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A NOVEL ACTIVE CLAMPED DUAL SWITCH FLYBACK CONVERTER

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ABSTRACT

In this paper, A novel ZVZCS active clamped dual switch flyback converter was proposed, whose main switches and auxiliary switch all realize zero-voltage turning-on, and the rectifier diode on the secondary side also achieves ZCS. It overcomes the demerit of high voltage stress on the main switch in conventional flyback converter; meanwhile the duty circle can extend to more than 50% by slope compensation. Thus it is favorable for high efficiency, wide input range capacity and suited for the high input voltage occasions. In addition, the converter makes full use of leakage inductor energy, no extra snubber is needed

Keywords: ZVZCS Flyback Converter

1.Introduction

The traditional dual switch flyback converter

of electrical engineering, power electronics must be placed on a level with digital, analog, and radio-frequency electronics if we are to reflect its distinctive design methods and unique challenges. The history of power electronics has been closely allied with advances in electronic devices that provide the capability to handle high-power levels. Only in the past decade has a transition been made from a ‘‘device-driven’’ field to an system To put this in perspective, consider that a typical American household loses electric power only a few minutes a year. Therefore, energy is available 99.999% of the time. A converter must be even better than this if system degradation is to be prevented. An ideal converter implementation will not suffer any failures over its application lifetime. In many cases, extremely high reliability can be a more difficult objective than that of high efficiency.

Power electronics is the study of electronic circuits for the control and conversion of

electrical energy. The technology is a critical part of our energy infrastructure, and supports

almost all important electrical applications. For power electronics design, we consider only those circuits and devices that, in principle, introduce no loss and can achieve near-perfect reliability. The two key characteristics of high efficiency and high reliability are implemented with switching circuits, supplemented with energy storage. Switching circuits in turn can be organized as switch matrices. This facilitates their analysis and design.

1. **Robert Watson, Fred C. Lee, Guichao C. Hua**, ware presented in their paper “Utilization of an Active-Clamp Circuit to Achieve Soft Switching in Flyback Converters” A variety of soft-switching techniques either passive-clamping or active-
2. clamping methods have been presented which have well solved the problem of voltage spike caused by leakage inductor, but the voltage stress is still so high that it is inapplicable to high voltage occasions.

2. Nikolaos P. Papanikolaou and Emmanuel

C. Tatakis in there paper “Active voltage Clamp in Fllyback Converters Operating in CCM Mode under Wide Load Variation” says that the energy of leakage inductor feedbacks to input source, no snubber is needed. However, the duty cycle this kind of topology can not exceed 50% and hard switching operation is commutated. Thus it can not be utilized in wide input voltage application. Though some improved topologies, which have the

4. K. Harada, H. Sakamoto, in the paper “Switched snubber for high frequency switch,” Explains without a snubber the leakage inductance of the Flyback transformer rings with stray capacitance in the circuit, producing large amplitude and high frequency waveforms

5. Sebastian J., Martinez J.A, Alonso,

J.M, et al, in there paper, Analysis of the Zero-current-swithched quasi-resonant flyback, SEPIC and Cuk used as Power factor preregulators with voltage-follower control, the quasi-resonant Flyback was also proposed but it is suitable to applications with fixed input source and not favorable for wide input application.

1.2Classification

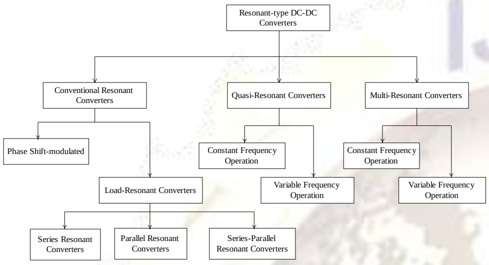


Fig.1.2 Classification

1.3. Resonant Switch

Prior to the availability of fully controllable power switches, thyristorswere the major power devices used in power electronic circuits. Each thyristor requires a commutation circuit, which usually consists of a LC resonant circuit, for forcing the current to zero in the turn-off process. This mechanism is in fact a type of zero-current turn-off process. With the recent advancement in semiconductor technology, the voltage and current handling capability, and the switching speed of fully controllable switches have significantly been improved. In many high power applications, controllable switches such as GTOs and IGBTs have replaced thyristors. However, the use of resonant circuit for achieving zero-current-switching (ZCS) and/or zero-voltage-switching (ZVS) has also emerged as a new technology for power converters. The concept of resonant switch that replaces conventional power switch is introduced in this section.
A resonant switch is a sub-circuit comprising a semiconductor switch S and resonant elements, L_r and C_r. The switch S can be implemented by a unidirectional or bidirectional switch, which determines the operation mode of the resonant switch. Two types of resonant switches, including zero-current (ZC) resonant switch and zero-voltage (ZV) resonant switches, are shown in Fig.3 and Fig.4, respectively.

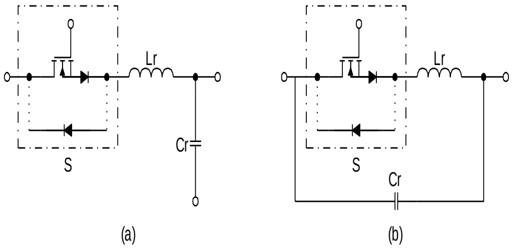


Fig.1.3 Zero-current (ZC) resonant switch.

