Design and Implementation of a Hybrid Output EMI Filter for High-Frequency Common-Mode Voltage Compensation in PWM Inverters

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Abstract—This paper presents a hybrid output electromagnetic interference (EMI) filter for common-mode voltage compensation in pulsewidth-modulation (PWM) inverters. Its structure is composed of a single-leg four-level active filter connected in series with a low-frequency blocking capacitor and a small passive LC filter that enhances the suppression of the common-mode voltage in the high-frequency range. The series capacitor helps in avoiding the core saturation of the common-mode coupling transformer caused by the low-frequency components of the common-mode voltage. Detection of PWM voltages by high-speed optoisolators is done at the inverter’s output to achieve better common-mode voltage detection and compensation characteristics. Analysis and design guidelines for the hybrid filter are also given to facilitate systematic design in the real implementation of the filter. Experimental results with a real drive system clearly illustrate a significant reduction of the common-mode voltage, the leakage current, the shaft voltage, and the bearing current.

Index Terms—Common-mode voltage compensation, common-mode voltage detection, hybrid electromagnetic interference (EMI) filter, single-leg four-level inverter.

I. INTRODUCTION

Using pulsewidth-modulation (PWM) inverters in adjustable-speed drives leads to a high $\frac{dv}{dt}$ common-mode voltage. This voltage not only generates electromagnetic interference (EMI) disturbances, but also causes leakage currents, shaft voltages, and premature failure of bearings. There are several approaches to reduce the common-mode voltage. One method is to use an alternative inverter topology or PWM control method [1]–[3]. A four-leg inverter discussed in [1] and [2] can decrease the common-mode voltage but with limited maximum modulation index and requires phase-shifted PWM carriers. A dual-bridge inverter topology introduced in [3] may also be used to eliminate the common-mode voltage if the motor is provided with dual windings. This approach is therefore not applicable to the standard PWM inverters.

On the other hand, passive output EMI filters are more widely applied to suppress the common-mode voltage at the output of the inverters [4]–[7]. However, the design of passive filters is quite difficult because it must attenuate the switching-frequency components, while at the same time, it must avoid being excited by the fundamental voltage and the triplen harmonics of the zero voltage injected by the PWM method. Normally, limitation on both the switching frequency and the PWM method must be imposed when using the passive output filter. Moreover, the common-mode filter must always be installed with a normal-mode filter, whether the normal-mode voltage is intentionally to be attenuated or not. Thus, the whole passive output EMI filter becomes bulky and costly.

In recent work [7]–[12], active EMI filters are investigated for possible common-mode voltage compensation. Ogasawara et al. [8] proposed an active circuit that has excellent cancellation of the common-mode voltage. However, it is difficult to implement the active filter in a typical inverter with the dc bus voltage above 500 V because of the uncommon voltage and current ratings of analog push–pull transistors. Xiang [12] introduced a concept of common-mode voltage cancellation by using a single-leg four-level inverter and a common-mode coupling transformer to construct an active EMI filter called therein Active Common-mode-voltage Compensator (ACCom). Since ACCom is based on the switching mode of power devices, it has less losses and is suitable for high-voltage drive systems. Although the ACCom scheme is theoretically promising, only simulation results were given in [12], and feasibility in real applications is still unconfirmed. In addition, there are several implementation issues that should be discussed, e.g., transformer saturation caused by both the low-frequency common-mode voltage and the dc-offset voltage of the four-level inverter, the required core size, the detection delay of the common-mode voltage, etc.

It is noted that in the common-mode voltage generated by the PWM inverter, the high-frequency components are most problematic and need to be eliminated. If the active EMI filter is designed to generate no low-frequency common-mode voltage, its coupling transformer will be subjected only to the high-frequency voltages and can be more compact because a smaller ferrite core can be used. Furthermore, the dc-offset voltage problem will also be solved.

