Cool Power

Switch mode power supplies: an introduction

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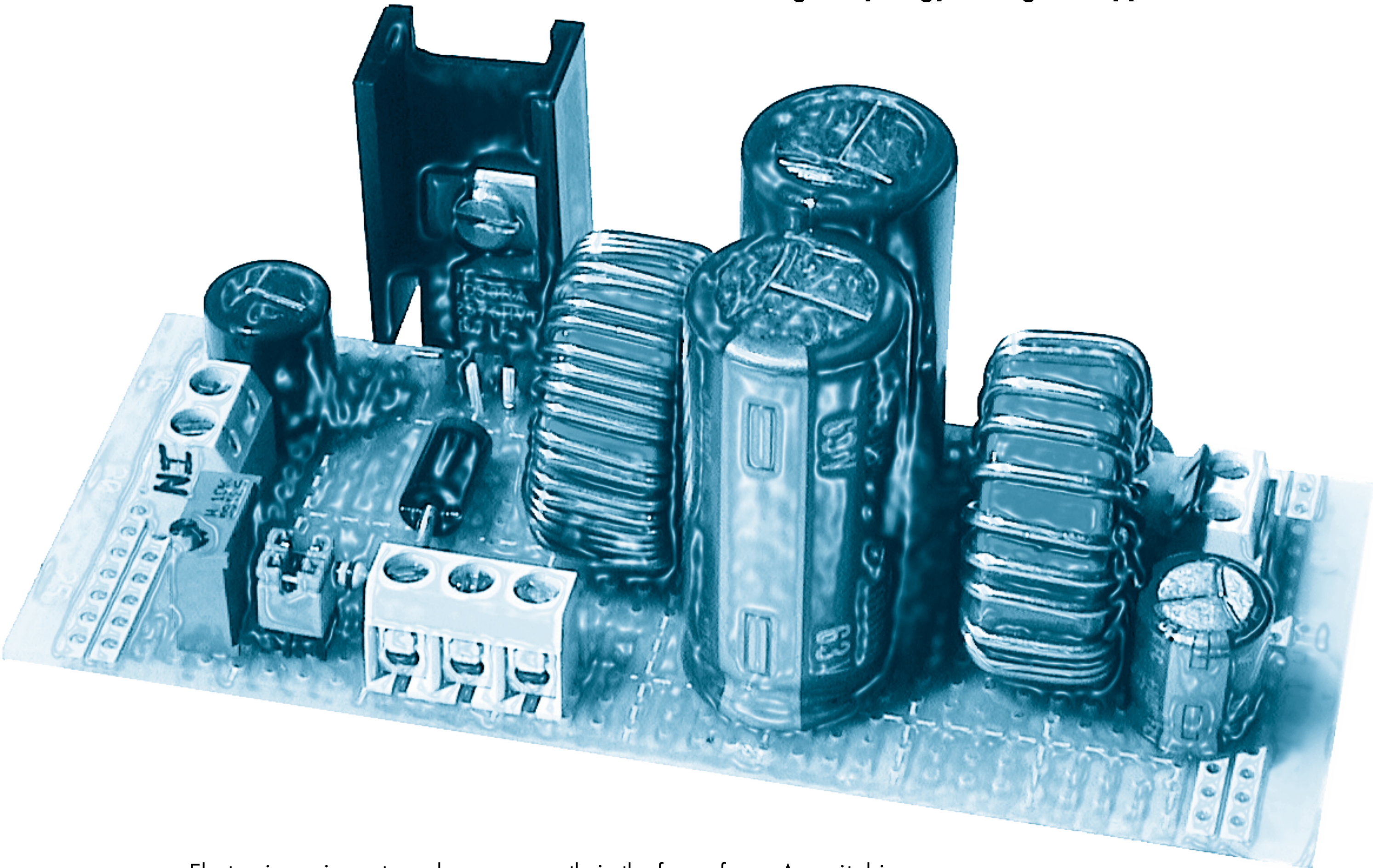
Switch Mode Power Supply Units (SMPSUs) are often too quickly put aside because of their complexity. This article presents an overview to SMPSU technology and a brief description of the most widely used topologies, but in a practical way, trying to demystify the topic and encourage the electronics entusiasts to start developing his or her own 'switchers', or at least know how to choose the right topology for a given application.



