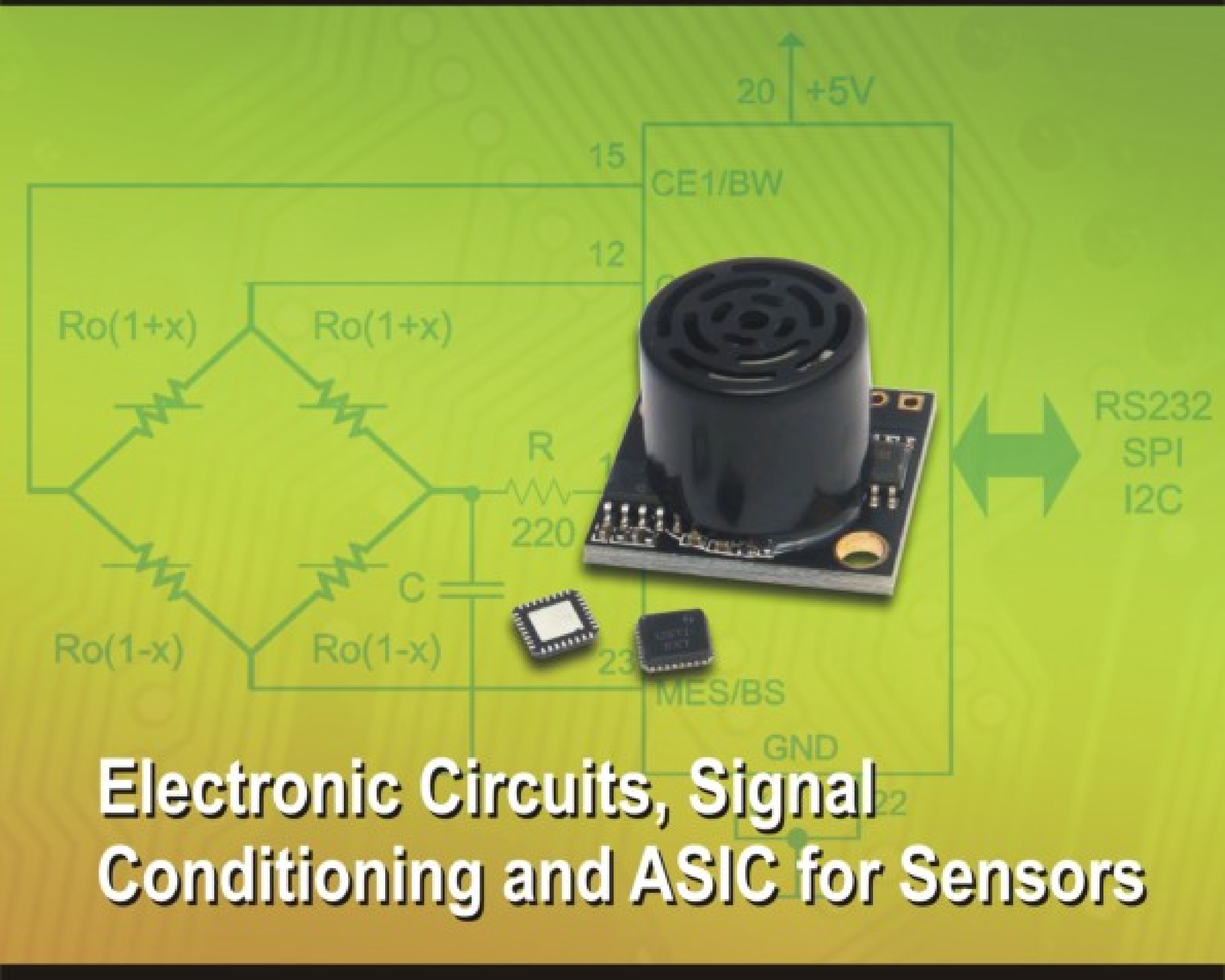
ISSN 1726-5479

SENSORS & TRANSDUCERS

6/12
vol. 141



Electronic Circuits, Signal Conditioning and ASIC for Sensors

International Frequency Sensor Association Publishing
IFSA

Sensors & Transducers

Volume 141, Issue 6,
June 2012
www.sensorsportal.com
ISSN 1726-5479

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Sensors & Transducers Journal

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Volume 141
Issue 6
June 2012

