Performance enhancement of powertrain DC–DC converter using variable inductor

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Abstract
Performance enhancement of a powertrain DC–DC converter is provided using a current-controlled magnetic device, the variable inductor (VI). The VI is composed of two groups of windings, main and auxiliary. The auxiliary winding is used to regulate the permeability of the magnetic core over a wide range of load variations. To regulate the saturation level of the magnetic core and avoid postsaturation operation of the power inductor, an appropriate level of auxiliary winding current is needed. To this end, a simple and dynamic strategy is proposed to estimate the auxiliary winding currents of the VI. The proposed approach is validated using simulation studies of a large-scale VI-based powertrain DC–DC converter. The results reveal the significance of the proposed approach in continuous regulation of the magnetic property of the power inductor core, enhancing powertrain DC–DC converter performance by up to 13.70% and reducing stress on switching devices by 17.87%. Furthermore, energy source current variation is reduced by 6.73%, leading to reduced stress on energy storage systems.

1 INTRODUCTION

In electrified vehicles, power electronic converters are crucial components that enable efficient conversion and management of electrical energy [1, 2]. Their size and characteristics affect driving performance and reliability. Consequently, these circuits are required to have small size, high power density, and high efficiency [3]. This is particularly important in the case of powertrain DC–DC converters, as they are required to meet stringent space limitations. Furthermore, the high power requirement, low voltage and wide operating range cause high currents that affect the electric and thermal stresses in both active and passive components [4].

The size, cost and performance of powertrain DC–DC converters is dominated by the magnetic components [5, 6]. The presence of bulky power inductors influences the overall size of such types of converters and limits space for energy storage systems (ESSs). Moreover, if they are forced to operate at higher currents, power inductors may saturate, leading to high current ripple, resulting in undesirable electromagnetic interference, mechanical vibration, and losses as well as affecting the lifetime of the ESS [7].

To address the aforementioned challenges, several magnetic control-based small-scale proof-of-concept converters are proposed in [8–13]. They can be categorised as variable inductor (VI), magnetorheological fluid, and permanent magnet-based converters. All approaches are useful for controlling the current ripple, regulating the inductance, and regulating the saturation level of magnetic components. The last two approaches have limited control flexibility, but the VI-based approach provides wider flexibility leading to dynamic regulation of the inductance, which makes it more suitable for applications with a wide range of load variations.
To this end, the current-controlled VI is introduced to reduce the size of the bulky magnetic component in a powertrain DC–DC converter, thereby improving performance and controlling current ripple while reducing size. The VI is composed of magnetic core materials and multiple windings with an interaction of main and auxiliary winding currents. These introduce design, materials selection, characterisation, control, and strategy challenges for the integration of VIs into power electronic converters for practical applications.

The detailed modelling and characterisation of a small-scale VI for its potential importance in enhancing current-handling capability is explored in [8, 14]. The detailed characterisation reveals the need for a suitable strategy that determines the suitable levels of auxiliary winding current to exploit the prominent merits of the current-controlled device. Consideration of its design in a real driving cycle is addressed in [15], and the selection of suitable magnetic core materials based on the weighted property method is dealt with in [16]. Furthermore, a simple strategy based on a heuristic approach that relates the currents between the two windings of the magnetic device is presented in [8].

Nonetheless, for smooth functionality of the VI and its large-scale integration into powertrain DC–DC converters, a systematic strategy that estimates the auxiliary winding current as a function of the loading situation is essential. Furthermore, the detailed analysis of the performance of such types of converters and the impacts of introducing VIs into the ESS is of paramount significance. These aspects are not dealt with in the literature.

The focus of this paper is to provide a simple dynamic strategy to enable dynamic control of the VI as a function of the loading situation of the converter and integrate it into the overall system. Accordingly, the performance improvements due to the integration of the VI in a large-scale powertrain DC–DC converter are evaluated, and its impact on the ESS lifetimes and switching devices stresses is assessed. The behaviour of the power inductor at different current levels is extracted from intensive finite element analysis (FEA) simulations (Infolytica MagNet), and the characteristics (linkage flux to main winding current and inductance to the main winding) are used to model the magnetic device. The model obtained from FEA simulations is used to explore the feasibility of a large-scale magnetic control-based converter to validate the proposed strategy and quantify the performance improvements.

