# Application of Integrated Magnetics in Resonant Converters

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***Abstract* -The physical size of inductors in resonant converters was examined in terms of inductor to transformer size ratio. Following the general discussion of the issue, the inductor size in series resonant DC-DC converter is examined in detail. It is shown that the inductor to transformer size ratio is a function of the peak voltage of the inductor to the rms voltage of the transformer and is independent of the inductance. The physical size of the inductor could be much larger than the transformer when the converter operates near resonance. Away from resonance, the inductor size becomes close to, and even smaller, than the transformer. Leakage inductance of the transformer can be used as the resonant inductance but if the leakage is associated with the secondary windings of the transformer, it will increase the peak reverse voltage of the rectifier diodes when a center-tapped rectifier topology is used. To solve this problem, an integrated transformer-inductor element with high primary leakage inductance and low secondary leakage is proposed and tested. The experimental unit was built for a power level of 1KVA and switching frequency range 30 - 60 kHz.**

## I. INTRODUCTION

The trend toward high power density in power electronics drives designers to look for high switching frequency solutions. Resonant and resonant transition inverters and converters are deemed to be more efficient at high switching frequency which in turn can reduce the size of the magnetics and filter capacitors [1]. However, resonant topologies require extra reactive components which diminish the return of the high switching frequency design. It is often argued that the extra components that need to be added are of small physical size and consequently the penalty is not that severe. Furthermore, since the required resonant inductor will normally be of small inductance value, the possibility of applying parasitic component, such as leakage inductance of the transformer, is also frequently brought up [2].

The objective of this study was to examine the issue of the expected physical size of inductors in resonant converters. We begin by developing the methodology for estimating the inductor to transformer size ratio. The method is then applied to examine in depth a more specific case: resonant inductors for series resonant converters. We then propose an integrated magnetics element that combines the inductor and transformer in one structure [3-6]. The element was built and tested experimentally. The measurements support the theory and suggest that an integrated magnetics structure is a viable engineering solution.

## II. THE INDUCTOR TO TRANSFORMER SIZE RATIO

The size of an inductor and of a transformer is a function of the core area product , $A_p$ [7]:

$$A_p = W_a A_c \tag{1}$$

where $W_a$ is the available window area and $A_c$ is the effective cross-sectional area.

The area product of an inductor($A_{p\ L}$) built around a ferromagnetic core with no DC flux can be shown to be:

$$A_{p\ L} = \frac{L I_{L\ rms} I_{L\ pk}}{B_m k_u J} \tag{2}$$

where L is the inductance, $I_{L\ rms}$ and $I_{L\ pk}$ are the rms current and the peak current of the inductor, $B_m$ is the maximum flux density, $k_u$ is the window utilization factor and J is the current density.

Similarly, the area product of a transformer core ($A_{p\ Tr}$ ) can be shown to be:

$$A_{p\ Tr} = \frac{I_{Tr1\ rms} \int_0^{\frac{Ts}{2}} v_{Tr1} dt}{B_m k_u J} \tag{3}$$

where $v_{Tr1}$ is the instant primary voltage, t is time, $T_s$ is the switching period and $I_{Tr1\ rms}$ is the primary rms current.

Assuming similar operation conditions for the inductor and transformer (i.e. same $B_m$, $k_u$ and J) the ratio between $A_{p\ L}$ and $A_{p\ Tr}$ can be found from (2) and (3):

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$$\frac{A_{p\ L}}{A_{p\ Tr}} = \frac{L I_{L\ rms} I_{L\ pk}}{I_{Tr1\ rms} \int_0^{\frac{Ts}{2}} v_{Tr1} dt} \qquad (4)$$

This general relationship is now used to analyze the case of the Series Resonant Converter (SRC) with capacitor output filter. The main assumptions are: switches, diodes and the transformer are ideal; the capacitance $C_o$ of the output filter is infinitely large and therefore the output voltage $V_o$ is constant. Under these assumptions the inductor current of the converter consists of ideal sinusoidal segments of the resonant frequency:

$$f_r = \frac{1}{2\pi\sqrt{LC}} \qquad (5)$$

where C is the resonant capacitance. The peak inductor voltage $V_{L\ pk}$ is related to the peak inductor current ($I_{L\ pk}$) by the simple relationship:

$$V_{L\ pk} = 2\pi\ f_r\ L\ I_{L\ pk} \qquad (6)$$

The primary transformer voltage is equal to the reflected output voltage $V_o'$:

$$v_{Tr1} = V_o' \qquad (7)$$

Therefore, in Continuous Current Mode (CCM) [8]:

$$\int_0^{\frac{Ts}{2}} v_{Tr1} dt = V_o' \frac{T_s}{2} = V_o' \frac{1}{2f_s} \qquad (8)$$

where $f_s$ is the switching frequency.

