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ZERO VOLTAGE SWITCHING USED FOR A REVERSIBLE DC-DC CONVERTER

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ABSTRACT

A novel bi-directional dc-dc converter with Zero Voltage Switching (ZVS) and interleaving for dual voltage systems in automobiles is presented. A variable frequency extended band hysteretic current control method is proposed. In comparison with classical fixed frequency current control PWM, the reverse polarity peak current needed for ZVS operation is kept constant. Inductor current ripple decreases with load reduction. Automatic changes in operation between buck and boost modes are accomplished without transient currents.

Converter simulations are carried out using Matlab/Simulink platform.

KEY WORDS

DC-DC Converter, Soft switching, ZVS Converter, Battery management, Converter control, Model simulation.

1. Introduction

The developments in dual-voltage automotive electrical systems (42 V and 14 V) include a high-voltage power distribution system (42 V), referred to as 42 V Power Net, which supplies the high power loads, along with a low-voltage system (14 V) which was preserved to ensure compatibility with the loads not yet re-designed for 42 V supply system, or unsuitable for high-voltage operation.

The most straightforward technique in dual voltage supply systems is to use a Lundell alternator as a generator for the 42 V system along with a dc/dc converter to supply the 14 V subsystem (Fig. 1). Since the latter represents an additional system component, the achievement of a cost-effective dc/dc converter constitutes a major research target in this field [1, 2, 3, 4].

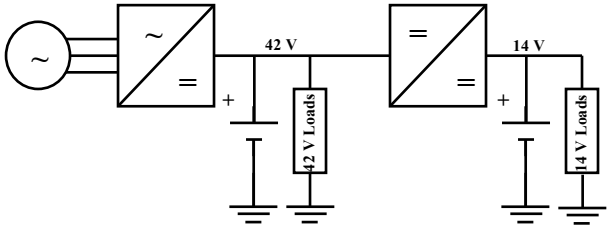

Fig. 1 Block diagram of the dual voltage system

At present, design and construction of these dc/dc converters are well documented in the field of power electronics, although aspects such as relatively high cost, size and weight have still to be overcome.

Apart from aspects related to size and cost, the dc/dc converter for automotive dual supply system must be able to operate under harsh temperature and vibration conditions and comply with regulations regarding EMC.

A problem directly connected with fuel consumption is the efficiency of this converter. Currently, efficiency improvement is obtained when using synchronous rectifiers in dc/dc converters. Apart from advantages such as high efficiency and a reduced number of power devices, the dc/dc converter with synchronous rectifier allows bi-directional operation without using other additional components. The relatively simple topology of this converter allows a high degree of component integration, with favourable effects on fabrication cost, heat dissipation, mechanical strength and the possibility of achieving a modular structure.

The reversible operation of the converter is necessary since it enables engine starting from the 14 V subsystem (jump starting) when the 42 V system battery is low, or when the vehicle must be started from an external battery. It can also transfer energy from one battery to another when necessary.

According to [16] a charging strategy consisting of a short big current charging pulse (I_p for t_p) is followed by a short discharging pulse (I_n for t_n) and a recovery (I_{cg}=0 for t_r) period, as shown in Fig. 2, allows to minimize the charging time and improve the battery cycle-life.

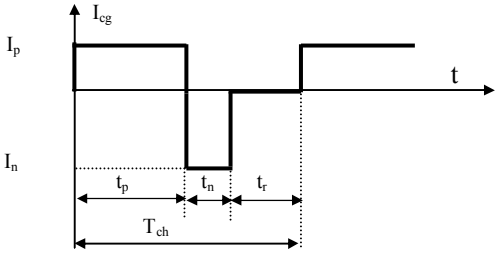

Fig. 2 Fast charging strategy

The bidirectional dc-dc converter can be easily controlled to generate positive and negative charging pulses without the need of an energy consumption load for the discharging mode.

During the charging and discharging pulses, the reference current can be kept constant. According to [3] and [4] the bandwidth of the converter switching frequency is much limited, and is dependent only on the variation of the input and output voltages.

In this strategy, the average pulse charging current is set to vary in a manner similar to the conventional constant current-constant voltage profile. The charging pulse period is pulse width modulated by the error signal of a voltage regulator, implemented in a microcontroller. For example, from a total charging cycle of around 500 ms, the maximum charging pulse period can be 450 ms (90%), the discharging pulse period is fixed at 35 ms (7%) and the difference is the recovery period.

