Acoustic Noise of Induction Motor With Low-Frequency Model Predictive Control

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ABSTRACT
This paper studies the acoustic noise of induction motor with low switching frequency model predictive control and compares its noise performance with the traditional pulse width modulation with proportional-integral controllers. First, the relationship among the air-gap magnetic field, the electromagnetic force, and the acoustic noise is modeled to describe the influence of control algorithm on the noise characteristics of motors. As the counterpart with the traditional pulse width modulation control, a finite control set model predictive control strategy with low switching frequency is proposed. The steady-state performance and switching frequency constraint ability of the proposed model predictive control strategy are verified on the experimental platform. Second, the noise spectrums of these two control strategies are sampled and compared in different operating conditions, and the influence of the weighting factor on the noise spectrum is also discussed. It is proved by the experimental results that model predictive control has better noise performance than the traditional sinusoidal pulse width modulation under the same switching frequency and the total harmonic distortion of stator current.

INDEX TERMS
Acoustic noise, motor control, model predictive control, pulse width modulation.

I. INTRODUCTION
The acoustic noise heard by human ears or captured by the acoustic detection device is one of the key aspects to evaluate the performance of a motor control system [1]. There are mainly three different types of noise according to the different noise sources. (1) The mechanical noise caused by the vibration of the machine body and rotor, (2) the ventilation noise generated by the cooling fan and rotor rotation, and (3) the electromagnetic noise transmitted in the form of high-frequency current and electromagnetic wave. Among these types of noise, the electromagnetic noise is considered as the most difficult one to avoid due to the nonlinearity of inverters in the variable-frequency motor drive system. Besides, this high-frequency noise is the main noise affecting people’s auditory perception [2].

The electromagnetic noise is generated by the switching process of power electronic devices. With the traditional pulse width modulation (PWM) control, the state switching of power electronic devices will generate voltage and current in a pulse-sequence form. Therefore, there are a large number of high-order time harmonics in stator current, which generate the high-speed rotating space harmonics in the air gap magnetic field with the frequency near the inverter’s switching frequency [3]. These space harmonics will affect the amplitude and frequency of the electromagnetic force wave in the air-gap magnetic field and may lead to the result that the frequency of the electromagnetic force is close to the natural frequency of the machine, and then causes resonance and increases the vibration and noise of the system.

Properly selecting the modulation and control strategy, the high-frequency electromagnetic noise in the drive system can be controlled within an acceptable range [4], [5]. Traditional noise suppression method usually adopts the modulation strategies that can reduce the current harmonics, especially the switching-frequency harmonics and its multiples, to suppress the high-frequency electromagnetic force. According to the Parseval theorem, the energy in the frequency domain is conserved as long as the energy in the time domain remains constant. This indicates that if the frequency band of the spectrum is extended, the peak value of the switching frequency components can be reduced and a fatter frequency
band usually has a better noise performance. At present, the main low noise modulation strategy for a two-level three-phase inverter system is random-frequency PWM [6]–[9]. It can spread the noise spectrum to a fatter form, so it is used on many occasions with strict noise requirements. However, the random-frequency PWM strategy may significantly deteriorate some performance of the system, such as increasing the stator current ripple and switching loss [10].

In addition to the modulation strategy, novel control algorithm can also improve the noise performance of motor control system. At present, the constrained control of induction motor is being developed from the traditional feedback-based control represented by the PI controller to optimize control such as model predictive control (MPC) [11]–[13]. With the improvement of the computing performance of digital processors and the pursuit of higher performance power electronic devices, MPC has attracted more and more attention and research in the field of power electronics [14]–[21].

According to whether there is modulation process or not, MPC in power electronics can be classified into two different types, the continuous control set model predictive control (CCS-MPC) and the finite control set model predictive control (FCS-MPC) [22]. CCS-MPC retains modulation as the traditional PWM strategy. Although it can improve the control accuracy, CCS-MPC will increase the complexity of the algorithm and increase the switching frequency of the system [23], [24]. By considering the limited states of the inverter, FCS-MPC does not need the modulation process and the control signals for inverter are directly output by minimizing the cost function in each control cycle. Compared with the traditional PI-SPWM strategy, MPC has many advantages. It is easy to contain nonlinear constraints, such as low switching frequency constraint, low loss constraint, and so on [25], [26]. In addition to these advantages, FCS-MPC has the inherent characteristic of the scattered current spectrum, which is beneficial to noise control [27].

