Voltage Control for Enhanced Power Electronic Efficiency in Series Hybrid Electric Vehicles

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Abstract—The paper presents a DC-link voltage control scheme by which the power losses associated with the power electronic converters of a series Hybrid Electric Vehicle (HEV) powertrain are reduced substantially. A DC-link commonly connects the three powertrain branches associated with series HEVs, presently interfaced by a three-phase rectifier, a three-phase inverter and a Dual Active Bridge (DAB) DC/DC converter. Dynamic efficiency models of the converters are developed and a methodology is proposed by which the DC-link voltage is varied with respect to its default value, based on the ratio between the battery and DC-link voltages. The voltage control scheme introduced varies the phase shift between the gating signals of the two DAB converter bridges, proportionally to the ratio of converter input voltage to output voltage referred to the transformer primary. This level of instantaneous control forces the converter to operate in boost mode when the battery charges and buck mode when the battery discharges, allowing the converter to persistently avoid Hard Switching (HS) losses over its entire operating range. The control scheme is tested in simulations with a full HEV model by comparing its performance with constant voltage and unity voltage conversion ratio PI control schemes. The scheme proves most effective for vehicles with high hybridization factor, driving in an urban environment.

I. INTRODUCTION

SINCE 1970, the global Greenhouse Gas (GHG) emissions from transportation have increased by more than a factor of two. The contribution from road vehicles towards transportation emissions has increased over this period from 59.85% to 72.06% [1]. The HEV represents a step towards sustainable transportation, by reducing fossil fuel consumption [2]. By minimizing power loss within the HEV powertrain, power demand is reduced for a given road load, and hence the sustainable impact of this technology is enhanced. As mentioned in [3], efficiency is one of the key drivers behind the next generation of vehicles, and significant progress has been made through intelligent supervisory controllers over the past decade [4]. This paper proposes a DC-link voltage control scheme which enhances the combined efficiency of the three power electronic converters of a series HEV powertrain, in a further step towards achieving this goal.

A wide range of research exists on HEVs for various architectures [5]. For topologies in which the voltage of the battery and DC-link are not equal, the inclusion of a DC/DC converter is necessary. The literature reviewed included standard boost converters [6], [7], three-level converters [8], and dual bridge isolated converters [9]–[16]. Transformer isolation provides advantages in terms of personnel safety and noise reduction [17]. Less common arrangements analyzed include the combined SEPIC and Luo converters detailed in [18]. The inclusion of a DC/DC converter enables control the DC-link voltage via the gating signals of the integral power electronic switches. Despite this, lack of physical space in the vehicle sometimes results in exclusion of the DC/DC converter in the HEV design [6]. An understanding of optimal voltage control methods lends itself to enhancing overall HEV efficiency, as the DC-link voltage influences the performance of all power electronic converters within a series powertrain.

Several approaches have been proposed for variable DC-link voltage and DC/DC converter control. A polynomial DC/DC converter control scheme is detailed in [19], which emphasizes the improvement of operation robustness as compared to PI control, rather than improving energy efficiency. Similarly, [20] also aims to improve operation reliability by implementing a constant modulation index control for the DC-link voltage. The work presented in [6] compares the performance of two electric drive systems. The first comprises of a battery-powered inverter, while the second incorporates a bidirectional boost converter between the battery and inverter, which facilitates DC-link voltage control. Both systems supply a Permanent Magnet Synchronous Motor (PMSM). A scheme is developed which varies the DC-link voltage proportionally with PMSM speed. The performance of the first topology is compared with that of the variable voltage bidirectional scheme, and the latter is shown to achieve improvements in terms of global drive, inverter and PMSM efficiencies. Although these single source models are less complex than the dual source model presented in this paper, the achievements of [6] have acted as inspiration for the work herein.

The main contribution of this paper is a proportional control law for the phase shift between the switching waveforms of the bridges of a DAB DC/DC converter which interfaces with the DC-link and battery of a series HEV powertrain, to control the dynamic DC-link voltage. The controller achieves a reduction of both DC/DC converter and total powertrain power electronic losses for an urban drivecycle, while achieving approximately the same level of total powertrain power electronic losses as the most efficient PI control scheme tested, for a highway drive cycle. A further contribution is the development of the

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Fig. 1. Block diagram of the series HEV powertrain used in this work.

Fig. 2. Control schematic showing the interconnection of the internal combustion engine (ICE), PMSG, rectifier, battery, DC/DC converter, inverter, PMSM, continuously variable transmission (CVT) and car, and the related control loops.

methodology to select the best nominal DC-link voltage to maximize the efficiency of the inverter and rectifier.

The paper is structured as follows. Section II details: (a) the base HEV model that is used to conduct the study, (b) respectively the modeling of the power loss and analysis of performance of the inverter and rectifier for varying motor and generator operating conditions, and (c) operating mode dependent switching and conduction loss models for the DAB converter. Section III describes the Constant Voltage PI (CVPI), Unity Ratio PI (URPI) and the proportional DC-link control schemes developed and tested in the paper. Simulation results are presented in Section IV, comparing the performance of the proposed control schemes and showing the energy savings achieved by the best scheme. Final conclusions drawn are presented in Section V.

II. MODELING

The HEV model used in this paper is a modified version of that presented in [21] and [22]. Changes are made to this model in relation to the inverter and rectifier modulation strategy, and the DC/DC converter model. Furthermore, dynamic efficiency modeling is also introduced for all the converters. Conduction and switching losses of the inverter and rectifier are dependent on converter modulation indices, which also represent boundaries in relation to feasible motor and generator operating regions. The switching loss of these converters is proportional to the switching frequency, and current through and voltage over the switched elements. Conduction loss of

the DC/DC converter is dictated by RMS and average currents. Switching varies between high loss and lossless depending on the operating variables. The HEV model and inverter, rectifier and DC/DC converter loss models are detailed in this section.

