Technical Engineering Notes

POWER CONVERSION - BASIC TOPOLOGIES

Introduction

It could be argued that there is no need for the user of power conversion products to have any interest in the principles of power conversion. All that is necessary is to match a "black-box" with the requirements of the application.

Provided output ratings, noise and ripple, input range, size, shape, connectors, safety and RFI standards etc., are all acceptable, what is inside the "black-box" is not important. The knowledge that busy design engineers are up against ever shorter development time targets lends sympathy to this view. Powerbox believe that some appreciation of the various circuit topologies in present day use is necessary to optimise selection of the power converter for a specific application.

Linear Power Supplies

Written off by many as "yesterday's technology", dubbed bulky, heavy and inefficient, they still power 50% or more of electronic equipment in use throughout the world.

The advantages of linear power supplies are listed as a reminder of their useful characteristics.

- Simple, rugged, very reliable, easy to maintain - low cost of ownership
- Moderate purchase price
- Excellent line and load regulation - 0.01% to 0.1%
- Very low output ripple and noise - 1mV to 10mV peak to peak
- Rapid recovery to load transients - 50µS
- Only one safety isolation barrier - mains transformer
- No AC mains on the PCB - therefore no creepage problems
- Very low levels of RFI - lower than class A without filters

Fig. 17 shows schematically the circuit of a linear power supply with a series dissipative regulator.

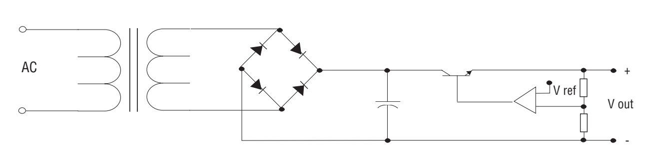

Fig 17 Linear power supply with series dissipative regulator

Switched Mode Power Supplies (SMPS)

Switched mode power supplies were originally developed for military and aerospace use in the 1960s where the bulk, weight and power consumption of linear units were unacceptable. Since then, many different switching circuit topologies and control techniques have been developed. Some of these, in general use in commercial and industrial grade SMPS, are discussed in the following subsections.

Input Rectification and Smoothing

The switching converters used in SMPS are DC/DC converters, therefore the AC supply must be rectified and smoothed to DC with an acceptable peak to peak ripple amplitude.

Almost all commercially available SMPS use the input circuit shown in Fig. 18a to allow operation from 90V to 132V AC RMS and 180V to 264V AC RMS for worldwide use. With the link S open the rectifier diodes are a full bridge configuration producing approximately 320V DC from a 230V AC RMS line input. With S closed, the circuit works as a voltage doubler giving approximately the same 320V DC from an AC line input of nominally 115V RMS. As a voltage doubler, on positive half cycles D1 and D2 conduct, charging C1 to a peak voltage of 160V (115 x ÷2). On negative half cycles C2 is charged to a negative 160V peak via diodes D3 and D4, so the voltage produced across C1 and C2 in series totals 320V.

To avoid overheating, the electrolytic capacitors must be low ESR types (Equivalent Series Resistance) and their working voltage must be at least 200V. The rectifying diodes must have a high PIV (Peak Inverse Voltage) rating of 600V minimum.

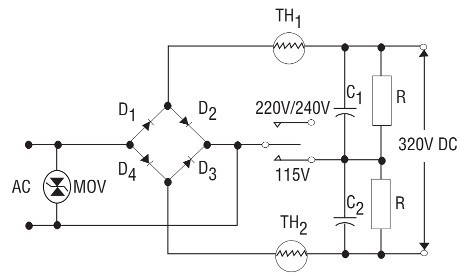

Fig 18a Voltage doubler input circuit used in most SMPS

Inrush Current

When switching on a SMPS with the input circuit of Fig. 18a, the impedance presented to the AC line is the ESR of the capacitors. Without additional series resistance the resulting first half cycle peak current would be hundreds of amperes. TH1 and TH2 are NTC (Negative Temperature Coefficient) thermistors limiting inrush current to a safe value. They have high cold resistance and low hot resistance. Self heating due to current flow causes the resistance to decrease until a low steady state resistance value is attained.

