Improved Single Current Sensing Method and Its Realization
Based on ADMCF341 DSP Controller

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Abstract

This paper describes a single current sensing technique that uses only one DC link current sensing resistor to measure the three phase motor currents, therefore, reducing the cost of whole system. However, in certain conditions, current pulse width in the shunt resistor is very narrow and the measurement becomes difficult. This paper proposed a new strategy to solve this problem by changing PWM switching frequency and using double reference voltage vector PWM technique. The realization of the proposed single current sensing method based on the ADMCF341 DSP, which is specifically designed for easy implementation of single current sensing technique, is described. Simulation and experimental results are presented.

1. Introduction

In order to implement the close-loop current control scheme, it is necessary to measure the motor line currents. The requirement for all these current sensors, such as CT sensor, is a large burden on the overall system size and cost. Also, because these sensors are known to drift in an unpredictable manner with operating temperature and time, variable gain and offset errors are introduced into the motor line current feedback. These errors have been shown to produce undesirable low frequency harmonic currents and oscillating torque components in the motor. In order to reduce the cost of the whole system, three shunt resistor current sensing technique has been introduced, where a single current sensing resistor is located in the bottom of each leg the inverter. However, because of three shunt resistors must be inserted in inverter three phases arms, the IPM (intelligent power module) can not be used.

Recently, current sensor reduction techniques have been developed[1]. In particular, it is shown that it is possible to eliminate the motor line current sensors and derive all necessary current feedback from a DC link single current sensor. Normally, a cheap shunt resistor can be used as the current sensor. However, certain difficulties still exist. Under certain operating conditions, the duration of one or both of the active states may become too small to guarantee reliable extraction of line current samples, so the motor line current can no longer be properly evaluated[2].

In this paper, we proposed a scheme to solve this problem. A variable switching frequency PWM technique and a double reference voltage vector PWM technique are proposed. For the easy realization of the single current sensing technique, a new series ADMCF341 DSP is developed. It has built-in sample/hold and over-current protection circuit, so the design of the DSP controller board can be performed easily. The details of the realization of the proposed single current sensing scheme using the high performance, low cost ADMCF341 DSP controller is described and the effectiveness of the proposed method is verified by simulation and experiment.

2. Limitation of Single Current Sensing Technique

Fig.1 shows the inverter-motor drive system and its corresponding output voltage space vector diagram.
In order to control most inverter systems, it is necessary to know all the three phase currents. As shown in Fig. 1, a single current sensing technique is to measure only the DC link current and then recombine the three phase currents using the inverter switching states. The relationship among the DC link current \(i_d\), the three phase current \(i_u\), \(i_v\), \(i_w\), and the inverter switching states can be given below:

\[
\begin{align*}
    i_d &= f(i_u, i_v, i_w, V_u, V_v, V_w) \\
    &= i_u, \quad \text{when } (V_u, V_v, V_w) = (100) \\
    &= -i_u, \quad \text{when } (V_u, V_v, V_w) = (011) \\
    &= i_v, \quad \text{when } (V_u, V_v, V_w) = (010) \\
    &= -i_v, \quad \text{when } (V_u, V_v, V_w) = (101) \\
    &= i_w, \quad \text{when } (V_u, V_v, V_w) = (001) \\
    &= -i_w, \quad \text{when } (V_u, V_v, V_w) = (110) \\
    &= 0, \quad \text{when } (V_u, V_v, V_w) = (111) \\
    &= 0, \quad \text{when } (V_u, V_v, V_w) = (000)
\end{align*}
\]

(1)

As an example, Fig. 2 shows the three phase inverter’s output voltage waveform (\(V_u, V_v, V_w\)) and the DC link current \(i_d\) waveform in one PWM cycle. From equation (1), it can be understood that it is not possible to make any measurement during the zero vector states (000) and (111), as no current is flowing in the DC line. Two different phase measurements can be made during states (100) and (110), so during \(T_1\), the measured phase current is \(i_u = i_d\), and during \(T_2\), the measured phase current is \(i_w = -i_d\) and the third phase current can be deduced as \(i_v = -(i_u + i_w)\).

The space vector output duration (\(T_1, T_2, T_0\)) can be calculated by:

\[
\begin{align*}
    T_1 &= T_{PWM} - a \times \sin \left( \frac{\pi}{3} - \Theta \right) \\
    T_2 &= T_{PWM} - a \times \sin \left( \Theta \right) \\
    T_0 &= \frac{1}{2} \times \left( T_{PWM} - T_1 - T_2 \right)
\end{align*}
\]

(2)\(\)\(\)\(\)

where \(\Theta\) is defined in Fig. 1, \(T_{PWM}\) is the period.

