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Robust Two Degree of Freedom Regulators for Velocity Ripple Elimination of AC Permanent Magnet Motors[1]

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Abstract

A general AC PM (Permanent Magnet) motor control system consists of a motion controller, a current tracking amplifier, a feedback encoder and the motor itself. The motion controller generates two analog commands to the current tracking amplifier and the three phase currents are reproduced at the motor terminals. However, DC offsets are always present at the motor terminals due to the DAC (Digital to Analog Converter) offsets of the motion controller and the current sensor offsets of the current tracking amplifier. These current offsets generate sinusoidal torque disturbance and hence produce velocity ripples. Such a disturbance cannot be rejected by using a simple PI (Proportional plus Integral) control. Furthermore, the current offsets drift with time and temperature so that an off-line compensation does not work satisfactorily. In this paper, a robust 2DOF (Two Degree of Freedom) regulator containing the internal model of the sinusoidal disturbance is proposed to accomplish disturbance rejection and constant speed tracking.

1 Introduction

Precision speed control systems are crucial in numerous industrial applications. For example, one typical application can be found in the feed control of machine tools in the manufacturing industry, where accurate smooth position and speed control are required for contour accuracy and small surface roughness of the products [1].

AC PM (Permanent Magnet) motors are attractive candidates for high performance industrial control applications such as the one stated above. In general, an AC PM motor control system consists of a motion controller, a current tracking amplifier, a feedback encoder and the motor itself. Figure 1 shows the block diagram of a typical AC PM motor control system.

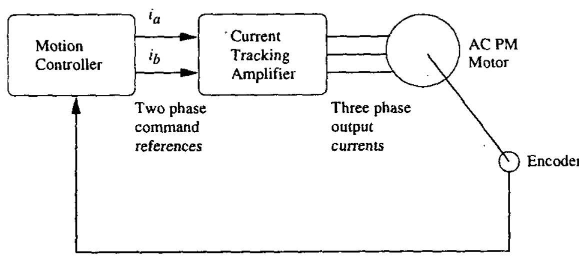

Figure 1: AC PM Motor Control System

However, DC offsets are always present at the motor terminals due to the DAC (Digital to Analog Converter) offsets of the motion controller and current sensor offsets of the current tracking amplifier. These current offsets generate sinusoidal torque disturbance and hence produce velocity ripples. Such a disturbance cannot be rejected by using a simple PI (Proportional plus Integral) control. Furthermore, the current offsets drift with time and temperature so that an off-line compensation does not work satisfactorily. In [2], an adaptive scheme based on the Lyapunov function method was developed to estimate the amplitudes of the periodic disturbances and then use the amplitude information to minimize the torque ripples of an AC PM motor. In [4], another adaptive scheme was developed to first identify the amplitude and the phase of the periodic disturbance and then use this information to cancel the repetitive vibrations. In this paper, a simple but effective method, based on a 2DOF (Two Degree of Freedom) control structure and IMP (Internal Model Principle) [5], is employed to solve the problem of robust disturbance rejection and tracking without estimating the amplitude and the phase of the sinusoidal disturbance explicitly.

The paper is organized as follows. Section 2 gives a brief review on the vector control of AC PM motors and the current offset disturbance modeling. In Section 3, the use of the IMP is proposed to solve the sinusoidal disturbance problem. Then a systematic 2DOF pole-zero placement controller design procedure

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is given to achieve a desired tracking performance and reject the sinusoidal disturbance simultaneously. Section 4 presents the simulation results of the proposed method. In Section 5, experimental results are compared with the simulation results to validate our control methodologies. Some concluding remarks are given in Section 6.

