

Article

Feasibility Evaluation of Capacitorless Active Switching Ripple-Suppressing Branch for Power Converters Interfacing Ripple-Sensitive Loads

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Abstract

Active ripple suppression branches (ARSBs) are widely employed in switching power converters interfacing ripple-sensitive devices such as batteries, supercapacitors, hydrogen electrolyzers, fuel cells, and photovoltaic panels. Conventional ARSBs share the main converter DC-link voltage and require inductance comparable to that of the primary power stage, resulting in high semiconductor voltage stress and bulky magnetic components. Recent studies have proposed supplying the ARSB from a lower auxiliary voltage source, significantly reducing both inductance value and semiconductor voltage ratings. This paper shows, however, that lowering the ARSB rating while keeping the series capacitance value unaltered inherently increases residual current ripple, degrading ripple-cancellation performance. It is then demonstrated that this limitation should be overcome by increasing the ARSB capacitance in inverse proportion to the rating reduction, thereby restoring ripple suppression performance. Furthermore, it is revealed that for converters operating at a fixed duty cycle, a unique operating point exists where the ARSB capacitor can be eliminated without sacrificing the ripple attenuation ability of the circuit. The resulting capacitorless implementation reduces component count, size, complexity, and cost while improving ripple suppression. Simulation and experimental results validate the theoretical analysis and confirm the feasibility and effectiveness of the proposed capacitorless open-loop operating ARSB.



Academic Editors: Michele De Santis
and Ehab Bayoumi

Received: 14 June 2026

Revised: 29 June 2026

Accepted: 2 July 2026

Published: 3 July 2026

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Keywords: switching power converter; current ripple; active ripple elimination; reduced rating system; device count

1. Introduction

Energy conversion accompanied by low switching ripple is crucial in a broad range of modern power-electronic applications. As power-electronic systems continue to penetrate virtually every sector of industry, transportation, and consumer electronics, the demand for converters capable of delivering high-quality electrical waveforms has become increasingly important. In particular, energy-storage systems based on batteries [1–10], supercapacitors [11,12], and hydrogen technologies [13–16] are highly sensitive to current and voltage

ripple, which can accelerate component degradation, reduce energy-conversion efficiency, and adversely affect overall system reliability. Repetitive ripple currents may increase conversion losses, generate additional thermal stress, and contribute to ageing mechanisms that shorten the useful lifetime of energy-storage devices. Similar concerns arise in numerous consumer and industrial applications, including LED lighting systems, power supplies, precision instrumentation, communication equipment, and other electronically regulated loads [17–19], as well as in photovoltaic generation systems, where excessive ripple may reduce the harvested energy and negatively affect maximum-power-point-tracking performance [20]. In such systems, excessive ripple not only deteriorates electrical performance but also increases electromagnetic interference (EMI), complicates compliance with electromagnetic compatibility requirements, and may shorten the operational lifetime of passive and active components. Furthermore, ripple-induced stress can adversely affect measurement accuracy, control-system performance, and overall power quality. Consequently, the development of converter topologies capable of delivering smooth current and voltage waveforms has remained an important research focus in power electronics over the past decade [21–24].

Traditionally, ripple attenuation has been achieved through passive filtering techniques employing discrete inductors, coupled inductors, or higher-order filter networks [25–31]. The fundamental principle of these approaches is to attenuate switching harmonics before they propagate to the source, load, or energy-storage interface. Although effective and widely adopted in commercial products, these solutions typically require bulky magnetic components and additional passive elements, resulting in increased converter volume, weight, and cost. The associated increase in stored magnetic energy may also limit transient performance and complicate converter integration in compact systems. Such drawbacks may become particularly restrictive in applications where power density is of major importance, including portable consumer electronics, electric vehicles, aerospace platforms, and distributed energy systems. An alternative approach consists of increasing the converter switching frequency so that the dominant ripple harmonics are shifted to higher frequencies and become easier to filter. Higher switching frequencies generally enable the use of smaller passive components and may improve dynamic response. However, this strategy may also result in elevated switching losses, increased thermal stress, greater electromagnetic emissions, and reduced overall efficiency. As a consequence, practical converter design is often governed by a compromise between ripple suppression, efficiency, power density, thermal performance, and implementation cost [32].

Among active ripple-mitigation techniques, interleaved converter architectures have become one of the most widely adopted solutions. In an interleaved converter, multiple parallel power stages operate with phase-shifted pulse-width-modulation signals, causing partial cancellation of the ripple components generated by individual phases [33]. This mechanism allows the effective ripple frequency observed at the converter terminals to increase while reducing the amplitude of low-order harmonics. Consequently, interleaving can significantly reduce output-current ripple, improve current-sharing capability, and decrease the size requirements of passive filtering components. Nevertheless, complete ripple cancellation is achieved only under specific duty-cycle conditions determined by the number of interleaved phases and their relative phase shifts. Outside these operating points, residual ripple remains present and may still require additional filtering. Furthermore, the requirement for multiple synchronized power stages increases semiconductor count, magnetic-component requirements, gate-drive circuitry, sensing hardware, and control-system complexity, thereby increasing both implementation cost and design effort [34]. The resulting increase in component count may also affect system reliability and manufacturing complexity.

To address these limitations, several alternative converter structures incorporating auxiliary ripple-cancellation mechanisms have been proposed in the literature. Notable examples include stacked converter configurations and auxiliary-leg topologies introduced in [35,36]. These solutions employ an additional active ripple suppression branch (ARSB) designed to generate a compensating current waveform that is phase-opposed to the ripple produced by the main power stage without delivering any energy. By appropriately controlling the auxiliary branch, the ripple generated by the primary converter can be substantially reduced or even cancelled before reaching the source or load. As a result, effective ripple cancellation can be achieved over the entire operating duty-cycle range rather than only at isolated operating points. This characteristic represents a significant advantage compared to conventional interleaving techniques. Since their introduction, numerous ARSB-based implementations have been reported, encompassing both single-phase and multi-phase arrangements and covering a variety of converter topologies [37–51]. These studies have demonstrated the applicability of the concept to different power levels and operating conditions while highlighting its ability to achieve substantial ripple reduction without requiring extensive passive filtering.

In its conventional form, the ARSB consists of a controlled switching leg terminated by a series LC network. The branch inductance is typically selected to match the inductance of the main converter stage, while the switching leg is connected directly to the converter high-side DC link. Consequently, ARSB switches must withstand the same voltage stress as the main power-stage devices, potentially resulting in relatively high voltage-rating requirements and increased semiconductor costs. In addition, the required inductance value may remain substantial, limiting the achievable reduction in converter size and complexity. Recently, the studies reported in [52,53] demonstrated that the ARSB switching leg can be decoupled from the high-side DC-link voltage of the main converter and supplied from an auxiliary voltage source of substantially lower magnitude. Such a modification enables a significant reduction in both the required ARSB inductance and the voltage rating of the associated semiconductor devices, thereby improving the practicality, compactness, and cost-effectiveness of incorporating a ripple-suppressing branch. The reduced-voltage concept also opens new opportunities for optimizing auxiliary-branch design and minimizing implementation overhead.

Building upon these developments, this paper further explores the concept of reduced-rating ARSB. It is shown that rating reduction deteriorates ARSB performance in terms of ripple-suppression ability when the value of the ARSB capacitor remains unchanged. This observation reveals an important trade-off between component rating reduction and ripple-cancellation effectiveness. On the other hand, it is demonstrated that the ripple-suppression capability can be fully restored if the value of the ARSB capacitor is increased inversely proportional to the reduction in ARSB inductance. This finding provides a practical design guideline for maintaining high-performance ripple cancellation while benefiting from reduced component ratings. In addition, it is demonstrated that, for operating conditions characterized by a fixed yet arbitrarily selectable duty-cycle value, the ARSB capacitor can be eliminated entirely without compromising ripple-suppression capability. The resulting capacitorless implementation preserves the fundamental ripple-cancellation mechanism while reducing component count, circuit complexity, physical volume, and implementation cost. Such simplification is particularly attractive for applications where compactness, reliability, and cost-effectiveness are of primary importance. The theoretical developments are validated through a comprehensive simulation framework, while the feasibility of the capacitorless ARSB concept is additionally verified experimentally. The obtained results exhibit excellent agreement with the corresponding analytical predictions, thereby confirming the validity of the proposed approach and its potential applicability in practical power-electronic systems.