A. Vehicle Model

The block diagram of the HEV powertrain is given in Fig. 1 and the control schematic is shown in Fig. 2. A diesel engine and DC battery respectively act as the primary and secondary HEV energy sources. The relative proportion of power supplied by the two sources is decided by a Supervisory Control System (SCS), based on the motor load and the battery State of Charge (SOC). The Thermostat SCS is given in [23] and [24], and works as follows. The engine will either be operating at its optimal point (25 kW, 1600 rpm) or producing zero power. When inverter demand exceeds 25 kW, both sources supply power together in hybrid mode. Engine only supply occurs when the battery SOC is at or below 50%, and will continue to supply until the SOC reaches 80%. At this stage, the battery discharges, in battery-only operation, until its SOC reaches 50% [25].

The PMSG supplies power to the rectifier and the PMSM accepts power from the inverter. PMSM torque is relayed to the wheels through a continuously variable transmission. As detailed in [21], for both the PMSG and PMSM, the direct current is controlled to 0 A. To achieve this, the reference currents $i_{dm,ref}$ and $i_{dg,ref}$ given in the powertrain control diagram (Fig. 2) are set to 0 A. The respective quadrature

Fig. 3. The series HEV powertrain includes the PMSG, rectifier, battery, DAB converter, inverter and PMSM, as illustrated by the circuit diagram. Symbols R , L and e represent phase resistances, inductances and induced emfs respectively. Subscripts a , b and c correspond to the individual phases, and subscripts g , m and ref correspond to ‘generator’, ‘motor’ and ‘reference’. E_{bat} and R_{bat} correspond to the battery emf and internal resistance, while L is the inductance associated with the DC/DC converter and C_0 is the DC-link capacitance. i_{PS} , i_{SS} and i_{PL} are the primary source (PMSG-branch), secondary source (battery-branch) and propulsion load (PMSM-branch) DC currents. The control signals are v_{dc} (the DC-link voltage), P_{PS} (the primary source power) and u_{car} (the forward speed of the car).

currents of the PMSG and PMSM are controlled to manipulate the machine torque to achieve desired PMSG power output and vehicle speed. The base model given in [21] assumes constant efficiencies of 94.61% for the inverter and rectifier, and 96% for the DC/DC converter. The work done in this paper, replaces these figures with dynamic efficiency models. Prior to the present work, PI control has been used to maintain a DC-link voltage of 700 V.

B. Inverter and Rectifier

The inverter and rectifier operate based on a sinusoidal Pulse Width Modulation (PWM) method, as opposed to the Space Vector Modulation (SVM) form of PWM assumed in the base model. As shown in Fig. 3 the rectifier design is the same as that of the inverter, although they generally realize opposite goals. The inverter converts DC to three-phase AC when power is demanded by the motor, only operating conversely during regenerative braking. The rectifier converts three-phase AC to DC. Switching is dictated by three-phase sinusoidal PWM (described in [26] and [27]), which compares three sinusoidal modulating signals (at the desired motor voltage frequency) with a common high-frequency triangular carrier

signal. Linear modulation occurs if the amplitude of the modulating signals is below that of the carrier signal. The converter modulation index (M) is the ratio of modulating and carrier signal amplitudes [27]. To avoid overmodulation, the following condition must be obeyed:

$$0 \leq M \leq 1 \quad (1)$$

For a given DC-link voltage (v_{dc}), the amplitude of AC phase voltage (v_{ph}) of the inverter and rectifier is limited by its modulation index [27]:

$$v_{ph} = \frac{M v_{dc}}{2} \quad (2)$$

For $M = 1$ this equation represents a constraint for the feasible operating region of the PMSG and PMSM, denoting the boundary between linear modulation and overmodulation. This constraint varies for a varying DC-link voltage. The conduction losses of the inverter are approximated by [28], [29]:

$$P_{cond} = 6 \left(i_{pk} v_{f0} \left(\frac{1}{2\pi} - \frac{M}{8} \right) + i_{pk}^2 r_f \left(\frac{1}{8} - \frac{M}{3\pi} \right) \right)$$

$$+ r_{ce} i_{pk}^2 \left(\frac{1}{8} + \frac{M}{3\pi} \right) + v_{c0} i_{pk} \left(\frac{1}{2\pi} + \frac{M}{8} \right) \right), \quad (3)$$

where i_{pk} is the peak AC current from the inverter, v_{f0} is the diode forward voltage corresponding to zero current, r_f is the diode forward resistance, r_{ce} is the IGBT collector emitter resistance and v_{c0} is the IGBT forward voltage corresponding to zero collector current. The power factor terms shown in [28] are omitted in Eq. 3. As mentioned previously, PMSM and PMSG direct currents are controlled to 0 A, which yields close to unity power factor for non-salient structures ($L_d = L_q$) except at very high loads [30], [31]. This approximation is sufficiently accurate for this work. The switching losses of the inverter are given by [29]:

$$P_{sw} = 6 \frac{f_i v_{dc} i_{pk}}{v_{ref} i_{ref} \pi} (E_{on,ref} + E_{off,ref} + E_{rr,ref}) \quad (4)$$

where f_i is the switching frequency (carrier signal frequency) of the inverter, $E_{on,ref}$ is the reference IGBT turn on energy loss, v_{ref} is the voltage at which reference energy loss is measured, i_{ref} is the current at which reference energy loss is measured, $E_{off,ref}$ is the reference IGBT turn off energy loss and $E_{rr,ref}$ is the reference diode reverse recovery energy loss. Reference values are obtained from the corresponding device datasheet. Summing the conduction and switching losses to acquire the total power loss P_{loss}^{inv} , the dynamic efficiency expression of the inverter is found. The expression differs based on the direction of power flow:

$$\eta_{inv}^{dc-ac} = \frac{T_{em} \omega_{rm} + \frac{3}{2} R_m i_{qm}^2}{T_{em} \omega_{rm} + \frac{3}{2} R_m i_{qm}^2 + P_{loss}^{inv}}, \quad (5)$$

