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A Digital Speed Filter for Motion Control Drives with a Low Resolution Position Encoder

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In motion control drives, the motor rotation speed measurement is generally obtained by means of an incremental encoder with a limited number of counts per revolution. Since encoders furnish a shaft position measure, a derivative operation is needed to obtain a speed measure. As consequence, especially at lower speeds, the rotation speed measure results rather noisy and a suitable filter must be used to reduce the noise.

After an overview on the speed measurement problems connected to the employment of an incremental encoder, the paper presents the algorithms of an original digital filter, based on a static Kalman filter; then, an efficient procedure for the implementation of such a filter on a 16-bit fixed point microcontroller is carried out. Proposed filter reduces the speed measurement noise to acceptable values and it can also provide an estimation of the acceleration. Finally, proposed filter has been employed on an industrial drive for machine tools.

Key words: encoder, digital filter, drives, Kalman filter, speed measurement, transducers

1 INTRODUCTION

To measure the motor rotation speed, motion control drives generally use an incremental encoder that provides two quadrature pulse trains, which frequency is proportional to the rotation speed. This pulse trains are sent to a Quadrature Encoder Pulse Circuit (QEPC), commonly inserted in modern microcontrollers used for the drive control, that count the number of rise and fall fronts of both the channels and furnishes the shaft position and speed. Due to the limited number of fronts in each sampling time, the shaft position measure is corrupted by a quantization noise; therefore the speed measure, is rather noisy, except when an incremental encoder with a great number of pulses per revolution is employed.

In general, to reduce the noise, a suitable FIR or IIR digital filter is used. This filter can lead to satisfactory noise values, but it introduces a significant lag, also when a moderate noise reduction is required; other types of filters, for example state variables filters, give similar results. A different solution consists of employing a state observer that provides a speed estimation based on the position measure; this observer can be of deterministic type [2] or of stochastic type, based on a Kalman filter [3–7]. The last solution allows a greater flexibility; so it was been preferred.

In a previous paper [7], the authors, have proposed to employ, for reducing the speed noise, a suitable static Kalman filter, based on a third order linear and time-invariant observation model. After a brief recall to the filter structure, the present paper deals with the problems connected to its implementation on a fixed point DSP controller and the influence of the scaling factors relevant to a fixed-point implementation and it concludes with the presentation of some experimental results.

2 SPEED MEASUREMENT PROBLEMS

Denoted with T the sampling period employed for the speed measurement, the rotation speed (in radians per second) can be obtained by the following expression:

$$\omega = \frac{2\pi n}{n_{tic} T}, \tag{1}$$

in which

- n is the number of pulses counted by the QEPC in the sampling period,
- n_{tic} is the number of pulses per revolution.

The digitalisation connected to the integer representation of n produces a quantization noise equal to

$$\Delta\omega = \frac{2\pi}{n_{tic} T}. \tag{2}$$

As an example, assuming $n_{tic} = 8\,192$ (i.e. two channel, each with 2 048 pulses per revolution) and a sampling period equal to 150 μs, the quantization noise is equal to

$$\Delta\omega = \frac{2\pi}{8192 \cdot 150 \cdot 10^{-6}} = 5.113 \ \text{rad/s}.$$

