15th Triennial World Congress, Barcelona, Spain

# A PLUG-IN ROBUST COMPENSATOR FOR IFOC INDUCTION MACHINE DRIVES

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Abstract: It is well-known that the system performance for an indirect field-oriented control (IFOC) induction motor drive degrades under parameter variations and in the presence of external load torque. In this paper, a novel plug-in robust compensator for position control enhancement of an indirect field-oriented induction machine drive is developed. This plug-in compensator, designed using the $\mathcal{H}_\infty$ loop shaping techniques, can be plugged into the existing controller without affecting the already satisfactory nominal tracking performance of the existing closed loop system, but can improve the system performance under plant parameter variations and in the presence of external disturbances.

Keywords: Induction machines, Position control, Robust control, Robust performance, Disturbance rejection

## 1. INTRODUCTION

A robust motor control system should exhibit good speed and position tracking and disturbance rejection responses even under plant parameter variations. Although indirect field-oriented control (IFOC) can transform a nonlinear induction motor into a linear system (Novotny and Lipo, 1998), it is well-known that under IFOC, the output response is sensitive to the plant parameter variations such as rotor resistance change (Nordin *et al.*, 1985). Different approaches have been applied to tackle the plant parameter variation problem.

In (Kao and Liu, 1992), $\mathcal{H}_\infty$ mixed-sensitivity optimization is applied to solve the plant parameter variation problem; however, weighting function selection is not an easy task and the order of the final controller, which is designed by $\mathcal{H}_\infty$ mixed-sensitivity optimization, is usually high. In (Heber *et al.*, 1997), a fuzzy logic control approach is employed to enhance the system robustness but the design of proper fuzzy rules is not straightforward and the knowledge from a skillful tuning operator may be needed to generate meaningful fuzzy rules. In (Ahmed-Ali *et al.*, 1998), an adaptive scheme is developed to estimate the rotor resistance so that the output speed performance can still be guaranteed when the rotor resistance changes during operation. However, the zero external torque assumption required by this adaptive controller may not be valid in real applications and its computational burden may be too demanding for a low-cost DSP/micro-controller. Furthermore, the tracking performance and the system robustness cannot be designed separately using the above controller design methodologies. As in many cases, the nominal/existing controllers for tracking performance are in operation or already designed properly by employing a low-order and simple controller such as PD, PI, PID or lead/lag compensators. However, these controllers may not have sufficient robustness against external disturbances or plant parameter variations such as the rotor resistance change in an IFOC induction motor system; hence, a good way to solve this problem is to design a plug-in compensator which can enhance the system robustness without affecting the nominal tracking performance. In this paper, a novel plug-in robust compensator, designed using $\mathcal{H}_\infty$ loop shaping techniques, is introduced so that the system robustness can be improved under plant parameter variations and

[1] This work is supported by Hong Kong Research Grants Council

in the presence of external disturbances. In addition, the proposed compensator does not affect the nominal tracking performance and can coexist with the nominal or existing controller without affecting the system stability.

## 2. IFOC OF INDUCTION MOTORS

The modeling and IFOC of induction motors are reviewed in this section. If a fast current-regulated voltage source inverter is employed in an induction motor drive, the stator currents become the new control input variables. IFOC is an effective linearization control algorithm for a highly nonlinear induction motor (Novotny and Lipo, 1998). With the rotor flux oriented at the $d$ axis, i.e. $\lambda_{qr}^e = 0$, the nonlinear motor equations can be transformed onto the $dq$ synchronous frame and described by the following equations:

$$L_r i_{qr}^e + L_m i_{qs}^e = 0 \quad (1)$$

$$L_r i_{dr}^e + L_m i_{ds}^e = \lambda_{dr}^e \quad (2)$$

$$R_r i_{dr}^e + p\lambda_{dr}^e = 0 \quad (3)$$

$$R_r i_{qr}^e + (\omega_e - \omega_r)\lambda_{dr}^e = 0 \quad (4)$$

$$\begin{aligned}\tau_e &= (3PL_m/4L_r)\lambda_{dr}^e i_{qs}^e \\ &= J_m p\omega_{rm} + B_m\omega_{rm} + \tau_l \quad (5)\end{aligned}$$