2 Vector Control and Disturbance Modeling

A three phase AC PM motor can be modeled in $d-q$ frame by the following equations [6]:

$$v_d = R_s i_d + \frac{d(L_d i_d + \lambda_m)}{dt} - \omega_e L_q i_q \quad (1)$$

$$v_q = R_s i_q + \frac{d(L_q i_q)}{dt} + \omega_e (L_d i_d + \lambda_m) \quad (2)$$

$$\tau_e = \frac{3}{2}\frac{P}{2}[\lambda_m i_q - (L_q - L_d) i_d i_q] \quad (3)$$

$$\tau_e - \tau_l = J_m \frac{d\omega}{dt} + B_m \omega \quad (4)$$

where the parameters and variables have the following meanings:

R_s	stator winding resistance
P	number of poles (even number)
L_d, L_q	$d-q$ frame stator inductances
J_m	moment of inertia
B_m	friction constant
λ_m	constant magnetic flux
v_d, v_q	$d-q$ frame stator voltages
i_d, i_q	$d-q$ frame stator currents
τ_e	electro-mechanical torque
τ_l	load torque
ω	rotor mechanical speed
$\omega_e = \frac{P}{2}\omega$	rotor electrical speed

If a sufficiently fast current tracking loop is used, Equations (1) and (2) can be eliminated. In this case, i_d and i_q become the system inputs. Furthermore, the vector control technique suggests to set $i_d = 0$. This converts the nonlinear AC PM motor system into a linear system:

$$\tau_e = \frac{3}{2}\frac{P}{2}\lambda_m i_q \quad (5)$$

$$\tau_e - \tau_l = J_m \frac{d\omega}{dt} + B_m \omega. \quad (6)$$

When the motor current amplifier is connected to the power source and the two current reference commands from the motion controller are kept at zero, a DC offset current induced by current sensor offsets and the motion controller DAC offsets may be present in one or both of the closed loop controlled phases and thus also in the third one [3]. Let I_a, I_b be the two DC current offsets present at the motor terminals due to motion controller DAC offsets and the current amplifier sensors offsets. Then $I_c = -(I_a + I_b)$ is the third phase current offset. Let i_a^*, i_b^* and i_c^* be the desired currents at the motor terminals. When the three phase currents with offsets enter into the motor, the actual currents, i_q, i_d and i_0 can be computed by using the Park (abc frame to $dq0$ frame) transformation [9] as follows:

$$[\, i_q \quad i_d \quad i_0 \,]' = \frac{2}{3}\begin{bmatrix} \cos(\theta_e - \frac{\pi}{2}) & \cos(\theta_e - \frac{7\pi}{6}) & \cos(\theta_e + \frac{\pi}{6}) \\ \sin(\theta_e - \frac{\pi}{2}) & \sin(\theta_e - \frac{7\pi}{6}) & \sin(\theta_e + \frac{\pi}{6}) \\ \frac{1}{2} & \frac{1}{2} & \frac{1}{2} \end{bmatrix} \times [\, i_a^* + I_a \quad i_b^* + I_b \quad i_c^* + I_c \,]' \quad (7)$$

where $\theta_e - \frac{\pi}{2}$ is the commutation angle for an AC PM motor and $\theta_e(t) = \theta_e(0) + \int_0^t \omega_e(t)dt$ is the electrical angle. Then i_q is equal to:

$$i_q = i_q^* + i_{qoff} \quad (8)$$

where

$$i_q^* = \frac{2}{3} i_a^* \cos(\theta_e - \frac{\pi}{2}) + \frac{2}{3} i_b^* \cos(\theta_e - \frac{7\pi}{6}) + \frac{2}{3} i_c^* \cos(\theta_e + \frac{\pi}{6}) \quad (9)$$

is the desired current value at q axis. The disturbance term i_{qoff} is equal to:

$$i_{qoff} = \frac{2}{3} I_a \left[\cos(\theta_e - \frac{\pi}{2}) - \cos(\theta_e + \frac{\pi}{6})\right] + \frac{2}{3} I_b \left[\cos(\theta_e - \frac{7\pi}{6}) - \cos(\theta_e + \frac{\pi}{6})\right]. \quad (10)$$

Similarly, for the d axis current,

$$i_d = i_d^* + i_{doff} \quad (11)$$

where i_d^* is the desired current value at d axis and is equal to 0 by employing vector control. The remaining disturbance term is equal to:

$$i_{doff} = \frac{2}{3} I_a \left[\sin(\theta_e - \frac{\pi}{2}) - \sin(\theta_e + \frac{\pi}{6})\right] + \frac{2}{3} I_b \left[\sin(\theta_e - \frac{7\pi}{6}) - \sin(\theta_e + \frac{\pi}{6})\right]. \quad (12)$$

