

Design and Optimization of a 98%-Efficiency On-Board Level-2 Battery Charger Using E-Mode GaN HEMTs for Electric Vehicles

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ABSTRACT

Most of the present EV on-board chargers utilize a three-stage design, e.g., AC/DC rectifier, DC to high-frequency AC inverter, and AC to DC rectifier, which limits the wall-to-battery efficiency to ~94%. To further increase the efficiency and power density, a matrix converter is an excellent candidate directly converting grid AC to high-frequency AC thereby saves one stage. However, its control complexity and the high cost of building the back-to-back switches are barriers its acceptance. Instead, this paper adopts the 650V E-mode GaN HEMTs to build a level-2 on-board charger using the indirect matrix topology. The input voltage is 80~260VAC, the battery voltage is 200~500VDC and the rated power is 7.2kW. Variable switching frequency is combined with phase-shift control to realize the zero-voltage switching. To further increase the system efficiency, four GaN HEMTs are paralleled to form one switching module with a novel gate-drive technology. The overall system efficiency is ~98% and the power density is 3.3kW/L.

Keywords: Gallium Nitride, High Electron Mobility Transistor, Battery Charger, Electric Vehicle, Matrix Converter, Wide-bandgap Semiconductor, Zero Voltage Switching

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I. INTRODUCTION

As the global vehicle fleet transitions to include more electric vehicles (EVs), the demand for a high-efficiency charging technology increases. Most present battery chargers adopt three stages, i.e., an AC/DC part to convert the grid voltage to DC voltage, a DC/AC part to transform the DC to a high-frequency AC (hundreds of kHz) on the primary side of the transformer, and another AC/DC part to rectify the induced high-frequency AC to DC for the battery charging. Shown in Fig. 1 is a typical on-board battery charger using the half-bridge resonant topology.

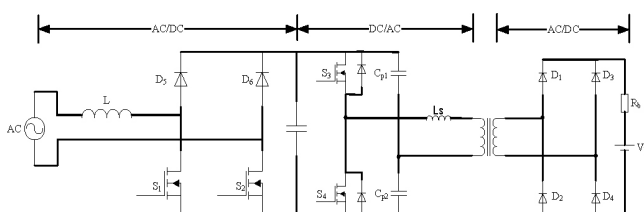


Fig.1. A Conventional Three-stage Isolated Charger

The grid-side AC/DC employs the boost-type power factor correction (PFC) control. L_s , C_{p1} and C_{p2} form the resonance circuit to realize zero voltage switching (ZVS). Here the transformer is required to provide the necessary galvanic isolation. If it is assumed that the PFC and the DC/AC part both have 98% efficiency, and the battery-side AC/DC and the transformer both have 99% efficiency, then the overall system efficiency will be $98\% \times 98\% \times 99\% \times 99\% = 94.1\%$.

To reach higher levels of efficiency, a single phase single-stage bidirectional isolated ZVS AC-DC converter is proposed in [1], shown as Fig.2 It is essentially a matrix converter (MC) adopting back-to-back connected switches, not available in the market today. An MC directly converts the grid voltage to high-frequency AC, saving one stage as compared to Fig.1 (PFC+DC/AC) and eliminating the DC-bus capacitor. However, control complexity and forming back-to-back switches become the major challenges.-

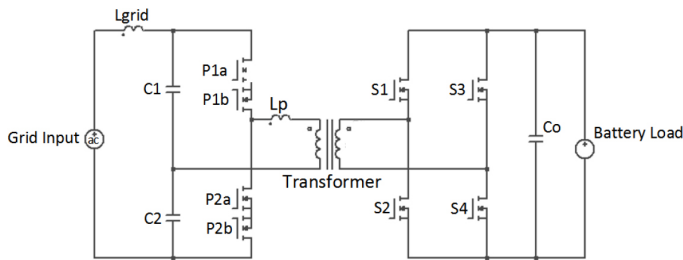


Fig. 2. Half-bridge Isolated ZVS Matrix Converter

Adopting a similar control algorithm like [1], this paper utilizes an indirect matrix converter topology [2] to realize a 7.2kW on-board charger, compatible to both 110VAC and 208VAC input as shown in Fig. 3. Different from Fig. 1, the front-end rectifier stage here ($M_1 \sim M_4$) is operated at 60Hz, which simply converts the sinusoidal waveform into a 120Hz, single-polarity DC voltage. Therefore, the efficiency of the front-end rectifier is extremely high.

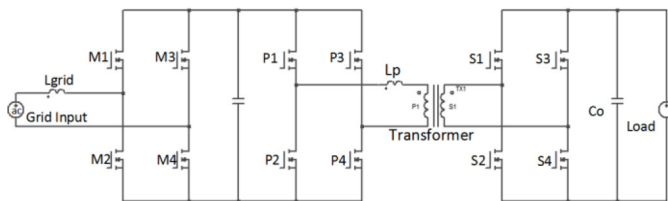


Fig. 3. Indirect Matrix Converter using GaN HEMTs

This topology also has no back-to-back switches. For switches ($P_1 \sim P_4$, $S_1 \sim S_4$), GaN HEMTs are excellent candidates due to their ultra-fast switching transitions and ultra-low gate-drive power as compared to Si MOSFETs [3,4,5,6]. At the present time, the cost of GaN HEMTs is ~ 5 times greater than the cost of the Si MOSFETs. However, compared to conventional Si devices, GaN HEMTs have much higher switching frequencies and higher efficiency, which allows for a much higher power density and a lower cost of the cooling the system and other related passive components. Section II will detail the system control algorithm. Section III, discusses the critical design criteria particularly as it relates to challenges and opportunities with the adoption of GaN HEMT, e.g., dead-band, parallel connection. Section IV, details the experimental results and Section V, provides the conclusion.

