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Abstract: In the current source inverter (CSI)–fed inductor motor (IM) drive system, DC–bus inductance and AC–filter capacitance have a direct impact on the dynamic response speed, power quality and power density. In addition, due to the addition of filter capacitors, the mathematical model formed by the IM and capacitor is a third–order system, which increases the difficulty of parameter tuning for the control loop. To solve the above problems in the CSI, a DC–bus inductance design method based on current response speed and the ripple of the DC–link current and an AC–filter capacitance design method based on current utilization and filtering characteristics are presented. Then, the analytical expressions between the open–loop cut–off frequency, phase margin and the PI controller parameters of the current loop and speed loop are derived. Finally, an experimental platform is established to validate the proposed method.

Keywords: inductor motor (IM); current source inverter (CSI); parameter design; tuning; control loop

1. Introduction

Inverters can be divided into voltage source inverter (VSI) and current source inverter (CSI) according to the characteristics of the power supply in the DC side [1,2]. For a long time, VSI has been a research hotspot and has been widely used in various AC drive and grid–connected applications [3,4]. However, the current ripple on the DC side of VSI is large, and it can only operate under the buck mode; deadtime is needed to prevent DC power supply from passing through [5]. Compared with VSI, CSI can boost the voltage and operate without short circuit protection [6]. Due to the presence of filter capacitors on the AC side, the voltage of the stator side of the motor has good waveform quality [7], so in many work situations, CSI is more suitable than VSI [8,9].

Although the CSI has the above advantages, due to some inherent defects, the application scope of CSI in motor drive systems is restricted. Instead of a constant current source, DC–link current is generated from a DC–bus inductor with a DC voltage source, and the DC–bus inductor is charged or discharged under different switching states, so the DC–link current of traditional CSI cannot be controlled [10,11]. Meanwhile, the CSI is unable to operate under the buck mode; otherwise, the DC–link current will increase continuously [12]. To achieve the control of DC–link current and a wide range of output voltage of CSI, many novel CSI topologies are proposed. In Refs. [13,14], a PWM current source rectifier is used as the front stage to control the DC–link current, but this topology is only suitable for the occasion of AC–DC–AC. In Ref. [15], a quasi–Z–source current source converter is added into the DC side, but the difficulty of control increases greatly with the addition of too many passive components. In Ref. [11], the bypass IGBT was paralleled on the DC–link inductor through the control of the switch device. Under the buck operation, the continuous increase of the DC–link current is effectively restrained. Compared to the above topologies, in Refs. [16,17], a bidirectional DC chopper is added in the front stage.
of the CSI; this topology is simple and consists of only a pair of switching tubes and diodes, and the DC–link current control can be achieved under the operations of boost, buck, and deceleration by regulating the switching state of the bidirectional DC chopper. Therefore, this paper takes the topology of CSI with bidirectional DC chopper as the research object.

In the AC side of CSI, due to the presence of filter capacitors, the mathematical model formed by the motor and capacitor is a third–order system, which increases the design difficulty of the control system of the inductor motor (IM). In recent years, the vector control technologies for the CSI–fed drive system have been studied by some scholars. In Ref. [13], the references of stator current in the d–q axis are obtained by the controllers of rotor flux and speed, respectively; then, the references of the output current of the inverter bridge in the d–q axis are calculated according to the steady state equation of the IM. However, this control strategy is more sensitive to the accuracy of the IM parameters. In Ref. [18], a small signal mathematical model of a CSI–fed permanent magnet synchronous motor is constructed through the state space averaging method; a speed control system based on duty cycle regulation is proposed, but the steady–state performance of this method is not ideal, and the stator current has obvious harmonic distortion. In Ref. [19], various decoupling methods and the active damping method are presented and compared, and a control structure of voltage inner loop and current outer loop are proposed. In addition to the method of decoupling and controlling, DC–bus inductance and AC–filter capacitance also have a direct impact on the dynamic response speed, power quality, and power density [20]; however, research on parameter design methods for DC–bus inductance and AC–filter capacitance is rarely reported.

In this paper, the design principles of DC–bus inductance and AC–filter capacitance are firstly studied. Then, to reduce the number of the control loops, a vector control technology without a voltage loop is proposed, and the analytical expressions of the control parameters are derived. Finally, under various conditions, experimental validation of the proposed method is carried out.

2. Topology and Mathematical Model

2.1. Topology of CSI–Fed IM Drive System with Bidirectional DC Chopper

To realize the decoupling control of the DC–link current and IM, a bidirectional DC chopper is added between the DC voltage source and DC–link inductor; the topology is shown in Figure 1. In the figure, $u_{dc}$ is the DC input voltage; $i_{Lc}$ is the current of DC–link inductor $L_c$; $S_1$–$S_6$ are six switching tubes of the inverter bridge; $D_1$–$D_6$ are the diodes connected in series with $S_1$–$S_6$; $C_s$ and $C_f$ are the third–phase filter capacitors in the AC side; $i_{sa}$, $i_{sb}$ and $i_{sc}$ are the three–phase output currents of the inverter bridge; $u_{sa}$, $u_{sb}$ and $u_{sc}$ are the stator currents of the IM; $u_{sa}$, $u_{sb}$ and $u_{sc}$ are the three–phase voltages of the filter capacitors. The bidirectional DC chopper is composed of the diodes $D_1$, $D_6$ and the switching tubes $S_1$, $S_6$.

