

DSP-Based Robust Control of an AC Induction Servo Drive for Motion Control

Ying-Yu Tzou, *Member, IEEE*

Abstract—A design method based on the internal model control (IMC) theory has been developed for the robust control of a low-power ac induction servo drive in application of high-performance motion control system. Indirect vector control with software current control technique has been developed to achieve fast response decoupling control of the induction motor. An equivalent model representing the control dynamics of the decoupled ac induction motor is used for the analysis and synthesis of the IMC controller. To improve the control performance of the servo drive with given frequency-domain robustness specifications, a design procedure for the servo controller is developed based on the IMC theory. The proposed control scheme has been realized using a single-chip digital signal processor (DSP) TMS320C14 from Texas Instruments. Experimental results are given to verify the proposed design method.

NOMENCLATURE

a, b, c	Subscripts denoting motor three phase axes.
d, q	Subscripts denoting d - q axes rotating synchronously with the rotor flux.
α, β	Subscripts denoting stator-fixed (stationary) axes.
s, r	Subscripts denoting stator and rotor, respectively.
*	Superscript denoting reference or command.
P	Number of poles.
i, v, λ	Current, voltage, and flux linkage.
L_s, L_r, L_m	Stator, rotor inductance per phase and mutual inductance.
σ	Total leakage factor $(1 - L_m^2/L_s L_r)$.
R_s, R_r	Stator and rotor resistance per phase.
τ_r	Rotor time constant (L_r/R_r) .
R_e	Equivalent resistance $(R_s + L_s/\tau_r)$.
J_m, B_m	Equivalent shaft inertia and viscous friction.
$\theta_r, \theta_e, \theta_m$	Rotor electrical, rotor flux, and rotor shaft angle.
$\omega_r, \omega_e, \omega_m$	Rotor electrical, rotor flux, and rotor shaft angular velocity.
K_T	Equivalent torque constant: $(P/2)(L_s/L_m)(L_m i_{ds})$.
K_E	Equivalent back emf constant: $(P/2)(L_m/L_r)(L_m i_{ds})$.
T_e, T_d	Electrical developed torque and torque disturbance.

Φ_R	Rated rotor flux.
K_I	Equivalent current control gain.
τ_m, τ_e	Equivalent mechanical and electrical time constant.

I. INTRODUCTION

MOTION control can be defined as a system integration technology in the application of control theory, power electronics, and microcomputer control to achieve precision control of torque, velocity, and/or position of mechanical systems [1]. Servo motors equipped with high-resolution encoders have made the incremental motion control system to be the major approach to achieve precision and robust control in today's industry automation. Therefore, motion control has been recognized as the key technology in factory automation. Various servo drives such as dc motor drives, variable-reluctance stepper drives, and brushless dc drives have been employed to achieve high-performance motion control [2]. However, with the great advances in field-oriented control and digital signal processor (DSP)-based control technology, induction drives currently provide a very competitive option for motion control [3], [4]. A well-tuned robust-controlled induction drive provides a very wide speed control range, very low torque ripple, inherent maintenance-free capability, explosion proof, high power density, and mechanically robust characteristics. It is an ideal solution for most high-performance motion control systems. This paper presents the application of internal model control theory using DSP realization for the robust control of an induction drive used for high-performance motion control systems.

Induction motors with their inherent nonlinear characteristics were traditionally used in low-performance adjustable speed applications using open-loop controlled pulse width modulation (PWM) inverters. However, with the invention of field-oriented vector control [5] and development of modern control techniques [6]–[8], the ac induction motor can be decoupled to behave just like an ideal external excited dc motor. With the great advances in microelectronics and very large scale integration (VLSI) technology, high-performance microprocessor and DSP's can be effectively used to realize advanced control schemes [9]. Although the DSP had been developed primarily for application in the field of communication and signal processing, it becomes more and more popular to be used as a control processor because of fast numerical computation capability. Due to its fast computation speed, the DSP has replaced much of the complex control hardware by ROM-based software. This development trend has made

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The author is with the Department of Control Engineering, National Chiao Tung University, 30050 Hsinchu, Taiwan, R.O.C.

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software control to reach new horizons in the applications of power electronics and motion control [10], [11]. In this paper, we have realized all the control functions of an induction servo drive using a single-chip DSP (TMS320C14) from Texas Instruments [12].

Good command tracking and high disturbance rejection are two basic requirements for every high-performance servo drive. To satisfy these requirements, an induction servo drive needs well-tuned decoupling control for torque control and robust servo control for speed regulation. To achieve well-tuned decoupling control of an induction motor, there are two major concerns: one is the exact information of the rotor flux is required, and the other is the stator currents have to be regulated instantaneously according to their references. Hysteresis control using analog circuit realization was traditionally used to achieve fast response current control [13]. However, this approach has drawbacks of unpredictable switching frequency, higher current ripples, and imposes higher switching stresses to the power converters [14]. Software current control provides a more flexible control approach in the decoupling control of an induction motor [15]. However, the limited sampling rate and finite word length of the microprocessor will also impose a limitation on the current loop bandwidth.

One effective method for detecting the rotor flux is the indirect vector control scheme [16] which eliminates the need for air-gap flux sensors and no flux distortion at low speed compared to the direct vector control scheme. However, the indirect vector control will be detuned due to motor parameters variation, especially the rotor time constant. This condition is serious for large-size motors but is insignificant for small-size horsepower range induction servo motors. In this paper, the design problems associated with parameters variation and the software current loop control are formulated as a servo controller design problem by developing an equivalent model that represents the decoupled induction servo drive with specified parameter uncertainties.

Frequency domain compensation techniques based on Bode plot, Nyquist plot, or Nichols chart have been widely adopted by practical engineers in the synthesis of loop compensators for servo drive design. However, these classical techniques cannot handle design problems with plant uncertainty. In recent years, linear controller synthesis methods based on parameterization and robust control theory have been developed for uncertain systems [17]–[19]. However, these modern design techniques are still scarcely employed in the design of practical industry control systems. One major reason may be the impractical design specifications for practical design problems. In this paper, we have developed a practical robust control scheme based on the internal model control (IMC) theory [20], [21] for the servo loop control of an induction servo drive.

This paper is organized as follows. Section II introduces the control structure of the proposed DSP-based fully digital vector-controlled ac induction servo drive. An equivalent model of the decoupled induction drive has also been developed for the synthesis of servo loop controller. Section III describes the operation principle of the IMC controller. An illustrated design procedure is also given in this section.

Section IV presents the DSP implementation of the designed IMC controller. Experimental verifications are also included in this section. Section V gives the conclusion.

II. DECOUPLING CONTROL AND MODELING OF AC DRIVES

The design of a high-performance ac servo drive consists of two major steps: the first is the field-oriented vector control to decouple the nonlinear dynamics inherent to the ac motors, and the second is the robust servo loop control to minimize sensitivity to parameter variations and unmodeled dynamics. Fig. 1 shows the block diagram of the proposed DSP-based fully digital-controlled ac induction drive. A single-chip DSP has been employed to realize all the control functions which includes 1) position/speed loop control; 2) field-weakening and indirect vector control; and 3) current control and PWM gating signal generation. The ac induction motor is driven by an insulated gate bipolar transistor (IGBT)-based voltage source inverter. The PWM gate pulse signals for the inverter are generated directly by the DSP according to the control processing. The feedback signals to the DSP-based controller are the measured stator currents and encoder pulse feedback of the motor shaft. The servo controller consists of a position loop controller (PLC) and a velocity loop controller (VLC). The position control loop of a motor drive is an inherent type-1 system due to the integration of velocity to position and it will have no steady-state error to a step position command. In the synthesis of the robust ac servo controller, the PLC is a pure gain which corresponds to the velocity error constant of the position servo system and is an important tuning parameter in practical applications. The VLC is synthesized based on an internal model control scheme.

