ULTRACAPACITOR/BATTERY HYBRID ENERGY STORAGE SYSTEMS FOR ELECTRIC VEHICLES

by

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Graduate Department of Electrical and Computer Engineering
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Abstract

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This thesis deals with the design of Hybrid Energy Storage System (HESS) for Light Electric Vehicles (LEV) and EVs. More specifically, a tri-mode high-efficiency non-isolated half-bridge converter is developed for the LEV based HESS applications. A 2 kW, 100 V interleaved two-phase converter prototype was implemented. The peak efficiency of 97.5% and a minimum efficiency of 88% over the full load range are achieved.

Furthermore, a power-mix optimizer utilizing the real-time Global Positioning System (GPS) data for the EV based HESS is proposed. For a specific design, it is shown that at the cost of less than 1.5% of the overall energy savings, the proposed scheme reduces the peak battery charge and discharge rates by 76% and 47%, respectively. A 30 kW bi-directional dc-dc converter is also designed and implemented for future deployment of the designed HESS into a prototype EV, known as A2B.
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Chapter 1

Introduction

1.1 Modern Electric Vehicles

During the last few decades, the negative environmental impacts of petroleum based vehicles have ignited the interest toward Electric Vehicles (EV) with low or zero emissions. An Electric Vehicle uses one or more electric motors for propulsion. There are three types of EVs [1] as follows,

1. Continuously powered from an external electric power station,

2. Powered by the electricity stored originally from an external electric power source,

3. Powered by an on-board electrical generator, such as an internal combustion engine (ICE). This vehicle category is called Hybrid Electric Vehicle (HEV). If the electric source of the HEV is also accessible from outside to recharge, the HEV is known as Plug-in HEV (PHEV). HEV can have a parallel design, series design or a combination of both with series configuration at low speeds and parallel configuration for highways and accelerations [2]:

- In a parallel design, the vehicle is powered by two independent mechanical drives, as shown in Fig. 1.1(a) [3]. The electric motor only assists the primary
engine in the events of acceleration, hill climbing, braking, etc [2].

- In a series design, the vehicle is powered only by one mechanical drive as shown in Fig. 1.1(b) [3] and the primary engine charges the battery, which drives the electric motor [2].

![Diagram of series hybrid electric vehicle (HEV) architecture.](image)

**Figure 1.1**: (a) Parallel, (b) series hybrid electric vehicle (HEV) architecture.

EVs can also be categorized based on their power rating, since their power rating can range from hundreds of watts in Light Electric Vehicles (LEV) to several hundreds of kilowatts in heavy-duty vehicles. The power rating of LEVs, such as e-bikes, scooters, segways™ and lawnmowers are less than several kilowatts. The next power level includes racing motorcycles [4] and sedan vehicles with a power rating in the range of 10-100 kW. The highest power EV category, with power rating over 100 kW, includes heavy-duty vehicles, such as buses and trucks.

EVs came into existence in the mid-19th century. Shortly after, they were driven to the side by the ICE based vehicles due to several reasons, such as level of comfort and
the energy-density of their storage element. Electric power remained a common source for vehicles like trains that are directly powered from the grid and do not require an energy storage. During the last few decades, oil price forecasts and environmental impacts renewed the interest toward electric transportation infrastructure [5]. The global oil price forecast shows a significant increase by 2035, according to the 2011 Annual Energy Outlook published by the Energy Information Administration (EIA) [6], as shown in Fig. 1.2. EVs can have a minimal environmental impact, since the energy consumed by the EVs can be generated from many sources, including clean renewable energy sources. Another key advantage to EVs is their ability to recover energy, that otherwise would be lost during braking, which is known as Regenerative (Regen) braking.

$\begin{figure}
\centering
\includegraphics[width=\textwidth]{figure12.png}
\caption{EIA forecasts of global oil price based on the 2011 Annual Energy Outlook [6].}
\end{figure}$

As of the date of this writing, the Toyota Prius, the first mass-produced hybrid gasoline-electric car, is the world’s best selling HEV. The Nissan Leaf is the world’s top selling highway capable EV and the Chevrolet Volt family is the best selling PHEV. According to Pike Research [7], HEVs and PHEVs will represent 3.1% and 5.1% of the
total worldwide and U.S. auto sales by 2017, respectively [8]. EVs, which are the main focus of this thesis, face many challenges in dominating the market. These challenges will be discussed in the following section and some of them will be addressed throughout this thesis.

1.1.1 Challenges in SESS EVs

One major challenge in the mass adoption of EVs and HEVs is their Energy Storage System (ESS) [9,10]. Classical electrochemical batteries have several limitations:

- Lower energy-density compared to petrol-based fuel, which affects the weight and the range of the vehicles. The specific-energy of petrol and Lithium-Ion (Li-Ion) batteries (the dominant battery chemistry in modern EVs) are 13 kWh/kg and 0.2 kWh/kg, respectively. Considering a typical energy efficiency of 20 % for the ICES and 90% for the electrical drives, the effective specific-energy of petrol is 14 times higher than that of Lithium-Ion batteries, resulting in a very limited range for EVs. The all-electric range of typical EVs is in the order of 100 km, based on the United States Environmental Protection Agency (EPA) [11]. For example, the all-electric range of the Chevy Volt and the Nissan Leaf is 56 km and 117 km, respectively.

- The charge time is another prohibitive factor [12] and charging time varies with the charging levels [13]. For instance, using an on-board 3.3 kW charger, it takes 7 hours to fully charge the Nissan Leaf’s 24 kWh battery pack.

- The battery lifetime is another major concern and is usually not sufficient for traction applications [12].

Numerous battery chemistries are available [14] and the best choice of battery is application dependant. A brief summary of available battery chemistries is provided below:
1. Lead-Acid Batteries:

Using Lead as the negative, Lead Oxide as the positive electrode and diluted Sulfuric Acid as the electrolyte, Lead-Acid batteries provide many advantages for HEV applications [14]; comparatively low cost, mature technology, and being mass produced are the advantages, while being limited to only 20% of the rated capacity in its Depth of Discharge (DOD), limited life cycle if operated with low State of Charge (SOC) and low specific-energy and power due to heavy Lead collectors [15, 16] are the main disadvantages of this battery chemistry.

2. Nickel-Metal Hydride (NiMH) Batteries:

Using an Alkaline solution as the electrolyte, NiMH batteries consist of Nickel Hydroxide as the positive and an engineered alloy of Vanadium, Titanium, Nickel, and other metals as the negative electrode [14]. With twice the specific-energy compared to the Lead-Acid batteries, the NiMH batteries are environmental friendly and are recyclable [17]. The NiMH battery is safe for high voltage operation and provides several advantages, such as long cycle life, wide operating temperature range and it is over charge and discharge resistance [18]. Repeatedly discharging at high load currents reduces the NiMH lifetime to 200-300 cycles. The DOD should be limited to 20-50% of the rated capacity, to ensure the best battery performance [19]. Memory effect in the NiMH batteries limits the usable power [14].

3. Lithium-Ion Batteries:

Li-Ion batteries provide excellent performance in portable applications [20]. The Positive electrode varies from Cobalt Oxide, Manganese Oxide, Iron Phosphate, Nickel Manganese Cobalt Oxide and Nickel Cobalt Aluminum Oxide [21]. Lithium salt in an organic solvent is used as the electrolyte and Carbon material as the negative electrode [14]. High specific-energy of over 100 Wh/kg, recyclability and long cycle life of several thousands cycles are the most important advantages of this
chemistry [22]. Fig. 1.3 shows a comparison of the most important characteristics of various Li-Ion based batteries. Lithium Nickel Manganese Cobalt Oxide technology is the preferred candidate for EVs [21], as it provides an excellent trade-off between these important aspects, as shown in Fig. 1.3.

![Figure 1.3: Characteristic comparison of different Lithium-ion based batteries.](image)

4. Nickel-Zinc (Ni-Zn) Batteries:

These batteries provide high specific-energy and power with low-cost materials with deep cycling capabilities [14]. They are environmentally friendly but suffer from poor lifetime due to the fast growth of dendrites [23].

5. Nickel-Cadmium (Ni-Cd) Batteries:

Ni-Cd batteries have a specific-energy of about 55 Wh/kg. Long lifetime and full discharge capability without any damage are the advantages coming at the cost of its higher price compared to other chemistries [14, 24, 25].

A comparison of the specific-energy of different battery chemistries is provided in Table 1.1 [21]. Although Lithium Cobalt has the highest specific-energy, Lithium Man-
ganese Oxide and Lithium Iron Phosphate are superior, based on power and thermal stability [21].

| Table 1.1: specific-Energy for Different Battery Chemistry |
|-----------------|----------------|----------------|-----------------|----------------|----------------|-----------------|
|                 | Lead-Acid | Ni-Cd | Ni-MH Phosphate | Li-MN Oxide | Li-Ni-Mn Co Oxide | Li-Co Oxide |
| Typical Specific Energy (Wh/kg) | 40 | 60 | 90 | 110 | 120 | 140 | 170 |

The power demanded by an EV is variable during a drive-cycle, with the peak power periods happening for short times during acceleration and braking. The ratio of the peak power to average power can be well over 10:1 [26]. Based on the Ragone plot shown in Fig. 1.4 [27], there is a trade-off between the specific-power and specific-energy within the available technology [28]. Furthermore, it is possible to optimize a given battery chemistry for better specific-energy or for better specific-power [29]. This trade-off was also shown in Fig. 1.3. The extracted energy from a Li-Ion battery is a function of the discharge current and the energy efficiency drops for higher currents [30,31].

Based on [26], two thirds of the total energy in an urban drive-cycle is processed in acceleration and deceleration transients. Therefore, the recovered energy from braking can substantially extend the vehicle range. However, with a lower limit on the charge current of the batteries compared to their discharge current, Regen energy recovery is limited [29].

The high cost of the batteries increases the importance of the battery life in the customer acceptance of the EVs in the market. High charge/discharge current rates shorten the battery life, including high-current Li-Ion batteries [29]. Furthermore, pulsed currents with the same average current result in increased cell temperature [32]. Lifetime reduction of Cobalt based Li-Ion cells cycled at high charge/discharge current is analyzed in [33]. In another study, the capacity fade of Li-Ion Sony US 18650 (1.4 Ah) after 300
cycles, discharging at 1C, 2C and 3C with a charging current of 1 A in all cases, were reported to be 9.5%, 13.2% and 16.9% respectively [34], where:

\[ 1C (A) = \frac{Battery \ capacity \ (Ah)}{1 \ hr}. \]  

(1.1)

The energy efficiency of the batteries depends on their Equivalent Series Resistance (ESR) that varies over time, based on the operating conditions. The ESR of the Sony US 18650 cycled at higher C rate after 300 cycles was found to be higher than the one cycled at lower C rates [34]. For example the cell cycled at 3C and 2C were found to have 14% and 5% higher ESR compared to the cell cycled at 1C. The cell cycled at 1C after 300 cycle found to have 12% higher ESR compared to its initial conditions.

Temperature dependant performance of the batteries is another challenge in the EVs, especially in environments with extreme weather conditions, such as Canada. The available power of the batteries reduces at lower temperature and causes a weak user experience during winter season.
With this limitations in mind, one can improve the ESS by combining different energy storage devices into the storage system. This motivates us to take a look at the hybrid energy storage systems (HESS) in the next section.

1.2 Hybrid Energy Storage System for EVs

Implementing low-cost, high-density and power-efficient energy storage system is one of the key technological hurdles for enabling future mass-adoption of EVs. Different storage technologies, such as fuel cell (FC), flywheel and ultracapacitor (u-cap), can be combined with the batteries to form a HESS for an EV. Fuel cell generates electricity from the reaction of the fuel and oxidant in an electrolyte. FC has several advantages including high conversion efficiency, zero or very low emission, quiet operation, fuel flexibility, durability and reliability. The main drawbacks of FCs are high cost and longer response time compared to batteries and u-caps [14].

A flywheel stores mechanical energy that can be transformed into electrical energy through an electrical machine with bi-directional operation capability. Flywheels have symmetric input and output specific-power in the range of 1-12 kW/kg [28,35]. Flywheels consist of environmental friendly materials and high number of charge/discharge cycles, high DOD capability, simple and reliable SOC monitoring are the main advantages of such storage systems [35,36]. However, flywheels are not suitable for power levels below 100 kW due to their higher capital cost compared to other storage systems, especially u-caps [28].

Electrochemical double-layer capacitors (EDLC), also known as u-caps, have been developed as a complimentary energy storage technology to batteries [37]. Today’s u-caps have symmetric input and output specific-power in the range of 0.5-25 kW/kg [38], which is at least one order of magnitude higher than typical Li-Ion based batteries [39]. The high specific-power and low ESR of modern u-caps allow their integration into smart,
electronically controlled HESS for EVs.

Effectively combining the high specific-power of u-caps and the high specific-energy of Li-Ion batteries is a major challenge. In conventional EVs, the excess Regen power that cannot be safely absorbed by the battery must be dissipated in the mechanical brakes. The most important goals of the u-cap/battery HESS are:

- Maximize the re-use of Regen energy,

- Reduce the wear on the mechanical brakes and

- Minimize the strain on the battery during rapid acceleration/deccelerations.

The HESS concept has been previously studied through system-level simulations [40–44] with reported driving-range improvements of up to 46 % [41].

