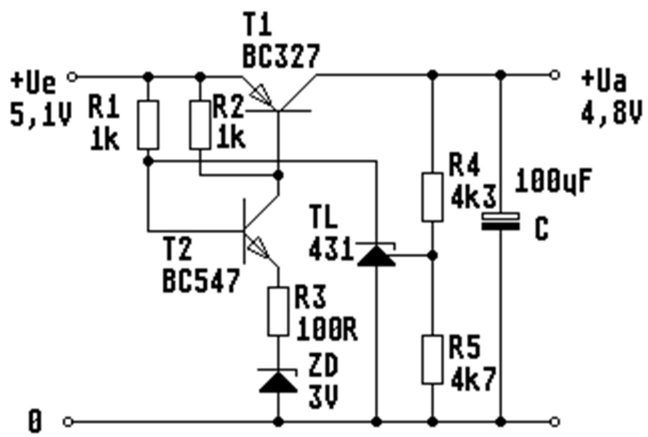
Picture 3.2.3 C
Regulator with current limiter

3.2.3 Image A shows the simplest regulating circuit, which is also found in many older voltage regulators. The resistor R1 is initially switched by the transistor T1 so that the output voltage Vout rises. Via the voltage divider R2 P-Ua is divided and given to the base of T2. With the Zener diode ZD increasing the threshold voltage of T2 artificially, so that the relative influence of the marked temperature dependence of threshold voltage of T2 is much smaller. By choosing a suitable positive temperature coefficient of Zener diodes some temperature compensation is even possible. If the output voltage is so high that the increased threshold voltage is applied to the base of T2, it begins to conduct. Characterized but the base current for over coming R1 T1 is discharged through the collector of T2. This reduces the output voltage to the base of T2 again, the threshold voltage is applied. With the potentiometer P can then set the desired output voltage. The lowest possible output voltage is the threshold voltage. It is then achieved when the base of T2 is directly connected to the output voltage. To avoid high frequency vibrations rule, the output voltage of the electrolytic capacitor C must be blocked to ground. If the circuit supply higher currents, it is necessary at T1, two to three in a row switch transistors in a Darlington circuit. For parallel connection of a plurality of transistors T1, it is necessary to isolate the emitters of the individual transistors each having a low resistance. The Entkoppelungswiderstände ensure a uniform current distribution on all transistors. T2 is replaced by a TL431 and ZD (Figure 3.2.3 B) not only simplifies the circuit, but it also receives a highly stable output voltage. The minimum output voltage now corresponds to the threshold voltage of the TL431, so 2.5 volts.
3.2.3 Image C shows a further improvement to the circuit by a current limitation. The output current flowing through R3, in this causes a voltage drop. If the voltage at 0.6 volts R3, T3 starts to conduct and T1 blocks. By the choice of R3 can be as any current limit can be selected and the controller is short circuit protected. In order not to disturb the voltage control by the voltage drop across R3, has the tap of the potentiometer P on the output voltage definitely behind R3, ie made directly at

the output. In these and the following circuits, there are basically two different ways to connect the potentiometer: If the potentiometer, as in these examples, works as a voltage divider with a variable tap, one obtains a linear potentiometer with a progressive characteristic. Ie, in the lower region, the voltage changes only slightly while achieving a large change in the voltage dependence of the Potidrehung at the top. Is the smallest voltage as 2.5 volts, it has turned up in semi pot just about 5 volts. This can be advantageous if you want to adjust small voltages more accurately than large ones. If you connect the wiper of the potentiometer directly with the connection to R2, the potentiometer works as an adjustable resistor. Since R2 is now abut getting 2.5 volts (for TL431), always flows through the potentiometer R2 and a constant current. The voltage across the potentiometer is proportional to the setting angle. Thus the relationship between setting angle and output voltage is linear.

3.2.4 Low-dropout regulator

Normal voltage regulator series pass transistor in Darlington, who works as an emitter follower, have between input and output a voltage drop of at least 1.5-2 volts.For some applications, it is important that a voltage regulator also still working properly when the input voltage as low as 0.5 volts higher than the desired value of the output voltage. In the following I give an example from my practice. The problem was that in a computer, a graphics chip had to be supplied with a stable 5V operating and reference voltage. The approximately 5.1 volts from the power supply were so dirty that one coming over the reference voltage of the D / A converter, the disturbances as an unpleasant flickering could see on the screen. Because of the power consumption of the chip of about 0.2 A, the generation of the 5 volts from the 12 volt power supply voltage by means of a series regulator additional cooling problems had brought with him. The solution was a series regulator, the down regulated the unclean stable 4.8 volts to 5.1 volts. The differential voltage of 0.3 volts was sufficient in order to correct all errors, and the graphics chip worked flawlessly even with 4.8 volts. In Figure 3.2.4 A diagram of the controller is used to see.

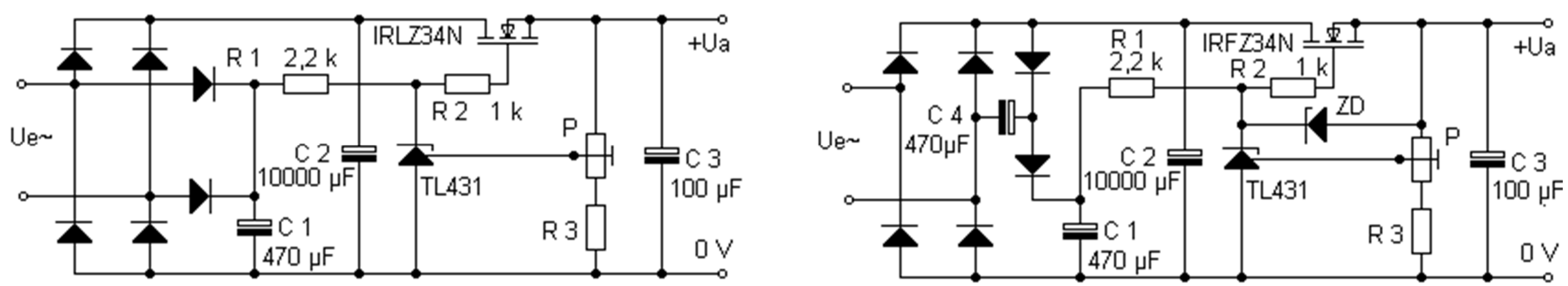


Picture 3.2.4 A stable low-drop regulator

The heart of the controller again is the popular TL431, which measures the output voltage. The voltage divider R4-R5 is such that the TL431 responsive output voltage at about 4.8 volts. The Zener diode ZD effected at the emitter of T2 to T2 begins to conduct only when its base voltage exceeds about 3.5 volts. ZD is required so that the lock can TL431 T2 without the minimum cathode operating voltage of about 2.5 volts is exceeded. The output voltage falls below 4.8 volts, the cathode voltage increases the TL431 and T2 begins to conduct. By the collector current of T2 then T1 begins to conduct, thereby increasing the output voltage again.Since T1 operates in the emitter circuit, the differential voltage between input and output voltage may be below 0.2 volt. However, the control loop is unstable by this circuit. Therefore, a particularly large decoupling capacitor of at least 100uF is required at the output.

A current limitation is not as easy as the normal linear regulator in this type of controller. A current sense resistor would also cause a voltage drop of up to 0.6 volts.The simplest option here is to limit the base current of T1. This has to flow through R3, may drop to the maximum of about 1 volt. Since the current gain of T1 can spread widely, R3 must be determined experimentally, if necessary. Integrated bipolar low-drop regulators are partly the same structure as the in Figure 3.2.4 A. In addition to the larger decoupling capacitor can therefore be also another peculiarity explain the diagram: If the minimum input voltage falls below the controller, the nominal voltage at the output tries to maintain obtained by full turns on the series transistor T1. Depending on the dimensioning of the controller can be used, the base current of T1 will be substantially greater than the normal operating current of the load. The power consumption of the controller to normal again when required for the use of minimum control voltage is exceeded. There is only a weak reliable power source available, it may so happen that the threshold for the minimum voltage can not be exceeded. But this mainly affects older design ICs.

Even with stabilized power supplies with low output voltages, it may be important to keep the voltage drop to a minimum. After all, the voltage drop reduces the efficiency at low voltages considerably. Relatively easy it is to implement such controllers when there is a slightly higher voltage available to drive the pass transistor.In a conventional power supply, the unregulated DC voltage is derived from a transformer winding of a 50-Hz transformer. I once shown in Figure 3.2.4 B are two simple ways to generate an increased supply voltage for the series transistor without an additional winding on the transformer is required.



Picture 3.2.4 B Low-Dropout-Regler/Netzteil with auxiliary voltage generation

Is the expected ripple voltage on the electrolytic capacitor C 2 well above 2 volts, the left variant can be used. The electrolytic capacitor C 1 is only slightly loaded and therefore invites to the positive peak of the input AC voltage. The scheme can only work properly if the minimum voltage at C 2 is still greater than the output voltage. If this is the case, but also the voltage on C 1 is always at least the value of the ripple voltage is higher than the output voltage. When a ripple voltage of about 2 volts is that enough to fully cycle through a logic-level MOSFET. At low load, while reducing the ripple voltage, but that is no problem, because indeed increased the voltage on C 2 and C first The right picture shows an improved variant. C 4 via the alternating voltage component is coupled to the rectifier bridge to a second rectifier input and adds to the voltage at C second At C 1 is then present at approximately twice the voltage as C 2nd This voltage can then be normally and normal control MOSFETs.
Unfortunately, a simple current limiting can be in low-dropout regulators not so easy to realize, because this would still cause a voltage drop of about 0.6 volts at maximum load again. It is sufficient in most cases also from to secure the secondary winding with a fuse, especially if it is not an experimental or laboratory power supply. In the right part of Figure 3.2.4 B I have drawn yet another way to limit the output current without generating an additional voltage drop. The Zener diode ZD limits the gate-source voltage. The saturation behavior of the MOSFET then limits the output current. This type of limitation is, however, imprecise and may even lead to the dropout voltage increases at high load.

4 Phase Angle

The regulators from the last chapter, have the disadvantage that the efficiency, depending on the voltage difference can be very bad. Should be regulated or controlled voltages at high power, the controller should also burn a very high power dissipation. This is not only inefficient but also poses significant problems in the disposal of waste heat with it. When an AC voltage is available at the input of a device, there are simple but still effective method to change the output voltage.

4.1 Power control and dimmer with phase control

In the control of resistive loads such as light bulbs and heaters, the input voltage can break easily at times. The effective voltage to the load is then obtained from the Einschalddauer of the switch and also the input voltage. When the heating is relatively easy because they are very slow and very slow to switch, can be used as relays. When the bulb is slightly more difficult, because these are mostly used for lighting and the metallic human eye is very sensitive to variations in brightness. To ensure a flicker-free light that flicker should not go below 100 Hz. Therefore, the tension inside each half sine wave has to be interrupted in the same way. Such fast switching is only possible with semiconductors. The first semiconductors, which were able to switch high currents and voltages were later thyristors and triacs. A major disadvantage of these components, however, is that they are no longer able to power when he was once turned to interrupt again. When operating on 50 Hz power line that is not a problem, since the current 100 times per second goes back to zero and the thyristor or triac has enough time to recover and block again. So when the switch is once "fired" per half-wave and remains switched until the next zero crossing, the effective voltage at the load can theoretically by the choice of the ignition continuously from 0 to 100% set.

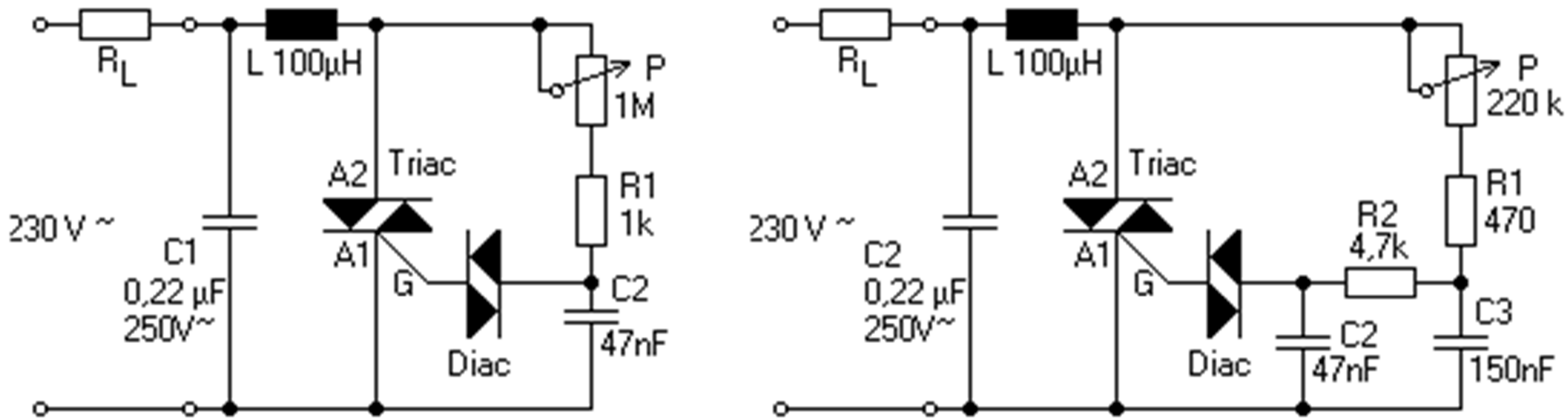


Figure 4.1 A Simple dimmer 4.1 B dimmer (almost) without hysteresis

Figure 4.1 A shows the simplest form of a dimmer with phase control. Here are two special components are used:

1. The triac is a further development of the thyristor. It has the advantage that it can turn by the current in both directions. He behaves like two anti-parallel connected thyristors, but has, except that it is only a single component, also the advantage that you can detonate it in both directions on a common control input (gate A1).
2. The diac is a trigger diode which works much like a neon lamp. Up to a certain threshold voltage of diac is not conductive. In just over 30 volts he suddenly begins to conduct and locks again when the holding voltage was reached, which is about 10 volts at the ignition voltage.

The diac now offers a very easy way to fire a triac. The triac needs it just a short trigger pulse and then remains until the next current zero crossing conductive. As long as the triac is not conducting, is the full mains voltage. The capacitor C2 charges up slowly through the resistors R1 P +. If the ignition voltage of the diac reached, it discharges the capacitor via the gate of the triac to approximately 10 volts and ignites it. After the next zero crossing of the triac is turned off again and the process starts over again, but with reversed polarity. Since triac and diac behave the same in both current directions, this process is symmetric to the zero line. By varying the potentiometer P is the ignition timing within a half wave over a wide range can be set. The limits are set by the firing voltage of the diac, because ignition can take place only at about 30 volts.

Unfortunately, there is a disturbing hysteresis effect in the simple circuit of Figure 4.1 A. If you slowly rotate the potentiometer from the stop in the direction of higher management, the dimmer is suddenly having significant power. Then you can go back in power again. The reason is that the capacitor charges with alternating polarity. Is the pot still very high impedance, the ignition voltage of the diac is never reached. If the pot but so low that it is being achieved, the diac fires and discharges the capacitor C2 to about 10 volts. This C2 will discharge faster while reversing the voltage and reaches the ignition voltage of the diac in the next half cycle much earlier. So the dimmer begins with increased initial performance can be continuously and then turn back again to almost zero. However, once exposed to the periodic ignition, the dimmer must be turned up again to the starting position.
Figure 4.1 B shows how this hysteresis can be largely suppressed. For this purpose the ignition capacitor C2 of the timer via R2-R1-P C3 is disengaged. When the ignition of the triac only C2 is discharged. The much larger C3 largely retains its charge and the ignition timing of the next half-cycle premature not so much.
An advantage of the dimmer is that he is bi-polar and therefore as may replace a light switch without changing the existing installation.
Although can be adjusted to the dimmer in a very simple manner the performance of large consumers there is also a disadvantage. The steep increase in current during the ignition of the triac the pipe network is heavily contaminated with harmonics. The higher frequency components are attenuated by the filter element L and C1. However, the lower frequency components are directly injected into the network. This is especially critical when such platforms shall be exposed and illuminated.The harmonics are caused by dimmer Namely in the audible range and are found frequently because of interference coupling on the audio signals of sound technology again.
An important question that comes up repeatedly in connection with dimmers, is the:
Which devices can be regulated at all with a simple dimmer?
Since I actually fall only two main applications:

1. Dimming of incandescent and
2. Speed control of universal motors

Bulbs behave broadly like an ohmic resistance. The effective lamp voltage is higher, the earlier the ignition point is located. The harmonic load on the line network is all the greater, the higher the lamp current. For large lamp powers should therefore be better to resort to switching regulator. Please avoid circuits, such as one sometimes finds them in power supplies of 80V Projector Lamp for Fotobelichter. There, these lamps are operated with a phase control directly to 230V. So that one is not only a high RMS current load and harmonic contamination of the network but it also risks a life of expensive lamps, which can already be terminated by a single misfire of the triac.
Universal motors can be controlled even better with phase control. The large self-inductance of the coil has a favorable effect on the disturbance response, since the current increase is much slower.
Power transformers may however not be the primary side dimmed with a dimmer so simple. It can, particularly in the idle or non-resistive load, lead to uncontrolled movements of the ignition timing due to the phase shift of the current. As I already wrote in Chapter 1.6, an unfavorable turn-on, especially at the beginning of a half-wave, bringing the core of a power transformer into saturation. Since this is repeated constantly in unfavorable adjustment of the dimmer, the transformer, despite lower secondary load can be easily overloaded. The remedy here is more modern with built-in dimmer protection.

Also resistive loads with high performance should not be regulated with a dimmer because of the harmonic contamination of the network. In general, these are to heaters with relatively large thermal time constant. It is much cheaper to switch the load with a zero voltage switch zero voltage crossing and leave for several seconds at full power before it is switched off again for several seconds. The rate of current rise is thus minimal and the harmonic load on the network correspondingly low.

Induction motors can also not control with a simple dimmer. Since the frequency of the rotating field remains constant, the large slip at low engine speeds may also cause motor overloading. The best speed control or control for induction motors is only possible with a frequency converter.

Also special dimmer required for fluorescent or energy saving lamps. The problem with these lights is that they need to re-ignite in each half wave. But that only works properly at full power. With a simple dimmer periodic ignition and may temporarily suspend the lamp begins to flicker heavily. Dimmer for fluorescent and energy saving lamps must be fitted with a special ignition aid. Firing can be carried out with a high-voltage pulse from an ignition coil or by an external electrode in the tube.

4.2 rectifiers with phase control

Also DC voltage regulator can be realized with a phase control. In the simplest case of the single-diode rectifier is replaced by a thyristor. While a diode automatically switched on in the region of the peak value and the Siebelko charges to approximately the peak value of the AC voltage, the thyristor is triggered only in time behind the voltage maximum, so that the only Siebelko charges to a lower voltage. By varying the ignition timing, the voltage can be regulated. This technique has been used for example in the approximately 70-years in power supplies of color televisions. As I have already mentioned in Section 2.2, the use of half-wave rectifiers is problematic at higher powers. Therefore, half-wave rectifier are now no longer used for rectifying 50-Hz AC voltage with higher performance.

An alternative would be a bridge rectifier, wherein the two diodes are each replaced by a thyristor in the positive branch. The control principle would be the same as the half-wave rectifier, except that both half-waves are used. Such regulators have actually been used to minimize the power loss of the subsequent linear regulator.However, a significant harmonic pollution and RMS current rating of the supply is connected to this control principle again. Now, however, they are technically obsolete by modern switching regulator. Therefore I will not elaborate on that too.

Often it is necessary to produce in a line powered circuits auxiliary voltages to the control electronics to be supplied. At currents below 50 mA, neither the use of a mains transformer nor the worth of a switching regulator. An ohmic resistor would produce too much power loss and a dummy resistor with a capacitor would possibly be too large and / or too expensive. Also in this case, an electronic switch may help the momentarily turns on the rectified mains voltage, only just before and just after the zero crossing, and precisely when the voltage just a few volts above the to be generated low voltage. The resulting voltage drops between the mains voltage and low voltage are then very small and the correspondingly low power dissipation. For currents above 50 mA, this method should not be applied, since it represents as a dimmer, a high RMS current and harmonic load for the pipe network to a greater extent. Ideally, the power consumption of the low voltage should be low relative to the main load on the network.

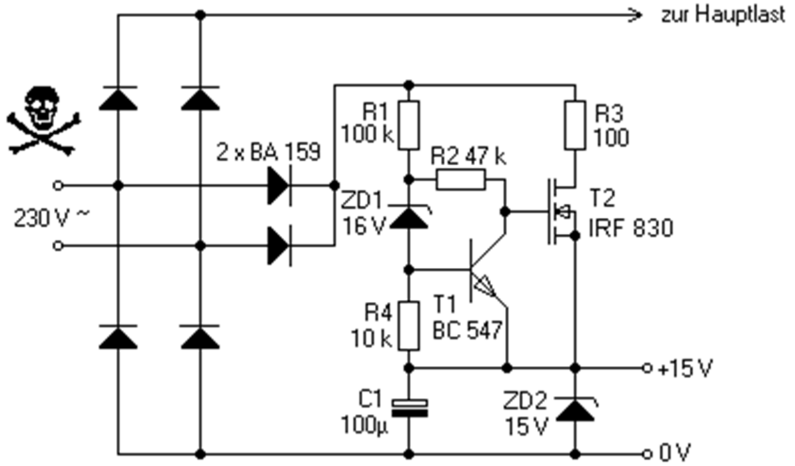


Figure 4.3 generate a 15 V auxiliary voltage

In Figure 4.3 shows a simple example of a powered 230 V mains voltage +15 V auxiliary voltage source. The supply voltage is first rectified by a bridge rectifier, before she gets on skin load, such as a switching power supply. The upper branch of the bridge rectifier is again simulated with two diodes (BA 159), so that the unfiltered voltage on R 1 and R 3 is present. This condition is necessary because it is not ensured that the voltage on the main load is always going back to zero, such as when a connected Siebelko.
Once the mains voltage approximately 5 V higher than the output voltage, T 2 gets through R 1 and R 2 is a sufficiently high gate voltage to turn on. About R 3 then a charging current to the electrolytic capacitor C 1 flows As soon as the voltage difference between power and output voltage exceeds 16 volts, ZD starts to conduct first In a little over 20 volts then T 1 begins to conduct and switches the Gatespanung of T 2 from, so this locks again. The circuit behaves like a constant current source charges the electrolytic capacitor C1 until the voltage by ZD 2 is limited. With the given values of the maximum output current is approximately 20 mA. It is easiest to adjust on R 3rd ZD 1 determines the length of the "Phasenstückchens", which turns on T 2nd If necessary. ZD 2 must be dimensioned a little stronger because it must accommodate the maximum output power in the absence of output load. Possibly also the Elko C 1 must be chosen larger.
Since at R 1 is applied practically in full line voltage, it is advisable either to take a more resilient type or switch two 47-k resistors in series. Otherwise it could happen that R 1 after prolonged use is high impedance.
Rules must be observed in this circuit is that it works correctly only with sinusoidal or similar waveform of the line voltage. For rectangular voltages, such as those supplied by some inverters, a function is not possible.

5 Switched-capacitor voltage converter

In this chapter it comes to voltage converters which generate using a high switching frequency which is usually above 20 kHz, an unregulated output voltage. These converters come from without coils and share input voltages with high efficiency, multiply or invert. Such transducers are found mainly in applications where low cost additional auxiliary voltages to be generated, for example in battery powered devices. This type of converter consists of two major components:

1. A square wave generator with the required output power
2. A rectifier or multiplier

The circuit is simplified considerably if you are on standard ICs for the square-wave generator, such as recourse to the NE555. I want to treat the different power and voltage ranges some standard circuits. At higher powers, especially at higher operating voltages, make sure that the uncharged capacitors can still draw a very high charging current after switching on. The generator is a corresponding current limit must be provided to protect the power transistors from damage. For very small loads, there is a widespread type IC ICL7660. This is a square wave generator and the rectifier included. Thus, small inverter or voltage doubler can be constructed.The special feature of the IC is that all switches are actively managed and therefore no forward voltage is lost to the rectifiers. In the data sheets to find enough examples, so I will not elaborate on here.

5.1 square generators for voltage transformers
Since the rectangle generators, at least for the multiplier and inverter circuits can be considered completely independent of the rectifier circuits, I would first treat the rectangle generators.

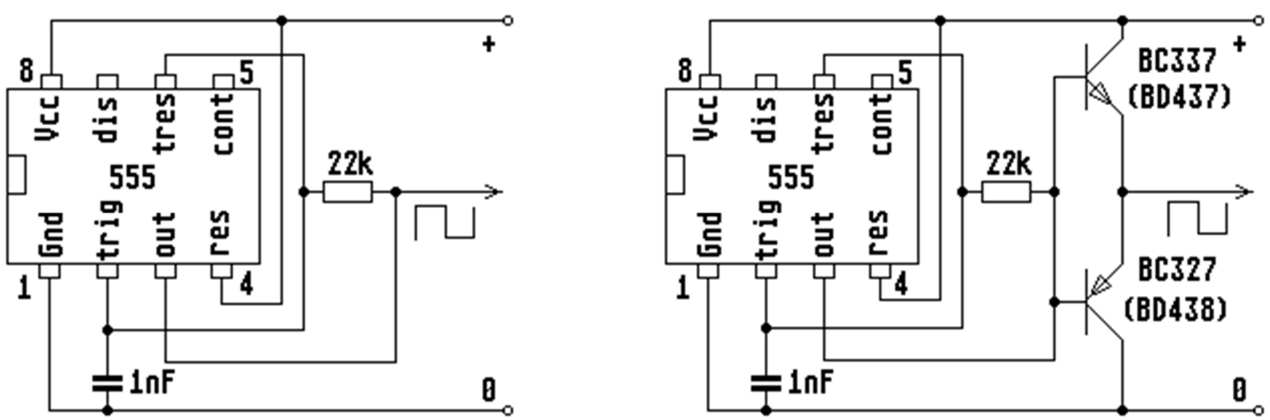


Figure 5.1 A square wave generator for small (left) and medium power (right)

In Figure 5.1 A two particularly simple generators for small and medium powers are shown. For the oscillator itself is needed, except for the IC, only a capacitor and a resistor. The specified values, the oscillator frequency is 20-25kHz. The 555 timer IC is available from various manufacturers as bipolar and CMOS version (the CMOS version can be identified by a "C" in the model name). The CMOS version operates from 2-15 volts and can hardly deliver more than 1 mA output current at 15 volts about 10 mA at 5 volts. The bipolar version operates from 5-15 volts and can deliver 200mA output current. For higher output currents a complementary driver must be followed, whereby the voltage is, however, reduced by 1-1.5 volts. With transistors BC337/BC327 the output current can be increased up to about 600 mA. With the types BD437/438 you reach 2-3 amps. If, for example, during battery operation, a very low idle power consumption is required, it is recommended to use a CMOS version (enlarge 22-kOhm resistor) to reduce the frequency significantly. If even higher performance and / or voltages are switched, the following circuit can be used in image B 5.1. Again, the oscillator 555 is used again. Serve as the final stage two N-channel MOSFETs. By using a P-channel FETs in the upper branch of the output stage Although would simplify the control, but P-channel types generally have a higher on-resistance. When it comes down to switch large currents as small as possible dissipation, the circuit should always be designed so that N-channel MOSFETs can be used. The transistors have shown a turn of about 40 mΩ and can deliver an output current of about 30 amps. For higher currents, and even more transistors may be used. The operating voltage should be at max. 24 volts. When the operating voltage to 15 volts, the voltage may be the 555 directly connected to the operating voltage of the amplifier.

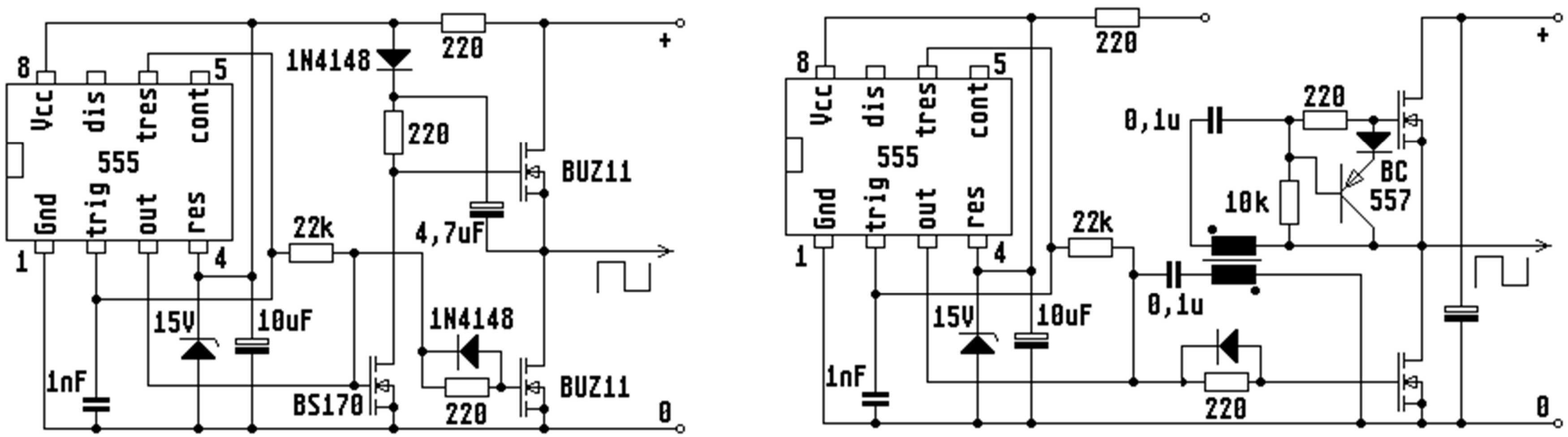


Figure 5.1 B rectangular generators for higher performance and higher operating voltage (right)

The control of the lower branch of the amplifier is particularly simple, since the source terminal of the MOSFET is located at the negative supply voltage. For the upper arm, an additional small-signal MOSFET is required, which is driven synchronously to the lower output transistor. When the transistor is turned on in the lower branch, and the BS170 turns on and sets the gate of upper MOSFET to zero volts. At the same time the 4.7 uF capacitor is charged through a diode to about 15 volts. If the output voltage of 555 zero disables the BS170 and the lower MOSFET. The 4.7 uF capacitor can now remove his voltage across the 220 Ω resistor to the gate of upper MOSFET turns on the full then. The absolute gate voltage of the transistor is then min. 10 volts above the operating voltage of the amplifier. The BS170 therefore must be able to lock this gate voltage increased absolute minimum. The control of the amplifier must ensure that both transistors are not switched simultaneously. This is ensured by the fact that the gates are each charged when turning over a 220Ω resistor and so the output transistors are turned slightly delayed.When you turn off the gates against discharge very low over the BS170 or a diode. If one of the output transistors turn on, so is the other one since min. 100ns locked (depending on the type of transistor). The circuitry of the output stage functions in principle even at higher voltages, such as 300 volts, however, the driving output stage of the upper branch is increasingly problematic with increasing voltage and frequency. For the control, you need a very fast switching, low-capacitance driver transistor. The problem can be largely avoided by using a driver transformer for the upper arm (right). Switching amplifiers for high operating voltages are therefore often driven by driver transformer in the upper branch. The driver transformer need only transmit very little power and can therefore be very small. The ratio should be about 1:1. Because the leakage inductance of the drive transformer may rapid discharge of the gate capacity prevents an additional discharge transistor (BC557) is mounted directly on the circuit breakers. A problem in the high-voltage amplifiers, the power supply of the oscillator. If necessary. an additional small 12-15V power supply must be installed.

A simple square wave generator can also have a completely discreet. Since the circuit is very simple in Figure 5.1 C, it provides an interesting alternative despite the low price of standard ICs (NE 555).