$$\eta_{inv}^{ac-dc} = \frac{T_{em} \omega_{rm} + \frac{3}{2} R_m i_{qm}^2 + P_{loss}^{inv}}{T_{em} \omega_{rm} + \frac{3}{2} R_m i_{qm}^2} \quad (6)$$

where T_{em} , ω_{rm} , i_{qm} and R_m are the electromagnetic torque, mechanical speed, quadrature current and stator resistance of the PMSM. The term proportional to the square of quadrature current is equal to the PMSM stator resistive power loss as per, while the numerator of Eq. 5 corresponds to the total power into the PMSM out of the inverter. Conduction, switching and total rectifier losses are calculated using the same expressions as for the inverter, but using rectifier variables. However, as PMSG power flow is unidirectional, there is a single expression for rectifier efficiency, in which P_{loss}^{rec} refers to the sum of rectifier conduction and switching loss:

$$\eta_{rec} = \frac{T_{eg} \omega_{rg} - \frac{3}{2} R_g i_{qg}^2 - P_{loss}^{rec}}{T_{eg} \omega_{rg} - \frac{3}{2} R_g i_{qg}^2}, \quad (7)$$

where T_{eg} (positive for power generation), ω_{rg} , i_{qg} and R_g are the electromagnetic torque, mechanical speed, quadrature current and stator resistance of the PMSG. Similarly to the PMSM equations, the term proportional to the square of quadrature current is equal to the PMSG stator resistive power loss, while the numerator of Eq. 7 is the total power out of the PMSG into the rectifier.

Inverter and rectifier performances are mapped for a number of design DC-link voltages. Both converter switching frequencies are 20 kHz. IGBT modules for these converters are selected for each design from the range offered by

TABLE I
IGBT MODULE SELECTION FOR RECTIFIER, INVERTER AND DAB CONVERTER

IGBT Module	v_{dc}^{max}	Converter
FF300R07ME4_B11	650 V	Inverter
FS300R12KE4	1200 V	Inverter
FS150R07PE4	650 V	Rectifier
FS150R12KT4_B11	1200 V	Rectifier
FF100R12RT4	1200 V	DAB
FF300R07ME4_B11	650 V	DAB

Infineon, by considering the voltage switched, and the RMS and peak current through each switch. Worst case RMS and peak currents are calculated based on maximum torque and speed of the PMSM and PMSG. Based on this means of switch selection, the rectifier and inverter IGBT modules used for different DC-link voltage designs are listed in Table I. Parameters were extracted from the corresponding datasheets for efficiency calculations. The efficiency of the inverter is plotted for varying PMSM load torque and mechanical speed. Maximum load torque and mechanical speed are 400 Nm and 5000 rpm [32]. These represent operating constraints.

Efficiency maximization is only considered for forward driving neglecting regeneration. Therefore, optimization region boundaries are present at 0 Nm and 0 rpm. The final operating constraint is that corresponding to the boundary of linear modulation. Design DC-link voltages were tested in 100 V intervals between 400 V and 1200 V inclusive, however 400 V results in overmodulation of the inverter for the drive cycles considered in Section IV. Results are given for 700 V and 500 V in Figs. 4 and 5 respectively. The lower the DC-link voltage, the smaller is the feasible operating region of the PMSM, but the higher is the low speed efficiency. The same

Fig. 4. Efficiency Map of Inverter ($v_{dc}^{ref} = 700$ V).

Fig. 5. Efficiency Map of Inverter ($v_{dc}^{ref} = 500$ V).

Fig. 6. Efficiency Map of Rectifier ($v_{dc}^{ref} = 700$ V).

Fig. 7. Efficiency Map of Rectifier ($v_{dc}^{ref} = 500$ V).

can be said for the rectifier and PMSG, based on Figs. 6 and 7. The diagonal constraint represents maximum generator torque (340 Nm) [33].

C. DC/DC Converter

The DC/DC converter in [21] has been changed from a Dual Half Bridge to a DAB topology as it is more appropriate for the required power range of the battery branch. The DAB converter is composed of two bridges of power electronic switches, connected by an isolation transformer. The primary referred circuit is shown in Fig. 3. An important parameter throughout this paper is the voltage conversion ratio (d). This is the ratio of the primary referred converter output voltage, to the converter input voltage (battery voltage):

$$d = \frac{v_o}{nv_i}, \quad (8)$$

where v_i is the converter input voltage, v_o is the converter output voltage and n is the transformer turns ratio. When d is greater/less than 1, the converter is said to be operating in boost/buck mode [34].

Another important parameter is the phase shift (ϕ) between the primary and secondary switch gating signals of the DAB converter, as it determines the direction of power flow. When the phase shift is positive, power flows in the forward direction from the battery to the DC-link, and reverses when phase shift is negative. Average converter power flow is given by the following expression [11], [35]:

$$P_{avg} = \frac{v_i v_o \phi (\pi - \phi)}{2n\pi^2 L f}, \quad (9)$$

where L is the primary referred leakage inductance of the intermediate isolation transformer and f is the switching

Fig. 8. A cycle of inductor current (forward power, boost mode) of the DAB DC/DC converter [10].

frequency of the DAB converter. As shown in [17], the transformer current is a piece-wise linear sinusoidal approximation based on the changing square wave voltages v_p and v'_s . These voltages result from the phase shifted bridge gating signals which have a constant duty cycle of 0.5. The current and voltage dynamics of the converter throughout a switching cycle is detailed in [36].

