CONSTANT FREQUENCY, FORWARD CONVERTER WITH RESONANT TRANSITION

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ABSTRACT

A single ended forward converter, operating at constant frequency and switching at zero voltage is presented. By using a secondary switch, the main transformer's core is symmetrically reset and a part of magnetizing energy is used to discharge the parasitic capacitances of the primary switch to zero. Zero voltage switching conditions are achieved over a broad input voltage and output current range. An experimental 500Khz converter, which delivers an output power of 200W at 5V is presented. Operating from an input voltage of 180 to 400 Vdc, the converter exibits an efficiency greater than 88% at full load.

INTRODUCTION

There is continuing increasing pressure in the power supply industry for higher power density and power processing efficiency. Operation at high frequency leads to size and weight reduction of the capacitive and magnetic components, which comprise a large percentage of the total volume. The resulting increase in power density , however, may lead to cooling problems unless high efficiency is maintained, with high operating frequency. Therefore, resonant circuit topologies have been chosen for low switching loss, which is the prime concern at these frequencies. While resonant-mode operation is popular for its low-loss switching at high frequency, it has an inherent disadvantage in its control through frequency modulation. In the case of quasi-resonant topologies, which have become popular in the last several years, the advantage of very low switching losses are balanced by the current or voltage stress on the main switch. As result the efficiency of quasi-resonant converters is comparable or even lower than conventional PWM techniques, with only a slight advantage in power density. Carefully designed and implemented PWM converters [1], can obtain better performance in efficiency and power density, over the quasi-resonant techniques. Recently there is a trend in power conversion development, which is a merge between PWM and quasi-resonant approaches, called constant frequency with resonant transitions. The resonant transition converter combines the best of both by allowing resonant switching while the primary power transfer is through pulse width modulation. The dc-dc converter described in this paper uses the resonant transition switching mechanism to provide zero voltage turn-on of the power MOSFETs. The resonance takes advantage of the parasitic components, such as the output capacitance and internal diode of the MOSFET, as well as the leakage and magnetizing inductance of the isolation transformer. This paper will show one application of this new power processing technique, which is a dc-dc converter delivering 5V at 40A of load current and operating from a 180 to 400 Vdc bus. An efficiency larger than 88% for a switching frequency of 500Khz, has been achieved.

RESONANT CIRCUIT WITH INITIAL CONDITIONS

The base element of the constant frequency with resonant transition forward converter is the "resonant circuit with initial conditions" which is depicted in fig 1. The circuit is composed of a voltage source Vin , an inductive element Lm and a resonant capacitor Cr, which at the initial moment is charged to a voltage Vr. Another initial condition is the current through the inductive element Im, which is directed towards the input source. This resonant circuit is characterized by the following three parameters: Zc, which is the characteristic impedance and is defined by the relationship (1), Ω is the natural frequency defined by (2), and the initial phase ϕ , which particularly characterizes only the resonant circuit with initial conditions (3). Based on these three parameters the behavior of the circuit, starting from the t=0, can be characterized. This is described by the equation (4) and (5). Considering the following example: Lm=140uH, Cr=220pF, Im=.9A, Vr=460v, Vin=270V. This example represents a case of the resonant circuit, employed by the forward converter with resonant transitions. The voltage across the Cr and the current through Lm are presented in Figure 2. The analysis shows that after a certain period of time, in this case approximately 110nS, the voltage across Cr reaches zero. Figure 3 presents a basic structure of an isolated converter, which can be structured as a forward or flyback.

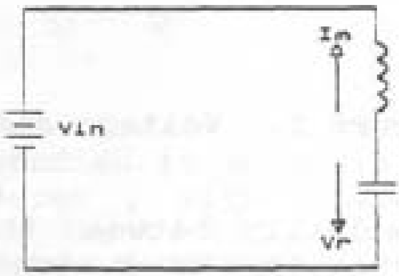

Fig #1

Resonant circuit with initial conditions

$$Zc = \sqrt{\frac{Lm}{Cr}} \quad (1) \qquad \Omega = \frac{1}{\sqrt{Lm \cdot Cr}} \quad (2)$$

$$\phi = \arctan\left[Zc \cdot \frac{I}{Vr - Vin} \right] \quad (3)$$

$$V_t = Vr - (Vr - Vin) \cdot \frac{\cos(\phi) - \cos[\phi + \Omega \cdot t]}{\cos(\phi)} \quad (4)$$

$$I_t = (Vr - Vin) \cdot \frac{\sin[\Omega \cdot t + \phi]}{Zc \cdot \cos(\phi)} \quad (5)$$