For Discontinuous Current Mode (DCM) [8]:

$$\int_0^{\frac{Ts}{2}} v_{Tr1} dt = V_o' \frac{1}{f_r} \qquad (9)$$

In SRC, the primary transformer current is also the resonant inductor current and therefore:

$$I_{Tr1\ rms} = I_{L\ rms} \qquad (10)$$

Inserting (8) - (10) into (4) and taking into account (6) we obtain:

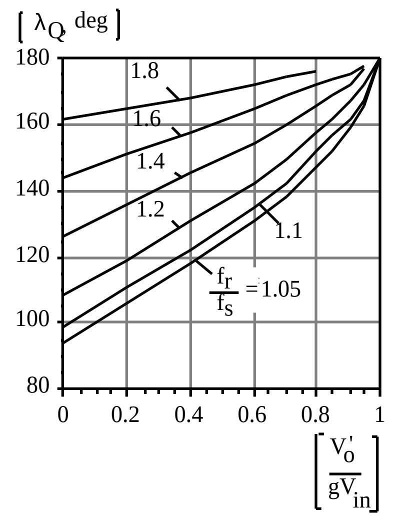
$$\frac{A_{p\ L}}{A_{p\ Tr}} = \frac{V_{L\ pk}}{V_o'} \frac{f_s}{\pi f_r} \qquad \text{(CCM)} \qquad (11)$$

$$\frac{A_{p\ L}}{A_{p\ Tr}} = \frac{V_{L\ pk}}{V_o'} \frac{1}{2\pi} \qquad \text{(DCM)} \qquad (12)$$

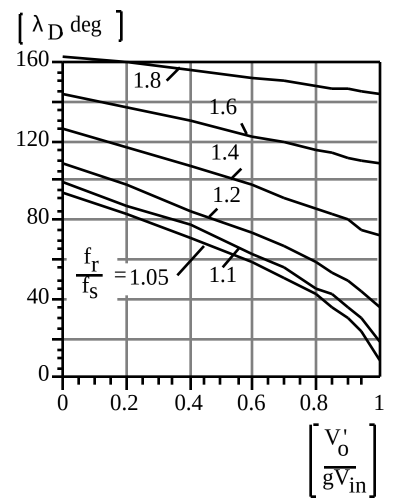
The ratio $\frac{V_{Lpk}}{V_o'}$ can be found from the equation [8]:

$$\frac{V_{Lpk}}{V_o'} = \left(\frac{gV_{in}}{V_o'} + 1\right) \frac{2}{(1-\cos\lambda_Q)} \frac{1}{\left(1 - \frac{1}{\tan(\frac{\lambda_Q}{2})\tan(\frac{\lambda_D}{2})}\right)} \qquad (13)$$

where $V_{in}$ is the input voltage, g is topology constant (g=1 for full-bridge and g=0.5 for half bridge); $\lambda_Q$ and $\lambda_D$ are transistors' and antiparallel diodes' conduction angles referred to the resonant frequency $f_r$. Plots illustrating $\lambda_Q$ and $\lambda_D$ as functions of DC transfer ratio and frequency ratio in CCM are given on Fig. 1 [8]. In DCM $\lambda_Q = \lambda_D = \pi$.




(a)




(b)

Fig. 1. Transistors' (a) and antiparallel diodes' (b) conduction angles as functions of the DC transfer ratio for various frequency ratios (SRC in CCM).

The plots $\frac{A_{pL}}{A_{pTr}}$ as functions of the output to input voltage ratio $\frac{V_o'}{gV_{in}}$ with the frequency ratio $\frac{f_r}{f_s}$ as a parameter calculated by (11) - (13) are given in Fig. 2. This presentation points out to the fact that the physical size of the inductor is independent of the inductance. That is, the size is a function of the operating conditions of the converter. The physical size of the inductor will be very large (as compared to the size of the transformer) when the switching frequency approaches the resonant frequency and/or when the output to input voltage ratio is small. The size of the inductor can be made smaller than that of the transformer when the series DC-DC converter operates far from resonance ($\frac{f_r}{f_s}$ > 1.6). Near resonance the peak of the inductor voltage is many times higher than the reflected output voltage $V_o'$ and therefore the necessary size of the inductor will be also much larger. For optimal design of the inductor, the DC voltage transfer ratio $\frac{V_o'}{gV_{in}}$ should be as large as possible since the size of the inductor increases when $\frac{V_o'}{gV_{in}}$ is small. Fortunately, this coincides with low conduction losses as detailed in [8].

