SMPSRM/D
Rev. 4, Apr-2014

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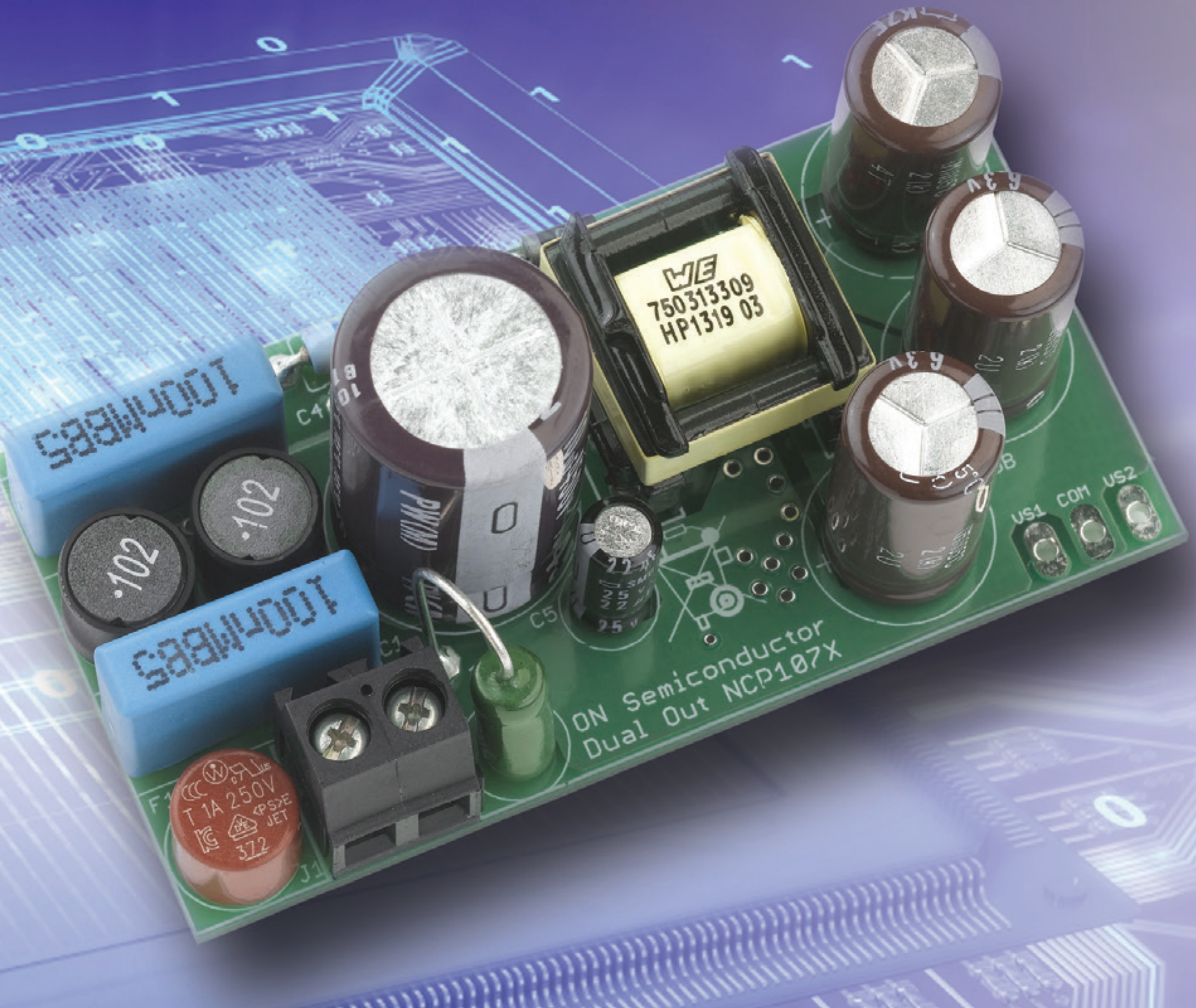
Energy Efficient Innovations

Switch-Mode Power Supply
Reference Manual



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Reference Manual

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Forward

Every new electronic product, except those that are battery powered, requires converting off−line 115 Vac or 230 Vac power to some dc voltage for powering the electronics. The availability of design and application information and highly integrated semiconductor control ICs for switching power supplies allows the designer to complete this portion of the system design quickly and easily. Whether you are an experienced power supply designer, designing your first switching power supply or responsible for a make or buy decision for power supplies, the variety of information in the *Switch−Mode Power Supply Reference Manual* should prove useful.

This reference manual contains useful background information on switching power supplies for those who want to have more meaningful discussions and are not necessarily experts on power supplies. It also provides real SMPS examples, and identifies several application notes and additional design resources available from ON Semiconductor, as well as helpful books available from various publishers and useful web sites for those who are experts and want to increase their expertise. An extensive list and brief description of analog ICs, power transistors, rectifiers and other discrete components available from ON Semiconductor for designing a SMPS are also provided.

For the latest updates and additional information on energy efficient power management and discrete devices, please visit our website at ***www.onsemi.com***.

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Introduction

The never−ending drive towards smaller and lighter products poses severe challenges for the power supply designer. In particular, disposing of excess heat generated by power semiconductors is becoming more and more difficult. Consequently it is important that the power supply be as small and as efficient as possible, and over the years power supply engineers have responded to these challenges by steadily reducing the size and improving the efficiency of their designs.

Switching power supplies offer not only higher efficiencies but also greater flexibility to the designer. Recent advances in semiconductor, magnetic and passive technologies make the switching power supply an ever more popular choice in the power conversion arena.

This guide is designed to give the prospective designer an overview of the issues involved in designing switchmode power supplies. It describes the basic operation of the more popular topologies of switching power supplies, their relevant parameters, provides circuit design tips, and information on how to select the most appropriate semiconductor and passive components. The guide also lists the ON Semiconductor components expressly built for use in switching power supplies.

Linear versus Switching Power Supplies

Switching and linear regulators use fundamentally different techniques to produce a regulated output voltage from an unregulated input. Each technique has advantages and disadvantages, so the application will determine the most suitable choice.

Linear power supplies can only step−down an input voltage to produce a lower output voltage. This is done by operating a bipolar transistor or MOSFET pass unit in its *linear* operating mode; that is, the drive to the pass unit is proportionally changed to maintain the required output voltage. Operating in this mode means that there is always a headroom voltage, Vdrop, between the input and the output. Consequently the regulator dissipates a considerable amount of power, given by (Vdrop Iload).

This *headroom loss* causes the linear regulator to only be 35 to 65 percent efficient. For example, if a 5.0 V regulator has a 12 V input and is supplying 100 mA, it must dissipate 700 mW in the regulator in order to deliver 500 mW to the load , an efficiency of only 42 percent. The cost of the heatsink actually makes the linear regulator uneconomical above 10 watts for small applications. Below that point, however, linear regulators are cost−effective in step−down applications.

A low drop−out (LDO) regulator uses an improved output stage that can reduce Vdrop to considerably less than 1.0 V. This increases the efficiency and allows the linear regulator to be used in higher power applications.

Designing with a linear regulator is simple and cheap, requiring few external components. A linear design is considerably quieter than a switcher since there is no high−frequency switching noise.

Switching power supplies operate by rapidly switching the pass units between two efficient operating states: *cutoff*, where there is a high voltage across the pass unit but no current flow; and *saturation,* where there is a high current through the pass unit but at a very small voltage drop. Essentially, the semiconductor power switch creates an AC voltage from the input DC voltage. This AC voltage can then be stepped−up or down by transformers and then finally filtered back to DC at its output. Switching power supplies are much more efficient, ranging from 65 to 95 percent.

The downside of a switching design is that it is considerably more complex. In addition, the output voltage contains switching noise, which must be removed for many applications.

Although there are clear differences between linear and switching regulators, many applications require both types to be used. For example, a switching regulator may provide the initial regulation, then a linear regulator may provide post−regulation for a noise−sensitive part of the design, such as a sensor interface circuit.

Switching Power Supply Fundamentals

There are two basic types of pulse−width modulated (PWM) switching power supplies, *forward−mode* and *boost−mode*. They differ in the way the magnetic elements are operated. Each basic type has its advantages and disadvantages.

The Forward−Mode Converter

The forward−mode converter can be recognized by the presence of an L−C filter on its output. The L−C filter creates a DC output voltage, which is essentially the volt−time average of the L−C filter's input AC rectangular waveform. This can be expressed as:

$$V_{out} \approx V_{in} \cdot \text{duty cycle} \qquad \text{(eq. 1)}$$

The switching power supply controller varies the duty cycle of the input rectangular voltage waveform and thus controls the signal's volt−time average.

The *buck* or *step−down converter* is the simplest forward−mode converter, which is shown in Figure 1.

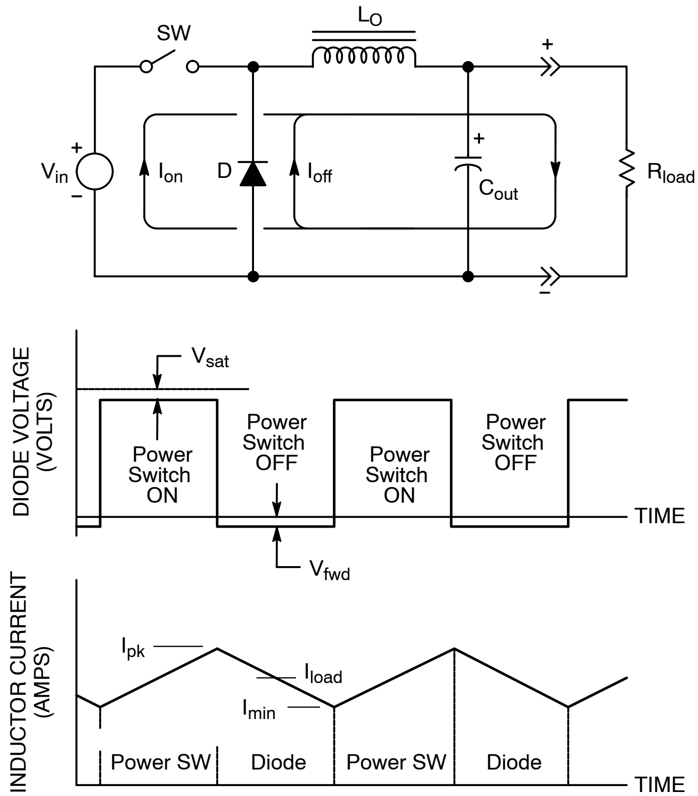


Figure 1. A Basic Forward−Mode Converter and Waveforms (Buck Converter Shown)

Its operation can be better understood when it is broken into two time periods: when the power switch is turned on and turned off. When the power switch is turned on, the input voltage is directly connected to the input of the L−C filter. Assuming that the converter is in a steady−state, there is the output voltage on the filter's output. The inductor current begins a linear ramp from an initial current dictated by the remaining flux in the inductor. The inductor current is given by:

$$i_{L(on)} = \frac{(V_{in} - V_{out})}{L} t + i_{init} \quad 0 \le t \le t_{on} \qquad \text{(eq. 2)}$$

During this period, energy is stored as magnetic flux within the core of the inductor. When the power switch is turned off, the core contains enough energy to supply the load during the following off period plus some reserve energy.

When the power switch turns off, the voltage on the input side of the inductor tries to fly below ground, but is clamped when the *catch diode* D becomes forward biased. The stored energy then continues flowing to the output through the catch diode and the inductor. The inductor current decreases from an initial value i_{pk} and is given by:

$$i_{L(off)} = i_{pk} - \frac{V_{out} t}{L} \quad 0 \le t \le t_{off} \qquad \text{(eq. 3)}$$

The off period continues until the controller turns the power switch back on and the cycle repeats itself.

The buck converter is capable of over one kilowatt of output power, but is typically used for on−board regulator applications whose output powers are less than 100 watts. Compared to the flyback−mode converter, the forward converter exhibits lower output peak−to−peak ripple voltage. The disadvantage is that it is a step−down topology only. Since it is not an isolated topology, for safety reasons the forward converter cannot be used for input voltages greater than 42.5 VDC.

The Flyback−Mode Converter

The basic flyback−mode converter uses the same components as the basic forward−mode converter, but in a different configuration. Consequently, it operates in a different fashion from the forward−mode converter. The most elementary flyback−mode converter, the *boost* or *step−up converter,* is shown in Figure 2.

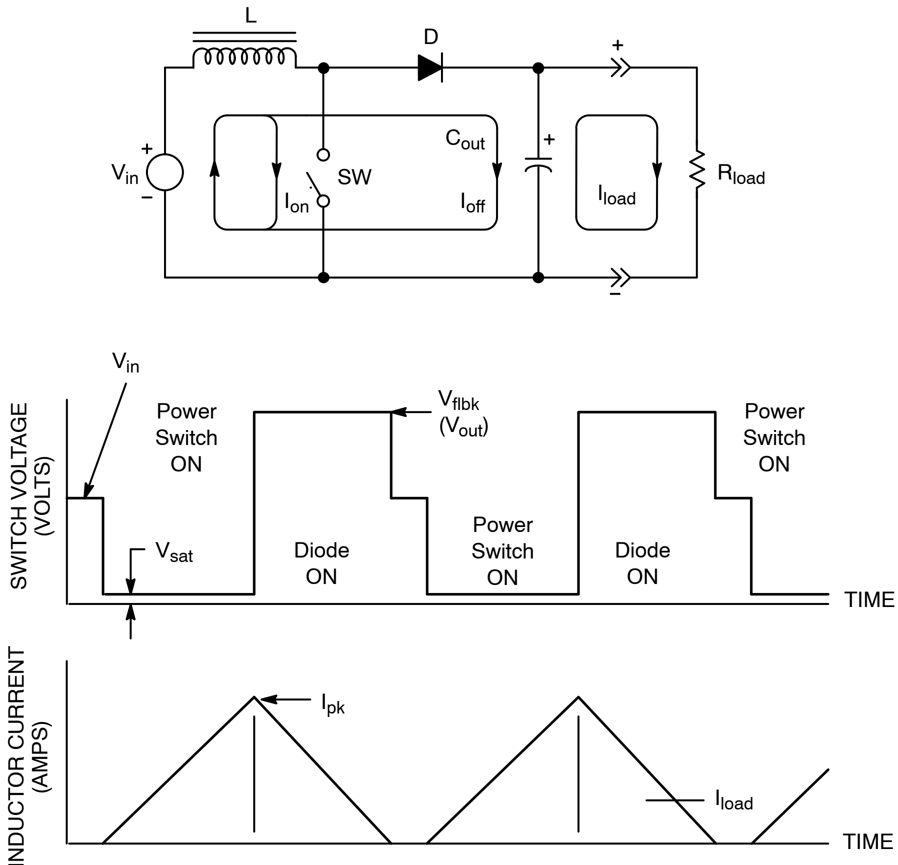


Figure 2. A Basic Boost−Mode Converter and Waveforms (Boost Converter Shown)

Again, its operation is best understood by considering the "on" and "off" periods separately. When the power switch is turned on, the inductor is connected directly across the input voltage source. The inductor current then rises from zero and is given by:

$$i_{L(on)} = \frac{V_{in} t}{L} \qquad \leq t \leq 0_{on} \qquad \text{(eq. 4)}$$

Energy is stored within the flux in the core of the inductor. The peak current, i_{pk}, occurs at the instant the power switch is turned off and is given by:

$$i_{pk} = \frac{V_{in}\, t_{on}}{L} \qquad \text{(eq. 5)}$$

When the power switch turns off, the switched side of the inductor wants to fly−up in voltage, but is clamped by the output rectifier when its voltage exceeds the output voltage. The energy within the core of the inductor is then passed to the output capacitor. The inductor current during the off period has a negative ramp whose slope is given by:

$$i_{L(off)} = \frac{(V_{in} - V_{out})}{L} \qquad \text{(eq. 6)}$$

The energy is then completely emptied into the output capacitor and the switched terminal of the inductor falls back to the level of the input voltage. Some ringing is evident during this time due to residual energy flowing through parasitic elements such as the stray inductances and capacitances in the circuit.

When there is some residual energy permitted to remain within the inductor core, the operation is called *continuous− mode*. This can be seen in Figure 3.

Energy for the entire on and off time periods must be stored within the inductor. The stored energy is defined by:

$$E_L = 0.5L \cdot ipk^2 \qquad \text{(eq. 7)}$$

The boost−mode inductor must store enough energy to supply the output load for the entire switching period (t_{on} + t_{off}). Also, boost−mode converters are typically limited to a 50 percent duty cycle. There must be a time period when the inductor is permitted to empty itself of its energy.

The boost converter is used for board−level (i.e., non−isolated) step−up applications and is limited to less than 100−150 watts due to high peak currents. Being a non−isolated converter, it is limited to input voltages of less than 42.5 VDC. Replacing the inductor with a transformer results in a flyback converter, which may be step−up or step−down. The transformer also provides dielectric isolation from input to output.

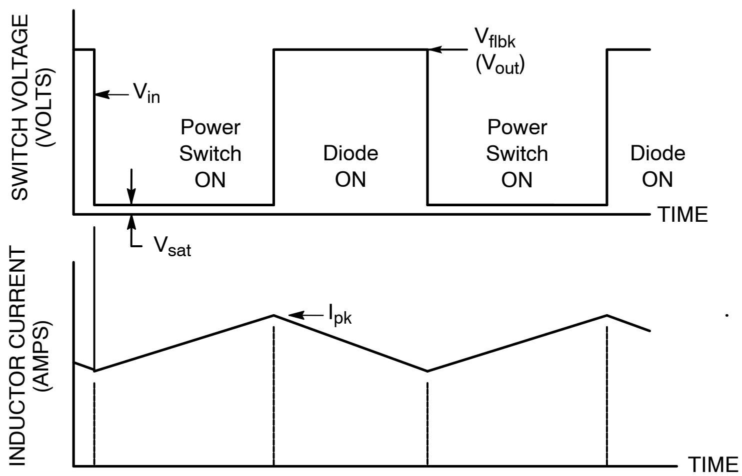


Figure 3. Waveforms for a Continuous−Mode Boost Converter

Common Switching Power Supply Topologies

A topology is the arrangement of the power devices and their magnetic elements. Each topology has its own merits within certain applications. There are five major factors to consider when selecting a topology for a particular application. These are:

1. Is input−to−output dielectric isolation required for the application? This is typically dictated by the safety regulatory bodies in effect in the region.
2. Are multiple outputs required?
3. Does the prospective topology place a reasonable voltage stress across the power semiconductors?
4. Does the prospective topology place a reasonable current stress upon the power semiconductors?
5. How much of the input voltage is placed across the primary transformer winding or inductor?

Factor 1 is a safety−related issue. Input voltages above 42.5 VDC are considered hazardous by the safety regulatory agencies throughout the world. Therefore, only transformer−isolated topologies must be used above this voltage. These are the *off−line* applications where the power supply is plugged into an AC source such as a wall socket.

Multiple outputs require a transformer−based topology. The input and output grounds may be connected together if the input voltage is below 42.5 VDC. Otherwise full dielectric isolation is required.

Factors 3, 4 and 5 have a direct affect upon the reliability of the system. Switching power supplies deliver constant power to the output load. This power is then reflected back to the input, so at low input voltages, the input current must be high to maintain the output power. Conversely, the higher the input voltage, the lower the input current. The design goal is to place as much as possible of the input voltage across the transformer or inductor so as to minimize the input current.

Boost−mode topologies have peak currents that are about twice those found in forward−mode topologies. This makes them unusable at output powers greater than 100−150 watts.

Cost is a major factor that enters into the topology decision. There are large overlaps in the performance boundaries between the topologies. Sometimes the most cost−effective choice is to purposely design one topology to operate in a region that usually is performed by another. This, though, may affect the reliability of the desired topology.

Figure 4 shows where the common topologies are used for a given level of DC input voltage and required output power. Figures 5 through 12 show the common topologies. There are more topologies than shown, such as the Sepic and the Cuk, but they are not commonly used.

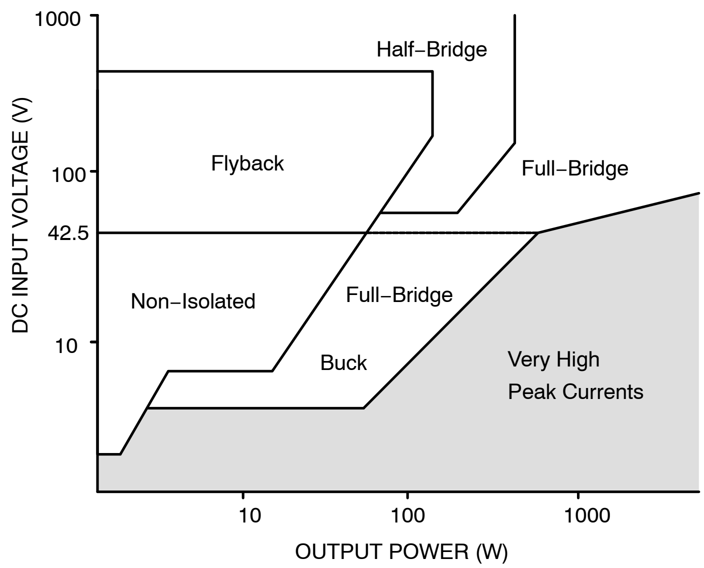


Figure 4. Where Various Topologies Are Used

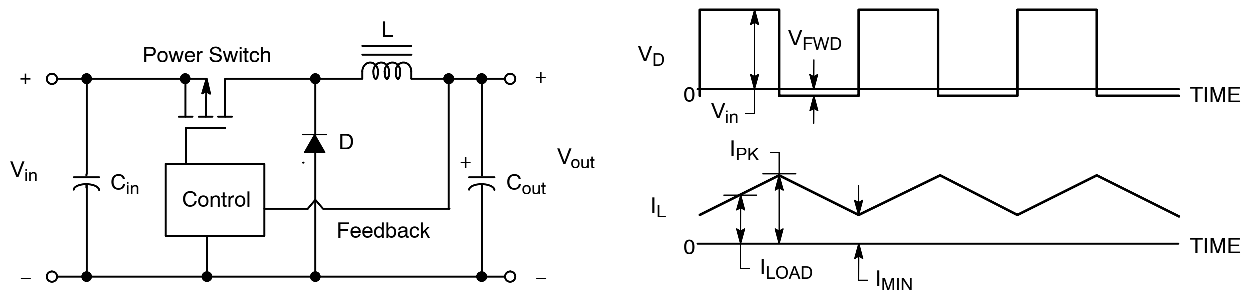


Figure 5. The Buck (Step−Down) Converter

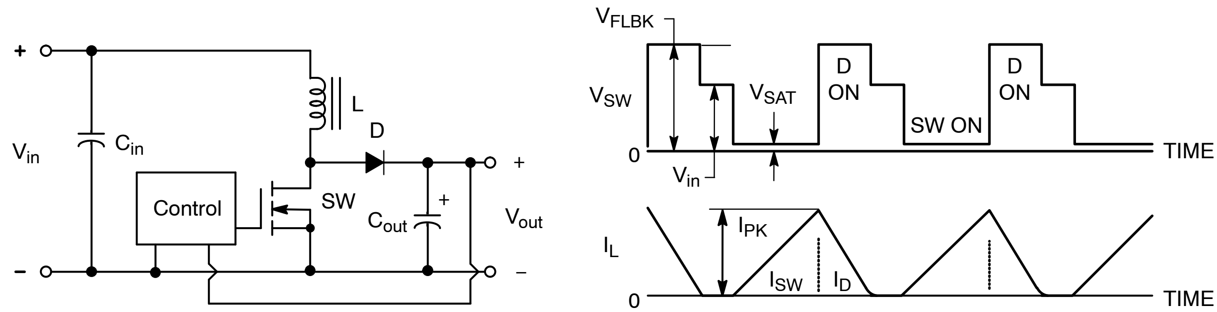


Figure 6. The Boost (Step−Up) Converter

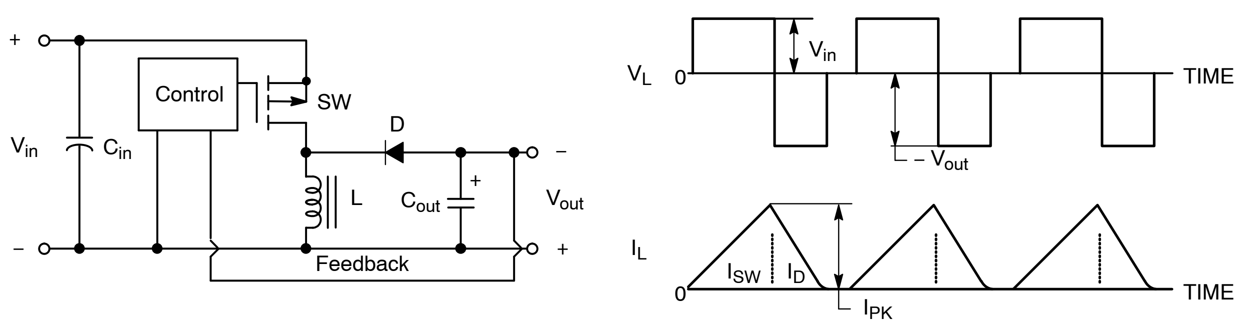


Figure 7. The Buck−Boost (Inverting) Converter

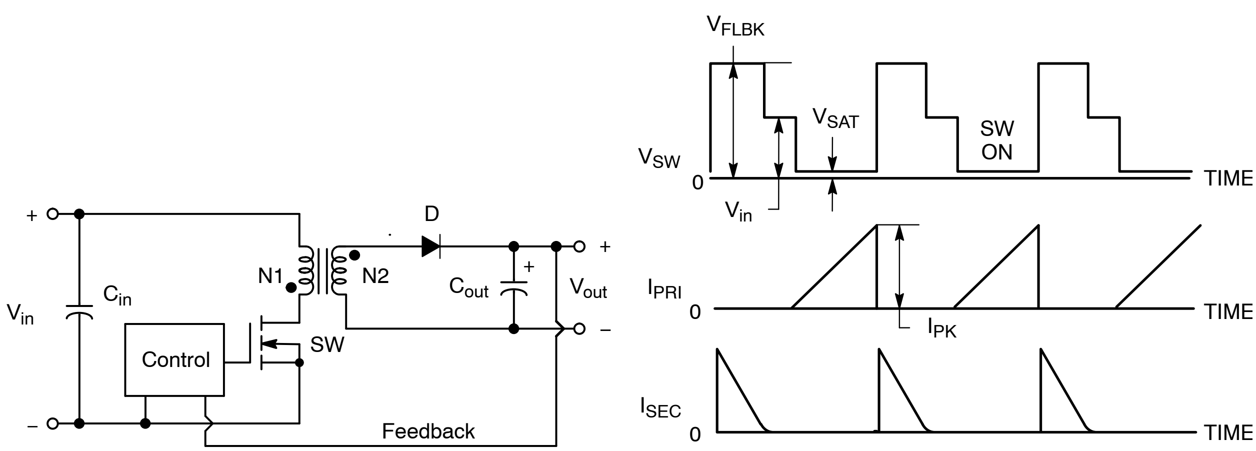


Figure 8. The Flyback Converter

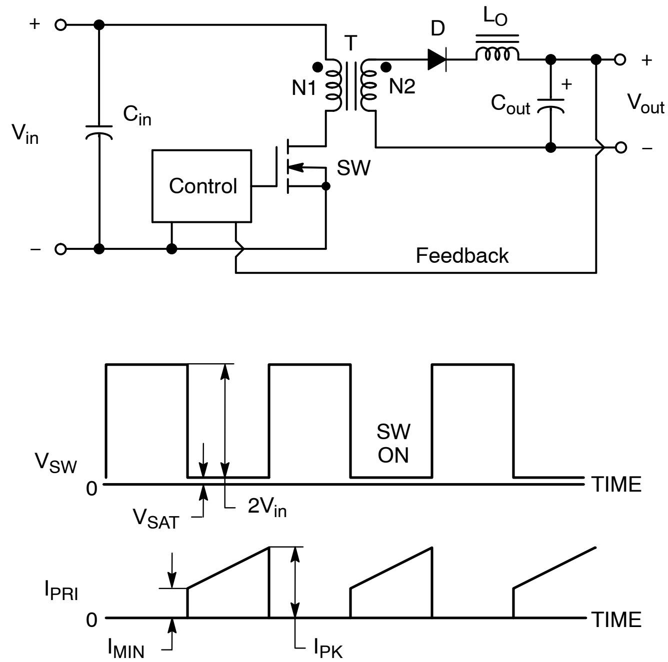


Figure 9. The One−Transistor Forward Converter (Half Forward Converter)

