SELF-SYNCHRONOUS DETECTION METHOD FOR MAGNETIC SUSPENSION DIGITAL CONTROL

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ABSTRACT

This paper deals with a detection method for picking up the amplitude component from a differential signal between two inductive type sensors in magnetic suspension control. The amplitude detection, which corresponds to measuring the displacement of the suspended body, is automatically completed in the software of a digital controller because the differential signal is synchronized with the sampling cycle and can be directly fed to the A/D converter. Such a digital synchronous detection scheme is embodied for a magnetic suspension device in which an electromagnet itself is used as an inductive type sensor. The inductive sensing is performed with a carrier current superimposed on the original exciting current of electromagnet, but the carrier frequency becomes much lower than that chosen for usual inductive type sensors. Thus a deterioration of the frequency response in the displacement measurement is inevitable. Besides, since the original exciting current itself varies the inductance of the electromagnet somewhat, a lowering of the measurement accuracy is inevitable as well. The present approach, which is named the self-synchronous detection method, holds the response delay to a minimum and reduces the measurement error remarkably. As the result, stable control with a crossover frequency of 25 Hz is achieved with a carrier frequency of only 2 kHz.

INTRODUCTION

Magnetic suspension devices often require position sensors which are low cost and compact in size. For this reason, the inductive type sensor with laminated steel core is employed widely. This sensor is usually used as a pair of gap sensors to secure a linearity between the output of the sensor and the displacement of the suspended body, e.g., as two sensors inserted into the differential sensing system of Fig. 1. In this sensing system, the oscillator feeds a carrier signal v_0 to both the sensors and the resistors. When the suspended body shifts from the equilibrium position, there is a difference in inductance between the two sensors and consequently the differential signal v appears according to the unbalance quantity. Therefore, if only the amplitude component of v is picked up, the displacement will be easily estimated. But, in the amplitude detection, the carrier component of v has to be sufficiently eliminated, especially when utilizing a digital controller, because it may make the magnetic suspension control unstable. The simplest approach is to use a low pass filter, together with a multiplier for multiplying the differential signal by the carrier signal. In this case, however, the low pass filter may bring a large time lag into the control system. The most advanced approach is to use a sample-hold circuit as a synchronous detector [1]. In this case, there is no need to eliminate the carrier component in the amplitude detection, but the detector becomes considerably complicated in composition and the unit cost of the sensor will increase.

The aforementioned problem may be settled by a synchronous detector which can be integrated well into a digital controller. In the sensing system, the differential signal is synchronized with the sampling cycle and can be directly fed to the A/D converter. The sign of the sampled data is reversed by each sampling cycle in the digital signal processor. This operation is equivalent to an amplitude detection for the differential signal. Therefore the hardware concerning amplitude detection is remarkably simplified. Though the fundamental idea of such a digital synchronous detection has been already proposed by the authors [1], the method is not yet established and the demonstration is left as a future subject. In this paper, a self-synchronous detection scheme is embodied for a magnetic suspension device in which an electromagnet itself is used as an inductive type sensor. The aim is to prove the strong points of the self-synchronous detection method. Though the conception of the sensorless control is based on the test signal injection method [2], the system construction is reconsidered from a different viewpoint because the materialization was not as easy as the authors thought originally.

SELF-SYNCHRONOUS DETECTION

Figure 2 shows a magnetic suspension device used in the present experiment. The arm is rotatable in a vertical plane while being supported at Point P. Electromagnet 3 is excited by a constant current so as to balance the attractive force with the gravitational force mg. On the other hand, Electromagnets 1 and 2 are excited by variable currents in order to keep the arm horizontal. Since they also are used as a pair of inductive type sensors, each exciting current must include a carrier component for sensing the arm displacement x. Since Electromagnets 1 or 2 has a large inductance, the carrier frequency becomes much lower than that chosen for usual inductive sensors. A lowering of the carrier frequency means a deterioration of the frequency response of the inductive sensing system. Therefore it is necessary to hold the deterioration to a minimum by some detection approach without low pass filter. In this point, it can be said that the device of Fig. 2 is a target suitable for clarifying the strong points of the self-synchronous detection method. Figure 3 shows a sensorless control system constructed for this device. The action is as follows.

