LETTER

A 12V-to-1V switched-capacitor-assisted hybrid converter with dual-path charge conduction and zero-voltage switching

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Abstract This paper proposes a direct 12V-to-1V hybrid converter using a switched-capacitor-assisted (SCA) topology, featuring dual-path output current conduction to lower the inductor current. In contrast to the existing dual-path structures, the proposed topology realizes a high step-down conversion with efficiency improvement in heavy-load conditions. Besides, the converter further achieves an efficiency boost in light-load conditions using zero-voltage switching. Designed in a 180-nm SOI BCD process, the proposed converter regulates an output of 0.7~1.3V from a 12-V input with a peak efficiency of 92%. Under a loading current of 2A, it attains an efficiency over 84% at the 1-V output. Compared with the conventional buck, it achieves an improvement of 9.8% in peak efficiency and 13.6% in maximum-loading efficiency.

Key words: hybrid buck converter, zero-voltage switching, high step-down, dual-path, DCR loss reduction

Classification: Power devices and circuits

1. Introduction

Conventional buck converters with compact volume suffer from significant power loss on the inductors, as well as heating problems due to the considerable intrinsic DC resistance (DCR) in high-power-density systems [1]. This issue can negatively impact the conversion efficiency, especially under a heavy loading current. To overcome the hard-tradeoff between the conversion efficiency and the size of the power components, different types of switched-capacitor (SC)-inductor hybrid converter topologies [2, 3, 4, 5, 6, 7, 8, 9, 10] have been proposed, targeting on reducing the inductor DCR loss through lowering the steady-state inductor DC current ($I_{L,DC}$). In [2, 3, 4], the multi-branch architectures using multiple inductors (e.g., double-step-down (DSD) topology) feature equal current flow in each branch, realizing a reduced total DCR loss due to the quadratic correlation between the inductor current and the

Fig. 1 The proposed SCA hybrid converter topology.

Fig. 2 Steady-state operations of the proposed topology.

power loss. However, increasing the number of the inductor can inevitably raise the system volume and cost. Although the topologies reported in [5, 6, 7, 8] can achieve $I_{L,DC}$ reduction without using additional inductors, they suffer from a limited voltage conversion ratio (VCR) range from 3:1 to 1:1, and therefore cannot fulfill the high step-down VCR scenarios, e.g., 12:1. The SC hybrid dual-path (DP) topologies presented in [9, 10] realize an extended VCR range together with $I_{L,DC}$ reduction, which unfortunately degrades as VCR increases, indicating their incompatibility for high step-down application [11, 12, 13, 14, 15] (i.e., the DCR loss can still dominate the efficiency under a high VCR).

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This paper proposes an SC-assisted (SCA) hybrid buck topology (as shown in Fig. 1) for efficient 12V-to-1V DC-DC conversion. Featuring enhanced inductive-capacitive DP output conduction in high step-down conversion, the proposed converter can significantly alleviate the current conduction loss on the power inductor. In contrast to a conventional buck converter, the proposed SCA hybrid buck topology can relax the duty ratio (D) requirement at a high step-down VCR, facilitating high-frequency operation for component size reduction while alleviating the control loop bandwidth requirement. The possibility of employing low-voltage transistors as the high-current power switches at the output for the proposed topology can also effectively reduce the switching loss and save the chip area. Apart from that, it also retains the advantage of full-range step-down VCR as the buck converter by grounding the inductor terminal in the two-phase operations, resolving the VCR limitation when compared with [5, 6, 7, 8]. Finally, the employment of zero-voltage-switching (ZVS) [16, 17, 18, 19, 20] in the light-load condition can effectively alleviate the hard switching loss and body-diode conduction loss of the HV power transistors, which further improves the light-load efficiency.

2. Proposed topology

The proposed SCA topology consists of a single inductor (L), single flying capacitor (C_F), and five switches, as shown in Fig. 1. Fig. 2 illustrates its steady-state operations in two phases (Φ_A and Φ_B). During Φ_A , with a phase duration of DT (where T is the switching period), S_1 , S_3 and S_5 are turned on while leaving S_2 and S_4 turn-off. Accordingly, L is magnetized with an average voltage difference of $(V_{IN} - V_{OUT})$ across it. Meanwhile, C_F is in parallel with C_{OUT} during its discharging phase. As a result, L and C_F together deliver charge to the converter output, resulting in $Q_{OUT,A} = Q_{L,A} + Q_{C,A}$. During the following phase Φ_B , with a duration of $(1-D)T$, all the switches invert the operating states, i.e., S_2 and S_4 are closed, while S_1 , S_3 and S_5 are open. L is demagnetized with $V_L = -(V_{OUT} + V_{CF})$, and C_F is charged by the inductor current I_L , which supplies the total output current during Φ_B , i.e., $Q_{OUT,B} = Q_{L,B}$. Due to the direct connection of C_F between V_{OUT} and GND during Φ_A , the average DC voltage across C_F is $V_{CF} = V_{OUT}$. Based on the conditions above, the VCR (described by M) expression can be derived according to the voltage-second balance of L in a switching period T , given as:

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

Fig. 3 Modeling comparison between the existing and the proposed topologies on (a) normalized inductor DC current, and (b) normalized conduction loss.