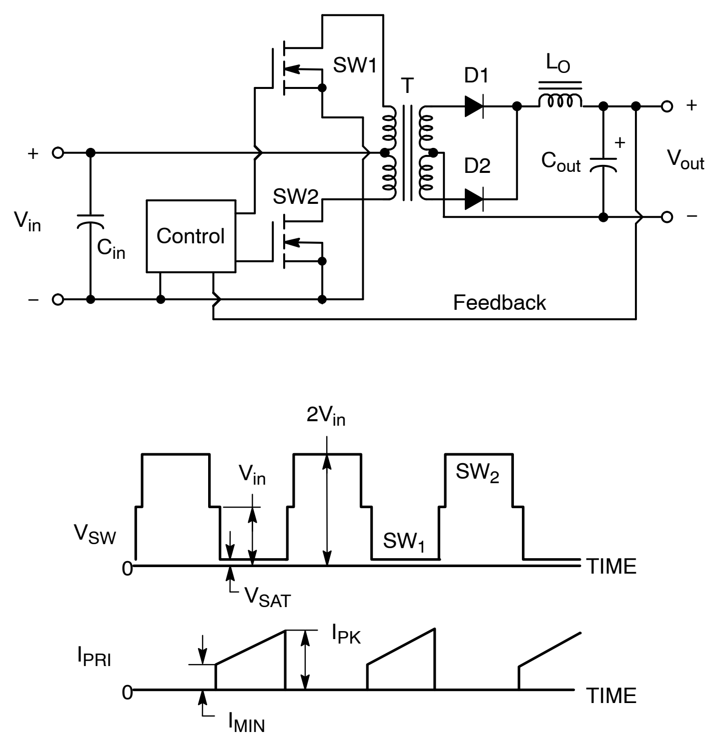


Figure 10. The Push−Pull Converter

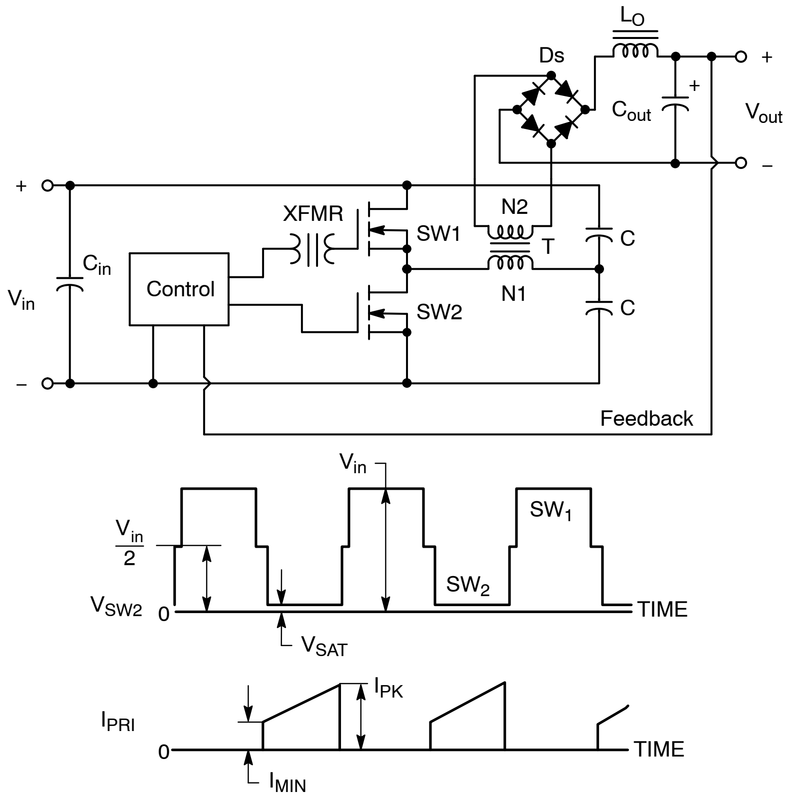


Figure 11. The Half−Bridge Converter

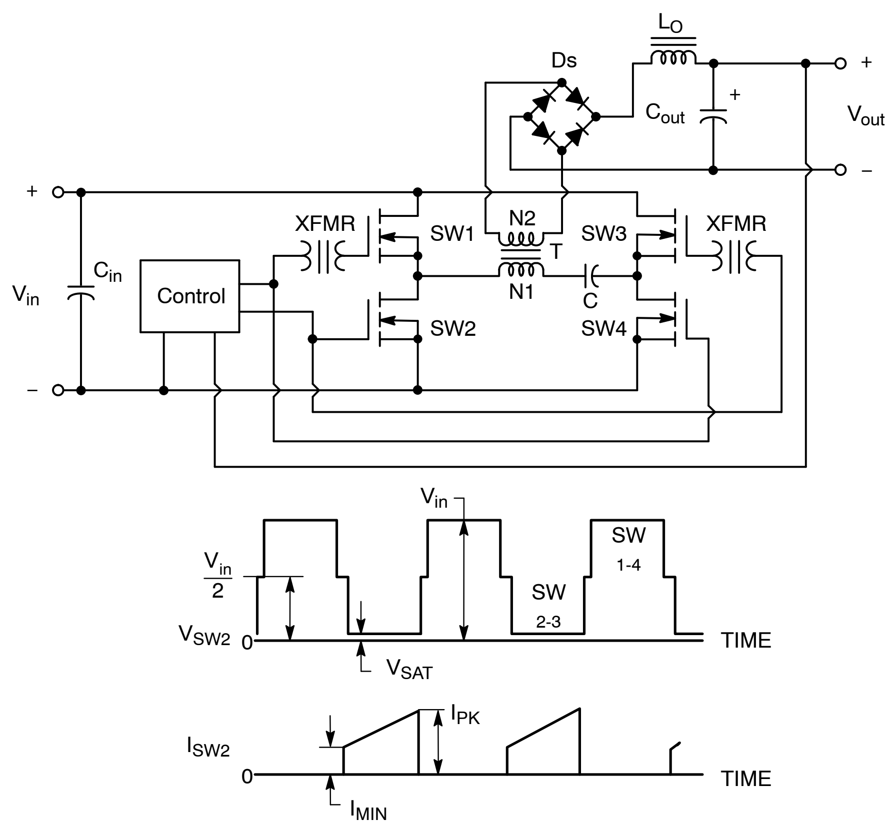


Figure 12. The Full−Bridge Converter

Interleaved Multiphase Converters

One method of increasing the output power of any topology and reducing the stresses upon the semiconductors, is a technique called interleaving. Any topology can be interleaved. An *interleaved multiphase* converter has two or more identical converters placed in parallel which share key components. For an *n*−phase converter, each converter is driven at a phase difference of 360/*n* degrees from the next. The output current from all the phases sum together at the output, requiring only I_{out}/n amperes from each phase.

The input and output capacitors are shared among the phases. The input capacitor sees less RMS ripple current because the peak currents are less and the combined duty cycle of the phases is greater than it would experience with a single phase converter. The output capacitor can be made smaller because the frequency of current waveform is *n*−times higher and its combined duty cycle is greater. The semiconductors also see less current stress.

A block diagram of an interleaved multiphase buck converter is shown in Figure 13. This is a 2−phase topology that is useful in providing power to a high performance microprocessor.

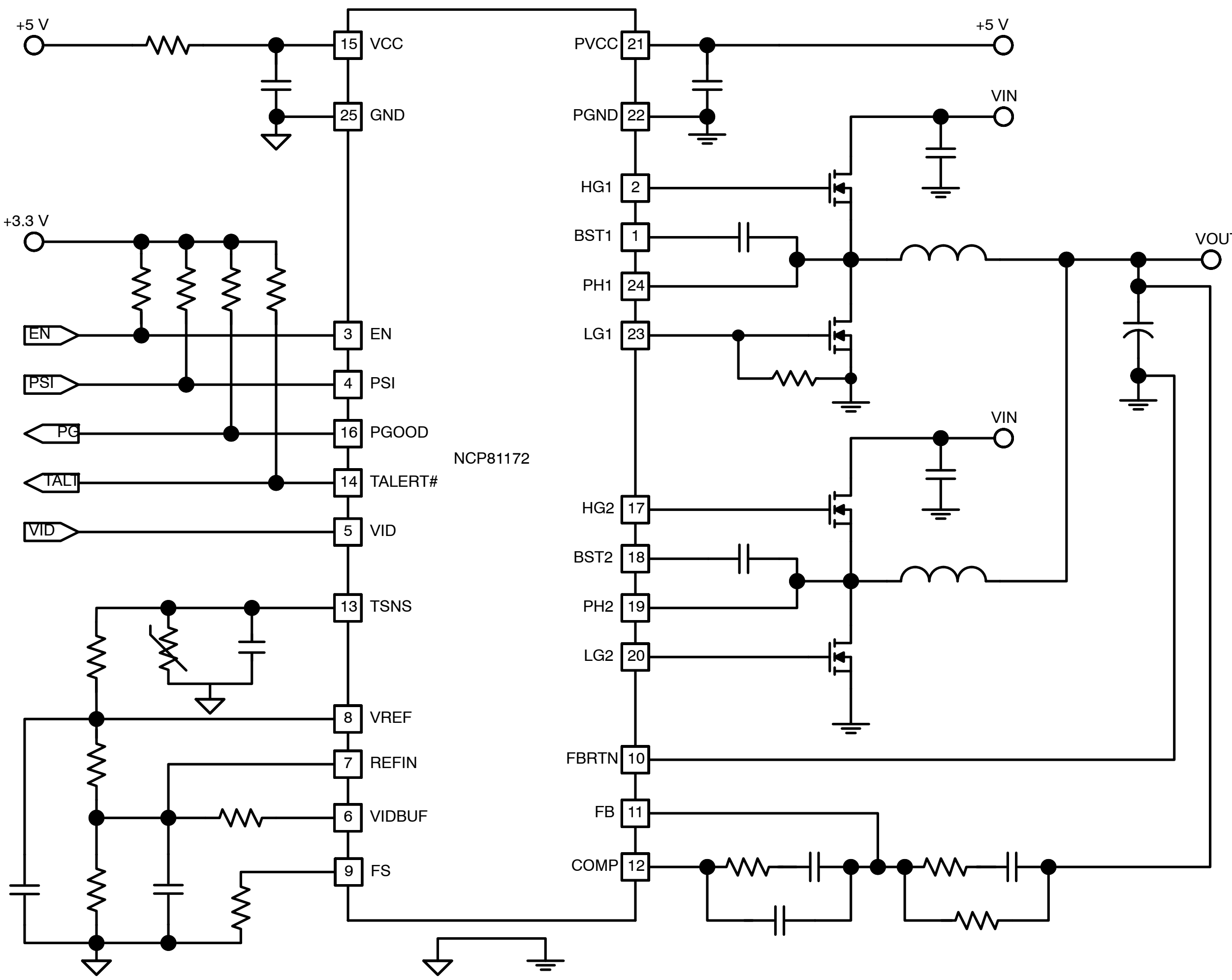


Figure 13. Typical Application of a Two−Phase Synchronous Buck Controller

Selecting the Method of Control

There are three major methods of controlling a switching power supply. There are also variations of these control methods that provide additional protection features. One should review these methods carefully and then carefully review the controller IC data sheets to select the one that is wanted.

Table 1 summarizes the features of each of the popular methods of control. Certain methods are better adapted to certain topologies due to reasons of stability or transient response.

Table 1. Common Control Methods Used in ICs

Control Method	OC Protection	Response Time	Preferred Topologies
Voltage−Mode	Average OC	Slow	Forward−Mode
	Pulse−by−Pulse OC	Slow	Forward−Mode
Current−Mode	Intrinsic	Rapid	Boost−Mode
	Hysteretic	Rapid	Boost & Forward−Mode
Hysteric Voltage	Average	Slow	Boost & Forward−Mode

Voltage−mode control (see Figure 14) is typically used for forward−mode topologies. In voltage−mode control, only the output voltage is monitored. A voltage error signal is calculated by forming the difference between Vout (actual) and Vout(desired). This error signal is then fed into a comparator that compares it to the ramp voltage generated by the internal oscillator section of the control IC. The comparator thus converts the voltage error signal into the PWM drive signal to the power switch. Since the only control parameter is the output voltage, and there is inherent delay through the power circuit, voltage−mode control tends to respond slowly to input variations.

Overcurrent protection for a voltage−mode controlled converter can either be based on the average output current or use a pulse−by−pulse method. In *average overcurrent protection,* the DC output current is monitored, and if a threshold is exceeded, the pulse width of the power switch is reduced. In *pulse−by−pulse overcurrent protection,* the peak current of each power switch "on" cycle is monitored and the power switch is instantly cutoff if its limits are exceeded. This offers better protection to the power switch.

Current−mode control (see Figure 15) is typically used with boost−mode converters. Current−mode control monitors not only the output voltage, but also the output current. Here the voltage error signal is used to control the peak current within the magnetic elements during each power switch on−time. Current−mode control has a very rapid input and output response time, and has an inherent overcurrent protection. It is not commonly used for forward−mode converters; their current waveforms have much lower slopes in their current waveforms which can create jitter within comparators.

Hysteretic control is a method of control which tries to keep a monitored parameter between two limits. There are hysteretic current and voltage control methods, but they are not commonly used.

The designer should be very careful when reviewing a prospective control IC data sheet. The method of control and any variations are usually not clearly described on the first page of the data sheet.

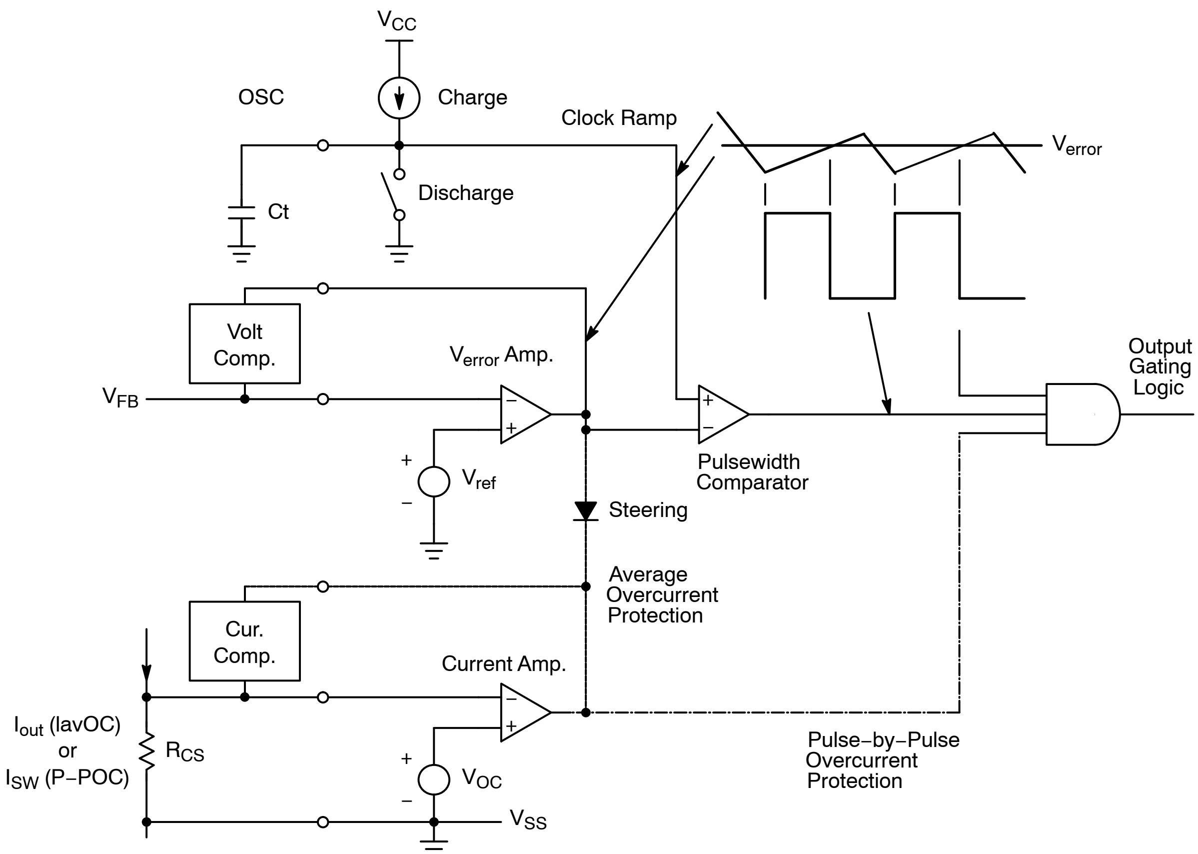


Figure 14. Voltage−Mode Control

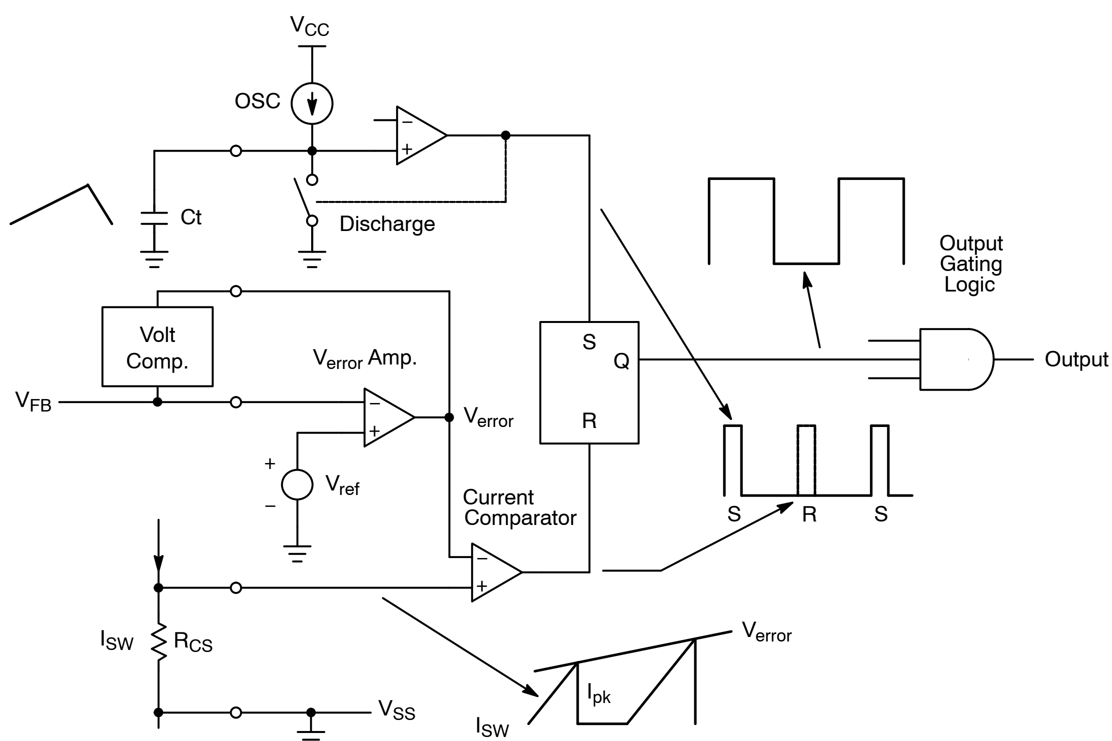


Figure 15. Turn−On with Clock Current−Mode Control

The Choice of Semiconductors

Power Switches

The choice of which semiconductor technology to use for the power switch function is influenced by many factors such as cost, peak voltage and current, frequency of operation, and heatsinking. Each technology has its own peculiarities that must be addressed during the design phase.

There are three major power switch choices: the bipolar junction transistor (BJT), the power MOSFET, and the integrated gate bipolar transistor (IGBT). The BJT was the first power switch to be used in this field and still offers many cost advantages over the others. It is also still used for very low cost or in high power switching converters. The maximum frequency of operation of bipolar transistors is less than 80−100 kHz because of some of their switching characteristics. The IGBT is used for high power switching converters, displacing many of the BJT applications. They too, though, have a slower switching characteristic which limits their frequency of operation to below 30 kHz typically although some can reach 100 kHz. IGBTs have smaller die areas than power MOSFETs of the same ratings, which typically means a lower cost. Power MOSFETs are used in the majority of applications due to their ease of use and their higher frequency capabilities. Each of the technologies will be reviewed.

The Bipolar Power Transistor

The BJT is a current driven device. That means that the base current is in proportion to the current drawn through the collector. So one must provide:

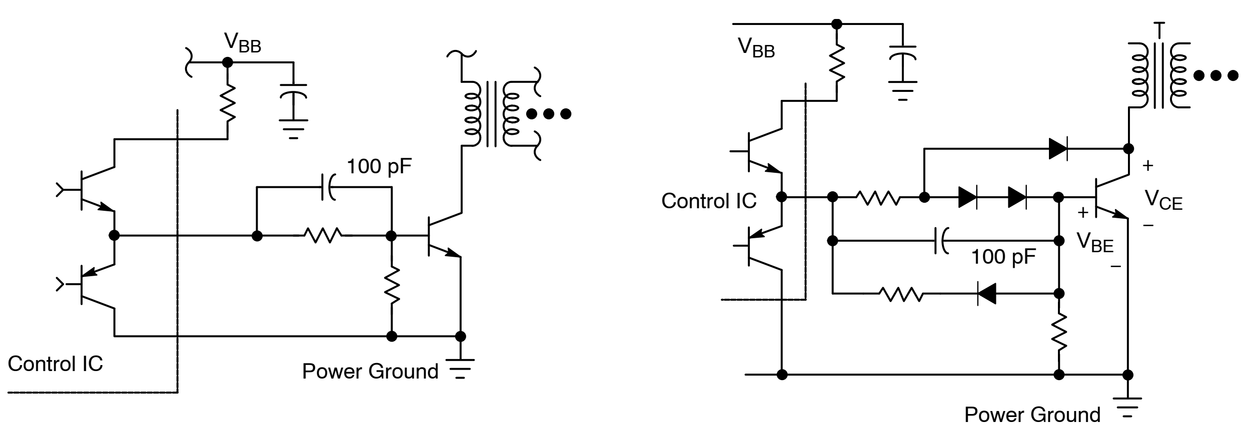
$$I_B > I_C / h_{FE} \qquad \text{(eq. 8)}$$

In power transistors, the average gain (h_{FE}) exhibited at the higher collector currents is between 5 and 20. This could create a large base drive loss if the base drive circuit is not properly designed.

One should generate a gate drive voltage that is as close to 0.7 volts as possible. This is to minimize any loss created by dropping the base drive voltage at the required base current to the level exhibited by the base.

A second consideration is the *storage time* exhibited by the collector during its turn−off transition. When the base is overdriven, or where the base current is more than needed to sustain the collector current, the collector exhibits a 0.3−2 μs delay in its turn−off which is proportional to the base overdrive. Although the storage time is not a major source of loss, it does significantly limit the maximum switching frequency of a bipolar−based switching power supply. There are two methods of reducing the storage time and increasing its switching time. The first is to use a base speed−up capacitor whose value, typically around 100 pF, is placed in parallel with the base current limiting resistor (Figure 16a). The second is to use proportional base drive (Figure 16b). Here, only the amount of needed base current is provided by the drive circuit by bleeding the excess around the base into the collector.

The last consideration with BJTs is the risk of excessive *second breakdown*. This phenomenon is caused by the resistance of the base across the die, permitting the furthest portions of the collector to turn off later. This forces the current being forced through the collector by an inductive load, to concentrate at the opposite ends of the die, thus causing an excessive localized heating on the die. This can result in a short−circuit failure of the BJT which can happen instantaneously if the amount of current crowding is great, or it can happen later if the amount of heating is less. Current crowding is always present when an inductive load is attached to the collector. By switching the BJT faster, with the circuits in Figure 15, one can greatly reduce the effects of second breakdown on the reliability of the device.



(a) Fixed Base Drive Circuit **(b) Proportional Base Drive Circuit (Baker Clamp)**

Figure 16. Driving a Bipolar Junction Transistor

The Power MOSFET

Power MOSFETs are the popular choices used as power switches and synchronous rectifiers. They are, on the surface, simpler to use than BJTs, but they have some hidden complexities.

A simplified model for a MOSFET can be seen in Figure 17. The capacitances seen in the model are specified within the MOSFET data sheets, but can be nonlinear and vary with their applied voltages.

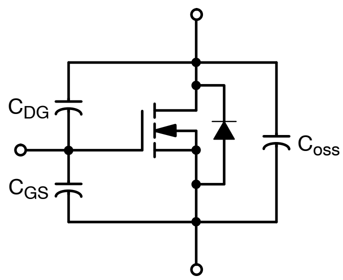


Figure 17. The MOSFET Model

From the gate terminal, there are two capacitances the designer encounters, the gate input capacitance (C_{iss}) and the drain−gate reverse capacitance (C_{rss}). The gate input capacitance is a fixed value caused by the capacitance formed between the gate metalization and the substrate. Its value usually falls in the range of 800−3200 pF, depending upon the physical construction of the MOSFET. The C_{rss} is the capacitance between the drain and the gate, and has values in the range of 60−150 pF. Although the C_{rss} is smaller, it has a much more pronounced effect upon the gate drive. It couples the drain voltage to the gate, thus dumping its stored charge into the gate input capacitance. The typical gate drive waveforms can be seen in Figure 18. Time period t1 is only the C_{iss} being charged or discharged by the impedance of the external gate drive circuit. Period t2 shows the effect of the changing drain voltage being coupled into the gate through C_{rss}. One can readily observe the “flattening” of the gate drive voltage during this period, both during the turn−on and turn−off of the MOSFET. Time period t3 is the amount of overdrive voltage provided by the drive circuit but not really needed by the MOSFET.

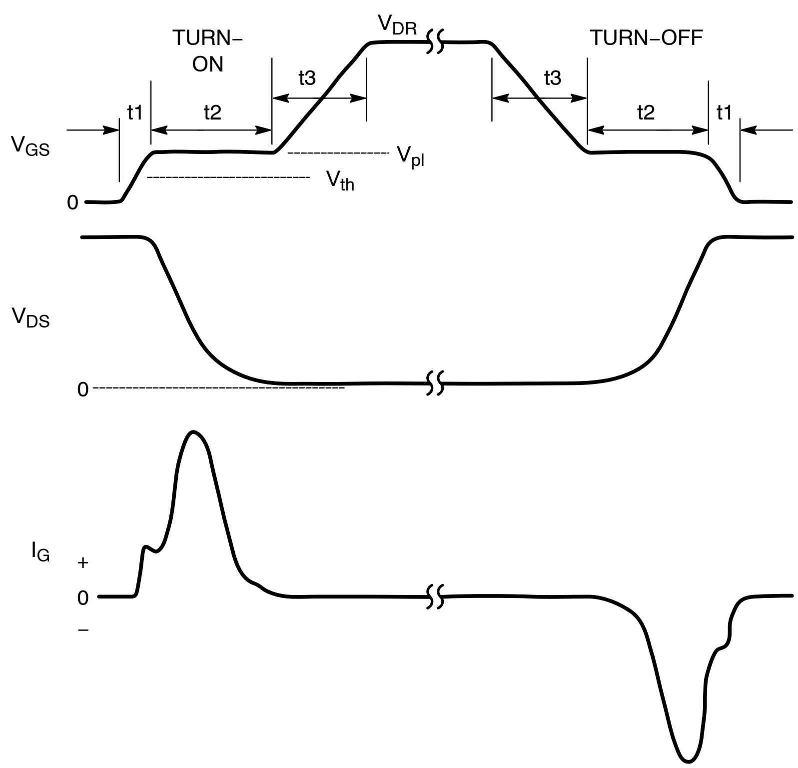


Figure 18. Typical MOSFET Drive Waveforms (Top: V_{GS}, Middle: V_{DG}, Bottom: I_G)

The time needed to switch the MOSFET between on and off states is dependent upon the impedance of the gate drive circuit. It is very important that the drive circuit be bypassed with a capacitor that will keep the drive voltage constant over the drive period. A 0.1 μF capacitor is more than sufficient.