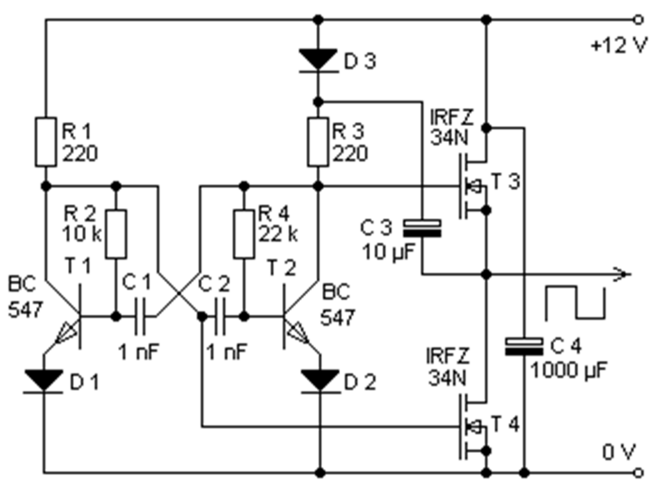


Figure 5.1 C Discrete wave generator

The generator is preferably provided for operating voltages in the range of 12 volts. For higher voltages, the relatively high power consumption of the oscillator has a negative effect. The oscillator is a classic, constructed with two NPN transistors astable multivibrator. The load resistors of the transistors are selected fairly low with

220 ohms. Although this causes a relatively high power consumption, but is necessary to the gate capacitances of the MOSFETs can load fast enough. If the switching frequency is selected to be lower and / or smaller gate capacitance of MOSFETs are used, the resistors can be much higher resistance to reduce power consumption. The emitter diodes D 1 and D 2 BE protect the routes of T 1 and T 2 from excessive reverse voltage. Unfortunately, keeping the BE diode bipolar transistors hardly more than 5 volts in reverse. When tipping the Mutltivibrators but takes the operating voltage is 12 volts in the reverse direction to the BE diode. At the base of T 1 occurs even twice the voltage, since C 1 is located directly on the increased gate control voltage of T 3rd Thus, the duty cycle of the output voltage does not excessively deviate from 50%, therefore, R 4 is much greater than R 2.
This circuit can be built very high impedance, so that it has an extremely low power consumption and then is ideally suited for battery operation. The circuit is located in the department circuitry / oscillators .

Who, despite the advantages of N-channel MOSFETs for easier wiring with P-channel decides to rely on the following circuit:

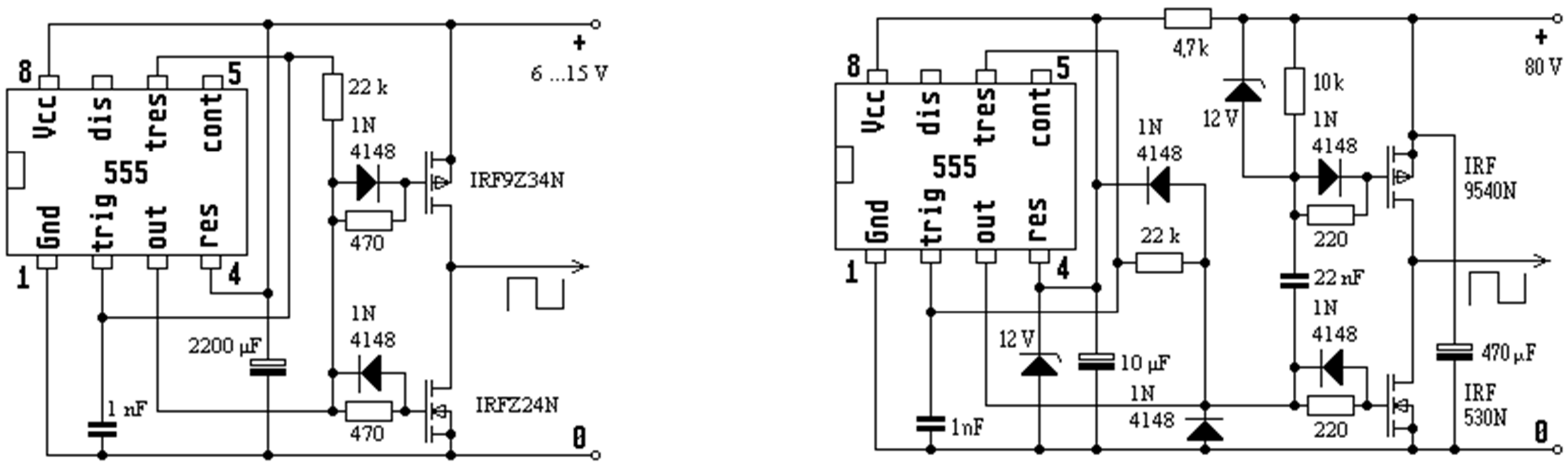


Figure 5.1 D rectangle generators for higher performance with P-channel MOSFETs

For operating voltages up to 15 volts, the simple circuit can be used on the left in Figure 5.1 D. The P-channel MOSFET can be controlled more directly by this timer 555. The gate diodes ensure again that the transistors are rapidly frozen and turned slowly. This ensures that during the switchover temporarily held locked both transistors are turned on.
For operating voltages above 15 volts, 80 volts, for example, the circuit is designed right in Figure 5.1 D. To drive the P-channel FET, the DC voltage at the source has to be decoupled from the control voltage of timer ICs. The easiest way is with a 22 nF capacitor. With a 10 k resistor and a 12-V Zener diode, the control voltage at the gate of P-channel MOSFET is forced back to a defined level with respect to source. Please note that turn on and off the operating voltage must not be too fast to

allow the charge and discharge of the 22-nF capacitor not destroy the FET gates or the timer output. The protective diodes on the timer output and the Zener diode to the gate of P-channel MOSFETs from these currents start largely. To ensure a clean driving the P-channel FETs is also important to ensure that at the source terminal, which is located on the positive supply voltage, no significant noise voltage may occur. This must be connected in parallel to a sufficiently large Stützelko operating voltage of the power amplifier.

With a series resistor and a Zener diode, the operating voltage of the timer is stabilized at 12 volts.

With the current state of the art low-cost P-channel MOSFETs can be had for blocking voltages up to 200 volts (eg IRF 9640). For operating voltages up to 180 volts provide P-channel MOSFETs in the positive branch of the power stage, therefore an economical alternative to the heavy controllable N-channel types dar. it will also be useful to use the niederohmigeren N-channel types with increasing output power.

A further alternative for the control of N-channel MOSFETs in the positive branch of the power amplifier, the use of gate driver ICs. So that very simple generators with high performance can be constructed. Since the main application area of the gate control ICs but is likely to be more in the range of SMPS Chapter 6, I will discuss it only there. By the gate driver ICs, the transformer driving the transistor is likely to be technically obsolete in the upper branch. On the other hand, I assume that there are to buy such Ansteuertrafos for MOSFETs and IGBTs in the future as an inexpensive standard components. Therefore, I will show in the following chapters both methods. The useful Ansteuertrafos are definitely areas where a complete electrical isolation of control and power supply is required.

5.2 Switched capacitor voltage multiplier

The simplest form of the multiplier is the doubler, I've dealt with in a similar form in the 50-Hz rectifier circuits. In Figure 5.2, a multiplier circuit for low operating voltages can be seen. I just picked this square-wave generator, because in this particular case, the simplified circuit. The electrolytic capacitor C2, which is intended for the multiplier circuit, while the gate voltage is used to generate the upper-stage transistor. The corresponding diode and the electrolytic capacitor in the generator circuit may be omitted. For higher voltages, the circuits described in Section 5.1 shall be used. In order to keep the loss in the low voltage diodes should be used at less than 40 volts, and operating voltages for higher power Schottky barrier diodes D1-D6. Standard types are eg 1N5817 (1A, 20V), 1N5819 (1A, 40V), 1N5822 (3A, 40V) or MBR1645 (16A, 45V).

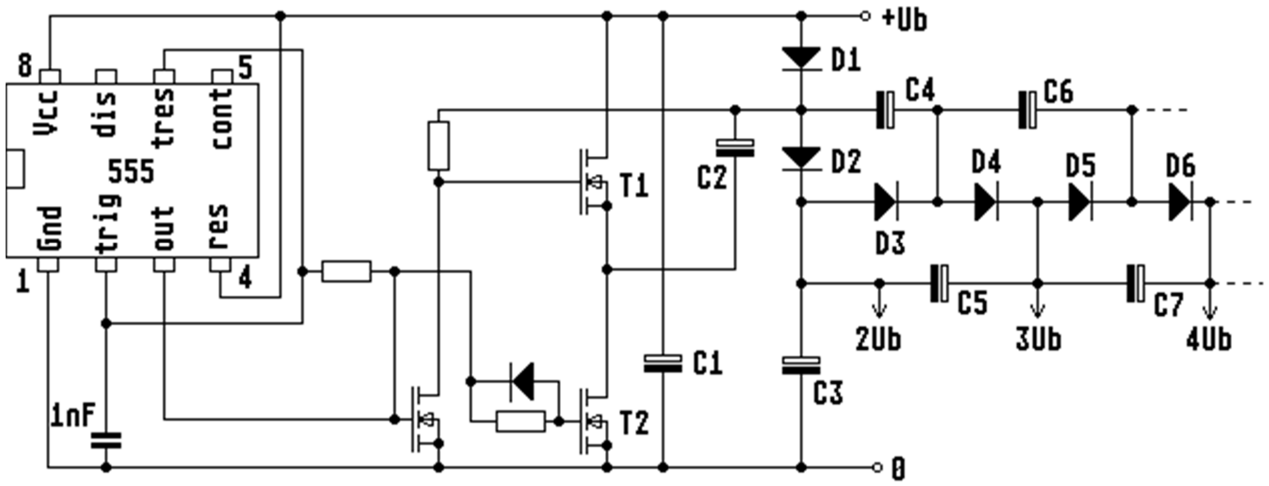


Figure 5.2 multiplier for low operating voltages of 8-15 volts

The selection of the transistors T 1 and T 2 depends also on the desired performance. Cheap standard types such as BUZ71A (0.1 Ω, 50 V) can switch currents up to about 10 A. If the transistor losses are kept small, the relatively inexpensive high capacity types such as IRFZ44N (24 mΩ, 55 V) can be used. With adequate cooling, these types can tolerate about 40 amps continuous current. If even that is not enough, several transistors should be connected in parallel. This is usually cheaper than using extremely resilient individual transistors. Due to the high switching frequency, the capacity is only of secondary importance in the design of electrolytic capacitors; important is the durability and the internal resistance. Good results can be achieved with low-ESR capacitors or when capacity and / or dielectric strength are abundant oversized.

For a doubler only the diodes D 1 and D 2 and the capacitors C 1-C 3 are required. For each additional stage which increases the output voltage Ub, two capacitors and two diodes are required. Theoretically, any number of steps can be followed. For large ratios, this method is then but possibly more expensive and larger than a switching regulator transformer.

5.3 A switched capacitor voltage inverter

Voltage inverters are circuits which generate, or vice versa from a negative to a positive operating voltage. Common application is for example the creation of a balanced operating voltage for analog circuits with operational amplifiers in battery powered devices. First, you look back from a generator with sufficient power.Depending on the required output voltage may then be followed by a single or a vervielfachender inverter.

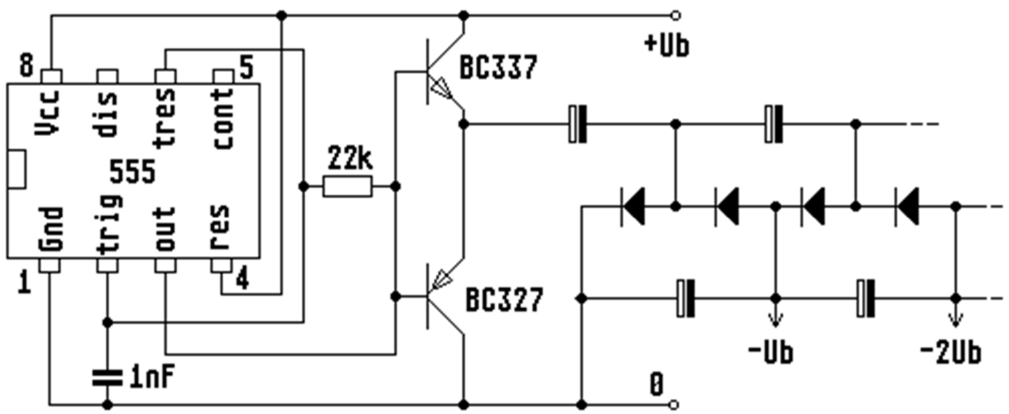


Figure 5.3 voltage inverter for low power

As seen in Figure 5.3, the circuit of the inverter of the multiplier is very similar. Just like the multiplier circuits and the normal inverter circuits with an arbitrary number of stages, and power can be built up.

5.4 Switched capacitor voltage divider

Less well known than the multiplier circuits are switched-capacitor divider circuits. One possible application would be for example the supply of 12 V consumers in a 24-V power supply. Of course, the effort is worth it just for higher power, where a linear regulator would burn too much power loss. As a useful division ratio actually just halving the input voltage comes into question.

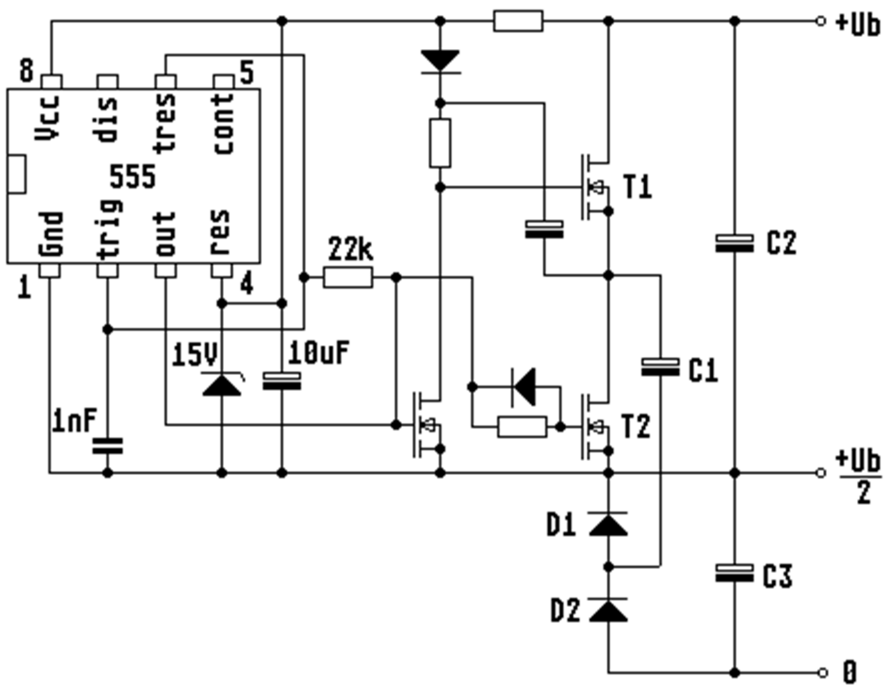


Figure 5.4 The Spannungshalbierer switched-capacitor

In Figure 5.4, such a voltage converter is shown. In principle, there is a voltage inverter, wherein the input voltage is between the positive supply voltage and the negative output voltage is applied. The output voltage is then removed from the negative output voltage and the initial mass. that adjusts the operating voltage to half of the original ground line can be explained as follows: Due to the stress of the output voltage of the electrolytic capacitor C 3 is discharged. At the same time the operating voltage of the generator circuit, whose value is the same as the voltage swing at the generator output increases. On C 1, D 1 and D 2, C 3 is charged to the value in the voltage provided that the voltage of C 3 is smaller than the rated voltage. If the voltage at C 3 is higher than the voltage, D 1 and D 2 are non-conductive and no current flows. Only when the voltage U C is decreased to $_b$ 3/2, the voltage of the generator is equal to the voltage at 3 C If the voltage is less than U of C 3 $_b$/ 2, the diodes start to conduct D 1 and D second Since there is no current limit, the output voltage drops below the operating voltage leads at the half to a massive increase in the input and output current. The output voltage is thus loaded with a high current. The output current is composed of the operating current of the generator and the coming via C 1, D 1 and D 2 identical output power of the generator. The output current is therefore twice as great as the input current, which consists of only the operating current of the generator. The series resistor for the oscillator must be adjusted to half the operating voltage may still be. For the dimensioning generally apply the same standards as for the multiplier. However, the input current is relevant for the design of components.

6 Voltage transformer with chokes

This very important and frequently used type of transducer used a so-called choke dc voltages to implement. The term choke is explained by the fact that a coil, like a capacitor that can store energy in the form of a current (W = ½ LI^{2}). While the capacitor stores energy in the form of a voltage (W = ½ CU^{2}) and can be charged or discharged by a current, the coil discharge by the applied voltage or ge. Formally, therefore behave in the same capacitor and coil, only voltage and current are both reversed. Compared to the switched capacitor converters with those who have the advantage that any division or multiplication factors can be achieved with them, the output voltage can be regulated and only one reactor is required with storage chokes. The efficiency of such converter is theoretically 100% and reaches practically almost always above 80%.

6.1 Buck converter with chokes

Also buck or English Step-Down Converters, buck converter called. Down converter should probably be the most commonly used transducers with choke. You can replace the lossy linear voltage converter without the rest of the circuit would have to be changed. As with the linear transducers, the output voltage is always smaller than the input voltage. For this, however, the output current is greater than the input current in the normal case, a logical consequence of the energy balance at a high efficiency. For the buck converter is needed first again a square wave generator with a sufficiently high output power. Basically, the storage inductor and the subsequent Siebelko form no more than a LC low-pass, which filters out the DC component of the square wave voltage. The ratio of output voltage to input voltage is then identical to the duty cycle of the square wave voltage.

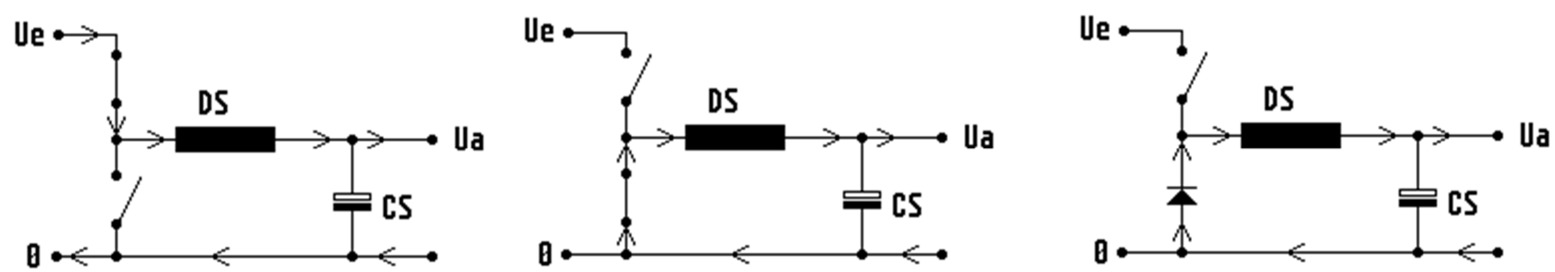


Figure 6.1 A The charging and discharging of the storage inductor with diode switches

In Figure 6.1 A, the principal function of the buck converter is shown. The output of the square-wave generator is composed of two electronic switches, the positive and the negative pole of the input voltage switch push-pull on the storage inductor. With a good dimensioning of the throttle, the inductor current is never zero, so always flows in the direction drawn. If the upper switch is closed, the inductor current increases as a function of U-Ua "charging voltage". If the lower switch is closed, the output voltage Vout with reversed polarity at the throttle, causing the inductor current is reduced again. The reactor is then as it were discharged-Ua the "discharge", while the current continues to flow in the same direction. By removing the lower switch, the coil current would cause a high negative voltage at the upper opening of the switch. This voltage can be short with a diode to the negative pole of the operating voltage. The lower switch can therefore also be replaced by a diode. To reduce switching losses, it will always try to use a Schottky diode. This is especially important when operating modes occur where the top switch turns on when a relatively large current still flows through the diode. The generator then needs only the upper branch of the output stage to drive. The use of an active switch for the lower branch is only useful if at low output voltages, particularly high efficiency is to be achieved, albeit at low output load no dead times to be created with undefined values for inductor current and voltage (zBbei converters without control) or if the converter to operate bidirectionally, ie, if at times also power to be transmitted from output to input.

When sizing a buck converter, a first switching frequency must be selected. The higher the frequency, the smaller inductor and electrolytic capacitor can fail. At high frequencies, however, cause problems with interference and the switching times of the components.At low frequencies can cause acoustic pollution. Are practical switching frequencies between 25 and 250 kHz. For low voltage (below 50 volts) and low to medium power, there is a wide range of toroidal chokes, Which are Ideally suited for the buck converter. These reactors have a special dust core, Which has a particularly high saturation field strength Because of its Relatively low permeability and requires no air gap. Commercially available storage chokes in the current capacity and the inductance is specified in the catalog. Here you need as a user not charge as much. Unfortunately, powder cores cause induction at high Voltages and high switching frequencies much higher losses than ferrite cores.THEREFORE chokes with ferrite core and air gap are Often used in the Appropriate applications. ferrites Since, like soft iron, has a very high permeability, the calculation of Ferritkerndrosseln is as easy as with the 50-Hz chokes. The inductance is calculated is with ≈ L in A / L ($\mu_0 = 4\pi \times 10^{-7}$ Vs / Am, N number of turns, A , l cross-sectional area and length of the air gap in m^2 and m), where again the restriction did with larger air gap lengths of the actual figure is much higher. Ferrite cores with built-in air gap are so frequently with at A_L -marked value. Represents This value inductance of a winding on said core. The inductance of a coil on this core then has the value $L = A_L N^2$. The A_L value has the advantage did it takes into account all the parameters of the core and, Therefore, allowed as Opposed to Purely theoretical calculation via the air gap, a fairly accurate calculation of the inductance. The maximum current of the storage inductor is then calculated is as well as with the 50-Hz restrictor with $I_{max} \approx Bl / N\mu$ .It Should Be Noted did the saturation field strength of the ferrite Wherein B is only about 0.4 Tesla. If in doubt, the manufacturer's datasheet is accurate information. Again, allow the A_L -value, a more accurate calculation of the maximum current $I_{max} = \Phi_{max} / NA_L$ .The maximum magnetic flux Φ_{max} is derived from the cross-sectional area of the core and $B_{max} \forall 0.4$ T. With a Induktivitätsmessgerät can the A_L -value of a deterministic core Easily mined. Shall be givenName by 10 turns of insulated wire around the core, and measures the inductance. The meter then displays exactly 100 times the A_L value in the core. The A_L value is, HOWEVER, only a nuclear-specific constant, if no additional air gap is inserted. An additional air gap Reduces the inserted A_L value.

At a minimum sizing of the reactor power during the discharge of ballast goes almost to zero before rising at the end of the charging cycle to Approximately twice the output current. The reactor core may THEREFORE double the output current does not yet reach saturation.
minimum inductance of the coil depends on the switching frequency from f For the calculation starts from a worst-case extremes did the input voltage is very high and that, accordingly, the duty of the upper branch of the switching stage short Compared to the discharge cycle of the storage inductor is negligible, Which is then Approximately equal to the period duration T = 1 / f Ideally, the throttle Should Be investigation did at the expected minimum output load current of the coil not quite back to zero during a period. Since the voltage across the inductor during the discharge cycle is almost constant, the current Decreases linearly, and the discharge can

be calculated is easily. Just as the discharge of a T with U_o charged capacitor at constant discharge current IT = U_o is C / I, the discharge of the coil may be Analogous to the formula T = I_o L / U are calculated is. Where U is the output voltage U_A and I_0 the maximum current of the choke, or about twice the output current I_a . If you have chosen A Certain throttle for the converter to be built, the minimum switching frequency f Accor ding to the formula f = 1 / T = U_a / 2I_A Lare calculated is. I. $_A$ is the smallest output current in normal operation. If the switching frequency is deterministic mined, the inductance L = U is $_a$ / 2 l $_a$ f are calculated is. In practice, one must assume did the inductance of the inductor Decreases Significantly due to saturation effects at higher currents. Since the inductance at higher currents may be smaller but did is no problem. Is the possibleness range of the output current very large, it can preventDefault the inductor current at low load breaks off before the end of the discharge cycle is hard. The result is then a slightly subdued high-frequency vibration, Which forms between the demolition of the current point and The Intended end of the discharge cycle (dead time). The resonant frequency is givenName by the parallel connection of parasitic capacitances of the inductor and switching stage reactor. The emergence of a dead time has the Disadvantage did the output voltage is Strongly DEPENDING on the load in to unregulated control of the switching stage and did may have interference suppression circuit is somewhat more complicated Because of the high frequency oscillation.The formation of the dead time can be Either constructive avoid to active switch in the lower branch of the switching circuit, or by using a non-linear inductor.Nonlinear chokes can for example build Characterized did in the air gap length is less than the full cross-sectional area of the core is the same. With small currents , the field lines can then pass through the region of the gap, Which is very short. Relatively quiet The inductance is large. For larger flows, then get the areas of the core, Which bridge the gap Partially to saturation. The field lines then have to resort to larger areas of the gap lengths, thereby Substantially reducing the inductance. Of course there are a number of integrated components, Which require some non-electrolytic capacitors and inductor, no external components are therefore Commonly used for this type of converter.

Relatively common and inexpensive should now be switching the controller from the Simple Switcher series from NSC. These are available in different performance classes LM2574 (0.5 A), LM2575 (1A) or LM2576 (3A). More are sure to follow. These types are then identified by the endings of each type designations, even with adjustable output voltage or with different fixed voltages. Besides the throttle and the electrolytic capacitors need these ICs have an external Schottky diode. A disadvantage of these ICs is that the oscillator is not externally accessible and, therefore, the switching frequency is not adjusted yet synchronized. It is set internally to 52 kHz. In most cases however, this should not be a problem.

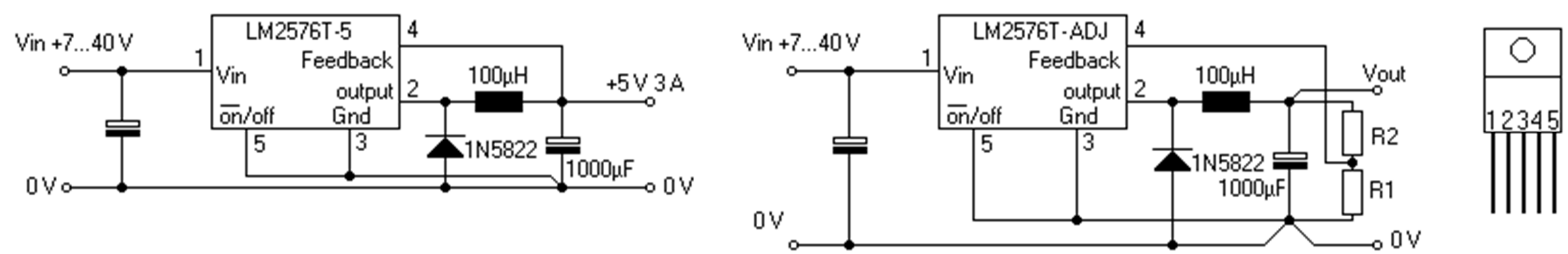


Figure 6.1 B Simple step-down controller with IC and less external circuitry

In Figure 6.1 B on the left side, the simplest version of an integrated switching regulator can be seen. In addition to the non-integratable components (coil and capacitors), only a Schottky diode is required. Specified in the diagram type 1N 5822 has a maximum reverse voltage of 40 volts. Since the reverse voltage of the diode must be at least as high as the input voltage, one should use a higher-barrier diode at more than 35 volts at the input. The voltage regulator has a measuring sensor input, which regulates the output voltage is fixed at +5 volts. There are also other types, identified by the last number in the name, the most important standard voltages (3.3, 12 and 15 volts) have fixed.

For a clean system, it is important silently did the wires going from the output capacitor to the IC, are normally as possibleness. Ie, the lines, where larger streams, Especially alternating currents flow must be Performed separately for Elko. In the diagram this is Indicated by a CORRESPONDING conductor guide. So, the wires for the output voltage Should Be picked separately at Elko, since the ripple is here at least.

More over, Should the lines In Which alternating currents to flow, Which are the interconnections to the IC pins 1 and 2 of the Eingangselkos and the Schottky diode, be as short as possibleness. Especially This is important at high output currents. Therefore I go from page 6 (Figure 6.1 E) elaborated upon.

In the event did the Desired output voltage to be regulated does not match the available default values or, there are versions of synthesis adjustable regulator ICs. The function of the variable is identical with the version of the fixed set. The reference voltage for the voltage sensor input, HOWEVER, is chosen to be very low with 1:23 volts. Diese Allows the output voltage to settle down to 1:23 volts. B in Figure 6.1 on the right side as you can see on adjustable buck. The voltage divider R 1, R 2 divides the output voltage down to 1:23 volts. HENCE the output voltage is calculated is as Vout = 1.23V (1 + R 2 / R 1). For R 1 is a value between 1 and k 5 is recommended.

HOWEVER, thesis ICs are not so long in the market and it is currently no standardization to detect. THEREFORE, I would like to employ in addition to synthesis ICs with solutions based on standard types. There are always cases where it is not reasonable to rely on ready-made solutions. The simplest converter with inexpensive standard components are the self-oscillating buck converter with abused as a switching linear regulators fixed voltage regulators.

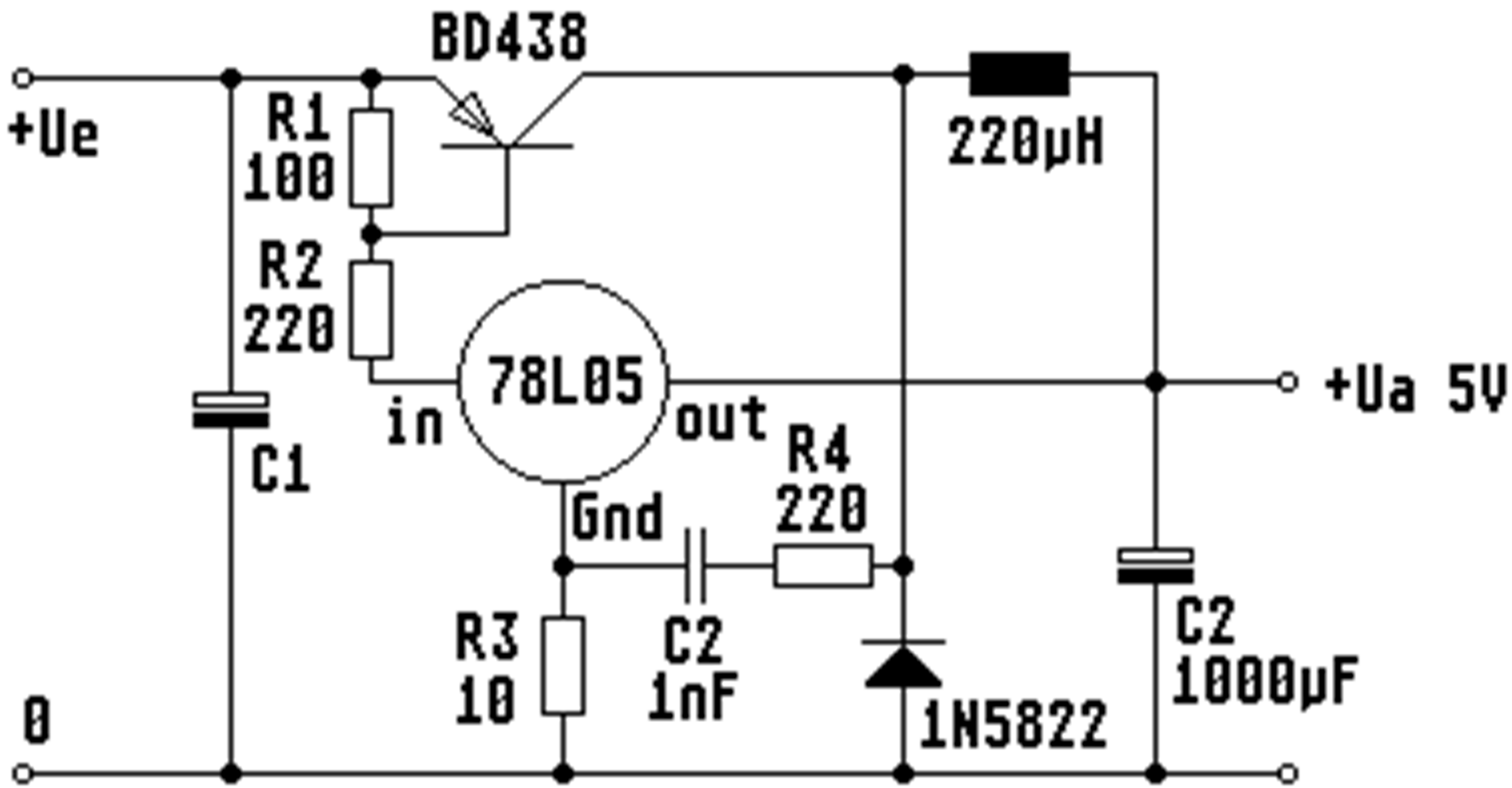


Figure 6.1 C Easy self-oscillating buck

C in Figure 6.1 is an example of seeking a controller with 5 V output voltage can be seen. At the heart of the circuit is a small, 100-mA linear voltage regulator with output voltage of 5 volts. While the output pin is as usual connected directly to the output voltage, the input current first passes through the base-emitter path of a PNP power transistor. The output voltage is somewhat Greater than 5 volts, the controller turns off and no current flows into the base of the transistor. Minor residual currents are derived from the base resistor R1 to the base over. If the output voltage but slightly less than 5 volts, the controller to increase enlarge the output current Attempts to reach the 5-volt level again. Resulting The input current flows through R2 and the base of the transistor. The controller itself is not to increase enlarge the output voltage capable. HOWEVER, by now connected transistor deterministic mines the input voltage to the throttle, Increasing the output voltage again. Eventually, the output voltage Exceeds 5 V, and the transistor is switched off, and the process begins again. In order to Improve the switching performance, is inserted through R 4, and C 2 is a positive feedback voltage to the ground pin of the controller IC. Approximately The maximum output current is 2 amps . circuits So simple but so have some disadvantages: Due to the lack of current limit the input voltage must be protected otherwise. The base resistor R 2 must be angepasst to the input voltage, and the transistor may be. R2 present at slightly less than the difference of input and output voltage. The current must be Sufficient to turn on the transistor safe even at maximum output current, but Should it not so Significantly higher than control. remains to note So did the linear regulators are not designed for this operating fashion. relevant for the proper switching operating characteristics of the ICs are guaranteed by any manufacturer. If necessary., The values of the components to be adjusted. For professional applications, I would advise against this controller version.