Since the inductor current I_L performs the C_F charging during Φ_B , under a small ripple approximation to the I_L (i.e., $I_L = I_{L,DC}$), the conducted charge by C_F in each phase is $Q_{CF} = (1-D)TI_{L,DC}$. Accordingly, the total delivered output charge in a whole period of T is: $Q_{OUT} = (2-D)TI_{L,DC}$, based on which the $I_{L,DC}$ can be derived as:

$$I_{L,DC} = \frac{I_{OUT}}{2-D} = \frac{M+1}{2} I_{LOAD}. \quad (2)$$

The above expression indicates that the $I_{L,DC}$ is suppressed by a factor of $(2-D)$ compared with that in a buck converter, which obviously reduces the DCR-related loss. In addition, the $I_{L,DC}$ reduction becomes more significant as the VCR shrinks, facilitating improving the high step-down conversion efficiency.

Figure 3 shows the theoretical modeling results regarding $I_{L,DC}$ (normalized to the I_{OUT}) and the total conduction loss (normalized to the buck converter loss) of the proposed SCA hybrid topology together with a comparison with that of an existing 3-level [21, 22, 23, 24, 25] and DSD topologies. The DCR is set to be $32R_{ON}$, which represents the switch on-impedance, according to the practical consideration of using a compact-volume inductor. From the results in Fig. 3(a)(b), in contrast to the 3-level structure, the proposed technique achieves evident advantages on $I_{L,DC}$ reduction and conduction loss over a wide VCR range. Especially, the loss reduction rate can be up to 58% under the VCR of 12:1. Although the DSD realizes effective I_L reduction with the

cost of doubling the number of inductors, the proposed topology still exhibits a conduction loss benefit up to 9% improvement in 12:1 conversion, as displayed in Fig. 3.

The state-space averaging (SSA) method is used to obtain the transfer function of the proposed topology. Based on the two-phase operation in Fig. 2, three equations can be obtained according to the current and voltage relations of the capacitors and the inductor. In Φ_A

$$C_O \frac{dV_{OUT}(t)}{dt} = \frac{V_{CF}(t) - V_{OUT}(t)}{r} - \frac{V_{OUT}(t)}{R_{OUT}} \quad (3)$$

$$C_F \frac{dV_{CF}(t)}{dt} = I_L(t) - \frac{V_{CF}(t) - V_{OUT}(t)}{r} \quad (4)$$

$$L \frac{dI_L(t)}{dt} = V_{IN}(t) - V_{CF}(t) \quad (5)$$

where R_{OUT} is the output load resistance, and r is the parasitic resistance of the flying capacitor path including the series parasitic resistance (ESR), on-resistance and so on. By defining the state vector and the input vector as

$$x = \begin{bmatrix} V_{OUT} \\ V_{CF}(t) \\ I_L(t) \end{bmatrix} \text{ and } u = [V_{IN}(t)]$$

respectively, the above equations can be presented in matrix form as:

$$\Phi_A : \dot{x} = A_1 x + B_1 u \quad (6)$$

where

$$A_1 = \begin{bmatrix} -\frac{1}{(r+R_{OUT})C_{OUT}} & \frac{1}{rC_{OUT}} & 0 \\ \frac{1}{rC_F} & -\frac{1}{rC_F} & \frac{1}{C_F} \\ 0 & -\frac{1}{L} & 0 \end{bmatrix}, \quad B_1 = \begin{bmatrix} 0 \\ 0 \\ \frac{1}{L} \end{bmatrix}$$

and $\dot{x}$ are the derivative matrices of x . Similar operation in Φ_B gives:

$$\Phi_B : \dot{x} = A_2 x + B_2 u \quad (7)$$

$$\text{with } A_2 = \begin{bmatrix} -\frac{1}{R_{OUT}C_{OUT}} & 0 & \frac{1}{C_{OUT}} \\ 0 & 0 & \frac{1}{C_F} \\ -\frac{1}{L} & -\frac{1}{L} & -\frac{r}{L} \end{bmatrix} \quad B_2 = \begin{bmatrix} 0 \\ 0 \\ 0 \end{bmatrix}$$