For forward power, boost operation, the inductor current is as shown in Fig. 8. The initial inductor zero-crossing of the cycle occurs at t^* . The inductor current at different stages in the cycle is represented by $i_L(0)$, $i_L(\phi)$, $i_L(\pi)$ and $i_L(\pi + \phi)$ and are easily quantified [37]. t^* can thus be quantified by acquiring the associated linear relations and calculating the time at which current initially crosses zero in the cycle. For forward power flow:

$$t^* = \frac{(v_i n + \frac{2}{\pi} v_o \phi - v_o)}{4f(nv_i + v_o)} \quad (10)$$

The average and RMS current through each of the IGBTs and diodes can be calculated by quantifying the shaded areas in Fig. 8, and using the relationship between operating mode and conduction of components throughout a switching cycle [38]. Average and RMS current expressions equivalent to those given in [10], were found for all combinations of power flow direction and operating mode. A conduction loss model was subsequently created by using these expressions within the basic transformer [14], IGBT and diode [28] conduction loss equations respectively given as follows:

$$P_{cond}^L = \left(R_p + \frac{R_s}{n^2} \right) i_{rms}^L{}^2 \quad (11)$$

$$P_{cond}^T = r_{ce} i_{rms}^T{}^2 + v_{c0} i_{avg}^T \quad (12)$$

$$P_{cond}^D = r_f i_{rms}^D{}^2 + v_{f0} i_{avg}^D \quad (13)$$

where i_{rms} and i_{avg} imply RMS and average current, and the superscripts L , T and D respectively refer to transformer leakage inductance, an active converter IGBT and an active converter diode.

Complete Zero Voltage Switching (CZVS), Incomplete Zero Voltage Switching (IZVS) and Hard Switching (HS) can all occur within the DAB converter. The boundaries between CZVS and IZVS can be derived with help from [38] and are shown on the $d - \phi$ plane in [34]. As stated in [34], at any primary switching instant, the minimum current through the inductance L for which CZVS can occur is given by:

$$i_{L,min}^{prim} = \frac{2\sqrt{v_i v_o'}}{Z_r}, Z_r = \sqrt{\frac{L}{C}}, \quad (14)$$

TABLE II
CONDITIONS FOR ZVS (IB=INPUT BRIDGE, OB=OUTPUT BRIDGE)

Forward	Reverse	$d < 1$	$d > 1$	$d = 1$
IB	OB	always	$ \phi \geq \frac{\pi(d-1)}{2d}$	always
OB	IB	$ \phi \geq \frac{\pi(1-d)}{2}$	always	always

where v_o' is equal to v_o referred to the transformer primary and C is the switch snubber capacitance. The minimum current for secondary CZVS is found by dividing by the turns ratio. As shown in [38], CZVS always occurs on one or both of the converter bridges. In the former case, which of the two bridges is capable of losing energy during switching is dependent on the direction of power flow and the mode of operation. In the latter case, the converter is said to be in CZVS mode as both bridges switch in a lossless manner. Whether the bridge will lose energy while switching can be determined by comparing the location of the operating point to the CZVS-IZVS and IZVS-HS boundaries on the $\phi - v_o$ or $\phi - d$ plane.

The two IGBTs which switch at lowest current magnitude ($i_L(\phi)$ and $i_L(\pi + \phi)$ in Fig. 8) in the bridge of interest contribute IZVS loss when the inductor current is too low. These switches can be identified using analysis from [38]. By equating the corresponding expression for this current value, to the minimum inductor current of interest (depending on the bridge of interest), linear CZVS-IZVS boundaries can be derived for all four quadrants of the $\phi - v_o$ plane.

Similarly, by equating the minimum inductor current of interest to 0 A, linear IZVS-HS boundaries can be derived for all four quadrants of the $\phi - v_o$ plane also. The nonlinear expressions for these constraints on the $\phi - d$ plane are given in Table II [34], [38]. Based on the location of the converter operating point on the $\phi - v_o$ plane, switching losses are calculated. CZVS is lossless as zero voltage exists over each IGBT at turn-on and turn-off. By analyzing the theoretical waveforms of the voltage over and current through the DAB IGBTs during turn-on and turn-off, it has been found that approximately zero voltage exists over each IGBT at turn-off in IZVS, though this is not the case for IGBT turn-on. Therefore, IZVS loss at a given switching instant is equal to the sum of IGBT turn-on and snubber capacitance loss. The IGBT turn-on loss [29]:

$$P_{sw,IGBT}^{IZVS} = E_{on,ref} \frac{i_L^{sw} v_x}{v_{ref} i_{ref}} \quad (15)$$

and the snubber turn-on loss [12]:

$$P_{sw,snubber}^{IZVS} = C \cdot v_x^2 \quad (16)$$

are respectively shown for a given current through the leakage inductance (i_L^{sw}), and the voltage over the switched IGBT (v_x) during turn-on of a single IGBT. Snubber capacitance loss is a result of a non-zero voltage over the IGBT snubber capacitance during turn-on.

v_x represents the non-zero voltage which is present over the IGBT at the turn-on instant in the case of IZVS. This is 0 V in the case of CZVS. For HS, v_x is equal to the full DC voltage of the converter bridge in question. However for

IZVS, the calculation of v_x depends on the quadrant of the $\phi - d$ plane in which the converter is operating. For forward power boost operation, IZVS occurring on the input bridge uses the following calculation of v_x :

$$v_x = v_i - \frac{i_L(0)t_x}{2C} \quad (17)$$

There are four different calculations of v_x - one for each quadrant. v_o is used instead of v_i , when output bridge IZVS occurs, and i_L^{sw} differs between $i_L(0)$, $i_L(\phi)$, $i_L(\pi)$ and $i_L(\pi + \phi)$ based on which of the four timesteps of the inductor current cycle IZVS occurs. In all IZVS cases, the calculation of v_x is made by subtracting from $v_{i/o}$, the drop in voltage over the IGBT in question, during the time t_x . t_x is the maximum time for voltage to drop to zero over the switch, while achieving CZVS. For input bridge IZVS, this occurs when $i_{L,min}^{prim}$ flows through the converter inductance at the switching instant. t_x is calculated by balancing the energy stored in the converter inductance at the switching instant with the integral of power discharge from the input bridge through the voltage v_o' . The energy balance is as follows:

$$\frac{L(i_{L,min}^{prim})^2}{2} = \int_0^{t_x} v_o'(t)i_L(t)dt = \frac{v_o' i_{L,min}^{prim} t_{x,prim}}{2} \quad (18)$$

