A Robust Speed Controller for Induction Motor Drives

Ghang-Ming Liaw, *Member, IEEE*, and Faa-Jeng Lin

***Abstract*— A speed controller considering the effects of parameter variations and external disturbance for indirect field-oriented induction motor drives is proposed in this paper. First a microprocessor-based indirect field-oriented induction motor drive is implemented and its dynamic model at nominal case is estimated. Based on the estimated model, an integral plus proportional (IP) controller is quantitatively designed to match the prescribed speed tracking specifications. Then a dead-time compensator and a simple robust controller are designed and augmented to reduce the effects of parameter variations and external disturbances. The desired speed tracking control performance of the drive can be preserved under wide operating range, and good speed load regulating performance can also be obtained. Theoretic basis and implementation of the proposed controller are detailedly described. Some simulated and experimental results are provided to demonstrate the effectiveness of the proposed controller.**

I. Introduction

GENERALLY, a high performance motor drive system must have good speed command tracking and load regulating responses, and the performances should be insensitive to the uncertainties of the drive system. The uncertainties usually are composed of plant parameter variations, external load disturbance, unmodeled and nonlinear dynamics of the plant. In the past decades, many control techniques [1]–[9] have been developed to let the field oriented induction motor drives can replace the dc motor drives in many industrial applications. The ways to deal with the plant uncertainties include the variable structure system control [5], [6], adaptive control [7], [8], and robust control [9], [10]. However, they are either very complex in theoretical bases or very difficult to be implemented. The robust control technique based on direct cancellation of uncertainties proposed in [11]–[13] is easy to derive and implement. However, the existence of an equivalent infinite gain loop makes it unable to be realized practically if system dead-time and other uncertainties are presented.

The object of this paper is to propose a simple robust speed controller with high performance for indirect field-oriented induction motor drives. The dynamic model of the drive at nominal case is first estimated. According to the estimated model and the prescribed speed tracking specifications, an IP controller is quantitatively designed. To reduce the performance degradation due to parameter variations and disturbance, a robust controller (RC), which is very suitable to be implemented, is proposed. In which, a dead-time compensator (DTC) [14] is applied such that the stability of the robust controlled system is enhanced. The stability analysis, the design and implementation of the proposed controller are presented. Good control performances both in the command tracking and load regulation characteristics can be achieved, and the performances are insensitive to parameter variations of the drive system. The effectiveness of the proposed controller is confirmed by some simulated and experimental results.

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II. Modeling of the Field-Oriented Induction Motor Drive

The block diagram of the indirect field-oriented induction motor drive system is shown in Fig. 1(a). Which consists of an induction motor loaded with a dc generator, a ramp comparison current-controlled PWM voltage source inverter, a field orientation mechanism, a coordinate translator, and a speed control loop. The induction motor used in this drive system is three-phase Y-connected two-pole 800 w 60 Hz 120 V/5.4 A.

By using the reference frame theory and the linearization technique, the field-oriented induction motor drive shown in Fig. 1(a) can be reasonably represented by the control system block diagram shown in Fig. 1(b). In which, $G_c(s)$ is an IP speed controller and

$$T_e = K_t i_{qs}^* \tag{1}$$

$$K_t = (3P/4)(L_m^2/L_r) i_{ds}^* \tag{2}$$

$$H_p(S) = \frac{1}{Js + B} \tag{3}$$

where

L_m = magnetizing inductance per phase
L_r = rotor inductance per phase referred to stator
P = number of poles
i_{qs}^* = torque current command generated from the speed controller
i_{ds}^* = flux current command
J = total mechanical inertia constant
B = total friction coefficient.

