Zeroing Transformer's DC Current in Resonant Converters with No Series Capacitors

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***Abstract* - DC current unbalance in the windings of a transformer may initiate a runaway process and ultimately drive the transformer into saturation. Such a situation could arise in a transformer coupled resonant converter that does not include a series capacitor. This issue was explored in this study theoretically, by simulation and experimentally. A method is proposed for automatically adjusting the transformer's DC current to zero by correcting the asymmetry of the drive. This balancing control is facilitated by a signal obtained from a DC/AC current sensor which senses the transformer's primary and secondary currents such that they cancel each other for the no DC case. The proposed method was tested on a self oscillating, parallel loaded DC-DC resonant converter of 12V input voltage 500V output voltage and 700W power level. The proposed approach allows the design of resonant converters without DC decoupling capacitors which could lead to a significant cost and size reduction.**

***Index Terms* - Resonant power conversion, magnetization current, digital control, transformer saturation.**

I. INTRODUCTION

Resonant converters [1- 4] can be divided into two groups with respect to the series capacitor that is normally placed between the switches and the transformer, those that apply a series resonant capacitor (e.g. LCL) and those that do not (e.g. PRC) [1- 4]. Even so, the usual practice is to place a large series DC blocking capacitor [1-5] to prevent possible saturation of the transformer. In low input voltage, high power applications, the required DC blocking capacitors are not only large and expensive, but in some instances utterly impractical due to the requirement of an ESR of milli-Ohms and below. The objective of this study was to explore the mechanism of transformer saturation in resonant converters with no series capacitors (resonant or DC blocking) and to investigate possible methods that will ensure safe operation without a series capacitor.

II. THE NEED FOR A SERIES CAPACITOR

A. The converter under study

We consider a seemingly "safe" case (with respect to transformer saturation) of a transformer isolated, parallel-loaded resonant converter with a capacitive filter at the output (Fig. 1). The resonant elements are C_R and the transformer's leakage inductance (denoted below as L_R, not shown in Fig. 1). Capacitor C_R represents the transformer's winding capacitance and an additional discrete capacitor (if used). The converter is driven by a full bridge (Q1-Q4) that operates under Zero Current Switching (ZCS) conditions (Fig. 2). This is assured by synchronizing the switching frequency to the zero crossing of the primary current (Fig. 2). In this exemplified case, the resonant current is fed to a digital circuit (Fig. 1) which turns off the active switches at the current cross-over instance, and after a proper dead-time, turns on the complimentary bridge diagonal pair.

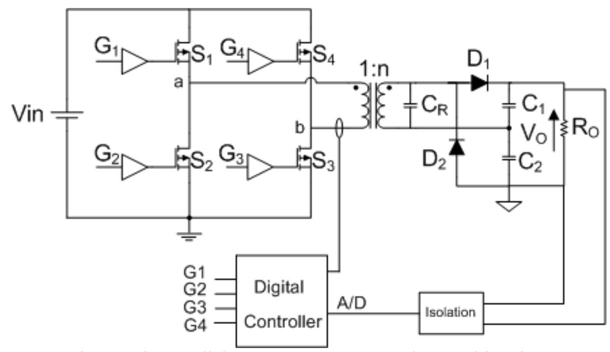

Fig. 1. The parallel resonant converter under consideration.

Parallel loaded resonant converters with capacitive output filter have been studied earlier, but normally for the general case of variable frequency control and for different operation conditions [1,2,4,7]. In the converter studied here, the reflected voltage Vrefl, is always higher than the input voltage. This facilitates ZCS operation by forcing a natural decrease of the inductor current when the output diodes are conducting (ILr, Fig. 2). Power control can be achieved by dithering. Details of this control approach are beyond the scope of this paper.