Some medium to high power SMPS employ active circuits with a current limiting resistor which is switched out of circuit by a triac shunt when the capacitors have become fully charged. Product specifications normally advise that inrush current is limited for cold starts (thermistor) or by active means.

Transient Protection

Apart from the line filtering which was discussed under the heading EMC, the other important built in protection device which is connected across the input is a transient line voltage suppressor. This protects the power converter against high voltage mains spikes caused by local inductive switching and electric storms. A suitable device is a metal oxide varistor (MOV) of appropriate voltage and current rating as shown in Fig. 18a.

Wide Input Range

Some equipment manufacturers need to be able to market a standard product worldwide without specifying AC operating range. Responding to this demand, power units are now available which automatically switch between the low and high ranges (in effect replacing the link S in Fig. 18a with an automatically controlled switch).

But low (to medium) power SMPS are becoming available with the ability to operate from 90 to 264V AC in a single continuous range. These units do not need the voltage doubler input circuit, and are usually easily recognisable by the single large 400V rated electrolytic capacitor in the input circuit.

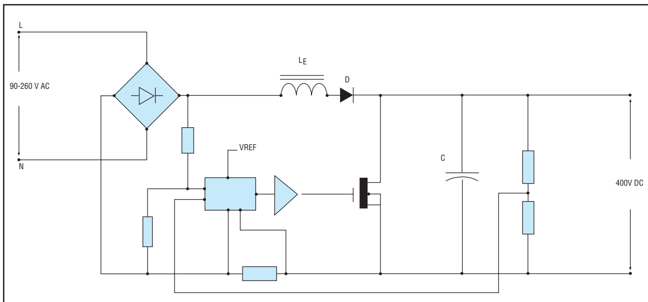

Fig 18b. Simplified schematic showing typical SMPS input circuit with HF switching PFC boost pre-regulator

PFC Input Circuit

Wide input range also comes as a bonus with the use of a switched mode boost pre-regulator for input harmonic current limitation to ensure compliance with EN61000-3-2. The typical scheme shown in Fig 18b functions essentially as follows. The full wave rectified input voltage is used as the reference input for a current mode switching regulator. This controls the rapid charge and discharge of the energy storage choke LE, forcing the average input current to track the reference waveform. An overall voltage control loop regulates the output voltage to approximately 400VDC. Initially, SMPS designers implemented such schemes with discrete component controllers. There are now available a number of integrated circuit PFC switching controllers. Typically they would comprise a current multiplier, precision voltage reference, error amplifier, oscillator, PWM with output gate drive, and input under/overvoltage detection and shutdown, all in a standard 16 pin DIP. Controllers of this type allow input voltage operation from 90V-260VAC and result in power factors from 0.95 to 0.99. A further benefit is that the output voltage across the storage capacitor C is constant; therefore the energy stored is constant. Output hold up times then depend only on the load. It is constant for fixed loads, not at all dependent on input voltage level.

Types of Switched Mode Converters

Converters based upon the following topologies may be found in both AC/DC SMPS and in DC/DC converters.

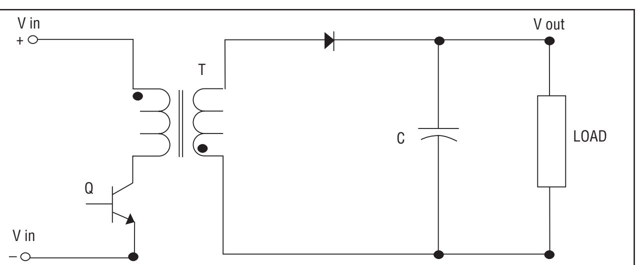

Fig 19 Flyback converter

Flyback Converter

Because of its circuit simplicity and low cost this has become the favoured topology for the majority of low power converters (up to 100 watts). Fig. 19 shows the fundamental principles of flyback conversion. When Q conducts, current rises linearly in the primary of the transformer. The transformer is designed to have a high inductance so that energy is stored as the flux builds up. Winding polarity ensures that diode D is reverse biassed during this period so no transformer secondary current flows. When Q turns off, flux in the transformer decays generating a secondary current which flows to the load and charges capacitor C. Energy is stored in the transformer field during the "on" period of the switch Q and transferred to the load during the "off" (flyback) period. Capacitor C is a reservoir maintaining voltage across the load during the "on" period. Regulation of the secondary voltage is achieved by comparing the output voltage with a reference voltage and using the difference to vary the "on" time of Q. Energy transfer to the secondary is therefore controlled to maintain secondary voltage independently of load and supply voltage changes.