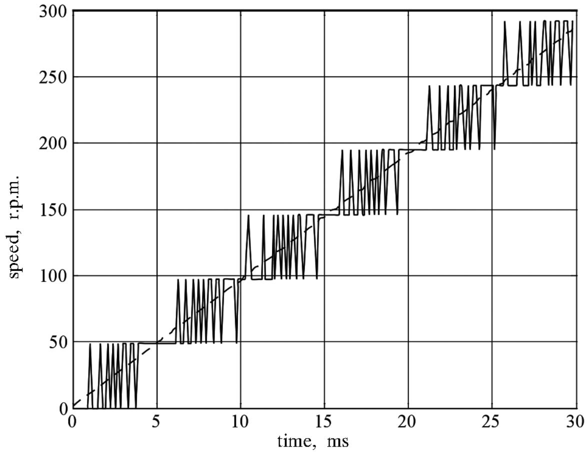
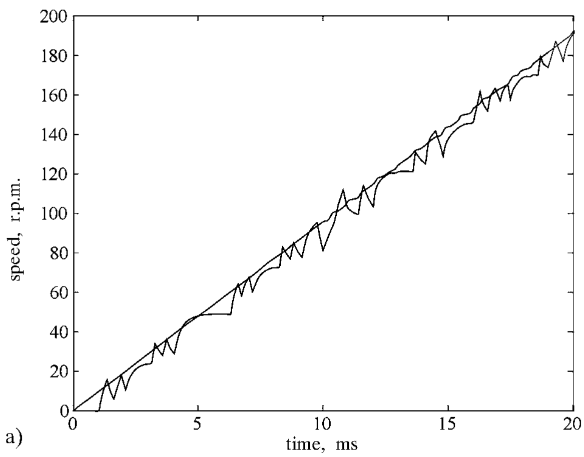
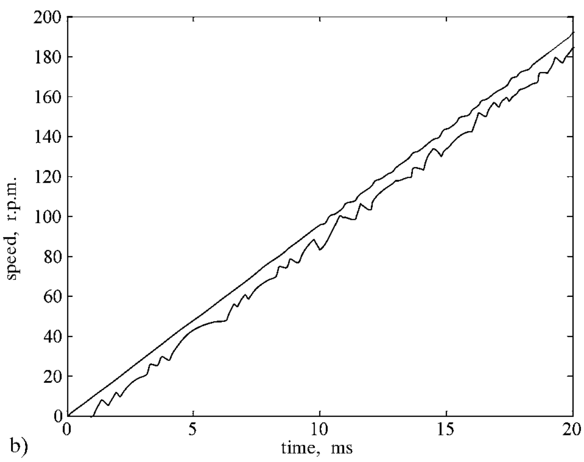


Fig. 1 Real and measured speed shapes

Figure 1 shows the real (dashed line) and measured (continuous line) shapes of the speed during a starting transient characterized by a constant acceleration equal to 1000 rad/s^2.



a)



b)

Fig. 2 a) Real and filtered speed shapes with 2T time constant filter, b) Real and filtered speed shapes with 5T time constant filter

If the measured speed is directly used for the speed control, the loop gain must be limited to a value much lower of the maximum gain that could be employed in noise absence. This is especially critical for AC drives because of the need for high current loop bandwidths to achieve proper instantaneous field orientation of the flux and current vectors [1, 2].

In general, to reduce the noise, a suitable digital filter is used. Figures 2a and 2b show the speed signal of Figure 1 filtered by a digital filter composed by a second order FIR filter and a transfer function obtained by digitalizing an analog filter with a time constant equal to $2T$ (Figure 2a) and to $5T$ (Figure 2b).

The figures shows that, the digital filter can reduce noise to satisfactory values, but it introduces a considerable lag; this lag is significant also when a moderate noise reduction is required (see Figure 2a). Other types of filters, as an example state variable filters, give similar results. It is to notice that also a lag occurrence in the feedback loop limits the value of the control system loop gain.

A different solution, which introduces a smaller delay, consists in employing a state observer that provides the speed estimation, starting from the position measure; this observer can be of deterministic type [2] or of stochastic type, based on a Kalman filter [3–6]. The last solution allows a greater flexibility; so it was been preferred.

3 KALMAN FILTER ALGORITHM

The Kalman filter proposed in [7] is based on the following stochastic discrete-time model of the mechanical system:

$$\begin{aligned} \boldsymbol{x_n} &= \boldsymbol{A}\boldsymbol{x}_{n-1} + \boldsymbol{b}\breve{a}_n + \boldsymbol{\xi}_n \\ y_n &= \boldsymbol{c}^T \boldsymbol{x}_n + \eta_n \end{aligned} \tag{3}$$

where

$$\boldsymbol{x} = \begin{bmatrix} \theta \\ \omega \\ \varepsilon \end{bmatrix} \quad \boldsymbol{A} = \begin{bmatrix} 1 & T & \frac{T^2}{2} \\ 0 & 1 & T \\ 0 & 0 & 1 \end{bmatrix} \quad \boldsymbol{b} = \begin{bmatrix} \frac{T^2}{2} \\ T \\ 0 \end{bmatrix}$$

$$\boldsymbol{c}^T = \begin{bmatrix} 1 & 0 & 0 \end{bmatrix}$$

n is the index of the sampling instant,

T is the duration of the sampling interval,

$\breve{a}$ is the value of the acceleration expected in interval

$(n-1, n)$ and computed on the basis of the torque desired value and of the estimated values of mechanical parameters,

θ is the value of the shaft position,

ω is the value of the angular speed,

ε is the difference between the angular acceleration and his expected value $\breve{a}$,

y is the output variable i.e. the measurement of shaft position θ,

ξ is the zero-mean independent Gaussian white-noise vector of the process state, with a covariance matrix $\boldsymbol{Q}$,

η is the zero-mean independent Gaussian white-noise scalar of the process measurement, with a variance r.

