

ANALYSIS AND DESIGN OF DC-DC CONVERTER FOR ELECTRIC VEHICLE APPLICATIONS

Yadlapalli Ravindranath Tagore^{1,*} and Kotapati Anuradha²

Received: September 11, 2020; Revised: June 10, 2021; Accepted: September 10, 2021

Abstract

Nowadays electric vehicles (EVs) have a prominent role for visualizing the pollution free environment. In view of this, there is a need to design efficient power conditioning units as per the present trends. Moreover, the fuel cells present a righteous solution for powering the EVs. However, the main drawback associated with each fuel cell is the production of very low output voltage. Hence, high step ratio DC-DC converters are required for generating the desired dc link voltage in order to feed the DC-AC converter. This paper exaggerates the analysis of an existing high step-up ratio DC-DC converter for the EV applications. This converter is designed with the voltage mode and average current mode control strategies. The PSIM simulation results are presented for the above two control strategies with an emphasis on the steady and dynamic load regulations. The converter efficiency curves are presented with respect to the different input and load conditions.

Keywords: Fuel cells, Electric vehicles, Power converters, Control strategies.

Introduction

The step-up converters are playing a key role in the design of power conditioning units for the fuel cell powered electric vehicle applications (Singh *et al.*, 2019). Today, the electric vehicles are saving thousands of billion dollars throughout the world (Zhang *et al.*, 2016). They have the advantages like zero CO₂ emissions, high energy efficiency and low maintenance. The electric vehicles are more preferable in the present transportation systems due to the dying of the fossile fuels. However, their usage depends on the availability of several battery charging stations. Furthermore, the recent approach is the application of the flash charge batteries with approximately sixty times fast charging as compared to the lithium-ion batteries. They have the extraordinary features like non-flammable, long life

and non-toxic. On the other hand, the fuel cells present an effective solution for the EV applications (Himadry Shekhar *et al.*, 2017; Ravindranath Tagore *et al.*, 2019). The fuel cells have the following prominent features.

- Low green house emissions
- Multiple sources
- Modularity
- Good reliability
- Durability
- Flexible siting and Scalability

However, it is important to design an efficient power conditioning unit for the fuel cells (Kim and Peng, 2007). Moreover, the downside associated with the fuel cell is the lower output voltage. This voltage is further affected by the various internal losses. These

¹ Department of EEE, RVR & JC CE, Guntur, A.P, India, 522019. Email: yrtagore@gmail.com

² Department of EEE, VNR VJIET, Hyderabad, T.S, India, 500090.

* Corresponding author

losses are ramified as the ohmic, activation and mass transportation effects. Furthermore, the output voltage regulation is the critical issue along with the good converter dynamics. Hence, a very attractive high step ratio DC-DC converter is to be designed for the production of the desired dc link voltage besides enabling the good voltage regulation. The output of the DC-DC converter is given as input to the DC-AC converter. Many step-up converters are reported in the literature by several researchers. The converter designed by Samavatian and Radan (2015) has a narrow conversion ratio. Wu *et al.* (2015) have implemented the converter with high efficiency. However, its usage is limited to the low power applications. A high power rating converter is designed by Zhang *et al.* (2017). This converter has the down side of larger size of the passive components. The converter reported by Kardan *et al.* (2017) has low efficiency. Thus, some converters are inferior in enacting the high conversion ratios. Also, the most unfavorable one is the device voltage stresses with a impact on the ratings. This further increases the cost of the overall system. In addition, the cost and efficiency are severely affected by the number of devices and the passive components. Hence, this necessitates a trade-off between the achievable conversion ratio and the power component number. This can fulfill the essential requirements of a dc-dc converter. This paper presents the analysis and design of a high step-up converter (HSC) for the fuel cell (FC) powered EV applications (Zhang *et al.*, 2018). This HSC topology has the notable features like fault tolerant operation, flexible for extended stages and high gain. The device stresses are only half of the load voltage. Also, the output and the input grounds are at the constant potential difference. This HSC is implemented with the voltage mode control (VMC) as a initial step. However, this paper highlights the average current mode control (ACMC) with the attractive features like fast dynamics, overcurrent protection and curtailed subharmonic oscillations (Yadlapalli and Kotapati, 2018). Hence, this paper emphasizes the VMC and ACMC based design for the above mentioned DC-DC converter reported in the literature. The PSIM simulation results are presented for the above two control strategies with an emphasis on the steady and dynamic load regulations. The converter efficiency curves are presented with respect to the different input and load conditions.

DC-DC Converter

A high gain step-up converter with the diode capacitor modules is depicted in the Figure 1 (Zhang *et al.*, 2018). HSC voltage relations are shown by

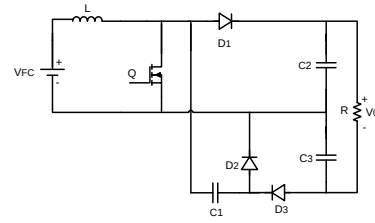
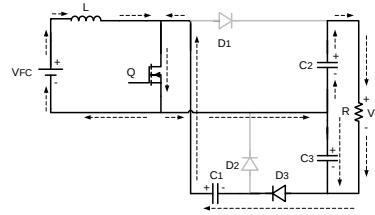
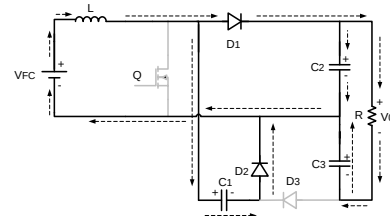


Figure 1. Schematic of the HSC.



(a) ON State of the HSC.



(b) OFF state of the HSC.

