Interleaving of a soft-switching Boost Converter operated in Boundary Conduction Mode

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Abstract—This paper presents the interleaved operation of a soft-switching boost converter operated in boundary conduction mode. First, the operating principle of the converter as well as the basic concept of the interleaving is presented. Then, the dynamic behavior is modeled by using the z-transform to obtain a converter model which is independent of the switching frequency. With the model, the stability of the closed loop system with a PI-controller is analyzed. It is shown that an adaptive PI-controller can easily be implemented to a minimal settling time over a wide operating range.

Finally, the controller is validated with two converters with 40 kW nominal output power and an output voltage of 3 kV. The tests at different output voltages under different load conditions show a stable interleaved operation.

I. INTRODUCTION

A compact and cost effective X-ray free electron laser system (SwissFEL) is currently built at the Paul Scherrer Institute (PSI) in Switzerland [1]. This laser system requires modulators with a pulse power of 127 MW for 3 μs. The high pulse power is provided by a capacitor bank which is recharged to 3 kV by two 40 kW boost converters with 1.3 kV input voltage between two consecutive pulses.

In the considered system, two interleaved converters (Fig. 1) [2], are used to charge the capacitor bank. The interleaved operation results in a reduced input current ripple. In PWM controlled systems with fixed switching frequency, the interleaving can be easily controlled by shifting the PWM signals relative to each other. However, this method does not work for converters operating in boundary conduction mode (BCM) that results in a variable switching frequency. Different open-loop and closed-loop control methods for the operation in BCM have been investigated (e.g. [3], [4]). However, those methods do not guarantee a soft-switched operation when the system is perturbed.

In this paper, a control strategy for interleaved boost converters operated in BCM is presented. The proposed controller guarantees zero voltage switching during the interleaved operation including startup and the synchronization of the converters.

The basic operation principle of the converter and the interleaving is presented in section II. In section III, the converter model and the controller model are introduced. Two different controllers are investigated and the implemented controller is shown. Finally, measurements with the proposed controller and two 40 kW, 3 kV interleaved boost converters are presented.

II. CONVERTER OPERATION

The basic circuit of the converter including the snubber circuits is shown in Fig. 2. Details of the converter and the feedback controller are presented and analyzed in [2] and [5]. In the following, the converter operation in BCM and the interleaved operation are explained.

A. Converter Operation

The series connected switches $S_{1a}$ to $S_{1n}$ are turned on at the beginning of interval $T_1$ (cf. Fig. 2). The input voltage $V_{in}$ is applied across inductor $L_1$ and inductor current $i_L$ starts to rise. When inductor current $i_L$ reaches the level $i_{LP}$, switches $S_{1a}$ to $S_{1n}$ are turned off at the beginning of interval $T_2$ (peak current control). Diodes $D_{1a}$ to $D_{1n}$, are not conducting at that time since capacitors $C_{Sn,D1a}$ to $C_{Sn,D1n}$ are still charged. Inductor $L_1$ and snubber capacitors $C_{Sn}$ form a resonant circuit. Since inductor current $i_L$ is positive, the capacitors in parallel to the diodes $D_{1a}$ to $D_{1n}$ are discharged to zero and the capacitors across the switches are charged to $V_{out}$. As soon as the voltage across the diodes reaches zero, they start to conduct (interval $T_3$). At that point of time, the difference...
between $V_{out}$ and $V_{in}$ is applied across $L_1$ and current $i_L$ decreases. At the beginning of interval $T_s$, $i_L$ reaches zero and the diodes stop conducting. The snubber capacitors and the inductor form again a resonant circuit. The voltage across the inductor is the difference between $V_{out}$ and $V_{in}$. Hence, the inductor current becomes negative and the capacitors across the switches are charged to zero if the difference between $V_{out}$ and $V_{in}$ is large enough. At then end of $T_s$, the body diodes of the switches start to conduct. At that point of time, the switches have to be turned on before the inductor current becomes positive and the next switching cycle starts.

The operation in BCM results in a switching frequency which depends on the input and output voltage, the output power as well as the inductance value.

