Design and Realization of a 12W Quasi-Resonant Flyback Converter

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Abstract – **In this paper the design of a quasi-resonant flyback converter is presented. The operating principles are briefly explained, including the functional block diagram. The prototype has been built and experimental results are presented to support the theoretical analysis and to demonstrate the converter performance.**

Keywords – **CCM, DCM, Flyback, EMI, Quasi-resonant**

I. INTRODUCTION

The most widely used isolated switch mode power supply topology is the flyback converter. Power levels are in low to mid power range, from 1 W to 50 W or even 100 W for low current outputs. The single magnetic component is actually not a transformer but a coupled inductor that combines functions of energy storage, energy transfer and isolation. The main advantages are wide input voltage range, with multiple outputs, higher or lower than the input voltage and simplicity. Depending on the design of the coupled inductor the flyback converter can operate either in Continuous Conduction Mode (CCM) or Discontinuous Conduction Mode (DCM). However flybacks are not perfect. Major drawbacks are poor magnetics utilization and higher electromagnetic interference as a result of hard switching with high peak currents.

II. QR FLYBACK CONVERTER BASICS

A quasi-resonant converter is a distant cousin of a resonant converter. However, the waveforms are not sinusoidal like at true resonant converter, in fact they have typical flyback shape. Also QR flyback power stage does not seem any different from a classical DCM flyback. The "quasi-resonance" is effective in the dead time after the core is demagnetized. The switching occurs at the valley of the resonant ringing generated by the circuit parasitics. There is no need to add inductors and capacitors because they are already there. Soft switching results in switching loss reduction because turning the MOSFET on at the valley of the resonant ringing lowers the losses associated with the output parasitic capacitance of the MOSFET. Another advantage is a reduction of conducted and radiated electromagnetic interference. Also the jitter that results from searching for valley of the ringing spreads the frequency spectrum reducing electromagnetic interference.

The QR flyback converter must operate in DCM because the core must be completely demagnetized. A specialized controller detects core demagnetization and resonant valley and initiates next turn on cycle. In a conventional flyback the constant frequency oscillator initiates the next turn on cycle, so turn on can be at any point of the resonant ringing. QR controllers rely on the switching waveform for valley detection, by detecting the change in slope or a zero-crossing threshold. Some controllers modulate both the switching frequency and the primary current over the most operating range and the others modulate the switching frequency but maintain a constant peak primary current over the most operating range.

The UCC28600 from Texas Instruments (Fig.1) is QR controller operating in different modes, modulating the peak primary current proportional to load and the switching frequency inversely proportional to load.

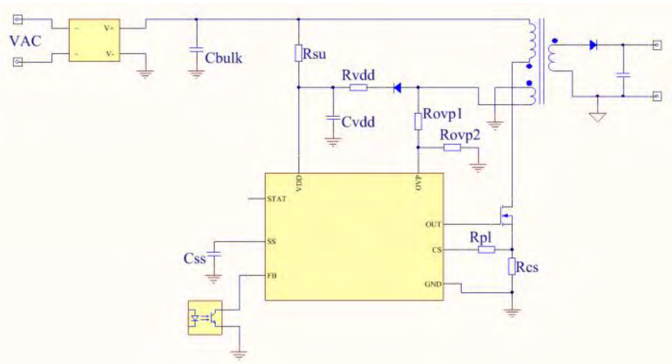

Fig.1. QR flyback converter

For loads below 10 % of rated power the controller will operate in green mode (GM) regulating the output using burst of 40 kHz pulses. The number of pulses increases until the converter is switching consistently at 40 kHz. Frequency fold-back mode (FFM) is reserved for loads between 10% and 30% of rated power. The output is regulated by modulating the frequency from 40 kHz to 130 kHz, holding the peak primary current constant. Finally, for higher loads between 30% and 100% of rated power the controller will operate in either DCM, where the peak primary current is modulated but the switching frequency is clamped to maximum (130 kHz), or quasi-resonant mode (QRM), where the peak primary current and the switching frequency are both modulated.

The above boundaries of steady-state operation are approximate because they are specific design dependent. They are programed by the flyback transformer and the four resistors R_{PL}, R_{OVP1}, R_{OVP2} and R_{CS} on the Fig.1.

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III. Design and Analysis

The task is to design a 12 W QR flyback converter using the UCC28600 controller with careful choice of operating parameters and components. The footprint size must be around 80x40mm. Design specifications are given in Table I.

TABLE I
DESIGN SPECIFICATIONS

		Min	Typ	Max	
Input voltage	V_{IN}	160	220	264	V_{AC}
Output voltage 1	V_O		+12		V
Output power	P		12		W
Full load efficiency	η		80		%
Output ripple voltage	V_{RIPPLE}		≤120		mV
Switching frequency	f	40		130	kHz

The transformer (inductor) must be sized at the minimum input voltage, maximum power and between frequency clamps (40 - 130 kHz). Knowing that, a good choice of core for the transformer is E20/10/6, N87 material from TDK. Reinforced insulation is a must, but the effectiveness of which must be verified by dielectric strength testing.