Based on (6) and (7), there is an averaged state-space equation, derived as

$$\begin{aligned} \dot{x} &= (DA_1 + (1-D)A_2)x + (DB_1 + (1-D)B_2)u \\ &= Ax + Bu \end{aligned} \quad (8)$$

By applying the perturbation analysis, the small-signal state equation can be obtained

$$\hat{\dot{x}} = A\hat{x} + b\hat{u} + E\hat{d} \quad (9)$$

where $E = (A_1 - A_2)x_0 + (B_1 - B_2)u_0$ (perturbed quantities are the hatted letters). x_0 and u_0 are the state

Fig. 4 System implementation of the proposed converter.

vector and input vector in steady-state, respectively, derived from (1) and (2). Accordingly, there is the transfer function (G_{vd}) from a duty input to the system output by applying Laplace transform to (9), as depicted below:

$$SX(s) = AX(s) + Ed(s). \quad (10)$$

Notice that the perturbed input vector does not affect the G_{vd} derivation and hence can be ignorable here. Based on (10) and assume $r = 0$, we can derive the expression of $G_{vd}(s)$ as given below:

$$\begin{aligned} G_{vd}(s) &= \frac{\hat{V}_{out}(s)}{\hat{d}(s)} \Big|_{\hat{u}=0, r=0} \\ &= \frac{V_{IN}[2(D-2)^2 R_{OUT} L s^2 + L s + R_{OUT}(D-2)^2]}{(D-2)^2 [(C_F + C_O) R_{OUT} L s^2 + L s + R_{OUT}(D-2)^2]} \end{aligned} \quad (11)$$

from which the pole-zero location is resemble to that in a buck converter, except for a right-half-plane (RHP)-zero which locates in high frequency domain in this design. As a result, conventional compensation strategies (e.g., type III compensation) for the feedback error amplifier (EA) are applicable to this design.

3. Circuit implementations and validation

Fig. 4 shows the system implementation overview for the proposed SCA hybrid buck converter. Implemented in a 180nm SOI BCD process, power switch S_1 and S_2 employ high-voltage (HV) LDMOS to withstand a V_{IN} up to 12V DC. $S_3 \sim S_5$ are all implemented using 1.8-V core devices according to the V_{OUT} design target of 0.7~1.3V. Regarding the switch control of S_2 and S_5 , direct driving by the buffered clock signals is feasible under the 5V and 1.8V supplies. The driving of S_1 and S_4 bases on the designed floating domain bootstrapping drivers to ensure proper switch-on/off. This

Fig. 5 Steady-state waveforms of the power stage.

Fig. 6 Key waveforms during ZVS operation.

work includes a customized cross-coupled charge pump design for the gate driving of S_3 [26, 27], as shown in Fig. 4, according to its fixed source-terminal voltage at V_{OUT} . The designed converter adopts a voltage-mode pulse-width modulation (PWM) strategy for its closed-loop load regulation control with variable clock dead-time generation [28, 29] to attain ZVS operation in lighter loading conditions. The variable clock dead-time generator consists of two 3-bit digital multiplexers for fine-grained dead-time adjustment. The first 3 bits are for the dead-time control during the V_{SW} charging phase, while the second 3 bits controls the dead-time during the discharging phase of the V_{SW} . As a results, it realizes soft switching-on operations with both high-side and low-side HV switches, and eliminates their body diodes conduction losses.

The passive components implementation in this design includes: $L = 1.5\mu\text{H}$, $C_F = 20\mu\text{F}$, and $C_{OUT} = 40\mu\text{F}$. The converter operates at a switching frequency $f_s = 1\text{MHz}$ to ensure a relatively small inductance demand for the continuous-conduction mode (CCM) operation. Fig. 5 shows the full-transistor level simulation results for the steady-state transient waveforms of the inductor current, C_F voltage, and V_{OUT} , which match well with analytical calculation except for the transition voltage glitches induced by the capacitor ESR. Fig. 6 displays the key waveforms of ZVS operation under the condition of $I_{LOAD} = 0.5\text{A}$. As observed, during the dead-times, the valley current of the

Fig. 7 Closed-loop frequency response of the proposed converter under heavy-load and light-load conditions.

Fig. 8 Simulated load transient step response.

inductor that is slightly lower than 0 charges the V_{SW} node up to the V_{IN} , while the inductor peak current discharges the node to 0 during the next phase transition state. As a result, it can alleviate both the switching loss and freewheeling diode conduction loss significantly.