In the current-controlled PWM inverter of the indirect field-oriented induction motor drive, the current commands in synchronous reference frame denoted by i_{ds}^* and i_{qs}^* must be transformed into the phase domain to yield the reference

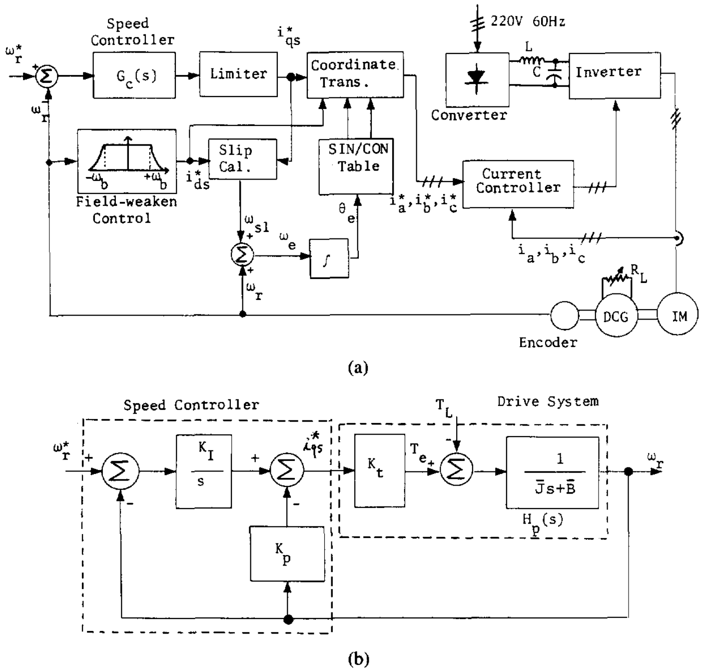

Fig. 1. (a) The configuration of an indirect field-oriented induction motor drive. (b) The control system block diagram of (a) and the IP speed controller.

currents. The unit vector $(\cos\theta_e + j\sin\theta_e)$ used in the transformation matrix is generated by using the measured rotor angular velocity ω_r and the following estimated slip angular velocity ω_{s1}:

$$\omega_{s1} = \frac{R_r i^*_{qs}}{L_r i^*_{ds}} \tag{4}$$

R_r is the rotor resistance referred to stator per phase.

Although the dynamic model can be found by the derivation introduced above, the accurate model of the whole drive system including the mechanical load is rather difficult to be obtained. To overcome this difficulty, the dynamic modeling based on measurements [15] is applied as an alternative to find the drive model at the nominal case ($\omega_{r0} = 1000$ rpm, $R_L = 156\ \Omega$). The results are

$$\overline{K_t} = 0.5443, \quad \overline{J} = 0.305, \quad \overline{B} = 0.2725. \tag{5}$$

The estimated drive parameters will be used in the design of the proposed controller.

III. Design of IP Controller

The integral plus proportional controller is shown in Fig. 1(b). It is applied here owing to its advantages [16], for example, the negligible overshoot in its step tracking response, good regulating characteristics as PI controller, and zero steady-state error.

Using the nominal drive model, the transfer function of rotor speed response to command input of Fig. 1(b) can be expressed by

$$\left.\frac{\omega_r(s)}{\omega^*_r(s)}\right|_{T_L(s)=0} = \frac{K_I K_t}{\overline{J}s^2 + (\overline{B} + K_p K_t)s + K_I K_t} \triangleq \frac{\omega_n^2}{s^2 + 2\zeta\omega_n s + \omega_n^2} \tag{6}$$

where

$$\zeta = \frac{\overline{B} + K_p K_t}{2(\overline{J} K_I K_t)^{1/2}}, \quad \omega_n = \left(\frac{K_I K_t}{\overline{J}}\right)^{1/2}. \tag{7}$$

Owing to the absence of zeros, the overshoot of the step response of (6) is avoided by setting the damping ratio $\zeta = 1$. Then one can find the unit-step response of (6) to be

$$\omega_r(t) = 1 - e^{-\omega_n t}(1 + \omega_n t). \tag{8}$$

For convenience of designing the IP controller quantitatively, the response time is defined as the time required for the step response to rise from 0 to 90% of its final value. By specifying the response time to be t_{re}, the following nonlinear equation is yielded:

$$0.9 = 1 - e^{-\omega_n t_{re}}(1 + \omega_n t_{re}). \tag{9}$$

Solve the above nonlinear equation to obtain ω_n, then from (6) one can find the parameters of IP controller as

$$K_I = \overline{J}\omega_n^2 / K_t, \quad K_p = (2\overline{J}\omega_n - \overline{B})/K_t. \tag{10}$$

Equation (6) also indicates that the tracking steady-state error is zero. It follows from the above analysis that the desired tracking specifications can be completely achieved by using the simple IP controller.