Electronic equipment needs power, mostly in the form of a static voltage (DC). The national grid delivers an alternating voltage (AC), so most electronic devices have a dedicated power supply (PSU) to convert this alternating voltage to the voltage that is needed by the circuitry. This conversion is never 100% efficient.

Generally, we have two types of power supplies: linear and switching. A linear power supply uses a pass element such as a transistor that dissipates the excess energy. This, of course results in energy loss due to the simultaneous voltage drop and current flowing through the component.

A switching power supply delivers its power in pulses or periodic 'packets' (at typically 10 kHz to 1 MHz rate) that are averaged to provide a smooth output. The power is transferred to the load (circuit) by means of non-dissipative components (at least theoretically), such as switches that operate only in two states, ON and OFF, inductors, capacitors and transformers. A theoretical switch doesn't dissipate power, ($P=V\times I$) because when it is open, no current flows through it, and when it is closed, it becomes a short-circuit (the voltage drop across it is 0). Of course there will be some losses in practice, but for sure we start from a better point than with linear regulators in terms of energy usage!

Why choose SMPSU?

Efficiency: one of the key advantages of SMPS (switch-mode power supplies) is that power can be converted in a useful way to the load and still keep losses very low, typically below 20% of the useful power. Generally this is way better than a linear PSU.
Size: despite a generally higher parts count, when all things are considered (the heat sink that may be required for this one, for example), a 'switcher's' size can be considerably smaller than a linear PSU. When transformers are involved, such as in AC/DC converters, the reduction is drastic, because switching transformers operate at much higher frequencies (10 kHz-1MHz) so their size can be greatly reduced.
Flexibility: switching regulators easily provide multiple voltages from a single input source, even higher in value or polarity-inverted. This is more difficult and costly to do with linear PSUs.

Disadvantages of SMPS

Noise and interference: the fast transitions occurring in a switching regulator can produce a lot of harmonics that are easily radiated. If not properly controlled, this may cause interference with nearby equipment or with the load itself. This noise can also be conducted in the forms of spurious noise or ripple. Fortunately, a lot of effort has been put into this issue and now it is possible to have very quiet SMPSUs even at high power levels.
Complexity/Reliability: the higher number of parts in a typical SMPSU, including the control circuit, has an impact in reliability; the more parts, the more possible failures. Among them, power semiconductors are the most prone to failure, although with careful design and thanks to the outstanding evolution of semiconductors, very high reliability can be obtained now.
Design difficulty: the design of a switching power supply is totally different and generally more complicated than an equivalent linear PSU, and the designer must know a lot of fields (power electronics, magnetics, EMI/RFI, feedback theory, etc).

Buck converter

The most basic switching converter is the buck converter. It takes a DC input and converts it to a lower DC output. The basic architecture can be seen in **Figure 1**.
The buck converter comprises a switch, a diode (that acts like a second switch), an inductor, and a smoothing capacitor. There are two possible status for the switch, ON and OFF, that are determined by a controller, whose main function is to get the output voltage to the desired value regardless of the input or load variations. When the switch is

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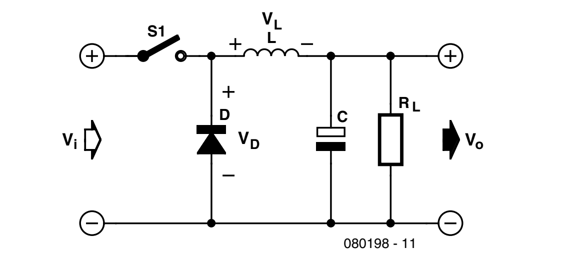

ON (closed), the diode is open as it is directly connected to the input and becomes reverse biased,. Then, $V_{in}=V_L+V_o$, so $V_L=V_i-V_o$. The basic equation of an inductor is $V_L=L dI_L/dt$, then

$$I_L=\int_0^{t_{on}}\frac{Vi-Vo}{L}dt$$

As $-(V_i-V_o)/L$ doesn't depend on time, the increment of the current is constant and it becomes:

$$\Delta I_{L(on)}=\frac{Vi-Vo}{L}t_{on}$$

(i.e. inductor current increases linearly during the ON phase).

