High-Performance Velocity Servo-System Design Using Active Disturbance Estimator

Boban Veselić[1] and Čedomir Milosavljević[2]

***Abstract* – This paper considers design of digitally controlled high-performance velocity servo-system, featuring fast response without overshoot. The proposed control structure contains main PI controller and active disturbance estimator, to further improve disturbance rejection dynamics. Resulting response meets the high-performance requirements. The designed control system is experimentally verified in induction motor velocity control.**

***Keywords* – Servo-system, Velocity control, Digital PI controller, disturbance estimator.**

I. INTRODUCTION

Advanced production technologies have imposed more rigorous demands on servo-systems performances. Velocity servo-systems in high-performance applications must comply with the requirements of fast response without overshoot, high steady state accuracy, good rejection of external disturbances and robustness to parameter perturbations. In general, a servo-system must have at least one pure integrator within the closed loop. Since control plant in a velocity servo-system does not have an integrating property, PI controllers are conventionally used. Such servo-system has no steady-state error on step references and completely rejects constant loads.

Beside standard simple regulation contour, two degrees-of-freedom controllers found their applications in servo-systems [1], in order to improve robustness. Reference and disturbance responses can be separately designed using this control structure. Similar results were obtained using internal model principle and internal model control, combined into IMPACT structure [2]. Another approach in disturbance compensation is introduction of disturbance estimators [3], which may be interpreted as a special case of the above two structures.

This paper deals with the design of a digitally controlled high-performance velocity servo-system. Both pole placement and zero-pole cancellation design methods of PI controller are considered. To further improve disturbance rejection dynamics an active disturbance estimator [4] is introduced. Fast response without overshoot and excellent disturbance rejection properties are ensured. The proposed servo-system is experimentally verified in induction motor velocity control.

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II. VELOCITY SERVO-SYSTEM STRUCTURE

Most of velocity servo-systems, regardless of the employed drive, may be described by the simplified generalized block-scheme depicted in Fig. 1. The cascade structure consists of two distinct control loops. Inner current control loop, responsible for adequate torque generation, is enclosed by a main speed control loop. Bandwidth of the inner loop is usually much higher then bandwidth of the speed loop, thus the current control subsystem dynamics may be ignored in the main controller design. Nevertheless, dynamic delay of the inner control subsystem has certain impact on overall system dynamics, acting as an unmodeled dynamics within the speed control loop. The current loop is commonly realized with a bandwidth around $1\,kHz$, implementing PI controller. Since exogenous disturbances, such as load torque $M_o(t)$, enter directly into the speed control loop, PI speed controller is usually employed in order to eliminate steady-state error in the case of step-like disturbances.

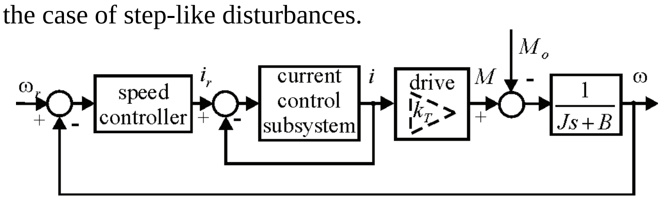

Fig. 1. Generalized velocity servo-system structure

Encouraged by huge technological advance of micro-controllers, digital implementation of control algorithms overcomes analog counterpart. Correct system analysis should be carried out in discrete-time domain. Block diagram of a digitally controlled velocity servo-system is given in Fig. 2. $G_r(z)$ is a discrete-time transfer function of digital controller, $G_{h0}(s)$ is sample and hold transfer function, k_T is a torque constant, J and B are moment of inertia and viscous friction coefficient, respectively. Hence, motor dynamics is described by first order function $G(s)=k_m/(1+sT_m)$, where $k_m = k_T/B$ is motor gain and $T_m = J/B$ is time constant.