With two energy sources inside the vehicle, it is possible to control the power-mix between the two sources depending on the HESS configuration and source interfaces. The power optimizer defines each source’s contribution to the load, as shown in Fig. 1.5. This is defined based on different parameters of the sources, such as their SOC and their charge/discharge current limits. The HESS improvements compared to SESS, depend intrinsically on how the sources are combined to exploit the strengths and avoid the weaknesses of each source [29].

![U-cap based HESS architecture for EVs.](image)
1.2.1 Detailed Comparison of Lithium-Ion Batteries vs. Ultracapacitors

In this section we describe battery and u-cap important characteristics and provide a comparison between the two sources based on these aspects. The capacitance and ESR versus mass for a wide range of commercially available 2.5 V and 2.7 V u-cap cells are shown in Fig. 1.6 and 1.7, respectively. The basic trend between the u-cap mass, $M_{uc}$, and the ESR, $R_{ucs}$, versus the capacitance, $C_{uc}$, for the 2.7 V u-cap technology can be approximated by (1.2) and (1.3), respectively.

\[
M_{uc} = 0.5873 \ C_{uc}^{0.8491} \quad (1.2)
\]
\[
R_{ucs} = 0.4667 \ C_{uc}^{-0.914} \quad (1.3)
\]

Figure 1.6: Mass versus capacitance for commercially available (a) 2.5 V and (b) 2.7 V u-caps.

The specific-energy and specific-power for the same u-caps as Figs. 1.6 and 1.7 are shown in Fig. 1.8. The best 2.7 V u-caps have a superior specific-power and specific-energy compared to the older 2.5 V technology. The data for both u-caps and batteries in this section is presented for a single packaged cell from which high-capacity modules
can be constructed. The addition of monitoring, protection and balancing circuits in these modules can increase the mass and the ESR by up to 50%. Based on Fig. 1.8, the best available u-cap technology as of this writing offers up to 5.95 Wh/kg [45] and 23.6 kW/kg [46] in separate devices. The specific-power is calculated based on the maximum extractable power when the u-cap is fully charged and is given by

\[
P_{ucc,max} = \frac{V_{uci}^2}{4 R_{ucs}},
\]

where \( V_{uci} \) is the internal capacitor voltage. This maximum extractable power, which is limited by \( R_{ucs} \), is seldom used in practice due to the high power losses. If we consider a short time interval such that \( V_{uci} \) remains constant then the u-cap efficiency, \( \eta_{uc} \), can be calculated as a function of the normalized extracted power, \( P_{uc0} = P_{uc}/P_{ucc,max} \),

\[
\eta_{uc} = 1 + \sqrt{1 - \frac{P_{uc0}}{2}}.
\]

It is interesting to note that \( \eta_{uc} \) is independent of the ESR when plotted against the normalized extracted power, as shown in Fig. 1.9. For example, the extracted power must be limited to 37% of \( P_{ucc,max} \) in order to achieve \( \eta_{uc} > 90 \% \). Operating at the maximum power, \( P_{uc} = P_{ucc,max} \) gives \( \eta_{uc} = 50 \% \).
Figure 1.8: Specific-power versus specific-energy for commercially available u-caps and Lithium-based rechargeable batteries.

Figure 1.9: U-cap efficiency versus normalized extracted power.

The specific-power and energy of commercially available 3.6 V Lithium based battery cells are also included in Fig. 1.8. Unfortunately the ESR of cells is seldom provided in datasheets. Furthermore, unlike u-caps, the maximum charge and discharge powers...
are not equal and are significantly lower than the predicted values when one considers the battery ESR, as in (1.4). It is apparent from Fig. 1.8 that there is a trade-off between batteries having a high specific-power and specific-energy, depending on the cell chemistry as was also shown in Fig. 1.3 for Li-Ion based batteries. Today’s high-power batteries have a discharge specific-power approaching that of the best u-caps. At first glance, it would appear from Fig. 1.8 that the benefit of introducing u-caps into a HESS is mainly to achieve a more symmetric charge/discharge power capability. In fact, the design considerations are considerably more subtle:

1. The long-term health of the battery is heavily dependent on the cumulative number of charge/discharge cycles, as well as the power profile. Even if the battery can handle the peak power requirements in the application, off-loading the power transients to the u-cap extends the battery lifetime [34].

2. The battery data in Fig. 1.8 is provided at nominal conditions. In practice, the energy capacity and ESR are strong functions of temperature [47]. The benefits of a HESS may therefore become much more attractive when considering the whole automotive temperature range. To further complicate the design, the maximum power capability of the u-cap depends on the state-of-charge (SOC).

3. Fig. 1.8 does not include economic considerations, which are rapidly evolving. In general, the cost of using a high-power Li-Ion battery technology has to be weighed against using a u-cap based HESS with a low-cost, high energy-density battery.

Today’s ultracapacitors offer improved cycling lifetime and withstand 500,000 to 1,000,000 cycles with less than 30% capacitance degradation and 100% ESR change [38, 45, 46], which is orders of magnitudes higher than the typical Li-Ion batteries with hundreds to thousands of cycles for 20% capacity fade [39, 48, 49]. U-caps do not suffer as severely from high DOD effects as Li-Ion batteries [50, 51]. [51] shows that by increasing
the DOD from 30% to 80%, a Li-Ion battery life is reduced from 2600 cycles to 1000 cycles.

Finally, u-caps can operate under a wider temperature range than batteries and when used together, u-caps can compensate for the reduction in the available power from the batteries in extreme weather conditions [29].

1.2.2 Possible Ultracapacitor/Battery HESS Configurations and Control Strategies

A HESS with two energy storage elements and one load has seven possible configurations, as shown in Fig. 1.10. The motor-drive load and all dc-dc converters have bi-directional power capability. Choosing the most appropriate configuration depends on the application and power level. The power optimizer control strategy is a strong function of the HESS configuration and therefore they are considered with their corresponding topologies in this thesis.

The simplest configuration choice is shown in Fig. 1.10(a), where the u-cap, battery and load are directly connected in parallel. The simplicity of this system is attractive, however it has several limitations:

1. The power sharing between the battery and u-cap is uncontrolled and dictated only by the parasitic elements [52,53], which are not well known and vary with aging.

2. The energy capacity of the u-cap is under-utilized, since the voltage is restricted to the narrow range of the battery, which typically ranges from 2.8 V to 4.2 V per cell.

3. There is no flexibility in the choice of nominal battery and u-cap voltage.

4. The bus voltage, $V_{bus}$, is un-regulated and varies depending on the battery voltage range, which impacts the inverter design in EV applications.
Figure 1.10: Seven possible configurations for a two-source, single-load HESS.

In Fig. 1.10(b) a dc-dc converter rated to handle the total load power regulates $V_{bus}$. However this configuration suffers the same limitations as the configuration in Fig. 1.10(a) except that it provides a regulated $V_{bus}$. The added dc-dc converter must handle the total load power, which increases the cost of this architecture.

The configurations shown in Fig. 1.10 (c)-(g) have one or more controlled power sources. More importantly, the battery and the u-cap can operate at different voltages, which allows the specific-power and specific-energy to be optimized using the best available technology.
The power sharing between the u-cap and the battery in the topology shown in Fig. 1.10(c) is well studied in the literature [3, 40, 41, 44, 54–58]. This topology offers a narrow voltage range on the bus and therefore a better inverter operating efficiency compared to Fig. 1.10(a), (d). In [40], a model predictive control of the HESS configuration shown in Fig. 1.10(c) was demonstrated to reduce the discharge intensity of the battery in a HEV to maximize its lifetime. In another study based on Fig. 1.10(c), the main control objectives for HESS were stated as [54]:

1. Setting the power mix based on the optimal operating point of each source.

2. Minimizing the losses of the overall system.

3. Optimizing the u-cap SOC.

The HESS improves the driving performance, reduces the energy consumption of the HEVs and improves the battery operating conditions based on the results published in [55]. Utilizing the same HESS configuration for a LEV, [56] states the main controller objectives to be the u-cap SOC control, while smoothing the battery current for optimal battery performance. [41] uses a fuzzy sliding mode controller to recover more Regen energy with the HESS assuming a battery charging current limit. A filter based control approach is adopted by [57] and augmented with a non-linear u-cap voltage controller as:

\[
\tilde{I} = G(s) \cdot I_{load}, \tag{1.6}
\]

\[
\hat{I}_{uc}(\tilde{I}, V_{uc}) = \tilde{I} \cdot f(V_{uc}, \tilde{I}) + a \cdot (V_{uc} - V_{mid}), \tag{1.7}
\]

where \(G(s)\) is a high-pass filter, \(f(V_{uc}, \tilde{I})\) is the u-cap voltage control factor and \(a \cdot (V_{uc} - V_{mid})\) is a linear additive correction term. The \(f(V_{uc}, \tilde{I})\) attenuates the charging currents at higher u-cap SOCs and attenuates the discharge currents at lower u-cap SOCs. With the same configuration, [58] shows that with an optimization method using a trained Neural Network based on results obtained from simulations of different drive-cycles, a 3.3% improvement is achieved over the SOC control strategies in terms of km/kWh for a
Chevrolet light utility vehicle. It also shows that with a primary source unable to receive Regen energy, the improvement would be as high as 28.7% in terms of km/kWh. In another study, [3] designed and implemented an ultracapacitor based HESS for an EV with a 54 kW drivetrain. Two control strategies based on heuristics and optimization model using neural networks are studied and the results were incorporated to an economic evaluation of the system. Results shows that a battery life extension of 50% is required to compensate for the HESS costs. Similar analysis were performed for a hypothetical fuel-cell based hybrid vehicle showing that the u-cap based HESS is the most cost effective solution.

In Fig. 1.10(d), the u-cap on \( V_{bus} \) provides the opportunity to efficiently capture the Regen energy [59,60]. However, the u-cap energy is under-utilized and the higher u-cap rated voltage introduces weight and balancing challenges. Hysteretic control schemes were applied to control \( V_{uc} \) in [59] for this configuration. It is shown that despite the improvement of battery current profile, it can experience high discharge currents if \( V_{uc} \) drops below \( V_{bt} \). The specific-power improvement of this configuration has also been studied for pulsed power applications in [60]. It is shown that power capability of the configuration shown in Fig. 1.10(d) can be nearly 3 times greater than that of Fig. 1.10(a) for a pulsating current at a rate of 0.2 Hz at a 10% duty ratio.

The configuration shown in Fig. 1.10(e) has flexible operating voltages and provides explicit power sharing control at the expense of adding two dc-dc converters [12,29,37,61,62]. A comparison of different control strategies such as:

1. Source resistance strategy that tries to minimize the losses when \( I_{load} > 0 \) and redirect all the energy to the u-cap when \( I_{load} < 0 \).

2. Vehicle acceleration strategy where the u-cap provides all the acceleration power while the battery provide the power to overcome global friction forces.

3. Filtration strategy by servicing the low-frequency component of the load power
from the battery.

4. Variable saturation current strategy to limit the battery charge/discharge current to predefined values based on electrical consumption.

is presented in [12], where sizing and expected life-time of the batteries were considered. For the same configuration, [29] uses the u-cap during transients to provide a smooth battery current profile and then restores the u-cap SOC to a desired level afterward. However, it does not describe the HESS system under Regen conditions. The configuration is also used in [62] to control the power-flow and avoid over-stressing the battery and the fuel-cell for a system containing a fuel-cell, battery and u-cap, in a hybrid vehicle. A self-optimizing, multi-objective optimization technique is proposed in [61] used to control a NiMH based HESS. The algorithm defines the importance of the two objectives: minimization of energy loss and maximization of the power reserve (i.e. the deviation of the u-cap SOC from its midpoint). The authors consider an application of a RailCab, containing predefined sets of short range load profiles (approx. 2 min), associated with recognized turns and paths. In our prior work [37], we proposed a predictive power optimization algorithm to control the power-mix in an HESS configured as Fig. 1.10(e) for a LEV where the entire drive-cycle is considered unknown, using a state-based approach, organized as a probability-weighted markov process to predict future load demands. A real-time global optimizer is then used to control the appropriate power mix based on the predictions and probabilities of state trajectories along with their associated system losses.

The configurations shown in Fig. 1.10(f) and (g) offer the same control flexibility as Fig. 1.10(d) and (c) but the extra dc-dc converter, dc-dc 2, must handle the total load power, increasing the cost. The rated and peak power capability of dc-dc converters in Fig. 1.10(c)-(e) can be lower than those of the load due to power sharing.

The salient features of the seven HESS configurations are summarized in Table 1.2. The topology shown in Fig. 1.10(e) is a good candidate for LEV based HESS [37], as
considered in Chapter 2, since it offers the most flexibility in operating voltages and provides effective control over the battery current, while minimizing the number of conversion stages from battery to load. The configuration in Fig. 1.10(c) is considered in Chapter 3 for high-power EV applications, where the additional cost of the second dc-dc converter and the energy losses due to the conversion stage from battery to load is prohibitive.