Finally i_0 in Equation (7) is still equal to zero since $i_a^* + i_b^* + i_c^* = 0$ and $I_a + I_b + I_c = 0$. It follows from Equation (3) that

$$\tau_e = \tau_e^* + \tau_{off} \quad (13)$$

where

$$\tau_e^* = \frac{3}{2}\frac{P}{2}\lambda_m i_q^* \quad (14)$$

is the desired torque and the offset torque is equal to

$$\tau_{off} = \frac{3}{2}\frac{P}{2}\lambda_m i_{qoff} - \frac{3}{2}\frac{P}{2}\lambda_m (L_q - L_d)(i_{doff} i_{qoff} + i_q^* i_{doff}). \quad (15)$$

For surface PM rotor type AC servo motors, which are popular in various industrial applications, we have $L_d = L_q$ [10]. With this nice property, Equation (15) can be further reduced to:

$$\tau_{off} = \frac{3}{2}\frac{P}{2}\lambda_m i_{qoff}. \qquad (16)$$

Substituting Equation (10) into Equation (16), we get:

$$\tau_{off} = \frac{P}{2}\lambda_m I_a \left[\frac{3}{2}\sin(\theta_e) - \frac{\sqrt{3}}{2}\cos(\theta_e)\right] - \frac{P}{2}\lambda_m I_b \left[\sqrt{3}\cos(\theta_e)\right]. \qquad (17)$$

When the motor is performing a constant speed tracking with reference ω_r, $\omega_e(t)$ can be approximated, at least after certain transient period, by a constant $\omega_d = \frac{P}{2}\omega_r$. This shows that τ_{off} can be approximated by a sinusoidal function $\tau_{off} = A_d\cos(\omega_d t - \phi_d)$, where A_d is the magnitude of the disturbance while ϕ_d is the phase of the disturbance.

For other PM rotor type AC servo motors with $L_d \neq L_q$, the offset torque, τ_{off}, in Equation (17) will contain extra terms with phase $2\theta_e$ because of the multiplication of i_{doff} and i_{qoff}; however, the amplitudes of these terms are usually small due to the multiplication of two small quantities together and could be neglected when compared the terms with phase θ_e.

In summary, after employing the vector control and the formulation of the sinusoidal disturbance, the model of a vector controlled AC PM motor is given by Figure 2. Here, $u = i_q^*$ is the input current, $y = \omega$ is the output velocity, $K_t = \frac{3}{2}\frac{P}{2}\lambda_m$ is the equivalent torque constant, $d = \frac{\tau_{off}+\tau_l}{K_t}$ where τ_l is the load torque which can be considered as an unknown constant disturbance, τ_{off} is the torque disturbance due to current offsets which can be approximated by a sinusoidal function with known frequency ω_d and unknown magnitude and phase. Our goal is to design a good controller so that the output speed tracks a constant reference and rejects the disturbance τ_{off} and τ_l. Such a good controller is required to be robust, i.e. to perform the tracking and disturbance rejection even when the the system parameters vary slightly, to have good transient response, and to have a simple structure, i.e. to have an order as small as possible.

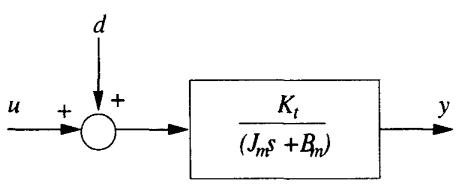

Figure 2: System Model with Disturbances

3 Controller Design

The problem to accomplish robust tracking and disturbance rejection is called a robust regulator problem. The key idea to solve a robust regulator problem is, based on the IMP, to have the controller to include the modes of the reference and disturbance. We also propose to use 2DOF controller structure to achieve better transient responses and easier designs. A 2DOF controller has a structure shown in Figure 3. One of its advantages, in comparison with the usual one degree of freedom or unity feedback structure, is that the tracking performance depends mainly on K_1, and the robustness and the disturbance rejection performance depends only on K_2. Hence K_1 and K_2 can be designed with different consideration. We also give a simple yet systematic pole-zero placement design procedure for the robust 2DOF regulators which yields low order controllers.