The state variables of the previous dynamic model can be estimated by the Kalman filter algorithm composed by the following steps:

1) prediction of state:
$$\tilde{\boldsymbol{x}}_n = \boldsymbol{A}\hat{\boldsymbol{x}}_{n-1} + \boldsymbol{b}\breve{a}_n$$

2) estimation of error covariance matrix:
$$\tilde{\boldsymbol{P}}_n = \boldsymbol{A}\hat{\boldsymbol{P}}_{n-1}\boldsymbol{A}^T + \boldsymbol{Q}$$

3) computation of Kalman filter gain vector:
$$\boldsymbol{g}_n = \tilde{\boldsymbol{P}}_n\boldsymbol{c}(\boldsymbol{c}^T\tilde{\boldsymbol{P}}_n\boldsymbol{c} + r)^{-1}$$

4) update of error covariance matrix:
$$\hat{\boldsymbol{P}}_n = (\boldsymbol{I} - \boldsymbol{g}_n\boldsymbol{c}^T)\tilde{\boldsymbol{P}}_n$$

5) correction of predicted state:
$$\hat{\boldsymbol{x}}_n = \tilde{\boldsymbol{x}}_n + \boldsymbol{g}_n(\theta_n - \boldsymbol{c}^T\tilde{\boldsymbol{x}}_n),$$

being θ_n the shaft position measured at sampling instant n.

This algorithm starts from an initial value $\hat{\boldsymbol{x}}_0$ and a matrix $\hat{\boldsymbol{P}}_0$ equal to the covariance matrix of $\hat{\boldsymbol{x}}_0$.

Since the dynamic system given by (3) is linear and time-invariant, the only variable element in the prediction and correction sections is gain vector $\boldsymbol{g}_n$. Therefore, it is possible to replace vector's sequence $\boldsymbol{g}_n$ with its limit $\boldsymbol{g}$ when $n \to \infty$; in this way a Steady-State Linear Kalman Filter (SSLKF) is utilised, with the advantage of a reduction in the number of math operations.

Once determined gain vector $\boldsymbol{g}$, the operations to be effected at every sampling interval are only those related to prediction and correction:

$$\begin{aligned}\tilde{\boldsymbol{x}}_n &= \boldsymbol{A}\hat{\boldsymbol{x}}_{n-1} + \boldsymbol{b}\breve{a}_n \\ \hat{\boldsymbol{x}}_n &= \tilde{\boldsymbol{x}}_n + \boldsymbol{g}(\theta_n - \boldsymbol{c}^T\tilde{\boldsymbol{x}}_n).\end{aligned} \quad (4)$$

In the SSLKF algorithm, gain vector $\boldsymbol{g}$ depends on measurement noise variance r and on dynamic system covariance matrix $\boldsymbol{Q}$; nevertheless, if r can be easily defined, it is not the same for $\boldsymbol{Q}$. On the other hand, the dynamic characteristics of the speed control loop are strongly influenced by the properties of the SSLKF algorithm and so by the gain vector. To avoid this trouble, an approach based on the allocation of the filter transfer function poles can be adopted; as shown in [7], the poles of the discrete time transfer function of the filter are equal to the eigenvalues of the following dynamic matrix:

$$\hat{\boldsymbol{A}} = \boldsymbol{A}(\boldsymbol{I} - \boldsymbol{g}\boldsymbol{c}^T). \quad (5)$$

Defining the following vector:

$$\boldsymbol{l} = \boldsymbol{A}\boldsymbol{g} = \begin{bmatrix} l_1 \\ l_2 \\ l_3 \end{bmatrix}, \quad (6)$$

dynamic matrix $\hat{\boldsymbol{A}}$ can be expressed as:

$$\hat{\boldsymbol{A}} = \boldsymbol{A} - \boldsymbol{l}\boldsymbol{c}^T = \begin{bmatrix} 1-l_1 & T & \frac{T^2}{2} \\ -l_2 & 1 & T \\ -l_3 & 0 & 1 \end{bmatrix}.$$

Imposing the determinant of matrix $\lambda I - \hat{\boldsymbol{A}}$ equal to zero, the following characteristic equation is obtained:

$$\lambda^3 + \lambda^2 c_2 + \lambda c_1 + c_0 = 0,$$

in which

$$\begin{aligned} c_2 &= l_1 - 3 \\ c_1 &= 3 - 2l_1 + l_2 T + l_3\frac{T^2}{2} \\ c_0 &= -1 + l_1 - l_2 T + l_3\frac{T^2}{2}. \end{aligned} \quad (7)$$