Varying the "on" time of Q may be carried out by pulse width modulation (PWM) driven at a constant frequency by a local oscillator, or, in the simplest and lowest cost schemes, by self oscillation. In this case as the load changes, the frequency varies to vary the "on" time.

Multiple Output Flyback Converter

Figure 20 illustrates the simplicity of adding isolated outputs to a flyback converter. For each output an extra secondary winding, a diode and a reservoir capacitor are the basic requirements. To provide load regulation for the auxiliary outputs, three terminal series regulators are often used, at the expense of decreased overall efficiency.

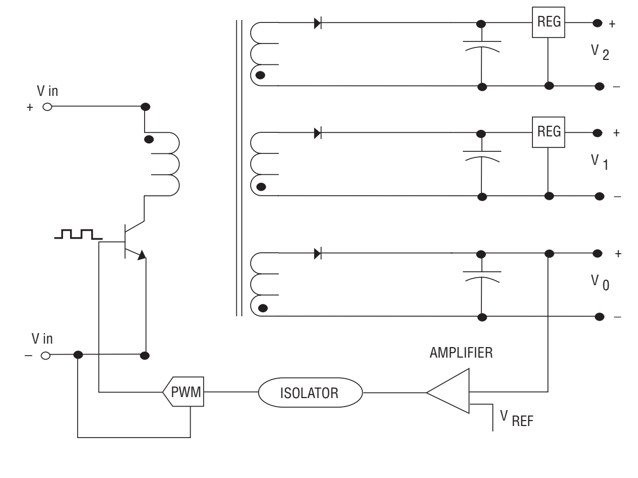

Fig 20 Multiple output flyback converter with regulated outputs

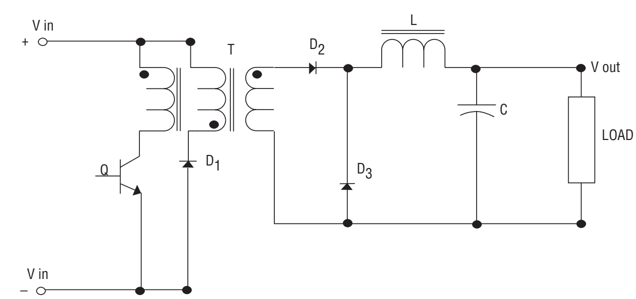

Fig 21 Forward converter

Forward Converter

The forward converter is cost effective from 100 to 250 watts. As Fig. 21 shows, it is more complex than the basic flyback circuit, but its operation is still fairly simple.When switching transistor Q is turned "on", current builds up in the primary and secondary windings simultaneously. Because winding polarity forward biases D2, secondary current flows via L to the load and flux builds up in L creating an

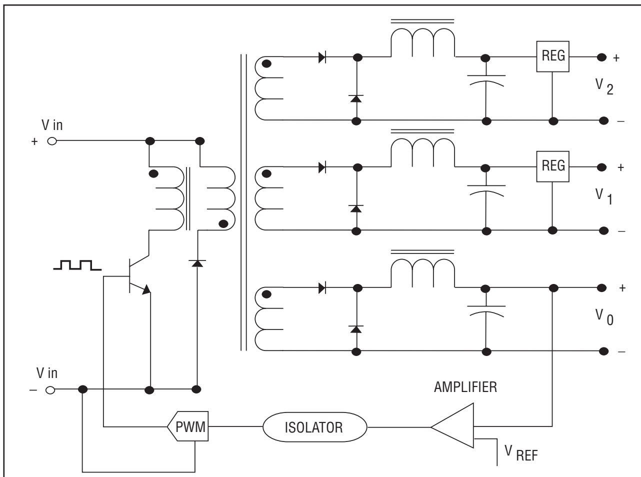

Fig 22 Forward converter with 3 fully regulated isolated outputs

energy store. Q turns “off”, primary current collapses reversing the secondary voltage. D2 is now reversed biased blocking current flow in the transformer secondary, but D3, the “flywheel” diode, now conducts allowing L to discharge into the load. The third winding shown allows energy remaining in the transformer leakage inductance at the end of the “on” cycle to be returned to the primary DC bus via D1. This winding is sometimes called the reset or demagnetisation winding. Unlike the flyback converter current flows from the energy storage inductor to the load during on and off times of the switch Q resulting in half the peak currents in the secondary diodes and inherently lower output ripple.