The rest of the paper is organized as follows. Existing ARSB structures are reviewed in Section 2, followed by the derivation of the proposed simplified topology. An example is elaborated in Section 3, accompanied by simulations and experiments. The paper is concluded in Section 4.

2. Materials and Methods

Consider a synchronous bidirectional DC-DC converter shown in Figure 1 [54,55], where the voltage source V_H exchanges energy with a general ripple-sensitive device (RSD), represented by a Thevenin equivalent consisting of a voltage source E and internal impedance Z (which may also be slowly varying in general). The converter switching node and the RSD are separated by inductance L . Assuming nearly ripple-free voltages V_H and V_L (with $V_H > V_L$), the converter operates with duty cycle d given by [56]

$$d = \frac{V_L}{V_H} \quad (1)$$

at switching frequency F_S with upper and lower transistors (Q_1 and Q_2) operated in complementary fashion, as shown in Figure 1. RSD current consists of DC and switching frequency components

$$i_{RSD}(t) = I_{RSD} + \Delta i_{RSD}(t), \quad (2)$$

where the former is given by

$$I_{RSD} \triangleq F_S \int_{F_S^{-1}} i_{RSD}(t) dt = \frac{V_L - E}{Z(\omega = 0)} \quad (3)$$

and may be either positive (for $V_L > E$) or negative (for $V_L < E$) according to the power flow direction. The switching node voltage is given by

$$v_{SW}(t) = \begin{cases} V_H, & 0 < t < dF_S^{-1} \\ 0, & dF_S^{-1} < t < F_S^{-1} \end{cases} \quad (4)$$

rendering inductor current as

$$i_L(t) = I_L + \Delta i_L(t), \quad (5)$$

with DC component

$$I_L = I_{RSD} \quad (6)$$

and switching ripple constituent expressed as [57,58]

$$\Delta i_L(t) = \begin{cases} -\frac{1}{2}\Delta I_L + \frac{V_H - V_L}{L}t, & 0 < t < dF_S^{-1} \\ \frac{1}{2}\Delta I_L - \frac{V_L}{L}(t - dF_S^{-1}), & dF_S^{-1} < t < F_S^{-1} \end{cases} \quad (7)$$

with

$$\Delta I_L = \frac{V_L}{LF_S}(1 - d) \quad (8)$$

denoting the peak-to-peak value of inductor current. Obviously (cf. Figure 1 and (2)),

$$\Delta i_{RSD}(t) = \Delta i_L(t). \quad (9)$$

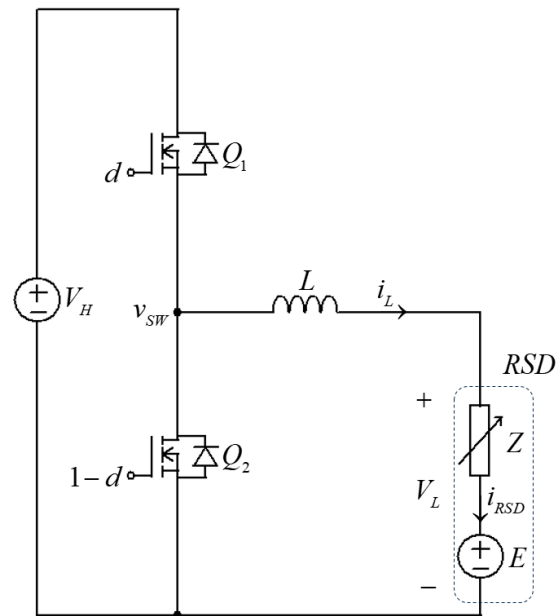


Figure 1. System under consideration.

In case the RSD current ripple restriction is defined by the maximum allowed peak-to-peak ripple value

$$\Delta I_{RSD} \triangleq |\max(i_{RSD}(t)) - \min(i_{RSD}(t))| \leq \Delta I_{MAX}, \quad (10)$$

the following must hold true (cf. (8) and (10)),

$$LF_S \geq \frac{(1-d)V_L}{\Delta I_{MAX}}. \quad (11)$$

On the other hand, inductor current value at the beginning of a switching cycle may be forced to remain below a certain negative value $-I_0$ to attain zero voltage switching of Q_1 [59], i.e.,

$$i_L(t=0) = I_{RSD} + \Delta i_L(t=0) = I_{RSD} - \frac{(1-d)V_L}{2LF_S} \leq -I_0 \quad (12)$$

must hold, imposing

$$LF_S \leq \frac{(1-d)V_L}{2(I_{RSD} + I_0)}. \quad (13)$$

It is well-evident that (11) and (13) would contradict each other if

$$2(I_{RSD} + I_0) > \Delta I_{MAX} \quad (14)$$

so that the RSD current ripple restriction and condition for zero voltage switching of Q_1 cannot be complied with simultaneously.

A common solution to the above issue is adding a high-frequency (ceramic or film) capacitance C in parallel with the RSD, as shown in Figure 2. Rewriting the inductor current switching ripple constituent (7) as a Fourier series given by

$$\Delta i_L(t) = \sum_{n=1}^{\infty} \Delta I_{Ln} \sin(2\pi n F_S t + \theta_n), \quad (15)$$

corresponding RSD current component becomes

$$i_{RSD}(t) = i_L(t) - i_C(t) = I_{RSD} + \sum_{n=1}^{\infty} \Delta I'_{Ln} \sin(2\pi n F_S t + \varphi_n) \quad (16)$$

with the capacitance addition, where

$$\Delta I'_{Ln} = \Delta I_{Ln} \left| \frac{(2\pi n F_S C)^{-1}}{(2\pi n F_S C)^{-1} + Z^{-1}(2\pi n F_S)} \right| < \Delta I_{Ln}. \quad (17)$$

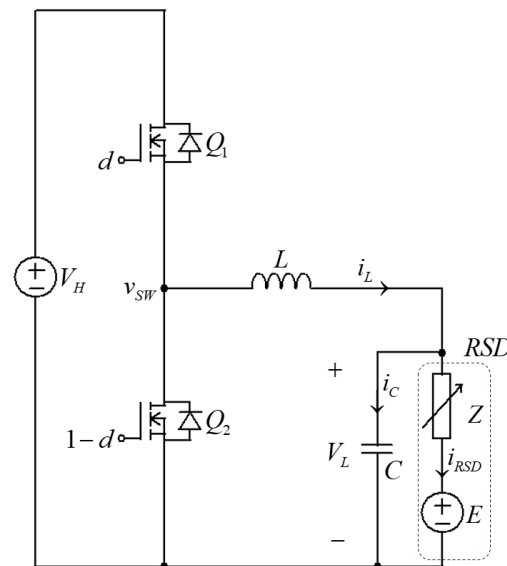


Figure 2. Switching ripple suppression by high-frequency capacitance.