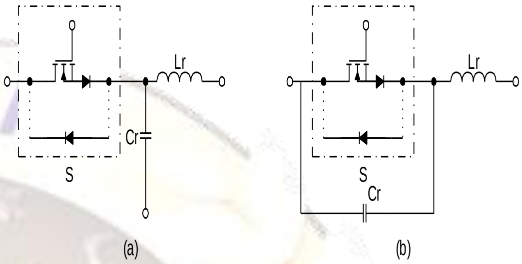


Fig.1.4 Zero-voltage (ZV) resonant switch.

1.4. ZC resonant switch

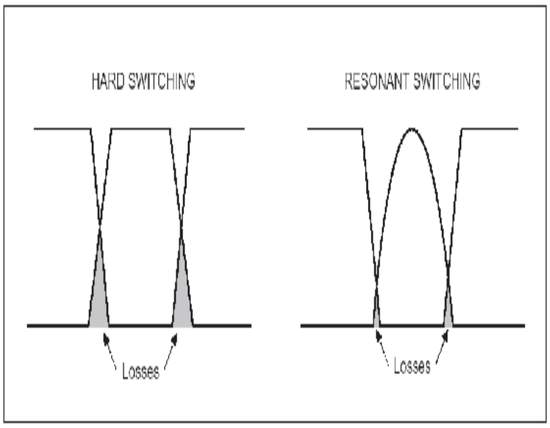
In a ZC resonant switch, an inductor L_r is connected in series with a power switch S in order to achieve zero-current-switching (ZCS). If the switch S is a unidirectional switch, the switch current is allowed to resonate in the positive half cycle only. The resonant switch is said to operate in half-wave mode. If a diode is connected in anti-parallel with the unidirectional switch, the switch current can flow in both directions. In this case, the resonant switch can operate in full-wave mode. At turn-on, the switch current will rise slowly from zero. It will then oscillate, because of the resonance between L_r and C_r. Finally, the switch can be commutated at the next zero current duration. The objective of this type of switch is to shape the switch current waveform during conduction time in order to create a zero-current condition for the switch to turn off.

1.5 ZV resonant switch

In a ZV resonant switch, a capacitor C_r is connected in parallel with the switch S for achieving zero-voltage-switching (ZVS). If the switch S is a unidirectional switch, the voltage across the capacitor C_r can oscillatefreely in both positive and negative half-cycle. Thus, the resonant switch can operate in full-wave mode. If a diode is connected in anti-parallel with the unidirectional switch, the resonant capacitor voltage is clamped by the diode to zero during the negative half-cycle. The resonant switch will then operate in half-wave mode. The objective of a ZV switch is to use the resonant circuit to shape the switch voltage waveform during the off time in order to create a zero-voltage condition for the switch to turn on.

1.6 Zero current switch:

Zcs consist of a switch S in series with the inductor L and the capacitor C connected in parallel if an output transformer is used in certain cases its parasitic inductance can be used as the resonant inductance when the switch S is off the resonant capacitor is charged up with more or less constant current, and so the voltage across it rises linearly. When switch is turned on the energy stored in the capacitor is transferred to the inductor, causing a sinusoidal current to flow in the switch during the negative half wave, the current flows to the anti parallel diode, and so in this period there is no current through or voltage across the switch: and it can be turned off without losses this type of switching is also known as thyristors mode, as it is one of the more suitable ways of using thyristors.



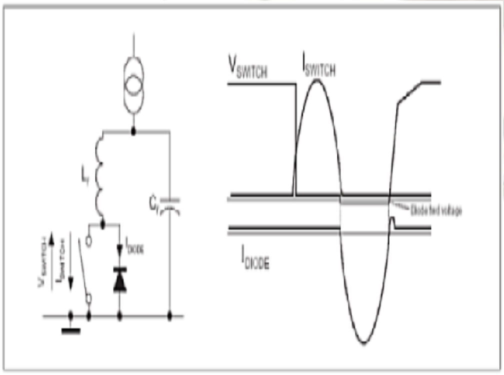


Fig 1.5 Zero current switch topology and waveforms

1.7. Zero voltage switch:

A zero voltage switch consists of a switch in series with a diode. The resonant capacitor is connected in parallel, the resonant inductor is connected in series, a voltage source connected in parallel injects the energy in to this system. When the switch is turned on a linear current flows through the inductor when the switch is turned off the energy stores in inductor flows in to the resonant capacitor the resulting voltage across the capacitor and the switch is sinusoidal the negative half wave of the voltage Is blocked by the diode during this negative half the current and voltage in the switch are zero and so it can be turned on without losses

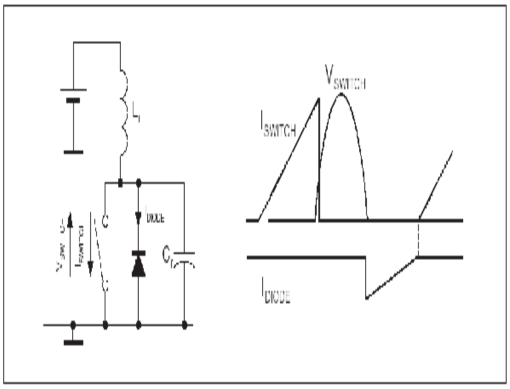


Fig 1.6Zero voltage switch topology and wave forms

1.8 Comparisons between ZCS and ZVS

ZCS can eliminate the switching losses at turn-off and reduce the switching losses at turn-on. As a relatively large capacitor is connected across the output diode during resonance, the converter operation becomes insensitive to the diode's junction capacitance. The major limitations associated with ZCS when power MOSFETs are used are the capacitive turn-on losses. Thus, the switching loss is proportional to the switching frequency. During turn-on, considerable rate of change of voltage can be coupled to the gate drive circuit through the Miller capacitor, thus increasing switching loss and noise. Another limitation is that the switches are under high current stress, resulting in high conduction loss. It should be noted that ZCS is particularly effective in reducing switching loss for power devices (such as IGBT) with large tail current in the turn-off process.