Driving MOSFETs in Switching Power Supply Applications

There are three things that are very important in the high frequency driving of MOSFETs: there must be a totem−pole driver; the drive voltage source must be well bypassed; and the drive devices must be able to source high levels of current in very short periods of time (low compliance). The optimal drive circuit is shown in Figure 19.

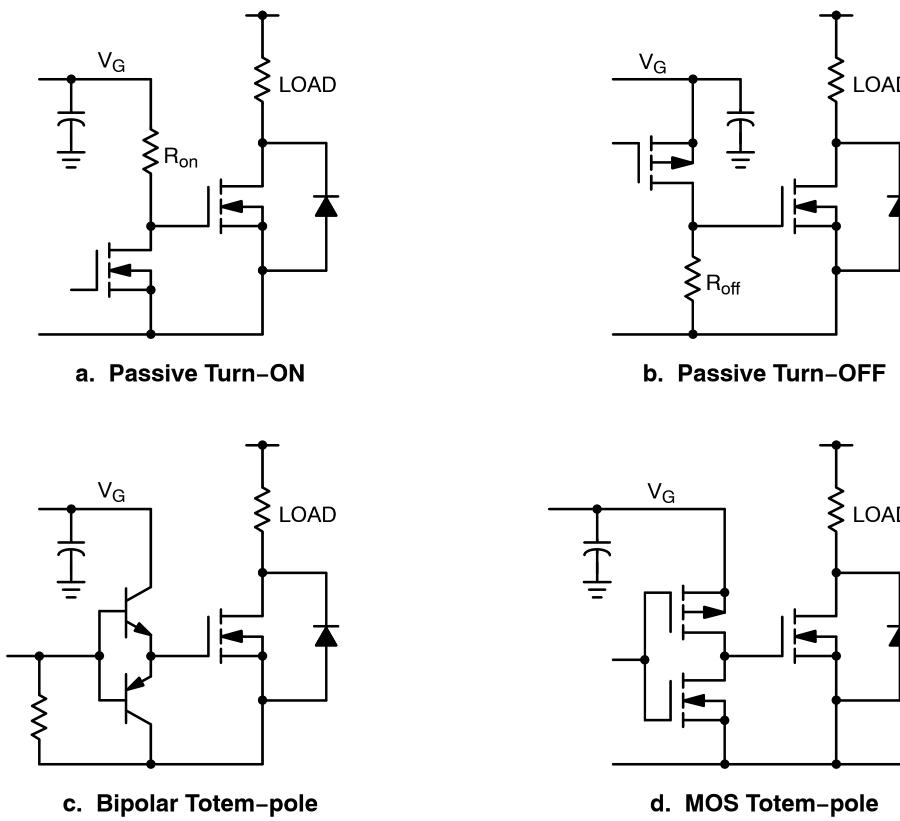


Figure 19. Bipolar and FET−Based Drive Circuits (a. Bipolar Drivers, b. MOSFET Drivers)

Sometimes it is necessary to provide a dielectrically−isolated drive to a MOSFET. This is provided by a drive transformer. Transformers driven from a DC source must be capacitively coupled from the totem−pole driver circuit. The secondary winding must be capacitively coupled to the gate with a DC restoration circuit. Both of the series capacitors must be more than 10 times the value of the C_{iss} of the MOSFET so that the capacitive voltage divider that is formed by the series capacitors does not cause an excessive attenuation. The circuit can be seen in Figure 20.

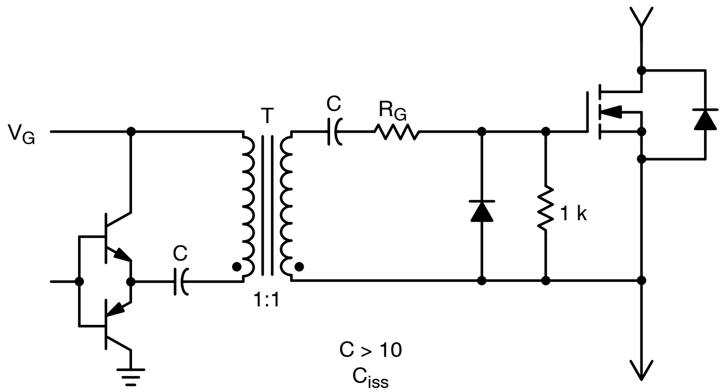


Figure 20. Transformer−Isolated Gate Drive

The Insulated Gate Bipolar Transistor (IGBT)

The IGBT is a hybrid device with a MOSFET as the input device, which then drives a silicon−controlled rectifier (SCR) as a switched output device. The SCR is constructed such that it does not exhibit the latching characteristic of a typical SCR by making its feedback gain less than 1. The die area of the typical IGBT is less than one−half that of an identically rated power MOSFET, which makes it less expensive for high−power converters. The only drawback is the turn−off characteristic of the IGBT. Being a bipolar minority carrier device, charges must be removed from the P−N junctions during a turn−off condition. This causes a “current tail” at the end of the turn−off transition of the current waveform. This can be a significant loss because the voltage across the IGBT is very high at that moment. This makes the IGBT useful only for frequencies typically less than 20 kHz, or for exceptional IGBTs, 100 kHz.

To drive an IGBT one uses the MOSFET drive circuits shown in Figures 18 and 19. Driving the IGBT gate faster makes very little difference in the performance of an IGBT, so some reduction in drive currents can be used.

The voltage drop of across the collector−to−emitter (V_{CE}) terminals is comparable to those found in Darlington BJTs and MOSFETs operated at high currents. The typical V_{CE} of an IGBT is a flat 1.5−2.2 volts. MOSFETs, acting more resistive, can have voltage drops of up to 5 volts at the end of some high current ramps. This makes the IGBT, in high current environments, very comparable to MOSFETs in applications of less than 5−30 kHz.

Rectifiers

Rectifiers represent about 60 percent of the losses in nonsynchronous switching power supplies. Their choice has a very large effect on the efficiency of the power supply.

The significant rectifier parameters that affect the operation of switching power supplies are:

- *forward voltage drop* (V_f), which is the voltage across the diode when a forward current is flowing
- the *reverse recovery time* (t_{rr}), which is how long it requires a diode to clear the minority charges from its junction area and turn off when a reverse voltage is applied
- the *forward recovery time* (t_{frr}) which is how long it take a diode to begin to conduct forward current after a forward voltage is applied.

There are four choices of rectifier technologies: standard, fast and ultra−fast recovery types, and Schottky barrier types.

A **standard recovery diode** is only suitable for 50−60 Hz rectification due to its slow turn−off characteristics. These include common families such as the 1N4000 series diodes. **Fast−recovery diodes** were first used in switching power supplies, but their turn−off time is considered too slow for most modern applications. They may find application where low cost is paramount, however. **Ultra−fast recovery diodes** turn off quickly and have a forward voltage drop of 0.8 to 1.3 V, together with a high reverse voltage capability of up to 1000 V. A **Schottky rectifier** turns off very quickly and has an average forward voltage drop of between 0.35 and 0.8 V, but has a low reverse breakdown voltage and a high reverse leakage current. For a typical switching power supply application, the best choice is usually a Schottky rectifier for output voltages less than 12 V, and an ultra−fast recovery diode for all other output voltages.

The major losses within output rectifiers are conduction losses and switching losses. The conduction loss is the forward voltage drop times the current flowing through it during its conduction period. This can be significant if its voltage drop and current are high. The switching losses are determined by how fast a diode turns off (t_{rr}) times the reverse voltage across the rectifier. This can be significant for high output voltages and currents.

The characteristics of power rectifiers and their applications in switching power supplies are covered in great detail in Reference (5).

The major losses within output rectifiers are conduction losses and switching losses. The conduction loss is the forward voltage drop times the current flowing through it during its conduction period. This can be significant if its voltage drop and current are high. The switching losses are determined by how fast a diode turns off (t_{rr}) times the reverse voltage across the rectifier. This can be significant for high output voltages and currents.

Table 2. Types of Rectifier Technologies

Rectifier Type	Average V_f	Reverse Recovery Time	Typical Applications
Standard Recovery	0.7−1.0 V	1,000 ns	50−60 Hz Rectification
Fast Recovery	1.0−1.2 V	150−200 ns	Output Rectification
UltraFast Recovery	0.9−1.4 V	25−75 ns	Output Rectification (Vo > 12 V)
Schottky	0.3−0.8 V	< 10 ns	Output Rectification (Vo < 12 V)

Table 3. Estimating the Significant Parameters of the Power Semiconductors

Topology	Bipolar Pwr Sw		MOSFET Pwr Sw		Rectifier	
	V_{CEO}	I_C	V_{DSS}	I_D	V_R	I_F
Buck	V_{in}	I_{out}	V_{in}	I_{out}	V_{in}	I_{out}
Boost	V_{out}	$\frac{(2.0\ P_{out})}{V_{in(min)}}$	V_{out}	$\frac{(2.0\ P_{out})}{V_{in(min)}}$	V_{out}	I_{out}
Buck/Boost	$V_{in} - V_{out}$	$\frac{(2.0\ P_{out})}{V_{in(min)}}$	$V_{in} - V_{out}$	$\frac{(2.0\ P_{out})}{V_{in(min)}}$	$V_{in} - V_{out}$	I_{out}
Flyback	$1.7\ V_{in(max)}$	$\frac{(2.0\ P_{out})}{V_{in(min)}}$	$1.5\ V_{in(max)}$	$\frac{(2.0\ P_{out})}{V_{in(min)}}$	$5.0\ V_{out}$	I_{out}
1 Transistor Forward	$2.0\ V_{in}$	$\frac{(1.5\ P_{out})}{V_{in(min)}}$	$2.0\ V_{in}$	$\frac{(1.5\ P_{out})}{V_{in(min)}}$	$3.0\ V_{out}$	I_{out}
Push−Pull	$2.0\ V_{in}$	$\frac{(1.2\ P_{out})}{V_{in(min)}}$	$2.0\ V_{in}$	$\frac{(1.2\ P_{out})}{V_{in(min)}}$	$2.0\ V_{out}$	I_{out}
Half−Bridge	V_{in}	$\frac{(2.0\ P_{out})}{V_{in(min)}}$	V_{in}	$\frac{(2.0\ P_{out})}{V_{in(min)}}$	$2.0\ V_{out}$	I_{out}
Full−Bridge	V_{in}	$\frac{(1.2\ P_{out})}{V_{in(min)}}$	V_{in}	$\frac{(2.0\ P_{out})}{V_{in(min)}}$	$2.0\ V_{out}$	I_{out}

The Magnetic Components

The magnetic elements within a switching power supply are used either for stepping−up or down a switched AC voltage, or for energy storage. In forward−mode topologies, the transformer is only used for stepping−up or down the AC voltage generated by the power switches. The output filter (the output inductor and capacitor) in forward−mode topologies is used for energy storage. In boost−mode topologies, the transformer is used both for energy storage and to provide a step−up or step−down function.

Many design engineers consider the magnetic elements of switching power supplies counter−intuitive or too complicated to design. Fortunately, help is at hand; the suppliers of magnetic components have applications engineers who are quite capable of performing the transformer design and discussing the tradeoffs needed for success. For those who are more experienced or more adventuresome, please refer to Reference 2 in the Bibliography for transformer design guidelines.

The general procedure in the design of any magnetic component is as follows (Reference 2, p 42):

1. Select an appropriate core material for the application and the frequency of operation.
2. Select a core form factor that is appropriate for the application and that satisfies applicable regulatory requirements.
3. Determine the core cross−sectional area necessary to handle the required power
4. Determine whether an airgap is needed and calculate the number of turns needed for each winding. Then determine whether the accuracy of the output voltages meets the requirements and whether the windings will fit into the selected core size.
5. Wind the magnetic component using proper winding techniques.
6. During the prototype stage, verify the component’s operation with respect to the level of voltage spikes, cross−regulation, output accuracy and ripple, RFI, etc., and make corrections were necessary.

The design of any magnetic component is a “calculated estimate.” There are methods of “stretching” the design limits for smaller size or lower losses, but these tend to be diametrically opposed to one another. One should be cautious when doing this.

Some useful sources for magnetics components are:

CoilCraft, Inc.
website: www.coilcraft.com/
email: info@coilcraft.com
Telephone: 847−639−6400

Coiltronics, Division of Cooper Electronics Technology
website: www.coiltronics.com
Telephone: 561−241−7876

Cramer Coil, Inc.
website: www.cramerco.com
email: techsales@cramercoil.com
Telephone: 262−268−2150

Pulse, Inc.
website: www.pulseeng.com
Telephone: 858−674−8100

TDK
website: www.component.talk.com
Telephone: 847−803−6100

Würth Elektronik
website: www.we−online.com
email: cbt@we−online.com
Telephone: +49 7940 946−0

Laying Out the Printed Circuit Board

The printed circuit board (PCB) layout is the third critical portion of every switching power supply design in addition to the basic design and the magnetics design. Improper layout can adversely affect RFI radiation, component reliability, efficiency and stability. Every PCB layout will be different, but if the designer appreciates the common factors present in all switching power supplies, the process will be simplified.

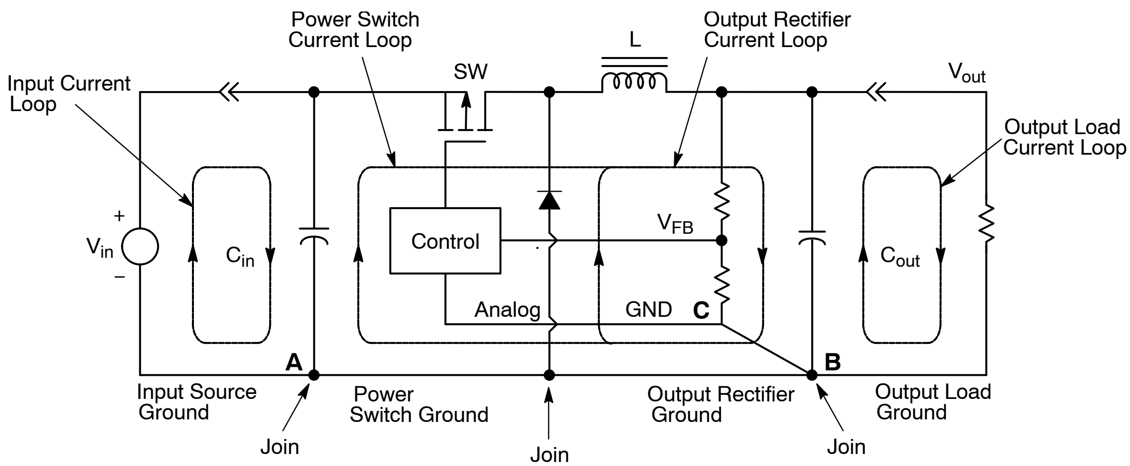
All PCB traces exhibit inductance and resistance. These can cause high voltage transitions whenever there is a high rate of change in current flowing through the trace. For operational amplifiers sharing a trace with power signals, it means that the supply would be impossible to stabilize. For traces that are too narrow for the current flowing through them, it means a voltage drop from one end of the trace to the other which potentially can be an antenna for RFI. In addition, capacitive coupling between adjacent traces can interfere with proper circuit operation.

There are two rules of thumb for PCB layouts: “short and fat” for all power−carrying traces and “one point grounding” for the various ground systems within a switching power supply. Traces that are short and fat minimize the inductive and resistive aspects of the trace, thus reducing noise within the circuits and RFI. Single−point grounding keeps the noise sources separated from the sensitive control circuits.

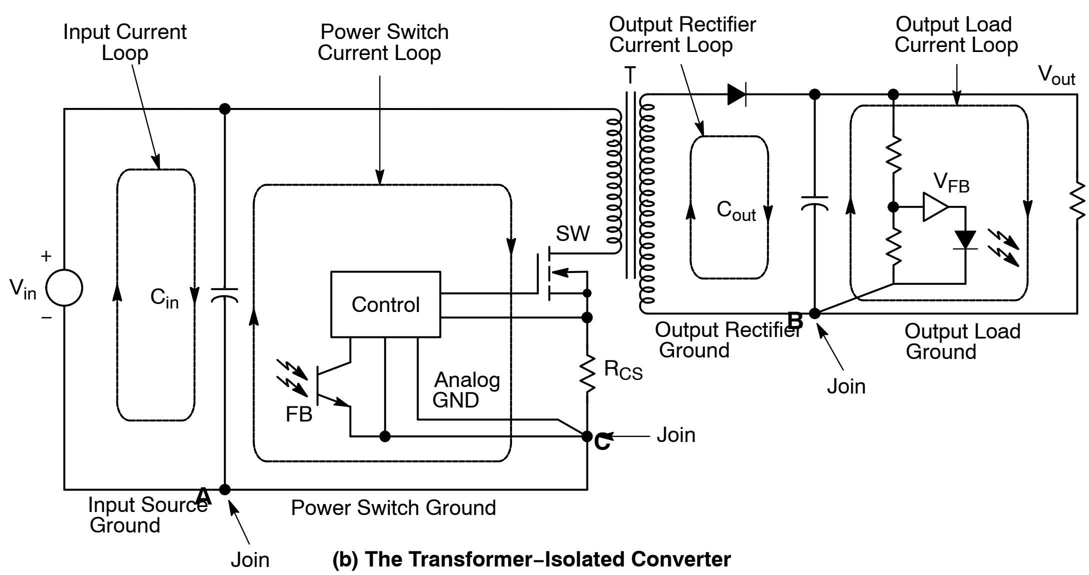
Within all switching power supplies, there are four major current loops. Two of the loops conduct the high−level AC currents needed by the supply. These are the power switch AC current loop and the output rectifier AC current loop. The currents are the typical trapezoidal current pulses with very high peak currents and very rapid di/dts. The other two current loops are the input source and the output load current loops, which carry low frequency current being supplied from the voltage source and to the load respectively.

For the power switch AC current loop, current flows from the input filter capacitor through the inductor or transformer winding, through the power switch and back to the negative pin of the input capacitor. Similarly, the output rectifier current loop's current flows from the inductor or secondary transformer winding, through the rectifier to the output filter capacitor and back to the inductor or winding. The filter capacitors are the only components that can source and sink the large levels of AC current in the time needed by the switching power supply. The PCB traces should be made as wide and as short as possible, to minimize resistive and inductive effects. These traces should be the first to be laid out.

Turning to the input source and output load current loops, both of these loops must be connected directly to their respective filter capacitor's terminals, otherwise switching noise could bypass the filtering action of the capacitor and escape into the environment. This noise is called conducted interference. These loops can be seen in Figure 21 for the two major forms of switching power supplies, non−isolated (Figure 21a) and transformer−isolated (Figure 21b).



(a) The Non−Isolated DC/DC Converter



(b) The Transformer−Isolated Converter

Figure 21. The Current Loops and Grounds for the Major Converter Topologies

The grounds are extremely important to the proper operation of the switching power supply, since they form the reference connections for the entire supply; each ground has its own unique set of signals which can adversely affect the operation of the supply if connected improperly.

There are five distinct grounds within the typical switching power supply. Four of them form the return paths for the current loops described above. The remaining ground is the low−level analog control ground which is critical for the proper operation of the supply. The grounds which are part of the major current loops must be connected together exactly as shown in Figure 21. Here again, the connecting point between the high−level AC grounds and the input or output grounds is at the negative terminal of the appropriate filter capacitor (points A and B in Figures 21a and 21b). Noise on the AC grounds can very easily escape into the environment if the grounds are not directly connected to the negative terminal of the filter capacitor(s). The analog control ground must be connected to the point where the control IC and associated circuitry must measure key power parameters, such as AC or DC current and the output voltage (point C in Figures 21a and 21b). Here any noise introduced by large AC signals within the AC grounds will sum directly onto the low−level control parameters and greatly affect the operation of the supply. The purpose of connecting the control ground to the lower side of the current sensing resistor or the output voltage resistor divider is to form a “Kelvin contact” where any common mode noise is not sensed by the control circuit. In short, follow the example given by Figure 21 exactly as shown for best results.

The last important factor in the PCB design is the layout surrounding the AC voltage nodes. These are the drain of the power MOSFET (or collector of a BJT) and the anode of the output rectifier(s). These nodes can capacitively couple into any trace on different layers of the PCB that run underneath the AC pad. In surface mount designs, these nodes also need to be large enough to provide heatsinking for the power switch or rectifier. This is at odds with the desire to keep the pad as small as possible to discourage capacitive coupling to other traces. One good compromise is to make all layers below the AC node identical to the AC node and connect them with many vias (plated−through holes). This greatly increases the thermal mass of the pad for improved heatsinking and locates any surrounding traces off laterally where the coupling capacitance is much smaller. An example of this can be seen in Figure 22.

Many times it is necessary to parallel filter capacitors to reduce the amount of RMS ripple current each capacitor experiences. Close attention should be paid to this layout. If the paralleled capacitors are in a line, the capacitor closest to the source of the ripple current will operate hotter than the others, shortening its operating life; the others will not see this level of AC current. To ensure that they will evenly share the ripple current, ideally, any paralleled capacitors should be laid out in a radially−symmetric manner around the current source, typically a rectifier or power switch.

The PCB layout, if not done properly, can ruin a good paper design. It is important to follow these basic guidelines and monitor the layout every step of the process.

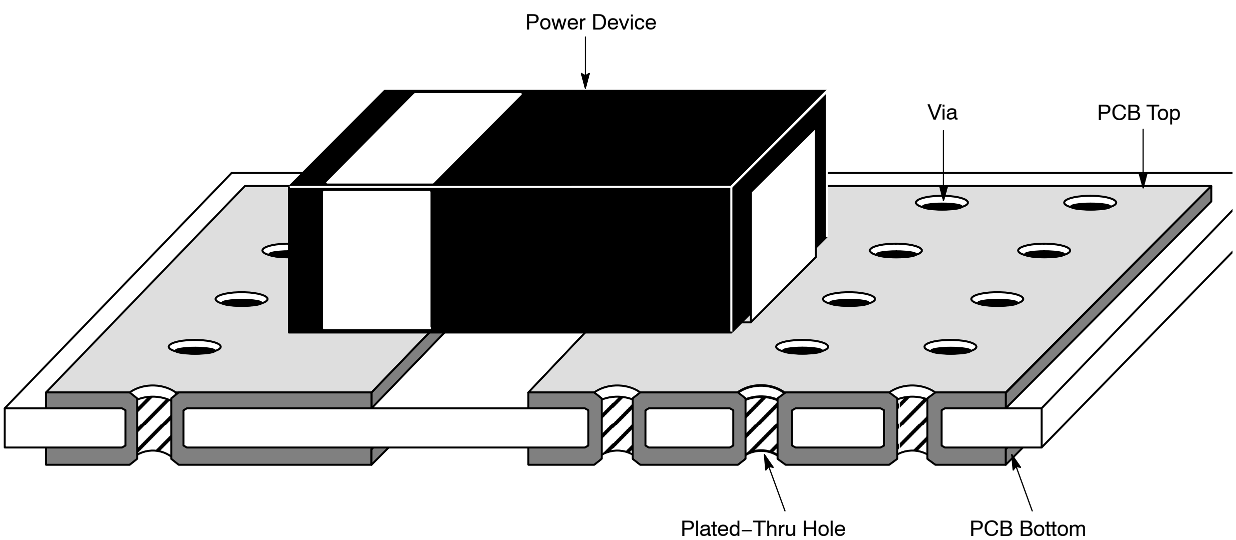


Figure 22. Method for Minimizing AC Capacitive Coupling and Enhancing Heatsinking

Losses and Stresses in Switching Power Supplies

Much of the designer’s time during a switching power supply design is spent in identifying and minimizing the losses within the supply. Most of the losses occur in the power components within the switching power supply. Some of these losses can also present stresses to the power semiconductors which may affect the long term reliability of the power supply, so knowing where they arise and how to control them is important.

Whenever there is a simultaneous voltage drop across a component with a current flowing through, there is a loss. Some of these losses are controllable by modifying the circuitry, and some are controlled by simply selecting a different part. Identifying the major sources for loss can be as easy as placing a finger on each of the components in search of heat, or measuring the currents and voltages associated with each power component using an oscilloscope, AC current probe and voltage probe.

Semiconductor losses fall into two categories: conduction losses and switching losses. The *conduction loss* is the product of the terminal voltage and current during the power device’s on period. Examples of conduction losses are the saturation voltage of a bipolar power transistor and the “on” loss of a power MOSFET shown in Figure 23 and Figure 24 respectively.

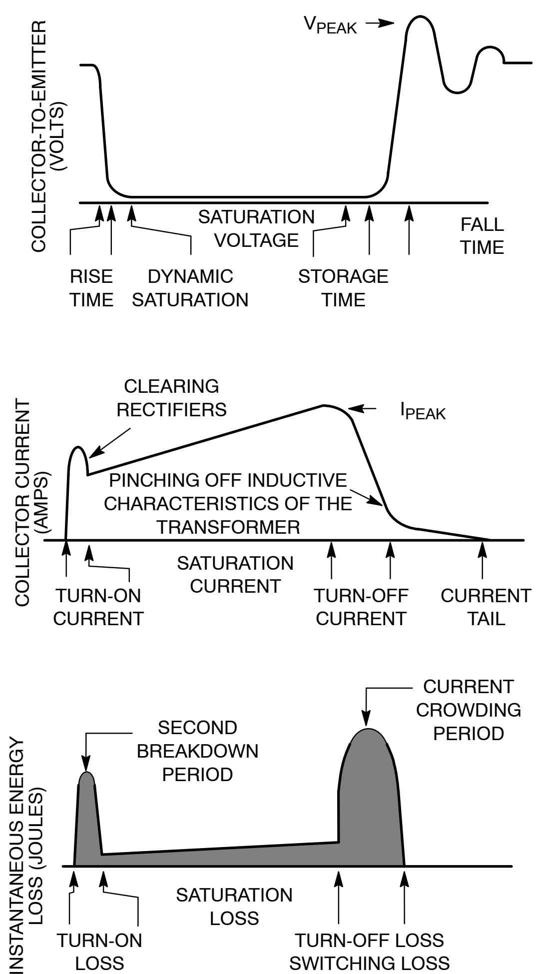


Figure 23. Stresses and Losses within a Bipolar Power Transistor

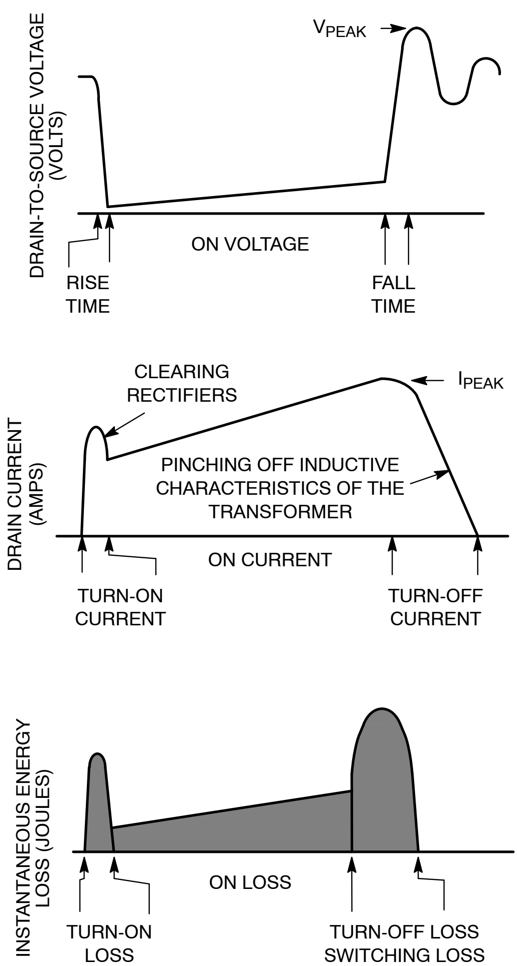


Figure 24. Stresses and Losses within a Power MOSFET

The forward conduction loss of a rectifier is shown in Figure 25. During turn−off, the rectifier exhibits a *reverse recovery loss* where minority carriers trapped within the P−N junction must reverse their direction and exit the junction after a reverse voltage is applied. This results in what appears to be a current flowing in reverse through the diode with a high reverse terminal voltage.