Figure 2. ON and OFF states of the HSC.

$$\frac{V_o}{V_{in}} = \frac{2}{1-D} \quad (1)$$

Where V_o and V_{in} are the output and input voltages, D is the duty cycle.

Modeling and Control Strategies of HSC

The dc-dc converters are modelled in view of analyzing the line and load dynamics. Many researchers have reported the different modeling techniques in the literature (Karaket and Bunlaksananusorn, 2011; Yadlapalli and Kotapati, 2014a; Yadlapalli and Kotapati, 2014b). The literature Highlights the very prominent modeling technique namely state-space averaging methodology. Figure 2 highlights the different operating states of the converter. When the switch is turned-on as shown in Figure 2(a), the inductor L takes energy from the source, and the diodes D_1 and D_2 are turned-off. Moreover, the capacitor C_1 supplies energy to the capacitor C_3 through the diode D_3 and the MOSFET switch. The capacitor C_2 delivers energy to the load. Equation (2) represents the state-space average model of the dc-dc converter during turn-on of the switch (Zhang *et al.*, 2018).

$$\begin{bmatrix} \frac{dv_{C1}(t)}{dt} \\ \frac{dv_{C2}(t)}{dt} \\ \frac{dv_{C3}(t)}{dt} \\ \frac{di_L(t)}{dt} \end{bmatrix} = \begin{bmatrix} 0 & -\frac{1}{rC} & 0 & \frac{1}{rC} \\ 0 & 0 & -\frac{1}{RC} & -\frac{1}{RC} \\ 0 & \frac{1}{rC} & -\frac{1}{RC} & -\frac{(R+r)}{rRC} \\ 0 & 0 & 0 & 0 \end{bmatrix} \begin{bmatrix} v_{C1}(t) \\ v_{C2}(t) \\ v_{C3}(t) \\ i_L(t) \end{bmatrix} + \begin{bmatrix} 0 \\ 0 \\ 0 \\ \frac{1}{L} \end{bmatrix} V_m(t)$$

$$V_o(t) = [0 \quad 1 \quad 1 \quad 0][v_{C1}(t) \quad v_{C2}(t) \quad v_{C3}(t) \quad i_L(t)]^T \quad (2)$$

Where, V_{C1} , V_{C2} , V_{C3} are the voltages across the capacitors C_1 , C_2 , C_3 , and I_L is the inductor current.

When the switch is turned-off as shown in Figure 2(b), the source and inductor L transfers energy to the capacitors C_1 and C_2 , whereas the diodes D_1 is on and D_3 off. The load gets energy through the combination of the source, inductor L and the capacitor C_3 . Equation (3) represents the state-space average model of the converter during turned-off of the switch.

$$\begin{bmatrix} \frac{dv_{C1}(t)}{dt} \\ \frac{dv_{C2}(t)}{dt} \\ \frac{dv_{C3}(t)}{dt} \\ \frac{di_L(t)}{dt} \end{bmatrix} = \begin{bmatrix} 0 & -\frac{1}{rC} & \frac{1}{rC} & 0 \\ \frac{1}{C} & \frac{1}{rC} & -\frac{(R+r)}{rRC} & -\frac{1}{RC} \\ 0 & 0 & -\frac{1}{RC} & -\frac{1}{RC} \\ 0 & 0 & -\frac{1}{L} & 0 \end{bmatrix} \begin{bmatrix} v_{C1}(t) \\ v_{C2}(t) \\ v_{C3}(t) \\ i_L(t) \end{bmatrix} + \begin{bmatrix} 0 \\ 0 \\ 0 \\ \frac{1}{L} \end{bmatrix} V_m(t)$$

$$V_o(t) = [0 \quad 1 \quad 1 \quad 0][v_{C1}(t) \quad v_{C2}(t) \quad v_{C3}(t) \quad i_L(t)]^T \quad (3)$$

Based on the Equation (2) and Equation (3), the state-space model of the converter is given as below.

$$\begin{bmatrix} \frac{dv_{C1}(t)}{dt} \\ \frac{dv_{C2}(t)}{dt} \\ \frac{dv_{C3}(t)}{dt} \\ \frac{di_L(t)}{dt} \end{bmatrix} = \begin{bmatrix} 0 & -\frac{1}{rC} & \frac{1-d(t)}{rC} & \frac{d(t)}{rC} \\ \frac{1-d(t)}{C} & \frac{1-d(t)}{rC} & -\left(\frac{1-d(t)}{rC} + \frac{1}{RC}\right) & -\frac{1}{RC} \\ 0 & \frac{d(t)}{rC} & \frac{1}{RC} & -\left(\frac{d(t)}{rC} + \frac{1}{RC}\right) \\ 0 & 0 & -\frac{1-d(t)}{L} & 0 \end{bmatrix} \begin{bmatrix} v_{C1}(t) \\ v_{C2}(t) \\ v_{C3}(t) \\ i_L(t) \end{bmatrix} + \begin{bmatrix} 0 \\ 0 \\ 0 \\ \frac{1}{L} \end{bmatrix} V_m(t)$$

$$V_o(t) = [0 \quad 1 \quad 1 \quad 0][v_{C1}(t) \quad v_{C2}(t) \quad v_{C3}(t) \quad i_L(t)]^T \quad (4)$$

Equation (4) is very useful for realizing the small-signal analysis of the converter by applying small perturbations to the control variables V_{C1} , V_{C2} , V_{C3} , I_L , V_{in} and D around the equilibrium point (Zhang et al., 2018). It includes the steady state components V_{C1} , V_{C2} , V_{C3} , I_L , V_{in} and D , and also the corresponding small ac variations $\tilde{v}_{C1}(t)$, $\tilde{v}_{C2}(t)$, $\tilde{v}_{C3}(t)$, $\tilde{i}_L(t)$, $\tilde{V}_m(t)$, $\tilde{d}(t)$.