ZVS eliminates the capacitive turn-on loss. It is suitable for high-frequency operation. For single-ended configuration, the switches could suffer from excessive voltage stress, which is proportional to the load. It will be shown in Section 15.5 that the maximum voltage across switches in half-bridge and full-bridge configurations is clamped to the input voltage.

For both ZCS and ZVS, output regulation of the resonant converters can be achieved by variable frequency control. ZCS operates with constant on-time control, while ZVS operates with constant off-time control. With a wide input and load range, both techniques have to operate with a wide switching frequency range, making it not easy to design resonant converters optimally.

2.Flyback Converter

2.1 Introduction

Flyback converter is the most commonly used SMPS circuit for low out put power applications. Where the out put voltage needs to be isolated from the input main supply the output power of Flyback type SMPS circuit may vary from few watts to less than 100 vats. The overall circuit topology of the circuit is considerably simpler than other SMPS circuits. Input to the circuit is generally unregulated Dc voltage obtained by rectifying the utility AC voltage followed by a simple capacitor filter. The circuit can offer single or multiple isolated output voltages and can operate over wide range of input voltage variation. In respect of energy-efficiency, Flyback power supplies are inferior to many other SMPS circuits but its simple topology and low cost makes it popular in low out power range.

The commonly used Flyback converter requires a singe controllable switch like MOSFET and the usual switching frequency is in the range of 100 KHz. A two switch topology exits that offers better energy efficiency and less voltage stress across the switches.

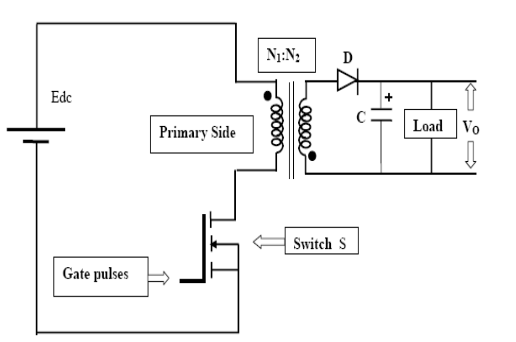


Fig: 2.1 Flyback Converter

2.2 Basic topology of Flyback converter

The above figure shows the basic topology of a Flyback circuit. Input to the circuit may be unregulated DC voltage derived from the utility AC supply after rectification and some filtering the ripple in DC voltage wave form is generally of low frequency and the overall ripple voltage waveform repeats at twice the AC mains frequency. Since SMPS circuit is operated at much higher frequency (in the range of 100 KW) the input voltage, in spite of being unregulated, may be considered to have a constant magnitude during any high frequency cycle. A fast switching device (S) like MOSFET is used with fast dynamic control over switch duty ratio to maintain the desired out put voltage. The transformer in figure used for voltage isolation as well as for better matching between input and output voltage and current requirements. Primary and secondary windings of the transformers are wound to have good coupling so that they are linked by nearly same magnetic flux.

In a normal transformer under load, primary and secondary windings conduct simultaneously such that the ampere-turns of primary winding is nearly balanced by the opposing ampere- turns of the secondary winding since secondary winding of Flyback transformer don't conduct simultaneously they are more like two magnetically coupled inductors and it May be more appropriate to call the Flyback transformer as inductor transformer.

Accordingly the magnetic circuit designing of a Flyback transformer is done like that for an inductor. the out section of the Flyback transformer, which consist of a voltage rectification and filtering, it is considerably simpler than in most other switched mode power circuit structure the secondary winding voltage is rectified and filtered using just a diode and a capacitor. Voltage across this filter capacitor is the SMPS output voltage.

The commonly used Flyback converter requires a single controllable switch like, MOSFET and the usual switching frequency is in the range of 100 KHz. A two switch topology exists that offers better energy efficiency and less voltage stress across the switches.

A more practical circuit will have provisions for out put voltage and current feedback and a controller for modulating the duty ratio of the switch. It is quite common to have multiple secondary winding for generating multiple isolated voltages. One of the secondary output may be dedicated for estimating the load voltage as well as for supplying the control power to the circuit.

For ease of under standing, some simplified assumptions are made. The magnetic circuit is assumed to be Lenoir and coupling between primary and secondary winding is assumed to be ideal. Thus the circuit operation is explained without consideration of winding leakage inductances. ON sate voltage drops of switches and diodes are neglected. the windings, the transformer core, capacitors etc. are assumed lossless. The input Dc supplies also assumed to be ripple-free.

2.3 Principle of operation

During its operation Flyback converter assumes deferent circuit configuration each of these circuit configurations have been refer here as modes of circuit operation. The complete operation of the power supply circuit is explained with the help of functionally equivalent circuits in these deferent modes.