Another interesting control IC from ON Semiconductor is the MC 34063A. For small output currents up to 500 mA and input Voltages up to 30 volts, the controller block, as shown in Figure 6.1 D, are used without driver stage. The IC Operates with input Voltages above 5 volts. The output voltage is deterministic mined by the voltage divider R 2, R 3rd The output voltage Adjusts itself so did the voltage at pin 2 of the IC is 1:25 volts. The result is again the well-known formula for the input voltage Vout = 1.25V (1 + R 2 / R 1).
Excluding driver stage can be output at current of about 500 mA to achieve achievement. The current limitation is Effected by the resistor Rsc. 34063 The MC switches off the output stage, as soon as the voltage difference between pin 6 and pin 7 Exceeds about 300 mV. 0:33 Add Rsc = Ω are the Approximately 1 A. The actual achievable output power but is always lower. For optimal dimensioning of the throttle when Power before switching the output transistor is not just going back to zero, is about 500 mA. If you oversize the throttle generous, almost 1 A can be Achieved. Is Too small for the throttle, and the switching frequency is too low, and the maximum output current is correspondingly smaller . did The reason is the inductor current to increase enlarge proceed rapidly after switching and THEREFORE only short activation times of the switching transistor are possibleness.

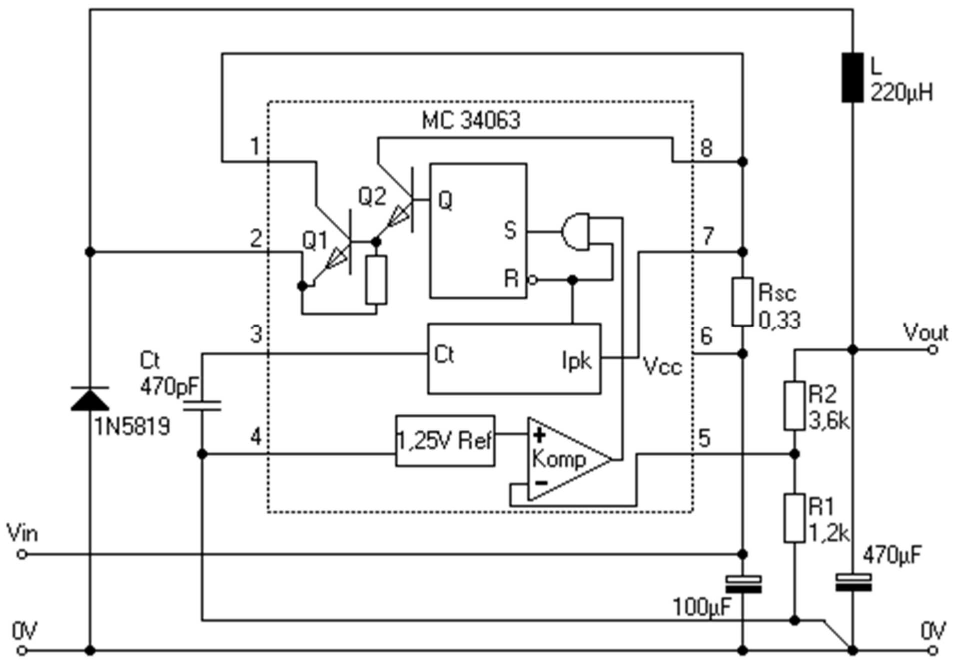


Figure 6.1 D single step-down controller for small power output

If higher output currents and power can be Achieved, Achieved with this can be to external switching transistor. Theoretically Could you size the circuit for any high output currents. HOWEVER, I would advise against it, as the MC 34063 does not allow real PWM modulation . For this, a variable gain amplifier having a downstream modulator PWM is required. This feature is found in the standard control ICs only in TL 494, SG 3524 and SG 3525th At higher power supplies are always THEREFORE Preferable to synthesis ICs, so I shall elaborate on below.

A true PWM controller generating rates at its output a square wave signal with a defined frequency, the duty cycle is always adjusted by the regulator, the output voltage Maintains its setpoint.

When MC 34063 that works not so good. acts directly to the control input of the output switch. This can lead to uncontrolled oscillations (control), Which are so felt as an unpleasant whistling and / or noise, Especially in the storage inductor. At low power Which is not so bad. For higher ratings, the associated decrease in efficiency leads to excessive heating of the components and to Increased noise radiation.

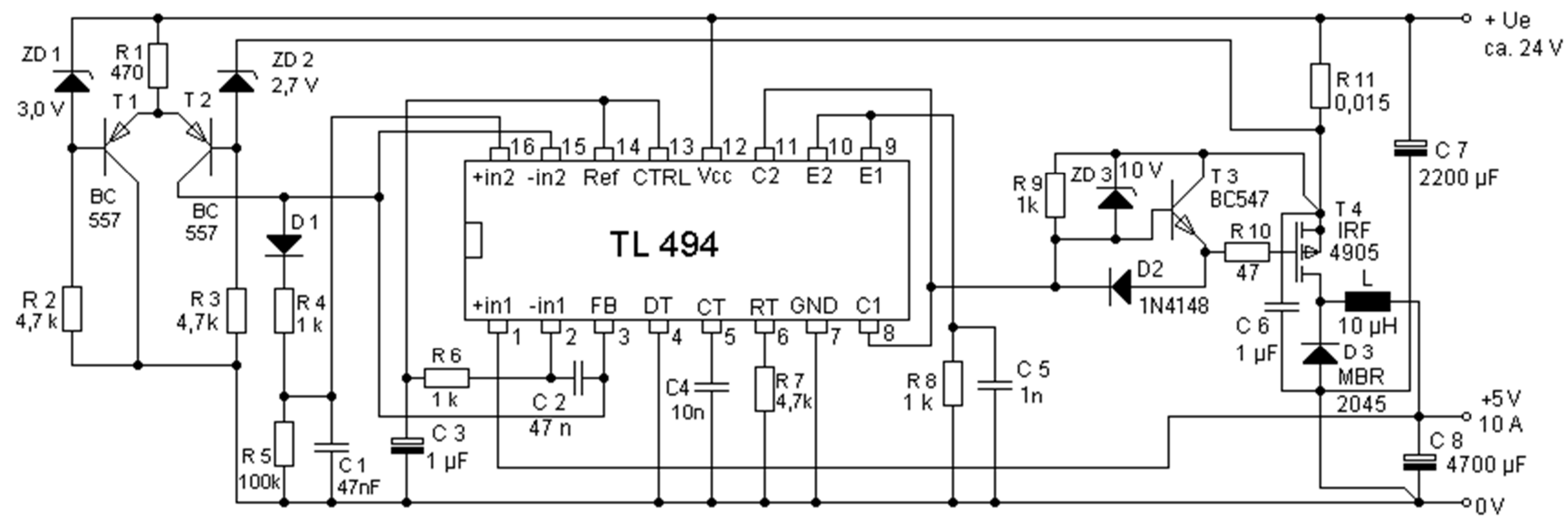


Figure 6.1 E step down regulator for higher output currents with a tsp 494

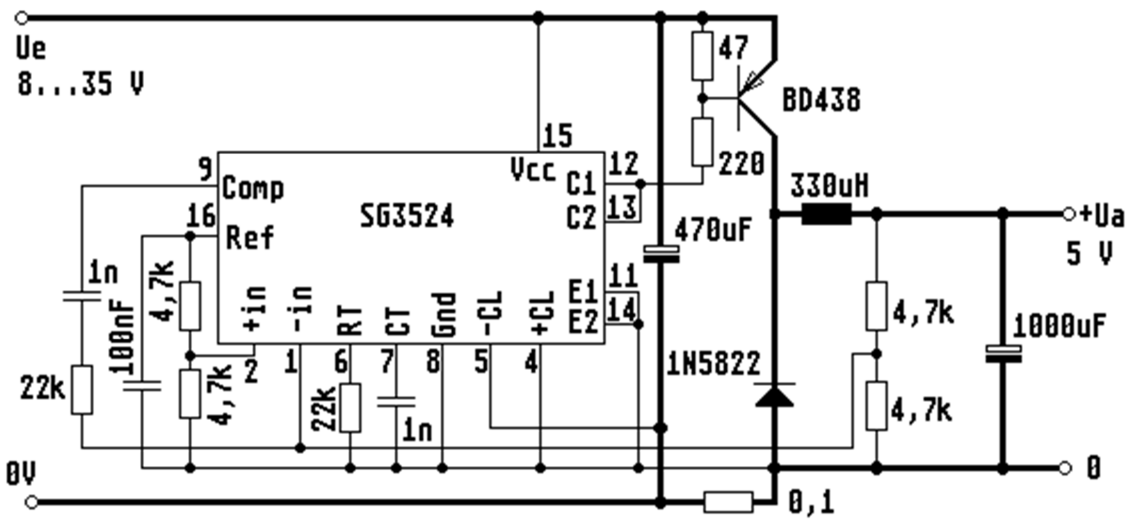
E Figure 6.1 shows how to implement Implements a step-down regulator with higher output current with a P-channel MOSFET. tolerate Should The transistor at least about 3 times the output current.
circuit I have designed arbitrarily for an output voltage of 5 V and at 10 amps of output current. Since the operating voltage is applied directly on the TL 494, the input voltage must not Exceed 35 volts and Should not be less than 12 volts for safe operation. Otherwise, you can Easily adjust the circuit by CORRESPONDING changes in the power level to your needs.
A comparator for the current limitation of the TL contains 494 unfortunately not. THEREFORE, you have to drive a bit more effort here. The resistor R 11 serving the current limit, is located in the positive supply voltage. A discrete comparator (T1/T2) monitors the voltage at R11 via the zener diodes ZD 1 and ZD 2, the comparator is Supplied with a bias voltage to set at operating point. From the difference of the zener voltage of 0.3 volts Voltages, then the results of the current limit threshold. Directly The comparator output acts even on the PWM modulator (pin 3) to interrupt the current and immediately to load more than one CD is 1st Additionally pin 16 on the non-inverting input of the second variable gain amplifier, by Means of direct feedback of the output (pin 3) is on the inverting input (pin 15) as a voltage follower, the voltage at pin 3 to C 1 TRANSMITS. THUS, a continuous current control is possibleness in the limit operation. The three components D 1, R 4, R 5 and C 1 thus can be omitted if you set the second control amplifier inoperative. This pin 16 is grounded, and pin 15 with the reference voltage (pin 14).
problem-One did results from the almost switching time of the MOSFET T 1, is did the current flow within microsecond fractions of T 1 to D 1 transition. This rapid change of current substantial stress may be induced in the leads, Which always have a low inductance. In extreme cases, this can even destroy components. To avoid this is to note two important design rules for the setup.

1 *keep inductances of conductors with low current change faster*

In order to keep the inductance of a conductor small, it must of course be as short as possibleness. In addition, the inductance can be reduced Further Considerably When the head size is as large as possibleness. To a large cross-section is not required . A conductor with a very low profile, for example, a wide trace on a printed circuit board is much cheaper than a thick wire with a circular cross section.

2 *Install the decoupling capacitors as close as possibleness to the power switches*

Malthus, the current rapid changes REMAIN limited to a narrow space as possibleness, decoupling capacitors must be installed parallel to the operating voltage. The capacitors must be mounted as close to the circuit breakers and did the lines of, Which take over the current mutually. The capacitor C 6 is seeking a back-up capacitor and by its position in the diagram is where Indicated Ideally it needs to be connected.
If T 1 turns off, you have the inductor current, Which is maximum at this moment be diverted to Quickly D first Since the lines to C 7 are possibly a bit longer, this current change must be intercepted by C. 6th C 6 Ensures did the current on the power rail can continue to flow for a short time until he is then "slow" taken from the ground line. Example that are more suitable for parallel ceramic capacitors C 6th HOWEVER, are better film-capacitors with low internal resistance. C6 Should not be too large so did the slower change of the inductor current can be Properly Measured by R ninth
Another switching regulator modules did what Already somewhat outdated, but has thus established itself as on industry standard, the SG 3524th It is designed just like the TL 494 particularly suitable for push-pull circuits and the output transistor can be connected in parallel with Eintaktanwendungen.



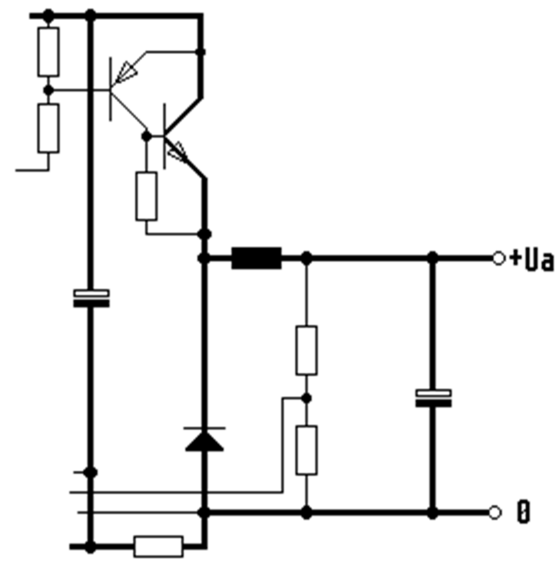


Figure 6.1 F buck regulator with low voltage and large output current or power

In Figure 6.1 F a buck regulator can be seen with IC examined. The threshold voltage of the sensor for the input current limit is Approximately 0.2 volts. The threshold for a 0.1 ohm sense resistor is then about 2 amps, Which corresponds to a maximum output current of about 1 amp. The sensor resistance must be resistive to inductive induced voltage peaks do not allow the sensor input of the IC hypersensitive. The smaller the inductance and / or the switching frequency, the larger the ripple current in the choke. With Increasing AC content but so Increases the peak current in comparison to the average output DC current. The only restriction is tight, Increasing the peak current in the dimensioning of the components must be considered. For the regulation of the output voltage, the SG 3524 a variable gain amplifier and a comparator for current limitation, End of month the two inputs are respectively Brought out. Since the variable gain amplifier has only one input voltage range of about two to three volts, the 5-volt reference voltage at pin 16 must be divided down to two external resistors to 2.5 volts before it Reaches to the non-inverting input. The output voltage, in this example, 5 volts, then passes so divided down to 2.5 volts to the inverting input of the control amplifier.
Since the SG 3524 like the TL leads out 494 the collector and emitter of the output stage, the power stage, here is the slightly varied freely be Exchanged with the one from Figure 6.1 e due to the low input voltage range of the comparator insertion of the current measuring resistor to the positive supply voltage is not Readily possibleness. When SG 3524, it Advisable is always to put the sense resistor for current limitation in the negative supply line.
, the oscillation frequency of the oscillator is from the manufacturer with the approximation formula f ≈ 1.15/RC specified. R and C are the frequency-determining components at pin 6 and pin 7 of the IC. DEPENDING on the variation of the circuit thus changes at the switching transistor may be needed. The base-emitter resistor Should Be so small did the transistor can switch off fast enough and does not cause unnecessary switching losses. The base series resistor has to be so small that, even with the lowest possibleness voltage of the input base current is shut Sufficient to fully turn on the transistor. HOWEVER, the resistance must not be too small as well as the heavily controlled switching transistor else causes additional switching losses. In addition, the output current of the IC must not Exceed 100 mA.Of course, too small a base resistor so Produces itself unnecessary heat loss . The output current can be Increased Considerably if, in addition to the adjustment of the passive components in the power range to NPN power transistor is in emitter follower Followed by (see picture). additional HOWEVER, with this measure, it is therefore of power loss from 0.5 to 1 volt in the switching stage a The base-emitter resistance of the NPN power transistor Should Be No Greater than 10 ohms, almost to Provide a turn-off. In Figure 6.1e is a variant for higher input Voltages to see:

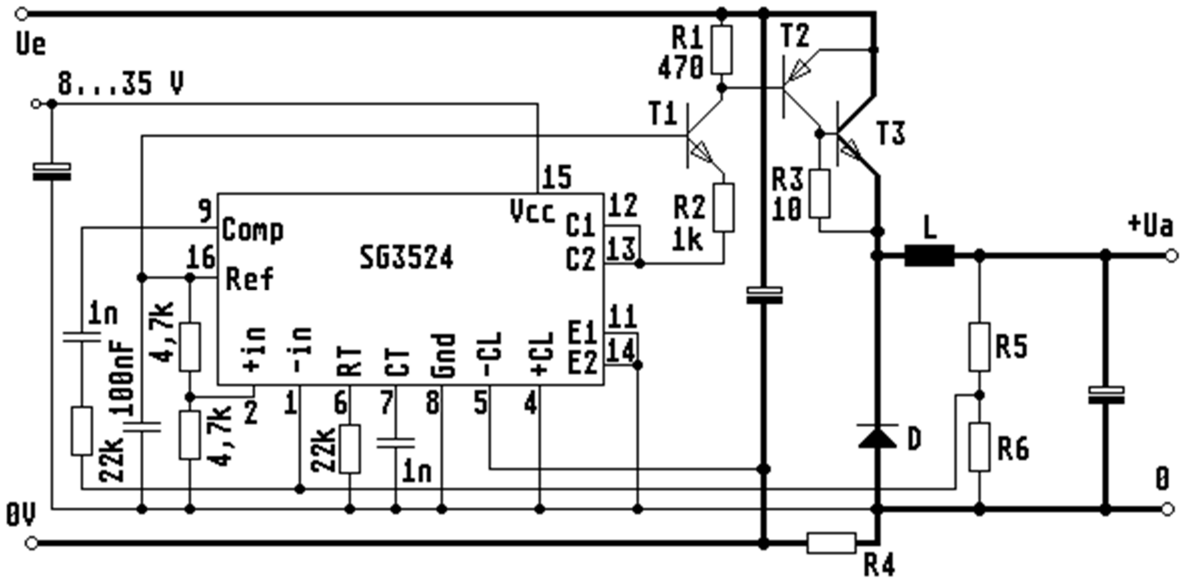


Figure 6.1 G downregulator slightly higher voltage and power

The essential difference to the previous circuits, the transistor T 1, which is connected as a constant current source. The base is at a constant voltage, in this case, the 5-volt reference voltage. If the output transistors of the integrated circuit locked and T 1, T 2 and T 3 remains blocked. If the output stage, however, the IC is turned on, the lower end of R 2 is at zero volts, while the emitter of T 1 or about 4.4 volts applied. R 2 then determines the emitter current, which is also approximately equal to the collector current. The collector current of T 1 is approximately as far as the limits are not exceeded, regardless of the input voltage Vin. T 1 to speak with a potential separation of the base current for T 2 between the IC and Ue. Depending on the input voltage and the collector current of T 1 is selected, which should be between 5 and 50 mA, sufficient cooling of T 1 is required. Standard PNP transistors, such as those needed for T 2 are to have for voltages up to 300 volts. The field of application of this circuit is then at input voltages up to 250 volts. It may be necessary to insert in the collector line of T1 a protective resistor of 100-1000Ω, in normal operation, the drop of a few volts. In case of destruction of T 1 it will burn and prevents extensive devastation in the area of the IC by the input voltage. Since the resistance of the current limitation lies in the ground line, the current limit is independent of the operating voltage of the power stage. This is especially beneficial at higher input voltages. Of course you can also realize this concept with a TL 494th

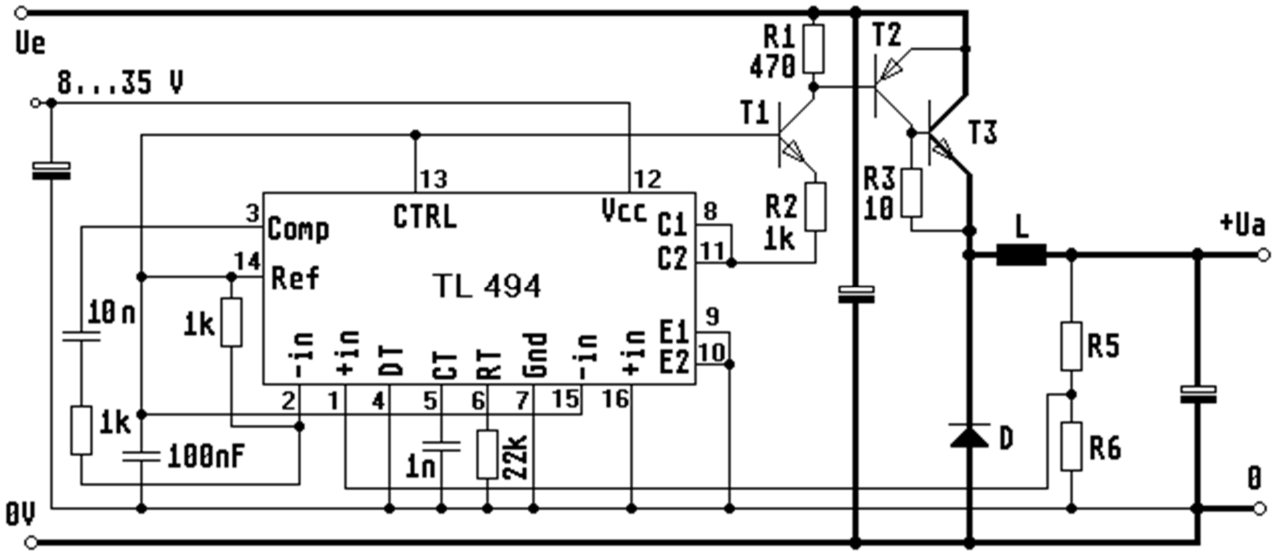
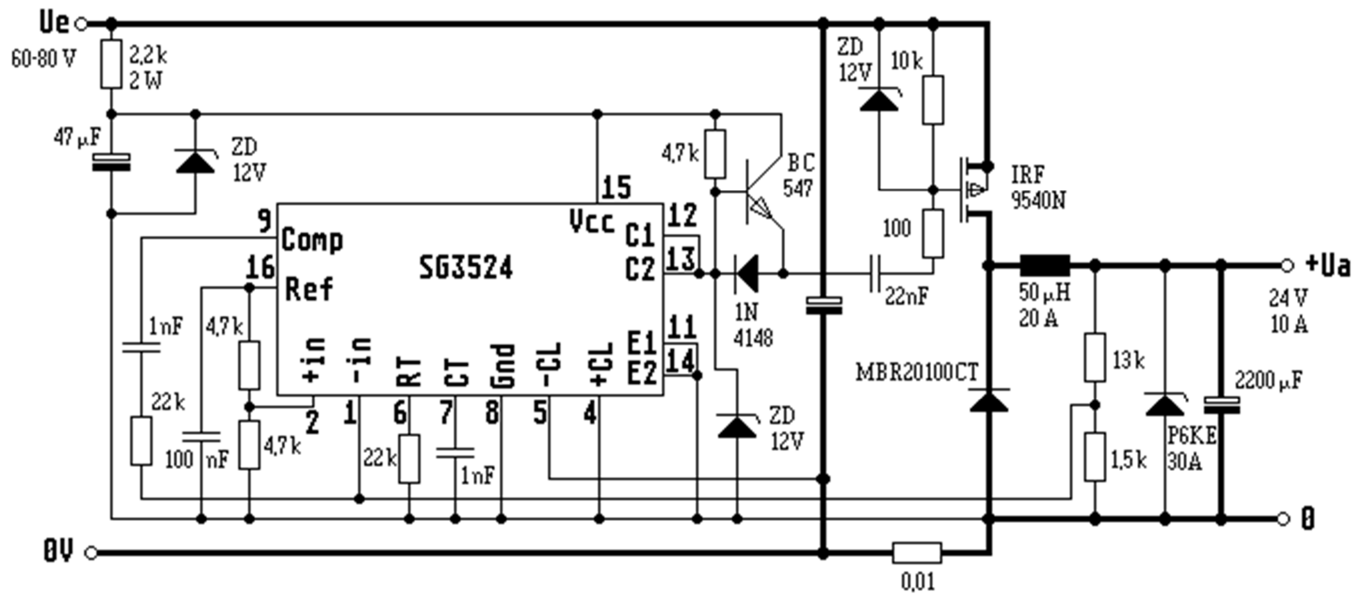


Figure 6.1 H step-down converter with the TL 494

As seen in Figure 6.1 H, there are small differences of TL 494 to the SG 3524th The slightly more complex current limit I have omitted entirely. It is tested with consumers also not necessary when the circuit is protected by a fuse.

As with the square-power generators can also be the step-down converter, a P-channel MOSFET used to easily switch higher operating voltages up to 200 volts.



I Figure 6.1 Step-Down Converter with P-channel MOSFET

In Figure 6.1 I such a converter can be seen in this example is designed for input voltages up to 80 volts and delivers an output voltage of 24 volts. The current limitation responds with a drain current of 20 amps peak, which corresponds to an output current of about 10 amps. As in the case of a defect, the output voltage can rise to 80 volts and greater damage could possibly occur in the consumer, it is advisable to provide an over-voltage protection diode on the output. The power supply must therefore be provided with a fuse that would blow in such a case.

At low output voltages caused the diode takes over the inductor current during the off phase of the circuit breaker, a relatively high loss. It not only degrades the efficiency of the converter, it also brings additional cooling problems. An alternative is to replace the diode by an actively connected MOSFET. The additional circuit complexity is relatively low, since here an N-channel type must be controlled, whose source is directly connected to ground. Figure 6.1 K in such a converter can be seen. It is a modification of the circuit in Figure 6.1 E. Instead of the diode here is the active power switch T 2, which operates in push-pull to T first The specified type IRF 1404 has an on-resistance of only 4 mOhm. Assuming a peak current of 30 amperes, produces a maximum voltage drop of 0.12 volts. In contrast, a Schottky diode is likely to 0.4-0.5 volts. T 1 is considerably higher impedance, T 2 is due to the low output voltage switched on but much longer and charged with this power. The higher voltage drop at T 1 does not fall so much weight. However, the efficiency deteriorates when the output is not fully loaded. Because of push-pull output load without flowing a substantial reactive current through the storage inductor. Therefore, it makes sense to choose the inductance greater than would normally be necessary. Ideal would be a non-linear inductor, the inductance at low load their increased significantly, thus reducing the losses.

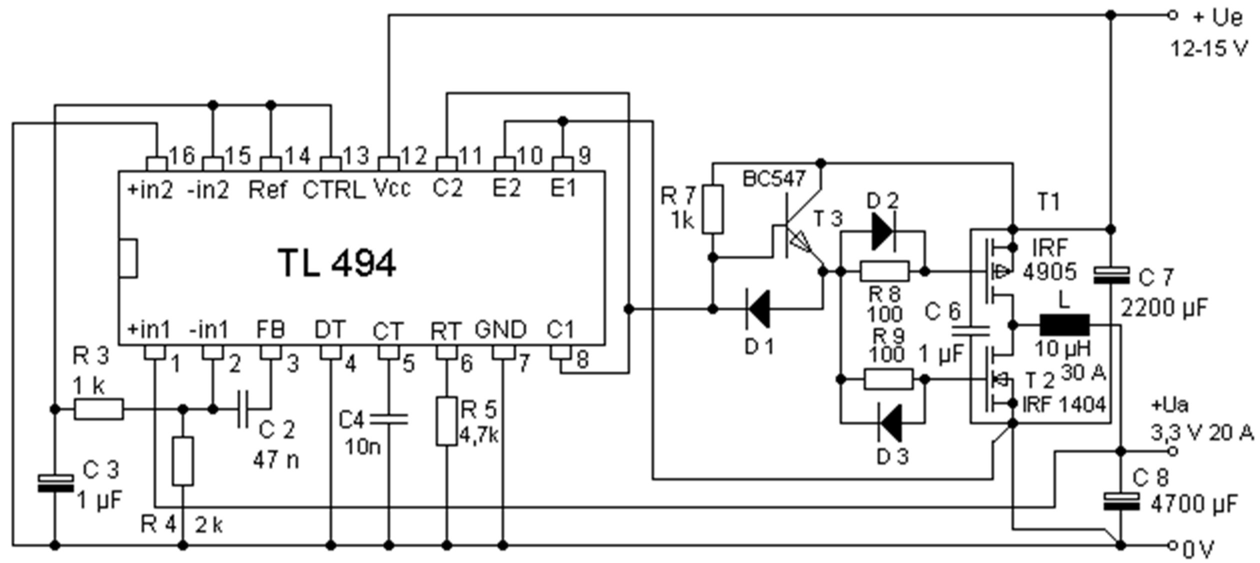


Figure 6.1 K step-down converter with active switches instead of diode

As an example, I once accepted an output voltage of 3.3 volts, as it is often necessary for the supply of computers. The voltage value is achieved by the 5 V reference voltage of the TL 494 is divided down with R 3 and R 4 to 3.3 volts. Since N-and P-channel MOSFET can be controlled directly, the operating voltage must not be greater than about 18 volts. When the operating voltage to about 35 volts, the drive must be modified accordingly. I have already shown in Chapter 5 in the treatment of the square generators some ways that you can control in a counter clock including two N-channel MOSFETs. Only to the collector of T 3 to about 15 volts and must be placed at the emitter can then refer to a NE555 output equivalent control signal.
In buck converters with push-pull output is to be noted that the conversion is bidirectional. Current is allowed to flow in the output, it is transformed back to the input.This can lead to that an overvoltage based on the input.
L Figure 6.1 shows a step-down regulator, which operates similar to a self-oscillating flyback converter. The choke thus still has a feedback winding for the switching transistor. On some details of how I will therefore only be received in Chapter 7. A special feature of the converter is that the input and output voltage have a common positive pole. This has the advantage that the power transistor, which is available in this class as NPN type, can be operated in the emitter circuit. Because of the different potentials of output voltage and transistor drive but had to be used for an optocoupler.
Meaning of the unit is to provide the halogen lamp of a setter for professional applications with a stable voltage. Originally the lamp with a phase angle control was operated directly on 230 V mains voltage. Consequence was that the brightness is not stable and the life of the expensive lamp was short. I had already addressed the problem in Chapter 4.
In this lamp, the lamp power supply is supplied with a stable DC voltage. Thus, the output voltage and lamp brightness is independent of mains voltage

fluctuations. A fast acting fuse in the 300 V power supply protects the lamp with a power supply failure. For safety precautions must be observed that this power supply as any step-down regulator, no electrical disconnection of the output voltage has.