The digital signal processor (DSP) produces a series of data in which the magnitude is constant and the sign alternates positive and negative. These data are outputted as a rectangular signal v_s through the left D/A converter (D/A 1). The signal v_s is applied to the oscillator through the band pass filter (BPF 1) with a quality factor of 10. As the result the oscillator generates a sinusoidal signal under a synchronization with the sampling cycle of DSP. The sinusoidal signal is converted to a carrier signal v_0 by the all pass filter (APF) with a phase lag ϕ_0. The carrier signal, together with the synchronizing rectangular signal, is shown in Fig. 4. The sampling frequency f_s of DSP is set to 4 kHz and the resultant carrier frequency f_0 becomes 2 kHz. The oscillator has an amplitude modulation function for examining the frequency response of this sensing system. But, if the purpose is only to produce a synchronized carrier signal, the oscillator is excluded because the output signal of BPF 1 can be regarded as a sinusoidal wave which is the most suitable for this sensing scheme. The carrier signal, together with a variable voltage command e, is supplied to the left power amplifier (PA 1) as a positive input signal and to the right power amplifier (PA 2) as a negative one. Since a dc voltage command e_0 is applied to PA 1 and PA 2 in common, the exciting currents of Electromagnets 1 and 2 can be approximately expressed as $i_1 = i_0 + i$ and $i_2 = i_0 - i$, respectively, if there is no carrier signal. Here i_0 is the bias current due to e_0 and i is the controlling current due to e. On the other hand, the contribution of v_0 appears as a carrier component superimposed on the exciting currents. The carrier component can be written as $s_1 = s_0 + s$ or $s_2 = - s_0 + s$, where s_0 is a carrier current at the equilibrium position and s is the unbalance quantity due to displacement.

The total currents of Electromagnets 1 and 2 are measured by means of the resistors and the buffer amplifiers (BA 1 & BA 2). The sum of the two total currents becomes 2 $(i_0 + s)$. Since the bias current i_0 in the addition signal is cut by the band pass filter (BPF 2) with a quality factor of 1, the output signal v of the instrument amplifier (IA) is due only to the unbalance quantity s. The signal v corresponds to the differential signal in this sensing system and can be directly fed to the left A/D converter (A/D 1). Therefore it is possible to sample the peak of v by adjusting the phase of v adequately with APF, as shown in Fig. 5. The peak detection is completed by reversing the sign of the sampled data by each sampling cycle in the DSP. Figure 6 shows a differential signal modulated with OSC and the peak detection result outputted by another D/A converter (D/A 3). On the other hand, the difference between the two total currents becomes 2 $(i + s_0)$. This subtraction signal can be directly fed to the right A/D converter (A/D 2) as well even if it includes the carrier current s_0. The data about the controlling current i is obtained by adding the sampled value at a sampling cycle to that at the previous cycle because the contribution of s_0 is canceled by doing so. Figure 7 shows a subtraction signal produced with a sinusoidal voltage command e and the addition result outputted by another D/A converter (D/A 4). The data about the controlling current will become necessary when introducing a state feedback scheme into the control system.

SENSING / CONTROLLING PERFORMANCES

A relation between the peak detection output and the arm displacement can be examined by changing the position successively while supporting the arm by micrometers. The results are summarized in Fig. 8. Here the bias current i_0 is set to the value indicated in this figure and the controlling current i is kept zero. A series of data for each case of i_0 is good in linearity, but the slopes of the fitting lines are different and the resultant sensitivity decreases from 8 V/mm to 5 V/mm as increasing i_0. This seems to be connected to a variation of inductance due to the exciting current. In general, the inductance of an electromagnet somewhat depends on the magnetic flux density in electromagnet core or in the suspended body and the dependence is stronger as the magnetic flux density approaches the saturation level. In this device, the magnetic flux density in the arm is closer to the saturation level and is roughly estimated to be 0.45 T at $x = 0$ mm and 0.6 T at $x = \pm$ 0.3 mm in the case $i_0 = 0.6$ A. Since the magnetic suspension control requires a bias current of 0.5 A at least, the exciting current will have a little influence on the inductance. The influence appears as an error in the displacement measurement. Figure 9 shows a relation between the peak detection output and the controlling current. Here the bias current is set to 0.5 A and the arm is fixed at the equilibrium position. From the negative slope, the error sensitivity is estimated to be 6 V/A. Since the detection sensitivity in the case $i_0 = 0.5$ A is 7 V/mm, the sensitivity ratio is close to 1 mm/A and cannot possibly be neglected. The error can be compensated with data about the controlling current, but it should be noted that the error sensitivity somewhat decreases as increasing the frequency of controlling current. So, in order to keep the compensation good over a wide frequency range, the current data was reformed according to an algorithm of a first-order low pass filter. Consequently it was possible to make the sensitivity ratio less than 0.01 mm/A in a frequency range up to 1 kHz.