Under certain conditions, the duration \(T_1\) or \(T_2\), namely, the line-line voltage pulse width are very narrow. In this case, due to the transistor commutation times and transient procedures, if the time \(T_1\) or \(T_2\) is less than certain value \(T_{min}\), the actual phase current is invisible on the DC link current. As a result, it is not possible to derive the motor phase currents from the DC link current.

Supposing the reference voltage vector is \(V^*\) and it lies in the sector surrounded by zero vector (000), (111) and vector (100) and (110), as shown in Fig. 3. The possible position of vector \(V^*\) can be the following four cases:

1. In case (a), the reference voltage vector \(V^*\) lies in the triangle area, where \(T_1, T_2 > T_{min}\). In this case, the three phase currents can be properly measured.

2. In case (b), the reference voltage vector \(V^*\) is very close to vector (110), it results in \(T_1 < T_{min}\). Because \(T_1 < T_{min}\), only one DC line current during
areas can be expressed in Fig.4, which includes a bar area corresponding to the case that the reference voltage vector $V^*$ is too close to the voltage space vector, and a star area corresponding to the case that the magnitude of reference voltage vector $V^*$ is too small. If the reference voltage vector $V^*$ lies in these areas, the correct measurement of the three phase motor currents is impossible.

3. Strategy of Avoiding Narrow DC Link Current Pulse

As mentioned above, for correct measurement of the motor current, it is necessary to keep the DC link current or line-line voltage pulse width ($T_1$ or $T_2$) greater than certain value $T_{\text{min}}$. This can be done by two techniques, such as variable switching frequency PWM technique and double reference voltage vector PWM technique, depending in which area the reference voltage vector $V^*$ lies.

(1) Reference Voltage Vector $V^*$ Lies in Bar Area

Variable switching frequency PWM technique is used. For easy understanding, we suppose the motor runs with open-loop V/f constant control and the reference voltage vector $V^*$ lies in the sector surrounded by zero vector (000), (111) and vector (100) and (110), which is illustrated in Fig. 5. The reference voltage vector $V^*$ moves along with a circle locus (dot line). The angle $\beta$ stands for the angle that if the angle $\theta$ of reference voltage vector $V^*$ is greater than $\beta$, it means that the vector $V^*$ lies in the bar area.

As shown in Fig. 5, at timing $t_1$, the angle $\theta$ of reference voltage vector $V^*$ is $\theta_1$. After one PWM cycle at next timing $t_1'$, where $t_1' = t_1 + \omega_0 \times T_{\text{PWM}}$, the angle $\theta$ becomes $\theta_2$. The angle $\beta$ is the minimum state time criterion. When $\theta > \beta$, the current can be measured.
\( \omega_0 \) is fundamental angular speed, \( T_{PWM} \) is PWM period, and if the angle \( \theta_2' \) of reference voltage vector \( V^* \) is \( \theta_2' > \beta \), the vector \( V^* \) lies in the bar area and it is impossible to derive the three phase motor currents properly.

In this case, we change the PWM switching frequency or period \( T_{PWM} \) and enforce vector \( V^* \) move out of the bar area, so at timing \( t_2 \), the angle \( \theta_2 \) of vector \( V^* \) can be given by,

\[
\theta_2 = \frac{2\pi}{3} - \beta
\]

(5)

The corresponding PWM period \( T_{PWM_{t2}} \) can be given below,

\[
T_{PWM_{t2}} = \frac{\theta_2 - \theta_1}{\omega_0}
\]

(6)

Obviously, \( T_{PWM_{t2}} > T_{PWM} \). To keep the average switching frequency unchanged, the vector \( V^* \) must be enforced to rotate a small angle when it lies in the outside of bar area.

The same approach can be applied to the close-loop current control algorithm. Because of the change of PWM period and if the sampling period is synchronized with the PWM period, the time parameters of controller must be modified.

(2) Reference Voltage Vector \( V^* \) Lies in Star Area

Double reference voltage vector PWM is used. As shown in Fig. 6, the reference voltage vector \( V^* \) lies in the star area. In this case, we use another two reference voltage vector \( V_1 \) and \( V_2 \) to generate PWM signals. The phase difference between vector \( V_1 \) and \( V_2 \) is \( \pi \) and the magnitude difference between them is \( | V^* | \). The mechanism of this technique can be explained below.

In the first PWM cycle, the vector \( V_1 \) is used to generate PWM. In the second PWM cycle, the vector \( V_2 \) is used to generate PWM. The average generated vector is \( V^* \). This process continues until the reference voltage vector \( V^* \) goes to the outside of the star area.

