UNBALANCE VIBRATION CONTROL OF MAGNETIC BEARING SYSTEMS USING ADAPTIVE ALGORITHM WITH DISTURBANCE FREQUENCY ESTIMATION

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ABSTRACT

This study is concerned with the adaptive forced rejection approach for a single periodic output disturbance with an unknown period and its application to unbalance vibration suppression in rotor-magnetic bearing systems. A discrete recursive technique is applied to estimate the unknown disturbance period and then to internally and adaptively generate a pseudofeedforward reference signal such as that this disturbance is eliminated. Simulations are carried out. The results indicate that both the estimation algorithm of an unknown period and reference signal adaptive generation algorithm are stable. Finally, the approach is applied to a problem of unbalance vibration suppression in rotor-magnetic bearing systems. Approximately 20dB and 7dB reductions in unbalance vibrations were experimentally obtained in the horizontal and vertical directions of the front and rear sides of the rotor, respectively. The experimental results show that the proposed algorithm is effective for achieving unbalance vibration suppression.

INTRODUCTION

Active magnetic bearings provide several advantages over conventional bearings for various practical industrial applications. These advantages include elimination of the lubrication system, friction free operation, decreased power consumption, operation at temperature, vacuum extremes and high rotation speed, and vibration control. However, active magnetic bearings are inherently unstable owing to the negative stiffness elements caused by the electromagnetic field. It is necessary to use feedback controller to stabilize the rotor-magnetic bearing systems. Recently there has been significant interest in digital control of magnetic bearing systems. The designs for such feedback controller have been developed using classic theory and state-space model approaches.

Vibration caused by mass unbalance are common problem in rotating machinery including the rotor-magnetic bearing systems. Perfect balancing is very costly and almost impossible if the unbalance distribution is changing during operation. Thus, residual unbalance always occur. If the unbalance vibration become large enough, the rotation will become impossible. It is also necessary to actively compensate for unbalance using active magnetic bearings, except feedback stabilizing. There are several reasons for unbalance compensation in AMB systems:

*Amplifier saturation can be avoided. When no unbalance compensation is used, the synchronous coil current caused by unbalance is approximately a quadratic function of rotation speed. Therefore, avoiding saturation becomes crucial for high-speed rotors.

*Housing vibrations and noise emission are reduced.

*Costs are reduced, because the required amplifier power is lower.

*Displacement orbits are reduced when the rotational speed crosses the rigid modes.

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Many researchers have investigated the control of unbalance response using magnetic bearings. One method to achieve unbalance response attenuation is through the addition of pseudo-state in observer-based controller, that is insertion of notch filter into the stabilizing control loop. Recent examples include leaning control method, disturbance cancellation control, and so on. The major drawback of this method is that the method requires an accurate plant model and the resulting controller order is usually quite large for the implementation on a single DSP. A completely different approach is the adaptive feedforward compensation of unbalance synchronous compensation signals are injected into the stabilizing control loop in order to cancel or minimize synchronous unbalance forces or displacements.

In the present paper, we propose a frequency estimation algorithm of a periodic signal with a single frequency and a novel Fourier coefficients adaptation algorithm which are use to generate the synchronous compensation signals. Then we combine these two algorithms and implement them on DSP and apply them to unbalance vibration suppression of rotor supported by magnetic bearing.

FREQUENCY ESTIMATION OF A SIGNAL WITH A SIGLE FREQUENCY

If ω is not precisely known, prior to applying adaptive corrective laws for $\alpha(k)$ and $\beta(k)$, the frequency estimation algorithm is necessary. Here we use the frequency estimation algorithm proposed by Tsao and Qian[4] to estimate the frequency of output disturbance.

Consider a periodic signal d(t) that is not identically zero with period τ^*. the frequency estimation algorithm is based on the gradient minimization of a quadratic function

$$J(\tau)=0.5\int_{t-T_{\max}}^{t}[d(s)-d(s-\tau)]^2 ds \tag{1}$$

where $T_{\max}>\tau^*$. For the following gradient adaptation algorithm:

$$\tau(t+1)=\tau(t)-h\frac{\partial J[\tau(t)]}{\partial\tau},\quad \tau(0)=\tau_0 \tag{2}$$

Tsao and Nemani have shown that there exists γ,σ ,h > 0 such that if $|\tau_0-n\tau^*|<\sigma$, then $\tau\rightarrow n\tau^*$ as $t\rightarrow\infty$ where n is an integer. The cost function J(τ) is periodic and has local minima at

$$\tau=n\tau^*,\quad n=\pm1,\pm2,\cdots\cdots \tag{3}$$

For the iteration Eq.(2) co converge to τ^* , the initial condition τ_0 must lie within the concave region containing τ^* .

