# DC Current Mode Control of Forward DC to DC Converter

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*Abstract*— The purpose of this project is to implement MOSFET based high precision low size prototype DC-DC converter for battery charging applications. Because of the disadvantages of linear power supplies, now most of the power supplies are replaced by switching mode power supplies (SMPS). In linear power supplies, a low-frequency (60-Hz) transformer is required. Such transformers are large in size and weight compared to high-frequency transformers. The transistor operates in its active region, incurring a significant amount of power loss. Therefore, the overall efficiencies of linear power supplies are usually in a range of 30-60%. Any SMPS basic requirements are Isolation between the source & the load, High-power density for reduction of size and weight, Controlled direction of power flow, High conversion efficiency, Input & output wave forms with a low harmonic distortion for small filters and Controlled power factor (PF) if the source is AC Voltage. The implementation of 400V dc to 108V dc converter using forward converter will fulfill the above requirements. The main advantage for selecting this topology is used for medium power applications, high switch utilization and number of switching devices is less. In this project current mode control is used in order to get the faster response, the switching frequency is high to get reduce the transformer core size and the harmonic components is reduced. The voltage spike across the switch is proportional to the switching frequency as well as the leakage inductance. By using of the snubber circuit voltage spike across the switch is reduced. The main objective of this project is to control the current and the voltage using current mode controller. The implementation of 400V dc to 108V dc converter using forward converter includes current mode controller UC3844. This will control the gating pulses to the MOSFET at the 65 KHz switching frequency in order to get the regulated output and the input current.

***Key words:*** Forward converter, PWM controller, MOSFET

## I. INTRODUCTION

### A. Switching Power Supply:

Switching power supplies gained popularity in the early 1970s, coinciding with the introduction of the bipolar power transistor. The basic theory of the switching power supply has been known since the 1930s.Since the 1930s, many evolutionary changes have occurred to make the switching power supply meet the needs of many diverse applications. The switch-mode DC to DC converters convert one DC voltage level to another, by storing the input energy temporarily and then releasing that energy to the output at a different voltage. The storage may be in either magnetic components (inductors, transformers) or capacitors. This conversion method is more power efficient (often 75% to 98%) than linear voltage regulation (which dissipates unwanted power as heat). This efficiency is beneficial to increasing the running time of battery operated devices. Variations have evolved, each with merits that make it better suited for particular applications.

The efficiency has increased in since the late 1980's due to the use of power FETs, which are able to switch at high frequency more efficiently than power bipolar transistors, which have more switching losses and require a more complex drive circuit. Another important innovation in DC-DC converters is the use of synchronous switching which replaces the flywheel diode with a power FET with low "On" resistance, thereby reducing switching losses. Other advantages include smaller size and lighter weight (from the elimination of low frequency transformers which have a high weight) and lower heat generation due to higher efficiency. Disadvantages include greater complexity, the generation of high amplitude, high frequency energy that the low-pass filter must block to avoid electromagnetic interference (EMI), and a ripple voltage at the switching frequency and the harmonic frequencies.

### B. Factors Affecting the Choice of an Appropriate Topology:

1) The peak primary current. This is an indication of how much stress the power semiconductors must withstand and tends to limit a particular configuration in the output power it can deliver and the input voltage over which it can operate.
2) How much of the input voltage can be placed across the primary winding of the transformer. This indicates how effectively power can be derived from the input line. Switching power supplies are constant-power circuits, so the more voltage supplied to the transformer or inductor, the less the average and peak currents needed in order to develop the output power.
3) How much of the B-H characteristic can be used within the transformer during each cycle. This indicates which configurations have physically smaller transformers for a rated output power.
4) DC isolation of the input from the load. This provides DC isolation of the output from the input and allows the designer to add multiple outputs with ease. Transformer isolation may also be necessary in order to meet the safety requirements dictated by the marketplace.
5) Cost and reliability. The designer wishes to select a configuration that requires the minimum parts without subjecting the components to undue overstress.