The two-pole cage-type ac induction motor employed for the servo drive design exhibits rotational symmetry, including constant effective air-gap. It has been shown [23] that the induction motor can be described by a set of four nonlinear differential equations

$$\vec{v}_s = R_s \vec{i}_s + L_s \frac{d\vec{i}_s}{dt} + L_m \frac{d}{dt} \cdot (\vec{i}_r e^{j\theta_r}) \quad (1)$$

$$0 = R_r \vec{i}_r + L_r \frac{d\vec{i}_r}{dt} + L_m \frac{d}{dt} \cdot (\vec{i}_s e^{-j\theta_r}) \quad (2)$$

$$J_m \frac{d\omega_r}{dt} + B_m \omega_r = T_e - T_d = \frac{P}{2} L_m \cdot \text{Im}[\vec{i}_s (\vec{i}_r e^{j\theta_r})^*] - T_d \quad (3)$$

$$\frac{d\theta_r}{dt} = \omega_r \quad (4)$$

and the torque equation can be expressed as

$$T_e = \frac{P}{2} \frac{L_m}{L_r} (i_{qs} \lambda_{dr} - i_{ds} \lambda_{qr}). \quad (5)$$

In the complex plane perpendicular to the motor axis, as shown in Fig. 2, the stator current vector is formed by combining the instantaneous three-phase quantities

$$\vec{i}_s = i_{as} + i_{bs} e^{j(2\pi/3)} + i_{cs} e^{j(4\pi/3)} = i_{\alpha s} + j i_{\beta s}. \quad (6)$$

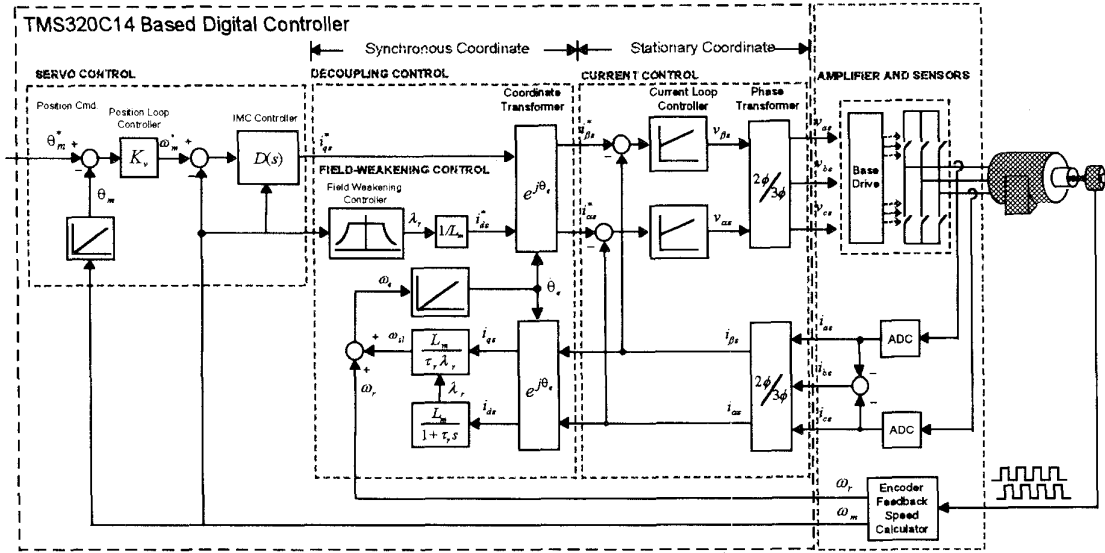


Fig. 1. Proposed DSP-based fully digital servo control of an ac induction motor.

This stator current vector in the stationary reference coordinate can be transformed to a synchronous rotating d - q coordinates as

$$\vec{i}_s e^{-j(\theta_e + \theta_r)} = i_{ds} + j i_{qs}. \quad (7)$$

This transformation also holds truth for other stator and rotor variables, such as voltage and flux linkage.

The basic concept in field-oriented vector control is to fix the rotating coordinate to a selected reference vector so that the generated electrical torque can be linearly controlled by a corresponding torque current component. There are various decoupling control mechanisms developed to achieve this goal in considerations of feedback quantities, computation complexity, and sensitivity. Compared to the direct vector control scheme, the indirect vector control has the advantages that it does not require to sense the air gap flux directly and can be operated to down stand-still. This is especially important for servo drives used for motion control. The main idea of the indirect vector control scheme is to fix the d -axis of the referenced d - q coordinates to an estimated rotor flux vector

$$\vec{\lambda}_r = L_r \vec{i}_r + L_m \vec{i}_s e^{-j\theta_r}. \quad (8)$$

This rotor flux vector can be calculated by a rotor flux model in the d - q coordinates using the feedback stator currents. The rotor flux model in the synchronous d - q coordinates is

$$\lambda_r = \frac{L_m}{s\tau_r + 1} i_{ds} \quad (9)$$

and the slip angular frequency is

$$\omega_{sl} = \frac{L_m}{\tau_r \lambda_r} i_{qs}. \quad (10)$$

This slip angular frequency is then used to generate the coordination transformation angle used for the decoupling control. If the d -axis can be exactly fixed to the rotor flux

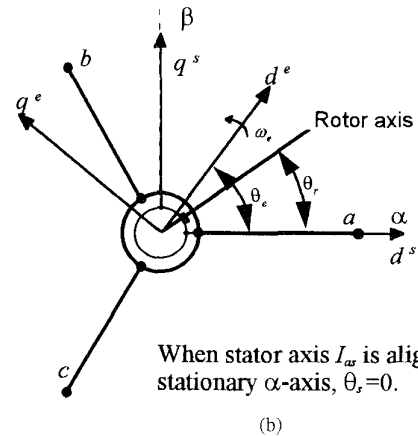
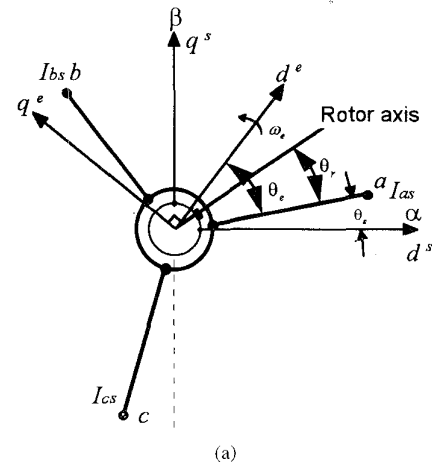


Fig. 2. Definition of current vector and coordinates.

vector, then

$$\lambda_{dr} = \lambda_r \quad \text{and} \quad \lambda_{qr} = 0, \quad \frac{d\lambda_{qr}}{dt} = 0 \quad (11)$$

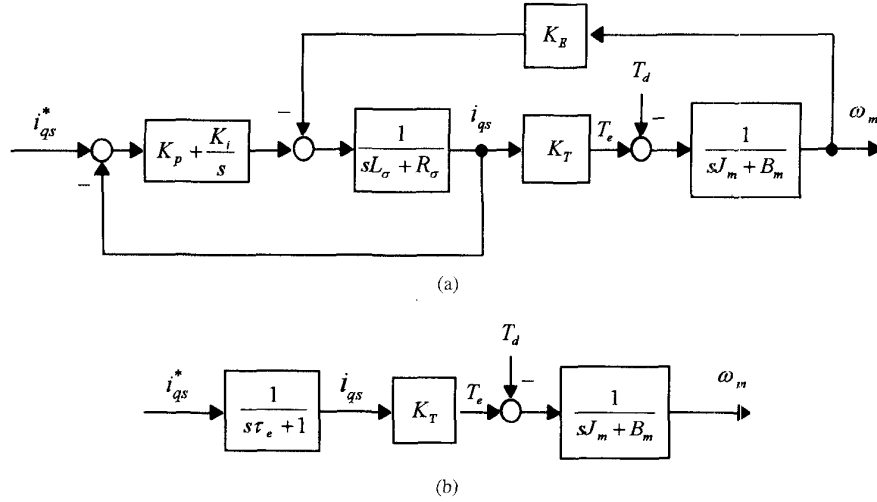


Fig. 3. (a) Mathematical model of the vector-controlled current-decoupled induction motor drive and (b) simplified block diagram.