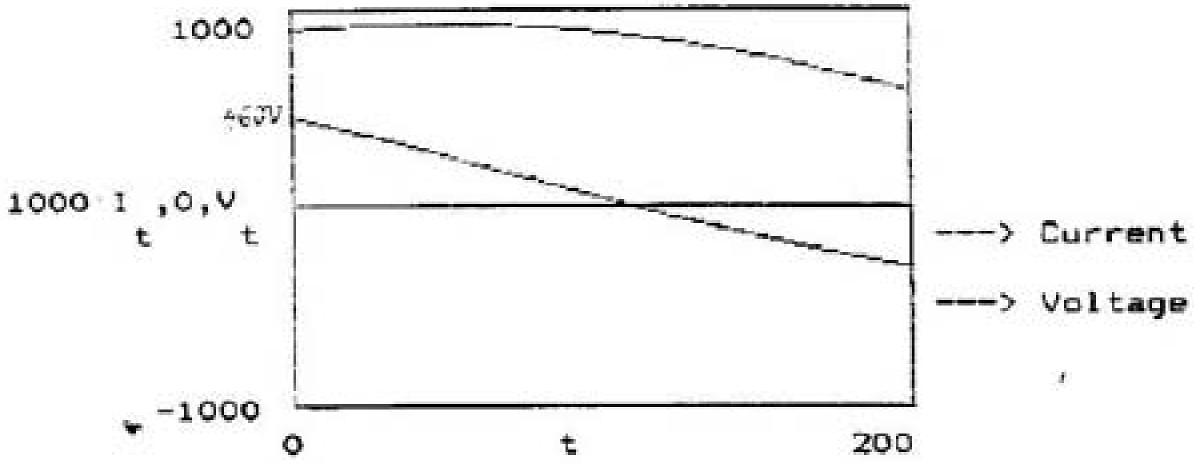

Figure 2. Voltage and current on the resonant capacitor

The similarity between the resonant circuit described above and this power converter structure can be noticed. There is a Vin, an inductive element Lm, which is the magnetizing inductance and a resonant capacitor which is the output capacitance of the switch. At the moment when the secondary side exhibits a large impedance reflected to the primary the circuit oscillates. For example in the case of a discontinuous mode flyback it is the moment when the output rectifier ceases to conduct, and for a forward topology it is after the core resetting cycle. This phenomenon is well known by design engineers, but usually they damp it out, instead of using it to their advantage. For this circuit the zero voltage switching condition can be accomplished only for Vr>2 Vin because there is no current flowing through Lm. Employing this zero voltage conditions , would lead in this case to a modulation in frequency, if regulation targets are pursued.

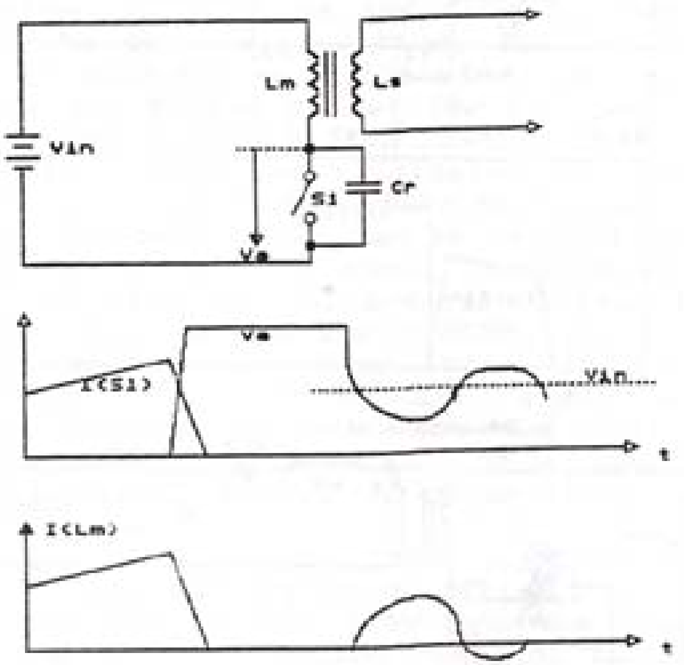

Figure 3. Resonant circuit with initial conditions as part of a power train

Using a secondary switch as is presented in figure 4, zero voltage conditions can be employed over large input - output conditions and operating at constant frequency.

