Department of Electrical Engineering
NIT Rourkela

Design and Implementation of ZCS Buck Converter

Project Report for Final Evaluation

Submitted by:

Gyana Ranjan Sahu(10602019)

Bimal Prasad Behera(10602044)

Rohit Dash (10602043)

Guided by:

Prof. A.K. Panda

Signature of the guide:

CERTIFICATE

This is to certify that the thesis titled – "**Design and Implementation of a Zero Current Switched Buck Converter**" submitted by **Sri Gyana Ranjan Sahu, Sri Rohit Dash** and **Sri Bimal Prasad Behera** in partial fulfillment of the requirements for the award of Bachelor of Technology Degree in Electrical Engineering at the National Institute of Technology, Rourkela (Deemed University) is an authentic work carried out by them under my guidance and supervision. The matter embodied in the thesis has not been submitted to or published in any other University / Institute for the award of any Degree or Diploma to the best of my knowledge and belief.

Date: 08/05/10

Prof. A.K.Panda
Department of Electrical Engg.
National Institute of Technology
Rourkela - 769008

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Date : 08/05/10

Gyana Ranjan Sahu

Rohit Dash

Bimal Prasad Behera

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Abstract

Buck converters are step-down DC-DC converters that are widely being used in different electronic devices like laptops,PDA's,cell phones and also electric vehicles to obtain different level of voltages. These converters are nothing but ,high frequency switching devices operating on PWM principle. The need for more and more lighter and smaller electronic devices propels the need for reduced size of converters operating at higher load currents. With all these inadvertent conditions the switching frequency has jumped from KHz range to MHz range.The switching devices are made to turn on and turn off the entire load current at high di/dt , and also withstand high voltage stress across them.Due to these two effects there occurs increased power losses in these converters and reduces the efficiency significantly. The reduction in efficiency is highly unacceptable as it leads to shorter battery life and derated device conditions.

The shortcomings explained above can be minimised and upto some extent eliminated if each switch is made to turn-on and turn-off when the voltage across it and/or current through it is zero at the instant of switching. The converter circuits which employ zero voltage and /or zero current switching are known as Resonant converters. In most of these converters some form of L-C resonance is used, that is why these are known as resonant converters.

In this project a detailed study of zero current switching buck converters is done and also practically implemented in hardware. In addition a mathematical analysis of switching loss occuring in MOSFET's is also presented and a short study of zero voltage switching is also appended. During the hardware implementation the T_{on},T_{off} and operating frequency were found out and thoroughly tuned through the IC555 circuit and various waveforms across inductors,capacitors,load resistor and test points were noted down. These waveforms were found to be in precise proximity of the theoretically observed waveforms.

CHAPTER 1

INTRODUCTION

DC-DC converters are electronic devices that are used to change DC electrical power efficiently from one voltage level to another. The use of one or more switches for the purpose of power conversion can be regarded as a SMPS. A few applications of DC-DC converters are where 5V DC on a personal computer motherboard must be stepped down to 3V, 2V or less. In all of these applications, we want to change the DC energy from one voltage level to another, while wasting as little as possible in the process. In other words, we want to perform the conversion with the highest possible efficiency. DC-DC Converters are needed because unlike AC, DC can't simply be stepped up or down using a transformer. In many ways a DC-DC converter is the DC equivalent of a transformer. They essentially just change the input energy into a different impedance level.

SWITCHING MODE REGULATORS

DC converters can be used as switching-mode regulators to convert a dc voltage, normally unregulated to a regulated dc output voltage. The regulation is achieved by PWM at a fixed frequency and the switching device is normally BJT, MOSFET, or IGBT. The output of dc converters contains harmonics and the ripple content is normally reduced by an LC filter.

Switching regulators are commercially available as integrated circuits. The designer can select the switching frequency by choosing the values of R and C of frequency oscillator. As a rule of thumb, to maximize efficiency, the minimum oscillator period should be about 100 times longer than the transistor switching time; for example, if a transistor has a switching time of 0.5µs, the oscillator period would be 50µs, which gives the maximum oscillator frequency of 20kHz. This limitation is due a switching loss in transistor. The transistor switching loss increases with the switching frequency and as a result the efficiency decreases. In addition, the core loss of inductors limits the high-frequency operation. Control voltage is obtained by comparing the output voltage with its desired value. The reference voltage can be compared with a saw-tooth voltage to generate the PWM control signal for the dc converter. There are three basic topologies of switching regulators.

- Buck regulators
- Boost regulators
- Cuk regulators

Furthermore, depending upon the direction of current and voltage flows, dc converters can be classified into five types:

- First quadrant converters
- Second quadrant converters
- First and second quadrant converters
- Third and fourth quadrant converters
- Four-quadrant converters

CHAPTER 2

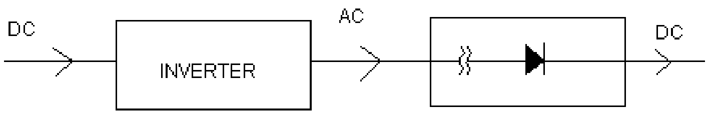
CHOPPER CIRCUITS

Many industrial applications require power from DC sources. Several of these applications, however, perform better in case these are fed from variable DC voltage sources. Examples of such DC system are subway cars, trolley buses, battery-operated vehicles, battery charging etc.

From an AC supply systems, variable DC output voltage can be obtained through the use of phase controlled converters or motor-generator sets. The conversion of fixed DC voltage to an adjustable DC output voltage through the use of semiconductor devices, can be carried out by the use of two types of DC to DC converters mentioned below.

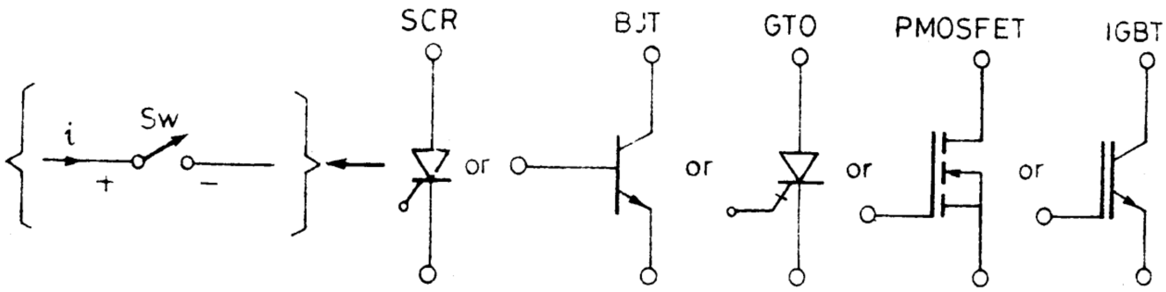
(1) AC link chopper :

In the ac link chopper dc is first converted to ac by an inverter (dc to ac converter), ac is then stepped-up or stepped-down by a transformer which is then converted back to a dc by a diode rectifier. As the conversion is in two stages, dc to ac and then ac to dc, the link chopper is costly, bulky and less efficient.



(a) AC link chopper

(b) dc chopper



(c)Representation of power semiconductor device

(2) DC Chopper :

A chopper is a static device that converts fixed dc input voltage to a variable dc output voltage directly. A chopper may be thought of as dc equivalent of an ac transformer since they behave in an identical manner. As choppers involve one stage conversion, these are more efficient.

Choppers are now being used all over the world for rapid transit systems. These are also used in trolley cars, marine hoists etc. The future electric automobiles are likely to use choppers for their speed control and braking. Chopper systems offer smooth control, high efficiency, fast response and regeneration. The power semiconductor devices used for a chopper circuit can be force-commutated thyristor, power BJT, power MOSFET, GTO or IGBT. Like the transformer, a chopper can also be used to step-down or step-up the fixed input voltage.

CHAPTER 3

PWM STEP DOWN OPERATION

The principle of step down operation is explained as follows. When a switch SW, known as the chopper is closed for a time t_1, the input voltage V_s appears across the load. If the switch remains off for a time t_2, the voltage across the load is zero. The waveforms for the output voltage and load current shown below. The converter switch can be implemented by using a (1)power bipolar junction transistor(BJT), (2) power metal oxide semiconductor field effect transistor (MOSFET) (3) gate turn-off thyristor(GTO), or (4) insulated-gate bipolar transistor(IGBT). The practical devices have a finite voltage drop ranging from 0.5 to 2V, and for simplicity we neglect the voltage drop of these power semiconductor devices.

The average output voltage is given by

$$V_a = (1/T)* \int_{t_0}^{t_1} V_0 \, dt = \frac{t_1}{T} V_s = f\, t_1 V_s = (kV_s)$$

and the average load current, $I_a = V_a/R = kV_s/R$

where T is the chopping period;

$k = t_1/T$ is the duty cycle of chopper;

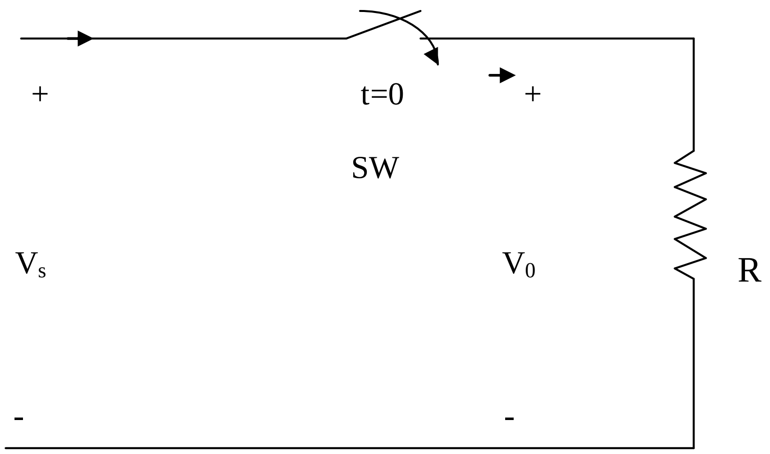
f is the chopping frequency.

The effective input resistance seen by the source is

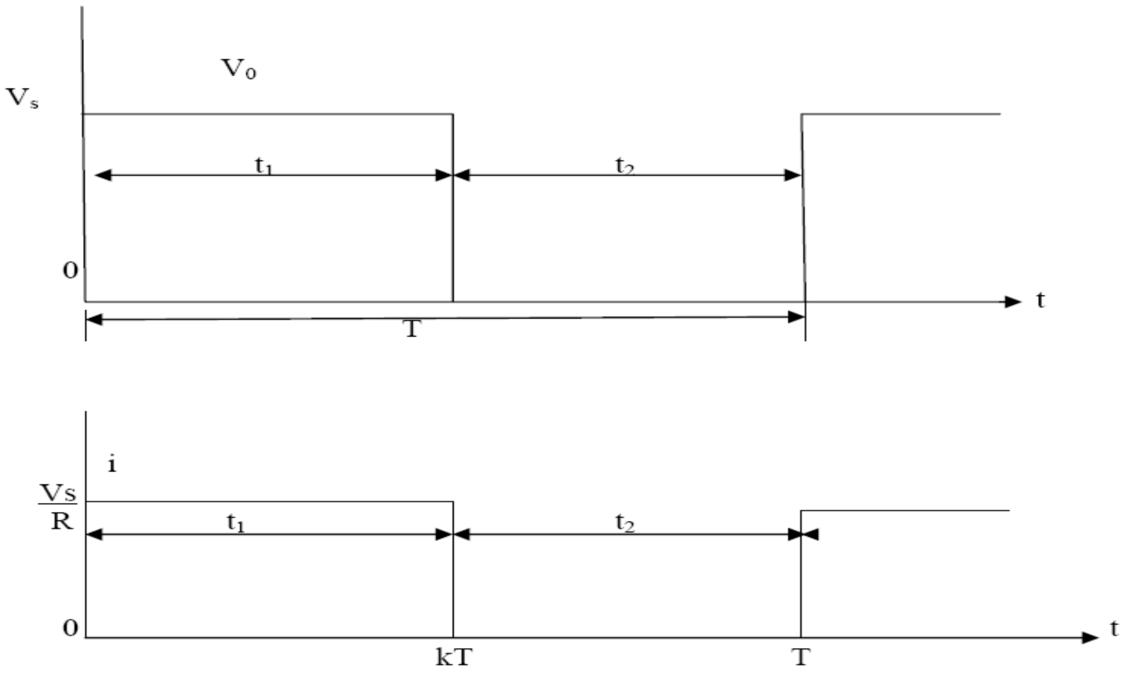
$R_i = V_s/I_a = V_s/(kV_s/R) = R/k$

Which indicates that the converter makes the input resistance R_i as a variable resistance of R/k.

The duty cycle k can be varied from 0 to1 by varying t_1, T or f. Therefore the output voltage V_o can be varied from 0 to V_s by controlling k, and the power flow can be controlled.



t=0
SW
+
+
V_s
V_0
R
-
-



V_0
V_s
t_1
t_2
0
t
T
i
$\frac{V_S}{R}$
t_1
t_2
0
kT
T
t

CHAPTER 4

CONTROL STRATEGIES

It is seen that the average value of Output voltage $\mathbf{V_0}$ can be controlled through Duty Cycle **D** by opening and closing the semiconductor switch periodically. The various control strategies for varying the duty cycle **D** are as follows:

1. Time ratio Control (TRC)
2. Current-limit Control

1. Time Ratio Control (TRC) :

As the name suggests, in this control scheme the duty cycle is varied. This is realized in two different control strategies :

(a) Constant Frequency System

(b) Variable Frequency System

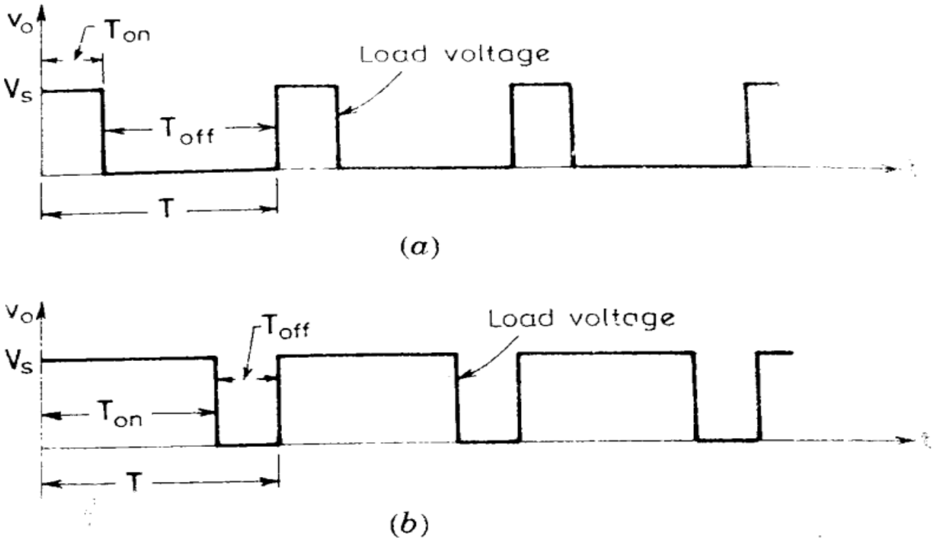
(a) Constant Frequency System :

In this scheme the ON time $\mathbf{T_{on}}$ is varied but the chopping frequency **f** (or the chopping period **T**) is kept constant. Variation of $\mathbf{T_{on}}$ means adjustment of pulse width, as such this scheme is also called **Pulse Width Modulation (PWM)** Scheme.