In the above figure, when switch S is on, the primary winding of the transformer gets connected to the input supply with its dotted end connected to the positive side. At this time the diode 'D' connected in series with the secondary winding gets reverse biased due to the induced voltage in the secondary (dotted end potential being higher). Thus with the turning on of switch S, primary winding is able to carry current but current in the secondary winding is blocked due to the reverse biased diode. The flux established in the transformer core and linking the windings is entirely due to the primary winding current. This mode of circuit has been here as MODE1 circuit operation

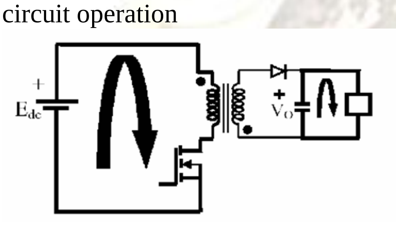


Fig: 2.2 Current path during mode1 of circuit op Flyback Converter eration

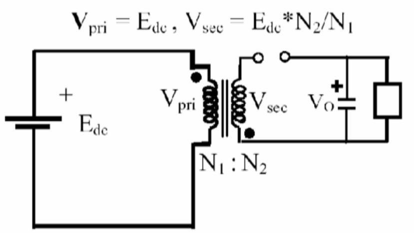


Fig: 2.3 Equivalent circuit in mode1

In the equivalent circuit shown, the conducting switch or diode is taken as an open switch. This representation of switch is inline with our assumption where the switches and diodes are assumed to have ideal nature, having zero voltage drop during conduction and zero leakage current during off state.

Under Mode1, the input supply voltage appears across the primary winding inductance and the primary current rises linearly. The following mathematical relation gives an expression for current rise through the primary winding.

$$E_{DC} = L_{Pri} \times \frac{d}{dt} i_{Pri} \text{------------------------------} (22.1),$$

Where E_{DC} is the input dc voltage, L_{Pri} is inductance of the primary winding and is the

Instantaneous current through primary winding

Linear rise of primary winding current during mode-1 is shown in Fig.22.5 (a) and Fig.22.5(b). As described later, the fly-back circuit may have continuous flux operation or discontinuous flux operation. The waveforms in Fig.22.5 (a) and Fig.22.5 (b) correspond to circuit operations in continuous and discontinuous flux respectively. In case the circuit works in continuous flux mode, the magnetic flux in the transformer core is not reset to zero before the next cyclic turning ON of switch 'S'. Since some flux is already present before 'S' is turned on, the primary winding current in Fig. 22.3(a) abruptly rises to a finite value as the switch is turned on. Magnitude of the current-step corresponds to the primary winding current required to maintain the previous flux in the core.

At the end of switch-conduction (i.e., end of Mode-1), the energy stored in the magnetic field of the fly back inductor-transformer is equal to $L_{Pri} I_P^2 / 2$, where I_P denotes the magnitude of primary current at the end of conduction period. Even though the secondary winding does not conduct during this mode, the load connected to the output capacitor gets uninterrupted current due to the previously stored charge on the capacitor. During mode-1, assuming a large capacitor, the secondary winding voltage remains almost constant and equals to

$$V_{Sec} = E_{DC} \times N_2 / N_1 .$$

During mode-1, dotted end of secondary winding remains at higher potential than the other end. Under this condition, voltage stress across the diode connected to secondary winding (which is now reverse biased) is the sum of the induced voltage in secondary and the output voltage

$$(V_{diode} = V_O + E_{DC} \times N_2 / N_1)$$

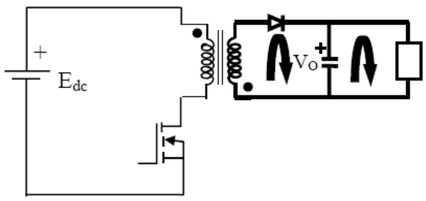


Fig: 2.4 Current path during mode2 circuit operation

Mode-2 of circuit operation starts when switch 'S' is turned off after conducting for some time. The primary winding current path is broken and according to laws of magnetic induction, the voltage polarities across the windings reverse. Reversal of voltage polarities makes the diode in the secondary circuit forward biased. Fig. 22.3(a) shows the current path (in bold line) during mode-2 of circuit operation while Fig. 22.3(b) shows the functional equivalent of the circuit during this mode.

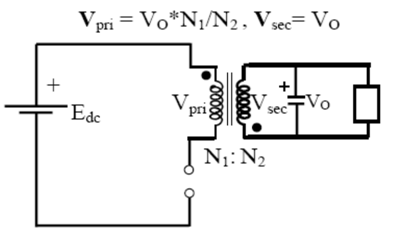


Fig: 2.5 Equivalent circuit in mode-2

In mode-2, though primary winding current is interrupted due to turning off of the switch 'S', the secondary winding immediately starts conducting such that the net mmf produced by the windings do not change abruptly. (mmf is magneto motive force that is responsible for flux production in the core. Mmf, in this case, is the algebraic sum of the ampere-turns of the two windings. Current entering the dotted ends of the windings may be assumed to produce positive mmf and accordingly current entering the opposite end will produce negative mmf.) Continuity of mmf, in magnitude and direction, is automatically ensured as sudden change in mmf is not supported by a practical circuit for reasons briefly given below. mmf is proportional to the flux produced and flux, in turn, decides the energy stored in the magnetic field (energy per unit volume being equal to , B being flux per unit area and μ is the permeability of the medium). Sudden change in flux will mean sudden change in the magnetic field energy and this in turn will mean infinite magnitude of instantaneous power, some thing that a practical system cannot support.

For the idealized circuit considered here, the secondary winding current abruptly rises from zero to as soon as the switch 'S' turns off. and denote the number of turns in the primary and secondary windings respectively. The sudden rise of secondary winding current is shown in Fig. 22.5(a) and Fig. 22.5(b). The diode connected in the secondary circuit, as shown in Fig.22.1, allows only the current that enters through the dotted end. It can be seen that the magnitude and current direction in the secondary winding is such that the mmf produced by the two windings does not have any abrupt change. The secondary winding current charges the output capacitor. The + marked end of the capacitor will have positive voltage. The output capacitor is usually sufficiently large such that its voltage doesn't change appreciably in a single switching cycle but over a period of several cycles the capacitor voltage builds up to its steady state value.