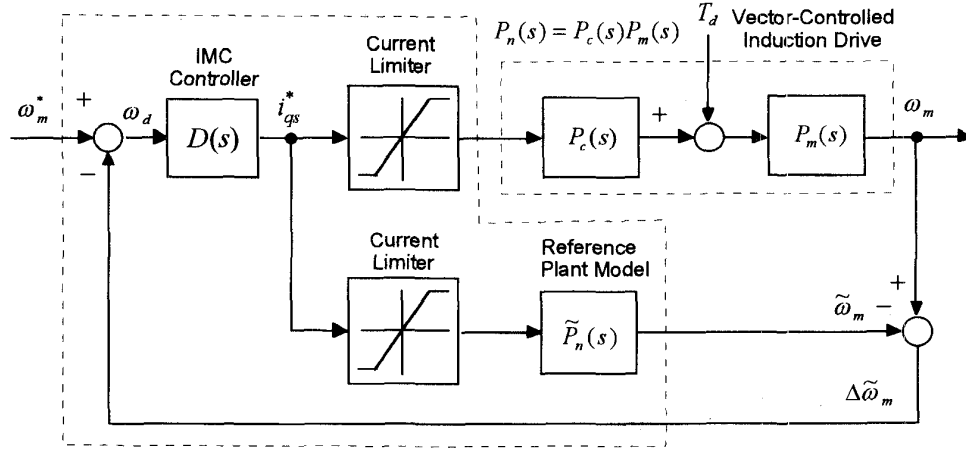


Fig. 4. IMC structure for the speed servo controller design.

and the torque equation (5) can be reduced to

$$T_e = \frac{P}{2} \frac{L_m}{L_r} \lambda_r i_{qs}. \quad (12)$$

This means the motor torque can be solely determined by a torque producing current i_{qs} , if the rotor flux is kept constant. In this condition, the stator current vector is decomposed to two orthogonal components, the flux-producing current i_{ds} and the torque-producing current i_{qs} , which can be controlled independently. However, this condition is under the assumption that the stator current can be controlled instantaneously. In practical condition there is always a bandwidth limitation to the current loop. As illustrated in Fig. 1, the software current control scheme is used to regulate the stator currents in the stationary α - β coordinates. The current controllers are implemented by discretizing the analog PI controllers

$$\begin{aligned} v_{\alpha s} &= \left(K_{p\alpha} + \frac{K_{i\alpha}}{s} \right) (i_{\alpha s}^* - i_{\alpha s}) \\ v_{\beta s} &= \left(K_{p\beta} + \frac{K_{i\beta}}{s} \right) (i_{\beta s}^* - i_{\beta s}). \end{aligned} \quad (13)$$

The software current feedback loop regulates the stator currents according to their reference values. The current references are generated by the indirect vector controller.

When the motor is operating within the constant-torque region, the rotor flux is kept constant to maintain an optimal efficiency. In this condition, the synchronous angular frequency of the rotor flux becomes

$$\omega_e = \omega_r + \frac{L_m}{\tau_r \Phi_o} i_{qs} \quad (14)$$

where $i_{ds} = I_o$, $\lambda_r = L_m I_o = \Phi_o$, and I_o is the rated magnetizing current. By converting the voltage vector to the d - q coordinates, the stator voltage equation (1) in the q -axis can be expressed as

$$R_e i_{qs} + \sigma L_s \frac{di_{qs}}{dt} = v_{qs} - K_E \omega_m. \quad (15)$$

Equation (15) together with the PI current control action (13) and the mechanical dynamics (3) forms the mathematical model of the vector-controlled current-decoupled induction motor drive. Fig. 3(a) shows the block diagram from the

torque-producing current reference to the rotor angular velocity. In usually condition, the current loop bandwidth is much higher than the open-loop mechanical pole. Therefore, the plant dynamics of Fig. 3(a) can be reduced to Fig. 3(b), where τ_e represents the time constant of the current loop and the equivalent model of the nominal plant is

$$P_n(s) = \frac{K_T/B_m}{(s\tau_e + 1)(s\tau_m + 1)}. \quad (16)$$

The plant model of vector-controlled induction drive has now been reduced to a second-order linear system to which the IMC method can be applied.

III. DESIGN OF THE IMC SERVO CONTROLLER

A. Principle of Internal Model Control

A control system is generally required to regulate the controlled variables to reference commands without steady-state error against unknown and unmeasurable disturbance inputs. Control systems with this property are called servomechanisms or servo systems [17]. In servomechanism system design, the internal model principle proposed by Francis and Wonham [20] plays an important role. The *internal model principle* can be viewed as a generalization of system type theory. The internal model principle states that the controlled output tracks a set of reference inputs without steady-state error if the *model* which generates these references is included in the stable closed-loop system.

In the control structure as shown in Fig. 4, in addition to the controller $D(s)$ it also includes the plant nominal model $\tilde{P}_n(s)$, therefore this feedback configuration is known as internal model control [10]. The design of a robust servomechanism system with plant uncertainty begins with three specifications: 1) definition of the plant model and associated uncertainty; 2) specifications of the inputs; and 3) desired closed-loop performance. The IMC theory provides a systematic approach in the synthesis of a robust controller for systems with specified uncertainties.

Fig. 4 shows the basic configuration of the IMC control system which consists of the plant $P_n(s) = P_c(s)P_m(s)$, a reference nominal model $\tilde{P}_n(s)$, an IMC controller $D(s)$, and limiters for the actuating signals. The transfer function $P_n(s)$ represents the uncertain plant, in this case, it is the current-decoupled ac drive with mechanical dynamics and T_d is the external torque disturbance.

It should be noted that the same actuator constraints are imposed on the controlled process and the reference model. The current constraint and limiter are due to the physical limitations of the voltage source inverter. From Fig. 4 we can see that if the model is exact and there are no disturbances, then the model output and the process output are the same, and therefore the feedback signal is zero. This illustrates that if a process is stable and all its disturbances are known exactly, there is no need for feedback control. The feedback signal $\Delta\tilde{\omega}(s)$ of the IMC control scheme represents the uncertainty of the controlled process and the purpose of the controller $D(s)$ is to compensate the process derivation from its nominal model.

TABLE I
LOAD AND PARAMETER VARIATIONS

	uncertainty bounds	nominal	unit
σL_s	15.7~19.2	17.5	mH
R_e	1.75~3.05	2.4	Ω
K_T	0.43~0.53	0.48	N-m/A
K_E	0.49~0.60	0.54	V-sec/rad
J_m	3.8~7.5	5.6	g-m ²
B_m	0.26~0.79	0.53	N-cm-sec/rad
K	67.1~203.8	90.5	rad/sec-A
τ_m	0.34~1.3	0.83	sec
τ_e	0.15~0.19	0.17	msec

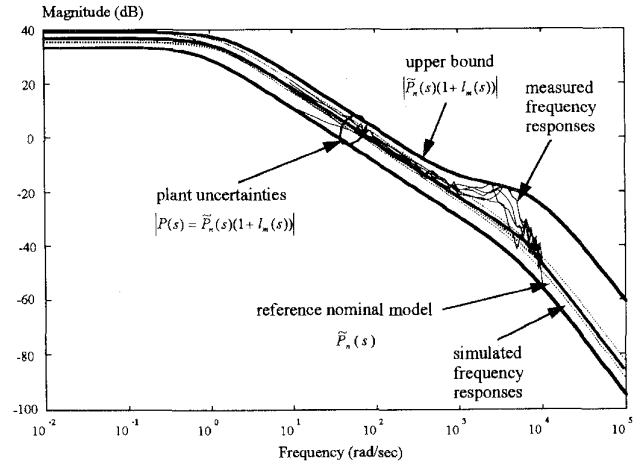


Fig. 5. Estimated and measured frequency responses of the uncertain plant.

The successful application of the IMC control scheme relies on the definition of a suitable nominal reference model $\tilde{P}_n(s)$ and the synthesis a loop compensator $D(s)$ so that robust stability and performance specifications can be achieved. There are two major advantages in applying the IMC control scheme in the synthesis of a servo controller. One is the closed-loop stability can be assured by choosing a stable IMC controller. The other is the closed-loop performances are related directly to controller parameters, which makes on-line tuning of the IMC controller very convenient.

