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Implementation of phase-locked loop control for MEMS scanning mirror using DSP

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Abstract

The present study has employed the phase-locked loop control method to ensure the operating of MEMS actuators at their resonant frequency. In this study, the control algorism was simulated by the MATLAB. Further, the digital signal processing (DSP) technique was adopted to implement the concept of phase-locked loop control algorithm. Thus, a wide VCO lock-in dynamic range was achieved. In applications, the optical scanners fabricated using the SOI wafer and MOSBE process were respectively employed to demonstrate the present technique. The test show that the resonant frequency was tracked for various driving voltages, loop gains, and initial frequency offsets. Thus, the variation of the scanning angle resulted from the offset of the resonant frequency of the devices can be prevented.

Keywords: Phase-locked loop; Scanning mirror; Optical MEMS

1. Introduction

The resonance behavior is a useful dynamic characteristic for MEMS devices. This characteristic has been extensively employed to enlarge the output displacement of micro actuators. For instance, the linear displacement of a comb-drive actuator [1] and the scanning angular displacement of a torsional mirror [2] are significantly increased during resonance. This characteristic has been extensively employed in the vibratory MEMS gyroscope [3], the fatigue testing of MEMS devices [4], and the resonant type Coriolis meter as well [5]. However, the resonant frequency of MEMS devices is very sensitive to variations of fabrication, such as the feature size of suspension and the thickness of proof mass. Some technologies are useful to compensate the shift of resonant frequency due to variations of fabrication. Laser trimming removed parts of device so that could compensate shift of resonant frequency [6]. The selective deposited poly-Si on MEMS structures also can compensate the process-induced shift of resonant frequency [7]. The DC bias applied to the actuator could provide the effect of negative spring to compensate the shift of resonant frequency [8]. Moreover, the more critical problems is that the resonant frequency of MEMS devices are also very sensitivity to the environment conditions, such as ambient pressure and temperature. The resonant frequency of MEMS devices will vary with operating time due to the fatigue problem. In addition, the resonant frequency of a non-linear dynamic system may even vary with the operating conditions such as driving voltage [9]. Hence, it is necessary to tune the driving frequency for the aforementioned devices if their resonant frequency is shifted.

Presently, the phase-locked loop (PLL) technology has extensive applications in the area of communication, motor control, etc. In general, a phase-locked loop consists of three main functional components such as voltage-controlled oscillator (VCO), phase detector (PD), and loop filter. A typical second-order spring-mass dynamic system will experience a phase change at its resonant frequency. Thus, the phase change can be detected by the PD, and then controlled. The PLL has frequently been implemented using analog IC, such as NE565 and CD4046. However, an external resistor and capacitor are required to set the frequency of voltage-controlled oscillator (VCO). The PLL can also be implemented by means of software, named software-PLL [10]. To this end, the functions including the mixer, loop filter, A/D converter, and VCO can be properly established and integrated.

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This study presents a software-PLL control method to ensure the operating of MEMS actuators at their resonant frequency. The digital signal processing (DSP) technique was adopted to implement the concept of phase-locked loop control algorithm. In applications, the scanning mirror operating at resonant frequency is employed to demonstrate the feasibility of the software-PLL control method. Applied proposed method, the resonant frequency of scanner will be tracked to ensure its performance of large scanning angle.

2. Concepts and modeling

The dynamics of a single-degree-of-freedom spring-mass-damping (k-m-c) system can be expressed as the following second-order differential equation,

$$m\ddot{x} + c\dot{x} + kx = F\sin\omega_0 t \tag{1}$$

where F and ω_0 are the amplitude and frequency, respectively, of the external excitation. In this study, the transient response of Eq. (1) is ignored, and the steady-state response of the dynamic system is:

$$x(t) = A\sin\theta(t) = A\sin(\omega_0 t + \phi) \tag{2}$$

where A and ϕ denote the amplitude and phase angle of the response, respectively, and $\theta(t)$ represents the phase of response. The typical variations of amplitude A and phase angle ϕ are shown in Fig. 1 for different ω_0. It is obtained from Fig. 1 that the phase angle will shift for 90° at the resonant point. This characteristic has been employed in this study to trace the resonant frequency of the MEMS device. Hence, the MEMS devices can be operated at their resonant frequency if their phase is locked by control algorithm. This study further implements the concept of software-PLL control algorithm by using a DSP board.

