Digital Demodulation using I/Q Signals and Optical Phase Control Applied to a Vibrometer

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Abstract—An efficient and cost effective optical phase detection vibrometer based on a modified closed loop homodyne Michelson interferometer is presented. Real-time phase demodulation is carried out, using an embedded platform that performs data acquisition, signal processing, PI (proportional-integral) control and the generation of signals that drive the electrooptic Pockels cell phase shifter and the piezoelectric actuator under test. Two phase quadrature signals are generated from a single interferometric output, using the interleaving action, in alternation, of a digitally generated modulating signal, and then the well-known differential-cross-multiplication technique is applied to perform the computation of the phase shift of interest. The quadrature condition is reached using the PI loop based on an error signal obtained from a Lissajous figure derived from out-of-phase signals. The vibrometer is capable of measuring nanometric displacements, and is simple, inexpensive, accurate, immune to fading and self-consistent. The new method was used to determine the displacement frequency response curves of two prototypes of multi-actuated flextensional piezoelectric actuators. Measurements were made between 500 Hz and 15 kHz, and the results agreed with those obtained by the standard SCM-Signal Coincidence Method.

Index Terms—Optical interferometry, Piezoelectric transducers, Signal processing, Metrology.

I. INTRODUCTION

O PTICAL interferometry has many applications in the measurement of sub-nanometer dynamic displacements. From the classical Michelson interferometer, numerous versions of the principle of measuring physical quantities using optical interference have been developed, such as [1, 2] Mach-Zehnder, Fabry-Pérot, Sagnac, etc.

Specifically, in sub-microscopic motion measurements, interferometry has several advantages over other traditional non-interferometric methods such as high sensitivity, resolution, signal-to-noise ratio, wide dynamic range, immunity to magnetic coupling and non-contact measurement mode. The task of displacement interferometry is to measure the phase shift of an optical wave ($\Phi(t)$) as a reflective target moves, and to infer the respective displacement ($\Delta L(t)$) by comparison with the known wavelength ($\lambda$) of the laser used: $\Delta L(t) = \lambda \Phi(t)/4\pi$, for the case of a Michelson interferometer [1].

Interferometric-based systems meet the increasing demand for high accuracy motion/position measurements in important industrial applications such as precision machining [3], nanotechnology [4], photolithography [5], semiconductor manufacturing [6], micro- and nano-mechanical-systems [7, 8]. Different methods of optical phase detection have been investigated in order to demodulate interferometric signals, such as homodyne [9] and heterodyne techniques [2, 3, 10], with spectral [11]–[13] or time domain analysis [14], as well as open-loop [15] or closed loop [16]–[18] approaches.

However, despite their remarkable performances, even state-of-the-art interferometers are susceptible to various sources of error such as noise, nonlinearities and phase ambiguities. Environmental influences can be important as they reduce repeatability, accuracy and reliability. Temperature variations, vibrations, and air turbulence lead to random variations in phase difference between the reference and sensor arms of the interferometer, leading to fading of the detected interferometer output signal [1].

To solve the fading problem, some schemes used dual interferometers, one for measurement and the other for path compensation [3, 10]. Other schemes use passive compensation of random optical phase variations, as occurs in detection methods based on the photodetected signal spectrum [12, 13], PGC [15] or I/Q-interferometry [11, 14]. However, these methods involve time-consuming off-line software algorithms or complicated optical/electrical hardware. Closed-loop interferometric techniques have demonstrated good performances such as PID (proportional-integral-derivative) control [9, 16], LMI (Linear Matrix Inequality) [17] and sliding mode control [18].

The detection scheme of the interferometric-based Optical Vibrometer (OV) presented in this paper is of the closed-loop type, but unlike those mentioned above, it does not need to use fiber Bragg grating or optical spectrum analyzer as in [9], does not require the previous and rigorous calibration of the piezoelectric transducer (PZT) phase controllers [16, 18], and does not require complicated optical hardware or feedback circuit/control method [17].

Among possible implementations of the optical phase shift method to stabilize closed-loop interferometers, one of the most popular is the use of PZTs, whereby a phase shift is obtained by introducing a variation in the optical path length. PZTs are simple to use, but have some drawbacks such as hys-
teresis, creep, nonlinearities and frequency resonances. If high phase shift accuracy is required, the cost of high-performance PZT controllers is significant. In some applications, the phase shifter, control electronics and the interface computer are the most expensive elements, as in [19]. The optical phase shifter used in this work is a bulk electro-optic phase modulator, based on lithium niobate (LiNbO
_3_). The Pockels cell, manufactured by the authors, and operates at voltages lower than 44 V. It has excellent linearity, large bandwidth and no hysteresis.

The OV presented in this work simply consists of a bulk Michelson interferometer, slightly modified to accommodate the electro-optic phase controller in one of its arms. Although only one photodetector is used in its output, the interferometric system employs the I/Q signals for phase demodulation, which would usually require two photodetectors. This is made possible by using the interleaving action of a digital signal generated in the detection process. This avoids the use of complicated polarimetric optical hardware commonly utilized to generate two phase quadrature signals [20]. In addition, the quadrature condition is reached by the action of the employed PI controller itself, with no time-consuming computationally intensive algorithms to correct the ellipticity of the corresponding Lissajous figure by applying least-square fitting methods [20, 21], learning algorithms [22], among others. Once phase quadrature terms are available, some authors usually use the ‘arctan’ demodulation method [23], which requires time-consuming phase unwrapping algorithms. Instead, the differential-cross-multiplying technique is used to demodulate the phase shift of the signal of interest [15, 23]. To accomplish these tasks, real-time measurement and processing are achieved through the use of embedded platforms and dedicated software for digital acquisition and digital signal processing.

In this paper, we propose a new method of nano-vibration measurement that is simple, cost-effective, accurate, immune to fading and self-consistent (independent of variations in system parameters such as fringe visibility, laser power, photodiode responsivity, etc.). The PI control system is well tolerant against electronic noise, as well as temperature induced variations in the Pockels cell half-wave voltage and amplifier gain.

