LETTER

A multi-path switched-capacitor-inductor hybrid DC-DC converter with reduced inductor loss and extended voltage conversion range

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Abstract This work proposes a hybrid DC-DC converter using a multi-path switched-capacitor-inductor (MPSCI) topology. Assisted by switched-capacitor branches, it features voltage-conversion range extension and inductor power loss reduction, and hence improving the conversion efficiency. Implemented in a 180-nm CMOS process, the proposed converter can regulate an output voltage of 1.8~3.3V from a 5-V input bus voltage using a 4.7μH inductor (DCR = 240mΩ) at a switching frequency of 800 kHz. It achieves a peak conversion efficiency of 93.7% when delivering an output current of 1A. Under a maximum loading of 3A, this design still attains an efficiency of up to 85.8%. Furthermore, in contrast to the existing dual-path and conventional buck topologies, the proposed converter realizes an inductor loss reduction up to 3.3 and 5.4 times, respectively, at the maximum loading conditions.

key words: DC-DC converter, DCR loss, hybrid topology, multi-path, switched-capacitor, voltage conversion range

Classification: Integrated circuits (analog)

1. Introduction

In recent years, the fast-growing demand for portable intelligent devices powered by low-voltage buses or batteries requires high-efficiency power converters with a small form factor [1 2 3 4 5 6 7 8]. The improvement of the functionality and performance of the loading systems brings more challenges to the DC-DC converter design to suffice the rising loading current requirements. Existing inductor-at-output topologies, such as buck or 3-level topologies [1 2 8 9 10 11 12 13 14 15 16 17 18], suffer from significantly climbed-up conduction loss on the power inductor when delivering a high loading current with a compact system volume. It is mainly due to the loss of inductor parasitic DC resistance (DCR) and can be critical for achieving high power conversion efficiency (PCE). As the tradeoff, a bulky

Fig. 1 Existing DPSCI topologies with VCR = 1/(2-D), (0 < D < 1) reported in (a) [1] and (b) [20].

inductor with a smaller DCR can contribute to the efficiency improvement by, however, scarifying the system compactness [2 3 4 13 16 19]. Within a specific volume, the DCR increases with the inductance and can be as high as hundreds of milliohms, which is dramatically higher than the switch conducting resistance (R_{ON}) and hence introduces considerable influence to the overall PCE. A common issue with the inductor-at-output converters is that the total output current (I_{OUT}) flows through the inductor, leading to a high I^2R loss (thermal dissipation) when delivering a large I_{OUT} .

To limit the converter size and simultaneously reduce the power dissipation on the inductor, a practical solution is to reduce the inductor current by adding more current-conducting paths to the output, as demonstrated in [1 20 21 22 23 24 25]. For example, Fig. 1 shows two existing dual-path (DP) switched-capacitor-inductor hybrid topologies proposed in [1 20]. These topologies reduce the inductor current by delivering a part of I_{OUT} through the switched-capacitor (SC) branches. However, these DP topologies suffer from a restricted voltage conversion ratio (VCR) of 1/2 to 1 due to the constraints on inductor charging and discharging voltages, limiting their application of converting a 5-V bus voltage to a 1.8-V load supply. Although this issue can be resolved by adding an extra power switch and one more switching phase, as discussed in [20], the effectiveness of the inductor current reduction degrades with enlarging the step-down VCR, resulting in a

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Fig. 2 (a) The Proposed MPSCI converter topology, and (b) its two-phase operating principle.

noticeable PCE drop.

This work proposes a multi-path switched-capacitor-inductor (MPSCI) converter topology, addressing the inductor DCR loss and VCR range limitation in high loading current conversions. It consists of a single inductor and dual flying capacitors to realize an extended step-down ratio of 1/3 to 1, featuring enhanced inductor current reduction in a higher-step-down (closer to 1/3) VCR condition.