The converter dynamics are highlighted by the following small-signal state-space equations (Karaket and Bunlaksananusorn, 2011).

$$\frac{d\tilde{x}(t)}{dt} = A_s \tilde{x}(t) + B_s \tilde{v}_{in}(t) + B_i \tilde{d}(t)$$

$$\tilde{y}(t) = C_s \tilde{x}(t) + E_s \tilde{v}_{in}(t) + E_i \tilde{d}(t) \quad (5)$$

Where, $\tilde{x}(t)$ and $\frac{d\tilde{x}(t)}{dt}$ are the small variations in the state vector and its derivative, $\tilde{y}(t)$ is the small variation in the output vector. Using Equation (4), the equivalent representation of the Equation (5) is given by

$$\begin{bmatrix} \frac{d\tilde{v}_{C1}(t)}{dt} \\ \frac{d\tilde{v}_{C2}(t)}{dt} \\ \frac{d\tilde{v}_{C3}(t)}{dt} \\ \frac{d\tilde{i}_L(t)}{dt} \end{bmatrix} = \begin{bmatrix} 0 & -\frac{1}{rC} & \frac{1-D}{rC} & \frac{D}{rC} \\ \frac{1-D}{C} & \frac{1-D}{rC} & -\left(\frac{1-D}{rC} + \frac{1}{RC}\right) & -\frac{1}{RC} \\ 0 & \frac{D}{rC} & -\frac{1}{RC} & -\left(\frac{D}{rC} + \frac{1}{RC}\right) \\ 0 & 0 & -\frac{1-D}{L} & 0 \end{bmatrix} \begin{bmatrix} \tilde{v}_{C1}(t) \\ \tilde{v}_{C2}(t) \\ \tilde{v}_{C3}(t) \\ \tilde{i}_L(t) \end{bmatrix} + \begin{bmatrix} 0 \\ 0 \\ 0 \\ \frac{1}{L} \end{bmatrix} \tilde{V}_m(t)$$

$$\tilde{V}_o(t) = [0 \quad 1 \quad 1 \quad 0][\tilde{v}_{C1}(t) \quad \tilde{v}_{C2}(t) \quad \tilde{v}_{C3}(t) \quad \tilde{i}_L(t)]^T \quad (6)$$

Here the matrices B_i , E_s and E_i are zero. The necessary small signal transfer functions are obtained by

$$\tilde{x}(s) = \begin{bmatrix} (SI - A)^{-1}B_s & (SI - A)^{-1}B_i \end{bmatrix} \begin{bmatrix} \tilde{u}(s) \\ \tilde{d}(s) \end{bmatrix}$$

$$\tilde{y}(s) = \begin{bmatrix} C_s(SI - A)^{-1}B_s + E_s & C_s(SI - A)^{-1}B_i + E_i \end{bmatrix} \begin{bmatrix} \tilde{u}(s) \\ \tilde{d}(s) \end{bmatrix} \quad (7)$$

Equation (8) and Equation (9) represent the required control transfer functions which are obtained based on the Equation (7) and Equation (8) gives the duty signal-to- i_L transfer function.

$$\frac{\tilde{i}_L(s)}{\tilde{d}(s)} = \frac{a_1 s^3 + b_1 s^2 + c_1 s + d_1}{a_2 s^4 + b_2 s^3 + c_2 s^2 + d_2 s + e_2} \quad (8)$$

Equation (9) gives the duty signal-to- v_o transfer function.

$$\frac{\tilde{v}_o(s)}{\tilde{d}(s)} = \left(\frac{as^3 + bs^2 + cs + d}{s^4 + a_2 s^3 + b_2 s^2 + c_2 s + d_2} \right) \quad (9)$$

The numerator and denominator coefficients are given in the appendix. The above control transfer functions are extremely useful in the design of the VMC and ACMC strategies. The block diagram of the voltage-mode control (VMC) is shown in Figure 3. It is easier to implement the VMC with one control loop. The feedback voltage is compared with the reference voltage and the error is fed to the controller. The PWM controller adjusts the duty cycle based on the control signal generated from the controller. However, the VMC has poor converter dynamics without the over-current protection. Therefore, the current-mode control strategies are playing a key role for enhancing the converter dynamics besides the excellent over-current protection. The ACMC is constituted by the inner current and the outer voltage loops with the block diagram shown in Figure 4. There should be a proper bandwidth relation between the two control loops. These two loops are designed such that the overall converter system should always be stable over the wide range of input or load variations. Different case studies are considered for analysing the robustness against the input voltage or load current regulations.

Design Aspects of HSC

The designed values of L and C are given by the Equation (10) and Equation (11).

$$L = \frac{V_{in} D}{\Delta I_L f_{sw}} \quad (10)$$

$$C = \frac{I_o D}{2 \Delta V_C f_{sw}} \quad (11)$$

Here, all the capacitor values are assumed to be equal in the converter analysis i.e. $C_1=C_2=C_3=C$. Where, f_{sw} and I_o are the converter operating frequency and the load current. The capacitor voltage ripple (ΔV_C) and the inductor current ripple (ΔI_L) are chosen as 1% and 15%. However, the values of L and C should be 25% larger than the critical values L_C and C_C as represented in the Equation (12).

$$L > \frac{RD(1-D)^2}{8f_{sw}} \quad (12)$$

$$C > \frac{D}{4f_{sw} R}$$

The selection of the inductor L is very crucial in case of the dc-dc converters. The high inductor current ripple (ΔI_L) can degrade the converter efficiency but with improved transient performance.