This paper is organised as follows. Section 2 provides the description of the study system and the design of the DC–DC converter considering a three-wheeled recreational vehicle. The modelling of the VI, the modelling and control of the DC–DC converter and the proposed strategy are provided in Section 3. Section 4 provides detailed simulations of the converter and evaluation of different parameters to quantify the improvements and validate the proposed strategy. Finally, concluding remarks and future directions are presented in Section 5.

2 | SYSTEM DESCRIPTION AND DESIGN

The study system comprises an ESS as a source, a half-bridge DC–DC converter module, an inverter, an electrical motor, and a control layer as shown in Figure 1. Several automotive manufacturers, such as Toyota with its Prius, commonly use this configuration. The traction DC–DC converter enables DC-link regulation, enhances the main ESS lifetime, extends the motor speed range and improves the efficiency of the electric drive [17]. In such types of converters, size and performance are dominated by the magnetic components. To develop a compact power-conversion unit with improved performance, the classical power inductor in the half-bridge DC–DC converter is replaced with a VI as shown in Figure 1. The VI topology used in the half-bridge converter is the two-core decoupled topology shown in Figure 1; detailed explanations of such topologies are presented in [8, 9].

FIGURE 1 Variable inductor-based power-conversion architecture
**TABLE 1** Converter specifications

| Parameter                | Symbol | Value |
|--------------------------|--------|-------|
| Input voltage range (V)  | $V_i$  | 40-85 |
| Nominal input voltage (V)| $V_{i,nom}$ | 48    |
| Output voltage range (V)| $V_o$  | 90-120|
| Nominal output voltage (V)| $V_{o,nom}$ | 96    |
| Current ripple\(^a\)    | $\Delta i$ | 30\% |
| Voltage ripple           | $\Delta V$ | 2\%   |
| Nominal power (kW)       | $P_{nom}$ | 33    |
| Peak power (kW)          | $P_{max}$ | 82    |
| Pulse-width modulation switching frequency (kHz) | $f_{sw}$ | 20    |

\(^a\)Selected within a range that minimises inductor size and maximises the flexibility of the auxiliary winding current control.

The control layer consists of the main converter control and a strategy layer. The strategy decides the mode of operation of the auxiliary winding and estimates the amount of auxiliary winding current needed to regulate the saturation level of the magnetic core. The proposed strategy helps maintain inductance at higher currents and stabilise the controllers.

Unlike the classic constant-value model of the power inductor, the inductance is modelled through a 2-D look-up table. The inductance versus current curve is obtained from FEA simulations. The inductance is calculated from the linkage flux obtained from the FEA. The model of the inductor, which shows the dependence of inductance on the main and auxiliary winding currents, was created in a 2-D look-up table in a Matlab/Simulink\textsuperscript{TM} environment and integrated into the overall study system.

From the given specifications of a desired application, the parameters of the DC–DC converter can be selected based on the design procedures presented in [9]. Based on the specifications provided in Table 1, the passive components and operating range of duty cycles are estimated as shown below.

These specifications are obtained from the requirements of the three-wheeled recreational vehicle [18] shown in Figure 2.

In boost/buck operating modes, the range of duty cycle operations, inductance and capacitance are estimated using the formulas presented in Table 2.

In Table 2, $D_{min}$, $D_{nom}$, $D_{max}$, $V_{i,min}$, $V_{i,nom}$, $V_{i,max}$, $V_{o, min}$, $V_{o, nom}$ and $V_{o, max}$ are the minimum duty cycle, nominal duty cycle, maximum duty cycle, minimum input voltage, nominal input voltage, maximum input voltage, minimum output voltage, nominal output voltage and maximum output voltage, respectively.