Under the above assumptions of ZCS synchronization, a reflected voltage that is always higher than the input voltage and lossless devices, the key parameters of the converter introduced in Fig. 1 can be described by the following equations:

$$t_R = \frac{1}{\omega_r}\cos^{-1}\left(\frac{1-M}{1+M}\right) \tag{1}$$

$$I_{Ll} = \frac{Vin}{z_r}\cdot 2\sqrt{M} \tag{2}$$

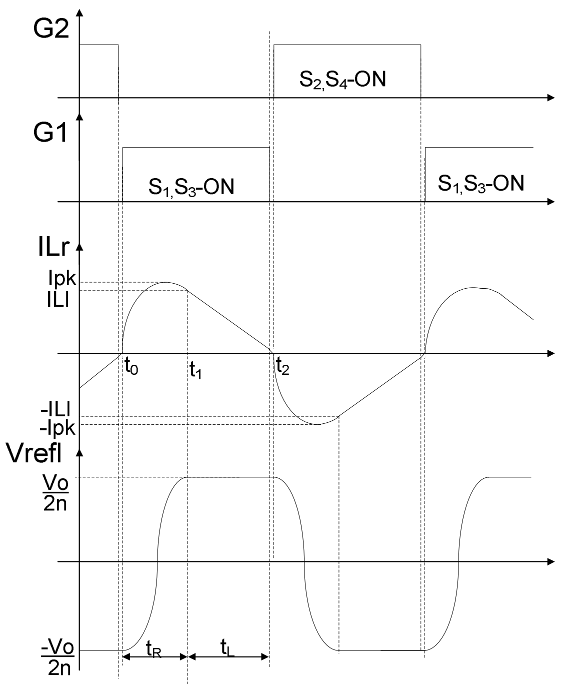

Fig. 2. Basic waveforms of the resonant converter of Fig.1. I_{Lr} is the transformer's primary current, V_{refl} is the reflected voltage to transformers primary side.

$$t_L = \frac{L_R \cdot I_{Ll}}{(M-1)\cdot Vin} \tag{3}$$

$$\frac{Vo^2}{R_O}\cdot(t_R + t_L) = \frac{{I_{Ll}}^2 \cdot L_R}{2} + \frac{I_{Ll}\cdot Vin \cdot t_L}{2} \tag{4}$$

where :

$$M = \frac{\mathrm{Vo}}{2\mathrm{n}\cdot\mathrm{Vin}} \quad ; \quad z_r = \sqrt{\frac{L_R}{n^2 \cdot C_R}} \quad ; \quad \omega_r = \frac{1}{n\sqrt{\cdot L_R \cdot C_R}} \quad ;$$

n is the turns ratio of the transformer and R_O is the load resistor. Other notations are consistent with Fig. 2.

This implicit set of equations is nonlinear, and since an analytic solution is not available, it needs to be solved by some numerical tools for specific private cases. It was found though that the voltage transfer ratio can be approximated by M_a:

$$M_a = (R_n)^{\frac{2}{3}} \tag{5}$$

Where: $R_n = \dfrac{R_O}{4\cdot n^2 \cdot z_r}$ is the normalized characteristic impedance.

The approximation (5) is obtained by assuming that areas marked '1' and '2' in Fig. 3 are equal. These areas are exactly the same when ILl=Ipk (Fig. 2). In all other cases, area '1' will be smaller than area '2' which will result in:

$$M_a \le M \tag{6}$$

That is, approximation (5) underestimates the voltage transfer ratio. The deviation of M_a from M depends on the operation point and in particular on the value of R_n. It should be noted that the lower boundary of R_n required for maintaining the reflected voltage higher than the input voltage is:

$$R_n > 1 \tag{7}$$

Comparing the approximate transfer function M_a to the exact one M (Fig. 4) one finds that the error increases for large values of R_n. It was found though, that the ratio M can be fitted to a slightly different power expressions (8), yielding a much lower error (Power fitting, Fig. 4):

$$M_{\text{fitting}} = 0.93\cdot(R_n)^{0.87} \tag{8}$$

Additional relationships can be used in the design stage of the converter. By some manipulation of (2-4) one finds that ZCS operation is maintained when the converter locks to the switching frequency:

$$f_s = \frac{\omega_r \cdot (M-1)}{4\cdot R_n} \tag{9}$$

Another approximate expression which can help in the design of the converter is the value of the primary RMS current. Due to the non trivial wave shape of the primary current, a closed form analytical solution for the crest factor is unavailable. Evidently, however, the crest factor lies