Nevertheless, the value of C required for RSD current ripple restriction compliance may sometimes be too high to be implemented in practice due to the relatively low capacitance density of high-frequency capacitors. In such a case, the concept of active ripple suppression may be employed as depicted in Figure 3, where an active ripple suppression branch (ARSB) is added in parallel with the RSD. Ideally, the ARSB would realize a current sink drawing the switching constituent of the inductor current so that

$$i_{ARSB}(t) = \Delta i_L(t) \Rightarrow i_{RSD}(t) = I_{RSD} \quad (18)$$

and the capacitance C is not required. In practice, (18) is not achievable and high-frequency capacitance has to be employed in addition to ARSB. Hence, the RSD current is

$$i_{RSD}(t) = i_L(t) - i_C(t) - i_{ARSB}(t) = I_{RSD} + \sum_{n=1}^{\infty} \Delta I''_{Ln} \sin(2\pi n F_S t + \psi_n), \quad \Delta I''_{Ln} < \Delta I'_{Ln}. \quad (19)$$

In some cases, ARSB-attained attenuation would be sufficient, minimizing the utilized value of C . The commonly used ARSB topology is depicted in Figure 4, consisting of switching legs (Q_3 and Q_4) sharing the high-side voltage V_H and transistors driving signals to the main converter. Yet, the switching signals are wired to the ARSB switching leg in complementary fashion, i.e., Q_4 is driven with the same signal as Q_1 and Q_3 is driven with the same signal as Q_2 [35,36]. ARSB switching node and the RSD are separated by a series inductance-capacitance branch. According to Figure 4, the ARSB switching node voltage is given by

$$v_{SW,ARSB}^{FR}(t) = \begin{cases} 0, & 0 < t < dF_S^{-1} \\ V_H, & dF_S^{-1} < t < F_S^{-1} \end{cases} \quad (20)$$

rendering the ARSB inductor current as

$$i_{ARSB}(t) = I_{ARSB} + \Delta i_{ARSB}(t), \quad (21)$$

with corresponding DC component of $I_{ARSB} = 0$ (due to the presence of a series-connected DC-blocking ARSB capacitor) and switching ripple constituent expressed as

$$i_{ARSB}(t) = \Delta i_{ARSB}(t) = \begin{cases} -\frac{1}{2}\Delta I_{ARSB} + \frac{V_L - v_{C,ARSB}^{FR}(t)}{L_{ARSB}^{FR}}t, & 0 < t < dF_s^{-1} \\ \frac{1}{2}\Delta I_{ARSB} - \frac{V_L - v_{C,ARSB}^{FR}(t) - V_H}{L_{ARSB}^{FR}}(t - dF_s^{-1}), & dF_s^{-1} < t < F_s^{-1} \end{cases} \quad (22)$$

with

$$\Delta I_{ARSB} = \frac{(V_L - v_{C,ARSB}^{FR}(t))d}{L_{ARSB}^{FR}F_s} \quad (23)$$

denoting the peak-to-peak value of ARSB current. The voltage across the ARSB capacitance is given in general by

$$v_{C,ARSB}^{FR}(t) = \int i_{ARSB}(t)dt = V_{C,ARSB}^{FR} + \Delta v_{C,ARSB}^{FR}(t), \quad (24)$$

with the DC component obtained as

$$V_{C,ARSB}^{FR} = V_L - F_s \int_{F_s^{-1}} v_{SW,ARSB}^{FR}(t)dt = V_L - (1-d)V_H. \quad (25)$$

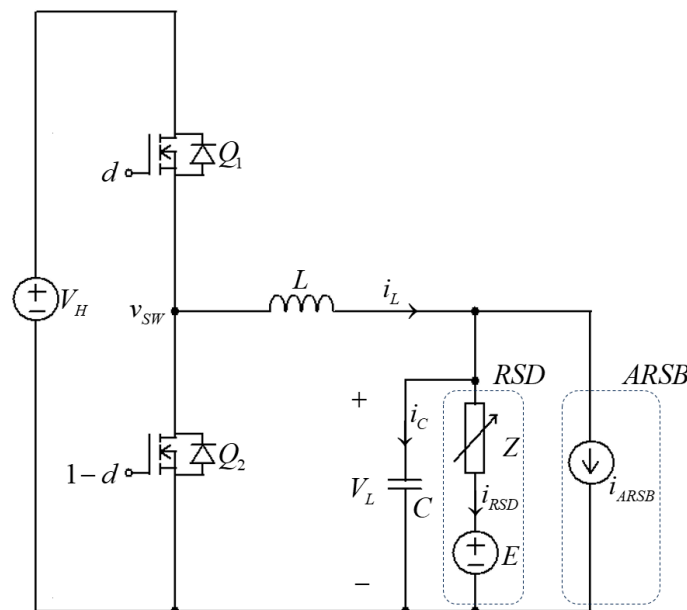


Figure 3. Switching ripple suppression by high-frequency capacitance and ARSB.

If the small ripple approximation

$$\Delta v_{C,ARSB}^{FR}(t) \ll V_{C,ARSB}^{FR} \quad (26)$$

is valid, then

$$v_{C,ARSB}^{FR}(t) \approx V_{C,ARSB}^{FR} \quad (27)$$

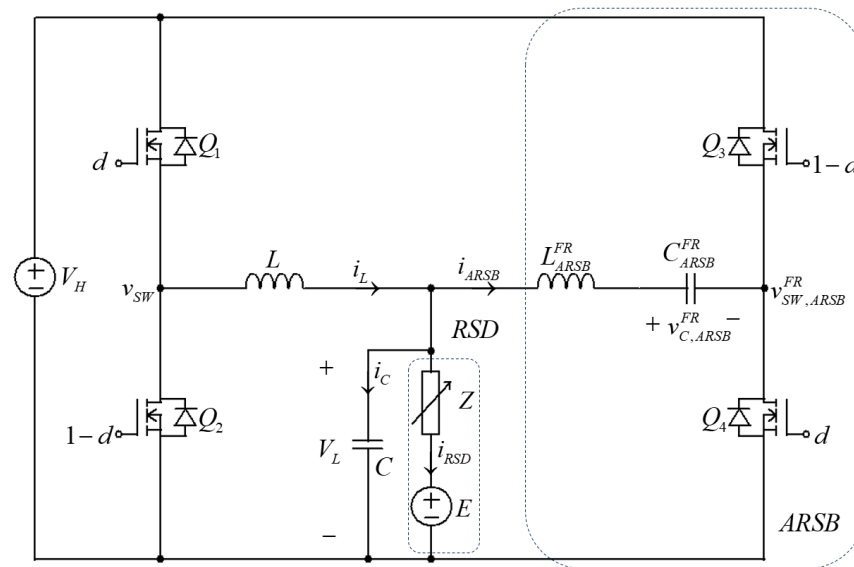


Figure 4. Full-rating ARSB topology.

Then, ARSB current in (22), (23) reduces to (cf. (1), (27))

$$i_{ARSB}(t) = \begin{cases} -\frac{1}{2}\Delta I_{ARSB} + \frac{V_H - V_L}{L_{ARSB}^{FR}}t, & 0 < t < dF_S^{-1} \\ \frac{1}{2}\Delta I_{ARSB} - \frac{V_L}{L_{ARSB}^{FR}}(t - dF_S^{-1}), & dF_S^{-1} < t < F_S^{-1} \end{cases} \quad (28)$$

with

$$\Delta I_{ARSB} = \frac{(1-d)V_L}{L_{ARSB}^{FR}F_S}. \quad (29)$$

Comparing (27) and (28) with (7) and (8) reveals that ARSB inductance should be sized to equal the main converter inductance

$$L_{ARSB}^{FR} = L \quad (30)$$

in order to attain (18). Since the voltage rating of ARSB transistors is equal to that of the main converter transistors and the value of the ARSB inductance equals that of the main converter, this topology is often referred to as “full-rating ARSB”.

It was recently realized that decoupling the ARSB switching leg DC-side from the high-side voltage of the main converter and feeding it from a separate, lower-valued voltage source $V_{ARSB} = k \cdot V_H$ (as shown in Figure 5) allows reducing both the voltage rating of ARSB transistors and the value of ARSB inductance for $k < 1$, forming the “reduced-rating ARSB”. According to Figure 5, the ARSB switching node voltage is given by

$$v_{SW,ARSB}^{RR}(t) = \begin{cases} 0 & 0 < t < dF_S^{-1} \\ V_{ARSB}, & dF_S^{-1} < t < F_S^{-1} \end{cases} \quad (31)$$

Hence, the DC component of the voltage across the ARSB capacitor is now given by

$$v_{C,ARSB}^{RR}(t) \approx V_{C,ARSB}^{RR} = V_L F_S - \int_{F_S^{-1}} v_{SW,ARSB}^{RR}(t) dt = V_L - (1-d)V_{ARSB} = V_L - (1-d)kV_H \quad (32)$$