![Figure 1. The topology of CSI-fed IM drive system with bidirectional DC chopper.](image-url)
When the control method proposed in Ref. [17] is adopted, the circuit of DC input side can be regarded as a controlled current source. In the process of mathematical modeling and control system design for the AC side of CSI, the influence on the DC–link current can be ignored, and the topology of the CSI–fed IM drive system with bidirectional DC chopper can be simplified as seen in the circuit shown in Figure 2.

\[ \begin{align*}
C_i & = C_a + C_b + C_c + C_s \\
C_s & = C_m + C_t
\end{align*} \]

**Figure 2.** The equivalent circuit of CSI–fed IM drive system with bidirectional DC chopper.

### 2.2. Mathematical Model of Filter Capacitors and IM

In the synchronous rotating frame (d–q axis) based on rotor flux orientation, the IM voltage equation can be expressed as

\[
\begin{align*}
\frac{d}{dt}i_d & = \frac{u_d}{L_d} - \frac{1}{\sigma} \left( L_{m} \frac{d}{dt} \psi_r + \psi_s \right) \\
\frac{d}{dt}i_q & = \frac{u_q}{L_q} + \frac{1}{\sigma} \left( L_{m} \frac{d}{dt} \psi_r + \psi_s \right)
\end{align*}
\]

where \( i_d \) and \( i_q \) are the stator currents in the d–q axis; \( u_d \) and \( u_q \) are the voltages of filter capacitor in the d–q axis; \( \psi_r \) is the rotor flux; \( \sigma \), \( R_o \), \( L_m \), \( L_s \) and \( w_s \) are the flux leakage coefficient, stator resistor, mutual inductance, rotor inductance and synchronous angular frequency, respectively.

According to Figure 2, the current equation in three–phase stationary coordinate frame can be expressed as

\[
\begin{align*}
C \frac{d}{dt}u_{as} & = i_at - i_as \\
C \frac{d}{dt}u_{bs} & = i_bt - i_{bs} \\
C \frac{d}{dt}u_{cs} & = i_ct - i_{cs}
\end{align*}
\]

where \( C \) represents the value of filter capacitors. Clarke transform and Park transform are carried out on Formula (3); in turn, the current equation in the d–q axis can be obtained and expressed as

\[
\begin{align*}
C \frac{d}{dt}u_{ds} & = i_{dt} - i_{ds} + \omega_q Cu_{qs} \\
C \frac{d}{dt}u_{qs} & = i_{qt} - i_{qs} - \omega_q Cu_{ds}
\end{align*}
\]

where \( i_{ds} \) and \( i_{qs} \) are the output current of the inverter bridge in the d–q axis. The rotor flux equation in the d–q axis is expressed as

\[
T_r \frac{d}{dt} \psi_r + \psi_r = L_m i_{ds}
\]

where \( \psi_r \) and \( L_m \) are the rotor flux and excitation inductor of IM, respectively, and \( T_r \) is the rotor time constant. The equations of torque and speed are expressed as
\[
\begin{align*}
T_c &= n_p \frac{L_m}{L_r} \psi_r i_{qs} \\
J \frac{d\omega_v}{dt} &= T_c - T_l
\end{align*}
\]

(5)

where \( J \) is the moment of inertia, \( \omega_v \) is the mechanical speed, and \( T_c \) and \( T_l \) are the electromagnetic torque and load torque, respectively.

In summary, the mathematical model of filter capacitors and IM in the AC side of CSI can be represented by Equation (1) to Equation (5).

3. Parameter Design of DC–Link Inductance and AC–Filter Capacitance

3.1. Selection Range of DC–Link Inductance

A strong output impedance and constant current source characteristics of the DC side in CSI can be obtained by a larger DC–link inductance. However, the current regulation becomes slower as the inductance increases, which leads to overmodulation and generates low–order current harmonics. Therefore, it is necessary to determine a selection range of DC–link inductance.

In the start–up stage, the DC–link current is too low to provide the required current for the IM, so the zero switching state is adopted, and \( L_{dc} \) is charged by \( u_{dc} \); the equivalent circuit is shown in Figure 3a. The state equation of \( i_{dc} \) is expressed as

\[
L_{dc} \frac{di_{dc}}{dt} = u_{dc}
\]

(6)

where \( T_{up, max} \) is the maximum allowable charging time, making sure that the charging time for \( i_{dc} \) to increase from zero state to maximum DC–link current \( i_{dc, max} \) is less than \( T_{up, max} \). \( L_{dc} \) should satisfy the following condition:

\[
L_{dc} < \frac{u_{dc} T_{up, max}}{i_{dc, max}}
\]

(7)

According to the conclusion in Ref. [21], the current fluctuation \( \Delta i_{dc} \) can be calculated when the IM is powered independently by \( L_{dc} \), which contains two active switching states in one carried period. For example, in sector I, the equivalent discharge circuits are shown in Figure 3b,c, and \( \Delta i_{dc} \) is expressed as

\[
\Delta i_{dc} = \frac{3T_s U \cos \theta}{2L_{dc} i_{dc}}
\]