### C. Non-Isolated Switching Power Supply Topologies:

The non-isolated type of switching power supplies are typically used when some external component provides the DC isolation or protection in place of the switching supply. These external components are usually 50-60-Hz transformers or isolated bulk power supplies. Their typical area of application is in local board-level voltage regulation. The non-isolated supplies are also easy to understand and thus are used as design examples by various manufacturers and subsequently overused by novice power supply designers. Non-isolated type configurations seldom are used by seasoned power supply designers simply because of the severity of the failure modes caused by the lack of the DC isolation. Also, isolated supplies add a degree of safety by having a second DC dielectric barrier to back up the 50-60-Hz transformer, which enhances the supply's degree of graceful degradation during any possible failures.

The non-isolated switching power supply topologies are Step-down (buck) converter, step-up (boost) converter, Step-down / Step-up (buck - boost) converter, Cuk converter and Full-bridge converter

Each topology generates and regulates an output voltage that is above or below the input voltage. Each also has only one output since it is not very practical to add additional outputs to them. Non-isolated supplies also have definite restrictions as to their application in regard to their input voltage respect to their output voltage

### *D. Isolated Switching Power Supply Topologies:*

As one may have seen in the non-isolated regulator topologies, only the semiconductors provide the DC isolation from the input to the output. Semiconductors have relatively low breakdown voltages and exhibit the worst mean time between failures (MTBF) of all the components within any given power supply. This is not because they were manufactured incorrectly but because of heat factors and sporadic adverse operating conditions such as transients. The isolated switching power supply topologies rely on a physical dielectric barrier provided by wire insulation and/or insulated tape. The energy passes through a non-conducting ferrite material prior to reaching the output. This transformer isolation can withstand many thousands of volts before it fails and does provide a second dielectric barrier in the event of a semiconductor failure. This greatly discourages the domino effect of failures once a failure does occur within the final product. On close inspection, the isolated regulators operate analogously to the non-isolated regulators.

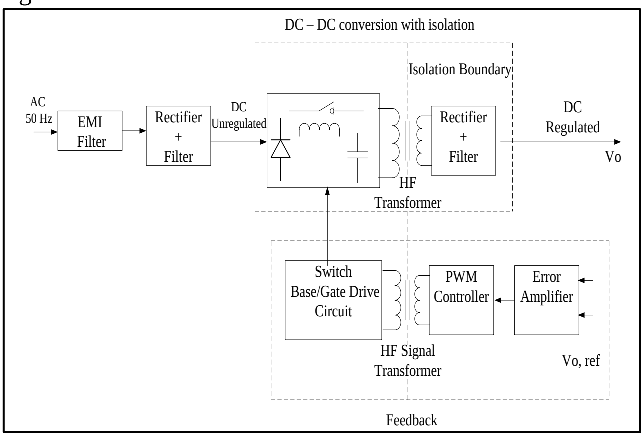


Fig. 1: schematic of switch-mode dc power supply.

As seen by the block diagram of Fig. 1, the electrical isolation in switching dc power supplies is provided by a high-frequency isolation transformer. Figure 2a shows a typical transformer core characteristic in terms of its B-H (hysteresis) loop. Here B_m is the maximum flux density beyond which saturation occurs and B_r is the remnant flux density.

Various types of dc-dc converters (with isolation) can be divided into two basic categories, based on the way they utilized the transformer core:

1) Unidirectional core excitation where only the positive part (quadrant 1) of $B-H$ loop is used.
2) Bidirectional core excitation where both the positive (quadrant 1) and the negative (quadrant 3) parts of $B-H$ loop are utilized alternatively.

### *E. Unidirectional Core Excitation:*

Some of the dc-dc converters (without isolation) disused above can be modified to provide electrical isolation by means of unidirectional core excitation. Two such modifications are Fly back converter (derived from buck – boost converter) & Forward converter (derived from step-down converter).The output voltage of these converters is regulated by means of the PWM switching scheme.