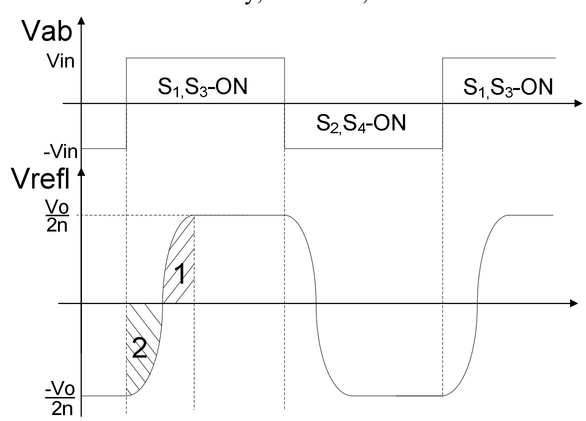

Fig. 3. Assumptions for equation (5)

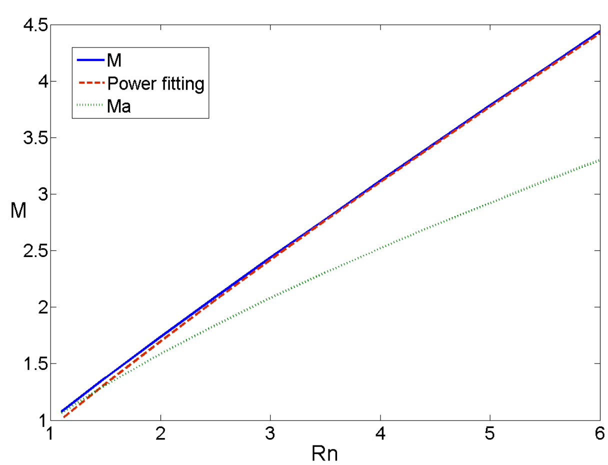

Fig. 4. Voltage transfer function M, approximate expression M_a and fitted curve"Power fitting"

between the sine wave crest factor, $\sqrt{2}$, and the triangular wave crest factor, $\sqrt{3}$. Consequently, a good approximation could be the average value of the two:

$$CF_R \approx \frac{\sqrt{2}+\sqrt{3}}{2} \tag{10}$$

Consequently, the approximate expression for the input rms current (I_{RMS}) will be:

$$I_{RMS} = \frac{I_{PK}}{CF_R} \tag{11}$$

where:

$$I_{PK} = \frac{(Vin + \frac{Vo}{2n})}{z_r} \tag{12}$$

B. Transformer saturation study

Even if one assumes that the zero crossing detection is ideal, ZCS by itself, will not protect the circuit against transformer saturation. This is due to the fact that the transformer's primary current includes both the transferred current component I_{T1} and the magnetization current I_{LM}, i.e.

$$I_{Lr} = I_{T1} + I_{LM} \tag{13}$$

Hence, commutating the input bridge when I_{Lr} is zero does not assure a balanced operation, that is, a zero DC input voltage to the transformer. This is illustrated in Fig. 5 in which the magnetization current is of positive polarity while commutation is kept at zero total current. The latter will cause asymmetry in the bridge output, increasing the positive voltage duration (Fig. 5). This asymmetry in transformer voltage constitutes, in fact, a positive feedback that will increase the magnetization current in the next cycle. This, in turn, will increase further the asymmetry of the bridge voltage, pushing eventually the transformer into saturation in the case of low parasitic resistances. This DC drift mechanism was further studied by simulation.