Therefore, a trade-off is required between the converter dynamics and the achievable efficiency. All the capacitors should be selected with the low ESR value in order to minimize the jumps in the output voltage.

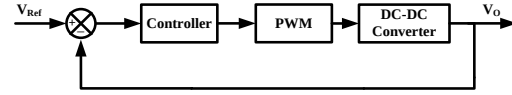


Figure 3. Block diagram of the VMC strategy based HSC.

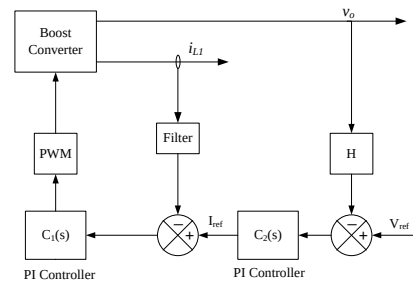


Figure 4. Block diagram of the ACMC strategy based HSC.

Simulation Results and Discussions

The performance of the HSC is analyzed by considering the various case studies.

The specifications of the HSC are presented in Table 1. The source and the load voltages are the same as specified in the literature (Zhang *et al.*, 2018). The values of L and C are obtained based on the Equation (10) and Equation (11).

Table 1. Specifications of the HSC.

Nominal Parameter	Value
V_{in}	45 V
L	225 μ H
C_3, C_4, C_5	450 μ F
ESR	30 m Ω
f_{sw}	20 kHz
V_o	400 V
Load $R_{min} - R_{max}$	100 - 400 Ω

Equation (13) shows the inductor current control transfer function (TF) of the HSC.

$$\frac{\tilde{i}_L(s)}{\tilde{d}(s)} = \left(\frac{854700s^3 + 121343539020s^2 + 4848691083538500s + 43242345934695349200}{s^4 + 141887s^3 + 5662980685s^2 + 50343098795928s + 2143796984794080} \right) \quad (13)$$

Equation (14) shows the output voltage control transfer function of the HSC.

$$\frac{\bar{v}_o(s)}{d(s)} = \left(\frac{68090s^3 - 7996221860s^2 - 200395051026720s + 1715091553911859200}{s^4 + 141887s^3 + 5662980685s^2 + 50343098795928s + 2143796984794080} \right) \quad (14)$$

The design of ACMC is realized with the usage of SISO tool in the MATLAB software. The control bandwidth (BW) of the system is always less than the switching frequency. Moreover, the current loop BW is always greater than the voltage loop BW. A trade-off is required between the current loop as well as the voltage loop dynamics in terms of the BW. The crossover frequency (f_c) of the current loop filter is predominantly equal to the f_{sw} . This can curtail the large amplitude oscillations in the output voltage. It is easier to tune the PI controller parameters until the system becomes stable. The initial tuning process is abided by the selection of the control architecture and the loop configuration before actualizing the graphical tuning process.

As explained earlier, the f_c of the current loop filter is equal to 20 kHz. The current loop low pass filter TF is given by

$$F(s) = \frac{12600}{(s + 12600)} \quad (15)$$

The current loop controller TF is given by

$$C_1(s) = \frac{1000(1 + 0.001s)}{s} \quad (16)$$

The voltage loop controller TF is given by

$$C_2(s) = \frac{10000(1 + 0.01s)}{s} \quad (17)$$

The PI controllers shown in Equation (16) and Equation (17) are designed to have high gain at low frequencies in view of minimizing the steady state error. In order to alleviate the control loop from the switching noise, the f_c of the current loop is always one to two orders of magnitude lesser than the operating frequency. The bode plot of both the loops is shown in Figure 5. The obtained gain and phase margins for the current and voltage loops are around 6.5 dB, 49° and 20 dB, 46.5°. The bode-plot highlights the fact that the f_c of the current loop is greater than the voltage loop.

As explained earlier, the PI controller design is very crucial for achieving the good dynamic response besides the good output voltage regulation of the HSC. The important issues are the load and supply voltage variations. In this simulation study, the HSC performance is verified over a wide range of input

and load conditions. It focuses on the different case studies that includes the worst operating conditions. The performance measurements during the steady state and dynamic operating conditions are the steady state load voltage deviation (SSLVD), transient load voltage deviation (TLVD) and the transient load voltage settling time (TLVST). During the input voltage or load changes, the output voltage should reach the nominal value very quickly.

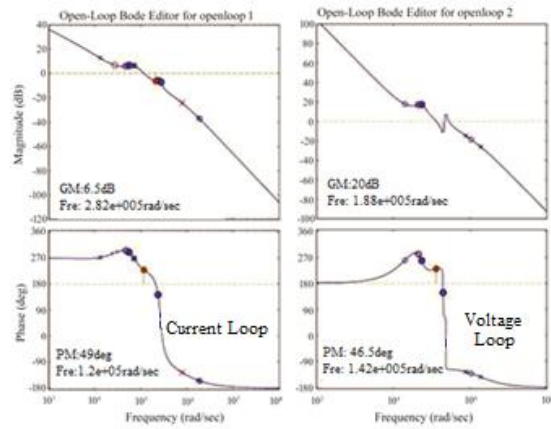


Figure 5. Bode plot for the current and voltage loops of the ACMC based HGC.

Performance comparison of the VMC and ACMC based HSC at an input voltage of 45 V

The VMC and ACMC based HSC waveforms are depicted in Figure 6(a) and Figure 7(a). At constant load current of 4A, the SSLVD is $\pm 3.5\%$ in case of the VMC as compared to $\pm 0.05\%$ with the ACMC, which are observed in the magnified waveforms of the Figure 6(b) and Figure 7(b).