Table 1.2: Comparison of HESS Architectures

<table>
<thead>
<tr>
<th></th>
<th>Fig. 1.10(a)</th>
<th>Fig. 1.10(b)</th>
<th>Fig. 1.10(c)</th>
<th>Fig. 1.10(d)</th>
<th>Fig. 1.10(e)</th>
<th>Fig. 1.10(f)</th>
<th>Fig. 1.10(g)</th>
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<tr>
<td>DC-DC Converters</td>
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<td>Controlled $I_{uc}$</td>
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<tr>
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<td>1</td>
<td>2</td>
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1.3 Thesis Motivation and Objectives

The goal of this work is to investigate and further improve the benefits of the ultracapacitor/battery based HESS in both LEVs and EVs. More specifically, we are addressing an efficiency optimization of the dc-dc converter for a LEV based HESS. Furthermore, we are investigating a HESS design for a prototype EV to evaluate the benefits of utilizing Real-Time (RT) Global Positioning System (GPS) data processing in the power-mix optimizer.

The first objective is to design a high-efficiency bi-directional dc-dc converter to interface the battery to the bus in the HESS architecture of a LEV, as shown in Fig. 1.10(e). The converter must meet the following requirements, as described in Chapter 2:

- Simple architecture with low cost, mass and volume,
- Digitally controlled,
- High efficiency over a wide operating range,
- Limited bandwidth, since the current commands are updated at a slow rate in the range of 1-10 Hz.

Although some research efforts were directed toward utilizing the information of the drive-cycle in the power-mix decisions, the possible benefits of real-time GPS data processing are not addressed in the literature. Therefore, the second objective of this thesis is to:

- Design a HESS for a prototype EV,
- Design a relatively simple real-time GPS data processing scheme to obtain the position of stop signs and traffic signals for a given current position of the vehicle,
- Design a power-mix optimizer with minimal computational requirements that utilizes the GPS data information,
• Simulate the system level architecture to evaluate the benefits of the approach,

• Finally, implement the designed HESS and integrate it into a prototype EV.

Complex computational processings should be avoided, since they need to be done in real-time, on a reasonably sized target processor based on the requirements of vehicular application. Assuming the effectiveness of the approach, the minimal incremental cost of the GPS data utilization will further justify the utilization of this scheme in HESS architectures.

This thesis is organized as follows, Chapter 2 introduces a high efficiency bi-directional dc-dc converter design for a LEV based HESS. A HESS design and a novel control strategy based on real-time GPS processing for a Canadian made prototype EV are then presented in Chapter 3. Furthermore, a 30kW bi-directional dc-dc converter is implemented for the proposed HESS and experimental results of this converter are also provided in Chapter 3. Finally, the conclusions and future work are discussed in Chapter 4.
References


REFERENCES


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Chapter 2

A DC-DC Converter Design for LEV Based HESS

2.1 Introduction

The main focus of this chapter is efficiency optimization in high-frequency non-isolated dc-dc converters. The target application is HESS for LEVs, as shown in Fig. 2.1. As Li-Ion batteries become more affordable, there is a growing number of LEV applications, including electric bicycles, scooters, motorcycles, small utility vehicles and personal mobility devices. The HESS contains both a Li-Ion battery pack for high-energy density and a u-cap module for high power-density [1]. The two energy sources are interfaced to the bus voltage, $V_{bus}$, using two bi-directional, non-isolated dc-dc converters. A current-command, $I_{bt*}$, is periodically generated by the system power optimizer with an update rate of 1-10 Hz in the targeted LEV application, based on the convergence time of the optimization algorithm presented in [2]. The current-command, $I_{bt*}$, is calculated based on the dynamic load requirements, and the battery/u-cap SOC. The u-cap converter directly regulates $V_{bus}$ using a standard voltage control loop. The system-level power optimizer is used to off-load the large load-transients from the battery to the u-cap, which has a
lower ESR and a higher cycle-life. The detailed operation of the power optimizer [1–4] is described in Chapter 3 and it is beyond the scope of this chapter, which deals exclusively with the bi-directional dc-dc converter that interfaces the battery to $V_{bus}$, as shown in Fig. 2.2. The converter operates with digital Average Current-Mode Control (ACMC). The capacitance $C_x$ represents the lumped parasitic capacitance at $v_x$, due to the power transistors. While several isolated and non-isolated topologies have been proposed for high-power EV applications [5–7], the half-bridge topology was chosen due to its low cost and high efficiency, especially when operated with soft-switching.

Figure 2.1: Architecture of a hybrid energy-storage system.

Operating the converter in Fig. 2.2 with a negative inductor valley-current, $I_v$, allows Zero-Voltage-Switching (ZVS) turn-on to be achieved in both switches, due to the resonant transitions of $v_x(t)$ during the dead-times $t_{d1}$ and $t_{d2}$, as shown in Fig. 2.3. Optimizing the dead-times, and hence eliminating the turn-on losses, for a range of load currents and battery voltages is essential for achieving the highest possible efficiency. The
main challenge in dead-time optimization is the precise timing of gate-drive signals to achieve ZVS in the presence of load changes, uncertainty in the parasitic components and gate-driver delays [8]. Dead-time optimization in high-frequency, non-isolated converters has been extensively studied, with various analog and digital on-line calibration schemes for fixed-frequency operation. Analog dead-time calibration based on a delay-locked-loop (DLL) has been carried out in [9–11]. A digital sensorless dead-time optimization has been introduced in [12], which has a simple implementation, but limited reaction speed during load transients. Another digital adaptive dead-time control scheme based on maximum efficiency tracking is introduced in [13]. A digital single-step dead-time correction was reported in [14], while an observer based method was used in [15]. More recently, digital dead-time optimization for a multi-phase dc-dc converter based on the phase current sharing was demonstrated [16]. All these works focus on dead-time control based on fixed-frequency operation. In contrast, this chapter describes a multi-mode scheme with variable frequency operation in the mid-load range, to improve the efficiency and simplify the dead-time control.

Consider the practical constraints on $i_L(t)$, as shown in Fig. 2.4. The second quadrant operation, where $I_L < 0$, is not shown for clarity. The converter must meet the constraint given by

$$i_L(t) < I_{L,sat} - \Delta I_m$$

(2.1)
Figure 2.3: Ideal waveforms for ZVS operation. The optimal $t_{d1}$ and $t_{d2}$ depend on $I_v$ and $I_P$ respectively.

under all conditions, where $I_{L,sat}$ is the inductor saturation current and $\Delta I_m$ is the safety margin required for dynamic excursions and sensing errors. In this HESS application, $\Delta I_m$ can be minimized, since the battery current-command, $I_{bt^*}$, is tightly controlled and is only periodically updated by the system power optimizer. Moreover, large dynamic load currents are off-loaded to dc-dc converter connected to the u-cap. This also helps to alleviate the potential Electro-Magnetic Interference (EMI) issues associated with operating the battery-connected dc-dc converter at variable frequency during those transients.

This chapter is organized as follows: in Section 2.2, the three operating modes are described and justified. A method to extend the dead-time over a limited portion of
the load curve is discussed in Section 2.3. Finally, the experimental results for a 2 kW prototype converter are reported in Section 2.4.

## 2.2 Mode Description and Control Strategy

Operating with \( I_v < 0 \) is required to achieve ZVS turn-on in both switches, which is referred to as dual-ZVS. Achieving dual-ZVS turn-on is unpractical at the full-load condition due to the large peak current, \( I_p \), which requires an inductor with a very high saturation current, leading to prohibitive cost. Operating with a large inductor current ripple, \( \Delta I_L \), in the full-load condition increases the conduction losses, \( P_{\text{cond}} \), given by

\[
P_{\text{cond}} = \left( I_L^2 + \frac{\Delta I_L^2}{3} \right) \left( R_L + R_{\text{on}} \right) + I_L^2 R_{\text{bt}} + \frac{\Delta I_L^2}{3} R_{C_{\text{in}}} \\
+ \left( \left( I_L - I_{\text{Load}} \right)^2 + \frac{\Delta I_L^2}{3} \right) D' + D I_{\text{Load}}^2 \right) R_{C_{\text{out}}},
\]

(2.2)

assuming \( R_{C_{\text{in}}} \ll R_{\text{bt}} \), where \( D \) is the steady-state duty-cycle of \( Q_1 \), \( D' = 1 - D \) and \( R_{\text{bt}}, R_L, R_{C_{\text{in}},} R_{C_{\text{out}}} \) are the ESRs of the battery, inductor, input capacitor and output capacitor, respectively. \( I_L \) and \( I_{\text{Load}} \) are the average inductor and load current, respectively, while \( \Delta I_L = I_L - I_v \). Assuming a switching-frequency of \( f_s \) and \( t_{d1,2} \ll T_s \), the converter can operate in dual-ZVS mode for \( I_L \) given by (2.3), where \( I_v \) is given by (2.4).

\[
I_L < I_{\text{crit}} = \frac{I_{L,\text{sat}} - \Delta I_m + I_v}{2}
\]

(2.3)

\[
I_v \approx -\frac{C_s V_{\text{bus}}}{t_{d1}}
\]

(2.4)

In this chapter, a tri-mode control scheme with variable switching frequency in the mid-load range is proposed for achieving a high efficiency over the full load range, while meeting the constraints of Fig. 2.4 and limiting the peak-to-peak inductor current. The ideal inductor current waveform is shown in Fig. 2.5 for a steadily decreasing power, in order to illustrate the three operating modes. The converter automatically transitions
from fixed-frequency operation at heavy-load, to near fixed valley current and variable frequency in the mid-load range, and finally, Pulse Frequency Modulation (PFM) at light loads.

Figure 2.5: Converter current command, \( I_{lt^*} \), and inductor current, \( i_L \), in different modes of operation.

2.2.1 Mode Description

The three operating modes shown in Fig. 2.5 are described in order of decreasing \( I_L \):

1. **Fixed-frequency (FF) Mode:**

   In this mode \( I_{L,QR,\text{max}} < |I_L| < I_{L,\text{max}} \). The converter operates at the maximum frequency, \( f_s = f_{\text{max}} \), which is the minimum operating frequency such that \( |i_L(t)| < I_{\text{sat}} - \Delta I_m \). In this mode, \( t_{d2} \) is controlled to achieve ZVS turn-on in \( Q_2 \). The transistor \( Q_1 \) experiences hard-switching, since \( I_v > 0 \) and \( t_{d1} \) is minimized to avoid cross-conduction. Operating with a fixed \( f_s \) at heavy-load conditions reduces the potential for EMI problems. At this switching frequency, inductor core losses as well as the eddy current losses become quite significant.
2. Variable Frequency Quasi-Resonant (VFQR) Mode:

In this mode $I_{PFM,max} < |I_L| < I_{QR,max}$. At $I_{QR,max}$, the frequency is reduced from $f_{max}$ to $f_{QR,min}$ such that $I_p = I_{sat} - \Delta I_m$ and the valley current, $I_v = I_{vr}$, provides sufficient charge to discharge $v_x$ from $V_{bus}$ to 0, in order to achieve ZVS turn-on in $Q_1$. The waveform of $i_L(t)$ at the edge of VFQR mode is shown in Fig. 2.5. In VFQR mode, $f_s$ is controlled such that a fixed dead-time $t_{d1}$ and a near-constant $I_{vr}$ are maintained as the battery current changes. The dead-time $t_{d2}$ is also adjusted such that dual-ZVS turn-on is achieved. Increasing $f_s$ while reducing $I_L$ in this mode is non-conventional, however in the absence of turn-on losses, it helps to maintain a near constant $I_v$. This approach has several important advantages: 1) at light-loads, for the same $I_L$, higher $f_s$ reduces the valley current and therefore prevents large negative currents into $C_{in}$, 2) it reduces $\Delta I_L$ and $P_{cond}$ at light-loads and 3) the constant dead-time is much easier to control in order to guarantee ZVS over a wide current range. In this mode, since $I_L(R_{bd} + R_L + R_{on}) << V_{bt}$, $f_s \propto 1/I_L$ as given by (2.5), where $D$ is given by (2.6) based on a time averaged model [17].

\[
\begin{align*}
    f_s & \approx \frac{V_{bt}D}{2L(I_L - I_v)} \quad (2.5) \\
    D & = 1 - \frac{V_{bt} - I_L(R_{bd} + R_L + R_{on})}{V_{bus}} \quad (2.6)
\end{align*}
\]

Three operating conditions in this mode are shown in Fig. 2.5.

3. Pulse-Frequency Modulation (PFM) Mode:

In this mode $I_{L,min} < |I_L| < I_{PFM,max}$. The converter operates in a fixed on-time PFM mode, as in [18–20]. Constant on-time is applied to $Q_1$ while $Q_2$ is controlled to avoid body-diode conduction as $i_L(t)$ decays to 0 as shown in Fig. 2.5. Operating in PFM results in higher efficiency below $I_{PFM,max}$, where the gate-drive losses, $P_{dr}$, and controller losses, $P_{cont}$, become significant. It can be shown that $f_s \propto I_L$ in
PFM mode [19, 20]. The low-efficiency region $|I_L| < I_{L,\text{min}}$ can be easily avoided by the system power optimizer, due to the presence of two energy sources. The limit $I_{L,\text{min}}$ is selected such that $f_{\text{min}} > 20 \text{kHz}$ lies outside of the audible band. Further efficiency improvements can be achieved using adaptive on-time PFM or burst-mode [21].

The above mode descriptions apply only to the first quadrant, where $I_L > 0$. In the second quadrant, the control strategy is identical, while $Q_1$, $Q_2$, $t_{d1}$ and $t_{d2}$ have opposite roles, as demonstrated in Section 2.4. In the second quadrant, the following changes apply: 1) In FF mode, $t_{d1}$ is controlled to achieve ZVS turn-on in $Q_1$, while $Q_2$ experiences hard-switching. 2) In VFQR mode, $t_{d2}$ is constant, while $t_{d1}$ and $f_s$ are adjusted to achieve dual-ZVS turn-on, resulting in a near-constant $I_p$. 3) In PFM mode, the fixed on-time is applied to $Q_2$ and diode emulation is achieved using $Q_1$. A summary of chosen switching frequency and dead-time control strategy of the tri-mode converter are illustrated in Fig. 2.6.