As a result, the switching ripple constituent of the ARSB inductor current may be expressed (under the small ripple approximation of the voltage across the ARSB capacitor) as

$$i_{ARSB}(t) = \begin{cases} -\frac{1}{2}\Delta I_{ARSB} + \frac{(1-d)V_{ARSB}}{L_{ARSB}^{RR}}t, & 0 < t < dF_S^{-1} \\ \frac{1}{2}\Delta I_{ARSB} - \frac{dV_{ARSB}}{L_{ARSB}^{RR}}(t - dF_S^{-1}), & dF_S^{-1} < t < F_S^{-1} \end{cases} \quad (33)$$

with

$$\Delta I_{ARSB} = \frac{d(1-d)V_{ARSB}}{L_{ARSB}^{RR}F_S} \quad (34)$$

denoting the peak-to-peak value of ARSB current. Comparing (33) and (34) with (7) and (8) reveals that ARSB inductance should be sized to equal the main converter inductance

$$\frac{L_{ARSB}^{RR}}{L} = \frac{V_{ARSB}}{V_H} = k \quad (35)$$

in order to attain (18). Consequently, both the ARSB transistor voltage rating and ARSB inductance value undergo a k -fold reduction compared to their full-rating ARSB counterparts. As to ARSB capacitor voltage stress, there is

$$|V_{C,ARSB}^{RR}| = \begin{cases} |2V_L - V_H|, & k = 1 \\ 0, & k^* = \frac{V_L}{V_H - V_L} \\ V_L, & k = 0 \end{cases} \Rightarrow \frac{|V_{C,ARSB}^{RR}|}{V_H} = \begin{cases} |2d - 1|, & k = 1 \\ 0, & k^* = \frac{d}{1-d} \\ d, & k = 0 \end{cases} \quad (36)$$

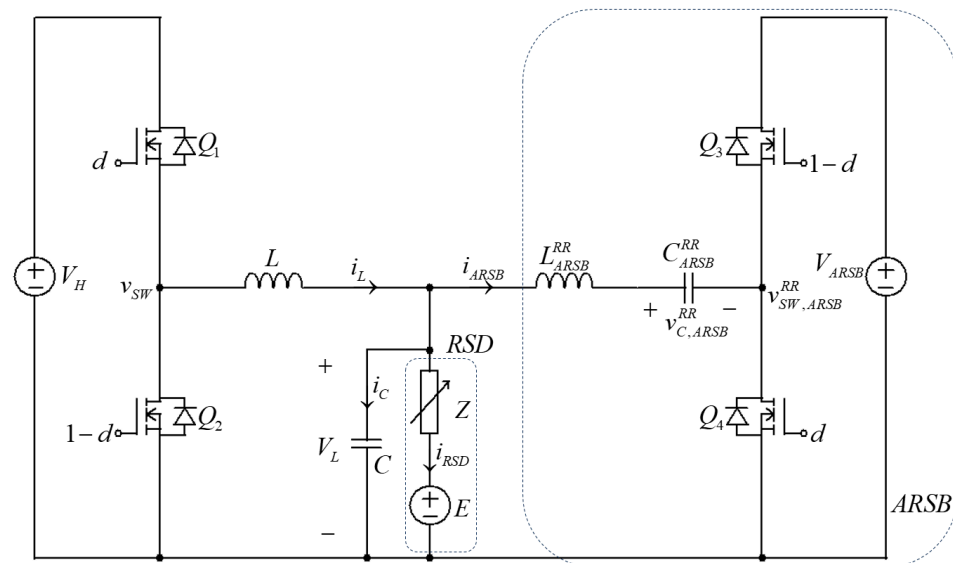


Figure 5. Reduced-rating ARSB topology.

It may be concluded that reducing k from unity to k^* reduces the voltage stress. Further reduction in k towards zero increases the voltage stress again. The normalized (w.r.t V_H) value of the DC component of the voltage across the ARSB capacitor versus k is depicted in Figure 6.

It must be emphasized that both (28) and (33) (with the former being a particular case of the latter for $k = 1$) were derived under the small ripple approximation of the voltage across the ARSB capacitor. In general, an inductor–capacitor pair imposes second-order behaviour on branch voltages and current. Referring to Figure 5, the voltage across the ARSB capacitor is governed by the following second-order differential equation,

$$C_{ARSB}^{RR}L_{ARSB}^{RR}\frac{d^2v_{C,ARSB}^{RR}(t)}{dt^2} + v_{C,ARSB}^{RR}(t) = \begin{cases} V_L, & 0 < t < dF_S^{-1} \\ V_L - kV_H, & dF_S^{-1} < t < F_S^{-1} \end{cases} \quad (37)$$

solved as

$$v_{C,ARSB}^{RR}(t) = V_L - \begin{cases} A_1 \cos(\omega_0 t) + B_1 \sin(\omega_0 t), & 0 < t < dF_S^{-1} \\ A_2 \cos(\omega_0(t - dF_S^{-1})) + B_2 \sin(\omega_0(t - dF_S^{-1})) + kV_H, & dF_S^{-1} < t < F_S^{-1} \end{cases} \quad (38)$$

with

$$\begin{aligned} \omega_0 &= \sqrt{\frac{1}{L_{ARSB}^{RR} C_{ARSB}^{RR}}}, \\ A_1 &= kV_H \frac{1 - \cos((1-d)\omega_0 F_S^{-1}) - \cos(\omega_0 F_S^{-1}) + \cos(d\omega_0 F_S^{-1})}{2(1 - \cos(\omega_0 F_S^{-1}))}, \\ A_2 &= kV_H \frac{\sin(d\omega_0 F_S^{-1}) + \sin((1-d)\omega_0 F_S^{-1}) - \sin(\omega_0 F_S^{-1})}{2(1 - \cos(\omega_0 F_S^{-1}))}, \\ B_1 &= -kV_H \frac{\sin(d\omega_0 F_S^{-1}) + \sin((1-d)\omega_0 F_S^{-1}) - \sin(\omega_0 F_S^{-1})}{2(1 - \cos(\omega_0 F_S^{-1}))}, \\ B_2 &= -kV_H \frac{1 - \cos(d\omega_0 F_S^{-1}) - \cos(\omega_0 F_S^{-1}) + \cos((1-d)\omega_0 F_S^{-1})}{2(1 - \cos(\omega_0 F_S^{-1}))}. \end{aligned} \quad (39)$$

Consequently, the ARSB current is given by

$$i_{ARSB}(t) = C_{ARSB}^{RR} \frac{dv_{C,ARSB}^{RR}(t)}{dt} = \sqrt{\frac{C_{ARSB}^{RR}}{L_{ARSB}^{RR}}} \cdot \begin{cases} A_1 \sin(\omega_0 t) - B_1 \cos(\omega_0 t), & 0 < t < dF_S^{-1} \\ A_2 \sin(\omega_0(t - dF_S^{-1})) - B_2 \cos(\omega_0(t - dF_S^{-1})), & dF_S^{-1} < t < F_S^{-1} \end{cases} \quad (40)$$

It is easy to verify that if

$$\omega_0 F_S^{-1} \ll 1 \quad (41)$$

holds, then (40) simplifies into

$$i_{ARSB}(t) = \sqrt{\frac{C_{ARSB}^{RR}}{L_{ARSB}^{RR}}} \begin{cases} A_1|_{\omega_0 F_S^{-1} \ll 1} \omega_0 t - B_1|_{\omega_0 F_S^{-1} \ll 1}, & 0 < t < dF_S^{-1} \\ A_2|_{\omega_0 F_S^{-1} \ll 1} (t - dF_S^{-1}) - B_2|_{\omega_0 F_S^{-1} \ll 1}, & dF_S^{-1} < t < F_S^{-1} \end{cases} \quad (42)$$

further reducing to (33) (and then to (28) for $k = 1$) for

$$\omega_0 F_S^{-1} = 0. \quad (43)$$

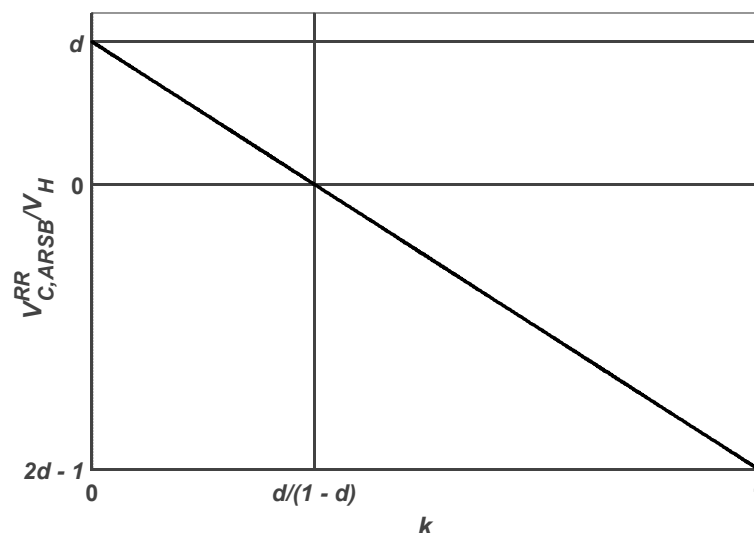


Figure 6. Normalized value of DC component of the voltage across ARSB capacitor versus k .