The *switching loss* is the instantaneous product of the terminal voltage and current of a power device when it is transitioning between operating states (on−to−off and off−to−on). Here, voltages are transitional between full−on and cutoff states while simultaneously the current is transitional between full−on and cut−off states. This creates a very large V−I product which is as significant as the conduction losses. Switching losses are also the major frequency dependent loss within every PWM switching power supply.

The loss−induced heat generation causes stress within the power component. This can be minimized by an effective thermal design. For bipolar power transistors, however, excessive switching losses can also provide a lethal stress to the transistor in the form of second breakdown and current crowding failures. Care should be taken in the careful analysis of each transistor's Forward Biased−Safe Operating Area (FBSOA) and Reverse Biased−Safe Operating Area (RBSOA) operation.

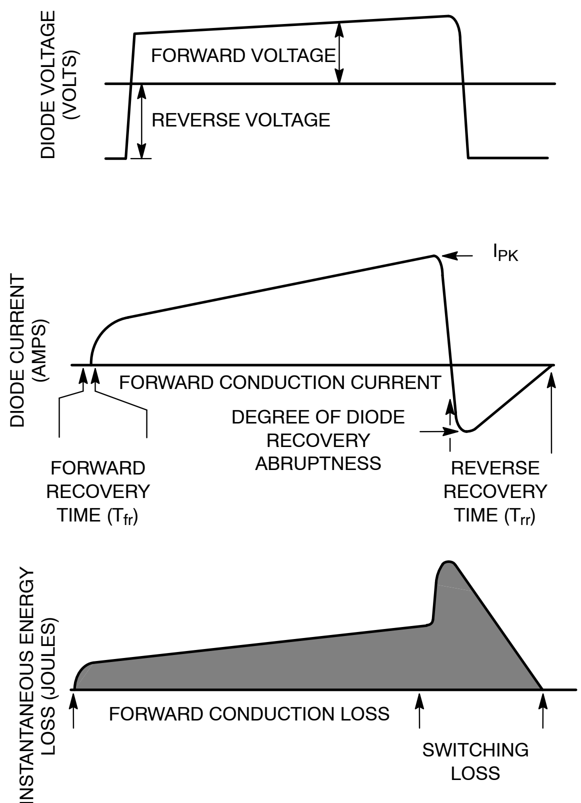


Figure 25. Stresses and Losses within Rectifiers

Techniques to Improve Efficiency in Switching Power Supplies

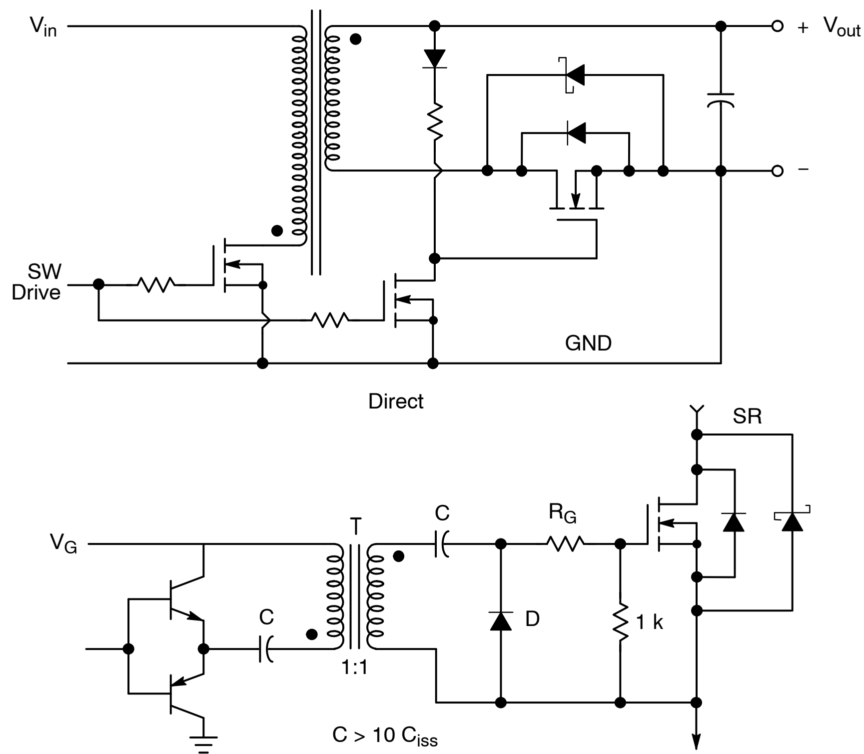
The reduction of losses is important to the efficient operation of a switching power supply, and a great deal of time is spent during the design phase to minimize these losses. Some common techniques are described below.

The Synchronous Rectifier

As output voltages decrease, the losses due to the output rectifier become increasingly significant. For V_{out} = 3.3 V, a typical Schottky diode forward voltage of 0.4 V leads to a 12% loss of efficiency. Synchronous rectification is a technique to reduce this conduction loss by using a switch in place of the diode. The synchronous rectifier switch is open when the power switch is closed, and closed when the power switch is open, and is typically a MOSFET inserted in place of the output rectifier. To prevent ”crowbar” current that would flow if both switches were closed at the same time, the switching scheme must be break−before−make. Because of this, a diode is still required to conduct the initial current during the interval between the opening of the main switch and the closing of the synchronous rectifier switch. A Schottky rectifier with a current rating of 30 percent of the MOSFET should be placed in parallel with the synchronous MOSFET. The MOSFET does contain a parasitic body diode that could conduct current, but it is lossy, slow to turn off, and can lower efficiency by 1% to 2%. The lower turn−on voltage of the Schottky prevents the parasitic diode from ever conducting and exhibiting its poor reverse recovery characteristic.

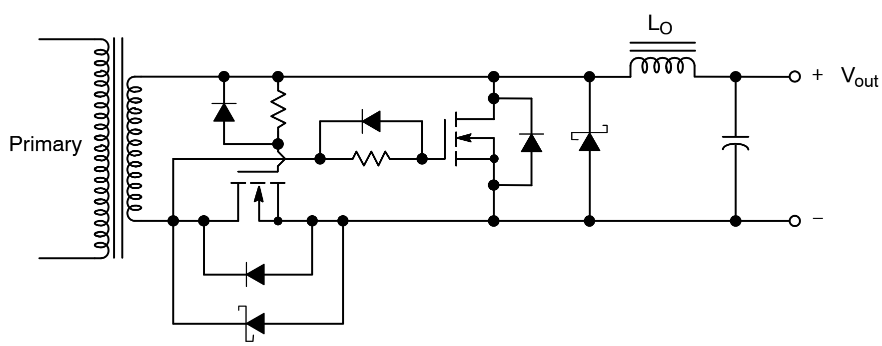
Using synchronous rectification, the conduction voltage can be reduced from 400 mV to 100 mV or less. An improvement of 1−5 percent can be expected for the typical switching power supply.

The synchronous rectifier can be driven either actively, that is directly controlled from the control IC, or passively, driven from other signals within the power circuit. It is very important to provide a non−overlapping drive between the power switch(es) and the synchronous rectifier(s) to prevent any shoot−through currents. This dead time is usually between 50 to 100 ns. Some typical circuits can be seen in Figure 26.



Transformer−Isolated

(a) Actively Driven Synchronous Rectifiers



(b) Passively Driven Synchronous Rectifiers

Figure 26. Synchronous Rectifier Circuits

Snubbers and Clamps

Snubbers and clamps are used for two very different purposes. When misapplied, the reliability of the semiconductors within the power supply is greatly jeopardized.

A snubber is used to reduce the level of a voltage spike and decrease the rate of change of a voltage waveform. This then reduces the amount of overlap of the voltage and current waveforms during a transition, thus reducing the switching loss. This has its benefits in the Safe Operating Area (SOA) of the semiconductors, and it reduces emissions by lowering the spectral content of any RFI.

A clamp is used only for reducing the level of a voltage spike. It has no affect on the dV/dt of the transition. Therefore it is not very useful for reducing RFI. It is useful for preventing components such as semiconductors and capacitors from entering avalanche breakdown.

Bipolar power transistors suffer from *current crowding* which is an instantaneous failure mode. If a voltage spike occurs during the turn−off voltage transition of greater than 75 percent of its VCEO rating, it may have too much current crowding stress. Here both the rate of change of the voltage and the peak voltage of the spike must be controlled. A snubber is needed to bring the transistor within its RBSOA (Reverse Bias Safe Operating Area) rating. Typical snubber and clamp circuits are shown in Figure 27. The effects that these have on a representative switching waveform are shown in Figure 28.

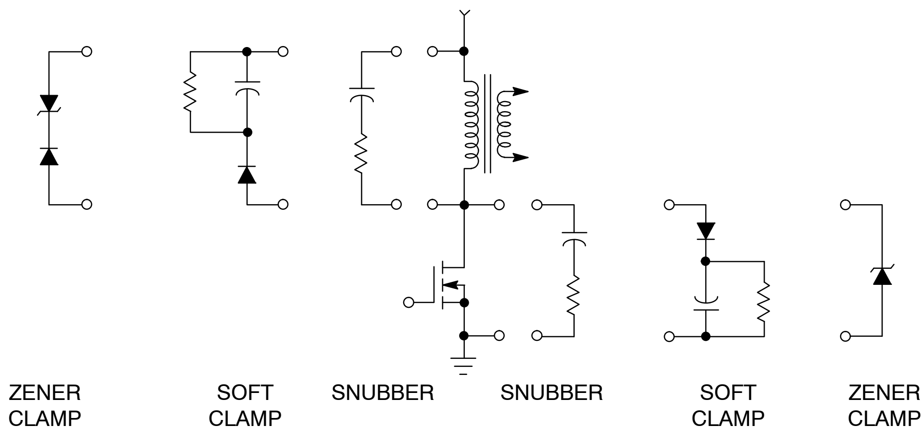


Figure 27. Common Methods for Controlling Voltage Spikes and/or RFI

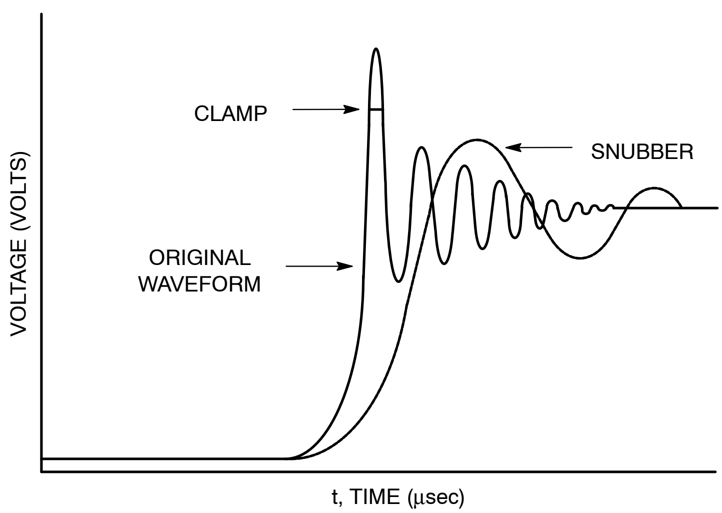


Figure 28. The Effects of a Snubber versus a Clamp

The Lossless Snubber

A lossless snubber is a snubber whose trapped energy is recovered by the power circuit. The lossless snubber is designed to absorb a fixed amount of energy from the transition of a switched AC voltage node. This energy is stored in a capacitor whose size dictates how much energy the snubber can absorb. A typical implementation of a lossless snubber can be seen in Figure 29.

The design for a lossless snubber varies from topology to topology and for each desired transition. Some adaptation may be necessary for each circuit. The important factors in the design of a lossless snubber are:

1. *The snubber must have initial conditions that allow it to operate during the desired transition and at the desired voltages.* Lossless snubbers should be emptied of their energy prior to the desired transition. The voltage to which it is reset dictates where the snubber will begin to operate. So if the snubber is reset to the input voltage, then it will act as a lossless clamp which will remove any spikes above the input voltage.
2. *When the lossless snubber is "reset," the energy should be returned to the input capacitor or back into the output power path.* Study the supply carefully. Returning the energy to the input capacitor allows the supply to use the energy again on the next cycle. Returning the energy to ground in a boost−mode supply does not return the energy for reuse, but acts as a shunt current path around the power switch. Sometimes additional transformer windings are used.
3. The reset current waveform should be band limited with a series inductor to prevent additional EMI from being generated. Use of a 2 to 3 turn spiral PCB inductor is sufficient to greatly lower the di/dt of the energy exiting the lossless snubber.

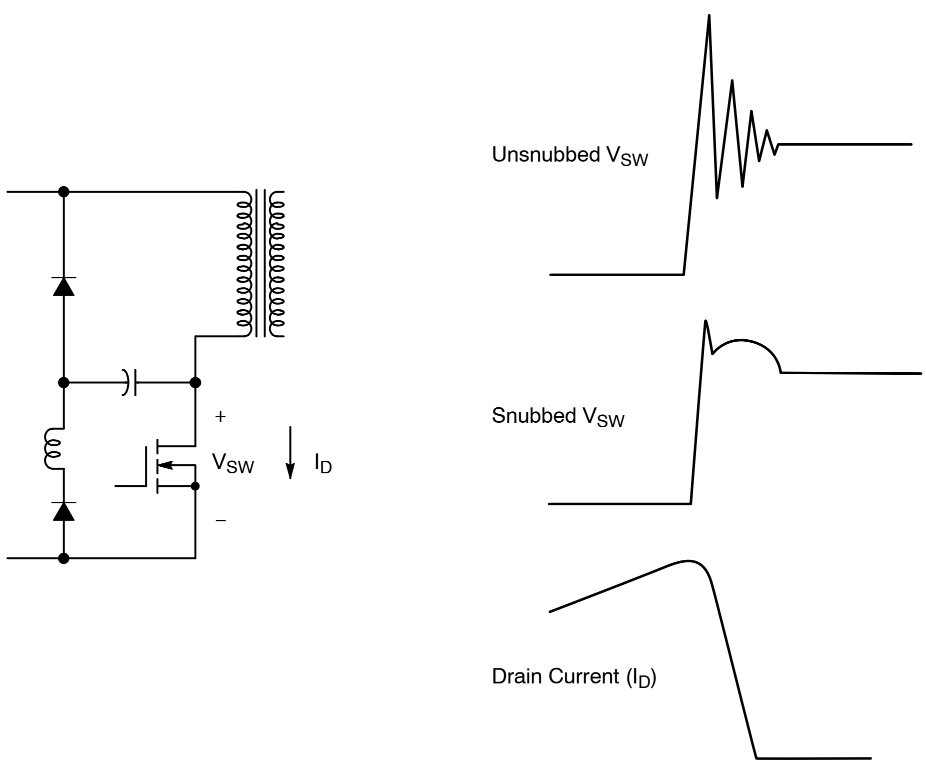


Figure 29. Lossless Snubber for a One Transistor Forward or Flyback Converter

The Active Clamp

An active clamp is a gated MOSFET circuit that allows the controller IC to activate a clamp or a snubber circuit at a particular moment in a switching power supply's cycle of operation. An active clamp for a flyback converter is shown in Figure 30.

In Figure 30, the active clamp is reset (or emptied of its stored energy) just prior to the turn−off transition. It is then disabled during the negative transition.

Obviously, the implementation of an active clamp is more expensive than other approaches, and is usually reserved for very compact power supplies where heat is a critical issue.

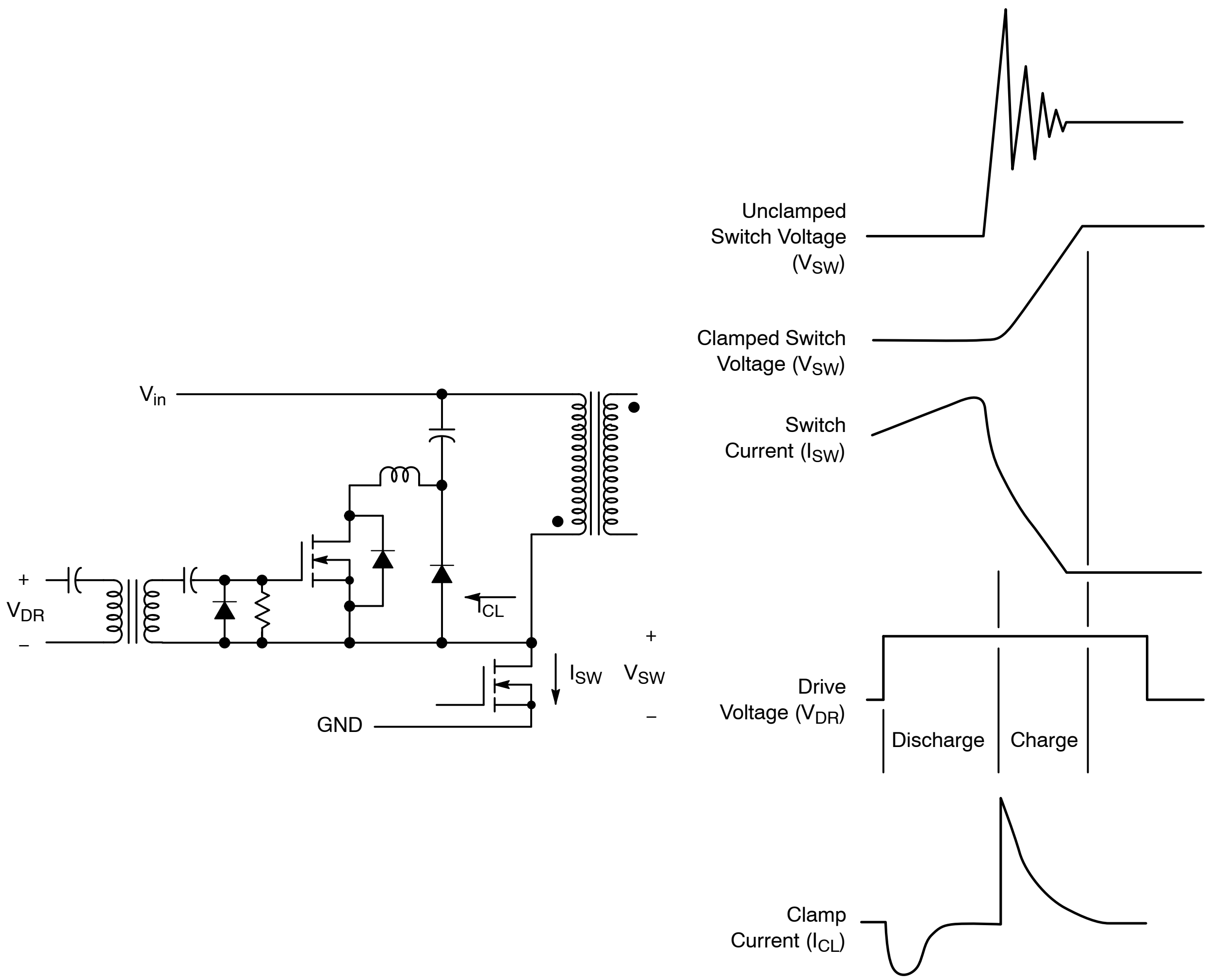


Figure 30. An Active Clamp Used in a One Transistor Forward or a Flyback Converter

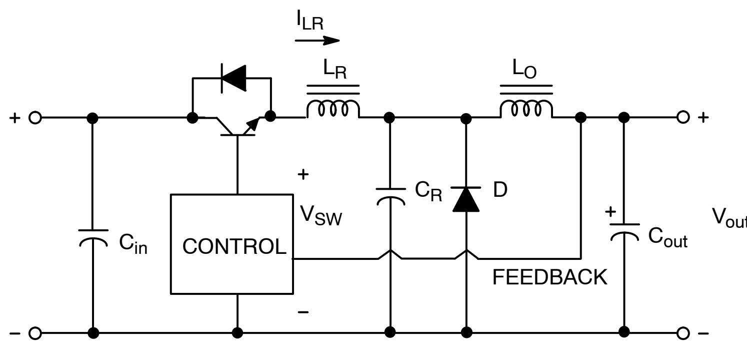
Quasi−Resonant Topologies

A quasi−resonant topology is designed to reduce or eliminate the frequency−dependent switching losses within the power switches and rectifiers. Switching losses account for about 40% of the total loss within a PWM power supply and are proportional to the switching frequency. Eliminating these losses allows the designer to increase the operating frequency of the switching power supply and so use smaller inductors and capacitors, reducing size and weight. In addition, RFI levels are reduced due to the controlled rate of change of current or voltage.

The downside to quasi−resonant designs is that they are more complex than non−resonant topologies due to parasitic RF effects that must be considered when switching frequencies are in the 100's of kHz.

Schematically, quasi−resonant topologies are minor modifications of the standard PWM topologies. A resonant tank circuit is added to the power switch section to make either the current or the voltage "ring" through a half a sinusoid waveform. Since the sinusoid starts at zero and ends at zero, the product of the voltage and current at the starting and ending points is zero, thus has no switching loss.

There are two quasi−resonant methods: zero current switching (ZCS) or zero voltage switching (ZVS). ZCS is a fixed on−time, variable off−time method of control. ZCS starts from an initial condition where the power switch is off and no current is flowing through the resonant inductor. The ZCS quasi−resonant buck converter is shown in Figure 31.



A ZCS Quasi−Resonant Buck Converter

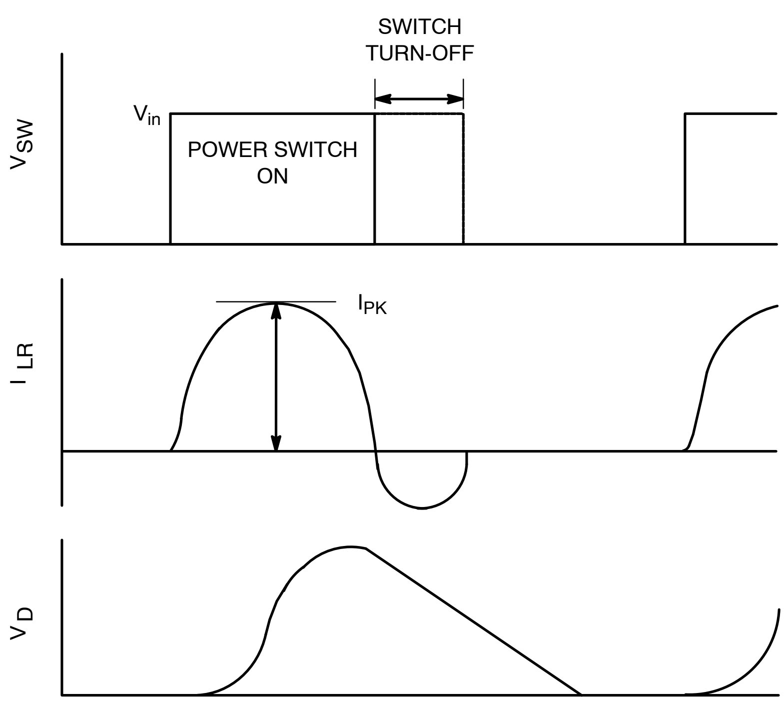


Figure 31. Schematic and Waveforms for a ZCS Quasi-Resonant Buck Converter

In this design, both the power switch and the catch diode operate in a zero current switching mode. Power is passed to the output during the resonant periods. So to increase the power delivered to the load, the frequency would increase, and vice versa for decreasing loads. In typical designs the frequency can change 10:1 over the ZCS supply's operating range.

The ZVS is a fixed off−time, variable on−time method control. Here the initial condition occurs when the power switch is on, and the familiar current ramp is flowing through the filter inductor. The ZVS quasi−resonant buck converter is shown in Figure 32. Here, to control the power delivered to the load, the amount of "resonant off times" are varied. For light loads, the frequency is high. When the load is heavy, the frequency drops. In a typical ZVS power supply, the frequency typically varies 4:1 over the entire operating range of the supply.

There are other variations on the resonant theme that promote zero switching losses, such as full resonant PWM, full and half−bridge topologies for higher power and resonant transition topologies. For a more detailed treatment, see Chapter 4 in the "Power Supply Cookbook" (Bibliography reference 3).

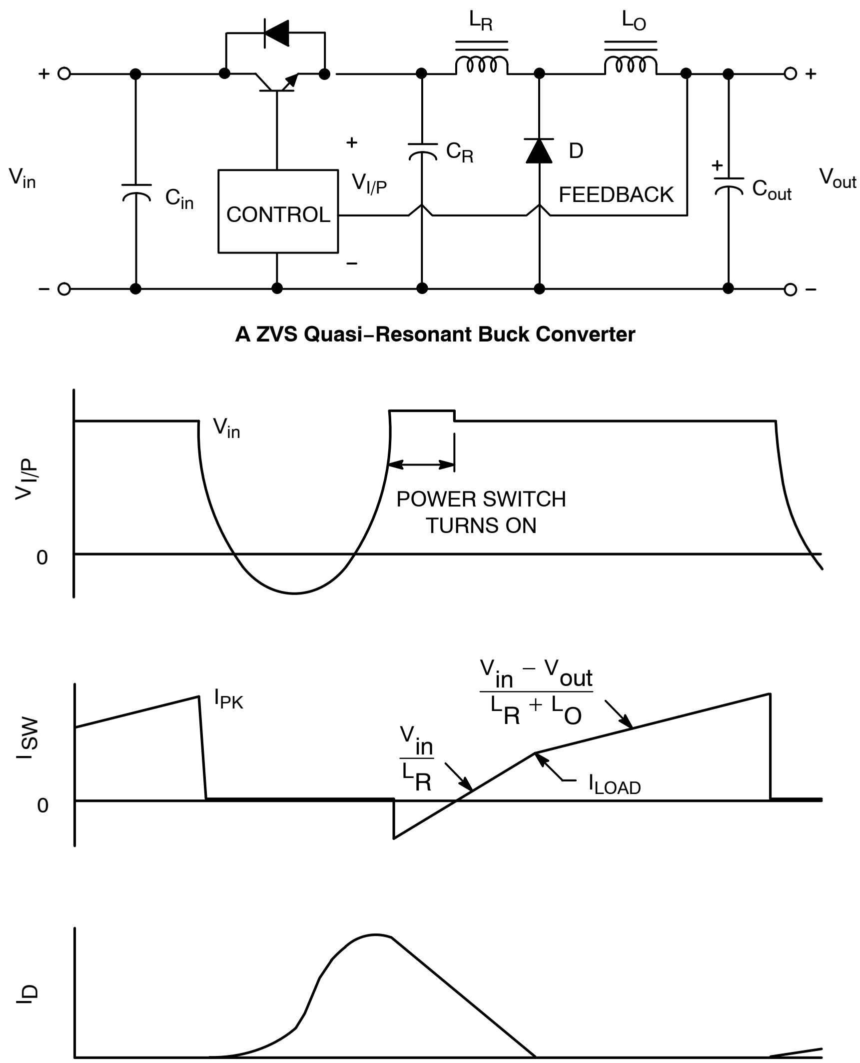


Figure 32. Schematic and Waveforms for a ZVS Quasi-Resonant Buck Converter

Power Factor Correction

Power Factor (PF) is defined as the ratio of real power to apparent power. In a typical AC power supply application where both the voltage and current are sinusoidal, the PF is given by the cosine of the phase angle between the input current and the input voltage and is a measure of how much of the current contributes to real power in the load. A power factor of unity indicates that 100% of the current is contributing to power in the load while a power factor of zero indicates that none of the current contributes to power in the load. Purely resistive loads have a power factor of unity; the current through them is directly proportional to the applied voltage.