By using this methodology, the hysteretic control of the current in every stage of the interleaved converter is straightforward, robust in terms of stability of operation, ensure good dynamic response and enables automatic changing from buck mode to boost mode.

2. Operating Principia

The topology of the power stage of the bi-directional 42 V-14 V dc/dc converter is presented in Fig. 3.

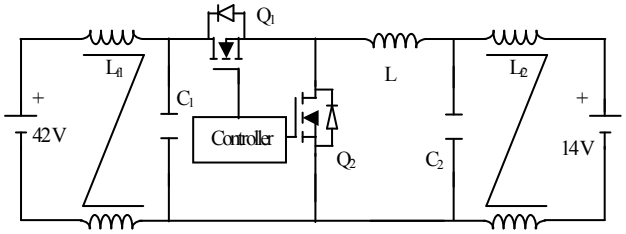

Fig. 3 Basic topology of the bi-directional dc/dc converter

This type of converter results by combining a buck converter with a boost one. The main advantages of this topology are: fewer components (low cost, high reliability) and the simplicity of control and driver circuits.

The bi-directional converter is employed in automotive dual voltage supply power systems in order to maintain optimum battery state of charge. When power is transferred from the 42 V system battery to the 14 V system battery, the converter operates in step-down mode (dc/dc buck converter) and, inversely, when power is transferred from the 14 V system battery to the 42 V system battery, it operates as a step-up converter (dc/dc *boost* converter).

For both operating modes, the role of the rectifier diodes is transferred to the MOSFETs (transistor Q_1 in the boost operating mode and transistor Q_2 in the buck operating mode, respectively).

Both MOSFETs and their body diodes operate by the principle of the synchronous rectifier. A MOSFET is turned on with its freewheeling diode already in on-state. Thereby, the MOSFET's switching process occurs at near zero voltage, which involves negligible switching power losses. In this case, the MOSFETs are operated in the 3rd quadrant of the volt-ampere characteristic (the current flows from source to drain). Once turned on, the MOSFET's very low channel resistance appears in parallel with its intrinsic diode and the current is transferred from the diode to the MOSFET. Since for low direct conduction resistance MOSFETs ($R_{DS(on)}$ around 10 mΩ), the voltage drop across the transistor, even for maximum load current, is less than the diode's on-state voltage, the largest part of the current will flow through the transistor.

Irrespective of the direction of power transfer, the operation of the bi-directional converter exhibits two states:

1. the active state when one of the MOSFETs that is operated as main switch (transistor Q_1 for the buck operating mode and transistor Q_2 for the boost operating mode) is turned on (the energy is drawn from one of the sources);
2. the passive state when the complementary transistor will operate as a synchronous rectifier (transistor Q_2 for the buck operating mode and transistor Q_1 for the boost operating mode) is turned on (the energy is delivered to the second source).

The MOSFETs' transition from the active to the passive state is always at ZVS and is achieved by the current in inductance *L*, which is switched to the diode of the complementary transistor.

The MOSFETs' transition from the passive to the active state depends on the operating mode of the bi-directional dc/dc converter and can be of three types:

1. Hard switching when the converter is operated in continuous current conduction without polarity change (the switch transistor catches the diode of the synchronous rectifier in its on-state). In order to minimize the negative effect of diode reverse recovery time $t_{rr,}$ in hard switching (power losses and parasitic oscillations) it is required to provide transistors with fast intrinsic diodes. This does not represent a major difficulty in the case of MOSFETs with maximum admissible peak voltage of around 100 V.
2. Zero Current Switching (ZCS) soft switching mode when the converter is operated in discontinuous current mode (at current zero-crossing, the complementary MOSFET must be turned off). Because MOSFETs have large output capacitances, the ZCS-mode is affected by some losses. Although the losses generated by the diode are eliminated, when the main switching transistor is turned on, its output capacitance is periodically discharged through the device and generates losses proportional with the switching frequency [5];
3. Zero Voltage Switching (ZVS) soft switching mode when the converter is operated in continuous current mode with zero-crossing (synchronous continuous conduction mode) [6-10].

The dc/dc buck converter with synchronous rectifier can be permanently operated in continuous output current mode through inductance *L*, since after zero-crossing, the current can change direction (the current can flow both directions through the Q_2 transistor). This feature allows the soft switching of transistors Q_1 and Q_2. For this reason, the value of the inductance *L* is determined so that, for a given switching frequency, the ripple of the current flowing through inductance *L* exceeds the double of the peak load current by a pre-established amount (the

current through the inductance L should cross below zero). When the transistor Q_2 is turned off, this current flowing from drain to source discharges the output capacitance of the transistor Q_1, which is then followed by the on-state switching of Q_1 transistor's body diode.