The steady-state magnitude of output capacitor voltage depends on various factors, like, input dc supply, fly-back transformer parameters, switching frequency, switch duty ratio and the load at the output. Capacitor voltage magnitude will stabilize if during each switching cycle, the energy output by the secondary winding equals the energy delivered to the load.

3 Active Clamped Duel Switch Flyback Converter

Flyback DC-DC converter, due to its advantages of simple topological structure, low cost, input and output isolation, etc, is widely used in auxiliary power supply, adapter, multiple output converters as well as other low and medium power applications. However, traditional single switch flyback DC-DC converter suffers from low utilization of transformer, high switch voltage stress and severe EMI. A variety of soft-switching techniques either passive-clamping or active-clamping methods have been presented in open literatures[1-5] which have well solved the problem of voltage spike caused by leakage inductor, but the voltage stress is still so high that it is inapplicable to high voltage occasions. The traditional dual switch flyback converter conquered the demerit of high switch voltage stress, whose two main switches just bear input voltage when they are off. Additionally, energy of leakage inductor feedbacks to input source, no snubber is needed. However, the duty cycle this kind of topology can not exceed 50% and hard switching operation is commutated. Thus it can not be utilized in wide input voltage application. Though some improved topologies, which have the advantage of wide duty cycle, are also proposed, meanwhile the demerits are obvious such as complicated control strategy or topology structure, one of the main switches inevitably subjecting to voltage spike. The quasi-resonant flyback [6-7] was also proposed but it is suitable to applications with fixed input source and not favorable for wide input application.

In this paper, active clamping method, which explored in detail for dual-switch forward converter [8-11], was applied in dual switch flyback converter, and a novel ZVZCS active clamped dual switch flyback converter was proposed, which inherits merit of low voltage stress of dual switch Flyback converter. Leakage inductor energy of this topology is utilized to achieve ZVS and rectifier diode realizes ZCS without the help of any other auxiliary circuit, thus switch-on loss is reduced and efficiency is improved. In addition, the duty cycle can extend to more than 50% by slope compensation. So it is recommended to wide input range, high input voltage and high performance applications, such as notebook adapter and auxiliary power supply for other wide range input converters. The basic operation process, characteristics and design principles will be analyzed in detail. Then experimental results are presented, which illustrate the converter function and verify the analysis presented.

3.2 Topology Structure and Operation Principles

S1 and S2 are main switches carrying load current, while S3 is auxiliary switch. The clamping capacitor (Cc). Connecting with S3 in series. Serves as a voltage source helping to reset transformers during (1-D) Ts period, Coss1, Coos2 and Coos3 are plastic output capacitors of S1-S2 correspondingly. The transformer includes magnetizing inductor Lm and Ls leakage inductor which place an important role in Zvs achieving. Diode Dc served to clamp the voltage stress on main switches and Dr is the rectifier diode on the secondary side

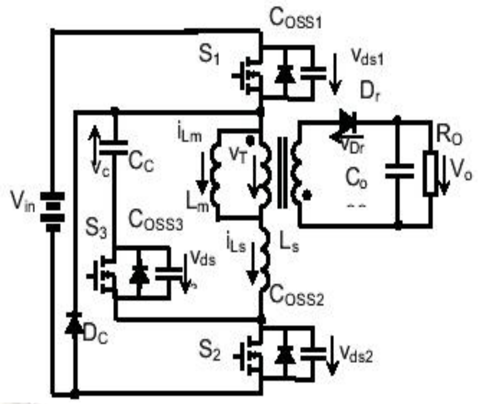


Fig 3.1: Active clamped duel switch flyback converter:

Power stage of the proposed converter is illustrated in Fig. 1. S1 and S2 are main switches carrying load current, while S3 is auxiliary switch. The clamping capacitor Cc, connecting with S3 in series, serves as a voltage source helping to reset transformer during (1-D) Ts period. Coss1, Coss2 and Coss3 are parasitic output capacitors of S1~S3 correspondingly. The transformer includes magnetizing inductor Lm and leakage inductor Ls which plays an important role in ZVS achieving. Diode Dc is served to clamp the voltage stress on main switches and Dr is the rectifier diode on the secondary side. S1 is supposed to turn off about 50ns earlier than S2 to assure the voltage stress of S1 be exactly input Voltage while S2 bears the same voltage stress as the clamping capacitor Cc.

The driving signals for switches and main principle waveforms are shown in Fig. 2. In order to simplify analysis, it is assumed that the circuit operation is in steady state, the output capacitor CO is large enough to be considered as a voltage source VO. The dead time t1-t3 and t5-t6 is actually rather short; they are lengthened on purpose in Fig. 2 for analysis convenience. There are about 7 stages in one switching period. Equivalent circuits for each stage are shown in Fig. 3. Operating processes for each stage are respectively described as follows:

Stage 1 [t0-t1]: The main switches S1 and S2 are both conducting, while S3 is off, transformer is clamped by input voltage Vin, magnetizing current iLm linearly increases, as well as the current of leakage inductor iLs. The Current through magnetic inductor and leakage inductor is equal. Meanwhile, the clamping diode Dc and the rectifier diode Dr are all reverse-biased. The output capacitor provides energy for output. In this stage, the operation mode is the same as the traditional flyback converter.