Following the above concept, this paper presents a hybrid output EMI filter for compensation of the high-frequency common-mode voltages in PWM inverters. Its structure is
composed of a single-leg four-level active filter connected in series with a low-frequency blocking capacitor and a passive $LC$ filter. The active EMI filter is responsible for cancellation in the range of switching frequency up to nearly 10 MHz, while the passive filter assists to suppress the resonance in the conducted EMI frequency range due to the stray capacitance of the motor and to improve the filtering characteristic in the higher frequency range as well. In contrast to the conventional active filter, the proposed EMI filter does not compensate the low-frequency common-mode voltage of fundamental and triplen harmonic frequencies, which is normally harmless to the motor. The size of the common-mode coupling transformer can thus be reduced without causing saturation due to the low-frequency components of the common-mode voltage. The low-frequency components are filtered out by insertion of a series capacitor between the four-level inverter and the coupling transformer. Furthermore, unlike the case of passive output $LC$ filter, the switching frequency and the operating frequency of the PWM inverter can be adjusted without exciting the resonance of the $LC$ filter. The space vector PWM method can also be adopted with no resonance problem caused by the injected zero voltage of triplen harmonics.

Additionally, since accurate common-mode voltage detection is indispensable for all active EMI filters, in this paper, a new method to sense the PWM voltages directly at the inverter’s output through high-speed optoisolators is adopted to generate the required common-mode voltage signal. The proposed method is easy to implement without any further electrical isolation and has minimal detection delay time. Analysis and design guidelines for the proposed hybrid filter will be given to support its practical implementation. Finally, experimental results with a real drive system will be shown to verify the feasibility of the proposed hybrid filter.

II. CHARACTERISTICS OF COMMON-MODE VOLTAGE

Fig. 1 shows the typical configuration of industrial PWM inverters. Waveforms of the measured common-mode voltage at the output of PWM inverter ($v_{cN}$) are shown in Fig. 2. The common-mode voltage $v_{cN}$ can be expressed as

$$v_{cN} = v_{cm} + v_{oN} = \frac{v_{uo} + v_{vo} + v_{wo}}{3} + v_{oN}$$

where $v_{oN}$ is the low-frequency (150 Hz) component, which can be observed in Fig. 2(a). This component is generated by the front-end rectifier of the PWM inverter. The first term in (1), i.e., $v_{cm} = (v_{uo} + v_{vo} + v_{wo})/3$, is the common-mode voltage (referred to the negative dc bus) generated by the PWM inverter. Its waveform is determined by the switching pattern of the PWM inverter. This common-mode voltage can be further decomposed into the following: 1) triplen (third) harmonic components and 2) switching frequency and its sideband components, as depicted in Fig. 2(b). The switching frequency and its sideband components generate the leakage and bearing currents and the shaft voltage, which are the leading causes of conducted EMI and machine failures. The high-frequency...
common-mode voltages are thus more harmful than the low-frequency ones and should therefore be eliminated. By considering the waveform in Fig. 2(b), it can be seen that the common-mode voltage is a four-level stepwise waveform. Therefore, an active filter with a single-leg four-level inverter can be used to generate the four-level voltage for compensation of the common-mode voltage.

III. PROPOSED HYBRID OUTPUT EMI FILTER

The structure of the proposed hybrid EMI filter is shown in Fig. 3. The PWM voltages at the inverter’s output are detected through optoisolators and are sent to the logic gate circuit to reconstruct the required common-mode voltage using the single-leg four-level inverter. The status of the six switches of the four-level inverter is given in Table I, and the four levels of dc voltages are provided by four capacitors connected in series between the dc bus. The compensating voltage is fed into the three-phase line via a common-mode coupling transformer. A series capacitor $C_S$ is inserted between the coupling transformer and the four-level inverter’s output in order to block the low-frequency components of the common-mode voltage. As a result, only the desired high-frequency common-mode voltage components are injected into the lines through the transformer, and the physical size of the transformer is reduced as compared to the case when all common-mode components are injected. To enhance the filtering characteristic, an additional capacitor $C_F$ and inductor $L_F$ will be included to form a passive LC filter with the common-mode leakage inductance of the coupling transformer. This results in a hybrid combination of the active and passive EMI filters.