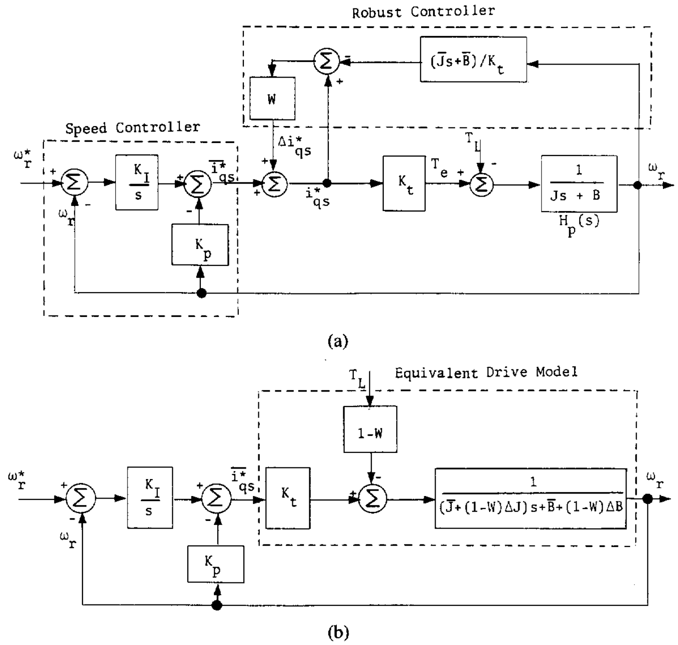

Fig. 2. (a) The block diagram of the robust control system. (b) The equivalent system of (a).

IV. The Robust Controller

A. Robust Control without Nonlinearities

A simple robust controller is proposed in Fig. 2(a). In which, the weighting factor W ($0 < W < 1$) is selected to be a compromise between the yielded performance and the required control effort. The mechanical inertia constant as well as the damping ratio are assumed to be subject to variations of ΔJ and ΔB, i.e.,

$$J \triangleq \overline{J} + \Delta J \tag{11}$$

$$B \triangleq \overline{B} + \Delta B. \tag{12}$$

The torque constant is relatively constant, so $K_t = \overline{K}_t$ is supposed in the following derivations. By augmenting the proposed robust controller, the dynamic equation corresponding to the drive model can be derived from Fig. 2(a) as

$$K_t\overline{i_{qs}}^* = (\overline{J} + (1-W)\Delta J)\dot{\omega}_r + (\overline{B} + (1-W)\Delta B)\omega_r + (1-w)T_L. \tag{13}$$

Accordingly, an equivalent block diagram is drawn in Fig. 2(b). Some comments about the proposed controller are: i) The disturbance of T_L and the variations of ΔJ and ΔB are reduced by a factor of $(1 - W)$; ii) The ideal case is obtained by setting $W = 1$, i.e., all the effects of disturbance and parameter variations on the control performance are completely eliminated. This ideal case is similar to that proposed in [11]–[13]. It is quite simple, however, the required control effort will be too demanding. Moreover, due to the existence of an infinite loop gain block [13], the closed-loop systems will be subject to stability problem. Particular for the systems which possess the nonlinearities, for example, limiter, dead-time element, etc.; iii) According to i) and ii), the scaling factor W introduced here can be adjusted such that the compromise between the performance and control effort is achieved. The detailed description about how to choose the value of W will be introduced as follows.

B. Robust Control Considering the Effect of System Dead-Time

Owing to the power processing of inverter circuit and the coupling characteristics of mechanical shaft, there will be some delay effect in the system dynamics. Accordingly, the drive model $H_p(s)$ in Fig. 2(a) is replaced by

$$H_{pd}(s) \triangleq H_p(s)e^{-\tau s} = \frac{1}{Js + B}e^{-\tau s} \tag{14}$$

with τ being the dead-time of the drive. In this case, the resulted closed-loop controlled system is quite easy to become unstable. So a smaller value of W must be adopted. However, the robustness and the control performance will be greatly degraded.