When the switch is OFF (open), the input voltage source gets disconnected from the circuit, and the diode gets forward biased, providing a patch for the current to flow. $V_L=-V_o$. So, similarly, we obtain

$$\Delta I_{L(off)}=\frac{-Vo}{L}t_{off}$$

These are only increment rates, not absolute values. The average value of the current, I_{avg}, depends on the load resistance, and will be the mean value of max. and min. currents.

Continuous vs. discontinuous operation

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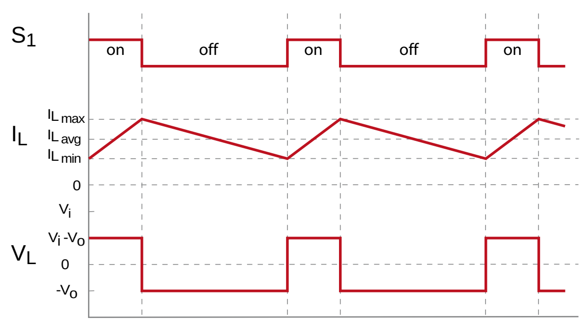

In **Figure 2** you can see the aspect of the inductor current, that is a triangle centered in the average current, rising and falling linearly when the switch is ON and OFF, respectively.

We assume that I_L never drops to zero. We will call this mode *"Continuous Conduction Mode (CCM)"*. Then in steady state it is clear that I_L at the start of the cycle ($t=0$) must be the same as I_L at the end of the cycle ($t=T$), because if not, the average current would indefinitely increase or decrease. Thus, the increments $\Delta I_{L(on)}$ and $\Delta I_{L(off)}$ must be equal but of opposite sign. That means that $\Delta I_{L(on)} = -\Delta I_{L(off)}$, so

$$\frac{Vi-Vo}{L}t_{on} = \frac{Vo}{L}t_{off}$$

If we assign *D* (duty cycle) to the portion of the period *T* (=1/F_s, switching frequency) that the switch is on, $D=t_{on}/T$, then $t_{on}=DT$, so the rest of the time the switch is off: $t_{off}=T(1-D)$. Substituting,

$$\frac{Vi-Vo}{L}DT = \frac{Vo}{L}(1-D)T$$

or, simplified:

$$\frac{Vo}{Vi} = D$$

D will be always less than or equal to 1, so V_o will always be less than or equal to than V_i. That's why the Buck converter is also called 'step down converter'.
Note that, in CCM, V_o only depends on the duty cycle and V_i. For example, if we want 5 V from a 12 V input, the controller will have to turn the switch on during 5/12=0.4166 (41.66%) of the time, regardless of the load, leaving it off the rest of the cycle.

If the current drawn by the load is not large enough, the former current waveform will evolve as shown in **Figure 3**, eventually reaching zero when the switch is OFF. This mode of operation is called Discontinuous Conduction Mode (DCM). In DCM we can't apply the above equation that the increments $\Delta I_{L(on)}$ and $\Delta I_{L(off)}$ have the same magnitude, and the math become more complicated. In this case the relation between V_o and V_i is not as simple as in CCM. For reference, the output voltage can be calculated then as:

$$Vo = \frac{Vi}{\frac{2L \cdot Iavg}{D^2 \cdot Vi \cdot T}+1}$$

As you can see, now it also depends on the switching period (T), the value of the inductor and the input voltage itself.
The minimal average current that guarantees that the inductor current doesn't drop to zero (so we keep the converter in CCM), is:

$$I_{avg(CCM)} \geq \frac{Vi(1-D)DT}{2L}$$

If one wants to keep control of a buck converter simple, a minimal load must be provided so it remains in CCM. There is no cause that impedes us running a buck regulator in DCM mode, although in this case it is sometimes better to use the variation of the switching frequency as the control parameter, instead of the duty cycle, (FM, frequency modulation instead of PWM, pulse width modulation).