(b) Variable Frequency System:

In this scheme the chopping frequency **f** (or chopping period **T**) is varied and either ON time $\mathbf{T_{on}}$ or OFF time $\mathbf{T_{off}}$ is kept constant. This method of controlling **D** is also called **Frequency Modulation Scheme.**

It is seen that PWM scheme is better than the Variable frequency scheme. PWM technique however has a limitation as $\mathbf{T_{on}}$ cannot be reduced to near-zero for most of the commutation circuits used in choppers. As such low range of **D** control is not possible in PWM. However this can be achieved by increasing the chopping period (decreasing the chopper frequency) of the chopper.



Principle of pulse-width modulation (constant T).

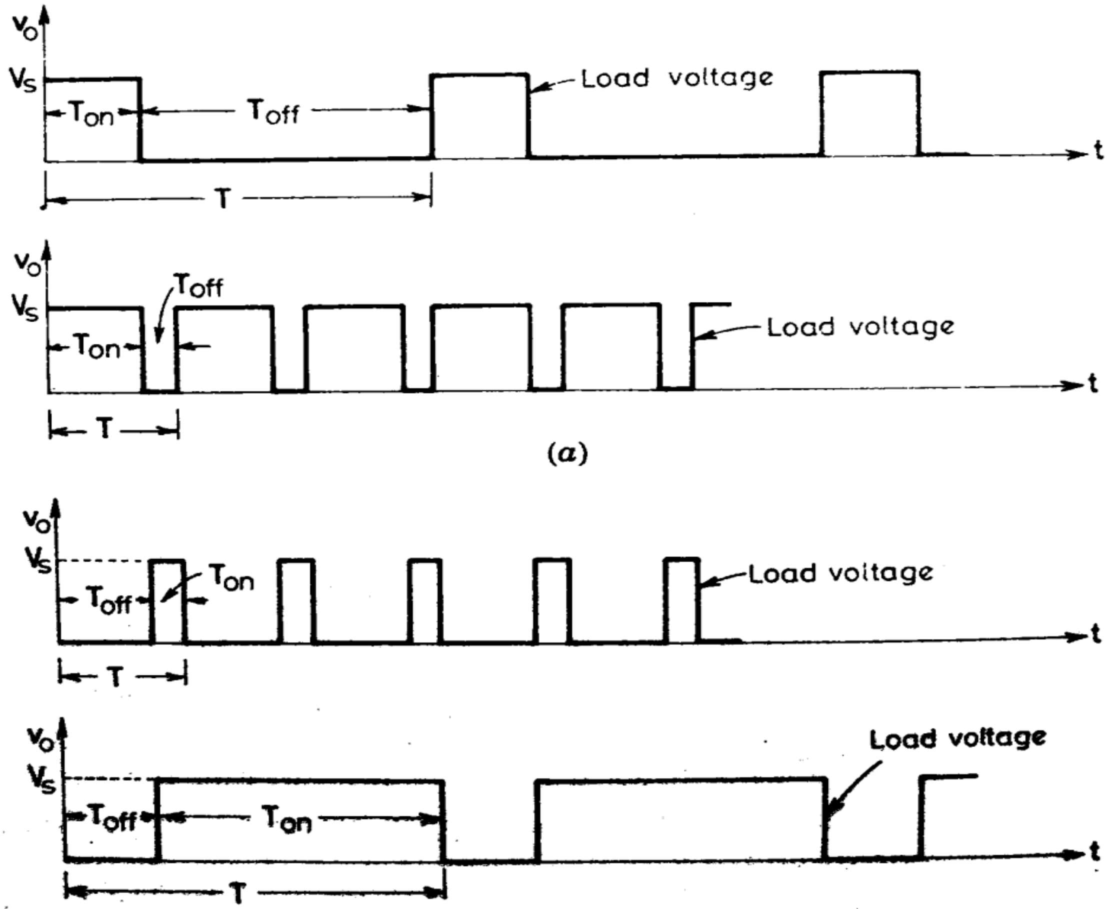


Fig 4.1 : (a) on-time Ton constant (b) off-time Toff constant

2. Current Limit Control :

In this strategy, the ON and OFF chopper circuit is guided by the previous set values of load current. These two set values are maximum load current $\mathbf{I_{0.max}}$ and minimum load current $\mathbf{I_{0.min}}$.

When load current reaches maximum limit the chopper is switched Off. Now load current free-wheels and begins to decay exponentially. When it falls to lower limit (minimum value), chopper is switched On and load current begins to rise as shown.

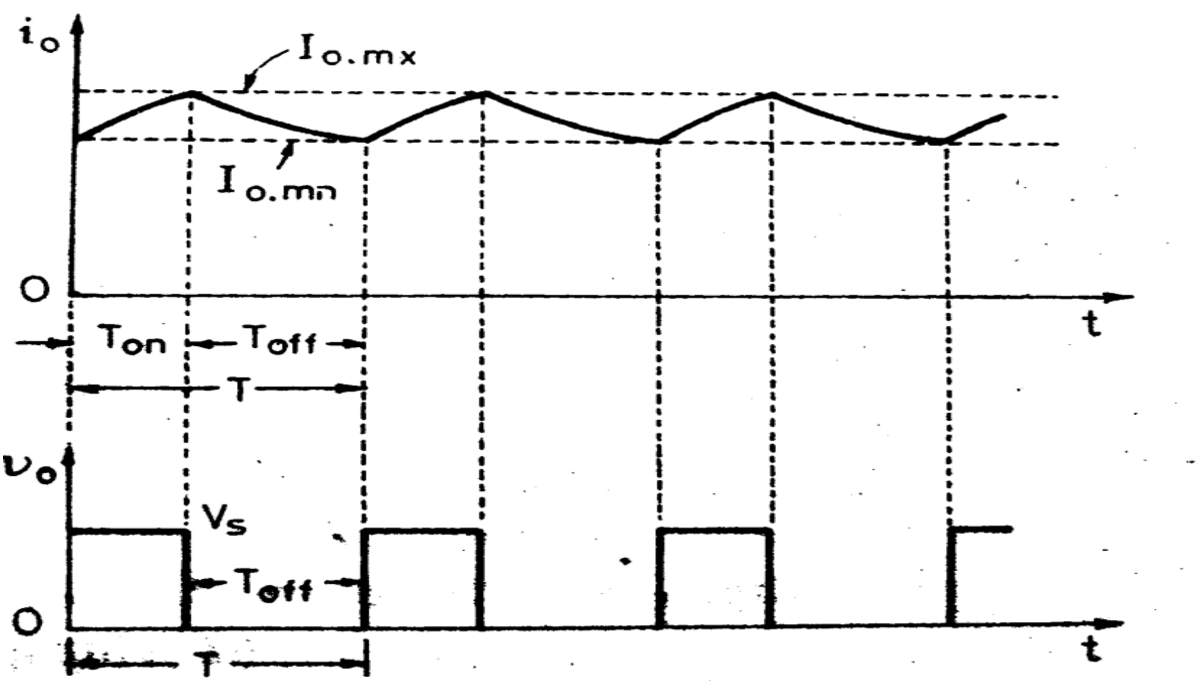


Fig 4.2 : Current Limit Control for Chopper

Switching frequency of chopper can be controlled by using $\mathbf{I_{0.max}}$ and $\mathbf{I_{0.min}}$. Ripple current (= $I_{0.max}$ - $I_{0.min}$) can be lowered and this in turn necessitates higher switching frequency and therefore more switching losses. Circuit limit control involves a feedback loop, the trigger circuitry for chopper is therefore more complex. PWM technique is, therefore, the commonly chosen control strategy of the power control chopper circuit.

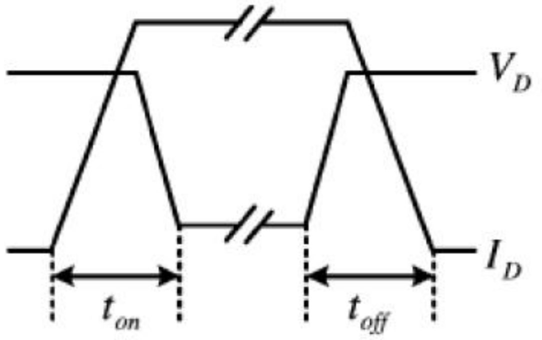
Chapter 5

Switching Losses in MOSFET

The swiching loss of power MOSFET is a major contributing factor towards the total power loss in a high frequency power converters. Calculation of switching losses occuring in a MOSFET is a relatively difficult task .It is because the complex switching behaviour of MOSFET's are difficult to model.The nonlinear characteristics arise due to the parasitic junction capacitance and inductance present in the MOSFET. It is comparatively easy to find out the switching losses by referring the parameters from the datasheet. A commonly used formula for estimating the MOSFET drain to source switching loss P_{SW} is given by:

$$P_{\mathrm{SW}} = \frac{1}{2} I_D V_D \left(t_{\mathrm{OFF}} + t_{\mathrm{ON}}\right) f + \frac{1}{2} C_{\mathrm{OSS}} V_D^2 f \quad \text{----------(1)}$$

where I_D , V_D , and f are the load current, input voltage, and switching frequency while t_{ON} and t_{OFF} are the MOSFET turn-on and turn-off times, respectively. Assuming a linear transition of i_{DS} and v_{DS} , the first term of (1) simply calculates the switching power loss as the area below i_{DS} and v_{DS} during the transition periods. The second term of (1) is often referred to as the output capacitance loss term. The motive of including the second loss term is to account for the loss of energy stored in the output capacitance that is internally dissipated through the MOS channel in the form of Joule heating during MOSFET turn-on. C_{OSS} is the output capacitance of the MOSFET and given by: $C_{OSS} = C_{GD} + C_{DS}$



Typical switching waveforms of a power MOSFET with an inductive load.

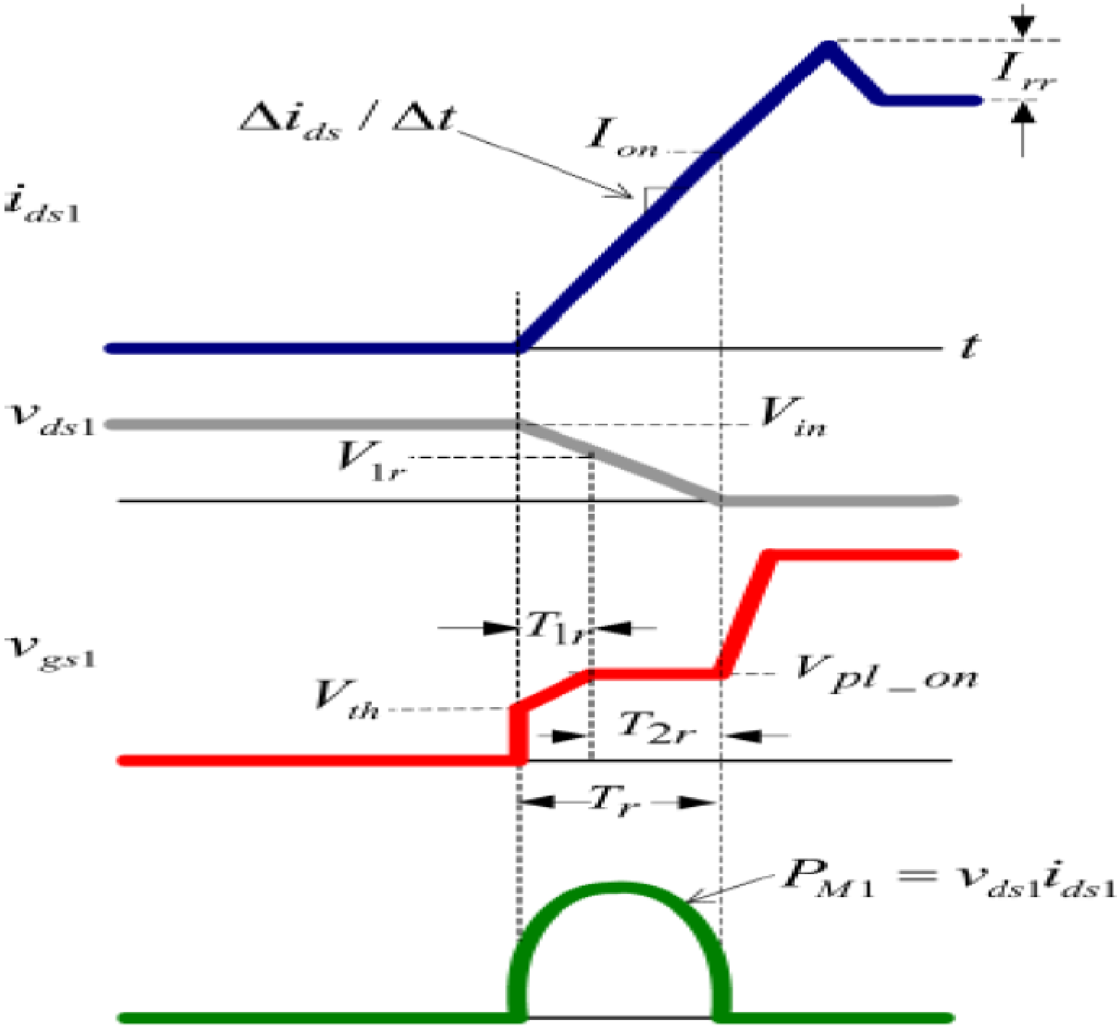
However the disadvantage of this method is that it predicts the magnitude of turn-on and turn-off loss as same. In a real converter operating at a high switching frequency, the model is highly inaccurate since turnoff loss is much larger due to parasitic inductances. In addition, the inductor ripple current decreases the current at turn-on and increases the current at turn-off, which further reduces turn-on switching loss and increases turn-off switching loss.