To solve the aforementioned problem, a dead-time compensator is added as an inner loop and the resulted block diagram is shown in Fig. 3(a). For convenience of study, the inner loop consisting of the drive model and the dead-time compensator is redrawn in Fig. 3(b). The major purpose of the dead-time compensator is to let the dead-time element be equivalently placed outside the closed-loop, and thus its effect on stability can be neglected. It is well known that this goal can be achieved only if the chosen reference model $H_{mc}(s)$ is exactly equal to the plant model $H_{pc}(s)$ [14]. Therefore, in

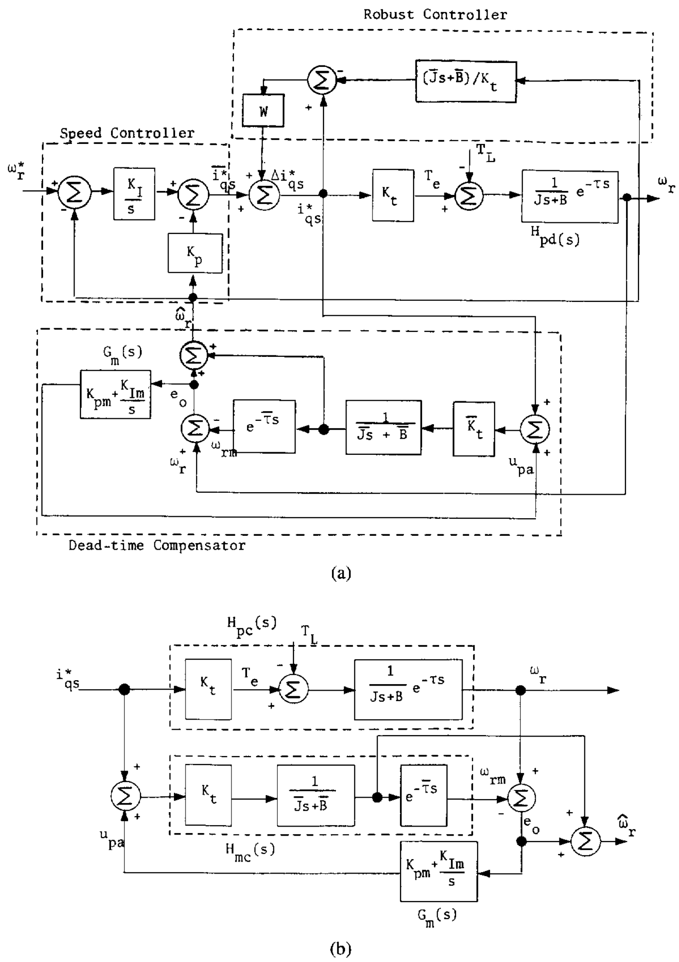

Fig. 3. (a) The block diagram of the robust control system with DTC. (b) The block diagram of the inner-loop DTC.

the case of plant subject to parameter variations, the on-line identification must be used to build up $H_{mc}(s)$. However, this is a rather difficult issue.

In this paper, the dead-time compensation inner loop is regarded as a model following control system. A model following controller $G_m(s)$ in Fig. 3(b) is added to let the model following error e_o be decayed asymptotically to zero as fast as possible compared with the dynamic characteristics of the outer loop. Although many types of controller can be applied for implementing $G_m(s)$, a PI controller is adopted here for the purpose of simplicity in realization and stability analysis of whole system. By careful derivation from Fig. 3(b), the model following error can be expressed by

$$e_o \triangleq \omega_r - \omega_{rm} = \frac{i_{qs}^*(H_{pc}(s) - H_{mc}(s))}{1 + G_m(s)H_{mc}(s)} \tag{15}$$

where

$$H_{pc}(s) \triangleq K_t \frac{1}{Js + B} e^{-\tau s} = K_t H_p(s) e^{-\tau s} \tag{16}$$

$$H_{mc}(s) \triangleq K_t H_m(s) e^{-\overline{\tau} s} \tag{17}$$

$$H_m(s) = \frac{1}{\overline{J}s + \overline{B}} \tag{18}$$

$$G_m(s) = K_{pm} + \frac{K_{\text{Im}}}{s} \tag{19}$$

and $\overline{\tau}$ denotes the nominal system dead-time. For the possible ranges of variations drive parameters (τ, J, and B), the parameters K_{pm} and K_{Im} of $G_m(s)$ are properly selected to obtain good model following characteristics. For simplicity, the trial-and-error approach is applied here to determine the values of K_{pm} and K_{Im}.