2.2.2 VFQR versus FFQR: Loss Comparison

The main purpose of this section is to specifically contrast VFQR mode with the more conventional Fixed-Frequency Quasi-Resonant mode (FFQR) in the medium-load range. While the discussion is limited to operation in the first quadrant for the sake of brevity, the arguments can easily be extended to both quadrants.

Assume that the converter covers a current range, $I_{L,QR,1} \leq I_L \leq I_{L,QR,2}$, in the QR mode. For a fair comparison, it is assumed that both schemes have same valley current, $I_{v,QR,2}$, at the the edge of QR mode:

$$I_{v,FFQR,2} = I_{v,VFQR,2} = I_{v,QR,2}. \quad (2.7)$$

In FFQR mode, $f_s$ is fixed and chosen such that QR operation is guaranteed over the stated current range, with a negative valley current, $I_v$. $f_s$ is therefore chosen based on
the maximum average inductor current, $I_{L,QR,2}$, as shown in Fig. 2.7 (a). The inductor current waveform at lower current, $I_{L,QR,1}$ is also shown in Fig. 2.7(a), while $f_s$ is constant, as shown in Fig. 2.7(b). The constant switching frequency in this mode is given by

$$f_{s,FFQR} = \frac{D \cdot V_{bt}}{2 \cdot L \cdot (I_{L,QR,2} - I_{v,QR,2})}. \tag{2.8}$$

Furthermore, the valley current can be calculated as a function of $I_L$ and $f_{s,FFQR}$,

$$I_{v,FFQR} = I_{L,FFQR} - \frac{D \cdot V_{bt}}{2 \cdot L \cdot f_s}. \tag{2.9}$$

Note that FFQR mode is not defined for $I_L >> I_{L,QR,2}$, since it no longer results in a negative valley current.

Now consider that the converter is operated in the VFQR mode, where $f_s$ is automatically adjusted to achieve soft turn-on of the low-side switch, as described in Section 2.2.
Figure 2.7: (a) Inductor current waveforms and (b) switching frequency in FFQR operation.

The corresponding inductor current waveforms and switching frequency are shown in Fig. 2.8 (a) and (b), respectively. The switching frequency is given by

\[
f_{s,FFQR} \approx \frac{D \cdot V_b}{2 \cdot L \cdot (I_{L,FFQR} - I_{v,QR,2})},
\]

and scales with the average inductor current, \(I_{L,FFQR}\). The valley current in VFQR mode stays relatively constant,

\[
I_{v,VFQR,1} \approx I_{v,VFQR,2} \approx I_{v,QR,2}.
\]

While FFQR mode offers a more predictable EMI noise spectrum than VFQR mode, it can be shown that VFQR has improved light-load efficiency, due to reduced conduction losses. Since both schemes offer ZVS turn-on in the low and high-side switches, the turn-on switching losses, \(P_{sw,on}\), are negligible and the turn-off switching losses are given by

\[
P_{sw,off} = \frac{1}{2} \cdot f_s \cdot (V_{bus} \cdot I_p + V_{bus} \cdot |I_v|) \cdot t_f = \frac{1}{2} \cdot f_s \cdot V_{bus} \cdot (I_p - I_v) \cdot t_f,
\]

where \(t_f\) is the switch fall-time. In order to calculate the turn-off losses, \(P_{sw,off}\), we need to find \(I_p\) and \(I_v\) in each mode for a certain average current, \(I_L\),

\[
I_L = k \cdot I_{L,QR,2},
\]
where $k$ is bounded by

$$\frac{I_{L,QR,1}}{I_{L,QR,2}} < k < 1. \quad (2.14)$$

In the FFQR mode, $I_p$ and $I_v$ are given by

$$I_{p,FFQR} \approx (k + 1) \cdot I_{L,QR,2} - I_{v,QR,2}, \quad (2.15)$$

$$I_{v,FFQR} \approx I_{v,QR,2} - (1 - k) \cdot I_{L,QR,2}. \quad (2.16)$$

Based on $f_s$ from (2.8), the turn-off losses in FFQR mode can be calculated using (2.12) as

$$P_{sw,off,FFQR} = \frac{1}{2} \cdot V_{bus} \cdot t_f \cdot \frac{D \cdot V_{bt}}{L}. \quad (2.17)$$

On the other hand, in the VFQR mode, $I_p$ and $I_v$ are given by

$$I_{p,VFQR} \approx 2 \cdot k \cdot I_{L,QR,2} - I_{v,QR,2}; \quad (2.18)$$

$$I_{v,VFQR} \approx I_{v,QR,2}; \quad (2.19)$$

and $f_s$ can be calculated based on (2.10) as follows,

$$f_{s,VFQR} = \frac{D \cdot V_{bt}}{2 \cdot L \cdot (k \cdot I_{L,QR,2} - I_{v,QR,2})}. \quad (2.20)$$
Finally, the turn-off losses in the VFQR mode using (2.12) are given by

\[ P_{sw,off,VFQR} = \frac{1}{2} \cdot V_{bus} \cdot t_f \cdot \frac{D \cdot V_{bt}}{L}. \]  \hspace{1cm} (2.21)

From (2.17) and (2.21), the turn-off losses are identical in VFQR and FFQR mode and therefore the switching losses are given

\[ P_{sw,VFQR} = P_{sw,FFQR} = P_{sw,on} + P_{sw,off} = 0 + \frac{1}{2} \cdot V_{bus} \cdot t_f \cdot \frac{D \cdot V_{bt}}{L}. \]  \hspace{1cm} (2.22)

Given that the gate-drive losses, \( P_{dr} \), are negligible in the mid-load range, VFQR mode offers improved efficiency due to its lower conduction losses and is hence adopted in this converter.

### 2.2.3 Closed-Loop Frequency and Dead-time Control Scheme

As outlined in Section 2.1, \( I_{bt*} \) is periodically calculated in the system power optimizer. Once a current command has been received by the dc-dc converter, the mode is selected based on the pre-defined mode limits, as shown in Fig. 2.6(a).

The digital controller architecture is shown in Fig. 2.9(a). Two auxiliary PWM signals, \( c'_1 \) and \( c'_2 \), are generated by the controller and delayed from the main PWM signals, \( c_1 \) and \( c_2 \), by \( \phi_1 \) and \( \phi_2 \), respectively. The auxiliary PWM signals are used to trigger the high-speed latched comparators that are sampled to detect the ZVS condition on both transistors. The delays \( \phi_1 \) and \( \phi_2 \) are programmed such that \( c'_1 \) and \( c'_2 \) are in phase with the actual \( V_{gs} \) crossing of \( V_{th} \) for \( Q_1 \) and \( Q_2 \), as shown in Fig. 2.9(b). For improved accuracy, \( \phi_1 \) and \( \phi_2 \) can even be programmed to vary with \( I_{bt*} \) in order to account for the current dependency of the gate-driver delays. The output of the comparators can thus be monitored to automatically control \( t_{d2} \) in first quadrant, \( t_{d1} \) in second quadrant operation, as well as \( f_s \) for ZVS operation.

The pre-calculated frequency and the dead-times are fed forward to the controller as the current command is updated. These approximate values may not be optimal for
achieving dual ZVS turn-on due to the parameter variations and uncertainties associated with mass-production. Automatic adjustment is therefore necessary to achieve the optimal efficiency under a wide range of conditions. In order to avoid the dead-time and frequency adjustments interfering with the current-loop, the fine-tuning of the dead-time and switching frequency is initiated when the current-loop has reached steady-state, based on the internal digital error signal in the current-loop, $e[n]$. Automatic calibration of the dead-time and switching frequency is done using a binary search algorithm around the calculated feed-forward values, based on the comparators outputs.

![Diagram](a)

Figure 2.9: (a) Simplified architecture of the VFQR controller. (b) Signal timing in VFQR mode.

## 2.3 Dead-Time Extension Using Controllable Resonant Capacitance

In this section we address one of the common complaints about QR converters, namely the challenge of achieving precise dead-time control in the presence of production spreads in all the components, comparator delays and other non-idealities. In particular, $t_{d2}$ in
the first quadrant and $t_{d1}$ in the second quadrant are extremely short under heavy load conditions, as shown in Fig. 2.6(a). Accurate dead-time adjustment under heavy load conditions is extremely important. Without the use of relatively expensive hardware, reliably achieving ZVS for $t_d < 10$ ns on a high-power, production-grade converter is prohibitive.

![Synchronous boost converter with auxiliary circuitry on the switching node.](image)

Figure 2.10: Synchronous boost converter with auxiliary circuitry on the switching node.

The most straightforward way to extend the dead-times into a comfortable range is to increase the effective capacitance $C_x$ using an auxiliary capacitor, $C_{QR}$. Unfortunately this reduces the efficiency at heavy loads, where hard-switching is used in FF mode. In order to actively control the effective resonant capacitance, a switch, $Q_3$, is added in series with $C_{QR}$, as shown in Fig. 2.10. It should be noted that $Q_3$ has the same blocking voltage and peak current as the main transistors, $Q_1$ and $Q_2$, however it has a very small average current and can thus be much smaller, cheaper and have a lower parasitic capacitance. The control of $Q_3$ is discussed in Section 2.4.3.

### 2.4 Experimental Results

A 2 kW prototype of the tri-mode bi-directional dc-dc converter was fabricated, as shown in Fig. 2.11. While the fabricated boost converter has two symmetric phases, the experimental results are reported for single-phase operation for simplicity. The frequency and dead-time control strategy can easily be extended to multi-phase operation, either by
simply replicating the control circuits, or by using the same dead-times in both phases and relying on matching. The nominal converter parameters are listed in Table. 3.2.

Figure 2.11: Fabricated two-phase 2 kW quasi-resonant tri-mode converter prototype.

A 16-bit microcontroller (dsPIC33FJ64GS606) is used to implement the digital ACMC scheme. This specialized microcontroller has a 1.04 ns resolution for adjusting the switching period, duty-cycle and dead-time.

2.4.1 Steady-State Measurements

The measured converter waveforms for VFQR mode are shown in Fig. 2.12. The inductor current has a peak value of 11 A, while the valley current reaches -2 A with $I_L = 4.5$ A and $f_s = 192$ kHz. The measured waveform of the converter operating in the PFM mode is shown in Fig. 2.13, with $I_L = 0.4$ A. The designed converter operates with a minimum
frequency of \( f_{\text{min}} = 68 \, \text{kHz} \), which lies safely outside of the audible band.

The measured \( t_{d1} \) and \( t_{d2} \) for FF and VFQR modes in both quadrants are shown in 2.14(a). The measured \( f_s \) of the tri-mode boost converter is shown in Fig. 2.14(b). \( f_s \) drops from 400 kHz to 90 kHz at the border of VFQR mode. In this application, the system power optimizer that periodically updates \( I_{bt} \) can easily be designed to avoid operating right at the boundary of FF and VFQR modes, to avoid frequent mode-switching and the associated EMI issues. Hysteresis is also implemented on the converter to provide noise immunity at the boundary of modes.

The measured efficiency is shown in Fig. 2.15(a) for the different operating modes. For this particular design, despite the higher conduction losses, VFQR mode provides a higher efficiency even at high load currents, however operating in this mode for \( I_L > I_{L,QR,max} \) violates the inductor constraints of Fig. 2.4. PFM offers a high efficiency at light-load

---

**Table 2.1: Experimental System Parameters**

<table>
<thead>
<tr>
<th>System Parameter</th>
<th>Value</th>
<th>Unit</th>
</tr>
</thead>
<tbody>
<tr>
<td>Bus Voltage, ( V_{bus} )</td>
<td>100</td>
<td>V</td>
</tr>
<tr>
<td>Battery Voltage, ( V_{bt} )</td>
<td>50</td>
<td>V</td>
</tr>
<tr>
<td>Max. Switching Frequency, ( f_{max} )</td>
<td>400</td>
<td>kHz</td>
</tr>
</tbody>
</table>

**Converter Parameter (Each Phase)**

<table>
<thead>
<tr>
<th>Parameter</th>
<th>Value</th>
<th>Unit</th>
</tr>
</thead>
<tbody>
<tr>
<td>Max. Power, ( P_{max} )</td>
<td>1</td>
<td>kW</td>
</tr>
<tr>
<td>Saturation Current, ( I_{L,sat} )</td>
<td>29</td>
<td>A</td>
</tr>
<tr>
<td>Current Margin, ( \Delta I_m )</td>
<td>5</td>
<td>A</td>
</tr>
<tr>
<td>Max. Current, ( I_{L,max} )</td>
<td>20</td>
<td>A</td>
</tr>
<tr>
<td>Min. Current, ( I_{L,min} )</td>
<td>0.4</td>
<td>A</td>
</tr>
<tr>
<td>Inductors, ( L )</td>
<td>10</td>
<td>( \mu \text{H} )</td>
</tr>
<tr>
<td>Inductor ESR, ( R_L )</td>
<td>2.7</td>
<td>m( \Omega )</td>
</tr>
<tr>
<td>MOSFET on-resistance, ( R_{on} )</td>
<td>5.9</td>
<td>m( \Omega )</td>
</tr>
<tr>
<td>Estimated Parasitic Capacitance, ( C_x )</td>
<td>2.6</td>
<td>nF</td>
</tr>
</tbody>
</table>
Figure 2.12: Measured steady-state waveforms of the converter operating in VFQR mode.

f-1: $V_{gs,Q2}$, 5 V/div; ch-2: $v_x$, 49.6 V/div; ch-3: $V_{gs,Q1}$, 5 V/div; f-4: $i_L$, 5 A/div.