Hence, a separate approach is presented here to calculate the switching losses in a MOSFET. The proposed model uses the piecewise linear approximations of the switching waveforms. Turn-on switching loss occurs during T_r*(rise time)* and turn-off switching loss occurs during T_f*(fall time)* . The key to the model is prediction of the turn-on current I_{ON}, the rise and fall times T_r and T_f , the reverse recovery current, I_{rr}, the magnitude of the rising current slope $\Delta i_{ds}/\Delta t$, and the current drop Δi_{1f} when v_{ds1} rises to V_{in} at turn-off. The goal of the proposed model is to calculate the switching loss with respect to load current, driver supply voltage, driver gate current, and total circuit inductance in a simple manner. The MOSFET parasitic capacitances are required in the model. They are estimated using the effective values, using datasheet specification values for V_{ds1} spec, $Crss1$ spec, and $Ciss1$ spec.

Turn-On Switching Loss Model

By definition, P_{ON} is derived using the simple integral, representing the average power over one switching period:

$$P_{\mathrm{ON}} = f_s \int_0^{T_r} v_{ds1} i_{ds}\, dt = \frac{1}{6} V_{\mathrm{in}} I_{\mathrm{ON}} T_r f.$$



(piecewise linearization approach during Ton)

Turn-Off Switching Loss Model

The circuit waveforms and knowledge of the circuit operation are used extensively in order to derive the turn-off loss P_{OFF}. The turn-off transition consists of two intervals T_{1f} and T_{2f} . During T_{1f} , the Miller capacitor C_{gd1} is discharged while v_{gs1} remains at $V_{pl\ OFF}$, and i_{ds1} is assumed to remain constant.

During this interval, v_{ds1} increases from zero to V_{in} . Therefore, from the geometry, the turn-off power loss P_{1OFF}, during T_{1f}, is given by:

$$P_{1\text{OFF}} = \frac{1}{2} V_{\text{in}} I_{\text{OFF}} T_{1f} f_s .$$

And during T_{2f} power loss is given by:

$$P_{2\mathrm{OFF}} = \left(\frac{1}{6}(V_p - V_{\mathrm{in}}+)I_{\mathrm{OFF}} + \frac{1}{2}V_{\mathrm{in}}I_{\mathrm{OFF}}\right)T_{2f}f_s.$$

'he total turn-off loss P_{OFF} is the sum of $P_{1\mathrm{OFF}}$ and $P_{2\mathrm{OFF}}$

$$P_{\mathrm{OFF}} = P_{1\mathrm{OFF}} + P_{2\mathrm{OFF}}.$$

Where:

$$V_{\mathrm{pl\_ON}} = V_{\mathrm{th}} + \frac{I_o - 0.5\,\Delta i_{Lf}}{g_{fs}}.$$

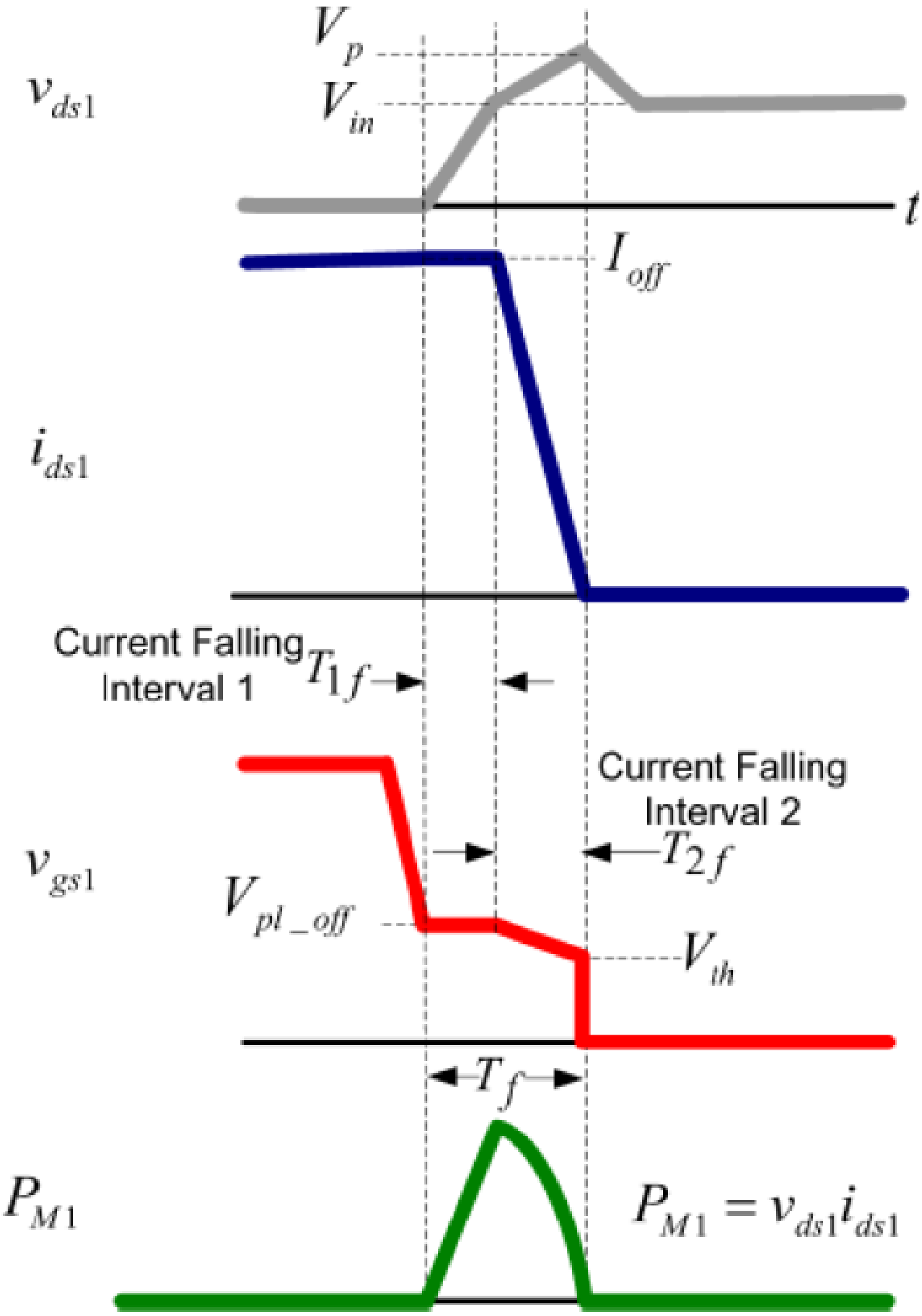


Fig 5.2 : piecewise linearization approach during T_{off}

CHAPTER 6

ZERO VOLTAGE SWITCHING RESONANT CONVERTERS

ZERO-VOLTAGE-SWITCHING RESONANT CONVERTERS

The switches of ZVS resonant converters turn on and off at zero voltage.

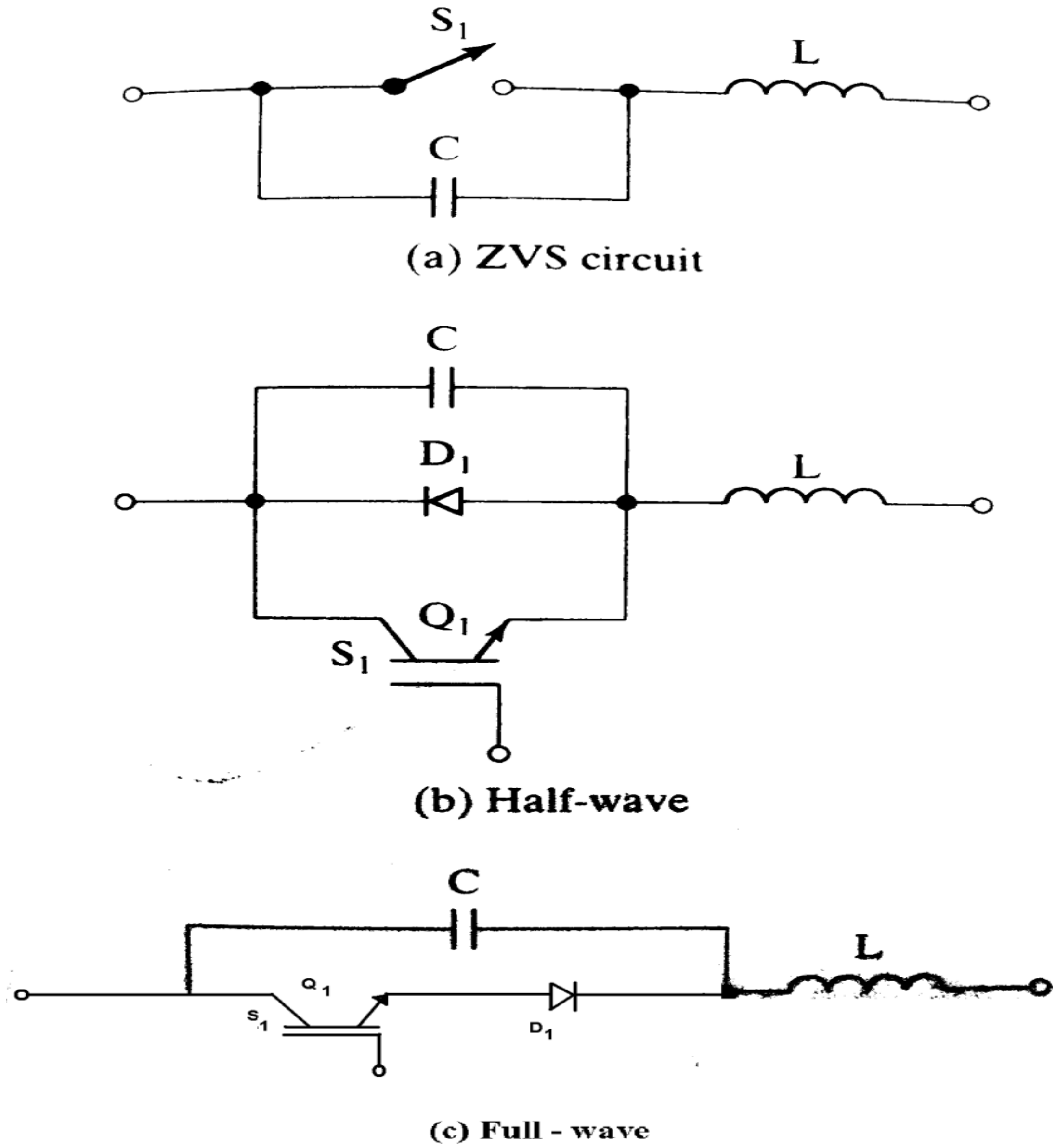
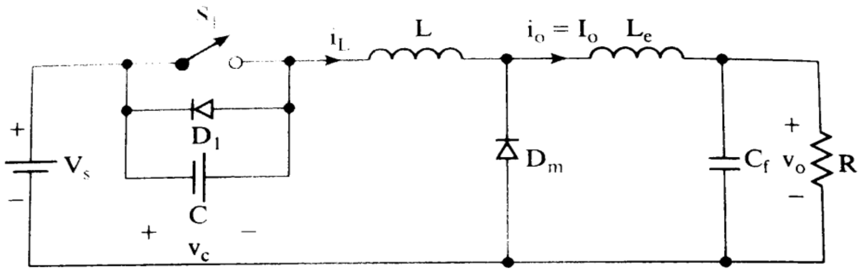
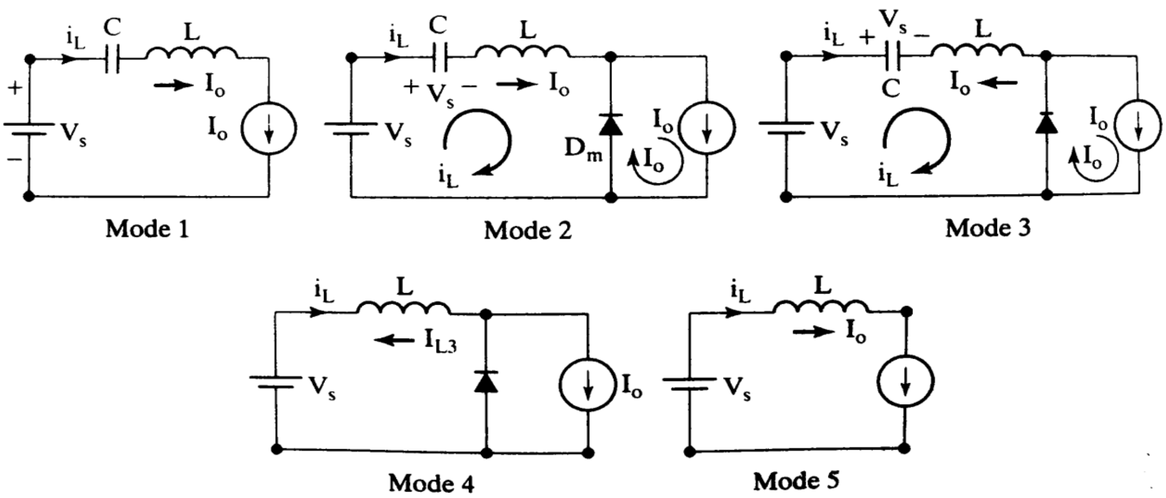


Fig 6.1 : Switch Configurations for ZVS Resonant Converters

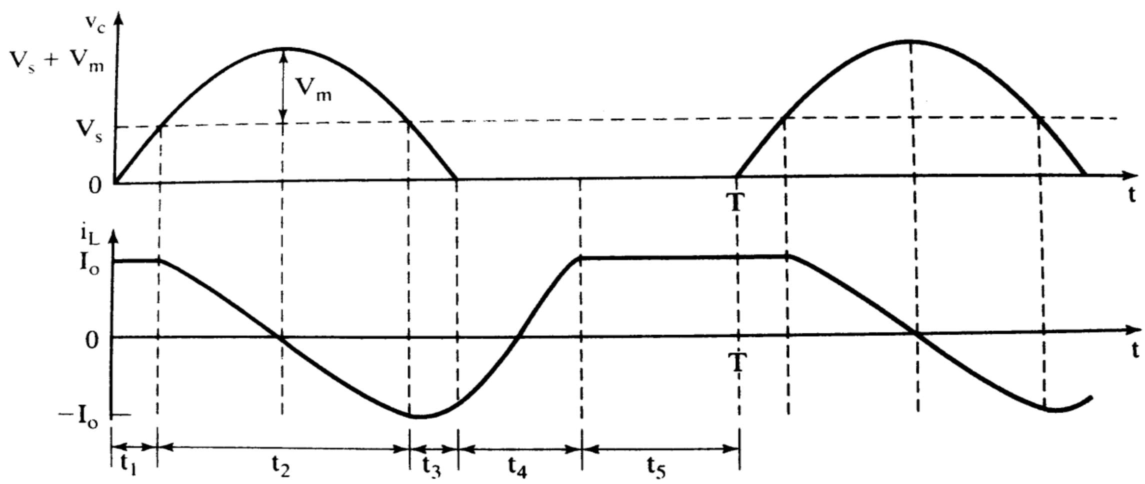
The capacitor **C** is connected in parallel with the switch **S_1** to achieve ZVS. The internal switch capacitance **C_j** is added with the capacitor **C** and it affects the resonant frequency only, thereby contributing no power dissipation in the switch. If the switch is implemented with transistor **Q_1** and an anti-parallel diode **D_1** as shown, the voltage across **C** is clamped by **D₁** and the switch is operated in half wave configuration. If the diode **D_1** is connected in series with **Q_1** as shown, the voltage across **C** can oscillate freely and the switch is operated in full wave configuration. A ZVS resonant converter is shown. A ZVS resonant converter is the dual of ZCS resonant converter.