The minimum inductance required for bidirectional operation can be selected as the maximum of the inductances in row 7 of Table 2, as shown in (1). To assure CCM operation, an inductor value greater than the minimum values in (1) should be selected:

$$L_{\text{min}} = \max(L_{\text{boost}}, L_{\text{buck}})$$  \(1\)

3 | MODELLING AND CONTROL

3.1 | Variable inductor modelling using finite element analysis

The modelling of the VI with different levels of main and auxiliary winding currents is obtained from thorough FEA simulations. A power inductor designed to meet the requirements of a powertrain DC–DC converter for a three-
wheeled recreational vehicle is presented in [18]. For different levels of currents, the inductance, linkage flux and permeability versus currents are investigated, and the required curves are obtained to provide the characterisation and model of the VI. The main winding currents vary from 0 to 2500 A in steps of 1 A, and the auxiliary winding currents vary from 0 to 5 A in steps of 0.5 A. The main currents are high due to the limit in the DC-link voltage in the three-wheeled recreational vehicle prototype. The inductance characterisation as a function of the main and auxiliary winding currents is shown in Figure 3. As is revealed in Figure 2, the inductance can be regulated continuously by injecting a DC current into the auxiliary winding.

The inductance variation of the VI is derived from the linkage flux information using the procedures presented in detail in [14]. The characteristics curves obtained in this section are used in the model of the power inductor to explore the features of the VI for improving the performance of powertrain DC–DC converters. Furthermore, the same model is integrated into the powertrain converter to evaluate the benefits of the VI in enhancing the lifetime of ESSs and reducing the stress in the switching devices. As is revealed in Figure 3, the saturation level of the power inductor is increased significantly as the levels of current in the auxiliary winding increase.

3.2 Converter modelling

Using the modelling procedures presented in [19], the modelling of the half-bridge converter is illustrated in detail in this section. The state variables are $i_L$, and $v_{dc}$. The two switches of the converter are controlled in a complementary manner. The state equations for the different operating modes of the converter can be derived as shown below.

Mode 1: $S_2$ is on and $D_1$ is off. Applying Kirchhoff’s laws in this mode of operation, the dynamic equations that describe the inductor voltage and the capacitor current can be expressed as

$$\frac{di_{s}(t)}{dt} = -\frac{v_{s}(t)}{L}$$

$$\frac{dv_{dc}(t)}{dt} = \frac{v_{dc}(t)}{RC_{dc}}$$

Mode 2: $S_2$ is off and $D_1$ is on. The inductor and capacitor voltages are expressed by a set of differential equations as in (4):

$$\frac{di_{s}(t)}{dt} = \frac{v_{s}(t)}{L} - \frac{(1-d(s))v_{dc}(t)}{L}$$

$$\frac{dv_{dc}(t)}{dt} = \frac{i_{s}(1-d(s))}{C_{dc}} - \frac{v_{dc}(t)}{C_{dc}R}$$

where, $i_s, v_s, v_{dc}, d_s, L$ and $R$ are the current across the power inductor, source voltage, DC-link voltage, duty cycle, inductance of the power inductor and equivalent resistance of the load, respectively. After linearising (4) and (5), neglecting the DC and second-order AC quantities, the average model of the converter can be expressed as

$$\frac{d\bar{i}_{L}(t)}{dt} = \frac{\bar{v}_{s}(t)}{L} - \frac{(1-D)\bar{v}_{dc}(t)}{L} + \frac{\bar{d}(t)V_{dc}}{L}$$

$$\frac{d\bar{v}_{dc}(t)}{dt} = \frac{(I_{L}\bar{d}(t) + \bar{i}_{L}(t) + \bar{i}_{L}(t)D)}{C_{dc}} - \frac{\bar{v}_{dc}(t)}{C_{dc}R}$$

After simplifying and taking the Laplace transform, the transfer functions can be expressed as

$$\frac{\hat{i}_{s}(s)}{d(s)} = \frac{(C_{dc}V_{dc}s + 2(1-D)I_s)}{(LC_{dc})s^2 + \frac{1}{R} + s + (1-D)^2}$$

$$\frac{\bar{v}_{dc}(s)}{\bar{i}_{s}(s)} = \frac{(1-D)V_{dc} - (LI_{s})s}{C_{dc}V_{dc}s + 2(1-D)I_s}$$