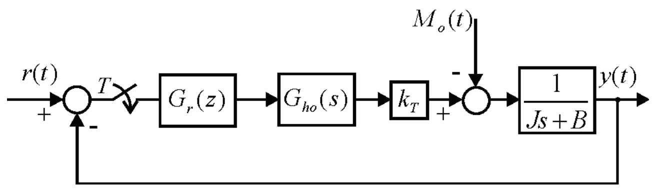

Fig. 2. Digitally controlled velocity servo-system

If load torque is a step function or is slowly varying between two consecutive sampling instants, which is true for small sampling periods, discrete-time model of the closed loop system is given as

$$Y(z)=\frac{G_r(z)G(z)}{1+G_r(z)G(z)}R(z)-\frac{G(z)}{1+G_r(z)G(z)}M_{oe}(z),$$
$$G(z)=\mathcal{Z}\{\mathcal{L}^{-1}\{G_{h0}(s)G(s)\}\}=\frac{k_m(1-a)}{z-a},\; a=e^{-T_s/T_m}, \tag{1}$$

where T_s denotes sample period and $M_{oe}=M_o/k_T$. As mentioned earlier, PI digital controller is employed within speed control loop. Controller transfer function is defined with

$$G_r(z)=k_p+k_i\frac{z}{z-1}=\frac{k_p}{b}\frac{z-b}{z-1},\; b=\frac{k_p}{k_p+k_i}, \tag{2}$$

where k_p, k_i are gains of proportional and integral action.

III. Controller Design

According to the high-performance requirements, recounted in the introduction, controller (2) parameters should be tuned to provide fast critically aperiodic response. Due to (1) (2), characteristic equation $1+G_r(z)G(z)=0$ is of second order, whose roots z_1, z_2 (poles of the closed loop system (1)) determine system response. As it is well known, second order system dynamics as well as nature of its response is defined by doublet ς, ω_n, which represents relative damping factor and undamped natural frequency, respectively. Closed loop poles are given by $s_{1,2}=-\zeta\omega_n\pm\omega_n\sqrt{1-\zeta^2}$. In the case of critically aperiodic response, $\zeta=1$ yielding double real pole $s_1=s_2=-\omega_n$. In discrete-time domain these poles are mapped into locations $z_1=z_2=e^{-\omega_n T_s}$, $0<z_1\le 1$. To obtain desired closed loop dynamics defined by z_1 using pole placement design technique [5], k_p and k_i of (2) should be

$$k_p=\frac{a-z_1^2}{k_m(1-a)},\quad k_i=\frac{1-2z_1+z_1^2}{k_m(1-a)}, \tag{3}$$

which results in a closed loop dynamics of the form

$$Y(z)=(1+a-2z_1)\frac{z-z_0}{(z-z_1)^2}R(z)-\frac{k_m(1-a)(z-1)}{(z-z_1)^2}M_{oe}(z), \tag{4}$$

showing that the desired poles are ensured. Consequently, a zero $z_0=(a-z_1^2)/(1+a-2z_1)$ arises in the response with respect to reference. Its location depends on the location of the desired pole z_1, which is plotted in Fig. 3. For zero z_0 not to be dominant, condition $z_0<z_1$ must hold. According to Fig. 3, valid selection of desired pole z_1 is $a<z_1<\sqrt{a}$ which gives $T_m<1/\omega_n<2T_m$. This indicates that the time constant of the closed loop system should be larger then motor time constant. This is unacceptable since notion of feedback control is to improve not to degrade system dynamics, so the desired pole must be located inside region $0<z_1<a$. This introduces dominant zero, which produces unwanted overshoot. Notice that system response with respect to load torque is free of undesirable zero.

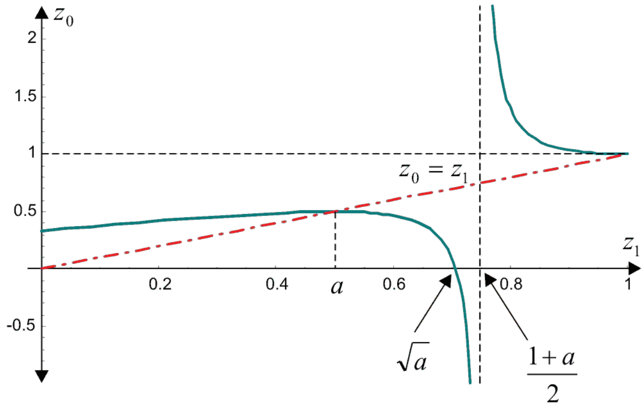

Fig. 3. Location of zero with respect to location of chosen pole

One approach in overshoot elimination is introduction of referent signal filtering, which would slow down the system. This is contradictory to the high-performance requirements.