(8)

where \( T_s \) is the carried period, equal to 100 \( \mu \)s; \( U \) is the fundamental amplitude of the stator voltage; \( I \) is the fundamental amplitude of the stator current; \( \theta \) is the phase deviation between stator voltage and stator current. In other sectors, the voltage of \( L_{dc} \) and the switching state are shown in Table 1, and the expression of \( \Delta i_{dc} \) is consistent with that of sector I, as shown in Equation (8). Making sure that \( \Delta i_{dc} \) is less than the maximum ripple \( \Delta i_{dc, max} \) under \( \theta = 1 \), \( L_{dc} \) should satisfy the follow condition:

\[
L_{dc} > \frac{3m_{max} m_{up, max} T_s u_{dc}}{2\Delta i_{dc, max}}
\]

(9)

where \( m_{max} \) and \( m_{up, max} \) are the maximum modulation ratio and boost ratio of CSI, respectively. Combining Equations (7) and (9), the selection range of \( L_{dc} \) is expressed as

\[
L_{dc} \in \left( \frac{3m_{max} m_{up, max} T_s u_{dc}}{2\Delta i_{dc, max}} \cdot \frac{u_{dc} T_{up, max}}{i_{dc, max}} \right)
\]

(10)
Figure 3. The equivalent circuits of charging and discharging for $L_{dc}$. (a) Equivalent charging circuit; (b) equivalent discharging circuit when $S_1$ and $S_6$ are turned on; (c) equivalent discharging circuit when $S_1$ and $S_2$ are turned on.

The parameters are set as follows: $m_{\text{imax}} = 1$, $m_{\text{umax}} = 1$, $u_{dc} = 24$ V, $\Delta i_{dc, \text{max}} = 1$ A, $i_{dc, \text{max}} = 50$ A and $T_{\text{up, max}} = 20$ ms. Substituting the above parameters into Equation (10), the minimum value and maximum value of DC-bus inductance are 3.6 mH and 9.6 mH, respectively.

Table 1. The voltage of $L_{dc}$ and switching state in sectors II–VI.

<table>
<thead>
<tr>
<th>Sector</th>
<th>On–State Switch</th>
<th>Voltage of $L_{dc}$</th>
</tr>
</thead>
<tbody>
<tr>
<td>II (active vector 1)</td>
<td>$S_1$ and $S_2$</td>
<td>$u_{ac}$</td>
</tr>
<tr>
<td>II (active vector 2)</td>
<td>$S_2$ and $S_3$</td>
<td>$u_{bc}$</td>
</tr>
<tr>
<td>III (active vector 1)</td>
<td>$S_2$ and $S_3$</td>
<td>$u_{bc}$</td>
</tr>
<tr>
<td>III (active vector 2)</td>
<td>$S_3$ and $S_4$</td>
<td>$u_{ba}$</td>
</tr>
<tr>
<td>IV (active vector 1)</td>
<td>$S_3$ and $S_4$</td>
<td>$u_{ba}$</td>
</tr>
<tr>
<td>IV (active vector 2)</td>
<td>$S_4$ and $S_5$</td>
<td>$u_{ca}$</td>
</tr>
<tr>
<td>V (active vector 1)</td>
<td>$S_4$ and $S_5$</td>
<td>$u_{ca}$</td>
</tr>
<tr>
<td>V (active vector 2)</td>
<td>$S_5$ and $S_6$</td>
<td>$u_{cb}$</td>
</tr>
<tr>
<td>VI (active vector 1)</td>
<td>$S_5$ and $S_6$</td>
<td>$u_{cb}$</td>
</tr>
<tr>
<td>VI (active vector 2)</td>
<td>$S_6$ and $S_1$</td>
<td>$u_{ab}$</td>
</tr>
</tbody>
</table>

3.2. Selection Range of AC–Filter Capacitance

The IM is represented by the T–type equivalent circuit; the phase–a circuit model of IM and the capacitor is shown in Figure 4, where $L_{\sigma}$ and $L_{\xi}$ are the stator leakage inductor and rotor leakage inductor, respectively; $R_r$ is the rotor resistor; $I_{ac}$ is the current vector of $C_a$; and $s$ is the slip ratio.

Figure 4. T–type equivalent circuit of IM with filter capacitor.

The vector diagram composed of $I_{at}$, $I_{as}$, $I_{ac}$ and $U_{as}$ is shown in Figure 5, and the following relationship can be obtained:
In order to maximize the AC current utilization rate, the following equation should be satisfied:

\[ I_{at} \leq I_{as} \tag{12} \]

According to the geometric relationship in Figure 5, \( I_{ac} \) should be satisfied as follows:

\[ I_{ac} \leq I_{as} \sin \theta \tag{13} \]

Substituting Equation (11) into Equation (13), the expression of maximum capacitance \( C_{\text{max}} \) can be obtained as

\[ C \leq C_{\text{max}} = \frac{2I_{as} \sin \theta}{\omega_s U_{as}} \tag{14} \]