B. Interleaved Operation

The inductor current waveforms for two interleaved converters are shown in Fig. 3. Since the switching frequency depends on $i_{Lp}$, the phase shift $\varphi = \frac{\Delta T_s}{T_s}$ can be controlled by reducing or increasing $i_{Lp,1}$ relative to $i_{Lp,0}$. The phase shift can be measured by detecting the inductor current zero crossings. One converter is used as a reference (master) whereas the other is measured by detecting the inductor current zero crossings. One converter is used as a reference (master) whereas the other is measured by detecting the inductor current zero crossings. One converter is used as a reference (master) whereas the other is measured by detecting the inductor current zero crossings.

In an ideal system with identical converters, the switching frequency for both converters are identical for the same current $i_{Lp}$. This is not the case in a real system because of component tolerances, temperature drifts, etc. In BCM, non-equal inductor values result in unbalanced inductor currents. Fig. 4 shows the relative deviation of the average inductor current between the master and the slave depending on the inductor mismatch at the nominal operating point of the investigated system (Table I). The deviation is smaller if the slave's inductor is larger than the one of the master. Therefore, the converter with the smallest inductor value should be selected as the master.

In order to model the interleaved operation, a general model with non-identical converters has to be developed.

III. FEEDBACK CONTROLLER

In this section, the interleaving model and the controller model are derived. Also, the optimal controller parameters are calculated. Afterwards, the implemented phase shift measurement and the controller are presented.

A. System Model

1) Plant Transfer-Function: The inductor current waveforms are shown in Fig. 3. It is assumed that the input- and output-voltage as well as $i_{Lp,0}$ remain constant. The dynamic behaviour of the time offset between the two inductor currents is modeled by calculating the time offset for the next switching cycle $\Delta T_{n+1}$ depending on the offset $\Delta T_n$ and the switching period lengths of the converters $T_{s,0}(i_{Lp,0})$ and $T_{s,1}(i_{Lp,1})$:

$$\Delta T_{n+1} = \Delta T_n - T_{s,0}(i_{Lp,0}) + T_{s,1}(i_{Lp,1}) \quad (1)$$

The switching periods $T_{s,k}(i_{Lp,k})$ can be calculated analytically by solving the differential equations describing the dynamic behaviour of the converter. These expressions are non-linear because of the resonant transitions during the time intervals $T_2$ and $T_3$. In order to simplify the model, $T_{s,1}(i_{Lp,1})$ is linearized at $i_{Lp} = i_{Lp,0}$:

$$T_{s,1}(i_{Lp,1}) \approx T_{s,1}(i_{Lp,0}) + \frac{dT_{s,1}(i_{Lp})}{di_{Lp}} \bigg|_{i_{Lp,0}} \Delta i_{Lp} \quad (2)$$

with $\Delta i_{Lp} = i_{Lp,1} - i_{Lp,0}$.

Since the continuous offset $\Delta T_n$ is discrete in time, the $z$-transform is applied:

$$Z\{\Delta T_{n+1}\} = zZ\{\Delta T_n\} = z\Delta T(z) \quad (3)$$

It is assumed that $i_{Lp,0}$ is constant. Hence, $T_{s,0}(i_{Lp,0})$, $T_{s,1}(i_{Lp,0})$ and $\frac{dT_{s,1}(i_{Lp})}{di_{Lp}} \bigg|_{i_{Lp,0}}$ are constant. By using (1) - (3) one obtains:

$$\Delta T(z) = G_d(z) + G_p(z)\Delta I_{Lp}(z) \quad (4)$$

with the disturbance transfer-function

$$G_d(z) = \frac{1}{z - 1} \left[ T_{s,1}(i_{Lp,0}) - T_{s,0}(i_{Lp,0}) \right] = \frac{1}{z - 1} \Delta T_s$$

and the plant transfer-function

$$G_p(z) = \frac{1}{z - 1} \left. \frac{dT_{s,1}(i_{Lp})}{di_{Lp}} \right|_{i_{Lp,0}} = \frac{1}{z - 1} T_{s,1}(i_{Lp,0})$$

Fig. 3. Interleaved inductor current waveforms.