Let τ be the known upper bound of the disturbance signal period and consider a time period $T_{\max}$ where $T_{\max}>\tau$. If ΔT is the sampling period, then let

$$T_{\max}=L\Delta T \text{ and } \tau=\eta\Delta T \tag{4}$$

Consider a known set of data $\{d_k\ ;\ 1\leq k\leq 2L+2\}$ and define $\lfloor\eta\rfloor$ as the integer part of the real number η. Then the two adjacent integers of η are $\lfloor\eta\rfloor$ and $\lfloor\eta\rfloor+1$. The partial derivatives of J in Eq.(2) at these integer points are

$$\frac{\partial J(\lfloor\eta\rfloor\Delta T)}{\partial\tau}\approx\sum_{k=L+1}^{2L}\left(d_k-d_{k-\lfloor\eta\rfloor}\right)\cdot\frac{d_{k-\lfloor\eta\rfloor+1}-d_{k-\lfloor\eta\rfloor-1}}{2} \tag{5}$$

and

$$\frac{\partial J[(\lfloor\eta\rfloor+1)\Delta T]}{\partial\tau}\approx\sum_{k=L+1}^{2L}\left(d_k-d_{k-\lfloor\eta\rfloor-1}\right)\cdot\frac{d_{k-\lfloor\eta\rfloor}-d_{k-\lfloor\eta\rfloor-2}}{2} \tag{6}$$

$$\frac{\partial J(\eta\Delta T)}{\partial\tau}\approx\frac{\partial J(\lfloor\eta\rfloor\Delta T)}{\partial\tau}(\lfloor\eta\rfloor+1-\eta)+\frac{\partial J[(\lfloor\eta\rfloor+1)\Delta T]}{\partial\tau}(\eta-\lfloor\eta\rfloor) \tag{7}$$

and Eq.(2) for the period as a function of η

$$\eta(t+1) = \eta(t) - \frac{h}{\Delta T}\frac{\partial J(\eta \Delta T)}{\partial \tau} \tag{8}$$

In real-time implementation, the iteration period need not coincide with the sampling period since the iteration operates in the background on a fixed set of sampled data.

ADAPTIVE REJECTION OF PERIODIC DISTURBANCE

Consider a active control system for periodic disturbance using adaptive algorithm as shown in Fig.1. Here plant is stable. Figure 2 shows the functional blocks of adaptive algorithm for generation compensation input. As shown in Fig.2, the adaptive algorithm generates a synchronous correction signal which is used to reduce or eliminate the response of the sinusoidal disturbance from the system output according to residual error.

At first, we can assume that a sinusoidal disturbance d is as follows:

$$d = \alpha_d \sin(\omega t) + \beta_d \cos(\omega t) \tag{9}$$

where α_d, β_d are the Fourier coefficients of the disturbance.
and the transfer function of the plant is given by G(s). Where, $G(j\omega)=Ae^{j\theta}$.

For the above disturbance, we can construct a synchronous correction signal of the following form

$$r = \alpha(t)\sin(\omega t) + \beta(t)\cos(\omega t) \tag{10}$$

here $\alpha(t)=\alpha(k)$, $\beta(t)=\beta(k)$, for $kT_s \leq t < (k+1)T_s$, k=0,1,2 are time varying Fourier coefficients computed on-line and updated at a sampling period of T_s. Then, the output signal of the plant and the residual error signal can be derived as Eq.(11) and Eq.(12).

$$y = A\left(\alpha(t)\sin(\omega t + \theta) + \beta(t)\cos(\omega t + \theta)\right) \tag{11}$$

$$e = A\left(\alpha(t)\sin(\omega t + \theta) + \beta(t)\cos(\omega t + \theta)\right) + \alpha_d \sin(\omega t) + \beta_d \cos(\omega t) \tag{12}$$

