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Simulation Comparison of Resonant Reset Forward Converter with Auxiliary Winding Reset Forward Converter

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ABSTRACT

This paper presents simulation results of Conventional axillary winding reset forward converter and resonant reset forward converter with synchronous rectifier operating at 250 KHz using LT8310. LT8310 is current mode controller, PWM signal generator. Operating frequency and soft-start time can be programmed by suitable selecting resistor and capacitor for LT8310. The limitation of 50% duty cycle in auxiliary winding reset forward converter can be overcome by using active clamp reset methods. Resonant reset method reduces power loss in the reset circuit. Replacing rectifier diode with self-driven MOSFET power losses on secondary side of transformer also reduces. Peak current of secondary side transformer for Resonant reset is clamped to higher value, thus this method is used for high current applications.

Keywords- Auxiliary winding Reset, Resonant Reset, current mode control, TL8310.

I.INTRODUCTION

In recent trends, Switch mode power supplies are designed to have high efficiency, low cost, light weight and small size. Weight and size of converter depends on capacitor filters and Transformer. Size of transformer reduces for high operating frequency. For output power in the range 100-300W, Forward converter topology is preferred [1]. Forward converter uses power transformer for energy transformation and circuit isolation [2]. In conventional forward converter auxiliary winding is to reset energy. Limiting factor for this technique is maximum duty cycle i.e. 50% to prevent saturation problems of transformer [2]. Duty cycle more than 50% can be achieved by using RCD reset circuit. The main drawback of this technique is power loss in the resistor which reduces efficiency. For high frequency operation reset of transformer can be achieved by resonance. Resonant circuit is formed by magnetizing inductance of transformer and capacitor. Resonant reset Forward converter has several advantages over other reset methods [3]:

-Low cost

-Leakage and magnetizing energy is fully recycled

-Duty cycle more than 50% can be achieved

In addition, current on secondary side of transformer is clamped. Rectifier diode can be replaced with MOSFET. This arrangement is known as synchronous Rectifier. Using advanced MOSFET technology, synchronous rectification become cost effective and can achieve maximum efficiency for low output voltage . However for the higher output voltage, voltage stress on switching device is too high which will increase the conduction losses. Therefore active clamp with synchronous rectifier is attractive topology for low output voltage converter.

II.CONVENTIONAL FORWARD CONVERTER WITH DIODE RECTIFICATION

The conventional Forward converter is shown in Fig.1.

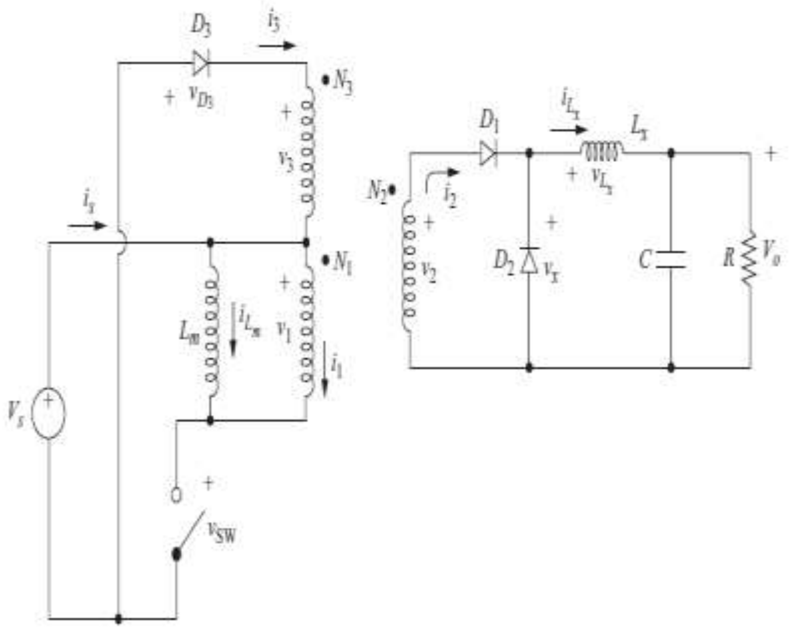

Fig.1: Auxiliary winding reset Forward converter

Basic power stage involve transformer with three windings. Winding 1 and 2 are used for Energy is transfer from source to load when the switch is closed for time DT using winding 1 and 2. Winding 3 is used to provide path for the magnetizing current when the switch is open for the period (1-D) T. Magnetizing current reduces to zero for this period, [4] where T is switching period. Waveforms of auxiliary winding current and various diode current are shown in Fig.2.