The other performance in this sensing system is the frequency response to a variable displacement of the peak detection output, but it is difficult to measure the frequency response directly. So, by modulating the differential signal with OSC as shown in Fig. 6, let us indirectly estimate the frequency response. Here the modulating signal is fed to another A/D converter (A/D 3) and is outputted straight by another D/A converter (D/A 5) so as to exclude a time lag due to DSP itself from the frequency response. Figure 10 shows the transmission characteristic from the modulating signal output to the peak detection output. The gain cutoff frequency is about 1 kHz and the phase lag is about 5° at 100 Hz and increases to about 50° at 1 kHz. In this sensing system, since no low pass filter is employed, the response delay is due to the DSP and BPF 2 essentially. It is known that the effect of a band pass filter on the amplitude component of an amplitude-modulated wave is equivalent to a first-order lag element with a time constant $Q / \pi f_0$ [1]. In the case of BPF 2, the cutoff frequency of the first-order lag element becomes 1 kHz. The characteristic of Fig. 10 nearly accords with this theoretical expectation. Therefore, if the quality factor Q is reduced, the sensing performance seems to be improved further. It should be noted, however, that the differential signal is still not released from the controlling current. Strictly speaking, when there is no carrier signal, the total exciting currents should be expressed as $i_1 = i_0 + i + \Delta i$ and $i_2 = i_0 - i + \Delta i$, where Δi is the unbalance quantity of controlling current due to displacement. Since the addition signal in Fig. 3 is corrected as $2 (i_0 + \Delta i + s)$, the unbalance quantity Δi may bring a new error into the displacement measurement. Considering this error and a deficiency of the aforementioned compensation, a regular limitation of frequency bandwidth due to BPF 2 seems to be necessary for the control system to work stably.

The feedback control in Fig. 3 is completed by producing the feedback signal e_f from the data about displacement according to an adequate control algorithm. In the present experiment, let us adopt the following algorithm in consideration for a voltage control of the electromagnet and a vibration mode of arm; a proportional and integral regulator, first-order high pass filter, band pass filter and band elimination filter. The frequency response of the digital controller is shown in Fig. 11. The phase margin is secured between 3 Hz and about 200 Hz for exciting the electromagnet and the sharp gain drop is set to about 400 Hz for preventing a vibration of the arm. By using this digital controller, stable control was achieved successfully. Figure 12 shows the frequency response to a sinusoidal voltage command e of the feedback signal output e_f, that is, the open loop frequency response of this control system. This response was measured with the reference signal e_r in Fig. 3. From this response, it is found that the phase margin is about 15° at a crossover frequency of 25 Hz and the gain margin is about 12 dB in the neighborhood of 100 Hz.

CONCLUSION

The self-synchronous detection method proposed by the authors has been applied to a magnetic suspension device without position sensor. In the sensing system, a carrier current superimposed on the original exciting current of electromagnet is used for measuring the displacement of the suspended body. The carrier current has to be made of a carrier signal with a relatively low frequency. Thus a deterioration of the frequency response in the displacement measurement is inevitable. Besides, since the original exciting current itself varies the inductance of the electromagnet somewhat, a lowering of the measurement accuracy is inevitable as well. The present approach holds the response delay to a minimum by excluding low pass filters from the sensing system and reduces the measurement error remarkably by compensating the measurement result with the data about exciting current. Consequently, stable control with a crossover frequency of 25 Hz was achieved under a carrier frequency of only 2 kHz.

Though the self-synchronous detection method is premised on the use of digital controller, the application enables us to optimize the flexible function of digital controller simultaneously. Furthermore, since the sensorless control system constructed in this study is very simple in composition, it is possible to mount all the electronic circuits except the power amplifiers on the DSP board. This will be useful to improve the cost-performance ratio of digital controller further.

REFERENCES

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[2] K. K. Sivadasan, "Analysis of self-sensing active magnetic bearings working on inductance measurement principle," *IEEE Trans. Magn.*, vol. 32, pp. 329-334, 1996.