Obviously, (43) is infeasible and discrepancies between (40) and (33) are always present in practice, as demonstrated by simulations in the subsequent section.

Observing (36), an interesting case arises for $k = k^*$ where the switching-cycle-averaged value of ARSB switching node voltage (31) equals the voltage drop V_L across RSD and

hence the two do not have to be decoupled, yielding zero average voltage across ARSB capacitor. Consequently, ARSB capacitor does not play any essential role here. On the contrary, it imposes second-order dynamics leading to deviation between (40) and (33), as mentioned above. It is then suggested to eliminate the ARSB capacitor in case the circuit is designed for operation with constant duty-cycle yielding $k = k^*$. The resulting reduced device count circuitry is depicted in Figure 7. According to Figure 5, ARSB switching node voltage is given by

$$v_{SW,ARSB}^*(t) = \begin{cases} 0 & 0 < t < dF_S^{-1} \\ V_{ARSB}^* & dF_S^{-1} < t < F_S^{-1} \end{cases} \quad (44)$$

with

$$V_{ARSB}^* = k^* V_H. \quad (45)$$

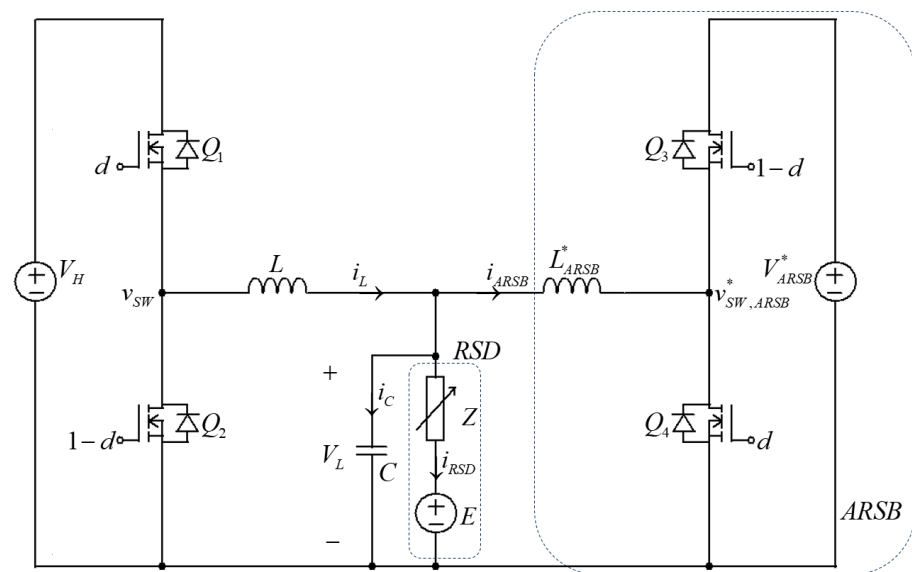


Figure 7. Reduced-rating capacitorless ARSB topology.

As a result, switching ripple constituent of ARSB inductor current may be expressed (due to ARSB capacitor absence) as

$$i_{ARSB}(t) = \begin{cases} -\frac{1}{2}\Delta I_{ARSB}^* + \frac{(1-d)V_{ARSB}^*}{L_{ARSB}^*}t, & 0 < t < dF_S^{-1} \\ \frac{1}{2}\Delta I_{ARSB}^* - \frac{dV_{ARSB}^*}{L_{ARSB}^*}(t - dF_S^{-1}), & dF_S^{-1} < t < F_S^{-1} \end{cases} \quad (46)$$

with

$$\Delta I_{ARSB} = \frac{d(1-d)V_{ARSB}^*}{L_{ARSB}^*F_S} \quad (47)$$

denoting the peak-to-peak value of ARSB current. Comparing (46) and (47) with (7) and (8) reveals that sizing ARSB inductance as

$$\frac{L_{ARSB}^*}{L} = \frac{V_{ARSB}^*}{V_H} = k^* \quad (48)$$

allows to simultaneously attain (18), reduce the device count (by eliminating ARSB capacitance) and decrease ARSB transistors voltage rating and inductance value by k^* . On the

other hand, it must be acknowledged that such a feature may only be realized in converters operating with constant duty cycle

$$d = \frac{k^*}{1 + k^*}. \quad (49)$$

3. Results and Discussion

3.1. Simulations

Consider the converter shown in Figure 1 with parameters summarized in Table 1. The RSD is represented by the voltage source of $E = 5$ V and the internal impedance Z is realized by parallel connection of 1.4Ω resistor with $100 \mu\text{F}$ capacitor, mimicking certain hydrogen electrolyzer behaviour [60]. According to (3) and (8),

$$I_L = I_{RSD} = 5 \text{ A}, \quad \Delta I_L = \frac{V_L}{L F_S} (1 - d) = 14 \text{ A} \quad (50)$$

are expected. PSIM2026 software is adopted in simulation framework. Fixed-step solver with $0.1 \mu\text{s}$ step was employed. Figure 8 presents simulated converter waveforms, accurately complying with (50).

Table 1. Synchronous buck converter parameters.

Parameter	Value	Units
V_H	48	V
V_L	12	V
I_L	5	A
d	0.25	---
F_S	100	kHz
L	6.4	μH

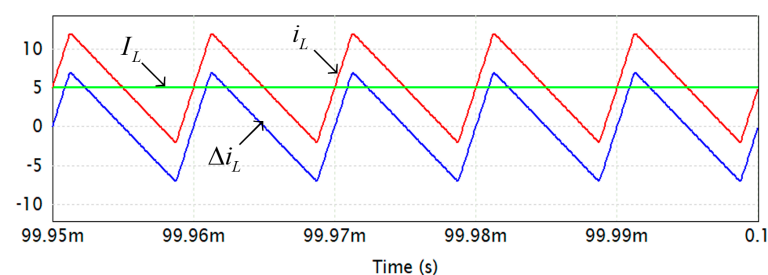


Figure 8. Current waveforms of the converter in Figure 1.

In case RSD current ripple restriction is defined by maximum allowed peak-to-peak ripple value of (cf. (10))

$$\Delta I_{MAX} = 1.4 \text{ A} = 0.1 \Delta I_L, \quad (51)$$

i.e., 10-fold switching ripple reduction is necessary, addition of high-frequency capacitance

$$C = 1000 \mu\text{F} \quad (52)$$

in parallel with the RSD would be necessary, well-validated by simulation results in Figure 9. However, 1 mF ceramic or film capacitance would be relatively expensive and corresponding physical size may be quite high. As a result, it could be more feasible to employ an ARSB discussed above.

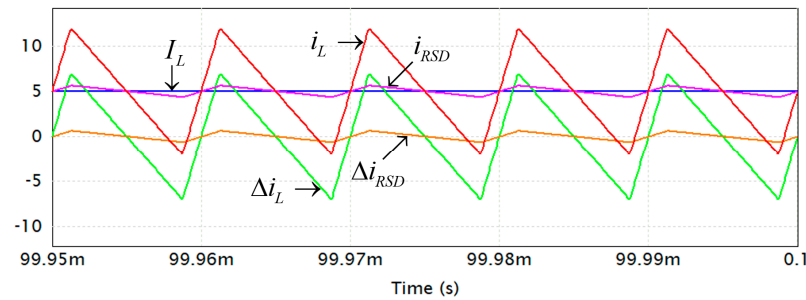


Figure 9. Current waveforms of the converter in Figure 2.