In some high-power applications such as railway traction systems, low switching frequency is required to reduce the switching power loss and the noise is strictly limited to provide passengers with a comfortable environment. Random-frequency PWM cannot guarantee a fixed low switching frequency, while FCS-MPC is a better choice to undertake this task. To the best of our knowledge, there are few researches focus on the noise performance of MPC in low switching frequency railway traction system. R. Lazar presents a preliminary simulation study between traditional PI control with MPC [28]. The author concludes that the noise of MPC is nicely spread and more pleasant to hear than fixed switching frequency. But the low frequency humming is not present in the simulation results, and the practical sound measurements are not conducted. M. Kronei et. al. add a noise constraint in the cost function of MPC to achieve a better noise performance [29], but it requires simultaneous measurements of the currents and the sound pressure, which are not suitable for industrial applications.

In this paper, to study the noise characteristics of low switching frequency traction system, the relationship among air gap magnetic field, electromagnetic force, and acoustic noise is established by mathematical model. According to the noise model, three main sources of electromagnetic noise are analyzed. Then, the current prediction model of induction motor is provided and a model predictive control with switching frequency constraint is proposed. Finally, the noise spectrum of PI-SPWM and FCS-MPC are obtained with different speeds and switching frequencies. Through the comparative study of the two strategies, it can be concluded that MPC has better noise performance than the traditional sinusoidal pulse width modulation control. It can provide a better solution for the noise suppression of traction system.

II. SOURCE ANALYSIS OF ELECTROMAGNETIC NOISE
A. MODEL OF ELECTROMAGNETIC FORCE WAVE
To analyze and control the electromagnetic noise, it is necessary to classify the sources of electromagnetic noise at first.

The radial electromagnetic force $p_n(\theta, t)$, generated by the air-gap magnetic field, is the main course of electromagnetic noise. According to Maxwell’s law, the electromagnetic force wave acting on the surface of the stator is proportional to the square of the air-gap flux density $b(\theta, t)$ [30]. It can be defined by (1),

$$p_n(\theta, t) = \frac{b^2(\theta, t)}{2\mu_0} \quad (1)$$

where $\mu_0 = 4\pi \times 10^{-7} H/m$. Ignoring the influence of magnetic circuit saturation, $b(\theta, t)$ can be expressed as the product of air-gap magnetomotive force (MMF) $f(\theta, t)$ and air-gap permeance $\lambda(\theta, t)$.

$$b(\theta, t) = f(\theta, t)\lambda(\theta, t) \quad (2)$$

The air-gap MMF $f(\theta, t)$ of three-phase inverter-driven induction motors consists of the following parts [2].

$$f(\theta, t) = f_1 + f_k + f_v + f_\mu + f_\nu + f_\kappa + f_{\mu k}$$

$$f_1 = F_1 \cos(\varphi - \omega_1 t - \varphi_1)$$

$$f_k = \sum_{k \neq 1} F_k \cos(\varphi - \omega_k t - \varphi_k)$$

$$f_v = \sum_{\nu \neq \mu} F_v \cos(\varphi - \omega_\nu t - \varphi_\nu)$$

$$f_\mu = \sum_{k \neq 1} F_\mu \cos(\varphi - \omega_\mu t - \varphi_\mu)$$

$$f_\nu = \sum_{k \neq 1} F_\nu \cos(\varphi - \omega_\nu t - \varphi_\nu)$$

$$f_\kappa = \sum_{k \neq 1} F_\kappa \cos(\varphi - \omega_\kappa t - \varphi_\kappa)$$

$$f_{\mu k} = \sum_{k \neq 1} F_{\mu k} \cos(\varphi - \omega_\mu t - \varphi_\kappa)\quad (3)$$

$p$ is the pole pairs, $F$, $\omega$, and $\varphi$ are MMF, angular frequency, and phase angle respectively. Subscript 1 and $k$ represent fundamental and harmonic order, while $v$, $\mu$, $\kappa$, and $\mu k$ are slot harmonics produced by stator/rotor current fundamentals and harmonics.

The first two terms represent the fundamental and first-order components of the air-gap MMF produced by
stator current. The following four terms are high-order MMF harmonics produced by the fundamental and current harmonics of stator and rotor respectively.