Figs. 4 and 5 are the common-mode equivalent circuits of Fig. 3 with and without the additional passive filter, respectively. $C_O$ denotes the stray capacitance between motor windings and its frame. $R_O$ and $L_O$ are the total resistance and inductance of the motor and cables. The common-mode voltage $v_{cm}$ is the one caused by the PWM inverter, while the compensating common-mode voltage $v_{com}$ is that generated by the single-leg four-level inverter. The common-mode transformer is represented by the model in the dotted area, where $L_{cm}$ is the magnetizing inductance, $L_{lp}$ and $L_{ls}$ are the primary and secondary leakage inductances, and $R_p$ and $R_s$ are the winding resistances of the transformer. Since the series capacitor $C_S$ blocks the low-frequency common-mode components in the common-mode voltage $v_{com}$, the magnetizing inductance $L_{cm}$ is subjected to only the high-frequency components, which will consequently be coupled to the inverter’s outputs for common-mode voltage cancellation. Though the low-frequency common-mode voltage remains impressed to the motor, its effect is normally not harmful due to the high common-mode impedance of the motor at the low frequencies.
IV. ANALYSIS AND DESIGN OF THE HYBRID OUTPUT EMI FILTER

Detail operations and design guidelines for the main components of the filter are described in the following.

A. Common-Mode Transformer and Series Capacitor for Low-Frequency Common-Mode Voltage Rejection

The compensating voltage is coupled to the PWM inverter by the common-mode transformer with four windings: one primary winding connected to the four-level inverter and the other three secondary windings connected to the three-phase PWM inverter’s output. The turn ratio between the primary and secondary windings is 1:1.

For the popular space vector PWM, low-frequency components of the common-mode zero voltage can lead to a large physical size of the transformer core because of the corresponding high volt-second value. The common-mode voltage of the PWM inverter consists of two main components.

1) The switching-frequency components \( v_{sw} \) as shown previously in Fig. 2(b). At the worst case when the modulation index is low, the common-mode voltage waveform becomes as shown in Fig. 6. The maximum magnetic flux \( \phi_{sw} \) in the ferrite core of the coupling transformer corresponding to this voltage is approximately given by

\[
\phi_{sw} = \frac{1}{N} \int_{0}^{1/4 f_{sw}} v_{com} \, dt = \frac{E_d}{8N \cdot f_{sw}} \tag{2}
\]

where \( E_d \) is the dc bus voltage, \( f_{sw} \) is the switching frequency of the inverter, and \( N \) is the number of turns per winding.

2) The triplen (third) harmonic components \( v_{third} \) of the zero voltage injected by the space vector PWM as shown in Fig. 7. At the maximum modulation index, the peak value of \( v_{third} \) is equal to \( E_d/(4\sqrt{3}) \). Therefore, the corresponding maximum magnetic flux \( \phi_{third} \) in the ferrite core due to this voltage can be approximated by

\[
\phi_{third} = \frac{1}{N} \int_{0}^{1/3 f_o} v_{third} \, dt \approx \frac{E_d}{96\sqrt{3} \cdot N \cdot f_o} \tag{3}
\]

where \( f_o \) is the operating frequency of the inverter.
If the transformer must carry all these magnetic fluxes, particularly the triplen harmonic flux $\phi_{\text{third}}$, its core size will become very large. From (2) and (3), if the transformer does not support the triplen (third) harmonic components, the size of the transformer core can be reduced to about one-tenth.

To achieve an implementation with a compact core size, the capacitor $C_S$ is thus inserted in series at the four-level inverter’s output. The magnetizing inductance $L_{cm}$ and the inserted $C_S$ form a high-pass filter that will filter out the low-frequency components around 0–150 Hz, while the higher frequency common-mode voltage components will be passed to the transformer.

Design guidelines of the common-mode transformer, which carries only $\phi_{\text{sw}}$, are derived under the following considerations.

1) Maximum Current of Common-Mode Transformer:
Since the low-frequency component is blocked by the decoupling capacitor $C_S$ and is not applied to the transformer, the peak value of the transformer magnetizing current $i_p$ is largest when the inverter’s output voltage is small, and its value can be calculated from the square-wave common-mode voltage [5], as shown in Fig. 6. Therefore, for a given maximum value of $i_p$, the magnetizing inductance $L_{cm}$ of the transformer must satisfy

$$L_{cm} = N^2 A_L \geq \frac{E_d}{8 \cdot i_p \cdot f_{sw}}.$$  \hspace{1cm} (4)

Here, $A_L$ is the inductance factor of the core.