Imposing that the characteristic equation must have a real eigenvalue $\lambda_1 = \rho_0$ and two complex conjugated eigenvalues $\lambda_{2,3} = \rho_1 \mathrm{Exp}(\pm j\varphi)$, the characteristic equation coefficients must be:

$$\begin{aligned} c_2 &= -2\rho_1\cos(\varphi) - \rho_0 \\ c_1 &= \rho_1[2\rho_0\cos(\varphi) + \rho_1] \\ c_0 &= -\rho_0\rho_1^2. \end{aligned} \quad (8)$$

The eigenvalues of $\hat{\boldsymbol{A}}$ can be imposed by discretization of a continuous filter characterized by a real negative pole of modulus p_0 and a couple of complex poles $p_{1,2} = -w(\cos(\phi) \pm j\sin(\phi))$.

The discretization gives:

$$\begin{aligned} \rho_0 &= \mathrm{Exp}(-p_0 T) \\ \rho_1 &= \mathrm{Exp}(-wT\cos(\phi)) \\ \varphi &= wT\sin(\phi). \end{aligned}$$

Different simulations have shown that it is convenient to select for p_0 and w values three, five times as high as the bandwidth (expressed in rad/s) foreseen for the drive and a value included between 40 and 60 degrees for ϕ.

Once chosen the poles of the discrete time transfer function of the filter, coefficients c_2, c_1 and c_0 are determined by (8) and so the elements of vector l can be obtained by (7) as:

$$\begin{aligned} l_1 &= c_2 + 3 \\ l_2 &= \frac{c_1 - c_0 - 4 + 3l_1}{2T} \\ l_3 &= \frac{c_1 + c_0 - 2 + l_1}{T^2}. \end{aligned} \tag{9}$$

Finally, from the definition (6) the elements of vector g can be computed as:

$$\begin{aligned} g_3 &= l_3 \\ g_2 &= l_2 - Tg_3 \\ g_1 &= l_1 - Tg_2 - g_3 \frac{T^2}{2}. \end{aligned} \tag{10}$$

4 CHOICE OF THE SCALING FACTORS

One of the main problems, concerning the implementation of a control procedure on a fixed point DSP controller, is the choice of the scaling factors employed to numerically represent the state variables and the multiplicative constants. This choice must meet two divergent requirements; in fact, at first care must be taken in order to avoid overflow when performing math operations and then it is necessary to minimize the truncation errors due to the discrete representation of the variables. Moreover, a suitable choice of the scaling factors can allow a reduction of the algorithm complexity and so of the computation time.

First of all, consider the prediction equation, which can be expressed as:

$$\begin{aligned} \tilde{\theta}_n &= \hat{\theta}_{n-1} + T\hat{\omega}_{n-1} + \frac{T^2}{2}(\hat{\varepsilon}_{n-1} + \breve{a}_n) \\ \tilde{\omega}_n &= \hat{\omega}_{n-1} + T(\hat{\varepsilon}_{n-1} + \breve{a}_n) \\ \tilde{\varepsilon}_n &= \hat{\varepsilon}_{n-1}. \end{aligned} \tag{11}$$

It is clear that its implementation on a fixed point DSP with words of 16 bits requires to specify suitable numeric representations of shaft position θ, angular speed ω and acceleration a (equal to $\varepsilon + \breve{a}$). In other words, it is necessary to select the following scaling factors:

$$S_\theta = \frac{\theta}{\theta^*}, \qquad S_\omega = \frac{\omega}{\omega^*}, \qquad S_a = \frac{a}{a^*}, \tag{12}$$

where θ^*, ω^* and a^* denote the numeric representations as 16-bit integer numbers of θ (radians), ω (rad/s) and a (rad/s^2) respectively.

Taking into account (11) and (12), the prediction equations, referred to the numeric representations of the variables become:

$$\begin{aligned} \tilde{\theta}^*_n &= \hat{\theta}^*_{n-1} + T\frac{S_\omega}{S_\theta}\hat{\omega}^*_{n-1} + \frac{T^2}{2}\frac{S_a}{S_\theta}(\hat{\varepsilon}^*_{n-1} + \breve{a}^*_n) = \\ &= \hat{\theta}^*_{n-1} + k_1\hat{\omega}^*_{n-1} + k_2(\hat{\varepsilon}^*_{n-1} + \breve{a}^*_n) \\ \tilde{\omega}^*_n &= \hat{\omega}^*_{n-1} + T\frac{S_a}{S_\omega}(\hat{\varepsilon}^*_{n-1} + \breve{a}^*_n) = \hat{\omega}^*_{n-1} + k_3(\hat{\varepsilon}^*_{n-1} + \breve{a}^*_n) \\ \tilde{\varepsilon}^*_n &= \hat{\varepsilon}^*_{n-1}. \end{aligned} \tag{13}$$