The mathematical relationship of \( U_{as}, I_{as} \) and \( \sin \theta \) can be obtained according to the T–type equivalent circuit of IM shown in Figure 4, and Equation (14) can be rewritten as

\[ C \leq C_{\text{max}} = \frac{2(\omega_s L_{as} + A/B)}{\omega_s \sqrt{(R_s + A/B)^2 + (\omega_s L_{as} + D/B)^2}} \tag{15} \]

where the specific forms of \( A, B \) and \( D \) are expressed as

\[
A = \left( \frac{R_t R_m}{s_t} - \omega_s^2 \sigma L_t L_m \right) \left( \frac{R_t}{s_t} + R_m \right) + \left( \frac{R_t \omega_s I_{m}}{s_t} + R_m \omega_s \sigma L_t \right) \left( \omega_s \sigma L_t + \omega_s L_m \right)
\]

\[
B = \left( \frac{R_t}{s_t} + R_m \right)^2 + (\omega_s \sigma L_t + \omega_s L_m)^2
\]

\[
D = \frac{R_t \omega_s I_{m}}{s_t} + R_m \omega_s \sigma L_t \left( \frac{R_t}{s_t} + R_m \right) - \left( \frac{R_t R_m}{s_t} - \omega_s^2 \sigma L_t L_m \right) \left( \omega_s \sigma L_t + \omega_s L_m \right)
\]

From Equations (15) and (16), \( C_{\text{max}} \) is a function of \( s_t \) and \( \omega_s \) is set as the rated angular frequency of 100π rad/s. Substituting the IM parameters listed in Table 2 into Equations (14) and (15), the relationship curve between \( C_{\text{max}} \) and \( s_t \) can be obtained, which is shown in Figure 6. It is obvious that a minimum value occurs in \( C_{\text{max}} \), so the AC–filter capacitance should be less than 2.37 mF.
Table 2. The parameters of the IM.

<table>
<thead>
<tr>
<th>Parameters</th>
<th>Description</th>
<th>Value</th>
</tr>
</thead>
<tbody>
<tr>
<td>$R_s$ (Ω)</td>
<td>Stator resistance</td>
<td>0.07</td>
</tr>
<tr>
<td>$R_r$ (Ω)</td>
<td>Rotor resistance</td>
<td>0.05</td>
</tr>
<tr>
<td>$L_s$ (mH)</td>
<td>Stator inductance</td>
<td>4.51</td>
</tr>
<tr>
<td>$L_r$ (mH)</td>
<td>Rotor inductance</td>
<td>4.63</td>
</tr>
<tr>
<td>$L_m$ (mH)</td>
<td>Mutual inductance</td>
<td>4.38</td>
</tr>
<tr>
<td>$σ$</td>
<td>Flux leakage coefficient</td>
<td>0.088</td>
</tr>
<tr>
<td>$P_n$ (kW)</td>
<td>Rated power</td>
<td>1.2</td>
</tr>
<tr>
<td>$U_n$ (V)</td>
<td>Rated voltage</td>
<td>48</td>
</tr>
<tr>
<td>$f_n$ (Hz)</td>
<td>Rated frequency</td>
<td>50</td>
</tr>
<tr>
<td>$J$ (kg·m²)</td>
<td>Rotational inertia</td>
<td>0.001</td>
</tr>
<tr>
<td>$n_p$</td>
<td>Number of pole pairs</td>
<td>2</td>
</tr>
<tr>
<td>$R_s$ (Ω)</td>
<td>Stator resistance</td>
<td>0.07</td>
</tr>
</tbody>
</table>

Figure 6. The relationship curve between $C_{max}$ and $s_r$.

Next, the influence mechanism on harmonic characteristics of the stator current caused by AC–filter capacitance will be discussed. The T–type equivalent circuit of IM is a third–order system; it is difficult to derive the transfer function and mathematical relationship between the capacitance and bandwidth. Since the high–order harmonic contents near the switching frequency of $i_{ds}$ ($i_{qs}$) are same as the stator currents in three–phase stationary coordinate, according to Equations (1) and (2), the equivalent circuits of the IM and AC–filter capacitor in the d–q axis based on rotor flux orientation are established, as shown in Figure 7. $i_{ds}$ and $i_{qs}$ are composed of a DC component and several high–order harmonics components near the switching frequency. The equivalent controlled current sources, voltage sources, and back EMFs in Figure 7 cannot amplify or attenuate harmonics, so the suppression effect of high–order harmonics is determined by $R_s$, $σL_s$ and $C$, and the transfer function $G_{i_2i_1}(s)$ from $i_{ds}$ ($i_{dq}$) to $i_{ds}$ ($i_{dq}$) can be expressed as

$$G_{i_2i_1}(s) = \frac{1}{σL_sCs^2 + R_sCs + 1}$$  \hspace{1cm} (17)

and the oscillation angular frequency $ω_n$ can be expressed as

$$ω_n = \frac{1}{\sqrt{σL_sC}}$$  \hspace{1cm} (18)
Figure 7. Second order equivalent circuit of the IM and AC–filter capacitor in the d–q axis: (a) d–axis; (b) q–axis.

Substituting the value of \( R_s \) and \( L_s \) into Equation (16), the Bode plot of \( G_{it2is}(s) \) is drawn under different AC–filter capacitance, which is shown in Figure 8. It can be seen that \( G_{it2is}(s) \) has a resonance spike, and the cutoff frequency decreases with the increase of \( C \). In order to obtain the high quality stator current, \( G_{it2is}(s) \) should effectively suppress the harmonic components in \( i_a \) near the switching frequency \( f_s \), so the oscillation frequency should be far below \( f_s \). This paper stipulates that the oscillation frequency is less than \( f_s/2 \); according to Equation (17), AC–filter capacitance should be satisfied as

\[
C > \frac{1}{\sigma L_s \pi^2 f_s^2}
\]  

(19)

Figure 8. The Bode plot of \( G_{it2is}(s) \) under different AC–filter capacitance.