Multiple Output Forward Converter (Fig. 22)
For each additional isolated output an extra secondary winding is needed on the transformer with two fast recovery diodes, an energy storage inductor and a filter capacitor. The diodes must each be capable of supplying the full output rating. The additional components make the circuit realisation more expensive than the simpler flyback converter. Series regulators can be used to improve the regulation of the auxiliary outputs.

Push Pull Converter (Fig 23)
To make more efficient use of the magnetic cores push pull converters have been developed. Essentially the push pull converter consists of two forward converters driven by antiphase inputs. The two diodes D1 and D2 in the secondary act as both forward and flywheel diodes. Ideally conduction times of Q1 and Q2 are equal, the transformer is driven symmetrically, and unlike the single ended converters, because there is no DC component in the drive no air gap is necessary in the magnetic path. This results in approximately half the core volume for the same power.

A major difficulty arises with this circuit because the transformer saturates if the characteristics of the switching transistors are not precisely matched. Core saturation results in rapid thermal runaway and destruction of one of the transistors. Solutions to this problem are complex and expensive. So for higher power (200 watt plus) applications, half bridge and full bridge converters have been developed.

Half Bridge Converter
This has become the most popular converter technique for the 200 to 400 watt power range, extending to 600 watts with integral forced cooling. From Fig. 24 it can be seen that the transformer primary is connected between the junction of the input bridge reservoir capacitors, which is floating at a nominal voltage of 160V DC (half peak rectified AC supply voltage) and the junction of Q1 emitter and Q2 collector. Antiphase pulsing of Q1 and Q2 alternately connects the transformer to the positive and negative input DC bus generating a 320V peak to peak square wave which is stepped down, rectified and filtered to produce the required DC output voltage.

One advantage is immediately obvious. Even with switching spikes unsuppressed the maximum voltage stress on the transistor switches is the peak rectified supply voltage (Vin) whereas in the basic flyback, forward and push pull converters it is 2Vin. Frequently 400V to 500V switching transistors are used in half bridge circuits whereas 800V to 1000V devices are common in single ended configurations. This explains why almost all circuits using MOSFET switches are based upon half bridge designs (or full bridge for very high power).

Unbalanced switching of Q1 and Q2, which in effect creates a DC bias leading to core saturation and transistor failure, is overcome by the series coupling capacitor C3. Regulation is normally achieved by comparing the output with a reference and using the difference to control pulse width. Most PWM control ICs currently available are suitable for use in half bridge converters.

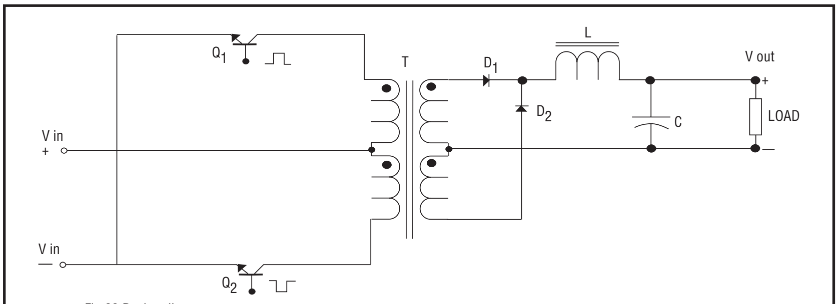

Fig 23 Push pull converter

Some of the advantages of half bridge converters are:

- Small magnetic cores
- No magnetic path gap therefore low stray magnetic field
- Frequency seen by secondary circuits is double the basic switching frequency
- Small filter components (L and C) in secondary circuits.
- Low output ripple and noise
- Multiple output configurations easily implemented
- Relatively low radiated noise, especially if the secondary inductors are toroidal cores.