To obtain $n_1(t)$ and $n_2(t)$ we demodulate e(t), filter the intermediate results with low pass filters having an impulse response h(t) as shown in Fig.3. The purpose of the low pass filter is to filter out the non-D.C. components. For this reason we choose its cut-off frequency as $\omega_B << 2\omega$. Furthermore, we choose the sampling frequency as $\omega_s = 2\pi/T_s < \omega_B$ so that $n_1(t)$ and $n_2(t)$ will reach steady prior to each sample. Now, if high frequency components having sufficiently small gain are neglected, then the outputs of the low pass filters can be approximately written as follows;

$$n_1(t) = 0.5\left(A\alpha\cos(\theta) + \alpha_d - A\beta\sin(\theta)\right) \tag{13}$$

$$n_2(t) = 0.5\left(A\beta\cos(\theta) + \beta_d + A\alpha\sin(\theta)\right) \tag{14}$$

Next, we will describe the adaptive correction laws for the Fourier coefficients $\alpha(k)$,$\beta(k)$ of the synchronous correction signal r.

a) Case in which θ is known

In this case we can derive the following adaptive correction laws for the Fourier coefficients $\alpha(k)$,$\beta(k)$

$$\alpha(k+1)=\alpha(k)-(\cos(\theta)n_1(k)+\sin(\theta)n_2(k))/A \tag{15}$$

$$\beta(k+1) = \beta(k)-(\cos(\theta)n_2(k)+\sin(\theta)n_1(k))/A \tag{16}$$

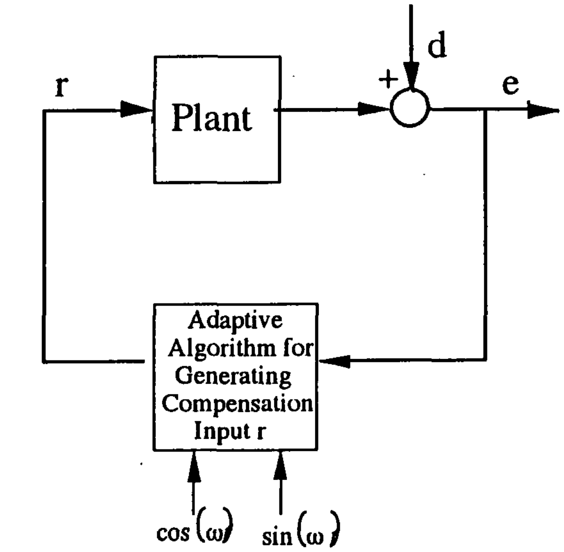

Fig.1 Block diagram of the control system rejecting periodic disturbance