The expressions in (8) and (9) are the control to input current, and current to output small signal transfer functions, respectively.
3.3 Control design using affine parameterisation

The inner-loop and outer-loop controller transfer functions are shown in (8) and (9). The inner-loop plant is a second-order system, and the outer-loop plant is a first-order system with non-minimum phase zeros (NMPZs). Both plants depend on inductance, and inductance itself is prone to variations caused by changes in the current running across it. This and the presence of NMPZs add additional control challenges. To deal with these issues, affine parameterisation is used. Unlike the Ziegler–Nichols method based on experimental tests, the affine parameterisation is a model-based simple method that can be used to deduce controllers easily in a systematic way [20]. Furthermore, it ensures the stability of the system with variations in operating conditions, which is the case for powertrain DC–DC converters. Unlike classic approaches, this method starts with the nominal closed-loop transfer function and enables open-loop design to closed-loop implementation [21].

Figure 4 reveals the standard closed-loop system with a negative feedback and affine parameterisation illustration.

For a given transfer function \( G(s) \) (it can be (8) when the inner loop is considered or (9) when the outer loop is considered) and a controller \( C(s) \), the closed-loop transfer function \( T(s) \) is expressed as in (10):

\[
T(s) = \frac{G(s)C(s)}{1 + G(s)C(s)} \quad (10)
\]

Introducing \( Q(s) \) in series with the transfer function \( G(s) \), as shown in Figure 4b, \( Q(s) \) is selected in such a way that it provides equivalence to the classical closed-loop transfer function as in (11):

\[
T(s) = Q(s)G(s) \quad (11)
\]

Depending on the relative degree (the difference of the degree of the denominator and degree of the numerator) and type of \( G(s) \), the stability situation and the presence of NMPZ, \( Q(s) \) is derived as

\[
Q(s) = F_Q(s)G'(s) \quad (12)
\]

where \( F_Q(s) \) is defined for the performance that is expected considering the dynamic characteristics of the plant. \( G'(s) \) is \( (G(s))^{-1} \) for stable systems with minimum phase zero plants. On the other hand, NMPZs are excluded from \( (G(s))^{-1} \). From (10) and (11), the controller can be deduced from the \( Q \) plant and the transfer function of the plant as shown in (13):

\[
C(s) = \frac{Q(s)}{1 - Q(s)G(s)} \quad (13)
\]

For the study system, the inner-loop current controller and outer-loop voltage controller are designed using the procedures presented in this section.

As shown in (8), the inner-loop transfer function has a relative degree of 1 with one minimum phase zero and two stable poles; subsequently, \( F_Q(s) \) with a relative degree of 1 is suitable for estimating the \( Q(s) \) function and can be expressed as

\[
F_Q(s) = \frac{1}{1 + \alpha s} \quad (14)
\]

where \( \alpha \) is a tuning parameter and is equal to the reciprocal of the desired natural frequency \( (\omega_v/\alpha) \) and desired damping factor \( (\nu_v/\alpha) \):

\[
Q_v(s) = \frac{s^2 + \frac{1}{RC_d} + \frac{(1-D)^2}{(LC_d)}}{2(1-D)/LC_d(1 + \alpha s)(C_d V_d/(1-D)/LC_d) s + 1} \quad (15)
\]

where \( Q_v(s) \) is the product of (14) and the inverse of (8). The outer loop has one pole and one NMPZ. For the types of plants shown in (9), the NMPZ is eliminated for deducing \( Q_v(s) \) [20]. In this case, \( Q_v(s) \) will be the product of (14) and the inverse of (9) excluding the NMPZ. When the NMPZ is eliminated, the relative degree of (9) is 1, hence \( F_Q(s) \) can be selected in the same way as (14) with values of \( \alpha \) different from (14), and \( Q_v(s) \) will be as shown in (16):

\[
Q_v(s) = \frac{1 + C_d V_d}{2(1-D)/LC_d} \quad (16)
\]