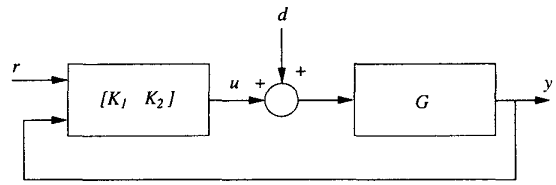

Figure 3: 2DOF Controller

3.1 Robust 2DOF Regulators and Pole-Zero Placement Design

Robust 2DOF regulators were discussed in [7] and [8], in which the disturbance and reference are assumed to have the same modes. In the following, we assume that they may have different modes, which may lead to simpler controllers. The main purpose of this section is to develop a simple pole-zero placement design method for robust 2DOF regulators.

Let G be a SISO (Single Input Single Output) plant described by a strictly proper transfer function $G(s) = \frac{b(s)}{a(s)}$, where

$$a(s) = s^{n_a} + a_1 s^{n_a-1} + \ldots + a_{n_a} \qquad (18)$$

$$b(s) = b_1 s^{n_a-1} + b_2 s^{n_a-2} + \ldots + b_{n_a} \qquad (19)$$

and it is assumed that $a(s)$ and $b(s)$ are coprime. The controller can be written as:

$$[K_1(s) \ K_2(s)] = \frac{1}{k(s)}[q(s) \ h(s)] \qquad (20)$$

where

$$k(s) = s^{n_k} + k_1 s^{n_k-1} + \ldots + k_{n_k} \qquad (21)$$

$$q(s) = q_0 s^{n_k} + q_1 s^{n_k-1} + \ldots + q_{n_k} \qquad (22)$$

$$h(s) = h_0 s^{n_k} + h_1 s^{n_k-1} + \ldots + h_{n_k}. \qquad (23)$$

The 2DOF control structure then becomes as shown in Figure 4. Let the unstable modes of r be the roots of monic polynomial $m_r(s)$ and those of d be the roots of $m_d(s)$. Let the least common multiple of $m_r(s)$ and $m_d(s)$ be $m(s)$. Then it is well-known that the robust regulator problem is solvable, i.e. it is possible to design a controller so that the disturbance rejection and reference tracking are achieved, if and only if $m(s)$ and $b(s)$ are coprime. Now assume that this condition is satisfied. Then according to the IMP, a solution to the robust regulator problem must satisfy

$$k(s) = g(s)m(s) \tag{24}$$

and

$$h(s) - q(s) = f(s)m_r(s) \tag{25}$$

where $g(s)$ and $f(s)$ are polynomials. Therefore, the design of the robust 2DOF regulator amounts to the determination of polynomials $f(s)$, $g(s)$ and $h(s)$. A simple yet systematic way to design $g(s)$ and $h(s)$ is to use pole placement. Denote

$$g(s) = s^{n_g} + g_1 s^{n_g - 1} + \dots + g_{n_g}. \tag{26}$$

The closed loop characteristic polynomial is

$$\begin{aligned} k(s)a(s) + h(s)b(s) &= \\ g(s)m(s)a(s) + h(s)b(s) &:= \delta(s). \end{aligned} \tag{27}$$

Notice that $\delta(s)$ is a monic polynomial of degree $n_g + n_m + n_a$. The purpose of pole placement is to design $g(s)$ and $h(s)$ so that the roots of $\delta(s)$ are arbitrarily assigned. Since $m(s)a(s)$ and $b(s)$ are coprime, it can be easily shown that the pole placement is possible if and only if $n_g \geq n_a - 1$. If this condition is satisfied, the coefficients of $g(s)$ and $h(s)$ can be obtained by equating the coefficients of both sides of Equation (27). To minimize the controller complexity, we can choose $n_g = n_a - 1$.

