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Oversampling of Resolver Data

POWER ENGINEERING

In literature on control technology, the resolver is often mentioned as a suitable feedback sensor. In fact, due to its design, it is highly tolerant to vibration and high temperatures. These features bring the manufacturers of servo drives to use resolvers for many critical and low cost applications. The accuracy achieved in position measurement by these sensors depends mostly on the quality of the analog signals, although resolver position data are not very precise: angular accuracies are in range from 1 ... 5 arc minutes with traditional data conversion algorithms.

There are some methods of excluding of systematic error, for example, calibration of resolver by precision optical encoder. But except of systematic error, the outputs of resolver also contain noise due to additive disturbance signals.

Typically, by the arctangent conversion method, resolver is excited by sinusoidal voltage and output data are read in extremums of excitation signal, once per sampling period. If the disturbance signals contain frequency components near any of the demodulation signal harmonics, then these components are aliased down to DC and add directly to the demodulated signal. In environments where significant high frequency noise is present, it is possible to filtering the input signals. But the amount of filtering is limited by the inherent phase shift exhibited by the filter at the reference frequency, as this phase shift adds to any reference phase shift, which is a source of position error.

The article performs methods of resolver oversampling and synchronous detecting due to additional multiplication of sampled signals by the carrier signal. This method may be used both for sinusoidal and pulse excitation. In both cases, the resolver outputs are read some times per sampling period and then are processed and decimated.

The resolver output signals have constant amplitude U_m and excitation frequency ω_0. During measuring shaft rotation, the output voltages have the amplitude modulation with suppressed carrier (AM-SC) with sine and cosine function of angle position:

$$U_S = U_m \sin\omega_0 t \cdot \sin\alpha;$$
$$U_C = U_m \sin\omega_0 t \cdot \cos\alpha, \quad (1)$$

where: α − the resolver rotor shaft position.

The most known way of angle data conversion consists in the fact that data sampling:

$$U_S = U_m \sin\alpha;$$
$$U_C = U_m \cos\alpha \quad (2)$$

is at the moments when $\sin\omega_0 t = 1$.

The (2) procedure of resolver data processing one may call as pulse-synchronous detection (PSD), because the reference voltage has the same frequency ω_0, as the carrier frequency. By PSD, there is no disturbances filtration, because of signal sampling during the short interval of carrier signal, and the disturbance averaging within exciting voltage period is absent.

Further, the angle may be calculated either by decision of equation:

$$U_m(\sin\alpha\cos\theta - \cos\alpha\sin\theta) = U_m \sin(\alpha - \theta) = 0 \quad (3)$$

or from ratio

$$\alpha = \arctan 2(U_S, U_C). \quad (4)$$

Here θ is the angle of digital converter which compensates angle of measuring shaft and arctan 2 is denoted the algorithm of angle calculation through arctan in 4 quadrants (from 0 to 2π).

But sometimes it is advisable to use not the pulse-synchronous detection, but digital realization of synchronous detection (SD). It is known [1], that for SD of amplitude modulated signal $S_{AM}(t)$ with carrier frequency ω_0 and amplitude $A(t)$:

$$S_{AM}(t) = A(t)\sin\omega_0 t \quad (5)$$

it is necessary to multiply it with $\sin\omega_0 t$:

$$y(t) = S_{AM}(t)\sin\omega_0 t = A(t)\sin^2\omega_0 t = \frac{1}{2}A(t)(1 - \cos 2\omega_0 t). \quad (6)$$

As a result one has two items. The first one is amplitude, the second one is amplitude modulated by the frequency $2\omega_0$. As usual, the high-frequency signal may be deleted by filtration of the whole signal through the filter.

But also it is easy to find, that in the case when the modulating signal amplitude does not changes within period of carrier signal, then after integration of the processed signal $y(t)$ during excitation period, the only amplitude of modulation signal will be as result:

$$y(t) = \frac{2}{T}\int_0^T A(t)\sin^2\omega_0 t \cdot dt = A(t), \quad (7)$$

where $T = 1/f_0 = \omega_0/2\pi$.

In this case the additional filtration is not required.