The premise of this equation is that for CZVS, the energy stored in the converter inductance is completely discharged prior to IGBT turn-on. This is the boundary condition between CZVS and IZVS as within the IZVS region, a portion of this energy will be lost in switching. Solving for $t_{x,prim}$:

$$t_{x,prim} = 2\sqrt{\frac{nLCv_i}{v_o}} \quad (19)$$

The drop in voltage during the time t_x , (shown in Eqn. 17) is calculated using the standard equation for current through a capacitor, under the assumption of linearly decreasing voltage. Note that the assumption of linearly decreasing voltage implies a minimum rate of decrease of voltage to achieve CZVS, which occurs at the aforementioned boundary condition. Eqn. 17 also assumes that half of the inductor current flows through both of the active bridge switches during an occurrence of IZVS on that bridge. These calculations can also be followed for output bridge IZVS. The complete expression for IZVS loss of a bridge is given by:

$$P_{sw}^{IZVS} = 4f \left(E_{on,ref} \frac{i_L^{sw} v_x}{v_{ref} i_{ref}} + C(v_x)^2 \right), \quad (20)$$

HS loss of a bridge is calculated in a similar manner. Analysis of the theoretical current and voltage waveforms associated with the voltage over and current through the DAB IGBTs during turn-on and turn-off, shows that HS power losses occur in relation to snubber capacitance, IGBT turn-on and diode reverse recovery. The occurrence of diode reverse recovery loss is detailed in [12]. These losses are proportional to the full converter input/output voltage, v , as opposed to v_x which represents a proportion of these voltages. The expressions for snubber capacitance loss [12], IGBT turn-on loss [29] and diode reverse recovery loss [12] are summed and scaled by

TABLE III
PARAMETERS OF INVERTER, RECTIFIER AND DAB CONVERTER

	Definition	Value (SI Units)
f_i	Inverter Switching Frequency	2×10^4
f_r	Rectifier Switching Frequency	2×10^4
f	DAB Switching Frequency	2×10^4
L	Transformer Leakage Inductance	8×10^{-6}
C	DAB Converter Snubber Capacitance	8×10^{-9}
n_{500}	Transformer Turns Ratio with $v_{dc}^{ref} = 500$ V	2.15
n_{700}	Transformer Turns Ratio with $v_{dc}^{ref} = 700$ V	3.01
N_1	Number of Primary Transformer Turns	10
R_p	Transformer Primary Winding Resistance	1×10^{-2}
R_s	Transformer Secondary Winding Resistance	2×10^{-2}
R_m	Motor Stator Winding Resistance	0.04
R_g	Generator Stator Winding Resistance	0.08
α	Transformer Core Loss Parameter	1
β	Transformer Core Loss Parameter	2.65
K	Transformer Core Loss Parameter	0.15
l_g	Transformer Air Gap Length	1.5×10^{-3}
V_c	Transformer Core Volume	3.72×10^{-5}
μ_0	Permeability of Free Space	$4\pi \times 10^{-7}$

4 times the switching frequency (as in Eqn. 20) to yield the complete expression for HS loss:

$$P_{sw}^{HS} = 4f \left(C(v)^2 + E_{on,ref} \frac{i_L^{sw} v}{v_{ref} i_{ref}} + E_{rr,ref} \frac{v}{v_{ref}} \right) \quad (21)$$

v is equal to v_i or v_o , depending on which bridge is in question. The factor of 4 is present because in a given cycle, the occurrence of IZVS/HS implies two IGBTs will both lose energy at two of the four switching intervals. Multiplication by switching frequency gives the energy loss per second, as opposed to energy loss per cycle. Transformer core loss is also included based on the Steinmetz equation given in [12], in which the core loss per unit volume is given as:

$$P_{core}^{density} = K f^\alpha (B_{pk})^\beta \quad (22)$$

where B_{pk} is the peak swing in flux density about its average value and K and α are Steinmetz Coefficients dependent on the type of material used in the transformer core. Scaling by core volume, the transformer core loss expression can be found. The peak flux density can be approximated by considering the flux associated with an RMS equivalent sinusoidal current waveform. Where I_{enc} is the current enclosed in the length l over which flux is integrated, Ampère's Law :

$$\oint B dl = \mu_0 I_{enc} \quad (23)$$

shows the peak flux density is to be equal to the expression in parentheses in the final transformer core loss expression:

$$P_{core} = K V_c f^\alpha \left(\frac{\sqrt{2} \mu_0 N_1 i_L^{rms}}{l_g} \right)^\beta \quad (24)$$

where the definitions of the various symbols are given in Table III. As will be shown in Section IV, the effect of core loss is negligible. Using the switching loss expressions, conduction loss model, and the transformer loss equation, efficiency maps corresponding to design DC-link voltages of 500 V and 700 V are given.

Table I shows IGBT modules chosen from the Infineon range, used in both the primary and secondary bridges of

Fig. 9. Efficiency Map of DAB Converter ($v_{dc}^{ref}=700$ V). IZVS-HS and CZVS-IZVS operating region boundaries are respectively denoted by bold black and grey lines respectively.

Fig. 10. Efficiency Map of DAB Converter ($v_{dc}^{ref}=500$ V). IZVS-HS and CZVS-IZVS operating region boundaries are respectively denoted by bold black and grey lines respectively.

the DAB converter for design DC-link voltages of 400 V to 1200 V in 100 V increments. Modules are selected based on switched voltage, and are capable of conducting worst case RMS, average and peak current in all operating conditions (assuming $\max |\phi| = 90^\circ$). For each design voltage, the turns ratio is selected to be the ratio of the DC-link voltage to the open circuit battery voltage, as indicated in Table III, to achieve $d = 1$ when the DC-link voltage is equal to the design value. As shown in [34], this results in avoidance of converter HS. This will be clear from Fig. 22 in Section IV, and agrees with Table II. The efficiency maps for the 700 V and 500 V designs are shown in Figs. 9 and 10. Operating modes corresponding to each $\phi - v_o$ quadrant are labeled.