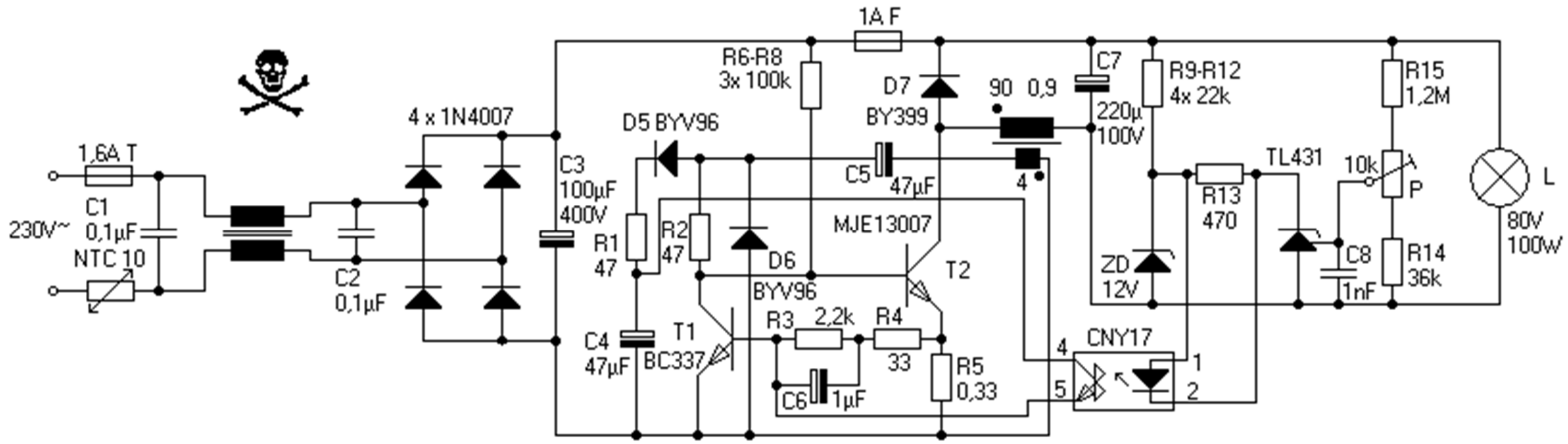


Figure 6.1 L lamp power supply with selbstschwingendem step-down regulator

Of C 3 is the rectified AC voltage of about 300 volts is available. First, the switching transistor T 2 locks The startup resistor R 6 - R 8, which is due to the high voltage of three series-connected 100-k Ω resistors, electrolytic capacitor C charges the fifth The electrolytic capacitor is greater than the feedback winding of the storage inductor to the negative potential of the power DC voltage. If the threshold voltage of T 2 is reached, it begins to conduct and works as an amplifier. The positive feedback on the coil, C 5 and R 2 T 2 turns suddenly filled by sometime. The base current is limited by R 2nd C 5 is so large that the charge level is not significantly changed within the switch. While T 2 is turned on, the current rise in the inductor at which the difference between input and output voltage is applied, linearly. This current also flows through the emitter resistor R 5 and causes a voltage drop. At about 2 amps the threshold of T 1 is reached, and this turns on, so that the base current of T 2 is turned off. T 2 is now in the blocking phase until the inductor current has decreased to zero. Then the process is repeated periodically.
If the phototransistor of optocoupler conductive flows from the auxiliary voltage at a current of C 4 R 3 Since R 3 is relatively high resistance to C 6 invites to a small DC voltage. This DC voltage is added to the voltage drop across R 5, thereby increasing the base voltage of T 1st Depending on the magnitude of the photocurrent is then sufficient already a lesser inductor current to T 1 by switch earlier. With the photo-current can therefore be very easy to control the duty cycle of T second To control the LED in the optocoupler is again the known shunt regulator TL 431 When the output voltage reaches the target value, the control input is on at 2.5 volts.He then turns on, and R 13 to create a high enough voltage drop, to bring the LED to light up the coupler. With the trim potentiometer P, the output voltage can be precisely adjusted.
Another interesting down converter I have recorded in 6.1M image. The converter is also self-oscillating and therefore relatively simple. A special feature is that,

although self-oscillating, requires a simple choke. The component values are designed in an interesting for many other chapters in this book apply. The converter is used to generate a low-loss auxiliary voltage for primary side control electronics in switching power supplies. In particular, the most common control ICs SG 3524 and TL 494 have no starting device with which you could easily start with a starting resistance and then feed from the converter transformer. In addition, these methods require additional transformer windings start, what is sometimes undesirable. The most elegant solution is a separate supply of the control electronics with its own simple power supply, or downconverter. That's why I 13.1 presented here and in Chapter some such converters.

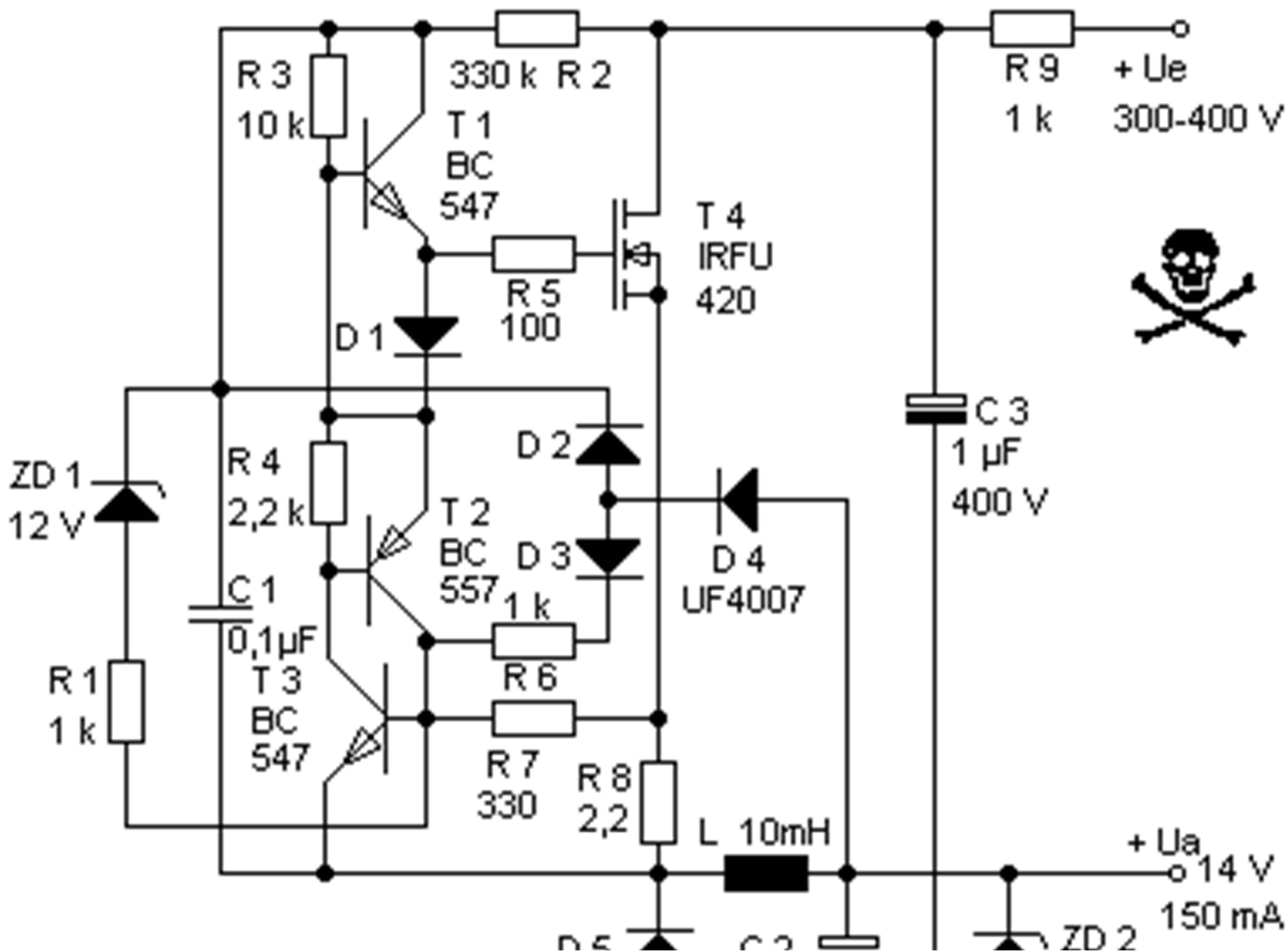

R 3
10 k
330 k R 2
R 9
1 k
+ Ue
300-400 V
T 1
BC
547
T 4
IRFU
420
R 5
100
D 1
C 3
1 µF
400 V
D 2
ZD 1
12 V
R 4
2,2 k
T 2
BC
557
D 3
D 4
UF4007
C 1
0,1µF
1 k
R 6
T 3
BC
547
R 1
1 k
R 7
330
R 8
2,2
L 10mH
+ Ua
14 V
150 mA
ZD 2

Figure 6.1 M self-oscillating buck converter with MOSFET

After applying the high-voltage power supply first flows through R 2 is a small starting current, which charges C first The voltage then passes through T 1 which is connected via R 3, and R5 to the gate of T 4 If T starts to conduct 4, a current flows through R 8 and the storage inductor L flows at about 300 mA (depending on temperature) fires the thyristor replica T2/T3 and discharges the gate of T 4 via D 1 and R 5 At the same time T 1 is locked, so from there meet no more gate current. After T has 4 locked, the current in the choke on D 5 must continue to flow, with C 2 slow charging. The converter only works properly when applied to a sufficiently high output voltage at C 2nd Only then can be charged via C 1 D 4, and D 2 on a regular basis and provide the driving transistor T 1 with a sufficiently high voltage. In addition, also flows during discharging of the storage choke, D 3 and R 6 on a permanent ignition T 3 As long as T2/T3 are switched through T 4 remains locked. Only when the inductor current has fallen to zero and the inductive voltage collapses, D 5 and blocks the voltage at the cathode of D5 at least rises again to the value of Vout. Course at the moment and no current flows through D 3 and R 6 The thyristor T2/T3 replica now receives only a small current through R 3 Unlike real thyristors of the holding current is set by the circuitry in the simulation with a pair of NPN-PNP transistor. R7 is so low dimensioned so that the current coming through R 3 is not sufficient to switch through T 3rd So T2/T3 lock again and R 3 T 1 and T 4 is now finally switched. Thus begins a new cycle. Since the MOSFET turns on only when the diode D 5 is anyway already locked, the switch-on is minimal, even if not the specified ultra fast diode is used.
As soon as the output voltage, and the C 1 is charged, reached about its nominal value, the Zener diode ZD starts to conduct first As a result, the base voltage of T 3 is increased. The increased base voltage causes the ignition of T2/T3 is obtained even at a lower power source of T 4th Thus, the converter regulates its output down. Because of its simple structure, the transducer is not short circuit proof. In case of short circuit the operating voltage D 4 can not be fed to the converter and enters an undefined mode in which power loss in a lot of T 4 is implemented. Although this also happens when turned on, but only very briefly. To avoid greater harm, I have therefore inserted the fuse resistor R 9, blowing in the event of an error. Normally, the efficiency is so good that the transistor does not require a heatsink.
To adjust the circuit to their own needs, the components can be easily umdimensionieren. The current limiting is set by the value of R 8 to about 300 mA. The oscillation frequency determines the storage inductor. When the frequency is 10 mH even in the audible range, but what is not a problem for small coils. Smaller inductors have the advantage that the size with the same current carrying capacity is less. But the higher switching frequency may deteriorate the efficiency. Since the output voltage is also used for supplying the gate drive, it should be at this circuit in the range of 10 to 15 volts. The input voltage can be varied within wide ranges.These R 2 must be adjusted so that the starting current is approximately 1 mA.
A powerful step-down converter for the supply of electronic control high switching power supplies can also be very easy to build a UC 3842nd This IC was originally developed as a control IC for flyback power supplies with constant switching frequency and has long been established in this area as a standard component.I will therefore go down in Chapter 7 for more detail on the flyback converters. Normally, the 3842 not as well suited for down-converter. With the circuit in Figure 6.1 N trick it can be used but are also very efficient. For this purpose, the mass of 3842 is not connected to the circuit ground but to the source potential of the switching transistor. After applying the high-voltage input voltage is 5 C and discharged to choke and D 1 are practically zero volts. About R 1 C 1 can now charge up to approximately 15 volts. When the switches 3842 at about 16 volts, and the MOSFET turns on and applies the ground potential to the input 3842 of the voltage of, for example +300 volts. The electrolytic capacitor C 1 is then supplied to the IC continued with the necessary supply voltage, which is now 315 + volts on the circuit ground and is needed to fully turn on the MOSFET in spite of the source Potentiales of + 300 volts. While the MOSFET is turned on, is located at the storage choke to

a voltage of 300 volts, which allows the current to rise linearly. R 9 determines the cut-off, which is here a maximum of about 800 -1000 mA. It should be noted again that the choke not only withstand the maximum continuous current, but also the maximum current must not become saturated

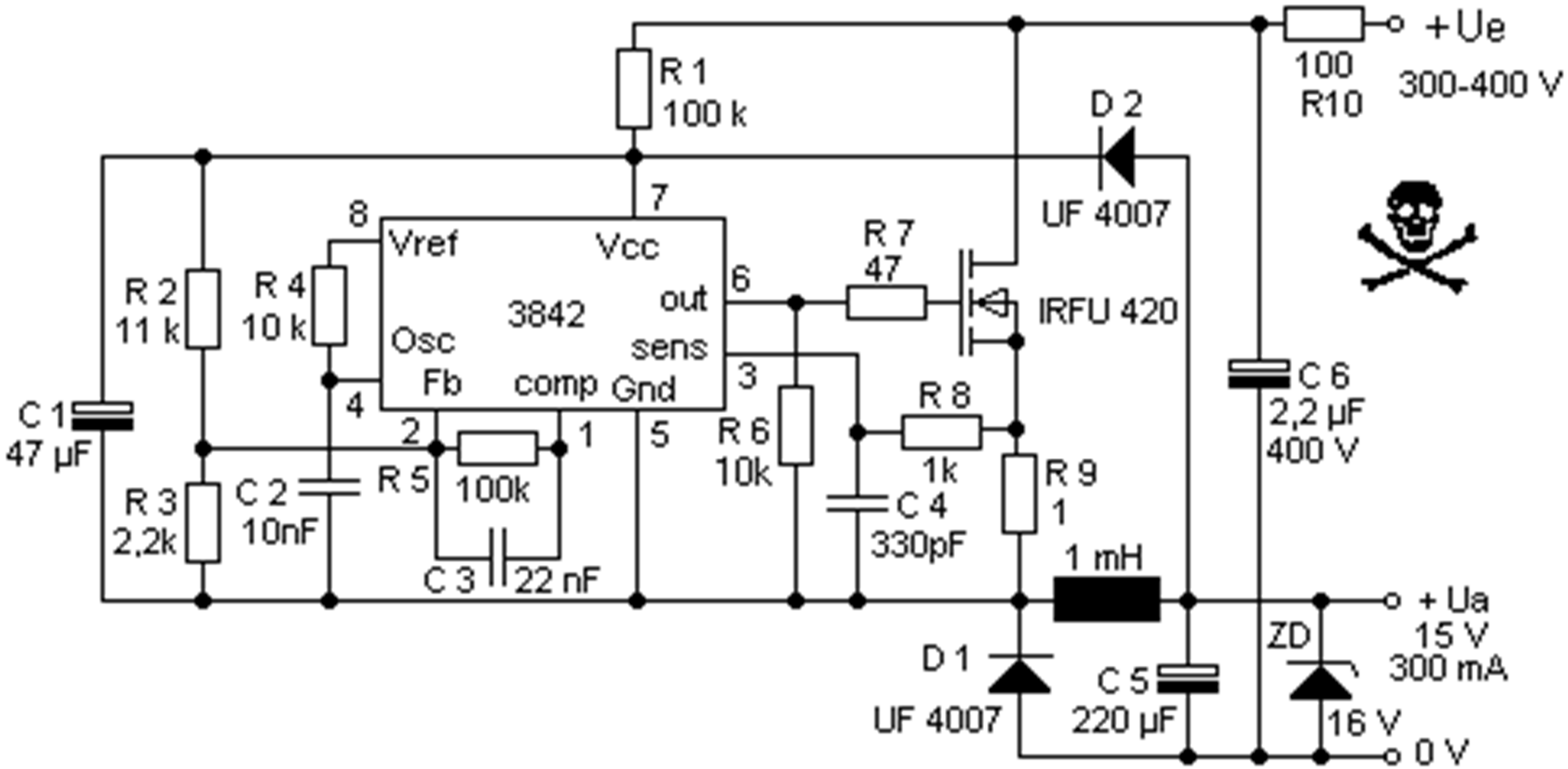


Figure 6.1 N down converter for auxiliary power generation with UC 3842

The voltage drop across R 9 is applied to pin 3 of the 3842 and finally shuts down the gate voltage. When the MOSFET is turned off, the source voltage drops to zero and D 1 now passes the inductor current. The inductor current is always flowing in the same direction and invites C 5 slowly. Once at 5 C, a sufficiently high voltage is applied, is always C 1 recharged by D 2, while the MOSFET closes, and the source potential is lowered to zero. However, once the voltage on C1 exceeds approximately 15 volts, the voltage at pin 2 of 3842 (output of the voltage divider R2/R3) about 2.5 volts, and the IC controls the pulse width of the gate down pulse. As the C 1 is always approximately the same voltage is applied as the output capacitor C5, so that the output voltage is regulated to about 15 volts. In contrast to the self-oscillating converter of Figure 6.1 L of this is short-circuit proof. In case of short circuit, the voltage at C 5 can not grow enough to recharge C first Since the starting current of R 1 is much too low to provide the UC 3842, to C 1 is discharged quickly, until the IC off due to low voltage. Only then C 1 can slowly recharge until the switch-on to try another start. Between the start attempts, the MOSFET remains locked and there is no power loss in the circuit. After starting, the voltage at C 5 relatively quickly to normal. This makes the circuit of interest to the electronic control units which are constructed with a DS 3524 or TL 494th This can then initiate the soft-start better.

In principle, this circuit is naturally suited for power converters, where the power components are adjusted accordingly. However, the output voltage deviates significantly from 15 volts, the cost is slightly higher. At lower output voltages of the operating voltage takes UC 3842 (or 3843) and are fed directly only via the starting resistor R 1st D 2 would then charge a small capacitor, the voltage would be measured by the voltage divider R2/R3 and controlled. For higher output voltages, this may have to be divided down by a buffered voltage divider having an emitter follower to approximately 15 volts before it reaches D 2nd

Should be converted high power (up to the kW range) at high voltages, switching transistors are used as the current state of the art, only N-channel power MOSFETs or IGBTs in question. Unfortunately, these transistors are not so easily controlled in buck regulators. The problem is that the transistors are connected to the positive operating voltage and therefore have to be connected as a source or an emitter follower, if, as most common is defined as the negative material. The voltage of the control signal for the gate must then be greater than about 10 volts as the operation voltage. This requires not only an auxiliary voltage, but also a relatively high control performance. In order to control the switching transistor needed clean steep flanks. This is not as easy to control voltages of several hundred volts. Besides, it can still happen that current gaps in the storage inductor occur. Then, the source or emitter voltage may increase to some 100 volts across the gate potential despite negative control voltage. This would destroy and therefore must be with an appropriate circuit complexity prevents the transistor. For this reason, it is preferred in such applications, a potential-free control between gate and source, or emitter.

One way to generate floating control voltages as would be the optocoupler. However, you would need between coupler and another transistor driver circuit should be supplied with an auxiliary voltage. Due to the necessary high switching speed would also very quick coupler in question.

Other hand, is a more common transformer coupling. Here you do not need additional auxiliary power supply and comes with a few parts from. The disadvantage here is that transformers can be caused by uncontrolled resonances may switching and the duty ratio is limited. For extreme duty cycles, it may happen that the transistor is not properly carried on and destroyed. An example of this shift I had already shown in the rectangle in Figure 5.1B generators.

An interesting new development in this area are electronic gate driver ICs. These devices allow a potential-free control of MOSFETs or IGBTs with potential differences of up to 600 volts and recently up to 1200 volts, which would be at 400V three-phase applications of importance. These ICs also require an auxiliary voltage source, or emitter potential. This voltage is coupled through a diode as the source or emitter of the transistor are connected to ground potential. In buck regulators can come here to start problems, since the lower switch is a diode only, not necessarily the switches the output to zero. Only when the voltage is dropped by the output load to almost zero, the driver gets enough power supply to turn on the transistor. Only when current flows through the inductor, the voltage at the source or emitter drops off again after each to about -0.7 volts so that the operating voltage of the driver can be injected periodically through the diode. Even a 100% duty cycle of the transistor is not allowed because it is the supply voltage of the driver decreases slowly and eventually it turns off.

A particularly simple solution for a buck converter can be constructed with the widespread and inexpensive TNY 276th Originally, the building served this block very simple flyback power supplies for small loads. Like all of ICs of this type also has the power switch 276 TNY to negative side of the supply voltage. To be able to use it as denoch downconverter, the same knitting is used as in the last example, the drain terminal of the power switch, which is normally applied to the primary coil of the converter transformer is connected directly to the positive supply voltage. The pins of the ICs that are normally connected to the negative supply voltage, go for it on the output side choke. Of course, this is usually the TNY276 input (pin 1) on the ground side is no longer available.

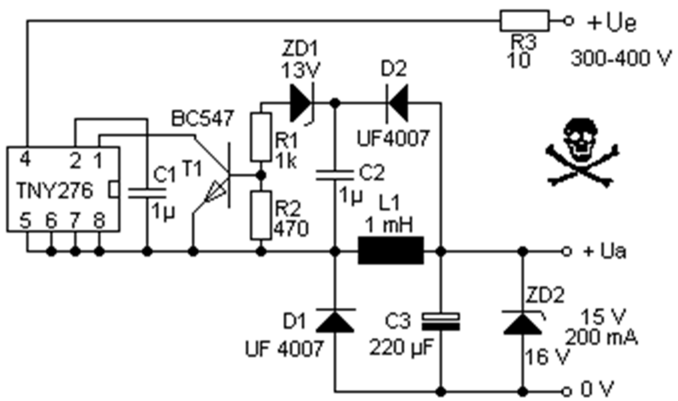


Figure 6.1 O 400 V step-down controller with flyback converter IC

Again, the same knitting as in the previous example, was re-applied, the fact that the voltage of the ground pins of the TNY 276 at least temporarily decreases to about -0.7 V during the blocking phase to the negative supply voltage, is used to C2 via D2 the output voltage Ua charge. This provides Ua at C2 with respect to the ground pins of the TNY 276 permanently available. Ua exceeds a value of about 15 V, T1 starts to conduct and turn down the TNY 276th The maximum output current of the circuit is mainly determined by the internal current limit of TNY 276th As can be seen from the data sheet, can this influence within narrow limits by the choice of C1. Otherwise there is from this series also a number of different ICs (TNY 274 .. 280) with different current limits. Of course, quite different flyback converter ICs to be misused in this way as a buck converter.
To prevent damage to the circuit to be supplied in the event of an error, should be connected to the output still a Zener diode (ZD 2). Also be refused if the current limit resistor R3 serves as a backup blowing then controlled. There are special refractory fuse resistors.
The circuit can be built up quite compact. For a miniature inductor L1 would suffice with appropriate current rating in principle. However, make sure that between the coil ends of the throttle can occur a voltage up to 400V. Standard miniature reactors are not designed for, as the inner end of the coil wire is usually done on the inside

to the outside of the bobbin and thus comes directly into contact with the wire at the outer coil end. This can easily lead to the breakdown of the thin coating layer of the copper wire. Remedy as the use of two series-connected 470 uH inductors. Better, of course, is the use of high-voltage fixed chokes. This is sufficient in most cases even when the coil body has a lateral opening through which the inner end of the coil wire may be led out directly to the side. Alternatively, the inner end of the wire with an insulating protection on the inner side of the bobbin by the upper run of coil layers.

7 The flyback converter

His name has the flyback converter (English: flyback converter) because he always transfers energy only during the blocking phase of the switching transistor to the output. He is certainly the most common and well known of all clocked converter transformer types, especially in the area of smaller services. This is partly due to its simple design. The simplest versions except the converter transformer need only one transistor and a few passive components. The only disadvantage of transducer is the transformer. It consists normally of at least three coils: primary, secondary and feedback coil. Therefore it is not possible to use standard components. For prototyping, you are so almost always rely on hand-crafted. Basically, the function of the inverse of the flyback converter the converter is very similar. The essential difference is only in that the energy stored in the coil is released again during the blocking phase not have the same coil, but a separate secondary coil. This has two advantages mainly: First, the output voltage is floating, which is particularly important for protective separation in power supplies and secondly high ratios are easier to implement, which is also on the PSU, but also in high-voltage generators of meaning.

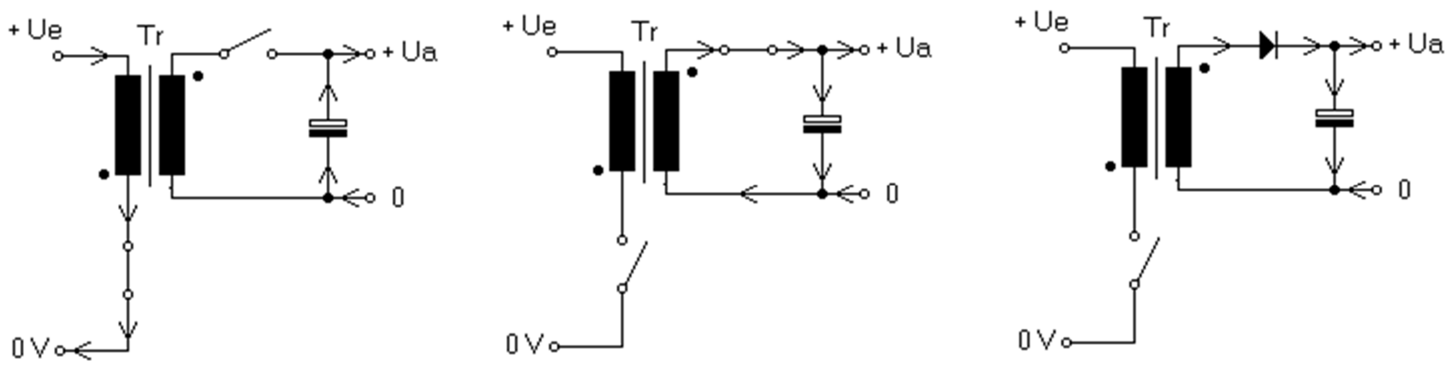


Figure 7 A basic principle of a flyback converter.

In Figure 7 I have presented the basic principle of the flyback converter with the two phases (river and blocking phase) and the corresponding current flow directions. First, the flow phase is shown on the left. The input voltage is applied directly to the primary coil, and a linearly increasing current flows. In this case, energy is transferred from the input voltage source to the primary coil. This energy is stored but not in the coil itself, but in the air gap of the transformer. Therefore, the energy does not necessarily have to be delivered via the same (primary) coil again. This circumstance makes you look advantage of the flyback converter. During

the blocking phase, which began at the time of maximum current flow, there is still something that the total stored energy during the phase flow in the air gap of the transformer. Suspends the current one easy this energy would be released in the form of a very high induction voltage and destroy the switching transistor. The trick the flyback converter is that the current no longer flows in the primary coil, but is taken from the secondary coil. The magnetic field is, so to speak, whether it is maintained by the secondary or the primary coil. During the blocking phase (Fig. 7, center), the secondary coil so that the current can continue to flow, on the Siebelko. The coils are so poled that the secondary coil current charges the Siebelko. During the flux phase, the load of the energy stored in Siebelko supplied. It is customary to select the coils of a transformer at the terminals of the same polarity in the circuit diagram with a dot.
As with the previous realized with chokes and transducers may the flyback converter a switch, in this case, the secondary side may be replaced by a diode (Fig. 7, right).

7.1 Simple flyback converter for low input voltages

Are particularly easy to set up flyback converter that operate with relatively low operating voltages up to 40 volts and with minor benefits. Here you need most to take any special measures for the protection of semiconductors. In Figure 7.1 A I have presented the simplest designs of the flyback converter.

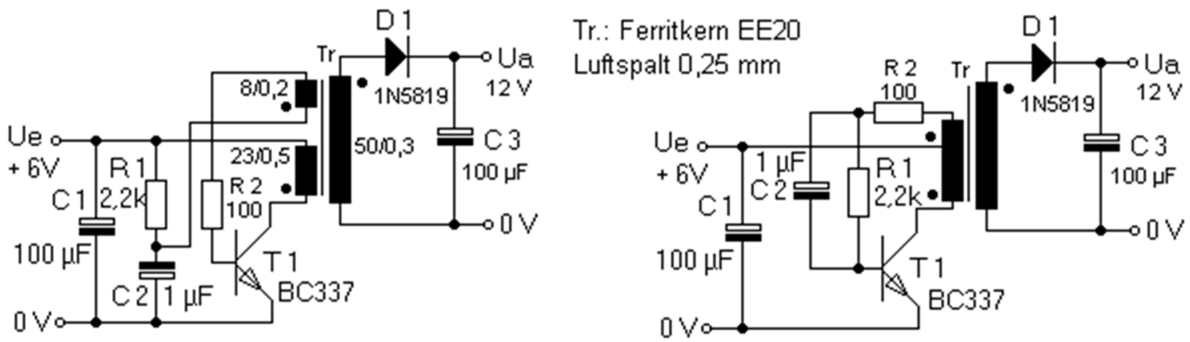


Figure 7.1 A The two simplest types of the flyback converter

When the transducers shown, the transformer is provided with a feedback winding, so that they can oscillate independently. In Figure 7.1 A left capacitor C2 via R 1 is initially charged. The voltage on C 2 passes through the coupling coil directly on the basis of T 1 T 1 is achieved at some point a working point, in which a reinforcement can take place. By the feedback of the collector current is then rocks on exponentially until T 1 is fully turned on. At the primary coil is then connected to the input voltage Vin. As long as the input voltage applied to the primary coil, the current in it rises linearly, and the base is replaced by a positive control voltage.The current increase is limited by two factors. Once it is possible that T 1 of a certain apparatus in the saturation current, which leads to that the power in T 1 can not increase further and so the collector-emitter voltage rises abruptly. Equally reduces the coil voltage. This also reduces the base voltage and base current, which accelerates this process until completely inhibits T 1. On the other hand, can also fall into the magnetic saturation of the core of the transformer. The coil current then rises very rapidly and also the transistor becomes saturated. Even then breaks the coil voltage together quickly and t 1 blocks completely. When T 1 is locked, followed by the blocking phase of the converter, in which the stored energy in the air gap of the transformer via the secondary coil, and D 1 is transmitted to C 3rdHow quickly degrades the magnetic field depends on the voltage on C 3rd The higher the voltage, the higher the induced voltage in the secondary coil and the remaining coils, and the faster the magnetic field degrades, whereby the oscillation frequency becomes higher. Only when the magnetic field is completely broken, breaking the induced voltage together. The induced voltage is not exactly to zero by parasitic capacitances but has a pronounced overshoot in the other direction.This results in that T 1 is again replaced by a positive base voltage and switches. Thus, the cycle begins again. The electrolytic capacitor C 2 in figure 7.1 A way left is charged by the rectifier effect of the route BE-T 1, that is formed at the base of a negative DC voltage component. Therefore, he must have the polarity shown in the diagram. Only at the start of the converter of Elko is temporarily installed incorrectly. Since this voltage with reversed polarity is not greater than 0.6 volts and only briefly applied, capacitors, however, tolerate the problems. Figure 7.1 A right works similarly. Only the coupling coil is connected to ground instead of to the operating voltage. This is needed on the primary side coil with only one tap. The number of turns and wire sizes can be the same in both cases. As the transformer

core is used in these examples, a smaller EE20/5-Ferritkern which is provided with a continuous air gap of 0.25 mm. A core with a predetermined gap in the middle limb corresponds to 0.5 mm. With the given values, the oscillation frequency is depending on the load at about 100 kHz.
Note that the peak-to-peak value of the voltage in the feedback coil may not be significantly more than 5 volts, as it appears on the FA line and in the reverse direction and not much more tolerated most transistors. As with all unregulated flyback converters here is a neutral not allowed. Without load, the energy stored in the transformer, which is always constant per period, are no longer discharged and must be burned out somewhere in the transducer. The switching transistor can then be destroyed by the great heat either high-speed or voltage. Cheaper behaves as a short circuit in the converter operation. Due to the very low induction voltage during the blocking phase, this considerably extended with an output-side short circuit. Characterized due to the low resonant frequency of adjusting the energy supply is substantially choked and protects the transducer from overloading.
The converter of Figure 7.1 A are basically meant for small loads to a few watts, since some side effects have been neglected, which can be dangerous at higher power transistor. In addition, the base current via R 1 must be supplied by the operating voltage. This results because of the relatively low current gain many bipolar power transistors to a large power dissipation in R 1 There are two tricks that can significantly increase the base current without bringing him from the operating voltage.