One can make sure, that if phase shift between carrier signal and reference signal is present, there will be error in calculated by detection signal amplitude. For example, when reference signal initial phase is $\pi/8$, the demodulated signal value will be:

$$y(t) = \frac{2}{T}\int_0^T Ay(t)\, sin\,\omega_0 t \cdot sin(\omega_0 t + \pi/8) \cdot dt = 0{,}92388 \cdot Ay(t), \tag{8}$$

but the value of error from (8) is amplitude A independent, so this phase shift will bring only scaling of signal amplitude.

Using the digital procedure of integration (7), one may try to increase the noise-immunity of resolver data by increasing of the AD-converter sampling frequency. Indeed, for such detection method, it is necessary to have information of signal value within the carrier signal period (resolver exciting voltage). That is why the disturbances may be averaged during the digital integration. One may proof, that if there is the even number of intervals in carrier signal period, the integration does not lead to error appearance in modulating signal amplitude calculations.

The important property of synchronous detection is the possibility to exclude the velocity error caused by back EMF. When the output windings of resolver have no load, there output voltages are represent as:

$$\begin{aligned}
U_S &= L_E\, sin\,\alpha \frac{dI_E}{dt} + I_E L_E\, cos\,\alpha \frac{d\alpha}{dt} = \\
&= -I_{Em} L_E\, sin\,\alpha \cdot sin\,\omega_0 t + I_{Em} L_E\, cos\,\alpha \cdot cos\,\omega_0 t \cdot \frac{d\alpha}{dt} = \\
&= -U_m\, sin\,\alpha \cdot sin\,\omega_0 t + U_\Omega\, cos\,\alpha \cdot cos\,\omega_0 t; \\
U_C &= L_E\, cos\,\alpha \frac{dI_E}{dt} - I_E L_E\, sin\,\alpha \frac{d\alpha}{dt} = \\
&= -I_{Em} L_E\, cos\,\alpha \cdot sin\,\omega_0 t - I_{Em} L_E\, sin\,\alpha \cdot cos\,\omega_0 t \cdot \frac{d\alpha}{dt} = \\
&= -U_m\, cos\,\alpha \cdot sin\,\omega_0 t - U_\Omega\, sin\,\alpha \cdot cos\,\omega_0 t,
\end{aligned} \tag{10}$$

where: $I_E = I_{Em}\, cos\,\omega_0 t$ – excitation current; L_E – amplitude of mutual inductance between excitation winding and signal winding.

One may see that there are transformer EMFs (position dependent signals):

$$\begin{aligned}
U_{ST} &= -U_m\, sin\,\alpha \cdot sin\,\omega_0 t; \\
U_{CT} &= -U_m\, cos\,\alpha \cdot sin\,\omega_0 t,
\end{aligned} \tag{11}$$

and back EMFs (velocity dependent signals):

$$U_{S\Omega} = U_{\Omega} \cos\alpha \cdot \cos\omega_0 t;$$
$$U_{C\Omega} = -U_{\Omega} \sin\alpha \cdot \cos\omega_0 t. \tag{12}$$

As EMF of each signal winding are algebraically summing up, so it is enough to calculate the define integral for back EMF only, multiplied by carrier signal with constant amplitude, as the following example:

$$Y_{\Omega} = \frac{2}{T}\int_0^T U_{\Omega} \cos\alpha \cdot \cos\omega_0 t \cdot \sin\omega_0 t \cdot dt = 0. \tag{13}$$

Here is again supposed, that during carrier signal period there are no fluctuations in $U_{\Omega}\cos\alpha$.

One may easy make sure that phase shift between excitation and reference signals leads to velocity error at synchronous detector output. The whole suppression of the velocity error is possible only at the common-mode conditions of excitation and reference signals. It is obviously, that the same results will be obtained under digital integration using rectangular formula.

The similar results one may obtain by pulse excitation voltage and triangle excitation current with zero offset. From equation (10) it follows, that output voltages will have the form of bipolar rectangular pulses, modulated as function of rotor angular displacement. They will contain two components: transformer EMF and back EMF. In Fig. 1, there are the results of one output winding voltage (U_S) multiplying by reference voltage (U_{ref}) within one carrier signal period and the average value of demodulated voltage (U_{ave}).