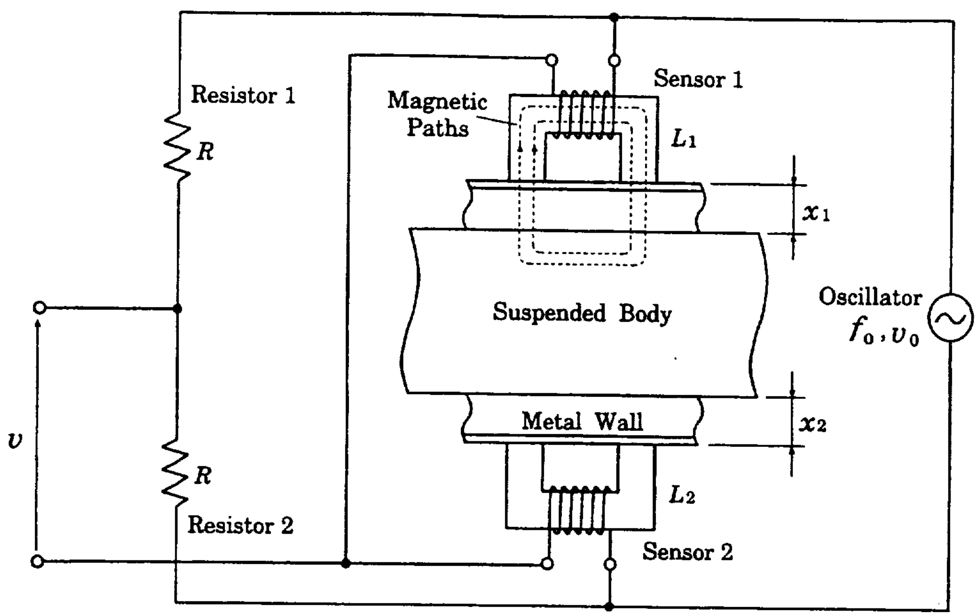

Fig. 1. Basic structure of inductive type sensors discussed in Ref. [1]. Though the metal wall is used for protecting the sensor, it is not the essential component for the inductive sensing system.

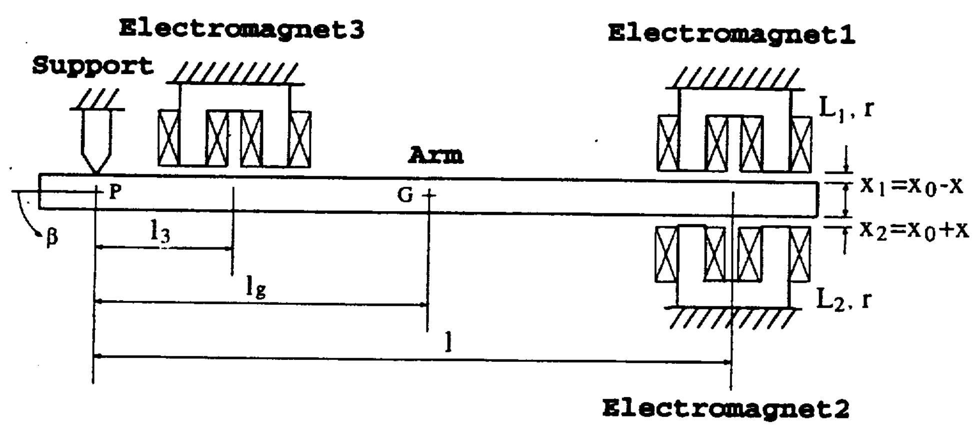

Fig. 2. A magnetic suspension device with one degree of freedom: l =193 mm, l_3 = 29 mm, l_g = 111 mm, x_0 = 1 mm, - 0.3 mm $\leq x \leq$ 0.3 mm, m = 1.3 kg, L_1, L_2 ~ 100 mH, r = 10 Ω.