www.sensorsportal.com

ISSN 1726-5479

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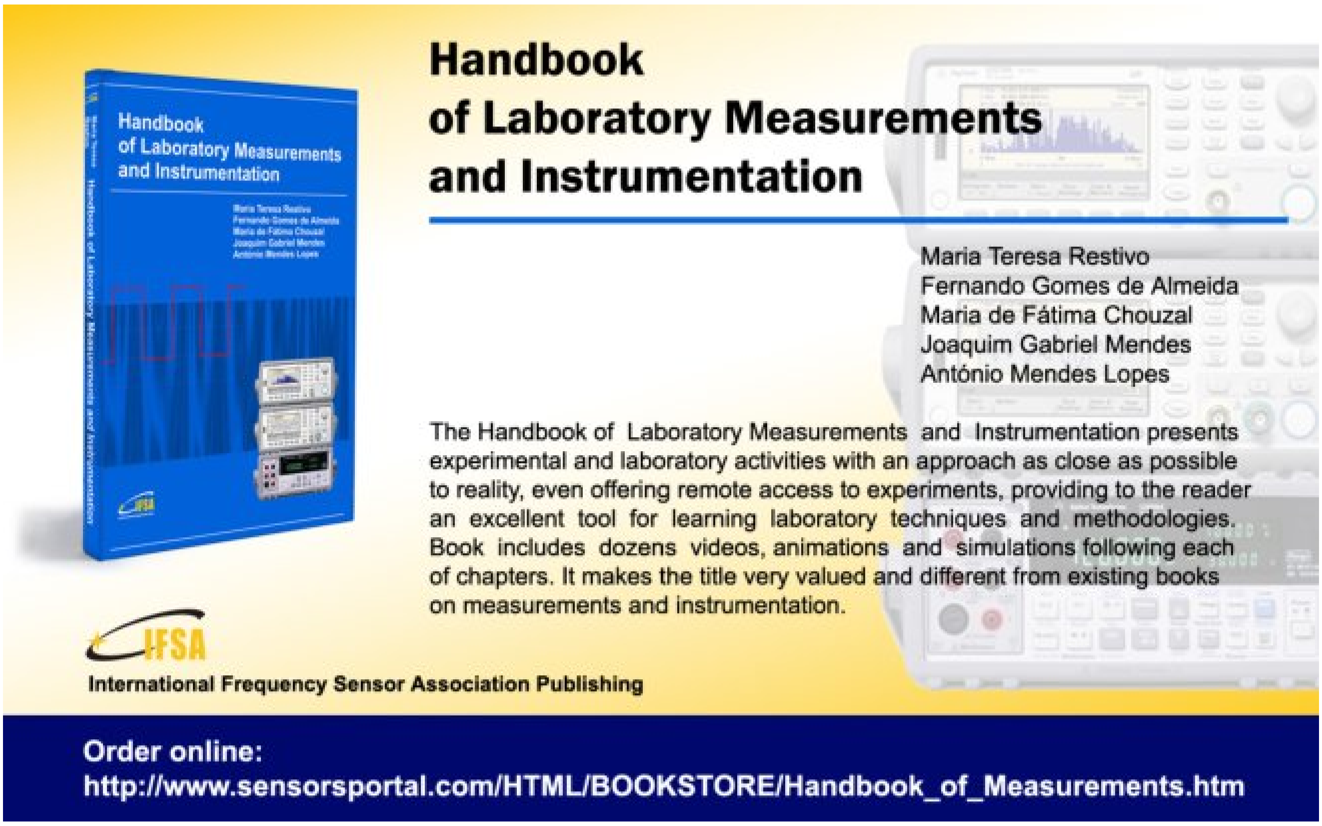

Sensors & Transducers Journal, Vol. 141, Issue 6, June 2012, pp. 71-91

Sensors & Transducers

ISSN 1726-5479

http://www.sensorsportal.com

Synchronization Based SAW Sensor Using Delay Line Coupled Dual Oscillator Phase Dynamics

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Received: 6 June 2012 /Accepted: 27 June 2012 /Published: 30 June 2012

Abstract: This paper presents an analysis of synchronization between two coupled nonlinear surface acoustic wave (SAW) delay line oscillators where coupling is provided via another linear phase SAW delay line. The analysis is aimed at determining the sensitivity of synchronization frequency for perturbations in the coupling SAW delay line, and then to explore whether the coupling SAW device can be used for making a better chemical sensor in comparison to the usual polymer-coated SAW delay line oscillator sensor. Two nearly identical free running SAW delay line oscillators with stable amplitude and weak coupling conditions are considered. The dynamics of coupled system is analyzed for small perturbations in limit cycles under phase approximation. The system represents nonlinear dynamics of a generic system of two phase coupled self-sustained oscillators with delayed-feedback and delayed coupling. The relations for synchronization frequency and sensitivity for delay perturbations are obtained. Considering the noise suppression characteristics and high sensitivity regions of synchronization it is found that the coupled SAW oscillators in synchronized states have potential for making high performance SAW sensors. *Copyright © 2012 IFSA.*

Keywords: Coupled SAW oscillators, Synchronization, SAW sensors, Coupled oscillators with delayed feedback and delayed coupling.

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1. Introduction

Synchronization is a particular state of coupled nonlinear oscillators wherein individual oscillators adjust their rhythms (frequency or phase) so that asymptotically their motion evolves to a common mode of oscillation. The essential components needed for synchronization to occur are that the oscillators are nonlinear and autonomous (having independent source of energy for self-sustained dynamics) and they interact via some mechanism through which energy flow can occur [1]. In case of limit cycle oscillators the phenomenon is often referred to as frequency entrainment or phase locking [2]. Modeling a system of interacting oscillators and analyzing their dynamics provides basis for understanding a variety of synchronization phenomena in nature [1, 2]. For example, circadian rhythm in living organisms is synchronization of internal biological clocks by environmental periodic signals such as luminance and temperature associated with day-night rotation of earth; cardiac cycle that yields periodic contraction of heart is synchronization of heart pacemaker cells with mean fields of regulatory nerves; locomotion in animals is synchronization of cellular activities under influence of control nerve signals [3, 4]. The understanding of dynamics of nonlinear oscillators and coupled oscillators system has provided basis for several novel engineering applications such as clock synchronization in communication networks [5], robotics control [6], encoding of information using chaotic signals and chaos synchronization for secure communication [7, 8], phase locking of relativistic magnetrons [9], synchronization of chaotic lasers and chemical oscillators [10], data mining [11]. In this work we aimed to analyze coupled phase dynamics of two nonlinear surface acoustic wave (SAW) oscillators and to explore the possibility of using synchronization advantageously for sensing applications.

The polymer coated SAW oscillators make an important class of chemical vapor sensors [12-15]; particularly, for making sensor arrays based electronic noses [14-17]. The SAW vapor sensors have been extensively investigated and developed for chemical sensing applications [17-20]. In most studies on SAW chemical sensors the focus has been on the polymer development and selection [13-15, 20], SAW device type selection and frequency of operation [12, 13]. The SAW platform design also has been the subject of study in some publications with objectives to improve sensitivity and detectivity [22-24].