II. FLEXURE-EXTENSIONAL ELECTROMECHANICAL TRANSDUCERS

Due to rapid development of semiconductor technology, nanotechnology, micro- or nano-electro-mechanical systems, and biotechnology, the demand for ultra-precise micro- and nanoscale positioning devices has increased. Flextensional Piezoelectric Actuators (FPA) are flexure-based mechanical drives, thus constituting appropriate platforms for micro or nano manipulation which can be used on these various technical fields [24]. Modern FPAs act as mechanical transformers, using a piezoelectric ceramic (such as PZT) as an active element bounded to a flexible metal structure whose function is to amplify and change the direction (from extensional to flexural) of the small mechanical motions of the piezoceramics. The stiffness and flexibility of the metal structure determine the rate of motion and generated force. In general, these flexure-based mechanisms are monolithic structures comprising solid links and compliant mechanisms where the action is given by the elastic deformation of the metal and not by the presence of joints and pins. Monolithic construction reduces assembly errors and ensures high accuracy positioning, on which we have published several studies [12, 25]. As a result, these mechanisms offer several advantages such as excellent motion resolution, negligible friction, zero backlash and ease of maintenance.

On the other hand, FPA used as driving sources have some drawbacks such as nonlinearities, hysteresis, creep and tracking error (due to spurious motions induced by high-frequency mechanical resonances) effects. The design of these actuators to overcome these problems is very complex. The topology optimization method is a powerful design technique which, combined with the finite element method, enables the design of FPAs pursuing the best possible combination of the range of motion and the load capability. A theoretical study of this FPA design technique and manufacturing is beyond the scope of this paper [26, 27].

MAFPAs consist of multi-flexible structures actuated by two or more piezoceramics, and thus, can provide more than one degree of freedom. This solution allows, for example, the development of nanopositioner XY stages [28], tilt-positioning mechanisms [29], piezoelectric grippers [30] and piezoelectric motors [31]. The MAFPA devices used in this paper were designed (using the topology optimization method) and manufactured utilizing two PZT piezoceramics, and therefore actuators with two degrees of freedom were obtained. Our main interest is that such devices can be used as piezoelectric motors.

At present, the main kind of commercially available devices for piezoelectric motor applications are of the following types: a) flexible hinge-based nanopositioner, where the solid links are generally rigid and the flexure hinges are compliant in bending about one axis but rigid about the cross axis [32]; b) inchworm motor, where the motor imitates the motion mode of an inchworm crawling, viz. the amount of continuous nanoscale movement occurs through a series of cyclic deformation motions, generally formed by three piezoelectric actuators [31, 33]; and c) ultrasonic motor [34, 35], where the stator formed in piezoelectric structures generates an ultrasonic vibration at frequencies of several kHz, and the friction motion can change this ultrasonic vibration into a rectilinear motion of the slider. All of these structures were composed of several piezoceramics or piezoelectric actuators embedded in a flexible metal structure.

Our ultrasonic type piezoelectric motor is based on a MAFPA XY nanopositioner, designed by the authors using the topology optimization method. The MAFPA shown in Fig. 1(a), MAFPAXY [13], consists of two piezoceramics PZT-5A (American Piezoceramics, dimensions of 6, 20 and 1 mm in the a-, b- and c-directions, poled along c-direction), \( P_{1X} \) and \( P_{1Y} \), and an aluminum flexible structure, which amplifies and redirects the induced vibrations when voltages are applied to the piezoceramics. The horizontal piezoceramic generates vertical movement of the MAFPA vertex (or tip), while the vertical piezoceramic, the horizontal movement. The bending vibration mode and stretching vibration mode of the
The OV experimental setup is shown in Fig. 2. The single-mode laser beam is provided by a low-cost laser diode (Mitsubishi, ML101J18) with the output power of 100 mW and a central wavelength of 658 nm. The laser beam is divided by beam splitter BS1 into two beams of equal intensity, which correspond to the sensor and reference arms of the interferometer. The laser beam is collimated so that it propagates in the LiNbO3 crystal (Crystal Technologies, dimensions of 10.0, 50.0 and 1.1 mm in the x-, y- and z-axis of the crystal, optical propagation in the y-direction) of the Pockels cell. A vertical polarizer, aligned with the z-axis of the Pockels cell, is used to ensure that the laser beams in each of interferometer arms have the same polarization. This permits only one propagation mode in the Pockels cell and, when the beams are recombined, achieves good fringe visibility. Unlike a conventional Michelson interferometer, the reference laser beam propagates only once through the Pockels cell and does not return to BS1. Instead, it is directed towards the beam splitter BS2 by plane mirrors M1 and M2. In the sensor arm, the lightwave is reflected in the mirror attached to the MAFPA to be characterized, directed back to BS1 and then to BS2. To improve the fringe visibility, a beam expander is used in the sensor arm. Finally, BS2 combines and directs both beams to an amplified PIN photodiode (Thorlabs, PDA55). The photodetected signal is filtered by a low-pass 80 kHz cutoff frequency anti-aliasing filter. The resulting signal, \( v_{PD}(t) \), is acquired and digitized by an I/O device (National Instruments, USB-6211) for real-time processing using a personal computer. In addition to acquiring the filtered photodetected signal, the I/O device generates two analog signals: the first signal is amplified by a linear amplifier (A. A. Lab Systems, A-301 Hs) and then filtered by a second order 80 kHz cutoff frequency low-pass analog filter. The resulting signal, \( v_P(t) \), is applied to the Pockels cell in such a way that the generated electric field is parallel to the vertically polarized laser beam, i.e., this electric field is induced along the z-axis. The second signal is filtered by a 150 kHz cutoff frequency low-pass filter. The resultant signal, \( v_{PZT}(t) \), is applied to the stator, generating elliptical motion on the contact surface of the stator. The elliptical motion should push the slider to move under the friction. In this case, one of the purposes of the topology optimization is to minimize undesirable coupling effects between generated and coupled displacements [36]. In the present work, only the horizontal generated movement was tested, as the main purpose here is to investigate the performance of our OV system in measuring nanometric motions, and not the MAFPA itself. Reflective tape (3M, Scotchlite 7610) was adhered at the measurement points (near the tip) to reflect a significant portion of the optical beam of the sensor arm of the interferometer.