$$(\omega_e - \omega_r) = (R_r L_m i_{qs}^e / L_r \lambda_{dr}^e) \quad (6)$$

where

| | |
|---|---|
| $i_{ds}^e$,$i_{qs}^e$ | $dq$ synchronous axis stator currents |
| $i_{dr}^e$,$i_{qr}^e$ | $dq$ synchronous axis rotor currents |
| $\lambda_{dr}^e$,$\lambda_{qr}^e$ | $dq$ synchronous axis rotor fluxes |
| $R_r$,$L_r$ | rotor resistance and inductance |
| $L_m$ | mutual inductance |
| $J_m$,$B_m$ | shaft inertia and friction constant |
| $\tau_e$,$\tau_l$ | generated torque and load torque |
| $p = \frac{d}{dt}$ | differentiation operator |
| $P$ | number of poles (even number). |

Fig. 1 shows the general block diagram of an indirect field-oriented induction machine drive. With the help of IFOC, a highly nonlinear induction motor can be converted into a linear system. Fig. 2 is the block diagram representation of the linearized induction motor. In the diagram, $u = \tau_e$ is the command torque input, $y$ is the mechanical position output $\theta_{rm}$, $d = \tau_l$ is the external disturbance and $\tau_l$ is assumed to be a constant load torque. The rotor flux $\lambda_{dr}^e$ is assumed to be kept at a constant value.

Note that the calculation of the slip frequency $(\omega_e - \omega_r)$ in (6) depends on the rotor resistance. Owing to the saturation and heating effects, the rotor resistance changes and hence the slip frequency is either over or under estimated. Eventually, the rotor flux $\lambda_{dr}^e$ and the stator $q$ axis current $i_{qs}^e$ will be no longer de-coupled in (5) and the instantaneous torque control is lost. Furthermore, the electro-mechanical torque generation is reduced at steady state under the plant parameter variation and hence the machine will work in a low-efficiency region (Nordin *et al.*, 1985). Finally, the variation of the parameters $J_m$ and $B_m$ is common in real applications. For instance, the

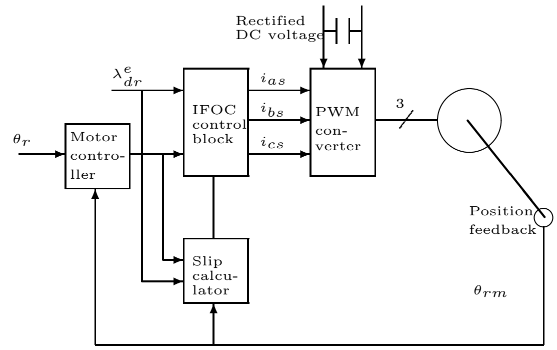


Fig. 1. Block diagram of an IFOC induction machine drive.

bearing friction will change after the motor has run for a period of time.

In order to solve the detuning problem, a novel plug-in robust compensator, using $\mathcal{H}_\infty$ loop shaping design technique is employed to compensate the system degradation under plant parameter variations and in the presence of external load torque.

## 3. CONTROLLER DESIGN

In a general controller design process, the system plant model is usually not perfectly known. The designer often only knows a nominal model of the plant and a simple controller can be designed to achieve a satisfactory tracking performance for the nominal plant. In this paper, we do not address the issue of how such a controller can be designed. We simply assume that such a controller has already been designed or, in some cases, is even already in operation in the real system. However, it is often the case that this controller may not work well when the plant is perturbed and/or external disturbances are present. In this situation, an additional controller is needed to improve the robustness of the overall system against plant uncertainties and external disturbances. It is desirable that this additional controller can be plugged into the existing control system without dismantling the existing controller and without affecting the already satisfactory nominal tracking performance. This is why we call such an additional controller a plug-in robust compensator. In this section, we investigate how to design such a plug-in robust compensator.