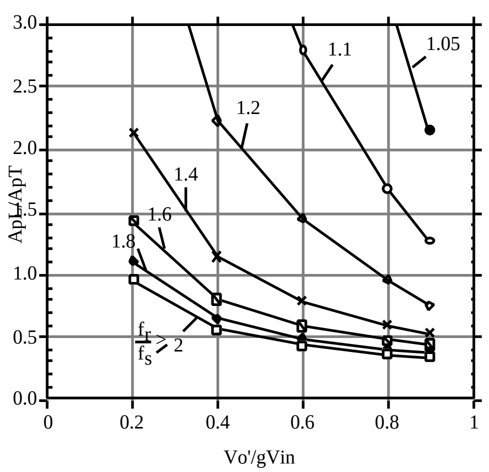



Fig. 2. The inductor to transformer area product ratio ($\frac{A_{pL}}{A_{pTr}}$) in SRC as functions of the DC voltage transfer ratio ($\frac{V_o'}{gV_{in}}$) with the frequency ratio ($\frac{f_r}{f_s}$) as a parameter.

## III. "GOOD" AND "BAD" LEAKAGE OF TRANSFORMER WINDINGS

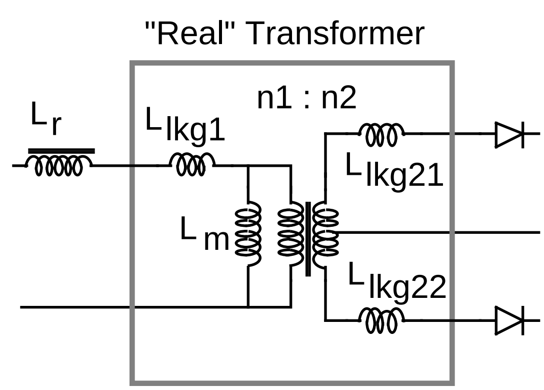
Leakage inductance of the transformer windings affect the operation of the converter (Fig. 3). In SRC, the leakage inductance of transformer windings is in series with the resonant inductor. Consequently, when leakage is present, the inductance of the resonant inductor can be decreased

proportionally. However, when the output rectifier has a center-tapped topology (Fig. 3a), the leakage inductance of the secondary windings $L_{lk2}$ will increase the peak reverse voltage on the rectifier diodes $V_{D\ pk}$:

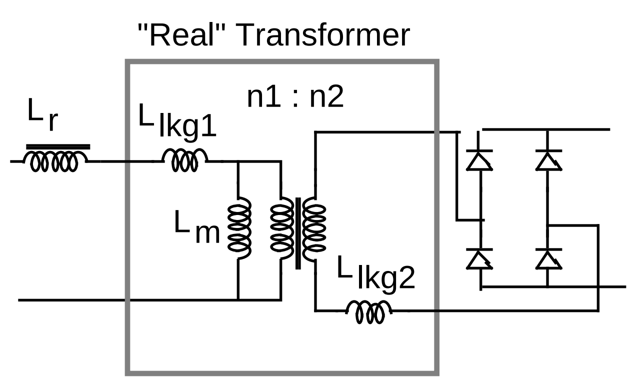
$$V_{D\ pk} = 2V_o + L_{lk2}\left(\frac{di_{Tr2}}{dt}\right)_{max} \tag{14}$$

where $i_{Tr2}$ is the instant secondary transformer current. Therefore, for an optimal design one should keep the leakage inductance of the secondary transformer winding as low as possible.

Simulation results of steady-state processes in SRC with center-tapped rectifier for two cases are given on Fig. 4. In the two cases shown, the total leakage inductance is the same. In Fig. 4a all the leakage is assumed to be in the primary. In Fig. 4b all the leakage is assumed to be in the secondaries. It is evident that the latter case is much worse. The peak reverse voltage of the rectifier diodes $V_{D\ pk}$ is much higher (120 V compared to 80 V) in presence of leakage in the secondary. It should be noted that extra $V_{D\ pk}$ stresses are expected when the diodes turn off due to the recovery of the reverse currents.



