A Novel Current Angle Control Scheme in a Current Source Inverter Fed Surface-mounted Permanent Magnet Synchronous Motor Drive

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Abstract—This paper describes a novel speed control scheme to operate a current source inverter (CSI) driven surface-mounted permanent magnet synchronous machine (SPMSM) for hybrid electric vehicles (HEVs) applications. The idea is to use the angle of the current vector to regulate the rotor speed while keeping the two dc-dc converter power switches on all the time to boost system efficiency. The effectiveness of the proposed scheme was verified with a 3 kW CSI-SPMSM drive prototype.

Keywords—automotive applications; pulse width modulation current source inverter; current angle control.

I. INTRODUCTION

As more and more power electronics systems (PESs) are packaged into the latest electric vehicles (EVs) and hybrid electric vehicles (HEVs), it is very important to keep the PESs at low cost and small volume while maintaining high reliability. The dc bus capacitors in the widely used voltage source inverter (VSI) is one of the weak links in the PESs in EV/HEVs because of their large volume, high cost, and undesirable deterioration characteristic of ripple current handling capability with temperature increase. Some of the earlier works on minimization of the dc bus capacitors are aimed for ac-de-ac converter-inverter systems and thus may not be applicable to HEVs [1–4]. Further, in some approaches an extra current sensor or control loop was added to the converter-inverter system, increasing the cost and complexity.

Some recent works on reducing dc bus capacitor ripple current of VSIs were reported in [5–8]. One approach aiming at EV/HEV applications utilized the switching synchronization between the buck-boost converter and VSI to reduce the dc bus capacitor current [5–6] and thus it is only applicable to VSI systems that have a front-end buck/boost converter. It was also shown that rather limited reduction in capacitor current was achieved. Others used a complicated pulse width modulation (PWM) scheme, in which zero voltage vectors were not used in certain conditions, to reduce capacitor current [7]. While this method can be applied to VSIs with or without a buck/boost front end, the experimental results showed very limited improvement in capacitor current. In [8], a three phase full bridge inverter was used to gain more control freedom in order to reduce capacitor current. A significant reduction was achieved at the expense of a doubled number of power switches. It might be noted that all these work are still based on the VSI.

Before the now widely-used VSI, the current source inverter (CSI) had been investigated intensively and used in many industrial applications including motor drives. Previous work on CSIs was mainly focused on large power industrial drive systems [9–18] and demonstrated the advantages of the CSI over the VSI such as elimination of dc bus capacitors, tolerance of phase leg short-circuit conditions, and better output voltage and current waveforms. All these advantages make the CSI an attractive candidate for EV/HEV traction drive applications.

In the HEV PES shown in Fig. 1 and described in [17–18], a CSI was used with a dc-dc converter to drive a permanent magnet synchronous motor (PMSM). The dc-dc converter was employed to regulate the inductor current to the desired value, and the CSI to generate the commanded d-axis and q-axis currents in the motor according to the speed controller. The dc-dc converter was also used to enable the CSI to charge the battery during regenerative braking. The total amount of capacitance of the input dc and three output ac filter capacitors for a 55 kW CSI was reduced to 200 µF from 1000 µF for a typical 55 kW VSI for HEV applications.

![Figure 1. System diagram of the CSI fed PMSM drive for HEV applications.](image)

In [19], a novel current-fed quasi-Z-source inverter was proposed and tested, in which a passive network of two inductors, two capacitors, and a diode was used to replace the dc-dc converter [17–18]. The inverter is able to achieve bi-directional power flow with output voltage boost functionality. However, the added weight and volume of the passive components may have a substantial impact on the inverter power density and specific power. In addition, the voltage

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boost factor is limited to 2. For automobile applications, a high voltage boost ratio is desirable, especially for permanent magnet motors, because it can significantly reduce the number of battery modules, optimize the PES efficiency at different speeds and state of charge of the batteries, and increase the motor output power at high speeds [20].