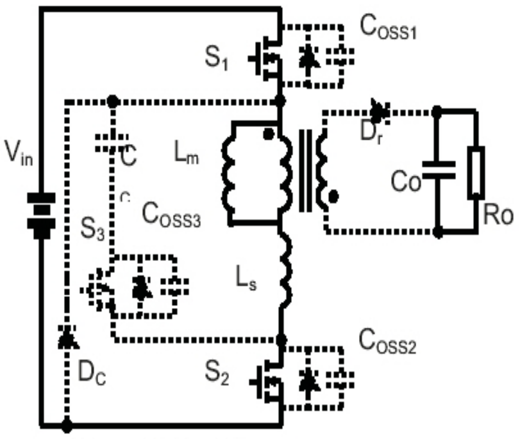


Fig 3.2 Stage1 (t0-t1)

Stage 2 [t1-t2]: At time t1, S1 turns off earlier than S2. iLm reaches the peak and it begins to charge capacitor Coss1. Coss3 is discharged at the same time till the voltage of Coss1 reaches input voltage. Then clamping diode Dc is freewheeling and the voltage stress is clamped to input voltage. Without Dc voltage stresses of S1 and S2 is uncontrollable. Magnetizing current and leakage current stay constant.

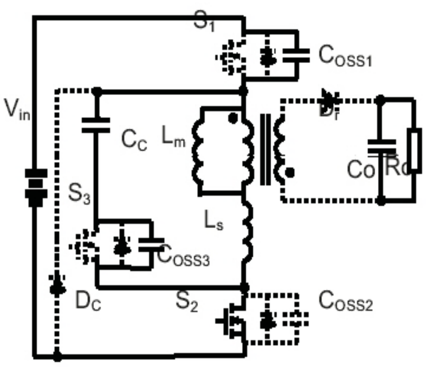


Fig.3.3 Stage 2 [t1-t2]

Stage 3 [t2-t3]: S2 turns off at t2 and S3 is still off, Coss2 is charged to Vc (the voltage of clamping capacitor) while Coss3 is discharged to zero. At this time, leakage current circulates through the parasitic diode of S3, and declines linearly under voltage Vc-nVO. Rectifier diode Dr conducts and magnetizing inductor is clamped by the output voltage. The turn-on signal of S3 can arrive at any time at this stage so long as iLs has not changed its direction.

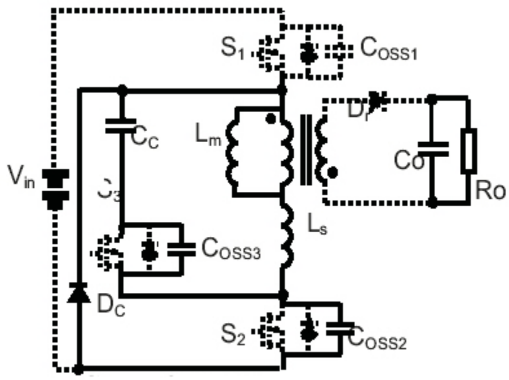


Fig. 3.4 Stage 3 [t2-t3]

Stage 4 [t3-t4]: S3 turns on under ZVS condition at time t3. iLs continues to circulate through the parasitic diode of S3, and declines under the voltage Vc-nVO . As soon as iLs reaches 0, stage 4 ends.

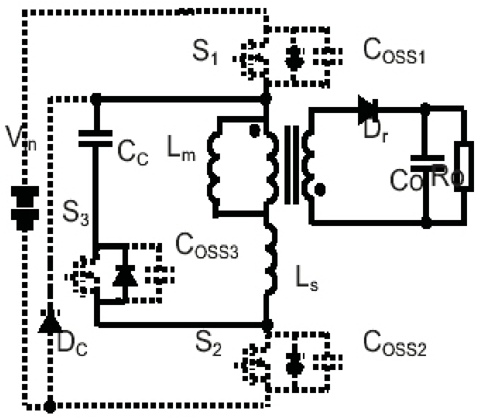


Fig.3.5 Stage 4 [t3-t4]

Stage 5 [t4-t5]: At this stage, ILs begins to circulate through S3 and increase reversely. The variation rate of Iis stays the same for the leakage inductor is still clamped by the voltage Vc-nVO. The energy of magnetizing inductor Lm continues delivering to secondary side. Rectifier diode remains conducting. When S3 turns off, ILs reaches its maximum value and stage 5 ends.

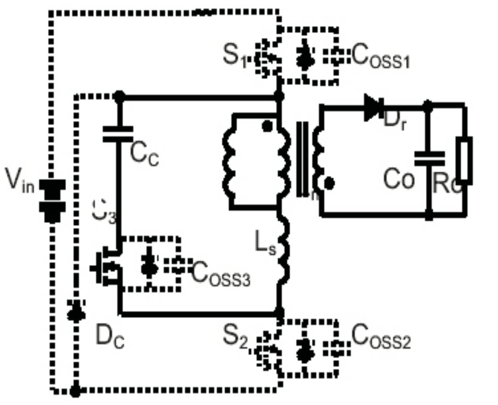


Fig.3.6 Stage 5 [t4-t5]

Stage 6 [t5-t6]: At time t5, S3 turns off. Since ILs is in reverse direction at this time, it starts to charge Coss3. Meanwhile, both Coss1 and Coss2 are discharged. At the end of stage 6, the voltage on Coss1 and Coss2 both declines to zero, and the parasitic diodes of S1 and S2 turn on naturally, which creates ZVS condition for S1and S2. Leakage inductor should have enough energy to discharge Coss1 and Coss2, or S1 and S2 can not achieve zero voltage turning-on. The current through the rectifier diode reaches its peak at the time t5 and beginning to decrease afterward.