Fig. 11. Block diagram showing the CVPI control.

Fig. 12. Block diagram showing the URPI control.

Fig. 13. Block diagram showing the PZVS control.

III. CONTROL OF DC-LINK VOLTAGE

This Section details the four control schemes tested within the paper, including parameter values obtained through trial and error tuning.

A. Constant Voltage PI Control

Two CVPI schemes are tested. These schemes represent the conventional approach of DC-link voltage control. They respectively control the DC-link voltage to 700 V (as in [21]) and 500 V, following the analysis in Section II-B. Both schemes are represented by Fig. 11, and $K_p = 0.16$ and $K_i = 0.016$ are used in this work.

1) *CVPI 700 V*: The input of the existing PI control scheme is the error between a constant 700 V reference and the measured DC-link voltage. This scheme manipulates the phase shift between the gating signals of the primary and secondary bridges of the DC/DC converter, which has the effect of controlling the power (by Eqn. 9) from the DAB converter such that a DC-link voltage of 700 V is maintained in steady state, regardless of the battery voltage.

2) *CVPI 500 V*: A 500 V CVPI scheme is also tested, in order to provide a benchmark for performance of the following two variable voltage schemes which entail designing about a reference of 500 V. This voltage value was found to be a suitable one to minimize the losses in the inverter and rectifier, as explained in Section II-B.

B. Unity Ratio PI Control

Based on the premise that DAB HS does not occur when $d = 1$, a URPI control scheme is implemented with the goal of achieving unity voltage conversion ratio throughout the tested drive cycles. This scheme is shown in Fig. 12. The battery voltage rises/falls while charging/discharging. The aim of this scheme is to vary the DC-link voltage in proportion to the variation of battery voltage. As with the previous control, the output of the implemented PI controller is ϕ . However,

the input is the error between the desired and instantaneous DC-link voltage. This control is tested for a powertrain designed for 500 V DC-link voltage, therefore simultaneously attempting to achieve efficiency enhancement of the inverter and rectifier through increased modulation indices compared to the previous level of control. $K_p = 0.1$ and $K_i = 0.01$ are used in this work.

C. Persistent Zero Voltage Switching Control

In order to provide a more deterministic relationship between d and ϕ that prevents DAB converter HS operation, the proportional control law

$$\phi = k(1 - d), \quad (25)$$

also shown in Fig. 13, is implemented. In this case also, the powertrain uses converters designed for a DC-link voltage of 500 V. k is a defined constant, such that:

$$-\frac{2}{\pi} \leq \frac{1}{k} \leq \frac{2}{\pi}. \quad (26)$$

The inequality constraints given in Eqn. 26 are acquired from the expressions in Table II to restrict operation in the ZVS region. The reasoning behind the scheme is twofold. Firstly, due to the relationship between DAB power flow and DC-link voltage, operating points in the Forward-Buck and Reverse-Boost regions tend towards the point $(\phi, d) = (0, 1)$, yielding a stable controller. Operating in these two regions only (instead of all four quadrants which Eqn. 26 is valid for) further limits the value of k , which is finally given by:

$$k \geq \frac{\pi}{2}. \quad (27)$$

Secondly, the memoryless nature of the controller means that HS never occurs during operation, even very near the origin $(\phi, d) = (0, 1)$ where even a very small overshoot, likely to be experienced by other control schemes, would be detrimental for the efficiency. The superiority of this Persistent Zero Voltage Switching (PZVS) scheme will be demonstrated by comparing its performance with that of both CVPI schemes and the URPI scheme. As with the previous controllers, the manipulated variable of the controller is ϕ . In order to simultaneously avoid HS and operation close to the boundaries of the regions of interest, $k = 10$ has been chosen in this work.

IV. SIMULATION RESULTS

The two drive cycles tested in this paper are the New York City Cycle (NYCC) and the Highway Fuel Economy Test (HWFET). Regardless of the DC-link voltage, the torque-speed operating locus of the PMSM for a respective drive cycle does not change, as it is only dependent on the road load. The NYCC and HWFET loci of PMSM operation are superimposed onto the inverter efficiency maps corresponding to DC-link 700 V and 500 V designs in Figs. 14-17. Similarly, the operating loci of the PMSG are superimposed onto the 700 V and 500 V rectifier efficiency maps in Figs. 18-21. Although during operation the engine operating point is fixed, the loci show engine turn off and turn on also. Note that for both the NYCC and HWFET, for the 500 V design, both the

Fig. 14. HWFET Inverter Simulation Results ($v_{dc}^{ref} = 700$ V).

Fig. 18. HWFET Rectifier Simulation ($v_{dc}^{ref} = 700$ V).

Fig. 15. HWFET Inverter Simulation Results ($v_{dc}^{ref} = 500$ V).

Fig. 19. HWFET Rectifier Simulation ($v_{dc}^{ref} = 500$ V).

Fig. 16. NYCC Inverter Simulation Results ($v_{dc}^{ref} = 700$ V).

Fig. 20. NYCC Rectifier Simulation ($v_{dc}^{ref} = 700$ V).

Fig. 17. NYCC Inverter Simulation Results ($v_{dc}^{ref} = 500$ V).

Fig. 21. NYCC Rectifier Simulation ($v_{dc}^{ref} = 500$ V).

rectifier and inverter experience higher efficiency throughout the entire cycles compared to the 700 V case. However, operation of both converters is closer to the linear modulation constraint, implying higher modulation indices. As constant DC-link voltage is assumed in these plots, these results do not guarantee that overmodulation will not occur for the variable voltage URPI and PZVS control schemes.