Transistor Q_2 is turned off at low voltage (owing to slow charging of its output capacitance), while transistor Q_1 is turned on at a very low drain-to-source voltage around zero, due to the freewheeling diode, already in on-state. This switching method that enables drastically reduced switching power losses and increased switching frequency is commonly referred as ZVS. The inductance L is chosen of relatively small value as to ensure ZVS for maximum load current, and then soft switching is maintained throughout the entire load current range.

Although ZVS ensures almost zero switching power losses in transistors Q_1 and Q_2, iron losses produced in the magnetic core of the inductance L are high (owing to the current ripple of around twice the load current). This is why this switching method is suitable in interleaved dc/dc converters [8].

The main advantage of interleaved dc/dc converters (Fig. 4) is the minimization of the reactive filter elements with direct consequences on the size of the converter. It can be assumed that a number $N = 4$ of phases [11] represents an optimum trade-off between circuit complexity and current ripple reduction degree.

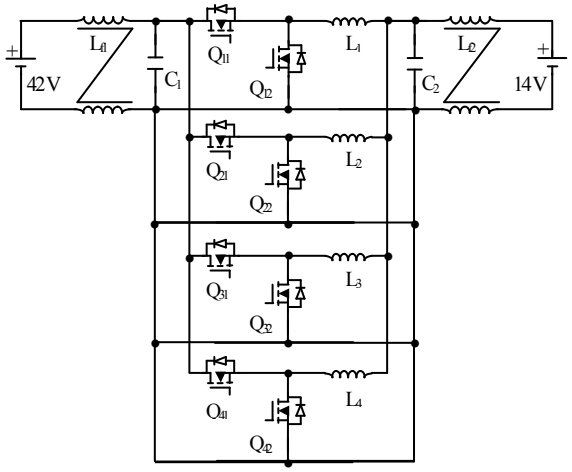

Fig. 4 Interleaved bi-directional dc/dc converter

Decreasing the magnitude of the filter inductance enables a fewer number of turns which results in a higher saturation current (for the same cross-sectional area of the magnetic core), lower winding losses and the possibility of inductor implementing in planar technology. A smaller filter inductance also contributes to an improved dynamic response of the converter.

3. Converter Control

To regulate the interleaved bi-directional dc/dc converter the current in each stage must be controlled [7, 8, 12, 13] to ensure operation with equal fractions of the total load current (current sharing). Current control implicitly ensures overload and short-circuit protection of the power transistors.

The reference current i_{ref} (Fig. 4) is generated by a voltage regulator and has positive or negative polarity depending on whether the converter operates in buck or boost mode respectively.

Following strategies can be adopted for current control:

- Fixed frequency current controlled PWM;
- Variable frequency extended band hysteretic current control.

The first of these strategies (fixed frequency current controlled PWM) can be implemented with general purpose, low-cost integrated circuits. Since these circuits are supplied from a single polarity source (generally +12 V), two circuits per stage are required in order to control current of both polarities. Also, when changing back and forth between buck and boost modes of operation, current surges occur due to controller saturation [14].

To avoid these current spikes, transition is achieved by including a switching pause or by implementing the PWM modulator with discrete circuits supplied with two polarities sources (±12 V dc).

Another disadvantage of the current control PWM is illustrated in Fig. 4, which shows the characteristic waveforms of this type of modulation for the buck mode operation [12].

With voltage sources applied at both input and output of the converter, the duty cycle $D=T_{on}/T$ (Fig. 5) is determined by the ratio of the voltages. For the ideal case of 42 V and 14 V, respectively, the duty cycle must be D=1/3 for the buck mode and D=2/3 for the boost mode.

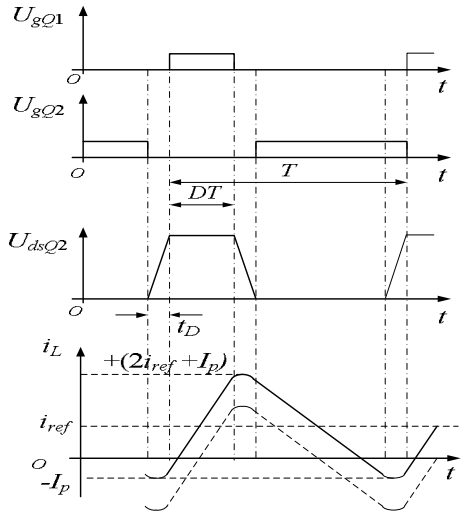

Fig. 5 Gating pulses; drain voltage and current through inductance L

In order to ensure ZVS, the inductance L is designed so that for maximum load current, a reverse current peak I_p will appear through the inductance. The maximum current peak is determined from the condition that the energy generated in L equates the energy required for charging the two output capacitors of the MOSFETs up to the input voltage (of 42 V nominal value). This condition is:

$$LI_P^2 / 2 = C_e U_i^2 / 2 = C_{oss\,eff} U_i^2 \qquad (1)$$

For a resonant transition, the optimum delay between the gating pulses applied to the two MOSFETs is equal to the quart of the period:

$$t_D = 0.5\pi\sqrt{LC_e} \quad (2)$$

Fig. 5 shows that for a constant value of the duty cycle D, as the load current decreases, the current ripple in the inductance remains constant, but the reverse current peak increases (dotted line). This surplus of current is re-circulated through the power switches and generates additional losses at light loads.

In the second strategy (variable frequency extended band hysteretic current control), the current in the output inductor is maintained within a band centred on the reference value $i_L = i_{ref} \pm \Delta i$, so that average inductor current equals reference current. Usually, this band $\pm\Delta i$ is very narrow. In order to ensure ZVS operation in the proposed control method, this band is extended to $\Delta i = i_{ref} + I_p$, so that the current changes polarity with a peak I_p, as shown in Fig. 6, in accordance with the buck or boost operating modes. The minimum value of I_p is determined by the equation (1). The current in the output inductor can be indirectly observed or directly measured.

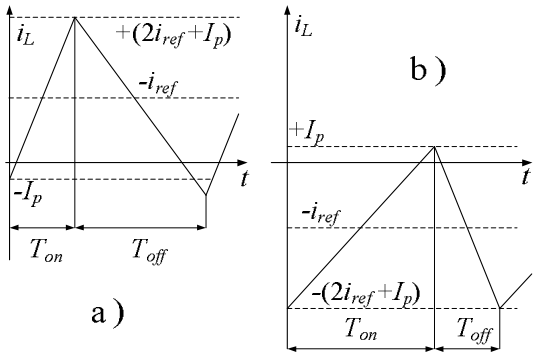

Fig. 6 The current waveform through inductance L: a) in *buck* mode; b) in *boost* mode.

For the buck operating mode [15] the lengths of the two states (active and passive) are given by the relations:

$$T_{on} = L\left(2i_{ref} + I_P\right)/\left(U_i - U_e\right) \quad (3)$$

and

$$T_{off} = L\left(2i_{ref} + I_P\right)/U_e \quad (4)$$

where U_e and U_i are ideally of 42 V and 14 V, respectively.

A first remark with respect to this method is that at light loads the current ripple in the output inductor decreases and the switching frequency increases. The variable switching frequency generates interleaving problems. These problems can be solved by assigning one stage as a master whose gating pulses that start the active state, are used as a clock signal for a PLL (Phase Locked Loop) circuit. The synchronizing pulses for the slave phases, at intervals equal with a quarter of the master period, are obtained from the logic outputs of a divide-by four circuit inserted in the PLL feedback loop. For each slave phase stage, the start of the active state is externally given by the PLL circuit, and the on time of the active switch is determined by comparing the estimated inductor current with $(2i_{ref} + I_P)$ level. The difference between average current among the stages will be mostly dependent of the output inductor tolerances.

In order to ensure zero voltage switching operation for all converter phases, the output inductor with the highest value must be selected for the master stage.

A very large switching frequency bandwidth can overpass the capture band of the PLL and create difficulties in the design of the EMI filters. In the same time, the increase of the switching frequency must be limited in order to reduce output inductances magnetic core losses.

To overcome these problems, a control strategy for bi-directional dc-dc converter is used that maintain an almost constant magnitude of the reference current.

The main aim of the bi-directional dc-dc converter is to be used as a high-efficiency fast battery charger.

4. Simulation Results

In order to validate the concept a simulation model for a single bi-directional dc/dc converter has been build using Matlab/Simulink platform. For the purpose of this validation, the power devices proposed for the implementation of the converter are International Rectifier Automotive series IRF2907ZS MOSFETs: $V_{DSS} = 75$ V; $I_D = 60$ A (package limited); $R_{DS(on)} = 3{,}8$ mΩ (at $T_j = 25$ºC); $t_r = 90$ ns; $t_f = 44$ ns; $C_{iss} = 7500$ pF; $C_{oss\ eff} = 1110$ pF; $Q_g = 260$ nC; $T_j = 175$ºC; $R_{\theta jC} = 0{,}5$ºC/W; $R_{\theta CS} = 0{,}5$ºC/W. From the specification sheet results $R_{DS(on)} = 2\times3{,}8$ mΩ =7,6 mΩ (at $T_j = 130$ºC).