Another piezoelectric transducer, XY05B [36], shown in Fig. 1(b), was also manufactured by the authors and it is composed by two piezoceramics PZT-5A (American Piezoceramics, dimensions of 5, 7 and 1 mm in the a-, b- and c-directions, poled along c-direction), \( P_{2X} \) and \( P_{2Y} \), bonded to an optimized aluminum structure, which minimizes the coupled movement.

The performance of the MAFPAs is investigated using our new OV system. The application of the MAFPAs as motors will be investigated in a future work.

III. PROPOSED METHOD

The OV experimental setup is shown in Fig. 2. The single-mode laser beam is provided by a low-cost laser diode (Mitsubishi, ML101J18) with the output power of 100 mW and a central wavelength of 658 nm. The laser beam is divided by beam splitter BS1 into two beams of equal intensity, which correspond to the sensor and reference arms of the interferometer. The laser beam is collimated so that
one of the transducer piezoceramics. The purpose of the two analog filters was to reduce digital-to-analog conversion noise. The photodetected signal \( v_{PD}(t) \) is [1],

\[
v_{PD}(t) = A\{1 + V \cos(\Phi(t))\},
\]

where \( V \) is the fringe visibility, \( A \) is a constant voltage proportional to the average received optical power. The total optical phase shift \( \Phi(t) \) is the sum of the optical phase shift \( \phi_P(t) \) induced by the Pockels cell due to the application of \( v_P(t) \), the dynamic optical phase shift \( \phi_{PZT}(t) \) of the MAFPA when a voltage \( v_{PZT}(t) \) is applied to one of its piezoceramics, and the random slowly-varying optical phase shift \( \phi_Q(t) \) that can cause fading of the interferometric signal. It will be shown that the optical phase demodulation method works independently of \( \phi_Q(t) \). \( \Phi(t) \) can be written as:

\[
\Phi(t) = \phi_P(t) + \phi_{PZT}(t) + \phi_Q(t).
\]

\( v_{PD}(t) \) is a phase-modulated signal where the objective is to obtain \( \phi_{PZT}(t) \) by means of a digital demodulation process. Consequently, the MAFPA input-output relationship, given in \( \text{rad/V} \), can be characterized for a specific frequency when \( v_{PZT}(t) \) is a sinusoidal signal. By varying the frequency of \( v_{PZT}(t) \), the MAFPA frequency response can be determined.

The relation between \( v_P(t) \) and the induced optical phase shift \( \phi_P(t) \) is given by [37],

\[
\phi_P(t) = \frac{\pi}{V_v} v_P(t),
\]

where \( V_v \) is the Pockels cell half-wave voltage, which has a nominal value of 44 V in the cell used in this work. In the signal processing method, \( v_P(t) \) is the sum of two signals: a principal modulating voltage \( v_M(t) \) and an auxiliary modulating voltage \( v_A(t) \):

\[
v_P(t) = v_M(t) + v_A(t).
\]

In principle, \( v_M(t) \) could be any periodic half-wave symmetric waveform because the phase control system will always adjust its amplitude in order to guarantee an amplitude difference of \( V_v/2 \) between two successive acquisition points. In this work, a square waveform was generated with a rate of two samples per cycle \( 1/f_M \).

At really, \( v_M(t) \) is a distorted square wave due to the low-pass filter and the amplifier, as shown in Fig. 3(c). In this work, \( v_M(t) \) has a fundamental frequency \( f_M = 30 \text{ kHz} \) and its amplitudes at the acquisition points are \( \pm V_M \). \( v_M(t) \) is used to induce an optical phase shift of \( \pm \pi/4 \) when \( V_M \) is given by:

\[
V_M = \frac{V_v}{4}.
\]

(5)

\( v_A(t) \) is a sinusoidal voltage,

\[
v_A(t) = V_A \sin(2\pi f_A t),
\]

where \( f_A = 20 \text{ Hz} \). As it will be seen, \( v_A(t) \) is used to carry out an optical phase control process, which is independent of the voltage applied to the piezoceramic of the MAFPA, to guarantee that (5) is always satisfied. Equation (3) can be rewritten as:

\[
\phi_p(t) = \frac{\pi}{V_v} v_M(t) + \frac{V_A}{V_v} \sin (2\pi f_A t). \quad (7)
\]

When (5) is satisfied, at the acquisition points \( t_i \), (7) is given by:

\[
\phi_p(t_i) = \pm \frac{\pi}{4} + \frac{V_A}{V_v} \sin (2\pi f_A t_i). \quad (8)
\]

\( v_{PD}(t) \) is acquired with a sampling frequency of \( f_s = 2f_M = 60 \text{ kHz} \). For the system to work, the acquisition of \( v_{PD}(t) \) has to be synchronized with the generation of \( v_M(t) \) such that the samples of \( v_{PD}(t) \) are acquired twice per each period of \( v_M(t) \). One sample of \( v_{PD}(t) \) is acquired when \( v_M(t) = +V_M \) and the another one when \( v_M(t) = -V_M \). This synchronization, illustrated in Fig. 3, can be done by generating a clock signal (Fig. 3(d)) with a frequency \( f_s = 60 \text{ kHz} \). Therefore, the samples of \( v_{PD}(t) \) (Fig. 3(a)-(b)) and \( v_M(t) \) (Fig. 3(c)) are simultaneously acquired and generated, respectively, during the rising edges of the clock signal. Once synchronization is achieved, the acquired photodetected signal (1) at integer multiples \( n \) of the sampling period \( T_S = f_s^{-1} \),
\( v_{PD}(nT_S) \), is given by:

\[
v_{PD}(nT_S) = A \left\{ 1 + V \cos \left[ \frac{\pi}{4} + \frac{V_A}{V_\pi} \sin(2\pi f_A nT_S) + \phi_{PZT}(nT_S) + \phi_Q(nT_S) \right] \right\}, \tag{9}
\]

where \( v_M(nT_S) = +V_M = +V_\pi/4 \), for \( n = 2k \), \( k \in \mathbb{Z}, k \geq 0 \),

and

\[
v_{PD}(nT_S) = A \left\{ 1 + V \cos \left[ -\frac{\pi}{4} + \frac{V_A}{V_\pi} \sin(2\pi f_A nT_S) + \phi_{PZT}(nT_S) + \phi_Q(nT_S) \right] \right\}, \tag{10}
\]

when \( v_M(nT_S) = -V_M = -V_\pi/4 \), for \( n = 2k + 1 \), \( k \in \mathbb{Z}, k \geq 0 \).