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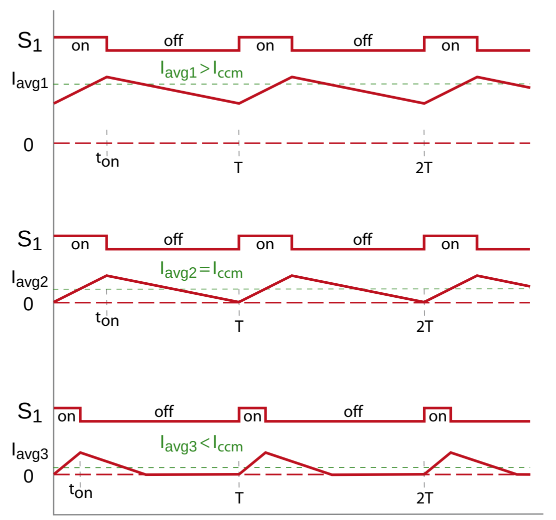

Boost converter

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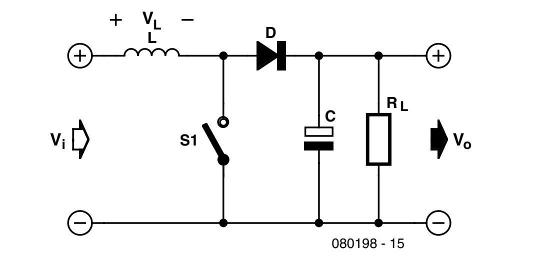


Figure 4 shows the basic architecture of a boost converter. The analysis is very similar to the buck converter, the main equations being $V_L = V_i$ (ON state), so:

$$\Delta I_{L(on)} = \frac{Vi}{L}t_{on}$$

Note that, during the ON state, the current to the load is supplied by the storage capacitor, thus the output voltage is smooth if it is large enough.

For the OFF state $V_L=V_i-V_o$, so:

$$\Delta I_{L(off)} = \frac{Vi-Vo}{L}t_{off}$$

Now we find that:

$$\frac{Vo}{Vi} = \frac{1}{1-D}$$

D will always be less than or equal to 1, so V_o will always be equal or greater than V_i. That's why the Boost converter is also called 'Step up converter'.

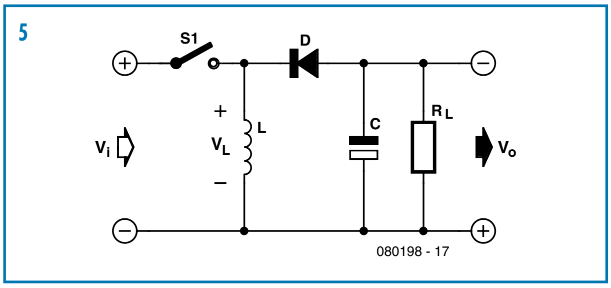
In CCM, output voltage only depends on the duty cycle and input voltage. For example, if we want 12 V from a 5 V input, the controller will have to turn the switch on during 58.3% of the time, regardless of the load, leaving it off the rest of the cycle.

However, as opposed to buck regulators, boost ones are more commonly used in DCM for stability reasons. The expression for the output voltage becomes:

$$\frac{Vo}{Vi} = 1 + \frac{V_i D^2 T}{2LI_{avg}}$$

In order to allow for core flux reset and avoid saturation, D is usually limited to around 0.8.

Buck-boost or inverting converter



The inverting converter is used to obtain negative voltages from a positive source. **Figure 5** shows the basic schematic. The analysis is very similar to the boost converter. The main equations for the ON state are V_L=V_i, so:

$$\Delta I_{L(on)} = \frac{Vi}{L} t_{on}$$

For the OFF state V_L=–V_o, so

$$\Delta I_{L(off)} = \frac{-Vo}{L} t_{off}$$

Being in parallel with the inductor, the capacitor gets charged to a negative voltage during OFF phase, and then provides current to the load when the switch goes ON again, allowing for assumedly constant output. The following relationship between V_o and V_i exists:

$$\frac{Vo}{Vi} = \frac{-D}{1-D}$$

So the buck-boost converter can provide a voltage from theoretically 0 to theoretically minus infinity from a positive input voltage. Of course, this converter can also operate in DCM if the load current I_{avg} is not large enough. In this case, analyzing the storage of energy in the coil, we can demonstrate that the V_o/V_i relation becomes:

$$\frac{Vo}{Vi} = -\frac{V_i D^2 T}{2LI_{avg}}$$

There are of course countless variations of these basic topologies, such as the "Cuk" converter, that provides the same V_o/V_i relationship as the Inverter topology but using two inductors.