A disadvantage is because they are working at half the supply voltage the switching transistors are working at twice the collector current compared with the basic push pull circuit.

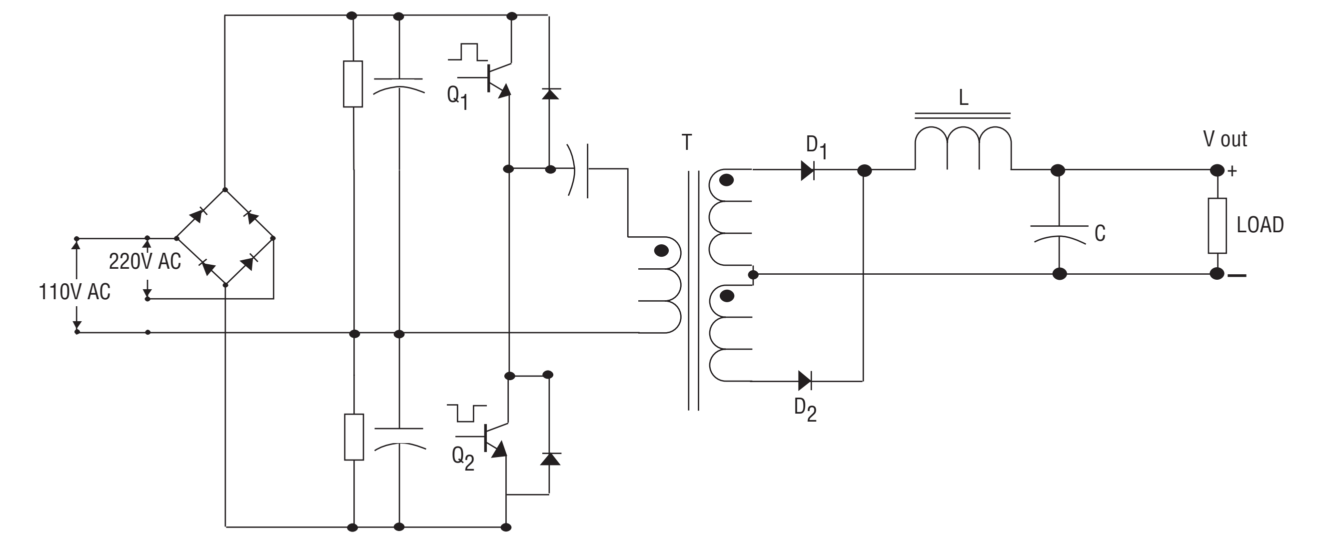

Fig 24 Half bridge converter

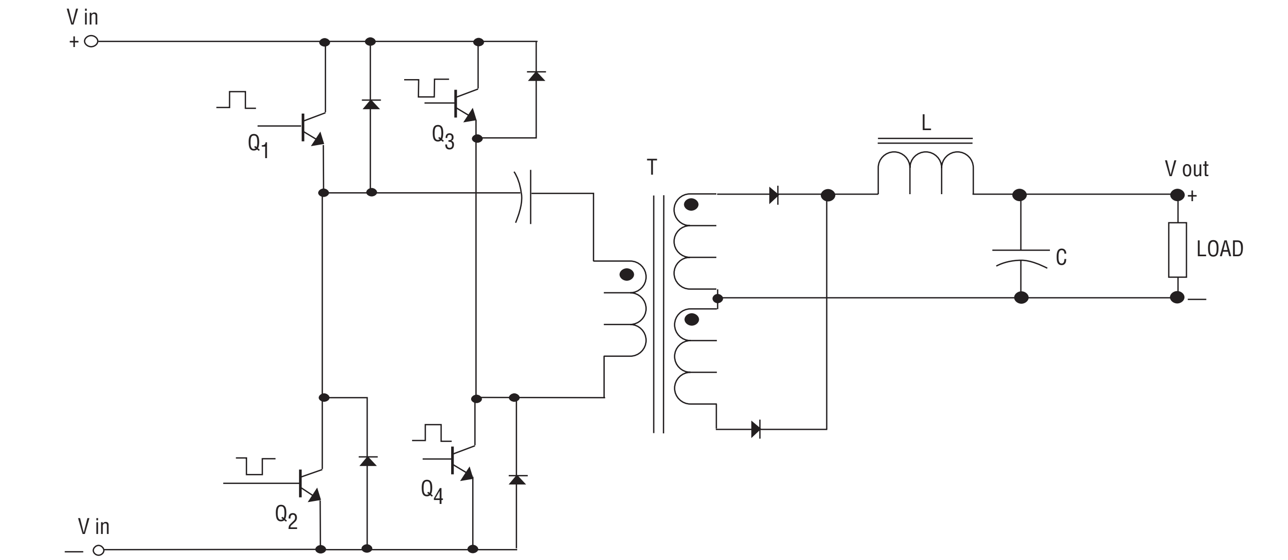

Fig 25 Full bridge converter

Full Bridge Converter

For powers in excess of 500 watts the collector currents of the switching transistors in half bridge converters become excessive. Fig. 25 shows the basic topology of a full bridge converter where the transistors in opposite arms of the bridge Q1 and Q4 are driven in phase and Q2 and Q3 are driven by the antiphase pulse. The transformer primary is therefore pulsed from +Vin to -Vin nominally 640V peak to peak), but the transistors are limited to Vce (Vce = collector emitter voltage) of Vin and the current is half that of an equivalent power half bridge. The use of four switching transistors which must each have an isolated drive makes the full bridge topology expensive.

The much simpler and lower power gate drive circuits required by MOSFETS when compared with the complex base drive requirements of bipolar transistor switches greatly favours their use in high power full bridge circuits. The higher switching frequencies achievable with MOSFETS gives considerable size advantages in the transformer core and in filter components. This is well illustrated by referring to the Power Ten range of high power rack mounted units.

Current Mode Control