An external snubber of 4 nF series with a very small resistor has been added for each transistor in order to minimize the variation of the internal capacitance ($C_{oss\ eff}$).

Considering that the selected converter control method is by the extended band hysteresis current with ZVS and the minimum frequency is about 50 kHz, an inductance L value of 4 µH is selected with a minimum saturation current of 25 A [7].

4.1 Buck Mode – Single unit

Fig. 7 shows the current through the inductor for a reference current of 20 A. Fig. 8 and 9 shows the same graph for a reference current of 15 A and 25 A respectively.

As can be noticed, the frequency of the inductor current varies between 48 and 67 kHz as the reference current changes between 25 A and 15 A respectively.

Another aspect investigated was the moment of transistors switching (see Fig. 10) and the voltages across the switches in the moment of triggering, Fig. 11 and 12 respectively; the reference current was 20 A and the amplitude of the gate signals have been conveniently amplified and vertical shifted in order to cache the right moments related with the current through inductor.

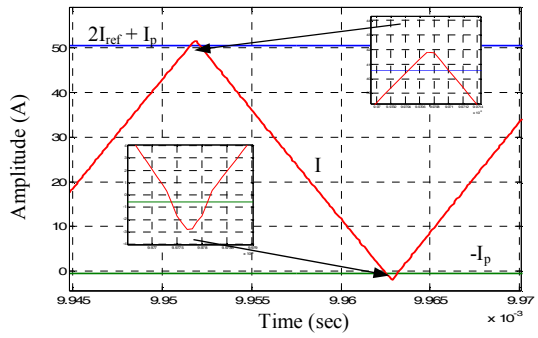

Fig. 7 Inductor current for I_{ref} = 20 A

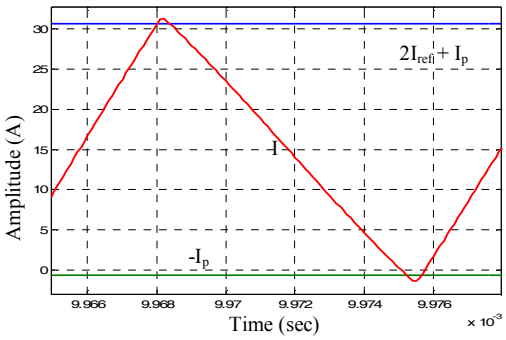

Fig. 8 Inductor current for I_{ref} = 15 A

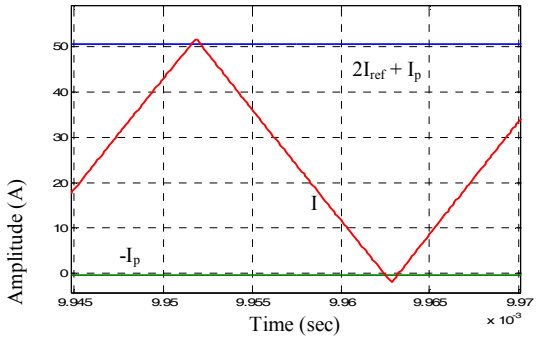

Fig. 9 Inductor current for I_{ref} = 25 A

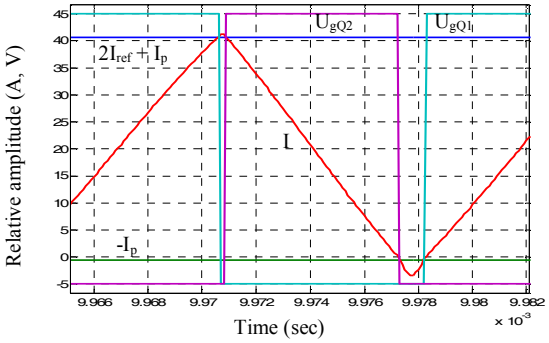

Fig. 10 Gate signals

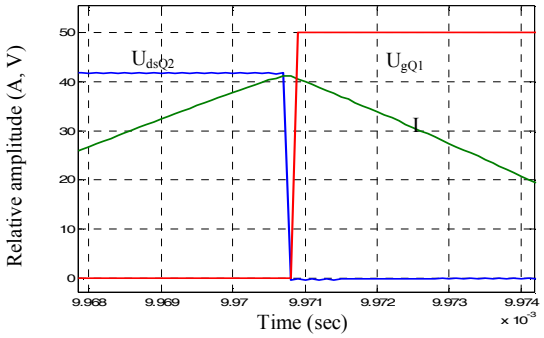

Fig. 11 Zero voltage for the lower transistor Q_2

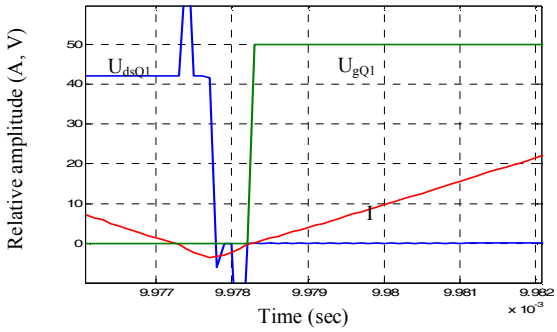

Fig. 12 Zero voltage for the lower transistor Q_1

4.2 Boost Mode – Single unit

In the boost mode, the current flow is in opposite direction from the 14 V battery to 42 V battery and the sense of it is negative. Fig. 13 shows the how the inductor current evolutes for a reference of 12 A.

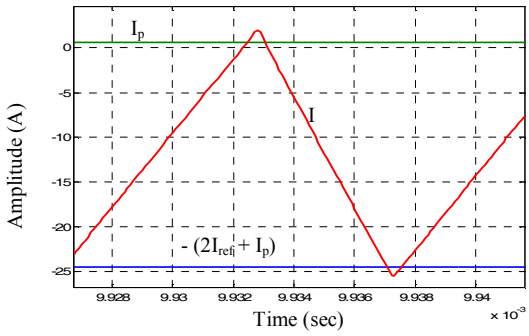

Fig. 13 Inductor current for I_{ref} = 12 A in boost mode