2) Maximum Saturation Level of the Transformer Core:
Since the amplitude of the square-wave common-mode voltage is determined by the dc bus level $E_d$ (Fig. 6), the condition to keep the core away from saturation is given by

$$N \geq \frac{E_d}{8 B_S \cdot A_c \cdot f_{sw}}$$  \hspace{1cm} (5)

where $A_c$ is the cross-sectional area of the core, and $B_S$ is the saturation flux density. Consequently, the minimum number of turns required is

$$N_{\text{min}} = \frac{E_d}{8 B_S \cdot A_c \cdot f_{sw}}.$$  \hspace{1cm} (6)

3) Window Area of the Transformer Core:
The possible maximum number of turns $N_{\text{max}}$ for each winding is related to the window area of the selected core by

$$A_w = \frac{N_{\text{max}}}{k_w} (a_{w1} + 3a_{w2})$$  \hspace{1cm} (7)

where $A_w$ is the window area of the core, $k_w$ is the winding factor, and $a_{w1}$ and $a_{w2}$ are the wire sizes of the primary and secondary windings, which depend on $i_p$ and the rated current of the PWM inverter. Taking into consideration the saturation constraint in (6), the minimum window area is achieved when

$$N = N_{\text{min}} = N_{\text{max}},$$

and the suitable ferrite core can be selected from

$$A_c \cdot A_w \cdot B_S \geq \frac{E_d}{8 f_{sw} \cdot k_w} (a_{w1} + 3a_{w2}).$$  \hspace{1cm} (8)

The design guidelines are summarized as a flowchart in Fig. 8.

In this paper, the common-mode transformer is designed for the inverter ratings of 2.2 kW, 50 Hz, 380 V ($E_d \approx 540$ V), 6.2 A, and $f_{sw} = 10$ kHz.

First, the peak value of the transformer current $i_p$ is specified as 0.5 A. $a_{w1} = 0.2$ mm$^2$ is then used, and SWG25 is selected for the primary winding. As for the secondary windings, $a_{w2} = 1.6$ mm$^2$ (SWG17) is chosen corresponding to the rated current of the PWM inverter. The ferrite core with the specification
shown in Table II is selected to satisfy (8), with $k_w = 0.4$. The minimum number of turns per winding is then calculated to be

$$N \geq N_{\text{min}} = \frac{E_d}{8B_S \cdot A_{L} \cdot f_{sw}} = 60 \text{ turns.}$$

For a comparison of the core sizes, if the transformer must support both the switching-frequency magnetic flux $\phi_{sw}$ and the triplen harmonic flux $\phi_{third}$ as in the conventional active filter, it will need four pieces of the ferrite core shown in Table III.

In order to assure a sufficient value of the magnetizing inductance of the common-mode transformer, the final number of turns is selected to be 90, and the inductance becomes $L_{cm} = N^2 A_L = 82 \text{ mH}$.

The so-designed magnetizing inductance $L_{cm}$ is found to satisfy the specification on the transformer current $i_p$ in (4). In the real implementation, the measured inductances of the common-mode transformer are found to be $L_{cm} = 82 \text{ mH}$, and the primary and secondary leakage inductances are $L_{lp} = 0.54 \text{ mH}$ and $L_{ls} = 0.36 \text{ mH}$, respectively.

Since the magnetizing inductance $L_{cm}$ and the inserted $C_S$ form a high-pass filter to block the low-frequency components away from the common-mode transformer, the cutoff frequency of the high-pass filter $f_{C1}$ should be located between the triplen harmonic frequency $(3f_O)$ and the switching frequency $f_{sw}$ to avoid the effect of resonance. The suitable cutoff frequency of the high-pass filter is selected to be 1 kHz, and the value of $C_S$ is calculated from (9) to be about 0.2 $\mu$F. We have

$$C_S = \frac{1}{(2\pi f_{C1})^2 L_{cm}}, \quad 3f_O < f_{C1} < f_{sw}. \quad (9)$$

B. Detection of Common-Mode Voltage

Precise detection of the common-mode voltage is a vital task that determines the effectiveness of voltage compensation. In [12], a Y-connected resistor circuit and high-voltage isolation amplifiers as shown in Fig. 9 are used to detect the common-mode voltage. However, the isolation amplifiers are costly, and their delays can cause imperfect common-mode voltage compensation. An alternative approach is to directly use the gate drive signals of the PWM inverter to generate the gate drive signals of the four-level inverter [12]. Although this approach is easy to implement, there exists some detection error due to the dead time of the PWM inverter.