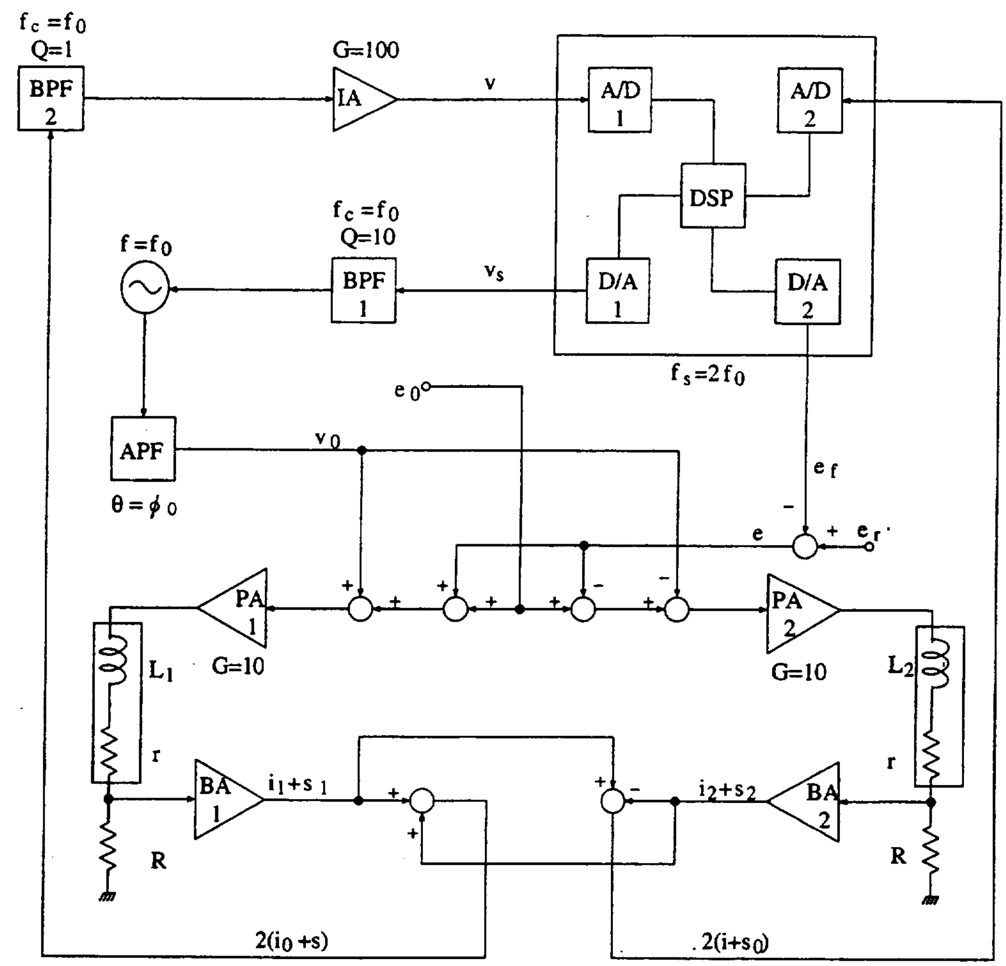

Fig. 3. Schematic diagram of a sensorless control system: R = 4 Ω, f_0 = 2 kHz, f_s = 4 kHz. The DSP board used as a digital controller has five A/D converters and five D/A converters in reality.

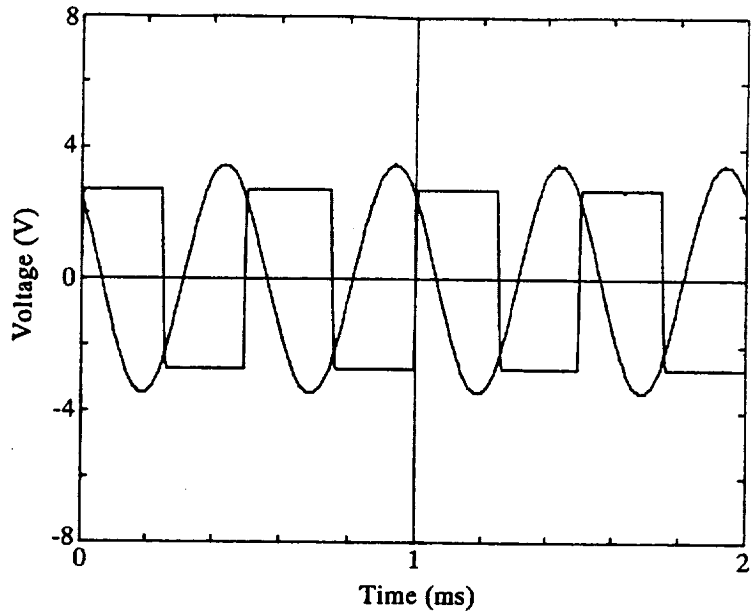

Fig. 4. A sinusoidal carrier signal and the synchronizing rectangular one. The phase difference between the two signals is equal to a phase lag given by APF.

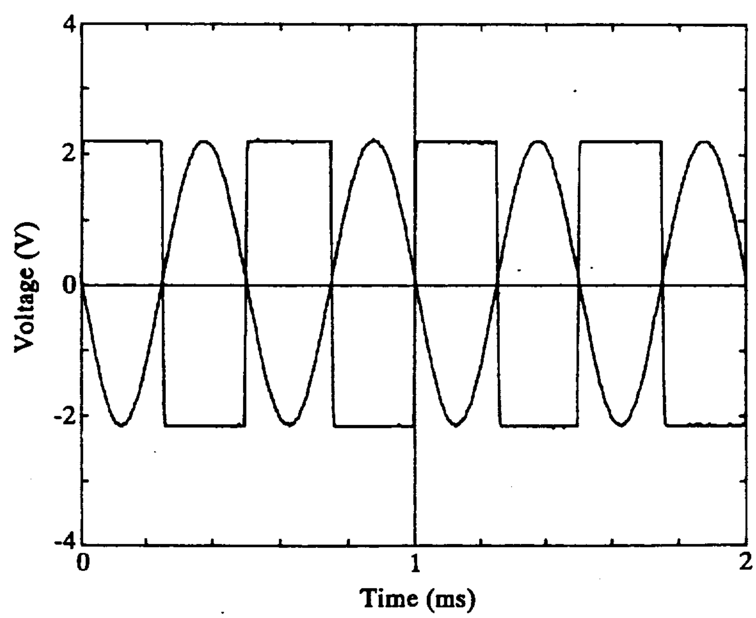

Fig. 5. A differential signal and the sampled data output (of D/A 3) when the arm is fixed at x = 0.3 mm.