The design procedure of the IMC controller consists of two major steps. In the first step, the IMC controller is designed to achieve so-called nominal performance without regard to plant uncertainty or, equivalently, with the assumption that $P_n(s) = \tilde{P}_n(s)$. In the second step, the IMC controller is augmented by a robustness compensator $F(s)$ to meet the robustness specifications. This robustness compensator is usually a low-pass filter since the plant uncertainty and unmodeled dynamics generally increases with frequency. Thus the IMC controller

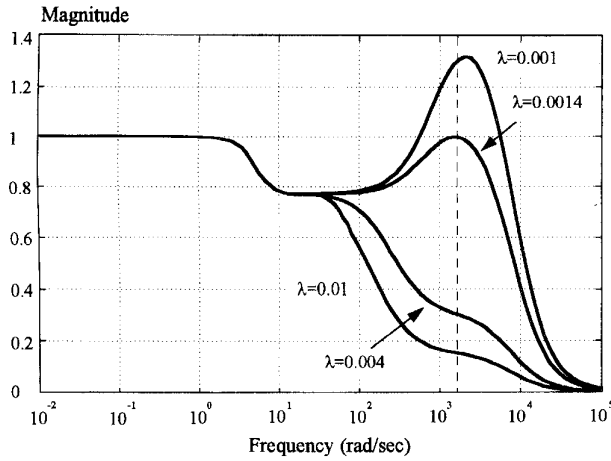


Fig. 6. Evaluation of robust stability for different filter time constants.

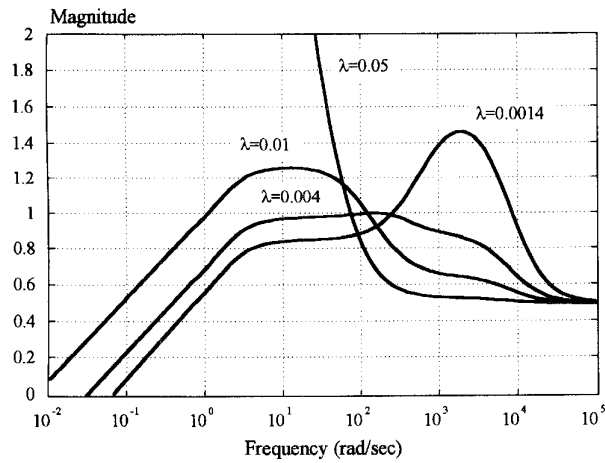


Fig. 7. Evaluation of robust performance for different filter time constants.

has the form

$$D(s) = D_n(s)F(s) \quad (17)$$

where $D_n(s)$ is an optimal controller obtained in the first design step. Within the operating bounds specified by the actuator constraint, the IMC control scheme is mathematically equivalent to a conventional feedback controller of the form

$$C(s) = D(s)(1 - \tilde{P}_n(s)D(s))^{-1}. \quad (18)$$

If the reference nominal model, the IMC controller, and the robustness compensator are appropriately designed, the IMC control scheme will produce a servo controller with desired robustness performances.

B. Specifications of Plant Uncertainty and Reference Model

It is important to incorporate plant uncertainty into the controller design procedure so that the designed control system can achieve the design objectives of robust stability and robust performance. The model uncertainty may have different sources, such as linearization, variation of operating point, variation of parameters, and unmodeled dynamics, etc. In a

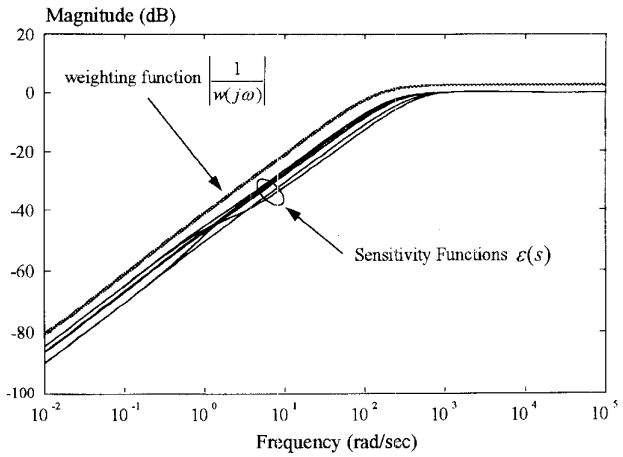


Fig. 8. Sensitivity functions bounded by the inverse of the performance weighting function.

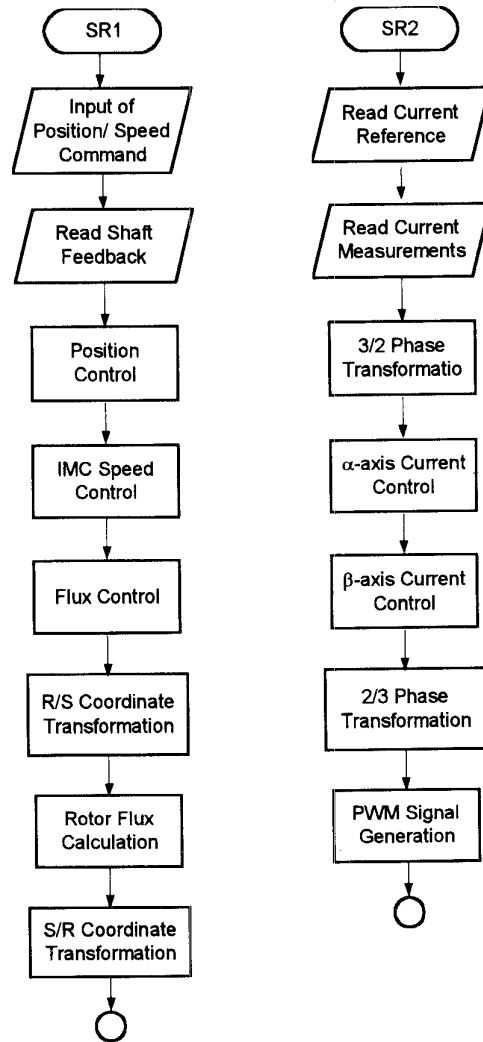


Fig. 9. DSP software flowchart for fully digital servo control of an induction motor.

microprocessor-controlled ac drive, plant uncertainties may come from nonlinearities of the power converter, fluctuation

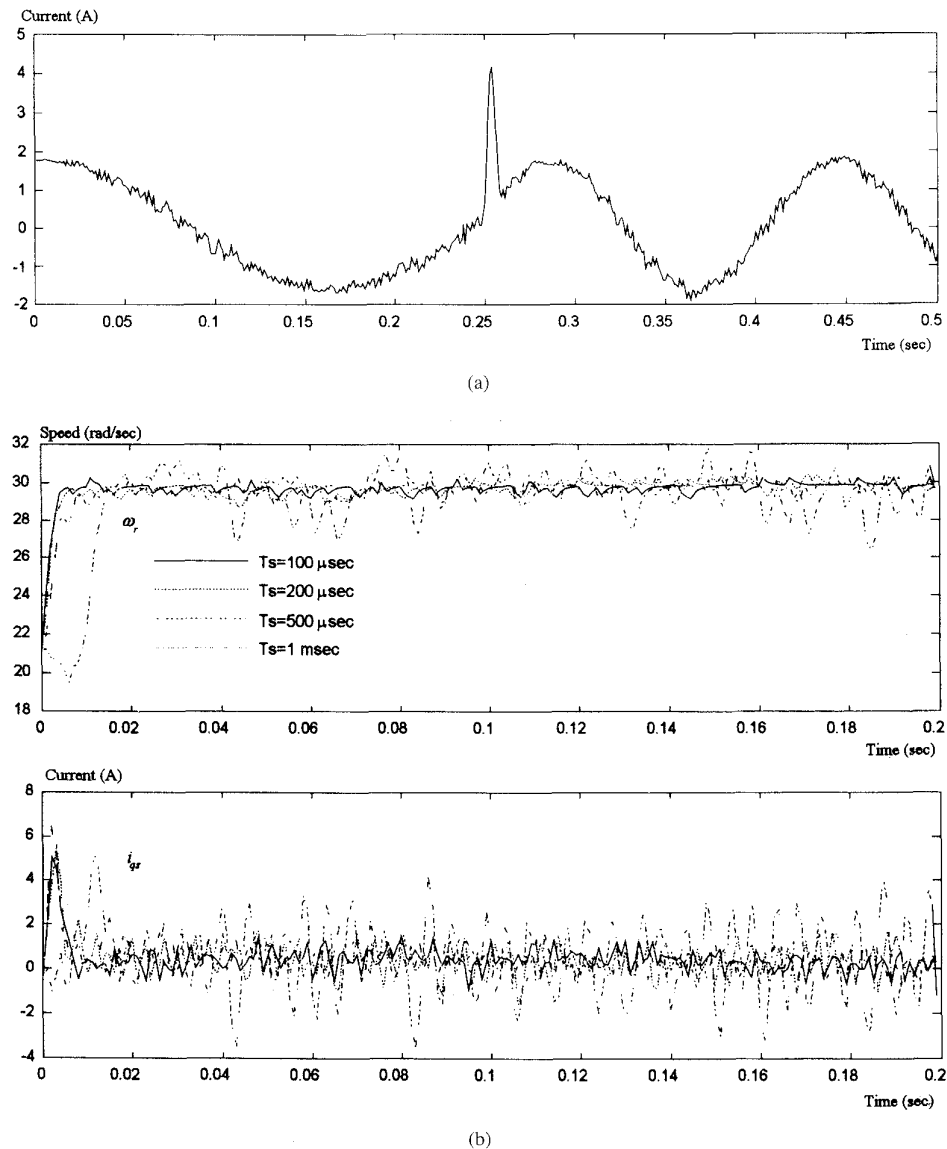


Fig. 10. Software current control results: (a) detected stator current with corresponding PWM signals and (b) speed and torque-producing current responses with different current loop sampling time.

of the dc link, motor parameters variation due to temperature change, inexact decoupling control, load variation, unmeasured torque disturbances, and high-frequency unmodeled dynamics. For example, if the rotor time constant varies due to changes in rotor temperature and saturation effect, the torque and voltage constants of the induction motor drive will be affected and the drive performance will be degraded.