The simulation was carried out by PSIM (Powersim Inc., USA) [6] on the model of Fig. 6 that emulates the converter under study (Fig. 1). To verify that the fixed step simulation of PSIM does not introduce extra errors, the simulation was repeated in the ORCAD environment, yielding the same

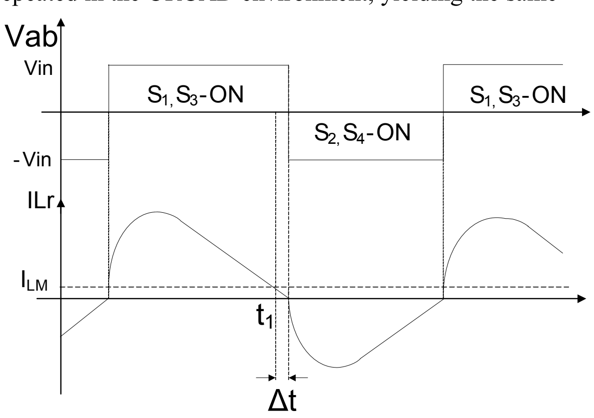

Fig. 5. Transformer primary current with none zero magnetization current.

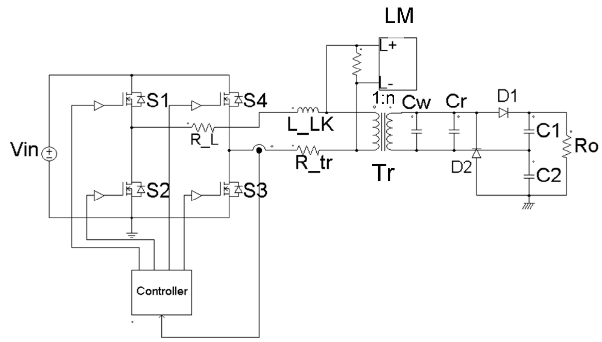

Fig. 6. PSIM simulation model used to explore the behavior of the resonant converter of Fig.1.

results. The four switches of the bridge, S1-S4 include an "on resistance" of 5mΩ each. The resistances R_L and R_tr introduce losses of the input bus (copper losses) and the transformer respectively. The inductance L_LK and capacitance Cw represent the leakage inductance and inter-winding capacitance of the transformer, respectively. It is assumed that the leakage inductance and the winding capacitance are independent of the transformer current. The capacitance Cr+Cw is the resonant capacitor, which along with inductance L_LK (assumed to be smaller than the magnetization inductance LM) constitute the resonant network. The output voltage doubler is assumed to be ideal. The 'Controller' block was realized in the simulation environment as an embedded C code block [6]. It senses the resonant current and commutates the bridge exactly at the current zero crossing point (ZCS). The block LM represents the magnetization inductance of the transformer and is implemented by a behavioral model of a non-linear inductor [7] depicted in Fig. 7. In this model, the inductance is a function of the current passing through the inductor. The expression of the non linearity parameter K was obtained by fitting a 'tanh' function to the data of the manufacturer (3F3, Ferroxcube). The definitions of the dependent sources are as follows:

$$\mathrm{G} \equiv \mathrm{Isc} = \mathrm{Ipr} \tag{14}$$

$$\mathrm{E} \equiv \frac{\mathrm{Vpr}}{\mathrm{K}} = \mathrm{Vsc} \tag{15}$$

$$\mathrm{L} = 1\mathrm{H} \tag{16}$$

$$\mathrm{K} = 81.3\times10^{-6} \cdot \tanh((0.443 \cdot \mathrm{Ipr})^{-1.6}) \tag{17}$$

The simulation results (Fig. 8) imply that ZCS by itself is not sufficient to ensure proper transformer operation. Fig. 8 shows a constant drift of the magnetization current (upper trace) and accumulation of 'Volt-seconds' on the primary of the transformer (third trace from top, INT(Vrefl)).

An additional phenomenon which can be related to the magnetization current buildup is the asymmetry of the transformer total current (second trace from top). The non linear nature of magnetization current makes the phenomena even worse. It can thus be concluded that even a perfect ZCS can not prevent transformer saturation when a series capacitor is not used. In the physical converter, additional asymmetry can be caused by errors in the zero current detection which may further increase the drifting of the transformer into saturation.