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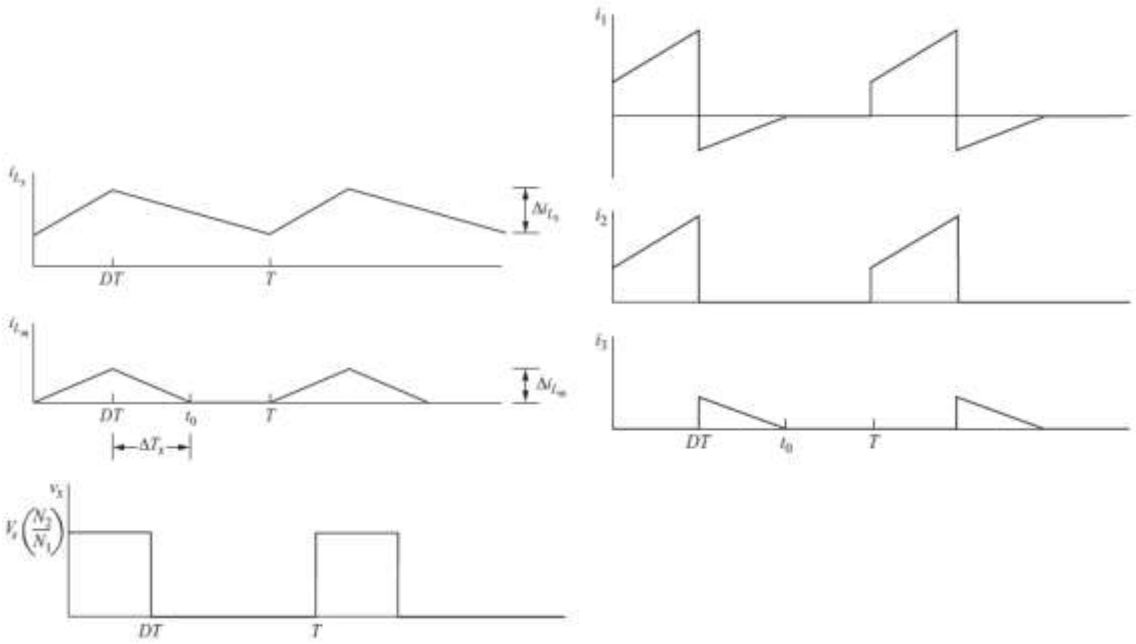

Fig.2: Current Waveforms

Winding 2 is connected to two diodes. Rectifier diode D1 turn on when switch is on and is reverse biased when the switch is off. Rectifier D2 is acts as freewheeling during switch is off [2]. In this method magnetizing current is reset to zero using winding 3 and due to resonance between inductance of winding 3 and parasitic capacitor of switch current continues in negative direction. This current plays key role for synchronous rectification [5].

III.FORWARD CONVERTER WITH RESONANT RESET AND SELF-DRIVEN SR

The forward converter with resonant reset and its waveforms are shown in Fig. 3 and Fig. 4 respectively.
In this method active clamp is formed with one clamp capacitor and one clamp switch which absorb the energy stored in the transformer and reduces stress on power switch. Energy absorption duration is almost entire off time of the primary switch (S). Which maximize the conduction time of transistor Q3.