In [18], the voltage boost capability of the CSI PES was tested experimentally. The CSI PES can increase the output voltage owing to its inherent boost capability attained through the use of the shoot-through mode. However, it was found that as the input dc (battery) voltage dropped the efficiency of the CSI PES fell, which must be overcome to maximize vehicle mileage. In this paper, a novel current angle control scheme is proposed to improve the efficiency of the CSI PES in the voltage boost mode. Modeling and experimental results is shown to verify the scheme.

This paper is arranged as follows: session II discusses the current control in the voltage boost mode with the traditional and proposed schemes; session III shows simulation results; session IV shows experimental results and a comparison with the traditional scheme; session V concludes the paper.

II. BOOST MODE CURRENT CONTROL IN A CSI-PES PM DRIVE SYSTEM

A. CSI Model

The system model of the CSI is expressed as

\[
\begin{align*}
    i_a &= S_{11} * I_{dc} - S_{12} * I_{dc} = m_a I_{dc} \\
    i_b &= S_{21} * I_{dc} - S_{22} * I_{dc} = m_b I_{dc} \\
    i_c &= S_{31} * I_{dc} - S_{32} * I_{dc} = m_c I_{dc}
\end{align*}
\]

(1)

where \( m_a, m_b, \) and \( m_c \) are the modulation signals for the three phase output currents, \( i_a, i_b, \) and \( i_c; I_{dc} \) is the dc link current; and \( S_{ij} (i=1,2,3, j=1,2,3) \) represents the switching function of each device. By choosing different switching functions, the three phase currents can be controlled accordingly. Because the three phase currents contain capacitor current components, which do not go into the motor, these capacitor currents need to be compensated so that accurate desired motor currents are supplied.

The inverter input voltage \( V_{in} \) can be calculated from the switching functions and the ac capacitor voltages by

\[
V_{in} = V_{an}(S_{11} - S_{12}) + V_{bn}(S_{21} - S_{22}) + V_{cn}(S_{31} - S_{32})
\]

\[= \frac{3}{2}(m_qv_{cq} - m_dx_{cd}) \]

(2)

where \( m_q \) and \( m_d \) are the \( q- \) and \( d- \) axis modulation indices for the three phase currents projected to the \( d-q \) frame; \( v_{cq} \) and \( v_{cd} \) are the \( q- \) and \( d- \) axis components of the capacitor voltages; and \( V_{an}, V_{bn}, \) and \( V_{cn} \) are the phase-to-neutral voltages of the CSI [5].

When a CSI generates a zero current vector, there is no current in the inverter output terminals to the filter capacitors. The two switches of a phase leg are turned on at the same time (referred as the shoot-through mode), which generates a boost effect for the output voltage due to the small dc link inductor, \( L \). The voltage boosting function by \( m \) reduction can be depicted in Eqs. (3–4) and details can be found in [17–18].

\[
m = \frac{i_{peak}}{I_{dc}}
\]

(3)

\[
V_{Out} = \frac{\pi}{3m} V_{dc}
\]

(4)

The aforementioned \( m \)-adjusting boost control scheme was verified experimentally [14]. The system diagram is shown in Fig. 2. (4) indicates that, when the dc voltage drops, to maintain the same output voltage, \( m \) must be reduced. In the meantime, (3) forces the inductor current to increase, leading to higher conduction and switching losses in the PES, which is not desirable. For example, it was found experimentally that there was an 8% reduction of the PES efficiency as the dc bus voltage dropped from 135 V to 100 V at 2,000 rpm with a load torque of 15 Nm. All these extra losses in the boost mode generate negative impact on the cooling system and need to be mitigated. The key is to limit the inductor current. It would be ideal to keep \( m = 1 \) in this situation to reduce the dc bus current.