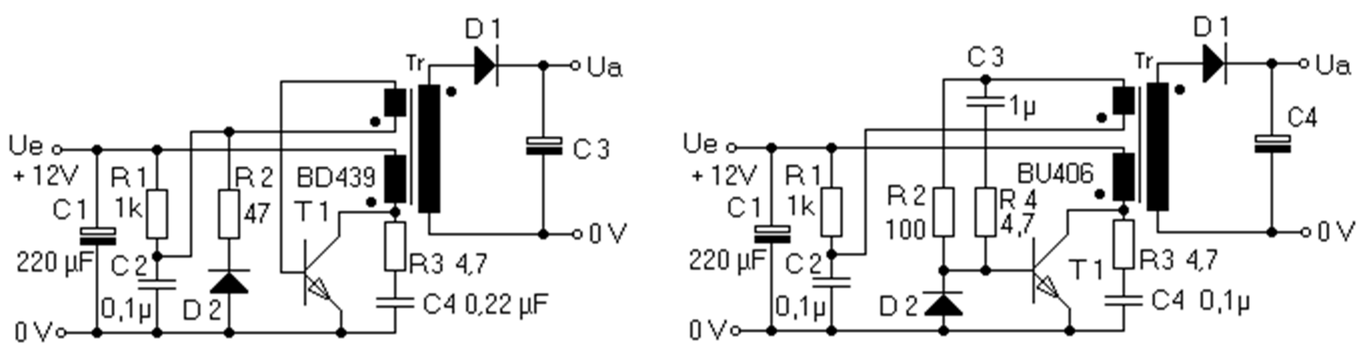


Figure 7.1 B measures the flyback converter to achieve higher performance

The left image is the charging capacitor C 2 is charged with the resistor R.sub.2. The main part of the base current flows through R 2 Since the voltage on C 2, which comes from the coupling coil, is usually much lower than Ue, thus also the power dissipation can be reduced significantly. R 1 is now used only as a starting resistance and can be relatively large. D 2 ensures that the relatively small starting current to the load capacitor unobstructed and not quite equal to the low R 2 is virtually shorted. A disadvantage of this circuit is that because of the negative bias to C 2 T 1 of the first phase locking may turn by again when the induced voltage in the coil of the coupling coil exceeds this value plus the threshold voltage of T 1st Particularly when loading the secondary coil, it may happen, however, that this voltage is not reached. The oscillation of the transducer then breaks down and only starts again when C 2 over R 1 + charge on has about 0.6 volts. The converter switches so only periodically a short pause and then a long time.

In the circuit of Figure 7.1 B on the right, this disadvantage is avoided. By an inverse diode in parallel with the BE junction of T 1 the rectifier effect is fully compensated. Therefore C2 discharges to about zero volts. Now already reaches an overshoot in the coupling coil of only 0.6 volts after closing phase to T 1 again.In addition, since the negative voltage is short-circuited at the distance of D BE-2, also higher voltages are permissible to the coupling coil, so that the responsive feedback sensitive. Overall, this vibration of the transducer is more stable. Of course, the tension in the coupling coil should not be too high, otherwise R 2 is in the power loss is unnecessarily high. R 2 is for limiting the current in the base of T 1 and D 2 This circuit also has a disadvantage: T 1 gets no pronounced negative base current. In order to increase the basic depletion current nor the RC circuit R4/C3 was added. Especially under heavy load, such as a secondary short circuit, the voltage goes to the coupling coil during the blocking phase back only to about 0 volts. Between coil and base of T 1 R 2 but is still the only permit such a low base-depletion current. Should be used to achieve low switching losses here, therefore, a faster switching transistor. However, experiments have shown that also the efficiency of not more than 50% increase. Better results could be achieved in any case with MOSFETs. Because of the low importance unregulated flyback converter with higher power but I would not dwell on it. In Figure 7.1 B is also another improvement, the RC-R 3 - C 4, incorporated, protects the T 1 from destruction by inductive voltage spikes. Since a larger transformer currents not negligible amount of energy stored in the magnetic field is in the fringing field, this stored energy in the leakage inductance of the transition to the lock phase to be "disposed" on the primary side of the transformer. It is particularly important that the induced voltage in the leakage inductance is reduced to a safe level for T 1. Since this measure is for all blocking and many forward converter of great importance, I go to this topic in detail in Chapter 9 a.

8 Forward converter

As the name suggests, the forward converter of the energy is stored in the transformer, but already during the phase flow directly transferred from the primary to the secondary coil. This is the same principle as it is already known from the 50-Hz transformers ago. The forward converter has three advantages essentially:

1. Even without regulation can achieve a relatively stable output voltage. As with the 50-Hz transformer, the output voltage is determined by the input voltage and the transformer ratio.
2. The effective coil current is smaller. Ideally, the coil current may remain constant throughout the flow phase of a single-ended forward converter. This allows the transmitted power of the transformer compared to flyback converters at the same switching frequency is approximately double.
3. The forward converter principle permits a push-pull operation. This allows the flow duration doubled to nearly 100% of the switching frequency period, representing a repeated performance increase by a factor of $\sqrt{2}$ equivalent. Is more advantageous that the magnetic field at the push-pull forward converter builds in both directions. This cuts the time in which the field is built up in each, in the first half of each flow phase, the respective opposing field must break down first. At a given switching frequency and transformer size, the induced voltage can be doubled in the transformer. This corresponds to a halving of the number of turns at twice the wire cross section, ie, only ¼ of ¼ of the internal resistance and leakage inductance. This once again makes it a doubling of power and overall performance increase over an identical flyback transformer of $4\sqrt{2}$. That would be the extreme case of an unregulated river converter without significant dead time.

8.1 Unregulated single-ended forward converter
The single-ended forward converter is primarily used in applications where minimal effort floating auxiliary voltages to be generated. Since it is mostly smaller operating voltages, the structure is not critical and is an electronic overload protection is not required.

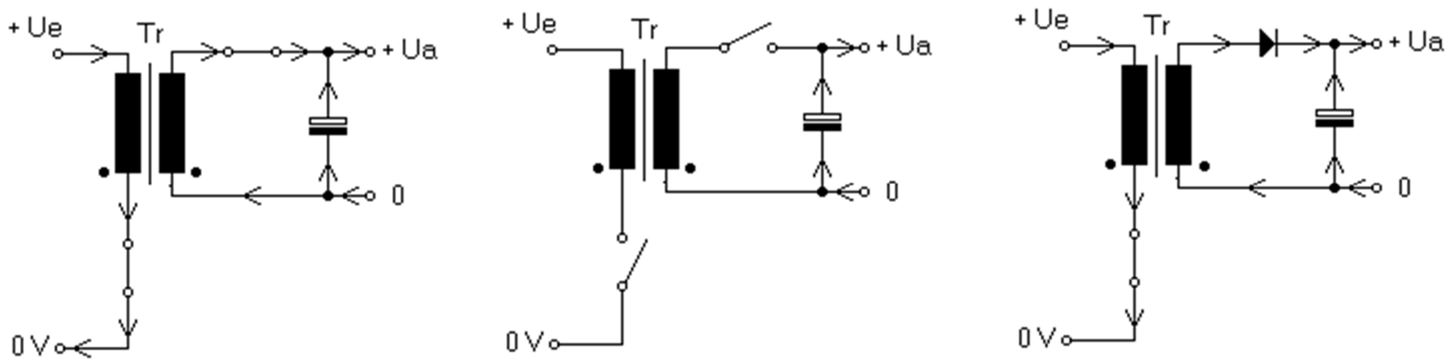


Figure 8.1 A flow blocking phase and the single-ended forward converter

The converter phases in Figure 8.1 A are very similar at first glance which of the flyback converter in Figure 7A. However, the two switches are now working synchronously. Either a current flows through the primary and secondary coil or it does no current flows in the transformer. It should also be noted now that the polarity of the coils is reversed to each other. Just as with the flyback converter is also produced when a single-ended forward converter unwanted stray field that has to be disposed on the primary side with a suitable attenuation element. In addition to the stray field of the flow transducer must be placed in the core stored energy but still. By avoiding an air gap, and use of materials highly permeable (soft iron or ferrite) but this energy can be minimized.
A forward converter can be operated self-oscillating or with a fixed frequency. When the self-oscillating converter of the end phase flow can not be controlled by the amount of the primary current, since it is directly a function of load. It is common for small transducers, easy to drive the transformer into saturation. Since a forward converter transformer usually has no air gap, the inductance is very large, and it flows until immediately before saturation of only a low primary current. The rapid increase in current also leads to the saturation of the switching transistor and initiates the stop phase.

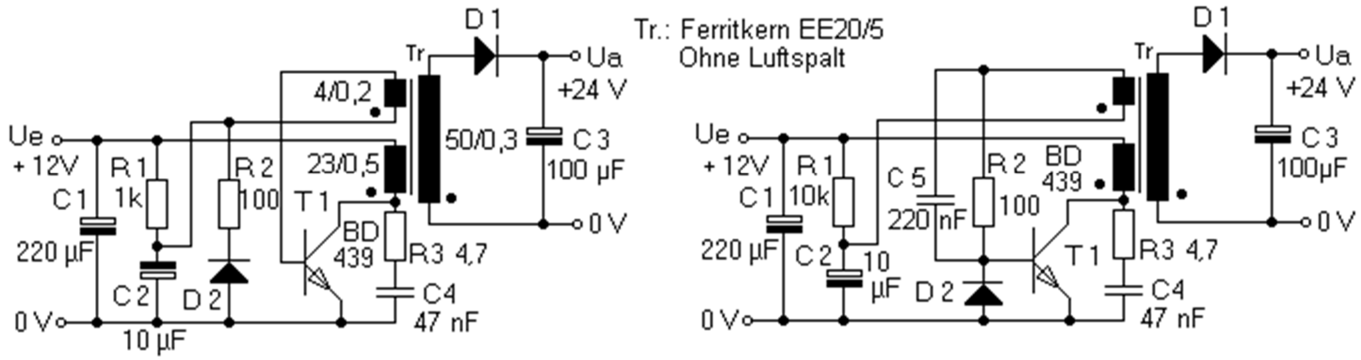


Figure 8.1 B Unregulated self-oscillating flow converter for low power

At first glance no different from the flyback converters from Figure 7.1 B is shown in Figure 8.1 B. The main difference lies in the transformer. First, the secondary coil is of different polarity and second transformer core has no air gap. The coils may remain unchanged in principle. In contrast to the nearly identical flyback converters, the output voltage is relatively stable under load for these flow transducers and idle. Simultaneously, a secondary-side load increases the power primary side. The converter accepts without regulation on only as much power as he just needs to set the output voltage to maintain. These converters are around 8 watts output power and an efficiency of about 80% can be achieved.
Even with a MOSFET can build a self-oscillating flow transducer. This then is largely identical to the regulated flyback converter of Figure 7.2 B, the regulation is not necessary.

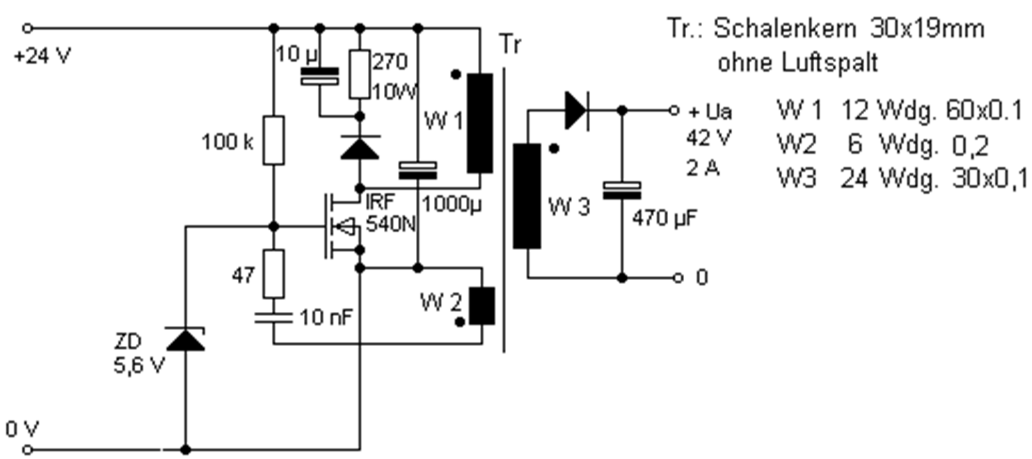


Figure 8.1 C. The self-oscillating forward converter with MOSFET

As seen in Figure 8.1 C, the converter is very simple. A zener diode limits the gate-source voltage to 5.6 volts, so that the FET having a fairly defined saturation behavior. If the transformer core becomes saturated at the end of the flow phase, therefore may increase the drain-source voltage and the lock phase can be initiated.With this converter can achieve around 80 watts of output power at an efficiency of 70-80%.
Of course you can operate the single-ended forward converter with a fixed frequency. This requires a clock for the NE offering 555th In Figure 8.1 D I recorded this converter. The clock is taken directly from Figure 5.1 B and produces a symmetrical square wave. Thus, the converter operates with a fixed duty cycle of about 50%.

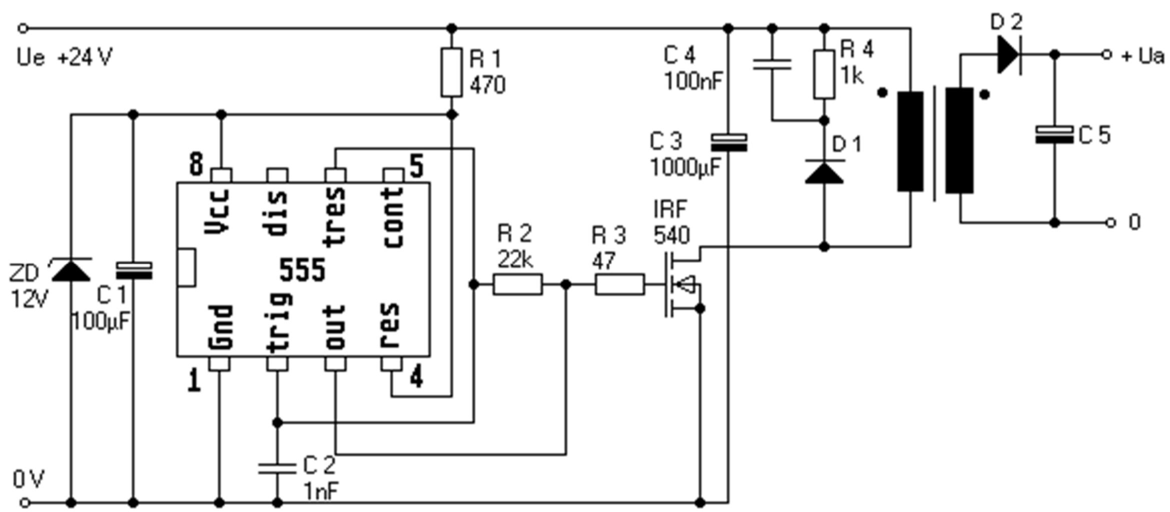


Figure 8.1 Unregulated D fixed-frequency forward converter for low operating voltages

With this device performances can be converted to about 100 watts, but I would recommend a push-pull converter from section 8.3 at powers above 50 Watts.

8.2 Regulated single-ended forward converter

To make an adjustable forward converter is an additional choke needed. Worth the effort really only at higher powers, such as in switching power supplies.

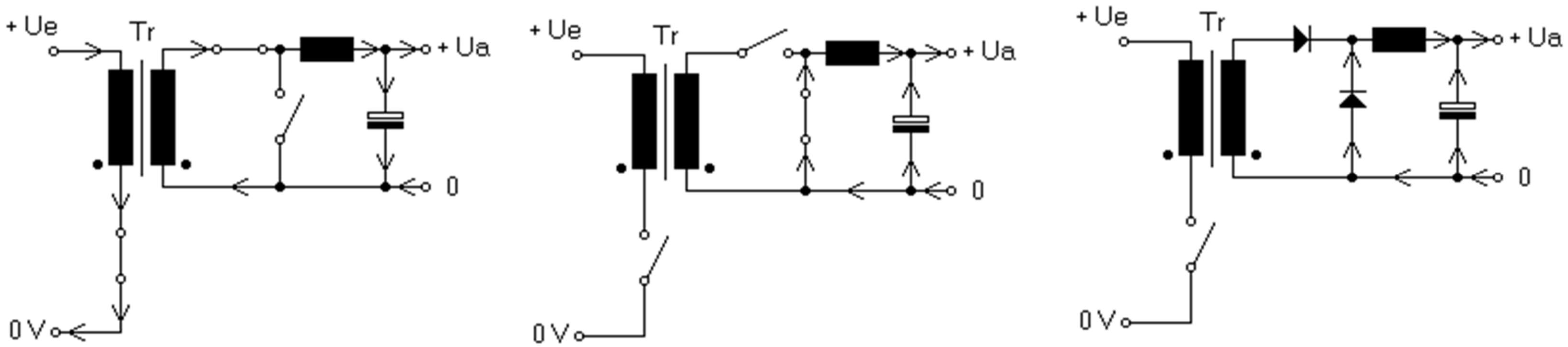


Figure 8.2 A flow-blocking phase and the single-ended forward converter controllable

The adjustable single-ended forward converter operates similar to a step-down transformer. In Figure 8.2 A, the phases of the converter are presented. During the flux phase, the unregulated secondary voltage is switched to the output voltage via a choke. As the secondary voltage is substantially higher than the output voltage, while the throttle is "charged". During the blocking phase of the reactor is separated from the transformer and instead connected to the negative terminal of the output voltage, so that they can give to the output their stored energy. Just as with the step-down transformer, the output voltage is determined by the duty cycle and the height of the secondary voltage. Since the transformer is min. Should allow 50% of the period for demagnetization, the upper limit of the duty cycle is 50%, ie, the secondary voltage must be at least twice as large as the intended output voltage. In order to have control reserves, but you should at least take into account the factor of 3.
Conveniently, the two secondary-side switches can be replaced by back diodes, so that controlling the duty cycle of the output voltage of the primary-side switch is possible.
As regulated single-ended forward converters are nowadays used not so often, there is no standard for this control IC. However, ICs for flyback converter or normal PWM controller ICs can be used. Commonly used ICs, which already have a built-in pulse width limitation of 50%. Would be used as the UC 3844th This is not as common as the 3842, but is largely identical to this one.

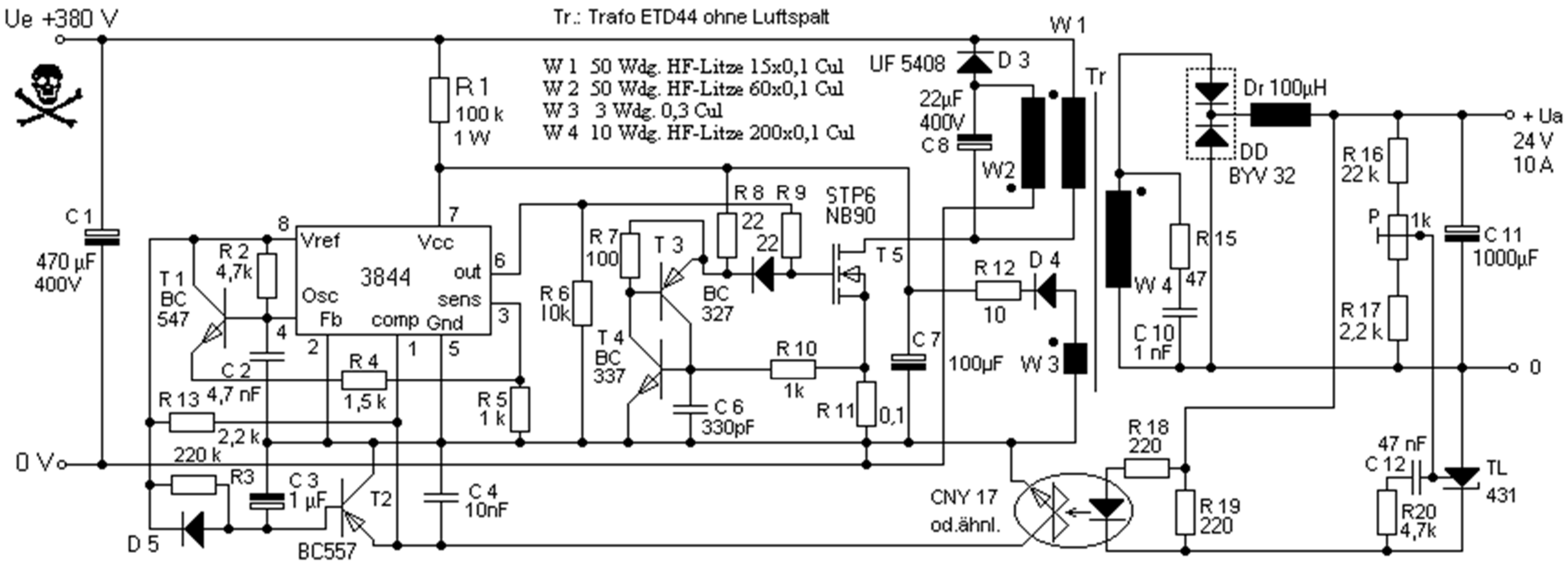


Figure 8.2 B-controlled single-ended forward converter power supply

The 3844 additionally contains a flip-flop, the control signal for the switching transistor suppresses every other clock oscillator. Accordingly, the switching frequency is only half the oscillator frequency (pin 4). The simplest form of a single-ended forward converter, power supply, I recorded in Figure 8.2 B. Thus, the 3844 can operate in PWM mode, he gets to pin 3, a "artificially" generated ramp voltage which is taken from the oscillator to pin 4 and coupled with the high impedance emitter follower T1. R 4 and R 5 divide the oscillator voltage down to the usual amplitude at pin 3. The duty cycle can be between 0 and nearly 50% are. Just like the 3842 and the 3844 has a soft-start circuit that turns off all internal functions of the IC so that C 7 can charge via R1 undisturbed until the starting voltage of about 16 volts.While usually anfährt flyback converter with maximum power, which the 3844 would also do, you should start a little more cautious in forward converters, as otherwise it could lead to congestion because of the still uncharged filter capacitors on the secondary side. Therefore, the circuit is also provided with a soft starter.Once the 3844 starts to operate, is connected to pin 8 of the reference voltage of 5V. Since C 3 is still uncharged, T 2 and shorts on by the control input of the IC pin 1 to ground. C 3 then charges via R3 until the output pin 6 an ever-widening pulse arises. The soft-start should not take too long, so that the converter starts to work before the voltage at C 7 dropped to the point that the 3844 off again. If the scheme of the optocoupler used, the voltage on pin 1 and the pulse width is smaller again. C 3 charges but further down to 5 volts and has no more influence on the control.

Since the primary current is not controlled by the IC, an additional protective circuit has been built. Once at R 11 is a voltage greater than 0.6 volt drop, which corresponds to a primary current of 6 amps, fires the thyristor replica T3/T4, which short-circuits the gate voltage of T 1 over D 2 immediately. Normally, T 3 and T 4 would lock again at the next gate control pulse. However, since too high a primary current indicates a serious problem, the power supply should be switched off

quickly. Through resistor R 8, a flow of the operating voltage of the IC can be connected through the T 3 and T 4 to C 7 discharged flow. The starting current through R1, which continues to flow, but not enough to make T 3 and T 4 is turned on. The holding current is set high with R 7 that T 3 and T 4 block as soon as C 7 unloaded. Then starts a new start. R 7 is selected to be higher, eg 1 ohm, the protection circuit will remain active until the unit for some time has been disconnected. The auxiliary winding W 3 of the transformer shall be such that 7 sets an operating voltage of 12-16 volts in normal operation on C. Since the coil in this example has only three turns, fine adjustment may be somewhat difficult. In this case, the winding number is incremented by one, and R 12 is adjusted so that the voltage is correct again.
A special feature is the demagnetization, which I'll talk more due to their fundamental importance in Chapter 9. The special feature is that the energy stored in the core and stray field energy is not converted into an attenuator into heat, but will be returned via D 3 of the operating voltage. The effort for larger power supplies, especially in forward converters is always worthwhile.

9 Stray field disposal at blocking and flow transducers

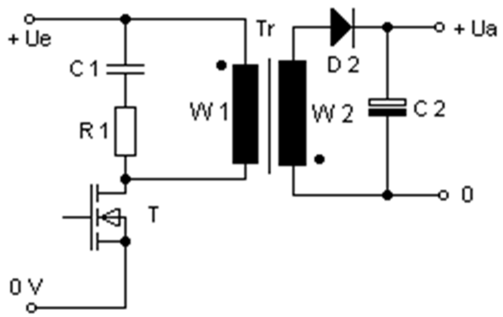
In flyback converters, forward converters and single-ended push-pull parallel power flow converters (one coil per transistor) is the processing of data stored in a significant stray field energy problem. For single-ended forward converters there is also the better the recovery or disposal of the magnetization energy. Because this subject is so important, I have devoted a separate chapter to the. I want to divide the techniques into two main groups:

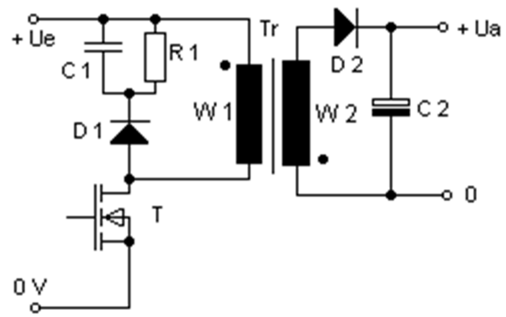
1. Thermal field disposal. The field energy is converted into heat is not required.
2. Energy feedback. The excess energy is fed back to the supply voltage.

9.1 Thermal stray field Disposal

The thermal disposal of surplus energy field circuitry is the easiest solution. It is mainly used for small transformers services, where the energy loss does not fall so much weight.
Basically three variations are common, which I recorded in Figure 9.1 AC. In Figure 9.1 A, the simplest version is to be seen with an RC attenuator, which usually also has the worst efficiency and is most difficult to predict. In addition to accumulating scattering power of the capacitor C 1 for each period must be each time charged to the primary voltage Ue and the transformed to the primary side the output voltage Vo '= Vo W1/W2 with W 1 = number of turns of the primary coil, and W 2 = number of turns of secondary coil. The measure of energy lost is $1/2\ C_l\ U_e{}^2 + 1/2\ C_l\ U_a{}'^2$ per period. The only way to influence the way that losses are reduced, is to let C 1 as small as possible. After turning off the transistor, the voltage at the transistor must be at least Ue Ua + 'to increase. This then adds still induced in the leakage inductance voltage. The leakage inductance Ls then together with C 1 a resonant circuit that can swing more freely now and the voltage adds up to Vin + Vout '. The resistor R 1 to dampen relatively high frequency oscillation as soon as possible to prevent high-frequency noise radiation. The optimal sizing is not so easy to define. On the one hand the capacitor as small as possible, so that the losses are low, on the other hand, the induction voltage appearing at the transistor the larger the smaller capacitor. Especially when you are MOSFETs with high blocking voltage and also low on-resistance and high current carrying capacity of a. It is therefore necessary to find a compromise, which of course you must first minimize the leakage inductance by a good magnetic coupling between the coils. Is to be

used, which is in fact often the case eg with a flyback power supply, a 600-volt transistor, one is initially based on a maximum power of 400 volts DC. In this case the 600 volts should not be exceeded. Now relies on Ua 'and the induced voltage in the leakage inductance with 100 volts, the frame is already exhausted. Is known, the maximum coil current Imax, which occurs immediately before turning off the transistor. Hence the energy stored in the leakage inductance results to $W_s = 1/2$ $L_s I^2_{max}$. In extreme cases, the entire energy of the scattered field can migrate into the capacitor C first Applies to the energy in the capacitor $W_c = 1/2$ $C_l U_c^2$. With complete energy transfer $L_s I_{max} = {}^2 C_l U_c^2$ is . Reference potential for the calculation of the energy in the capacitor is the plateau voltage Vin + Vout ', is added to this voltage, since the induced voltage of the leakage inductance. Uc in this example is therefore 100 volts to accept. Since now all the other sizes are known, one can calculate C 1, by dissolving the formula for C 1-$C_l = L_s I_{max}^2 / {}_c^2 U$. For the value of R 1, there is an upper limit which is determined by the resistance of the resonant tank circuit Ls/C1. At $R_l = \sqrt{(L_s / C_l)}$ the resonance circuit is critically damped, and a further increase of R 1 takes place a higher damping only a higher voltage at the transistor. If it is determined after the calculation of R 1, that after switching off the transistor, the coil current causes a significant voltage drop across the resistor, R 1 should rather be somewhat smaller. Although the voltage swings from then on some periods, but this is not as bad as too high a voltage on the transistor. Much better and more common, the circuit in Figure 9.1 as it has the advantage that it can be optimized in the negative edge and no losses during the flow phase produces the desired efficiency at full load, this circuit can be relatively easy to calculate. To do this select C 1 so large that its voltage does not change significantly within a period, ie, $C_l R_l$ »1 / f .





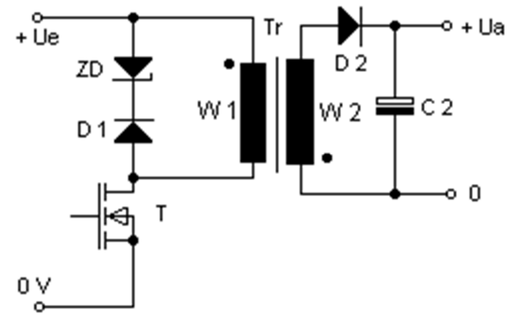


Figure 9.1 A

Figure 9.1 B

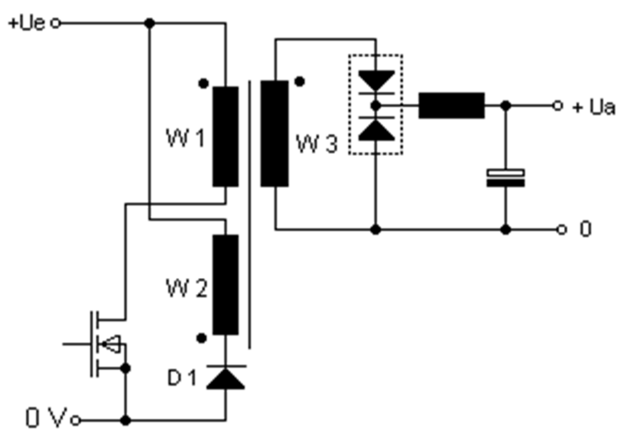
Figure 9.1 C

Three common variants of the thermal scattering field Disposal

From the calculation of the converter to the switching frequency f and the coil current Imax know how high to full load. If the leakage inductance is known, the stray field energy is obtained for $W_s = 1/2\ L_s I_{max}^2$ and thus to be disposed of from the stray field performance to $P_s = 1/2\ f\ L_s I_{max}^2$. If the minimum appearing at the transistor voltage (Vin + Vout approx ') be R 1 can be sized so that it converts a multiple of that power when the voltage Ua him' present. Since Ua present 'always on R 1, this power is converted even at idle. The efficiency can be improved if one allows for a higher voltage at R 1 and C first This R 1 is selected larger, so that the power loss is particularly low at low load. At full load, the voltage then rises over Ua, up to R1, the total scattered field performance can be implemented.The power is converted into R1 are added together due to the induced voltage of the inductance, but still much larger than the stray field performance. What is the optimal sizing, so also depends on the particular application. Even easier sizing in Figure 9.1 C. The voltage spike is simply trapped with a zener diode, or better a.The zener voltage is slightly larger than Ua 'is chosen so that the diode is not conducting the regular induction voltage of the primary side inductance. The necessary power dissipation of the zener diode is always significantly greater than the fringe field power at full load. Exactly like the picture 9.1.B RCD network is the fact that to the induced voltage of the leakage inductance still adds the voltage of the inductance. The advantage of the Zener diode, however, is that at low load only little power needs to be implemented in the Zener diode. As the zener diode is reverse-biased, the diode D 1 is not to be connected in series. She holds during the flow phase away the input voltage Vin of the zener diode, since Ue would otherwise be short-circuited by the zener diode connected in the forward direction.

9.2 Regenerative stray field Disposal

In order to improve the efficiency and to avoid cooling problems, it is particularly at higher powers attempt to return the energy stored in the stray field. There are various options that are naturally more expensive than the thermal disposal techniques. Most of them I have already anticipated in the corresponding transducers. It is best, of course, if the conversion principle already a return "free" supplies. Unfortunately, this is only the case for half and full bridge circuits.