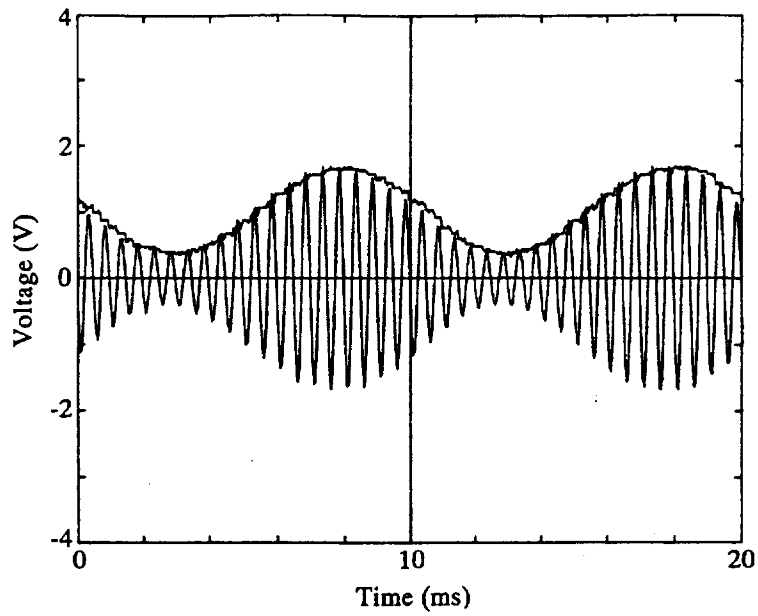

Fig. 6. An amplitude-modulated differential signal and the peak detection output (of D/A 3) when x = 0.3 mm. The modulating frequency is 100 Hz.

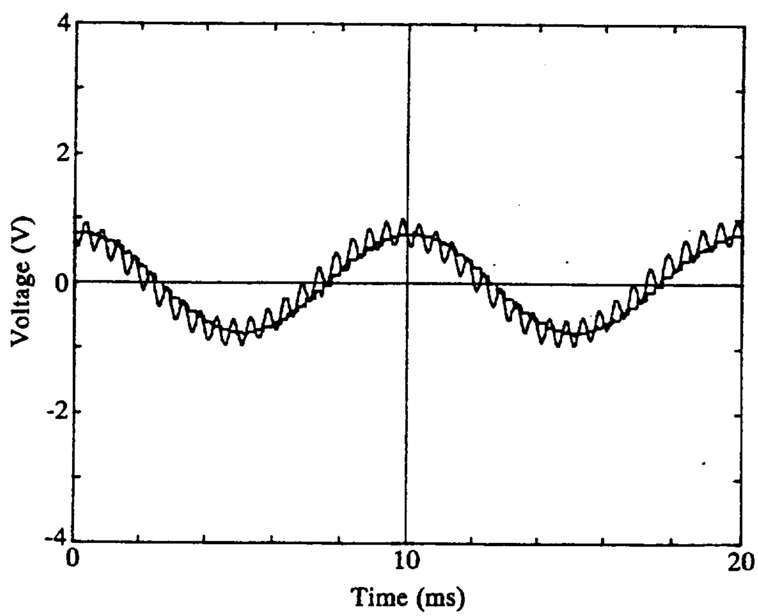

Fig. 7. A subtraction current signal and the output (of D/A 4) for the controlling current with a frequency of 100 Hz.