Fig. 4. Current balancing of the average inductor currents depending on inductance mismatch $\Delta L = \frac{L_{Lp,0} - L_{Lp,1}}{L_{Lp,0}}$. 
2) Closed Loop Transfer-Function: In this section, the closed-loop system shown in Fig. 5 is analyzed. The system model including the controller is:

\[
\Delta T(z) = \frac{G_c(z)G_p(z)}{1 + G_c(z)G_p(z)}\Delta T_{set}(z) + \frac{G_d(z)}{1 + G_c(z)G_p(z)}\Delta T_{dis}
\]

\[= G_{cl}(z)\Delta T_{set}(z) + G_{dis}
\]

(5)

where \(G_{cl}(z)\) represents the plant closed loop transfer-function and \(G_{dis}\) the disturbance closed loop transfer-function.

B. Proportional-Integral Controller

It is known from the control theory that a system is stable if and only if its poles are inside the unit circle of the complex plain. Since a proportional (P) controller is not able to drive the system to the desired phase-shift without a static deviation, a proportional-integral (PI) controller is investigated.

1) Stability: The closed loop transfer-function for a PI-controller \(G_{cl}(z) = K_p + K_i\frac{z}{z-1}\) is

\[G_{cl,PI}(z) = \frac{(K_{I,n} + K_{P,n})z - K_{P,n}}{z^2 + ((K_{I,n} + K_{P,n}) - 2)z - K_{P,n} + 1}\]

\[K_{P,n} = K_p T_{s,1}^\prime(i_{Lp,0})\]

\[K_{I,n} = K_i T_{s,1}^\prime(i_{Lp,0})\]

(6)

(7)

(8)

Hence, the dynamic behaviour of the system depends on \(K_{P,n}\) and \(K_{I,n}\), providing an operating point independent dynamic behaviour. \(G_{cl,PI}(z)\) has two poles at

\[p_1,PI = \frac{-K_{P,n} + K_{I,n} - \sqrt{(K_{P,n} + K_{I,n})^2 - 4K_{I,n}}}{2} + 1\]

(9)

and

\[p_2,PI = \frac{-K_{P,n} + K_{I,n} + \sqrt{(K_{P,n} + K_{I,n})^2 - 4K_{I,n}}}{2} + 1\]

(10)

The system is stable for

\[0 < K_{I,n} < 4 - 2K_{P,n}\]

(11)

and

\[0 < K_{P,n} < \frac{4 - K_{I,n}}{2}\]

(12)

2) Settling Time: The settling time of the system with PI-controller is shown in Fig. 6. The plot shows a minimum settling time of 2 switching cycles around \(K_{P,n} = 1\) and \(K_{I,n} = 1\).
controller output to ensure ZVS and to avoid excessive thermal stress.

IV. MEASUREMENTS

A. Test System

The interleaved operation is tested with two 40 kW, 3 kV converters. The converter parameters are shown in Table I. The two converters are connected in parallel to an ohmic load as shown in Fig. 9.

B. Results

The interleaved operation was successfully tested with two different ohmic loads of 1.4 kΩ and 215 Ω. The output voltages was set between 750 V and 3 kV. The operation at low output power is more critical than the operation with high output power because of the higher switching frequency and because the relative error of the linearized term is bigger at input currents close to the free-wheeling current. As an example, the measured current waveforms of the inductor current and the resulting phase shift at 2.5 kV with a 1.4 kΩ load are shown in Fig. 10.

The deviation of the set-point during the operation can be explained by the high switching frequency and the jitter of the zero-crossing detection and the accuracy of the internal current measurement of the converter. A phase-shift of 10° corresponds to a shift of approximately 200 ns at 140 kHz switching frequency. The used current sensor has a measurement range of 100 A. An accuracy of 0.5 % already corresponds to a deviation of 50 ns of the on-time of the switch at the measured operating point.

V. CONCLUSION

In this paper, the interleaved operation of a soft-switching boost converter operated in boundary conduction mode is investigated. The soft-switching operation as well as the interleaving concept is presented.

A general model for interleaved operation is derived by using the z-transform in order to obtain a switching frequency independent model. Afterwards, the stability of the closed loop system with a PI-controller is investigated. Also, it is shown that the dynamic behaviour of the small signal model of the system is independent of the operating point for normalized controller coefficients. An adaptive interleaving controller is presented.

Finally, the controller has been implemented and successfully tested with two converters at different loads at an output voltage ranging from 750 V to 3000 V.

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REFERENCES