Fig. 7 displays the simulated closed-loop frequency response for the proposed converter under heavy- and light-load conditions. The results show that the converter loop operates with a phase margin of 40.2° and 49° at loading current $I_{LOAD} = 2\text{A}$ and 0.5A , respectively. The simulated load transient step-response shown in Fig. 8 verifies the stability of the control loop. During a I_{LOAD} rising step of 0.5A to 2A in 1ns , the voltage droop is 54mV and the recovery time is about $8\mu\text{s}$. When I_{LOAD} changes from 2A to 0.5A in 1ns , the voltage overshoot is 43mV and the recovery time is about $9\mu\text{s}$.

Fig. 9 shows the conversion efficiency versus I_{LOAD} under a fixed input-output conversion ($V_{IN} = 12\text{V}$, $V_{OUT} = 1\text{V}$) with different DCR values. For comparison, the plots also include the results of a buck converter simulated under the same parameter setting. According to the results in Fig. 9, this design achieves an efficiency up to 84.1% at $I_{LOAD} = 2\text{A}$ and $\text{DCR} = 70\text{m}\Omega$. As I_{LOAD} decreases, the efficiency rises due to the reduced conduction losses. The efficiency in a light-load range below 1A gets improved in this design through switching the converter operation to the ZVS mode, using the variable clock dead-time generator. In heavy-load

Fig. 9 Simulated efficiency versus I_{LOAD} under a fixed $V_{OUT} = 1V$ in the case of (a) $DCR = 70m\Omega$ and (b) $160m\Omega$.

Fig. 10 Simulated efficiency versus V_{OUT} under fixed I_{LOAD} .

conditions where conduction loss dominates, the ZVS operation contributes little to the overall efficiency, on the contrary, it may worsen the conduction loss due to the increased inductor current ripple. Hence, the proposed converter operation switches back to the non-ZVS mode when I_{LOAD} is higher than a specific value, i.e., 1A in this work. The results in Fig. 9 demonstrate obvious efficiency improvements in light load conditions. The converter attains an efficiency of 90% when delivering an I_{LOAD} of 0.7A, and an efficiency of 80% under a maximum load condition of 3A. By further shrinking the inductor volume, which results in an increased DCR of 160mΩ, the converter can still exhibit a peak efficiency of 87.42% at $I_{LOAD} = 0.7A$ and 80.15% at $I_{LOAD} = 2A$, with respective improvement of 9.8% and 13.6% compared to buck converter, as shown in Fig. 9(b).

Fig. 10 shows the simulated conversion efficiency with V_{OUT} varying from 0.7 to 1.3V under $I_{LOAD} = 2A$ and $I_{LOAD} =$

Table I Performance summary and comparison

	[2]	[3]	[30]	This Work
Topology	SCB	TS-DSD	3-Level	SCA
Technology	N/A	180nm	180nm	180nm
V_{IN} (V)	12	12-24	12-100	12
V_{OUT} (V)	1.2	1	10	0.7-1.3
Max I_{LOAD}	10A	3A	0.5A	3A
f_s (MHz)	10	0.1~1	2	1
Inductor (DCR mΩ)	220nH×2 (10)	560nH×2 (N/A)	1.5μH (N/A)	1.5μH (70)
# of Switches	4	7	4	5
C_F	2.2μF	1μF×2	1μF	20μF
C_{OUT}	147μF	10μF	4.7μF	40μF
ZVS Operation	No	No	Yes	Yes
η_{peak}	85%	91.2%	90.2%	92
$\eta_{max,load}$ ($V_{OUT}=1V$)	80%	75%	N/A	80%

0.7A. As observed, a peak efficiency of 92% is achieved when $V_{OUT} = 1.3V$ and $I_{LOAD} = 0.7A$. The efficiency drops as V_{OUT} decreases due to the reduction in the total output power. Table I summarizes the overall performance and the comparison with state of the art.

4. Conclusion

This article presents a direct 12V-to-1V DC-DC converter using the proposed SCA hybrid buck topology, featuring a reduced inductor current specialized for high step-down applications and, hence, a reduced DCR loss by the dual-path output current conduction property using a single inductor and single flying capacitor. It delivers an I_{LOAD} as high as 3A while still maintaining an 80% efficiency. By adjusting the clock dead-time for ZVS operations, this design achieves a peak efficiency of 92% when delivering an I_{LOAD} of 700mA at $V_{OUT} = 1.3V$. Thus, the proposed solution realizes the efficiency boost for heavy- and light-load operations under a high step-down conversion ratio.

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