### *3.1 Controller Structure*

Consider the feedback system shown in Fig. 3. Assume $P$ is an SISO strictly proper nominal system. Initially assume that the two degree of freedom (2DOF) controller $K = [K_1 \quad -K_2]$ is given by $K = C = [C_1 \quad -C_2]$ which is either already available or in operation with a satisfactory nominal tracking performance, i.e. the transfer function from reference $r$ to output $y$

$$\frac{Y}{R} = \frac{C_1 P}{1 + C_2 P}$$

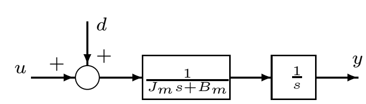


Fig. 2. Linearized induction motor model.

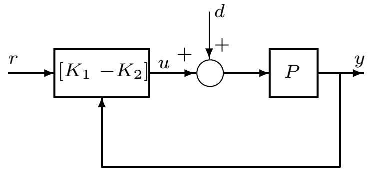


Fig. 3. General 2DOF controller.

is satisfactory. How to design $C$ is not the concern of this paper. It can be a simple PI or PD controller tuned in an online fashion or it can be designed by any other methods such as the one given in (Gan and Qiu, 2000). Let a coprime factorization of $P$ be given as

$$P = \frac{N}{M}$$

where $M, N \in \mathcal{H}_\infty$. Since $C$ is a stabilizing 2DOF controller for $P$, for any coprime factorization

$$C = \frac{[X_1\ -X_2]}{Y_0}$$

where $X_1$, $X_2$ and $Y_0 \in \mathcal{H}_\infty$. It is shown in (Vidyasagar, 1985, Theorem 15 of Section 5.6) that all 2DOF stabilizing controllers can be parameterized as

$$[K_1\ -K_2] = \frac{[S\ -(X_2 + QM)]}{(Y_0 - QN)} \qquad (7)$$

where $Q \in \mathcal{H}_\infty$ and $S \in \mathcal{H}_\infty$ are arbitrary stable systems. If we plug $K = [K_1\ \ -K_2]$ to the feedback system, then the transfer function from $r$ to $y$ becomes

$$\frac{Y}{R} = \frac{NS}{Y_0 M - X_2 N}$$

which is independent of $Q$. Since we are satisfied with the original transfer function from $r$ to $y$ when $C = [C_1\ \ -C_2]$ is used, it then follows that we can choose $S = X_1$ since

$$\frac{Y}{R} = \frac{NX_1}{Y_0 M - X_2 N} = \frac{C_1 P}{1 + C_2 P}.$$

Therefore, the set of all stabilizing 2DOF controllers which gives the same nominal tracking performance is given by

$$[K_1\ -K_2] = \frac{[X_1\ -(X_2 + QM)]}{(Y_0 - QN)}.$$

The loop property of the feedback system, which depends on $K_2$ and $P$ only, now depends on $Q$ only. For any stable system $Q$, which can even be nonlinear and time varying, the nominal tracking performance is unaffected and the closed loop stability is guaranteed (Zhou and Ren, 2000). Suppose that a $Q$ is chosen, theoretically there are two ways to implement the new controller $K$. One is to explictly obtain $K$ from (7) and implement as in Fig. 3. The other way is to use the structure in Fig. 4. Clearly, the first way requires the dismantling of the original controller $C = [C_1\ \ -C_2]$. It is the use of the structure

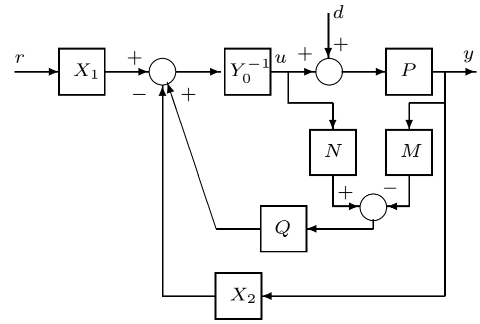


Fig. 4. Proposed plug-in robust compensator.

in Fig. 4 that gives us the plug-in feature of our additional controller. Now we give one method to design a $Q$ for enhancing the robustness of the control system.