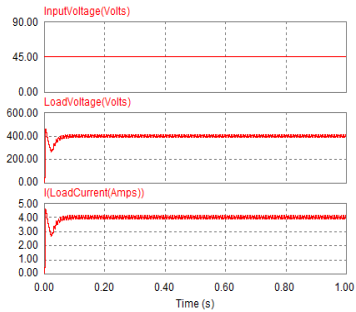
Performance comparison of the VMC and ACMC based HSC for the load transition at an input voltage of 45 V

Figure 8(a) and Figure 8(b) present the VMC based HSC waveforms. The TLVDs and TLVSTs during the step-down and step-up load transitions are +1.6%, -2.7% and 2msec, 3.5msec respectively. Whereas the ACMC based HSC is exhibiting +0.62%, -0.7% and 1msec, 1msec respectively as shown in the Figure 9(a) and Figure 9(b).

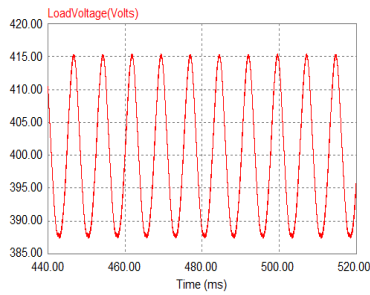
Performance comparison of the VMC and ACMC based HSC for different step input voltages

Figure 10 and Figure 11 present the VMC and ACMC based HSC waveforms. Two different case studies are considered with the application of 45 V

to 30 V and 45 V to 100 V step input voltages. The ACMC based HSC exhibits the excellent performance as compared to the VMC based HSC without any oscillations or peaks.

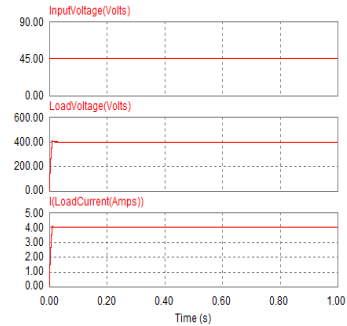


(a) Unmagnified

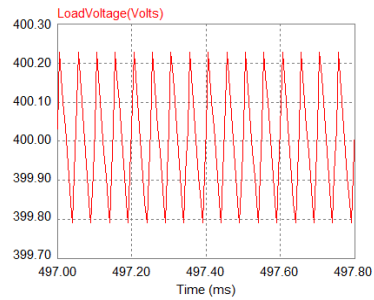


(b) Magnified

Figure 6. VMC based HSC waveforms at the rated operating conditions.

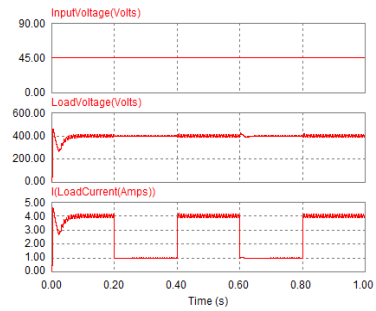


(a) Unmagnified

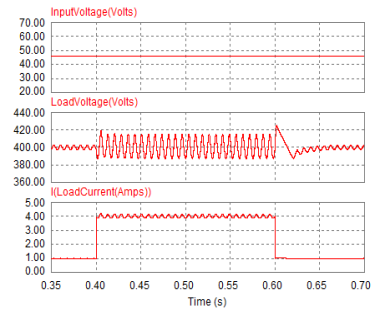


(b) Magnified

Figure 7. ACMC based HSC waveforms at the rated operating conditions.

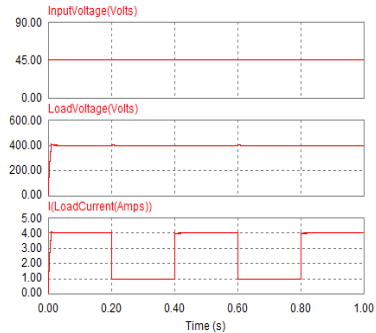


(a) Unmagnified

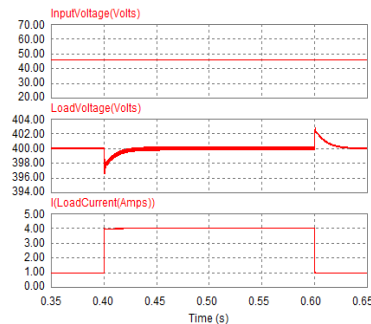


(b) Magnified

Figure 8. VMC based HSC waveforms during the periodic load transition.

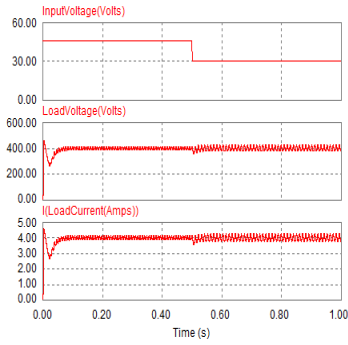


(a) Unmagnified

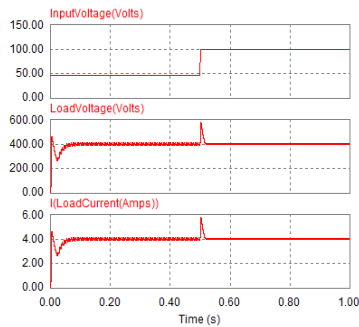


(b) Magnified

Figure 9. ACMC based HSC waveforms at the rated operating conditions.