C. Practical limitations of the use of series capacitors

The practical problem of applying series capacitors in low input voltage high power converters stems from the difficulty of finding a commercial capacitor with a very low ESR to keep the capacitor's dissipation to a reasonable value. This is illustrated by considering the experimental resonant converter with a nominal input voltage of 12V and nominal power of 700W. It is estimated that the rms value of the transformer's primary current will reach 80A. This implies that if a DC blocking capacitors is used, its ESR should be around 1mΩ to keep the capacitor losses to about 6W. The capacitance value of such a capacitor should be 400-500uF, in order to keep a voltage drop under 1-1.5V and to not interfere with the resonant process.

Table 1 displays the characteristics of some commercial film capacitors designed for power applications from different vendors. It demonstrates the fact that low ESR can be obtained only by parallel connections of several capacitors since off the shelf single capacitors are not appropriate for the task. Parallel connection will increase the volume (and cost) and will require meticulous layout to insure even current distribution. For higher rms currents, say 200A and above, application of DC blocking capacitors in resonant converters is seemingly unrealistic.

III. PROPOSED APPROACH

A. Measuring the magnetization current

A cardinal requirement for any system that will automatically correct for the transformer's DC component is the availability of a signal that is a function of the DC component of the magnetization inductance. Ideally, the signal should replicate the flux density state of the transformer's core. This conceivably could by accomplished by Hall effect sensors or other magnetic field sensors (e.g. magneto-resistors). This, however, will not

Table 1. CAPACITORS COMPARISON

Manufacturer	A	B	C	D
Capacitors	6x75uF	3x170uF	1x350uF	1x400uF
Cap Volume	6x100cm^3	3x607cm^3	91.03cm^3	607cm^3
ESR/capacitor	2.9mΩ	6mΩ	1.14mΩ	5mΩ
Total ESR	0.48mΩ	2mΩ	1.14mΩ	5mΩ
Power dissipation per capacitor	0.5W	4.26W	7.3W	32W

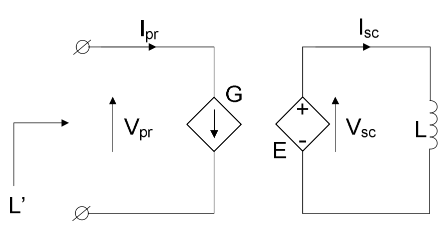

Fig. 7. Behavioral model of a non-linear inductor used in PSIM simulation. See text for definition of dependent sources.

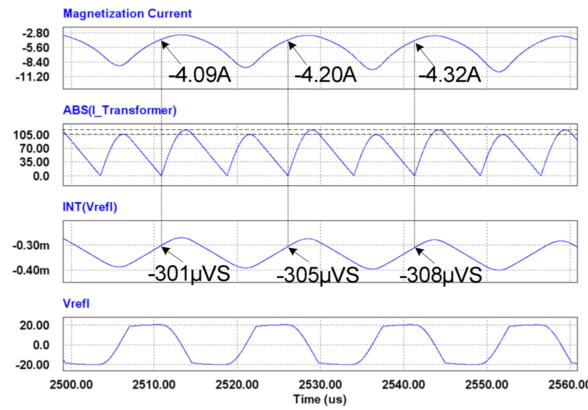

Fig. 8. Simulation results of the resonant converter of Fig. 6. Upper trace – transformer's magnetization current, second trace – rectified transformer primary current, third trace – 'Volt-seconds' of transformer's primary voltage, bottom trace – transfomer's primary voltage.

only be complex and expensive, but will require a redesign of the physical construction of the transformer with possible ill effects on its performance. The approach adopted in this study is based on external measurements around the transformer.