Full-rating ARSB depicted in Figure 4 is considered first. Here, the ARSB inductor is sized $6.4 \mu\text{H}$ (cf. (30)) to equal the main converter inductance. To comply with the RSD current ripple restriction (51), the ARSB capacitance of

$$C_{ARSB}^{FR} = 3.5 \mu\text{F} \quad (53)$$

is required. Corresponding simulation results are demonstrated in Figure 10. According to (25), average value of the voltage blocked by ARSB capacitance is expected to be

$$|V_{C,ARSB}^{FR}| = 24 \text{ V}, \quad (54)$$

accurately supported by the results. On the other hand, non-negligible ripple component is also visible in ARSB capacitance voltage, preventing complete elimination of RSD current ripple.

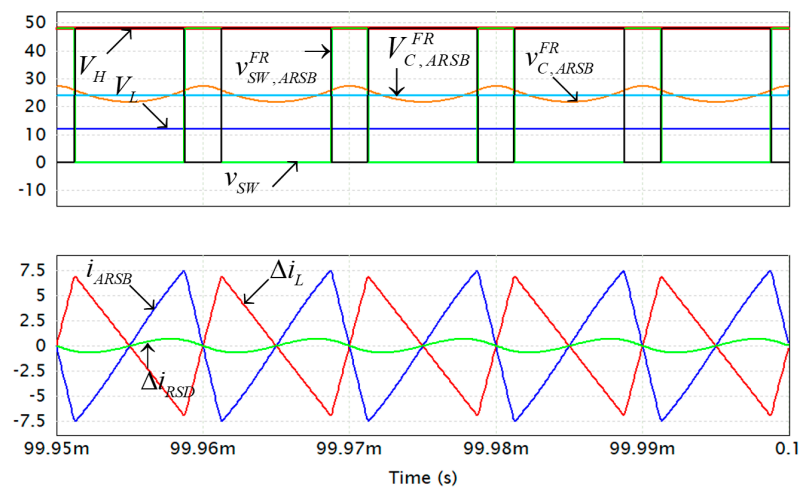


Figure 10. Waveforms of the converter in Figure 4.

Decoupling the ARSB switching leg from V_H while feeding it from a separate 24 V voltage source (cf. Figure 5) with corresponding two-fold ARSB inductance reduction, i.e.,

$$k = 0.5 \Rightarrow V_{RSB} = 0.5V_H = 24 \text{ V}, \quad L_{ARSB}^{RR} = 0.5L = 3.2 \mu\text{H} \quad (55)$$

while keeping the ARSB capacitance value (53) yields simulation results in Figure 11. It is well-evident that the circuit is incapable of complying with RSD current ripple restriction (51) since

$$\Delta I_{RSD} = 2.8 \text{ A} = 2\Delta I_{MAX}, \quad (56)$$

i.e., attaining twice the allowed peak-to-peak value. Such a behaviour was not mentioned in the literature so far. Yet, it is consistent with (40) and hence calls for increasing ARSB capacitance in order to restore the desired RSD ripple value.

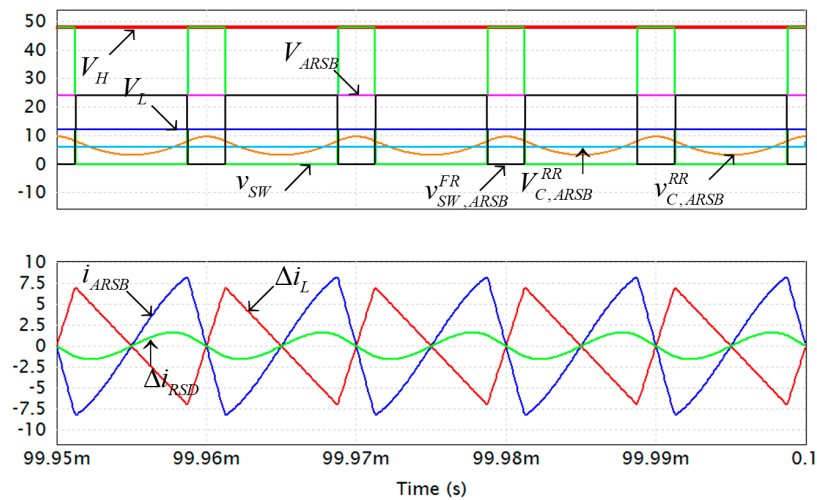


Figure 11. Waveforms of the converter in Figure 5 with $k = 0.5$ and $C_{ARSB}^{RR} = C_{ARSB}^{FR}$.

As a result, RSD current ripple was reduced to the allowed value only after increasing ARSB capacitance value to

$$C_{ARSB}^{RR} = 7 \mu\text{F} = 2C_{ARSB}^{FR}, \quad (57)$$

as shown in Figure 12, which is a two-fold increase compared to the previous case. To verify the repetitivity of the phenomenon, consider further ARSB rating reduction to

$$k = 1/6 \Rightarrow V_{RSB} = 1/6V_H = 8 \text{ V}, \quad L_{ARSB}^{RR} = 1/6L = 1.067 \mu\text{H}. \quad (58)$$

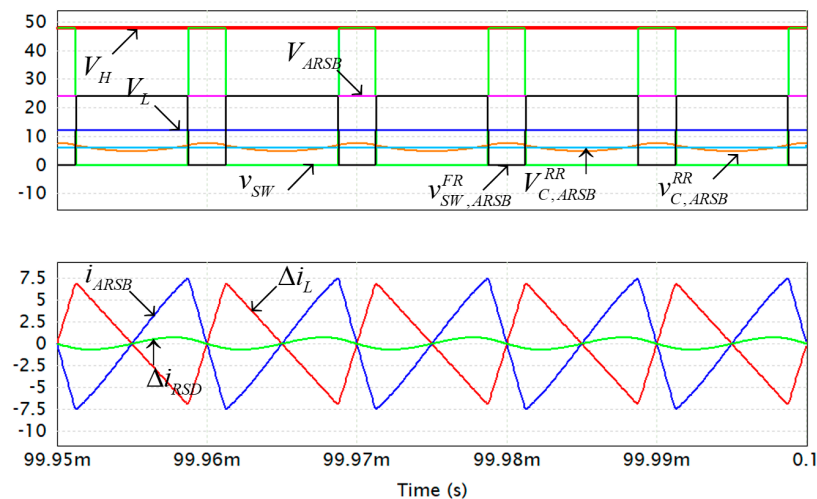


Figure 12. Waveforms of the converter in Figure 5 with $k = 0.5$ and $C_{ARSB}^{RR} = 2C_{ARSB}^{FR}$.

In case the ARSB capacitance value (53) is employed, the system attained 6-times the allowed peak-to-peak value (corresponding results are not shown for brevity). RSD current ripple was reduced to the allowed value after increasing ARSB capacitance value to

$$C_{ARSB}^{RR} = 6C_{ARSB}^{FR} = 21 \mu\text{F}, \quad (59)$$

as shown in Figure 13. It may then be concluded that in order to maintain the allowed RSD ripple while reducing ARSB rating by k , ARSB capacitance value must undergo corresponding increase, i.e., sized according to

$$C_{ARSB}^{RR} = \frac{C_{ARSB}^{FR}}{1/k}. \quad (60)$$

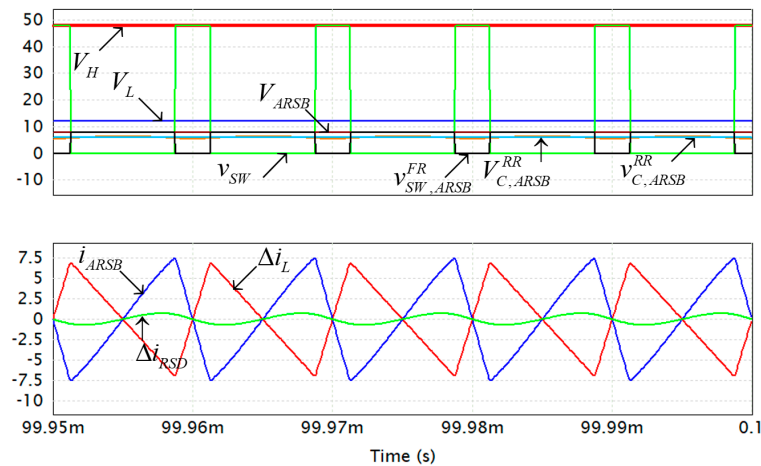


Figure 13. Waveforms of the converter in Figure 5 with $k = 1/6$ and $C_{ARSB}^{RR} = 6C_{ARSB}^{FR}$.