As can be observed from (9) and (10), \( v_M(t) \) allows us to obtain two signals from \( v_{PD}(t) \), which have a sampling frequency of \( f_S/2 = 30 \text{ kHz} \) so their maximum frequencies are \( 15 \text{ kHz} \). As will be discussed, after normalizing, these two signals will generate Fig. 4.

Equations (9) and (10) do not take values simultaneously since their samples are alternately acquired. Both signals are subjected to an interpolation process, by which three samples are created between two existing samples, effectively increasing the sampling frequency to \( 120 \text{ kHz} \). Because the acquired samples are alternately taken with a sampling frequency of \( f_S = 60 \text{ kHz} \), it is necessary to delay one of the interpolated signals by a time \( T_S \) with respect to the other to obtain two signals that are always in quadrature as shown in the inset of Fig. 4.

After the interpolation and sample-delay processes, \( v_1[n] \) and \( v_2[n] \) are given by:

\[
v_1[n] = A \left\{ 1 + V \cos \left[ \frac{\pi}{4} + \frac{V_A}{V_\pi} \sin(2\pi f_A nT'_S) + \phi_{PZT}(nT'_S) + \phi_Q(nT'_S) \right] \right\}, \tag{11}
\]

and

\[
v_2[n] = A \left\{ 1 + V \sin \left[ \frac{\pi}{4} + \frac{V_A}{V_\pi} \sin(2\pi f_A nT'_S) + \phi_{PZT}(nT'_S) + \phi_Q(nT'_S) \right] \right\}, \tag{12}
\]

where \( T'_S = T_S/2 \) is the new sampling period.

The constant \( A \) can be calculated as:

\[
A = \frac{v_{PD_{\text{max}}} + v_{PD_{\text{min}}}}{2}, \tag{13}
\]

where \( v_{PD_{\text{max}}} \) and \( v_{PD_{\text{min}}} \) are, respectively, the maximum and the minimum values of \( v_{PD}(nT_S) \). Subtracting \( A \) from \( v_1[n] \) and from \( v_2[n] \), results:

\[
x_1[n] = AV \cos \left( \Omega(nT'_S) \right), \tag{14}
\]

and

\[
x_2[n] = AV \sin \left( \Omega(nT'_S) \right), \tag{15}
\]

where,

\[
\Omega(nT'_S) = \frac{\pi}{4} + \frac{V_A}{V_\pi} \sin(2\pi f_A nT'_S) + \phi_{PZT}(nT'_S) + \phi_Q(nT'_S). \tag{16}
\]

Fig. 4. Normalized signals \( y_1[n] \) and \( y_2[n] \). The inset shows the interpolated samples and the original ones.

\( x_1[n] \) and \( x_2[n] \) are in phase quadrature, therefore,

\[
x_1^2[n] + x_2^2[n] = (AV)^2. \tag{17}
\]

The discrete-time phase shift \( \Omega(nT'_S) \) can be obtained by using the differential-cross-multiplication method. Let \( x_1(t) \) and \( x_2(t) \) be the continuous time-domain representations of \( x_1[n] \) and \( x_2[n] \), given by (14) and (15), respectively, with \( t = nT'_S \). Their time derivatives are

\[
\frac{dx_1(t)}{dt} = -AV \Omega(t) \sin(\Omega(t)), \tag{18}
\]

and

\[
\frac{dx_2(t)}{dt} = AV \Omega(t) \cos(\Omega(t)), \tag{19}
\]

therefore:

\[
x_1(t) \frac{dx_2(t)}{dt} - x_2(t) \frac{dx_1(t)}{dt} = (AV)^2 \frac{d\Omega(t)}{dt}. \tag{20}
\]

Taking into account (17) and (20), the time derivative of \( \Omega(t) \) is obtained as:

\[
\frac{d\Omega(t)}{dt} = \frac{x_1(t) \frac{dx_2(t)}{dt} - x_2(t) \frac{dx_1(t)}{dt}}{x_1^2(t) + x_2^2(t)}. \tag{21}
\]

The phase shift \( \Omega(nT'_S) \) can be obtained by integrating (21) and, according to (16), it consists of the constant term \( \pi/4 \), the quasi-static phase drift \( \phi_Q(nT'_S) \), the low-frequency term \( \pi V_A/V_\pi \sin(2\pi f_A nT'_S) \) and the phase shift \( \phi_{PZT}(nT'_S) \). To avoid interference between the low-frequency band and the MAPFA operation band, the frequency response measurement is limited from 500 Hz to 15 kHz. It is then possible to obtain \( \phi_{PZT}(nT'_S) \) from \( \Omega(nT'_S) \) by using a high-pass filter.
this way, this control process allows the correct demodulation and is immune to signal fading caused by random
variations in $\phi_{\text{PZT}}$. As the main interest in the work is to obtain the frequency response of MAFPAs, only the particular case of sinusoidal signals is considered.

$\Delta V$ can be determined from (1) as,

$$\Delta V = \frac{v_{\text{PD}} \max - v_{\text{PD}} \min}{2}.$$  

The $x_1[n]$ and $x_2[n]$ normalized signals, as shown in Fig. 4, are given by:

$$x_1[n] = \frac{x_1[n]}{AV} = \cos (\Omega (nT_S'))$$  

and

$$x_2[n] = \frac{x_2[n]}{AV} = \sin (\Omega (nT_S')).$$

According to (24) and (25), the Lissajous figure corresponding to the points $(y_1[n], y_2[n])$ is a unit circle, shown in Fig. 5, since:

$$y_1^2[n] + y_2^2[n] = 1.$$  

At this point, although the demodulation method works correctly and is immune to signal fading caused by random variations in $\phi_Q(t)$, in practice, electronic noise due to the amplifier, the ADC and DAC processes is present, which can cause variations in the value of $V_M$, making (5) invalid. Consequently, there is a need for a control process to maintain (5). In this way, this control process allows the correct demodulation of the optical phase shift $\phi_{\text{PZT}}(nT_S')$.