2. The proposed MPSCI converter topology

2.1 Steady-State Operations

Regarding the inductor-voltage analysis for the existing DP topologies depicted in Fig. 1, the VCR limitation is determined by the voltage across the inductor in both charging and discharging phases. Based on the inductor current balance in steady state, applying $V_{IN} - 2V_{OUT} < 0$ in Φ_2 for the inductor discharging indicates that $V_{OUT}/V_{IN} > 1/2$ for both topologies shown in Fig. 1. Accordingly, a higher voltage difference across the inductor during its current decaying phase can give a lower VCR boundary and, hence, realize a wider conversion range. An effective way to extend the step-down conversion range is to increase the voltage difference across the inductor from $(V_{IN} - 2V_{OUT})$ to $(V_{IN} - 3V_{OUT})$ during Φ_2 by using two SC cells, coming up with the proposed MPSCI topology as illustrated in Fig. 2. According to the operating states shown in Fig. 2(b), the steady-state capacitor voltages on C_1 and C_2 are $(V_{IN} - V_{OUT})$ and V_{OUT} , respectively. Based on the steady-state volt-second-balance relation of the inductor, the corresponding VCR expression

Fig. 3 Conceptual waveforms for the steady-state operation of the proposed MPSCI topology.

is:

$$M = \frac{V_{OUT}}{V_{IN}} = \frac{1}{3-2D}, 0 < D < 1. \quad (1a)$$

$$D = \frac{3M-1}{2M} \quad (1b)$$

As observed, the duty ratio D varying between 0 and 1 results in a VCR from 1/3 to 1.

During Φ_1 , the current i_L ramps up as the inductor voltage $V_{L,\Phi1} = V_{IN} - V_{OUT}$. Paralleled with the inductor, C_1 is charged up to the same voltage as the $V_{L,\Phi1}$. At the same time, capacitor C_2 is discharged by the converter output and, hence, $V_{C2} = V_{OUT}$ in steady-state. During Φ_2 , L , C_1 , and C_2 are connected in series. Thus, by applying a small ripple approximation to the inductor, the total charges flowing to C_1 and C_2 are

$$Q_{C1} = Q_{C2} = (1-D)TI_{L,DC}, \quad (2)$$

where $I_{L,DC}$ is the DC average of the i_L , and T is the switching period. i_L decays as $V_{L,\Phi2} = V_{IN} - 3V_{OUT}$ during Φ_2 , while the charging/discharging states for C_1 and C_2 are opposite to Φ_1 . According to the above operations, three current-conducting branches (through L , C_1 , and C_2) are connected to the output during Φ_1 , realizing a multi-path output-charge delivery. In contrast to single-path structures, the proposed scheme effectively reduces the $I_{L,DC}$. Based on charge conservation and the expression in (2), the $I_{L,DC}$ is derived as below:

$$\begin{aligned} TI_{L,DC} + Q_{C1} + Q_{C2} &= Q_{OUT} \\ \Rightarrow I_{L,DC} &= \frac{I_{LOAD}}{3-2D} = MI_{LOAD}. \end{aligned} \quad (3)$$

From above, the inductor current can be reduced by a factor of M when compared with the existing inductor-at-output topologies.

Fig. 3 exhibits the conceptual waveforms for the steady-

state operations of the proposed topology. According to the information of $V_{T1} - V_{T2}$ in either Φ_1 or Φ_2 , the inductor current ripple Δi_L has the following form:

$$\Delta i_L = \frac{(V_{IN} - V_{OUT})DT}{L} = \frac{TV_{OUT}(3M-1)(1-M)}{L \cdot 2M^2}. \quad (4)$$

From (4), the peak value of the Δi_L occurs at $M = 0.5$, and the corresponding maximum inductor current ripple amplitude equals $0.5TV_{OUT}/L$. Thus, in the conversion range from $1/3$ to $1/2$, the proposed topology achieves a lower inductor current for DC and AC components than a buck.