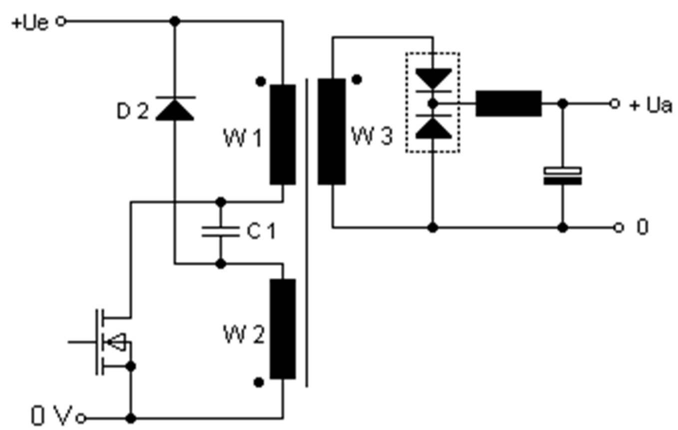


Figure 9.2 A

Figure 9.2 B

Return of the magnetization and stray field energy

As such transducers are generally used only at higher powers, an additional circuit must return the stray field and possibly the magnetization energy (with flow transducers) with smaller transducer performance. In Figure 9.2 A simple execution of a return of the magnetization energy is to see how it is used in single-ended forward converters. During the blocking phase of the power W of the degaussing coil 2 is fed back to the supply voltage. Since the magnetic coupling of W 1 and W 2 are not ideal, however, still remains a stray field that must be disposed of directly to W 1st It is therefore additionally one of the circuits from the images required 9.1. Better the circuit is shown in Figure 9.2 B. The windings W 1 and W 2 have exactly the same number of turns. The coils are connected so that the signals are exactly in phase at the points where the capacitor C 1 is connected. About C 1 then they are firmly linked together. The coupling capacitor C 1 closes, so to speak, the leakage inductance between W 1 and W 2 are short. Now, both the magnetizing energy as well as the stray field energy are returned directly via the diode D 1 of the supply voltage. Since C 1 is connected in parallel to the leakage inductance in the equivalent circuit, it can theoretically be undesirable resonant vibrations. This is prevented by C 1 is chosen such that the resonance frequency of said combination is far below the switching frequency. The same problem exists with push-pull converters with parallel feed, ie with separate coils for each transistor.

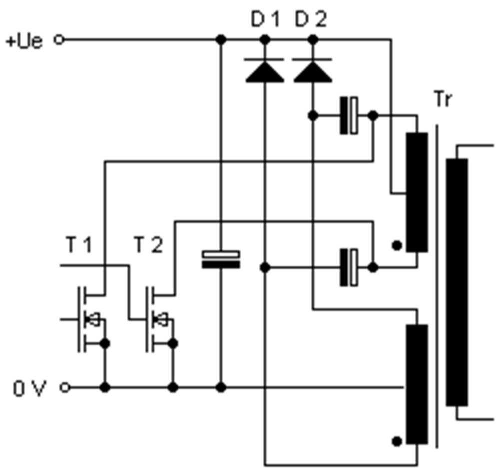
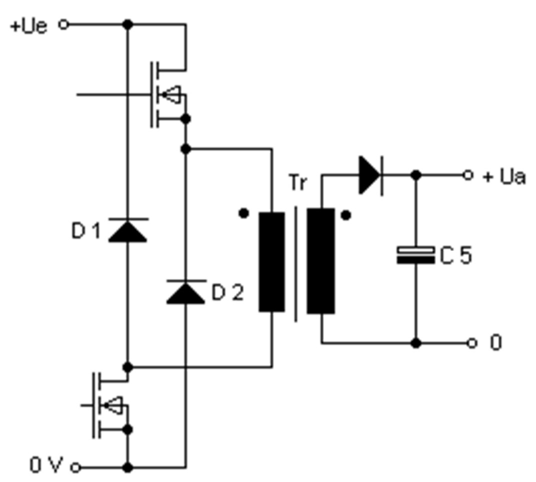


Figure 9.2 C Energy recycling in the push-pull converter

As seen in Figure 9.2 C, the principle is exactly the same. Everything is symmetrical, built or twice. The energy returns so far described all have the disadvantage that the transformer requires a relatively large windings and is therefore quite expensive to manufacture and also the available winding space is not optimally utilized. Even more interesting is a technique in which no additional coil is required for recycling. This is possible with a bridge circuit. As seen in Figure 9.2 D / E, the bridge consists of two diodes and transistors that sit across diagonally. Both transistors are turned on synchronously in the river stage and off line again during the blocking phase. A transistor defines the primary coil can be connected to ground and the other at the other end at the same operating voltage. In the blocking stage, the polarity of the voltage in the primary coil may be reversed, and the energy in the transformer via the diodes D 1 and D 2, the operation voltage can be reduced. The transistors need only the simple operation voltage tolerated, has a favorable effect in MOSFETs on the current carrying capacity of the usable types.
A disadvantage of the bridge circuit is again the difficult controllability of the transistor in the upper bridge arm. Suitable control circuits with and without transformer I have already presented to suffice. An interesting variation of the control is shown in Figure 9.2 e. The control signal for T 1 can be taken directly from

the converter transformer. When T 2 is controlled by the source voltage decreases T 1 R 1 via the gate of T 1 is now brought to a bias voltage, so that T 1 begins to conduct. If, on T 2, this causes a change in current and voltage in the primary coil and the auxiliary winding. About R 3 and C 1 T 1 is then properly turned on or off depending on the current phase. The startup resistor R 1 is designed for operating voltages of over 100 volts and must be reduced accordingly, so that T 1 gets enough gate voltage at lower operating voltages. Normally, R 2 can be omitted even. Then R 1 is also not needed to be adjusted to the operating voltage. The voltage across the auxiliary winding should be about 20 volts.



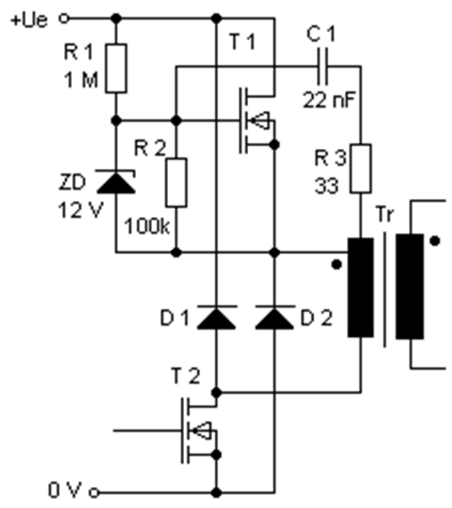


Figure 9.2 D

Figure 9.2 E

Energy return with a bridge circuit

A one of the circuits used for the recovery of the stray field energy is to be noted that the induced voltage during the blocking phase can never be greater than the operating voltage. Therefore, the duty cycle of the switching transistor can never exceed 50%. Otherwise, a complete demagnetization of the transformer core during the blocking phase is no longer possible and there is a risk that the core is driven into saturation. This can especially happen with single-ended forward converters, since the threat of saturation not announce for cores without air gap. Therefore, only control ICs should be used that limit the duty cycle to 50%. This is eg the UC 3844 and UC 3845th Also, the standard ICs for push-pull control (SG 3524, SG 3525 and TL 494) are suitable if you just use only a push-pull branch.

10th Resonant converter

When resonant converter is for blocking and forward converters to the third major group of primary switched converter with galvanic isolation of input and output voltage. Resonant converter is available in many variants, which I will not go into all of them. First, a distinction is made according to the switching time of the transistors. You try to turn in either voltage or current zero crossing, so that switching losses are minimized and you can work efficiently at high switching frequencies, therefore, even at very high performance yet. Depending on whether one calls the converter ZVS (Zero Voltage Switching) or ZCS (Zero Current Switching). A self-oscillating ZVS converters are all described in Chapter 11 sine oscillators , or as described in Chapter 13.2 Inverter for CCFL tubes . Additional information is found in the Department circuitry / oscillators powerful ZVS Royer oscillators sine and

In Figure 10 A, the basic circuit of a ZCS converter is shown. First, a square-wave voltage is generated at a half-bridge. At the output is a series resonant circuit Lr / Cr, where the actual isolation transformer Tr is also connected in series. The transformer itself works as a forward converter, the output voltage is rectified and filtered directly with the output capacitor Ca. The maximum output current supplied by the converter when the square wave generator to the exact resonance frequency $f = 1/2\pi \sqrt{(LrCr)}$ is tuned. In an output short-circuit, the resonant circuit unchecked would turn, and the square wave generator with a very high current load until it was destroyed, unless protective action is not taken. The diodes D 1 and D 2 limit the voltage at the resonant circuit capacitor Cr and lead to excess energy back to the supply voltage. The voltage limit of Cr also means a current limitation in case of short circuit. The maximum current can flow when the voltage across Cr is about in phase opposition to the excitation voltage of the square wave generator in the case of resonance. If the components tolerate this maximum current continuously, the resonant converter is basically short circuit proof. This is a major advantage of the resonant converter, for the slightly more challenging current limit of half-bridge output stages can complete entfallen.Es it should be noted that there may be a significant phase shift of the output current under overload by the nonlinear distortion caused by D1 and D2, so the switched at a specified time not switch transistors in the current zero crossing. The resonant converter is only so long, short circuit protection, as the transistors' switch hard "tolerate under full load.

If you omit the diodes D 1 and D 2, still a current limitation is possible. By a precise adjustment of the oscillator could achieve that the resonance frequency is never reached exactly. However, this has the disadvantage that you can easily destroy the output stage due to careless attitude and above all to dispense with the low loss switchover in the current zero crossing ..

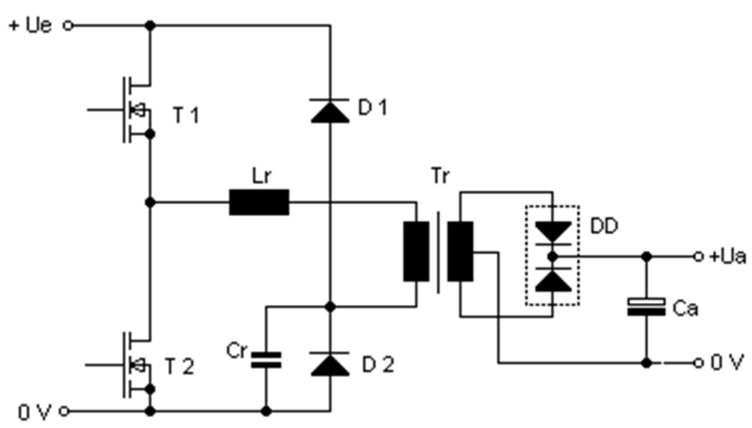


Figure 10 A basic circuit of a resonant converter ZCS

The optimal operating point is reached when the transformer applied straight + / Ue / second The voltage of Cr then swings at full load just between 0 and Ue as it sets idle on a DC value of approximately Ue / second In case of overload, the voltage across Cr beyond the supply voltage attempts to swing out, but this is prevented by the diodes D1 and D2.

As you can see in the block diagram, the coil Lr in series with the isolation transformer. Thus, one can also think of as leakage inductance Lr of the isolation transformer. This means on the one hand, that Lr could be omitted if the transformer had a sufficiently high leakage inductance and on the other hand you need to build the transformers do not pay attention to a low leakage inductance, which makes a good isolation between the primary and secondary coil. Unfortunately, there is not (yet) ready ferrite cores for transformers scattered, so that the separation of Lr and the transformer is likely to continue to be necessary. Another possibility would be to dispense with the closed core transformer completely. Especially at high switching frequencies, you could even build a transformer coils of air, then from the principle already had a high leakage inductance. At what frequency air coil transformer is economically, however can not be said so easily.

The transistors of the half bridge to be ideally controlled so that they are alternately switched for the duration of a half period of the resonance frequency of Lr and Cr. To turn down the resonant converter, only the frequency must be adjusted downward, and the dead time will be extended accordingly. In the following diagram I once represented the progress of the load current and the voltage of Cr under full load and at about 50% load. Between the curves, the gate voltages of the transistors T1 and T2 are shown.

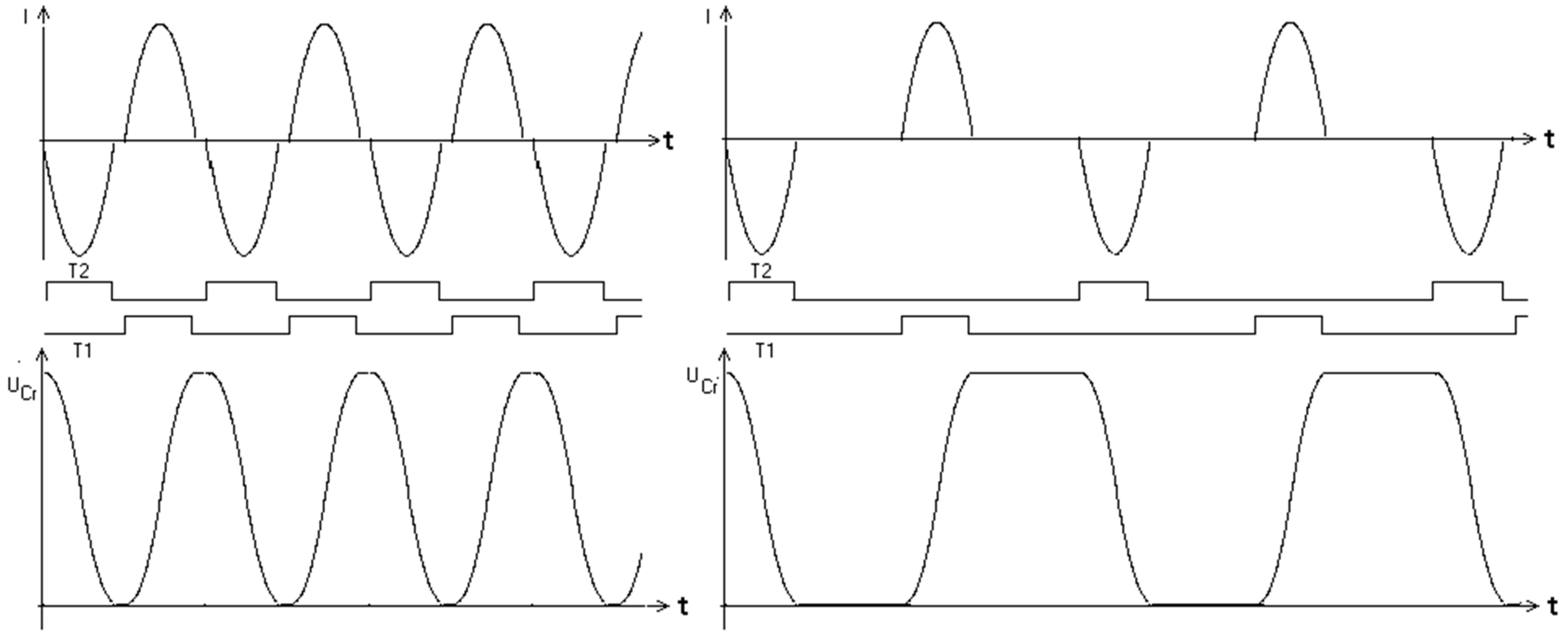


Figure 10 b ideal current and voltage characteristics of ZCS resonant converter under full load (left) and 50% load (right)

At full load, the transistors are controlled by a small dead-time in the push-pull at the resonant frequency of the resonant circuit. Accordingly, sinusoidal voltage waveform is at a Cr (almost) a. The primary-side load voltage should it be about Ue / second Since the transistors are turned on for the duration of a half period of the resonant frequency, the current in the resonant circuit is in the on and off of the transistor just always at the zero crossing, causing the wanted low switching losses.If the performance of the resonant converter can be shut down, just the switching frequency is reduced at a constant duty cycle. In principle, the resonance circuit is then still resonates but the vibration for the period of dead time in which both transistors lock, downright frozen and then continued after the end of the dead time at the same place. It can be seen that the voltage waveform of Cr (bottom right): the sine wave will be frozen for the duration of the peak value of the dead time, wherein Cr stores charge up to the end of the dead time. The switching frequency of the resonant converter can be shut down at low load to 0 Hz.
In principle, it would also be possible to operate the transducer always and only at the resonant frequency to vary the duty cycle. The whole thing would be a kind of push-pull forward converter with passive thermal protection. But you then dispense with the advantage of low switching losses at variable frequency generator. Another advantage you can also use optimal only at variable frequency generator, results from the ZCS technology: It occurred without rapid current changes in the

leads of the transistors. This not only reduces the EMC problem, but also makes the interconnect design much less critical.
Unfortunately, it is not possible with conventional PWM switching regulator ICs to achieve an absolute constant duty cycle. However, since this also is not necessarily so exactly necessary, a tendential synchronization of switching frequency and duty cycle ranges. This can be done easily with a SG 3524. The DC 3524 has the data sheet of the three standard-mode ICs (SG 3524/25 TL494) is the most stable oscillator frequency and is therefore best suited for a resonant converter.

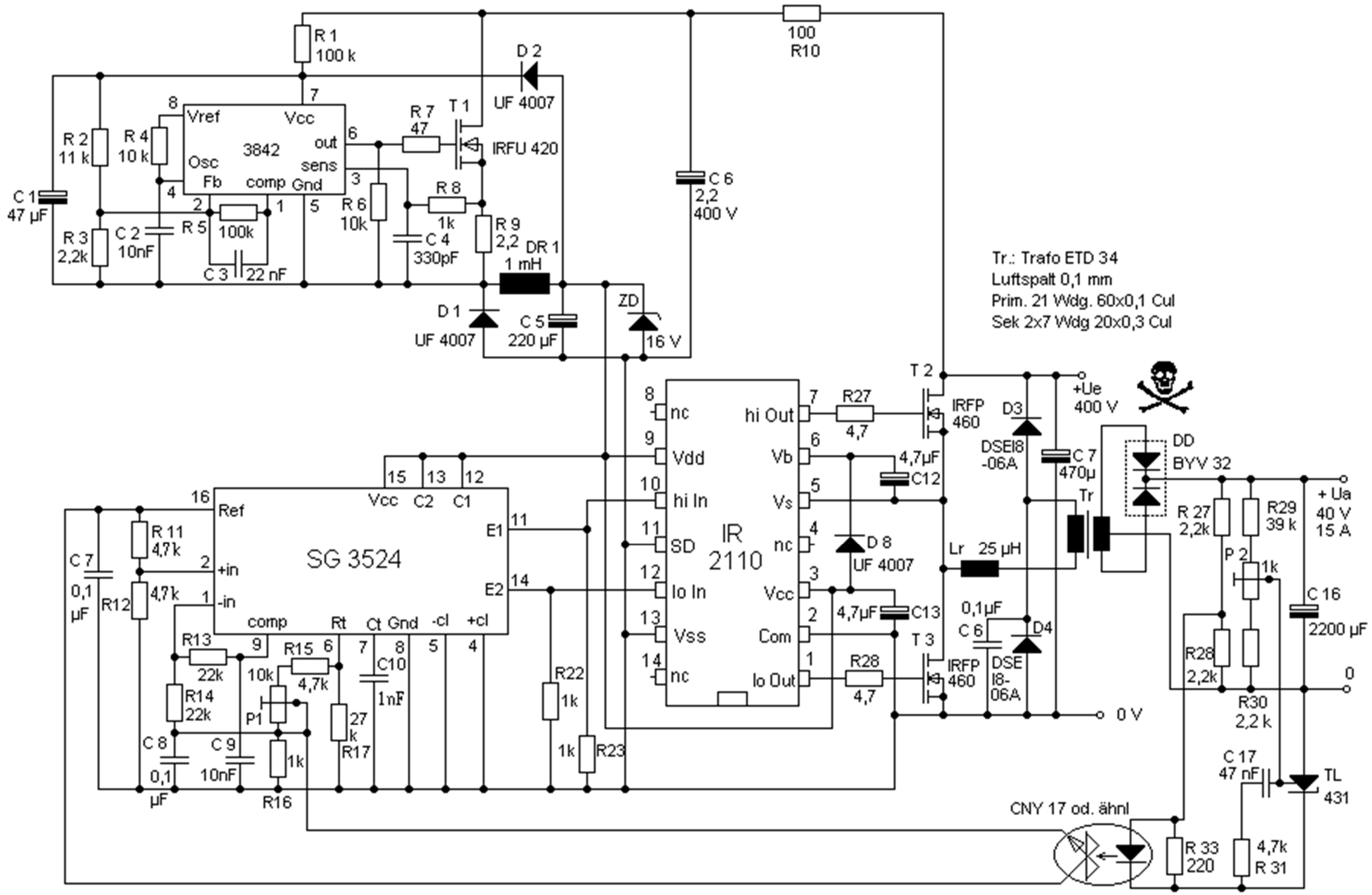


Figure 10 C 600-watt resonant converter for high load control range

Figure 10 shows a C laid out for about 600 watt transformer can be seen. Since 3524 the SG has no strong output driver and therefore can neither control nor control transformers transistors directly, here lends itself to a gate driver IC. The power level and the power supply module is otherwise identical to already described.
The oscillator frequency depends on the SG 3524 capacitor C 10 and the current flowing from the pin 7 (Rt) to ground. In the uncontrolled state, the voltage on C 16 and R 8 is relatively low. The oscillator operates at its maximum frequency, which is set with P 1 on the resonance frequency. The control of the amplifier 3524 is connected as an inverting amplifier as a virtual ground at 2.5 volts. In the uncontrolled case, at the output of the control amplifier (pin 9) of the maximum voltage and the internal PWM comparator generates a maximum duty cycle. When the control starts, a current from the reference voltage of the optical coupler 3524 via 16 flows by R If the voltage is increased to R 16 and C8, the output voltage of the control amplifier 9 pin behaves exactly opposite directions and the duty cycle is reduced accordingly. At the same time also the current from pin 6 (Rt) via R 15 and P 1 is reduced flows because now the difference between the voltage applied to R 15 and P 1 is lower. R 17 is selected so that with decreasing frequency, the first duty absolutely and finally is getting shorter approximately constant until the pulse disappears. Thus, an energy-efficient load balancing to reduce possible to 0% without the switching frequency is audible as annoying whistling. However, the efficiency is poor due to the shortening of the turn.
The resonant frequency is selected to be relatively tall, about 100 kHz. MOSFETs have no problems. If the output of the converter, however, be significantly increased, they will try quick IGBTs, best use with built-FRED. Eventually, the frequency has to be reduced when using powerful IGBTs. This is simply C 10 is increased, thereby simultaneously also increased the minimum dead time (see data sheet). Otherwise, the converter can be dimensioned unchanged for almost any services. Only the power components of the voltages and currents which must withstand.

11th Sinusoidal and trapezoidal converter

A major drawback of most transducers, the steep flanks of the transformer voltages, which cause not only radio interference, but also high switching losses in the transistors. If it is possible to work with (almost) sinusoidal voltages, so you can immediately kill two birds with one stone; sinusoidal voltages have (almost) no harmonics and increase the speeds are moderate. The easiest way to generate sinusoidal voltages, is a parallel resonant circuit of the primary coil and a capacitor.Otherwise, such a converter similar to be built as a self-oscillating flyback converter. It makes sense to a sinewave inverter is self-oscillating, so he always vibrates exactly at the resonant frequency of the resonant circuit. A further advantage of the sine converter and the elimination of the stray field of waste disposal. The stray field energy is again fed back into the resonant circuit. In simple converters can waive the provision itself in network operation, since the parallel resonant circuit stabilized amplitude and frequency. When the sinusoidal waveform distorted by the power supply and / or secondary-side rectifier, or flattened, the voltage waveform looks more trapezoidal. In such cases, one speaks then of more trapezoidal transducers. Common to these types of converters, the switching transistor on and off, while the voltage difference of close to zero at the transistor. Sinusoidal and trapezoidal transducers are therefore to ZVS (zero voltage switching)-resonant converters.
Also sinewave inverter can be built as single-ended or push-pull converter. Single-ended converter can be very simple, but are not as well suited for high performance. In order for the single-ended swing stable, the reactive power in comparison to the output power must be relatively high. At high powers, a cost and space problem can result from this already. In addition, the switching transistor needs to be oversized, so that it is not overloaded during the initial phase, when the

resonant circuit capacitor is still uncharged.
Push-pull sine wave converter, specifically the Royer oscillator can be against it, similar to ZCS resonant converter, any build for high performance.

11.1 single-ended sine wave converter
As with the flyback converters and the low-voltage sinewave inverter is particularly simple. In Figure 11.1 A operated with such a low voltage transformer can be seen. The starting resistor R 1 3 C charges up a base current flows. The transistor then operates in the amplifier mode, and the converter will begin to oscillate due to the feedback. The base-emitter diode of T 1 C load 3 with a negative voltage so that only the very tip of the feedback sine wave voltage can activate T by first This happens precisely at the moment when the collector-emitter voltage is just zero. By the collector current of the sinusoidal voltage at the resonant circuit will be slightly flattened. The higher the saturation current of T 1 with respect to the reactive current of the resonant circuit, the greater flattening of the lower and the higher the voltage peak of the upper half-wave. When sizing is to make sure that this voltage peak does not exceed the maximum reverse voltage of T first The feedback winding should be such that the peak-to-peak value of this induced voltage is a maximum of about 6 volts. Higher values would lead to an unacceptably high base-emitter junction voltage in T first

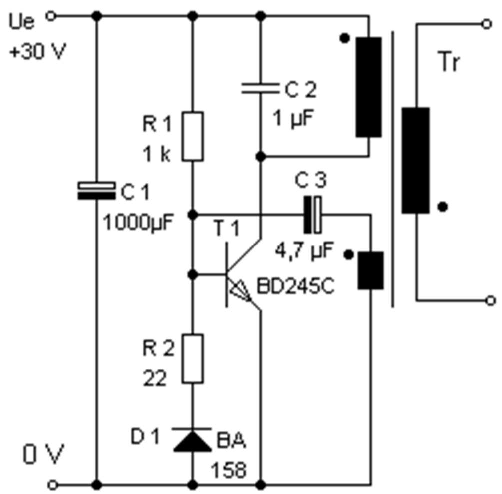


Figure 11.1 A low-voltage sine wave generator unregulated

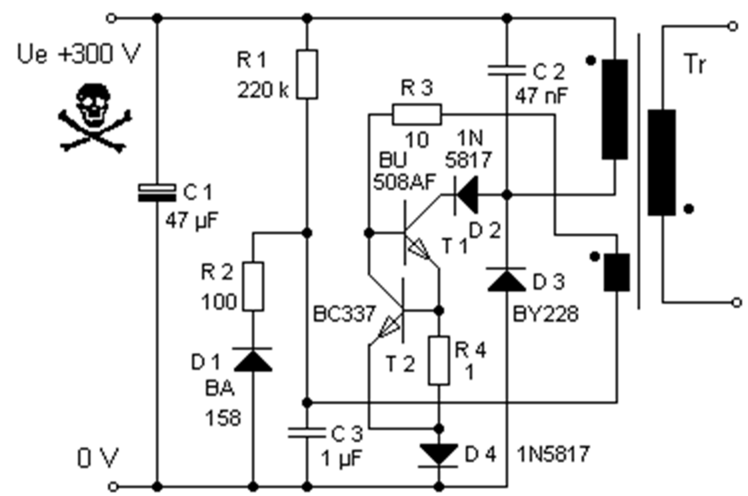


Figure 11.1 B high-voltage sine wave generator unregulated

A little more effort you have to drive to mains powered sine converters to ensure safe operation of the switching transistor. In Figure 11.1 B a sinewave inverter is shown for an operating voltage of about 300 volts. Important is a defined current limit consisting of R 4 and T 2, and responsive at about 0.6 amps. This power, you also need to calculate the maximum collector-emitter voltage. At a relatively undistorted sine it is about 2 Ue. In this case, would be about 600 volts and 1000 volts common transistor would be entirely adequate. This also applies to D 3; here then extends as a BY 399 In highly distorted sine voltage, the voltage can be much higher. To calculate the maximum voltage that can be applied to the conservation of energy. Since the voltage on C_2 does not change the cut-off time of T 1, no charge / discharge current and the collector current flows in the coil flows directly. If T 1 off with a current of Ic = 0.6 A, is located in the resonant circuit, the energy

$W_{tot} = 1/2\, C_2 U_e^2 + 1/2\, LI^2_c$, Where L is the inductance of the primary coil.

The voltage maximum, this energy is entirely in C 2nd According to the energy conservation law $W_{tot} = 1/2\, C_2 U_e^2 + 1/2\, LI_c^2 = 1/2\, CU_c^2 \rightarrow C_2 U_e^2 + LI_c^2 = C_2 U_c^2 \rightarrow U_c = \sqrt{(U_e^2 + I^2_c L / C_2)}$.

To T 1 is the sum of the maximum operating voltage Ue and charging voltage Uc $U_{CE} = U_e + U_c$ to. Is thus the peak voltage at the transistor

$U_{CE} = U_e + \sqrt{(U_e^2 + I^2_c L / C_2)}$. If I now assume L with 1 mH, I'm with the values from Figure 11.1 B to around 613 volts and an oscillation frequency of approximately 23 kHz. The sinusoidal voltage is therefore only slightly distorted. It should be noted that even in the idling in the resonant circuit is an effective reactive current of about 1.5 amps flows. Capacitor and primary coil must be sufficiently resilient. The use of an unregulated area sine converter is very versatile. It ranges from simple auxiliary voltage generator to the electronic ballast for gas discharge lamps. In the latter case, the transformer must have such a high leakage inductance, the transformer is short circuit protected. The auxiliary winding must be located close to the primary coil, so that the converter can swing clean.
T1 turns on so as exactly as possible at the time when the collector voltage is approximately zero, it needs a high basic control voltage from the feedback winding. To be practical, a value of about 30 V_{SS} has been established. However, this means that the base voltage up to - may drop 30 volts. Thus, the maximum base-emitter junction voltage of approximately 5 volts can not be exceeded, the diode D 4 is inserted in the emitter yet. If the base is too negative, and the emitter potential can assume negative values, so that the BE-blocking voltage is not so large.
When the vibration is largely undamped, a reverse current flows, with the part of the stored energy in the tuned circuit goes back into the voltage source during the first half of the T switch of the first In the second half of the switch the current flows in the normal forward direction of T 1 and again performs the previously extracted energy into the resonant circuit. The resonant circuit was taken from elsewhere energy, the reverse flow of current phase is correspondingly shorter. The diode D 3 increases the reverse current to complete.
The diode D in the collector 2 will ensure that during the reverse-phase current, the collector potential does not become negative. This would also pull down the base voltage and disrupt the control signal.
If a sine wave transducers are used as switching power supply, you will usually provide for a system of amplitude. Circuitry, it is constructed similar to a flyback converter. The efficiency is usually much higher than in a comparable flyback converter. The switching losses are reduced because the collector voltage of the switching off of T 1 rises slowly and because they are already zero, when it turns on. The stray field energy remains in the resonant circuit and must not be disposed of. Moreover, the frequency is idling is not high, which significantly reduces power consumption at low load. The amplitude of the sinusoidal voltage can not be less than Ue. This limits the control range unfortunately a bit. The negative half wave always has the constant displacement Ue. The greater the negative half-wave is

flattened, the higher the positive half-wave. The system is therefore only possible, if the half-wave rectified during the blocking phase of T 1 on the secondary side. In Figure 11.1 C, a primary side regulated sinewave inverter can be seen.