B. Proposed Current Angle Control Method

In steady state, the \( d \) and \( q \) axis voltages, \( v_{ds} \) and \( v_{qs} \), and currents, \( i_{ds} \) and \( i_{qs} \), can be related by Eqs. (5–6), where \( \omega_c \) is the rotor speed, \( L_d \) and \( L_q \) are \( d \) and \( q \) axis inductance, \( v_{in} \) is the input voltage, and \( \theta \) is the angle of the current vector.

\[
\begin{align*}
    v_{qs} &= r_s i_{qs} + \omega_c \lambda_m + \omega_L L_{ds} i_{ds} \\
    v_{ds} &= r_s i_{ds} - \omega_c L_{qs} i_{qs}
\end{align*}
\]

(5)

\[
 v_{in} = \frac{3}{2} m (\sin \theta v_{qs} + \cos \theta v_{ds})
\]

(6)

In a large surface-mounted permanent magnet synchronous motor, we should have equal \( d- \) and \( q- \) axis inductances, i.e., \( L_{ds} = L_{qs} \). The stator resistance can be neglected (\( r_s \approx 0 \)). Equation (6) can be simplified as

\[
 v_{in} = \frac{3}{2} m \omega_c \lambda_m \sin \theta .
\]

(7)

From (7), it is seen that for a given dc bus voltage the speed of the SPMSM is only determined by the current angle \( \theta \), rotor flux \( \lambda_m \), and modulation index \( m \). In other words, the current angle \( \theta \) can be used to regulate the rotor speed, if the rotor flux \( \lambda_m \) and modulation index \( m \) are fixed. This is the basis of the proposed current angle control method. It may be noted that as
Under the proposed angle control scheme, the dc-dc converter is kept on all the time \((m_{dc} = 1)\) and the modulation index of the CSI is set to 1.0 (maximum) to minimize the dc bus current, so that higher efficiencies can be achieved. This is due to the fact that a smaller dc bus current produces smaller conduction and switching losses in all the switches in the CSI. The speed regulation of the motor is achieved by regulating the current vector angle directly, as indicated in the control block diagram in Fig. 3. It has two loops: an outer loop to regulate the rotor speed and an inner loop to generate PWM signals for the CSI according to the rotor position, modulation index, \(m\), and dc current. Several PWM schemes can be used for the CSI [12]; for this study, we selected the sinusoidal PWM because of its simplicity.

\[
\theta = \tan^{-1}\left(\frac{q}{d}\right)
\]

Fig. 3 shows steady-state waveforms at 1500 rev/min with a load torque of 20 Nm. The output voltage and current are quite sinusoidal. The dc inductor current is 37 A. Fig. 6 shows the steady-state waveforms at 2000 rev/min with 15 Nm load. The power supply internal resistance and reactance were ignored in this simulation.

In Fig. 3, only one PI controller was used to control the rotor speed and consequently just one rate limiter is needed to constrain the change in phase angle for better system stability.

Fig. 4 shows waveforms of rotor speed, dc inductor current, output line current to the motor and angle of the current vector at motor starting. The controller changes from the traditional control mode to the proposed control mode at 0.17 Sec. The motor settles at a speed of 1500 rev/min after 0.6 sec. It is seen that there is inrush current around the switching-point. However, the amplitude is not very high and the duration is short (about 0.2 sec.). This figure shows that the proposed scheme can regulate the rotor speed by the angle controller in high speed range.

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The advantages of the proposed scheme are smaller dc inductor currents and zero switching loss in the dc-dc interface circuits. The conduction and switching losses in the CSI are also reduced.

III. MODELLING RESULTS

Based on the control scheme discussed in the previous section, a model was built and simulated using the commercial software, PSIM to test the CSI-PES at different speeds and load levels. The sampling frequency for both the dc-dc converter and CSI is set at 7.5 kHz while 1.5 kHz is chosen for the speed loop. The dc voltage was set to 100 V. The motor is an 8.2 kW PMSM with equal \(d\)- and \(q\)-axis inductance. The parameters of the PMSM are given in the Appendix.

The controller is a combination of Fig. 2 and Fig. 3. At speeds lower than the base speed, the control diagram in Fig. 2 is used and at higher speeds, the diagram in Fig. 3 is used. In Fig. 2, two proportional-integral (PI) controllers are used, one for control of the outer speed loop and the other for control of the inner current loop. As explained in [18], two rate limiters are used to make the changes in the reference current and phase angle not too fast for better system stability.
Figure 5. Simulated steady state waveforms at 1500 rev/min with 20 Nm load-modeling results. From top, input dc voltage, two line-to-line voltages, dc inductor current and two line currents.