Such an outcome was not reported in the literature as well.

Next, consider the special case of

$$k = k^* = \frac{d}{1-d} = 1/3. \quad (61)$$

Considering (48) and (60), there is

$$V_{RSB}^* = k^* V_H = 16 \text{ V}, \quad L_{ARSB}^* = k^* L = 2.13 \text{ } \mu\text{H}, \quad C_{ARSB}^* = \frac{C_{ARSB}^{FR}}{1/k^*} = 10.5 \text{ } \mu\text{F}. \quad (62)$$

Corresponding results are given in Figure 14, indicating that the system accurately complies with the RSD current ripple restriction (51). On the other hand, the average value of the voltage blocked by the ARSB capacitance is

$$\left| V_{C,ARSB}^{RR} \right| = 0 \text{ V}, \quad (63)$$

as predicted, while some ripple is still present, preventing complete elimination of RSD current ripple. The ARSB capacitor is then eliminated from the circuit, as suggested in the preceding Section, yielding the reduced device count circuitry depicted in Figure 7. Eliminating the ARSB capacitance is equivalent to having an infinite ARSB capacitance

$$C_{ARSB}^* = \infty, \quad (64)$$

which reduces corresponding voltage ripple to zero. The results are demonstrated in Figure 15. It is well-evident that RSD current ripple reduces significantly below the required value (51)

$$\Delta I_{RSD} \ll \Delta I_{MAX}, \quad (65)$$

as predicted at the end of the preceding section.

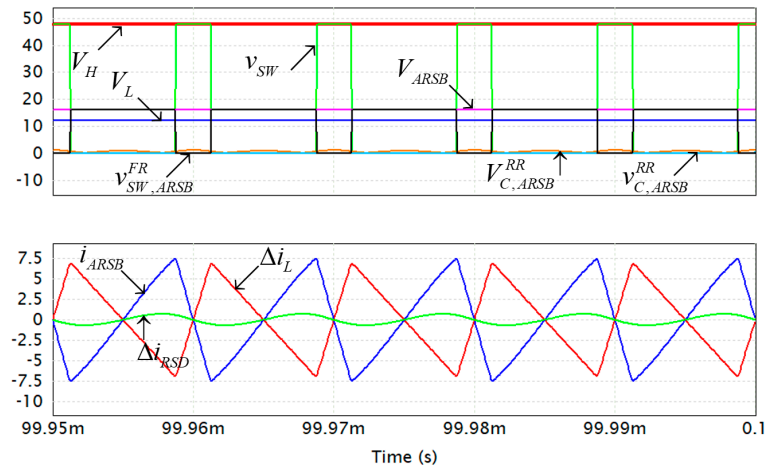


Figure 14. Waveforms of the converter in Figure 5 with $k = 1/3$ and $C_{ARSB}^{RR} = 3C_{ARSB}^{FR}$.

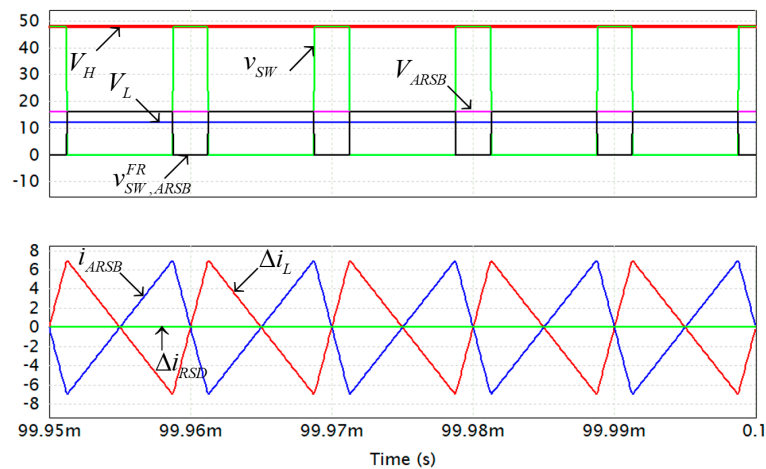


Figure 15. Waveforms of the converter in Figure 7.

3.2. Experiments

To demonstrate practical operation of the converter equipped with capacitorless ARSB (cf. Figure 7), an experimental setup matching the values in Table 1 was established (see Figure 16) with main components listed in Table 2. The TMS320F2800049 DSP Launchpad was used to drive the converter via two PWM outputs.

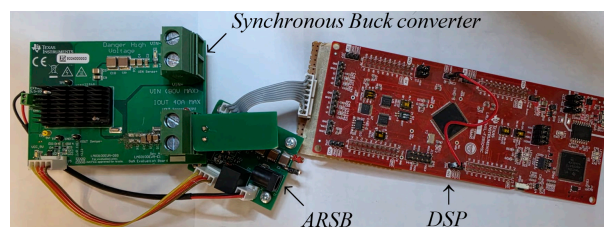


Figure 16. Experimental setup.

Three experiments were carried out. In the first, the ARSB was disabled so that the remaining circuitry matched that of Figure 1. The results are depicted in Figure 17, closely matching (50) and the corresponding simulation in Figure 8. In the second experiment, the ARSB was enabled so that the circuitry matched that of Figure 7. The results (under

both 0 A and 5 A load levels) are depicted in Figure 18. It may be concluded that the peak-to-peak RSD ripple is 2.5 times lower than the allowed value.

$$\Delta I_{RSD} = 0.58 \text{ A} \approx \frac{\Delta I_{MAX}}{2.5}, \quad (66)$$

accurately complying with the expectations under both loading extremes. In the third experiment, the input voltage was varied by $\pm 3 \text{ V}$ and the duty cycle was adjusted accordingly to keep the output voltage at the rated value to observe ripple suppression ability degradation under duty-cycle deviation. The results are depicted in Figure 19, supporting the expected performance deterioration. On the other hand, it is shown that the residual ripple still complies with the allowed value even under quite severe duty-cycle deviations from the design value.

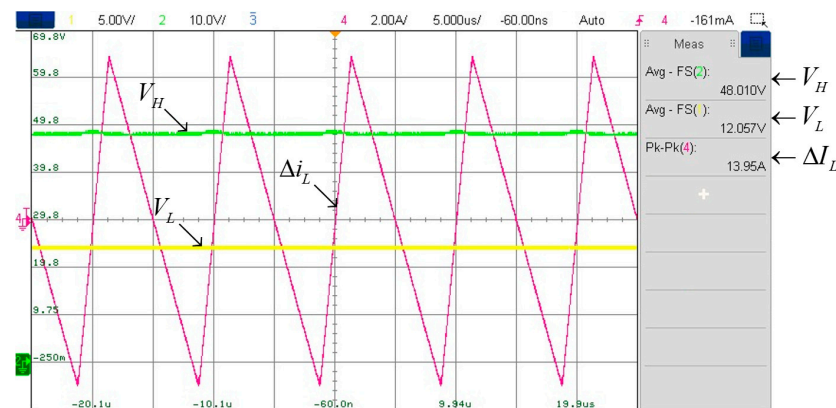
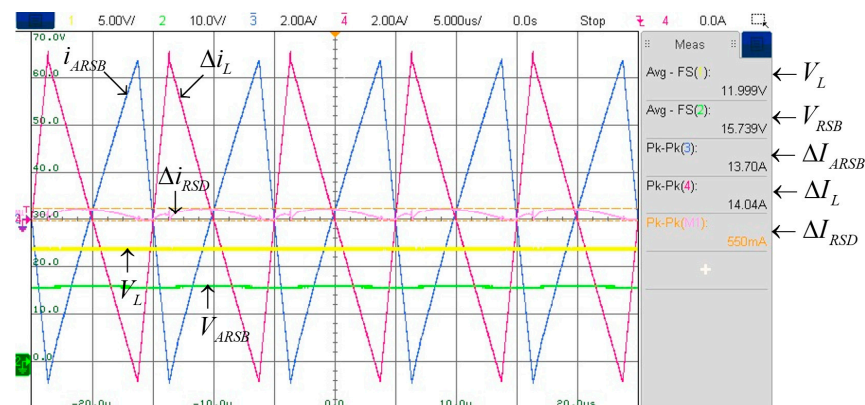
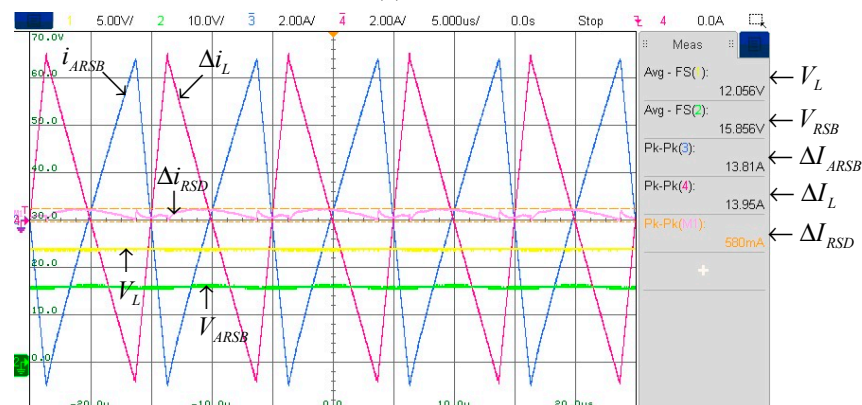