Using a switching frequency of 80 kHz we can now calculate the maximum primary inductance using the following equation

$$L_{P\max} = \left(\frac{V_{DC\min}(V_O + V_F)N \times 0.925T}{V_{DC\min} + N(V_O + V_F)} \right)^2 \left(\frac{f}{2P_{IN\max}} \right) \quad (1)$$

We will use 650V MOSFET so the reflected voltage must be around 100V. If we adopt 110V, we can easily calculate the primary to secondary turn ratio using equation

$$N = \frac{N_P}{N_S} = \frac{V_R}{V_O + V_F} \quad (2)$$

In order to prevent the saturation of the core we must obtain inductance factor to be $A_L = 200\text{nH/T}^2$. For that value we need air-gap in center leg to be g = 0.2 mm. Because the grinding of the center leg is problematic, we will place non-conducting spacers between core halves. The spacer thickness should be half the value used for the center leg gap. In our case we have s = 0.1 mm.

Now we can calculate the number of primary turns using equation

$$N_P = \sqrt{\frac{L_P}{A_L}} \quad (3)$$

Skin depth in mm is given by the following equation

$$\delta = \frac{76}{\sqrt{f}} \quad (4)$$

As a result the wire diameter must be

$$d < 2\delta \quad (5)$$

At the end the number of secondary turns is given by

$$N_S = \frac{N_P}{N} \quad (6)$$

The results are given in Table II.

TABLE II
BASIC PARAMETERS

		---	Typ	---	
Max. primary inductance	L_P		1.2		mH
Number of prim. turns	N_P		78		
Number of sec. turns	N_S		9		
Primary peak current	I_{PPEAK}		0.5		A
Primary RMS current	I_{PRMS}		0.2		A
Secondary RMS current	I_{SRMS}		1.6		A
Max. wire diameter	d		0.42		mm

Now it is time to wind the transformer. We will use a single 0.2 mm enamelled copper wire for the primary and a triple-insulated bundle of 7 twisted wires with 0.3mm enamelled copper wires for the secondary, in order to minimize copper losses taking into account the skin effect and to fulfil demands for reinforced insulation. Using triple-insulated wire instead of layers of tape barrier between the windings we will minimize the leakage inductance which must not exceed 4% of the primary inductance. Also to minimize the leakage inductance the bias and secondary winding are interleaved between the primary halves. The bias winding is designed not only to supply the current for the controller, but also to detect the valley of the resonant ringing

Knowing specific core losses we can now calculate the loss in magnetic component (Table III). Total transformer power loss is 0.354 W. Since the thermal resistance of chosen transformer is 46°C/W the temperature rise will be approximately 17°C above ambient temperature. Satisfied with the results, we will keep the chosen core geometry.

TABLE III
TRANSFORMER LOSSES

		---	Typ	---	
Core effect. volume	V_E		1.49		cm^3
Specific core losses	P_V		0.15		W/cm^3
Core loss	P_{CORE}		223		mW
Primary resistance	R_{PRI}		1.73		Ω
Primary loss	P_{PRI}		86		mW
Secondary resistance	R_{SEC}		14		mΩ
Secondary loss	P_{SEC}		46		mW

For the output diode we will use Schottky diode despite much higher junction capacitance which can cause ringing with parasitic inductance. The rated voltage of used diode is 100V in order to accommodate the sum of the reflected primary voltage and the output voltage including margin for the leakage inductance spike

IV. Realization

AC/DC converter was built on two layer FR-4 substrate with 35µm copper with a footprint of 80x40mm. The transformer is wounded on through-hole coil former according to the calculations. Output voltage is further filtered out by the added LC filter. All electrolytic capacitors are low ESR organic polymer electrolytic capacitors.

Furthermore we have measured efficiency at various loads at various input voltages. For practical reasons we have used input voltages from 250 to $350V_{DC}$. The results are given in Fig.2. The efficiency is over 84% at full load.