This paper proposes a new detection method as shown in Fig. 10. The PWM voltages at the inverter’s output are directly detected to avoid the dead-time effect, and the delay of the detection circuit is reduced by using high-speed logic gates and optoisolators. From the experimental results, it is found that the delay time of the detection circuit is about 200 ns. The overall delay time between the common-mode voltage at the output of PWM inverter $v_{cm}$ and the compensating common-mode voltage $v_{com}$ is roughly 500 ns, as shown in Section V. It includes all the delays caused by the detection circuit, the driver circuit, and the dead time of the four-level inverter.

C. Single-Leg Four-Level Inverter

Since the compensating common-mode voltage $v_{com}$ is generated by the single-leg four-level inverter depicted in Fig. 3, which operates in switch mode, the active filter can thus generate a high-frequency common-mode voltage without much difficulty as compared to the conventional analog push–pull topology [8]. In addition, in the OFF state, each power device must handle the voltage of only $E_d/3$, so it is not hard to find the power devices with the required voltage rating. In the implementation, small and fast power MOSFETs IRF730 are used with a 100-ns dead time. Since ideally, the four-level inverter generates the compensating common-mode voltage and supplies only the corresponding magnetizing current to the transformer’s core, it will not supply any active power. Therefore, the average current per switching period of each dc-bus capacitor will be equal to zero and the problem regarding the dc-bus voltage imbalance is negligible.
To provide the required common-mode voltage levels for each switching state, the capacitors $C_1 - C_4$ must satisfy

$$C_1 = C_4 \quad C_2 = C_3 = 2C_1.$$  \hfill (10)

The final values of $C_1 - C_4$ are determined by considering the allowable ripple $\Delta v_{\text{com}}$ (peak value) in the compensating voltage. The worst case ripple occurs when $v_{\text{com}}$ is a square wave, as shown in Fig. 6. In this situation, the switching state of four-level inverter is 111 or 000 when $v_{\text{com}}$ is positive or negative, respectively.

From Fig. 3, it can be shown that half of the transformer magnetizing current will flow through the capacitors $C_1$ and $C_2$. Therefore, the peak value of the charge/discharging current will be half of that in Fig. 6, i.e.,

$$\frac{i_p}{2} = \frac{E_d}{16L_{\text{cm}} \cdot f_{\text{sw}}}. \hfill (11)$$

Since the charging time of the capacitors $C_1$ and $C_2$ is $1/(4f_{\text{sw}})$, with (10), the ripple voltage $\Delta v_{\text{com}}$ across the capacitors $C_1$ and $C_2$ caused by the charging/discharging current becomes

$$\Delta v_{\text{com}} = \left(\frac{3}{2}\right) \frac{1}{C_1} \cdot \frac{i_p}{2} \cdot \frac{1}{4f_{\text{sw}}}. \hfill (12)$$

Consequently, if the ripple voltage $\Delta v_{\text{com}}$ is given, the capacitor $C_1$ can finally be determined from

$$C_1 = \frac{3E_d}{128 \cdot \Delta v_{\text{com}} \cdot L_{\text{cm}} \cdot f_{\text{sw}}^2}. \hfill (13)$$

In this paper, the ripple voltage $\Delta v_{\text{com}}$ is specified as 0.5% of $E_d/2$ ($= 1.35$ V). With the parameters of the single-leg four-level inverter ($E_d = 540$ V, $f_{\text{sw}} = 10$ kHz) and the inductance $L_{\text{cm}}$ ($= 82$ mH), all capacitors of the four-level inverter are designed to be $C_1 = C_4 = 3.3 \mu$F and $C_2 = C_3 = 6.6 \mu$F.