Using (15) and (16), the current and voltage controllers are designed using the equations shown in Table 3. The PID controller shown in (17) is commonly used in industrial applications. Unlike with other methods, the type of controller is deduced from (13). The voltage controller obtained using the affine parameterisation is equivalent to the PI controller, and the controller parameters are calculated as in Table 3. Similarly, the controller for the current controller is equivalent to the PID controller, and its parameters are determined as shown in column 2 of Table 3.
TABLE 3 Controllers for the main converter

| Voltage controller parameters | Current controller parameters |
|------------------------------|------------------------------|
| $a_{p/v}$ | $2i_{p/v}$ | $2i_{p/v}$ |
| $\tau_{p/v}$ | 0.707 | 0.707 |
| $K_{p/v}$ | $\frac{\tau_{p/v}}{K_{p/v}}$ | $\frac{2\pi V_{dc}}{2V_{dc}}$ |
| $K_{i/v}$ | $\frac{i_{p/v}}{K_{i/v}}$ | $\frac{\pi V_{dc}}{\pi V_{dc}}$ |
| $K_{d/v}$ | 0 | $\frac{\pi V_{dc}}{\pi V_{dc}}$ |
| $T_{d/v}$ | 0 | $\frac{\pi V_{dc}}{\pi V_{dc}}$ |

where $K_{p/v}, K_{i/v}, K_{d/v}$ and $T_{d/v}$ are the proportional gain, integral gain, differential gain and filter time constant in respective order.

Relating the gain parameters of the PID to the controller obtained using affine parameterisation, the values of the controllers’ gains are derived as shown in Table 3.

In Table 3, $K_{a,v} = \frac{V_{dc}}{s^2}, \tau_{p,v} = \frac{C_{a,v}V_{dc}}{2\pi(1-D)V_{dc}}, \tau_{i,v} = \frac{(1-D)}{(1-D)} \sqrt{\frac{LC_{a,v}}{\omega}}, K_{o,j} = \frac{2((1-D)\beta)}{2(1-D)\beta}$, and $\beta = \frac{C_{a,v}V_{dc}}{2(1-D)\beta}$ are the parameters of the transfer functions.

3.4 Current follower strategy

The proposed current follower strategy determines the amount of auxiliary winding current based on the main winding levels of the current. It enables smooth operation of the VI depending on the levels of the main winding currents. The strategy is developed based on the linkage flux ($\lambda$) versus main winding current ($i_{s}$) characterisation curves obtained from offline FEA simulations.

The $\lambda$-$i_{s}$ characterisation curves of the magnetic core material are obtained from a series of FEA simulations for different current levels in the main and auxiliary windings. For each $\lambda$-$i_{s}$ curve, the point of maximum value of the slope of the change of linkage flux to the change of the main winding current ($\frac{d\lambda}{di}$) is determined. A moving slope estimation approach is used to determine the maximum slope of each curve. Accordingly, the corresponding values of main and auxiliary winding currents, which provide the maximum slope of each curve, are used to determine the strategy that relates the two currents. A simple mathematical relationship as in (18) is deduced using the curve-fitting tool in Matlab®. This strategy enables smooth operation of the power inductor for different loading situations by controlling the magnetic properties of the magnetic core material.

In normal operation, that is, for currents lower than the saturation level of the magnetic core, the auxiliary winding will be idle. As a result, the current follower strategy sets the reference current of the auxiliary winding to zero. For current levels higher than the saturation current, the auxiliary winding current reference will be set according to the analytical expression shown in (18). The proposed strategy is simple, and it can be easily implemented in microcontrollers for real-time applications:

$$i_{b}(i_{s}) = \begin{cases} 
0, & i_{s} < i_{s,\text{sat}} \\
ai_{s}^2 + bi_{s} + c & i_{s,\text{sat}} \leq i_{s} \leq 2500 
\end{cases} \quad (18)$$

where $a = 3.52.10^{-6}, b = 0.0159$ and $c = -12.96$.