In the present study we seek ways for improving performance of the SAW oscillator sensors for a given state-of-the-art SAW oscillator technology. We consider two nearly identical nonlinear SAW delay line oscillators and couple their outputs through another linear phase SAW delay line device. The analysis presents first a theoretical analysis of synchronized states, and then considers the coupling SAW device as sensing platform. The sensitivity of synchronization frequency and their fluctuations (noise) are then calculated to assess the potentiality of this configuration for chemical vapor sensing. The dependencies of synchronized states on the coupling parameters and on delay perturbations are analyzed. The analysis demonstrates that the coupled SAW oscillator system can make a better sensor by adjusting their operating conditions so that the two autonomous SAW oscillators remain pulled to the synchronized state while the coupling SAW device generates sensor output.

2. Synchronization of Linearly Coupled Nonlinear SAW Oscillators

2.1. Nonlinear SAW Delay Line Oscillator

Autonomous means the system has its own independent source of energy and its state of oscillation is self-sustained by adjusting amplitude and phase dynamics. We consider a SAW feedback oscillator whose frequency control comes from the phase shift across a simple SAW delay line device as shown in Fig. 1. The output voltage from the amplifier $v(t)$ is input to the SAW delay line. The voltage transfer function of the SAW device $H(\omega) = |H(\omega)|e^{-j\omega\tau}$ determines the input to the amplifier $v_{in}(t) = H(\omega)v(t)$ where $\omega = 2\pi f$ denotes the angular frequency of the oscillation and τ denotes the time delay. The $|H(\omega)|$ determines the amplitude attenuation and $\Phi_{SAW} = -\omega\tau$ determines the phase shift of the feedback signal in going from the amplifier output to its input.

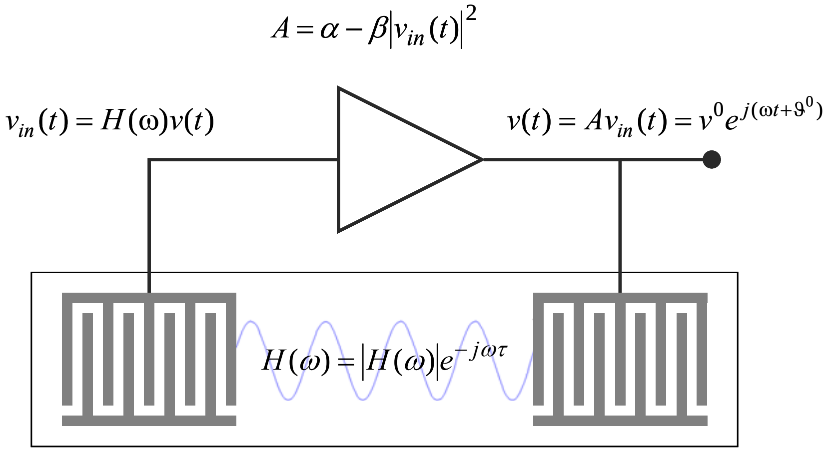

Fig. 1. SAW delay line oscillator supported by loop amplifier with cubic nonlinearity.

The output voltage from the amplifier $v(t)$ is input to the SAW delay line. The voltage transfer function of the SAW device $H(\omega) = |H(\omega)|e^{-j\omega\tau}$ determines the input to the amplifier $v_{in}(t) = H(\omega)v(t)$ where $\omega = 2\pi f$ denotes the angular frequency of the oscillation and τ denotes the time delay. The $|H(\omega)|$ determines the amplitude attenuation and $\Phi_{SAW} = -\omega\tau$ determines the phase shift of the feedback signal in going from the amplifier output to its input. The voltage gain of the amplifier is assumed to be nonlinear having cubic nonlinearity of the form $A = \alpha - \beta|v_{in}(t)|^2$; that is, $v(t) = \alpha v_{in}(t) - \beta|v_{in}(t)^2|v_{in}(t)$ where α and β define linear and nonlinear parts of the voltage gain. The self-sustained oscillations occur at that frequency for which the Berkhausen criteria for limit cycle oscillations: close-loop gain $A|H(\omega)| \geq 1$ and phase shift $\Phi_{loop} = \Phi_{SAW} + \Phi_A = 2\pi n$, are satisfied ($\Phi_A$ denotes the phase shift due to amplifier, n is an integer).

The output voltage from the amplifier $v(t)$ is input to the SAW delay line. The voltage transfer function of the SAW device $H(\omega) = |H(\omega)|e^{-j\omega\tau}$ determines the input to the amplifier $v_{in}(t) = H(\omega)v(t)$ where $\omega = 2\pi f$ denotes the angular frequency of the oscillation and τ denotes the time delay. The $|H(\omega)|$ determines the amplitude attenuation and $\Phi_{SAW} = -\omega\tau$ determines the phase shift of the feedback signal in going from the amplifier output to its input. The voltage gain of the amplifier is assumed to be nonlinear having cubic nonlinearity of the form $A = \alpha - \beta|v_{in}(t)|^2$; that is, $v(t) = \alpha v_{in}(t) - \beta|v_{in}(t)^2|v_{in}(t)$ where α and β define linear and nonlinear parts of the voltage gain. The self-sustained oscillations occur at that frequency for which the Berkhausen criteria for limit cycle oscillations: close-loop gain $A|H(\omega)| \geq 1$ and phase shift $\Phi_{loop} = \Phi_{SAW} + \Phi_A = 2\pi n$, are satisfied ($\Phi_A$ denotes the phase shift due to amplifier, n is an integer).

2.2. SAW Delay Line Coupled Dual SAW Nonlinear Oscillators

Fig. 2 shows two identical self-sustained SAW delay line oscillators coupled by a simple SAW delay line. The coupling SAW device is assumed to have the same centre frequency as that of the devices in the feedback loop of the two oscillators. The passband width and insertion loss of this device are however assumed to be larger so that it provides a flat pass and weak coupling to any frequency of oscillation supported by the two coupled oscillators. The strong coupling situation is avoided here as it may lead to instability of the oscillators system, and equations of motion can not be approximated as being pure phase coupling. Considering perturbations to amplitudes also will make the whole treatment quite complex. The simplifying assumptions of weak coupling are important to our present interest in phase synchronization and its potential use for making sensors.