These equations show that a suitable choice of the scaling factors can allow obtaining a good solution of both the above mentioned implementation problems. In particular, it is clear that the implementation of the first two equations of (13) doesn't introduce errors if coefficients:

$$k_1 = T\frac{S_\omega}{S_\theta}, \qquad k_2 = \frac{T^2}{2}\frac{S_a}{S_\theta}, \qquad k_3 = T\frac{S_a}{S_\omega},$$

can be exactly represented by integer numbers; moreover, if they are equal to powers of 2, the prediction values of the angular position and the speed can be simply computed by addition and shift operations.

4.1 Choice of the shaft position scaling factor

As regards the shaft position, it is convenient to select a scaling factor S_θ equal to:

$$S_\theta = \frac{\theta}{\theta^*} = \frac{2\pi}{2^{16}}. \tag{14}$$

In fact, by this scaling factor, when angle θ changes from 0 to 2π its representation θ^* assumes values from 0 to 2^{16}. This representation results very useful because it automatically limits the value of θ inside the round angle. Besides, if an encoder is chosen with a number impulses equal to a power of 2, also the transfer from the impulse number counted by the QEPC and the integer representation of the shaft position requires only shift operations.

4.2 Choice of the angular speed scaling factor

Denoting as $\omega_{\min}$ the rotation speed corresponding to an unitary increment of the angle representation in a sampling period, i.e.:

$$\omega_{\min} = \frac{1}{T}S_\theta = \frac{1}{T}\frac{2\pi}{2^{16}},$$

it's opportune to represent this speed value with an integer number expressed by a power of 2:

$$\omega^*_{\min} = 2^{k_\omega}.$$

In fact, by this choice, speed scaling factor S_ω and coefficient k_1 become:

$$S_\omega = \frac{\omega_{\min}}{\omega^*_{\min}} = \frac{2\pi}{2^{k_\omega} 2^{16} T} = \frac{2\pi}{2^{(16+k_\omega)} T}$$
$$k_1 = 2^{-k_\omega}. \tag{15}$$

The exponential value k_ω must be chosen so that the maximum speed representation $\omega^*_{\max}$ doesn't exceed a value equal to 2^{15}, i.e.:

$$k_\omega = 14 - \text{int}\left(\log_2 \frac{\omega_{\max}}{\omega_{\min}}\right) \tag{16}$$

where int(α) denotes the integer part of α.

4.3 Choice of the angular acceleration scaling factor

A suitable angular acceleration scaling factor can be achieved from the following considerations.

First of all, on the basis of (15), a speed variation

$$\Delta\omega = \omega_{\min} = \frac{2\pi}{2^{(16+k_\omega)} T}$$

corresponds to an unitary variation of the speed integer representation i.e. $\Delta\omega^* = 1$.

Therefore, considering at every sampling time, the above speed variation $\Delta\omega$, the minimum acceleration value $a_{\min}$ results:

$$a_{\min} = \frac{2\pi}{2^{(16+k_\omega)} T^2}.$$

Imposing $a^*_{\min} = 2^{k_a}$, the acceleration scaling factor results:

$$S_a = \frac{a_{\min}}{a^*_{\min}} = \frac{2\pi}{2^{(k_\omega + k_a + 16)} T^2}. \tag{17}$$

Consequently, coefficients k_2 and k_3 become:

$$k_2 = 2^{-(1+k_a+k_\omega)} \qquad k_3 = 2^{-k_a}$$

that are powers of 2, as desired.

As for k_ω, the value of k_a must be chosen so that the maximum acceleration representation $a^*_{\max}$ doesn't exceed a value equal to 2^{15}, i.e.:

$$k_a = 14 - \text{int}\left(\log_2 \frac{a_{\max}}{a_{\min}}\right). \tag{18}$$

In order to guarantee a good convergence of the integration procedure of the observation system, it is convenient to use in prediction and correction equations a double precision representation (32 bit) of its state variables θ, ω and ε, whereas for prediction error $e^*_n = \theta^*_n - \tilde{\theta}^*_n$ a 16-bit representation is adequate.