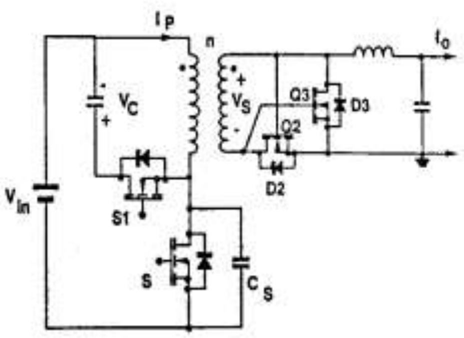

Fig.3: Forward Converter with resonant reset and self-driven SRs

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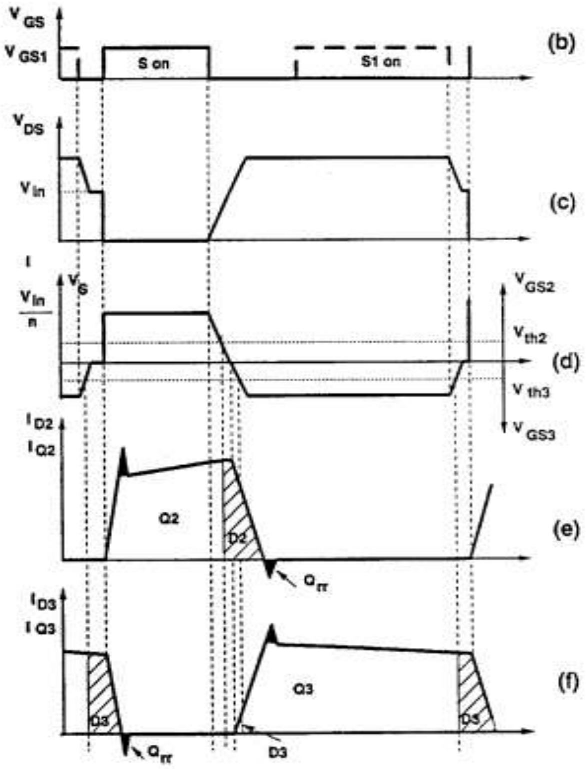

Fig.4: Forward Converter with active clamp and self-driven SRs: (b)gate-drive signals; (c) drain-to-source voltage of control switch: (d)secondary winding voltage; (e) current though SR2; and (f) current through SR3.

Since this approach minimize the voltage stress on primary switch (S) and can be turn on at zero voltage by adjusting magnetizing inductance of transformer [6].

Self-driven SR is implemented by cross-coupling of transistor gate terminals as shown in Fig 3. Here secondary gate drive signals are developed independently from the primary- side control i.e. no primary side information.

When primary switch is on, supply voltage appears across transformer primary, transformer secondary also follows supply voltage reduced by transformer turn ratio. As the voltage across secondary builds up, current flows through body diode of Transistor 2 until transformer secondary voltage turn on Q2. When primary switch is off, transformer secondary detects building of negative voltage. Secondary current initial flows through body diode of Q3 and places a positive voltage on the gate of Q2. Body diodes conduct on every switching cycle. Since transformer secondary voltage requires finite time between power transfer and reset, there is an unavoidable dead time. During turn on period Q2 experiences supply voltage reduced by the transformer turn ratio and during turn off period Q2 sees same voltage in negative direction. Hence for higher output voltage, voltage stress on switches is very large. Thus maximum input voltage range depends on the output voltage. Therefore this method feasible for application with a relatively small input voltage range and low output voltage.

IV. DESIGN AND IMPLEMENTATION OF FORWARD CONVERTER

Specifications and design considerations are given below

4.1Specification

- Input voltage range : 35 V to 55 V
- Switching Frequency : 248 KHz
- Output Parameters : 9.6 V/1 A
- Efficiency : ≥75%
- Line and Load regulation : ≤1%
- Ripple and Noise : <1% of output voltage.

4.2Abbreviation:

$R_{ds(on)}$: On State Resistance of MOSFET

I_{pft} : Peak current Pulse

t_{on} : On time of MOSFET = 1/f

f : Switching Frequency

T : Total time period.

V_f : Forward diode drop

V_R : Reverse recovery voltage of diode

Vgate : Amplitude of gate signal

ΔV : Output voltage Ripple

ΔI : Output Current Ripple

Vout : Output Voltage

Iout : Output Current

Dmin : minimum duty cycle

L4 : Demagnetizing inductance

C6 : Demagnetizing Capacitance

4.3Power Losses.