Supposing the minimum voltage vector lying in the star area is \( V_{min} \). To keep the motor current ripple as small as possible, the reference voltage vector \( V_1 \) and \( V_2 \) can be given below,

\[
\begin{align*}
|V_2| &= |V_{min}| \\
\angle V_2 &= \angle V^* + \pi \\
|V_1| &= |V_2| + |V^*| \\
\angle V_1 &= \angle V^*
\end{align*}
\]

(7)

(8)

Fig.6. Reference voltage vector lies in star area.

3. Realization Based on ADMCF341 DSP

For the purpose of simplifying the realization of the single current sensing technique, a new series Analog Devices ADMCF341 motor control DSP has been developed. Basically, this is achieved by the unique design of Isense input analog-to-digital conversion (ADC) block of the ADMCF341, which is shown in Fig.7.

![Fig.7. Isense ADC circuit of DSP ADMCF341.](image-url)
The ADC of the ADMCF341 DSP[3] is based on the so-called single slope conversion technique. This topology converts the analog signal by simply timing the crossover between the analog input and a reference ramp voltage. This reference ramp voltage is made by charging an external capacitor $C$ with a programmable current source $I_{\text{CONST}}$. Comparing each ADC input to the reference ramp voltage, and timing the comparison of the two signals, performs the conversion process.

The following three principal features of the Isense ADC of the ADMCF341 DSP make the realization of the single current sensing technique simple and reliable.

1. **Built-in bipolar Isense amplifier.**
   This bipolar Isense amplifier (gain=$-2.5$) is designed to acquire the voltage on a current sensing resistor, whose voltage can be positive or negative with respect to the power supply ground rail.

2. **Built-in programmable Sample / Hold circuit.**
   This allows the user to determine the desired sampling time, which is programmable from 150 ns to 3.28 ms.

3. **Built-in over-current protection circuit.**
   The signal output of Isense amplifier is used to judge whether over-current occurs or not. If it happens, the PWM trip signal is generated to shunt down the PWM.

For the verification of the proposed single current sensing scheme and the DSP relevant functions, an experimental system is developed and Fig.8 shows its connection circuit. It can be seen that the voltage signal from the resistor is connected directly to two ADC channels (Channel 2 and 3) without any extra interface circuit.

Fig.9 shows the ADC conversion scheme for the single current sensing technique realized by the ADMCF341 DSP. In the first half of PWM period, shunt resistor voltage is sampled the twice during the time $T_1$ and $T_2$ at different timing ($T_{SH,1}$, $T_{SH,2}$) by ADC Ch2 and Ch3, respectively. As shown in Fig.9, two sampled data are $i_u$ and $-i_w$. From these two data and PWM switching states, the three phase motor currents can be derived.

The sampling time can be given below,

$$T_{SH,1} = T_0 + T_1 - \Delta T_1$$  \hspace{1cm} (9)

$$T_{SH,2} = T_0 + T_1 + T_2 - \Delta T_2$$  \hspace{1cm} (10)

where, $\Delta T_1$, $\Delta T_2$ is depended on the design of the DSP controller board and can be determined by experiment. Considering the possible delays and switching transients, it is proposed to take sample of the DC link current towards the end of the active state times

In the second half of PWM period, the sampled data is converted and is available for the next PWM cycle.

![Fig.8. Experimental system.](image)

![Fig.9. ADC conversion scheme.](image)
4. Simulation and Experimental Results

Fig.10 is a simulation result. It shows the waveform of the twice sampled data, which are defined as $i_{d1}$ obtained from ADC channel 2 and $i_{d2}$ obtained from ADC channel 3, and the motor phase current waveform $i_u$, which is derived from $i_{d1}$ and $i_{d2}$.

Fig.10. Simulation result, sampled DC link current and derived motor current waveform.

Fig.11 shows the corresponding experimental result with PWM switching frequency 8 kHz. For comparison, the current waveform measured by CT sensor ($i_{u,CT}$) is also illustrated.

Fig.11. Experimental result, sampled DC link current and derived current waveform.

5. Conclusions

In this paper, an improved single current sensing technique is proposed. For its simple realization, a new series ADMCF341 DSP has been developed. The main features of ADMCF341 DSP related to the realization of the single current sensing technique is introduced and the actual implementation of the proposed one current sensing scheme based on ADMCF341 DSP is presented and its effectiveness is confirmed by the simulation and experimental results.

References:


3. ADMCF341 Datasheets, Analog Devices Inc.