### D. Passive Filter for Improvement of Filtering Characteristics

By using the equivalent circuit in Fig. 4, the common-mode voltage filtering characteristic of the pure active filter with the series capacitor $C_S$ can be investigated from the frequency response $v_{\text{mot}}/v_{\text{cm}}$ shown as a dotted curve in Fig. 11. The term “no filter” used throughout the paper refers to the case without any filters and the coupling transformer. Note that without filters, a resonance exists at the frequency $f_{\text{C0}} = 1/(2\pi \sqrt{L_O C_O})$ due to the motor’s leakage inductance and the stray capacitance.

With the measured parameters $C_O = 0.17$ nF and $L_O = 26 \mu$H, $f_{\text{C0}}$ becomes about 2.4 MHz.
When the pure active filter with the common-mode coupling transformer and the series capacitor $C_S$ is installed, the frequency response of $v_{mot}/v_{cm}$, taking into account the detection delay time of 500 ns, becomes a dashed curve shown in Fig. 11. In general, the active filter provides a good attenuation in the range of 2 kHz to 30 MHz and a satisfactory attenuation at the switching frequency (10 kHz) of about $-30 \text{ dB}$. However, two resonances can be observed in the frequency response. The lower resonance frequency $f_{C_1} = 1/(2\pi \sqrt{L_{cm} C_S})$ belongs to the high-pass filter formed by the series capacitor $C_S$ and the magnetizing inductance $L_{cm}$, and it functions to block the low-frequency common-mode component, as discussed previously. On the other hand, the higher resonance frequency $f_{C_2} = 1/(2\pi \sqrt{(L_{lp} + L_{ls}) C_O})$ arises from the total leakage inductance of the common-mode transformer $L_{lp} + L_{ls}$ and the stray capacitance $C_O$. As this latter resonance occurs around 400 kHz in the conducted EMI range, it causes degradation of the filtering property. To alleviate this problem, a small passive filter is applied to improve the filtering characteristic, particularly around this resonance frequency $f_{C_2}$.

To realize the passive filter, $L_F$ and $C_F$ are added as shown in Fig. 3, with its equivalent circuit depicted in Fig. 5, and the active filter becomes a hybrid one. The characteristic of the passive filter is governed by the capacitor $C_F$, the total leakage inductance $L_{lp} + L_{ls}$, and the transferred inductance of $L_F$. The location of the inductance $L_F$ of the passive filter in this circuit has an advantage over the conventional LC filter that is connected directly to the three-phase power lines because only a small common-mode current passes through the inductance winding.

In order that the passive filter can function to reduce the effect of the resonance in the conducted EMI range, the cutoff frequency of the passive filter $f_{C_3} = 1/(2\pi \sqrt{(L_F + L_{lp} + L_{ls}) C_F})$ must be designed to be lower than 150 kHz. Considering the frequency response in Fig. 11, the cutoff frequency of the passive filter is selected to be 90 kHz. Another constraint for the design of this passive filter is that $C_F$ should be small to prevent a high normal-mode current. In this paper, $C_F = 0.3 \text{ nF}$ and $L_F = 2.9 \text{ mH}$ are selected, and the frequency response of the hybrid filter is shown as a solid curve in Fig. 11. In comparison to the pure active filter, the filtering property of the hybrid filter is enhanced, and a better attenuation in the conducted EMI frequency range (150 kHz to 30 MHz) is obtained.

V. PERFORMANCE EVALUATION OF HYBRID ACTIVE FILTER

In this section, the performance of the proposed active filter will be evaluated through the measurements of the common-mode voltages, the leakage current, the motor’s shaft, and frame voltages as well as the bearing current. The experimental setup is given as shown in Fig. 1, wherein the motor’s bearings are modified and insulated by Teflon films for bearing current measurement [10] and a carbon brush is used for shaft voltage measurement.