### *3.2 $\mathcal{H}_\infty$ Loop Shaping Plug-in Compensator Design*

Since the purpose of $Q$ is to improve the loop property of the feedback system, the tracking issue is not of concern in its design. With the reference injection part ignored, Fig. 4 can be simplified to Fig. 5 with $K_2 = \frac{X_2+QM}{Y_0-QN}$. Our idea in the design of a stable $Q$ is to design a stabilizing $K_2$ and then back substitute to get $Q$ using

$$Q = \frac{K_2 Y_0 - X_2}{M + K_2 N} \qquad (8)$$

which is obtained from (7). Since all stabilizing $K_2$ are obtained from

$$K_2 = \frac{X_2 + QM}{Y_0 - QN}$$

over all stable $Q$, it follows that $Q$ obtained from (8) for a stabilizing $K_2$ has to be stable.

The design of the controller $K_2$ is further divided into two steps. The first step is to choose a proper pre-filter $W_1$ and post-filter $W_2$ so that the shaped plant, $P_s = W_1 P W_2$, has a desired open loop frequency response according to some well-defined specifications such as the bandwidth and steady state error requirement. Then an $\mathcal{H}_\infty$ optimal robust controller $K_3$ is found to minimize

$$\left\| \begin{bmatrix} I \\ K_3 \end{bmatrix} (I + P_s K_3)^{-1} [\, I\ \ P_s \,] \right\|_\infty. \qquad (9)$$

This can be done using the solution in (McFarlane and Glover, 1990) or the command *ncfsyn* of the MATLAB $\mu$-Analysis and Synthesis Toolbox (Balas *et al.*, 1994). The controller $K_2$ is

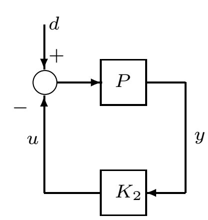


Fig. 5. Standard feedback configuration.

obtained by combining the pre-filters $W_1$, $W_2$ and the $\mathcal{H}_\infty$ controller $K_3$ as $K_2 = W_1 K_3 W_2$. Finally, the block $Q$ can be found in (8).

The main advantage of the $\mathcal{H}_\infty$ loop shaping controller over the $\mathcal{H}_\infty$ mixed-sensitivity controller is that the selection of the weighting functions, which is very difficult to assign properly in practice, is avoided in the design process. Instead, the pre-filter and post-filter are used to shape the open loop plant to achieve a desired frequency response according to some well defined design specifications such as the bandwidth and steady state error requirement. Furthermore, the $\mathcal{H}_\infty$ norm in (9) actually includes the balance of the sensitivity and complementary functions as the design objective in the $\mathcal{H}_\infty$ mixed-sensitivity optimization. Therefore, the $\mathcal{H}_\infty$ loop shaping controller can perform or outperform what the $\mathcal{H}_\infty$ mixed-sensitivity controller can do by choosing $W_1$ and $W_2$ properly.

### *3.3 Position Controller Design for the Induction Motor Control System*

In a position control system, a desired nominal tracking response can be achieved easily using a simple 2DOF PD controller. The nominal controller $C$ we employ in this section is a simple 2DOF PD controller with a low-pass filter:

$$[C_1(s) \ -C_2(s)] = \frac{1}{\delta_1 s + 1}[c_{10}s + c_{11} \ -(c_{20}s + c_{21})].$$

The design of the coefficients $c_{10}$, $c_{11}$, $c_{20}$ and $c_{21}$ are based on the $\mathcal{H}_2$ optimization such that the step output can track the one of a desirable first order system (Gan and Qiu, 2000). The coefficient $\delta_1$ is a small constant for the differentiation noise filtering in a practical implementation.