The current in an ac line can be thought of as consisting of two components: real and imaginary. The real part results in power absorbed by the load while the imaginary part is power being reflected back into the source, such as is the case when current and voltage are of opposite polarity and their product, power, is negative.

It is important to have a power factor as close as possible to unity so that none of the delivered power is reflected back to the source. Reflected power is undesirable for three reasons:

1. The transmission lines or power cord will generate heat according to the total current being carried, the real part plus the reflected part. This causes problems for the electric utilities and has prompted various regulations requiring all electrical equipment connected to a low voltage distribution system to minimize current harmonics and maximize power factor.
2. The reflected power not wasted in the resistance of the power cord may generate unnecessary heat in the source (the local step−down transformer), contributing to premature failure and constituting a fire hazard.
3. Since the ac mains are limited to a finite current by their circuit breakers, it is desirable to get the most power possible from the given current available. This can only happen when the power factor is close to or equal to unity.

The typical AC input rectification circuit is a diode bridge followed by a large input filter capacitor. During the time that the bridge diodes conduct, the AC line is driving an electrolytic capacitor, a nearly reactive load. This circuit will only draw current from the input lines when the input's voltage exceeds the voltage of the filter capacitor. This leads to very high currents near the peaks of the input AC voltage waveform as seen in Figure 33.

Since the conduction periods of the rectifiers are small, the peak value of the current can be 3−5 times the average input current needed by the equipment. A circuit breaker only senses average current, so it will not trip when the peak current becomes unsafe, as found in many office areas. This can present a fire hazard. In three−phase distribution systems, these current peaks sum onto the neutral line, not meant to carry this kind of current, which again presents a fire hazard.

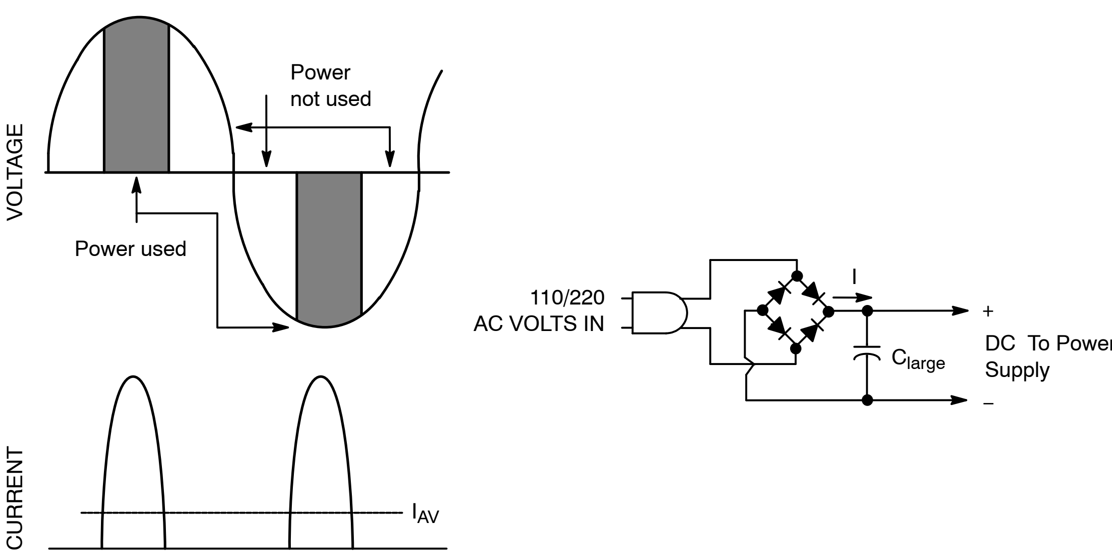


Figure 33. The Waveforms of a Capacitive Input Filter

A Power Factor Correction (PFC) circuit is a switching power converter, essentially a boost converter with a very wide input range, that precisely controls its input current on an instantaneous basis to match the waveshape and phase of the input voltage. This represents a zero degrees or 100 percent power factor and mimics a purely resistive load. The amplitude of the input current waveform is varied over longer time frames to maintain a constant voltage at the converter's output filter capacitor. This mimics a resistor which slowly changes value to absorb the correct amount of power to meet the demand of the load. Short term energy excesses and deficits caused by sudden changes in the load are supplemented by a "bulk energy storage capacitor", the boost converter's output filter device. The PFC input filter capacitor is reduced to a few microfarads, thus placing a half−wave haversine waveshape into the PFC converter.

The PFC boost converter can operate down to about 30 V before there is insufficient voltage to draw any more significant power from its input. The converter then can begin again when the input haversine reaches 30 V on the next half−wave haversine. This greatly increases the conduction angle of the input rectifiers. The drop−out region of the PFC converter is then filtered (smoothed) by the input EMI filter.

A PFC circuit not only ensures that no power is reflected back to the source, it also eliminates the high current pulses associated with conventional rectifier−filter input circuits. Because heat lost in the transmission line and adjacent circuits is proportional to the square of the current in the line, short strong current pulses generate more heat than a purely resistive load of the same power. The active power factor correction circuit is placed just following the AC rectifier bridge. An example can be seen in Figure 34.

Depending upon how much power is drawn by the unit, there is a choice of three different common control modes. All of the schematics for the power sections are the same, but the value of the PFC inductor and the control method are different. For input currents of less than 150 watts, a *discontinuous−mode* control scheme is typically used, in which the PFC core is completely emptied prior to the next power switch conduction cycle. For powers between 150 and 250 watts, the *critical conduction mode* is recommended. This is a method of control where the control IC senses just when the PFC core is emptied of its energy and the next power switch conduction cycle is immediately begun; this eliminates any dead time exhibited in the discontinuous−mode of control. For an input power greater than 250 watts, the *continuous−mode* of control is recommended. Here the peak currents can be lowered by the use of a larger inductor, but a troublesome reverse recovery characteristic of the output rectifier is encountered, which can add an additional 20−40 percent in losses to the PFC circuit.

Many countries cooperate in the coordination of their power factor requirements. The most appropriate document is IEC61000−3−2, which encompasses the performance of generalized electronic products. There are more detailed specifications for particular products made for special markets.

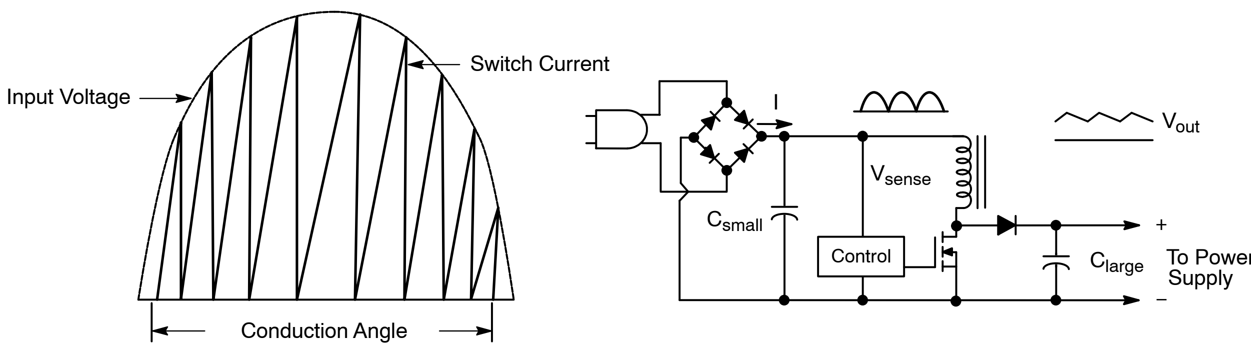


Figure 34. Power Factor Correction Circuit

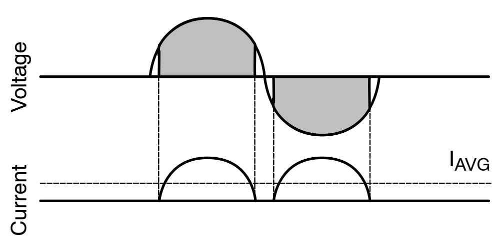


Figure 35. Waveform of Corrected Input

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5. Martin, Robert F., “Harmonic Currents,” *Compliance Engineering − 1999 Annual Resources Guide*, Cannon Communications, LLC, pp. 103−107.
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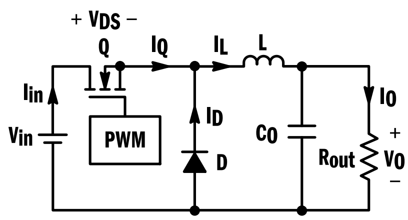
7. Christophe Basso, *Designing Control Loops for Linear and Switching Power Supplies: A Tutorial Guide*, Artech House, 2012, ISBN 978−1−60807−557−7.

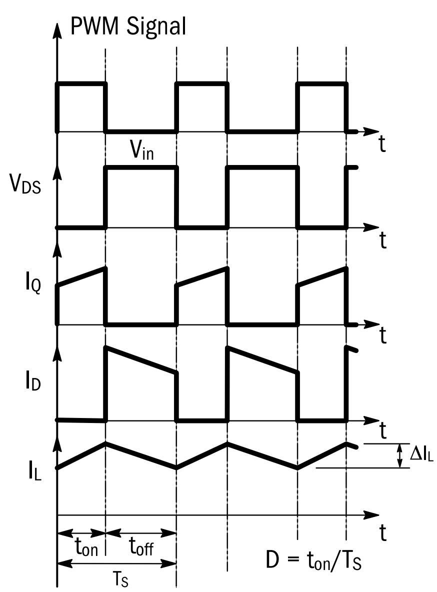
Topology Overview

NON-ISOLATED TOPOLOGIES

BUCK



The buck converter is used when the required output voltage is less than the input voltage. When the output voltage requirement is low, the diode D is replaced by a switch which is called a synchronous rectifier.



Input-Output Relationship

$V_O = DV_{in}$

Switch (Q) Ratings

Drain Voltage, $V_{DS} = V_{in}$

Peak Current, $I_Q = I_O + \frac{\Delta I_L}{2}$

Diode Ratings

Voltage $= V_{in}$

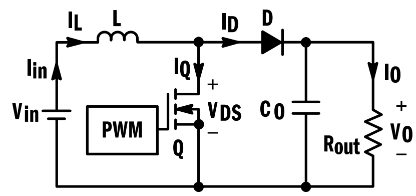
Peak Current, $I_D = I_O + \frac{\Delta I_L}{2}$

ΔI_L

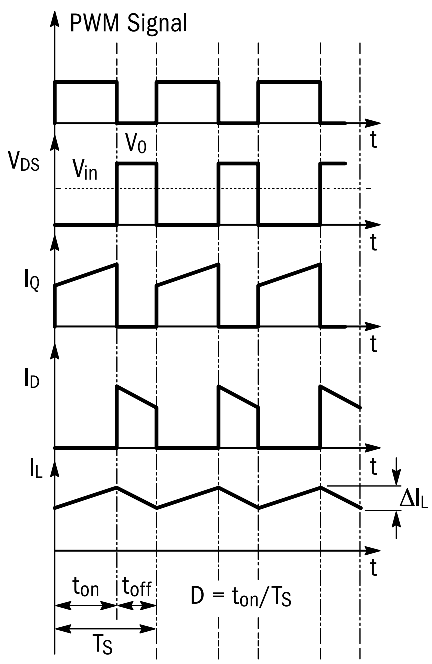
$$\Delta I_L = \frac{V_O\,(1\text{–}D)\,T_S}{L}$$

NON-ISOLATED TOPOLOGIES

BOOST — CCM



The boost converter is used when the required minimum output voltage is greater than the maximum input voltage. It can be used in Continuous Conduction Mode (CCM) to minimize the peak current or in Discontinuous Conduction Mode (DCM) to minimize the inductor value. It is also employed as a pre-converter for Power Factor Correction applications.



CCM

CCM Input-Output Relationship

$$V_0 = \frac{V_{in}}{(1 - D)}$$

CCM Switch (Q) Ratings

Drain Voltage, $V_{DS} = V_0$

Peak Current, $I_Q = \dfrac{I_0}{(1 - D)} + \dfrac{\Delta I_L}{2}$

CCM Diode Ratings

Voltage $= V_0$

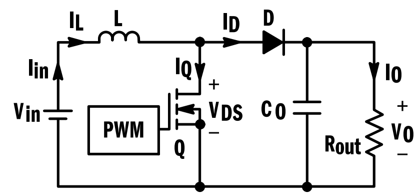
Peak Current, $I_D = \dfrac{I_0}{(1 - D)} + \dfrac{\Delta I_L}{2}$

ΔI_L

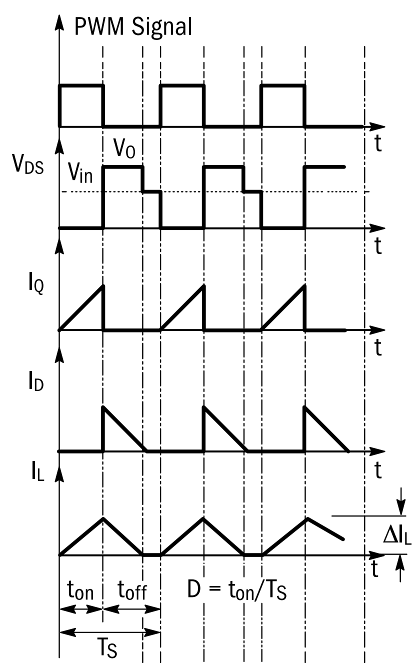
$$\Delta I_L = \frac{(V_0 - V_{in})(1 - D)T_S}{L}$$

NON-ISOLATED TOPOLOGIES

BOOST — DCM



The boost converter is used when the required minimum output voltage is greater than the maximum input voltage. It can be used in Continuous Conduction Mode (CCM) to minimize the peak current or in Discontinuous Conduction Mode (DCM) to minimize the inductor value. It is also employed as a pre-converter for Power Factor Correction applications.



DCM

DCM Input-Output Relationship

$$V_O = V_{in} \frac{1}{2} \left[1 + \sqrt{\frac{2\,R_{out}\,D^2}{L\,f_{sw}}} \right]$$

DCM Switch (Q) Ratings

Drain Voltage, $V_{DS} = V_O$

Peak Current, $I_Q = \dfrac{V_{in}\,t_{on}}{L}$

DCM Diode Ratings

Voltage = V_O

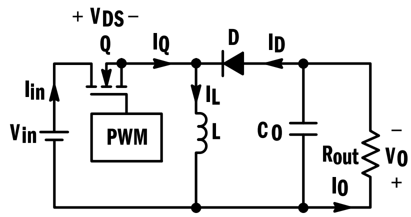
Peak Current, $I_D = \dfrac{V_{in}\,t_{on}}{L}$

ΔI_L

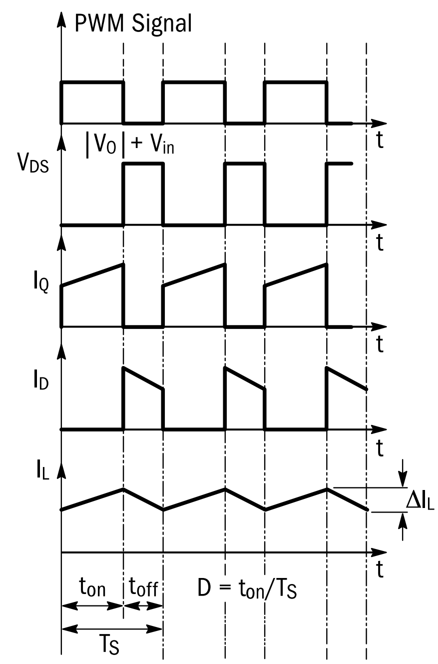
$$\Delta I_L = \frac{V_{in}\,t_{on}}{L}$$

NON-ISOLATED TOPOLOGIES

INVERTING BUCK-BOOST



The inverting buck-boost converter is used when the absolute value of the output voltage is greater or lesser than the applied input voltage and the user wants a negative output voltage.



Input-Output Relationship

$$V_O = -\frac{D}{1-D} V_{in}$$

Switch (Q) Ratings

Drain Voltage, $V_{DS} = |V_O| + V_{in}$

Peak Current, $I_Q = \frac{I_O}{(1-D)} + \frac{\Delta I_L}{2}$

Diode Ratings

Voltage = $V_{in} + |V_O|$

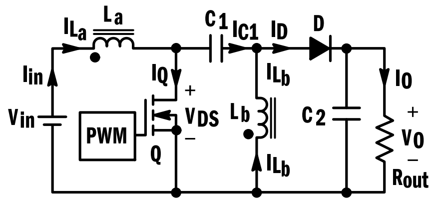
Peak Current, $I_D = \frac{I_O}{(1-D)} + \frac{\Delta I_L}{2}$

ΔI_L

$$\Delta I_L = \frac{V_o T_s}{L}(1\text{-}D)$$

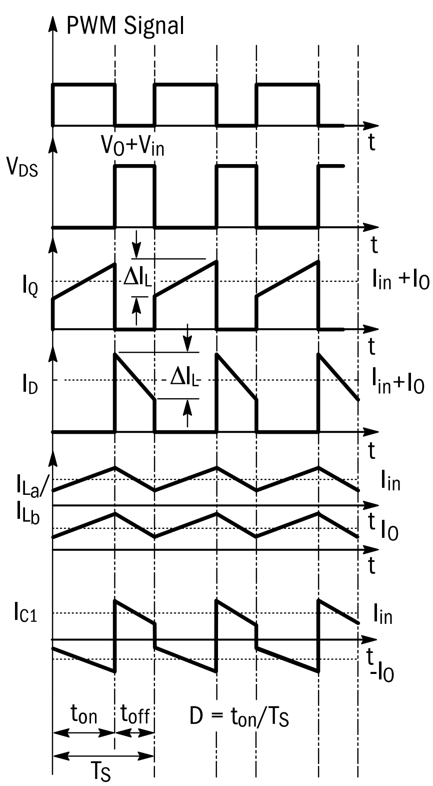
NON-ISOLATED TOPOLOGIES

SEPIC



(Note: L_a and L_b are coupled inductors)

The SEPIC circuit is used when the input voltage must cover a range that is both greater than and less than the output voltage and of the same polarity. As in a boost circuit, the input current is smooth, and the output current (into the output capacitor) is discontinuous. Capacitor ripple current is large.



Input-Output Relationship

$$V_O = \frac{D}{1 - D} V_{in}$$

Switch (Q) Ratings*

Drain Voltage, $V_{DS} = V_O + V_{in}$

Peak Current, $I_Q = \left(1 + \frac{r_a}{2}\right) \frac{I_O}{1 - D}$

Diode Ratings*

Voltage = $V_O + V_{in}$

Peak Current, $I_D = \left(1 + \frac{r_a}{2}\right) \frac{I_O}{1 - D}$

ΔI_L ($L_a = L_b$, $L = L_a + L_b$)

$$\Delta I_L = \frac{V_{in}}{L} (DT_S)$$

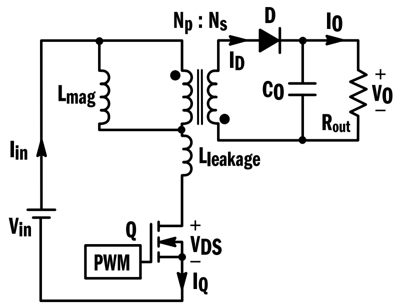
*Applies to Switch and Diode Ratings

$$r_a = \frac{\Delta I_Q}{I_{in} + I_O}$$

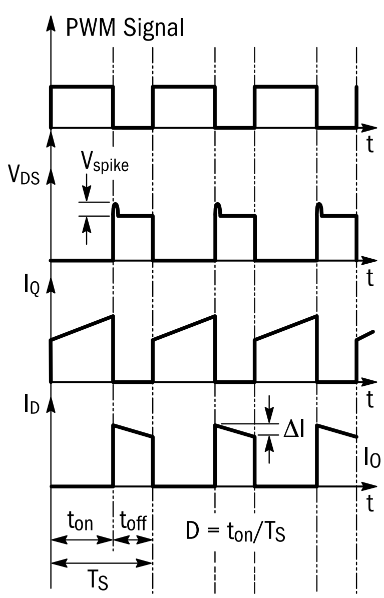
$I_O\, D$

ISOLATED TOPOLOGIES (HARD SWITCHING)

FLYBACK – CCM



The flyback converter, derived from the buck-boost converter is normally used at power levels below 150 W. It has a higher switch rating due to the reflected output voltage appearing across the switch in addition to the applied input voltage. It is also used for Single Stage Power Factor Correction applications.



CCM

CCM Input-Output Relationship

$$V_O = \left(\frac{N_s}{N_p}\right)\left(\frac{D}{1 - D}\right) V_{in}$$

CCM Switch (Q) Ratings

Drain Voltage, $V_{DS} = V_{in} + V_{out}\left(\frac{N_p}{N_s}\right) + V_{spike}$

Peak Current, $I_Q = \left(\frac{N_s}{N_p}\right)\left(\frac{I_O}{(1-D)} + \frac{\Delta I}{2}\right)$

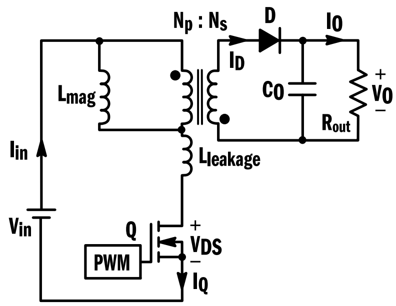
CCM Diode Ratings

Voltage $= V_O + \left(\frac{N_s}{N_p}\right) V_{in}$

Peak Current, $I_D = \frac{I_O}{(1 - D)} + \frac{\Delta I}{2}$

ISOLATED TOPOLOGIES (HARD SWITCHING)

FLYBACK – DCM



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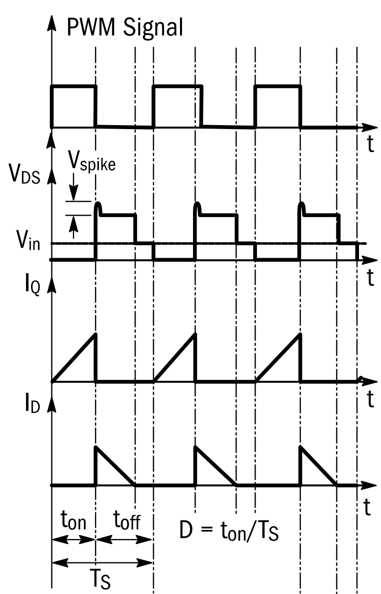


DCM

DCM Input-Output Relationship

$$V_O = \left(\frac{N_s}{N_p}\right) D \sqrt{\frac{R_{out}\, T_S}{2L_{mag}}}$$

DCM Switch (Q) Ratings

Drain Voltage, $V_{DS} = V_{in} + V_{out}\left(\frac{N_p}{N_s}\right) + V_{spike}$

Peak Current, $I_Q = \frac{V_{in}\, t_{on}}{L_{mag}}$

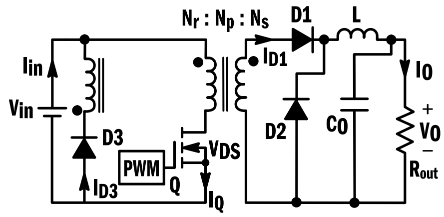
DCM Diode Ratings

Voltage $= V_O + \left(\frac{N_s}{N_p}\right) V_{in}$

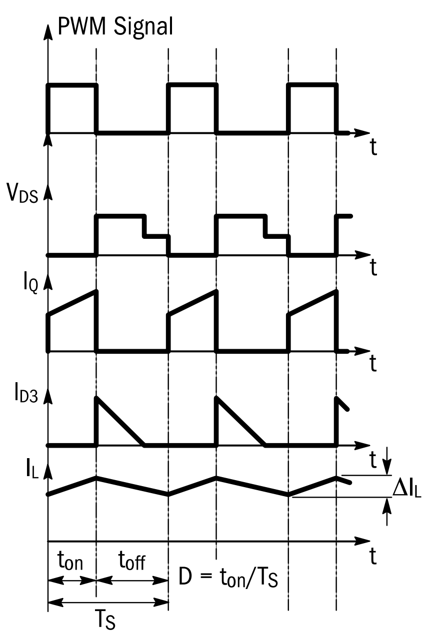
Peak Current, $I_D = \left(\frac{V_{in}\, t_{on}}{L_{mag}}\right)\left(\frac{N_p}{N_s}\right)$

ISOLATED TOPOLOGIES (HARD SWITCHING)

FORWARD



The forward converter, derived from the buck converter is used at power levels typically in the range of 150 - 400 W. It is widely used for step down applications.



Input-Output Relationship

$$V_O = \left(\frac{N_s}{N_p}\right) D V_{in}$$

Switch (Q) Ratings

Drain Voltage, $V_{DS} = V_{in}\left(1 + \frac{N_p}{N_r}\right)$

Peak Current, $I_Q = \left(I_O + \frac{\Delta I_L}{2}\right)\left(\frac{N_s}{N_p}\right)$

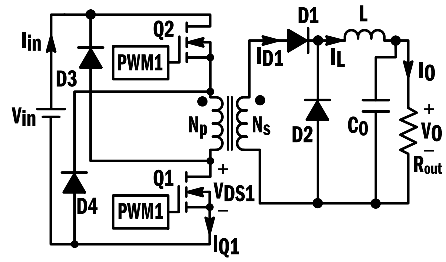
Diode Ratings

Voltage $= V_{D1} = V_{D2} = V_{in}\left(\frac{N_s}{N_p}\right)$

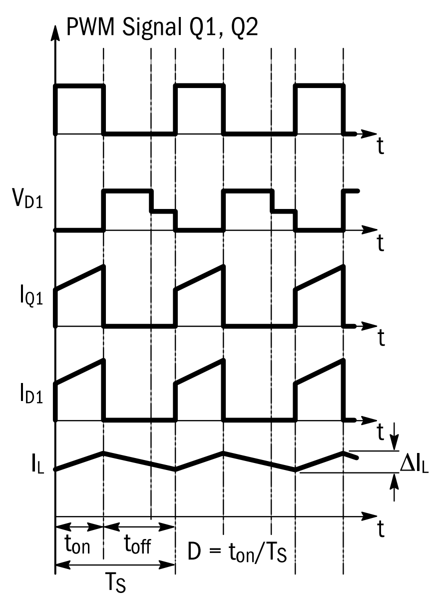
Peak Current, $I_{D1} = I_{D2} = I_O + \frac{\Delta I_L}{2}$

ISOLATED TOPOLOGIES (HARD SWITCHING)

TWO-SWITCH FORWARD



Two-switch forward converter, a variation of the forward converter can be used at higher power levels and does not have an additional reset winding. The voltage stress on the switches is less in this case when compared to a single switch forward converter.