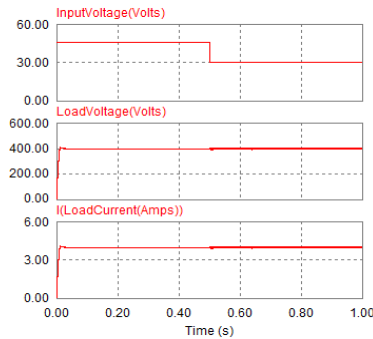


(a)

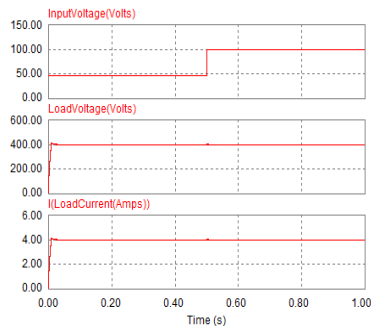


(b)

Figure 10. VMC based HSC waveforms with the step input voltages of (a) 45 V to 30 V and (b) 45 V to 100 V.



(a)

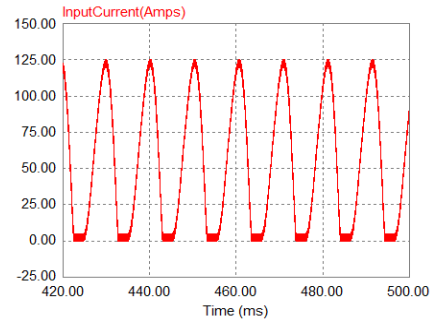


(b)

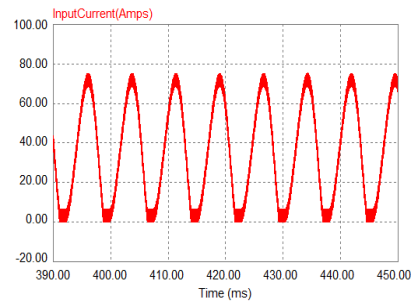
Figure 11. ACMC based HSC waveforms with the step input voltages of (a) 45 V to 30 V and (b) 45 V to 100 V.

Input current ripple of the VMC and ACMC based HSC at an input voltage of 45 V and 30 V

As shown in Figure 12 and Figure 13, the input or source current ripple is observed to be $\pm 110.9\%$ in case of the VMC based HSC as compared to $\pm 9.392\%$ with the ACMC based HSC. Therefore, the ACMC can enhance the reliability and lifetime of the fuel cell.



(a) 30 V



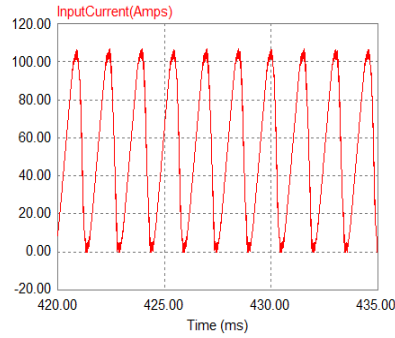
(b) 45 V

Figure 12. The input current ripple waveforms of the VMC based HSC at an input voltages of 30 V and 45 V.

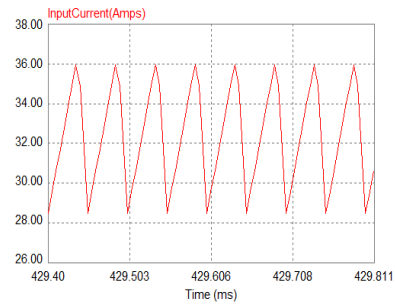
The HSC performance is shown in Table 2 with the VMC as well as the ACMC based HSC at different input voltage conditions of 30 V, 45 V and 100 V. Here, the SSLVD is estimated by considering the each input voltage individually with the load current ranging from 1A (minimum load) to 4 A (maximum load). At the rated load condition of 4 A, the SSLVD is found to be $\pm 0.05\%$ with the ACMC based HSC as compared to $\pm 3.5\%$ in case of the VMC based HSC. Figure 14 shows the comparison of the load voltage ripple with the VMC and ACMC based HSC at an input voltage of 45 V. Furthermore, the ACMC exhibits the superior performance at the other load conditions of 2.5 A and 1 A. Table 3 shows the dynamic performance interpretation of the HSC at various input voltages of 30 V, 45 V and 100 V. The load regulation is estimated with the step-load variation from 1 A to 4 A and vice-versa. At rated input voltage of 45 V,

the TLVD and TLVST during the step-up load changes are -0.7% and 1 msec with the ACMC as compared to -2.7% and 3.5 msec in case of the VMC. These indices during step-down load changes are +0.62% and 1 msec with the ACMC as compared to +1.6% and 2 msec in case of the VMC. The dynamic performance comparison of the VMC and ACMC based HSC is shown in Figure 15 at an input voltage of 45 V. It is also evident that the ACMC exhibits the superior dynamic performance at the other input voltages of 30 V and 100 V. This work highlights the least TLVST of 1 msec as compared to the existing literature Zhang *et al.* (2018) with a TLVST of 8 msec. Table 4 shows the source current ripple of the HSC with the VMC and

ACMC strategies at an input voltages of 30 V, 45 V and 100 V, the output current is varied from 1 A to 4 A. At rated load current of 4 A, the input current ripple is found to be $\pm 9.392\%$ with the ACMC based HSC as compared to $\pm 110.9\%$ in case of the VMC based HSC. The ACMC exhibits the superior performance at the other load conditions of 1 A and 2.5 A. Therefore, the ACMC based HSC is phenomenal during the steady and transient operating conditions as compared to the VMC based HSC. The efficiency curves are shown in Figure 16 at different input conditions of 30 V, 45 V, 60 V, 70 V and 100 V. The maximum converter efficiency is found to be 93% at an input voltage of 30 V.



(a) 30 V



(b) 45 V

Figure 13. The input current ripple waveforms of the ACMC based HSC at an input voltages of 30 V and 45 V.