4 SIMULATION AND EVALUATION

4.1 Simulation system

The vehicle under study is a three-wheeled sports vehicle, and detailed specifications of the vehicle are presented in [18]. The objective of the simulations is to illustrate the impact of the variable inductor and the proposed strategy for powertrain DC–DC converters for such types of vehicles. Furthermore, simulations are used to evaluate the controllers of the main converter and the proposed strategy for a full-scale VI-based converter. The full-scale parameters of the converter are used to evaluate the suitability of the proposed current follower strategy and the controllers considering the varying nature of the target application.

To validate the model, controllers and strategy, the simulations are carried out in Matlab/Simulink™ as shown in Figure 5.

4.2 Evaluation

To verify the suitability of the strategy for controlling the variable inductor for powertrain DC–DC converters and the proposed control approach, several parameters such as the current ripple ratio, inductance variation, efficiency, and component stresses are considered to evaluate the merits of the VI for powertrain DC–DC converters.

Inductance curves and current ripple variations are essential parameters that can be used to evaluate the performance improvements of the developed system and control strategy. Figure 6 shows the demanded load power, inductance currents, inductance and auxiliary winding current profiles in respective order. As revealed in Figure 6b, current ripple is significantly reduced when an auxiliary winding current is applied in accord with the proposed strategy. Similarly, inductance can be controlled in accord with the load variations as shown in Figure 6c. The dotted plot shows that inductance is regulated even if the main winding currents are higher than the saturation currents of the power inductor. As revealed in Figure 6d, the profile of the auxiliary winding current is determined in accord with the demanded load based on the strategy in (18).
To further explore the merits of the VI and the proposed strategy, switching device stress, battery current variation and efficiency are evaluated.

The switching stress in DC–DC converters is quantified as switch stress parameter ($SS_{P_i}$), which is expressed for the $i$th switch in the converter as in (19):
current and is responsible for losses in power electronic converter components:

\[ I_{\text{rms}} = \sqrt{\frac{1}{n} \sum_{k=0}^{n} (I_s(k))^2} \]  

(20)

The standard deviation of the battery current, which is estimated as in (21), is also an important parameter. It shows the spread of current variation from the continuous value of the battery current. As one of the key parameters, it affects the lifetime of energy storage devices. A higher value reveals that the battery is exposed to a wider range of current variations, which does not bode well for its lifetime.

\[ \sigma = \sqrt{\frac{1}{n} \sum_{k=0}^{n} (I_s(k) - \mu)^2} \]  

(21)

where \( \mu \) is the mean current of the battery for the time under consideration.

The coefficient of variation, which provides the ratio of the standard deviation to the mean of the battery current, is an essential parameter that gives the degree of variation of the battery current, which is related to the lifetime of the battery. A smaller value signifies that the degree of variation of the battery current is small; hence, the impact on the lifetime of the battery is smaller. The coefficient of variation is expressed as

\[ C_v = \frac{\sigma}{\mu} \]  

(22)

The evaluations of the switching stress and energy storage stress parameters are summarised in Figure 7a. The normalised values of \( I_{\text{rms}} \), \( \sigma \), SSP and \( C_v \) for the converter under study with conventional inductance and no control strategy (without control) and with VI and control strategy (with control) are shown in Figure 7a. For the considered simulation time and load profile shown in Figure 6a, decreases of 6.15% in \( I_{\text{rms}} \), 10.55% in \( \sigma \), 17.87% in SSP and 6.73% in \( C_v \) are noted in Figure 7a. This reveals significant improvements in reducing the switching stress on switching devices and ESSs due to the magnetic control and the proposed strategy.

By dynamically regulating the inductance of the power inductor with the help of the proposed strategy, current ripple of the power inductor is regulated. As a result, the peak-to-peak switching current will be controlled, leading to a reduction in the switching stress of the switching devices as shown in Figure 7a.