Uncertainty can be described in many different ways: bounds on the parameters of a linear model, bounds on nonlinearities, frequency bounds, time domain bounds, etc. In this paper uncertainty bounds on the parameters of the decoupled linear model are used to study the plant uncertainties of the ac drive. To reflect the changing dynamics caused by the plant uncertainty, extreme operating conditions have to be considered in the design of a servo drive. The load torque is usually an unmeasurable disturbance from the load.

However, in steady state, the load torque disturbance can also be viewed as changes of the equivalent torque constant. The uncertainty bounds of the corresponding parameters of the designed ac drive are listed in Table I. These test conditions include shaft load changes, varying parameters, and different flux levels and operating speeds.

Fig. 5 shows some frequency responses of the uncertain plant under various operating conditions obtained by computer simulation using PC-MATLAB and frequency response measurement using dynamic signal analyzer HP3562A, respectively. To cover all possible uncertainty of the controlled plant, an uncertainty set Π based on the second-order plant transfer function (16) is defined as

$$\Pi = \{P(s) | P(s) = \frac{K}{(s\tau_m + 1)(s\tau_e + 1)} P_h(s)\}$$

$$K = K_o(1 + \delta_1), \tau_m = \tau_{mo}(1 + \delta_2), \tau_e = \tau_{eo}(1 + \delta_3) \quad (19)$$

where K_o, τ_{mo} , and τ_{eo} stand for the nominal values, K represents the equivalent dc gain, and a $P_h(s)$, of the form $(sa_z + 1)/(sa_p + 1)$, is included to cover the uncertainty in high frequency ranges. Once the uncertainty bounds of the plant are specified, we can employ the IMC control scheme to synthesize a robust controller that will achieve the desired performance and guarantee system stability for all possible operating conditions confined by the uncertainty set II.

Model uncertainty can be defined using either additive or multiplicative uncertainty. Multiplicative uncertainty is usually employed to define uncertainties in the frequency domain. According to the nominal model $\tilde{P}_n(s)$, the model uncertainty can be specified by a multiplicative uncertainty $l_m(s)$ with an upper bound $\bar{l}_m(s)$ as follows:

$$l_m(s) = \frac{P_n(s) - \tilde{P}_n(s)}{\tilde{P}_n(s)}, \quad |l_m(j\omega)| \leq \bar{l}_m(\omega), \quad \forall \omega. \quad (20)$$

The magnitude plot of the nominal model and upper bound of the uncertain plant are shown in Fig. 5.

In the given design example, the reference model is defined as the nominal model of the current-decoupled ac drive, therefore

$$\tilde{P}_n(s) = \frac{K_o}{(s\tau_{eo} + 1)(s\tau_{mo} + 1)}. \quad (21)$$

C. Synthesis of the IMC Controller

Once the reference model is specified, a nominal controller $D_n(s)$ is designed such that it is H_2 -optimal and achieves the nominal performance. That is, $D_n(s)$ is designed to solve the following H_2 -optimal problem for a particular input v

$$\min_{D_n} \|e\|_2 = \min_{D_n} \|(1 - \tilde{P}_n D_n)v\|_2. \quad (22)$$

The input v can be a specific input or specified by a set of bounded inputs

$$\mathbf{v} = \left\{ v: \|e\|_2^2 = \left\| \frac{v}{w} \right\|_2^2 = \frac{1}{2\pi} \int_{-\infty}^{\infty} \left| \frac{v(j\omega)}{w(j\omega)} \right|^2 d\omega \leq 1 \right\} \quad (23)$$

where $w(j\omega)$ is a weighting function used to normalize the bounded inputs. Since the reference model $\tilde{P}_n(s)$ of the servo system is stable and minimum-phase, (22) reaches its absolute minimum with $D_n(s) = \tilde{P}_n^{-1}(s)$. However, because $\tilde{P}_n^{-1}(s)$ is unrealizable, a second-order low-pass filter is added, and the nominal controller becomes

$$D_n(s) = \frac{\omega_o^2(s\tau_{eo} + 1)(s\tau_{mo} + 1)}{K_o(s^2 + 2\xi\omega_o s + \omega_o^2)} \quad (24)$$

where $\omega_o \gg 1/\tau_{eo}$ and $\xi = 0.707$.

After the nominal controller $D_n(s)$ has been determined, a robustness compensator $F(s)$ should be synthesized to ensure robust stability and performance when the plant is subjected to a set of specified uncertainty bounds. This design process can be simplified by first fixing the compensator structure and

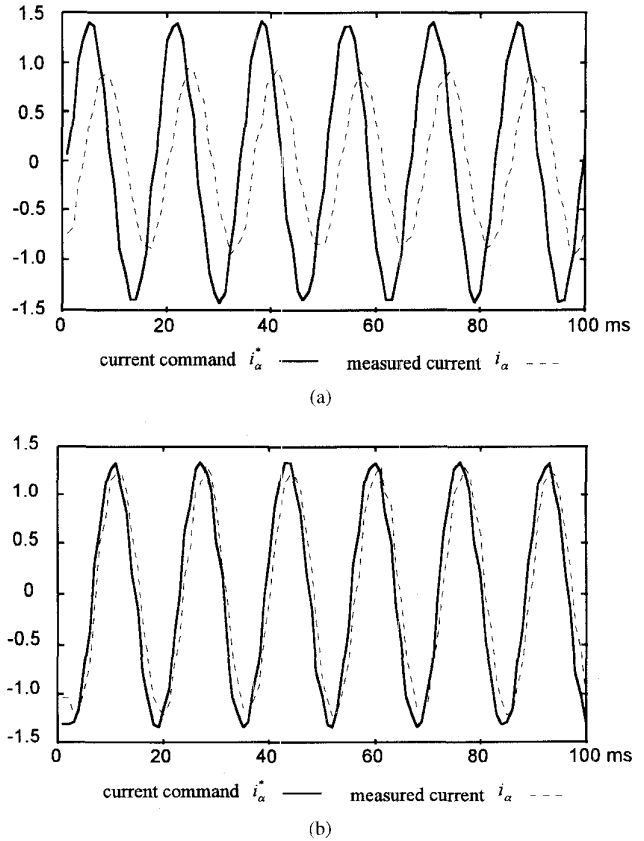


Fig. 11. Steady-state currents response using digital control scheme with (a) locked rotor and (b) free running at 3500 r/min.

then searching over its parameters to obtain desired robustness characteristics.