CIRCUIT OPERATION

The power train of the dc-dc converter studied is shown in Figure 4. It is a single ended forward converter , with a secondary reset switch Q2, which has two purposes. One of its function is to offer a complete and symmetrical reset of the transformer's core. A second role is to steer the magnetizing current at the end of reset cycle towards the source. This creates the initial conditions for the resonant circuit described in previous chapter. The reset capacitor Cr, automatically charges after a few cycles to a voltage level which insure a complete and symmetrical reset of the core. The reset current is symmetrical to zero, and any asymmetry, will lead to an adequate change of the voltage across Cr, which will force back the symmetry. This resetting technique, of using a secondary reset switch is presented in Figure 4. During Q1 conduction, the magnetizing current builds up in the magnetizing inductance. At the moment, when the primary switch Q1 ceases to conduct, the magnetizing current flows through CR1 into reset capacitor Cr.

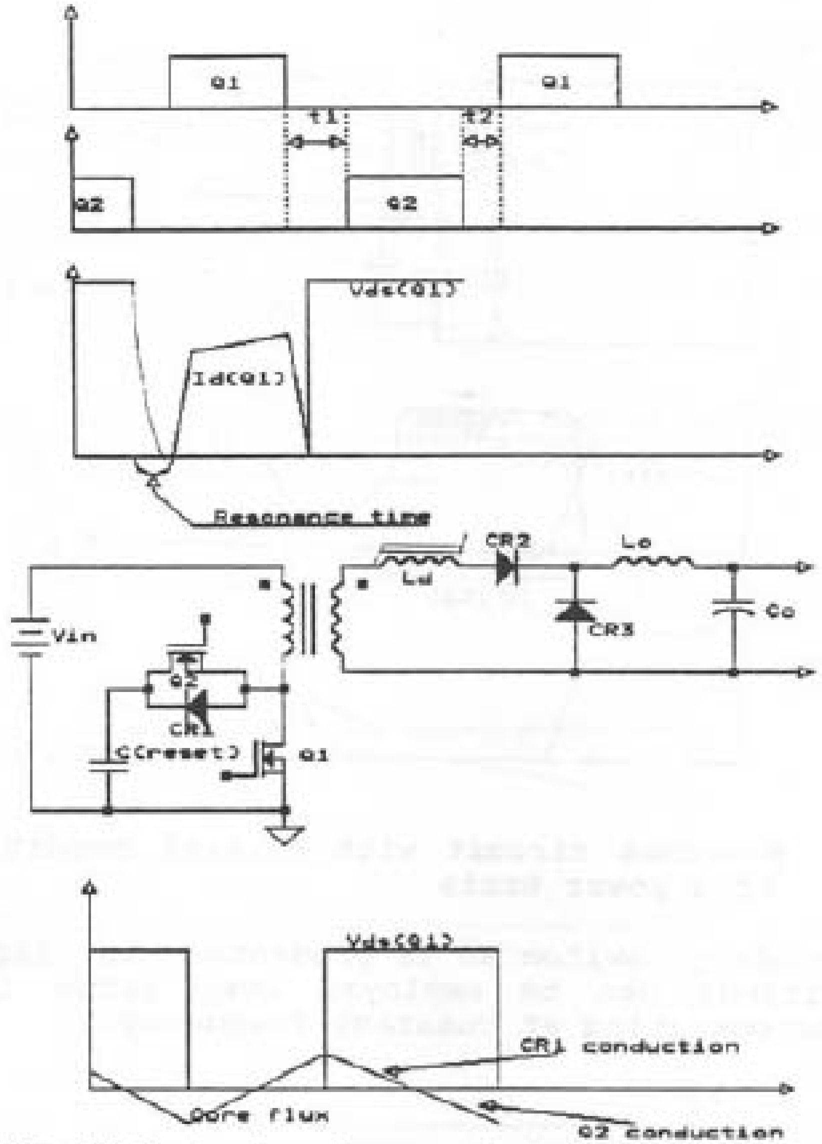

Figure 4. Simplified structure of forward converter with resonant transitions.