Various power losses in hard switching converters are

- Switching and conduction losses associated with control switch
- Diode losses i.e. conduction and reverse recovery losses
- Copper loss and core losses related to magnetic materials
- Output capacitor ESR losses

Power losses associated with switches are

Conduction losses of MOSFET is given in equation (1)

$$P_{Cond(Q1-Q4)} = R_{ds(on)} * I^2_{Q(rms)} \qquad (1)$$

Where $Rds_{(on)}$ is On State Resistance of MOSFET

$I^2_{Q\ (rms)}$ is RMS value of pulsed current waveform and calculated using equation (2)

$$I_{Q\ (rms)} = I_{pft} * (t_{on}/T)^{1/2} \qquad (2)$$

Conduction losses of body diode is given in equation (3)

$$P_{cond(D1-D4)} = V_f * I_{D(av)} \qquad (3)$$

Where $I_{D\ (av)}$ is Average value of pulsed current waveform and calculated using equation (4)

$$I_{D\ (av)} = I_{pft} * t'_{on/T} \qquad (4)$$

Reverse recovery losses of diode = $f * V_R * Q_f$

Q_f is calculated from reverse recovery voltage (Q_{rr}), peak reverse current (I_{rm}), and slope of the current during turn-off $_{using}$ equation (5)

$$Q_f = Q_{rr} - I^2_{rm}/ (2di_f/dt) \qquad (5)$$

Q_f varies between 0.5 and 0.25 of the Q_{rr}.

Gate losses of MOSFET = $_f * Q_{gate} * V_{gate}$

Where Q_{gate} are specified in the manufacturer data sheets [7]

4.4 Output Filter [2]

Output voltage ripple depends on output capacitor value and ESR. Capacitor value is determined using equation (6)

$$C = \frac{K*Iout}{8*f*\Delta V} \qquad (6)$$

Output current ripple depends on inductor value and determined using equation (7).

$$L = \frac{Vout*(1-Dmin)*T}{2*K*Iout} \qquad (7)$$

Where k is ripple factor given by equation (8)

$$K = \frac{\Delta I}{2*Iout} \qquad (8)$$

4.5 PWM controller

PWM signals can be generated in two methods [2]

-Voltage mode control

-current mode control

Current mode control has more advantage over voltage mode control:

-Fast response time

-high gain bandwidth product

-current limit foe each pulse can be achieved.

To implement current mode control LT8310 IC is used. It has additional features [8]

-Short circuit overcurrent protection

-Programmable soft switch and operating frequency

-Programmable OVLO and UVLO with hysteresis

-low shutdown current <1uA

4.6 Programming the Switching Frequency:

Switching frequency varies between 100 KHz and 500 KHz to optimize efficiency and performance. Higher frequency operation yield smaller component size, but increase switching losses and gate driving current. It also decreases magnetizing current, which reduces the minimum load requirement under duty cycle mode control. Switching frequency can be adjustable by connecting resister between R_T pin and ground. Value of Resistor determined using equation (9)

$$\mathrm{Rt} = \frac{1000\,\mathrm{KHz}}{\mathrm{fsw}} * 10\mathrm{K} = \frac{\mathrm{tsw}}{1\mathrm{us}} * 10\mathrm{K} \qquad (9)$$

4.7 Programmable current sense:

System is protected from excessive load current by controlling primary side switch current. When V_{SENSE} exceeds 125mV (nom), the system shutdown and attempts a restart after a slow wake-up period. Primary side switch current is given by equation (10)

$$\mathrm{Iswitch} = \frac{\mathrm{IL1}}{\mathrm{Np/Ns}} + \mathrm{Iu,p} \qquad (10)$$

Where Np/Ns is transformer turn ratio, determined using ratio of output voltage to the input voltage.

Resister Rsense connected between the SENSE and GND pin converts the switch current to voltage. Rsense that accounts for the minimum sense threshold is given by equation (11)

$$Rsense \leq \frac{115mV}{1.4 * Iswitch(max)} \qquad (11)$$

115mV accounts for 10% less than maximum load current.

4.8 Programmable soft-start:

It has inbuilt soft-start to reduce inrush current spike and output voltage overshoot at startup. Startup time is programmed by connecting capacitor from SS pin to GND. Css is calculated using equation (12)

$$Css = 50nF * \frac{tss\,[ms]}{1ms} \qquad (12)$$

4.9 Auxiliary Inductance

Current in the magnetizing inductor should cross zero before starting of next period to avoid saturation of transformer core. Auxiliary inductance dependent on peak current of magnetizing inductor. Inductor value is calculated using equation (13)

$$Ld = \frac{Vin(\min)}{2 * Ip} ton \qquad (13)$$

V.SIMULATION

Simulation diagram for both conventional Forward converter and resonant reset forward converter with Synchronous rectifier is shown in Fig.5 and Fig.6 respectively. Simulation is carried in LT Spice simulation application. Component and their values are listed Table.1. Simulation is carried for 7m seconds. Operating frequency is set to 248.7 KHz using Rt. Startup delay of 2m seconds is provided using Css to avoid overvoltage in startup.