2.2 Conduction Loss Analysis

Under heavy load conditions, the conduction losses from inductor DCR and switches R_{ON} dominate the converter efficiency [26–27]. The corresponding theoretical loss analysis is as below

$$I_{L,rms} = \sqrt{\frac{1}{T} \int_0^T i_L^2(t) dt} = \sqrt{I_{L,DC}^2 + \frac{\Delta i_L^2}{12}}. \quad (5)$$

By substituting (3) into (5), we can derive the loss power expression accordingly, i.e.,

$$\begin{aligned} P_{loss,L} &= I_{L,rms}^2 R_{L,DC} \\ &= R_{L,DC} \left\{ M^2 I_{OUT}^2 + \frac{1}{12} \left[\frac{V_{OUT}(3M-1)(1-M)}{Lf_s} \right]^2 \right\}. \end{aligned} \quad (6)$$

From (6), the DC part of the conduction loss with the MPSCI is M^2 times smaller than that of the inductor-at-output converters. Besides, the AC part of the conduction loss due to the inductor current ripple decreases with M . From Fig. 3, the charge amount $Q_{C1,2}$ flowing through the capacitors $C_{1,2}$ equals to the Q_L during Φ_2 , estimated as

$$Q_{C1,2} = (1-D)TI_{L,DC} \quad (7)$$

According to the steady-state charge conservation with the capacitor, the average current through the $C_{1,2}$ during Φ_1 is

$$I_{C1,2,\Phi_1} = \frac{Q_{C1,2}}{DT} = \frac{(1-D)}{D} I_{L,DC} \quad (8)$$

Based on (8) and the operating states shown in Fig. 2(b), we can estimate the passing current through different switches in each phase, as:

$$\Phi_1: \begin{cases} I_{RS1} = \frac{1}{D} I_{L,DC} \\ I_{RS3} = \frac{(1-D)}{D} I_{L,DC} \\ I_{RS4} = \frac{1}{D} I_{L,DC} \\ I_{RS5} = \frac{(1-D)}{D} I_{L,DC} \end{cases} \quad (9)$$

$$\Phi_2: \begin{cases} I_{RS2} = I_{L,DC} \\ I_{RS6} = I_{L,DC} \end{cases} \quad (10)$$

where I_{RSi} represents the passing current through the switch S_i . The overall switch conduction loss expression is:

Table I Topology modeling parameters

V_{IN}	f_{SW}	I_{OUT}	L	C_F	C_{OUT}	R_{ON}	DCR
5V	500kHz	2A	4.7μH	10μF	20μF	10mΩ	250mΩ

Fig. 4 Modeling results and comparison between the proposed and the existing topologies on (a) conduction-loss-only conversion efficiency and (b) the conduction loss power using the parameters given in Table I.

$$\begin{aligned} P_{loss,SW} &= D(I_{RS1}^2 R_{S1} + I_{RS3}^2 R_{S3} + I_{RS4}^2 R_{S4} + I_{RS5}^2 R_{S5}) \\ &\quad + (1-D)(I_{RS2}^2 R_{S2} + I_{RS6}^2 R_{S6}) \\ &= I_{L,DC}^2 \left[\frac{1}{D} (R_{S1} + R_{S4}) + \frac{(1-D)^2}{D} (R_{S3} + R_{S5}) + (1-D)(R_{S2} + R_{S6}) \right] \end{aligned} \quad (11)$$

where R_{Si} denotes the switch-on resistance of the switch S_i , as shown in Fig. 2. Based on the above theory, we can establish an analytical model of the conduction loss power to evaluate and compare the performance of different topologies. Using the corresponding parameters given in Table I, we can obtain the modeling results, as displayed in Fig. 4, for the proposed MPSCI, existing DP, and the conventional buck topologies. From the highlighted region in the figures, the efficiency and power loss difference between the proposed and the existing topologies become more noticeable as the VCR enlarging.