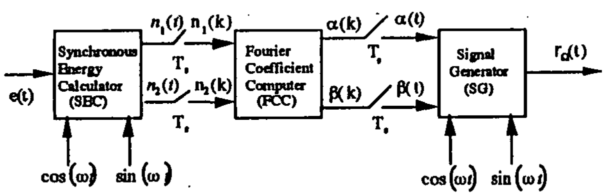

Fig.2 Block diagram of adaptive algorithm internally generating reference signal

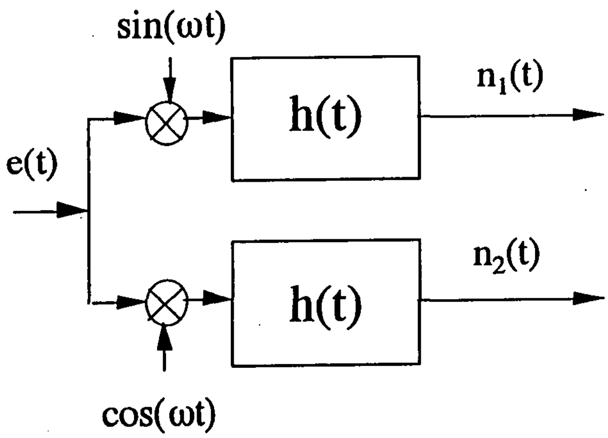

Fig.3 Details of SEC

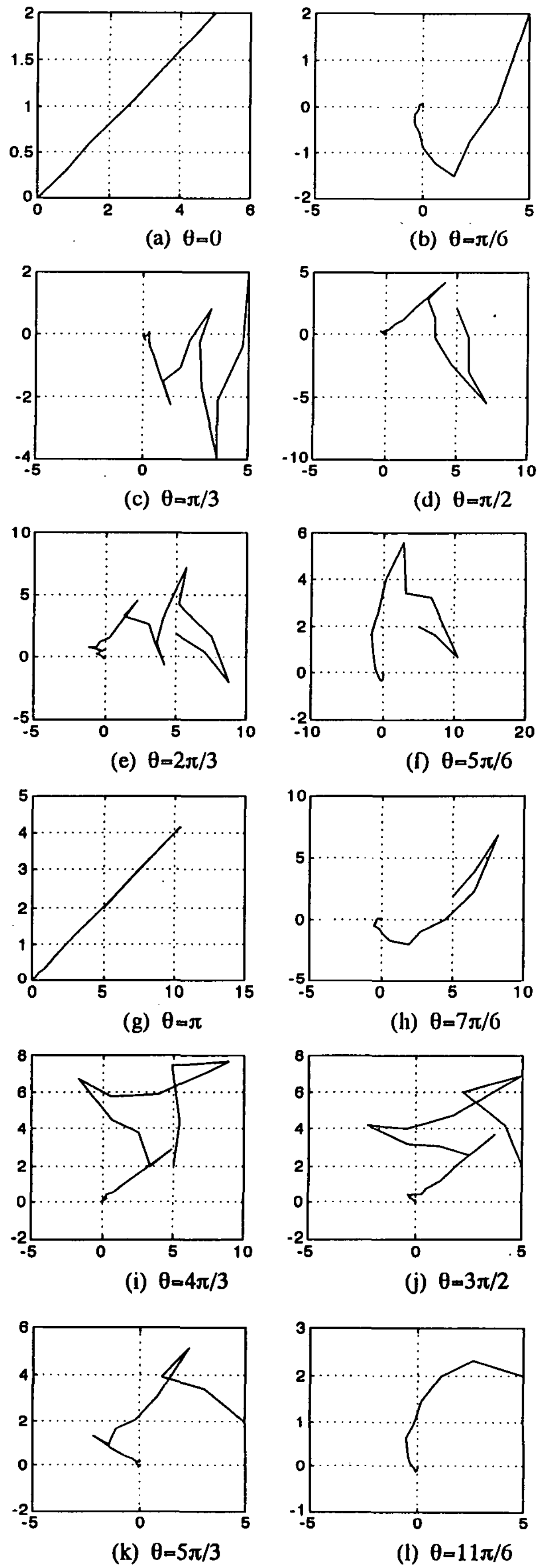

Fig.4 Phase trajectories of nonlinear discrete system Eqs.(21)~(24) with different θ

We substitute Eqs.(15) and (16) into Eqs.(13) and (14), and $n_1(t)$ and $n_2(t)$ at a sampling period of T_s are demoted by $n_1(k)$ and $n_2(k)$ respectively, then a discrete-time system of the forms

$$n_1(k+1) = 0.5n_1(k) \tag{17}$$

$$n_2(k+1) = 0.5n_2(k) \tag{18}$$

can be derived.

It is easily seen that the system is asymptotically stable, because the two eigenvalues of the system are both -0.5 within unit circle. However, there are many cases in which frequency response of a plant can not be obtained, or although the one of a plant can be obtained, the implementation of Eqs.(15) and (16) on-line is very difficult, because q is dependent of frequency. Furthermore, when there is uncertainty in θ , if adaptive corrective laws Eqs.(15) and (16), the system may become unstable. Therefore, it is necessary to derive adaptive corrective laws which are independent of θ.

b) Case in which θ is unknown

In this case, we propose following adaptive corrective laws $\alpha(k)$ and $\beta(k)$ which are independent of θ.