4.3 Interleave mode

Fig. 14 shows the output current (I_o) for interleaving four units in buck mode and an individual reference of 20 A. As can be noticed, the ripple of the output current is much smaller as for an individual unit. Fig. 15 shows the output current (I_o) for interleaving four units in boost mode and four times the individual reference of 12 A.

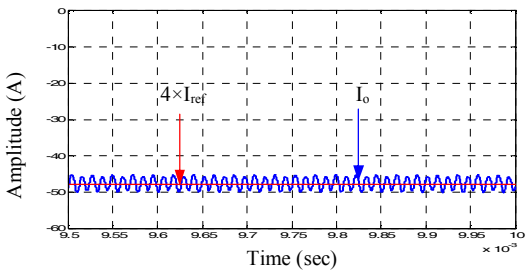

Fig. 14 Output current for interleaving in buck mode

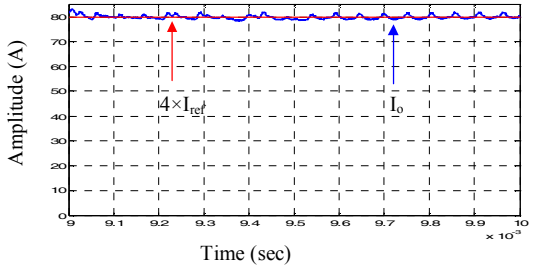

Fig. 15 Output current for interleaving in boost mode

5. Conclusion

Employing the bi-directional dc/dc converter for power transfer from the 42V supply system to the 14V supply system and vice versa, in order to achieve a balanced supply system in dual voltage system vehicles represents a viable solution. A four stage interleaved buck-boost topology was selected. One module is set to operate under ZVS by extended band hysteretic current control. The gating signals of the master module are used to generate synchronizing signals for the slave modules by means of a PLL circuit. The slave modules are commanded to operate under the ZVS peak current control, to achieve active current sharing and low output ripple.

Simulation results of applying the proposed method to a bi-directional dc/dc buck-boost converter confirm the analytical results.

Interleaving makes possible a modular design, power loss distribution and high power applications. The interleaving structure offers additionally advantages: fast transient response, reduced ripple current and filter component miniaturisation.

The next research objectives are to evaluate the converter operation and to verify the control strategy by experiment.

The simple layout of the bi-directional dc/dc converter power module is adequate for a high degree of integration and allows high-volume manufacturing with direct effects on cost reduction and reliability increase.

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