The harmonic angular frequencies are [30]

\[
\omega_{mk} = \omega_k \left[ 1 + \frac{q_2}{p} \left( 1 - s_k \right) \right]
\]

(4)

where \( s_k \) is the slip of \( k \)th harmonic and

\[
s_k = (\omega_k - p\omega_r)/\omega_k
\]

(5)

Ignoring the small coupling between the slotting effect of rotor and stator, the air-gap permeance \( \lambda(\theta, t) \) can be simplified as (6) considering the slot of stator and rotor [2].

\[
\lambda(\theta, t) = \Lambda_0 \left( 1 + \frac{\lambda_1}{\Lambda_0} \right) \left( 1 + \frac{\lambda_2}{\Lambda_0} \right)
\]

\[
\approx \Lambda_0 + \sum \lambda_1 + \sum \lambda_2
\]

(6)

\( \Lambda_0 \) is the invariant part of the permeance.

When considering the stator slot and the rotor is regarded as smooth, the harmonic permeance is

\[
\sum \lambda_1 = \sum \Lambda_{k1} \cos k_1 Z_1 \theta
\]

(7)

While considering the slot of rotor and the stator is regarded as smooth, the harmonic permeance can be presented as

\[
\sum \lambda_2 = \sum \Lambda_{k2} \cos k_2 Z_2 [\theta - \omega_1/p (1 - s) t]
\]

(8)

\( Z_1 \) and \( Z_2 \) are the slot number of stator and rotor. \( \Lambda_{k1} \) and \( \Lambda_{k2} \) are the amplitude of stator and rotor slot permeance.

**B. ACOUSTIC NOISE SOURCE**

Vibration force wave, vibration response, and acoustic noise have the same frequency spectrum due to the homology of sound and vibration. Therefore, the source of noise can be classified by analyzing the coupling form of vibration force wave. To describe coarsely, the sinusoidal waves are represented as their amplitudes.

Neglecting the rotor eccentricity and magnetic circuit saturation, three main noise sources of the PWM-drive system can be obtained from the three coupling forms of vibration force wave.

1) SOURCE A

The vibration force produced by the coupling of stator/rotor MMF and slot permeance,

\[
\langle (F_1 \Lambda_0) + (F_1 \Lambda_{k1}) + (F_1 \Lambda_{k2}) \rangle
\]

The electromagnetic force wave generated by this coupling mode is mainly generated by the fundamental current which is similar to the wave generated by a sinusoidal power supply. Its frequency and order are as follows:

When vibration mode \( m^+ = \mu + \nu \),

\[
f = f_1 \left[ q_2 Z_2 \left( 1 - s_1 \right) + 2 \right]
\]

(9)

When vibration mode \( m^- = \mu - \nu \),

\[
f = f_1 \left[ q_2 Z_2 \left( 1 - s_1 \right) \right]
\]

(10)

\( s_1 \) is the slip of fundamental and \( q_2 = 0, \pm 1, \pm 2, \cdots \)

2) SOURCE B

The first order vibration force harmonic produced by fundamental and current harmonics,

\[
\langle (F_{k1} \Lambda_0), (F_{k2} \Lambda_0) \rangle
\]

where \( k_1 = 1 \) or \( k_2 = 1 \) represents the fundamental component.

When supplied by PWM inverter, this kind of noise is inevitable and cannot be suppressed by optimizing modulation or control strategy.

The electromagnetic force wave generated by the first-order component is large and distributed near the switching frequency and its multiples where

\[
f = |(\pm f_{k1}) - (\pm f_{k2})| \text{ and } m = 0
\]

(11)

where \( f_k \) is the frequency of \( k \)th harmonic and

\[
f_k = f_0 + f_0 \pm 4f_1, f_0 \pm 2f_1, f_0 \pm 3f_1, \cdots
\]

(12)

3) SOURCE C

The vibration force produced by high order filed harmonics,

\[
\langle (F_{1} \Lambda_{k1} \Lambda_{k2}), (F_{k1} \Lambda_{0}) \rangle, \quad \langle (F_{1} \Lambda_{k1} \Lambda_{k2}), (F_{k1} \Lambda_{0}) \rangle
\]

The electromagnetic force wave generated by this coupling mode is more important when the load is heavy with a low order vibration order. At this time, frequency of the electromagnetic force wave is

\[
f = \left\{ \frac{|(\pm f_k - f_1)|}{f_k \pm (f_1 + q_2 Z_2 \left( 1 - s_1 \right) f_1)}, \quad m = 0, 2 \right\}
\]

(13)

**C. NOISE ASSOCIATED WITH INVERTER POWER SUPPLY**

Noise sources B and C are associated with non-sinusoidal power supply brought by inverters.