The transformer DC current detection method explored in this study is based on external measurements of the transformer's primary and secondary currents. By doing so, the transformer is left as is, as there is no need to modify it in any way. The proposed detection method is illustrated in Fig. 9. The detector is a DC/AC current sensor that is based on a Hall effect element or some other technology. The sensor includes two windings that carry the transformer's primary and secondary currents. The turns ratio of the sensor's windings corresponds to that of the transformer that is being measured. For an ideal transformer (infinitely large magnetization inductance and zero winding capacitance), the output signal of the current sensor will be zero since the primary and secondary Amper-turns cancel each other. In the case of a real transformer, the output signal of the current sensor may comprise both DC and AC elements while the former is proportional to the unbalanced magnetization current and the latter is due to the parasitic elements: L_M, C_W. This AC component will introduce ripple on the DC components of interest.

B. Ripple cancellation

The major components of the AC part include the AC current of the transformer's magnetization current (real and imaginary) and the reactive current due to the inter-winding capacitance.

As was noted before, the AC component of the magnetization current is an unwanted component of the measurement since it introduces a ripple on the useful signal. The "brute force" solution to the ripple current is to filter the signal by some filtering techniques. This solution is simple and intuitive but will limit the control bandwidth. Another possible solution for attenuating the ripple is to null the AC current seen at the input winding by a parallel resonant network. That is, to cancel the effect of the lagging magnetization-inductance current by a leading capacitance

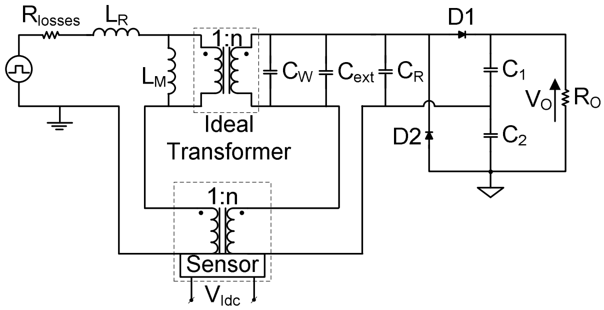

Fig. 9. Proposed method for measuring transformer's DC current.

current. This approach is explained by Fig. 10 which shows the original parallel elements involved: the magnetization inductance, L_M and inter-winding capacitance, C_W which is reflected to the primary side (C'_W) for the sake of simplicity.

As is, there is already some cancellation of the reactive currents $I_{C'W}$ and I_{LM}. Better cancellation can be achieved by tuning C'_W, and L_M to the switching frequency f_s. It was found that in practical cases this can be achieved by adding an external capacitance C'_{EXT} in parallel to C'_W:

$$C'_{EXT} = \frac{1}{4 \cdot \pi^2 \cdot f_s^2 \cdot L_M} - C'_W \tag{18}$$

To make the value of the required capacitance smaller, the extra capacitor can be placed at the secondary side of the transformer, yielding:

$$C_{EXT} = \frac{C'_{EXT}}{n^2} \tag{19}$$

Although perfect cancellation is not possible to achieve, the AC component can be considerably attenuated. This will significantly simplify the filtering requirements and improve the signal to noise ratio. It should be noted that for the tuned case both $I_{C'W}$ and I_{LM} are neutralized and the converter's resonant frequency is practically unaffected by them.

The availability of a signal that is proportional to the transformer's DC current makes it possible to balance out the DC by one of several possible control methods. For example, by injecting a counter DC current into an auxiliary winding.

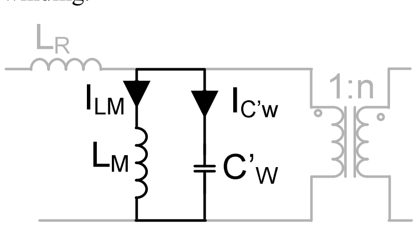

Fig. 10. Components that cause ripple current when primary and secondary currents are summed. Interwinding capacitance (C_W) is reflected to primary (C'_W).

The approach adopted in this study was to introduce some asymmetry in the bridge switching. This was achieved by introducing an offset to the primary current signal such that it will shift the commutation instance of one of the bridge legs.