Another design approach is zero-pole cancellation method, [5]. Namely, controller gains should be set in such manner that controller zero b cancels plant pole a, ($b=a$), and gain k_p determines location of the desired closed loop pole z_3. Hence, controller parameters are obtained in the form

$$k_p=\frac{a(1-z_3)}{k_m(1-a)},\quad k_i=\frac{1-z_3}{k_m}. \tag{5}$$

Closed loop system behavior is then described by

$$Y(z)=\frac{1-z_3}{z-z_3}R(z)-\frac{k_m(1-a)(z-1)}{(z-a)(z-z_3)}M_{oe}(z). \tag{6}$$

The system has first order dynamics defined by desired pole z_3 with respect to reference, providing fast response without overshoot. However, complete cancellation does not occur in disturbance related term, which is described by second order dynamics. Furthermore, behavior with respect to disturbance is determined by "slow" pole of the plant $z=a$, which is now dominant. Reference response is quite satisfactory, whereas disturbance rejection dynamics is unacceptable. In order to obtain high-performance servo-system it is necessary to improve disturbance rejection performance.

IV. Active Disturbance Estimator

A way to improve system robustness to parameter perturbation and exogenous disturbances is introduction of disturbance estimator. The concept of disturbance estimator is that the external disturbances and model uncertainties, usually regarded as an equivalent disturbance, can be efficiently compensated by feedback of the estimated value.

Consider the control structure in Fig. 4, consisting of a real plant $G(z)$ and disturbance estimator in the local loop. Equivalent disturbance q is evaluated inside the disturbance estimator employing discrete transfer function of the plant nominal model $G_n(z)$. A local feedback for the disturbance compensation is closed via digital filter $G_k(z)$. Due to the uncertainties of the plant parameters, the mismatch between real plant and nominal model inevitably exists. The real plant may be described as $G(z)=G_n(z)(1+\delta G(z))$, where the perturbation is limited by the multiplicative bound of uncertainty $\left|\delta G(e^{j\omega T})\right|\le\gamma(\omega),\ \omega\in\left[0,\pi/T\right]$. Plant output is

$$\begin{aligned}Y(z)=&\frac{G_n(z)(1+\delta G(z))}{1+G_k(z)G_n(z)\delta G(z)}U(z)\\&+\frac{G_n(z)(1+\delta G(z))(1-G_k(z)G_n(z))}{1+G_k(z)G_n(z)\delta G(z)}M_{oe}(z).\end{aligned} \tag{7}$$

Suppose that $G_k(z)=G_n^{-1}(z)$, i.e., digital filter represents nominal plant inverse dynamics. Using (7) plant output becomes $Y(z)=G_n(z)U(z)$, which indicates that the disturbances are completely rejected and the nominal plant behavior is obtained. Unfortunately, such $G_k(z)$ is not a causal filter, which cannot be realized. It is evident from (7) that the model perturbation $\delta G(z)$ affects the stability of the system. Robustness of the proposed structure against model uncertainties is, therefore, limited to the level of the model perturbation $\delta G(z)$ quantified in a suitable way for which input-output transfer function (7) remains stable.

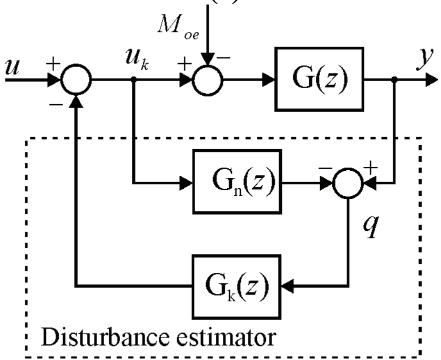

Fig. 4. Structure of digital disturbance estimator

In [4] an active disturbance estimator is proposed, where passive digital filter $G_k(z)$ is replaced with an active control subsystem, Fig. 5. The signal $\hat{q}$ is an estimate of the compensated part of the equivalent disturbance. If controller $G_{r2}(z)$ ensures $\hat{q}=q$, $U_e(z)=G_n^{-1}(z)Q(z)$ holds, which is equivalent to the passive structure with ideal digital filter $G_k(z)=G_n^{-1}(z)$. Hence, nominal plant behavior is obtained, $Y(z)=G_n(z)U(z)$. From the control design aspect, problem of equivalent disturbance compensation is here transformed into tracking control problem with referent signal $q(k)$. Depending on the applied controller inside estimator certain error between q and $\hat{q}$ exists in general case, implying that complete equivalent disturbance rejection cannot occur and the obtained plant behavior is almost nominal.