NOTE: There is another topology also called 'Buck-boost', consisting of a buck converter followed by a boost converter, with the additional difference that it doesn't invert polarity.

Isolated converters

When the input or output voltage is high enough to be dangerous, an isolated topology must be used, where the separation between input and output is not accomplished only by a semiconductor, but by a physical dielectric barrier. The circuit is then split in two parts: the *primary* (that gets power directly from the source) and the *secondary* (where the output(s) are connected). International regulation dictates some norms about the required distances (named clearance and creepage) between both parts in order to ensure safety. These norms must be had in mind when designing an isolated converter.

The typical application of isolated converters (although not the only one) is AC/DC (mains input) PSUs, also called "offline converters". There are several topologies, some of them directly derived from the basic regulators explained before, while others are a bit more complicated and appeared in order to provide higher power levels.

Flyback converter

The basic schematic can be seen in **Figure 6**. It is clearly derived from the buck-boost (inverting) converter: the input inductor has been substituted by a transformer, with the diode and capacitor connected in its secondary. The dots and diode polarity have been rearranged so the output voltage is positive. Finally, the connection between primary and secondary has been removed to provide galvanic isolation.

When S1 turns ON, the transformer current starts to ramp up linearly at a rate $t \times V_i/L_{pri}$, where L_{pri} is the primary inductance. As primary and secondary have opposite polarity due to the different position of the dots, the secondary voltage is multiplied by the turns ratio N_2/N_1 and inverted in polarity, so the diode gets reverse biased. The load is

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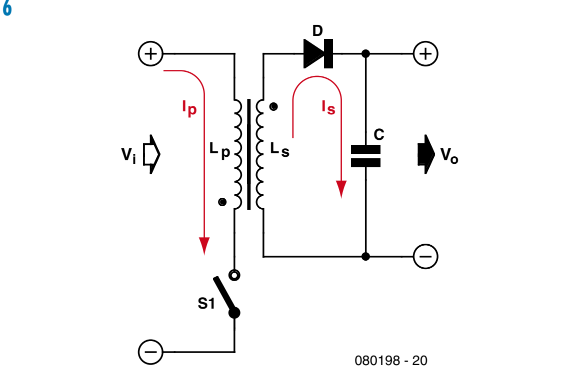

supplied current by the capacitor alone, that we assume previously charged.

When S1 turns OFF, there is no primary current, but the polarity reverses, and so does the secondary voltage: the diode gets now forward biased and hence the core energy can be discharged through the secondary (see **Figure 7**), at a linear rate of $-t \times V_o/L_{sec}$. During this time, the load receives current, and so does the capacitor, that replenishes its charge for the next ON time.

During OFF time, the secondary voltage is reflected back ("flies back") to the primary, multiplied by the turns ratio, so the rating of S1 must be at least $V_{in}+V_o \times N_1/N_2$ plus some margin.

Unlike in other types of isolated converters, the flyback transformer stores energy itself (as well as the capacitor that supplies the load during the first part of the cycle), and is usually constructed with ferrite materials, with a 'gap' or separation between the core halves that increases energy storage capacity.

Using the expression of the energy stored in an inductor during the ON time, and assuming that it fully discharges during the OFF time (discontinuous mode), the expression for the output voltage is:

$$\frac{Vo}{Vi} = D \cdot \sqrt{\frac{T \cdot V_o}{2 I_{avg} L_{pri}}}$$

The controller can adjust the output voltage by means of the duty cycle *D*.

Note that the voltage at the primary is always positive, so only one quadrant of the transformer *B/H* curve is used, leading to inefficient core use. Other topologies such as push-pull or half/full-bridges can get at least double the power from the same core volume. But even when a flyback is useful only for <200 W typically, it has lower cost (doesn't need output inductors, uses a single primary switch and secondary diode, etc), and that's one of the main reasons for its popularity. Flyback converters can be found in every TV set, monitors, laptop chargers, small adapters, etc.