When the MAFPA operates in its linear region, phase $\phi_{\text{PZT}}(t)$ is proportional to the sinusoidal voltage $v_{\text{PZT}}(t)$. Then, $\phi_{\text{PZT}}(t)$ can be written as:

$$\phi_{\text{PZT}}(t) = \phi_{\text{PZT}} \sin (\omega_{\text{PZT}} t),$$  

where $\omega_{\text{PZT}} = 2\pi f_{\text{PZT}}$ is the angular frequency of the MAFPA, $f_{\text{PZT}}$ its operation frequency and $\phi_{\text{PZT}}$ the amplitude.

We define $\phi_{L1}(t)$ and $\phi_{L2}(t)$ as the low-frequency optical phases

$$\phi_{L1}(t) = \frac{\pi}{4} + \frac{V_A}{V_\pi} \sin (2\pi f_A t) + \phi_Q(t)$$  

and

$$\phi_{L2}(t) = -\frac{\pi}{4} + \frac{V_A}{V_\pi} \sin (2\pi f_A t) + \phi_Q(t).$$  

Using (27), (28) and (29), (14) and (15) can be rewritten as,

$$x_1[n] = AV \cos (\phi_{L1}(nT_S')) + \phi_{\text{PZT}} \sin (\omega_{\text{PZT}} nT_S'))$$  

and

$$x_2[n] = AV \cos (\phi_{L2}(nT_S')) + \phi_{\text{PZT}} \sin (\omega_{\text{PZT}} nT_S')).$$

Equations (30) and (31) are phase-modulated signals, which can be expanded as series of Bessel functions of the first kind and rank $m$:

$$x_1[n] = AV \cos (\phi_{L1}(nT_S')) [J_0(\phi_{\text{PZT}}) + 2 \sum_{m=1}^{\infty} J_{2m}(\phi_{\text{PZT}}) \cos (2m \omega_{\text{PZT}} nT_S')]$$

$$- AV \sin (\phi_{L1}(nT_S')) \times 2 \sum_{m=1}^{\infty} J_{2m-1}(\phi_{\text{PZT}}) \sin ((2m - 1) \omega_{\text{PZT}} nT_S').$$

From (32), the frequency spectrum of $x_1[n]$ consists of the spectrum of the phase-modulated term $AVJ_0(\phi_{\text{PZT}}) \cos (\phi_{L1}(nT_S'))$ and the high-frequency sidebands centered at integer multiples of $f_{\text{PZT}}$. Similarly, from (33), the spectrum of $x_2[n]$ consists of the spectrum of $AVJ_0(\phi_{\text{PZT}}) \cos (\phi_{L2}(nT_S'))$ followed by the high-frequency sidebands centered at integer multiples of $f_{\text{PZT}}$. In addition, $AVJ_0(\phi_{\text{PZT}}) \cos (\phi_{L1}(nT_S'))$ and $AVJ_0(\phi_{\text{PZT}}) \cos (\phi_{L2}(nT_S'))$ can also be expanded in series of Bessel functions of the first kind. Thus, taking into account (28) and (29), results:

$$AVJ_0(\phi_{\text{PZT}}) \cos (\phi_{L1}(nT_S')) = AVJ_0(\phi_{\text{PZT}}) \times \cos \left( \frac{\pi}{4} + \phi_Q(nT_S') \right) [J_0 \left( \frac{V_A}{V_\pi} \right)]$$

$$+ 2 \sum_{m=1}^{\infty} J_{2m} \left( \frac{V_A}{V_\pi} \right) \cos (2m \omega_{\text{AN}} nT_S')$$

$$- AVJ_0(\phi_{\text{PZT}}) \sin \left( \frac{\pi}{4} + \phi_Q(nT_S') \right) \times 2 \sum_{m=1}^{\infty} J_{2m-1} \left( \frac{V_A}{V_\pi} \right) \sin ((2m - 1) \omega_{\text{AN}} nT_S'),$$
and

\[
AVJ_0(\phi_{PZT}) \cos(\phi_{L2}(nT_S')) = AVJ_0(\phi_{PZT}) \times \cos \left(-\frac{\pi}{4} + \phi_Q(nT_S')\right) + 2 \sum_{m=1}^{\infty} \left(J_{2m} \left(\frac{V_A}{\sqrt{\pi}}\right) \cos(2m\omega_A n T_S') - AVJ_0(\phi_{PZT}) \sin \left(-\frac{\pi}{4} + \phi_Q(nT_S')\right) \times 2 \sum_{m=1}^{\infty} J_{2m-1} \left(\frac{V_A}{\sqrt{\pi}}\right) \sin((2m - 1)\omega_A n T_S')\right),
\]

where \(\omega_A = 2\pi f_A\).

From (34) and (35), since \(f_A = 20 \text{ Hz}\), the first ten spectral components of \(AVJ_0(\phi_{PZT}) \cos(\phi_{L1}(nT_S'))\) and \(AVJ_0(\phi_{PZT}) \cos(\phi_{L2}(nT_S'))\) are contained in the frequency band between 0 Hz and 200 Hz. Then, without a significant loss of their frequency spectrum, \(x_1[n]\), as well as \(x_2[n]\), can be filtered by a 200 Hz cutoff frequency low-pass digital filter. The filtered signals \(x_{1f}[n]\) and \(x_{2f}[n]\) are given by the first terms of (32) and (33), respectively:

\[
x_{1f}[n] \approx AVJ_0(\phi_{PZT}) \cos(\phi_{L1}(nT_S'))),
\]

and

\[
x_{2f}[n] \approx AVJ_0(\phi_{PZT}) \cos(\phi_{L2}(nT_S')).
\]

Taking into account (28) and (29), (37) can then be rewritten as:

\[
x_{2f}[n] \approx AVJ_0(\phi_{PZT}) \sin(\phi_{L1}(nT_S')).
\]

Equations (36) and (38) can be normalized by calculating the constant term \(AVJ_0(\phi_{PZT})\) from either \(x_{1f}[n]\) or \(x_{2f}[n]\) in the same way as \(AVJ\) was calculated in (23). Then:

\[
y_{1f}[n] = \frac{x_{1f}[n]}{AVJ_0(\phi_{PZT})} = \cos(\phi_{L1}(nT_S')), \quad (39)
\]

and

\[
y_{2f}[n] = \frac{x_{2f}[n]}{AVJ_0(\phi_{PZT})} = \sin(\phi_{L1}(nT_S')). \quad (40)
\]