In the given design example, a first-order low-pass filter with an augmented integrator was chosen to make the IMC controller proper and ensure the offset-free tracking property. The defined robustness compensator $F(s)$ has a form of

$$F(s) = \frac{1}{s(s\lambda + 1)} \quad (25)$$

and the IMC controller becomes

$$D(s) = D_n(s)F(s) = \frac{\omega_o^2(s\tau_{eo} + 1)(s\tau_{mo} + 1)}{K_o(s^2 + 2\xi\omega_o s + \omega_o^2)} \cdot \frac{1}{s(s\lambda + 1)}. \quad (26)$$

This is a fourth-order controller. Considering the realization problem using integer arithmetic, this compensator can be reduced to third-order at the same time maintains its low-frequency characteristics. The reduced IMC controller becomes

$$D(s) = D_n(s)F(s) = \frac{\omega_o^2(s\tau_{eo} + 1)(s\tau_{mo} + 1)}{K_o s(s\lambda + 1)(s\tau_{me} + 1)} \quad (27)$$

where $\tau_{me} = \sqrt{\tau_{eo}\tau_{mo}}$ is the realization time constant. Now the only parameter of the IMC controller yet to be determined is the time constant λ of the robustness compensator. This

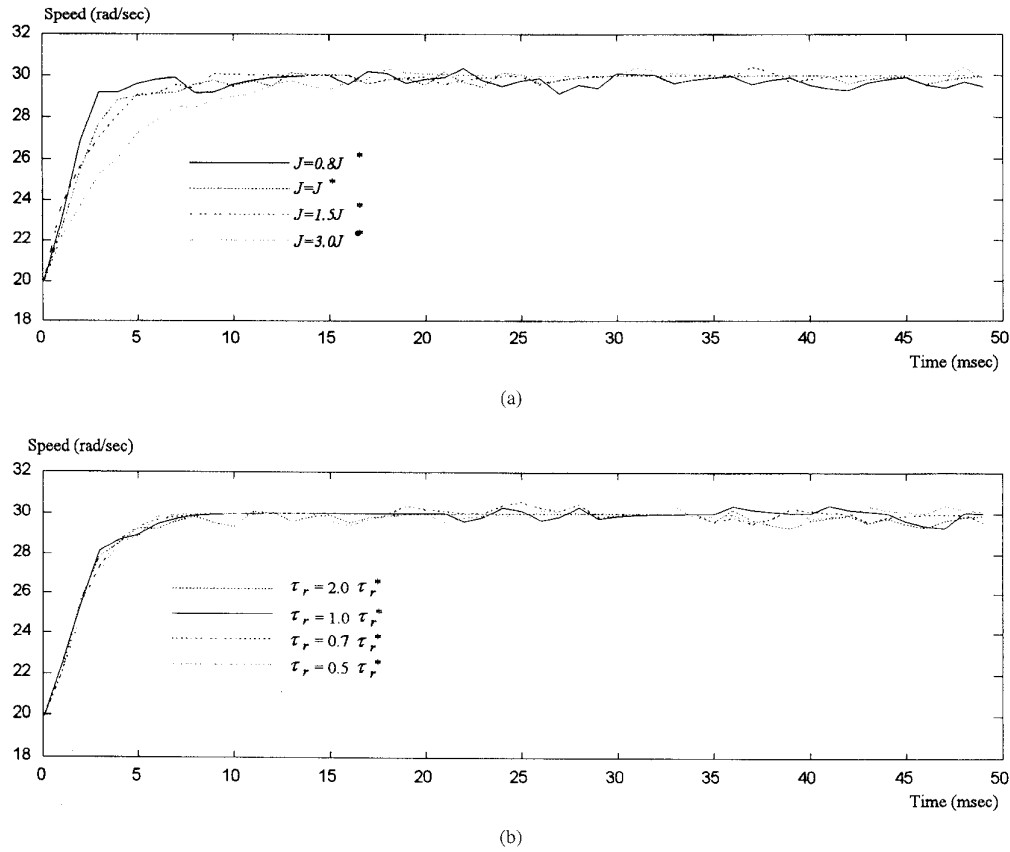


Fig. 12. Speed step responses with (a) variations of load inertia and (b) variations of rotor time constant.

parameter is determined by satisfying the stability and performance specifications of the H_∞ objective function and is discussed in the following.

D. Robust Stability and Robust Performance

Robust stability and robust performance are pursued by every control system designer in the synthesis of a robust control system. However, in practical applications, it is always hard to formulate a proper way in defining specifications of robust stability and robust performance. A discussion of the robust stability and robust performance of a control system based on H_∞ and μ theory can be found in [22]. The complementary sensitivity function expresses the effect of measurement noise and model uncertainty on the output of a control system and is of primary importance in judging the stability of a feedback controller. The nominal complementary sensitivity function of the designed ac drive is defined as

$$\eta_n(s) = \tilde{P}_n(s)D(s). \quad (28)$$

It can be shown that robust stability is guaranteed for the uncertain plant confined by the uncertainty set Π if and only if

$$\|\eta_n l_m\|_\infty = \sup_\omega |\eta_n(j\omega)| l_m(\omega) < 1. \quad (29)$$

With the given upper bound of the plant uncertainty, the time constant $\lambda \geq 0.0014$ is found to satisfy the robust stability criterion. Fig. 6 shows the gain plot of the robust stability given

by (29) for different design parameters. It can be observed that when the λ reaches 0.0014 the corresponding gain reaches to unit around 2000 rad/s. Since the plant uncertainty increases with the frequency, violations of the stability criterion will occur in the high-frequency region if this time constant is too small.

The sensitivity function expresses the effect of the external disturbance on the output of a control system and is of primary importance in judging the performance of a feedback controller. The sensitivity function of the induction servo drive system is

$$\varepsilon(s) = \frac{1 - \tilde{P}_n(s)D(s)}{1 + (P_n(s) - \tilde{P}_n(s))D(s)}. \quad (30)$$

The robust performance of the servo system is imposed by placing an upper bound on the magnitude of the sensitivity function

$$\|\varepsilon w\|_\infty = \sup_\omega |\varepsilon(j\omega)w(j\omega)| < 1, \quad \forall P \in \Pi \quad (31)$$

where $w(s)$ is the performance weighting function chosen to convey the characteristics of the inputs to be tracked and the disturbances to be rejected.

An appropriate form for $w(s)$ can be defined as

$$w(s) = \frac{\alpha(\beta s + 1)}{\beta s} \quad (32)$$

which supports the specifications of a lower limit on bandwidth ($1/\beta$) and an upper limit on peak magnitude ($1/\alpha$) [21]. According to the design specifications of the servo drive, $\alpha = 0.5$ and $\beta = 0.01$ are chosen for the performance weighting function. The time constant of the robustness compensator can be obtained by optimizing the robust performance index

$$\min_{\lambda} \sup_{\omega} (|\eta_n|l_m + |(1 - \eta_n)w|) < 1. \quad (33)$$

Fig. 7 shows the gain plot of the robust performance of (33) as a function of frequency for different λ . As a result, $\lambda = 0.004$ is obtained for the robust performance without violating the robust stability. The IMC controller is then obtained as

$$D(s) = 1.46 \frac{(0.83s + 1)(0.017s + 1)}{s(0.037s + 1)(0.004s + 1)}. \quad (34)$$

The performance weighting function in (32) can be viewed as an inverse bound on the magnitude of the sensitivity function. Fig. 8 shows the gain plot of the sensitivity functions of the compensated servo drive with specified model uncertainties. It can be observed that all the sensitivity functions are bounded by the inverse of the performance weighting function. This guarantees the robust performance under specified plant uncertainties.

IV. DSP REALIZATION AND EXPERIMENTAL RESULTS

After a thoughtful consideration of performance, price, simplicity in hardware design, and software support, we choose a single-chip DSP (TMS320C14) from Texas Instruments to realize the digital ac servo controller. The TMS320C14 has many good features which make it a good candidate to realize digital control for power converting systems, such as multiple independent programmable timers, 200-ns instruction cycle, 16-bit parallel multiplier, and on-chip RAM and ROM, etc. [12]. The hardware circuit of the proposed DSP-based fully digital servo controller consists of a 16-bit TMS320C14 single-chip DSP, six-channel multiplexed 12-bit analog-to-digital converter, 4K-word external program memory, and an RS232 computer interface. The TMS320C14 includes the on-chip peripherals necessary for industrial control. These peripherals include four 16-bit timers, two general-purpose timers, a watchdog timer, a baud-rate generator, 16-bit programmable input-output, a serial port, and an event manager with six-channel PWM output. The event manager consists of a six-output-compare subsystem and a four-input-capture subsystem. The PWM output waveform can vary from eight-bit of resolution at 100 kHz to 14 bit at 1.6 kHz.