3. Converter design and implementation

The overview implementation of the proposed MPSCI converter system composes of the power stage, gate drivers, and the feedback control loop, as shown in Fig. 5. All the power switches are based on 5-V transistors in a 180-nm CMOS process to ensure operating reliability. Regarding the power passive selection, the capacitance of C_1 and C_2 is 10μF, and C_{OUT} is 20μF with an equivalent series resistance (ESR) of 10mΩ for each. The converter adopts an inductor of 4.7μH with DCR=240mΩ and a dimension of 2.5mm × 2mm × 1mm. It features regulating a V_{OUT} of 1.8~3.3V from an input of 5V based on voltage-mode PWM control. A typical type-III compensated error amplifier (EA) is adequate for stabilizing the load regulation control loop. For proper switch on/off operations, the gate driving voltage for all the switches is 5V. Except for the input switch S_1 , which is implemented by a PMOS transistor to ease the driving

Fig. 5 Implementation overview of the proposed converter system.

Fig. 6 Circuit implementation of the level shifters adopted in the proposed converter for switch control.

requirement, all the other switches are implemented by NMOS for better conductance and lower silicon area.

We designed the gate driver circuits in three categories for different power switches according to the individual operating voltage conditions. Specifically, S_2 and S_5 are GND-connected switches, an inverter-based buffer chain with a stage size increase by a factor of 10 that ensures an effective fan-out is applicable for the gate driving. A starving-resistor-based buffer is used to drive the last stage of the buffer chain to prevent shoot-through current. Due to the fixed-DC-terminal property with the operation of S_1 and S_4 , charge pump-based level shifters (CPLS), as given in Fig. 6, with different types of cross-coupled transistors can be used to fulfill the driving requirements [28 29 30]. Considering S_1 as an example, the generated turn-on and -off gate-driving voltage are $V_{IN} - V_{DD}$ and V_{IN} , respectively. A similar control principle applies to S_4 . Since S_3 and S_6 are floating switches (i.e., with no terminal connected to a DC voltage), bootstrapping driving techniques are necessary to generate floating voltage domains for the individual switches. The junction isolation level shifter (JILS), as shown in Fig. 6, shifts up the clock signal from $0 \sim V_{DD}$ to $V_{SSH} \sim (V_{SSH} + V_{ch})$ domain, $V_{ch} = V_{DD}$ according to the node switching voltages of V_{B1} and V_{B2} .

Fig. 7 Complete loss breakdown comparison between the proposed MPSCI and conventional buck converters under a specific operating condition of $V_{OUT} = 1.8V$ and $I_{OUT} = 3A$.

4. Simulation results

The converter design validation is through full-transistor-level simulation at a switching frequency of $f_{sw} = 800kHz$. The concerned losses include inductor DCR loss, switch R_{ON} conduction loss, switching loss, switch gate driving loss, and shunt loss from the control building blocks. Considering the line-conduction loss caused by the practical parasitic effect, each power switch is in series with a line impedance of $\sim 10m\Omega$. A simulated loss-power breakdown is summarized in Fig. 7 for a 5V-to-1.8V conversion with an output current of 3A. It validates that the proposed MPSCI converter achieves improved overall efficiency compared with a buck converter implemented with the same total switch area and power inductor. From the pie chart, in contrast to the buck, the DCR loss reduction in the proposed converter is significant, and hence it attains a higher efficiency under a heavy load condition.

Fig. 8 shows the simulated efficiency of the proposed MPSCI converter under different loading conditions and VCRs. From Fig. 8(a), this design ensures a PCE above 90% over a conversion range of 5V to 1.8-3.3V under 1A loading current. The peak efficiency can be up to 93.7%. When delivering a current of 2A, the total conduction loss increases, and the converter can still reach a 90% peak efficiency and maintain the overall PCE above 86%. Targeting on a maximum loading of 3A, the simulated result demonstrates a peak efficiency of 85.8%. Fig. 8(b) displays the simulated curves of PCE and DCR loss power together with a comparison between the proposed converter and the existing ones under a continuous loading current range and different V_{OUT} of 1.8 and 2.5V. In contrast to the existing DP and conventional buck converters, the proposed MPSCI realized a DCR loss reduction up to 3.3 and 5.4 times and hence a PCE benefit of 7.1% and 15%, respectively, at a 1.8-V output. As the increase of the V_{OUT} level, the DCR loss power takes less proportion among the total loss. On the other hand, although the relative DCR loss advantage attained by the proposed topology shrinks as the increase of the D, according to (3), it still realize a PCE advantage up to 2.3% and 11%, and a DCR loss reduction up to 1.5 and 3

Fig. 8 Simulated results and comparison between the proposed MPSCI and the existing topologies on (a) the efficiency versus V_{OUT} under different I_{OUT} and (b) the efficiency and DCR loss versus I_{OUT} under different V_{OUT} .

times when compared with the existing DP and conventional buck converter, respectively.