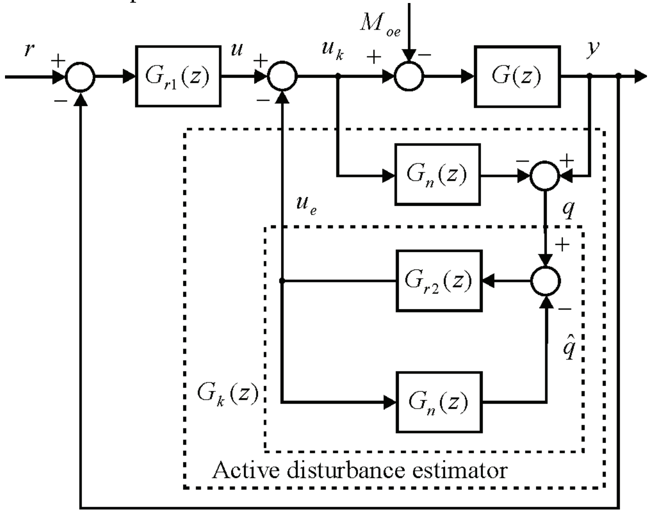

Fig. 5. Servo-system with active disturbance estimator

V. Servo-system synthesis

The proposed servo-system is depicted in Fig. 5. Both controllers, in the main loop $G_{r1}(z)$ and within estimator $G_{r2}(z)$, governs nominal plant and model, respectively, since disturbance estimator forces the real plant to exhibit nominal behavior. Hence, identical controllers designed for a nominal plant may be used in the main loop as well as in the estimator.

For sake of expressions simplicity suppose that plant parameter identification is done correctly and parameter uncertainties are not significant, ($\delta G\approx 0$). The proposed servo-system dynamics is then given by

$$\begin{aligned}Y(z)=&\frac{G_n(z)G_{r1}(z)}{1+G_n(z)G_{r1}(z)}R(z)\\&-\frac{G_n(z)}{(1+G_n(z)G_{r1}(z))(1+G_n(z)G_{r2}(z))}M_{oe}(z).\end{aligned} \tag{8}$$

Clearly from (8), system performance with respect to reference is directed only by the main controller and identical to the system without disturbance estimator (Fig. 2, eq. (1)). Since cancellation design method results in a satisfactory response to the reference, the main controller may be realized as PI type (eq. (2)) with the parameters tuned according to (5).

However, both controllers participate in disturbance rejection. Since the main controller already has integral action, in case of step-like exogenous disturbances it is sufficient for estimator controller to be P type, $G_{r2}(z)=k_{p2}$. Gain k_{p2} is tuned using pole placement technique, where z_4 is desired pole introduced by the estimator. Consequently,

$$k_{p2}=\frac{a-z_4}{k_m(1-a)}. \tag{9}$$

With the controllers tuned according to (5) and (9), servo-system dynamics become

$$Y(z) = \frac{1-z_3}{z-z_3}R(z) - \frac{k_m(1-a)(z-1)}{(z-z_3)(z-z_4)}M_{oe}(z). \qquad (10)$$

It is obvious that the proposed structure with active disturbance estimator offers possibility to obtain both responses, with respect to the reference and exogenous disturbances, acceptable for high-performance servo-systems. Overshoot does not arise and system dynamics is completely defined by the freely adopted poles z_3, z_4.