$$\alpha(k+1) = \alpha(k) - \mu_1(k+1)\, n_1(k) \tag{19}$$

$$\beta(k+1) = \beta(k) - \mu_2(k+1)\, n_2(k) \tag{20}$$

Where μ_i is step size. And

$$\mu_1(k+1) = \mu_1(k)\ \mathrm{sgn}\,(n_1^2(k-1) - n_1^2(k)) \tag{21}$$

$$\mu_2(k+1) = \mu_2(k)\ \mathrm{sgn}\,(n_2^2(k-1) - n_2^2(k)) \tag{22}$$

We substitute Eqs.(19) and (20) into Eqs.(13) and (14), and $n_1(t)$ and $n_2(t)$ at sampling period of T_s are denoted by n_1 (k) and n_2 (k) respectively, then a discrete-time system of the forms

$$\begin{aligned} n_1(k+1) = n_1(k) + 0.5A(\mu_2(k+1)n_2(k)\sin(\theta) \\ - \mu_1(k+1)\, n_1(k)\cos(\theta)) \end{aligned} \tag{23}$$

$$\begin{aligned} n_2(k+1) = n_2(k) - 0.5A(\mu_2(k+1)n_2(k)\cos(\theta) \\ + \mu_1(k+1)n_1(k)\sin(\theta)) \end{aligned} \tag{24}$$

can be derived.

The system given by Eqs.(23),(24) and Eqs.(21), (22) is nonlinear discrete-time system. Here we verify the asymptotic stability of the system through a great number of computer simulations. Phase trajectories of the nonlinear discrete-time system (21)~(24) are computed for various different θs.

Figure 4 shows phase trajectories of the nonlinear discrete-time system Eqs.(21)~(24) for several representative θ s. In above simulation, we choose $\mu_i(0)=0.9 / A$, i = 1,2. From Fig.4, it is clear that the nonlinear discrete-time system Eqs.(21)~(24) is stable. Furthermore we also compute the phase trajectories of the nonlinear discrete-time system Eqs.(21)~(24) for various different θs and several $\mu_i(0)$s satisfying $0< \mu_i(0) < 1 / A$, i=1,2 . As the result, we can conclude that the nonlinear discrete-time system Eqs.(21)~(24) is stable as if $\mu_i(0)$ satisfies $0 < \mu_i(0) < 1 / A$, i = 1,2.

From Eqs.(12)~(14), we can note that if $n_1(t)$ and $n_2(t)$ both converge 0 , then indeed the residual error signal e(t) also converges 0. Moreover, although besides sinusoid component at frequency ω, disturbance signal d contains sinusoid components at other frequencies, as if the difference between the frequency ω and the other frequency is larger than ω_B, Eqs.(13) and (14) still hold and the residual error signal converges the sinusoid components at other frequencies. Hence, we can conclude that the sinusoid components at other frequencies do not almost give effect on dynamic of plant.

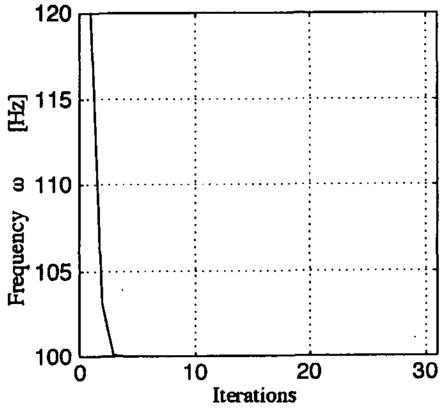

Fig.5 Estimation of frequency ω

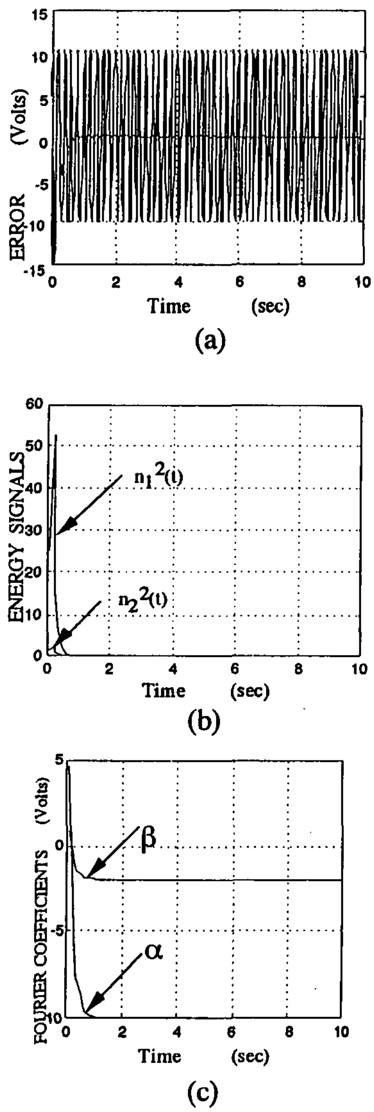

Fig. 6 Transient response of the error signal and Fourier coeffciients and energy signals by simulation when $A = 1, \theta = 0$