The plug-in compensator design for position control, is more difficult than that for speed control because the plant is now a SISO strictly proper unstable system. However, our controller design algorithm is still capable of handling such a system. The coprime factorization of the $P$ is first performed and it follows that $N(s) = \frac{1}{(\delta_2 s+1)(J_m s+B_m)}$ and $M(s) = \frac{s}{(\delta_2 s+1)}$ where $\delta_2$ is any positive constant. $P(s) = \frac{1}{s(J_m s+B_m)}$ is the nominal plant model. To minimize the order of the final controller, the inverse of the nominal plant stable mode $(J_m s + B_m)$ is included in $W_1(s)$. Hence, we can simply choose $W_1(s) = \frac{\alpha(J_m s+B_m)}{s}$ and $W_2(s) = 1$. For this position control design example, $Y_0(s) = \frac{\delta_1 s+1}{c_{20}s+c_{21}}$, $X_1(s) = \frac{c_{10}s+c_{11}}{c_{20}s+c_{21}}$ and $X_2(s) = 1$ can be first assumed, then the pre-filter is assigned as $W_1 = \frac{\alpha(J_m s+B_m)}{s}$. The shaped plant is equal to $P_s = \frac{\alpha}{s^2}$ finally. Following the solution in (McFarlane and Glover, 1990), the $\mathcal{H}_\infty$ controller, $K_3$, is a first order stable transfer function and the final solution $Q$ can be found from (8).

## 4. IMPLEMENTATION AND EXPERIMENTAL RESULTS

The hardware setup of our motor control system is constructed as in Fig. 1. A dSPACE DS1102 DSP controller board is used as our motion controller to implement the position controller and the IFOC algorithm. 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 phase current reference commands to the current tracking amplifier. An Advanced Motion Controls Inc. S30A40B current tracking driver is used and the 1.5kW three phase induction motor is from Baldor Inc. with the parameters listed in Table 1. A Magtrol Inc. dynamometer is used to generate the load torque in this experiment.

Table 1. Motor parameters.

| | |
|---|---|
| $J_m$ | 0.01111kg·m$^2$ |
| $B_m$ | $7.355 \times 10^{-4}$Nm/rad·s$^{-1}$ |
| $R_r$ | 0.675Ω |
| $R_s$ | 0.76Ω |
| $L_m$ | $0.2176H$ |
| $L_r$ | $0.2235H$ |
| $L_s$ | $0.2248H$ |
| $P$ | 4 |
| Encoder resolution | 4096 counts/rev |

Assume that the 2DOF nominal performance controller for position loop control

$$[C_1(s) \ -C_2(s)] = \frac{1}{0.001s + 1}[0.9028s + 50 \ -(1.5307s + 50)]$$

had already been employed to track a step reference and reject a constant external disturbance. Owing to the need for the differentiation noise filtering in a practical implementation, the low pass filter is added. The corresponding closed-loop poles are approximately located at $-84.7262$ and $-53.1176$ respectively while the closed-loop zero is approximately equal to $-55.38$.

For the design of the block $Q$, defined in Section 3, the pre-filter $W_1(s) = \frac{10^6(J_m s+B_m)}{s}$ is chosen. As described in the previous section, the constant $\alpha$ is chosen to be $10^6$ so that the cross-over frequency of the shaped plant is around 160Hz, which is adequate for the torque rejection loop because the bandwidth of the outer position loop is in the order of ten Hertz generally. Fig. 6 shows the Bode plot of the shaped plant $P_s$.

Using the command *ncfsyn* of MATLAB $\mu$-Analysis and Synthesis Toolbox again, the $\mathcal{H}_\infty$ controller $K_3(s) = \frac{2.414(s+414.21)}{(s+2.41\times10^3)}$ is found and then from (8), the optimal $Q$ is equal to

$$Q(s) = \frac{0.0165(s + 825.73)(s + 526.22)(s + 1000)(s + 0.0067)}{(s + 32.6648)(s + 993.1)(s^2 + 1417.14s + 1.0072 \times 10^6)}.$$

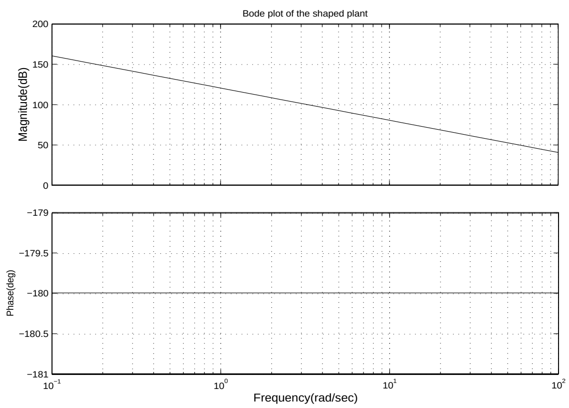


Fig. 6. Bode plot of the shaped plant $P_s = W_1 P$.