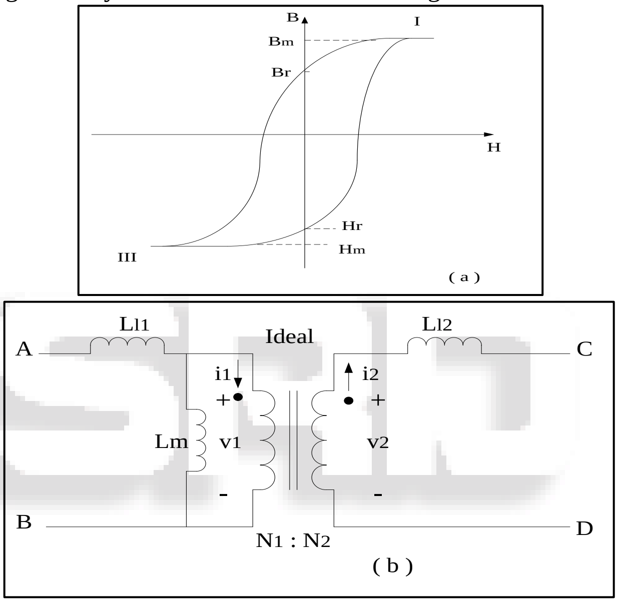


Fig. 2: Transformer representations: (a) typical B-H loop of transformer core; (b) Equivalent circuit of two winding transformer.

### *F. Bidirectional Core Excitation:*

To provide electrical isolation by means of bidirectional core excitation, the single phase switch mode inverter can be used to produce a square wave AC at the input the high frequency isolation transformer as shown in Fig1. The switch mode power supply can be of the following types are Push-pull, Half bridge & Full bridge

The transformer also provides a second great advantage, the ease of adding multiple outputs to the power supply without adding additional separate regulators for each output. All these factors make the transformer-isolated regulator topologies an attractive choice for virtually all applications.

## II. CONTROL STRATEGY

### *A. Basic Operation:*

Forward converter topology is shown in Fig 3.a. It is probably the most widely used topology for medium power applications.

 

As for the forward, Fig 3.a shows a master output $V_{om}$ and two slaves, $V_{s1}$ and $V_{s2}$. A negative-feedback loop is closed around the master and controls the $Q_1$ on time so as to keep $V_{om}$ constant against line and load changes. With an on time fixed by the master feedback loop, the slave outputs $V_{s1}$ , $V_{s2}$ are kept constant against input voltage changes but only partly (to about 5 to 8 percent) against load changes either in themselves or in the master. The circuit works as follows.

In the forward converter the MOSFET $Q_1$ is turned on, the dot end of the primary power winding $N_p$ and of all secondaries go positive with respect to their no dot ends. Current and power flows into the dot end of $N_p$. All rectifier diodes $D_2$ to $D_4$ be forward-biased and current and power flows out of the dot ends of all secondary's to the $LC$ filters and the loads.

Note that power flows to the loads when the power MOSFET $Q_1$ is turned on – thus the term *forward converter*. The push-pull and buck regulator also deliver power to the loads when the power MOSFET's are on and are also forward types. In contrast, the boost regulator, the polarity inverter (buck-boost) and the flyback type store energy in an inductor or transformer primary when the power MOSFET is on and deliver it to the load when the MOSFET turns off. Such energy storage topologies can operate in either the discontinuous or continuous mode. They are fundamentally different from forward topologies.

Now, for an $Q_1$ on time $T_{on}$ the voltage at the master rectifier cathode (Fig. 3b) is at a high level for $T_{on}$. Assuming a 1-V on voltage for $Q_1$ and a rectifier on forward drop of $V_{D2}$, that high-level voltage $V_{omr}$ is