II. CLOSE- LOOP CONTROL

As shown in Fig. 4, the indirect MC based, on-board, level 2 battery charger consists of one rectifier stage, a dual-active-bridge (DAB) stage and a buck-type active filter. Different from the conventional MC, a 60Hz, full bridge is employed in order to rectify the sinusoidal grid voltage. Therefore the regular Si MOSFET with small R_{DSon} can be adopted without considering the switching performance, which may lead to an overall lower system cost. The power factor correction (PFC) control is achieved in the DAB stage, while the instantaneous power transferred from the primary side to the secondary side is controlled through the phase shift and switching frequency. In this paper, 650V GaN HEMTs provided by GaN Systems Inc are employed to increase the switching frequency (150kHz $\sim$ 500kHz) and power density.

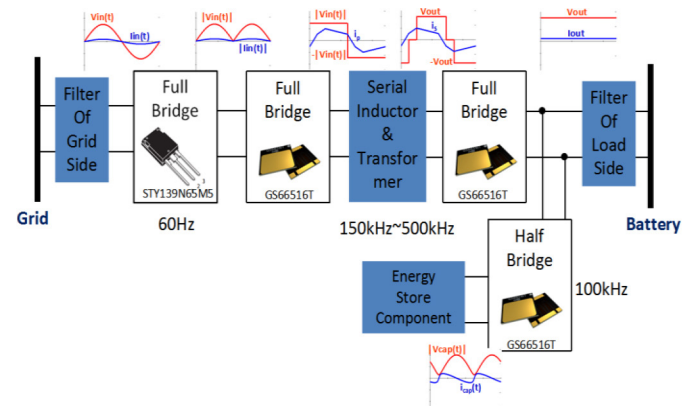


Fig. 4. Indirect Matrix Converter Based On-Board Level 2 Battery Charger

Previous literature exists on the DAB control strategies [7,8,9,10,11], most of which adopts fixed switching frequency and constant DC input methodologies. In this paper, the DAB converter is controlled by both the phase-shift and the switching frequency. Shown in Fig. 5, the transformer primary-side voltage is v_p , secondary-side voltage is v_s , and the transformer primary current is i_L . Different from the conventional DAB control, only the primary side voltage v_p is a square wave in this circuit. The two legs at the secondary side have some phase-shift between each other. During one switching period, the gate signals of each device are shown in Table 1. i_L during one switching period is given by Eqn. (1). Note, in this control algorithm, there are three control variables $g(t)$, $w(t)$ and switching frequency $f_s(t)$, these are the major difference from the previous control algorithm. $w(t)$ is decided by $g(t)$, shown as Eqn.(2). To achieve ZVS, the control variable $g(t)$ should be larger than $g_{min}(t)$ given by Eqn. (3), where $C_{eq,p}$, $C_{eq,s}$ are equivalent output capacitances of the switch on the primary side and secondary side, respectively. Detailed control algorithm could refer to [1].