Determination of Weighting Factor: Having designed the dead-time compensator in inner loop, the weighting factor W of the proposed robust controller is determined based on the following procedure: i) The stability boundaries in parameter plane with τ, J, and W as varying parameters are sketched. Where the variation of B is not considered owing to its relative small magnitude in variation. According to the results

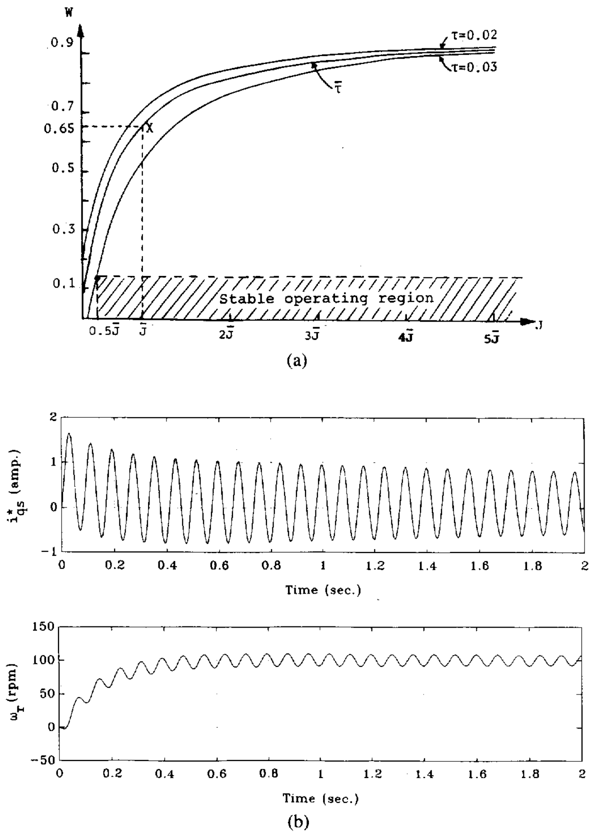

Fig. 4. (a) The parameter plane of the robust control without DTC for stability analysis. (b) The simulation results for marginally stable condition.

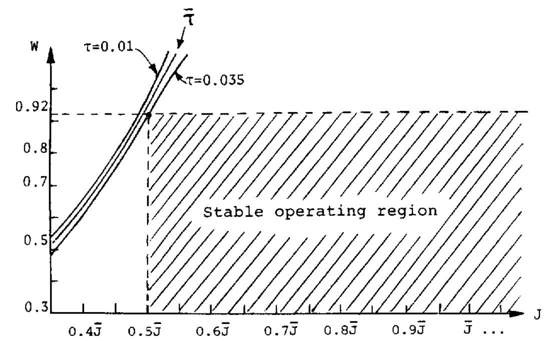

Fig. 5. The parameter plane of the robust control system without DTC for stability analysis.

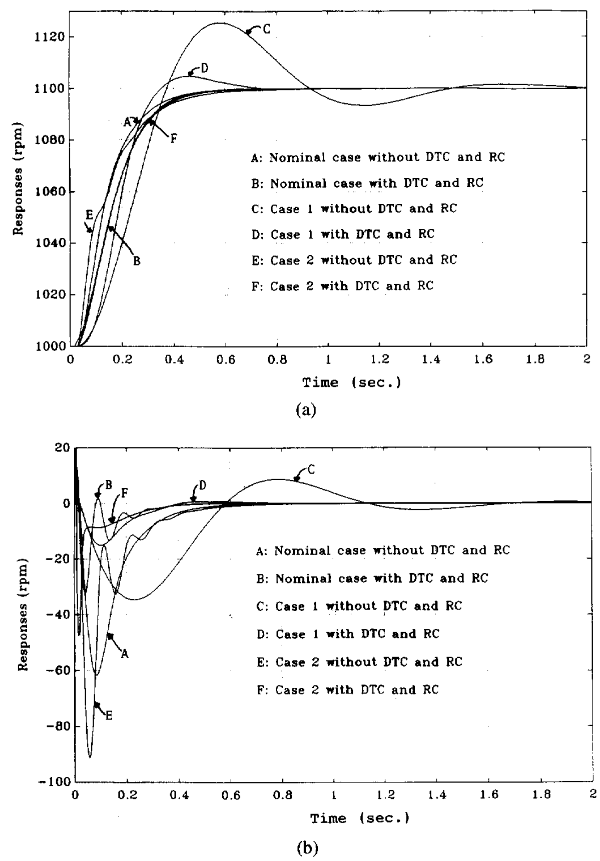

Fig. 6. Comparison of simulated rotor speed responses. (a) Step command tracking responses. (b) Responses due to unit-step load torque change.