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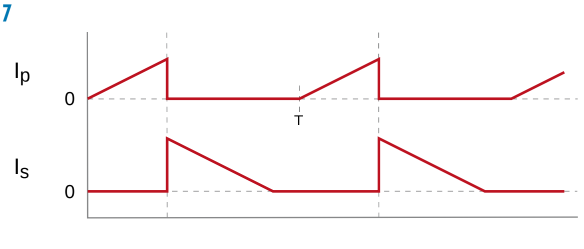

The flyback can be constructed with many output voltages with the only addition of separate secondaries. Cross-regulation (regulation of each output when the load of another output changes) is particularly good in this topology, that's another reason of its success.

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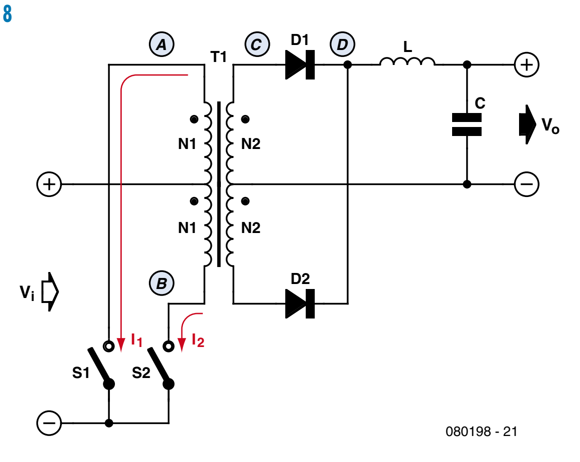

Push-Pull converter

When higher power is needed, better transformer utilization is required, and hence a topology that uses two quadrants of the *B-H* curve (provides positive and negative voltage swing to the primary). The push-pull is one of these topologies, see **Figure 8**.

The transformer has a centre-tap in the primary, connected to the input source V_i, so it really has two identical primaries in series, each having N_1 turns. The same happens in the secondary, there are two in series each one having N_2 turns.
Both switches are activated by a control voltage with duty cycle varying between 0 and 50%. Both switches can never be ON at the same time. **Figure 9** shows the basic waveforms of this converter.

When S1 or S2 turns ON, its corresponding primary is set to almost 0V, so it primary 'sees' V_i. The total primary voltage then swings from $-V_i$ to $+V_i$. The current of each primary ramps up linearly during the corresponding ON time due to the primary inductance. The primary voltage is multiplied by N_2/N_1 and applied to the secondary. The corresponding diode gets forward biased, so when any of both switches is on, there is current in one of the secondaries, so the inductor current, that supplies the load (and capacitor) ramps up.
When both switches S1 and S2 are OFF, the diodes block and the only current to the load is provided from the output inductor (and the smoothing capacitor), that starts ramping down at a rate $-t \times L_1 \times V_o$. The expression for the output voltage is

$$\frac{Vo}{Vi} = 2D \cdot \frac{N_2}{N_1}$$

The controller can adjust the output voltage by means of the duty cycle. Note that we can get an output voltage that is lower or higher than the input, depending on the transformer construction.

There is a potential problem with push-pull converters that has limited their use: if the flux swing magnitude is not exactly the same for both half-primaries, the core will eventually 'walk' into saturation. Its inductance drops drastically, behaving nearly as a short-circuit, so the switches will be destroyed. This can be detected because the current waveforms of the switches don't have the same amplitude, and when the situation is really critical, one of the waveforms can start to curve upwards at the end of its ON time.
This is less of a problem with MOSFETs, that provide some auto-correction due to their negative temperature coefficient ($R_{ds(on)}$ increases with current, so primary voltage drops due to the higher $V_{ds(on)}$).

Note that the voltage each switch withstands is twice the input voltage, so they are not very suitable for high power off-line converters (each switch would have to be rated at nearly 1KV and also high current, being expensive). This kind of converter is preferred for lower $V_{i.}$
A typical application of push-pull converters are step-up inverters for powering audio amplifiers from car batteries, up to 1KW. The primary currents are huge, but the voltage rating of the MOSFETs is only 30-60 V, so there are many high-current devices readily available.

Half-bridge and Full-bridge converters

For 230 VAC off-line converters, the voltage in the push-pull MOSFETs may become unpractical. Half-bridge and full-bridge converters, on the other hand, allow for a more relaxed rating of the switches, while still providing high output power and good use of the transformer. The schematic of the simpler of both, the half bridge, can be seen in **Figure 10**.