![Fig. 12. Waveforms of common-mode voltages. (a) Switching-frequency time scale. (b) Fundamental-frequency time scale. (c) Spectra in the low-frequency range ($f_{sw} = 10 \text{ kHz}$; $f_o = 50 \text{ Hz}$).](image)

A. Common-Mode Voltage Detection and Generation Performance

First, the performance of the common-mode voltage detector and the generation of the compensating voltage by the four-level inverter will be investigated. Fig. 12 shows the waveforms
of common-mode voltages at various points in the system presented in both the fundamental- and switching-frequency time scales to highlight the triplen harmonic and switching frequency components, respectively. From Fig. 12(a), it is clear that the common-mode voltage $v_{\text{cm}}$ of the PWM inverter is correctly detected as can be compared with the voltage $v_{\text{com}}$ generated by the four-level inverter. The four-level inverter’s output voltage $v_{\text{com}}$ is then coupled to the secondary sides of the common-mode coupling transformer and becomes the final compensating voltage that cancels out the common-mode voltage $v_{\text{cm}}$ of the PWM inverter. From the common-mode voltage waveforms in Fig. 12(b) and their spectra emphasized in the low-frequency range in Fig. 12(c), it can be seen that the waveform of the common-mode voltage $v_{\text{mot}}$ at the motor terminals after compensation contains mainly the low-frequency components of the triplen harmonics that are generated by the space vector PWM and the diode rectifier. In this case, the dominant triplen voltages of both the inverter and the rectifier are at the same frequency of 150 Hz. These results indicate that although the four-level inverter generates the compensating voltage $v_{\text{com}}$, which duplicates the whole spectrum of the common-mode voltage $v_{\text{cm}}$ of the PWM inverter, only the high-frequency components of the common-mode voltage $v_{\text{com}}$ above 1 kHz are allowed to pass through the coupling transformer, while the low-frequency components are blocked by the series capacitor as designed. It can be concluded then that the proposed hybrid filter functions correctly as expected by canceling only the high-frequency common-mode voltage of the PWM inverter and leaving the harmless low-frequency components uncompensated.

Figs. 13 and 14 are the experimental results that illustrate how the hybrid active filter performs when the operating or the switching frequency is varied. In Fig. 13, the operating frequency is increased to 75 Hz. From the spectra of Fig. 13(c), the dominant triplen harmonic of the diode rectifier is still at 150 Hz, while that of the space vector PWM moves to 225 Hz, corresponding to the output frequency of 75 Hz. Similar to Fig. 12, the hybrid active filter still effectively suppresses the high-frequency components and leaves the triplen harmonic common-mode voltage uncompensated. The change in the operating frequency thus does not affect the performance of the active filter.

In the same manner, when the switching frequency is reduced to 5 kHz, the hybrid filter still works satisfactorily, as can be seen from the waveforms and the spectra shown in Fig. 14.

It is concluded then that the hybrid active filter works well even though the switching frequency may come close to the triplen harmonics of the PWM inverter and the diode rectifier. This is a significant advantage of the active filter over the conventional passive output filter, which normally suffers from the resonance excited by those frequency variations. The resonance excitation will normally result in poor filtering performance or even impossible filter design. It should be noted, however, that variation ranges of the switching and operating frequencies must be taken into consideration in the design stage to avoid the saturation of the common-mode transformer’s core.

Figs. 15 and 16 compare the resultant common-mode voltages measured at the motor terminals in various configurations,
Fig. 14. Performance of the hybrid filter when the switching frequency is varied ($f_{sw} = 5$ kHz; $f_o = 50$ Hz). (a) Switching-frequency time scale. (b) Fundamental-frequency time scale. (c) Spectra in the low-frequency range.

Fig. 15. Waveforms of common-mode voltages at the motor terminal $v_{mot}$ with (top) no filter, (middle) the pure active filter, and (bottom) the hybrid filter.

Fig. 16. Spectra of common-mode voltages at the motor terminal $v_{mot}$ with (top) no filter, (middle) the pure active filter, and (bottom) the hybrid filter.

As expected from the assistance of the additional passive filter. Viewed from the frequency domain, the spectra in Fig. 16 show that the pure active filter has a good attenuation roughly over the frequency range of 10 kHz up to 10 MHz, except around the vicinity of 100–700 kHz, where the resonance between the leakage inductance of the coupling transformer and the stray capacitance occurs. However, with the hybrid filter, the attenuation gain around this frequency range is further improved, and the measured spectra are in good consistency with the analysis results shown in Fig. 11.