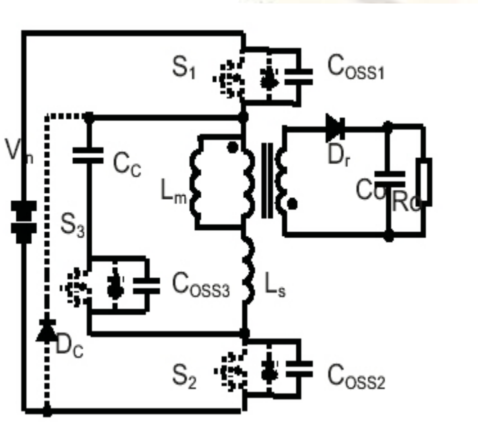


Fig 3.7 Stage 6 [t5-t6]

Stage 7 [t6-t7]: S1 and S2 turn on under the condition of ZVS at time t6. ILs increases quickly to be positive again under voltage Vin+nVO. As long as ILs equals to ILm, the current through Dr reaches zero and rectifier diode Dr achieves zero-current turning-off. It is the end of stage 7 and the beginning of the next switching period.

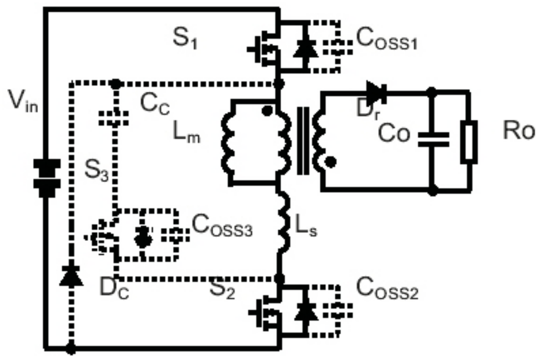


Fig 3.8 Stage 7 [t6-t7]

According to the previous analysis, main switches S1 and S2, as well as auxiliary switch S3, all achieve ZVS. The rectifier diode Dr achieves ZCS as well. The voltage stress on each semiconductor switch in the proposed topology is categorized in Tab. 1. Leakage inductor Ls, inherited in the transformer, servers as an independent component during (1-D) Ts, whose energy is used to achieve soft switching after exchanged energy with clamping capacitor Cc, no extra snubber is needed and no extra energy is consumed. Thus, the topology structure is simplified while high efficiency is obtained.

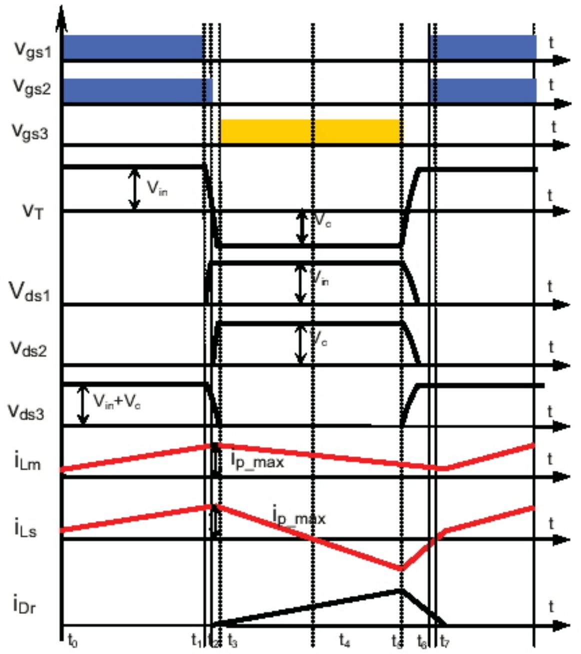


Fig 3.9 Operation Principle Wave Form

TABLE 1
THE VOLTAGE STRESS ON EACH COMPONENT

S_1	S_2	S_3	D_r	D_c
V_{in}	V_c	$V_c + V_{in}$	$V_O + \frac{V_{in} L_m}{(L_s + L_m) n}$	V_{in}

Design of output voltage VO and clamping voltage Vc

Actually the switching transient times, defined as t1-t3 and t5-t6, is rather short compared with the whole switching period and it will be ignored here for discussion convenience. Capacitor Cc and CO are large enough to be considered as voltage sources. Based on the assumption, the relationship between output voltage VO and input voltage Vin can be derived from the following equation:

$$L_m \Delta i_{Lm} = \frac{V_{in} D}{n f_s} \frac{L_m}{L_m + L_s} = \frac{n V_O (1-D)}{f_s} \qquad (1)$$

Where D is duty ratio; fs is frequency; Ls is the value of leakage inductor; Lm is the value of magnetizing inductor; n is the turn ratio of primary to secondary. iLm is the variation of current iLm during DTs. Then the expression of output voltage VO can be get:

$$V_O = \frac{V_{in}}{n} \frac{D}{1-D} \frac{L_m}{L_m + L_s} \qquad (2)$$

In stage 4, clamping capacitor Cc is charged by iLs and is discharged in stage 5. By ampere-second balance on the clamping capacitor, the peak values of leakage inductor's