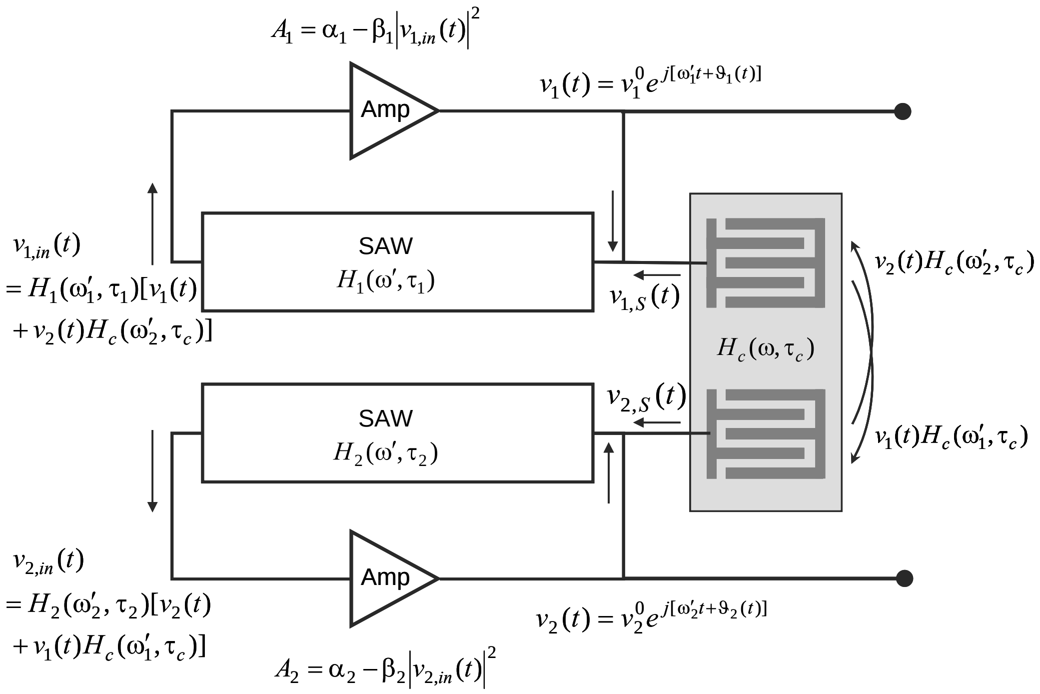

Fig. 2. Configuration for mutual interaction between two nonlinear SAW delay line oscillators coupled by a simple linear phase SAW delay line. The primed frequencies ω'_1 and ω'_2 indicate instantaneous values.

All the three SAW devices are assumed to simple delay lines with linear phase transfer characteristics. The nonlinearity comes from the loop amplifiers which are assumed to have cubic nonlinear response to input characterized by amplitude dependent gain of the form $A = \alpha - \beta |v_{in}(t)|^2$. Thus, the system considered here represents a case of *'linearly-coupled nonlinear oscillators with delayed feedback and coupling'*. The width of passband of the SAW coupling device is assumed large enough so that its amplitude transfer characteristic (insertion loss) can be taken to be constant over the frequency range of self-sustained oscillations of both interacting oscillators. This is easily achieved by adjusting the number electrode pairs in the interdigital transducers of the coupling SAW device [25]. All the quantities are labelled with subscripts 1 or 2 to designate which oscillator they pertain to. The quantities associated with the coupling device are labelled with subscript *c*.

Prior to coupling, each oscillator is assumed to be on the limit cycle having stable sinusoidal oscillations of the form $v(t) = v^0 e^{j(\omega t + \vartheta^0)}$ where v^0 denotes time-invariant amplitude, ω initial frequency and ϑ^0 initial phase. The assumption of weak coupling allows us to describe mutual interactions between them purely in terms of phase flow via the coupling SAW delay line. The influence of coupling on phases of the two oscillators can be expressed by attributing to them a phase function of the form $\varphi(t) = \omega' t + \vartheta^0$ where ω' denotes their instantaneous frequencies. Here, the phase perturbation due to coupling is assumed not to disturb the sinusoidal character of the limit cycle oscillations. The theory of mutual synchronization of weakly coupled phase oscillators has been described in detail in [2, Ch. 7, 8]. Here, we largely adapt that description with the difference that in the present case we have a specific type of oscillators system whose oscillations are sustained by delayed self-feedback and interactions occur via delayed coupling.

2.3. Coupled Phase Dynamics

2.3.1. Phase Equations of Oscillation under Weak Coupling

Considering that both oscillators are autonomous systems having stable limit cycles with uncoupled state frequencies ω_1 and ω_2, and the coupling is weak so that the amplitude perturbations are negligible, we can obtain the equations for phase dynamics as outlined in [2, Ch. 8]. Each oscillator in the uncoupled stable state is defined by the following input-output relations (referring to oscillator 1):

$$v_1(t) = v_1^0 e^{j(\omega_1 t + \vartheta_1^0)} \tag{1a}$$

$$v_{1,in} = H_1(\omega_1, \tau_1) v_1(t) \tag{1b}$$

$$A_1(t) = \alpha_1 - \beta_1 \left| v_{1,in}(t) \right|^2 \tag{1c}$$

$$v_1(t) = A_1 v_{1,in}(t) \tag{1d}$$

The equation of motion for this oscillator can be obtained by differentiating Eq. (1d) with respect to time and making substitutions for the associated differentials from Eqs. (1a)-(1c) as follows:

$$\begin{aligned} \dot{v}_1(t) &= \dot{A}_1(t) v_{1,in}(t) + A_1(t) \dot{v}_{1,in}(t) \\ &= [\dot{A}_1(t) + j\omega_1 A_1(t)] H_1(\omega_1, \tau_1) v_1(t), \\ &\cong j\omega_1 A_1 H_1(\omega_1, \tau_1) v_1(t) \end{aligned} \tag{2}$$