IV. Experimental Setup and Results

The proposed DC correction method was investigated on a low input voltage, medium power, resonant converters with no series capacitors that was controlled by a dsPIC30F2020 microcontroller (MicroChip USA). The relevant DC cancellation circuitry is depicted in Fig. 11. V_{Idc} is the transformer's unbalance signal obtained by the method described in Fig. 9. The function of the control circuit of Fig. 11 is to shift the commutation instance of one leg of the bridge while keeping the commutation instance of the other leg intact. This is achieved by adding a fraction of the error voltage (V_{Idc}) to the signal of the "resonant current measurement" of the relevant phase. When the combined voltage is compared to the reference voltage (Ref), the modified cross over point will now lead or lag (depending on the polarity of V_{Idc}) as compared to the original resonant current signal. By this, an asymmetry in the bridge commutation is created which constitute in fact a feedback loop that will suppress the transformer's DC current to zero (assuming of course a correct offset voltage polarity).

The operation of the experimental inverter (Table 2) with no DC balancing (Fig. 12) shows an unbalanced DC current of about 1.55A. Slight asymmetry in the input current can also be seen (Fig. 13). A direct measurement of the DC voltage across the transformer's primary gave the value of about 6mV which is consistent with a primary resistance of 4mΩ. Applying the proposed DC balancing control circuitry that adjusts the switching instances of one leg, the transformer's DC component was significantly reduced back to an acceptable lower value of 240mA (Fig. 14). Under the DC closed loop condition, the transformer current became almost symmetrical, (Fig. 15) and the DC voltage across the transformer's primary dropped to about 0.8mV.

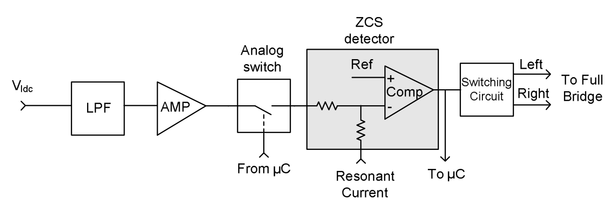

Fig. 11. Experimental circuit.

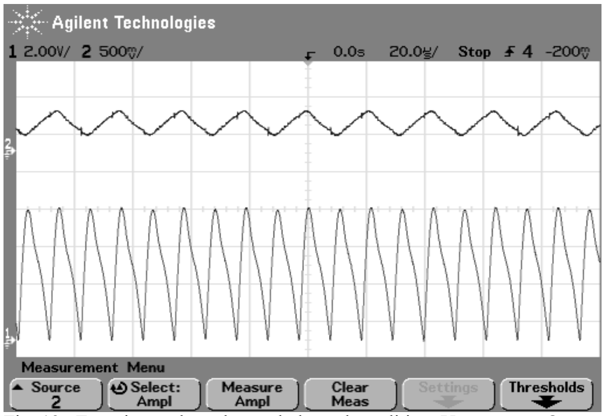

Fig. 12 Experimental results - unbalanced condition. Upper trace: Output of DC sensor, 2.27A/div; Lower trace: Rectified transformer's primary current, 20A/div, Horizontal scale 20us/div.

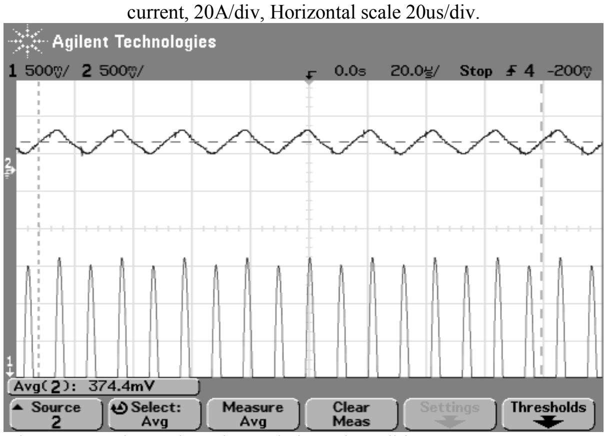

Fig. 13 Experimental results - unbalanced condition. Upper trace: Output of DC sensor, 2.27A/div; Lower trace: Rectified transformer's primary current (zoom of the corresponding trace in Fig. 12), 5A/div. Horizontal scale 20us/div.