Figure 6. Simulated steady state waveforms at 2000 rev/min with 15 Nm load. From top, input dc voltage (100 V/div), two line-to-line voltages (200 V/div), dc inductor current (50 A/div) and two line currents (50 A/div).

IV. EXPERIMENTAL RESULTS

A CSI-based PES was constructed for experimental verification. It mainly consists of a liquid-cooled cold plate, an insulated gate bipolar transistors (IGBT) dc-dc converter, an inductor of 330 µH, a CSI bridge with 6 IGBTs and three ac film capacitors of 30 µF each. All the power modules are mounted on the cold plate. Two types of IGBTs were used. Two normal IGBTs were used in the dc-dc interface. The reverse-blocking IGBT (RB-IGBT) was not available at the time of hardware design. Therefore, six custom-made IGBTs were used in the CSI, each of which was constructed with a normal IGBT and an extra diode in series. The addition of the extra diodes in series in the custom-made IGBT modules to achieve reverse-blocking capability for the IGBTs results in higher conduction losses than that would result from the use of the RB-IGBTs. The PES hardware is shown in Fig. 7.

A Texas Instruments digital signal processor (DSP), TMS320F2812, and a Xilinx complex programmable logic device (CPLD) chip were selected as the controller and a TMS320F2812 board designed in house was used to carry out the real-time control algorithms and generate gating signals for all the IGBTs. The diagram of the DSP card can be seen in Fig. 8. Armed with the CPLD and DSP chips, the board was able to carry out sophisticated control algorithms and generate advanced PWM control signals for the CSI and dc-dc converter.

Figure 8. Block diagram of the DSP controller board.

An 8.2 kW SPMSM was used as the load to the PES with an absolute encoder to obtain the rotor position information. An adjustable dc power supply was used to power the PES. A dynamometer with a continuous power of 3 kW and a short-time maximum power of 5 kW was used to load the SPMSM with different load torques at various speeds.

The system was tested with a fixed input voltage of 100V. The switching and sampling frequencies are exactly the same as those used in modeling. A PZ4000 power meter was used to measure the efficiency of the PES.

A. Representative waveforms at 1500 and 2000 rev/min with the proposed control method

Fig. 9 shows experimental results at 1500 rev/min with 20 Nm with the proposed angle control method. It is seen that the output voltages and currents are sinusoidal. The dc inductor current is only 39 A. The output voltage is 97 Vrms and the output current is 21 Arms. All the current and voltage waveforms are in close agreement with those shown in the simulation results.

Fig. 10 shows experimental results at 2000 rev/min with 15 Nm. The output voltages and currents are still quite sinusoidal. The inductor current is about 37 A. The output voltage is 97 Vrms and the output current is 21 Arms. All the current and voltage waveforms are in close agreement with those shown in the simulation results.

B. Typical waveforms with the traditional control method

In Figs. 11-12, representative waveforms under the traditional control scheme are shown to compare with those in Figs. 9-10 under the proposed angle control method. The d-axis current was set to zero to achieve maximum-torque-per-ampere control. Figs. 9 and 11 have the same operating conditions of speed, input voltage and load torque. It is obvious
that with the proposed control method, the dc inductor current was reduced significantly from 62 A to 39A, the current ripple of proposed control is also smaller. Similarly, in Figs. 10 and 12, the dc inductor current was reduced from 66A (Fig. 12) to 38 A (Fig. 10).

Figure 9. Typical waveforms with 20 Nm load at 1500 rev/min with the proposed control method. From top, input dc voltage (100 V/div), two line-to-line voltages (200 V/div), dc inductor current (50 A/div) and two line currents (50 A/div).

Figure 10. Typical waveforms with 15 Nm load at 2000 rev/min with the proposed control method. From top, input dc voltage (100 V/div), two line-to-line voltages (200 V/div), dc inductor current (50 A/div) and two line currents (50 A/div).