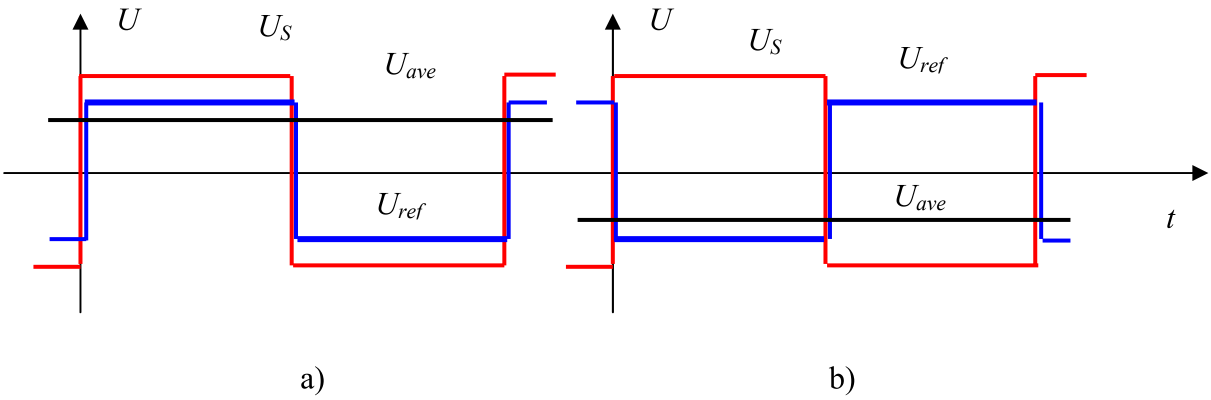

Fig. 1. Output signal voltage, multiplied by reference voltage

The profile in Fig. 3, a shows the situation when the phase of transformer EMF coincides with the reference voltage phase, i.e. demodulated voltage is positive. At the case, when the phases of reference voltage and signal windings transformer EMF are in opposite (Fig. 3, b), the demodulated voltage is negative.

The back-EMF has $\pi/2$ phase shift from reference voltage (Fig. 2). One may see, that within the first and third quadrants, the product of reference voltage and transformer EMF is positive, but within the second and the fourth quadrants this product is negative and it has the same volume. Hence, during reference voltage period, the rotating EMF average value is zero.

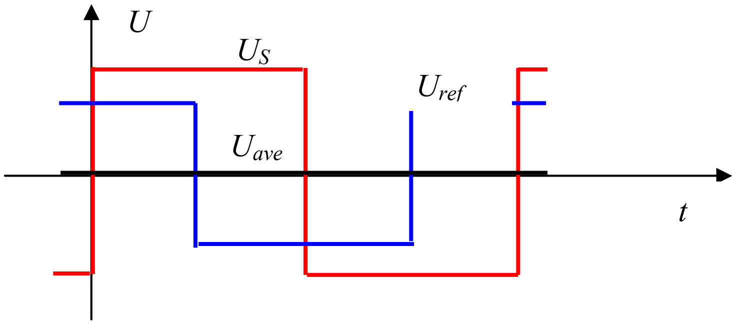

Fig. 2. Mutual shift of rotating EMF and reference voltage

So, when resolver is excited by triangle current, for output voltages synchronous detecting it is necessary to multiply them by reference rectangular voltage and to calculate the define integral of this product in carrier signal period. There is finite number of resolver voltages samples is excitation signal period, so it is necessary to hold same samples number within positive and negative reference voltage half-waves.

CONCLUSION

The theoretical research of digital synchronous detection of resolver output voltages shows that integration of demodulated signals allows essential weakening of resolver velocity error and increases the measuring system noise-immunity.

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[1] Sergienko A.B. The signal digital processing. – Saint-Petersburg.: Peter, 2002. – 608 p.
[2] Axenenko V.D. Conversion of the Resolver Output Information. – Proceedings of The First International Conference on Mechatronics and Robotics, Saint-Petersburg, 2000, vol. 1 p. 175.

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