Under light load, noise sources A and B are the main sources of the acoustic noise. Under heavy load, noise sources C also occurs. When the frequency and vibration mode of the electromagnetic force wave close to the natural frequency and match the mode order of the motor, the motor will produce large noise. From the above analysis, it can be found that the noise of PWM motor drive system is related to the system electromagnetic, mechanical, acoustic and control parameters, such as the number of motor pole pairs, slot number of stator and rotor, stator current frequency, PWM switching frequency, natural frequency, and vibration mode of the system.
Although the natural frequencies and vibration modes of the system are very rich, the analysis of noise source B shows that the main electromagnetic noise excitation generated by PWM inverter is the $m = 0$ pulsing mode, the component with the largest noise amplitude will appear near the switching frequency of the inverter. Therefore, by selecting appropriate control or modulation strategy, the resonance caused by noise source B can be mitigated and the noise can be then reduced.

III. PREDICTIVE MODEL AND FREQUENCY CONTROL

A. WORKING PRINCIPLE OF FCS-MPC

Unlike the traditional motor drive strategy with PI controller and pulse width modulation, FCS-MPC directly selects the switching combination from the candidate vectors.

There are mainly three parts in an FCS-MPC controller, including the current predictive model, the cost function, and the rolling execution process. In each control period, the predictive model predicts the current for each candidate switch combination based on the sampled values such as stator current and rotor speed. Then, the cost of each candidate will be calculated with the cost function. The switching combination that achieves the minimum cost will be selected as the optimum vector and output as the drive signals.

The FCS-MPC control diagram in a two-level three-phase inverter motor drive system is shown in Fig.1. The number of possible switching combinations is very small in this case. We use ‘1’ denotes the state that top transistor is ‘ON’ and ‘0’ denotes for ‘OFF’, there will be eight possible control states corresponding to switching combinations belongs to $S_{abc} = \{000, 001, 010, 011, 100, 101, 110, 111\}$.

The sampling values of stator current and rotor speed are obtained from the current and speed sensors. The external speed loop obtains the reference of the internal current control loop and then the switch combination with the smallest cost is selected among the eight candidates. Finally, the optimal switching combination is directly adopted to drive the switching devices of the inverter. The above steps are rolling executed in each control cycle to make the stator currents track their reference curves.

![FIGURE 1. Control diagram of the proposed model predictive control.](image)

B. CURRENT PREDICTIVE MODEL

The state-space equation for induction motor in the $\alpha$-$\beta$ coordinate system can be written as (14),

$$
\begin{align*}
\mathbf{p} &= \begin{bmatrix} i_{\alpha} \\ i_{\beta} \\ \psi_{ra} \\ \psi_{rb} \end{bmatrix} \\
&= A \begin{bmatrix} i_{\alpha} \\ i_{\beta} \\ \psi_{ra} \\ \psi_{rb} \end{bmatrix} + B \begin{bmatrix} u_{\alpha} \\ u_{\beta} \end{bmatrix} \\
A &= \begin{bmatrix} L_{11} & 0 & L_{13} & L_{14} \\ 0 & L_{11} & -L_{14} & L_{13} \\ L_{31} & 0 & L_{33} & L_{34} \\ 0 & L_{31} & L_{34} & L_{33} \end{bmatrix} \\
B &= \frac{-L_r}{L_m^2 - L_sL_r} \begin{bmatrix} 1 & 0 \\ 0 & 1 \\ 0 & 0 \\ 0 & 0 \end{bmatrix}
\end{align*}
$$

where

$$
L_{11} = (L_{2r}^2R_r + L_{2s}^2)/(L_r(L_m^2 - L_sL_r))
$$

$$
L_{13} = -L_mR_r/L_r(L_m^2 - L_sL_r)
$$

$$
L_{14} = -L_m\omega_r/L_m^2 - L_sL_r
$$

$$
L_{31} = L_mR_r/L_r
$$

$$
L_{33} = -R_r/L_r
$$

$$
L_{34} = -\omega_r
$$

$i$ and $u$ are the stator current and voltage. $\psi$ is the rotor flux. $L_m$, $L_s$, and $L_r$ are mutual inductance, stator, and rotor inductance respectively. $R_s$, $R_r$ are stator and rotor resistance.