$V_{omr}$ = [ $(V_{dc} - 1)\frac{N_m}{N_p}$ - $V_{D2}$……………………1

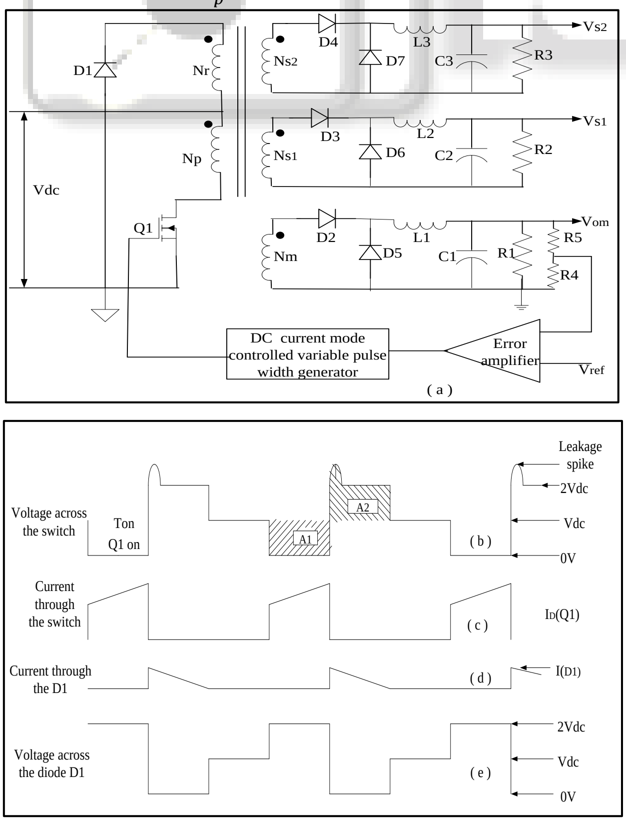


Figure 3(a) Forward converter topology, (b) Voltage across the switch, (c) Current through the switch, (d) Current through the diode, (e) Voltage across the diode D1.

Diodes $D_5$ to $D_7$ act like the freewheeling diode. When $Q_1$ turns off, current stored in the magnetizing inductance of $T_1$ reverses the polarity of the voltage across $N_p$. Now all the dot ends of primary and secondary windings go negative with respect to their no-dot ends. If not for "catch" diode $D_1$, the dot end of $N_r$ would go very far negative and since $N_p$ and $N_r$ have equal turns (usually), the no-dot end of $N_p$ would go sufficiently positive to avalanche $Q_1$ and destroy it.

However, catch diode $D_1$ catches the dot end of $N_r$ at one diode drop below ground. If there were no leakage inductance in $T_1$, the voltage across $N_p$ would equal that across $N_r$. Assuming that the 1-V forward drop across $D_1$ can be neglected, the voltage across $N_r$ and $N_p$ is $V_{dc}$ and the voltage at the no-dot end of $N_p$ and at the $Q_1$ drain is then $2V_{dc}$.

Recall, now, that within one cycle, if a core has moved in one direction on its hysteresis loop, it must be restored to exactly its original position on the loop before it moves in the same direction again in the next cycle. For otherwise after many cycles the core will be pushed in one direction or another into saturation. It will not be able to support the applied voltage and the MOSFET will be destroyed.

In fig. 3, it is seen that when $Q_1$ is on for a time $T_{on}$, $N_p$ is subjected to volt-second product $V_{dc}T_{on}$ with its dot end positive. That volt-second product is the area $A_1$ in Fig. 3. by Faraday's law $(E = NA_e(dB/dt) * 10^{-8})$, that volt-second product causes-say, a positive-flux change $dB = (V_{dc}T_{on}/N_pA_e)10^{-8}$ gauss.

At turnoff, when the magnetizing inductance has reversed the polarity across $N_p$ and kept its no dot-end at $2V_{dc}$ long enough for the volt-second area product $A_2$ in Fig. 3 to equal area $A_1$, the core has been restored to its original position on the hysteresis loop and the next cycle can safely start. In the common jargon expression, the "reset volt-seconds" has equaled the "set volt-seconds". Critical secondary currents in forward converter, each secondary has the characteristic ramp-on-a-step wave shape because of the fixed voltage across the output inductor and its constant inductance. It ramps up and down about the DC output current. Primary current is the sum of all the ramp-on-a-step secondary currents reflected by their turns ratios into the primary current is then also at ramp-on-a-step waveform.