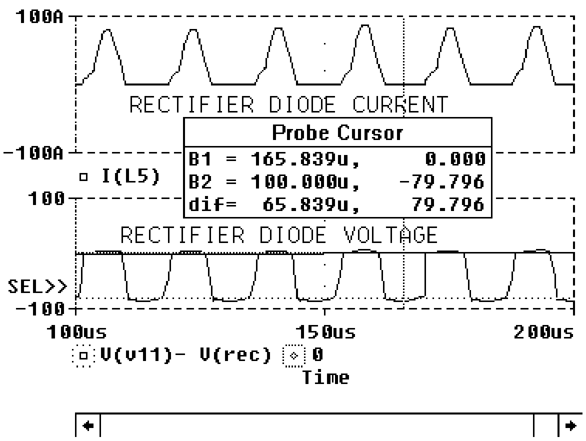

(a)



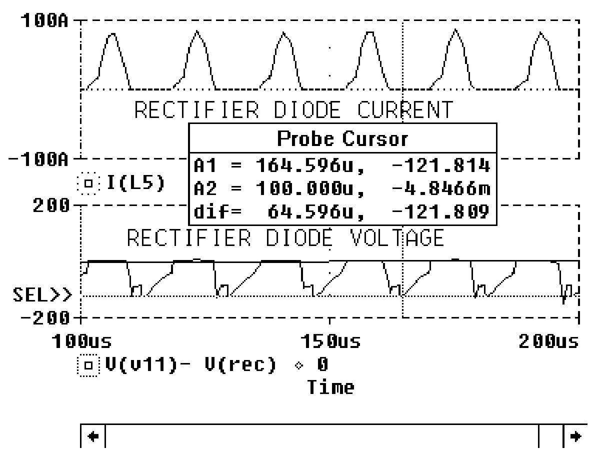
(b)

Fig. 3. Simple model of leakage inductances in a transformer for center tap rectification (a) and full bridge rectification (b), $L_m$- magnetization inductance.

When the output rectifier has a bridge topology (Fig. 3b), the leakage inductance of the secondary does not increase the peak reverse voltage on the rectifier diodes.




(a)




(b)

Fig. 4. Rectifier diode current and voltage in a SRC with center-tapped rectification topology (simulation results). Upper traces: rectifier diode current. Lower traces: rectifier diode voltage; (a) - all leakage inductance in primary, (b) - leakage inductance in the two secondaries.

## IV. INTEGRATED MAGNETICS

Notwithstanding the fact that the size of the inductor in SRC is a dependent variable and may be rather large in some cases (Fig. 2), size reduction of the total system is still possible by applying integrated magnetics [3-6]. The first and obvious saving is in the interconnection. Combining two rather major elements into one structure simplifies the wiring task and normally reduces the total foot print area. The series inductance can be realized by leakage inductance, namely by increasing the distance between the primary and secondaries

This realization is limited to small inductances. For larger inductances one would prefer to base the design of the inductance on an auxiliary ferromagnetic core. This will reduce the copper losses as compared to air leakage realization. In this study we explored the two approaches.

## V. PROPOSED INTEGRATED MAGNETICS COMPONENTS

An integrated transformer-inductor 1KVA, 30 - 60kHz was designed and constructed (Fig. 5). It replaces a transformer with primary voltage $V_{Tr1}$=150V, turn ratio $n_1$:$n_2$=3:1 and a resonant inductor (19 μH). The design was for an input current of 6A rms. The integrated magnetic element is in fact a transformer with high leakage inductance at the primary. To achieve this, the primary is displaced from the secondaries. The leakage inductance is used as the resonant inductance, disposing of the need to use an extra magnetic component for the resonant inductor. The secondary winding is tightly wound around the center post of the core to reduce leakage between both its sections as much as possible. Consequently, the overall weight and size of the SRC can be reduced.