The rotor speed can be regarded as constant within a control period since the sampling and control interval is far less than the electrical and mechanical time constant of the machine. By using the forward Euler approximation, the discretized state space equation can be expressed as

$$
\begin{align*}
\begin{bmatrix}
i_{\alpha}(k+1) \\
i_{\beta}(k+1) \\
\psi_{ra}(k+1) \\
\psi_{rb}(k+1)
\end{bmatrix} &= \left( A T_s + I \right) \\
&+ T_s BD \begin{bmatrix} S_{a}(k) \\ S_{b}(k) \\ S_{c}(k) \end{bmatrix}
\end{align*}
$$

$T_s$ is the sampling period, $I$ is the unit diagonal matrix, $Sa, Sb, Sc = \{0 \ ior \ 1\}$ is the switching sequence, and $D$ is the coordinate transform matrix where

$$
D = \frac{2}{3} U_{dc} \begin{bmatrix} 1 & -1/2 & -1/2 \\ 0 & \sqrt{3}/2 & -\sqrt{3}/2 \end{bmatrix}
$$

Equation (15) is the current predictive model with the input of load current, rotor speed, and the switch function of candidate vectors.

C. COST FUNCTION AND LOW SWITCHING FREQUENCY CONSTRAIN

The cost function is another key point of model predictive controller design. The control objective of the MPC controller is the same as that of the traditional PI controller.
which is to realize a good steady-state and dynamic performance of the internal current loop with the regulation of the external flux loop and speed/torque loop. A good steady-state performance requires that the current can track the amplitude and frequency of the reference accurately, while the dynamic performance refers to the fast-tracking when reference changes [21], [32]. Therefore, the cost function in MPC should be designed to achieve the minimum current ripple and fast response. It can be defined as the sum of the absolute value of the errors between the predicted current and the reference in the α-β stationary coordinate system,

$$g = |i^*_a(k + 1) - i_a(k + 1)| + |i^*_b(k + 1) - i_b(k + 1)|$$  \(17\)

In the motor drive application, we must make a balance between two conflicting objectives. On the one hand, the stator current distortions have to be kept as low as possible and the distortions can be mitigated by increasing the switching frequency in most of the cases. On the other hand, high switching frequency produces high power losses and stress on the semiconductor devices which means we need to put some restrictions on the switching frequency. Therefore, there is an unavoidable tradeoff between these two criteria. How to balance the harmonic distortion and switching frequency of stator current is an important issue in designing the cost function. The cost function proposed in this paper is (18).

$$g = |i^*_a(k + 1) - i_a(k + 1)| + |i^*_b(k + 1) - i_b(k + 1)| + \lambda * \Delta S$$  \(18\)

For low switching frequency applications, the switching frequency constraint is required. \(\lambda\) is the weighting factor, which is used to balance the current tracking performance and the switching frequency. \(\Delta S\) is the difference between the switching combination of the last control period and the candidate ones,

$$\Delta S = |S^k_A - S^{k-1}_A| + |S^k_B - S^{k-1}_B| + |S^k_C - S^{k-1}_C|$$  \(19\)

IV. ACOUSTIC NOISE ANALYSIS WITH PULSE WIDTH MODULATION AND MPC

To obtain the noise spectrum under different control strategies, both PI-SPWM and MPC control strategies are implemented on the experimental platform under different speeds and switching frequencies. It is proved in [2] that the resonance peaks of the noise spectrum of SPWM and SVPWM are identical. Therefore, we adopt the SPWM modulation strategy as the counterpart of MPC.

The digital signal processor TMS320f28377d is used as the controller to control a 10-kW inverter to drive the motor. The system is shown in Fig. 4 and the system parameters are shown in Table 1. The acoustic noise spectrum acquisition is realized by the LMS SCADAS mobile signal acquisition system and two high-precision microphones. The microphones are placed 20 cm above and in front of the controlled motor.
In the cost function, the weighting factor $\lambda$ is set as 0.001, and the THD of stator current phase a is 8.77%.

By adjusting the weighting factor $\lambda$, we can balance the current tracking performance and switching frequency. Fig.5 and Fig.6 show the waveform of stator current with different weighting factors. When $\lambda = 0.01$, compared with $\lambda = 0.001$ in Fig.4, the average switching frequency reduces to 450Hz and the THD of stator current increases to 11.82%. When $\lambda = 0.02$, the switching frequency further reduces to 350Hz and the THD increases to 12.54% accordingly.

It is worth noting that the moderating effect of the weighting factor is not linear. Due to the lack of theoretical guidance for the selection of $\lambda$, many experiments are conducted to find the optimal balance between switching frequency and current distortions. Results show that $\lambda = 0.001$ is a reasonable optimum value.