(a) ZVS circuit



(b) Equivalent circuits



(c) Waveforms

Fig 6.2 : ZVS Resonant Converter

The circuit operation can be divided into 5 modes whose circuits are shown. We shall redefine the time origin t=0, at the beginning of each mode.

Mode 1 :

This mode is valid for $0 \leq t \leq t_1$. Both switch $\mathbf{S_1}$ and diode $\mathbf{D_m}$ are off. Capacitor **C** charges at a constant rate of load current $\mathbf{I_0}$. The capacitor voltage $\mathbf{V_c}$ which rises is given by

$$V_c = I_o.t / C$$

This mode ends at time $t = t_1$ when V_c ($t = t_1$) = V_s. That is $t_1 = V_s.C / I_0$.

Mode 2 :

This mode is valid for $0 \leq t \leq t_2$. The switch $\mathbf{S_1}$ is still off, but diode $\mathbf{D_m}$ turns on. The capacitor voltate $\mathbf{V_c}$ is given by

$$V_c = V_m \sin \omega_0 t + V_s \quad \text{where } V_m = I_0\sqrt{(L/C)}.$$

The peak switch voltage which occurs at $t = (\pi/2)\sqrt{LC}$, is

$$V_t(pk) = V_c(pk) = I_0 \sqrt{(L/C)} + V_s$$

The inductor current i_L is given by

$$i_L = I_0 \cos \omega_0 t$$

This mode ends at $t = t_2$ when V_c ($t = t_2$) = V_s, and $i_L(t = t_2) = -I_0$. Therefore, $t_2 = \pi\sqrt{LC}$.

Mode 3 :

This mode is valid for $0 \leq t \leq t_3$. The capacitor voltage that falls from V_s to zero is given by

$$V_c = V_s - V_m \sin \omega_0 t$$

The inductor current i_L is given by

$$i_L = -I_0 \cos \omega_0 t$$

This mode ends at $t = t_3$ when V_c ($t = t_3$) = 0, and i_L ($t = t_3$) = i_{L3}. Thus,

$$T_3 = \sqrt{(LC)} \sin^{-1} x$$

Where, $x = V_s/V_m = (V_s/I_0)\sqrt{(C/L)}$.

Mode 4 :

This mode is valid for $0 \leq t \leq t_4$. Switch $\mathbf{S_1}$ is turned on and diode $\mathbf{D_m}$ remains on. The inductor current which rises linearly from I_{l3} to I_0 is given by

$$i_L = I_{L3} + (V_s/L)t$$

This mode ends at time $t = t_4$ when i_L $(t = t_4) = 0$. Thus $t_4 = (I_0 – I_{L3})(L/V_s)$. I_{L3} has a negative value.

Mode 5 :

This mode is valid for $0 \leq t \leq t_5$. Switch $\mathbf{S_1}$ is on but $\mathbf{D_m}$ is off. The load current $\mathbf{I_0}$ flows through the switch. This mode ends at time $t = t_5$, when the switch S_1 is turned off again and the cycle is repeated. That is $t_5 = T – (t_1 + t_2 + t_3 + t_4)$.

The waveforms for i_L and V_c are shown. The equation

$$V_t(pk) = V_c(pk) = I_0 \sqrt{(L/C)} + V_s$$

shows that the peak switch voltage $V_t(pk)$ is dependent on the load current I_0. Therefore a wide variation in the load current results in a wide variation of the switch voltage. For this reason, ZVS converters are used only for constant-load applications. The switch must be turned on only at zero voltage. Otherwise, the energy stored in **C** can be dissipated in the switch. To avoid this situation, the antiparallel diode $\mathbf{D_1}$ must conduct before turning on the switch.

CHAPTER 7

ZERO CURRENT SWITCHING RESONANT CONVERTERS

Why to go for ZCS

The control switches in all the PWM dc-dc converter topologies, operate in a switch mode, in which they turn whole load current on and off during each switching. This switch-mode operation subjects the control switches to high switching stress and high switching power losses. To maximize the performance of switch-mode power electronic conversion systems, the switching frequency of the power semiconductor devices needs to be increased, but this results in increased switching losses and electromagnetic interference (EMI). To eradicate these problems, Zero voltage switching (ZVS) technique and zero current switching (ZCS) technique are two conventionally employed soft switching methods. These techniques lead to either zero voltage or zero current during switching transition, significantly decreasing the switching losses and increasing the reliability for the converters. The ZVS technique eliminates capacitive turn-on losses, and decreases the turn-off switching losses by slowing down the voltage rise, thereby lowering the overlap between the switch voltage and the switch current. However, a large external resonant capacitor is needed to lower the turn-off switching loss effectively for ZVS. Conversely, ZCS eliminates the voltage and current overlap by forcing the switch current to zero before the switch voltage rises, making it more effective than ZVS in reducing switching losses, especially for slow switching power devices. For high efficiency power conversion, the ZCS topologies are most frequently adopted.

Variable frequency control can achieve output regulation of the resonant converters in both traditional ZCS and ZVS approaches. Traditional ZCS converters operate with constant on-time control, while the traditional ZVS converters operate with constant off-time control. Both approaches need to operate with a wide switching frequency range when given a wide input source and load range, making the filter circuit design difficult to optimize. Many high efficiency converter topologies with ZCS have been explored and proposed. The primary design feature of novel ZCS/ZVS PWM power converters is the incorporation of an auxiliary switch in the traditional quasi-resonant circuit. The resonance of the novel converters is dominated by the auxiliary switch, which generates resonance and temporarily stops a period that can be regulated, thus circumventing the disadvantages of fixed conduction or cut-off time in a traditional quasi-resonant power converter.

ZERO CURRENT SWITCHING RESONANT CONVERTERS

The switches of Zero Current Switching (ZCS) resonant converters turn ON and OFF at zero current. The resonant circuit that consists of switch **S_1**, inductor **L**, and capacitor **C** is shown. The inductor **L** is connected in series with power switch **S_1** to achieve ZCS. It is classified into two types – **L** type and **M** type. In both the types the inductor **L** limits the di/dt of the switch current and **L** and **C** constitute a series resonant circuit. When the switch current is zero there is a current $i=C_j.dv_t/dt$ flowing through the internal capacitance **C_j** due to finite slope of switch voltage at turn off. This current flow causes power dissipation in the switch and limits the high switching frequency.

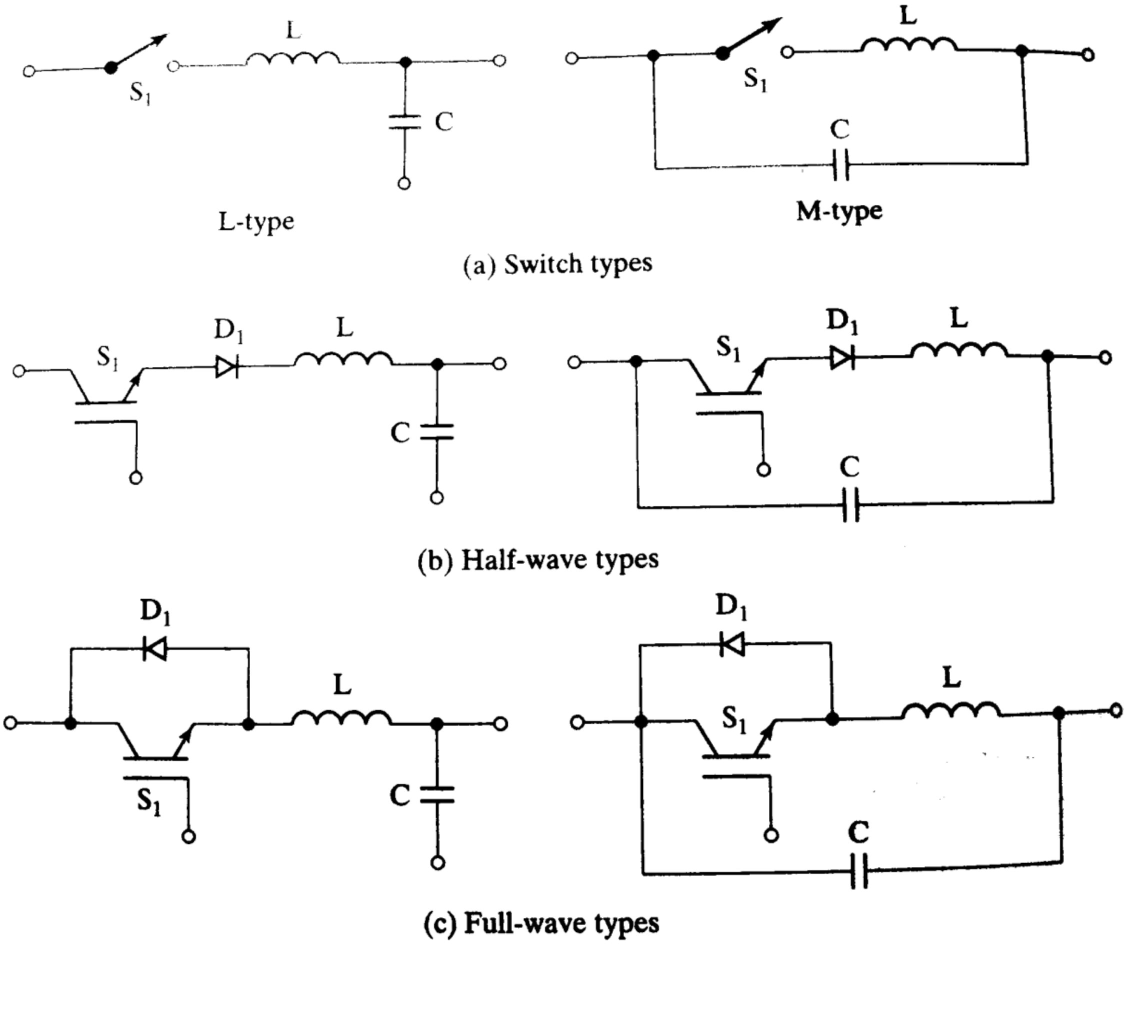


Fig 7.1 : Switch configurations for ZCS Resonant Converters

The switch can be implemented either in half wave configuration where diode **D1** allows unidirectional current flow or in full-wave configuration where the switch current can flow bidirectionally. The practical devices do not turn off at zero current due to their recovery time. As a result, an amount of energy can be trapped in inductor **L** of the **M**-type configuration and voltage transients appear across the switch. This normally favours **L**-type configuration over **M**-type one.

The name “Buck Converter” presumably evolves from the fact that the input voltage is bucked/chopped or attenuated in amplitude and a lower amplitude voltage appears at the output. A buck converter or step-down voltage regulator provides non-isolated, switch-mode dc-dc conversion with the advantages of simplicity and low cost. Figure below shows a simplified non-isolated buck converter that accepts a dc input and uses pulse-width modulation (PWM) of switching frequency to control the output of an internal power MOSFET. An external diode together with external inductor and output capacitor, produces the regulated dc output. Buck or step down converters produce an average output voltage lower than the input source voltage.

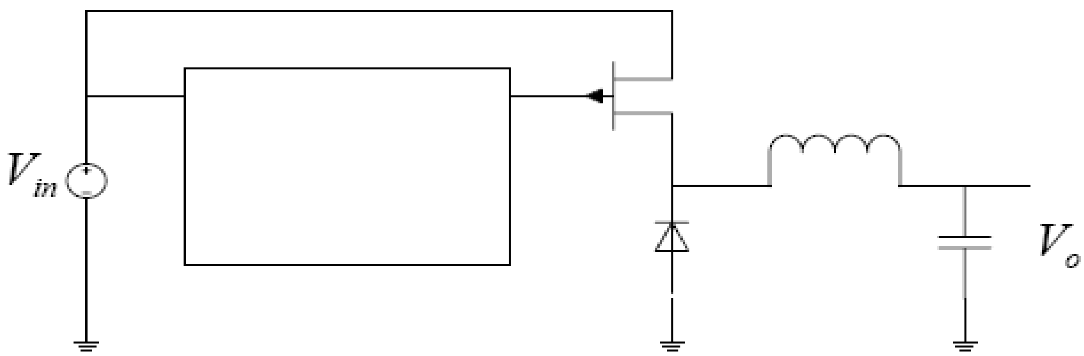


Fig 7.2 : Complete Switching Regulator Topology

Evolution of Buck Converter

The buck converter here onwards is introduced using the evolutionary approach. Let us consider the circuit in Figure 7.3 containing a single pole double-throw switch.