Fig. 22 shows the DAB DC/DC converter operating loci

for the four control schemes studied in this work, for the NYCC. In the case of the HWFET, the types of switching (i.e. CZVS, IZVS, HS) which respectively correspond to each of the four control schemes, are the same as those which occur for the NYCC, although the extent of occurrences vary. The bold black lines are the IZVS-HS boundaries as given in Figs.

Fig. 22. DAB DC/DC converter loci on the $\phi - d$ plane for the NYCC drive cycle for four DC-link control schemes.

Fig. 23. Inverter efficiency for a window of NYCC for four control schemes.

9 and 10. The IZVS region is close to the border of the ZVS region, although the CZVS-IZVS boundaries are not shown as they change dynamically with system operating variables. From Fig. 22, it is clear that both of the CVPI schemes result in HS of the DAB converter. Although $d = 1$ is desirable, the URPI control scheme implemented in an attempt to achieve this goal also results in HS. Although the desired DC-link voltage is known at any given time by the URPI controller, it is unable to instantaneously achieve this voltage. This is due to the integral time lag which causes the operating point of the converter to cross the boundary into loss regions at low power. In contrast, the PZVS control law implemented avoids HS throughout both drive cycles, transitioning from Forward-Buck and Reverse-Boost operating modes directly through the $(\phi, d) = (0, 1)$ point. This results in high efficiency operation of the DAB converter, by varying the DC-link voltage.

Fig. 23 shows the variation of inverter efficiency for each of the control schemes, simulated on the NYCC cycle. Rectifier efficiencies are not plotted, as the rectifier is mostly inactive for the NYCC. As indicated by the previous efficiency maps, the inverter efficiency corresponding to the 700 V CVPI

Fig. 24. DC/DC converter efficiency for a window of NYCC for four control schemes.

TABLE IV
DAB LOSS BREAKDOWN, NYCC PZVS

Operating Mode	Bridge	IZVS Loss (kJ)	% Time in IZVS
Forward Buck	Input	6.42	43.89
Reverse Boost	Input	25.14	31.25
Forward Buck	Output	35.46	47.50
Reverse Boost	Output	5.64	30.97

scheme is lower than those corresponding to the other schemes designed about 500 V. However, the inverter efficiency plots all follow a similar pattern dictated by the speed and torque of the PMSM which also determine the three-phase voltage level of the inverter. As the operating point of the engine is mostly fixed when operating a series HEV with a Thermostat SCS, variation of rectifier efficiency is dictated by variation of DC-link voltage. This is the primary reason as to why within the following results, the PZVS scheme is seen to yield better results for the NYCC as opposed to the HWFET, when compared to the other schemes. For HWFET, the PZVS scheme causes high DC link voltage while the battery is recharging, whereas the engine rarely provides power during the NYCC, as the battery does not discharge fully during this drive cycle.

Similarly, Fig. 24 shows the variation of DAB converter efficiency for each of the control schemes, simulated on the NYCC drive cycle. However, a significant insight cannot be garnered from this plot. Efficiencies of 98% to 99% are achievable within the CZVS region with all schemes. However, at times, efficiency also drops substantially for each of the schemes. For the CVPI and URPI schemes this is often related to HS. For PZVS, this is due to the occurrence of IZVS when power demand is very low. The associated losses differ significantly. Therefore without simultaneously showing the power demand, analyzing this efficiency plot will provide no information relating to the associated levels of losses and hence the overall powertrain efficiency.

Tables IV, V and VI give a better understanding of the effectiveness of the PZVS, URPI and 500 V CVPI schemes, by presenting detailed DAB loss information. The percentage of the drive cycle spent within each IZVS and HS region of the $\phi - d$ plane is also provided. The following conclusions can be drawn from this information. Although HS is avoided when using the PZVS scheme, IZVS occurs on both the input

TABLE V
DAB LOSS BREAKDOWN, NYCC URPI

Operating Mode	Bridge	IZVS Loss (kJ)	HS Loss (kJ)	% Time in IZVS	% Time in HS
Forward Buck	Input	0.86	0.00	8.87	0.00
Forward Boost	Input	5.79	0.6	36.05	0.22
Reverse Buck	Input	0.00	0.00	0.01	0.00
Reverse Boost	Input	147.40	0.00	33.01	0.00
Forward Buck	Output	5.00	0.00	9.07	0.00
Forward Boost	Output	29.02	0.00	36.24	0.00
Reverse Buck	Output	0.00	0.00	0.01	0.00
Reverse Boost	Output	1.87	4.06	11.47	21.52

TABLE VI
DAB LOSS BREAKDOWN, NYCC CVPI 500 V

Operating Mode	Bridge	IZVS Loss (kJ)	HS Loss (kJ)	% Time in IZVS	% Time in HS
Forward Buck	Input	0.86	0.00	5.76	0.00
Forward Boost	Input	1.85	60.43	20.04	21.59
Reverse Buck	Input	0.09	0.37	4.08	0.13
Reverse Boost	Input	40.44	0.00	17.76	0.00
Forward Buck	Output	4.35	6.75	4.72	1.03
Forward Boost	Output	23.01	0.00	35.18	0.00
Reverse Buck	Output	1.52	0.00	5.19	0.00
Reverse Boost	Output	1.19	0.83	11.94	5.80

and output bridge of the converter for a significant percentage of the NYCC. The losses remain relatively low as this IZVS only occurs at low power, very close to the $(\phi, d) = (0, 1)$ point. As the scheme is confined to operate in two quadrants of the $\phi - d$ plane, it has only two operating modes. As Fig. 22 suggests, for the URPI scheme, the operating point strays into the HS region for boost operation. However, input bridge IZVS during Reverse-Boost operation is responsible for the majority of DAB losses. Although the input bridge spends a comparable amount of time operating within the IZVS region of the Forward-Boost quadrant, the resulting losses are lower due to the corresponding power demand of the DAB converter. Although it is clear from Fig. 22 that the 500 V CVPI scheme yields CZVS at high power, this can be said for each of the schemes. The occurrence of HS is clear at low power during boost operation. 60.43 kJ of energy is lost in the input bridge during Forward-Boost HS. Efficiency plummets during this period of operation.