where $(\dot{.}) \equiv d/dt$ and the last equality has been reached by considering that the amplitude is stable, that is, $\left| v_{1,in}(t) \right| \cong \text{constant, hence, } \dot{A}_1(t) \cong 0$. A similar equation can be written for the oscillator 2 in uncoupled equilibrium state as

$$\dot{v}_2(t) \cong j\omega_2 A_2 H_2(\omega_2, \tau_2) v_2(t). \tag{3}$$

The coupling of their outputs through the SAW delay line having voltage transfer function $H_c(\omega, \tau_c) = \left| H_c(\omega, \tau_c) \right| e^{-j\omega\tau_c}$ allows a fraction of output from oscillator 2 to be fed into the input of feedback SAW device in oscillator 1, and vice versa, as shown in Fig. 2. In view of weak coupling assumption the effects of this signal flow on the amplitudes are small. The dominant effect is on the phases of the oscillations. The stable limit cycle of an autonomous dynamical system has the property of being invariant under time shift $t \to t - \Delta t$. This is equivalent to the limit cycle invariance under phase shifts $\Delta\phi = \omega\Delta t$ [2, section 7.1.1]. Therefore, the phase shifts of oscillator 2 due to signal flow from the oscillator 1 and the phase shift of the oscillator 1 due to signal flow from the oscillator 2 are assumed not to alter the limit cycles of the respective oscillators. The equations of motion [Eqs. (1) through (3)] will be modified according to the nodal values of the various voltages involved. Fig. 2 indicates appropriate instantaneous values of the respective oscillators. As mentioned above, it has been assumed here that both the oscillators retain sinusoidal character during the course of phase adjustment for synchronization. In the following we use unprimed symbols for denoting unperturbed frequencies and primed symbols for the instantaneous values. Thus,

$$\Delta = \omega_2 - \omega_1 \tag{4}$$

defines the detuning parameter at onset of coupling, and

$$\Delta' = \omega_2' - \omega_1' \tag{5}$$

represents the difference of instantaneous frequencies during the course of synchronization.

At some arbitrary instant t after the onset of coupling interactions the signals at the input of SAW feedback devices are

$$v_{1,S}(t) = v_1^0 e^{j(\omega_1' t + \vartheta_1^0)} + H_c(\omega_2', \tau_c) v_2^0 e^{j(\omega_2' t + \vartheta_2^0)} \text{ with } \omega_1' = \omega_1'\big|_t \text{ and } \omega_2' = \omega_2'\big|_{t-\tau_c} \tag{6a}$$

$$v_{2,S}(t) = v_2^0 e^{j(\omega_2' t + \vartheta_2^0)} + H_c(\omega_1', \tau_c) v_1^0 e^{j(\omega_1' t + \vartheta_1^0)} \text{ with } \omega_2' = \omega_2'\big|_t \text{ and } \omega_1' = \omega_1'\big|_{t-\tau_c} \tag{6b}$$

Note that the second terms in these equations represent the signal from the other oscillator transferred via coupling SAW delay line. Therefore, in these terms the relevant instantaneous value of the frequency of the other oscillator at $t - \tau_c$ must be used. The signal at the output of SAW feedback device (or at the input of nonlinear amplifier) appears τ_1 or τ_2 later. Therefore, input signal to the amplifier at time t is that which was at time $t - \tau_1$ or $t - \tau_2$ at the input of the SAW feedback device. This is accounted for by the complex transfer function of these devices, and one can write

$$v_{1,in}(t) = H_1(\omega_{1,2}', \tau_1) v_{1,S}(t) \quad \text{with } \omega_1' = \omega_1'\big|_{t-\tau_1} \text{ and } \omega_2' = \omega_2'\big|_{t-\tau_1-\tau_c} \tag{7a}$$

$$v_{2,in}(t) = H_2(\omega_{1,2}', \tau_2) v_{2,S}(t) \text{ with } \omega_1' = \omega_1'\big|_{t-\tau_1} \text{ and } \omega_2' = \omega_2'\big|_{t-\tau_1-\tau_c} \tag{7b}$$

where $\omega_{1,2}'$ takes appropriate instantaneous values depending on the frequency components in $v_{1,S}(t)$ and $v_{2,S}(t)$ as appearing in Eqs. (6a) and (6b). That is, $\omega_1' = \omega_1'\big|_{t-\tau_1}$ and $\omega_2' = \omega_2'\big|_{t-\tau_1-\tau_c}$ in (7a), and $\omega_2' = \omega_2'\big|_{t-\tau_2}$ and $\omega_1' = \omega_1'\big|_{t-\tau_2-\tau_c}$ in (7b).

Applying Eqs. (2) and (3) under weak coupling and fixed amplitude conditions one can obtain the equations of motion as

$$\begin{aligned} \dot{v}_1(t) &\cong A_1 \dot{v}_{1,in}(t) \\ &= j\omega_1' A_1 H_1(\omega_1', \tau_1) v_1(t) + j\omega_2' A_1 H_1(\omega_2', \tau_1) H_c(\omega_2', \tau_c) v_2(t) \end{aligned} \tag{8a}$$

$$\begin{aligned}\dot{v}_2(t) &\cong A_2\dot{v}_{2,in}(t)\\ &= j\omega_2' A_2 H_2(\omega_2',\tau_2)v_2(t) + j\omega_1' A_2 H_2(\omega_1',\tau_2)H_c(\omega_1',\tau_c)v_1(t)\end{aligned} \tag{8b}$$

with ω_1' and ω_2' as specified in Eqs. (7a) and (7b) respectively.