After $g(s)$ and $h(s)$ are designed, the closed loop transfer function from r to y becomes

$$\frac{y(s)}{r(s)} = \frac{q(s)b(s)}{\delta(s)} = \frac{[h(s) - f(s)m_r(s)]b(s)}{\delta(s)} \tag{28}$$

We wish to design $f(s)$ so that the undetermined n_k zeros of this transfer function, i.e., the roots of $h(s) - f(s)m_r(s)$, can be assigned in a desirable way. In general, $f(s)$ does not have enough degree of freedom to assign the roots of $h(s) - f(s)m_r(s)$. However, if the reference is a step function, then $m_r(s) = s$ and the $n_k - 1$ coefficients of $f(s)$ can be used to assign all but one coefficients of $h(s) - f(s)m_r(s)$ and hence all roots of $h(s) - f(s)m_r(s)$. Therefore, in the case when the reference is a step function, n_k zeros of $\frac{y(s)}{r(s)}$ can be arbitrarily assigned.

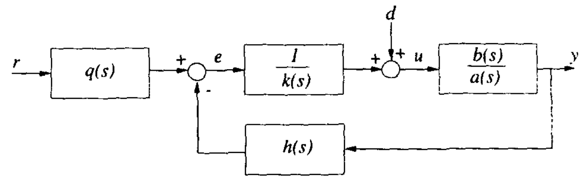

Figure 4: 2DOF Pole-Zero Placement Regulator

3.2 Design for the PM Motor Control System

For our PM motor control system, in reference to Figure 4, we have $a(s) = s + \frac{B_m}{J_m}$ and $b(s) = \frac{K_t}{J_m}$. Since r is a step reference, it follows that $m_r(s) = s$. Since d contains a sinusoidal function of frequency $\omega_d = \frac{P}{2}\omega_r$ and a constant function, it follows $m_d(s) = s(s^2 + \omega_d^2)$. Therefore, $m(s) = s(s^2 + \omega_d^2)$. Since deg $a(s) = 1$, we can choose $n_g = 0$ as $n_a - 1 = 0$. This leads to a controller of order equal to deg $m(s)$, which is the lowest possible to achieve robust regulator. Hence

$$k(s) = m(s) = s(s^2 + \omega_d^2) \tag{29}$$

and $h(s)$ and $q(s)$ have the following forms:

$$\begin{aligned} h(s) &= h_0 s^3 + h_1 s^2 + h_2 s + h_3 && (30) \\ q(s) &= q_0 s^3 + q_1 s^2 + q_2 s + q_3 && (31) \\ &= h(s) - f(s)s && \\ f(s) &= f_0 s^2 + f_1 s + f_2. && (32) \end{aligned}$$

Choose the four closed loop poles of the system according to transient response specification so that the closed loop characteristic polynomial is

$$\delta(s) = s^4 + \delta_1 s^3 + \delta_2 s^2 + \delta_3 s + \delta_4. \tag{33}$$

Then by equating the coefficients of both sides of

$$\delta(s) = k(s)a(s) + b(s)h(s) \tag{34}$$

we can obtain

$$\begin{aligned} h_0 &= \frac{J_m}{K_t}\left(\delta_1 - \frac{B_m}{J_m}\right), \quad & h_1 &= \frac{J_m}{K_t}\left(\delta_2 - \omega_d^2\right) \\ h_2 &= \frac{J_m}{K_t}\left(\delta_3 - \omega_d^2 \frac{B_m}{J_m}\right), \quad & h_3 &= \frac{J_m}{K_t}\delta_4 \end{aligned} \tag{35}$$

Finally, choose $f(s)$ so that the three zeros of the transfer function from r to y, which is $\frac{h(s) - f(s)s}{\delta(s)}$, cancels three of its poles. In this way, the system from r to y is turned to a first order system with a pole determined by the remaining pole of $\delta(s)$.

4 Simulation Results

A 200W AC PM motor is used in our simulations and experimental tests. The motor parameter is listed in Table 1.

To show our controller effectiveness, a traditional PI controller is always used for comparison. In addition, a -0.1A current offset is assumed presenting at phase 1 of the motor terminal and a 0.05A current offset is assumed presenting at phase 2 of the motor terminal.