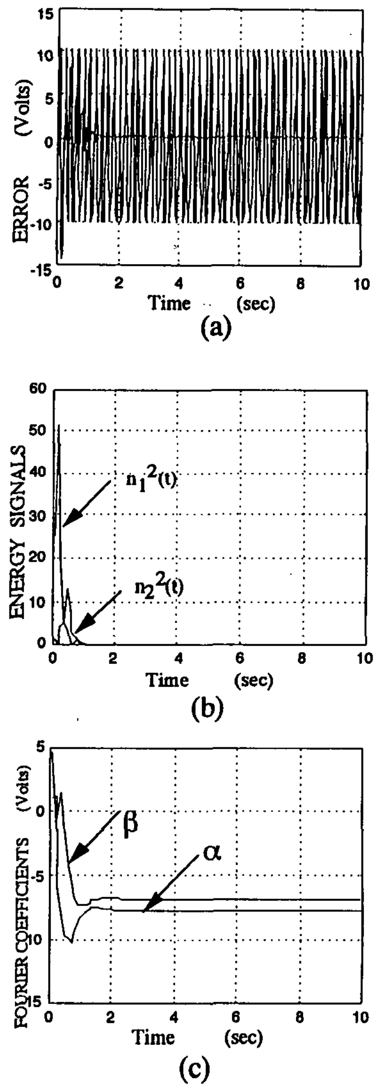

Fig. 7 Transient response of the error signal and Fourier coefficients and energy signals by simulation when $A = 1, \theta = {}^{11\pi}/_{6}$

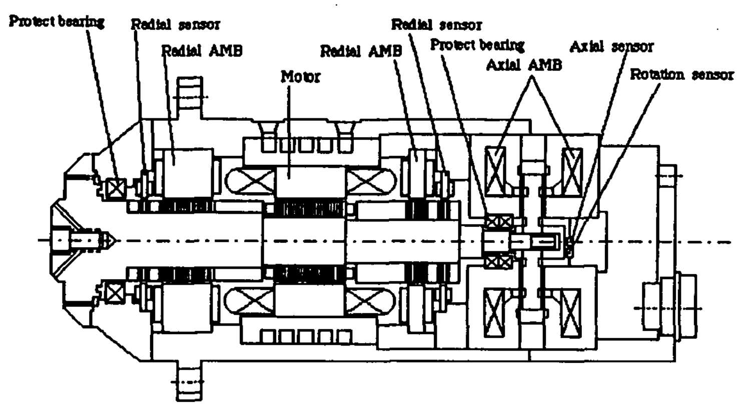

Fig.8 Test apparatus of AMB system

SIMULATIONS

At first, for a signal with frequency 100Hz, taking initial condition ω_0 =120Hz, we conducted simulation using the frequency estimation algorithm presented in section 2. Figure 5 shows the simulation result. It is seen that the frequency estimation algorithm is exact and fast. Next, we performed a great number of simulations of disturbance rejection using adaptive corrective laws for $\alpha(k)$ and $\beta(k)$ presented in section 3 for various θ. For restriction of paper, we only give the simulation results for A = 1, θ = 0 and A = 1, $\theta = 11\pi/6$ in Fig.6 and Fig.7. From Fig.6 and Fig.7, it is clear that Fourier coefficients $\alpha(k)$, $\beta(k)$ converge to constants and the residual error signals converge to zero within 1 second.