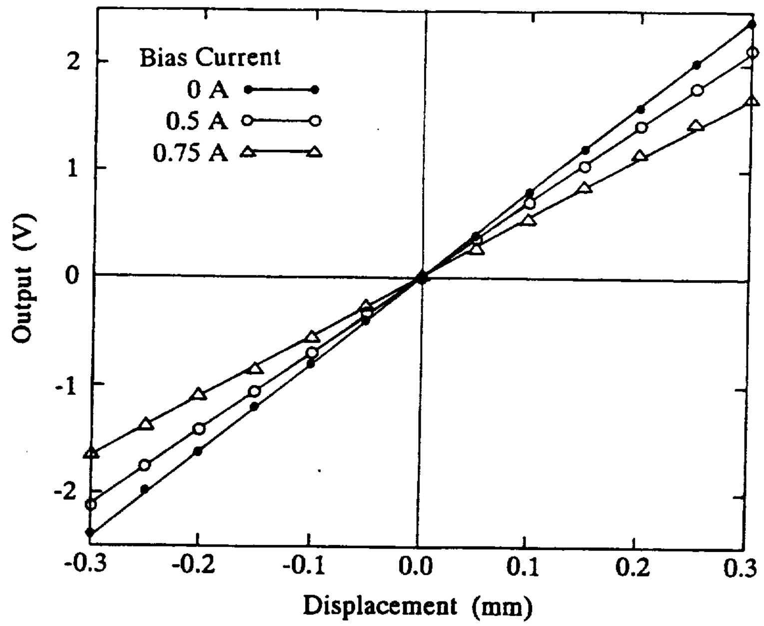

Fig. 8. Relation between the peak detection output (of D/A 3) and the arm displacement when $i = 0$ A. The arm displacement is measured by micrometers with a resolution of 10 μm.

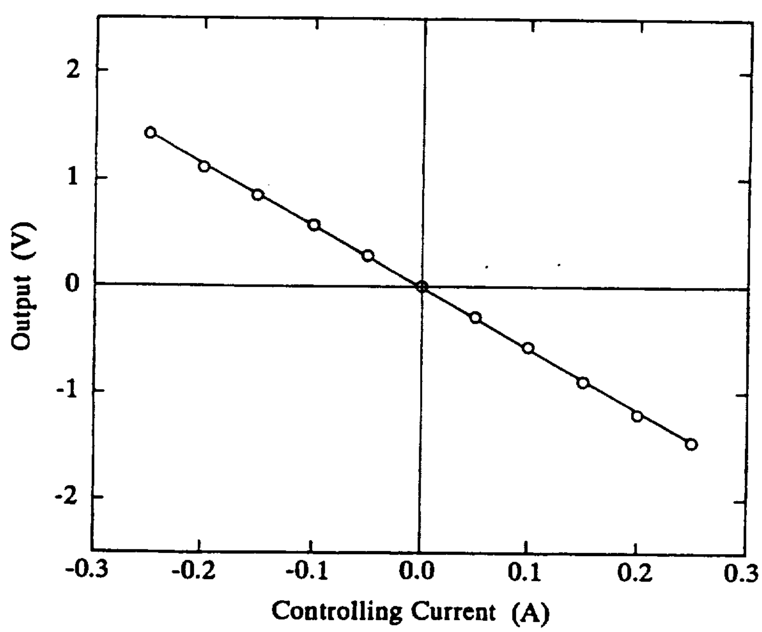

Fig. 9. Relation between the peak detection output (of D/A 3) and the controlling current when $i_0 = 0.5$ A and $x = 0$ mm.

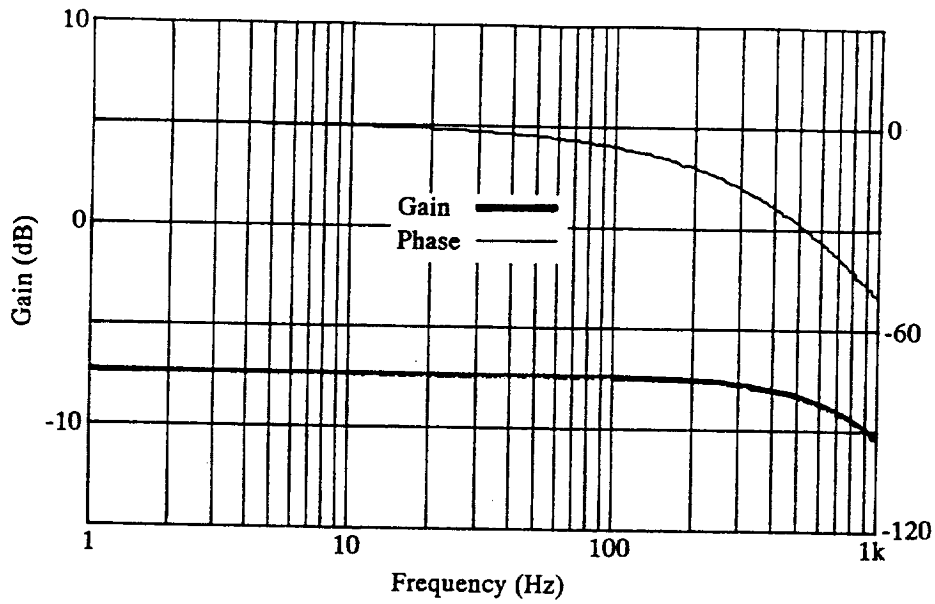

Fig. 10. Transmission characteristic from the modulating signal output (of D/A 5) to the peak detection output (of D/A 3) when i_0 = 0 A and x = 0.3 mm.