Figure 2.13: Measured steady-state waveforms of the converter operating in PFM mode.

f-1: $V_{gs,Q2}$, 5 V/div; ch-2: $v_x$, 49.6 V/div; ch-3: $V_{gs,Q1}$, 5 V/div; f-4: $i_L$, 2 A/div.
conditions, as shown in Fig. 2.15(a). The efficiency of the tri-mode boost converter with optimal mode-switching based on $I_{BL*}$ is shown Fig. 2.15(b). Peak efficiencies of 97.5% and 97% are achieved in the first and second quadrant, respectively. A minimum efficiency of 88% is maintained over the full load range and an efficiency improvement of 2.6% is achieved using VFQR mode over the FF mode at the boundary. Given that the high-side and low-side switches are identical, the efficiency curve in first and second quadrant should be identical, however the slight difference in the measurements can be attributed to variations in parasitic elements of $Q_1$ and $Q_2$, as well as the finite accuracy of the ZVS control scheme.
Figure 2.15: Measured efficiency in (a) different modes of operation and with (b) tri-mode converter.

### 2.4.2 FFQR and VFQR Comparison

In order to confirm the relative benefits of VFQR mode, the converter was also tested in FFQR mode over a range of $2 < I_L < 12.5$ A. The measured $f_s$ in both modes is shown in Fig. 2.16. $f_s$ changes from 330 kHz to 90 kHz in VFQR mode over the specified current range.

The measured efficiency is shown in Fig. 2.17. As the load current is reduced, the efficiency in the FFQR mode starts to suffer from its high conduction losses. The efficiency of the two modes converge as $I_L$ increases, since both modes have the same valley current at the boundary of the QR range, as in (2.7).

Infrared (IR) images taken using a FLIR T300 camera are shown in Fig. 2.18, to
illustrate the temperature distribution of the prototype converter in different operating modes, at the same operating point, $I_L = 10\, \text{A}$. The IR camera provides peak temperature for the area defined around the phase 1 power-stage. Figs. 2.18 (a) and (c) clearly show the effectiveness of the VFQR mode compared to the FF mode in the mid-load range, as the dominant losses, and hence the high temperatures, are shifted from the active semiconductor switches to the passive inductor. Fig. 2.18 (b) and (c) show that the lower ripple, $\Delta I_L$, of the VFQR mode compared to the FFQR mode result in 12.6 °C decrease in the inductor temperature, as a result of the reduced conduction losses.
2.4.3 Dead-Time Extension Measurements

The dead-time, efficiency and switching frequency of the prototype with the added auxiliary circuit of Fig. 2.10 was also evaluated. The auxiliary capacitance, $C_{QR}$, was first set to 2.4 nF, which is in the same range as $C_2$, and then to 8.2 nF, for the sake of comparison. The transistor parameters of $Q_3$, parameters are given in Table. 2.2.

The dead-time measurement of the converter in VFQR mode with and without the auxiliary circuit is shown in Fig. 2.19. Only $t_{d2}$ is shown in Fig. 2.19 as $t_{d1}$ is constant in the first quadrant VFQR mode. As expected, by adding $C_{QR}$, the shorter dead-time,
Table 2.2: Auxiliary MOSFET Parameters

<table>
<thead>
<tr>
<th>System Parameter</th>
<th>Value</th>
<th>Unit</th>
</tr>
</thead>
<tbody>
<tr>
<td>Break down voltage, $V_{dss}$</td>
<td>200</td>
<td>V</td>
</tr>
<tr>
<td>Continuous drain current, $I_D$</td>
<td>24</td>
<td>A</td>
</tr>
<tr>
<td>Pulsed drain current, $I_{DM}$</td>
<td>96</td>
<td>A</td>
</tr>
<tr>
<td>MOSFET on-resistance, $R_{on}$</td>
<td>100</td>
<td>mΩ</td>
</tr>
<tr>
<td>Total gate charge, $Q_{gt}$</td>
<td>57</td>
<td>nC</td>
</tr>
</tbody>
</table>

$t_{d2}$ in the first quadrant and $t_{d1}$ in the second quadrant, has effectively increased. It is important to notice that the $C_{QR}$ provides a well defined dead-time range, since the non-linear parasitic capacitance, $C_x$, can vary drastically across different products. By increasing $C_{QR}$, we can further increase the dead-time and also further decouple the duration of the dead-time from the size of the parasitic capacitance.

![Figure 2.19: Comparison of $t_{d2}$ for the converter with and without the auxiliary circuitry in VFQR mode.](image)

The closed-loop $f_s$ in VFQR mode with and without the auxiliary circuit is shown in Fig. 2.20. The effect of $C_{QR}$ on $f_s$ is reduced as $I_L$ increases. This phenomenon can be explained by considering $I_v$ in the first quadrant. As described in (2.4), $I_v$ in the first quadrant depends on the duration of the corresponding dead-time, $t_{d1}$, and the switching node capacitance. Increasing this capacitance further decreases $I_v$, which, in turn, causes $\Delta I_L$ to increase. From (2.5), $f_s$ also decreases and if $I_L$ is small, the effect of the $C_{QR}$ on
$f_s$ increases, as shown in Fig. 2.20.

![Graph of measured switching frequency with and without the auxiliary circuit in VFQR mode.](image)

Figure 2.20: Measured switching frequency with and without the auxiliary circuit in VFQR mode.

Finally, the efficiency of the prototype converter in the VFQR mode with and without the auxiliary capacitance on the switching node is shown in Fig. 2.21. As anticipated, by increasing the auxiliary capacitance, the efficiency decreases. Similar to the converter switching frequency, the effect of the auxiliary circuit on the converter efficiency will decrease as $I_L$ increases. It should be noted that at higher currents, the conduction losses are the dominant source of losses and since the switching frequency of the converter as well the current ripple do not change drastically when $I_L$ is larger, the conduction losses and therefore the efficiency of the converter with and without the auxiliary circuit does not change noticeably.

![Graph of measured efficiency with and without the auxiliary circuit in VFQR mode.](image)

Figure 2.21: Measured efficiency with and without the auxiliary circuit in VFQR mode.
As shown in Figs. 2.19–2.21, the auxiliary circuit can increase the dead-time without drastically changing the switching frequency and efficiency of the converter at heavy loads. Having this in mind, we can enable the auxiliary circuit in the higher load currents, while effectively negating its effect by turning off $Q_3$ at light loads. As result, the converter achieves both high-efficiency through the entire VFQR range and high ZVS accuracy through dead-time extension. A summary of the measured characteristics in the VFQR mode with this proposed scheme is shown in Fig. 2.22, where $C_{QR} = 2.4 \text{ nF}$. The $t_{d2}$ is effectively doubled when $Q_3$ is turned on, while the efficiency drops less than 1%.

Figure 2.22: Measured (a) dead-time, (b) switching frequency and (c) efficiency of the converter in VFQR mode with auxiliary circuit turned on only in the heavy load range.
2.4.4 Dynamic Measurements

The transient response of the converter due to a step in $I_{bt}*$ from 15 A to 6.5 A is shown in Fig. 2.23. The converter first operates in FF mode at 15 A and automatically switches to VFQR mode following the command-step. The measured converter waveforms are shown in Fig. 2.23(a). The measured $f_s$ in VFQR mode is shown in Fig. 2.23(b), during the automatic adjustment process. $f_s$ is first updated based on a feed-forward and then fine-tuned using a binary search algorithm, as described in Section 2.2. The dead-time and frequency fine-tuning begin 7 ms after the command-step, when steady-state is detected. The frequency and dead-time adjustment are done in less than 15 ms, which is an order of magnitude faster than the update rate of $I_{bt}*$ in the full system. While operating in FF mode at 15 A, $f_s = 400 \text{ kHz}$ while the optimized $f_s$ in VFQR mode at 6.5 A is 154 kHz with a valley current, $I_v \approx -2 \text{ A}$. 
Figure 2.23: Measured transient response of the converter due to a step in $I_{bt}$ from 15 A to 6.5 A. (a) Inductor current and internal digital control signals. ch-4: $i_L$, 10 A/div. $f_s$ adjustment = 0 indicates that the frequency is decreasing. Dead-time adjustment = 0 indicates that the dead-time is increasing. (b) Switching frequency during calibration process.
The converter’s transient response due to a change in $I_{bt^*}$ from 3 A to 9 A is shown in Fig. 2.24. In this case, the converter operates in VFQR mode for both values of $I_{bt^*}$. The measured waveforms are shown in Fig. 2.23(a). The switching frequency of the converter, which is regulated using the method described in Section 2.2, is shown in Fig. 2.24(b). The optimized $f_s$ at 3 A and 9 A is 227 kHz and 110 kHz, respectively, while the converter maintains a nearly constant valley current of $I_v \approx -2$ A.

![Graph showing transient response](image)

Figure 2.24: Measured transient response of the converter due to a step in $I_{bt^*}$ from 3 A to 9 A. (a) Inductor current and internal digital adjustment signals. ch-4: $i_L$, 10 A/div. $f_s$ adjustment = 0 indicates that the frequency is decreasing. Dead-time adjustment = 0 indicates that the dead-time is increasing. (b) Switching frequency during calibration process.
2.5 Chapter Summary and Conclusions

This chapter targets the design of a high efficiency bi-directional dc-dc converter for the HESS applications in LEVs. The concept of variable-frequency quasi-resonant operation was demonstrated for maintaining a constant dead-time and a near constant inductor valley current in the mid-load range. We have compared this scheme with the conventional fixed-frequency quasi-resonant scheme and it is shown that the VFQR mode provides higher efficiency. Fixed frequency and PFM operating modes are combined with the VFQR operation to obtain a tri-mode bi-directional converter that can operate with a high efficiency over the full load range. The proposed auxiliary circuit is effective in extending the dead-times for improved ZVS accuracy, without significantly deteriorating the efficiency. The implemented prototype dc-dc converter achieves the peak efficiency of 97.5% and maintains the efficiency of 88% over a wide load range. The benefits of VFQR over FFQR mode and effectiveness of the proposed dead-time extension scheme are confirmed in this power range using infrared images and efficiency measurements.
References


Chapter 3

HESS Design for EVs with RT GPS Based Control Method

3.1 Introduction

Implementing low-cost, high-energy and high power-density storage that performs reliably for 10-15 years is crucial for the mass adoption of EVs. Advantages of U-cap based HESS, including life-time extension of its storage elements, were discussed in details in Chapter 1. To summarize, the main objectives of an automotive HESS are to (1) minimize the battery stress during rapid accelerations in order to limit long-term capacity fading and (2) maximize the capture of the Regen, while reducing the wear on the mechanical brakes. Accurately predicting the battery lifetime extension due to the reduction in dynamic currents under real drive-cycle conditions is a major challenge, and is currently under investigation.

This chapter targets a u-cap based HESS for the custom EV prototype, known as A2B, shown in Fig. 3.1 [1]. The chosen HESS topology that is shown in Fig. 3.2 was discussed in Section 1.2.2 and also shown in Fig. 1.10. The topology includes a non-isolated bi-directional dc-dc converter between $V_{uc}$ and $V_{bt}$. This architecture allows (1)
effective power sharing control within the HESS, (2) flexible voltage swing and thus good utilization of the u-cap energy, and (3) minimal number of conversion stages from the battery to the load. Several control strategies for this topology are presented in the literature, as reviewed in Chapter 1.

Figure 3.1: EV Prototype design by Project Eve (The “A2B”).

Figure 3.2: Utilized HESS architecture.

3.1.1 The A2B Vehicle

A2B is a Canadian made full-electric vehicle, with locally sourced materials and manufacturers, developed by Project Eve [1], Canada’s multi-company consortium formed to
advance electric mobility. Companies that specialize in electric motors, drivetrains, battery management systems, lithium-ion battery recycling, etc. The vehicle’s specifications are listed in Table 3.1.