APPLICATION TO UNBALANCE VIBRATION SUPPRESSION OF A ROTOR SUPPORTED BY MAGNETIC BEARING

Figure 8 shows test apparatus of AMB system under consideration. An induction motor rotor is located in the middle of the shaft and two radial magnetic bearings are located on each side of the motor rotor. A thrust magnetic bearing is located at the left end of the shaft. The radial magnetic bearings together control two rotational and two translational degrees of freedom. The thrust magnetic bearing controls the displacement in the axial direction. Eddy-current-type proximity sensors are set up for the radial magnetic bearings at both sides of the bearings, but inner sensors are only used in tests.

The block diagram of the control system implemented digitally is shown in Fig.9. A digital signal processor TMS320C30 with high-speed A/D converter and D/A converter is used to accomplish control algorithm and access data from or to host personal computer.

In this experimental setup, at first, a linear multi-variable analog PID controller is applied to stabilize and decouple the rotor-magnetic bearings system. The four magnetic bearings in x direction and y direction can be considered as decoupled and as four independent SISO systems.

Then we directly applied the combination of the proposed frequency estimation algorithm and the proposed adaptive algorithm for periodic disturbance rejection to suppress the unbalance vibration of the rotor. Figure 10 shows the comparison of the trajectories of the geometric center positions between before applying proposed control

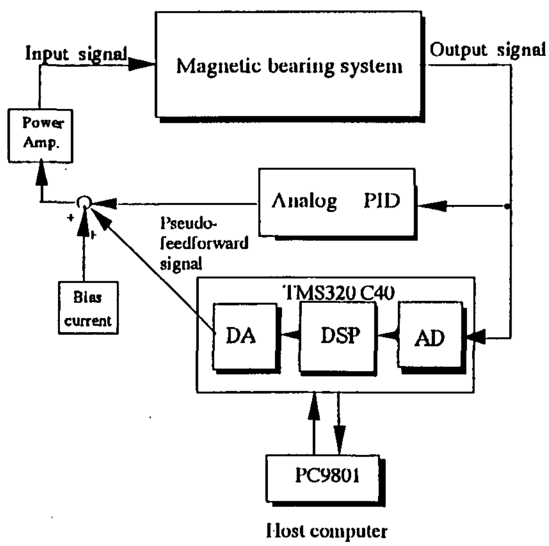

Fig.9 Block diagram for control system

and after applying proposed control when rotating at 6204 rpm. Figure 11 shows the stationary response of the geometric center positions when rotating at 6204 rpm. Figure 12 shows the spectra of stationary response of the geometric center positions without applying the proposed control when rotating at 6204 rpm. Figure 13 shows the spectra of stationary response of the geometric center positions after applying the proposed algorithm when rotating at 6204 rpm. The amplitudes of unbalance vibration of front end and rear end of the shaft in x-direction and y-direction are approximately 0.3vp-p, 0.16vp-p, 0.3vp-p and 0.2 vp-p respectively, before applying proposed

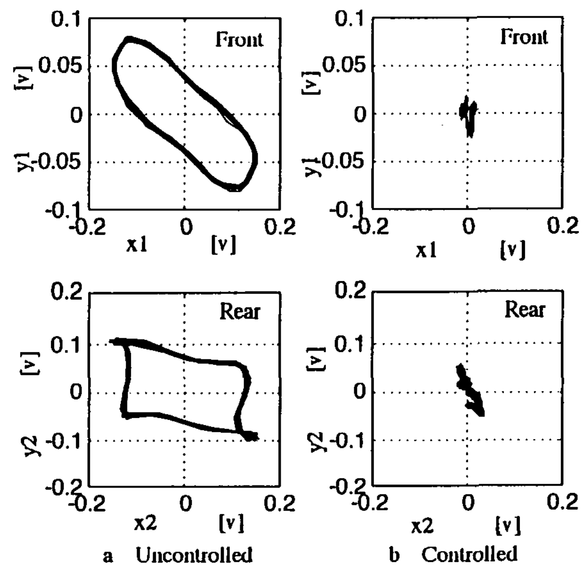

Fig.10 The trajectories of the geometric center positions when rotating at 6204 rpm