As for the leakage current, the waveforms and spectra in Figs. 17 and 18 show that both the pure and the hybrid active filters can mitigate the leakage current efficiently. Again, better attenuation is achieved with the hybrid filter over the frequency range of 10 kHz to 10 MHz. Therefore, it is confirmed that the proposed hybrid filter can reduce both the common-mode voltage and the leakage current of the motor with good agreement with the analysis results.
The performance limitation of the hybrid filter above 10 MHz is due to the following: 1) the compensation delay time in real implementation and 2) the limited bandwidth of the common-mode transformer and the inductor $L_F$ because of the parasitic capacitances between the windings.

The effects of the compensation delay time of about 500 ns are illustrated in Fig. 19, where the response of the leakage current is focused during the switching instant of the four-level inverter. Fig. 19(a) is the case with no filters. In comparison to Fig. 19(a), Fig. 19(b) reveals that insertion of the coupling transformer alone reduces the leakage current as a result of the leakage inductance. The results when the active filter is activated are shown in Fig. 19(c). The 500-ns delay time is responsible for the increase of the oscillation of the leakage current, as compared to that of Fig. 19(b). Fig. 19(d) shows that the hybrid filter can attenuate the oscillation of the leakage current with the help of the additional passive filter, and the effects of the 500-ns delay time are alleviated.

C. Reduction of Shaft Voltage and Bearing Current

Measured waveforms of the shaft voltage and the bearing current without any filters are shown in Fig. 20. The motor shaft voltage of about 10 Vpeak can cause the electric field breakdown of a thin oil film existing between the bearing race and the balls. As a result of the breakdown, a high spike-like bearing current will flow and, in the long term, can result in bearing failure. Both the shaft voltage and the bearing current problems can be alleviated using the proposed hybrid filter. Fig. 21 shows that when the hybrid filter is used, the shaft voltage is well reduced and the bearing current spike is completely eliminated.
D. Prevention of Electric Shock Hazard

To investigate the effectiveness of the hybrid filter regarding prevention of electric shock hazard when the motor frame is ungrounded, the motor frame voltage is measured using the circuit network shown in Fig. 22. In Fig. 22, the measured object is the PWM inverter and the motor, with their connections to the Earth removed. In order to comply with the IEC 60335 standard, the frame voltage \( v_{12} \), as measured between terminals 1 and 2, must be less than 0.25 Vrms.

Measurement result in Fig. 23 is for the case when there is no filter installed. It is seen that the frame voltage \( v_{12} \) is induced by the common-mode voltage \( v_{\text{mot}} \) at the motor terminals, and in this case, the rms value of \( v_{12} \) is 0.45 V, which fails to comply with the standard. On the contrary, the waveforms of the common-mode voltage and the frame voltage when the hybrid filter is installed are shown in Fig. 24. Since with the hybrid filter, the common-mode voltage is filtered out, the frame voltage is also reduced. Although there remains some high-frequency ripple voltage, the rms value of the frame voltage \( v_{12} = 0.095 \) V is lower than the limitation of the standard. Therefore, it is concluded then that the hybrid filter can be used to prevent the electric shock hazard from the ungrounded motor frame as well.

VI. Conclusion

To cancel the common-mode voltage generated by the PWM inverter in the drive system, a novel hybrid structure of an active output EMI filter, which combines a pure active filter with a small passive filter is proposed in this paper. Noting that the harmful common-mode voltages are in the high-frequency range, the proposed hybrid active filter is connected in series with a blocking capacitor to focus its compensation only on the switching frequency and its higher sideband components. The core size of the common-mode coupling transformer is then tremendously reduced by a factor of about one-tenth as compared to that of the conventional active filter wherein the low-frequency common-mode voltage is also compensated. With the assistance of the additional passive filter, the attenuation characteristic of the active filter is further improved, and the resonance between the stray capacitance of the motor frame and the leakage inductance of the coupling transformer is suppressed. To support its real implementation, the design procedures of all relevant components of the hybrid filter are also given. Finally, the effectiveness of the hybrid active filter is verified by experiments. Further improvement of the filtering characteristic, particularly in the frequency range above 10 MHz, may be obtained by using ultra-high-speed electronic components to reduce the detection delay time and by refinement of the winding arrangement together with the use of a superior magnetic core material to broaden the bandwidth of the coupling transformer and the inductors.
REFERENCES


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