To completely reject ramp-like exogenous disturbances estimator controller must be of PI type with the parameters defined with eq. (3). System dynamics is then described by

$$Y(z) = \frac{1-z_3}{z-z_3}R(z) - \frac{k_m(1-a)(z-1)^2}{(z-z_3)(z-z_1)^2}M_{oe}(z). \qquad (11)$$

VI. Experimental Investigation

The effectiveness of the proposed control structure has been investigated by experiments, which have been conducted on a servo-system with a three phase, 50Hz, 0.37kW, Seiber LS71 induction motor with 1.3 Nm nominal torque. Control part of the servo-system is realized by dSPACE DS1104 R&D controller board. Indirect rotor flux oriented vector control of induction motor is employed. The control scheme contains measurement of two line currents and rotor shaft angle position, coordinate transformations, decoupling circuits, electrical angle estimator, two local current control loops with 200 Hz bandwidth and 10 kHz sampling frequency, and a main speed control loop with 1 kHz sampling frequency.

By neglecting current control loops along with other unmodeled dynamics and nonlinearities, present in such complex system, in certain approximation speed control system may be considered as one in Fig. 2. In this representation motor parameters are: $k_T = 0.6481$ Nm/A, $J = 3.5\cdot10^{-4}$ kgm^2, $B = 3\cdot10^{-4}$ Nms/rad. Reference signal is given with $r(t) = 100\cdot h(t-0.5)$ rad/s, and the system is subjected to load torque $M_o(t) = 0.65\cdot h(t-1.5)$ Nm, which is 50% of nominal torque.

In the first experiment (Fig. 6., trace (1)), active disturbance estimator is deactivated and the main PI controller is designed using pole placement. Desired dynamics is defined by $\zeta = 1$, $\omega_n = 20$ rad/s ($z_1 = 0.98$). $k_{p1} = 0.0207$, $k_{i1} = 2.118\cdot10^{-4}$ are obtained using (3). An unwanted overshoot arises due to the dominant zero $z_0 = 0.99$.

The main controller is then tuned by cancellation method in the second experiment (Fig. 6., trace (2)). Desired bandwidth of 10 Hz results in pole $z_3 = 0.939$, which according to (5) gives $k_{p1} = 0.033$, $k_{i1} = 2.819\cdot10^{-5}$. System has good response to reference, but very slow dynamics of disturbance rejection, caused by not cancelled plant pole a.

Finally, active disturbance estimator is activated in the third experiment (Fig. 6. trace (3)). The main PI controller remains unchanged from the previous experiment, while P controller in the estimator is set by applying pole placement under condition $z_4 = z_3$, resulting in $k_{p2} = 0.0324$ by virtue of (9). The proposed servo system is superior to the other two, with fast velocity response without overshoot and equally fast dynamics in disturbance rejection.

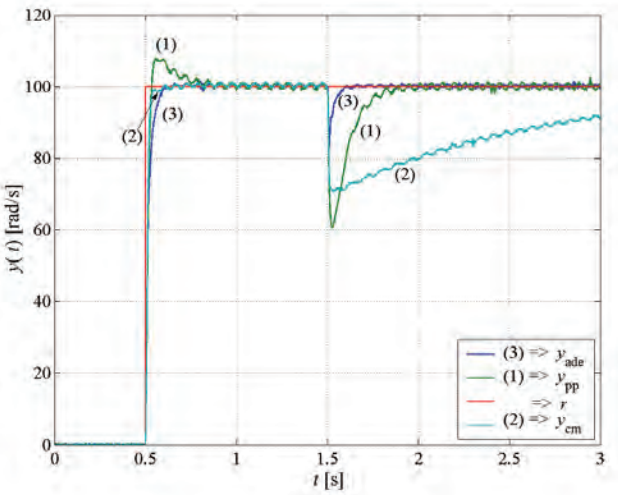

Fig. 6. Velocity servo-system step responses

VII. Conclusion

The paper considers design of high-performance velocity servo-system with active disturbance estimator, whose introduction drastically improves system behavior with respect to exogenous disturbances. This ensures fast response without overshoot and system dynamics is completely defined by freely adopted poles. Analytically predicted performance has been experimentally verified in case of induction motor servo–system, in which significant modeling error and parameter uncertainties exist. The proposed servo-system has exhibited excellent exogenous disturbance rejection property as well as robustness to parameter perturbations.

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