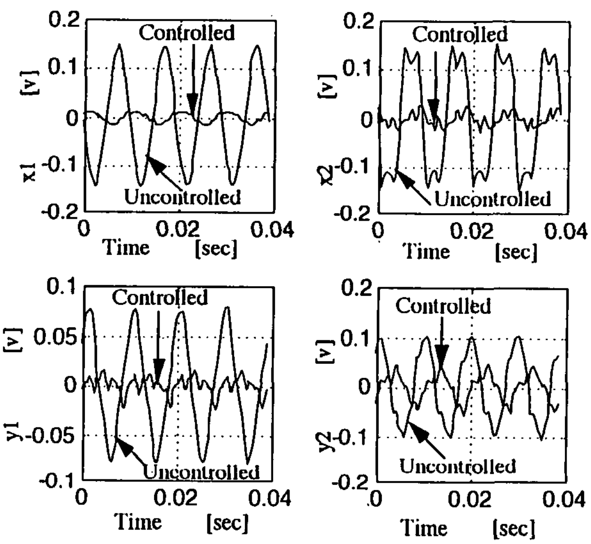

Fig.11 The stationary response of the geometric center positions when rotating at 6204 rpm

adaptive algorithm for periodic disturbance rejection. But after applying proposed adaptive algorithm for periodic disturbance rejection, they are reduced to about 0.035vp-p, 0.042vp-p, 0.052vp-p and 0.1vp-p. From Fig.12 and Fig.13, it is seen that the vibrations of the shaft at the fundamental frequency corresponding with rotating speed 6204 rpm are reduced by 22dB, 20dB and 8dB respectively. This means that the combination of the proposed frequency estimation algorithm and the proposed adaptive algorithm with gain scheduling for periodic disturbance rejection can be effectively used to suppress the unbalance vibration of the rotor.

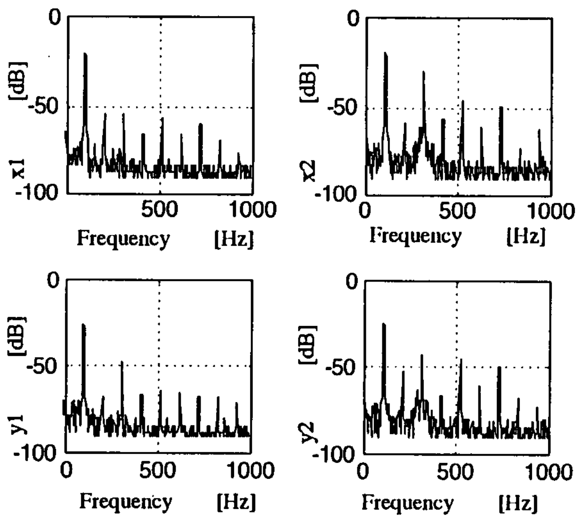

Fig.12 The spectra of stationary response of the geometric center positions without applying the proposed algorithm when rotating at 6204 rpm

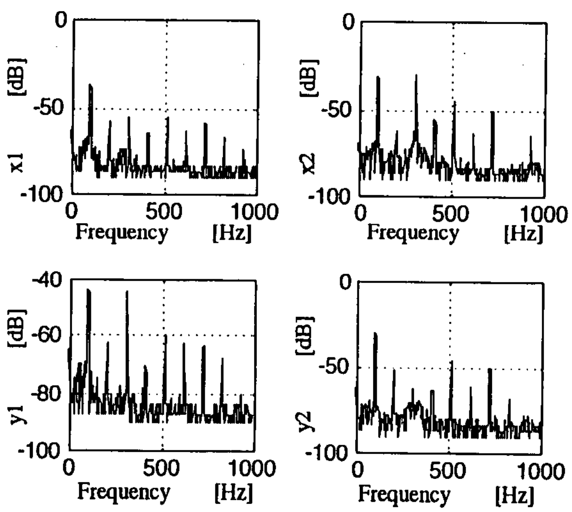

Fig.13 The spectra of stationary response of the geometric center positions with applying the proposed algorithm when rotating at 6204 rpm

CONCLUSIONS

The simulation and experimental results clearly demonstrate the high degree of unbalance response attenuation that can be achieved with the combination of the proposed frequency estimation algorithm and the proposed adaptive algorithm with gain scheduling for periodic disturbance rejection. However, the proposed frequency estimation algorithm will not perform well, if the sensor noise is large. If the sensor noise can be reduced by sensor itself and/or filter algorithm, the degree of unbalance response attenuation can be further improved.

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