To verify the difference in noise characteristics of the motor under different control strategies and working conditions, experiments are carried out under different conditions. First of all, it should be pointed out that the noise of the motor is the sum of electromagnetic noise, mechanical noise, and ventilation noise. Since we only care about the high-frequency electromagnetic noise, we first test the background noise of the motor under the direct sinusoidal power supply and the noise spectrum is shown in Fig.7.

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**B. INFLUENCE OF THE WEIGHTING FACTOR ON NOISE SPECTRUM**

From the analysis in the previous section, it can be seen that different weighting factors will change the switching frequency of the system. To verify whether the different weighting factors will change the noise spectrum of the motor, different values of $\lambda$ are selected at 500 rpm and 800 rpm respectively, and the noise spectrum is compared and shown in Fig.8 and Fig.9.
Fig. 8 shows the noise spectrum when the rotor speed is 500 rpm and the weighting factors are 0.001 and 0.02 respectively. Fig. 9 is the noise spectrum when the rotor speed is 800 rpm with weighting factors of 0.001, 0.01, and 0.02 respectively. From the experimental results, it can be seen that although the switching frequency change, the amplitude and frequency distribution of the noise spectrum do not change greatly when choosing different weighting factors at different speeds. Therefore, it can be concluded that adjusting the weight factor in the MPC cost function within a certain range will not affect the noise response of the system. This is because, unlike PI-SPWM, FCS-MPC has no modulation process, and the spectrum of stator current is scattered. Although the average switching frequency of MPC changes, it will not affect the distribution of resonance peaks. So it will not change the frequency of vibration force wave of noise source B and then their noise spectrum will not change greatly.

C. NOISE RESPONSE DIFFERENCE ANALYSIS OF MPC AND PI-SPWM UNDER THE SAME CONDITIONS

To compare the noise response of the two different control strategies, a controlled test is carried out under the same working conditions. When the rotor speed is set as 800 rpm, the MPC weighting factor is set as 0.001, the corresponding average switching frequency is 950 Hz, and the switching frequency of PI-SPWM is 1 kHz. Other system parameters remain the same.

Fig. 10 and Fig. 11 are the results of the controlled experiment. The upper figures are stator currents, the middle figures are stator currents spectrum, and the lower figures are the noise spectrum. It can be seen from the comparison that
1) The switching frequency of PI-SPWM is 1000Hz whereas the switching frequency of MPC is 950Hz.
2) The stator current THD of PI-SPWM is 11.55% whereas MPC is 8.77%. This means that MPC can produce better stator current with lower switching frequency than PI-SPWM.
3) In the stator current spectrum, the peak magnitude of PI-SPWM is about 6.8% of fundamental whereas MPC
FIGURE 16. Noise spectrum comparison of PI-SPWM and MPC at rotor speed 1500 rpm.

The noise spectrums are also compared in low-frequency band (0-1.2kHz) and high-frequency band (3k-20kHz) in Fig.12 and Fig.13 respectively.

It can be seen that, due to the dispersion of the current spectrum of MPC, it may be easy to resonate with the natural frequency of the motor in the low-frequency band, so the noise of MPC in the low-frequency band is slightly higher than that of PI-SPWM.

The human ear is more sensitive to higher frequency noise, whereas in the high-frequency band, MPC can significantly reduce the decibel number, so we can conclude that MPC has better noise performance than the traditional PWM control strategy within the entire speed range.

D. THE NOISE SPECTRUM DIFFERENCE BETWEEN MPC AND SPWM AT DIFFERENT SPEEDS

The noise peak value of MPC and PI-SPWM at different rotor speeds are shown in Fig.17. It can be seen from the results that the noise amplitude of MPC is much lower than that of PI-SPWM at lower rotor speed, but their noise amplitudes are close to each other at higher speed.

V. CONCLUSION

This paper compares the noise difference of two control strategies in the field of motor control. It can be concluded that MPC has better noise performance than the traditional PWM control which can be reflected in the following aspects.

1) The noise peaks of MPC and SPWM appear near the switching frequency, but the noise spectrum of SPWM has many ripples which are multiple of switching frequency, while MPC does not have.
2) The noise spectrum of MPC is smoother within the entire speed range.
3) Adjusting the weighting factor of MPC will not affect the noise spectrum.
4) The noise performance advantage of MPC is more obvious in the high-frequency band.

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