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<tr>
<th>Vehicle Parameter</th>
<th>Value</th>
<th>Unit</th>
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<tr>
<td>Max. Vehicle Speed, $\nu_{\text{max}}$</td>
<td>116</td>
<td>km/h</td>
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<td>Car Mass (without HESS)</td>
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<td>kg</td>
</tr>
<tr>
<td>Estimated Range</td>
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<td>km</td>
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<tr>
<td>Gearbox Ratio, $K_g$</td>
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<tr>
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<table>
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<td>Number of Series modules</td>
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<td>Pack Mass</td>
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<tr>
<td>Pack Volume</td>
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<td>L</td>
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<table>
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<th>Unit</th>
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<tr>
<td>Nominal Voltage, $V_{\text{bt,nom}}$</td>
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<td>V</td>
</tr>
<tr>
<td>Module Capacity</td>
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<td>Ahr</td>
</tr>
<tr>
<td>Module ESR</td>
<td>6</td>
<td>mΩ</td>
</tr>
<tr>
<td>Specific-Energy</td>
<td>89.1</td>
<td>Wh/kg</td>
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<tr>
<td>Specific-Power</td>
<td>0.432</td>
<td>kW/kg</td>
</tr>
<tr>
<td>Cycle Life (20% Capacity degradation)</td>
<td>2,800</td>
<td>Cycles</td>
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<table>
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<tr>
<th>Drivetrain Parameter</th>
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<tr>
<td>Max. Motor Speed, $\omega_{\text{max}}$</td>
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<td>rpm</td>
</tr>
<tr>
<td>Max. Torque (continuous), $T_{\text{Max}}$</td>
<td>65</td>
<td>Nm</td>
</tr>
<tr>
<td>Max. Torque (for 30 sec), $T_{\text{peak}}$</td>
<td>170</td>
<td>Nm</td>
</tr>
<tr>
<td>Max. Power (continuous), $P_{\text{Max}}$</td>
<td>37</td>
<td>kW</td>
</tr>
<tr>
<td>Max. Power (for 30 sec), $P_{\text{peak}}$</td>
<td>80</td>
<td>kW</td>
</tr>
<tr>
<td>Efficiency at Nominal Operation</td>
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<td>%</td>
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<td>Operating Voltage</td>
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<td>36</td>
<td>kg</td>
</tr>
<tr>
<td>Inverter Weight</td>
<td>34</td>
<td>kg</td>
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</tbody>
</table>

The vehicle battery pack consists of 24 Valence U24-12XP Lithium Iron Magnesium Phosphate (LiFeMgPO4) battery modules [2] in series. This chemistry is considered safe
and energy efficient with a long life cycle [2]. The pack has a total mass and energy of 380 kg and 33.8 kWh, respectively. The module capacity drops 20% after 2800 cycles with 20 A charge and discharge rate as shown in Fig. 3.3. The module discharge curves are shown in Fig. 3.4. These curves are used to model the battery module and ultimately the battery pack using the approach presented in [3].

Figure 3.3: Valence U24-12XP capacity degradation over charge/discharge cycles [2].

The drivetrain, MΦTIVE A, is designed by TM4 [4] and its specifications are listed in Table 3.1. Typical performance curve of the drivetrain is shown in Fig. 3.5. It should be noted that the efficiency of the drivetrain varies with input voltage, operating speed and torque and is not shown in this thesis for confidentiality purposes. However, the efficiency curves are used in the system modeling for our simulation purposes.

The purpose of this chapter is to develop the full system model, including the vehicle’s mechanical system and the HESS consisting of the battery pack, the u-cap pack and the dc-dc converter. Then to propose a novel control scheme for the HESS based on the real-time GPS data processing. The dc-dc converter design and testings are provided in this chapter, while full system integration is beyond the scope of this work and will be
Figure 3.4: Valence U24-12XP discharge curves for various discharge currents [2].

Figure 3.5: TM4 drivetrain performance curve [4].
This Chapter is organized as follows. Data acquisition and analysis of a 3 hour, 61 km (round trip) typical urban drive-cycle is discussed in Section 3.2. HESS design and control strategies are discussed in Section 3.3. A bi-directional dc-dc converter design for this specific HESS is also provide in this Section. Section 3.4 discusses the simulation procedure and results of the proposed HESS based on the experimental drive-cycle introduced in Section 3.2. Lab experimental results of the designed dc-dc converter are provided in Section 3.5.

3.2 Drive-Cycle Data Acquisition

As the vehicle information is available on the Controller Area Network (CAN) bus, one can record this information for further analysis and modelings. Since a prototype vehicle is considered in this work, modeling of the car is an important part of the system simulation and design. The Regen braking is currently disabled in the vehicle to protect its battery pack against high charging currents, increasing the pack’s lifetime.

A 3 hour, 61 km (round trip) typical urban drive-cycle was performed with this vehicle in downtown Toronto in April 2012. All internal parameters in the mechanical/electrical sub-systems, as well as the GPS trajectory were recorded to develop a complete electromechanical model of the vehicle. The available mechanical parameters are, the vehicle speed (km/h), motor speed (rpm), motor torque (Nm), gas pedal position (%) and brake pedal status (0 or 1). Electrical parameters available on the CAN bus includes the battery voltage, current and state of charge.

The measured GPS trajectory is shown in Fig. 3.6 using Google Maps [5].
Figure 3.6: Vehicle trajectory for the 3 hour urban drive-cycle.
3.2.1 Drive-Cycle Analysis

The measured vehicle speed, motor speed and motor torque are shown in Fig. 3.7. The motor torque is always positive, since the Regen braking is disabled in the vehicle as shown in Fig. 3.7(c).

The battery voltage and current measured by the Battery Management System (BMS), and the calculated load power are shown in Fig. 3.8. The estimated battery SOC, broadcasted on the CAN bus by the BMS is shown in Fig. 3.9.

The vehicle traveled 61 km and consumed 11.97 kWh net electrical energy. The results show an average energy consumption of 19.62 kWh per 100 km (706 kJ/km), which compares favorably with the Chevy Volt [6] with an official EPA measurement of 22.4 kWh per 100 km (810 kJ/km).
Figure 3.7: Measured (a) vehicle speed, (b) motor speed and (c) motor torque during the urban drive-cycle.
Figure 3.8: Measured (a) battery voltage, (b) battery current and (c) load power during the urban drive-cycle.
3.3 HESS Description

The HESS system specifications are listed in Table 3.2. The HESS is loaded by an inverter and a 37 kW three-phase permanent magnet electric motor. Three u-cap modules (BMOD0165) are connected in series, building a 144 V, 55 F with 18.9 mΩ of ESR pack. The u-caps are then interfaced to the bus using a 30 kW bi-directional dc-dc converter. The design of this converter is discussed in more details in Section 3.3.3.

The embedded system control is split into two control-targets, contained inside a 9025 CompactRIO (CRI0) module with a 9114 Chassis from National Instruments. The real-time controller features an 800 MHz processor with 4 GB of nonvolatile storage and 512 MB of DDR2 memory. The 8-slot reconfigurable embedded chassis features a Xilinx Virtex-5 reconfigurable I/O (RIO) FPGA. The high-level control, which includes the vehicle CAN bus and GPS monitoring, street map analysis and power-mix calculation are done in the CPU. The high-speed IGBT gating signals, digital average current-mode control compensator and protection functions are implemented in the FPGA for minimum latency.
Table 3.2: HESS System Parameters

<table>
<thead>
<tr>
<th>U-cap Pack</th>
<th>Value</th>
<th>Unit</th>
</tr>
</thead>
<tbody>
<tr>
<td>Number of Series modules</td>
<td>3</td>
<td></td>
</tr>
<tr>
<td>Pack Mass</td>
<td>40.5</td>
<td>kg</td>
</tr>
<tr>
<td>Pack Volume</td>
<td>43.5</td>
<td>L</td>
</tr>
</tbody>
</table>

<table>
<thead>
<tr>
<th>U-cap Module (BMOD0165)</th>
<th>Value</th>
<th>Unit</th>
</tr>
</thead>
<tbody>
<tr>
<td>Nominal Voltage, $V_{uc,nom}$</td>
<td>48</td>
<td>V</td>
</tr>
<tr>
<td>Module Capacitance</td>
<td>165</td>
<td>F</td>
</tr>
<tr>
<td>Module ESR</td>
<td>6.3</td>
<td>mΩ</td>
</tr>
<tr>
<td>Specific-Energy</td>
<td>3.9</td>
<td>Wh/kg</td>
</tr>
<tr>
<td>Specific-Power</td>
<td>6.77</td>
<td>kW/kg</td>
</tr>
<tr>
<td>Cycle Life (20% Capacitance degradation)</td>
<td>1,000,000</td>
<td>Cycles</td>
</tr>
</tbody>
</table>

<table>
<thead>
<tr>
<th>4 Phase DC-DC Converter</th>
<th>Value</th>
<th>Unit</th>
</tr>
</thead>
<tbody>
<tr>
<td>Input Voltage</td>
<td>0-144</td>
<td>V</td>
</tr>
<tr>
<td>Output Voltage</td>
<td>240-350</td>
<td>V</td>
</tr>
<tr>
<td>Converter Mass</td>
<td>12</td>
<td>kg</td>
</tr>
<tr>
<td>Converter Volume</td>
<td>13.5</td>
<td>L</td>
</tr>
<tr>
<td>Maximum Power (continuous)</td>
<td>30</td>
<td>kW</td>
</tr>
</tbody>
</table>

<table>
<thead>
<tr>
<th>Phase Parameter of the DC-DC Converter</th>
<th>Value</th>
<th>Unit</th>
</tr>
</thead>
<tbody>
<tr>
<td>Inductors, $L$</td>
<td>90-170</td>
<td>µH</td>
</tr>
<tr>
<td>Inductor Saturation Current, $i_{L, sat}$</td>
<td>80</td>
<td>A</td>
</tr>
<tr>
<td>Inductor Max DC Current, $I_{L, max}$</td>
<td>60</td>
<td>A</td>
</tr>
<tr>
<td>Inductor ESR, $R_L$</td>
<td>5.9</td>
<td>mΩ</td>
</tr>
<tr>
<td>IGBT Saturation Voltage, $V_{CE, sat}$</td>
<td>0.75</td>
<td>V</td>
</tr>
<tr>
<td>IGBT on Resistance, $R_{on,Q}$</td>
<td>4.5</td>
<td>mΩ</td>
</tr>
<tr>
<td>IGBT Turn on Losses ($V_{CE} = 300$ V, $I_C = 200$ A)</td>
<td>3.3</td>
<td>mJ</td>
</tr>
<tr>
<td>IGBT Turn off Losses ($V_{CE} = 300$ V, $I_C = 200$ A)</td>
<td>8.3</td>
<td>mJ</td>
</tr>
<tr>
<td>Diode Reverse Recovery Energy ($V_{CE} = 300$ V, $I_C = 200$ A)</td>
<td>4.7</td>
<td>mJ</td>
</tr>
</tbody>
</table>
3.3.1 Power Optimizer Algorithm

Our objective is to introduce a new HESS power-mix optimization approach incorporating real-time GPS data. The prediction of upcoming stops and subsequent bursts of Regen helps to optimize the u-cap SOC.

As demonstrated in [7], the dc-dc converter efficiency varies with the u-cap SOC, which makes it more lossy (or expensive from the optimization point of view) to draw energy from the u-cap as its voltage drops. Therefore keeping the u-cap SOC at an optimal level, based on vehicle speed, is one of the key controller objectives. This requirement conflicts with the instantaneous optimum power mix, based solely on the ESRs and the dc-dc converter losses at each point in time. Predicting the load current using GPS data is therefore extremely valuable in the u-cap SOC considerations, as exploited in this work.

As lithium batteries dominate the EV system cost, extending the pack lifetime significantly with minimum incremental cost helps to increase effective mass EV adoption. For example, the Chevy Volt battery is expected to lose 10-30% of its capacity within 8 years or 160,000 km, depending on driving habits. The effect of charge/discharge rates on the batteries cycle life was shown in [8] and thus limiting the peak battery current is another major controller objective. The following power-mix optimization cost-function is minimized every second by the controller

\[
f_{\text{cost}} = P_{\text{bt;loss}} \cdot m_{\text{bt}} + (P_{\text{uc;loss}} + P_{\text{conv;loss}}) \cdot m_{\text{uc}} + P_{\text{mech;loss}},
\]  

(3.1)

under the following constraints:

\[
0.6 < \text{SOC}_{\text{uc}} < 1, \quad (3.2)
\]

\[
-216 < I_{\text{uc}} < 216, \quad (3.3)
\]

where \(P_{\text{uc;loss}} = I_{\text{uc}}^2 R_{\text{uc}}\) is the u-cap ESR loss, \(P_{\text{bt;loss}} = I_{\text{bt}}^2 R_{\text{bt}}\) is the battery ESR loss, \(P_{\text{conv;loss}}\) is the load-dependent dc-dc converter loss and \(P_{\text{mech;loss}}\) is the mechanical brak-
The constants \(m_{uc}\) and \(m_{bt}\) are adaptive multipliers introduced to influence the power-mix, based on the u-cap SOC and battery current management.

As described, one of the controller objectives is to limit the peak battery charge and discharge currents to the predefined values of \(I_{min}\) and \(I_{max}\), respectively. However, the limits can be violated if the u-cap can not safely handle the excess requested/generated power of the load. This is done by penalizing the higher battery peak currents by marginally increasing the cost of the battery contribution. The five operating modes based on the u-cap SOC are shown in Fig. 3.10. \(m_{uc}\) and \(m_{bt}\) are adjusted based on the \(SOC_{uc}\), such that in mode 1, the system performs most efficiently:

\[
m_{uc} = \begin{cases} 
1, & \text{if } 0 \leq SOC_{uc} < 0.05 \\
1 + \frac{I_{bt}}{I_{max}}, & \text{if } 0.6 < SOC_{uc} < 0.64 \\
\infty, & \text{if } SOC_{uc} \geq 0.64 
\end{cases}
\]

(3.4)

\[
m_{bt} = \begin{cases} 
1, & I_{min} < I_{bt} < I_{max} \\
1 + \frac{I_{bt}}{I_{max}}, & I_{bt} > I_{max} \\
\infty, & I_{bt} < I_{min}
\end{cases}
\]

(3.5)

while it performs most conservatively in mode 5 to increase the u-cap voltage.