Input-Output Relationship

$$V_O = \left(\frac{N_s}{N_p}\right) DV_{in}$$

Switch (Q) Ratings

Drain Voltage, $V_{DS1} = V_{in}$

Peak Current, $I_Q = \left(I_O + \frac{\Delta I_L}{2}\right)\left(\frac{N_s}{N_p}\right)$

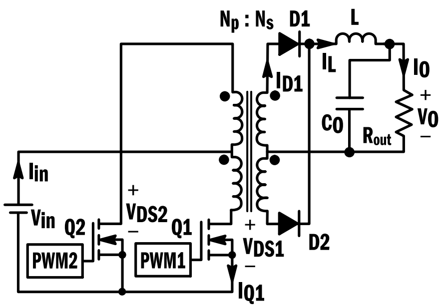
Diode Ratings

Voltage = $V_{D1} = \left(\frac{N_s}{N_p}\right) V_{in}$

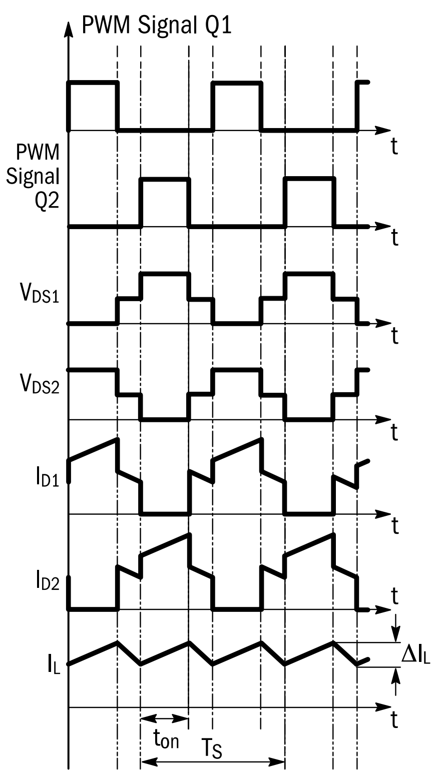
Peak Current, $I_{D1} = I_{D2} = I_O + \frac{\Delta I_L}{2}$

ISOLATED TOPOLOGIES (HARD SWITCHING)

PUSH-PULL



Push-pull converter is used when there is a wide variation in the input and when the output voltage is lesser than the input voltage. It can be used at power levels in the range of few hundred watts to 1 kW.



Input-Output Relationship

$$V_O = 2\left(\frac{N_s}{N_p}\right) DV_{in}$$

Switch (Q) Ratings

Drain Voltage, $V_{DS1} = 2V_{in}$

Peak Current, $I_{Q1} = I_{Q2} = \left(\frac{N_s}{N_p}\right)\left(I_O + \frac{\Delta I_L}{2}\right)$

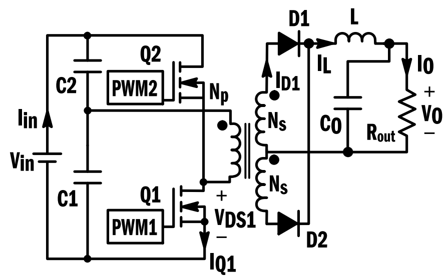
Diode Ratings

Voltage $= V_{D1} = V_{D2} = \left(\frac{2\,V_{in}\,N_s}{N_p}\right)$

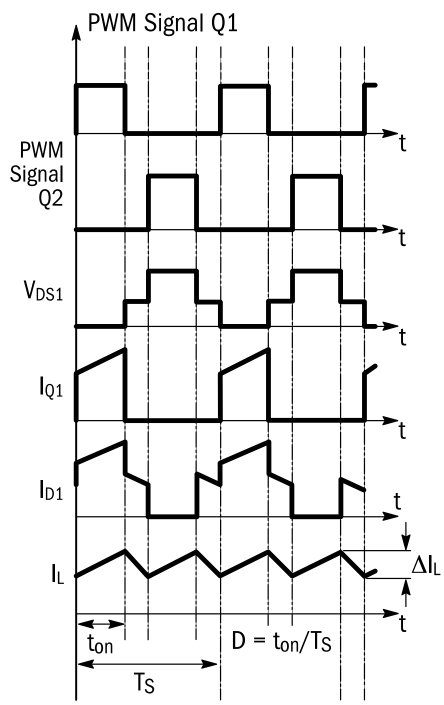
Peak Current, $I_{D1} = I_{D2} = I_O + \frac{\Delta I_L}{2}$

ISOLATED TOPOLOGIES (HARD SWITCHING)

HALF-BRIDGE



Half-bridge converter is used at medium power levels in the range of 200-700 W for step down applications. It has lower part count when compared to the full-bridge converter but the transistor switches have to handle twice the current compared to a full-bridge converter.



Input-Output Relationship

$$V_O = \left(\frac{N_s}{N_p}\right) D \left(\frac{V_{in}}{2}\right)$$

Switch (Q) Ratings

Drain Voltage, $V_{DS1} = V_{in}$

Peak Current, $I_Q = \left(\frac{N_s}{N_p}\right)\left(I_O + \frac{\Delta I_L}{2}\right)$

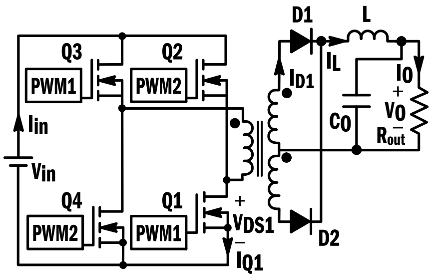
Diode Ratings

Voltage $= V_{D1} = V_{D2} = \left(\frac{V_{in}}{2}\right)\left(\frac{N_s}{N_p}\right)$

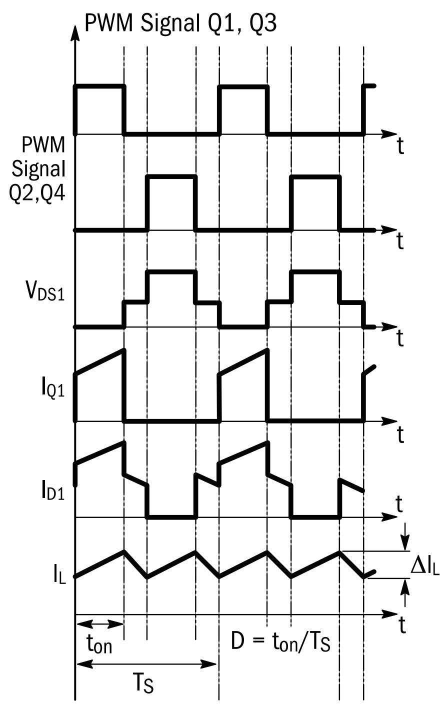
Peak Current, $I_{D1} = I_{D2} = I_O + \frac{\Delta I_L}{2}$

ISOLATED TOPOLOGIES (HARD SWITCHING)

FULL-BRIDGE



A full-bridge converter is used at high power levels in the range of a kW and is normally not used at lower power levels due to its higher part count. It is normally used in step-down applications.



Input-Output Relationship

$$V_O = \left(\frac{N_s}{N_p}\right) D V_{in}$$

Switch (Q) Ratings

Drain Voltage, $V_{DS1} = V_{in}$

Peak Current, $I_Q = \left(\frac{N_s}{N_p}\right)\left(I_O + \frac{\Delta I_L}{2}\right)$

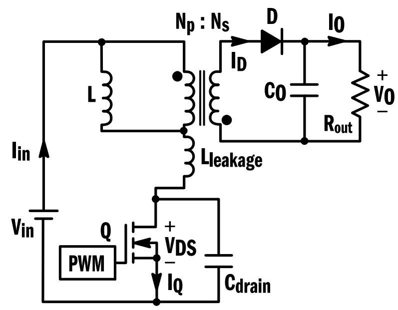
Diode Ratings

Voltage $= V_{D1} = V_{D2} = \left(\frac{V_{in}}{2}\right)\left(\frac{N_s}{N_p}\right)$

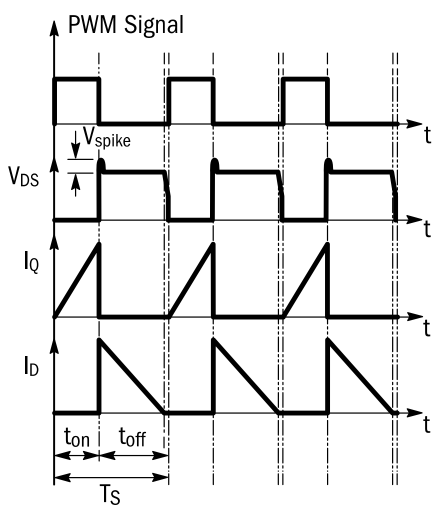
Peak Current, $I_{D1} = I_{D2} = I_O + \frac{\Delta I_L}{2}$

ISOLATED TOPOLOGIES (SOFT SWITCHING)

QUASI-RESONANT FLYBACK



This topology is used at power levels which are lesser than 150 W. It allows operation at variable frequencies.



Input-Output Relationship

$$V_O = \left(\frac{V_{in}\,D}{1 - D}\right)\left(\frac{N_s}{N_p}\right)$$

Switch (Q) Ratings

Drain

$$\text{Voltage, } V_{DS} = V_{in} + V_{out}\left(\frac{N_p}{N_s}\right) + V_{spike}$$

$$\text{Peak Current, } I_Q = \frac{V_{in}\,t_{on}}{L_{mag}}$$

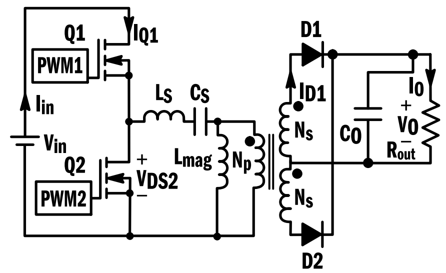
Diode Ratings

$$\text{Voltage} = V_{out} + \left(\frac{N_s}{N_p}\right)V_{in}$$

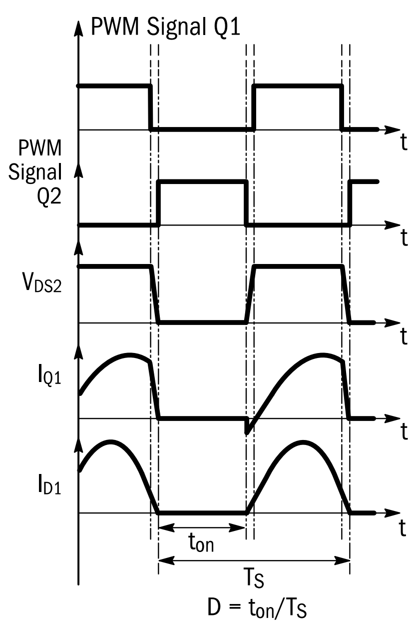
$$\text{Peak Current, } I_D = \left(\frac{V_{in}\,t_{on}}{L_{mag}}\right)\left(\frac{N_p}{N_s}\right)$$

ISOLATED TOPOLOGIES (SOFT SWITCHING)

HALF-BRIDGE RESONANT



It offers high efficiency and requires no inductor at the output. However, it has high ripple current in the output and is difficult to be operated at light loads.



Input-Output Relationship

$$V_O = \left(\frac{V_{in}}{2}\right)\left(\frac{N_s}{N_p}\right)$$

NOTE: Valid at resonant frequency

Switch (Q) Ratings

Drain Voltage, $V_{DS} = V_{in}$

Peak Current, $I_{Q1} = \left(\frac{I_O\,\pi}{2}\right)\left(\frac{N_s}{N_p}\right)$

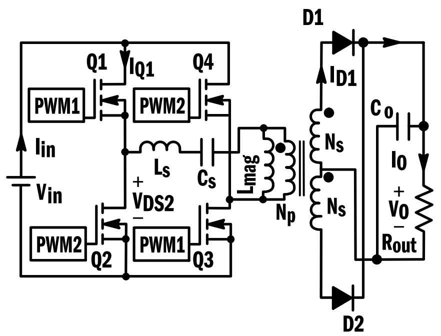
Diode Ratings

Voltage $= V_{D1} = V_{D2} = 2V_O$

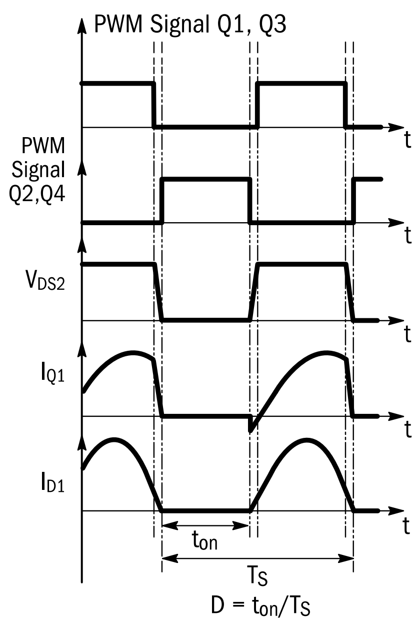
Peak Current, $I_{D1} = I_{D2} = \left(\frac{I_O\,\pi}{2}\right)$

ISOLATED TOPOLOGIES (SOFT SWITCHING)

FULL-BRIDGE RESONANT



It offers high efficiency and requires no inductor at the output. However, it has high ripple current in the output and is difficult to be operated at light loads. Full-bridge version provides two times higher gain compare to half-bridge resonant topology and is usually used for high output power.



Input-Output Relationship

$V_O = V_{in} \left(\frac{N_s}{N_p}\right)$ **NOTE: Valid at resonant frequency**

Switch (Q) Ratings

Drain Voltage, $V_{DS1} = V_{in}$

Peak Current, $I_{Q1} = \left(\frac{I_O \pi}{2}\right)\left(\frac{N_s}{N_p}\right)$

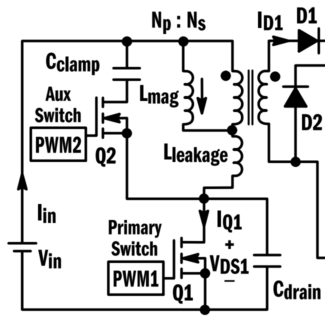
Diode Ratings

Voltage $= V_{D1} = V_{D2} = 2V_O$

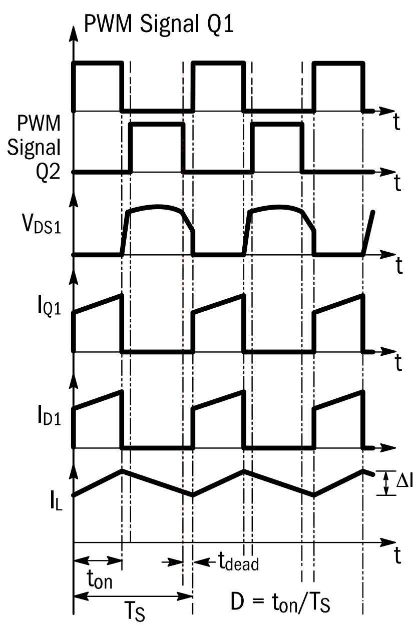
Peak Current, $I_{D1} = I_{D2} = \left(\frac{I_O \pi}{2}\right)$

ISOLATED TOPOLOGIES (SOFT SWITCHING)

ACTIVE-CLAMP FORWARD



It is normally used at high power levels in the range of few hundred watts. It has a higher parts count and requires a dedicated controller, but offers a higher efficiency due to soft switching. It can tolerate wide input voltage variations.



Input-Output Relationship

$$V_O = \left(\frac{N_s}{N_p}\right) DV_{in}$$

Switch (Q) Ratings

Drain Voltage, $V_{DS1} = \dfrac{V_{in}}{(1 - D)}$

Peak Current,

$$I_{Q1} = I_{mag} + \left(\frac{N_s}{N_p}\right)\left(I_O + \frac{\Delta I_L}{2}\right)$$

Diode Ratings

Voltage = $V_{D1} = \left(\dfrac{V_{in}\,D}{1 - D}\right)\left(\dfrac{N_s}{N_p}\right)$

Voltage = $V_{D2} = V_{in}\left(\dfrac{N_s}{N_p}\right)$

Peak Current, $I_{D1} = I_{D2} = I_O + \dfrac{\Delta I_L}{2}$

Switch−Mode Power Supply Examples

This section provides both initial and detailed information to simplify the selection and design of a variety of Switch−Mode power supplies. The ICs for Switching Power Supplies figure identifies control, reference voltage, output protection and switching regulator ICs for various topologies.

Page

ICs for Switching Power Supplies 52
Integrated circuits identified for various sections of a switching power supply.

Real SMPS Applications
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25 to 150 W Flyback Switch−Mode Power Supply 54
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Some of these circuits may have a more complete application note, spice model information or even an evaluation board available. Consult ON Semiconductor’s website (**www.onsemi.com**) or local sales office for more information.

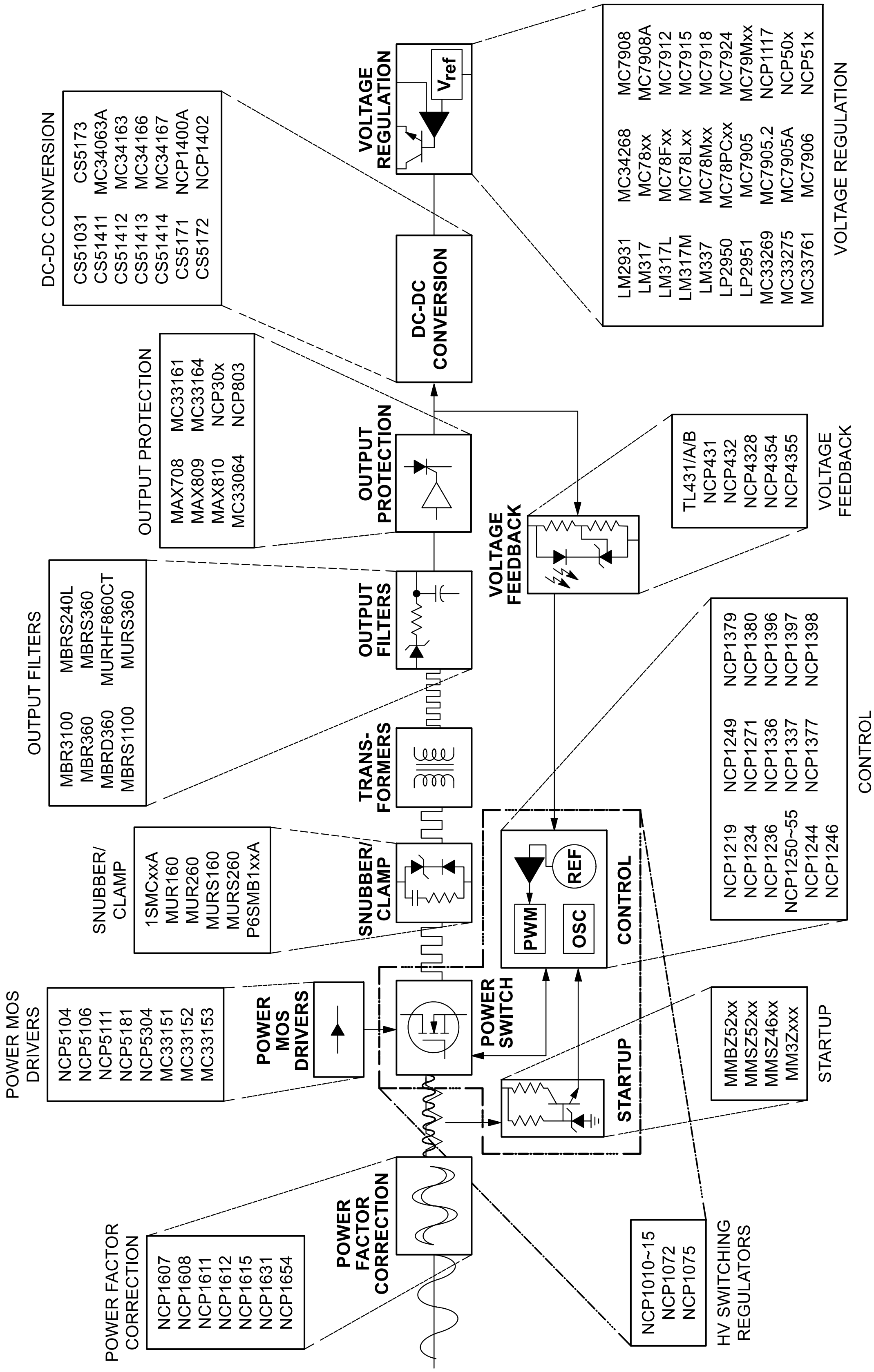


Figure 36. Integrated Circuits for Switching Power Supplies

Up to 25 W Flyback Switch−Mode Power Supply

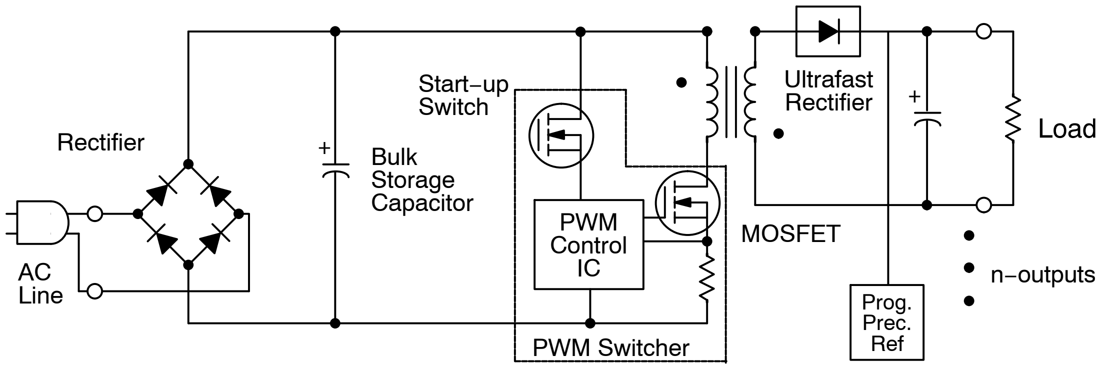


Figure 37. Up to 25 W Flyback

Table 1. High Voltage Switching Regulator Selection

Product	Description	Technical Documentation and Design Resources						
		App Notes	Design & Development Tools	Design Notes	Eval Board	Reference Designs	Videos	White Papers
NCP1014	High Voltage Switching Regulator for Offline SMPS	✓	✓	✓	✓	✓	✓	✓
NCP1013	High Voltage Switching Regulator for Offline SMPS	✓	✓	✓	✓			
NCP1012	High Voltage Switching Regulator for Offline SMPS	✓	✓	✓	✓			
NCP1072	High Voltage Switching Regulator for Offline SMPS		✓	✓	✓			
NCP1075	High Voltage Switching Regulator for Offline SMPS		✓	✓	✓		✓	
NCP1011	High Voltage Switching Regulator for Offline SMPS	✓	✓	✓				
NCP1010	High Voltage Switching Regulator for Offline SMPS	✓	✓					

Table 2. PWM Switching Controllers Selection

Product	Description	Technical Documentation and Design Resources			
		App Notes	Design & Development Tools	Design Notes	Eval Board
NCP1250	PWM Controller, Current Mode, for Offline Power Supplies	✓	✓	✓	✓
NCP1251	PWM Controller, Current Mode, for Offline Power Supplies	✓	✓	✓	✓
NCP1253	PWM Controller, Current Mode, for Offline Power Supplies		✓		

25 to 150 W Flyback Switch−Mode Power Supply

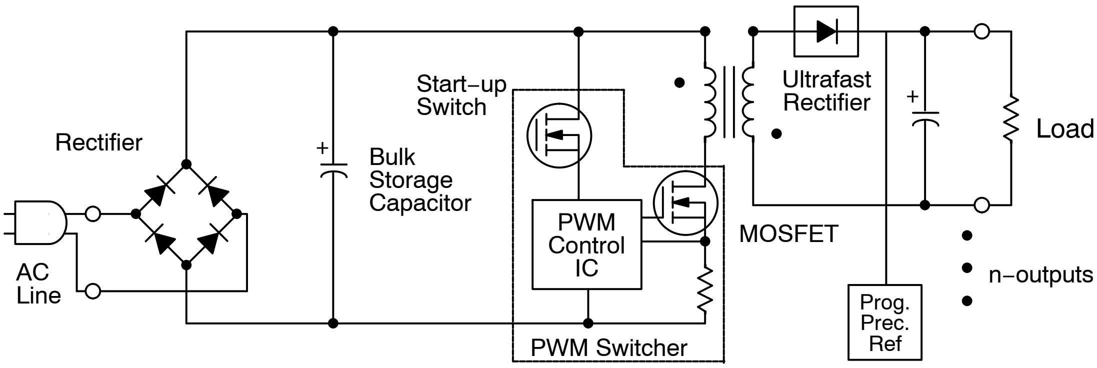


Figure 38. 25 to 150 W Flyback

Table 3. Fixed Frequency Switching Controllers Selection

Product	Description	Technical Documentation and Design Resources				
		App Notes	Design & Development Tools	Design Notes	Eval Board	Videos
NCP1219	PWM Controller with Adjustable Skip Level and External Latch Input	✓	✓	✓	✓	
NCP1234	Controller, Fixed Frequency, Current Mode, for Flyback Converters			✓		
NCP1236	Controller, Fixed Frequency, Current Mode, for Flyback Converters			✓	✓	
NCP1244	Controller, Fixed Frequency, Current Mode, for Flyback Converters					
NCP1246	Controller, Fixed Frequency, Current Mode, for Flyback Converters				✓	✓
NCP1249	PWM Controller Featuring Peak Power Excursion and HV Startup	✓			✓	
NCP1255	PWM Controller, Current Mode, for Offline Power Supplies Featuring Peak Power Excursion	✓			✓	✓
NCP1271	PWM Controller, Soft−Skip™ Standby, with Adjustable Skip Level and External Latch	✓	✓	✓	✓	
NCP1927	PFC and Flyback Controller for Flat Panel TVs			✓		

Table 4.Quasi−Resonant Switching Controllers Selection

Product	Description	Technical Documentation and Design Resources				
		App Notes	Design & Development Tools	Eval Board	Reference Designs	Videos
NCP1336	Controller, Quasi−Resonant, Current Mode, with HV Start−Up					
NCP1337	Controller, Quasi−Resonant Current Mode, with Overpower Compensation	✓	✓	✓	✓	✓

Table 4.Quasi−Resonant Switching Controllers Selection

Product	Description	Technical Documentation and Design Resources				
		App Notes	Design & Development Tools	Eval Board	Reference Designs	Videos
NCP1338	PWM Controller, Free Running Quasi−Resonant Current Mode	✓	✓		✓	
NCP1377	Controller, Free Running Quasi−Resonant Current Mode	✓	✓			
NCP1379	Controller, Quasi−Resonant, Current Mode		✓		✓	
NCP1380	Controller, Quasi−Resonant, Current Mode	✓	✓	✓		✓

150 to 500 W Resonant Switch−Mode Power Supply

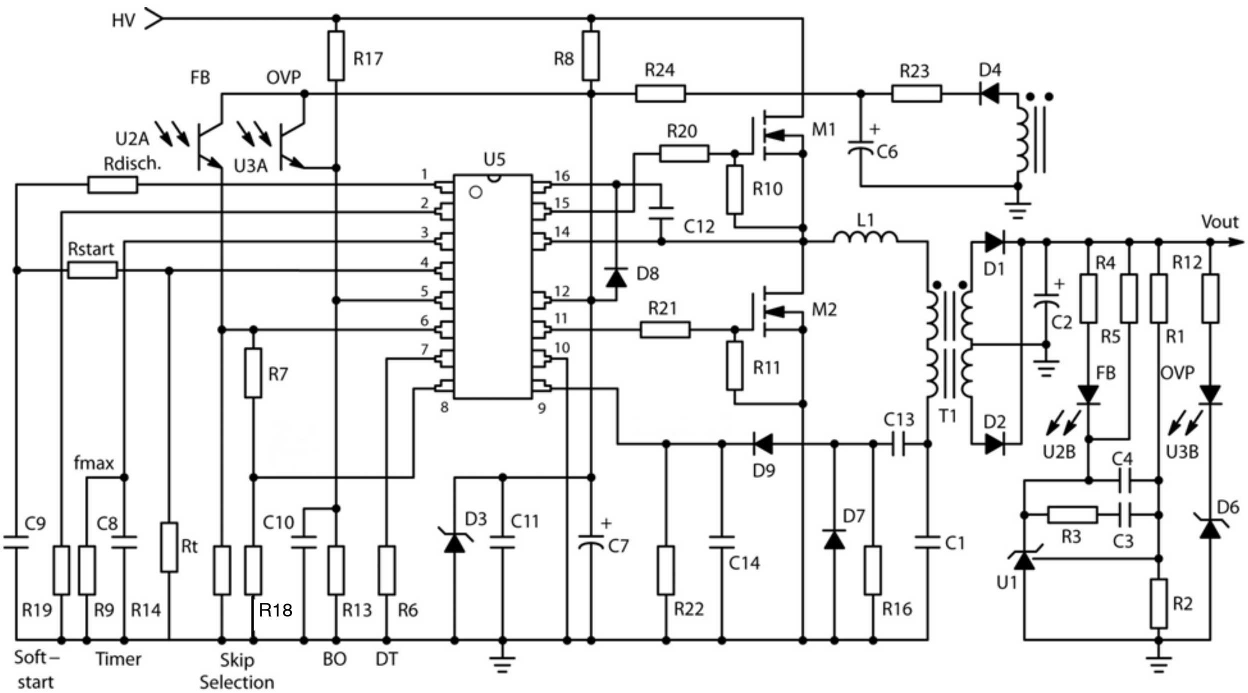


Figure 39. Typical Resonant Mode Controller Application

Table 5. Resonant Switching Controllers Selection

Product	Description	Technical Documentation and Design Resources				
		App Notes	Design & Development Tools	Eval Board	Reference Designs	Videos
NCP1392	MOSFET Driver, High Voltage, Half Bridge, with Inbuilt Oscillator	✓	✓			
NCP1393	High Voltage Half−Bridge MOSFET Driver with Inbuilt Oscillator Featuring Brown−out and Latched Enable Input	✓	✓			
NCP1395	Controller, High Performance Resonant Mode	✓	✓		✓	
NCP1396	Controller, High Performance Resonant Mode, with High and Low Side Drivers	✓	✓		✓	
NCP1397	Controller, High Performance, Resonant Mode, with Integrated High Voltage Drivers	✓	✓		✓	✓
NCP1398	Controller, High Performance, Resonant Mode, with Integrated High Voltage Drivers	✓		✓		✓
NCP1910	High Performance Combo Controller for ATX Power Supplies	✓	✓	✓		

100 W Boost CRM PFC

The NCP1607 is a voltage mode power factor correction controller designed to drive cost−effective converters to meet input line harmonic regulations. The device operates in Critical Conduction Mode (CRM) for optimal performance in applications up to about 300 W. Its voltage mode scheme enables it to obtain unity power factor without the need for a line sensing network. The output voltage is accurately controlled with a built in high precision error amplifier. The controller also implements a comprehensive array of safety features for robust designs.