It is assumed that the reset capacitor Cr is charged to the steady-state voltage level, which is Vin/(1-D). The magnetizing current is reflected back towards the source during the second half of the OFF time. At the moment when the reset switch Q2, ceases to conduct, the magnetizing current will continue to flow towards the source via the output capacitance of the main switch Q1. The resonant circuit with initial condition described in the previous chapter, is created by the magnetizing inductance, output capacitance of Q1, the input source Vin. The initial conditions are the voltage across the Q1 and the magnetizing current flowing towards the input source. If the power train design is done in accordance with the equation (4 & 5) for zero voltage switching conditions, the magnetizing current flowing towards the source, will discharge the parasite capacitance of the main switch Q1 to zero. The magnetizing current will flow further through the body diode of the main switch Q1. In this way the conditions are created for zero voltage turn on of the main switch.

Similar zero voltage turn ON conditions is experienced for the reset switch, which can turn on any time on the first half of reset cycle when the Cr is conducted. Turning the switches ON under zero voltage conditions, eliminates the turn on switching losses, eliminates the Miller effect for the gate drive circuit, decreases the driving currents, and lowers significantly the EMI level. The turn off losses are eliminated by forcing the gate voltage below the gate threshold, prior to the voltage across the drain starting to build-up. This can be accomplished with a fast turn OFF gate drive circuit. A small capacitance between drain and source will also slow the voltage rise across the MOSFET. The supplementary capacitor across the MOSFET, will not affect turn ON losses due to zero voltage switching conditions.

During the resonant transition to zero voltage across the main switch, when the voltage in secondary becomes positive, the forward rectifier starts to conduct. This shunts the primary resonant circuit, forcing a part of magnetizing current to flow towards the load.

There are two ways to overcome this unwanted phenomenon. One solution is to increase the amplitude of the magnetizing current, to such a level that the current flow into the secondary does not totally deplete the current flowing towards the source, via the parasitic capacitance of the main switch. This approach will increase the conduction losses in the primary side switching elements. The power dissipation saved by eliminating the switching losses, will be partially wasted by increasing the conduction losses.

A second approach is to delay the current flow into the secondary by 30 to 60ns, besides the delay introduced by the leakage inductance. The delay cell can be implemented by using a constant volt-second Mag Amp or by using a synchronized rectifier.

EXPERIMENTAL RESULTS

The power train configuration and the driving circuit is presented in Figure 6. The reset circuit is composed of Cr,Q2 and D5. There is a resistor R3, which acts as a flux sensor. Information about the flux density is supplied to the control circuit. Flux feedback is an important part for reliable operation of the converter. One of the major drawbacks of this power processing technique is the risk of core saturation during transients, until the reset capacitor Cr charges to the proper voltage. Through the flux feedback scheme, information about the core flux is supplied to the control circuit to modify the duty cycle accordingly. The duty cycle has to be modified in a such way that the flux density in the transformer's core should not exceed a set level. The controller receives information about current through the main switch and the output voltage. The goal of this paper is to present the power train topology and the performance characteristics of this power circuit. The control circuit will be the subject of a future presentation.