Substituting the IM parameters listed in Table 2 into Equation (19), the minimum value of \( C \) is 2.5 \( \mu F \).


According to Section 1, the mathematical model formed by the IM and capacitor is a third–order system; when the three closed–loop control strategy is adopted, more voltage sensors are needed, and the dynamic response to torque is slower than the VSI–fed IM drive system. In this section, a novel vector control strategy without a voltage inner loop for the CSI–fed IM drive system will be proposed, which includes current loop, flux loop and speed loop.

4.1. Analysis and Parameter Design for Current Loop

The switching delay and digital control delay are equivalent to a first–order transfer function, which is expressed as
\[ G_d(s) = \frac{1}{1 + T_s s} \]  

Due to the presence of high-frequency noise in the stator current sampling signal, a filter circuit is adopted in this paper, and its equivalent transfer function is

\[ G_{d}(s) = \frac{1}{1 + T_c s} \]  

where \( T_c \) is the filter time constant, equal to \( 6 \times 10^{-5} \) in this paper.

Ignoring the coupling terms in Equations (1) and (3), such as \( \omega_0 L_{atqy}, \omega_0 L_{atqa}, \omega_0 C_{atq} \), and \( \omega_0 C_{atq}, \) in the d–q axis, the control diagram of the current loop without stator voltage sampling and control is shown in Figure 9, where \( k_p \) and \( \tau_d \) are the proportional coefficient and integral coefficient of PI controller for adjusting the stator current, \( \dot{i}_{st}^* \) and \( \dot{i}_{qs}^* \) are the reference value of output current of the inverter bridge in the d–q axis, \( \dot{i}_{ds} \) and \( \dot{i}_{qs} \) are the reference value of the stator currents in the d–q axis.

**Figure 9.** The control diagram of the current loop.

In this paper, the target cut–off frequency and phase margin of current loop are represented as \( \omega_{cc} \) and \( \phi_{cc} \). According to Figure 9, \( Q_{c,amp} \), the amplitude of the current open–loop at \( \omega_{cc} \), can be derived as follows:

\[ Q_{c,amp} = \frac{1}{\sqrt{[1 + (T_s \omega_{cc})^2][1 + (T_c \omega_{cc})^2][1 - \sigma L_s C_{dc}^2 + R_s^2 C^2 \omega_{cc}^2]}} \]  

\( \phi_{c,phase} \), the phase of the current open–loop at \( \omega_{cc} \), can be derived as follows:

\[ \phi_{c,phase} = -\arctan(T_s \omega_{cc}) - \arctan(T_c \omega_{cc}) - \arctan\left(\frac{R_s C_{dc} \omega_{cc}}{1 - \sigma L_s C_{dc}^2 \omega_{cc}^2}\right) \]  

(23)

To ensure that the phase angle margin of the current loop with the PI controller is greater than \( \phi_{cc} \), \( \tau_d \) should be satisfied as follows:

\[ \tau_d \geq \frac{\tan(\phi_{cc} - \phi_{c,phase} - \frac{\pi}{2})}{\omega_{cc}} \]  

(24)

To ensure that the amplitude of the current loop with the PI controller is equal to 1, \( k_p \) should be satisfied as follows:

\[ k_p = \frac{\omega_{cc}}{Q_{c,amp} \sqrt{1 + \tau_d^2 \omega_{cc}^2}} \]  

(25)

In order to achieve fast dynamic response and the effective suppression of high-order harmonics in the output current of the inverter bridge, \( \omega_{cc} \) is set to 2000\( \pi \) rad/s, which is one-tenth of the switching frequency. Meanwhile, to obtain a better stability, \( \phi_{cc} \) is set to \( \pi/4 \) rad. Substituting the parameters listed in Table 2 into Equation (22) to Equation (25), \( k_p \) and \( \tau_d \) can be calculated as 0.98 and 22.15, respectively. The Bode plot of the
current closed–loop transfer function \( G(s) \) is shown in Figure 10; the bandwidth of \( G(s) \) is 1kHz, and the amplitude remains 0 dB when the frequency is less than 1kHz. When the frequency is greater than 1 kHz, the amplitude decays with a slope close to \(-60 \text{ dB/dec}\), which can effectively suppress the high–order harmonics generated in the switching process. This explains that a good steady–state and dynamic performance can be achieved by the PI parameter optimization proposed in this paper. In order to facilitate the subsequent parameter design of the external loop control system, \( G(s) \) is simplified to a second–order transfer function and expressed as

\[
G_i(s) = \frac{3.13 \times 10^4}{1.65 \times 10^{-3}s^2 + 37.1s + 3.13 \times 10^7}
\]  

(26)

**Figure 10.** Bode plot of transfer function of current closed loop.

### 4.2. Analysis and Parameter Design for Flux Loop

A PI controller is adopted to adjust rotor flux \( \psi_r \) and is combined with Equation (4); the diagram of the flux loop with a PI controller is shown in Figure 11. \( k_{\psi p} \) and \( \tau_{\psi r} \) are the proportional coefficient and integral coefficient of the PI controller.