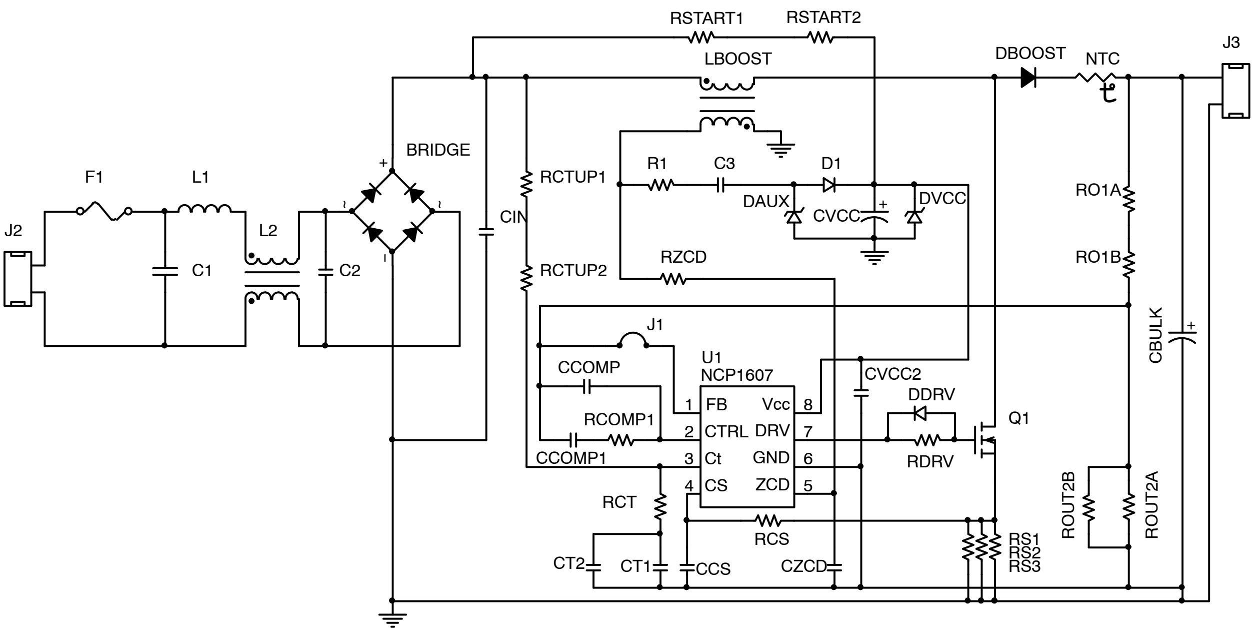


Figure 40. 100 W Boost CRM PFC

Table 6. Key Devices

Designator	Description	Value	Footprint	Manufacturer Part Number
U1	NCP1607	—	SOIC−8	NCP1607BDR2G
D1	Diode, General Purpose	100 V	SOD−123	MMSD4148T1G
DAUX	Diode, Zener	18 V	SOD−123	MMSZ4705T1G
DBOOST	Diode, Ultrafast	4 A, 600 V	Axial	MUR460RLG

For more details please refer to
http://www.onsemi.com/PowerSolutions/evalBoard.do?id=NCP1607BOOSTGEVB

120 W High Efficiency, High Voltage, Active X2, <30 mW No Load PFC

The NCP1615 is a high voltage PFC controller designed to drive PFC boost stages based on an innovative Current Controlled Frequency Foldback (CCFF) method. In this mode, the circuit classically operates in critical conduction mode (CrM) when the inductor current exceeds a programmable value. CCFF maximizes the efficiency at both nominal and light load. In particular, the standby losses are reduced to a minimum. An innovative circuitry allows near¿unity power factor even when the switching frequency is reduced.

The integrated high voltage startup circuit eliminates the need for external startup components and consumes negligible power during normal operation. Housed in a SOIC−16 package, the NCP1615C also incorporates the features necessary for robust and compact PFC stages, with few external components.

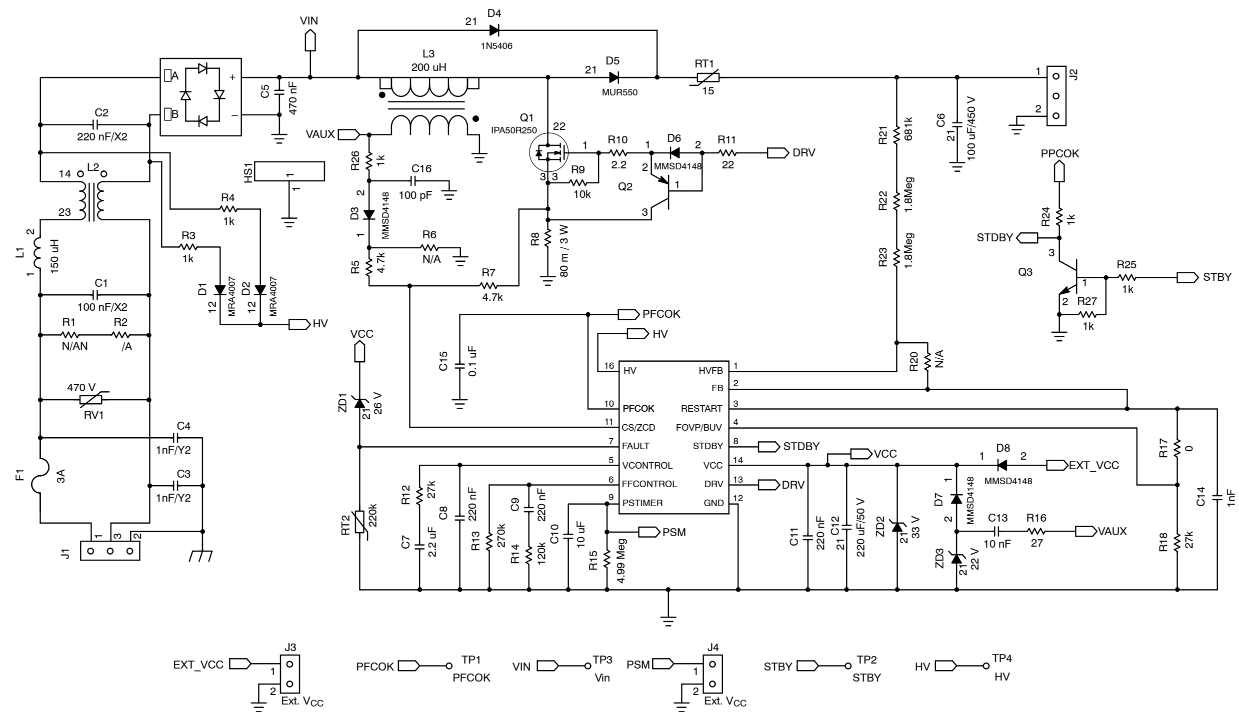


Figure 41. 120 W High Efficiency, High Voltage, Active X2, <30 mW No Load PFC

Table 7. Key Devices

Designator	Description	Value	Tolerance	Footprint	Manufacturer Part Number
Q3	BJT, NPN	—	?? V/200 mA	SOT−23	MMBT3904LT1G
Q2	BJT, PNP	—	30 V/1 A	SOT−23	MMBT589LT1G
D4	Diode, Fast Acting	—	600 V/3 A	DO201AD	1N5406G
D3, D6, D7, D8	Diode, General Purpose	—	?? V/200 mA	SOD−123	MMSD4148T1G
ZD3	Diode, Zener	22 V	500 mW	SOD−123	MMSZ22T1G
ZD1	Diode, Zener	27 V	500 mW	SOD−123	MMSZ27T1G
ZD2	Diode, Zener	33 V	500 mW	SOD−123	MMSZ33T1G
D1, D2	Diode, Rectifier	—	1000 V/1 A	SMA	MRA4007T3G
D5	Diode, Fast Acting	—	520 V/5 A	DO201AD	MUR550APFRLG
U1	PFC Controller	—	—	SO−16NB	NCP1615C

For more details please refer to
http://www.onsemi.com/PowerSolutions/evalBoard.do?id=NCP1615GEVB or
http://www.onsemi.com/pub_link/Collateral/DN05057−D.PDF

300 W Compact Continuous Conduction Mode PFC

The NCP1654 is a Power Factor Controller to efficiently drive Continuous Conduction Mode (CCM) step−up pre−converters. The NCP1654 allows operation in Follower Boost mode to drastically lower the pre−converter size and cost, in a straight−forward manner. The NCP1654 is a continuous conduction mode and fixed frequency controller (65 kHz).

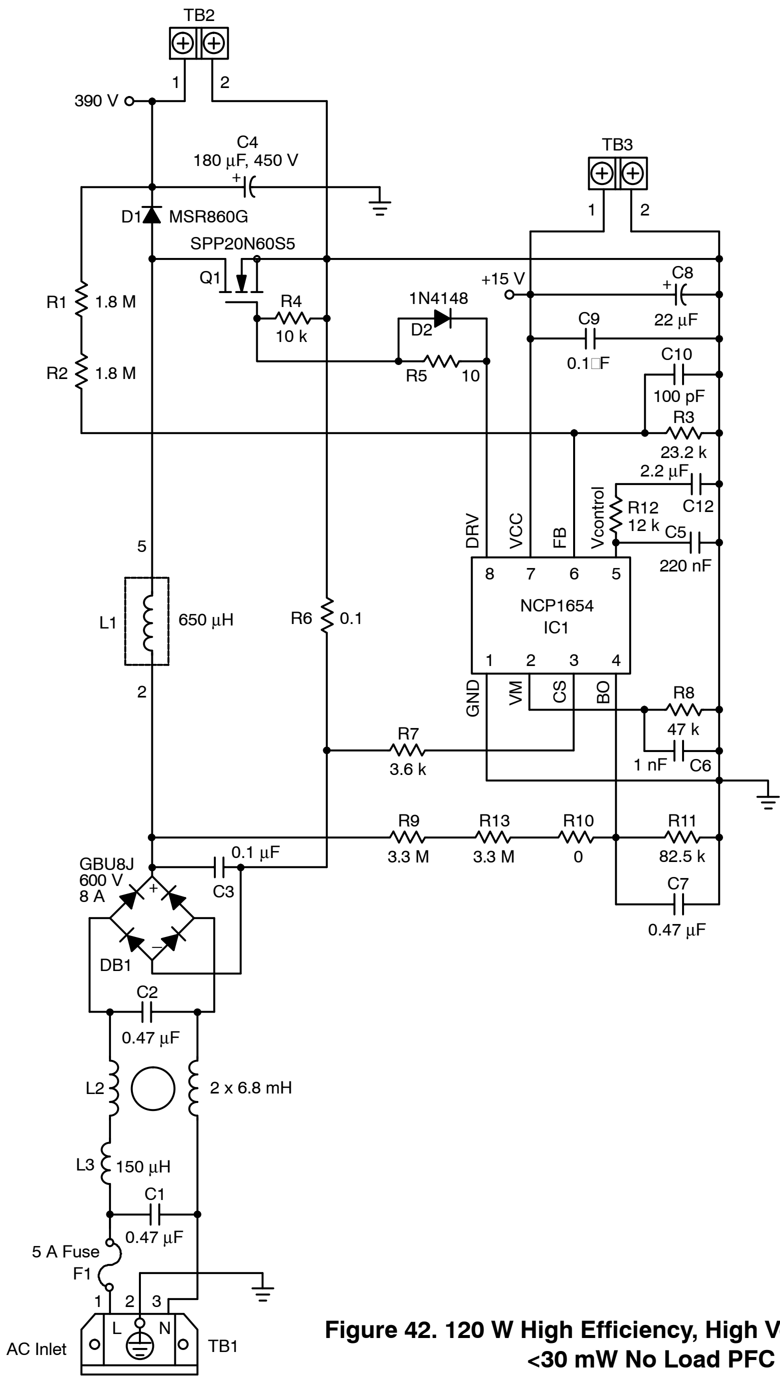


Figure 42. 120 W High Efficiency, High Voltage, Active X2, <30 mW No Load PFC

For more details please refer to
http://www.onsemi.com/PowerSolutions/evalBoard.do?id=NCP1654PFCGEVB

Up to 15 W AC−DC Compact Power Supply

The NCP1072 / NCP1075 products integrate a fixed frequency current mode controller with a 700 V MOSFET. Available in a PDIP−7 or SOT−223 package, the NCP1072/5 offer a high level of integration, including soft−start, frequency−jittering, short−circuit protection, skip−cycle, a maximum peak current set point, ramp compensation, and a Dynamic Self−Supply (eliminating the need for an auxiliary winding).

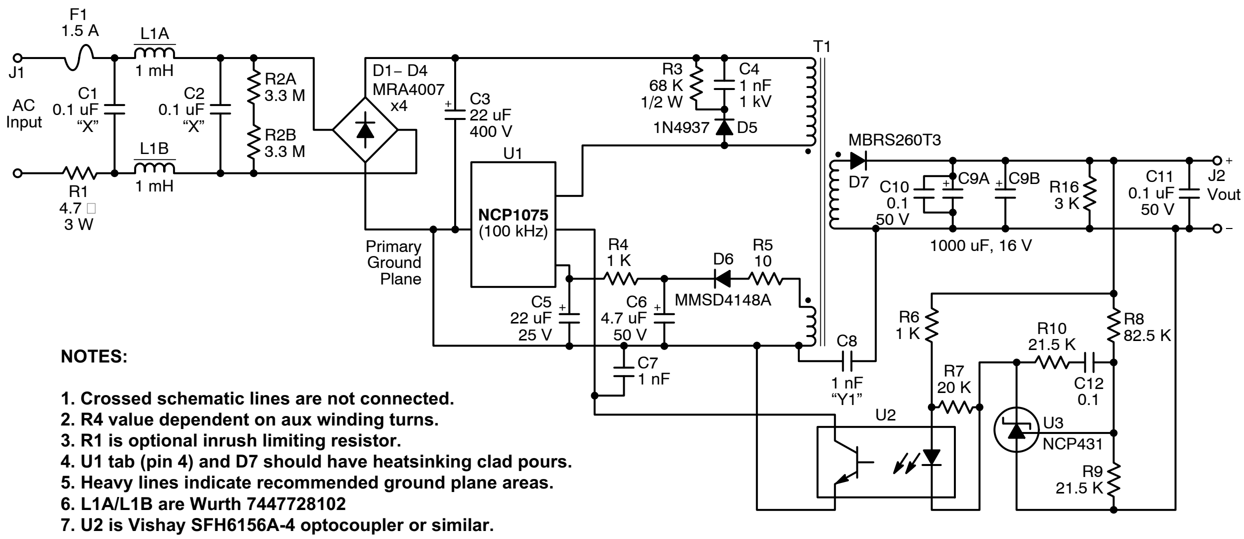


Figure 43. Up to 15 W AC−DC Compact Power Supply

Table 8. Key Devices

Designator	Description	Value	Footprint	Manufacturer Part Number
D7 (12 V Out)	Schottky Diode	2 A, 60 V	SMB	MBRS260T3
D7 (5 V Out)	Schottky Diode	3 A, 40 V	SMB	MBRS2040L
D1, D2, D3, D4	Diode, 60 Hz	1 A, 800 V	SMA	MRA4007
D5	Diode, Fast Recovery	1 A, 800 V	Axial Lead	1N4937
D6	Signal Diode	100 mA, 100 V	SOD−123	MMSD4148A
U3	Programmable Zener	2.5 V	SOIC−8, SOT−23	NCP431A
U1	Switcher IC	100 kHz	SOT−223	NCP1075STBT3G

For more details please refer to
http://www.onsemi.com/PowerSolutions/evalBoard.do?id=NCP1075SOTGEVB

25 W Offline Power Supply High Voltage Switcher

The NCP1129 switcher offers everything needed to build reliable and compact AC−DC switching power supplies with minimal surrounding elements. Incorporating an avalanche rated 650 V MOSFET, converters built with the NCP112X can be safely designed for international conditions without jeopardizing the overall reliability. The NCP112X implements peak current mode control with adjustable ramp compensation that ensures stability in Continuous Conduction Mode (CCM) operation. With an external resistor, the maximum peak current is adjustable, allowing the designer the ability to inject ramp compensation to stabilize CCM power supplies.

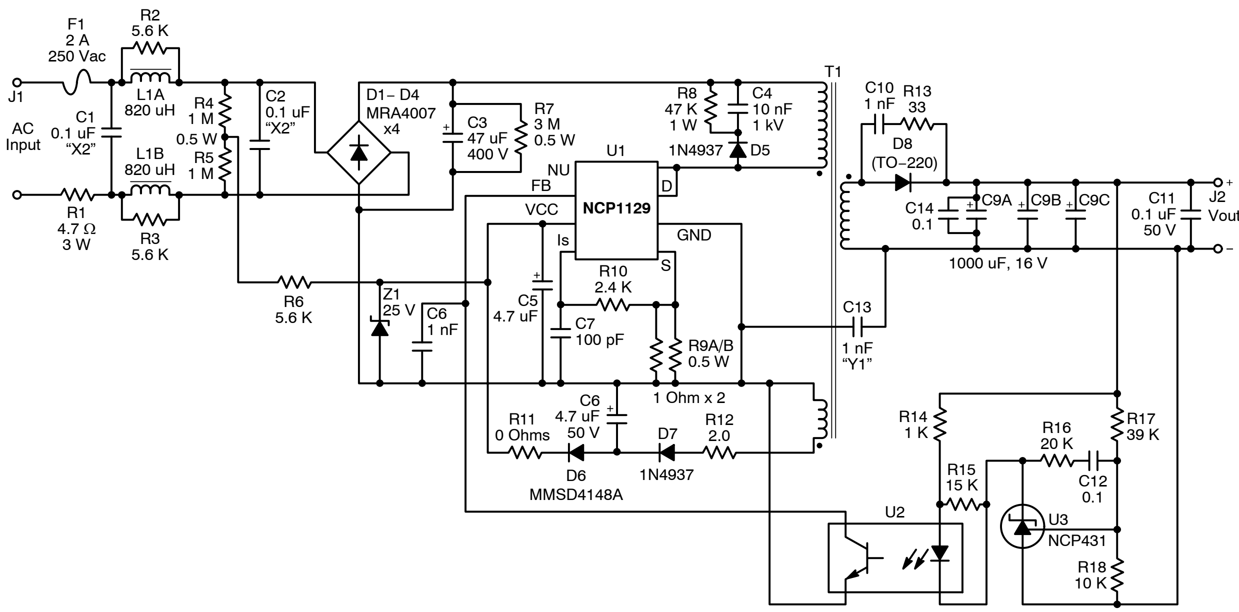


Figure 44. 25 W Offline Power Supply High Voltage Switcher

Table 9. Key Devices

Designator	Description	Value	Footprint	Manufacturer Part Number
D8	Schottky Diode	20 A, 100 V	TO−220	NTST20100CTG
D1, D2, D3, D4	Diode, 60 Hz	1 A, 800 V	SMA	MRA4007
D5, D7	Diode, Fast Recovery	1 A, 600 V	Axial Lead	1N4937
D6	Signal Diode	100 mA, 100 V	SOD−123	MMSD4148A
Z1 (Optional)	Zener Diode	25 V	SOD−123	MMSZ5253B
U3	Programmable Zener	2.5 V	SOT−23	NCP431A
U1	Controller	65 kHz	DIP−8	NCP1129

For more details please refer to
http://www.onsemi.com/PowerSolutions/evalBoard.do?id=NCP1129DIPGEVB

65 W High Efficiency Notebook Adapter

The NCP1236 is a new fixed−frequency current−mode controller featuring Dynamic Self−Supply (DSS). This device is pin−to−pin compatible with the previous NCP12xx families. The DSS function greatly simplifies the design of the auxiliary supply and the VCC capacitor by activating the internal startup current source to supply the controller during transients. Due to frequency foldback, the controller exhibits excellent efficiency in light load condition while still achieving very low standby power consumption. Internal frequency jittering, ramp compensation, and a versatile latch input make this controller an excellent candidate for converters where components cost is the key constraints.

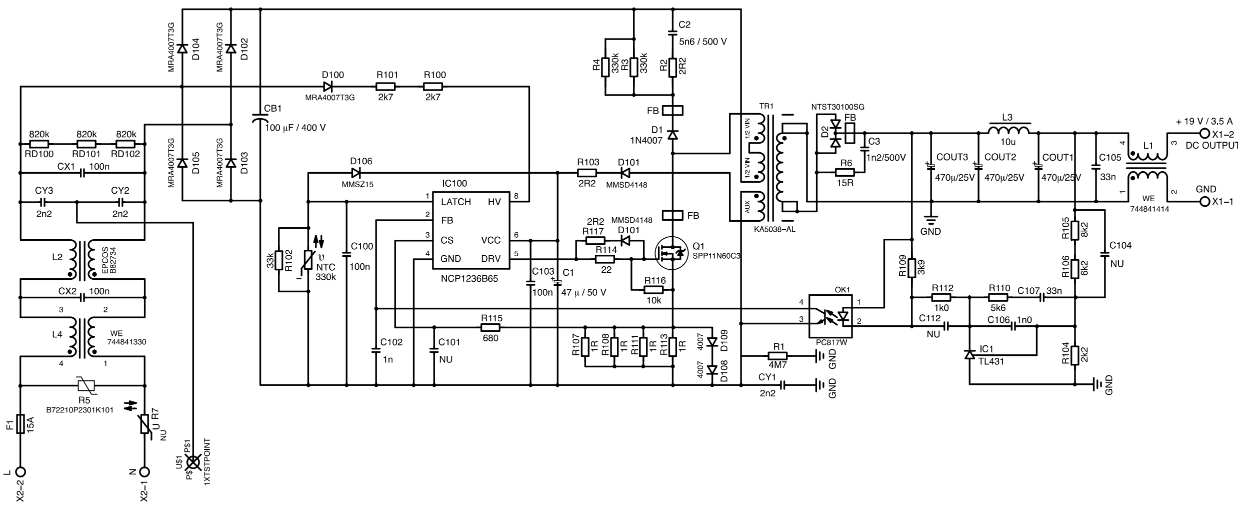


Figure 45. 65 W High Efficiency Notebook Adapter

Table 10. Key Devices

Designator	Description	Tolerance	Footprint	Manufacturer Part Number
D1	Standard Recovery Rectifier	—	DO−41−10B	1N4007G
D100, D102, D103, D104, D105, D108	Standard Recovery Rectifier	—	SMA	MRD4007T3G
D101, D107	Diode	—	SOD−123	MMSD4148T3G
D106	Zener Diode	5%	SOD−123	MMSZ15T3G
D2	Diode Schottky, 150 V, 15 A	—	TO−220	NTST30100SG
IC1	Programmable Precision Reference	—	TO−92	TL431BCLPG
IC100	SMPS Controller	—	SOIC−8	NCP1236BD65R2G

For more details please refer to
http://www.onsemi.com/PowerSolutions/evalBoard.do?id=NCP1236B65NBGEVB

15 W Nominal − 40 W Peak Adapter

The NCP1255 is a highly integrated PWM controller capable of delivering a rugged and high performance offline power supply in a SOIC−8 package. With a supply range up to 35 V, the controller hosts a jittered 65−kHz switching circuitry operated in peak current mode control.

The controller architecture is arranged to authorize a transient peak power excursion when the peak current hits the limit. At this point, the switching frequency is increased from 65 kHz to 130 kHz until the peak requirement disappears. The timer duration is then modulated as the converter crosses a peak power excursion mode (long) or undergoes a short circuit (short).