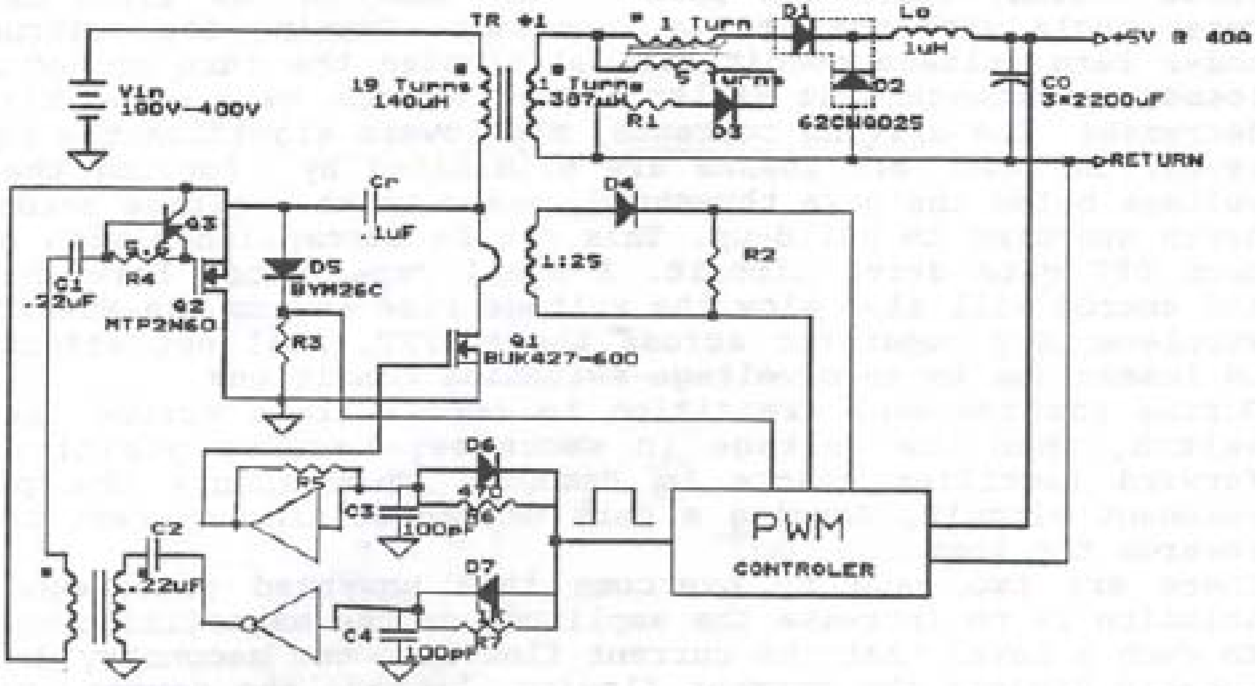

Figure 5. Power train and driving circuit for the constant frequency, resonant transition converter

The secondary side delay presented in the previous chapter is implemented by using a amorphous material toroid (MB 8x7x4.5), with a one turn primary and five turns on the reset winding. The reset circuit consists of D3 and R1. The delay time varies between 30nS to 50nS, as a function of the input voltage. The output choke is implemented by using a low profile core from Magnetics (P-43618-EC) with one turn only. The reset capacitor and D5 creates an effective clamp circuit. The leakage inductance energy is stored in Cr and during the OFF time of the main switch Q1, is recycled back to the source, and a part of it is used to discharge the output capacitance of Q1 and Q2. The main switch Q1 turns ON, after the voltage across it reaches zero or the lowest point. The zero voltage sensor is based on the current flowing through gate-drain capacitance, when the voltage in drain starts to fall. A part of this current flows through R5 and C3, delaying the turn ON of Q1. This circuit is a very effective and low cost zero or lowest voltage detector.

In figure 6,7 and 8 is presented the current and the drain to source voltage of main switch Q1. It is noticeable the large duty cycle operation >50%, at low line. The voltage across Q1 does not vary much over the input voltage range as is depicted in Figure 12. Over the input voltage range of 180V to 400V the duty cycle varied between 62% and 30%, and the drain to source voltage varied between 473V and 554V.

The resonant transitions to zero voltage is suggestive presented in Figure 9 at a scale of 100nS per division. The high efficiency which characterize this topology, is due to the fact that circulating currents, characteristic to the resonant converters are kept to a minimum. As is presented in Figure 10, the resonant current which discharge the parasitic capacitance of the switches is only .8A. Figure 10 and 11 presents the current through reset elements, Q2 and respective D5. The initial spike of current through D5 is due to the leakage inductance of the main transformer. Figure 10 presents the current through controled element of the reset switch. As can be seen, the magnetizing current is returned back to the power source during the second part of reset cycle. For 270V input and 5V @ 40A output, the measured efficiency is 88.2%. The most dissipative elements are the output rectifiers with 18.3W, where 16W is due to a voltage drop of .4V and leakage losses of 2.3W. The second dissipative active element is the main switch Q1 with 2.71W at 90 C, junction temperature. The reset switch Q2 dissipates .46W and the reset diode approximately .25W. The transformer dissipates around 1.8W were the core losses are calculated to be .5W, for a peak flux density of 570 Gauss at 500Khz. The core used was PC50PQ26/25Z-12 from TDK.

Having a single layer of 19 turns in primary and a one turn copper strip in secondary, the proximity losses were kept relatively low. The copper losses calculated using a specialized computer program were 1.47W. The power processing efficiency versus input voltage is plotted in Fig. 13. The efficiency decreases slighty with higher input voltage due to the turn ON losses, switching occcuring near zero at high input voltage.