These signals are shown in Fig. 6, where it can be observed that the high-frequency components were removed by the low-pass filters. An experimental Lissajous figure, when (5) is satisfied, consisting of the points \((y_{1f}[n], y_{2f}[n])\) is shown in Fig. 7. On the other hand, when (5) is not satisfied, \(v_M(t)\) has an amplitude \(V_M' = V_\pi/4 + \Delta V_M\), where \(\Delta V_M\) is the variation of \(V_M'\) from the ideal value \(V_\pi/4\). Then, taking into account (3), \(v_M(t)\) no longer induces an optical phase shift of \(\pi/4\). Instead, there will be a phase error \(\delta\) such that:

\[
\frac{\pi}{4} + \frac{\delta}{2} = \frac{\pi V_\pi/4 + \Delta V_M}{V_\pi},
\]

therefore:

\[
\delta = 2\pi \frac{\Delta V_M}{V_\pi}.
\]

Thus, a variation \(\Delta V_M\) of the amplitude of \(v_M(t)\) induces a phase shift \(\delta\) between \(y_{1f}[n]\) and \(y_{2f}[n]\). Equations (39) and (40) can be rewritten as:

\[
y_{1f}[n] = \cos \left(\phi_{L1}(nT_S') + \frac{\delta}{2}\right), \quad (43)
\]

and

\[
y_{2f}[n] = \sin \left(\phi_{L1}(nT_S') - \frac{\delta}{2}\right). \quad (44)
\]

It can be shown that the Lissajous figure of the points \((y_{1f}[n], y_{2f}[n])\) is no longer a circle, instead, it is a rotated ellipse:

\[
y_{1f}^2[n] + y_{2f}^2[n] + 2y_{1f}[n]y_{2f}[n] \sin \delta = \cos^2 \delta. \quad (45)
\]

In order to calculate the angle of rotation \(\beta\) of the ellipse, a coordinate system transformation is carried out:

\[
y_{1f}[n] = y_{1f}'[n] \cos \beta - y_{2f}'[n] \sin \beta, \quad (46)
\]

and

\[
y_{2f}[n] = y_{1f}'[n] \sin \beta + y_{2f}'[n] \cos \beta, \quad (47)
\]

where \((y_{1f}'[n], y_{2f}'[n])\) are the ellipse points referred to the rotated coordinate system. Inserting (46) and (47) in (45), it can be shown that:

\[
2 \sin \delta \cos 2\beta = 0. \quad (48)
\]

However, there is a non-zero phase error \(\delta\) when (5) is not satisfied, thus:

\[
\beta = \frac{\pi}{4}, \quad (49)
\]

and

\[
y_{1f}'^2[n](1 + \sin \delta) + y_{2f}'^2[n](1 - \sin \delta) = \cos^2 \delta. \quad (50)
\]
In other words, when (5) is not satisfied, the Lissajous figure of the points \((y_{1f}[n], y_{2f}[n])\), which is shown in Fig. 8, is an ellipse with its principal axes rotated by \(\pi/4\) rad.

In order to obtain an equivalent expression of \(\delta\), the error signal \(E(\delta)\) is defined as the difference between the mean squared magnitudes of the vectors \((y_{1f}[n], y_{2f}[n])\) of the odd and even quadrants. A given point \((y_{1f}[n], y_{2f}[n])\) has a squared magnitude:

\[
r^2[n] = y_{1f}^2[n] + y_{2f}^2[n].
\]  

(51)

Due to the low value of \(f_A\) (20 Hz) with respect to the sampling frequency of \(x_1[n]\) and \(x_2[n]\) (120 kHz), it allows to take a high number of points \((x_{1f}[n], x_{2f}[n])\), specifically 120 \(\times 10^3/20 = 6 \times 10^3\) points, per each cycle of \(v_A(t)\). Therefore, (51) can be rewritten as:

\[
r^2(\theta) = y_{1f}^2(\theta) + y_{2f}^2(\theta),
\]  

(52)

where \(y_{1f}(\theta)\) and \(y_{2f}(\theta)\) satisfy (45) and are given by:

\[
y_{1f}(\theta) = \cos \left(\theta + \frac{\delta}{2}\right),
\]  

(53)

and

\[
y_{2f}(\theta) = \sin \left(\theta - \frac{\delta}{2}\right).
\]  

(54)

Equation (52) can be rewritten as:

\[
r^2(\theta) = 1 - \sin 2\theta \sin \delta,
\]  

(55)

and the mean squared magnitude can be approximated as:

\[
r^2_{\text{odd}} = \frac{2}{\pi} \left(\int_0^{\pi/2} r^2(\theta)d\theta + \int_{3\pi/2}^{2\pi} r^2(\theta)d\theta\right),
\]  

(56)

and

\[
r^2_{\text{even}} = \frac{2}{\pi} \left(\int_{\pi/2}^{\pi} r^2(\theta)d\theta + \int_{5\pi/2}^{3\pi/2} r^2(\theta)d\theta\right),
\]  

(57)

where \(r^2_{\text{odd}}\) and \(r^2_{\text{even}}\) are the mean squared magnitudes of the odd and even quadrants, respectively.