**Figure 11.** The control diagram of the flux loop.

To simplify the design method of the control parameters, zero–pole cancellation is adopted, and \( \tau_{\psi r} \) is set to \( T_r \). Then, the root locus plot is utilized to determine \( k_{\psi p} \), which is shown in Figure 12. It is obvious that the condition for the stability of the flux loop is as follows:

\[
k_{\psi p} \in (0, 6.21 / L_m)
\]  

(27)

Finally, \( k_{\psi p} \) is obtained by the Sisotool box of Matlab, where \( k_{\psi p} \) is equal to \( 3.15/L_m \); the Bode plot of flux open–loop is shown in Figure 13. The cut–off frequency is 600 Hz, and the phase margin is \( 60^\circ \).
Figure 12. Root locus plot of rotor flux loop.

Figure 13. Bode plot of rotor flux open loop.

4.3. Analysis and Parameter Design for Speed Loop

The speed of IM is sampled by an orthogonal encoder, and the sampling frequency is half of $f_s$; the transfer function of speed sampling and delay is expressed as

$$G_d(s) = \frac{1}{1 + 2\tau_s s} \quad (28)$$

In accordance with Equations (5) and (6), the block diagram of the speed loop under the PI controller is shown in Figure 14, where $k_p$ and $\tau_i$ are the proportional coefficient and integral coefficient of the PI controller for adjusting the stator current, respectively. $\omega^*_s$ is the rated angular frequency. It is noted that the speed loop contains variable $\psi_r$, so the cut-off frequency of the speed open loop should be less than that of the flux open loop, and $\psi_r$ is set as 0.05 Wb. The open-loop transfer function of speed $G_{s,\text{open}}(s)$ can be expressed as

$$G_{s,\text{open}}(s) = \frac{3n_p L_w \psi_r k_w G_1(s)(1 + \tau_i)}{2JL_s s^2(1 + 2\tau_i s)} \quad (29)$$

The close-loop transfer function of speed $G_{s,\text{close}}(s)$ is expressed as
\[ G_{s,\text{close}}(s) = \frac{G_{s,\text{open}}(s)(1+2T_s)}{1+G_{s,\text{open}}(s)} \]  

(30)

\[ \omega_s = \frac{\tan(\phi_{s,\text{phase}} - \phi_{s,\text{amp}}) - \pi}{2} \]

\[ \tau_{ai} = \frac{\omega_{s,\text{zo}}}{k_{sp}} \]

\[ r_{ai} = \frac{\omega_{s,\text{zo}}}{Q_{s,\text{amp}} \sqrt{1 + \tau_{ai}^2}} \]

where \( Q_{s,\text{amp}} \) and \( \phi_{s,\text{phase}} \) are the amplitude and phase of the speed open loop at \( \omega_{s,\text{zo}} \) before the PI controller, respectively. \( k_{sp} \) and \( \tau_{ai} \) are obtained by substituting the equation \( \omega_{s,\text{zo}} = 200\pi \) rad/s and \( \phi_{s,\text{zo}} = \pi/6 \) rad into Equation (31). The specific expression of \( G_{s,\text{close}}(s) \) is as follows:

\[ G_{s,\text{close}}(s) = \frac{30.89s^4 + 8.5 \times 10^3s^3 + 9.34 \times 10^5s^2 + 2.93 \times 10^3s + 2.93 \times 10^1}{5.42s^4 + 2.716 \times 10^5s^3 + 0.6s^2 + 7045s^4 + 4.29 \times 10^5s^3 + 1.01 \times 10^7s^2 + 2.93 \times 10^6s + 2.93 \times 10^2} \]

(32)

The Bode plots of open-loop and closed-loop transfer functions are shown in Figure 15. The cut-off frequency of the speed open-loop transfer function is 100 Hz, and the phase margin is 30°. The closed-loop transfer function has a good following characteristic in the bandwidth of 40 Hz. The unit step response of the speed closed-loop transfer function is shown in Figure 16; the rise time is only 2 ms, the overshoot is little, and no steady-state error occurs.

Figure 14. The control diagram of the speed loop.

Figure 15. Bode plot of speed open loop and closed loop.
5. Simulation Results

The simulation model of the CSI-fed IM drive system with bidirectional DC chopper is built on MATLAB Simulink, including the power module, control module and modulation module. The parameters of IM adopted are consistent with Table 2, and other simulation parameters are shown in Table 3.

Table 3. Simulation parameters.

<table>
<thead>
<tr>
<th>Parameters</th>
<th>Value</th>
</tr>
</thead>
<tbody>
<tr>
<td>DC-link inductor $L_{dc}$</td>
<td>4 mH</td>
</tr>
<tr>
<td>Filter capacitor $C_a, C_b, C_c$</td>
<td>66 μF</td>
</tr>
<tr>
<td>DC-bus voltage $u_{dc}$</td>
<td>20 V</td>
</tr>
<tr>
<td>Switching frequency</td>
<td>10 kHz</td>
</tr>
<tr>
<td>Sampling frequency</td>
<td>10 kHz</td>
</tr>
</tbody>
</table>

5.1. Load Acceleration Condition

The value of load torque is set to 3 N·m, the initial reference speed is 1500 rpm, and the reference speed is 3000 rpm when the time is 0.8 s. The simulation waveforms of stator phase a current, stator currents in the d–q axis, electromagnetic torque, motor speed and rotor flux in the whole simulation process are shown in Figure 17.