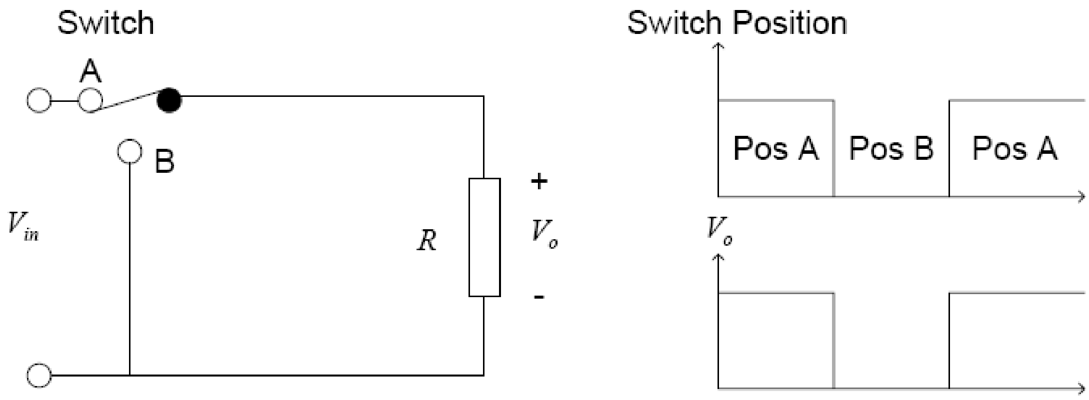


Fig 7.3 : A resistor with a single-pole double-throw switch

For the above circuit, the output voltage equals the input voltage when the switch is in position A and it is zero when the switch is in position B. By varying the duration for which the switch is in position A and B, it can be seen that the average output voltage can be varied, but the output voltage is not pure dc. The circuit in Figure 7.3 can be modified as shown in Figure 7.4 by adding an inductor in series with the load resistor. An inductor reduces ripples in current passing through it and the output voltage would contain less ripple since the current through the load resistor is the same as that of the inductor. When the switch is in position A, the current through the inductor increases and the energy stored in the inductor increases. When the switch is in position B, the inductor acts as a source and maintains the current through the load resistor. During this period the energy stored in the inductor decreases and its current falls. It is important to note that there is continuous conduction through the load for this circuit. If the time constant due to the inductor and load resistor is relatively large compared with the period for which the switch is in position A or B, then the rise and fall of current through inductor is more or less linear as shown in Figure 7.4.

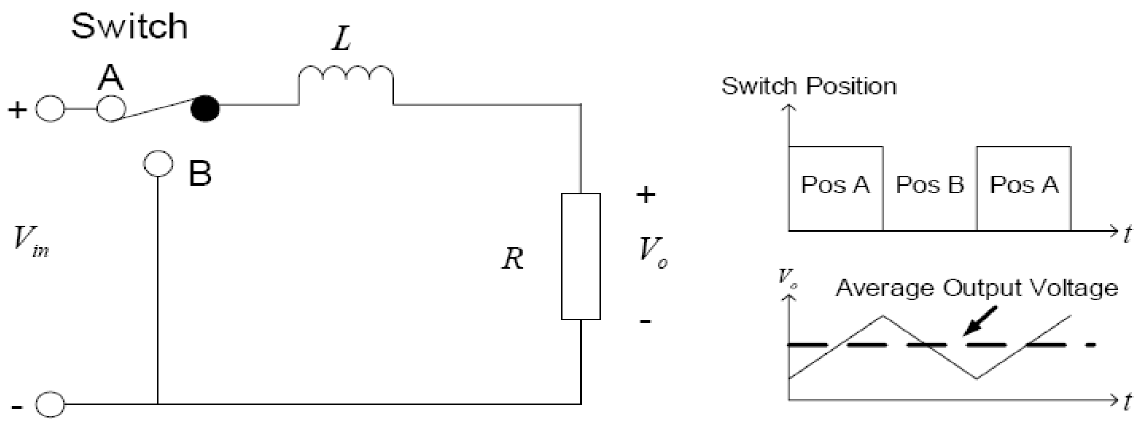


Fig 7.4 : Effect of an Inductor

The next step in the evolutionary development of the buck converter is to add a capacitor across the load resistor and this circuit is shown in Figure 7.5. A capacitor reduces the ripple content in voltage across it, whereas an inductor smoothes the current passing through it.
The combined action of LC filter reduces the ripple in output to a very low level.

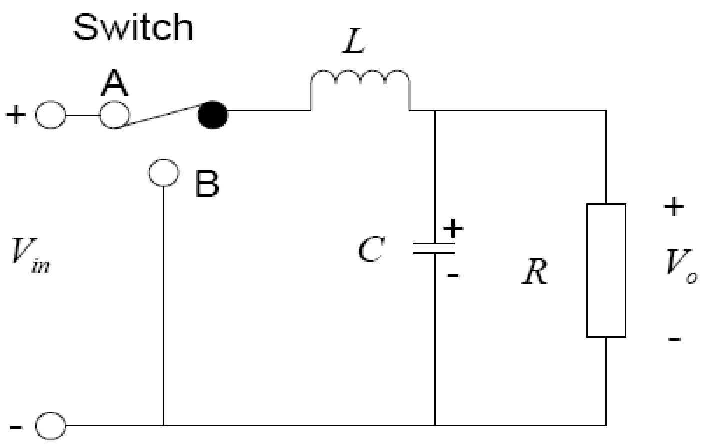


Fig 7.5 : Circuit with an LC Filter

With the circuit in Figure 7.5 it is possible to have a power semiconductor switch to correspond to the switch in position A and a diode in position B.The circuit that results is shown in Figure 7.6. When the switch is in position B, the current will pass through the diode. The important thing now is the controlling of the power semiconductor switch.

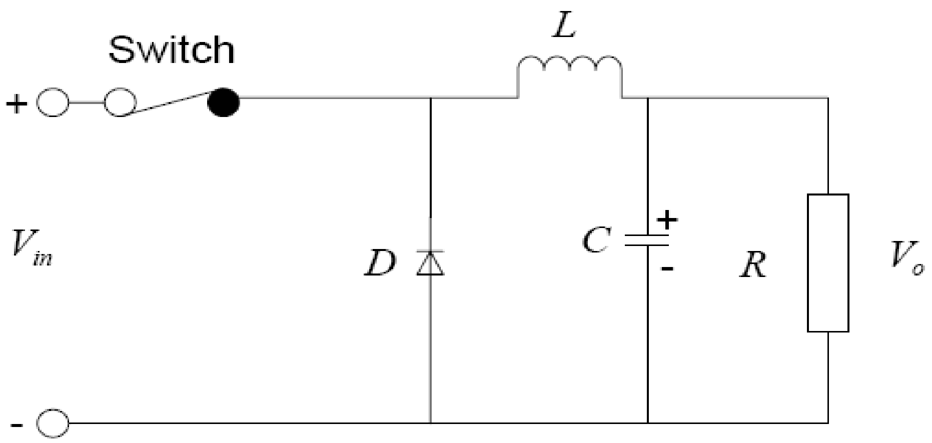


Fig 7.6 : Buck Converter with Load Resistor

The circuit in Figure 7.6 can be regarded as the most elementary buck converter without a feedback. The Buck Converter transfers small packets of energy with the help of a power switch, a diode and an inductor and is accompanied by an output filter capacitor and input filter. All the other topologies such as the Boost, Buck-Boost Converter etc vary by the different arrangement of these basic components.This circuit can be further modified by adding the feedback part which is integral for a SMPS because based on the feedback it stabilizes the output. Such a circuit is shown in the Figure 7.7.

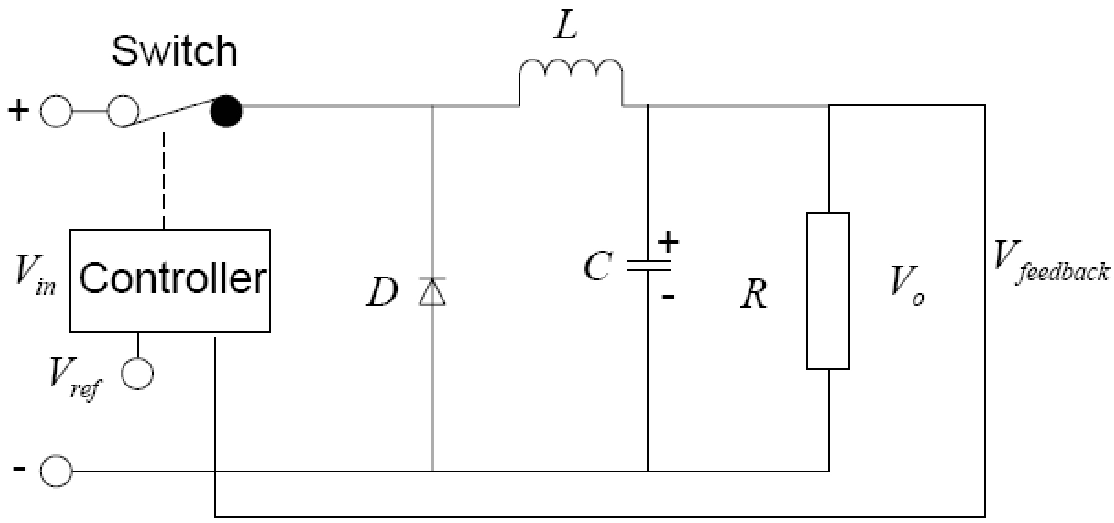


Fig 7.7 : Step down Switch mode Power supply

Purpose of different components in the Buck Converter

As just seen in the previous section that any basic switched power supply consists of five standard components:

- Pulse-width modulating controller
- Transistor switch (active switch)
- Inductor
- Capacitor
- Diode (passive switch)

Switch

In its crudest form a switch can be a toggle switch which switches between supply voltage and ground. But for all practical applications which we shall consider we will deal with transistors. Transistors chosen for use in switching power supplies must have fast switching times and should be able to withstand the voltage spikes produced by the inductor. The input on the gate of the transistor is normally a Pulse Width Modulated (PWM) signal which will determine the ON and OFF time. Sizing of the power switch is determined by the load current and off-state voltage capability. The power switch (transistor) can either be a MOSFET, IGBT, JFET or a BJT. Power MOSFETs are the key elements of high frequency power systems such as high-density power supplies**.** Therefore MOSFETs have now replaced BTJ's in new designs operating at much higher frequencies but at lower voltages. At high voltages MOSFETs still have their limitations. The intrinsic characteristics of the MOSFET produce a large on-resistance which increases excessively when the devices breakdown voltage is raised. Therefore the power MOSFET is only useful upto voltage ratings of 500V and so is restricted to low voltage applications or in two-transistor forward converters and bridge circuits operating off-line. At high breakdown voltages (>200V) the on-state voltage drop of the power MOSFET becomes higher than that of a similar size bipolar device with similar voltage rating. This makes it more attractive to use the bipolar power transistor at the expense of worse high frequency performance. As improvements in fabrication techniques, new materials, device characteristics take place than MOSFETs are likely to replace BJTs.

Operating Frequency

The operating frequency determines the performance of the switch. Switching frequency selection is typically determined by efficiency requirements. There is now a growing trend in research work and new power supply designs in increasing the switching frequencies.The higher is the switching frequency, the smaller the physical size and component value. The reason for this is to reduce even further the overall size of the power supply in line with miniaturisation trends in electronic and computer systems. However there is an upper frequency limit where either magnetic losses in the inductor or switching losses in the regulator circuit and power MOSFET reduce efficiency to an impractical level. Higher frequency also reduces the size of the output capacitor. E.g. the capacitance required is
67µF at 500 KHz, but only 33µF at 1MHz.The ripple current specification remains unchanged.

Inductor

The function of the inductor is to limit the current slew rate (limit the current in rush) through the power switch when the circuit is ON. The current through the inductor cannot change suddenly. When the current through an inductor tends to fall, the inductor tends to maintain the current by acting as a source. This limits the otherwise high-peak current that would be limited by the switch resistance alone. The key advantage is when the inductor is used to drop voltage, it stores energy. Also the inductor controls the percent of the ripple and determines whether or not the circuit is operating in the continuous mode. Peak current through the inductor determines the inductor's required saturation current rating, which in turn dictates the approximate size of the inductor. Saturating the inductor core decreases the converter efficiency, while increasing the temperature of the inductor, the MOSFET and the diode. The size of the inductor and capacitor can be reduced by the implementation of high switching frequency, multiphase interleaved topology, and a fast hysteric controller. A smaller inductor value enables a faster transient response, it also results in larger current ripple which causes higher conduction losses in the switches, inductor and parasitic resistances. The smaller inductor also requires a larger filter capacitor to decrease the output voltage ripple. Inductors used in switched supplies are sometimes wound on toroidal cores, often made of ferrite or powdered iron core with distributed air-gap to store energy.

A DC-DC converter transfers energy at a controlled rate from an input source to an output load, and as the switching frequency increases, the time available for this energy transfer decreases. For example, consider a buck converter operating at 500 kHz with a 10 μH inductor. For most DC-DC converters, changing the frequency to 1 MHz allows use of exactly one half the inductance, or 5μH.

Capacitor

Capacitor provides the filtering action by providing a path for the harmonic currents away from the load. Output capacitance (across the load) is required to minimize the voltage overshoot and ripple present at the output of a step-down converter. The capacitor is large enough so that its voltage does not have any noticeable change during the time the switch is off. Large overshoots are caused by insufficient output capacitance, and large voltage ripple is caused by insufficient capacitance as well as a high equivalent-series resistance (ESR) in the output capacitor. The maximum allowed output-voltage overshoot and ripple are usually specified at the time of design. Thus, to meet the ripple specification for a step-down converter circuit, we must include an output capacitor with ample capacitance and low ESR.

The problem of overshoot, in which the output-voltage overshoots its regulated value when a full load is suddenly removed from the output, requires that the output capacitor be large enough to prevent stored inductor energy from launching the output above the specified maximum output voltage. Since switched power regulators are usually used in high current, high performance power supplies, the capacitor should be chosen for minimum loss. Loss in a capacitor occurs because of its internal series resistance and inductance. Capacitors for switched regulators are partly chosen on the basis of Effective Series Resistance (ESR). Solid tantalum capacitors are the best in this respect. For very high performance power supplies, sometimes it is necessary to use parallel capacitors to get a low enough effective series resistance.

Freewheeling Diode

Since the current in the inductor cannot change suddenly, a path must exist for the inductor current when the switch is off (open). This path is provided by the freewheeling diode (or catch diode).The purpose of this diode is not to rectify, but to direct current flow in the circuit and to ensure that there is always a path for the current to flow into the inductor. It is also necessary that this diode should be able to turn off relatively fast. Thus the diode enables the converter to convert stored energy in the inductor to the load. This is a reason why we have higher efficiency in a DC-DC Converter as compared to a linear regulator. When the switch closes, the current rises linearly (exponentially if resistance is also present). When the switch opens, the freewheeling diode causes a linear decrease in current. At steady state we have a saw tooth response with an average value of the current.