Fig. 9 shows the software flowchart for the fully digital control of an ac induction motor. The control processing is performed by two subroutines: SR1 performs the IMC control and vector control algorithm and SR2 executes the software current control in dq -axis. The subroutine SR1, which is interrupted at a sampling rate of 1 kHz, first reads the servo command and motor encoder feedback, and then performs the IMC and field-weakening control algorithm to generate the desired torque current command i_q^* and field current command i_d^* in the synchronous rotating reference frame. The SR1 then performs the indirect vector control to calculate the

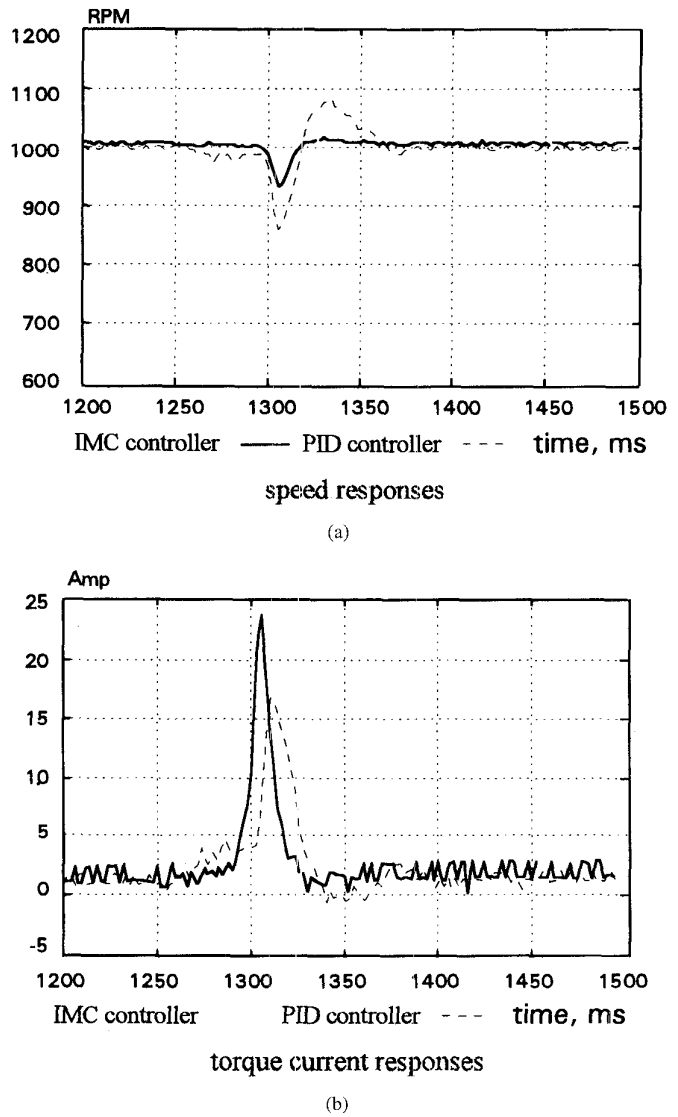


Fig. 13. Speed regulation with variation of step torque disturbance (a) speed and (b) torque current.

rotor flux and transforms the current components between the synchronous and stationary coordinates. Subroutine SR2, which is interrupted at a sampling rate of 10 kHz, reads the measured stator currents from the A/D converters and transforms them to stationary α - β coordinates to execute the current regulation algorithm.

As a result of the current regulation, stator voltage components are obtained and used for the PWM gating signal generation. Detection of the shaft speed and generation of the PWM signals are implemented by using the DSP event manager. The input-capture subsystem is used to count the transients of the shaft encoder signals and measure the time difference between them. This time difference is then used to calculate the shaft rotation speed. The timing of the PWM gate pulses is controlled by the output-compare subsystem with 11 bit of resolution at 20 kHz of carrier frequency.

The designed IMC controller in the s -domain must be transformed to the z -domain for discrete realization. There

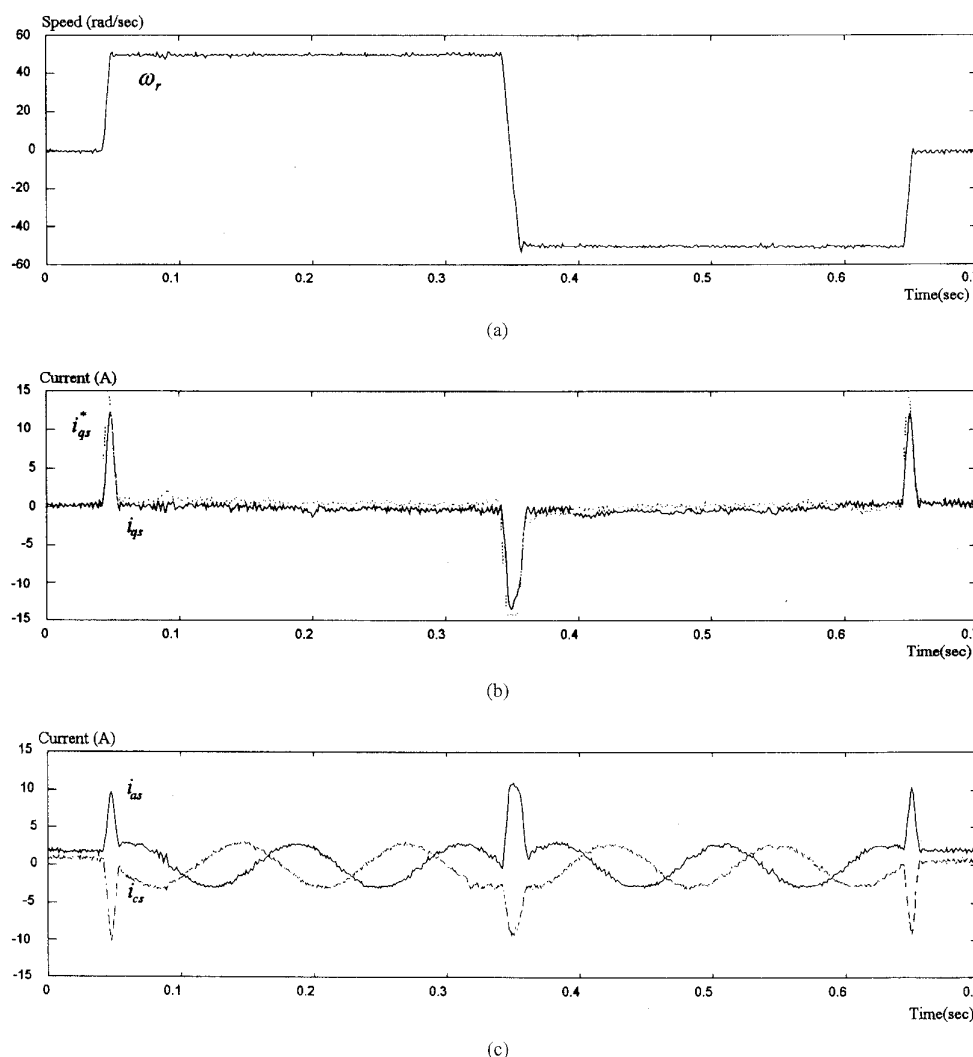


Fig. 14. Characteristics of four-quadrant speed operation: (a) rotor speed, (b) torque-producing current and its reference, and (c) stator currents.

are many methods for the discretization of analog controllers. However, it should be noted that transformation from the analog domain to the digital domain may also destroy some important system properties. This is particularly true when the sampling rate is not far above the crossover frequency of the analog controller. The bilinear transformation is generally preferred when the frequency response characteristics of the analog controller are to be preserved in the transformed digital controller.

Due to the integer arithmetic of the selected DSP, selection of the controller structure and scaling of the control parameters are especially important in the realization of the transformed digital controller. The direct form two has been used to realize the transformed digital controller. The direct forms have the lowest complexity in terms of the number of nontrivial multiplications and additions, and can be programmed with fast processing speed. However, it should be noted they are sensitive to small changes in coefficients and should not be used to realize high-order controllers. It should also be noted that digital control algorithms become particularly sensitive

to numerical errors when systems are controlled under a fast sampling rate. And to minimize the numerical errors induced by the finite word-length effect, the IMC controller is first decomposed to three cascaded controllers and then realized using the direct form two structure. Some multiplications can be implemented using shift instructions to achieve high-speed signal scaling. To avoid offset due to round-off error, variables are stored with double precision.

The motor under test is an 800 W three-phase induction servo motor with two poles and Y-connected windings. The rotor shaft of the induction motor was connected to a current controlled magnetic brake. A programmable current source is connected to the magnetic brake to simulate viscous friction and load disturbances. The position and speed of the motor shaft were detected by a rotary encoder with 2000 pulses per revolution (PPR).