As regards the correction procedure, the corrected values of the state variables result

$$\begin{aligned} \hat{\theta}^*_n &= \tilde{\theta}^*_n + g^*_1 e^*_n \\ \hat{\omega}^*_n &= \tilde{\omega}^*_n + g^*_2 e^*_n \\ \hat{\varepsilon}^*_n &= \tilde{\varepsilon}^*_n + g^*_3 e^*_n, \end{aligned} \tag{19}$$

where coefficients g^*_1, g^*_2 and g^*_3 correspond to coefficients g_1, g_2 and g_3 given by (10) and take into account the scaling factors.

In particular, following relationships are obtained:

$$g^*_1 = g_1 \qquad g^*_2 = \frac{S_\theta}{S_\omega} g_2 \qquad g^*_3 = \frac{S_\theta}{S_a} g_3. \tag{20}$$

Finally, the values of coefficients g^*_1, g^*_2 and g^*_3 must be further scaled to allow a representation by sufficiently significant integer numbers; consequently each of the three multiplication operations require also a suitable shift.

5 IMPLEMENTATION AND RESULTS

The described filter has been implemented on a TMS320F240 DSP controller, specifically designed for digital motor control, with a sampling interval $T = 150$ µs. An optical encoder with two channels, each with 2048 pulses per revolution, has been employed to obtain the speed measure.

The implementation has been experimented on a drive characterized by a bandwidth of about 250 rad/s, a maximum rotation speed equal to 6000 r.p.m. and an angular acceleration lesser than 50000 rad/s^2.

The specifications on the maximum values of speed and acceleration lead to choose $k_\omega = 5$ and $k_a = 6$ (see equations (16) and (18)). As consequence, coefficients k_1, k_2 and k_3 assume the following values:

$$k_1 = 2^{-5} \qquad k_2 = 2^{-12} \qquad k_3 = 2^{-6}$$

and the scaling factors become:

$$S_\omega = \frac{\omega_{\min}}{\omega^*_{\min}} = \frac{2\pi}{2^{(16+k_\omega)} T} = \frac{\pi}{2^{20}} \frac{10^6}{150}$$

$$S_a = \frac{a_{\min}}{a^*_{\min}} = \frac{2\pi}{2^{(16+k_\omega+k_a)} T^2} = \frac{\pi}{2^{26}} \frac{10^{12}}{150^2}.$$

The drive bandwidth suggests the employment of values for p_o and w of about a thousand. In particular, choosing $p_o = w = 1000$ and $\theta = 40$, from (5), (6) and (7) the following gain factors are achieved:

$$g_1 = 0.31601 \quad g_2 = 315.106 \quad g_3 = 124212$$

Using (18), the gains to be employed in the implementation result

$$g_1^* = 0.31601 \quad g_2^* = 1.5125 \quad g_3^* = 5.7237,$$

conveniently represented as

$$g_1^* = 20710 \cdot 2^{-16} \quad g_2^* = 24780 \cdot 2^{-14} \quad g_3^* = 23444 \cdot 2^{-12}.$$

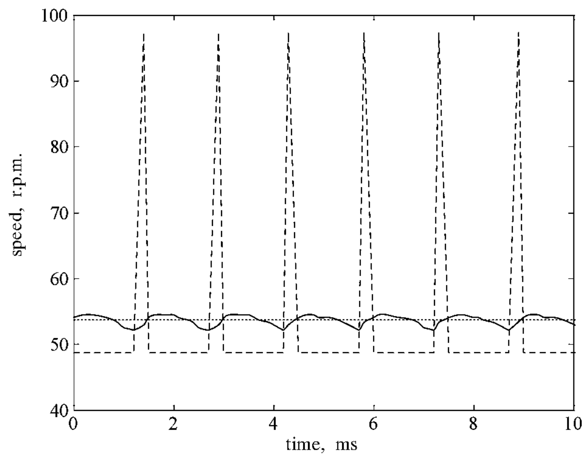


Fig. 3 Real, non-filtered and filtered speed shapes

Figure 3 shows the non-filtered (dashed line) and filtered (continuous line) shapes of the measured speed during steady-state operation when the speed is equal to 1.1 $\omega_{\min}$ (dotted line). The figure points out the drastic reduction of the measurement noise.

Figures 4 and 5 show two filtered speed shapes during the same starting transient of Figures 2a and 2b. In Figure 4 expected acceleration value $\breve{a}$ is assumed equal to zero, because no information is given. Comparing the shape of Figure 4 with those of Figures 2a and 2b, it is possible to assert that the proposed filter introduces a very small lag, except in presence of rapid acceleration variations. More accurate results are obtained, by supposing that an approximate acceleration estimation is available as shown in Figure 5, in which it is assumed that the acceleration expected value is equal to half of the real one. As it is possible to see, an approximate knowledge of the acceleration reduces the lag due to rapid acceleration variations.