Feedback

Feedback and control circuitry can be carefully nested around these circuits to regulate the energy transfer and maintain a constant output within normal operating conditions. Control by pulse-width modulation is necessary for regulating the output. The transistor switch is the heart of the switched supply and it controls the power supplied to the load.

L-TYPE ZCS RESONANT CONVERTER

The circuit operation can be divided into 5 modes whose equivalent circuits are shown. We shall redefine the time origin, t=0, at the beginning of each mode.

Mode 1 :

This mode is valid for $0 \leq t \leq t_1$. Switch S_1 is turned on and diode D_m conducts. The inductor current i_L which rises linearly is given by

$$I_L = (V_s/L)t$$

This mode ends at time $t=t_1$ when i_l $(t=t_1) = I_0$. That is $t_1 = I_0L/V_s$,

Mode 2 :

This mode is valid for $0 \leq t \leq t_2$. Switch S_1 remains on but diode D_m is off. The inductor current i_L is given by

$$i_L = I_m \sin \omega_o t + I_0$$

where $I_m = V_s\sqrt{(C/L)}$ and $\omega_0 = 1/\sqrt{LC}$. The capacitor voltage V_c is given by

$$V_c = V_s(1 - \cos\omega_0 t)$$

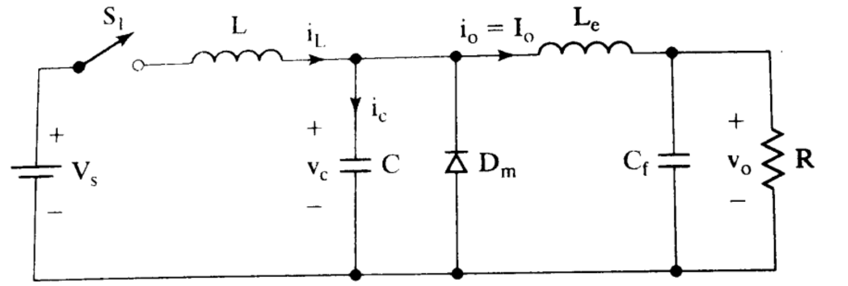
The peak current which occurs at $t = (\pi/2)\sqrt{LC}$ is

$$I_p = I_m + I_0$$

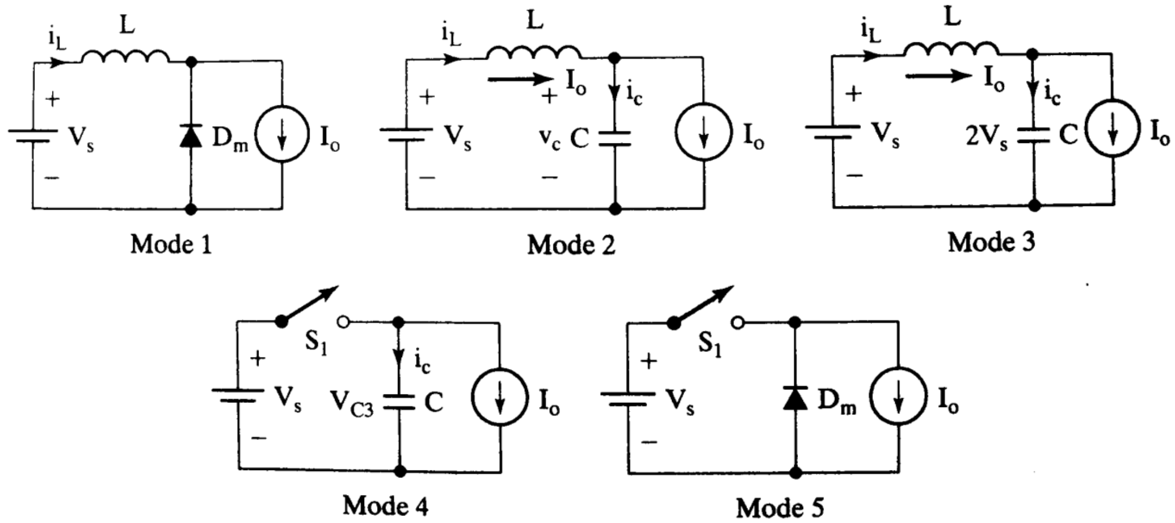
The peak capacitor voltage is given by

$$V_c(pk) = 2V_s$$

This mode ends at $t = t_2$ when $i_L(t = t_2) = I_0$ and $Vc(t = t_2) = V_{c2} = 2V_s$. Therefore $t_2 = \pi\sqrt{LC}$.



(a) Circuit



(b) Equivalent circuits

Mode 3 :

This mode is valid for $0 \leq t \leq t_3$. The inductor current that falls from I_0 to zero is given by

$$i_L = I_0 - I_m \sin\omega_0 t$$

The capacitor voltage is given by

$$V_c = 2V_s\cos\omega_0 t$$

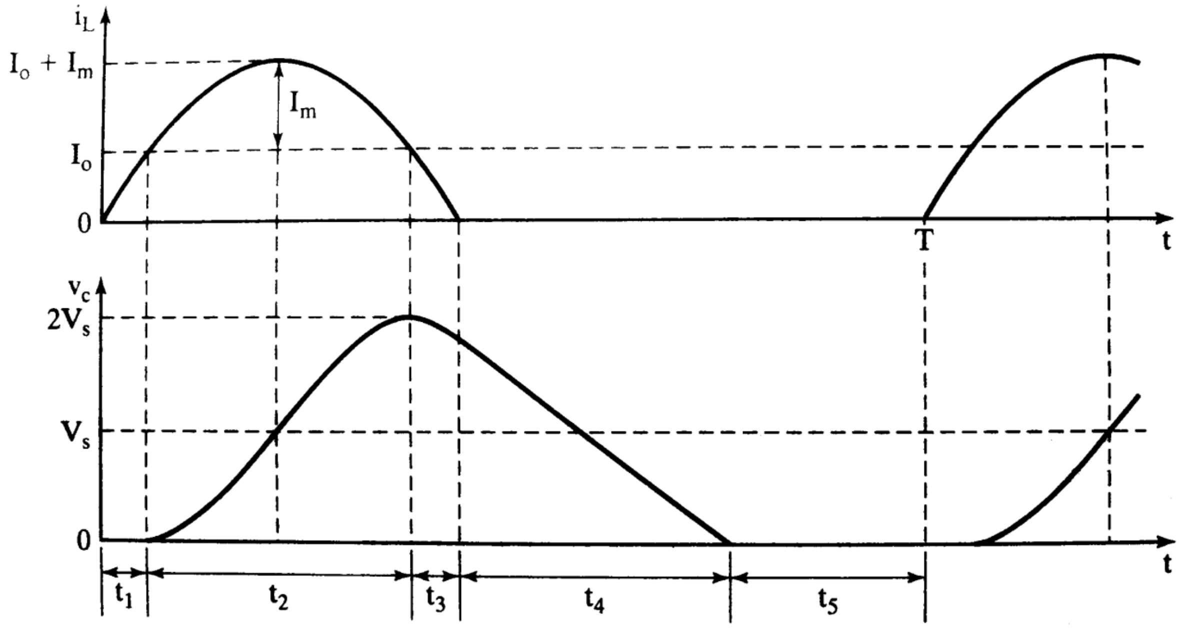
This mode ends at $t = t_3$ when i_L $(t = t_3) = 0$. And $V_c(t = t_3) = V_{c3}$. Thus $t_3 = \sqrt{LC}\sin^{-1}(1/x)$ where $x = I_m/I_0 = (V_s/I_0)\sqrt{C/L}$

Mode 4 :

This mode is valid for $0 \leq t \leq t_4$. The capacitor supplies the load current I_o and its voltage is given by

$$V_c = V_{c3} - (I_0/C)t$$

This mode ends at $t = t_4$ when V_c $(t = t_4) = 0$. Thus $t_4 = V_{c3}C/I_0$



(c) Waveforms

Fig 7.7 : L-Type ZCS Resonant Converter

Mode 5 :

This mode is valid for $0 \leq t \leq t_5$. When the capacitor voltage tends to be negative, the diode D_m conducts. The load current I_o flows through diode D_m. This mode ends at time $t=t_5$ when the switch S_1 is turned on again and the cycle is repeated i.e. $t_5=T-(t_1+t_2+t_3+t_4)$.

The waveform for I_l and V_c are shown. The peak switch voltage equals the dc supply voltage because the switch current is zero at turn on and turn off, the switching loss, which is the product of V and I, becomes very small. The peak resonant current I_m must be higher than the load current I_o and this sets a limit on the minimum value of load resistance **R**. However by placing an anti-parallel diode across the switch, the output voltage can be made insensitive to load variations.