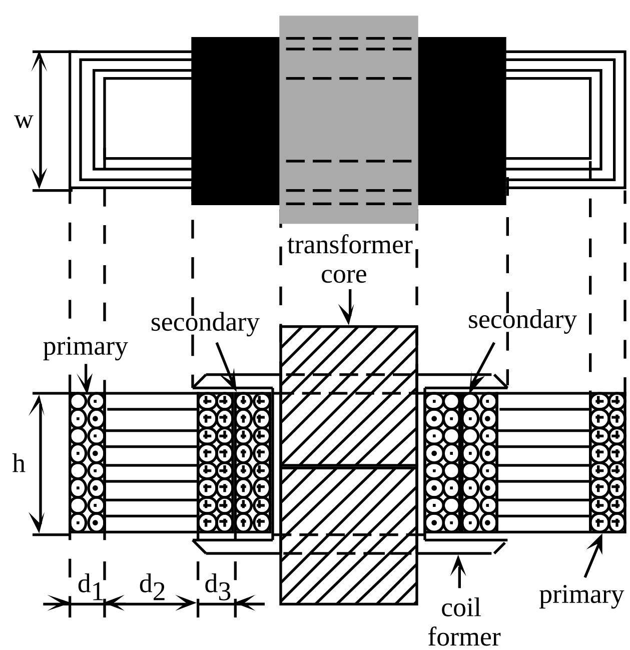



Fig. 5. An integrated transformer-inductor with air leakage inductances.

Transformer-inductor was designed like regular transformer with extra of leakage inductance. The ferrite core was Philips E65/32/27 with $A_c$=532mm$^2$.

The leakage inductance can be estimated by [9]:

$$L_{lk1\text{-}2} = \frac{4\pi\, 10^{-7}\, n_1^2\, w}{h} \left(\frac{d_1}{3} + d_2 + \frac{d_3}{3}\right) \cdot 2 \quad (15)$$

where $n_1$ - number of turns in the primary, $d_1$ - thickness of primary, $d_2$ - thickness of secondary, $d_3$ - distance between primary and secondary, w, h - width and height of windings.The multiplying factor 2 takes into account both sides of the integrated arrangement.

If $d_1$ and $d_3$ are appreciably smaller than $d_2$, equation (15) can be simplified:

$$L_{lk1\text{-}2} \quad \frac{8\pi\, 10^{-7}\, n_1^2\, w}{h}\, d_2 \quad (16)$$

Calculated and measured leakage inductance $L_{lk1\text{-}2}$ were about 11µH (w=25mm, h=35mm, $d_2$=40mm). The inductance was further increased to 19µH by placing an auxiliary gaped core (ETD29/16/10) on one side of the primary winding. An alternative approach would be to base the design on ferrite cores [10] as depicted in Fig. 6. In this embodiment, the total length of the copper wire is shorter by 2*h (Fig. 5) as compared to the conventional two elements arrangement. This alternative design is particularly advantageous when the inductor to transformer ratio (Fig. 2) is large.

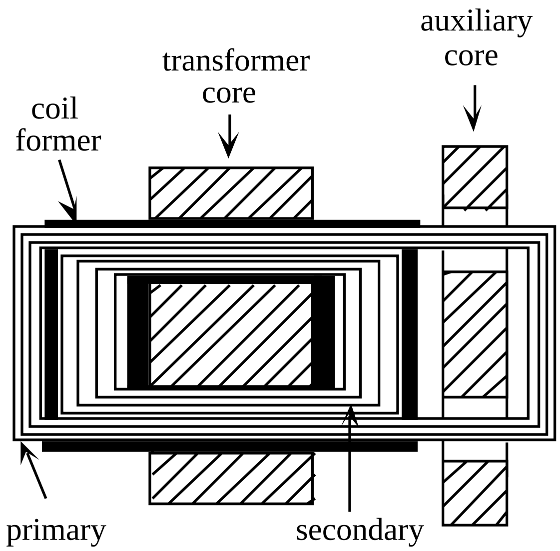



Fig. 6. A cross-section of integrated transformer-inductor with auxiliary core. The cut is in parallel to the windings plane.

## VI. EXPERIMENTAL

An experimental SRC with an integrated transformer-inductor (Fig. 7) was investigated in continuous and discontinuous current modes at power levels to 1kW and switching frequency in the range 30 - 60kHz. Experimental current and voltage waveforms (Fig. 8) were found to be in good agreement with simulation results (Fig. 4).

## VII. CONCLUSIONS

The physical size of the inductor in series resonant converter is a dependent parameter and cannot be reduced by decreasing the inductance value. The only possible way to reduce this ratio is by operating the converter away from resonance and/or designing it for a high DC voltage transfer ratio.