Therefore, the error signal \(E(\delta)\) is given by:

\[
E(\delta) = r^2_{\text{odd}} - r^2_{\text{even}}.
\]  

(58)

Replacing (55) in (56) and (57), we get:

\[
r^2_{\text{odd}} = 2 - \frac{4}{\pi} \sin \delta,
\]  

(59)

and

\[
r^2_{\text{even}} = 2 + \frac{4}{\pi} \sin \delta.
\]  

(60)

From (58), \(E(\delta)\) is

\[
E(\delta) = -\frac{8}{\pi} \sin \delta.
\]  

(61)

For the particular case when \(\delta \approx 0\), \(E(\delta)\) exhibits an approximate linear dependence with \(\delta\), as follows:

\[
E(\delta) \approx -\frac{8}{\pi} \delta.
\]  

(62)

Equations (61) and (62) are shown in Fig. 9. The control signal \(v_C\) is introduced to compensate \(\Delta v_M\) in such a way that \(V'_M\) reaches its ideal value \(V_\pi/4\). Thus, in closed-loop, \(V'_M\) is redefined as:

\[
V'_M = \frac{V_\pi}{4} + \Delta v_M + v_C,
\]  

(63)

and, by similarity with (42), \(\delta\) is now given by:

\[
\delta = 2\pi \frac{\Delta v_M + v_C}{V_\pi},
\]  

(64)

such that \(\delta = 0\) when \(\Delta v_M = -v_C\). Replacing (64) in (62), we get the new error expression:

\[
E = -\frac{16}{V_\pi} \Delta v_M + v_C.
\]  

(65)

The phase control process is shown in Fig. 10. \(v_C\) is obtained as the output of a PI block and it compensates the disturbance \(\Delta v_M\). However, (65) and, consequently, the phase control process assumes that \(\Delta v_M\) varies slowly in time in
such a way that the time delay introduced by the demodulation method is negligible. Under these considerations, the control loop is described in terms of Laplace transforms as:

$$E(s) = -\frac{16}{V_\pi} \left[ \Delta V_M(s) + \left( \frac{K_P + K_I}{s} \right) E(s) \right],$$

where $K_P$ and $K_I$ are the proportional and integral gains, respectively. Equation (66) can be rewritten as:

$$E(s) = -\frac{A_1 s}{s + A_2} \Delta V_M(s),$$

with:

$$A_1 = \frac{16}{V_\pi} \frac{K_P}{1 + 16 \frac{V_P}{V_\pi}},$$

and

$$A_2 = \frac{16 K_I}{V_\pi} \frac{K_P}{1 + 16 \frac{V_P}{V_\pi}}.$$  

From (67), and since $K_P$ and $K_I$ take positive values, the control loop has a real single pole at $s = -A_2 < 0$. Therefore, the system is locally stable. To obtain the value of $E(\delta)$ in steady-state, the slow variation of $\Delta V_M(s)$ is considered to model it as an unknown constant $A_3$. Therefore, applying the final value theorem to $E(s)$, we get:

$$\lim_{t \to \infty} E(\delta(t)) = \lim_{s \to 0} sE(s) = \lim_{s \to 0} \frac{-A_1 A_3 s}{s + A_2} = 0.$$  

This shows that $E(\delta)$ can be used to obtain a control signal $v_C$, which compensates $\Delta V_M$. $E(\delta) = 0$ is a locally stable equilibrium point. In other words, $\delta = 0$ is a stable operation point. For example, if $\Delta V_M > 0$, when $-\pi/2 < \delta < 0$, $E(\delta) > 0$ so that the controller output $v_C$ ($v_C < -\Delta V_M$) increases until $\delta \approx 0$. On the other hand, when $0 < \delta < \pi/2$, then $E(\delta) < 0$, thus $v_C$ ($v_C > -\Delta V_M$) decreases in order to guarantee $\delta \approx 0$. A similar behavior occurs when $\Delta V_M < 0$.

The parameters $K_P$ and $K_I$ were manually tuned. In this way, $v_A(t)$ enables the operation of optical phase control that is independent of $v_{PZT}(t)$. This control process provides robustness to the system since it is able to work in the presence of electronic noise, variations in the amplifier gain and the actual value of $V_\pi$. Taking into account (21), the differential-cross-multiplication method makes the system immune to variations in $AV$, i.e., variations in fringe visibility or in photodiode responsivity. In addition, from (22), it can be seen that the optical phase demodulation process is immune to interferometric signal fading caused by random variations in $\phi_Q(t)$. A block diagram of the signal processing method, optical phase demodulation process and phase control, is shown in Fig. 11.

IV. EXPERIMENTAL RESULTS

The OV is aimed at the measurement of the optical phase shift $\phi_{PZT}(t)$ caused by the vibration of the MAFPA when subjected to a sinusoidal voltage $v_{PZT}(t)$. Therefore, the input-output relationship of two MAFPAs, named as FPA-XY and XY05B, were characterized in this work for the band of frequencies between 500 Hz and 15 kHz. The results were validated by using the Signal Coincidence Method (SCM) described in ISO 6063-41 [38, 39].

The stability of the optical phase control was initially tested by setting the amplitude of $v_M(t)$ to zero and operating the OV with no feedback (open loop). For this case, the faded photodetected signal is shown in Fig. 12(a), which simply consists of a cosine function modulated by the sum of the phases induced by $v_A(t)$ and $v_{PZT}(t)$, plus the quasi-static phase shift $\phi_Q(t)$. In the inset, the phase shift induced by the vibration of the piezoelectric transducer is shown. When feedback is enabled, $\phi_{PZT}(t)$ is successfully demodulated as shown in Fig. 12(b), which demonstrates the immunity of the
OV to the photodetector signal fading induced by random variations in $\phi_Q(t)$.

To test the measurement precision of the OV, 200 measurements were carried out, each one consisting of 1600 cycles of the auxiliary modulating signal, over different periods within a day, for various frequencies in the operation band of the two MAFPAs. The measured magnitude frequency responses were compared with those obtained using the SCM as described in the following subsections. To evaluate the performance under severe conditions, a 4 V amplitude, 6 kHz sinusoidal voltage was applied to the XY05B. This choice of frequency constitutes an interesting case study since it is located right after the first mechanical resonance of the actuator, although small, it still presents a modulation index high enough for the SCM to be applied. By applying the SCM an actuator sensitivity of 0.0115 rad/V was obtained, and so the corresponding optical phase shift is 0.046 rad. In comparison, the OV measures a phase shift of 0.044 rad and therefore an actuator sensitivity of 0.011 rad/V. Therefore the measurement error of the OV relative to the SCM is 4.35%. Fig. 13 shows the demodulated optical phase shift along with the sinusoidal voltage (converted to the phase domain) applied to XY05B. As the discrepancy is small (even under such a severe condition), this demonstrates the accuracy and the precision of the OV. To evaluate the noise performance of the OV, the Minimum Detectable Phase Shift (MDPS) criterion was used [40]. The amplitude of $v_{PZT}(t)$ was slowly decreased from its initial value (4 V) until the demodulated output signal cannot be distinguished from noise. At this point, we can say that the phase signal of interest and the phase noise (demodulated output from the closed loop vibrometer when no signal is applied to the actuator) have the same level, which by definition corresponds to the MDPS. The measured MDPS value of the OV was 0.002 rad.