The improvements available with the acceleration estimation can be underlined considering a sinusoidal speed shape having a frequency comparable with the drive bandwidth. As an example, Figure 6

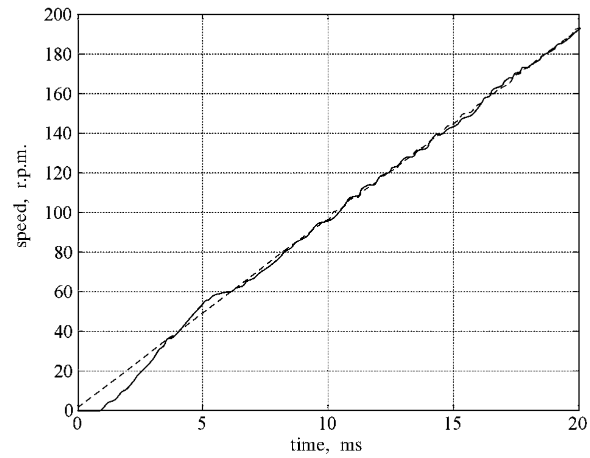


Fig. 4 Measure of the speed obtained by the proposed filter when $\breve{a} = 0$

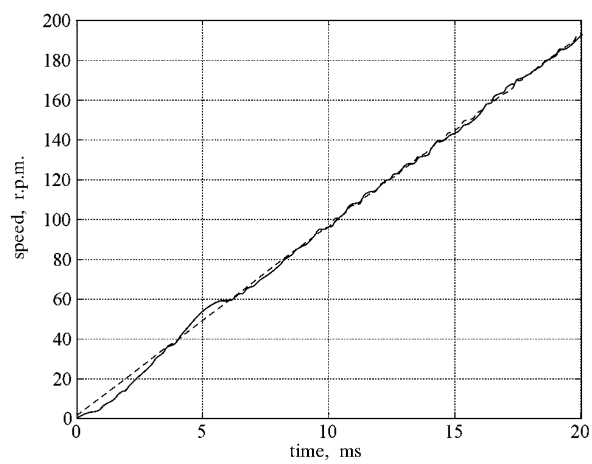


Fig. 5 Measure of the speed obtained by the proposed filter when $\breve{a} = 0.5a$

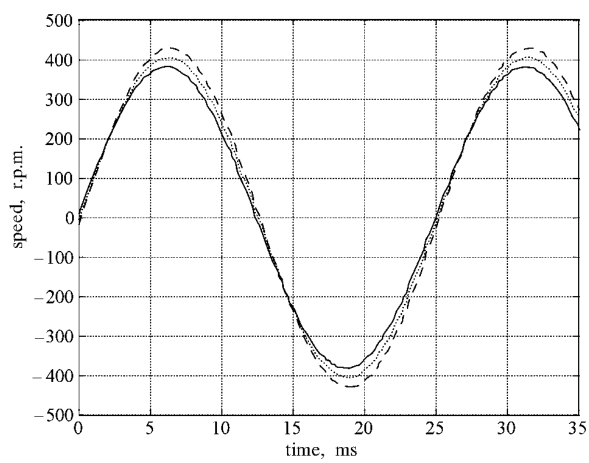


Fig. 6 Responses of the filter to a sinusoidal speed shape

shows the shapes of the speed (continuous line) and of its estimations, gotten with $\breve{a}=0$ (dashed line) and with $\breve{a}=0.5a$ (dotted line), when the speed varies with an angular frequency equal to $w/4$ i.e. equal to the drive bandwidth.

The proposed approach for the speed estimation allows obtaining also an acceleration estimation, even though rather noisy, which can be determined adding the estimated acceleration error $\breve{\xi}$ to expected value $\breve{a}$. Figure 7 shows the acceleration estimation gotten when $\breve{a}=0$ (dashed line) and when $\breve{a}=0.5a=500$ rad/s^2 (continuous line), during the starting transient already taken into account. As it is possible to see, the noise overlapped to the acceleration estimation is sufficiently limited; moreover, at the end of the initial transient, the two curves practically coincide.