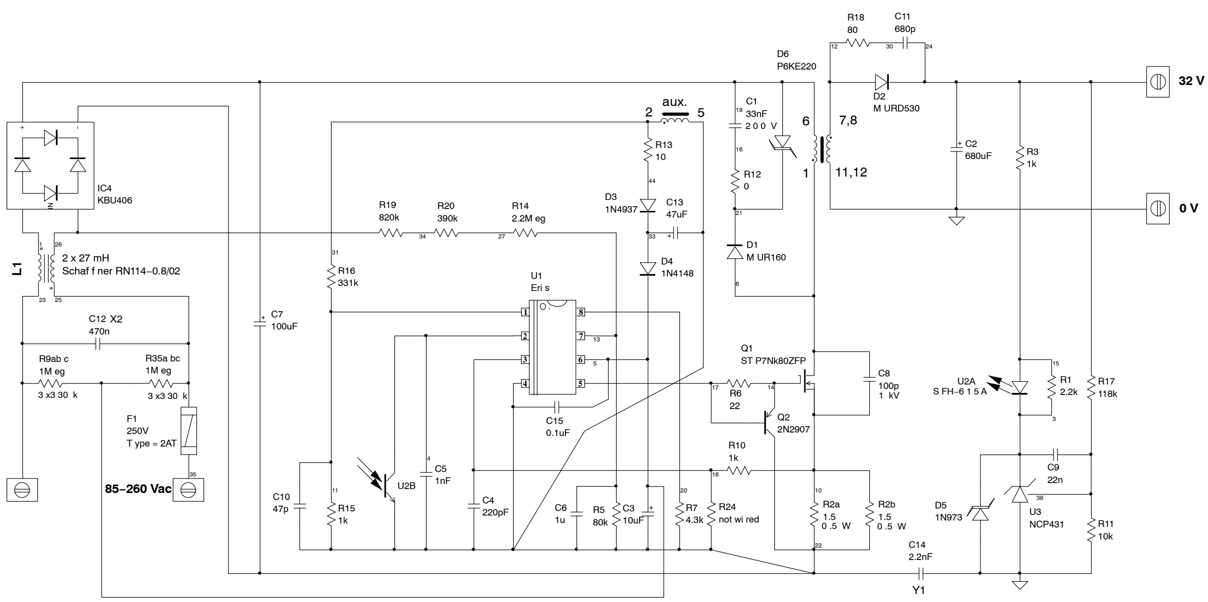


Figure 46. 15 W Nominal − 40 W Peak Adapter

Table 11. Key Devices

Designator	Description	Footprint	Manufacturer Part Number
D1	Ultrafast Diode	Axial	MUR160RLG
D2	Rectifier	DPAK−4	MURD530
D3	Rectifier	Axial	1N4937RLG
U1	PWM Controller	SOIC−8	NCP1255BD65R2G
U3	Shunt Regulator	SMD	NCP431ACSNT1G

For more details please refer to
http://www.onsemi.com/PowerSolutions/evalBoard.do?id=NCP1255PRNGEVB

220 W LED−TV Design Power Supply

The NCP1398 is a high performance controller for half bridge LLC resonant converters. The integrated high voltage gate driver simplifies layout and reduces external component count. A unique architecture, which includes a 750 kHz Voltage Controlled Oscillator whose control mode permits flexibility when an ORing function is required allows the NCP1398 to deliver everything needed to build a reliable and rugged resonant mode power supply.

In association with NCP1607/08 CRM Boost PFC Controller and NCP1027 high integrated HV switching regulator they offer a complete solution.

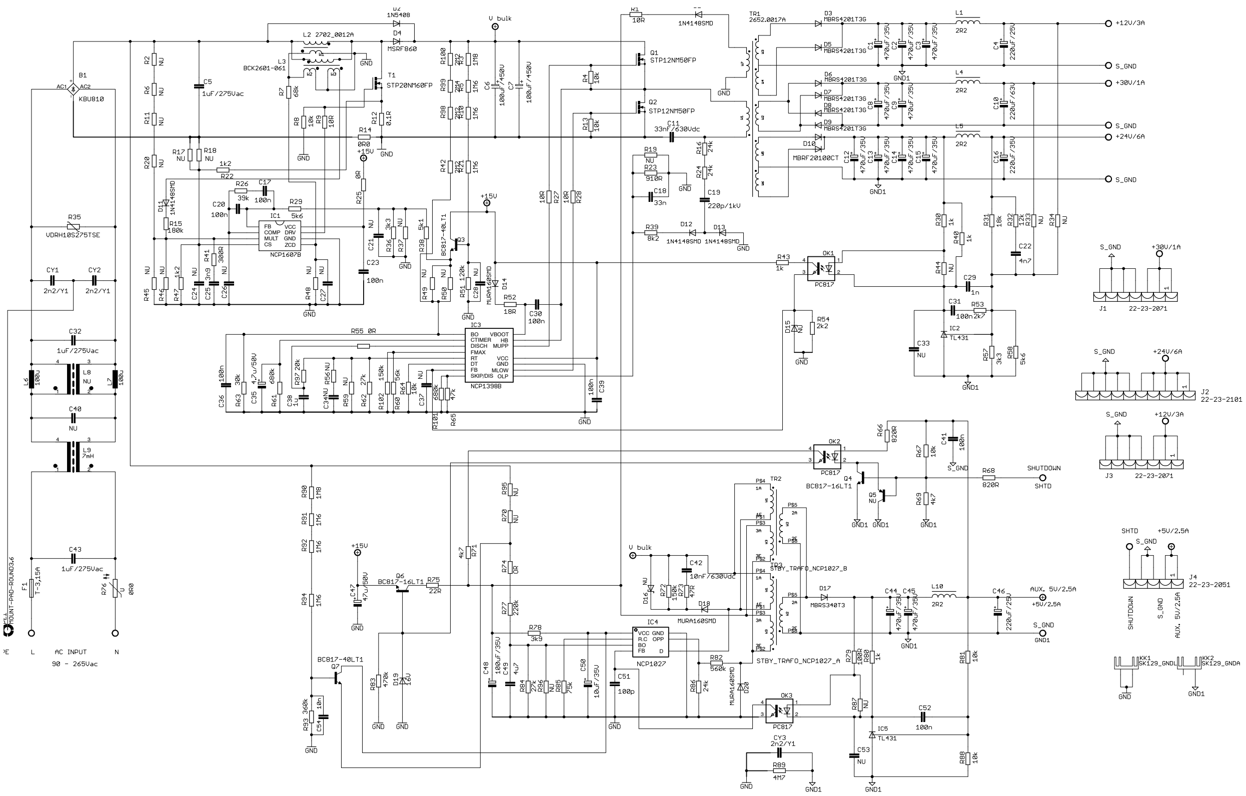


Figure 47. 220 W LED−TV Design Power Supply

Table 12. Key Devices

Designator	Description	Tolerance	Footprint	Manufacturer Part Number
D1, D11, D12, D13	Switching Diode	—	SOD−123	MMSD4148T3G
D2	Standard Recovery Rectifier	—	Axial Lead	1N5408RLG
D3, D5, D6, D7, D8, D9	Diode	—	SMC	MBRS4201T3G
D4	Soft Recovery Rectifier	—	TO−220FP	MSRF860G
D10	Schottky Rectifier	—	TO−220FP	MBRF20100CTG
D17	Diode	—	SMC	MBRS320T3G
D19	Zener Diode	5%	SOD−123	MMSZ16T1G
IC1	PFC Controller	—	SOIC−8	NCP1607BDR2G
IC2, IC5	Programmable Precision Reference	—	TO−92	TL431BCLPG
IC3	Resonant Controller	—	SOIC−16	NCP1398BDR2G
IC4	HV Switcher for Medium Power Offline SMPS	—	PDIP−8	NCP1027P065G

For more details please refer to
http://www.onsemi.com/PowerSolutions/evalBoard.do?id=NCP1398LCDGEVB

High Efficiency and Below 10 mW Standby PFC + QR Combo Chip for 90 W SMPS

This combination IC integrates power factor correction (PFC) and quasi−resonant flyback functionality necessary to implement a compact and highly efficient Switched Mode Power Supply for an adapter application.

The PFC stage exhibits near−unity power factor while operating in a Critical Conduction Mode (CrM) with a maximum frequency clamp. The circuit incorporates all the features necessary for building a robust and compact PFC stage while minimizing the number of external components.

The quasi−resonant current−mode flyback stage features a proprietary valley−lockout circuitry, ensuring stable valley switching. This system works down to the 4th valley and toggles to a frequency foldback mode with a minimum frequency clamp beyond the 4th valley to eliminate audible noise.

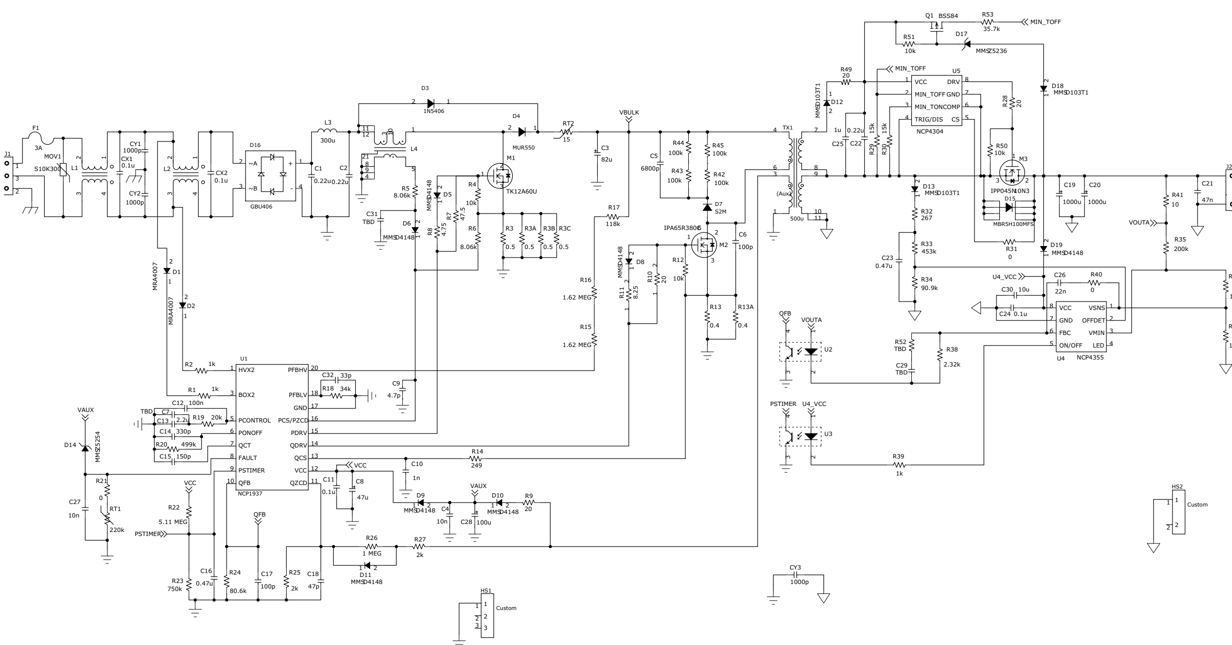


Figure 48. High Efficiency 90 W SMPS Based on NCP1937

Table 13. Key Devices

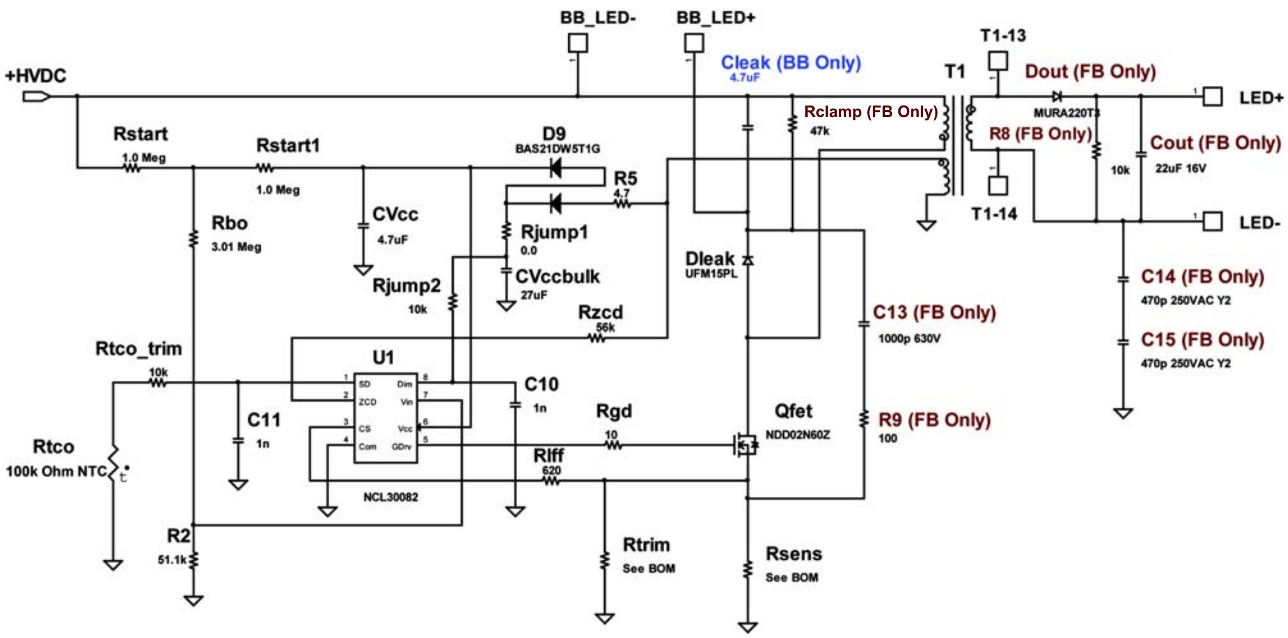
Designator	Description	Value	Footprint	Manufacturer Part Number
D1, D2	Diode, Standard Recovery	1000 V, 1 A	SMD, SMA	MRA4007T3G
D3	Diode, Standard Recovery, Vertical Mount, Teflon Tube Required on Exposed Lead	600 V, 3 A	DO−201AA	1N5406G
D4	Diode, Switch−mode Rectifier, Vertical Mount, Teflon Tube Required on Exposed Lead	520 V, 5 A	DO−201AA	MUR550APFG
D5, D6, D8, D9, D10, D11, D19	Diode, Switching	100 V, 200 mA	SMD, SOD−123	MMSD4148T1G
D12, D13, D18	Diode, Switching	250 V, 100 mA	SMD, SOD−123	MMSD103T1G
D14	Diode, Zener	27 V, 500 mW	SMD, SOD−123	MMSZ5254BT1G
D15	Diode, Schottky	100 V, 5 A	SMD, SO−8FL	MBR5H100MFST1G
D17	Diode, Zener	7.5 V, 500 mW	SMD, SOD−123	MMSZ5236BT1G
U1	Controller, PFC−QR Combo	—	SMD, SOIC−20NB	NCP1937A1
U4	Controller, Sleepmode, Secondary Side	—	SOIC−8	NCP4355B
U5	Synchronous Rectification Driver	—	SOIC−8	NCP4304A

For more details please refer to
http://www.onsemi.com/PowerSolutions/evalBoard.do?id=NCP1937BADAPGEVB

10 W Passive PFC Flyback and Buck−Boost Dimmable LED Driver

The NCL30083 is a PWM current mode controller targeting isolated flyback and non−isolated constant current topologies. The controller operates in a quasi−resonant mode to provide high efficiency. Thanks to a novel control method, the device is able to precisely regulate a constant LED current from the primary side. This removes the need for secondary side feedback circuitry, biasing and an optocoupler. The device is highly integrated with a minimum number of external components.

A robust suite of safety protection is built in to simplify the design. This device is specifically intended for very compact space efficient designs. It supports step dimming by monitoring the AC line and detecting when the line has been toggled on−off−on by the user to reduce the light intensity in 5 steps down to 5% dimming.



NOTE: Components labeled **FB** are only populated for the flyback version and components labeled **BB** are only required for the Buck-boost configuration

Figure 49. 10 W Passive PFC Flyback and Buck−Boost Dimmable LED Driver

Table 14. Key Devices

Designator	Description	Value	Footprint	Manufacturer Part Number
Dout	Ultrafast Power Rectifier	—	SMA−2	MURA220T3
D9	250 V Switching Diode	—	6−TSSOP (5−Lead)	BAS21DW5T1G
Qfet	MOSFET N−Ch 600 V IPAK	600 V	TO−251−3 Short Leads	NDD02N60Z
U1	—	—	—	NCL30083BDMR2G

For more details please refer to
http://www.onsemi.com/PowerSolutions/evalBoard.do?id=NCL30083FLYGEVB

Additional Documentation

Table 15. Design Notes

Document Title	Document ID
Compact 200−265 Vac HiPF Boost LED Driver	DN05062/D
NCP1011: Efficient, Low Cost, Low Standby Power (<100 mW), 2.5 W CCCV Charger	DN06017/D
NCP1012: 1 W, Dual Output, Off−line Converter	DN06002/D
NCP1012: 3.6 W Auxiliary Power Supply for Appliances / White Goods	DN06013/D
NCP1013: Universal Input, 5 W, LED Ballast	DN06027/D
NCP1014, NSI45025: Universal Off−Line LED String Driver with PFC	DN06065/D
NCP1014: 10 W, 3−Output Off−line Power Supply	DN06005/D
NCP1014: 10 W, Dual Output Power Supply	DN06020/D
NCP1014: 12 Watt Mini Boost Power Factor Corrector for LED applications	DN06064/D
NCP1014: 5 V, 400 mA Current Boosted Buck Converter	DN06052/D
NCP1014: 5 W, CCCV Cell Phone Battery Charger	DN06009/D
NCP1014: 8 W Off−Line Non−Isolated Buck Regulator	DN06011/D
NCP1014: 8 W, 3−Output Off−Line Switcher	DN06003/D
NCP1014: Improving the Power Factor of Isolated Flyback Converters for Residential ENERGY STAR® LED Luminaire Power Supplies	DN06051/D
NCP1014: Low Power, Off−Line Buck, CVCC Power Supply	DN06037/D
NCP1014: Low Power, Off−Line, Constant Voltage Non−Isolated Power Supply	DN06066/D
NCP1014: Universal Input, +/− 12 V Output, 8 Watt Power Supply	DN06034/D
NCP1027: 1 A, 12 W Constant Current Off−Line LED Driver	DN06006/D
NCP1027: 10 W, 24 V / 5 V Off−Line Power Supply	DN06012/D
NCP1027: 16 W, 12 Vdc Modem Power Supply	DN06021/D
NCP1028: Single Stage, Off−line, Isolated 12 V, 800 mA Converter with High Power Factor	DN06069/D
NCP1050: Extremely Low AC−DC Power Supply Used with GFCI	DN05037/D
NCP1055, MC33340: Universal AC 15 W charger for 4 NiMH/NiCd cells	DN06023/D
NCP1055: 7 W, non−isolated buck ac−dc converter	DN06028/D
NCP1055: Simple, 3 W, Non−isolated Bias Supply	DN06053/D
NCP1071, NCP1075, MCP1077: 12 Vout, Off−line Buck Regulator	DN05058/D
NCP1072: 10 W Primary Side Regulation (PSR)	DN05036/D
NCP1075, NCP4328: Compact 90−135 Vac HiPF Boost LED Driver	DN05055/D
NCP1075: Non−isolated, 8 W Dual Output, Off−line Power Supply	DN05038/D
NCP1075: Universal AC Input, 12 Vout,10 W E−Meter Power Supply	DN05018/D
NCP1076/7: Universal Input, 5 V or 12 V Output, up to 24 W Power Supply	DN05049/D
NCP1077: 12 Vout, 6 W, Off−line Buck Regulator Using a Tapped Inductor	DN05059/D
NCP1124 : 10 W Universal Input, 5 V output power supply	DN05061/D
NCP1129: 12 W, Off−line Buck Regulator	DN05053/D
NCP1129: Universal AC Input, 5 V or 12 V output, 20 W − White Goods/Industrial	DN05043/D
NCP1136 : 5 V / 20 W Universal Input Power Supply (800 V MOSFET)	DN05054/D
NCP1216: 24 V, Off−line Flyback Converter for Motor and Solenoid Applications	DN06049/D
NCP1216: 7 W, 90−135 Vac, 500 mA LED Driver	DN06050/D
NCP1217: 18 to 24 V, 85 W, Off−line PSU with Short Term, High Surge Current Capability	DN06038/D
NCP1217: 200 W, single output power supply	DN06010/D
NCP1217: Universal AC 40 W Programmable Output	DN06024/D
NCP1219: Universal AC Input, Non−isolated, 6 W Electric Meter Buck Power Supply	DN05014/D
NCP1236: 65 W Off−Line Adapter	DN06074/D
NCP1251: 12 V, 30 W, Off−line Mini−Forward Converter	DN05029/D

Table 15. Design Notes

Document Title	Document ID
NCP1251: 20 W Flyback with Primary Side Regulation (PSR)	DN05033/D
NCP1251: 20 W Output Offline Power Supply	DN05012/D
NCP1251: 20 W Output Offline Power Supply Using Synchronous Output Rectification	DN05028/D
NCP1251: A 24 Vin, 40 Watt, Low Cost, DC−to−DC Converter	DN05032/D
NCP1251: Low Power, Ultra−Wide AC Input Range, Electric Meter Power Supply	DN05017/D
NCP1251: Ultra Wide Input Triple Output 15 W (E−Meter)	DN05039/D
NCP1271: Ultra−Wide Input Range Bias Power Supply	DN06058/D
NCP1308, LM2575: Universal Input, 50 W, 5 Output Quasi−Resonant Flyback Converter	DN06029/D
NCP1308: $\pm$18 V Dual Output Power Supply	DN06008/D
NCP1308: 170 Vdc output, quasi−resonant power supply	DN06014/D
NCP1615: High Efficiency, High Voltage, Active X2, <30mW no load PFC	DN05057/D
NCP1927: 140 W LCD TV Power Supply	DN05001/D
NCP1937: High Efficiency, <10 mW Standby PFC + QR Adapter	DN05044/D

Table 16. Tutorials

Document Title	Document ID	Revision	Revision Date
Thermal Runaway	TND327/D	0	Feb, 2007
Achieving High Power Density Designs	TND325/D	0	Sep, 2007
LED Lighting Definitions & Characteristics	TND328/D	1	Sep, 2007
Standby Power Reduction Techniques	TND324/D	1	Sep, 2007
SMPS Overview	TND342/D	0	Jun, 2008
SMPS Topologies	TND343/D	0	Jun, 2008
AD2 − Adapter Less than 75 W	TND355/D	0	Oct, 2008
ATX − Designing High−Efficiency ATX Solutions	TND356/D	0	Oct, 2008
FB1 − DC−DC Converters Feedback and Control	TND352/D	0	Oct, 2008
MOF − Multi−Output Flyback Off−Line Power Supply	TND351/D	0	Oct, 2008
PFC1 − Advanced Power Factor Correction	TND344/D	0	Oct, 2008
QR − Analysis and Design of Quasi−Resonant Converters	TND348/D	0	Oct, 2008
SIM − Simulating Power Supplies with SPICE	TND349/D	0	Oct, 2008
TEL − DC−DC Telecom and Networking Solutions	TND347/D	0	Oct, 2008
ADAP1 − Fixed Frequency Adapter with Very Low Power Consumption	TND376/D	0	Nov, 2009
ADAP2 − QR Adapter Design	TND377/D	0	Nov, 2009
AiO − 200 W Power Supply for All−in−One PC	TND383/D	0	Nov, 2009
F2S − 2 Switch Forward Converter	TND378/D	0	Nov, 2009
FB2 − The TL431 in Switching Power Supplies	TND381/D	0	Nov, 2009
HB − Half−Bridge Drivers, a Transformer−Based Solution or an All−Silicon Drive	TND379/D	0	Nov, 2009
LCD − AC−DC Power Architecture in LCD TV	TND353/D	1	Nov, 2009
LED − Driving High Brightness LEDs in the General Lighting Marketplace	TND345/D	2	Nov, 2009
MAG − Magnetics in Switched−Mode Power Supplies	TND350/D	1	Nov, 2009
OVR − Overview of Efficient Power Supply Solutions	TND357/D	1	Nov, 2009
PFC2 − Compensating a PFC Stage	TND382/D	0	Nov, 2009
Electromagnetic Compatibility	TND389/D	0	Dec, 2009
Switcher Efficiency and Snubber Design	TND396/D	0	Dec, 2009

Table 17. Application Notes

Document Title	Document ID
100 Watt Universal Input PFC Converter	AND8106/D
19 V, 3 A, Universal Input AC−DC Adaptor Using NCP1271	AND8242/D
2 Switch−Forward Current Mode Converter with the NCP1252	AND8373/D
300 W, Wide Mains, PFC Stage Driven by the NCP1654	AND8324/D
300 W, Wide Mains, PFC Stage Driven by the NCP1653	AND8185/D
32 W, 32 V Universal Input AC−DC Printer Adapter using the NCP1237	AND8454/D
48 W, 24 V/7.5 V Universal Input AC−DC Printer Adapter Using the NCP1219	AND8393/D
5 Key Steps to Design a Compact, High−Efficiency PFC Stage using the NCP1611	AND9062/D
5 Key Steps to Design a Compact, High−Efficiency PFC State Using the NCP1612	AND9065/D
5.0 V, 2.0 A Flyback Converter	AND8099/D
700 mA LED Power Supply Using Monolithic Controller and Off−Line Current Boosted (Tapped Inductor) Buck Converter	AND8328/D
8 W DVD Power Supply with NCP1027	AND8262/D
90 W, Single Stage, Notebook Adaptor	AND8209/D
90 W, Universal Input, Single Stage, PFC Converter	AND8124/D
A 12V/40W Demonstrator with NCP1250	AND8486/D
A 160 W CRT TV Power Supply Using NCP1337	AND8246/D
A 30 W Power Supply Operating in A Quasi−Square Wave Resonant Mode	AND8129/D
A 36 W Ballast Application with the NCP5181	AND8244/D
A 48 V/ 2 A High Efficiency, Single Stage, Isolated Power Factor Corrected Power Supply for LED Drivers and Telecom Power	AND8394/D
A 5 V/2 A Standby Power Supply for INTEL Compliant ATX Applications	AND8241/D
A 6 W/12 W Universal Mains Adapter with the NCP101X	AND8142/D
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A 90 W/19V High Efficiency, Notebook Adapter Power Supply with Inherent Power Factor Correction	AND8397/D
A Quasi−Resonant SPICE Model Eases Feedback Loop Designs	AND8112/D
A Simple 12 Vout, 22 W, Off−line Forward Converter Using ON Semiconductor’s NCP1027/1028 Monolithic Switcher	AND8489/D
A Simple DC SPICE Model for the LLC Converter	AND8255/D
A Simple Low−Cost Non−Isolated Universal Input Off−Line Converter	AND8078/D
Adjusting the Over Temperature Protection Trip Point in NCP1250−Based Adapters	AND9057/D
An Innovative Approach to Achieving Single Stage PFC and Step−Down Conversion for Distributive Systems	AND8147/D
An Off−Line, Power Factor Corrected, Buck−Boost Converter for Low Power LED Applications	AND9043/D
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Implementing Simple Brown−Out Solution for NCP101X Series	AND8272/D
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Key Steps to Design an Interleaved PFC Stage Driven by the NCP1631	AND8407/D
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Safety Tests on a NCP1612−Based PFC Stage	AND9079/D
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Synchronizing NCP1207 (NCP1377/B, NCP1378)	AND8187/D

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The NCP1250, a Versatile Tiny Controller for Offline Converters	AND8469/D
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Tips and Tricks to Build Efficient Circuits with NCP1200	AND8069/D
Understanding Loop Compensation with Monolithic Switchers	AND8334/D
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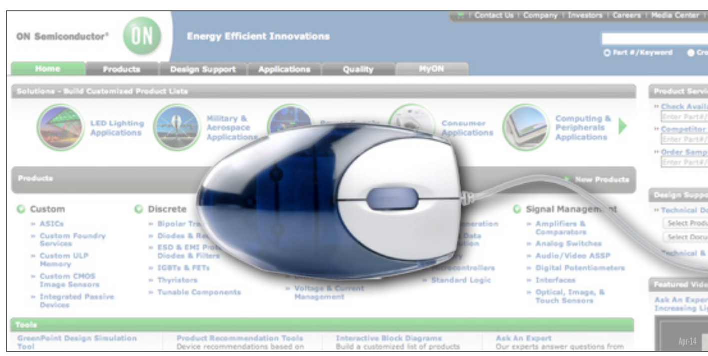

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