A. Frequency Response of the OV System

For the frequency response measurements of the OV system to be independent of the MAFPA used in the sensor arm of the interferometer, a third signal, $v_i(t)$, is applied to the Pockels cell, and $v_{PZT}(t) = 0$, which implies that the MAFPA now behaves as a fixed mirror (see Fig. 2). To obtain the system frequency response, the frequency of the sinusoidal input $v_i(t)$ was varied between 500 Hz and 15 kHz and its amplitude set to $V_i/8$. Instead of (4), now $v_p(t)$ is given by:

$$ v_p(t) = v_M(t) + v_A(t) + v_i(t), $$

being

$$ v_i(t) = V_i \sin(2\pi f_i t), $$

where $V_i$ and $f_i$ are the amplitude and frequency of $v_i(t)$, respectively. Equation (3) can be rewritten as:

$$ \phi_p(t) = \frac{\pi}{V} (v_M(t) + v_A(t) + v_i(t)). $$

Since $\phi_{PZT}(t) = 0$ the total optical phase shift is:

$$ \Phi(t) = \phi_p(t) + \phi_Q(t). $$

Using (7) and (73),

$$ \Phi(t) = \phi_p(t) + \frac{v_i(t)}{V} + \phi_Q(t). $$

So, by varying $f_i$ between 500 Hz and 15 kHz, the system demodulates the optical phase induced by $v_i(t)$, and, in this way, it is able to provide its input-output characteristic, given in rad/V, for that frequency band. The normalized system frequency response is shown in Fig. 14. The $-3$ dB bandwidth is approximately 13.8 kHz.

The OV system linearity was verified by varying the amplitude for several frequencies between 500 Hz and 14.3 kHz and demodulating the induced phases. In particular, for the frequencies 2 kHz and 7 kHz, the results are shown in Fig. 15 in terms of the $v_i(t)$ and the amplitude of the associated induced phase shift.
B. Frequency response of the transducers

As an application of the OV, \( v_{PZT}(t) \) was applied only to the piezoceramic \( P_{1X} \) (Fig. 1(a)) and the optical phase shift \( \phi_{PZT}(t) \) due to the direct vibration of FPA-XY was measured for frequencies between 500 Hz and 10 kHz. Since the OV operation starts when (5) is not satisfied, only the measurements obtained for \( \delta \approx 0 \) were considered. These measurements were saved for post-processing when the error \( |E(\delta)| \leq 0.03\% \). The results were validated by the SCM and are shown in Fig. 14. Once the phase shift \( \phi_{PZT}(t) \) is known, the mechanical displacement is calculated using \( \Delta L(t) = \lambda \phi_{PZT}(t)/4\pi \), where \( \lambda = 658 \) nm, as shown in the vertical axis on the right side of Fig. 16. The measured resonance frequencies were the same for the OV as well as for the SCM: 1.51, 3.1, 5.1, and 7.0 kHz.

For the XY05B, its magnitude response due to the direct movement of the piezoceramic \( P_{2X} \) (Fig. 1(b)) when a voltage \( v_{PZT}(t) \) is applied to it was measured for the frequencies between 500 Hz and 14.5 kHz. The results were validated by the SCM as shown in Fig. 17. As can be observed from the magnitude response, the ratio between the demodulated optical phase shift \( \phi_{PZT}(t) \) and the applied voltage \( v_{PZT}(t) \) is much smaller than the values obtained for FPA-XY. In particular, the maximum magnitude measured by the system for XY05B was 0.38 rad/V, while for FPA-XY is 13.58 rad/V. This makes the application of the SCM difficult since very high voltages are required in order to obtain a suitable interferometric signal. Therefore, the SCM was only applied around the resonance frequencies, where the input-output relationship takes higher values. The measurements show that the resonance frequencies are, approximately, 5.5, 7.1, 9.8, and 12.9 kHz.

V. Conclusions

An innovative optical phase detection method, applied to the measurement of nanometric movements has been proposed, modeled, and successfully verified. The method uses a simple, slightly modified, homodyne Michelson interferometer to accommodate a Pockels cell phase shifter utilizing a low-cost laser diode. The Pockels cell has been modulated with principal modulating signal such that two phase quadrature signals can be generated from a single interferometric output using the interleaving action. This made it possible to perform accurate measurement of the optical phase, even under high modulation index, by using the technique of differential-cross multiplication.

However, the half-wave voltage of a Pockels cell varies slightly with temperature, thereby moving the interferometer away from quadrature. PI control system was used to stabilize...
the interferometer, which proved to be robust and locally stable; the demodulation method works under electronic noise and undesirable variations of the Pockels cell half-wave voltage and amplifier gain. A criterion for the quadrature error has been proposed, based on the difference between the lengths of the main axes of the ellipse generated by the signals to be taken to the phase quadrature. This leads to an expression that exhibits linear behavior relative to the phase error around the desired operation point.

In addition, an auxiliary modulating signal was introduced in order to: a) allow the system to set the frequency of the auxiliary signal. The only two restrictions are that, first, such frequency should be distant from the PZT operation bandwidth in order to avoid aliasing in the pre-control filtering. Second, the frequency has to take low values (around 20 Hz) to maximize the quantity of available points for the control process; b) carry out a pre-control filtering and, after that, to develop an independent-input optical phase control process to keep the interferometer in quadrature; c) obtain the system frequency response, by applying the input sinusoidal signal to the Pockels cell with a fixed mirror in the sensor arm of the interferometer, that is, with no signal applied to the MAFPA.

The OV response has bandwidth of 13.8 kHz, and the ratio between the demodulated optical phase and the input amplitude is constant. The new demodulation method is immune to signal fading. It is also self-consistent, that is, the term $AV$ is constant. The new demodulation method is immune to variation of parameters of the interferometer, that is, with no signal applied to the MAFPA.

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