Extensive simulations had been performed using MATLAB/SIMULINK to verify the proposed plug-in robust compensator. In addition, to further validate the practical implementation of the proposed compensator, three experiments are performed and the results are described below.

The first experiment is conducted without any load disturbance and plant parameter variation. Fig. 7 shows the position tracking performance comparison with and without the plug-in compensator; it can be observed that the transient tracking response does not affected with the addition of the plug-in compensator. This verifies the claim that the plug-in compensator can coexist with the nominal controller and without affecting the nominal transient tracking response. The current tracking response for the phase A winding is shown in Fig. 8 for one reference cycle when the plug-in compensator is turned on. The actual current tracks closely with the command value and falls within the driver and motor limit. This proves the internal electrical current is still stable when the plug-in compensator is on.

Next, a step load torque is applied using a dynamometer. Fig. 9 shows the experimental results, it is clear that the system controlled by the proposed compensator did not have any significant position variation under a 2 Nm load torque; however, a 0.1 rad position variation is recorded if only the nominal controller is used. The $i_{qs}^e$ current command comparison with and without the plug-in compensator is shown in Fig. 10. It can

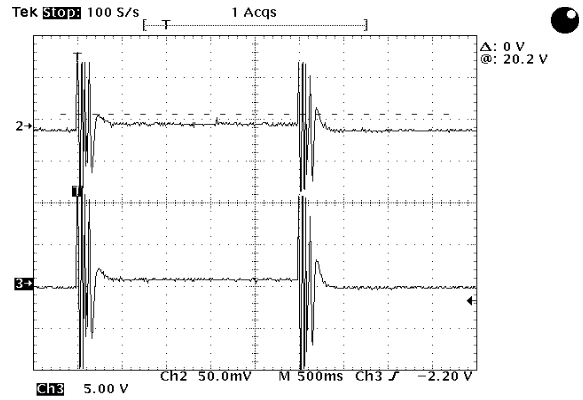


Fig. 8. Channel 2 — actual phase A current (10A/div). Channel 3 — phase A current command.

be observed that the torque compensation using the controller with the plug-in compensator, is much faster than that using the nominal controller alone; hence, the position regulation can still be maintained in the presence of the external load torque.

The final experiment in this section is to verify the robustness enhancement of the proposed plug-in compensator. In this experiment, the value $R_r$ in the controller is artificially doubled and then the output position response is measured. Fig. 11 shows the experimental results. There are overshoots and oscillations in the transient tracking response if the nominal controller is used alone. However, a good tracking response with no overshoot is achieved by turning on the plug-in compensator although the rise time is prolonged slightly.

From the above experimental results, it is clear that the proposed plug-in robust compensator can improve the disturbance response and system robustness of an IFOC induction motor drive. The plug-in robust compensator can coexist with the existing controller, does not affect the nominal tracking performance and the overall system stability.

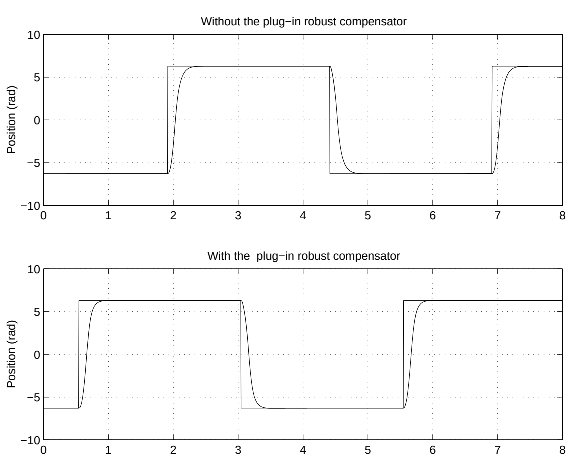


Fig. 7. Transient tracking response comparison.