Table 1: Motor Parameter

J_m	$0.144 \times 10^{-4}\text{kg}\cdot\text{m}^2$
B_m	$5.416 \times 10^{-4}\text{Nm/rad}\cdot\text{s}^{-1}$
λ_m	0.0283Wb
L_d, L_q	11.5mH
P	8
$K_t = \frac{3}{2}\frac{P}{2}\lambda_m$	0.1698Nm/A
Encoder resolution	8000 counts/rev

4.1 Simulation Results for Robust 2DOF Regulator

Assume that the reference speed is $\omega_r = 100$ rpm $= 10.472$ rad$\cdot$s^{-1}. Then $\omega_d = \frac{8}{2} \times \omega_r = 41.88$ rad$\cdot$s^{-1}. Suppose four closed loop poles are chosen to be -40, -50, -60 and -80. Furthermore, our overall desired closed loop system is a first order system with a pole at -80. According to the design procedure in Section 3.2, the final controller is equal to

$$\begin{aligned} k(s) &= s(s^2 + 1754.6) \\ h(s) &= 0.0165s^3 + 1.4964s^2 + 55.0641s + 814.1343 \\ q(s) &= 0.0068s^3 + 1.0177s^2 + 50.2049s + 814.1343. \end{aligned}$$

The upper figure of Figure 5 shows the simulation results comparison between the proposed robust 2DOF regulator and a PI controller with $k_p = 0.01$ and $k_i = 0.08$. It can be observed that the velocity ripples due to DC current offset can be rejected by the proposed controller completely. Another controller is

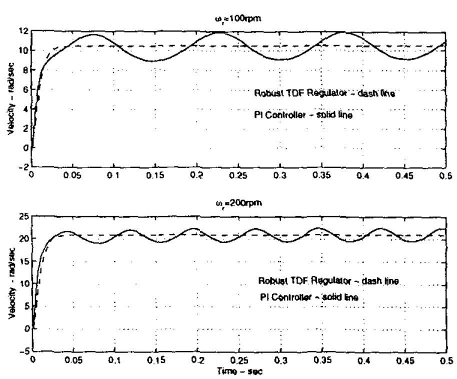

Figure 5: Simulation Results Comparison between Robust 2DOF Regulator and PI Controller, upper figure - $\omega_r = 100$ rpm, upper figure - $\omega_r = 200$ rpm

designed for $\omega_r = 200$ rpm. Again, the final controller is obtained by following the design procedure in Section 3.2:

$$\begin{aligned} k(s) &= s(s^2 + 7018.4) \\ h(s) &= 0.0165s^3 + 1.05s^2 + 39.1115s + 814.1343 \\ q(s) &= 0.0068s^3 + 1.0177s^2 + 50.2049s + 814.1343. \end{aligned}$$

The lower figure of Figure 5 shows the simulation results comparison between the proposed robust 2DOF regulator and a PI controller with $k_p = 0.01$ and $k_i = 0.08$. The velocity ripples at $\omega_r = 200$ rpm are also eliminated by the proposed controller.

In summary, the robust 2DOF regulator can reject the sinusoidal disturbance with guaranteed output tracking response. On the contrary, a PI speed controller fails to reject the sinusoidal disturbance. This is expected since the PI controller doesn't contain all modes of the disturbance.

5 Experimental Results

Experiments are performed to verify the effectiveness of our proposed controllers. Figure 1 shows the basic setup of our experiment. A dSAPCE DS1102 DSP controller board is used as our motion controller. In connection with MATLAB real time workshop and SIMULINK, a fast prototyping working environment is achieved and hence the code development time can be saved. The DSP controller implements all control algorithms with a sampling frequency 2kHz. In every control cycle, the controller reads the motor encoder, performs the control algorithm calculation and then outputs two current reference commands i_a and i_b to the current tracking amplifier. An Advanced Motion Controls Inc. S30A40B current tracking driver is used and the three phase AC PM motor is from Sanyo Denki with the parameters listed in Table 1.