Now when $Q_1$ has turned off, the dot ends of all secondaries go negative with respect to their no-dot ends. Current in all output inductors $L_1$ to $L_3$ try to decrease, since current in inductors cannot change instantaneously, the polarity across all inductors reverses in an attempt to maintain constant current. The front ends of the inductors try to go far negative but are caught at one diode drop below output ground by free-wheeling diodes $D_5$ to $D_7$ (Fig. 2.1) and rectifier diodes $D_2$ to $D_4$ be reverse-biased. Inductor current now continues to flow in the same direction through its output end, returning through the load, partly through the filter capacitor, and up through the free-wheeling diode back into the inductor.

Voltage at the cathode of the main diode rectifier $D_2$ is then as shown in Fig. 3.a. It is high at a level of [ $(V_{dc} - 1)(N_m / /N_p)$ ] - $V_{D2}$ for time $T_{on}$, and for a time $T - T_{on}$ it is one free-wheeling diode $(D_5)$ drop below ground. The $LC$ filter averages this waveform, and assuming that the forward drop across $D_5$ equals that across $D_2(=V_d)$, the $DC$output voltage at $V_{om}$ is

$$V_{om} = [( (V_{dc} - 1) \frac{N_m}{N_p} ) - V_d] \frac{T_{on}}{T} \text{...........2}$$

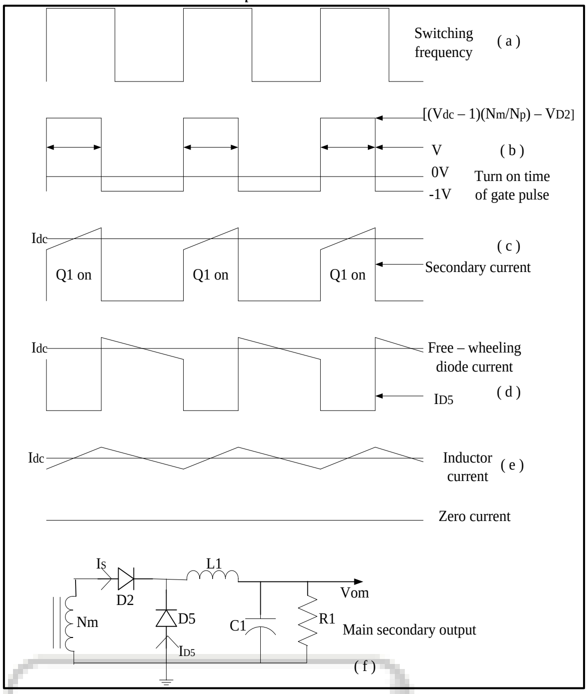


Fig. 4: (a) Switching frequency, (b) Turn on time of gate pulse, (c) Secondary current,(d) Free-wheeling diode (D5) current, (e) Inductor current, (f) Main secondary output.

*B. MOSFET Voltage Stress and Leakage Inductance Spikes:*

It can be seen from the transformer dots in Fig.3.a that when MOSFET is off, the MOSFET drain is subjected to at least twice the DC supply voltage since both primary and reset winding have an equal number of turns.

However, the maximum stress in somewhat more than twice the maximum DC input voltage the added contribution comes from the so-called leakage inductance spikes shown in Fig.3.b. These come about because there is an effective small inductance (leakage inductance $L_l$ ) in series with primary as shown in Fig.5.

At the instant of turnoff, current in the MOSFET falls rapidly at a rate $dI/dt$ causing a positive-going spike of amplitude $E_{1s}$ = $L_l\, dI/dT$ at the bottom end of the leakage inductance. Conservative design practice is to assume the leakage inductance spike may be as much as 30 percent more than twice the maximum DC input voltage. The MOSFET's should then be chosen so that they can tolerate with some safety margin, a maximum voltage stress ($V_{ms}$) of

$$V_{ms} = 1.3(2\overline{T_{on}}) \text{.........................3}$$

Leakage inductance spikes can be minimized by addition of a capacitor, resistor, and diode (CRD) combination to the transistor collector as shown in Fig. 5. Such CRD configurations also serve the important function of reducing AC switching losses due to the overlap of falling MOSFET current and rising voltage at the drain