CHAPTER 8

IMPLEMENTATION WORK & OBSERVATIONS

Circuit Diagram of ZCS Buck Converter

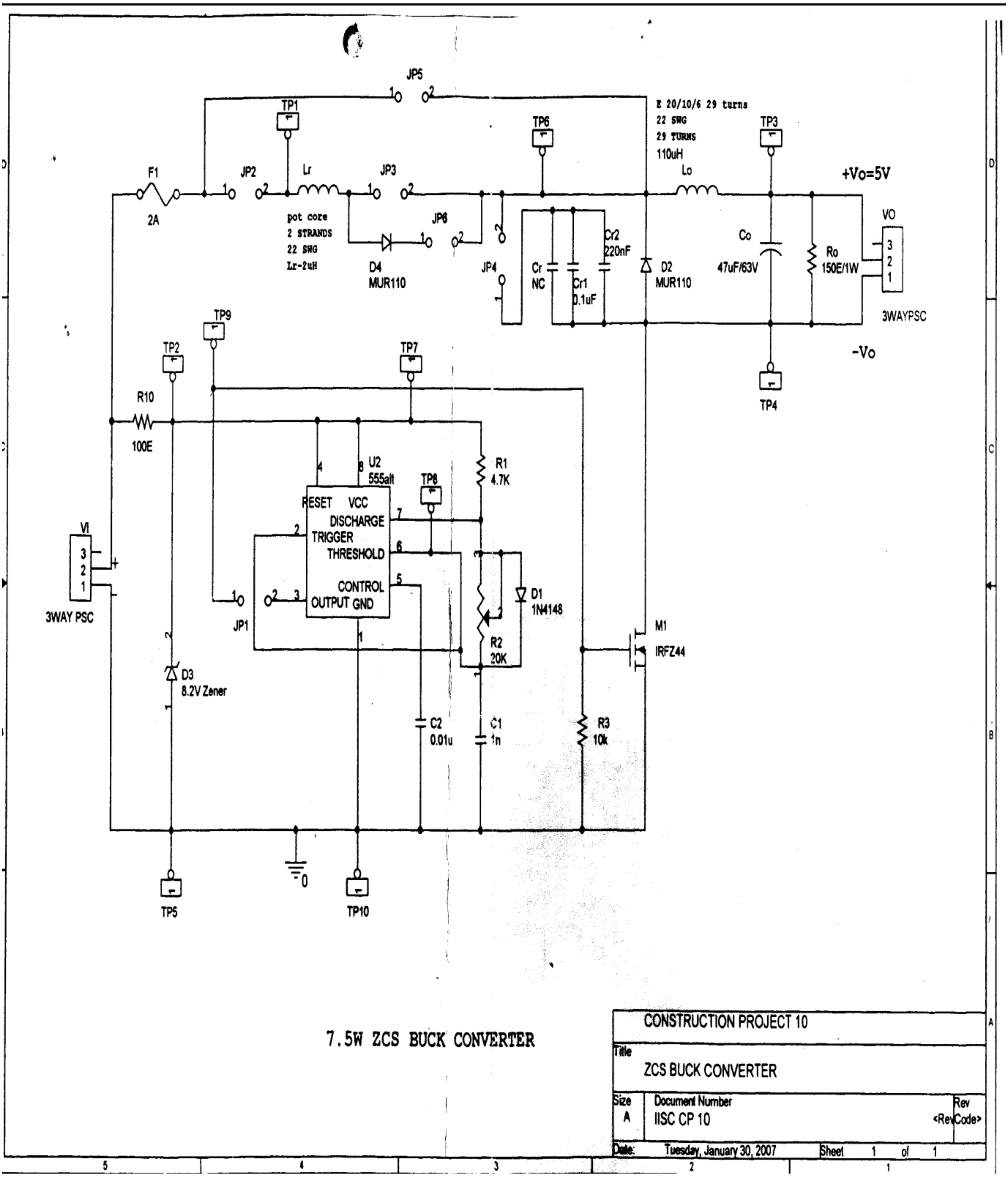


Fig 8.1 : 7.5W ZCS BUCK CONVERTER

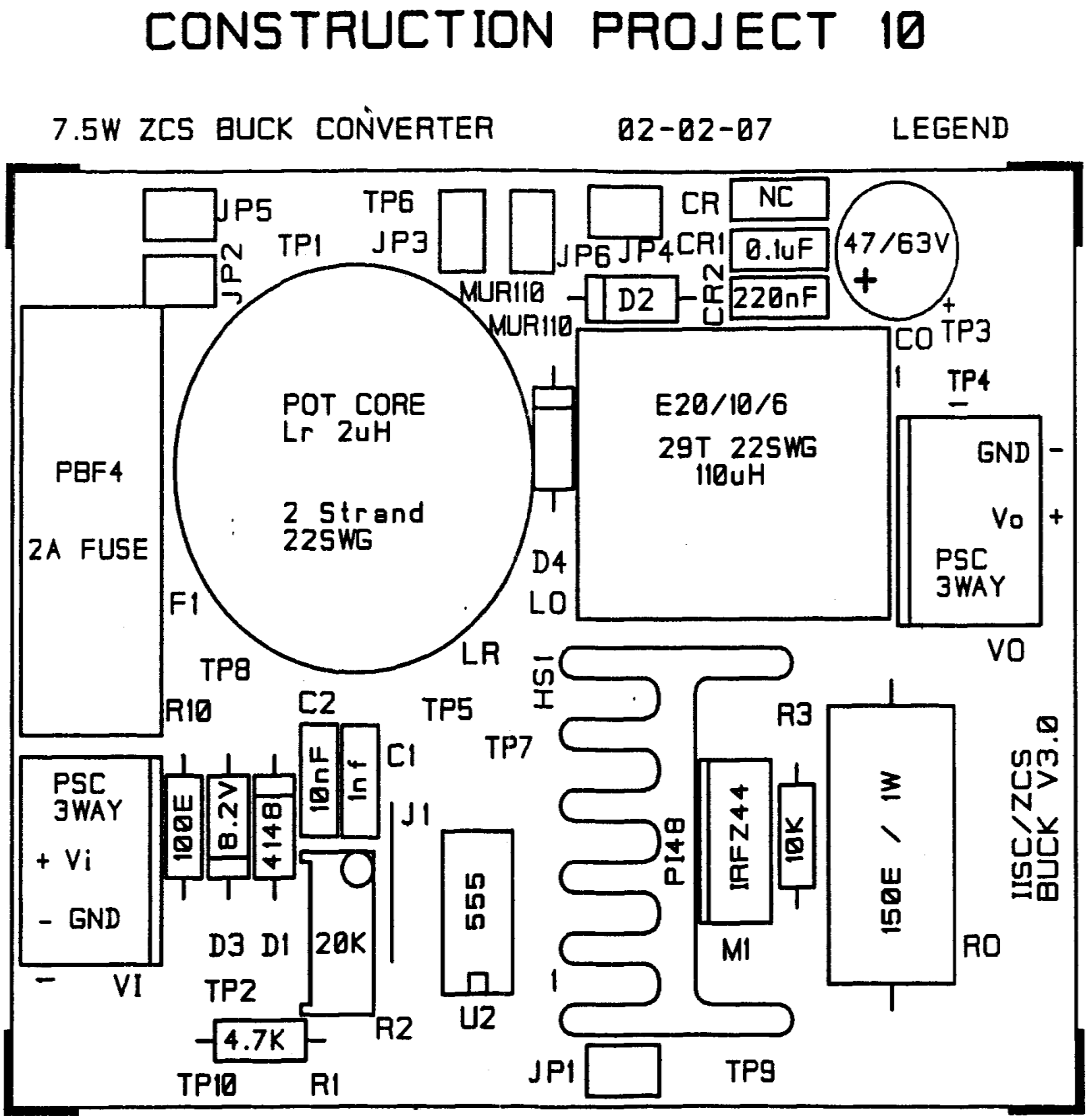


Fig 8.2 : The Construction Project PCB

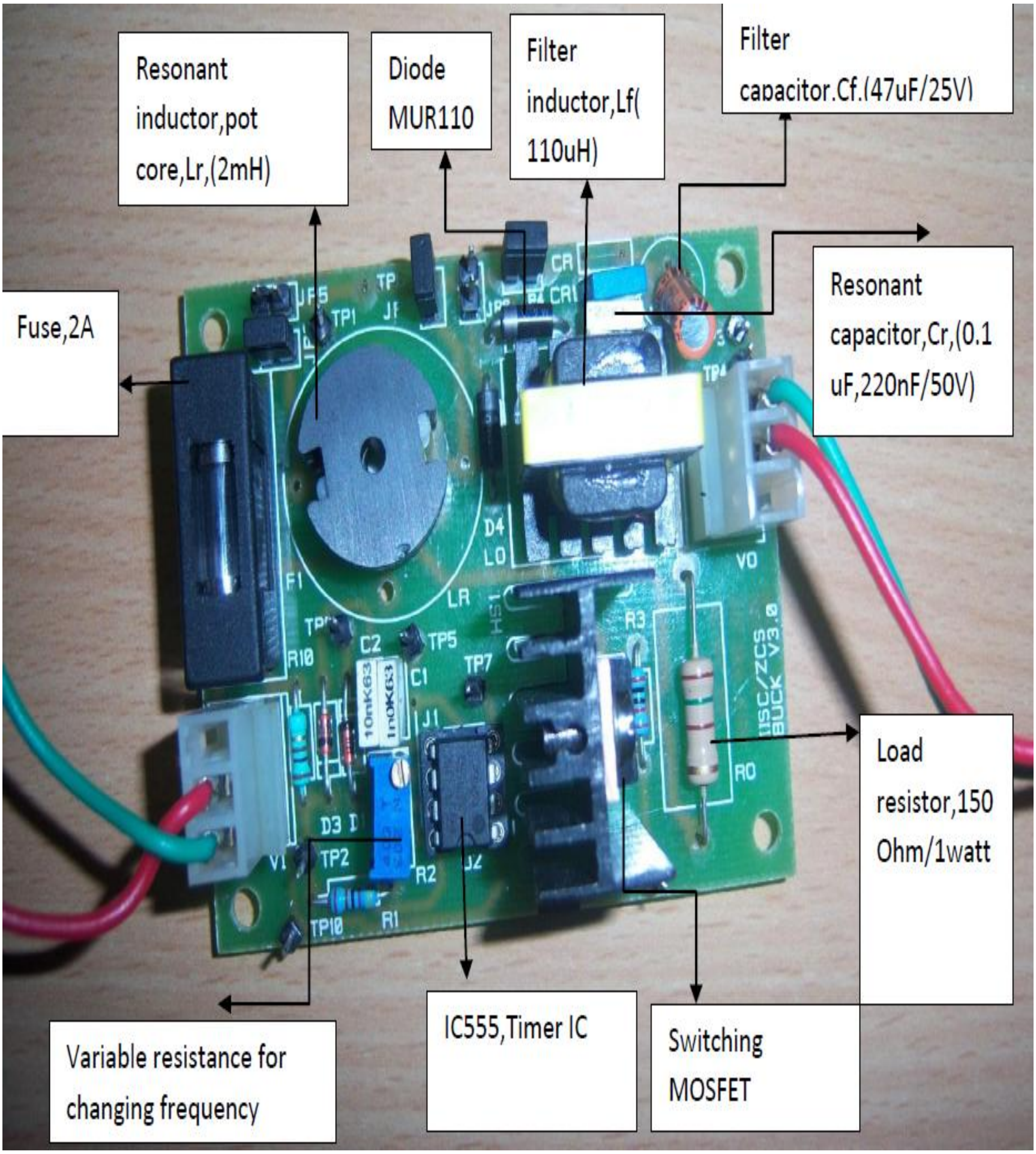
CONSTUCTION PROJECT 10 — 7.5W ZCS BUCK CONVERTER V3.0 — 03/02/0

Bill Of Materials

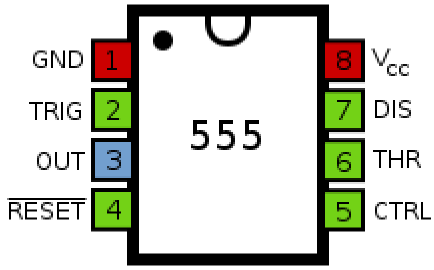
SL. NO.	REFERENCE	DESCRIPTION	VALUE	QTY	RATE	AMOUNT
1	Vi,Vo	CONNECTOR	3 WAY PSC	2	3.00	6.00
2	Ro	RESISTOR CFR	150E/1W	1	3.50	3.50
3	R1	RESISTOR MFR	4.7K	1	0.35	0.35
4	R2	RESISTOR	20K(POT 3296)	1	7.00	7.00
5	R3	RESISTOR MFR	10K	1	0.35	0.35
6	R10	RESISTOR MFR	100E	1	0.35	0.35
7	Co	ELECTROLYTIC	47uF/63V	1	2.00	2.00
8	Cr	CAPACITOR	NC	1	1.50	1.50
9	Cr1	CAPACITOR	0.1uF/100nF/50V	1	1.50	1.50
10	Cr2	CAPACITOR	0.2uF/220nF/50V	1	1.80	1.80
11	C1	CAPACITOR	0.001uF/1nF	1	1.10	1.10
12	C2	CAPACITOR	0.01uF/10nF/50V	1	1.10	1.10
13	D2,D4	DIODE	MUR110	2	8.00	16.00
14	D1	DIODE	1N4148	1	0.40	0.40
15	D3	DIODE	8.2V ZENER	1	1.00	1.00
16	F1	FUSE	2A	1	1.50	1.50
17	FUSE HOLDER	FUSE HOLDER	PBF4	1	2.80	2.80
18	M1	MOSFET	IRFZ44	1	16.00	16.00
19	U2	IC	555 TIMER	1	3.6	3.60
20	Lo	INDUCTOR	110uH E20/10/6 29T,22SWG	1	25.00	25.00
21	Lr	INDUCTOR	POT CORE,2STRANDS 4T,2uH,22SWG	1	30.00	30.00
22	TP1,TP2,TP3,TP4,TP5 TP6,TP7,TP8,TP9,TP10	TEST POINT	1PIN BERG	10	0.10	1.00
23	JP1,JP2,JP3,JP4,JP5,JP6	TEST POINT	2PIN BERG	6	0.20	1.20
24	HS1(25mm)	HEAT SINK	PI48/25MM	1	3.20	3.20
25	IC BASE	ROUND	8 PIN	1	2.20	2.20
26	WIRING,CRIMPING			4	2.00	8.00
27	SHORTING LINKS			6	1.00	6.00
28	SCREWS		M3/8	2	0.11	0.22
29	PLANE WASHER		M3	1	0.10	0.10
30	FIBRE WASHER		M3	1	0.15	0.15
31	PCB-SS/SM/LP	IISC ZCS BUCK V3.0 CONVERTER	7.5X5.5cm	1	22.00	22.00

1 Price of components are before tax
2 Minimum order quantity - 10
3 All Resistors are 1/4 W Metal Film type unless otherwise specified
4 Non polarised Capacitors are Polyster except values in pf range which are disc
5 PCB - Single side with Mask & Legend