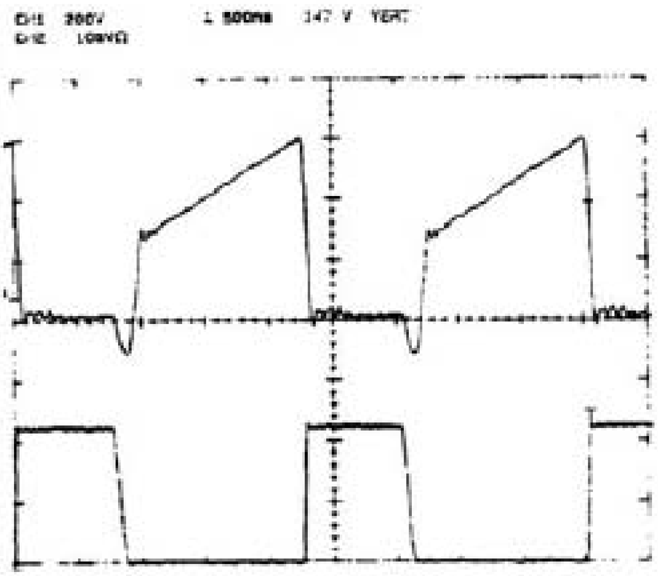

Figure 6. Vin=200V

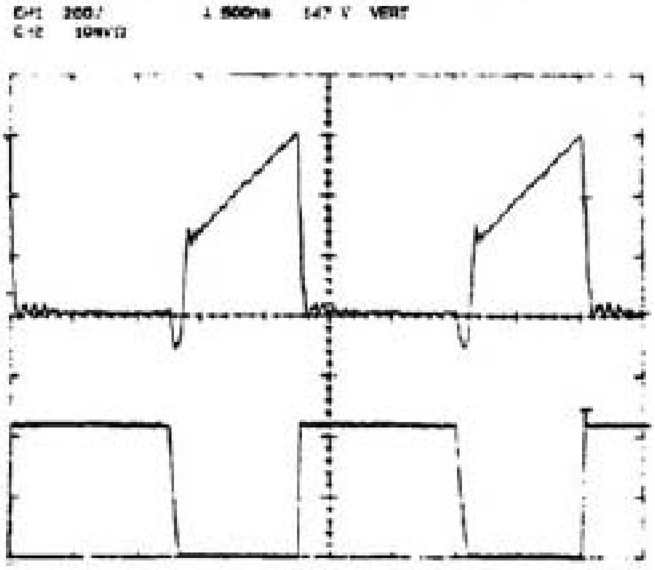

Figure 7. Vin=270V Io=40A

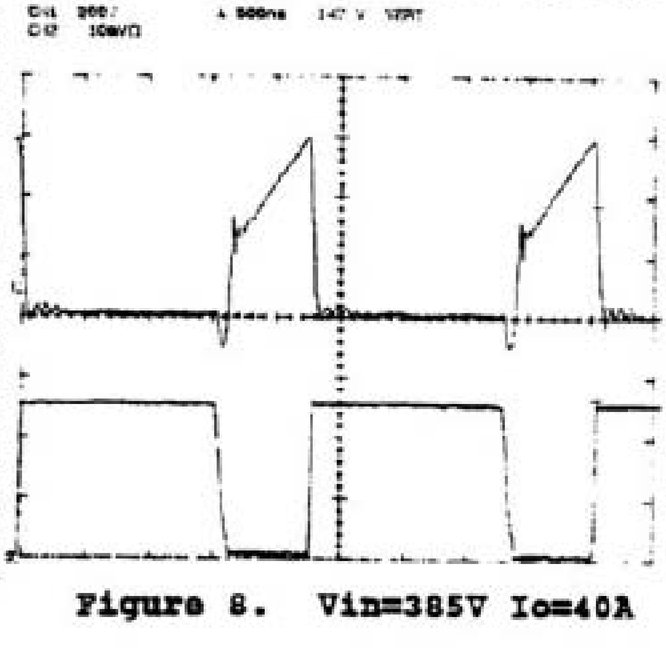

Figure 8. Vin=385V Io=40A

For example for Vin=385V, the turn ON occures at Vds=85V.

A larger current delay in the secondary would lead to a zero voltage turn ON for high line conditions, but would increase the power dissipation in the saturable reactor.