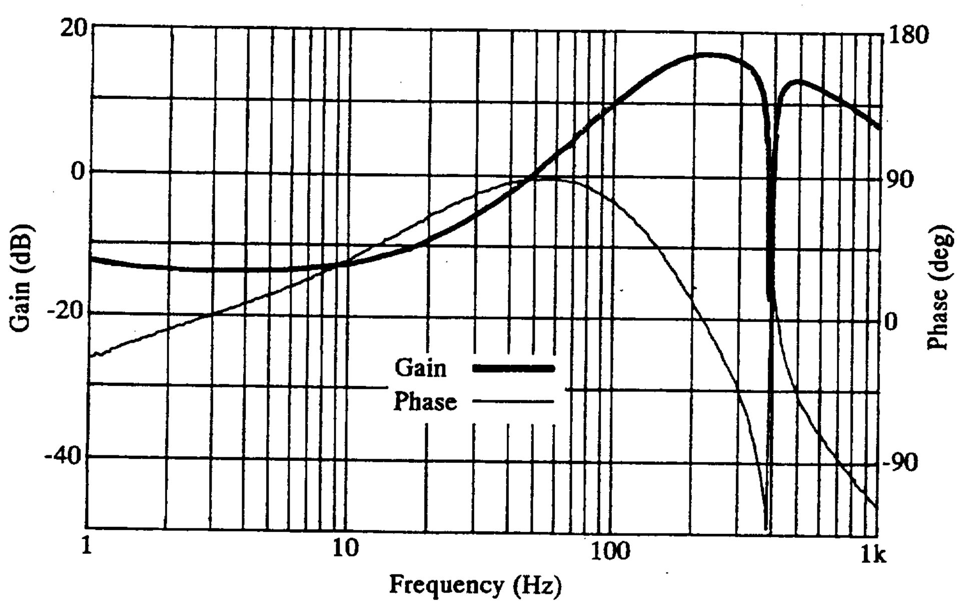

Fig. 11. Frequency response to a sinusoidal signal input (of A/D 1) of the feedback signal output (of D/A 2).

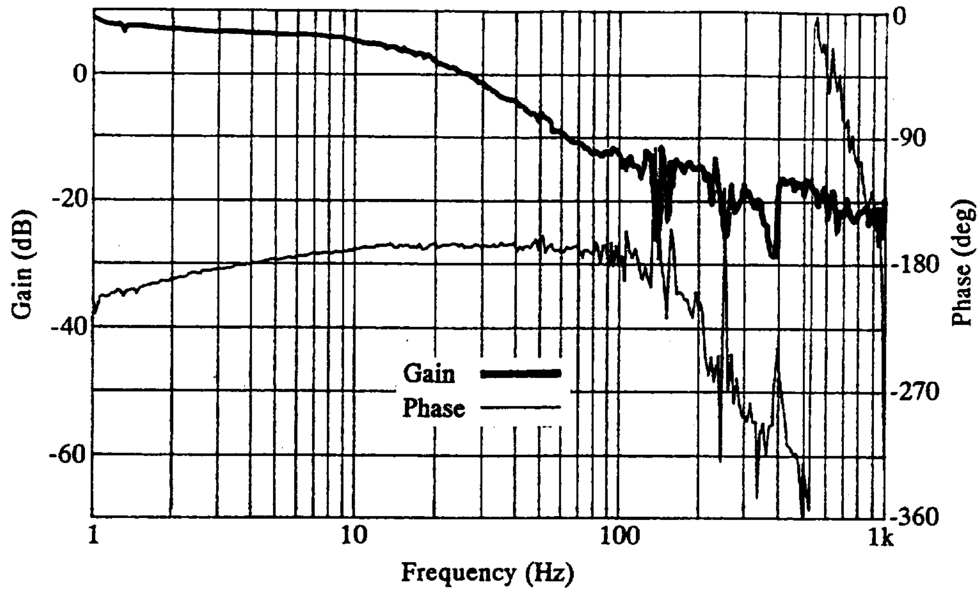

Fig. 12. Frequency response to a sinusoidal voltage command of the feedback signal output (of D/A 2). The arm is magnetically suspended by Electromagnets 1 and 2 with a bias current of 0.5 A.