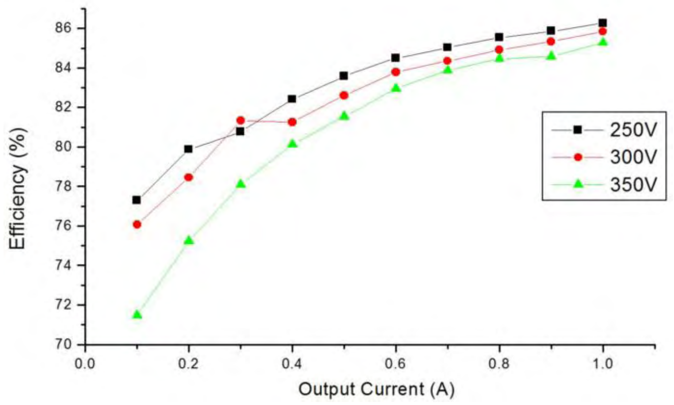

Fig.2. Efficiency as a function of load

The drain waveform of the primary MOSFET in green-mode at 0.1A output load is given in Fig.32. With raising load the converter goes to frequency fold-back mode (Fig.4). At maximum load (1A) the converter operates in quasi-resonant mode (Fig.5). All three waveforms are captured at 300V DC input.

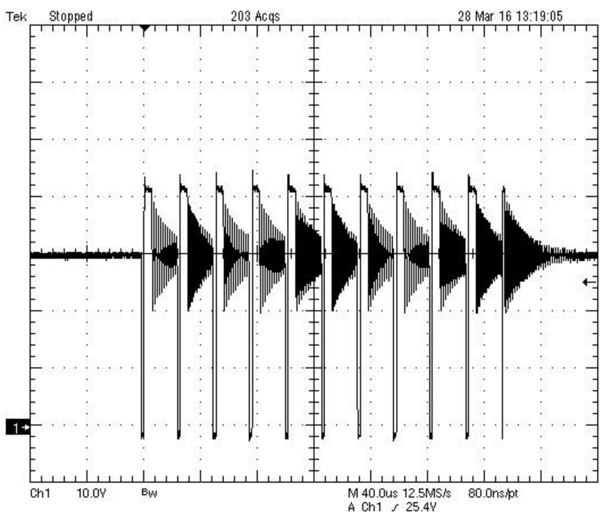

Fig.3. Drain voltage in green-mode (burst)