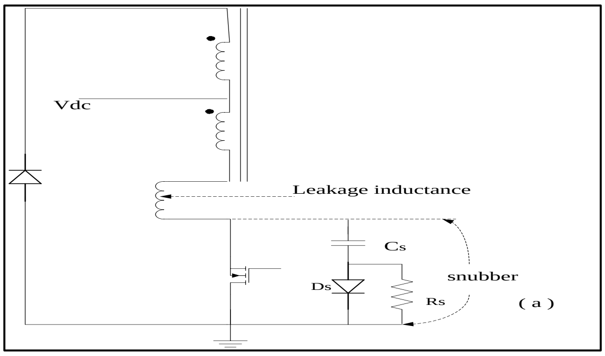


Fig. 5: Leakage inductances cause the spikes of Fig.3.a

*C. PWM Controller:*

In this project Current Mode PWM Controller (UC3844) has been selected to provide better protection for power MOSFET and also to ease the control loop design.

Current mode has two feedback loops: an outer one which senses DC output voltage and delivers a DC control voltage to an inner loop which senses peak power MOSFET currents and keeps them constant on a pulse-by-pulse basis.

*D. Feedback- Loop:*

Consider the negative-feedback loop for a typical forward converter. The essential error-amplifier and PWM functions are contained in all pulse-width-modulating chips. The chip also provides many other functions, but for understanding the stability problem, only the error amplifier and pulse-width modulator need be considered.

For slow or DC variations of the output voltage$V_o$, the loop is of course, stable. A small, slow variation of error amplifier EA via the sampling network, and compared to a reference voltage at the non-inverting EA input. This will cause a small change in the DC voltage level $V_{EA}$ at the EA output and at the EA input to the pulse-width-modulator PWM.

*E. Transformer Design & Output Inductor Design:*

The high frequency transformers are largely used in converter and inverter applications.

$V_{in}$ = 400V, $V_0$ = 108V, $I_0$ = 3A, $B_m$ = 0.2 T , J = 5 A/$mm^2$, $D_{max}$ = 0.5, $K_w$ = 0.4 , η = 85%, $V_{in\ max}$ = 400 V, $V_{in\ min}$ = 385V, $f_s$ = 65 KHz, $\Delta I$ = 1% to 10% of $I_0$

$$A_p = \frac{\sqrt{D_{max}}\, P_{02}(1 + \frac{1}{\eta})}{K_w\, J\, B_m\, f_s}$$

Chose a suitable core from the datasheets which has a $A_p$greater than the value calculated Ap.ETD 39/20/13 is a proper choice ($A_c$=128.679 $mm^2$, $A_w$=234.3 $mm^2$ $A_p$ = 30,149.48 $mm^4$).

For the forward converter inductor (L) = $\frac{V_0\,(1 - D_{min})T}{\Delta I}$

*F. MOSFET Selection:*

The practical maximum output power limit for a forward converter whose maximum DC input voltage is under 400V is about 324 W. thus consider a 324-W forward converter for the industry where the specified minimum and maximum input voltages are 385 and 400V, respectively.

Peak primary current is $I_{pft} = \frac{2.767 * P_o}{V_{dcmin}}$ = 2.328 A

Maximum off voltage stress is

$$V_{ms} = 2.5\, V_{dcmax} = 1000 \text{ V}$$

To provide the safety margin, the device with at least a 1000-V raring would be used to provide protection against input voltage transients which could drive the input above the maximum steady-state value of 400V.