Fig. 2a illustrates the block diagram of a conventional PLL consisting of phase detector, loop filter and VCO. The block diagram in Fig. 2b illustrates the concept to apply the PLL to track the resonant frequency of MEMS resonator [4]. In the present study, the phase angle will vary with time during the tracking of resonant frequency, thus the phase $\theta(t)$ of Eq. (2) can be rewritten as [11],

$$\theta(t) = \omega_0 t + \theta_1(t) \tag{3}$$

where $\theta_1(t)$ is the time varying phase angle. The frequency ω_0 is also regarded as the initial frequency of VCO. In this case, the VCO will provide an output phase $\theta_v(t)$ until the phase angle reach 90°, so as to ensure the MEMS resonator driving at its resonant frequency. The time derivative of $\theta_v(t)$ is expressed as,

$$\frac{d\theta_v(t)}{dt} = \omega_0 + K_2 e(t) \tag{4}$$

where K_2 is the sensitivity (gain) of the VCO. The VCO acts as an integrator, so that the Eq. (4) becomes [11],

$$\theta_v(t) = \omega_0 t + \theta_2(t) \tag{5}$$

where $\theta_2(t)$ is the phase angle of the VCO. To simplify the calculation, an additional gain of $\sqrt{2}$ will add to $y(t)$ before it input to PD. As indicated in Fig. 2b, the output $y(t)$ from PD (which acts as a multiplier) becomes,

$$y(t) = AK_1\{\sin[\theta(t) - \theta_v(t)] + \sin[\theta(t) + \theta_v(t)]\} \tag{6}$$

where K_1 is the VCO output gain. The loop filter, which employed to eliminate the double frequency in Eq. (6), will determine the DC component of the multiplied signals. Thus, the output $e(t)$ from loop filter becomes,

$$e(t) = e(0) + \int_0^t y(t-u) f(u)\,du \tag{7}$$

where $f(t)$ is the impulse function of the loop filter; in addition, the initial condition of loop filter $e(0)$ are zero in this study. After substituting of Eqs. (6) and (7) into Eq. (4) results in,

$$\frac{d\theta_v(t)}{dt} = \omega_0 + K_2 \int_0^t f(t-u) AK_1 \sin[\theta(u) - \theta_v(u)]\,du \tag{8}$$

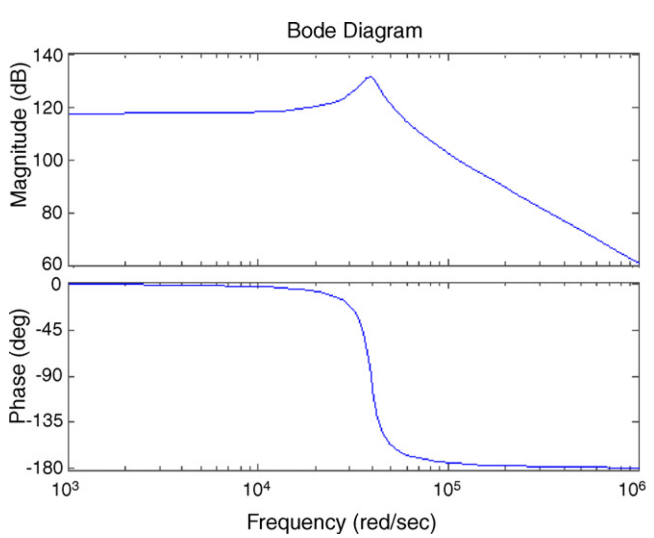

Fig. 1. A typical frequency response and phase change of a dynamic system.

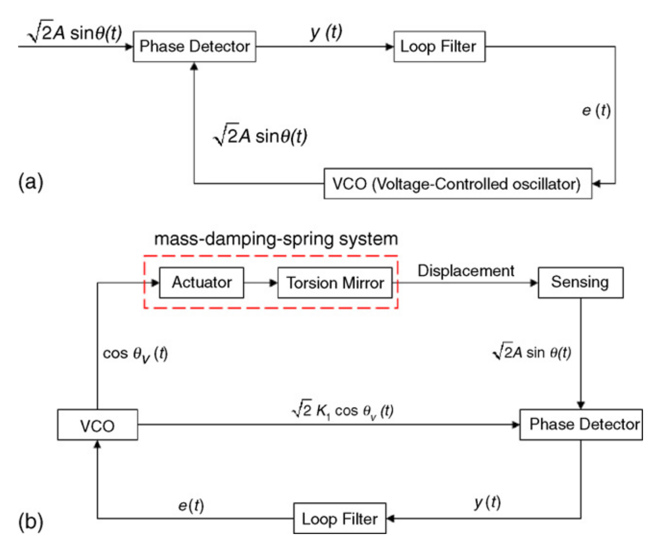

Fig. 2. (a) A typical block diagram of phase-locked loop, and (b) the modification block diagram of phase-locked loop.