of stability analysis, a value of W is chosen from the view point of stability; ii) With the selected value of W in i), the maximum value of required control effort is checked. Under the condition of without leading to saturation, a larger value of W, which also satisfies the stability requirement, can be selected.

V. Design of the Proposed Controller

The design methodology of the proposed controller introduced in the previous section will be illustrated numerically as follows.

Design of IP Controller: Using the parameters listed in (5) and setting t_{re} to be 0.3 s, the parameter of the IP speed controller can be found from (9) and (10) as

$$K_p = 14.0242, \quad K_I = 94.1637. \tag{20}$$

The dead-time of the drive implemented in this paper is estimated following the approach developed in [17] to be

$$\overline{\tau} = 0.0235s. \tag{21}$$

For investigating the effectiveness of the proposed robust controller, suppose that the mechanical inertia constant J is significantly changed to let the transfer function model $H_p(s)$ be changed to

$$\text{Case 1:} \quad H_{p1}(s) = \frac{1}{1.525s + 0.2725} \quad (J = 5 \times \overline{J}) \tag{22}$$

$$\text{Case 2:} \quad H_{p2}(s) = \frac{1}{0.1525s + 0.2725} \quad (J = 0.5 \times \overline{J}) \tag{23}$$

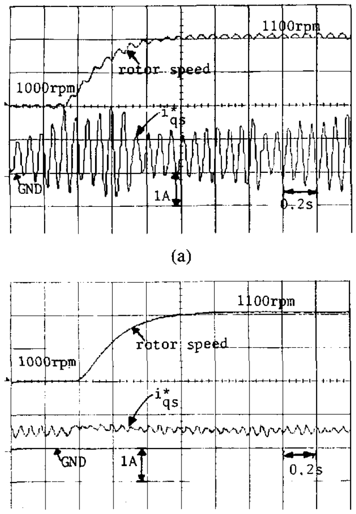

Fig. 7. The measured results for marginally stable condition. (a) Without DTC. (b) With DTC.

Robust control without DTC: The closed-loop command tracking transfer function of Fig. 3(a) without the connection of dead-time compensator can be derived as

$$H_c(S) \triangleq \left.\frac{\omega_r(S)}{\omega_r^*(S)}\right|_{T_L(s)=0} = \frac{K_I K_t}{\Delta(s)} \tag{24}$$

with

$$\begin{aligned}\Delta(s) = {} & [\overline{J}W + (1-W)(\overline{J}+\Delta J)e^{\tau s}]s^2 \\ & + [\overline{B}W + (1-W)(\overline{B}+\Delta B)e^{\tau s} + K_t K_p]s \\ & + K_I K_t.\end{aligned} \tag{25}$$

For convenience of analysis and simulation, the dead-time transfer function $e^{\tau s}$ is approximated by

$$e^{\tau s} \approx \frac{1+\tau s/2}{1-\tau s/2}. \tag{26}$$

The stability limit of the control system represented by (24) can be obtained by finding the conditions of existence of imaginary roots in $\Delta(s)$ with varying system parameters. The dominant parameters selected to be studied are the mechanical inertia J, the system dead-time τ, and the weighting factor W of the robust controller. Using the values in (5), (20), and (21), the stability boundaries for several values of τ are drawn in Fig. 4(a). The area to the right of the boundary is the stable region, and conversely, the area to the left of the boundary is the unstable region. For verification, the simulated step responses of i_{qs}^* and ω_r using the parameter of operating point X indicated in Fig. 4(a) are shown in Fig. 4(b). From which, the phenomena of marginally stable condition are obvious. The stability analysis made in Fig. 4(a) indicates that the weighting factor W cannot be chosen too large for wide range of inertia constant variations from the stability point of view. For example, if τ and J are assumed to be varied from 0.02 s to 0.03s and $0.5\overline{J}$ to $5\overline{J}$, respectively, the weighting factor must be chosen to be $W < 0.15$ and the stable operating region shown is very narrow. It follows from Fig. 2(b) that the improvement achieved by robust controller will be limited for such small value of W.