To improve DC–link current utilization and reduce losses and harmonic distortion, according to the MTPA method proposed in Ref. [17], the references of rotor flux and DC–link current are adjusted with the change of operating point; the specific principle is as follows:

A. In the dynamic process, in order to provide greater torque, the reference DC–link current is set to 50 A, which is the maximum current of the DC–link inductor.

B. In the steady state, the references of rotor flux $\psi^*_r$ and DC–link current $i^*_d$ are adjusted according to the load torque and speed, which are shown in Table 4.

Table 4. The reference of rotor flux and DC–link current under different load torque.

<table>
<thead>
<tr>
<th>$T_e$ (N·m)</th>
<th>Speed (rpm)</th>
<th>$\psi^*_r$ (Wb)</th>
<th>$i^*_d$ (A)</th>
</tr>
</thead>
<tbody>
<tr>
<td>0</td>
<td>1500</td>
<td>0.03</td>
<td>11.5</td>
</tr>
<tr>
<td>0</td>
<td>900</td>
<td>0.031</td>
<td>12</td>
</tr>
<tr>
<td>1</td>
<td>1500</td>
<td>0.039</td>
<td>14.8</td>
</tr>
<tr>
<td>2</td>
<td>1500</td>
<td>0.055</td>
<td>20.9</td>
</tr>
<tr>
<td>3</td>
<td>1500</td>
<td>0.068</td>
<td>23.5</td>
</tr>
<tr>
<td>3</td>
<td>3000</td>
<td>0.064</td>
<td>22.8</td>
</tr>
</tbody>
</table>
Figure 17. Simulation waveform under load torque of 3 N·m. (a) $i_{dc}$; (b) $i_a$; (c) $i_d$ and $i_q$; (d) $\psi_r$; (e) $T_e$; (f) Speed.

The whole simulation process is divided into three stages.

Start-up stage: in order to make the motor speed reach the reference speed as soon as possible, the hysteresis control strategy of DC–link current is adopted to regulate $i_{dc}$ to 50 A, which is the max reference, so as to meet the demand of current. The optimal reference value of rotor flux $\psi^*_r$ is 0.068 Wb. After 0.05 s, $\psi^*_r$ reaches the reference value, $i_{dc}$ is around 15 A, $i_q$ is around 43 A, and $T_e$ is 8 N·m. After 0.4 s, the speed is adjusted to 1500 rpm, and $T_e$ is reduced to 3 N·m.

Steady–state operation stage at 1500 rpm: $i_{dc}$ is regulated to 23.5 A, as shown in Figure 17a. It can be seen from Figure 17e that when $i_{dc}$ is reduced to 23.5 A, the ripple of torque is significantly reduced.

Second acceleration stage: at the time of 0.8 s, the value of step speed is 3000 rpm. $i_{dc}$ reach to 50 A again, which is the same as the start-up stage. When $\psi^*_r$ is adjusted to 0.064 Wb, $i_{dc}$ is reduced immediately, $\psi_r$ follows the given value quickly, $i_q$ and $T_e$ are the same as the start-up. Finally, at 1.08 s, the speed reaches to 3000 rpm.
5.2. Deceleration Condition with No Load

This section simulates the deceleration condition of the motor with no load. When the IM speed reaches the rated speed of 1500 rpm, the reference speed is changed to 900 rpm at 0.6 s. The simulation waveforms of electromagnetic torque, speed, DC-link current and output power of DC power supply are shown in Figure 18, respectively. When the speed decelerates, \( i_{dc} \) is around 30 A. The electromagnetic torque plays a braking role and reaches to \(-2.3 \) N·m, and the speed reaches stability again after 0.28 s. The DC power supply intermittently absorbs the energy feedback from the motor decelerating process, and the output power is less than zero, so the energy feedback is realized.

![Simulation waveform under no-load deceleration](image)

**Figure 18.** Simulation waveform under no-load deceleration. (a) \( i_{dc} \); (b) output power of DC power supply; (c) \( T_e \); (d) speed.

5.3. Load Disturbance Condition

Figure 19 are the simulation waveforms of load disturbance, when the speed remains at 1500 rpm, the load torque is reduced from 3 N·m to 1 N·m at 0.1 s, and the speed controller makes the electromagnetic torque drop rapidly to 1 N·m, where the speed remains unchanged. At the same time, the \( \psi_r \) and \( i_a \) are reduced to the optimal value of 0.039 Wb and the minimum value of 14.8 A, respectively. When the load torque is increased to 2 N·m at 0.3 s, the speed is stabilized at 1500 rpm after 0.16 s, and the maximum error of speed is only 14 rpm. The \( \psi_r \) and \( i_a \) are adjusted to 0.055 Wb and 20.9 A, respectively.
6. Experimental Results

In order to verify the feasibility of parameter design for DC–link inductance and AC–filter capacitance and the effectiveness of the vector control strategy proposed in this paper, the experimental platform was established, as shown in Figure 20; the experimental parameters are shown in Table 5. The value of DC–link inductance and AC–filter capacitance are satisfied with the selection range specified in Section 3. Because the motor speed, rotor flux, and stator d–q axis current cannot be measured by the oscilloscope, in order to facilitate debugging as well as the observation and storage of these signals, this paper develops a host computer program on the GUI platform of MATLAB to realize the data communication with DSP.

Table 5. Models and parameters of the experimental platform.