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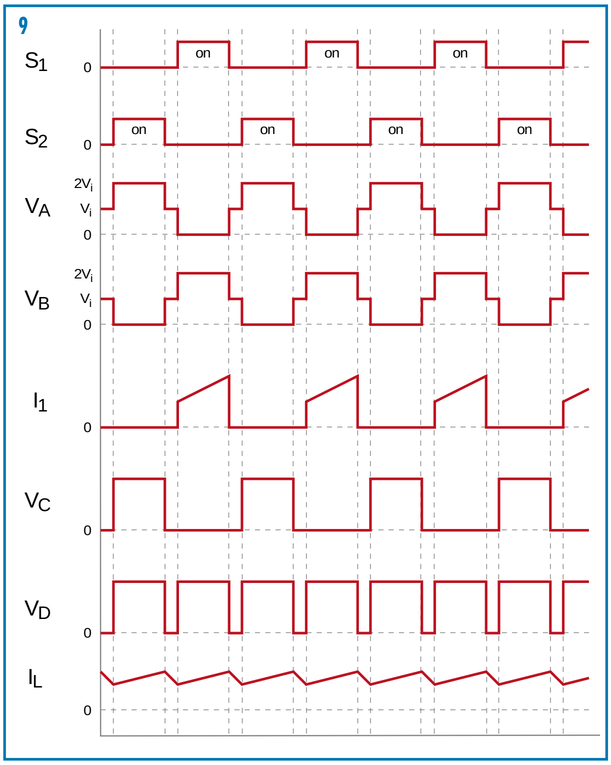

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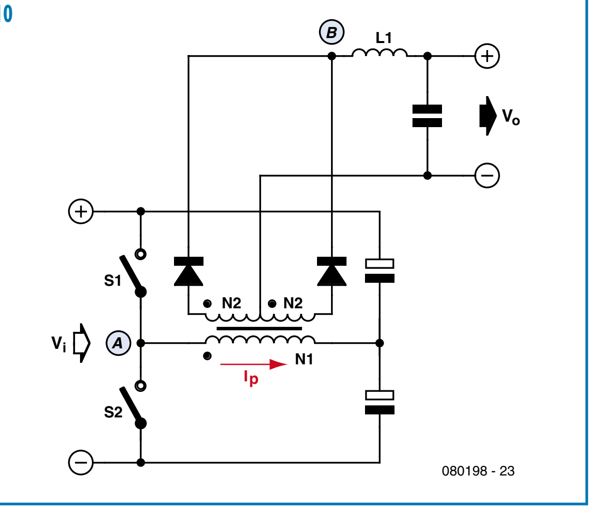

This topology also uses two switches, two rectifier diodes and 1 output inductor. The transformer has a single primary with N_1 turns, and two center-tapped secondaries, each one having N_2 turns. Note that the other transformer primary leg is connected to $V_i/2$, built with a capacitive voltage divider.

Both switches are activated by a control voltage with duty cycle varying between 0 and 50%. Both switches can never be ON at the same time. **Figure 11** shows the basic waveforms of this converter.

When S1 or S2 turns ON, the transformer leg that is connected between them is switched either to V_i or 0V, and as the other leg is fixed at $V_i/2$, the total voltage in the primary swings from $-V_i/2$ to $+Vi/2$. The current of the primary ramps up (in magnitude) linearly during the ON times. The primary voltage is multiplied by N_2/N_1 and applied to the secondary. The corresponding diode gets forward biased, so when any of both switches is on, there is current in one of the secondaries and hence one of the diodes is forward-biased, so the inductor current that supplies the load (and