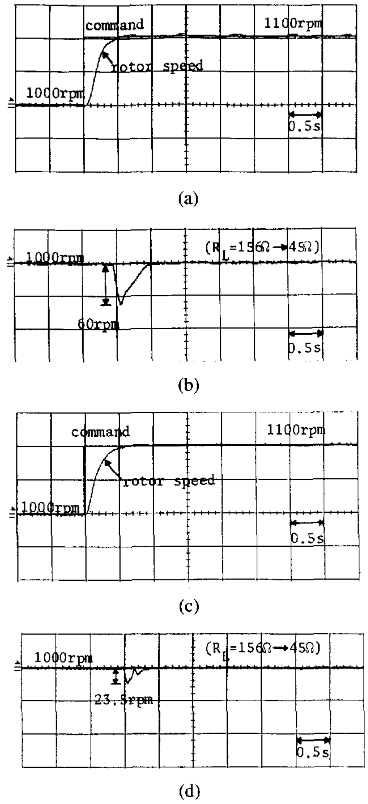

Fig. 8. The measured rotor speed response at nominal case: (a) due to step command change without RC and DTC; (b) due to step load resistance without RC and DTC; (c) due to step command change with RC and DTC; (d) due to step load resistance change with RC and DTC.

Robust control with DTC: Now the dead-time compensator in Fig. 3(a) is connected, the parameters of $G_m(s)$ are selected as follows:

$$K_{pm} = 30, \quad K_{\mathrm{Im}} = 15. \tag{27}$$

The parameters of $H_{mc}(s)$ are set to the nominal values of $H_{pc}(s)$. Following the same procedure in the previous paragraph, the stability boundaries of the system shown in Fig. 3(a) are drawn in the (J, τ, W) parameter space shown in Fig. 5. For enlarging the illustration, the variation of τ is assumed to be from 0.01 to 0.035 s. It is obvious from Fig. 5 that the stable operating region is greatly broadened compared with that of Fig. 4(a). Now for the same range of J $(0.5\overline{J} < J < 5\overline{J})$, the weighting factor can be chosen up to 0.92. However, for preserving some stability margin and also to lessen the control effort, $W = 0.8$ is set in the implementation.

Simulation Results: Before performing the hardware implementation, some simulations are made to test the validity of the proposed controller. At the nominal case, the step rotor speed tracking and step load regulating responses of the drive system shown in Fig. 3(a) without and with the connection of robust controller and dead-time compensator are shown in Fig.

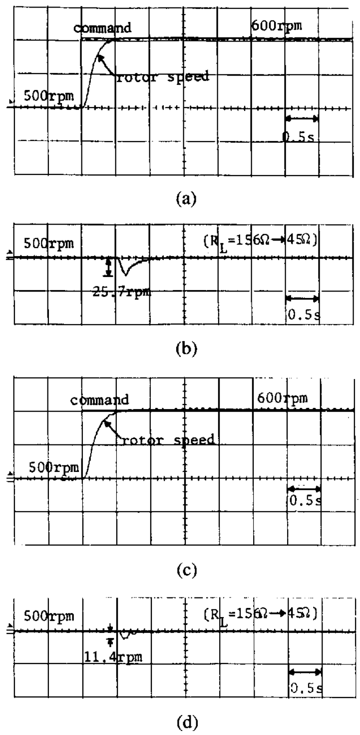

Fig. 9. The measured rotor speed responses at the operating condition of ($\omega_{r0} = 500$ rpm, $R_{L0} = 156\ \Omega$): (a) to (d) the same as that in Fig. 8.