If we denote the phases of the individual oscillators in the coupled state by $\varphi_1(t)$ and $\varphi_2(t)$ then we can alternately write for their outputs,

$$v_1(t) = v_1^0 e^{j\varphi_1(t)} \tag{9a}$$

$$v_2(t) = v_2^0 e^{j\varphi_2(t)} \tag{9b}$$

$$\dot{v}_1(t) = j\dot{\varphi}_1(t)v_1(t) \tag{9c}$$

$$\dot{v}_2(t) = j\dot{\varphi}_2(t)v_2(t) \tag{9d}$$

with

$$\varphi_1(t) = \omega_1' t + \vartheta_1^0 \tag{10a}$$

$$\varphi_2(t) = \omega_2' t + \vartheta_2^0 \tag{10b}$$

defined by instantaneous values of ω_1' and ω_2' at instant t. Therefore, by using Eq. (9c) on the left hand side of Eq. (8a) and after a little manipulation we can write for the phase equation of motion for the oscillator 1 as

$$\dot{\varphi}_1(t) = \left[\omega_1' + K_1 e^{j[\varphi_2(t)-\varphi_1(t)]}\right] A_1 H_1(\omega_1',\tau_1) \tag{11a}$$

With

$$K_1 = \omega_2' \frac{H_1(\omega_2',\tau_1)H_c(\omega_2',\tau_c)v_2^0}{H_1(\omega_1',\tau_1)v_1^0} \text{ with } \omega_1' = \omega_1'\big|_{t-\tau_1} \text{ and } \omega_2' = \omega_2'\big|_{t-\tau_1-\tau_c}\ . \tag{11b}$$

Similarly, by using Eq. (9d) in Eq. (8b) one can obtain the phase equation of motion for the oscillator 2 as:

$$\dot{\varphi}_2(t) = \left[\omega_2' + K_2 e^{-j[\varphi_2(t)-\varphi_1(t)]}\right] A_2 H_2(\omega_1',\tau_2) \tag{11c}$$

with

$$K_2 = \omega_1' \frac{H_2(\omega_1',\tau_2)H_c(\omega_1',\tau_c)v_1^0}{H_2(\omega_2',\tau_2)v_2^0} \text{ with } \omega_2' = \omega_2'\big|_{t-\tau_2} \text{ and } \omega_1' = \omega_1'\big|_{t-\tau_2-\tau_c}\ . \tag{11d}$$

These equations are in general form derived for an arbitrary instant t during the course of mutual phase interaction between oscillators. The factors K_1 and K_2 define the strength of interaction. The expression for these factors can be simplified by using the following characteristics of the coupled oscillators. Since they are assumed to be nominally identical, therefore $\nu_1^0 \approx \nu_2^0$. The feedback SAW delay lines have flat amplitude response so that $|H_1(\omega_2', \tau_1)| \approx |H_1(\omega_1', \tau_1)| = H_1$ (say) and $|H_2(\omega_1', \tau_2)| \approx |H_2(\omega_2', \tau_2)| = H_2$ (say). The close-loop conditions (Berkhausen criteria) for gain and phase for stable sinusoidal oscillations are always closely maintained so that $A_1 H_1(\omega_1', \tau_1) \approx A_2 H_2(\omega_2', \tau_2) \approx 1$, or $A_1 H_1 = A_2 H_2 \approx 1$ and $\omega_1' \tau_1 \approx 2\pi n$, $\omega_2' \tau_2 \approx 2\pi m$ with n and m being integers .The coupling SAW delay line has flat amplitude transfer over frequencies of interest so that $H_c(\omega_2', \tau_c) = |H_c(\omega_2', \tau_c)| e^{-j\omega_2' \tau_c}$ and $H_c(\omega_1', \tau_c) = |H_c(\omega_1', \tau_c)| e^{-j\omega_1' \tau_c}$ with $|H_c(\omega_2', \tau_c)| \approx |H_c(\omega_1', \tau_c)| = H_c$ (say). Thus, Eqs. (11b) and (11d) become

$$K_1 = \omega_2' H_c e^{j[\omega_1' \tau_1 - \omega_2' (\tau_1 + \tau_c)]} \tag{12a}$$

$$K_2 = \omega_1' H_c e^{j[\omega_2' \tau_2 - \omega_1' (\tau_2 + \tau_c)]} \tag{12b}$$

and the phase equations (11a) and (11c) can be written as

$$\dot{\varphi}_1(t) = \omega_1' + \omega_2' H_c e^{j[\{\varphi_2(t) - \omega_2'(\tau_1 + \tau_c)\} - \{\varphi_1(t) - \omega_1' \tau_1\}]} \tag{13a}$$

$$\dot{\varphi}_2(t) = \omega_2' + \omega_1' H_c e^{j[\{\varphi_1(t) - \omega_1'(\tau_2 + \tau_c)\} - \{\varphi_2(t) - \omega_2' \tau_2\}]}. \tag{13b}$$

Recall that the values of ω_1' and ω_2' in these equations are as defined in Eqs. (11b) and (11d).

Let us consider the factor $\{\varphi_2(t) - \omega_2'(\tau_1 + \tau_c)\}$ in (13a). In this: the $\varphi_2(t) = \omega_2' t + \vartheta_2^0$ as defined by (10b) represents the phase of oscillator 2 at time t where $\omega_2' = \omega_2'|_{t=t}$ is the instantaneous frequency at time t which has resulted from the coupling interactions in time interval $(\tau_1 + \tau_c)$; and $\omega_2'(\tau_1 + \tau_c)$ with $\omega_2' = \omega_2'|_{t-\tau_1-\tau_c}$ represents the phase change of the oscillator 2 in time interval $[t,\ t - \tau_1 - \tau_c]$ had it continued oscillating at $\omega_2' = \omega_2'|_{t-\tau_1-\tau_c}$. But due to interactions the instantaneous frequency of this oscillator has changed from $\omega_2' = \omega_2'|_{t-\tau_1-\tau_c}$ to $\omega_2' = \omega_2'|_{t=t}$. Therefore, $\{\varphi_2(t) - \omega_2'(\tau_1 + \tau_c)\}$ can be interpreted as $\varphi_2(t - \tau_1 - \tau_c)$ with instantaneous frequency $\omega_2' = \omega_2'|_{t=t}$. Using similar arguments for the other phase difference terms in Eqs. (13a) and (13b) we can rewrite them as

$$\dot{\varphi}_1(t) = \omega_1' + \omega_2' H_c e^{j[\varphi_2(t - \tau_1 - \tau_c) - \{\varphi_1(t - \tau_1)]} \tag{14a}$$

$$\dot{\varphi}_2(t) = \omega_2' + \omega_1' H_c e^{j[\varphi_1(t - \tau_2 - \tau_c) - \varphi_2(t - \tau_2)]}. \tag{14b}$$