TABLE.1: COMPONENT VALUE

Component	Auxiliary winding FC	Resonant Reset FC
Inductor, Lx	33uH	33uH
Output Capacitor, Cx	200uF	200uF
Load, Ro	5ohm	5ohm
Demagnetizing inductor, Lm	420uH	-
Resonant Capacitor, Cs		100pF

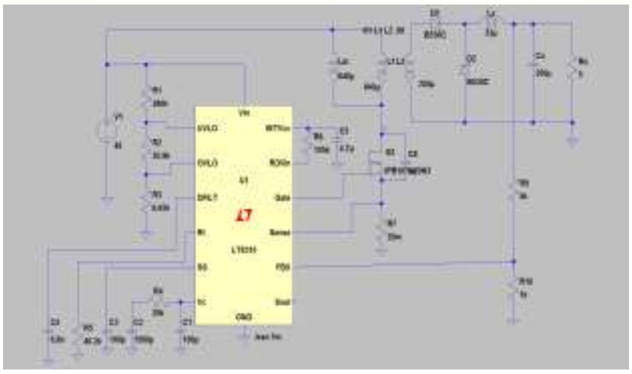

Fig.5: Simulation of Auxiliary Winding Forward Converter

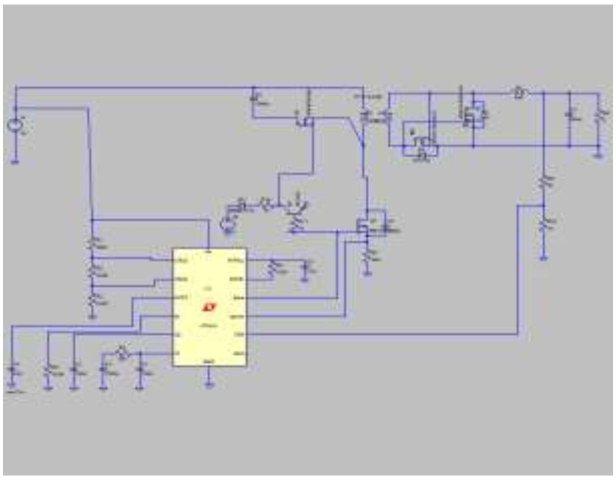

Fig.6: Simulation of Resonant Reset Forward Converter

Load regulation is computed in Table.2 and Table.3.

Fig. 7, 8, 10 and 11 Shows current waveforms of Transformer for auxiliary winding reset Forward converter and resonant reset Forward converter respectively. Fig.9 and 12 shows MOSFET Drain-source Voltage waveform for auxiliary winding reset Forward converter and resonant reset Forward converter respectively.

Table.2: Load Regulation for Auxiliary winding reset FC

Vin in Volts	Output Voltage in Volts		Load Regulation
	No load	Full load	
35	9.6002	9.601	-0.0083
40	9.6	9.6015	-0.0156
55	9.6009	9.6012	-0.0031

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Table.3: Load Regulation for Auxiliary winding reset FC

Vin in Volts	Output Voltage in Volts		Load Regulation
	No load	Full load	
35	9.6003	9.6005	-0.00208
40	9.6005	9.6011	-0.00625
55	9.6004	9.6007	-0.00312