Two experiments are conducted with command reference speeds equal to 0.4 Hz square waves with amplitudes 100 rpm $= 10.472$ rad$\cdot$s^{-1} and 200 rpm $= 20.944$ rad$\cdot$s^{-1}. The poles are chosen as in Section 4, i.e. -40,-50,-60 and -80, and the desired overall system transfer function is a first order system with the pole at -80. The upper figure of Figure 6 shows the velocity output when only a PI controller with $k_p = 0.01$ and $k_i = 0.08$ is used. It is clear that the output velocity contains ripples with the peak value equals to 30% of the command value. The lower figure of Figure 6 shows the velocity output when the robust 2DOF regulator is used. The velocity output ripples are reduced to the motor encoder resolution, which is, the best we can do. In our system, the motor encoder resolution is equal to 8000 counts/rev and our servo loop sampling frequency is 2kHz; therefore, the smallest velocity ripples are equal to $\frac{2000 \times 2 \times \pi}{8000} = 1.5708$ rad$\cdot$s^{-1}. The output velocity ripples are believed to be further reduced if a higher resolution encoder is used.

The upper figure of Figure 7 shows the velocity output at 200 rpm when only a PI controller with $k_p = 0.01$ and $k_i = 0.08$ is used, in which, the velocity ripples are present. On the contrary, The lower figure of Fig-

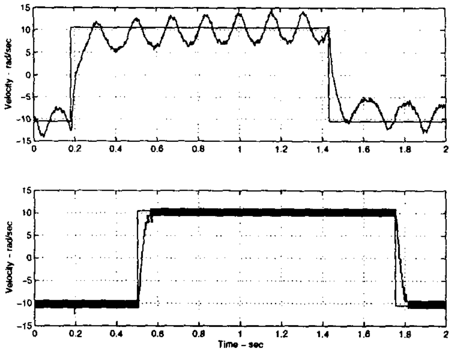

Figure 6: Experimental Velocity Output - $\omega_r = 100$ rpm, upper figure - PI Controller, lower figure - Robust 2DOF regulator

ure 7 shows the velocity output at 200 rpm when the robust 2DOF regulator is used. Both the sinusoidal disturbance rejection and tracking performance specifications can be met.

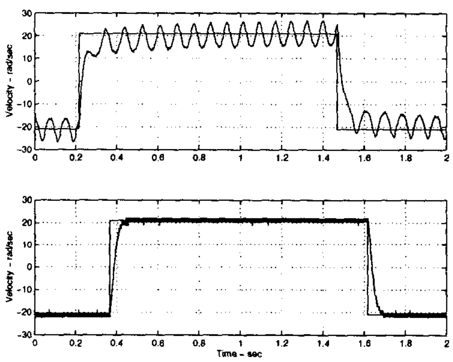

Figure 7: Experimental Velocity Output - $\omega_r = 200$ rpm, upper figure - PI Controller, lower figure - Robust 2DOF regulator

Our experimental results match well with the simulation results in the last section. They validate that the controller proposed in Section 3 can reject the sinusoidal disturbance and achieve a desired output tracking performance at the same time.

6 Conclusions

In this paper, the robust 2DOF regulator for AC PM motors based on the IMP is demonstrated to be an effective method to eliminate the velocity ripples that are created by DC current offsets. Furthermore, by following our proposed systematic robust 2DOF regulator design method, both the velocity tracking and the disturbance rejection requirement can be achieved simultaneously. A velocity ripple-free output is crucial to some constant speed requirement applications such as machine feed control and assembly line application, etc.

In this paper, a pole-zero placement technique is employed in the robust 2DOF regulator design. Currently we are also studying the applications of optimal design methods such as $\mathcal{H}_2$ and $\mathcal{H}_\infty$ methods, and the results will be reported soon.

Notice that the controller design in Section 3 depends on the reference speed, ω_r. In the pole-zero placement design of the robust 2DOF regulator for different speed references, the only controller parameters that need to be adjusted are h_1 and h_2 in Equation (35) and one coefficient of $k(s)$ in Equation (29). This reveals a potential method to implement the adaptive version of the proposed controller for online adjustment in response to the change in speed reference. We are also working on this direction so that an adaptive controller can be developed for varying speed reference.

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