Figs. 25 and 26 respectively show converter modulation indices for NYCC and HWFET, corresponding to the PZVS

Fig. 25. Inverter and rectifier modulation indices when NYCC is simulated.

Fig. 26. Inverter and rectifier modulation indices when HWFET is simulated.

Fig. 27. DC-link voltage and NYCC speed profile.

scheme. In both cases, overmodulation of the inverter never occurs over a sustained period. Transient overmodulation occurs briefly within the rectifier for both drive cycles. During transient periods (such as engine turn on), the direct axis PMSG current deviates from its desired value of 0 A. This value is quickly achieved after this transient thanks to the integral action of the PMSG PI control. However, the rectifier efficiency plots both assume constant direct axis current of 0 A. Therefore, the operating point strays into the region of overmodulation briefly during harsh transients. Due to the fact that the level of overmodulation is so slight, and occurrence rare, the loss expressions given in Eqns. 3 and 4 are assumed to be accurate.

Figs. 27 and 28 show the variation of DC-link voltage caused by the PZVS scheme for the NYCC and HWFET respectively. The drive cycles themselves are also shown. Although, the converters have been designed for a DC-link voltage of 500 V, the switches chosen can endure up to 650

Fig. 28. DC-link voltage and HWFET speed profile.

V; see Table I. As shown, this maximum voltage is never exceeded for either of the drive cycles. Fig. 28 can be used to explain the voltage variation most effectively. The battery SOC falls to 50% SOC at 300 s. The engine turns on and supplies until SOC of 80% is reached at 700 s. The open-circuit battery voltage is approximately 232 V in the typical SOC range of operation. While charging, this voltage increases and while discharging it decreases. This is reflected in the DC-link voltage. As shown, when the battery is charging in the 300 – 700 s interval, the DC-link voltage is above its design value, and while discharging the DC-link voltage drops below its design value. As d is not controlled to a constant value, the variation of battery and DC-link voltage is not a constant relationship, but rather is a function of ϕ . The same applies to the NYCC in Fig. 27, but occurs at a more sporadic rate.

Figs. 29 and 30 compare the power electronic energy loss data of each scheme, for the NYCC and HWFET respectively. Rectifier losses for the NYCC are almost zero as the engine very rarely supplies during the cycle. This is not the case for the HWFET. It is clear that power loss is dominated by switching for the inverter and rectifier. By comparing the switching losses of the inverter and rectifier (for HWFET) respectively for the 700 V CVPI scheme and the 500 V CVPI scheme, the benefit of designing about a lower voltage is clear. As the DC-link voltage varies with the movement of the DAB operating point, for the variable voltage schemes the rectifier and inverter switching losses are dependent on the proportion of battery charging and discharging associated with the control schemes. As the battery discharges during the NYCC, DC-link voltage is lower than the design value of 500 V and hence inverter efficiency is higher compared to the 500 V CVPI simulation. However, the battery charges from 300s to 700s of the HWFET, and therefore the DC-link voltage is above 500 V and as a result, the inverter is less efficiency for the PZVS scheme as for 500 V CVPI control. This is not surprising, as the primary design goal of the PZVS scheme was to lower DAB switching losses. Although the PZVS scheme achieves this goal for both drive cycles, the energy saving is offset by an increase in rectifier switching losses in the case of the HWFET. For a given engine power rating, this indicates that the energy savings offered by the PZVS scheme increase with the energy storage capability of the HEV battery. By comparing the DAB switching losses corresponding to URPI and PZVS, the superiority of the latter scheme can be quantified. For both drive cycles, the URPI scheme does not deliver the expected powertrain energy savings due to excessive IZVS on the input bridge during reverse boost operation. Core loss is clearly negligible. Also, conduction losses are relatively low in all converters. This loss is related to the level of current flowing through each of the converters in order to satisfy the PMSM load which is dependent on the drive cycle. For the NYCC, the order of performance of the schemes (in order of increasing total powertrain energy loss) is as follows: PZVS, 500 V PI, URPI, 700 V PI. For the NYCC, the percentage energy saving achieved by the PZVS scheme compared with the original 700 V PI scheme is 52.5%. For the HWFET, the overall performance of the PZVS, URPI and 500 V CVPI schemes are roughly equivalent.

Fig. 29. Converter energy loss data by implemented control for the NYCC.

Fig. 30. Converter energy loss data by implemented control for the HWFET.

V. CONCLUSION

The DC-link voltage in a series HEV imposes a constraint in relation to the feasible operating region of the PMSM and PMSG, corresponding to unity modulation index. However, it also offers possibilities for improving the overall efficiency of the power electronics in the powertrain. The efficiency of the inverter and rectifier are maximized by increasing modulation indices as close as possible to unity (corresponding to a drop in DC-link voltage). The losses of the inverter and rectifier are dominated by switching. Therefore, efficiency is increased by dropping the switching voltage. Switching losses of the DAB converter vary between CZVS (zero loss), IZVS (low loss) and HS (high loss). DAB conduction losses are reduced by keeping power flow (and therefore ϕ) low. Switching losses are dependent on the relationship between the voltage conversion ratio and ϕ . By manipulating ϕ , the DC-link voltage can be controlled in such a way that maximizes the occurrence of lossless switching within the DAB converter. This PZVS control scheme is the most desirable scheme tested for the NYCC as power electronic loss was minimized, compared

to the other control schemes tested. Such benefits were not found for the HWFET simulation, due to the higher rectifier switching losses which occur during battery recharging. The results indicate that the PZVS control is most useful for vehicles with high hybridization factor, for application in an urban environment.

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