Although not a new concept, this method of controlling switched mode converters has only recently become popular. This is because of the widespread availability of integrated circuit control elements containing PWM and current mode control circuits in a single inexpensive IC.

Current mode control uses dual feedback loops. There is the conventional output voltage feedback loop via an error amplifier, and a second loop which senses the current in the output inductor (primary of the converter transformer) and compares this with the error amplifier output. The switching transistor is switched on by the start of the clock pulse, but ends its conduction phase when the inductor current is sufficient to null the output of the error amplifier. Fig. 26 shows a simple schematic circuit illustrating the main features of current mode control.

Advantages of current mode control are improved transient response and dynamic stability, inherent current limiting and easy current sharing in parallel schemes. A basic voltage regulated switched mode converter is a third order control system. This requires compensation to limit the loop gain at high fre-

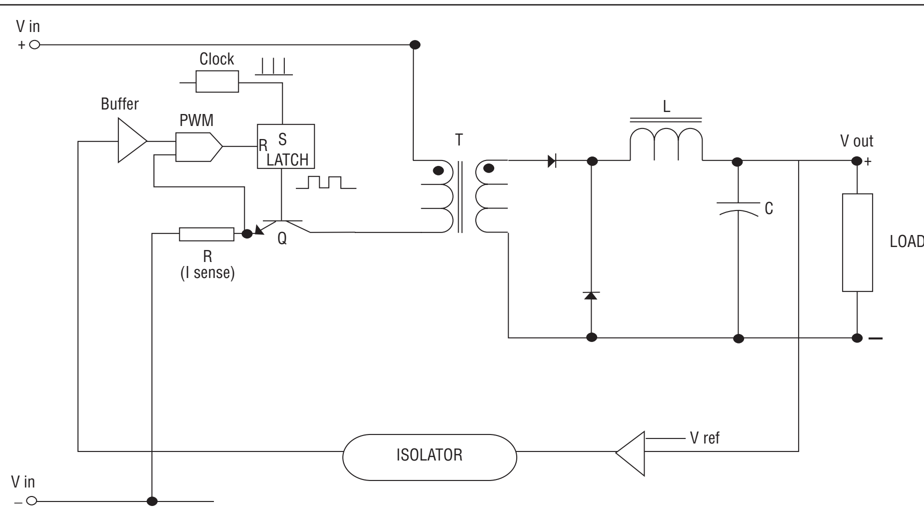

Fig 26 Dual loop current mode control

quencies. Therefore dynamic performance is poor and there is a tendency to “ringing” when subjected to transient disturbances. Adding current mode control converts the open loop to a first order system (one dominant pole) making stabilisation easy and allowing a higher gain band width product. This improves transient response without any tendency to “ring”.