The maximum \(V_{uc}\) that allows the u-cap to fully absorb a Regen burst, \(SOC_{uc,des,h}\), is calculated based on the vehicle speed. A band of 10% is defined according to \(SOC_{uc,des,h}\), in order to set the desired lower limit, \(SOC_{uc,des,l}\). Operating in Modes 2 and 3, within
Chapter 3. HESS Design for EVs with RT GPS Based Control Method

±5% of $SOC_{uc,des}$, is optimal from the Regen, acceleration and dc-dc converter efficiency point of view. The $m_{uc}$ and $m_{bt}$ vary linearly in this region from the most efficient case at $SOC_{uc,des,h}$ to highly conservative values at $SOC_{uc,des,l}$. In Mode 4, the u-cap is only activated to limit the battery current, while it tries to absorb most of the Regen energy, despite the dc-dc converter losses for low $V_{uc}$. Mode 5 is a critical mode, where the u-cap is charged at the expense of higher battery current and should ideally be avoided. The real-time GPS data is used to adaptively adjust $m_{uc}$ and $m_{bt}$, based on the prediction of upcoming stops. This helps to minimize the losses over the mighty bursts of power in the next 200 m, since the upcoming Regen energy from braking will compensate the $SOC_{uc}$. It should be noted that without the GPS information, decision based on minimum losses causes an undesired effective drop in the $SOC_{uc}$.

3.3.2 Open Street Map and GPS Data Processing

The incremental cost of using the GPS data is minimal, since the hardware is standard in all modern EVs. A vectorized street map of the city of Toronto [9] is stored locally in the vehicle controller using the following method. A specific area can be selected and its data can be downloaded as a text file using [9]. The map consists of three main elements, nodes, ways and relations. We have only considered the nodes as they are sufficient for our purpose. A node in the map describes a single geospatial point using its latitude and longitude. Each node has a tag that defines it to be a building, stop sign, parks, highways, traffic signals and etc. The map also contains some unnecessary information that are eliminated using Perl [10], a high level programming language, and only the tag, latitude and longitude of the nodes with stop sign and traffic signal tags are stored to make a smaller database to check the real-time position against.

Real-time GPS data processing is then performed to detect the presence and relative position of stop signs and traffic lights that are within 200 m of the car trajectory as shown in Fig. 3.19(a) for the measured drive-cycle. There are numerous challenges in
accurately predicting when the EV will actually come to a complete stop, due to the traffic conditions, the exact location of traffic stops and the fact that the state of the red/green lights is unknown.

As an example, a hypothetical situation is shown in Fig. 3.11. Points A and C represent traffic signals at which the vehicle might have to stop depending on the state of the red/green lights. Points B and F represent stop signs which might be falsely detected since the locations of the signs are in the direction of movement, but these stop signs are not actually intended for the route that the vehicle is moving along. Point D shows the specific pedestrian crossing point and depending on the pedestrian’s traffic, a stop might not be necessary. Point E falls within the 200 m detecting zone of the GPS data processor. However, it can be neglected as it is out of the vehicle moving direction. This is achieved by calculating the vehicle moving direction and the displacement vector for the position of the stop sign/traffic signal. Point G shows a stop sign that might be falsely picked up, like points B and F. Although at this point the vehicle is turning, the vehicle does not necessarily need to stop.

A detailed analysis of the experimental drive-cycle showed that the GPS system correctly detected actual stops in 66% of the cases for stop signs and 49% of the cases for traffic lights, which is deemed sufficient for this work.

3.3.3 Bi-directional DC-DC Converter

A non-isolated 30 kW bi-directional, liquid-cooled dc-dc converter was fabricated to deploy the HESS system in the car, as shown in Fig. 3.12. The converter box is mounted on top of the three Maxwell 165 F u-cap modules. The complete converter measures $30 \times 30 \times 15$ cm, including the chill-plate and excluding the breakers, contactor and CRIO module. Fig. 3.12(a) shows the half assembled converter while the inductors and IGBTs are visible, and Fig. 3.12(b) shows the full assembled converter with the IGBTs covered by their driver and the inductors covered by the input and output capacitor banks. The
Chapter 3. HESS Design for EVs with RT GPS Based Control Method

Figure 3.11: A hypothetical driving situation with the stop signs and traffic signals shown over a map. The route is shown in blue and the possible points that the GPS data processor may pick up are marked in purple.

The converter was designed and simulated using the new co-simulation methodology [11].

The converter has four interleaved IGBT half-bridge phases as shown in Fig. 3.13 (a). Multi-phase interleaved architecture provides lower current and voltage ripple and operates at higher efficiency when the gate drive loss is not the dominant source of losses. The inductor currents are shown in Fig. 3.13 (b). Phase shedding is implemented to decrease the switching losses and improve the efficiency at light loads. The switching frequency, $f_s$, is set to 20 kHz. A summary of converter activated phases as a function
Figure 3.12: Fabricated 30 kW 4-phase dc-dc converter (a) lower deck, half assembled top view. (b) full assembled top and side view.
of u-cap current is shown in Fig. 3.14.

![Diagram of multi-phase bidirectional dc-dc converter](image_url)

**Figure 3.13:** (a) Multi-phase bidirectional dc-dc converter. (b) Ideal inductor current in a 4 phase boost converter.

![Diagram of converter activated phases](image_url)

**Figure 3.14:** Converter activated phases.

### Controller Design

The converter modeling is done using the approach introduced in [12]. A single-phase equivalent model of the multi-phase converter is derived and averaged switch modeling [13] is then used to find the equivalent circuit of the converter for the control design purposes as shown in Fig. 3.15. The equivalent parameters in Fig. 3.15 are calculated as
follows:

\[
\begin{align*}
    f_{s,eq} &= 4f_s, \\
    L_{eq} &= \frac{L}{4}, \\
    R_{eq} &= \frac{R}{4}, \\
    V_{eq} &= \frac{V}{4},
\end{align*}
\]

where \( f_s \) is the phase switching frequency, \( L \) is the phase inductor, \( R \) is the sum of the effective parasitic resistances and \( V \) is the effective voltage drop due to the \( V_{ce} \) of the IGBT and the \( V_F \) of the anti-parallel diode. \( Z_{i,eq} \) and \( Z_{o,eq} \) represent the input and output capacitor banks, respectively. A unified controller using the approach proposed in [14] is then used to design a PI controller for both buck and boost mode and provide smooth transition between modes.

Figure 3.15: Equivalent circuit of the multi-phase half-bridge converter.

The converter symmetry simplifies the controller implementation where the PI controller is used to control phase 1 current by adjusting its duty cycle. Other phases operates around the same duty cycle with slight variations due to parasitics and component mismatches. The control architecture is shown in Fig. 3.16. Phase currents are allowed to be within an optimal acceptable range of the phase 1 current to avoid unnecessary oscillation in the duty cycles. It is important to note that perturbing one phase results in some perturbations in other phases due their physical coupling.
3.4 System Simulations based on Experimental Drive-Cycle

MATLAB is used for system level simulations to investigate the benefits of using the proposed GPS based HESS approach, using the experimental drive-cycle data from the A2B. System is modeled in MATLAB, several test scenarios are provided and finally the simulation results are provided.

3.4.1 System Modeling

Several components of the system are modeled using the device datasheets and the collected data as described below:

1. Battery Modeling:

The battery module is modeled using the discharge curves provided in Fig. 3.4 based on the modeling approach proposed by [3]. The pack is ultimately modeled using the equivalent circuit shown in Fig. 3.17 where the terminal voltage can be

Figure 3.16: Simplified controller implementation utilizing system symmetry.
found using:

\[ V_{bt} = V_{bti} - R_{bt} \cdot I_{bt}, \]  
(3.10)

\[ V_{bti} = E_0 - K \cdot \frac{Q}{Q - \int_0^t I_{bt} \, d\tau} + A \cdot e^{-B\int_0^t I_{bt} \, d\tau}, \]  
(3.11)

where \( R_{bt} \) is the battery ESR, \( E_0 \) is the battery constant voltage (V), \( Q \) is the battery capacity (Ah), \( K \) is the polarisation voltage (V), \( A \) is the exponential zone amplitude (V) and \( B \) is the exponential zone time constant inverse (Ah\(^{-1}\)).

![Battery Model Diagram]

Figure 3.17: Utilized battery model.

2. U-cap Modeling:

The u-cap bank is modeled by the equivalent circuit shown in Fig. 3.18 consists of an ideal capacitor in series with an equivalent resistance. The u-cap terminal voltage can be calculated using:

\[ V_{uc} = V_{uci} - R_{uc} \cdot I_{uc}, \]  
(3.12)

\[ V_{uci} = V_{uci,0} - \frac{1}{C_{uc}} \int_0^t I_{uc} \, d\tau, \]  
(3.13)

where \( R_{uc} \) is the u-cap ESR.

3. DC-DC Converter Modeling:

The converter is modeled to obtain the losses associated with its operation. Conduction losses and switching losses are considered and the gate drive losses are
neglected for such high power converter. Switching losses including the IGBT turn-on, turn-off losses and the diode reverse recovery losses are modeled using the IGBT module’s datasheet. The model is derived based on the number of activated phases, as shown in Fig. 3.14.

4. Vehicle Modeling:

Since Regen is disabled in the vehicle, a vehicle mechanical model has to be established in order to generate the negative torque data associated with the braking. The net force is given by

\[ \vec{F}_{net} = \vec{F}_{acc} + \vec{F}_F + \vec{F}_D, \]  

(3.14)

where \( \vec{F}_{acc} \) is the needed force for acceleration, \( \vec{F}_F \) is the rolling friction resistance and \( \vec{F}_D \) is the drag resistance. The acceleration force can be calculated using

\[ F_{acc} = m \cdot \frac{d\nu}{dt}, \]  

(3.15)

where \( m \) is the vehicle’s mass and \( \nu \) is the vehicle speed. Rolling friction force is modeled by:

\[ F_F = C_{rr} \cdot N, \]  

(3.16)

where \( C_{rr} \) is the dimensionless rolling resistance coefficient and \( N \) is the normal
force to the surface that the wheels are rolling on. Drag force is also modeled using:

\[ F_D = \frac{1}{2} \rho \cdot v^2 \cdot C_D \cdot A, \]  

where \( \rho \) is the mass density of air, \( C_D \) is the drag coefficient dependant on the vehicle’s geometry and \( A \) is the orthogonal projection area of the vehicle to the direction of motion.

For the purpose of modeling, it is assumed that the vehicle is running on a flat surface and the torque delivered to the wheels are modeled using a simplified equation as follows:

\[ \tau_{net} = R_w \times F_{net}, \]  

where \( R_w \) is the displacement vector. Based on typical values of \( C_{rr} \) and \( C_D \) for standard sedan vehicles, \( C_{rr} = 0.015 \) and \( C_D = 0.4 \) are considered in our modeling.

The torque delivered to the wheels is a function of mechanical system efficiency as described in:

\[ \tau_{net} = K_G \cdot \eta_m \cdot \tau_m, \]  

where \( \tau_m \) is the torque available on the motor shaft, \( K_G \) is the gearbox ratio and \( \eta_m \) is the mechanical system efficiency that depends on the operating point. A linear approximation as

\[ \eta_m = \eta_0 + \frac{|Tm|}{T_{max}} (\eta_{T_{max}} - \eta_0), \]  

is deployed in this work where \( \eta_{T_{max}} \) and \( \eta_0 \) are optimized using the least square method, such that the calculated motor torque best matches the available measured motor torque from the drive-cycle. Using these coefficients, the regenerative torque can be calculated for the drive-cycle. It should be noted that using a more sophisticated analysis, taking the road inclination and declinations into account, can further augment the system modeling but are beyond the scope of this work.
Now that we have the motor speed and motor torque, the load current can be calculated using the efficiency data of the drivetrain by:

\[ I_{bt} = \frac{P_e}{V_{bt}}, \]  
\[ P_e = \frac{P_m}{\eta_D}, \]

where:

\[ P_m = \tau \cdot \omega, \]
\[ \eta_D = f(\tau, \omega). \]

### 3.4.2 Simulation Procedure and Test Scenarios

Now that both the electrical and mechanical systems are modeled, the benefits of the HESS based EVs compared to the conventional SESS based EVs can be exploited. Several test scenarios are considered for simulations, as described below:

1. **SESS with No Regen**:

   This case is similar to the experimental drive-cycle measurement and is used as a reference. A comparison between the simulation result of this case versus the experimental results verifies the accuracy of system modelings.

2. **SESS with \(-0.25 C < I_{bt}\)**:

   This scenario represents a vehicle with Regen capability. However, the battery is unable to absorb all the Regen energy due to its limited charging current that is set in accordance with battery lifetime considerations. In these cases, all Regen energy for which \(-0.25 C < I_{bt}\) is returned to the electrical system, while the rest is dissipated in the mechanical brakes.
3. **SESS with No Limits:**

   Similar to the previous case, with a battery capable of absorbing the Regen energy at higher charging rates.

4. **HESS without GPS Data Processing:**

   This scenario utilizes the HESS with the vehicle Regen enabled. The battery charging current is limited to \(-0.25\, C\) and therefore having a u-cap with enough room for the charge helps to recover more energy. The battery discharge current is unlimited, but the power optimizer tries to limit battery discharge current to \(0.75\, C\). This limit can be violated if the u-cap SOC is too low and the u-cap is not capable of handling the load demand with the limited battery current. This feature maintains the user experience of the EV while provides improved battery lifetime. The u-cap is targeted to operate within \(0.6 < SOC_{uc} < 1\).