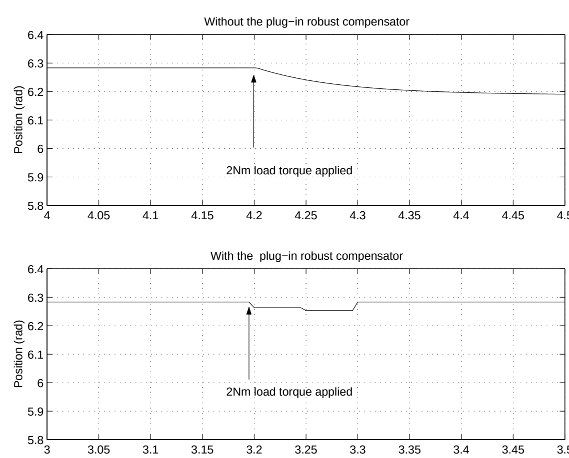


Fig. 9. Disturbance rejection comparison.

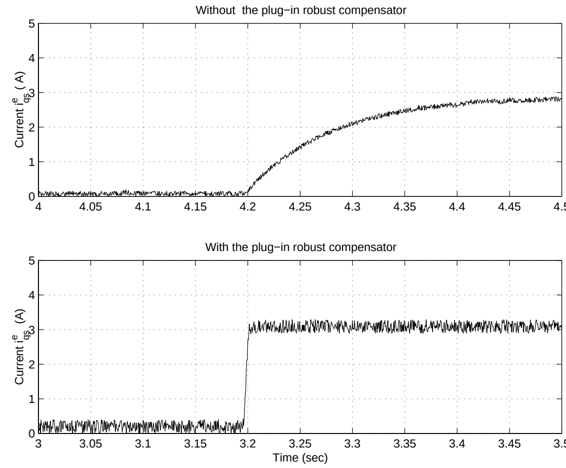


Fig. 10. Current command $i^e_{qs}$ comparison.

## 5. CONCLUSIONS

In this paper, a novel, simple, but effective plug-in robust compensator is shown to be a potential candidate which can achieve a good external disturbance rejection response and compensate the plant parameter variations for an IFOC induction motor drive system. The design of the proposed compensator is optimized in every step systematically. The controller implementation is simple since all blocks are fixed linear systems.

The novel plug-in feature of the proposed robust controller is unique, in comparison to the other robust controllers, so that the plug-in robust compensator can be designed with the existing industrial controllers such as PID and lead-lag controllers. The output performance gained by the existing controller is not affected by the proposed plug-in compensator. The proposed compensator design procedure can be applied to a speed or position control IFOC induction motor drive system such that the robustness of the overall system can be enhanced.

The proposed plug-in robust compensator is not only suitable for IFOC induction motor drive systems, but is also capable of handling any system with disturbance and modeled/un-modeled uncertainties. The proposed plug-in compensator had

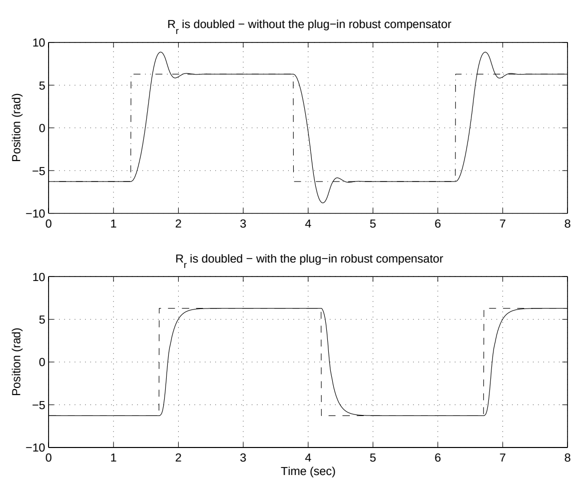


Fig. 11. Position tracking response comparison under plant parameter variation.

been successfully applied to an accurate position control of a linear switched reluctance motor (Gan *et al.*, 2001). Finally, the adaptive implementation of the blocks $Q$ and $P_0$ are being studied to further improve the system performance.

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