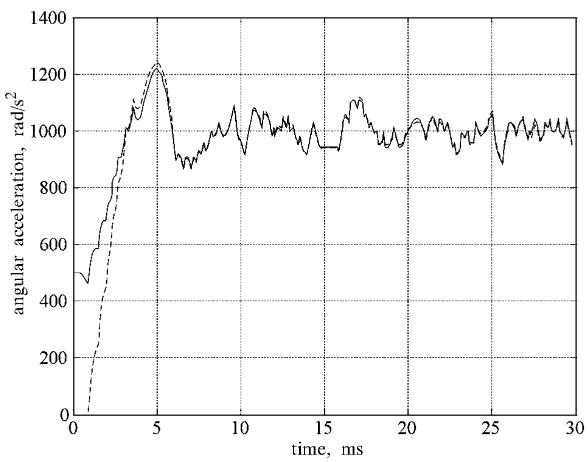


Fig. 7 Acceleration estimation

The knowledge of the acceleration can allow subsequent developments as different control techniques or identification of the drive mechanical parameters.

6 CONCLUSIONS

In motion control drives, the speed measurement is usually achieved employing an incremental encoder and deriving the speed value by the position variations during each sampling period. Because of the discretization connected to the computation of the pulses supplied by the encoder, the speed signal is very noisy and cannot directly be used for the drive control but it requires a suitable filtering.

The paper proposes to utilize, for the filtering, a SSLKF. To avoid the insertion on the speed estimation of a significant lag during the starting and stop transients, which generally occur at practically constant acceleration, the filter makes use of a third order model, whose state variables are the estimated values of position, speed and acceleration. Often, in motion control drives the acceleration value can be predicted on the basis of the desired torque value and of the mechanical parameters estimation; in this case, the acceleration predicted value can suitably be used as input of the Kalman filter to improve its behaviour, limiting the estimation to the acceleration error.

As regards the selection of the gain matrix, it was been underlined that only the measurement noise covariance can be easily defined while the dynamic system noise covariance cannot. On the other hand, the dynamic characteristics of the speed control loop are strongly influenced by SSLKF ones and, therefore, by the gain matrix. To overcome this problem, the paper has proposed an approach for the choice of the gain vector, different from the usual one and based on the poles allocation of the filter transfer function.

The paper has, also, presented the implementation problems of the proposed filter on a fixed point microprocessor. Particular care has been taken on the choice of the scaling factors employed to numerically represent the state variables and the multiplicative constants. In fact it is necessary both to avoid numerical overflows and to minimize the truncation errors due to the discrete representation of the variables. Moreover, a suitable choice of the scaling factors allows a significant reduction of the algorithm complexity and so of the computation time.

The filter has been implemented, at first, on a TMS320F240 DSP based prototype board; some tests have shown a perfect coincidence among the results furnished by the prototype and those gotten by simulation. Successively, the filter has been substituted to the original digital filter on an actual two axis control board allowing a reduction of the trajectory inaccuracy of about 10 times. Finally, the filter furnishes an estimation, even rather noisy, also for the acceleration; this last information can be used to achieve an identification of the drive mechanical parameters [8].

The proposed filter can be employed also in drives using an electromagnetic resolver; in this drives, it is nevertheless convenient to use a different approach inserting the Kalman filter in a Phase Looked Loop structure [9].

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Digitalni filtar brzine vrtnje za upravljane pogone s pozicijskim enkoderom niske rezolucije. Mjerenje brzine vrtnje pri upravljanju pogonima općenito se obavlja pomoću inkrementalnih enkodera s ograničenim brojem impulsa po okretaju. Budući da enkoderi mjere zakret vratila, potrebno je derivirati taj signal kako bi se dobio signal brzine vrtnje. Tako dobiveni signal brzine vrtnje sadrži šum, osobito izražen pri niskim brzinama vrtnje, kojega je potrebno smanjiti prikladnim filtrom. Nakon razmatranja problema mjerenja brzine vrtnje uporabom inkrementalnih enkodera, u radu je prikazan algoritam izvornog digitalnog filtra zasnovanog na stacionarnom Kalmanovom filtru. Prikazan je i učinkoviti postupak realizacije toga filtra u 16-bitovnom mikrokontroleru s aritmetikom fiksne točke. Predloženi filtar smanjuje šum mjernog signala do prihvatljive vrijednosti te se tako filtrirani signal može iskoristiti i za estimaciju akceleracije. Eksploatacijska valjanost filtra pokazana je primjenom u industrijskom pogonu na alatnom stroju.

Ključne riječi: enkoder, digitalni filtar, Kalmanov filtar, mjerenje brzine vrtnje, pogon, pretvornik

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