Furthermore, current variations and stress levels affect the lifetime of the energy storage (battery). The peak current variations, RMS current, continuous current, and coefficient of variation are used to quantify the energy storage stress and its lifetime.

RMS current, which is estimated as in (20), depends on current ripple levels, is related to the heating effect of the
important metric of comparison, efficiency is considered in evaluating the VI-based converter and the proposed strategy.

In estimating efficiency, the losses due to passive components including the auxiliary winding, main winding, energy source power losses and active component power losses are considered.

Similarly, the conduction losses of the power switches will be affected by the levels of current ripples considering the insulated-gate bipolar transistor (IGBT) switching device [22]. This can be illustrated using the mathematical relationships that follow. Assuming that \( r \) is the ratio of peak-to-peak ripple current \( I_{p-k-pk} \) to the average current \( I_{savg} \) across a power inductor,

\[
P_{C,IGBT} = 2I_T V_{CIST} + 2(1 - D)I_a V_F + 2r(I_{T,RMS})^2
\]  
(23)

\[
I_{T,RMS} = I_{savg} \sqrt{D \left( 1 + \frac{r^2}{12} \right)}
\]  
(24)

where \( I_{IGBT} \) is the average IGBT current, \( V_{CIST} \) is the voltage drop across the IGBT, \( I_a \) is the output current, \( V_F \) is the forward diode voltage, \( r_c \), is the collector emitter on state resistance and \( I_{T,RMS} \) is the IGBT RMS current.

Similarly, the conduction losses in a diode can be expressed as

\[
P_{C,D} = 2V_D I_D + 2r_D(I_{D,RMS})^2
\]  
(25)

where \( V_D \), \( I_D \), \( r_D \) and \( I_{D,RMS} \) are the diode forward voltage, average diode current, diode forward resistance and diode RMS current, respectively. In the same fashion, current ripple affects the last term in (25). Figure 7b reveals the efficiency curves of the powertrain DC–DC converter with VI and control strategy (dotted lines) and with classical inductor and no strategy (solid line) for the load profile shown in Figure 6a. With the VI and proposed strategy, efficiency improvements of up to 13.70% are obtained for currents much higher than the rated current of the power inductor.

The results shown in Figure 7a, b summarise the effectiveness of the proposed strategy and the benefits of VI integration to powertrain DC–DC converters.

5 | CONCLUSION

A simple strategy is presented for controlling the magnetic properties of the VI, and the viability of the approach in improving the performance of large-scale powertrain DC–DC converters is explored. The proposed strategy is introduced to estimate auxiliary winding current as a function of main winding current, hence enabling the smooth functionality of VIIs in powertrain DC–DC converters. Integration of the VI into a large-scale powertrain DC–DC converter, its control and the proposed performance improvement strategy are evaluated through simulations. The results reveal significant improvements in the ageing index of ESSs (6.73% in the battery current coefficient of variation) and switching stress parameter (17.87% reduction). Furthermore, efficiency gains of up to 13.70% were noted. Future research will focus on experimental validation of the proposed strategy and implementation of the large-scale VI-based powertrain DC–DC converter for the three-wheeled recreational vehicle prototype.

ACKNOWLEDGEMENTS

This work was supported in part by Grant 950-230672 from the Canada Research Chairs Program, in part by Grant RGPIN-2017-05924 from the Natural Sciences and Engineering Research Council of Canada, in part by the Arbour Foundation Scholarship, University of Sherbrooke, Faculty of Engineering Excellence Scholarship and by projects UIDB/00308/2020 and by the European Regional Development Fund through the COMPETE 2020 Program and FCT – Portuguese Foundation for Science and Technology within projects ESGRIDS (POCI-01-0145-FEDER-016434) and MAneGER (POCI-01-0145-FEDER-028040).

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How to cite this article: Beraki MW, Trovão JPF, Perdigão MS. Performance enhancement of powertrain DC–DC converter using variable inductor. IET Electr. Syst. Transp. 2021;11:161–170. https://doi.org/10.1049/els2.12014