5. **HESS with GPS Data Processing:**

   The final studied scenario is similar to the previous case with the GPS data being fed to the power optimizer where \(m_{bt}\) and \(m_{uc}\) are adjusted according to the prediction of stop signs and traffic signals.
3.4.3 Simulation Results and discussion

System simulations were performed in MATLAB to investigate the benefits of using the proposed GPS based HESS approach, using the experimental drive-cycle data from the A2B, as discussed in Section 3.2.

The comparison of the HESS with/without GPS data processing, and a battery-only SESS for the measured 3 hour urban drive-cycle is presented in Table 3.3, where the simulated energy consumption based on the EV model was within 0.4% of the measured experimental SESS data from the 3 hour drive-cycle. Due to the low battery ESR in the prototype vehicle, the HESS with GPS data processing consumed 0.2% more energy compared to the SESS with no limits, but this only correspond to less than 1.5% of the overall energy savings. The HESS with GPS data processing decreases the peak battery charge and discharge rates by 76% and 47%, respectively. The peak charging current is reduced by 37%, at the cost of less than 1% reduced overall energy savings, when the HESS power optimizer utilizes the GPS information. The reduction in required peak battery current can translate into a different choice of battery chemistry with higher specific-energy and lower peak power requirement, based on the trade-off discussed in Chapter 1.

<table>
<thead>
<tr>
<th></th>
<th>SESS (No Regen)</th>
<th>SESS ($I_{bt} &gt; -\frac{C_4}{4}$)</th>
<th>SESS (No limits)</th>
<th>HESS without GPS ($I_{bt} &gt; -\frac{C_4}{4}$)</th>
<th>HESS with GPS ($I_{bt} &gt; -\frac{C_4}{4}$)</th>
</tr>
</thead>
<tbody>
<tr>
<td>Net Electric Energy Usage in Drive-Cycle (kWh)</td>
<td>12.01</td>
<td>10.40</td>
<td>10.26</td>
<td>10.27</td>
<td>10.28</td>
</tr>
<tr>
<td>Energy Benefits (%)</td>
<td>0</td>
<td>13.4</td>
<td>14.6</td>
<td>14.5</td>
<td>14.4</td>
</tr>
<tr>
<td>Battery Current Range, $I_{bt,max/min}$ (A)</td>
<td>154.7/0</td>
<td>154.7/-27.5</td>
<td>154.7/-68.9</td>
<td>82.5/-27.0</td>
<td>82.5/-16.8</td>
</tr>
</tbody>
</table>

The partial simulation result for the SESS with no current limits and the HESS with/without GPS are shown in Fig. 3.19(b). It can be seen that the HESS effectively limits the battery current, based on the $I_{bt}$ histogram, as shown in Fig. 3.20(a). Fig. 3.20(b) shows the battery current based on the $SOC_{bt}$. In this work, the charge/discharge limits of the battery current for the HESS system are fixed, but they can be decreased at lower $SOC_{bt}$, improving the battery operating condition.
Figure 3.19: (a) Measured vehicle speed, (b) real-time GPS stop/traffic signal flag during the 3-hour urban drive-cycle. (c) Simulated $I_{bt}$ for SESS with no limit (blue), HESS without GPS (red) and HESS with GPS (green). (d) Simulated U-cap SOC for HESS without GPS (red) and HESS with GPS (green).
Figure 3.20: (a) Simulated battery current histogram and (b) simulated battery current versus \( SOC_{bt} \) for the 3 hour urban drive-cycle.
3.5 Experimental Results of the DC-DC converter

The converter efficiency as a function of output current for multiple input voltages is shown in Fig. 3.21. The output power was limited to 4.8 kW during these measurements due to the 5 kW maximum power limit of the Electronic Load (e-load) available in our laboratory. As shown in Fig. 3.21, 1 phase operation has higher efficiency at light loads. However, at higher output powers, the 2 phase operation provides higher efficiency. Utilizing the optimal number of phases at each operating points results in:

- Efficiency over 90% for operating points above 1 kW,
- Almost flat efficiency curve for operating points above 2 kW, with efficiency over 93%,
- Peak efficiency of 96% at maximum input voltage.

The high converter losses are due to significant inductor core losses associated with high-ripple and high-frequency operation. The ripple can reach up to 45 A peak to peak in the worst case condition, i.e. maximum input and output voltages. The core losses could not have been estimated prior to the converter implementation, due to the limited information available through the inductor’s datasheet.

The power optimizer automatically avoids the low efficiency operating points of the converter due to its substantial cost of losses. It is interesting to note that higher $V_{uc}$ results in lower efficiency at light loads due to higher current ripple, but results in higher efficiency at mid and heavy load conditions due to smaller conversion ratio.

It should be noted that lower $V_{uc}$ not only reduces the efficiency but also limits the nominal converter power due to inductor saturation, as shown in Fig. 3.22. For instance, at $SOC_{uc}=0.6$, the nominal power of the converter is limited to 17.7 kW which is 60% of the maximum nominal power of the converter. As mentioned earlier, due to some thermal issues and low efficiency, the converter can not reach its nominal power for continuous operations. As a result the following limits are used in our testings:
Figure 3.21: Measured converter efficiency at $V_{bt} = 300$ V with different $V_{uc}$.

- At $V_{uc} = 90$V, the converter can achieve 2.25 kW (9 kW) output power per phase (total) for continuous operation, while it can safely operate at the maximum current with 4.86 kW (19.4 kW) output power per phase (total) for 1 minute.

- At $V_{uc} = 140$V, the converter can only achieve 1 kW (4 kW) output power per phase (total) for continuous operation.

The measured u-cap voltage and current based on a sequence of u-cap current commands for an emulated drive-cycle are shown in Fig. 3.23, where time steps of 10 seconds are chosen to provide visible $V_{uc}$ excursion.

The measured converter step response based on a sequence of u-cap current commands for an emulated drive-cycle is shown in Fig. 3.24. Steps are applied every 250 ms to clearly show the details of the step response in the bi-directional operation. The converter has a 70 ms settling time, and the response is over-damped to avoid saturation in the inductors. The measured u-cap current is shown in Fig. 3.24(a), while Fig. 3.24(b) shows the averaged inductor current logged using the CRIO.
Figure 3.22: Converter nominal power versus u-cap voltage.

Figure 3.23: Measured $V_{uc}$ and $I_{uc}$ for a sequence of u-cap current commands. ch-2: $V_{uc}$, 15 V/div; ch-3: $I_{uc}$, 10 A/div.
Figure 3.24: Converter transient response for reference current steps. Measured converter
(a) Total input current. ch-3: $I_{uc}$, 10 A/div. (b) Logged inductor current using CRIO.
Chapter 3. HESS Design for EVs with RT GPS Based Control Method

3.6 Chapter Summary and Conclusions

In this Chapter, an HESS with a power-mix optimizer, that utilizes the real-time GPS data for load distribution decisions, is developed. The full system modeling of the A2B and the proposed HESS has been developed to provide a powerful tool for system simulations. The benefits of introducing the GPS data processing in the power-mix optimizer of the HESS for the A2B was then investigated. It is shown that utilizing the GPS data in the power optimizer can reduce the battery charging peak current by 37%, without significantly compromising the system efficiency and without substantially increasing the cost, since the GPS is already available in the EV. The main advantage of the HESS aside from increasing the battery lifetime by reducing the peak currents, is to allow the use of batteries optimized for high specific-energy, while the peak power demands are met by the u-cap.

The 30 kW bi-directional dc-dc converter used in the HESS was tested and lab experimental result of the converter shows stable closed-loop response in the presence of reference steps. The converter has an over-damped characteristics with a 70 ms settling time for steps in the reference current. The converter efficiency is greater than 93% for load powers higher than 2 kW. The high converter losses are due to significant inductor core losses associated with the high-ripple and high-frequency operation. Due to some thermal management issues related with the core of the inductors, we were unable to test the converter at the full power, which is 7.5 kW per phase. Therefore, the presently used inductors need to be replaced with new parts suitable for high ripple, high frequency operations. The integration of the proposed HESS in the A2B for further verification of the proposed method are considered for future work.
References


Chapter 4

Conclusions

4.1 Thesis Summary and Contributions

The focus of this work is to address one of the main challenges in the mass adoption of EVs, namely implementing low-cost, high energy and high power-density storage system that performs reliably for 10-15 years. The ultracapacitor/battery based HESS benefits in LEVs and EVs are investigated and further improved in this thesis.

In Chapter 2, a digitally controlled bi-directional half-bridge converter is designed specifically to meet the requirements of the battery interfacing dc-dc converter of a LEV based HESS. This architecture was used for its low cost and high efficiency specifically when operated with soft-switching. The novel contributions of this chapter include:

- Demonstrating the concept of variable-frequency quasi-resonant operation to maintain a constant dead-time and a near constant inductor valley current in the mid-load range,

- Showing the superior efficiency of the VFQR scheme compared to the FFQR operation,

- Improving the dead-time adjustment accuracy with the dead-time extension tech-
nique, using the proposed auxiliary circuit without significantly deteriorating the efficiency,

- Combining fixed frequency and PFM operating modes with the VFQR mode to obtain a tri-mode bi-directional converter that operates with high efficiency over a wide load range.

The prototype converter achieves the measured peak efficiency of 97.5% and maintains the efficiency of 88% over the full load range. The benefits of VFQR mode over the FFQR mode are shown with efficiency measurements and infrared images. Measured dead-times are effectively doubled, using the proposed scheme at the heavy load conditions during VFQR mode, to provide dead-time adjustment accuracy.

Chapter 3 is devoted to a HESS design for a Canadian prototype EV, known as A2B. The novel contributions of this chapter include:

- Proposing the concept of utilizing real-time GPS data processing in the power-mix optimizer. The current location of the vehicle is processed to obtain some information about upcoming stop signs and traffic signals based on a preloaded map of a city.

- Proposing a computationally simple power-mix optimizer, that utilizes the real-time GPS data. The proposed scheme adjusts the cost function multipliers based on the GPS information. This simplified control architecture can be done on a reasonably sized target processor. In this project an 800 MHz processor inside a 9025 CompactRIO (CRIIO) module is used.

The following verification steps are also covered in this chapter:

- Full system model of the vehicle including the mechanical and electrical subsystems are derived to provide a base for the system level simulations,
• System level simulations are provided for different scenarios to investigate the benefits of the proposed approach,

• A bi-directional dc-dc converter is designed and implemented to integrate the HESS into the A2B and verify the simulation results with experimental measurements.

System level simulations shows that utilizing the GPS data can further reduce the battery charging peak current without significantly compromising the system efficiency and without substantially increasing the cost, since the GPS is already available in the EV.

The HESS with GPS data processing reduces the peak battery charge and discharge rates by 76% and 47% compared to the SESS with no limits, respectively. However, due to the low battery ESR in the A2B, the energy saving with the SESS with no limits exceeds that of the HESS with GPS data processing by 0.2%. It should be noted that, this difference is less than 1.5% of the overall energy savings. The HESS with GPS data processing also reduces the peak battery charging current by 37% compared to the HESS design without the GPS data processing.

The half-bridge architecture is used for the 30 kW bi-directional dc-dc converter of the proposed HESS. This architecture was chosen due to its low cost and high efficiency. Four phase interleaved topology with phase shedding is utilized to provide high efficiency over a wide load range. Lab experimental results of the converter shows stable closed-loop response in the presence of reference steps. The converter shows an over-damped characteristics and a 70 ms settling time for steps in the reference current. Bi-directional operation of the converter was tested using the u-cap modules clearly showing the charge and discharge events. The present converter design faces a thermal issue in the inductors, due to the significant core losses which could not have been estimated with the limited datasheet information prior to the implementation. The inductors need to be replaced with new parts, with high-ripple high-frequency capabilities, prior to integration of the HESS into the A2B.
As discussed in chapter 1, there is a trade-off between the specific-energy and specific-power within the battery technologies. Therefore, the main advantage of the HESS presented in Chapter 3, aside from increasing the battery lifetime by reducing the peak currents, is to allow the use of batteries optimized for high specific-energy, while the peak power demands are met by the u-cap.

The research work presented in Chapter 2 is published as a conference publication [1], while the content of Chapter 3 has been submitted for future publication [2].

4.2 Future Work

Based on the presented research work in this thesis, the following suggestions are provided for further explorations:

- The dead-time extension technique using the proposed auxiliary circuit can be applied to other quasi-resonant converter topologies. The trade-off between the increased ZVS accuracy versus efficiency reduction should be studied in these topologies.

- Altitude information of the GPS data can be used for determining the road inclination and declinations and therefore providing a better load calculations. However, this can further complicate the calculations that should be done in real-time.

- Real-time GPS data processing can be used in conjunction with more complex power-mix optimizers, such as the one proposed in our prior work [3]. The GPS data can be used to adjust the probability of the pathes and enhance the overall prediction accuracy and system performance.

- Experimentally quantifying the capacity fade of batteries subjected to different dynamic current profiles can be done to further estimate the long term benefits of the HESS.
• The research work presented in Chapter 3, needs to be completed by integrating the designed HESS into the A2B and verifying the results of the simulations.
References