Figure 11. Typical waveforms with 20 Nm load at 1500 rev/min with the traditional control method. From top, input dc voltage (100 V/div), dc inductor current (50 A/div), two line currents (50 A/div) and a line-to-line voltages (200 V/div).

Figure 12. Typical waveforms with 20 Nm load at 1500 rev/min with the traditional control method. From top, input dc voltage (100 V/div), dc inductor current (50 A/div), two line currents (50 A/div) and a line-to-line voltages (200 V/div).

Figure 13. A comparison of the PES efficiencies under the conventional and proposed control schemes is given in Fig. 13. It shows the system efficiency of the dc-dc converter and the CSI, measured with a PZ4000 power meter. There are four curves in the figure. The two dashed lines indicate the CSI PES efficiencies under the conventional m-adjusting control. The efficiency ranged from 82 % (2,000 rpm with 15 Nm) to 86 % (1,500 rpm with 5 Nm). It is seen that with the increase of load torque or rotor speed, the efficiency drops. With the proposed current angle control indicated by the two solid lines, the efficiency is from 91.5 % (1,500 rpm with 5 Nm) to 92.5 % (1,500 rpm with 10 Nm). Moreover, the efficiency with the proposed scheme does not change much with either speed or load level. More importantly, there is a significant increase in efficiency at all load levels and speeds under the proposed scheme, especially at high speeds and heavy load conditions. A maximum efficiency improvement of about 10 percentage points is achieved at 2,000 rpm with 15 Nm load while a

It may be noted that the output voltage of the proposed scheme in Fig. 9 (97 V rms) is smaller than that shown in Fig. 11 (104 V rms), even they have the same speed and load torque. This is also true with Fig. 12 and 10. This is mainly because the proposed scheme generates a field-weakening current which reduces the magnetic field in the motor. This also means if the motor takes the same load torque, the stator winding current will be higher under the proposed scheme, although the increase is not very significant: from 21 A in Fig. 11 to 22A in Fig. 9; from 16.7A in Fig. 12 to 20.9 A in Fig. 10.
minimum efficiency improvement of about 5 percentage points is achieved at 1500 rpm with 5 Nm load.

Fig. 14 shows the drive system efficiency with the traditional and proposed angle control methods. The load torque as output of the drive system was measured by a dynamometer. This includes all the losses in the PES and PM motor. In the figure, the solid lines show the efficiencies with the proposed scheme and the dashed lines for the traditional control scheme. It shows again that the proposed angle control scheme consistently achieves substantially higher system efficiency than the traditional control scheme at all testing conditions.

C. Limitations of the proposed scheme

As shown in the experimental results, the proposed scheme is very effective in boosting the PES efficiency. However, it has some limitations. Firstly, the proposed scheme can only be used at high speeds, as indicated in (7-8). The motor controller has to have different control algorithms for low speed operation. Secondly, it is noted that the stator current is higher with the proposed scheme, which leads to higher copper loss in the stator windings. The increase is, however, not significant in our test results to impact the overall drive system efficiency.

V. CONCLUSIONS

In this paper, a novel current angle control scheme was proposed to significantly improve the system efficiency of a CSI PES. Its effectiveness was verified by modeling and experimental results. An increase of drive system efficiency by as high as 10 percentage points was observed.

APPENDIX

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<th>Table I. Parameters of the SPMSM</th>
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<tr>
<td>Name of the Parameters</td>
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<td>------------------------</td>
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<tr>
<td>Rated output power (Watt)</td>
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<tr>
<td>Rated speed (rev/min)</td>
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<tr>
<td>Rated voltage (Volt, rms)</td>
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<tr>
<td>Rated current (A, rms)</td>
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<tr>
<td>Back EMF constant (V/1000rpm, L-L voltage, 0 to peak)</td>
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<tr>
<td>Pole numbers</td>
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<tr>
<td>Rated torque (Nm)</td>
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<td>Stator resistance (Ω)</td>
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<tr>
<td>Stator inductance (mH)</td>
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REFERENCES