Experimental results of the software current-controlled ac drive during a transient response are presented in Fig. 10. Fig. 10(a) shows the controlled motor phase current and the corresponding PWM gating signals when the motor speed is

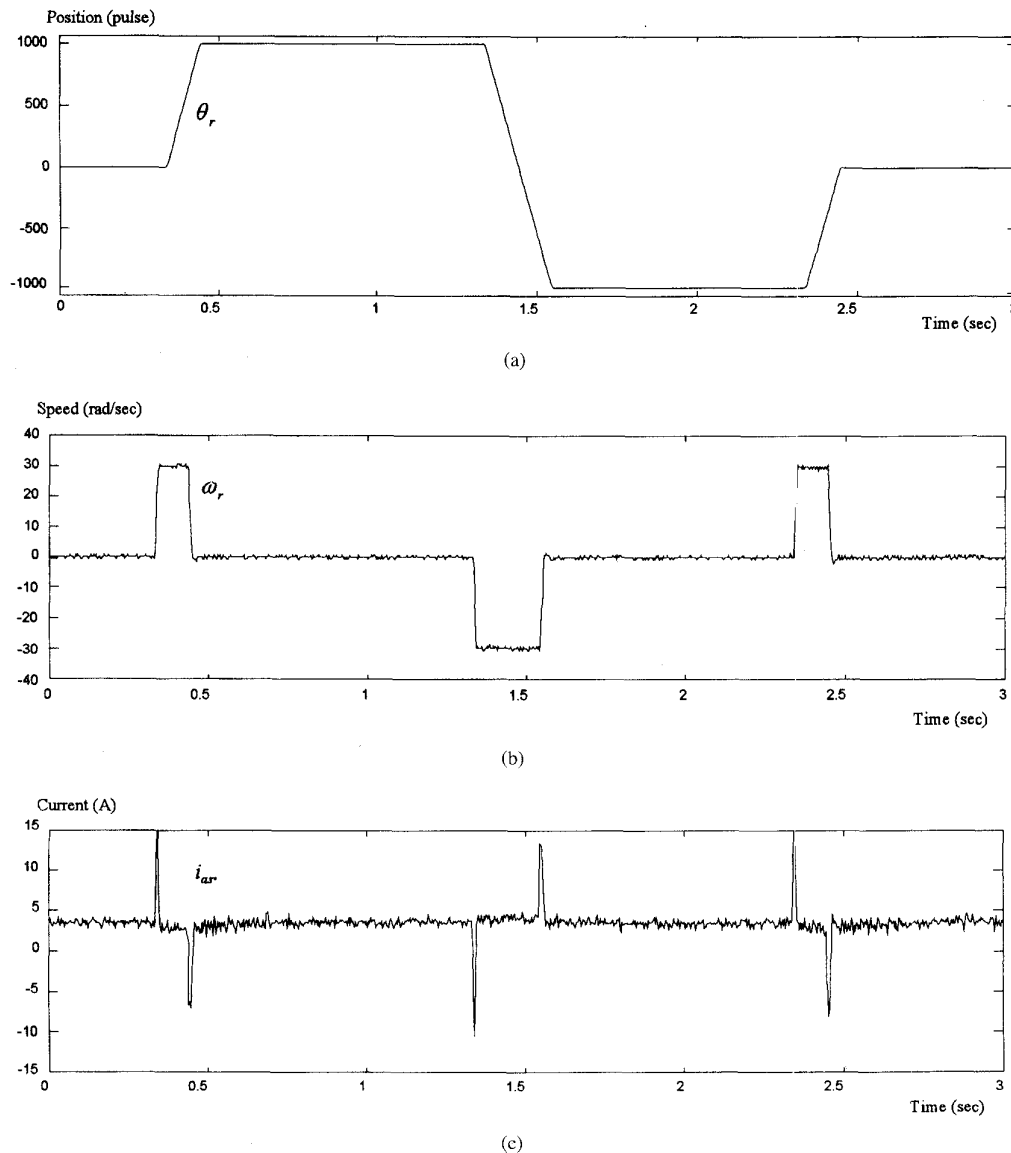


Fig. 15. Position step response: (a) rotor position, (b) rotor speed, and (c) stator current.

changed from 20 to 40 rad/s. The duration of the current peak for speed acceleration is about 5 ms. This result reveals that the software current control scheme is not only feasible, it also can achieve fast dynamic response. Fig. 10(b) illustrates the influence of the current-loop sampling rate on the transient and steady-state responses when the motor speed changes from 20 rad/s to 30 rad/s. The drive performance suffers from inferior current regulation and large current ripple when the software current loop sampling time is extended from 100 μ s to 1 ms. This shows that a fast sampling frequency is needed for the software current feedback loop. Fig. 11 shows steady-state current responses under digital current regulation with sampling time set at 100 μ s. Fig. 11(a) shows the current responses of the motor under closed-loop speed regulation with speed command set at 3500 r/min. The measured stator current has an angular velocity of about 66 Hz. In Fig. 11(b), the same current command has been applied to the digital current

controller with disconnected speed loop and locked rotor. The measured phase current has a phase lag of 18°. Experimental results show that the controlled ac drive will result a larger current phase lag due to the back emf.

Fig. 12 shows the robustness of the step responses under variations of rotor inertia and rotor time constant. In Fig. 12(a) the rotor inertia is changed from 0.8 to 3.0 of its nominal value and in Fig. 12(b) the rotor time constant is changed from 0.5 to 2.0 of its nominal value. It can be observed that the IMC control scheme can achieve robustness to both load and motor parameter variations. Fig. 13 shows the speed regulation due to a step load torque disturbance generated by a magnetic brake. Experimental results of the designed IMC controller and a well-tuned PID controlled have been compared. Experimental results reveal that the IMC controller will exert a larger torque correct command to reduce the unknown external torque disturbance and a more robust speed

regulation can be achieved using the proposed control scheme.

Fig. 14 shows experimental results for four-quadrant operation when the speed command changes from zero to 50 rad/s in forward rotation and then changes to reverse direction at the same speed. Fig. 15 shows the time responses of the position servo under a step position change of 180° , corresponding to 1000 pulses for a 2000 PPR encoder. Experimental results show satisfactory dynamic responses for both speed and position control can be achieved using the proposed IMC control scheme.

V. CONCLUSION

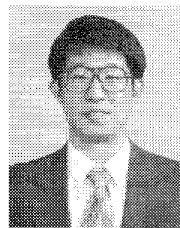
In this paper, we have illustrated the design and implementation of a DSP-based fully digital-controlled ac induction servo drive using the proposed IMC control scheme. With knowledge of the norm-bounded plant uncertainty, the IMC structure effectively clarifies the design problems related to the plant uncertainty and generates a conventional controller of the desired order by using the modern optimal sensitivity theory. With the given design procedure and obtained experimental results, we can see the IMC control scheme can achieve robust stability and robust performance with specified uncertainty bounds and input constraints. However, one should also note that an improper specification of the uncertainty bound or an inappropriate selection of the performance weighting function may lead to a conservative design or even no solution. The DSP-based fully digital control scheme eliminates the inherent drawbacks of the analog circuits used in conventional current-loop control and PWM gating signal generation. This paper reveals the feasibility of implementing advanced control scheme using DSP in the design of high-performance ac servo drive.

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Ying-Yu Tzou (S'81-M'88) was born in Taiwan, R.O.C., on February 13, 1956. He received the B.S. and M.S. degrees in control engineering from the National Chiao Tung University, Hsinchu, Taiwan, and the Ph.D. degree in electrical engineering from the Institute of Electronics Engineering of National Chiao Tung University in 1978, 1983, and 1987, respectively.

During 1980 to 1981 he was with the Electronic Research and Service Organization (ERSO) of the Industry Technology Research Institute (ITRI) as a Design Engineer in the Control System Department for the Factory Automation. During 1983 to 1986 he was with Microtek Automation, Inc., as a Project Manager for the development of a computer numerical controller (CNC) for machine tools. He is currently an Associate Professor of the Department of Control Engineering of National Chiao Tung University and also serves as an Industrial Consultant. He was the Director of the Institute of Control Engineering during 1992 to 1994. He has authored numerous papers on power electronics, motion control, ac servo drives, parallel processing and control systems, and DSP applications in the control of power converters and motor drives.