A further claim for current mode control is that it is “EMC friendly”, because the tight current control prevents the transformer saturating. The major source of radiated noise in conventional switched mode converters is the transformer flux leakage which is maximised when the transformer core saturates (on peak current demands for example). Another noise source is “ringing” under transient conditions and as pointed out above this is minimised by the good stability characteristics of current mode control. Current mode control is relatively easy to implement with flyback and forward converters, but more difficult with push pull and bridge types especially for multiple output configurations.

Resonant Mode Conversion

In the quest for higher power densities, designers have been increasing switching speeds in power converters. 100kHz is now relatively common in converters using conventional PWM control. Above this frequency switching losses, component limitations and EMI pose problems which are difficult to overcome at reasonable costs. Some of the drawbacks encountered by square wave switching at high frequencies are virtually eliminated by using resonant mode techniques.

The two important characteristics of resonant mode conversion are that switching occurs at zero current so there are no switching losses, and the current waveform is essentially sinusoidal resulting in low dI/dt, low switching stresses in components and no broadband EMI. There are many different ways of implementing resonant mode conversion. When combined with a fixed frequency PWM control the term quasi-resonant switching is often used. Most commercially available resonant converters achieve regulation by fixing the on or off time of the switch and modulating the frequency.

There are two fundamental resonant mode topologies, series loaded and parallel loaded as shown in Fig. 27. The LC combination is known as the resonant tank circuit and could be on either the primary or secondary side of the converter.

Primary side resonance as shown in Fig. 28 is more commonly used. Benefits of resonant mode techniques start at switching frequencies around 200kHz and continue up to 2MHz. With zero voltage switching as an alternative approach to zero current switching experimental models operating at switching frequencies up to 10MHz have been reported.

One of the disadvantages of resonant mode conversion is that multiple output units based upon a single converter are not practicable. The only practicable solution is one converter per output which is expensive by comparison to conventional converters. A quasi-resonant parallel converter configured as a half bridge driving an LC tank circuit in the primary side is a good compromise which does allow multiple outputs while conferring the benefits of zero current switching, sinusoidal current waveform and high frequency operation.

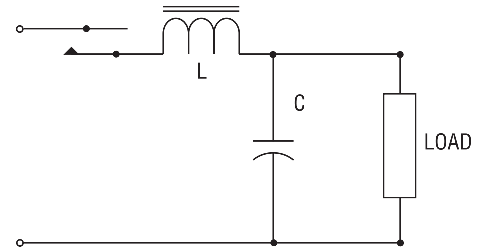

Fig 27a Series resonant tank - parallel loaded

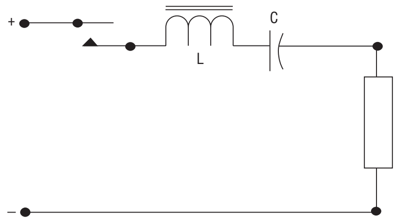

Fig 27b Series resonant tank - seriesloaded

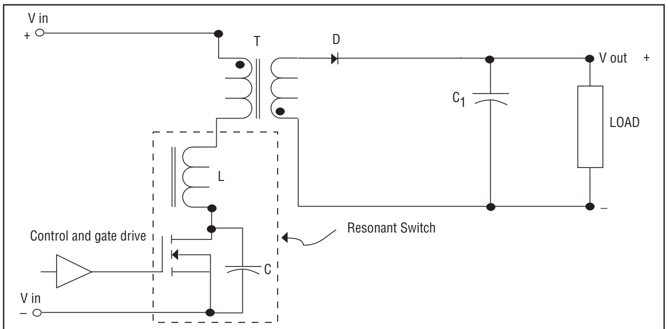

Fig 28 Primary side resonant tank circuit (parallel loaded)