Next, note that in arriving at these equations we considered the coupled system at an arbitrary time point on the time scale continuing from the uncoupled state to the onset and progress of the coupling interactions. To describe the coupled phase dynamics beginning with the onset of coupling we should shift the time origin to the start of coupling where $\omega_1' = \omega_1$ and $\omega_2' = \omega_2$. Therefore,

$$\dot{\varphi}_1(t) = \omega_1 + \omega_2 H_c e^{j[\varphi_2(t-\tau_1-\tau_c)-\{\varphi_1(t-\tau_1)]} \tag{15a}$$

$$\dot{\varphi}_2(t) = \omega_2 + \omega_1 H_c e^{j[\varphi_1(t-\tau_2-\tau_c)-\varphi_2(t-\tau_2)]} . \tag{15b}$$

Finally, the phase equations of motion will be given by the real part of the right hand side of Eqs. (15a) and (15b). Thus, we obtain the coupled phase equations of the oscillators system presented in Fig. 2 as

$$\dot{\varphi}_1(t) = \omega_1 + \omega_2 H_c \cos[\varphi_2(t-\tau_1-\tau_c) - \varphi_1(t-\tau_1)] \tag{16a}$$

$$\dot{\varphi}_2(t) = \omega_2 + \omega_1 H_c \cos[\varphi_1(t-\tau_2-\tau_c) - \varphi_2(t-\tau_2)] . \tag{16b}$$

We can simplify these equations by noting that $\omega_1 >> \omega_2 H_c$, $\omega_2 >> \omega_1 H_c$ and $\omega_1 \approx \omega_2$. Therefore, we can write the coupling strength factor by $\omega_m H_c$ with $\omega_m = (\omega_1 + \omega_2)/2$ denoting the mean frequency of uncoupled oscillators. From the structure of phase equations (16) it is apparent that they describe any coupled oscillator system having similar delayed feedback and coupling arrangement. Therefore, we rewrite these phase equations in general form by denoting the coupling strength $\omega_m H_c$ by a new symbol K. Thus,

$$\dot{\varphi}_1(t) = \omega_1 + K \cos[\varphi_2(t-\tau_1-\tau_c) - \varphi_1(t-\tau_1)] \tag{17a}$$

$$\dot{\varphi}_2(t) = \omega_2 + K \cos[\varphi_1(t-\tau_2-\tau_c) - \varphi_2(t-\tau_2)] . \tag{17b}$$

where $K = \omega_m H_c$ describes the SAW coupled system.

2.3.2. Phase Difference Equations

In order to determine the conditions for phase locking we can set up an equation of motion for the phase difference between the two interacting oscillators by subtracting Eq. (17a) from (17b), and by defining a new variable for the phase difference

$$\psi(t) = \varphi_2(t) - \varphi_1(t) . \tag{18}$$

Differentiating this with respect to time we obtain

$$\dot{\psi}(t) = (\omega_2 - \omega_1) + K[\cos A - \cos B] \tag{19}$$

where $A = \varphi_1(t-\tau_2-\tau_c) - \varphi_2(t-\tau_2)$ and $B = \varphi_2(t-\tau_1-\tau_c) - \varphi_1(t-\tau_1)$.

The Eq. (19) can be further manipulated by using the trigonometric identity $\cos A - \cos B = -2\sin\frac{A+B}{2}\sin\frac{A-B}{2}$ and using the definition for oscillator phase $\varphi(t) = \omega' t + \vartheta^0$. This gives

$$\begin{aligned} \frac{A+B}{2} &= -\omega_m \tau_c + \frac{1}{2}(\omega_2 - \omega_1)(\tau_2 - \tau_1) \\ &= -\omega_m \tau_c + \tfrac{1}{2}\Delta\tau_d \end{aligned} \tag{20a}$$

$$\begin{aligned}\frac{A-B}{2} &= -\varphi_2(t)+\varphi_1(t)+\tfrac{1}{2}(\omega_2-\omega_1)(\tau_1+\tau_2+\tau_c) \\ &= -\psi(t)+\tfrac{1}{2}\Delta\tau_s\end{aligned}, \tag{20b}$$

where $\tau_d = \tau_2 - \tau_1$ is the delay detuning between SAW feedback devices and $\tau_s = \tau_1 + \tau_2 + \tau_c$ is the open loop delay of the coupled system. Thus we obtain

$$\dot{\psi}(t) = \Delta - 2K\sin(\omega_m\tau_c - \tfrac{1}{2}\Delta\tau_d)\sin[\psi(t) - \tfrac{1}{2}\Delta\tau_s]. \tag{21}$$

In general form,

$$\frac{d\psi}{dt} = \Delta - \varepsilon\sin(\psi - \psi_0), \tag{22}$$

where $\varepsilon = 2K\sin\left(\omega_m\tau_c - \tfrac{1}{2}\Delta\tau_d\right)$ represents the interaction strength and $\psi_0 = \tfrac{1}{2}\Delta\tau_s$. In the case of present SAW coupled system

$$\varepsilon = 2\omega_m H_c \sin\left(\omega_m\tau_c - \tfrac{1}{2}\Delta\tau_d\right) \tag{23}$$