Forward and reverse current are equal. Peak current value ip-max and its relationship with clamping voltage Vc can be expressed as follows:

$$i_{p\text{-}max} = \frac{I_O L_m}{n(1-D)(L_m + L_s)} + \frac{V_{in} D}{2 f_s (L_m + L_s)} \qquad (3)$$

$$2 i_{p\text{-}max} L_s = (V_c - n V_O)(1-D) T_s \qquad (4)$$

Where IO is load current. Then put (3) to (4), get:

$$V_c = \frac{V_{in} D}{1-D} + \frac{2 I_O L_s L_m f_s}{n (1-D)^2 (L_s + L_m)} \qquad (5)$$

(2) and (5) indicate that higher output voltage and higher load current lead to higher clamping voltage, and the voltage stresses of S2 and S3 ascend correspondingly. Fig. 4 gives a direct view of relationship between Vc and IO&Vin. Based on the analysis above, voltage fluctuation of Vc can also be obtained:

$$\Delta V_c = \frac{(1-D) i_{p\text{-}max}}{4 C_c f_s} \qquad (6)$$

Clamping capacitor Cc can be regarded as a constant voltage source when it is large enough that voltage fluctuation is neglect able. However, too large clamping capacitor will lead to the converter bulky, costly and slow dynamic response when load switches. On the contrary, if clamping capacitor is too small, then voltage fluctuation Vc will be large, voltage stress on Cc, S2 and S3 will increase either.

6.3 Application

- Adapter.
- Low and medium power applications.
- Auxiliary power supply.
- Battery charger.

6.4 Advantages of Proposed system

- High efficiency.

- Simple topological structure
- Wide input range capacity and suited for the high input voltage occasions.
- No extra snubber is needed.
- Low cost
- Input and output isolation is possible.

4. MATLAB SIMULINK DIAGRAMS & RESULTS OF DUAL SWITCH FLYBACK CONVERTER TOPOLOGY

4.1 Simulink Model

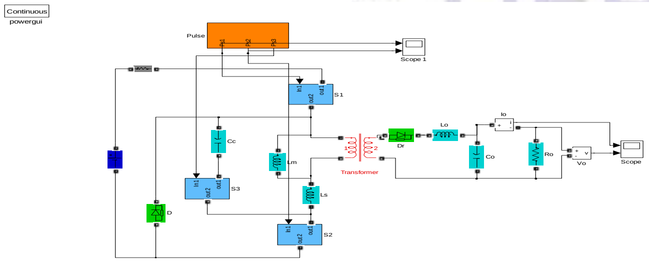


Fig 4.1 Simulink Model

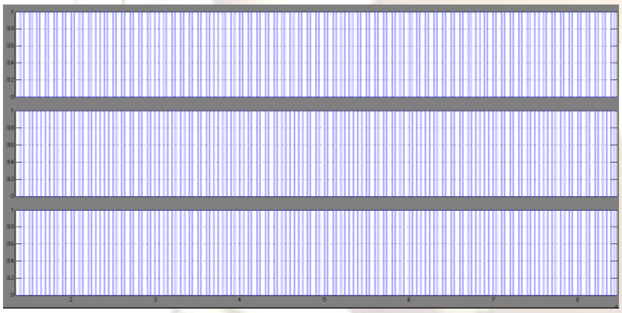


Fig 4.2 Pulses Wave Form

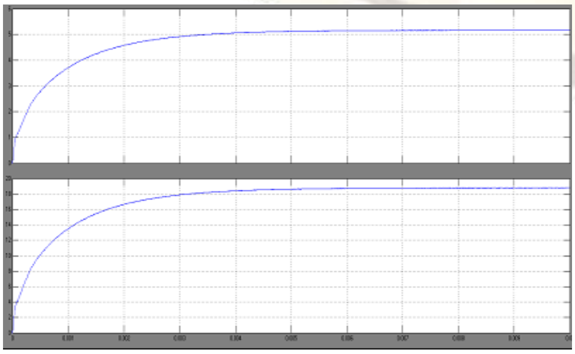


Fig 4.3 Output Voltage & Current Waveforms

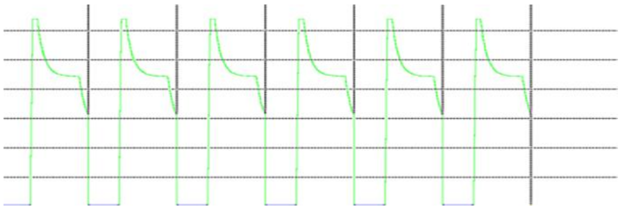


Fig 4.4 Zero Voltage and Zero Current Switching Wavveforms

5. Conclusions and Future Scope of Work

5.1 Conclusion

1.A Novel Active Duel Switch Flyback Converter with high efficiency is proposed to in this paper. The topology uses only three MOSFET devises for DC-DC conversion.

The proposed converter compared with the conventional Flyback converter eliminates the usage of snubber circuit

- Therefore this converter is fevered as the choice of low and medium power applications
- Comparing to the Conventional converter the major design and advantages of the proposed converter is quite favorable for improvement of conversion efficiency and power density
- The proposed Flyback converter is very attractive for high input, wide range and high efficiency practical application of small and medium power.

5.2 Future scope

As future studies, the controller must be designed based on the PWM technique with closed loop for high power application

Appendix-A

The main parameters are

Input voltage range (Vin)	127-330Vdc
Load current (Io)	5A
Output voltage (Vo)	18.6V
Operation frequency (Fs)	100K
Transformer Turns Ratio (n)	7:1
Magnetizing Inductance (Lm)	628uH
Leakage Inductance (Ls)	68uH
Clamping Capacitor (Cc)	2uF

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