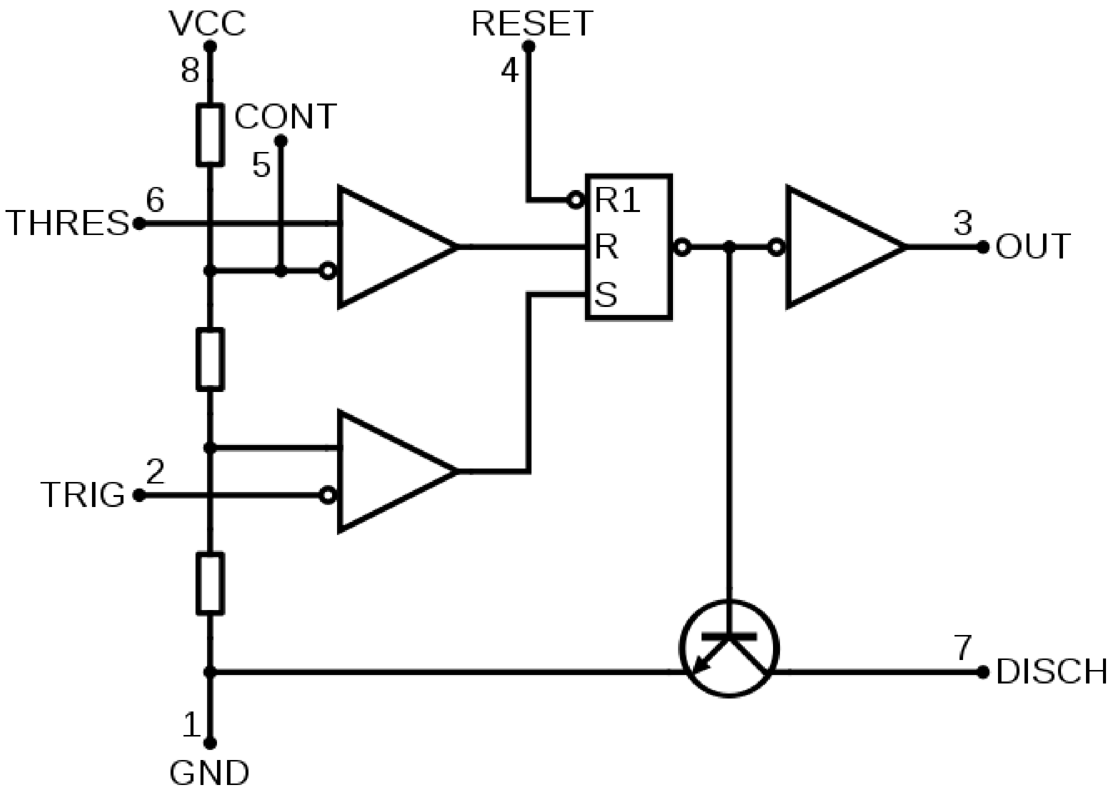
Fig 8.3 : Table of Bill Of Materials



Implemented Hardware Circuit of ZCS Buck Converter

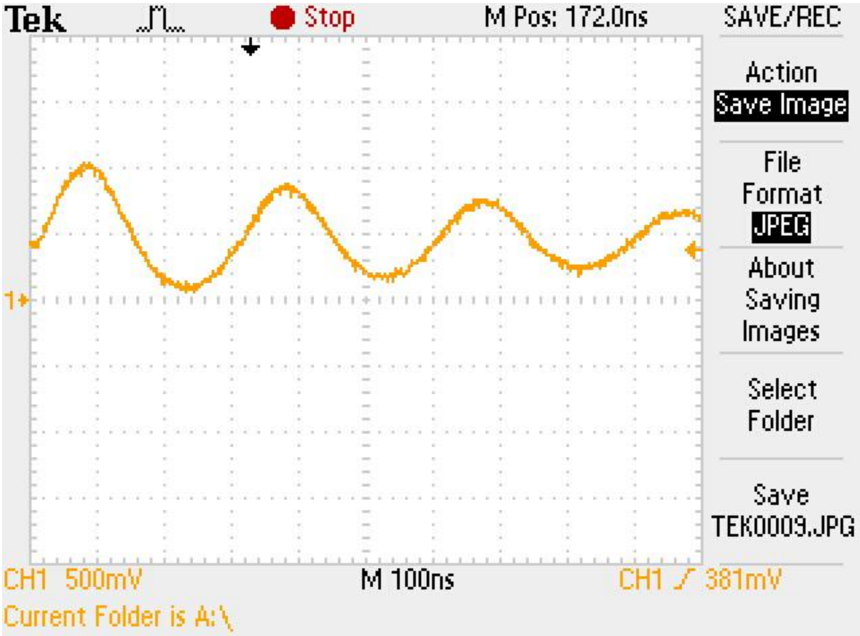


NE555 IC Pinout



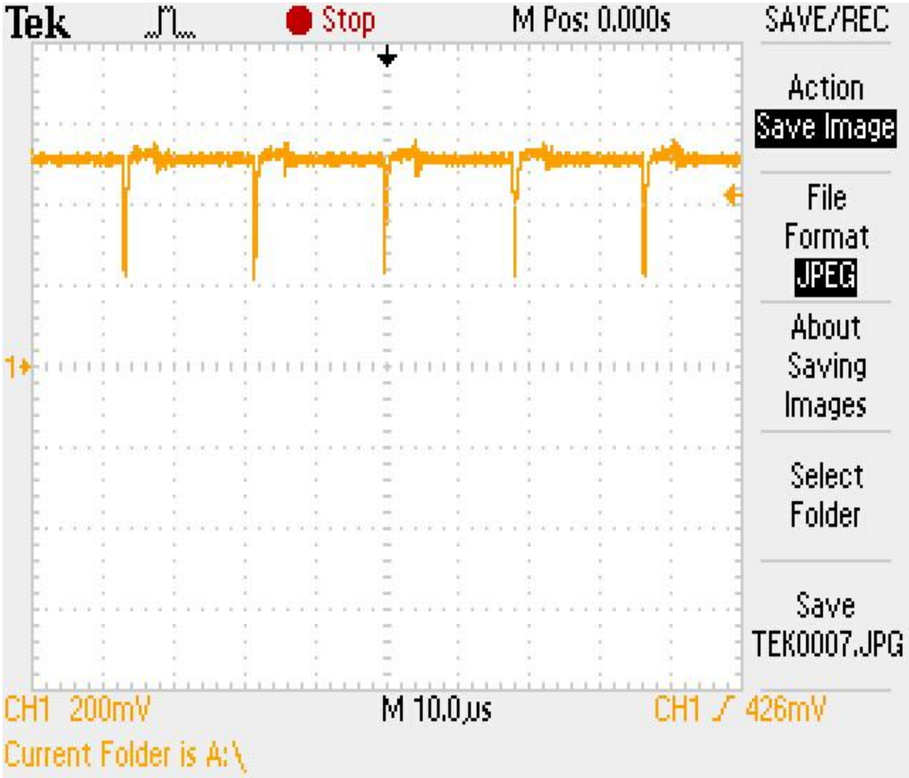
NE555 IC Circuit Schematic

Observed Waveform

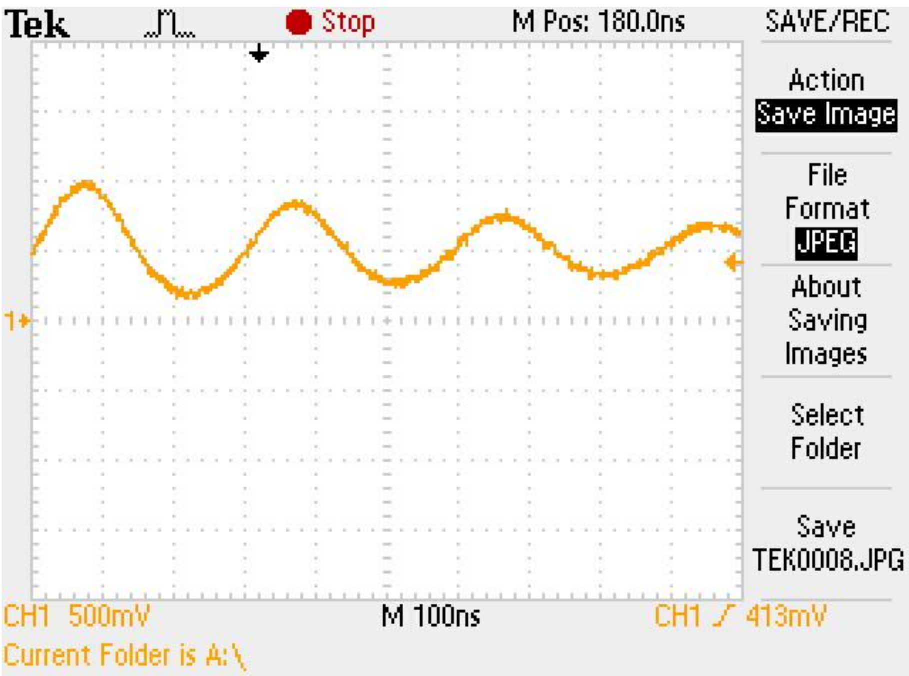


Voltage across Resonant Capacitor

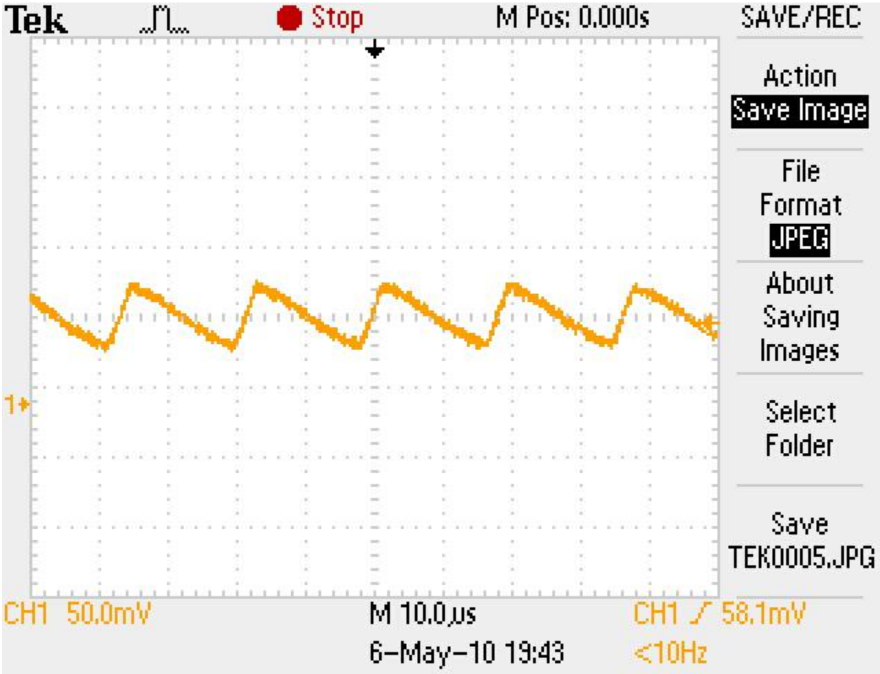
Output voltage across Load



Waveform across Zener Diode at Test Point 2

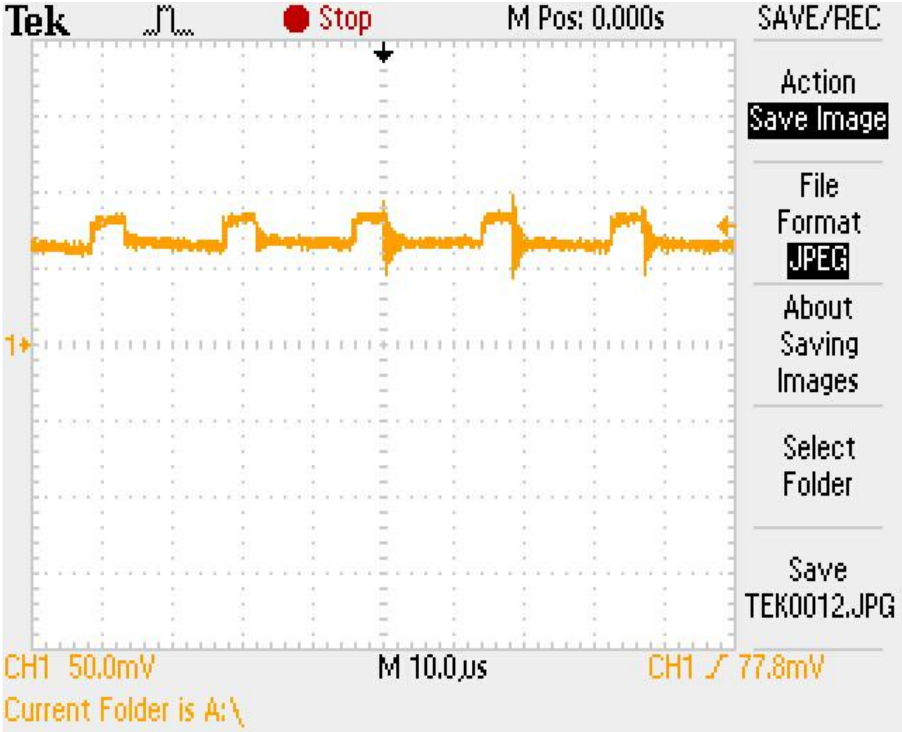


Waveform at Test Point 3

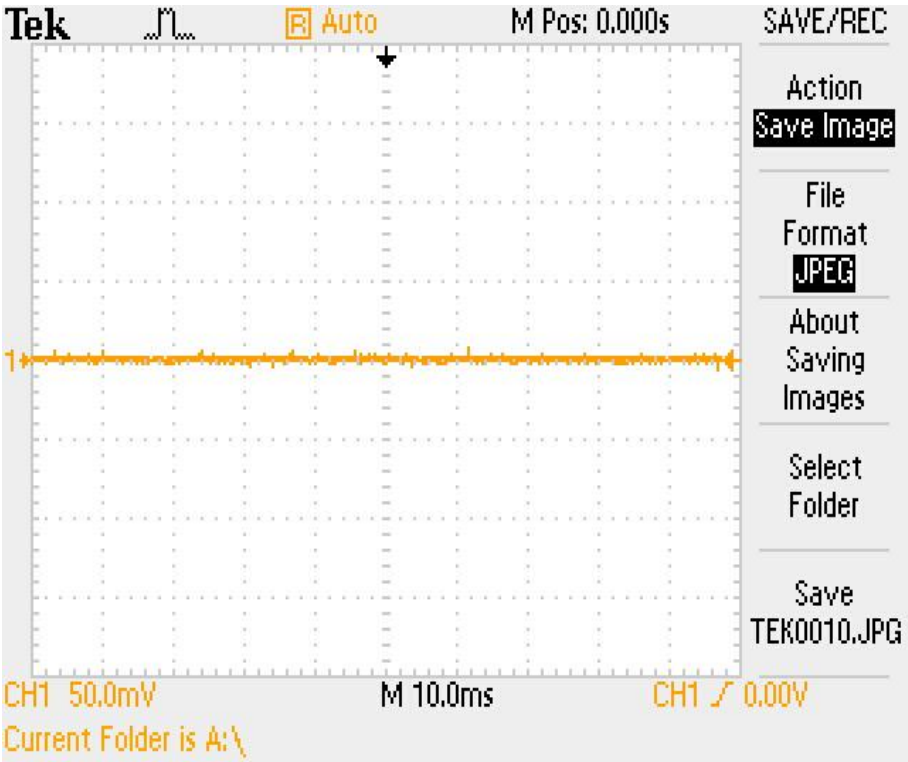


Waveform at Test Point 4

Waveform at Test Point 5



Waveform at Test Point 7



Waveform at Test Point 8

Mathematical Analysis of Modes of Operation

We operate from a regulated DC voltage supply and as the capacitor rating is a maximum voltage of 25V we select V_s=5V.

Assuming a purely resistive load the approximate current flowing in the load resistance is obtained as: I_o = 5/150=0.034A.

Resonant inductor = 2mH; Resonant capacitor = 0.304uF

Mode 1 :

This mode is valid for $0 \leq t \leq t_1$. Switch S_1 is turned on and diode D_m conducts. The inductor current i_L which rises linearly is given by $I_L = (V_s/L)t$.

This mode ends at time $t=t_1$ when i_l $(t=t_1)$ = 0.034.

That is $t_1 = I_0L/V_s$ = (0.034*2*e-3)/(5)=1.36xe-5 sec.

Mode 2 :

This mode is valid for $0 \leq t \leq t_2$. Switch S_1 remains on but diode D_m is off. The inductor current i_L is given by $i_L=I_m \sin \omega_o t + I_0$,

where $I_m = V_s\sqrt{(C/L)}$= 5(0.304*e-6/2*e-3)^0.5=0.0616.

The capacitor voltage V_c is given by $V_c = 5(1 - \cos\omega_0 t)$.

The peak current which occurs at $t = (\pi/2)\sqrt{LC}$ is $I_p = I_m + I_0$ = 0.034+0.0616=0.956A

The peak capacitor voltage is given by $V_c(pk) = 2V_s$ =10V.

This mode ends at $t = t_2$ when $i_L(t = t_2) = I_0$ and $Vc(t = t_2) = V_{c2} = 2V_s$=10V.

Therefore $t_2 = \pi\sqrt{LC}$ = 0.314*(2*e-3*0.314*e-6)^0.5 =7.74xe-5 sec.

Mode 3 :

This mode is valid for $0 \leq t \leq t_3$. The inductor current that falls from I_0 to zero is given by $i_L = I_0 - Im \sin\omega_0 t$. The capacitor voltage is given by $V_c = 10\cos\omega_0 t$. This mode ends at $t = t_3$ when i_L $(t = t_3) = 0$.

And $V_c(t = t_3) = V_{c3}$. Thus $t_3 = \sqrt{LC}\sin^{-1}(1/x)$

where $x = I_m/I_0 = (V_s/I_0)\sqrt{C/L}$= 8.26xe-5

Mode 4 :

This mode is valid for $0 \leq t \leq t_4$. The capacitor supplies the load current I_o(0.034) and its voltage is given by $V_c = V_{c3} - (I_0/C)t$

This mode ends at $t = t_4$ when V_c $(t = t_4) = 0$. Thus $t_4 = V_{c3}C/I_0$;

V_{c3}=8.3V; t_4=7.42xe-5 sec.

Mode 5 :

This mode is valid for $0 \leq t \leq t_5$. When the capacitor voltage tends to be negative, the diode D_m conducts. The load current Io flows through diode D_m. This mode ends at time $t=t_5$ when the switch S_1 is turned on again and the cycle is repeated i.e. $t_5=T-(t_1+t_2+t_3+t_4)$.

For the sake of simplicity we assume $t_5=0$.

Therefore $T_{on} = t_1+t_2+t_3= 17.36xe-5$

$T_{off} = t_4+t_5= t_4 = 7.42xe-5$

$T= T_{on} + T_{off} = 24.78xe-5$

Duty ratio = T_{on}/T = 70.05%

Frequency of operation = 1/T = 4KHz.

Conclusion

ZCS buck converter is an efficient step down DC-DC converter used in numerous electronics devices. The same was implemented as a hardware project and an output voltage of 2V was obtained with an input of 5.32V DC supply. The time analysis of various modes of zcs buck converter was done and tuning of the IC555 was done accordingly. Also the waveforms across capacitors and various test points were obtained, studied and compared with the theoretical waveforms. The waveforms were found to be in precise proximity of theoretical waveforms.

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