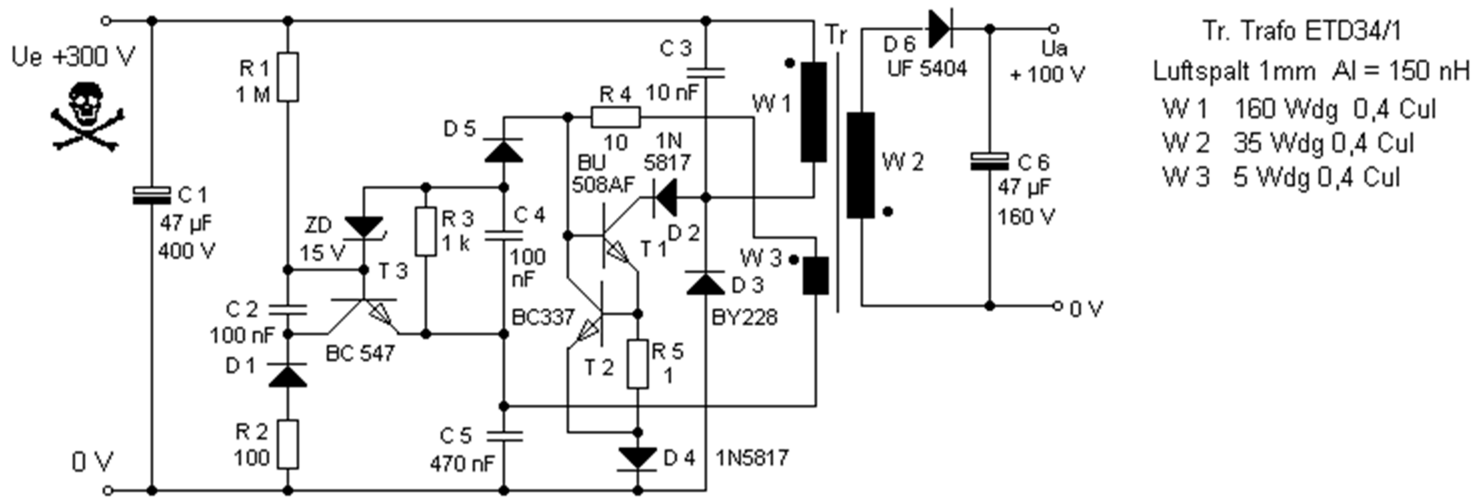


Figure 11.1 C primary side regulated high-voltage sine wave converter

As soon as the operating voltage is Vin, a small current flows through R 1 in the base of T 3 This base current of T 3 then flows through the emitter to C 5 and charges this. Through the feedback winding W 3 of the voltage on C 5 R 4 is applied to the base of transistor T 1 When the base voltage of T 1 is sufficiently high, this starts to conduct and to operate as amplifiers. In this state, then the converter starts to oscillate. By the rectifying action of the base-emitter diode of T 1 C 5 starts to charge negatively. Thus, a significant base current can flow in T 1, the negative charge of C 5 must be counteracted. To a discharge current flows through R 2, D 1 and D 3 Requirement that the current can flow through T 3 is the upcoming base current of R 1.
Representing the output voltage of the negative half-wave of the base control voltage for T 1 of D 5 is rectified and filtered with C 4th When the voltage at C 4, such that the Zener diode ZD begins to conduct the base current of T 3 is branched, and it begins to turn off. This C 5 can be something negative charge, whereby the entire base control signal is slightly negative. In T 1 then flows a lesser base current and the saturation current is also lower. The lower the saturation current, the voltage, the sine is undistorted and the lower the positive half-wave. The amount of the positive half cycle can be controlled. On the secondary side, the positive half-wave rectified and therefore determines the amount of output voltage. With heavy loads, the positive half-wave is flattened by the rectifier diode D 6th The oscillation frequency becomes lower and the sinusoidal voltage degenerate into more of a trapezoid shape. The converter then operates almost like a flyback converter. With the given component values, output power of 40-50 watts can be achieved. If the sine converter can be controlled on the secondary side, just needs of the phototransistor of the optocoupler is connected in parallel with zener diode ZD (collector of the phototransistor in base of T 3). ZD can be omitted in principle, since the output voltage is limited by the current limitation anyway. For the secondary-side control of the optocoupler one of the standard control circuits used already in the previous

power supplies can be used.
Of course you can also build sinewave inverter with MOSFETs. Corresponding basic circuits I have in the circuit design / oscillators described.

An interesting application of the trapezoidal converter consists of a simple DC-DC converter for high input voltages up to 800 V. These voltages such as can occur in the areas of photovoltaics, electric cars and of course in the intermediate circuits of 400 V AC power supplies. Transducers are used there are often high power, but require an auxiliary low power voltage for the control electronics. In this power range are normally used small flyback converter. With input voltages up to 800 V, however, result in significant problems since. First, the procurement of adequate voltage resistance MOSFETs some problems. Although MOSFETs are available for blocking voltages up to 1500V. But these are relatively expensive and may not always be procured at short notice. Furthermore, such a disturbs the high Rdson high voltage MOSFETs. To make matters worse, that non-resonant flyback converter switch. That is, the energy of the parasitic capacitance has to be burned in at each power MOSFET. Since the energy stored in the parasitic capacitances increases with the square of the voltage, the loss component alone at 800 V 6-7 times as high as the operation of normal mains voltage. Remedy a trapezoidal transformer primary side unregulated.

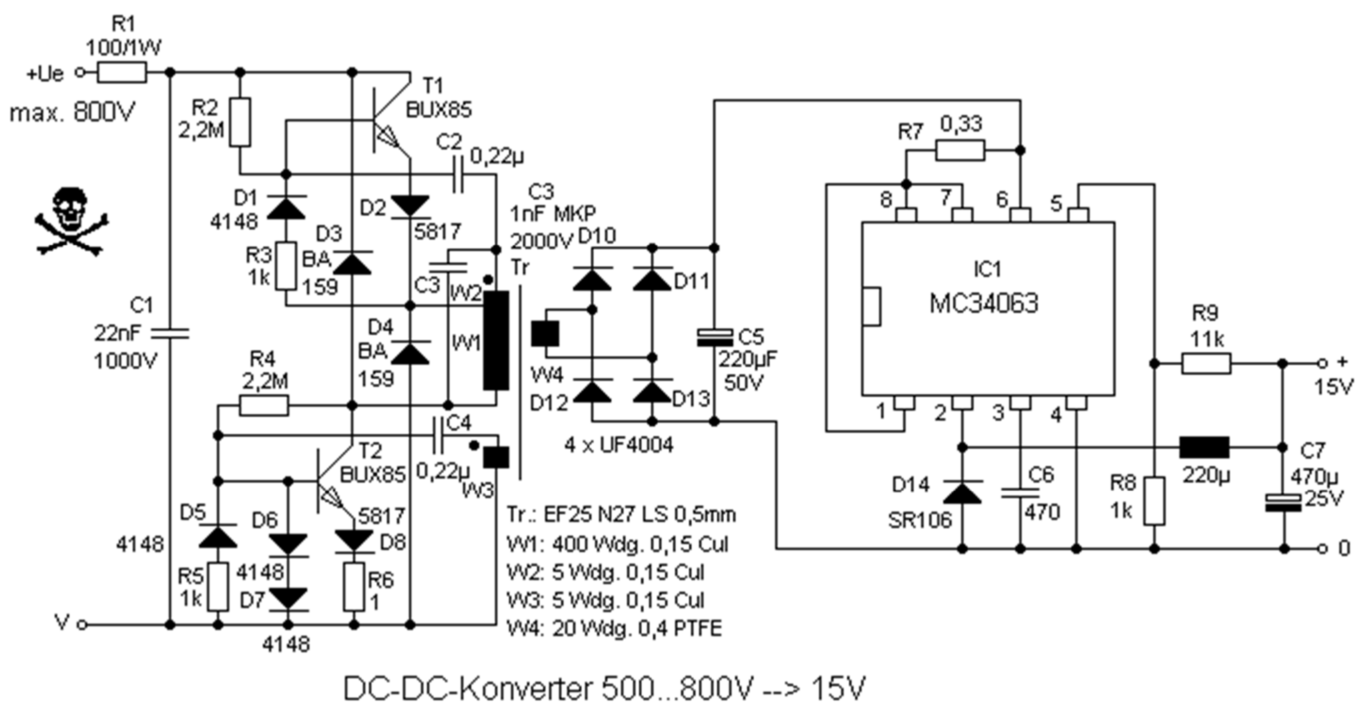


Figure 11.1 D secondary side regulated high-voltage harness converter

In order to build the converter with readily available transistors for blocking voltages up to 1000 V, both coil ends of the primary coil are connected and clamped each with a freewheeling diode to the opposite pole of the operating voltage. This has the advantage that flow and off phase can be used for power transmission in both the secondary coil. A self-oscillating converters are particularly easy to build with bipolar transistors. To this end, the still well-obtain high-voltage switching transistors of the type suitable BUX85. This can add up to 1000 V block, which would allow a theoretical maximum operating voltage of 1000 V in this topology. In practice, however, incorporate a reserve in the interest of long-term reliability and should the converter with max. 800 V to operate. In the initial phase, the transistors are biased by the starting resistances R2 and R4, so that the transistors can go into the amplifier operation. Both transistors have their own feedback winding (W2 and W3), via which the light coupled into the primary coil signal is fed back to the bases. Thus, a vibration in the resonance frequency of the primary oscillating circuit C3 and W1 used in which the transistors in the common mode on and off. After the oscillation build the transistors go into the pure switching operation. The original sine wave is thereby strongly flattened on both sides, so that the typical trapezoidal shape. During the flux phase which is formed by the flattening Stromflus in the switching transistors, and during the locking phase by the free wheeling diodes D3 and D4. The trapezoidal shape of the coil voltage (quasi-) resonant switching of the transistors at low voltage CE is possible. This is achieved by clamping the feedback voltage at the bases of the switching transistors. The BE of the transistors diodes act here as clamping diodes to charge the coupling capacitors C2 and C4 to a DC voltage, which is so large that the transistors conduct precisely when the plateau of the trapezoidal voltage (flow phase) is obtained. Once the keystone is leaving the plateau voltage (start the blocking phase), the transistors are blocked very quickly, while the trapezoidal voltage rises slowly. The transistors, therefore, produce little switching losses, which allows a relatively high efficiency of this circuit.
Since the contact resistance is not sufficient base current supply circuit for the normal operation, the coupling capacitors are charged via the resistors R3 and R5. This load current must flow into the flow bases of the transistors of the stage and serves to clean switching of the transistors. The diodes D1 and D5 permit free of charge by the start-coupling capacitors resistors R2 and R4 during the startup phase. The emitter diodes D2 and D8 have the same function as in other sine converters already described: keep the fairly high negative base voltage of the diode-BE remotely without degrading the switching speed of the transistors.

The lower switching transistor T 2 is additionally provided with a current limiter consisting of D6, D7, and R6. So you can control a controlled end of the flow phase before the onset of core saturation. Secondary side of the voltage is removed using a bridge rectifier. Thus, locking and phase flow can be used for energy transmission. A primary-side regulation of the output voltage is of course not possible. This is done here, for example, a downstream switching regulator type MC34063. The maximum output current of the converter is approximately 500 mA.

Should also be noted that the resistors R2 and R4 are permanently exposed to a voltage of up to 800V. Here you should use special high voltage resistors and at least four standard resistors in series.

Special attention deserves the protection of the circuit. In case of short circuit, the operating voltage must be reliably separated from the circuit, which is not easy at 800 V DC. For simple glass fuses, the wire would burn through, but the current continues to flow through an arc. Depending on the internal resistance of the voltage source can cause devastation caused in the circuit and in the surrounding area with significant risk of injury to persons present. Although the resistance R1 can be used as a fuse resistance, but it does not replace a proper protection of the operating voltage with an appropriate fuse.

12th Passive and active power filter / power factor correction

Power supplies cause high and low-frequency noise, which must be removed with suitable filter circuits. High-frequency interference can be eliminated with good passive LC filters, while one needs active filtering for low power electronics.

12.1 Passive Noise Filter

In particular, Switching Power Supply cause radio interference, which can easily travel through the power supply lines to the grid, where they can be spread over long distances. Of course, such high-frequency interference may not leave the AC adapter. Direct noise radiation prevents a closed metal housing. However, this alone is not enough. The pulsed current drawn by the switching circuit, which is absorbed in part by the Siebelko, the low operating current, a high frequency noise current is superimposed. This noise current can filter out with a LC low-pass. The interference current flows between N and L wire of the mains voltage. In Figure 12.1 is such a universal noise filter to see how it is so similar in part or in simplified form in almost all mains powered devices, are potential sources of interference.

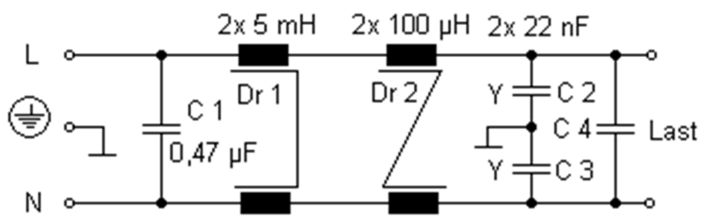


Figure 12.1 Standard Noise Filter for Mains powered devices with about 1 kW

The LC low pass formed by the so-called X-capacitor C 1 and the throttle Dr second For the noise current of the two coils are connected in series and Dr 2 form, as they are wound on a common core, a total inductance of 400 Fixed Inductors. Frequently found in filter circuits instead of a double or twin choke chokes, which are cheaper and easier to obtain. A single reactor has the disadvantage that the filter is asymmetrical and symmetrical interference voltages can pass to the outside may asymmetrically. With twin throttles the symmetry is indeed given, but it is altogether more copper wire needed, which is disadvantageous for compact filters.Since the load current magnetises the core of Dr 2, may be used herein high permeability material without any air gap, in order to obtain a large inductance.Commonly used to niedrigpermeable powder toroids. The cut-off frequency which corresponds approximately to the resonant frequency of C 1 and Dr. 2 (400µH), should be well below the switching frequency being used. Dr 1 is a current-compensated choke, the coils are connected so that the load current is opposite to canceling magnetic fields generated. For the load current 1 Dr has therefore at least theoretically no inductance. Remains practically left a residual leakage inductance, which is on the order of about 1/100 of the inductance of a coil (depending on design). The basis for calculating the cutoff frequency inductance therefore increased slightly from 400 to 450 Fixed Inductors Fixed Inductors.

In addition to the symmetrical interference voltages that are superimposed on the load current, there is still asymmetric disturbances which are in common to both L- and on the neutral conductor. For such voltages, the current-compensated choke Dr 1 has its full inductance of 5 mH. The high inductance of Dr 1 represents an insurmountable barrier for asymmetrical interference voltages dar. Dr 1 is not replaceable by twin reactors. In order to achieve high inductance coils are insulated

from each wound on a high permeability ferrite core without any air gap. Thus, the core does not come by the load current in the saturation, it flows in opposite directions through the two coils. Thus, the high inductance of Dr 1 is maintained at least for the asymmetrical interference voltages.
The so-called Y-capacitors C 2 and C 3 are between housing and the L and N. Asymmetric frequency interference voltages are shorted so directly against the housing. With 22 nF, the upper limit has already been reached, which may have these capacitors. Common values are 2.2 to 4.7 nF. Since in normal operation one of the leaders, in the worst case even both conductors are located on 230 Volts AC, flows through the Y-capacitors, a residual current of about 3 mA. Much larger Y-capacitors would already cause dangerous high fault currents, quite apart from the fact that adding the residual currents at multiple connected devices and would trigger the GFCI. For devices with protection insulation, the Y capacitors, if any, associated with the device ground. Since the Y-capacitors, the power supply separate from the low-voltage range, they must meet very high safety requirements. Suitable capacitors are marked with the appropriate safety symbols (VDE).
The capacitor C 4 is optional and usually not additionally necessary because anyway, a capacitor or electrolytic capacitor in the power supply itself is parallel to the line voltage and already captures the worst.
On the secondary side, it is usually enough if the operating voltage ground to the shield, and the metal casing of the power supply is connected.
The fuse for securing the device to the mains side is sensibly in front of the filter. This ensures that even the X capacitor is protected in case of error. Furthermore, you can switch a ZnO varistor (surge) parallel to N and L wire on the load side yet. This can at least absorb short overvoltage spikes.
The values given in Figure 12.1 refer to a device with about 1 kW. In individual cases, the dimensioning depends not only on the performance but also on the type of noise source. Usually you will first fall back on standard Suppression, it staggered for mains voltage, for electricity, to buy as individual components or as a complete equal encapsulated filter modules. For discrete filters, the appropriate values then need to be determined empirically.
If in addition to the high-frequency noise components and low frequency harmonics caused mainly by rectifier circuits are passively filtered out, one needs very large chokes. A disadvantage is also that the output DC voltage is strongly load dependent. Usually content themselves with relatively small chokes that iron out the current spike at the apex somewhat.

12.2 Active power filter / power factor correction

If an AC power grid removed, a portion is withdrawn as active and the other part as a reactive power. Although the reactive power is fed back to the network and ignored by the electricity meter, but causes additional power losses at the expense of the utility. Therefore, it is often mandatory, especially at high power levels, the reactive power is minimal. With inductive or capacitive loads, the reactive power only causes a phase shift between voltage and current. For loads with inductive reactive power, most typically, the reactive power can be eliminated by a compensation capacitor. Conversely, power can be blinded by a free capacitive loads connected in parallel coil. The goal is always that the voltage and current are in phase, ie, a consumer behaves like an ohmic resistance.
In power supplies now there is a case that is strongly distorted with a bridge rectifier with a downstream Siebelko the current, ie, he is no longer sinusoidal (see Figure 2.2). In this case, the reactive power can not be easily compensated with a capacitor or a coil. A passive LC filter that filters out the fundamental clean, would certainly be much larger, heavier and more expensive than the actual power supply. In newer and especially larger power supplies (> 100 watts) one finds therefore a so-called active power filter for power factor correction (PFC Data Sheet Power Factor Correction). In principle, these are simple boost converter, as I have already described in chapter 6.2. The converters are supplied with the rectified but unfiltered mains voltage and convert it to approximately 400 volts (at 230 V AC).The control of the converter now ensures firstly that taken from the mains instantaneous current is proportional to the instantaneous voltage and secondly it regulates the

rms value of the current is so high that the network exactly the required power is removed to an average output voltage of approximately 400 volts to receive. The up-converter operates at a switching frequency well above the power supply frequency. Therefore, the switching frequency is relatively easy with a passive LC filter (see Figure 12.1) away from the mains. A in Figure 12.2 shows the block diagram of a conventional power factor correction can be seen. As you already can see at first glance, is not so simple, though only the most important functions are located. First, the grid voltage is applied to a rectifier bridge and from there, to the capacitor C 1, replacing, in principle, the capacitor C 4 in the line filter of Figure 12.1. Then follows a normal up-converter consisting of the storage inductor Dr, the switch T, the diode D and the second Ausgangssiebelko C Thus, the converter can operate clean, the output voltage must be significantly higher than the peak value of the mains voltage. Very popular are the output voltages to 400 volts, for you then filter capacitors with min. 450 volt needs. Since it is sometimes difficult to get large electrolytic capacitors with 400 volt dielectric strength, occasionally trying to put on about 380 volts and the output voltage. With an rms value of the mains voltage from 230 volts to get to a peak of about 325 volts. That is very scarce, exposing eg at elevated supply voltage, but just about acceptable, especially with an increase in peak voltage exceeds the setpoint, the output voltage, only the filter function more or less. The actual function of the overall device is not affected. Is critical in this circuit, the switch-on. Since the boost converter, the output voltage can not be lower than the input voltage, the initially uncharged Capacitor C 2 is charged and about Dr D on power with a high charging current. Just as with conventional power rectifiers therefore also a limitation must be provided. Also, make sure that the converter does not start, as long as the choke core is still saturated by the inrush current. Often the throttle and the diode D by an additional power diode of C 1 to C 2 is bridged. In order to start the converter, a starting pulse is required. The can for example generate a watchdog timer that is not shown here. If the converter suspends some time, the RS flip-flop is set and T connected through. To Dr then lies on the instantaneous value of the AC voltage. The inductor current flows through R 3, where it causes a linearly increasing voltage. Once the voltage across R 3 exceeds the multiplier output voltage, the comparator sets the flip-flop back and Comp 2 blocks T again. While T blocks, the current continues to flow through DR and D and C 2 to loads. When the magnetic field is broken down into Dr, also breaks the tension together. To recognize this time is still an auxiliary winding on the throttle. Following the collapse of the induced voltage in Dr Comp 1 detects a zero crossing in the auxiliary winding, and sets the flip-flop, so that a new cycle can begin.

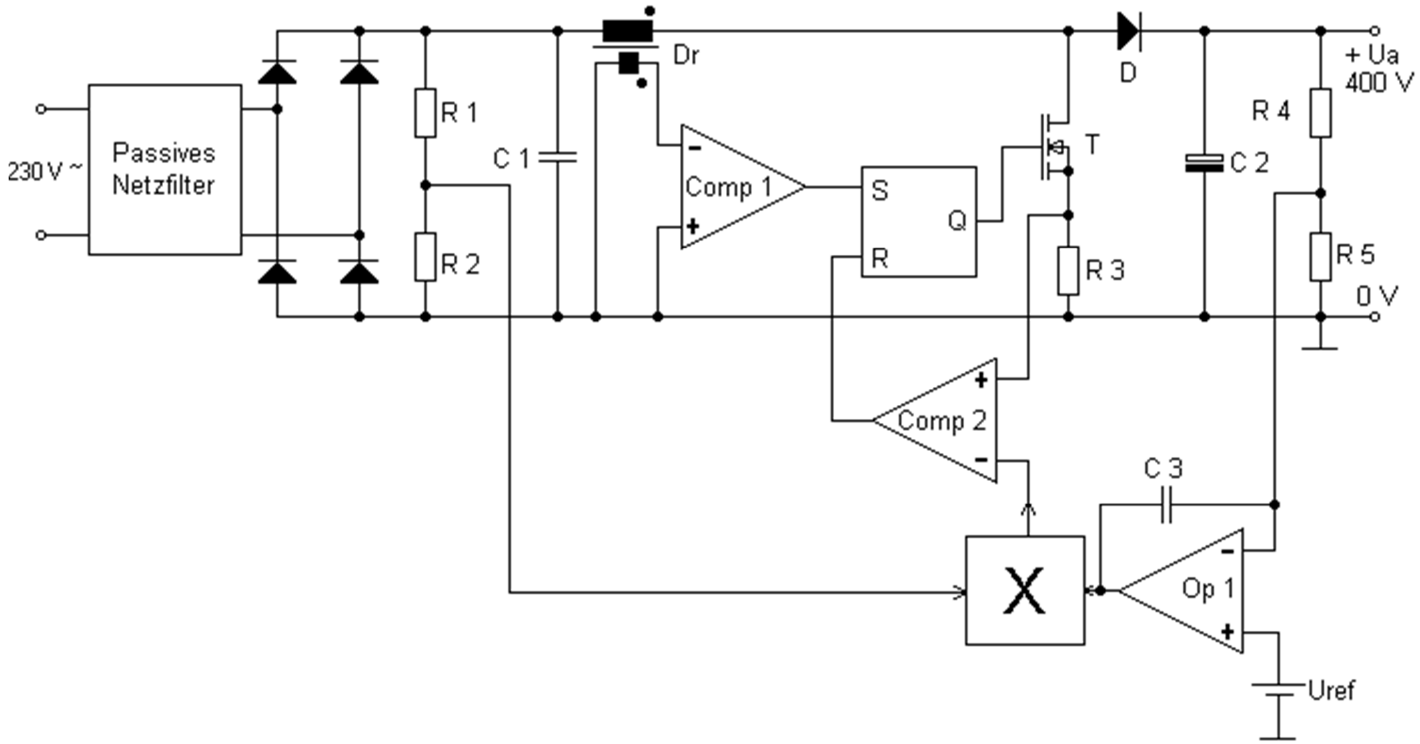


Figure 12.2 A block diagram of a conventional power factor correction

As the inductor current is a triangular gaps because the operation of the transducer and again starts from zero, the measured peak current of R 3 is exactly twice as large as the average (based on the switching frequency) line-side load current. The Power taken from mains power is proportional to the voltage at the inverting input of Comp 2, which comes from a multiplier. At one input of the multiplier is divided by the voltage of the voltage divider R1/R2. This ensures that the current drawn from the power is proportional to the instantaneous mains voltage and the circuit behaves as a resistor. Thus, only so much current flows as needed to maintain the output voltage maintains a controller is still required. The control amplifier Op 1 with R4/R5 compares the divided output voltage with a reference voltage, such as 2.5 volts. The integrating amplifier so that the output voltage changes only slowly. This is necessary so that can set a fairly stable voltage at the output of Op 1, despite the still existing 100-Hz ripple voltage on C 2 The level of the output voltage of Op 1 and thus also removed the power to regulate so the magnitude of the output voltage. Since the regulation at light loads can be slightly unstable and the integration capacitor C 3 is quite slow, even a surge protector for the output voltage should be installed, which turns the transistor with overvoltage immediately.

If the power margin of the converter is large enough, it can also be operated at voltages as low as about 100 volts AC. The power factor correction is therefore likely to be used for voltage matching of universal power supplies (with). Overall, the cost of the circuit is so high that it hardly makes sense to build it discreetly. Circuits for power factor correction are therefore almost always built with special PFC controller ICs. Unfortunately still no standardization in this area is evident. Many large manufacturers have such controller ICs in the program. But unfortunately these are not mutually compatible. To inform themselves about the latest types and switching examples I therefore recommend the websites of the major semiconductor manufacturers.

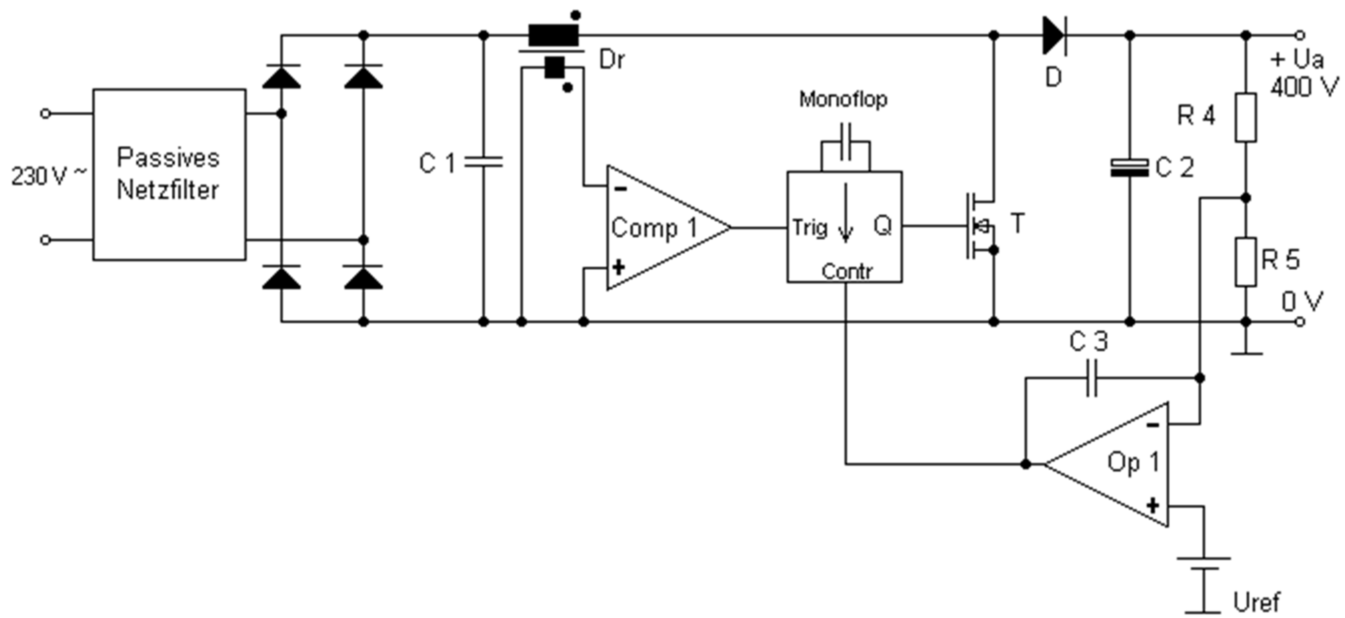


Figure 12.2 B simplified block diagram of a power factor correction

Fortunately, the PFC controller, which not always are quite cheap and make the power supply manufacturer may of individual semiconductor manufacturers depend tricky deal, namely it is also easier. You can just turn on for a constant time of the switching transistor. Following the collapse of the induced voltage in the storage inductor, the zero crossing detector Comp 1 provides the triggering of the monostable multivibrator, which switches on the transistor for a predetermined time. Just as in the previously described "conventional" power factor correction, the inductor current is zero, in this case, starting and triangular gaps. Since the time is constant, the voltage is a constant time on the choke. The rate of inductor current increase is, however, proportional to the applied voltage. At a constant duty cycle, this means that the peak current and hence the current taken from the network is proportional to the instantaneous value of the mains voltage. A measurement of the voltage and the inductor current is eliminated. B in Figure 12.2 is to see how the circuit has considerably simplified.
Another important advantage of the simplified power factor correction is the elimination of the multiplier, which greatly simplifies the construction with standard components. In order to adjust the input current, the duty cycle of the monoflop only needs to be changed, which is relatively easy to perform with a control voltage.The control amplifier, which is identical to the latest version, compares the actual with the desired tension and so regulates the pulse width of the monostable

multivibrator of the input current and the output voltage. Such a transducer can already build quite well with standard components. First, a power factor correction must be preceded by a noise filter and a rectifier. Since a step-up converter, the output voltage can never be less than the input voltage is obtained, as in normal Siebelko rectifier circuits, the problem of inrush current for charging the output electrolytic. For smaller power supplies ranges for most of an NTC as shown in Figure 12.2 C on the right can be seen. For larger power supplies, it is useful to use the slightly more complex circuit the left. It includes a charging current limit for the output capacitor, and a protection circuit. To charge the output capacitor, the current must initially by the rectifier via R 12, T 6 and T 7 flow. For the control of T 6 and T 7, an auxiliary voltage is generated in C 2, which is stabilized by ZD 1 to 12 volts. The auxiliary voltage is supplied via the resistors R 1, R 2 and D 1 of the unfiltered mains voltage. Thus, the life expectancy of the resistors does not suffer too much, R 1 and R 2 are connected in series. If you were to take only a single resistor, the DC power would there be present continuously, often leads to failure in normal resistors.

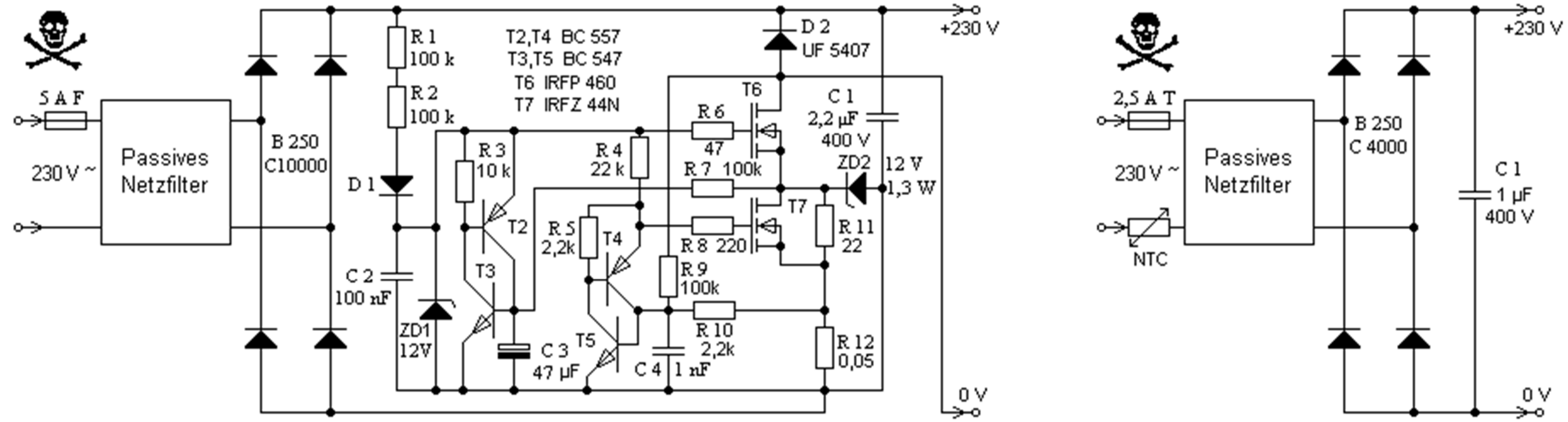


Figure 12.2 C rectifier for power factor correction with and without protective circuit

Since a significant drop after switching voltage difference at t 6, the thyristor is ignited via replica T4/T5 R 9, and closes the gate voltage of M 7 short, so that it remains locked. About R 6 is located at the gate of T 6 to the auxiliary voltage. The drain current of T6 causes a voltage drop across R 11, where raising the potential of the source T 6, that the current through R 11 is limited to 200 mA. The power loss in T 6 is then well below 100 watts and there is no danger of overload during the loading phase of the output electrolytic. If the voltage drops below about 6 to T 30 volts, the base voltage 5 is no longer sufficient to T of the thyristor for switching and replica T4/T5 locks again. Thus, also the gate of about 7 T R 4 can charge to 12 volts, so by T7 fully switched on and R 11 shorts. Then at T 6 is the full gate-source voltage of 12V and no more current limiting instead. Only when the current rises above about 10 amps, the voltage drop across R 12 is so high that light T4/T5 and T 7 closes again. Once the output capacitor has been charged with the peak value of the mains voltage, the current through R 20 and the voltage at T is too low to ignite 6 to T4/T5. In normal operation, therefore, T 6 and T 7 remain fully connected. If the output capacitor due to an overload or a short circuit on the primary side

can not charge, and the voltage of T 6 remains so high that T4/T5 remain connected and can only flow through the limited charging current of about 200 mA. If this state lasts much longer than a second, the C 3 R 7 can charge about the extent that the other thyristor replica T2/T3 lights and the gate voltage of both MOSFETs shorts permanently. Since T2/T3 have a low holding current, the current through R ranges from 10 to permanently switched through. A restart of the converter is only possible after an interruption of the mains voltage. The diode D2 limits the induced voltage that is generated during the current interruption in the storage inductor of the power factor correction.
Another advantage of the current limitation is that in the switch-on current flows never significantly longer than in normal operation. The circuit can therefore be well to secure a speedy fuse, which also does not have to be oversized.
The circuit shown on the left in Figure 12.2 C to use with the specified values for power up to about 1 kW. But principle, this will also achieve much higher performance. Purpose, only the components in the power circuit must increase accordingly, C 1 increases, R 11 and R 12 are reduced. Please note that this electronic charging current limit is not working at full load. The DC load on the output capacitor power factor correction may only see the full power when the Elko has reached approximately the nominal voltage of 380 volts in this case. This can be achieved eg by a time-delay as soft-start or an undervoltage detector you include in the actual converter will start the conversion only if a certain minimum voltage is reached at the output capacitor. Although a full load, suitable for electronic charging current limit is also possible, but much more expensive.