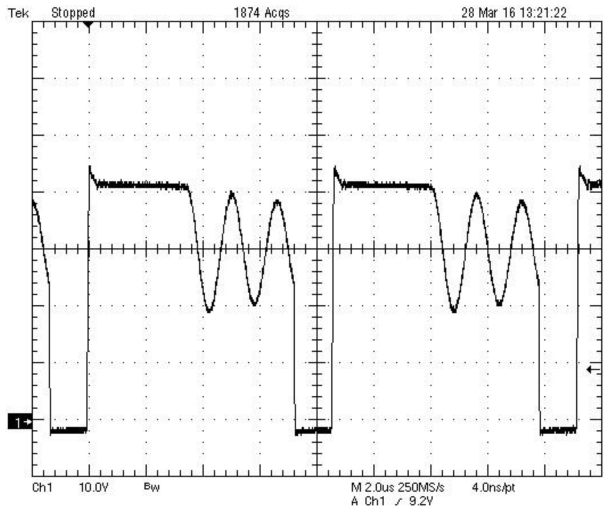

Fig.4. Drain voltage in frequency fold-back mode

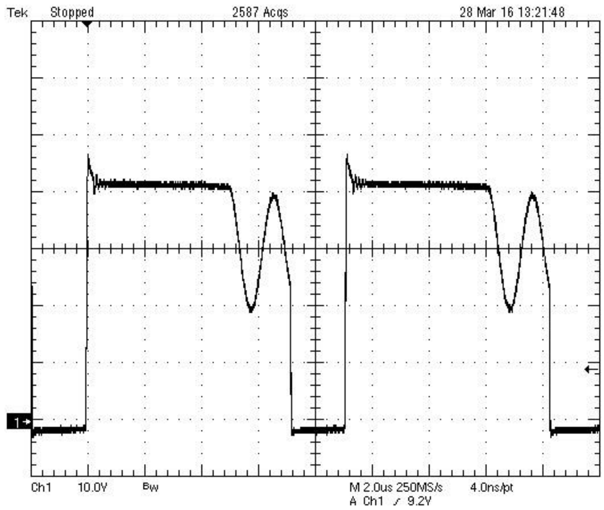

Fig.5. Drain voltage in quasi-resonant mode

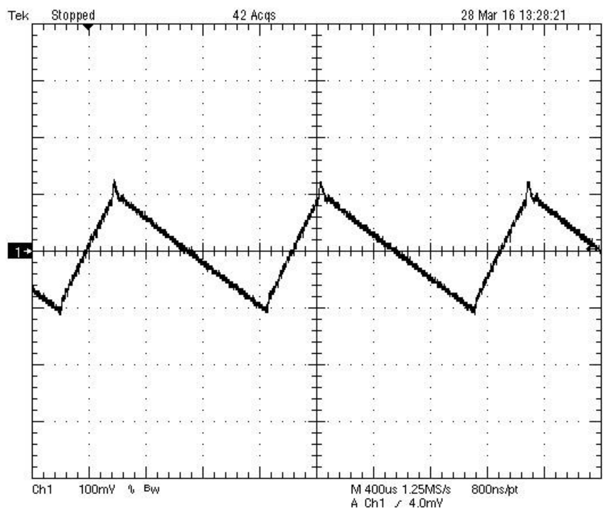

Fig.6. Output ripple in green-mode

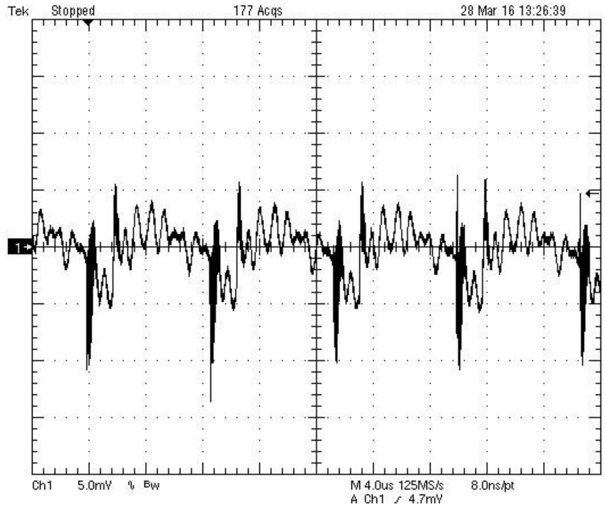

Fig.7. Output ripple in quasi-resonant mode

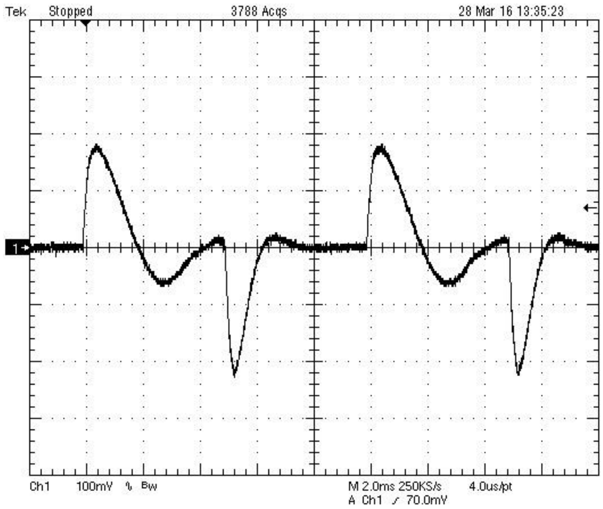

Fig.8. Output voltage load transient response

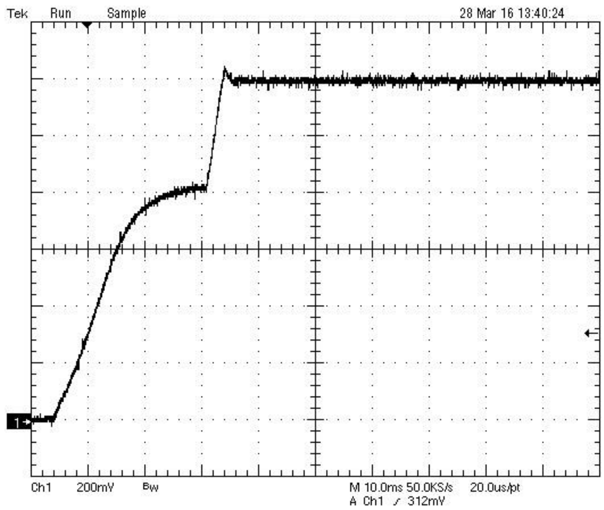

Fig.9. Output voltage rise into 1A load

The output voltage ripple in green-mode is around 230mV$_{pp}$ (Fig.6) with triangular shape. In quasi-resonant mode output ripple has more complex shape (Fig.7), but much lower level. The basic ripple is about 10 mV$_{pp}$ and the spikes are about 20 mV$_{pp}$.

Output voltage load transient response was measured with load current changed between 0.2A and 0.8A at 5ms period and at 0.5A/µs. Fig.8 shows the AC coupled output voltage under this load transient condition. As we can see over/undershoot is under 200mV.

Output voltage rise (Fig.9) was measured into the electronic load with 1A load at 300V input voltage. The rise is monotonic with small 400mV overshoot.

The picture of the converter prototype is given in Fig.10.

Fig.10. Converter prototype

V. Conclusion

In this paper the design and analysis of 12W QR flyback converter are presented. The prototype was built and tested. The results verified that the efficiency between 50% and 100% of load is over 80%. With minor changes we can make different versions of this power supply.

Acknowledgement

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