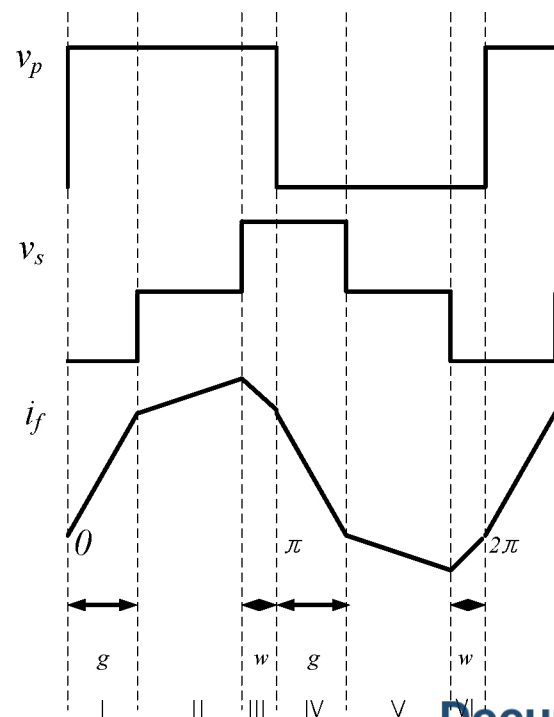


Fig. 5. Waveforms in one switching period

$$i_L(t) = \begin{cases} \frac{v_m + n \cdot V_b}{L} t - \frac{\frac{v_m}{2} + (g-w) \cdot n \cdot V_b}{2 \cdot f_s \cdot L} & 0 \leq t < g \cdot T_s \\ \frac{v_m}{L} (t - g \cdot T_s) + \frac{(2 \cdot g - \frac{1}{2}) v_m + (g+w) \cdot n \cdot V_b}{2 \cdot f_s \cdot L} & g \cdot T_s \leq t < (\frac{1}{2} - w) \cdot T_s \\ \frac{v_m - n \cdot V_b}{L} (t - (\frac{1}{2} - w) \cdot T_s) + \frac{(\frac{1}{2} - 2 \cdot w) v_m + (g+w) \cdot n \cdot V_b}{2 \cdot f_s \cdot L} & (\frac{1}{2} - w) \cdot T_s \leq t < \frac{1}{2} T_s \\ \frac{-v_m - n \cdot V_b}{L} (t - \frac{1}{2} T_s) + \frac{\frac{v_m}{2} + (g-w) \cdot n \cdot V_b}{2 \cdot f_s \cdot L} & \frac{1}{2} \cdot T_s \leq t < (\frac{1}{2} + w) \cdot T_s \\ \frac{-v_m}{L} (t - (\frac{1}{2} + w) \cdot T_s) - \frac{(2 \cdot g - \frac{1}{2}) v_m + (g+w) \cdot n \cdot V_b}{2 \cdot f_s \cdot L} & (\frac{1}{2} + w) \cdot T_s \leq t < (\frac{1}{2} + g) \cdot T_s \\ \frac{-v_m + n \cdot V_b}{L} (t - (\frac{1}{2} + g) \cdot T_s) - \frac{(\frac{1}{2} - 2 \cdot w) v_m + (g+w) \cdot n \cdot V_b}{2 \cdot f_s \cdot L} & (\frac{1}{2} + g) \cdot T_s \leq t < T_s \end{cases} \quad (1)$$

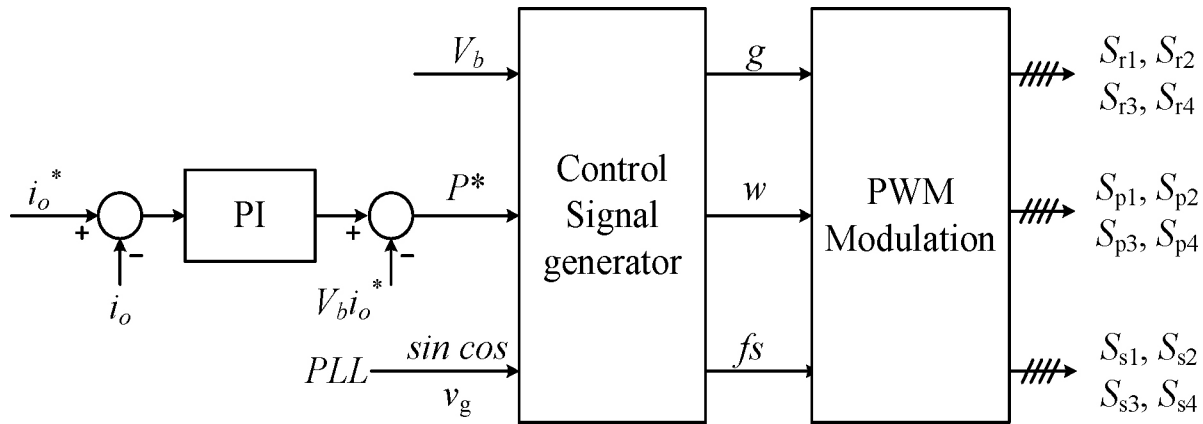


Fig. 6. Close loop control diagram.

Table 1. Gate signals of each device during one switching period

State	Sp1, Sp4	Sp2, Sp3	Ss1, Ss4	Ss2, Ss3
I	0	1	1	0
II	0	1	0	0
III	0	1	0	1
IV	1	0	0	1
V	1	0	0	0
VI	1	0	1	0

$$w(t) = \frac{v_m(t) \cdot (1 - 2 \cdot g(t))}{2 \cdot n \cdot V_b} \quad (2)$$

$$g_{\min}(t) = \frac{4 \cdot f_s \cdot L \cdot I_s}{2 \cdot v_m + 2 \cdot n \cdot V_b} \quad I_s = \max \left\{ \frac{v_m}{\sqrt{C_{eq,p} \cdot L}}, \frac{V_b}{\sqrt{C_{eq,s} \cdot L}} \right\} \quad (3)$$

The instantaneous power in one switching period transferred from the AC to DC side is given by Eqn. (4). Considering the limit of the maximum instantaneous power, the maximum $g(t)$ is given by Eqn. (5) [1].

$$p(t) = \frac{v_m \cdot n \cdot V_b (g - 2 \cdot g^2 + w - 2 \cdot w^2)}{2 \cdot f_s \cdot L} \quad (4)$$

$$g_{\max}(t) = \frac{v_m^2 - \frac{1}{2} v_m \cdot n \cdot V_b + \frac{1}{2} \cdot n^2 \cdot V_b^2}{2 \cdot (v_m^2 + n^2 \cdot V_b^2)}$$