Figure 17. Experimental waveforms of the converter in Figure 1.



(a) 0 A load.



(b) 5 A load.

Figure 18. Experimental waveforms of the converter in Figure 7 under nominal duty cycle.

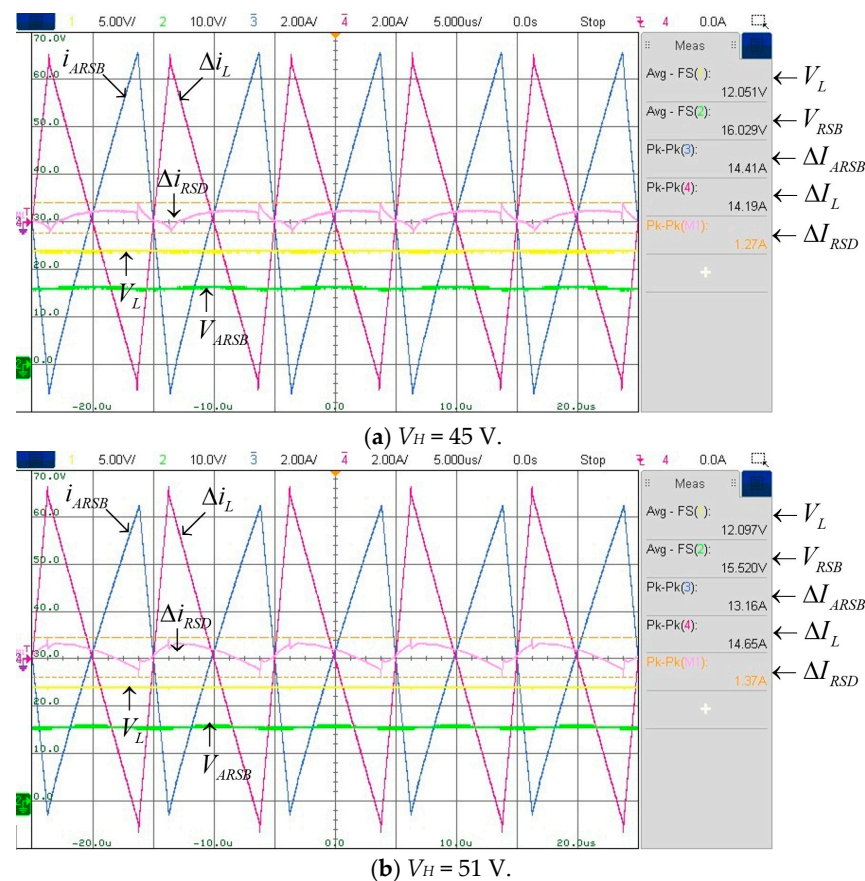


Figure 19. Experimental waveforms of the converter in Figure 7 under deviated duty cycle.

Table 2. Main experimental setup components.

Parameter	Serial Number	Manufacturer
Q_1, Q_2	LMG3100	Texas Instruments (Dallas, TX, USA)
L	B82559A5682A033	TDK (Tokyo, Japan)
Q_3, Q_4	IPD060N03LGATMA1	Infineon (Neubiberg, German)
L_{ARSB}^*	SRP1038A-2R2M	Bourns (Riverside, CA, USA)

It must be emphasized that even though all derivations above were carried out while neglecting parasitic resistances of the devices and mismatches between nominal and actual inductance values, the proposed circuit still operates as expected, complying with the requirements.

4. Conclusions

In this work, a comprehensive analysis of a reduced-rating active ripple suppression branch (ARSB) has been presented, and a reduced-component, capacitorless topology has been proposed for fixed-duty-cycle DC–DC converter operation. The study systematically examined the impact of reducing the ARSB rating, highlighting the associated benefits, including a decrease in the required voltage rating of the auxiliary switching devices and a reduction in the inductance value of the ARSB. These modifications contribute to improved implementation feasibility, reduced component stress, and enhanced overall converter compactness. However, the analysis also demonstrated that such a rating reduction is inherently coupled with a degradation in ripple attenuation performance unless appropriately compensated. It was shown that maintaining the desired ripple-suppression capability requires

a corresponding increase in the ARSB capacitance value, which compensates for the reduction in inductance and restores the effectiveness of the ripple-cancellation mechanism. This relationship establishes a clear design trade-off between component rating reduction and reactive energy storage requirements within the auxiliary branch. Furthermore, the study identified a specific operating condition in which the converter operates at a fixed but arbitrarily selectable duty-cycle value. Under this condition, the ARSB capacitance can be eliminated without compromising the ripple attenuation performance. This finding enables a simplified converter structure characterized by reduced device count, lower control and implementation complexity, and improved power density, while simultaneously preserving or even enhancing ripple suppression capability. The proposed concepts were validated through detailed simulation studies, confirming the analytical predictions across a range of operating conditions. In addition, the feasibility of the reduced-rating capacitorless ARSB topology was experimentally verified, with measurement results showing close agreement with both theoretical and numerical models. This consistency across analytical, simulation, and experimental domains confirms the correctness of the proposed approach and its applicability to practical DC–DC converter systems. Future research will focus on extending the analysis to include the sensitivity of residual output ripple to component tolerances, parasitic resistances, and non-ideal switching behaviour as well as the development of a suitable closed-loop control algorithm. Moreover, guidelines for designing an ARSB-based converter from scratch will also be established.

Author Contributions: Conceptualization, R.M. and A.K.; methodology, V.Y. and R.H.; software, V.Y. and R.H.; validation, V.Y., R.H. and R.M.; formal analysis, V.Y. and R.H.; investigation, V.Y., R.H. and R.M.; resources, M.M.P. and A.K.; data curation, A.K., M.M.P. and V.G.; writing—original draft preparation, V.Y.; writing—review and editing, R.M. and A.K.; visualization, M.M.P., A.K. and V.G.; supervision, M.M.P., A.K. and V.G.; project administration, M.M.P., A.K. and V.G.; funding acquisition, M.M.P. and A.K. All authors have read and agreed to the published version of the manuscript.

Funding: This research was funded in part by the Israeli Ministry of Innovation, Science and Technology.

Institutional Review Board Statement: Not applicable.

Informed Consent Statement: Not applicable.

Data Availability Statement: The original contributions presented in this study are included in the article. Further inquiries can be directed to the corresponding author.

Conflicts of Interest: The authors declare no conflicts of interest. The funders had no role in the design of the study; in the collection, analyses, or interpretation of data; in the writing of the manuscript; or in the decision to publish the results.

Abbreviations

The following abbreviations are used in this manuscript:

RSD Ripple-Sensitive Device
ARSB Active Ripple Suppression Branch

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