## III. EXPERIMENTAL RESULTS

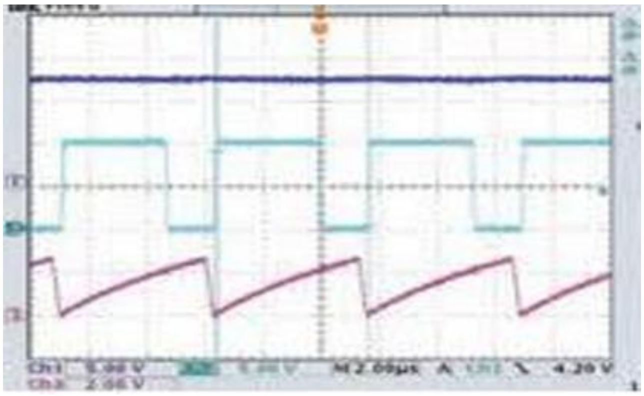

Fig. 6: (a) Output voltage waveform, (b) Output waveforms of PWM controller UC3844 at pin 6. (c) Saw tooth waveform at $R_T/C_T$ pin in PWM controller.

The output voltage must be held constant within a specified tolerance for changes within a specified range in the input voltage and the output loading. In this project the input voltage varies from 385V to 400V and the regulated output is 108 +/-1V.

The output waveform of the PWM controller UC3844 at pin 6. The total time period is 15.38$\mu sec$. These pulses give it to drive the MOSFET. In this wave form the peak to peak amplitude is 16.2 volts. The PWM controller maximum duty cycle is 70%.

The oscillator waveform, in this oscillator selection pin 4 the $R_T$ and $C_T$ set as 13.3$K\Omega$ and 1nF. The timing capacitor should be connected to the device ground.

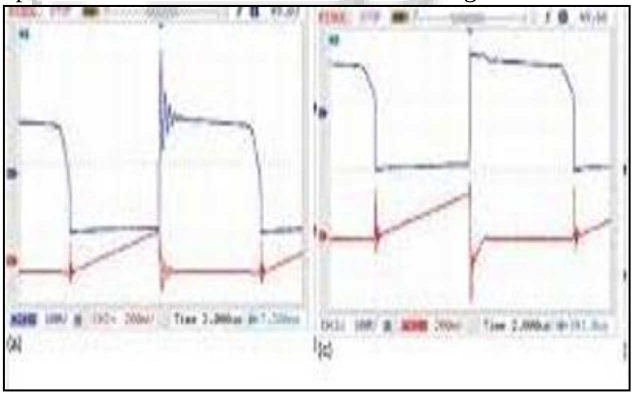

Fig. 7: (a) Voltage waveform across the switch & Input current waveform, (b) Voltage waveform across the switch & Input current waveform with snubber circuit

It can be seen from the transformer dots in Fig.3 that when MOSFET is off, the MOSFET drain is subjected to at least twice the DC supply voltage since both primary and reset winding have an equal number of turns.

However we are seen that the voltage wave across the MOSFET is more than the twice the input voltage, we can avoid the leakage inductance spikes can be minimized by addition of a capacitor, resistor, and diode (CRD) combination snubber circuit to the MOSFET drain as shown in Fig.5 Such CRD configurations also serve the important function of reducing AC switching losses due to the overlap of falling MOSFET current and rising voltage at the drain.

The current drawing from the in the switch is measuring through the current transformer (CT), as shown in Fig 7(a) & 7(b). This input current sense to the PWM controller to generate the gating pulses with respect to current drawing from the supply.

## IV. CONCLUSION

The analysis of various isolated PWM topologies has been made on basis of their power level, benefits, drawbacks, and cost. Considering above factors, it was concluded that for a medium power applications the forward converter is best suitable.

PWM controller with current control for both input side and output side of the forward converter is designed and applied. A high frequency isolation transformer is used, so the size and weight of converter are reduced.

A MOSFET based high frequency dc-dc converter has been designed for battery charging applications. A laboratory prototype was made and tested.

The setup was able to regulate the desired output voltage of 100V dc for an input of 400V dc at 65 KHz frequency and up to 3A load current. The expected results were satisfactorily obtained.

## V. FUTURE WORK

The work was done on single-ended forward converter. It has been observed that for medium power applications, in the single-ended forward converter the voltage across the switch is twice that of input voltage during it's off condition. Further during turn-off there is a leakage inductance spike across the switch. In order to overcome these disadvantages a double-ended forward converter may be proposed.

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