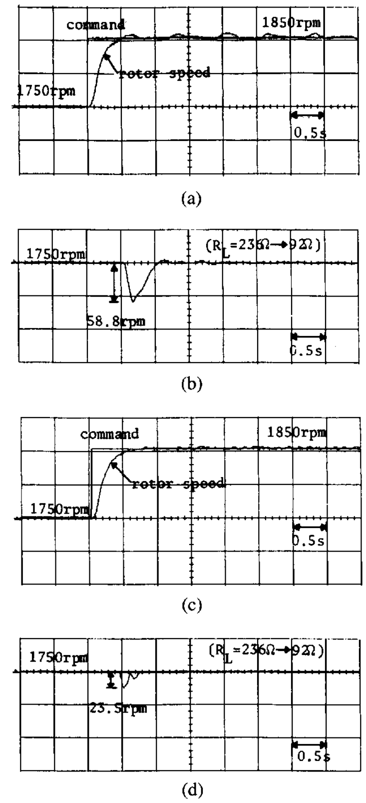

Fig. 10. The measured rotor speed responses at the operating condition of ($\omega_{r0} = 1750$ rpm, $R_{L0} = 236\ \Omega$): (a) to (d) the same as that in Fig. 8.

6(a) and (b) by curves A and B, respectively. For the tracking condition, the dead-time is equivalently moved outside the control loop, and the regulating performance is improved. Suppose that the mechanical inertia constant J is changed according to (22) and (23), the rotor speed responses due to step command change and step load torque change without and with the connection of robust controller and dead-time compensator are compared in Fig. 6(a) and (b). Significant performance improvements both in the tracking and regulating responses by the proposed controller are observed from the results shown in Fig. 6(a) and (b).

VI. Implementation and Experimental Results

Having confirmed the performance of the proposed controller by simulation, the hardware implementation of the drive system and the software realization of the proposed controller are carried out. The field-oriented mechanism and the proposed controller are realized in a 80386/80387 personal computer. The sampling interval is selected to be 10 ms. The D/A card and multifunction card are used as the interfaces between the computer and the drive. The sampling interval is selected to be 10 ms. The pure differentiator may amplify the high-frequency noise, so the operating stability of the closed-loop controlled drive will be greatly affected. Thus in practical implementation, a filter is used as an alternative. It is designed to behave as a pure differentiator for the main low-frequency dynamic signal, but it becomes a low-pass filter for high-frequency signals.

To verify the phenomena under marginally stable condition, the control system shown in Fig. 3(a) without the connection of dead-time compensator is implemented and the W is set according to point X in Fig. 4(a) ($W = 0.65$). The measured rotor speed ω_r and torque current i_{qs}^* due to a step command of 100 rpm are shown in Fig. 7(a), which are rather close to the simulation results shown in Fig. 4(b). After connecting of the dead-time compensator, the results shown in Fig. 7(b) indicate that the oscillating phenomena have been eliminated and good speed tracking performance is obtained.

Owing to lacking suitable means for changing the mechanical inertia constant, this paper provides only the measured tracking and regulating responses at various operating conditions to show the robust characteristics of the proposed controller. First, the measured rotor speed responses due to step command change and step load resistance change without and with the DTC and RC are compared in Fig. 8(a) through (d). The results show that while the tracking responses of these two cases are almost identical, the regulating response is significantly improved by augmenting the DTC and RC. In addition, the drive rotor speed tracking and regulating responses without and with the DTC and RC at another operating condition are shown in Fig. 9(a) through (d). Better control performance yielded by the proposed controller is obvious from the results. Now let the drive be operated at a higher speed, the measured results compared in Fig. 10(a) through (d) show that the improvement in tracking response by the proposed controller is also very evident.

VII. Conclusions

A simple robust speed controller for indirect field-oriented induction motor drives is successfully designed and implemented in this paper. First, an IP controller is quantitatively designed according to the estimated nominal drive model and the prescribed speed tracking specifications. Then a dead-time compensator and a simple robust controller are designed and augmented to reduce the effects of parameter variations and external disturbances. The stability analysis, design, and implementation of the proposed controller are detailedly described. Some practical considerations have been taken into account in the implementation. The simulated and experimental results show that good speed tracking and load regulating responses are yielded by the proposed controller, and the performances are rather insensitive to parameter and operating condition changes. The proposed control technique can also be applied to other types of motor drives.

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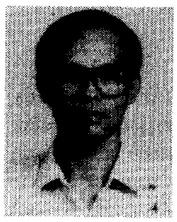

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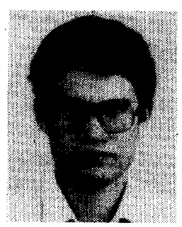

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