CONCLUSIONS

A constant frequency PWM forward converter with resonant transition has been proposed. The new converter has the following advantages over the conventional PWM forward converter or quasi-resonant topologies:

- Duty cycle can be larger than 50%, which allows a larger input voltage range operation or a much lower voltage stress on the secondary rectifiers.
- The magnetizing energy is not dissipated, being used to discharge the parasite capacitances of the primary switches, and returns to the input source.
- The flux through the core is symmetrical, maximizing the available flux swing.
- It minimizes the voltage stress on the main switch over the input voltage range.
- The primary switches turn on at zero voltage, which eliminates the turn on losses and the Miller effect on the driving circuit, lowering drive current and reducing noise.
- The voltage across the switches and the output rectifiers exhibits lower dV/dT and result in lower recovering currents and EMI.
- There is no need for a snubber across the main switch, which is effectively clamped by the reset capacitor.

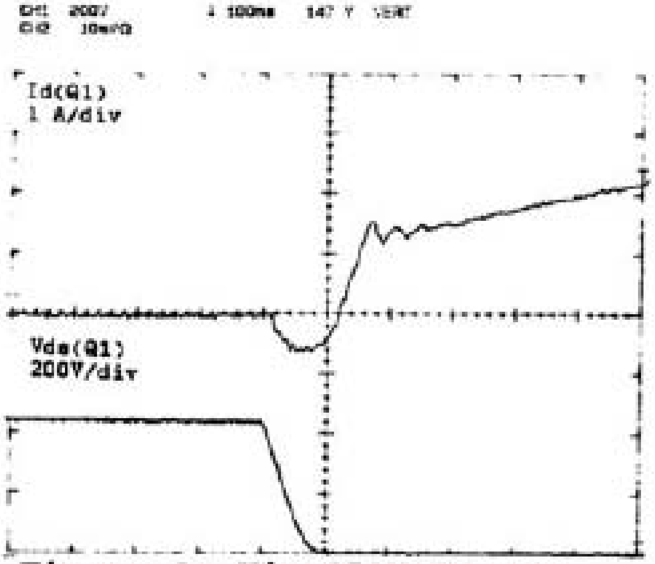

Figure 9. Vin=270V Io=40A

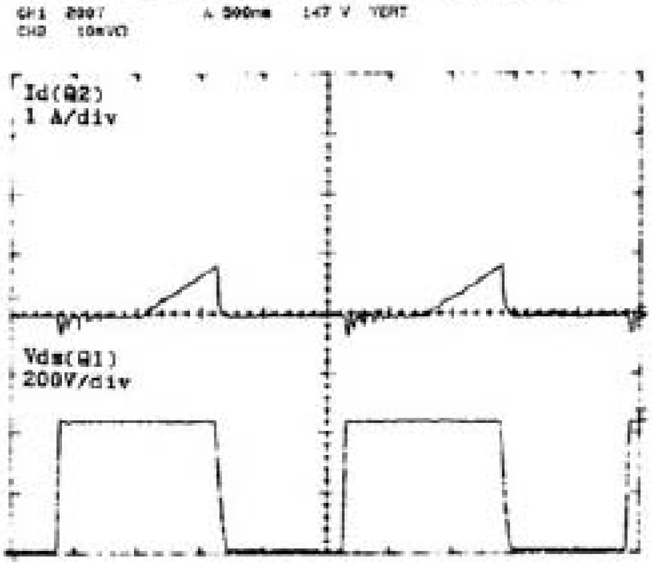

Figure 10. Vin=270V Io=40A

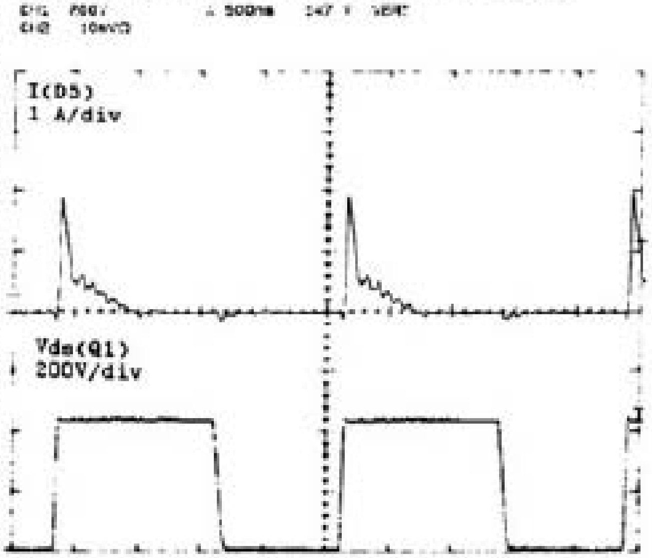

Figure 11. Vin=270V Io=40A

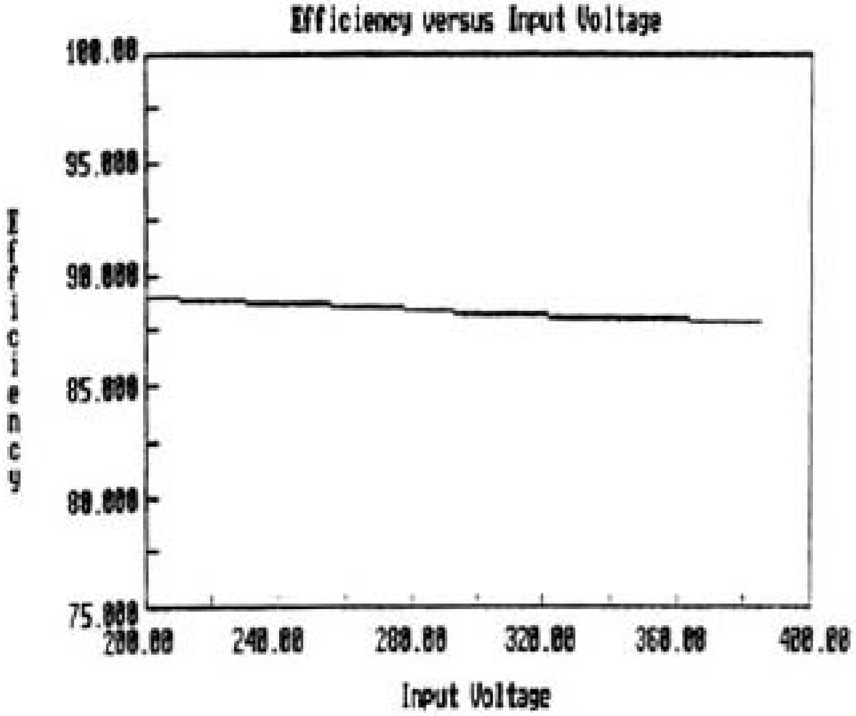

Figure 12.

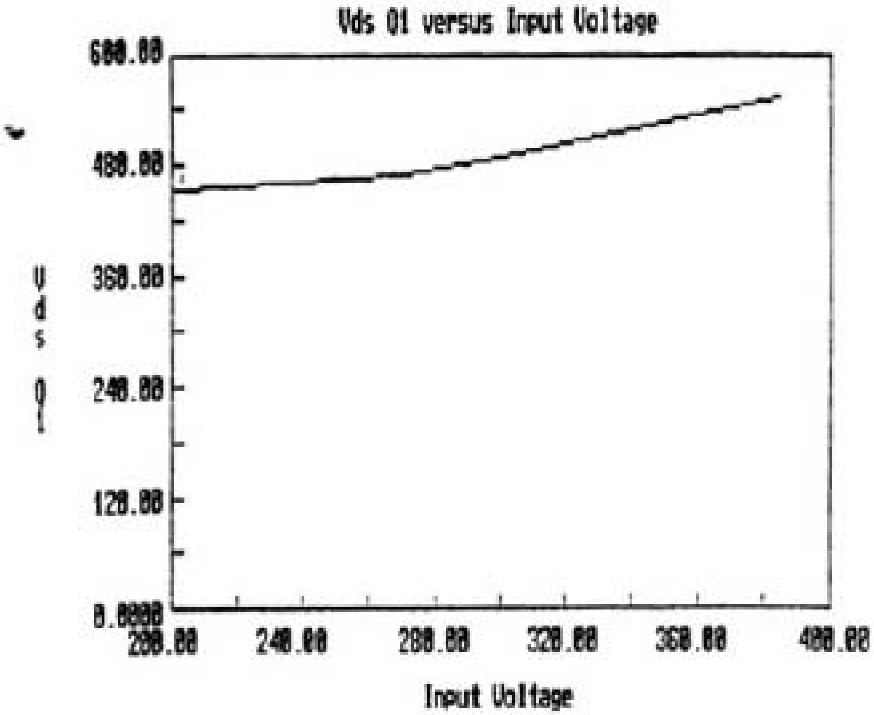

Figure 13.

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