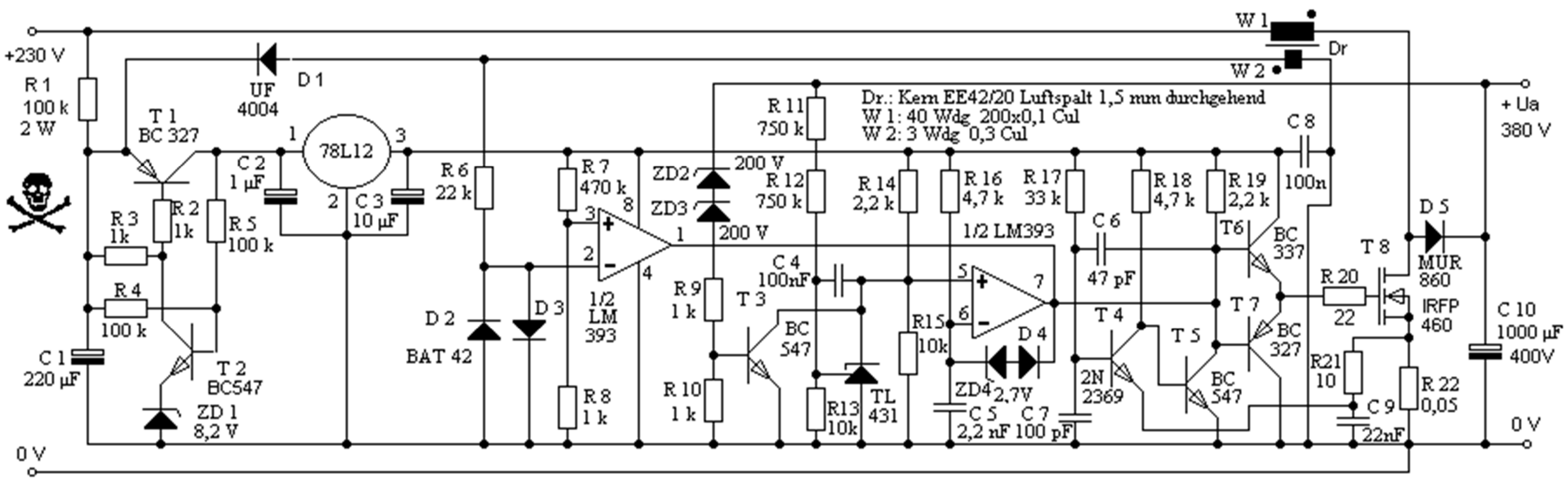


Figure 12.2 D Simplified Power Factor Correction with standard components

In Figure 12-2, the D of the actual power factor correction circuit is shown. The control circuit looks at first glance quite expensive. However, since, apart from the few power components, is only small electronic signal, the electronic control unit can be built compact. First, the circuit must be supplied with an operating voltage of about 12 volts. Since the operating voltage can be taken from the auxiliary winding of the storage inductor in normal operation, a start-up circuit is needed. Once the power is applied, the electrolytic capacitor C 1 is charged slowly through the starting resistor R first Since T 1 and T 2 first lock, C 1 is not loaded and can be recharged without hindrance. About R 4 and R 5 are only a small current flows, so that the electrolytic capacitor C 2 builds no significant voltage. On the basis of T 2 or about half the charging voltage of C 1 at. The Zener diode ZD 1 ensures that T 2 begins to conduct only when present about 9 volts at its base. This corresponds to a charging voltage of approximately 18 volts at 1 C Once a sufficiently high collector current flows in T 2, T 1 is also activated by and can increase the voltage on C 2. This increases the positive feedback (R 5), in turn the second base voltage of T In this way, T 1 and T 2 are abruptly switched on reaching the minimum charging voltage of about 18 volts at C 1 and C 2 is also located about 18 volts. Now that both R 4 and R 5 are at full operating voltage, T 2 and T 1 can lock again when this has dropped to about 9 volts. However, before this occurs, the converter is running and supplying the electrolytic capacitor C 1 through the rectifying diode D 1 with a sufficiently high operating voltage. Thus, the electronic control unit has defined characteristics, the voltage is not stabilized with an integrated series regulator to 12 volts.

The switching transistor is controlled over three modules: The two integrated comparators of the LM 393 and the monostable multivibrator T4/T5. All three have an open collector output and are connected such that T 8 blocks when at least one of the three modules output a logic zero. After establishing the operating voltage give all three a logical one and T 8 full turns through. The zero crossing detector (pin 1-3) gets through the auxiliary winding of the choke first no input voltage at Pin 2 The non-inverting input (pin 3) is biased by R 7 and R 8 to about 24 mV, resulting in a logical one at the output. The second comparator lies with its non-inverting input (pin 5) delivers its maximum output voltage of approximately 10 volts at the output of the control amplifier, which comprises a TL 431, and since the reference voltage has not been reached at the output. At the inverting input (pin 6) is the RC timer R16/C5. 5 C charges through R16 and eventually exceeds the voltage at pin 6, the voltage at pin 5 and the comparator output (pin 7) goes to logic zero. Before starting, however, can actually go to zero, C 5 is immediately a bit dead, so the output voltage can not fall substantially below 7 volts above ZD 4 and D 4th Also, the third assembly, the monoflop T4/T5, normally outputs a logic one. However, it is triggered when lowering the output voltage of the comparators and goes for at least 1us to logic zero and thus inhibits T 8 T 8 is first of all locked in a voltage of the storage choke, ie also in the auxiliary winding is induced if a current is flowed in advance. The induction voltage is fed from R 6 to the inverting input of the zero crossing detector, and has a positive sign if T blocks 8. The diodes D 2 and D 3 on the input of the comparator limit the voltage to a safe level. For a Schottky diode D 2 must be used, so that the input voltage does not become negative. Otherwise, a proper function of the comparator is no longer guaranteed. The induction voltage of the comparator determines the output to logical zero. During this time, then C5 is discharged to about 4 volts. Just as in the self-oscillating flyback converter, the next turn-on is initiated by the zero crossing at the auxiliary winding of the storage inductor and in the power factor correction. If the inductor current has decayed, collapses the induced voltage and the voltage at pin 2 falls below the voltage at pin 3 of the comparator. Thus, the comparator output goes to logic one and turned back by T 8. Now C can charge 5 unhindered until its voltage exceeds the output voltage of the error amplifier and the next phase of blocking T 8 initiates. The monostable multivibrator T4/T5 ensures that a reverse phase lasts at least 1 microseconds and thus ensures that C 5 is discharged to its start value to generate a clean blocking phase. Since the charging curve C is determined only by the components 5, 8 T of the duty cycle is only dependent on the output voltage of the control amplifier. If the control has not yet been used, the output voltage of the control amplifier and the duty cycle is maximum. The choke must be dimensioned so that it can not reach saturation at the maximum input voltage (350 volts) and duration (about 10 ms). Since it is may be difficult to achieve this, a current limit has been

inserted yet responsive to the maximum inductor current. R 22 is dimensioned so that it (here about 12 amperes) drop about 0.6 volts at the maximum inductor current. This voltage is fed from the RC element R21/C9 to the emitter of T 4 With an emitter voltage of about 0.6 volts from T 4 T 5 can not be locked so that it turns off the gate voltage of T 8th
The variable gain amplifier comprises a TL 431 has an integrating effect and by the negative feedback with C 4th This control is so sluggish that it averages out largely on the output capacitor, the residual ripple voltage. The voltage divider R 11, R 12 and R 13 is dimensioned so that the output voltage of about 380 volts to 2.5 volts is divided down. For R11, R 12 and R 13 should be used with 1% tolerance in any case metal film resistors. Since the scheme is quite slow, a surge protection was still drying installed. When the output voltage increases beyond 400 volts, the high-voltage zener diodes ZD2 and ZD3 are turned on and off by T 3. This includes the output voltage of the control amplifier, briefly, causing a shutdown of t eighth This scram mainly protects the expensive output capacitor C 10, from brief voltage spikes that can easily lead to its destruction because of the very tight dimensions. To ensure a sufficient dielectric strength even at higher operating temperatures, C10 should therefore be designed for 105 ° C. When using a 450-volt capacitors for C 10, the output voltage can be increased to about 400 volts. To R 11 and / or R12 can be increased to 820 kΩ, and a further Zener diode ZD2/ZD3 with about 30 volts needs to be connected in series.

13th Voltage transformers for special applications

In many areas, voltage transformers are optimized for specific purposes and can not be more clear in one of the previously treated group, the default converter lane.In this chapter, I will summarize and describe such converters.

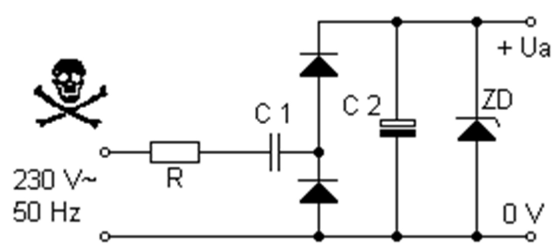
13.1 Auxiliary power generators in mains-powered devices

The need for a low operating voltage for the control of power electronics has been shown in switching power supplies. For space and cost reasons, it is often not possible to install a conventional auxiliary power supply to a device. As with switching power supplies is not even a DC isolation is also required for many other applications. Because of the high voltage difference is a linear regulator or resistor out of the question. Even at only 10 mA load current would have to be converted into heat at mains voltages of 230 volts 2-3 watts. An electronic control system for switching power supplies but can also easily accommodate 100 mA current times.50-Hz transformers are usually too big and takes a autotransformer at the high voltage difference too few benefits. An auxiliary voltage generation means of phase control I have already described in Chapter 4. The oldest and best known method to avoid power losses, the use of dummy resistors on the AC line voltage. Since power reactors would be even greater than transformers with the same output power, just as resistors capacitors come into question. With output currents above about 50 mA but which are also very large and expensive. A further possibility of smaller coils and / or capacitors to be used is to operate at high frequencies. These can be either specially created or taken from a switching power supply after it is started by a start-up resistor. In Chapter 6 Figure 6.1 M / N / O I have already described some very powerful downconverter for such applications. Moreover, the simple converter for such purposes described in other chapters are suitable. In the next section I want to present the converter in 50-Hz technology.

13.1.1 Auxiliary power generation in 50-Hz technology

As I wrote already, only capacitive reactive resistors come in 50-Hz applications in question. 13.1.1 A picture in the simplest of these circuits is shown. The resistance

R is the only limitation to the diodes to protect. He has some 100 ohms and is therefore so low that it is hardly noticeable compared to the reactance of C first To calculate the output current is easily assumed that C-1 of 50 per second is charged to + Û Û times while the charge Q = C * Û 2 via the diode D is applied to the electrolytic capacitor C 1 second When the charge on C 1 of + Û Û on-the charge of the diode D 2 is shorted to ground. The average current is given by I = f 2 = f * Q * C * Û.
For example, flows with Û = 320 volts (equivalent to approximately 230 V ~), f = 50 Hz and C = 0.47 uF, a current of approximately 15 mA. Less power is removed from the circuit, with the excess current must be dissipated of a Zener diode ZD to limit the output voltage.



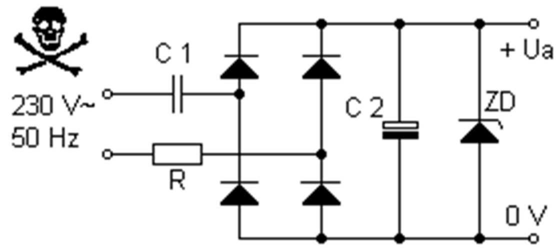


Image 13.1.1 A half-wave rectifier bridge rectifier B 13.1.1 image

If now, as B is shown in figure 13.1.1, the reactive current is on a bridge rectifier, both transhipment of C 1 C charge to be passed as second Thus, the current is doubled to I = 4 f * C * Û. Unfortunately, in this circuit, the output voltage must be potential-free compared to the mains voltage. For powering the control electronics, such as switching power supplies, it is therefore inappropriate. When a circuit that is supplied with a mains rectifier, also requires an auxiliary voltage, the AC voltage needs to be picked up directly at the input of the bridge rectifier. With respect to the negative terminal of the rectifier, which is the power grounding the circuit to be supplied, the voltage oscillates at the input between 0 and Û.

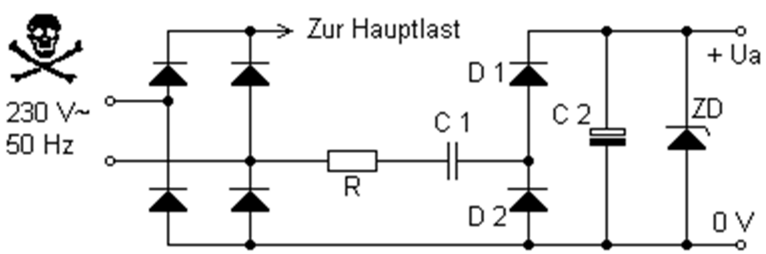
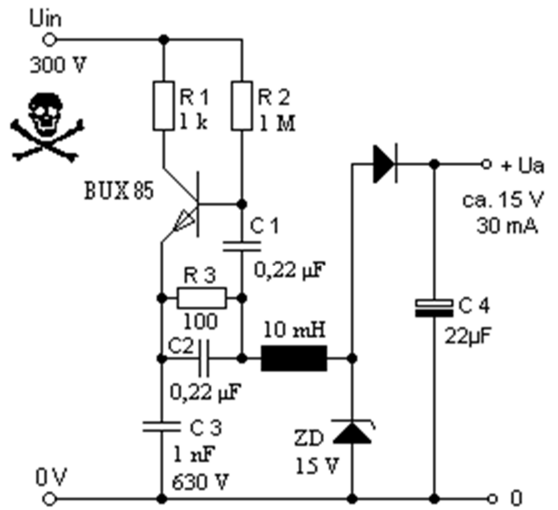


Image 13.1.1 C Power supply for control electronics

Because of the lower voltage swing compared to the last two circuits, an output current of I = f * C * Û results. Although the AC load of C1 is also lower, although C 1 must tolerate the full mains voltage Û. At the same capacitor C 1, the circuit thus provides only half of the current as shown in Figure 13.1.1 A and maybe only a quarter of the example in Figure 13.1.1 also remains to note that the bridge rectifier must be loaded from the main load Otherwise, there is at the entrance of the bridge rectifier, a DC voltage with respect to ground would build.

13.1.2 Auxiliary power generation with high operating frequency

Since the needed capacity at 50 Hz quickly assume unwieldy proportions when larger currents are required, it is useful to work with much higher frequencies. 13.1.2 A picture in a particularly simple RF generator for auxiliary power generation is seen. A special feature is that only a simple choke is required. The circuit operates as a somewhat modified Colpitts oscillator is connected such that the resonant circuit is at ground potential. The positive half-wave of the reactive current in the inductor is then simply coupled with a diode to C 4th A zener diode limits the output voltage to about 15 volts. The actual resonant circuit with a resonance frequency of approximately 50 kHz and C 3 form the throttle C 1 C 2 forms a voltage divider which ensures that the base of the transistor is applied as a somewhat larger amplitude at the emitter. C 1 is only a coupling capacitor which bridges the DC voltage difference between the base and emitter. R 2 provides the base current and sets the operating point of the transistor at the start of the oscillator. The resistor R 3 takes over the DC component of the emitter current and should not be larger than necessary. However, it must be relatively large compared to the reactance of C 2, which in this case is approximately 15 ohms. R 1 is an additional fuse resistance, would blow in the event of an error and prevents the capacitors and the inductor are destroyed. The transistor is so little burdened that he usually gets along without heatsink. In order to keep the reactive power losses, a correspondingly high throttle should be used. C 3 must be a low loss capacitor type MKP, FKP or FKC. For C 2 reaches a normal film capacitor C 1 and a multilayer ceramic capacitor. The circuit is short circuit proof, since it would vibrate in case of short circuit of the resonant circuit with low attenuation. The output current is calculated exactly as in the 50-Hz circuit in Figure 13.1.1 A to I = 2 * f * C Û Û with Uin = and f = 1 / (2π √ (LC $_{3))}$. Summarized results then I = V / π √ (C $_3$ / L)



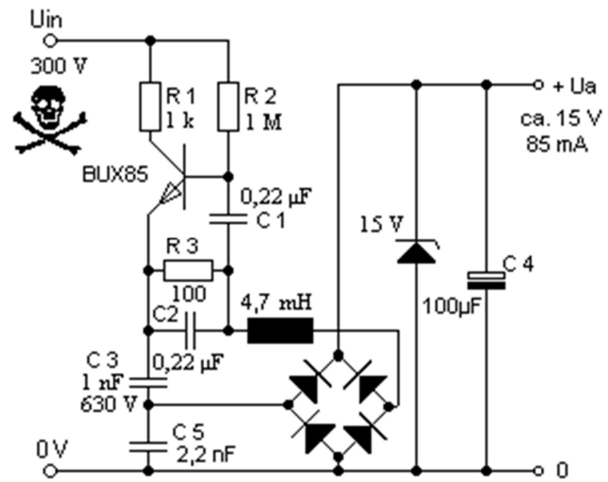
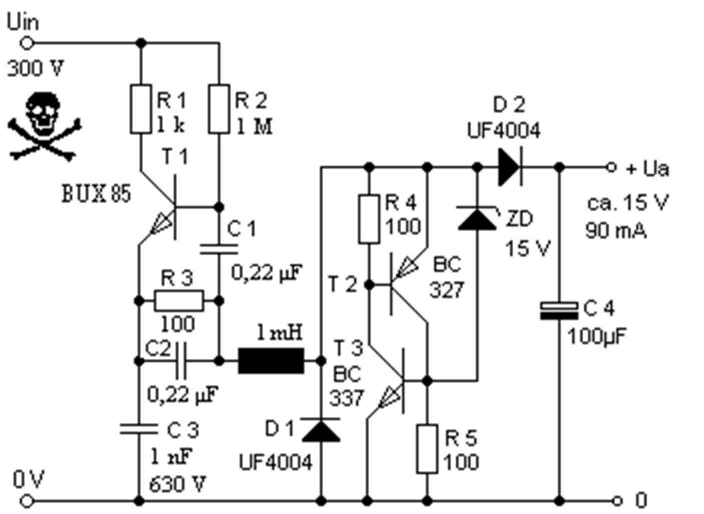


Image 13.1.2 A simple transformerless auxiliary RF voltage generator

Image 13.1.2 B generator with increased output power

If a higher output current is required, although you can dimension the circuit so that the reactive current in the resonant circuit is higher, but I recommend first mitzunutzen also the capacitive reactive power. In Figure 13.1.2 B such better utilization of reactive power can be seen. In principle, a bridge rectifier is simply inserted into the resonant circuit, which doubles the output current. Halving the inductance also increases the resonance frequency and thus also the reactive power.The output current is then calculated as $I = 2U / \pi \sqrt{(C_3 / L)}$. 5 C during the starting phase ensures that the diodes are capacitively bridged or high impedance and so the resonant circuit is closed. On the throttle side is not necessary because the direct current in the inductor, a diode is already conducting and thus alternating current is the lower end of the throttle to ground. A disadvantage of the two last-mentioned circuits is that the power supply is not regulated. At low load, the excess power output of the Zener diode ZD must be burned. That would be B in Figure 13.1.2 already around 1 Watt power dissipation in the zener diode. With increasing output power circuit is a simple rule would be desirable. In Figure 13.1.2 C I combined the simplest auxiliary voltage generator with an also very simple control circuit. The transistors T 2 and T 3 form of a thyristor, which ignites when a voltage higher than 15 volts is applied across the diode D 1st During the positive half-wave of the reactive current by the reactor, flows through the load 2, and D to C 4. C during 4 is charged, the voltage increases slightly to D 1. As soon as the output voltage rises above about 15 volts, the voltage increase leads during the current flow phase of ignition of the thyristor replica T2/T3 that the voltage shorts to D 1 and takes care of the power factor for the remainder of the positive half-wave. The higher the output voltage, the sooner T2/T3 is switched through and the less the time during which the current can flow via D2 to the output. During the negative half-wave of the reactive current is then enough time for T 2 and T 3 to lock again.



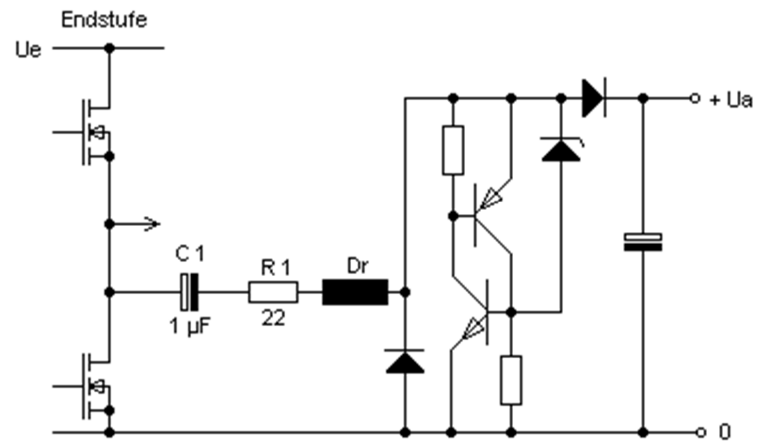


Image 13.1.2 C simple regulated supply voltage generator

13.1.2 Regulated D image generator with "external power supply"

In switching power supplies with half-bridge circuit course you can also use the square wave voltage at the output of the half bridge as an AC voltage source, as shown in Figure 13.1.2 D. Condition then is a starting circuit that starts the power supply to the auxiliary power generator can take over the power supply of the control circuit. Furthermore, it is necessary that a reasonably constant square-wave voltage applied to the half-bridge. For example, this is not the case for a regulated flow transducer. At low load, the pulse width would approach zero, and the auxiliary voltage transformer could not make any more power. At resonance transformers, this only works if the frequency range is limited enough and the control reserve is sufficiently large. Optimally, however, the converter operates when supplied with a symmetrical square with a constant frequency. Then you could also do without a control. The calculation of the components is relatively easy. C 1 is only the decoupling of the DC share, which always rests on a half-bridge and is uncritical terms of size. It must be so large that the resonance frequency of the series resonant circuit that it forms with the throttle Dr, is well below the working frequency (> 10 times). But he should not be unnecessarily large also, so that the charging current pulse is not too large. The resistor R 1 only limits the charging current of C 1 and is also critical. It should be small enough so that the load current does not cause a substantial power loss. Must be calculated only the inductance. When the AC voltage is symmetrical rectangular (± Ue / 2), the inductor current is triangular.In the positive half period of the square-wave voltage, the inductor current of Imax +, Imax rises. The increase in current in the inductor is dI / dt = 0.5 Ue / L. I Multiply that by the time the voltage is applied consistently, I have the entire current change in half a period, ie Imax (Imax-) = 2 | Imax | = 1/2 U_{e} / LT / 2 . As the current increases approximately linearly and falls, the average output current can be easily calculated. A half period, no current flows to the output and the other half of the current increases linearly from zero to Imax and then back to zero. Thus, the output current Ia = Imax ¼. It is then

$8I_{a} = 1/2\ {}_{e} U / L\ T / 2$ and $TU_{e} / L \rightarrow I_{a} = U_{e} / 32FL$.

At Ue = 300 V, f = 100 kHz and L = 1 mH would then deliver to the circuit output current to approximately 90 mA.

14th Formulary

The following is a summary of the main formulas that are relevant for the dimensioning of the components in all chapters.

Transformers and reactors generally
Flow and flux density

$\Phi = B * a$ and $\Phi_{max} = B_{max} * a$

Φ = magnetic flux in a coil or in a core in Vs

Φ_{max} = saturation limit of the magnetic flux in a core

B = Magnetic. Flux density of the field in a coil or in a core in Tesla = Vs / m^2

B_{max} = saturation limit of the field strength in a core (iron ~ 1.5 T, 0.3 T ~ ferrite)

A = cross-sectional area of the core of the coil or in m^2

Estimating the transmitted power of a 50-Hz transformer commercial EI-side length L of the iron core (yoke length in cm):

$P \sim_t (l / cm)^{3.5} * 0{,}038$ VA

Calculation of the maximum voltage sinusoidal current of an iron core in V / Wdg:

$\hat{U}_1 = 2\pi f \quad = 2\pi fa$

f = frequency in Hz, a cross-sectional area of the iron core in square the maximum magnetic field strength in Tesla (1.5 T at soft iron)

At f = 50 Hz and = 1.5 T is:

$\hat{U} \sim_1 470a * [V / m^{2]}$ or for the effective voltage $U_{eff} \sim a * 333 [V / m^{2]}$

(Scattered) inductance L of an inductor, a cylindrical coil or air (litter) transformer:

L> ìN A / l or L A_L N A = $_L{}^2$ at a known value

Maximum coil current at the onset of saturation

$I_{max} \approx Bl / (N\mu_0)$ (B = saturation field strength of 1.5 T and 0.3 T with soft iron with ferrite)

or $I_{max} = \Phi_{max} /_L$ NA with known A_L value

a coil inductive reactance X_L $X_L = 2\pi fL$

stored energy W_L a choke or a transformer: $W_L = 0.5 LI^2$

with I coil current, $\mu_0 = 4\pi \times 10^{-7}$ Vs / Am (Magnetic field constant), N = number of turns of the coil, A = cross-sectional area of the coil or of the air gap magnetic flux Φ_{max} = max l = length of the core and the coil, and the air gap.

Calculate the required transmission power of an autotransformer:

Up autotransformer $P_t = P_a (1 - {}_{U\,i} / U_{a)}$

Down Autotransformer $P_t = P_a (1 - U_a / U_{e)}$

P_t = power actually transformed the transformer

P_a = Ein-/Ausgangsleistung

Rectifier and filter circuits

No-load direct voltage to a Siebelko behind a rectifier: $U_{max} = \sqrt{2}\ V_{rms}$

Three-phase bridge rectifier $U_{max} = \sqrt{6}\ U_{eff}$

with U_{eff} = effective voltage between a phase and neutral

Residual voltage ripple Siebelko $U_{br} = I_a T / C$

$_{br}$ U peak-to-peak value of the residual voltage ripple Siebelko, I_a load current, C is the capacitance of the filter capacitors and T the period of the ripple voltage (10 ms at 50 Hz bridge rectifier)

Wish rule: If 1 amp load current, a 1000 uF electrolytic capacitor discharges at 1 V / ms

Cut-off frequency of an RC filter network $f_t = 1 / (2\pi RC)$

Cut-off frequency of an LC ladder filter $f = 1 / (2\pi \sqrt{(LC)})$

Transformer with chokes

Recommended minimum switching frequency f of a down-converter or inverse

$f = U_a{}^2 / (U_a L_e 2I)$ or $_a$ minimum inductance $L = U^2 / (V_{s\,a} f 2I)$ if f is defined

applies << U_e | for | U_a

f = | U_a / $_{(a}$ L 2I) | or minimum inductance L = | U_a | / $_{(a}$ f 2I) if f is defined

I_a = minimum output current at which the inductor current will not be continuous

Recommended minimum switching frequency f of a boost converter

U f = $_e{}^2 / (U_a L_e 2I)$ Minimum inductance L or U = $_e{}^2 / (U_{a\,e} f 2I)$ if f is defined

applies to U_a U_e <<

U f = $_e / (2I_s)$ Or minimum inductance $L = U_s / (2I_e f)$ if f is defined

I_a = minimum output current at which the inductor current will not be continuous

Primary switched converter
maximum duty cycle T_{max} flow of a single-ended flyback converter or

$T_{max} = N \Phi_{max} / U_b$ or $T_{max} = LI_{max} / U_b$

maximum duty cycle T_{max} or minimum switching frequency f_{min} of a push-pull converter

Converter with full bridge or parallel feed:

$T_{max} = 2N \Phi_{max} / U_b$ and $f_{min} = 1/4\ U_b / (N\Phi_{max})$

Half-bridge converter with:

$T_{max} = 4N \Phi_{max} / U_b$ and $f_{min} = 1/8\ U_b / (N\Phi_{max})$

with U_b = operating voltage of the converter, N = number of turns of the primary coil, and a primary coil with parallel feed, Φ_{max} = maximum magnetic flux of the core, L = inductance of the primary coil and I_{max} = maximum current in the primary coil

Power dissipation P_S is the thermal scattering field disposal of a flyback converter at full load

$P_S = 1/2\ f\ L_S\ I_{max}^2$

where f = switching frequency, L_S = primary leakage inductance and I_{max} = primary current at the end of the flux phase at full load.

maximum short-circuit current I of a primary ZCS resonant converter

$I \approx U_b \sqrt{(C/2L)}$ L / C = inductance / capacitance of the resonant circuit

U_b = supply voltage of the half bridge

CE voltage at the transistor of a single-ended sine wave converter

$_{CE}U = U_b + \sqrt{(U_b^2 + {}_c^2 I\ L / C)}$

Effective reactive current I in the resonant circuit with low distortion

$$I = V_b \sqrt{(C/2L)}$$

with U_b = operating voltage of the converter, I_C = collector saturation current, L = inductance of the primary coil and C = capacitance of the resonance capacitor amplitude at the center tap of a resonator coil of a push-pull sine converter shown with power supply inductor as shown in Figure 11.2 B.

$U_m = U_b * \pi / 2$ and maximum collector voltage $V_{CE} = \pi * U_b$

Effective reactive current $I = U_b \pi \sqrt{(C / (2L))}$

Other converters
Maximum output current I of a mains-powered auxiliary power supply with upstream capacitor C with half-wave rectification:

$I = 2 * f * C \hat{U}$ and with bridge rectifier $I = 4 f * C * \hat{U}$

When the voltage at the input of the bridge rectifier for the main load is removed (see Figure 13.1.1 C)

$I = f * C * \hat{U}$

f = line frequency (50 Hz) and $\hat{U}$ = peak value of the mains voltage (325V)

Maximum output current I of an auxiliary power supply with RF Sine Generator

with half-wave rectification $I = V_b / \pi \sqrt{(C / L)}$ and bridge rectifier $I = 2U_b / \pi \sqrt{(C / L)}$

with U_b = Power, C and L, capacitance and inductance of the capacitor and coil of the resonant circuit.

Maximum output current I, one of a half-bridge supplied with a symmetrical square wave voltage auxiliary power generator (see Figure 13.1.2 D)

$I = V_b / (32FL)$

with U_b = operating voltage and f = switching frequency of the half-bridge. L = inductance of the series reactor

Maximum ratio of an ideal transformer Tesla (see Figure 13.3.2 C)

$U_a / U_e = \sqrt{(C / C_k)}$

with U_a = output voltage of the Tesla coil, U_e = Input voltage at Primärresonator, C = capacitance of the Primärresonators and C_k = head capacity of the Tesla coil.