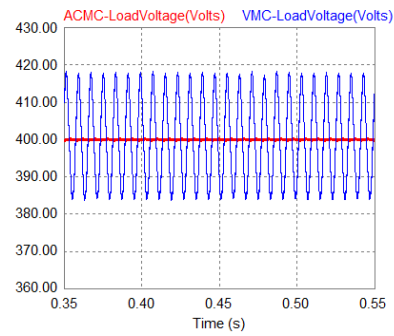


Figure 14. The comparison of the load voltage ripple with the VMC and ACMC based HSC at an input voltage of 45 V.

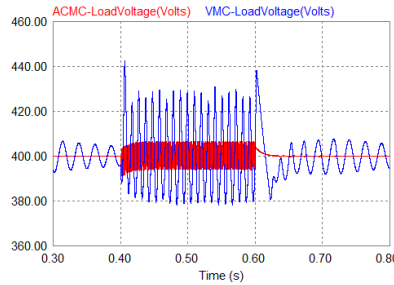


Figure 15. The comparison of the dynamic performance with the VMC and ACMC based HSC at an input voltage of 45 V.

Table 2. Steady state performance of the HSC at different load currents of 1 A, 2.5 A and 4 A.

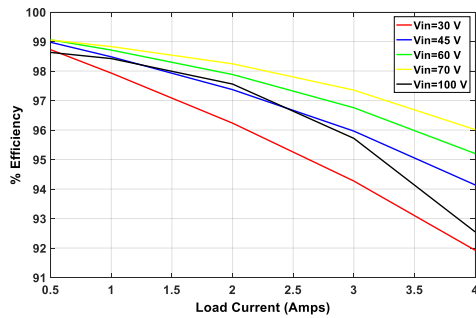
Load Current	Input voltage	SSLVD (Δv_o)	VMC based HSC	ACMC based HSC
			$\frac{\Delta v_o}{\text{Desired output voltage } (v_o)} \times 100\%$	$\frac{\Delta v_o}{v_o} \times 100\%$
4 A ($R_{\min} = 100 \Omega$)	30 V	20 V	± 5.0	5 V
	45 V	14 V	± 3.5	0.2 V
	100 V	6 V	± 1.5	0.2 V
2.5 A ($R_{\text{avg}} = 160 \Omega$)	30 V	14 V	± 3.5	0.147 V
	45 V	9 V	± 2.25	0.141 V
	100 V	2.7 V	± 0.675	0.121 V
1 A ($R_{\max} = 400 \Omega$)	30 V	5 V	± 1.25	0.059 V
	45 V	2.5 V	± 0.625	0.055 V
	100 V	0.039 V	± 0.00975	0.035 V

Table 3. Dynamic regulation of the HSC.

Input voltage	VMC based HSC				ACMC based HSC			
	Step-Up Load Transition		Step-Down Load Transition		Step-Up Load Transition		Step-Down Load Transition	
	TLVD	TLVST (msec)	TLVD	TLVST (msec)	TLVD	TLVST (msec)	TLVD	TLVST (msec)
30 V	-10 V (-2.5%)	15	12 V (+3.0%)	60	-8 V (-2.0%)	1	8 V (+2.0%)	2
45 V	-11 V (-2.7%)	3.5	6.5 V (+1.6%)	2.0	-3 V (-0.7%)	1	2.5 V (+0.62%)	1
100 V	-21 V (-5.2%)	40	11 V (+2.7%)	58	-2 V (-0.6%)	2	2 V (+0.6%)	2

Table 4. Input current ripple of the HSC at different load currents of 1 A, 2.5 A and 4 A.

Load Current	Input voltage	Input current ripple for VMC based HSC	Input current ripple for ACMC based HSC
4 A ($R_{min.} = 100 \Omega$)	30 V	$\pm 134.4 \%$	$\pm 93.51 \%$
	45 V	$\pm 110.9 \%$	$\pm 9.392 \%$
	100 V	$\pm 80.63 \%$	$\pm 33.94 \%$
2.5 A ($R_{avg.} = 160 \Omega$)	30 V	$\pm 125.02 \%$	$\pm 9.90 \%$
	45 V	$\pm 111.50 \%$	$\pm 14.98 \%$
	100 V	$\pm 108.0 \%$	$\pm 54.7 \%$
1 A ($R_{max.} = 400 \Omega$)	30 V	$\pm 124.9 \%$	$\pm 24.76 \%$
	45 V	$\pm 112.6 \%$	$\pm 37.23 \%$
	100 V	$\pm 113.75 \%$	$\pm 113.5 \%$

**Figure 16. Efficiency Vs Load current curves at different input voltages.**

Conclusions

This paper emphasizes the prominent features and performance analysis of a high conversion ratio DC-DC converter with the VMC and ACMC strategies for the EV applications. The PSIM simulation results are interpreted for the steady as well as dynamic regulations. In the different case studies, the figures of merits under consideration are the SSLVD, TLVD and TLVST. The ACMC exhibits the phenomenal performance in all the above aspects as compared to the VMC strategy. The various losses of the devices are considered while realizing the efficiency analysis. This converter can be the righteous option for the fuel cell powered EVs

and also provides insightness for the design engineers. This work can be extended further for the high power applications.