The structural complexity of series resonant converters can be reduced by applying an integrated transformer-inductor

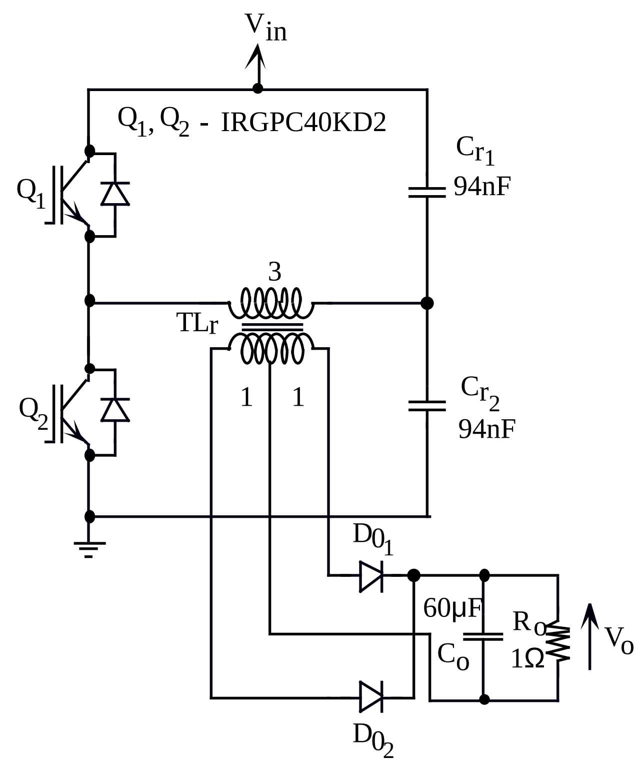
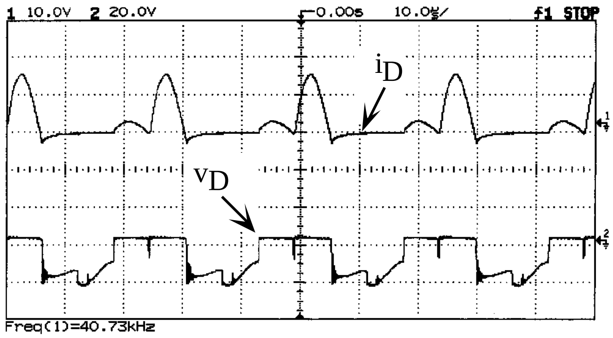


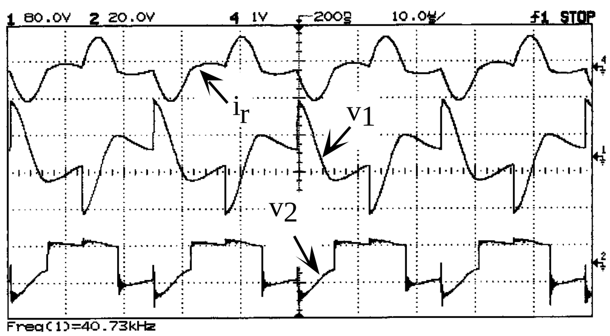

Fig. 7. Experimental 1kW series resonant converter with integrated transformer-inductor.

built on a common magnetic structure. If the output rectifier is of the center-tapped topology, the leakage inductances of the integrated transformer-inductor must be concentrated in the primary (between the primary and the two secondary windings) because a leakage inductance between the secondary windings increases the voltage stresses of the output rectifier.

The integrated magnetic structure can be realized by simply increasing the distance between the primary and secondaries. This would be practical in cases that call for a small inductance. In this arrangement, copper losses may increase due to the longer wire length and a possible increase in ac resistance in the poorly coupled portions of the windings. The second alternative (Fig. 6) could be useful in cases that call for a relatively large inductance. In this case the expected copper losses are smaller than the two element approach. The penalty will be a somewhat more complex production.




(a)




(b)

Fig. 8. Current and voltage waveforms of the experimental converter with integrated transformer-inductor ($gV_{in}$=50V, $V_o$=9V, $n_1$:$n_2$=3:1, $\frac{f_r}{f_s} \approx 2$): (a) - rectifier diode current ($i_D$) and rectifier diode voltage ($v_D$); (b) - current of the series resonant circuit ($i_r$), primary voltage ($v_1$) and secondary phase voltage ($v_2$) of the transformer-inductor.

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