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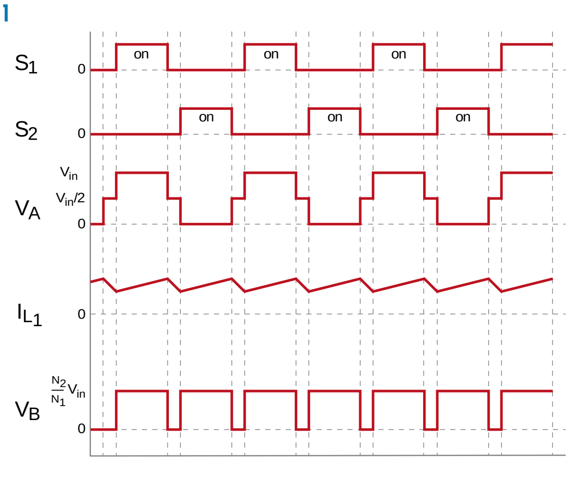

capacitor) ramps up.
When both switches S1 and S2 are OFF, the diodes block and the only current to the load is provided by the output inductor (and the smoothing capacitor), that starts ramping down at a rate $-t \times L_1 \times V_o$.
The expression for the output voltage is similar to the push-pull topology, but as the voltage swing of the transformer is halved:

$$\frac{Vo}{Vi} = D \cdot \frac{N_2}{N_1}$$

The controller can thus adjust the output voltage by means of the duty cycle. Note that we can also get an output voltage that is lower or higher than the input.
The full bridge is very similar, but the transformer primary is connected between two sets of switches, for a total of 4. The left side top turns ON simultaneously with the right side bottom, and conversely. The capacitor voltage divider doesn't exist, and the voltage swing in the transformer primary is doubled (and hence its use is fuller). The output/input relationship becomes:

$$\frac{Vo}{Vi} = 2D \cdot \frac{N_2}{N_1}$$

Sometimes, a coupling capacitor C_c is added in series with the primary, in order to remove any DC in the transformer windings and avoid saturation. C_c must have a value large enough so that it doesn't produce significant (<5%) voltage "droop" on the top of the transformer primary voltage waveform. A high voltage polypropylene of good quality ceramic cap should be used. The voltage-divider capacitors in half-bridge must also be properly dimensioned, with a high-valued (20-100 k) resistor in parallel with each one to ensure proper balancing at $V_{bus}/2$.
Half-bridge converters are commonly used for medium-power (250 W-1 KW) offline applications, the main one being PC PSUs. Another example of this kind of PSUs can be found in the **SAPS-400 Audio SMPSU** described elsewhere in this issue. This one has two symmetric outputs with coupled inductors for good cross-regulation, as well as a couple of auxiliary windings.
Although full-bridges have four switches, the control circuit is the same as for half-bridge or push-pull, as the MOSFETs turn on in pairs so only two signals have to be provided. However, isolation is required for each gate-source voltage, as the sources of each MOSFET are not common.
Full-bridges are more costly, so they are reserved for powers above >1 KW.

Losses in the converters

Although the equations obtained above don't show this, there are of course some losses during the conversion in all this kind of converters, that have two main components:
Conduction losses occur when the switch is ON, due to its ON resistance, $R_{ds(on)}$ in MOSFETs. In formula:

$P_{cond} = I^2_{switch} \times R_{ds(on)} \times D$

(the switch is only closed during a part of the cycle D)
Switching losses: real switches don't change instantaneously from ON to OFF, there are rise and fall times (t_r and t_f). During these times both current and voltage drop is developed at the switch simultaneously, producing power dissipation. These losses can be approximated by:

$$P_{switching} = \frac{V_{switch} I_{switch}}{2} (t_r + t_f) f$$

There are other kinds of losses, such as the gate driver losses, the losses associated to the diodes "recovery time", losses in the wire and core of the magnetic components, etc., but they are usually smaller.
A typical simple converter can quite easily reach efficiencies of around 90%. As a comparative example, if one needs 5 V from a 12 V input, with 2 A load current (P_{out}=10 W), a linear regulator will dissipate (12–5)×2=14 W. A buck converter would do the job dissipating only around 1.2 W.

What topology to choose?

On the Elektor website we present a table and a flowchart for free downloading. With it you can easily choose a topology that's best suited to the application. Topics as safety, input/output voltage, output power and costs will be covered also.

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Web Links (reference & tutorials)

www.smps.com

www.smps.us

www.onsemi.com/pub_link/Collateral/SMPSRM-D.PDF

http://sound.westhost.com/project89.htm

http://schmidt-walter.eit.h-da.de/smps_e/smps_e.html

http://members.tripod.com/valveaudio/Membuatsendiri.htm