Appendix

The coefficients of the numerator and denominator of the Equation (8) and Equation (9) are given below.

$$\begin{aligned}
 a_2 &= LC^3R^2r^3 \\
 b_2 &= 2LC^2R^2r^2 + 2LC^2Rr^3 \\
 c_2 &= C^2D^2R^2r^3 - 2C^2DR^2r^3 + C^2R^2r^3 - 2LCD^2R^2r - LCD^2Rr^2 + 3LCDR^2r + 3LCRr^2 \\
 d_2 &= (CD^3R^2r^2 - CD^2R^2r^2 + CD^2Rr^3 - CDR^2r^2 - 2CDRr^3 + CR^2r^2 + CRr^3 - LD^3R^3 + LD^3Rr + LD^2R^2 - 3LD^2Rr + 3LDRr + LDR^2 - 2LD^2r^2) \\
 e_2 &= -D^4Rr^2 + D^3R^2r + 2D^3Rr^2 - 2D^2R^2r + DR^2r - 2DRr^2 + Rr^2 / (C^3LR^2r^3)
 \end{aligned}$$

$$\begin{aligned}
 a_1 &= (C^3R^2r^3) / (LD) \\
 b_1 &= (2C^2R^2r^2 + 2C^2Rr^3) / (C^3LR^2r^3) \\
 c_1 &= [(2CD^2R^2r - CD^2Rr^2 + 3CDR^2r + 3CRr^2) / (C^3LR^2r^3) + (Cr^2 / C^2LRr^2) + ((CDR^2r - CR^2r) / (C^2LR^2r^2)) + ((RC_2r^2 + RCDr + Cr^2) / (C^2LRr^2))] \\
 d_1 &= [((D^3R^2 + D^3Rr + D^2R^2 - 3D^2Rr - 2D^2r^2 + 3DRr + Dr^2) / (C^3LR^2r^3)) - ((RD^2 - RD + r) / (C^2LRr^2)) + ((D - 1) / Cr) + ((D^2R^2 + DR^2 + DRr + Dr^2 - Rr) / (C^2LR^2r^2)) - ((D - 1)C + (Cr - (Cr - D^2r + RD + r)) / (C^2LRr^2))]
 \end{aligned}$$

$$\begin{aligned}
 a &= LRC^2r^2 - RCDr \\
 b &= 2RC^2r^2 - ((LRCDr + LCr^2 + 2LRCr) / (C^2LRr^2)) \\
 c &= ((D^2S - DS + S) / (C^2Rr)) - ((RCDr + Cr^2S + RCr - D^2r + RD + r) / (C^2Rr^2)) - ((D^2LR + 2DLR + DLR) / (C^2LR^2r)) + ((CDr - 1) / (C^3LRr)) + RCD^2r^2 + RCr^2S + LRD^2 - 2RDr \\
 d &= (-2D^2Rr + D^3Rr) / (C^2LR^2r) + ((2D - 2D^2 + D^3 - 1) / (C^3LRr))
 \end{aligned}$$

References

- Himadry Shekhar, D., Chee Wei, T., and Yatim, A.H.M. (2017). Fuel cell hybrid electric vehicles: A review on power conditioning units and topologies. *Renewable and Sustainable Energy Reviews*, 76(1):268-291.
- Karaket, K. and Bunlaksananusorn, C. (2011). Modelling of a quadratic buck converter. In: *Electrical Engineering*,

- Electronics, Telecommunications and Information Technology (ECTI) Conf., Khon Kaen, p. 764-767.
- Kardan, F., Alizadeh, R., and Banaei, M.R. (2017). A new Three Input DC/DC Converter for Hybrid PV/FC/Battery Applications. *IEEE Trans. Power Electron.*, 32(9):1,771-1,778.
- Kim, M.J. and Peng, H. (2007). Power management and design optimization of fuel cell/battery hybrid vehicles. *J. Power Sources.*, 165:819-832.
- Ravindranath Tagore, Y., Anuradha, Y., Vijay, Babu., A.R., and Manoj Kumar, P. (2019). Modelling, simulation and control of a fuel cell powered laptop computer voltage regulator module. *Int. J. Hydrogen Energy.*, 44(21):11012-19.
- Samavatian, V. and Radan, A. (2015). A high efficiency input/output magnetically coupled interleaved buck-boost converter with low internal oscillation for fuel-cell applications: ccm steady state analysis. *IEEE Trans. Ind. Electron.*, 62(9):5,560-5,568.
- Singh, K.V., Bansal, H.O., and Singh, D. (2019). A comprehensive review on hybrid electric vehicles: architectures and components. *J. Mod. Transport.*, 27:77-107.
- Wu, G., Ruan, X., and Zhihong, Ye. (2015). Non-Isolated high step-up dc-dc converters adopting switched-capacitor cell. *IEEE Trans. Ind. Electron.*, 62(1):383-393.
- Yadlapalli, R.T. and Kotapati, A. (2014a). A fast-response sliding-mode controller for quadratic buck converter. *Int. J. of Power Electronics.*, 6(2):103-130.
- Yadlapalli, R.T. and Kotapati, A. (2014b). An efficient sliding-mode current controller with reduced flickering for quadratic buck converter used as LED lamp driver. *Int. J. of Power Electronics.*, 6(4):345-375.
- Yadlapalli, R.T. and Kotapati, A. (2018). Modelling, design and implementation of quadratic buck converter for low power applications. *Int. J. of Power Electronics.*, 11(3):322-338.
- Zhang, F., Zhang, X., Zhang, M., and Edmonds, A.S.E. (2016). Literature review of electric vehicle technology and its applications. In: 2016 5th International Conference on Computer Science and Network Technology (ICCSNT), Changchun, p. 832-837.
- Zhang, Y., Shi, J., Zhou, L., Li, J., Sumner, M., Wang, P., and Xia, C. (2017). Wide input-voltage range boost three-level dc-dc converter with quasi-z source for fuel cell vehicles. *IEEE Trans. Power Electron.*, 32(9):6,728-6,738.
- Zhang, Y., Zhou, L, M., Sumner, M., and Wang, P. (2018). Single-Switch, wide voltage-gain range, boost dc-dc converter for fuel cell vehicles. *IEEE Trans. Vehicular Technol.*, 67(1):134-145.