<table>
<thead>
<tr>
<th>Category</th>
<th>Part Number</th>
<th>Parameters</th>
</tr>
</thead>
<tbody>
<tr>
<td>DC–link inductor</td>
<td>custom–made</td>
<td>4mH</td>
</tr>
<tr>
<td>Filter capacitor</td>
<td>MKP1847610354P4</td>
<td>66μF</td>
</tr>
<tr>
<td>DC–bus voltage</td>
<td>Battery</td>
<td>24V</td>
</tr>
<tr>
<td>Current sensors</td>
<td>LEM LA28–P</td>
<td>/</td>
</tr>
<tr>
<td>Diode modules</td>
<td>RM300HA–24F</td>
<td>240V/300A</td>
</tr>
<tr>
<td>IGBT modules</td>
<td>PM400HSA120</td>
<td>400V/120A</td>
</tr>
<tr>
<td>DSP</td>
<td>TMS320F28335</td>
<td>/</td>
</tr>
<tr>
<td>Encoder</td>
<td>OIH, 2500C/T</td>
<td>/</td>
</tr>
</tbody>
</table>
Next, the experimental results will be carried out under the three operation conditions: acceleration with load, deceleration with no-load and change of load.

6.1. Performance under Acceleration with Load of 3 N\cdot m

Experimental waveforms of accelerated conditions are shown in Figure 21, under the load of 3 N\cdot m. The experimental process consists of two stages. The reference speed changes from 1500 rpm to 3000 rpm, in order to reduce the harmonics and losses, and to improve current utilization, the references of $i_{dc}$ and $\psi_r$ are also adjusted according to the calculation method for the minimum DC-link current and the optimal flux proposed in Ref. [17]. In the dynamic process, to obtain a larger torque and reduce the dynamic response time, the reference $i_{dc}$ is set to 50 A. In the first stage, the reference speed is set to 1500 rpm, and the reference $i_{dc}$ and $\psi_r$ are set to 23.5 A and 0.068 Wb, respectively. In the second stage, the reference speed is set to 3000 rpm, and the reference $i_{dc}$ and $\psi_r$ are set to 22.5 A and 0.065 Wb, respectively. From Figure 21, the rise time of $i_{dc}$ increasing from 0 A to 50 A is only 15 ms, and the ripple of $i_{dc}$ is less than 0.5 A, which demonstrates that the parameter design method of the DC-link inductor is feasible. In two stages, the speed and rotor flux can quickly track the references, and the THD of the stator current is less than 2%.

![Figure 20. Prototype for CSI-fed IM drive system with bidirectional DC chopper.](image)
6.2. Performance under Deceleration with No Load

The experimental waveforms of the reference speed decreasing from 1500 rpm to 900 rpm with no load are shown in Figure 22. Unlike with the operational condition where the load is 3 N·m, the amplitude of the stator voltage is 18.5 V, which is less than \( u_{dc} \). \( i_{dc} \) remains at 12 A with little ripple in the steady state, and during the dynamic process of speed decreasing from 1500 rpm to 900 rpm, the reference of \( i_{dc} \) is set to 30 A, and the mechanical energy from IM is recovered by the DC voltage source. Due to a fast dynamic and little ripple of \( T_e \), dynamic and steady performance of speed can be obtained, the rotor flux can quickly track the references, and the stator current has very little harmonic distortion.
Figure 22. Experimental waveforms under deceleration with no load: (a) DC–link current; (b) Rotor flux; (c) Torque; (d) Speed; (e) Stator current in d–q axis; (f) Stator voltage of phase–a under speed of 900 rpm; (g) Stator current of phase–a under speed of 1500 rpm; (h) Stator current of phase–a under speed of 900 rpm.

6.3. Performance under Different Loads

The reference speed is set to 1500 rpm. The IM starts with the load of 3 N·m, then the load is suddenly reduced to 1 N·m, and finally the load is increased to 2 N·m. The experimental waveforms are shown for the reference speed decreasing from 1500 rpm to 900 rpm with no load are shown in Figure 23. Since the torque can be quickly adjusted when the sudden change occurs on the load, the fluctuation of the speed is less than 3%. Meanwhile, \(i_{dc}\) and \(\psi_r\) can quickly track the references.

Figure 23. Experimental waveforms under different loads: (a) DC–link current; (b) Rotor flux; (c) Torque; (d) Speed.
7. Conclusions

In order to improve the performance of the current source inverter–fed inductor motor drive system, this paper presents a novel design principle of DC–bus inductance and AC-filter capacitance, and a vector control technology without voltage loop. The feasibility and correctness are verified by the simulation and experimental results under various operating conditions. The main contributions can be summarized as follows:

1. Under consideration of the dynamic response speed and current ripple, the range expression of the DC–link inductance is derived.
2. Under consideration of the DC–link current utilization and harmonic suppression, the range expression of the AC–filter inductance is derived.
3. The simulation and experimental results show that the fast dynamic performance of speed and flux with no overshoot and a stator current with low harmonic distortion can be obtained by the vector control strategy without voltage loop for current source inverter–fed inductor motor drive systems. The analytical expressions of the control parameters derived in this paper are also suitable for other IM parameters.

Author Contributions: Conceptualization and methodology, P.S.; simulation, Y.L.; validation and experiment, P.S. and C.L.; writing and supervision, P.S.; review and editing, C.L. funding acquisition, P.S. All authors have read and agreed to the published version of the manuscript.

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