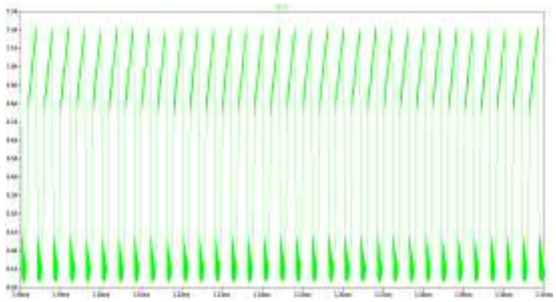

Fig.7: Primary side Transformer Current Waveforms

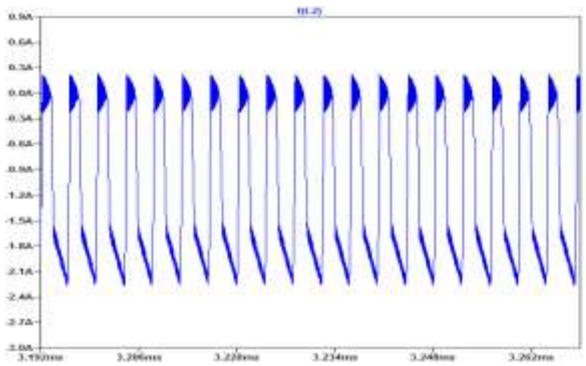

Fig.8: Secondary side Transformer Current Waveforms

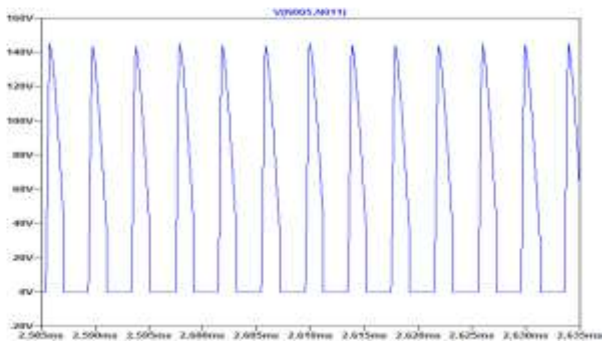

Fig.9: MOSFET Drain-Source Voltage Waveforms

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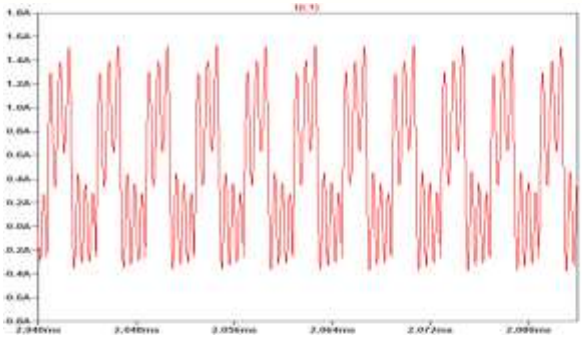

Fig.10: Primary side Transformer Current Waveforms

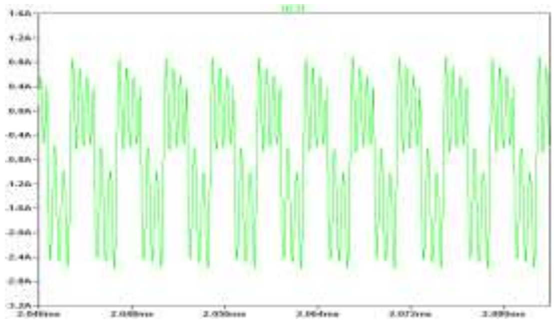

Fig.11: Secondary side Transformer Current Waveforms

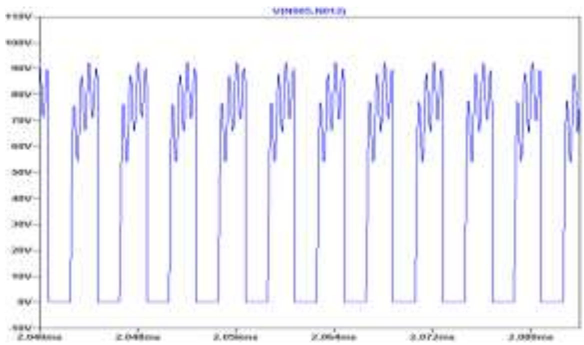

Fig.12: MOSFET Drain-Source Voltage Waveforms

VI.CONCLUSION

This paper compare two reset techniques for Forward Converter. MOSFET Voltage Stress is higher for Auxiliary winding reset Forward Converter, around 142V whereas MOSFET Voltage stress for resonant reset FW is around 95V. Output Voltage Ripple for both converter are less than 1%. Since peak current of secondary side transformer for Resonant reset is clamped to higher value, this method is used for high current applications.

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