Fig. 9 shows the frequency-domain characteristics of the proposed converter. From the Bode plot, the power stage double poles locate around 16kHz, and the loop phase margin is 54° after the compensation. In contrast, the power stage phase margin is only 7° in the open-loop condition. Accordingly, the type-III compensator is adequate for stabilizing the voltage-mode control loop of the implemented converter system.

Fig. 10 depicts the time-domain transient waveforms of the proposed converter under a step response with the loading current from 0.5 to 3A. A $\sim 10\%$ V_{OUT} overshoot/undershoot happens during the step-current transition and the recovery time is about $35\mu s$. As observed, during the dead-time, the body diode of the power MOSFETs in the inductor current path is in forward conducting to provide a freewheeling path for the I_L , causing the diode voltage drop observed in the waveform. Besides, the V_{OUT} ripple is about 30mV, which is majorly due to the ESR influence mainly.

The overall performances of the proposed converter are summarized in Table II and compared to the state of the art. From the table, this work achieves high efficiency and a high loading current level using a large-DCR inductor. In addition, the proposed topology achieves a higher inductor current reduction theoretically under a high step-down condition and hence demonstrates an advantage of delivering heavier

Fig. 9 Frequency-domain characteristics of the proposed converter in open and closed-loop conditions.

Fig. 10 Simulated time-domain waveforms of proposed converter with step responses of the output current as well as the steady-state power stage internal waveforms.

Table II Performance summary and comparison

	ISSCC'15 [15]	APEC'17 [2]	ISSCC'17 [14]	JSSC'19 [20]	This Work
Process	0.35um	Discrete	28nm	0.18um	0.18um
Topology	Buck	S-hybrid	3-Level	DPDC	MPSCI
V_{IN} [V]	2.7-4.5	5	3-4.5	4.5	5
V_{OUT} [V]	2	3.3-3.8	0.8-1.45	0.8-4	1.8-3.3
Max. output current [A]	1	3.9	1	1.6	3
f_{SW} frequency [MHz]	1	2	3	1	0.8
I_{LDC} reduction	No	Yes $I_L = I_{OUT}/(2-D)$	No	Yes $I_L = I_{OUT}/(2-D)$	Yes $I_L = I_{OUT}/(3-2D)$
# of discrete components	1L, 1C (4.7uH, 10uF)	USB 3.0 Cable, 4C (10 $\mu F \times 4$)	1L, 1C (2.2uH, 2uF)	1L, 2C (4.7uH, 10uF $\times 2$)	1L, 3C (4.7uH, 10uF $\times 2$, 20uF)
Inductor DCR	38m Ω	141.4m Ω	NA	250m Ω	240mΩ
Efficiency @ 1A loading	88.5%	88.0%	80.0%	94.0%	93.7%
Full load efficiency	—	84% @ 3.9A	—	92% @ 1.6A	85.8% @ 3A

loading current. It features significance for the efficiency optimization, owing to the improvement of the DCR loss.

5. Conclusion

This letter proposes a hybrid DC-DC converter technique that features multi-path output-charge conduction to achieve effective inductor current reduction and, therefore, a lower DCR loss. This property ensures the possibility of adopting a large-DCR power inductor with a shrunk volume in high-output-current delivery scenarios. A systematic analysis, design, and implementation methodology have been

discussed. Through the simulation validation, in contrast to the existing techniques, the proposed topology demonstrates an extended VCR range down to 1/3 and a significant efficiency advantage when delivering a loading current up to 3A. Furthermore, the converter implementation attains a peak efficiency of 93.7% at 1A and 85.8% at 3A conditions.

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