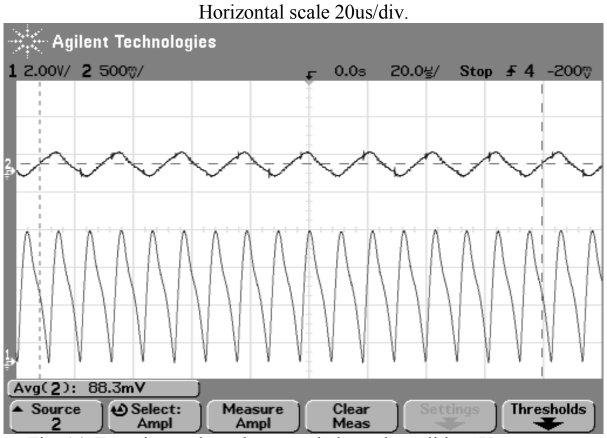

Fig. 14 Experimental results - near balanced condition. Upper trace: Output of DC sensor, 2.27A/div; Lower trace: Rectified transformer's primary current. Horizontal scale 20us/div.

V. Discussion and Conclusions

The present study explored the mechanism of DC unbalance in a transformer coupled resonant converter without a series capacitor. It was found that a runaway situation may arise and drive the transformer into saturation if the DC current persists. Due to the fact that the transformer's primary current includes both the resonant current and the magnetization current, zero current switching (ZCS) by itself cannot assure that the transformer will not go into saturation. Furthermore, once a DC unbalance is created, the ZCS operation will, in fact, introduce positive feedback that will cause a runaway situation. To overcome the problem created by the unavailability of very high current DC coupling capacitors, a method is proposed for automatically adjusting the transformer's DC current to zero. This balancing control is driven by a signal obtained from a DC/AC current sensor which senses the transformer's primary and secondary currents such that they cancel each other for the no DC case. It as found that the ripple of this sensed signal can be significantly reduced by creating a parallel resonant network around the magnetization inductance of the transformer, that is tuned to the switching frequency. The proposed DC balancing control scheme applies the error signal obtained by the DC/AC current sensor, to modify the commutation instance of one of the bridge legs such that it will introduce DC offset in the transformer's drive to correct the original unbalance.

The method was simulated by a PSIM model and tested on a digitally controlled 12V input 500V output converter at a power level of about 700W. The simulation and experimental results verify the viability of the proposed active transformer DC balancing approach.

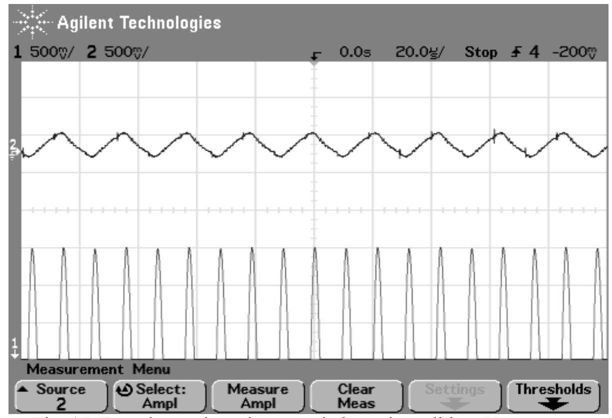

Fig. 15 Experimental results - near balanced condition. Upper trace: Output of DC sensor, 2.27A/div; Lower trace: Rectified transformer's primary current, 5A/div. No assymetry detected. Horizontal scale 20us/div.

Table 2. Key Power Stage Components.

Component	Value
L_R	0.7 µH
L_M	81 µH
C_w	150pF
C_{ext}	1.6nF
C_R	76nF
D_1,D_2	MUR460
C_1,C_2	1 µF
S_1,S_2,S_3,S_4	IRFP3077
R_O	265Ω
core	E65, 3F3
n	9
Current sensor	LA50P

The proposed approach allows the design of resonant converters without DC decoupling capacitors which could lead to a significant cost and size reduction in low input voltage high power DC-DC converters.

VI. Acknowledgment

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