2.3.3. Synchronization

The synchronization occurs when the phase difference between the two oscillators becomes time-invariant, that is, $\dot{\psi} \approx 0$. In other words, the rates of their phase variations become equal, $\dot{\varphi}_1(t) = \dot{\varphi}_2(t)$. In this condition, both the oscillators have pulled their frequencies to a common frequency called the frequency of synchronization. Let it be denoted by Ω. If ψ_d denotes the phase difference in locked condition then from Eq. (21) by setting $\dot{\psi} = 0$ one obtains

$$\begin{aligned}\psi_d &= \psi_0 + \sin^{-1}\left(\frac{\Delta}{2K\sin(\omega_m\tau_c - \frac{1}{2}\Delta\tau_d)}\right) \qquad \text{for } \sin^{-1}x > -\psi_0 \text{ (in-phase)} \\ &= \pi - \psi_0 - \sin^{-1}\left(\frac{\Delta}{2K\sin(\omega_m\tau_c - \frac{1}{2}\Delta\tau_d)}\right) \qquad \text{otherwise (anti-phase)}\end{aligned}, \tag{24}$$

where $x = \dfrac{\Delta}{2K\sin(\omega_m\tau_c - \frac{1}{2}\Delta\tau_d)}$.

The synchronization frequency Ω can be obtained by writing $\Omega = \dot{\varphi}_1 = \dot{\varphi}_2$ and calculating $\dot{\varphi}_1 + \dot{\varphi}_2$ by using Eqs. (17) and (20) under synchronization condition. We obtain,

$$\Omega = \omega_m + K\cos(\omega_m\tau_c - \tfrac{1}{2}\Delta\tau_d)\cos[\psi_d - \tfrac{1}{2}\Delta\tau_s]. \tag{25}$$

This equation can also be expressed in terms of ε, K and Δ by using $\cos x = \sqrt{1-\sin^2 x}$ and Eqs. (22) and (23)

$$\begin{aligned}\Omega &= \omega_m + K\sqrt{1-\frac{\varepsilon^4-\Delta^2(\varepsilon^2-4K^2)}{4\varepsilon^2K^2}} \\ &\cong \omega_m + \sqrt{1-\frac{\varepsilon^4+4K^2\Delta^2}{4K^2\varepsilon^2}} \qquad \text{(in view of } \Delta << 2\text{K)}\end{aligned} \tag{26}$$

Equation (25) with Ψ_d from (24) fully specifies the synchronized state of the coupled oscillator system. The parametric dependencies of the synchronization frequency can be noted. It can be seen that $\Omega = f(\omega_m, \Delta, \tau_d, H_c, \tau_c)$. That is, the synchronization frequency depends on the natural oscillation frequencies of the interacting oscillators and coupling SAW device. In principle, the change in synchronization frequency due to change in delay across the coupling SAW device can be made the basis for making SAW sensors. In the following, we examine the sensitivity and noise characteristics of this configuration to assess its potentiality for sensor applications.

3. Sensing by Synchronization

In the coupled dual oscillator configuration analyzed above we consider the coupling SAW delay line as the sensing platform. For making a gas sensor the SAW propagation surface can be sensitized by depositing a thin polymer coating as in normal SAW sensors. The solution for synchronization frequency, Eq. (25), shows that any perturbation to propagation delay τ_c affects the synchronization frequency. Therefore, the present configuration must work like a traditional SAW vapor sensor with change in synchronization frequency as its output. However, to compare its performance with traditional SAW sensors we must analyze sensitivity of Ω to perturbation of τ_c, influence of synchronization on oscillator noise and its robustness to spurious perturbations.

3.1. Sensitivity

The sensitivity of a sensor is defined as change in sensor output for unit change in stimulant. In the present case [in view of Eq. (25)] it is convenient to define sensor signal referred to the mean frequency. Therefore, in parallel to the traditional sensors we define the sensor signal as the fractional change in $\Omega' = \Omega - \omega_m$, and sensitivity as

$$y = \frac{1}{\Omega'}\frac{d\Omega'}{d\tau_c}. \tag{27}$$

By defining, $A_1 = \omega_m\tau_c - \tfrac{1}{2}\Delta\tau_d$, $A_2 = \psi_d - \tfrac{1}{2}\Delta\tau_s$ in Eqs. (24) and (25) we rewrite these as

$$A_2 = \sin^{-1}\left(\frac{\Delta}{2K\sin A_1}\right) \tag{28}$$

$$\Omega' = K\cos A_1 \cos A_2. \tag{29}$$

and calculate

$$\frac{dA_1}{d\tau_c} = \omega_m \tag{30}$$

$$\frac{dA_2}{d\tau_c} = -\frac{\Delta\omega_m}{2K}\frac{\cos A_1}{\cos A_2}\frac{1}{\sin^2 A_1} \tag{31}$$

and then, differentiating Eq. (29) with respect to τ_c we obtain

$$\frac{1}{\Omega'}\frac{d\Omega'}{d\tau_c} = -\beta\omega_m , \tag{32a}$$

where the factor

$$\beta = \frac{\sin A_1 - (\tan A_2 / \tan A_1)^2}{\cos A_1} = \tan A_1 - \sec A_1 (\tan A_2 / \tan A_1)^2 . \tag{32b}$$

A normal SAW mass sensor is a single free running polymer-coated SAW oscillator. The frequency change of oscillator is described the well known Sauerbrey`s relation [26, 27, 12 ch.5]

$$\frac{\Delta\omega}{\omega} = -\kappa\omega h \Delta\rho , \tag{33}$$

where κ is a constant depending on the SAW substrate material and propagation mode, ω denotes the oscillator operating frequency, h is polymer film thickness and $\Delta\rho$ is the polymer mass density. The vapor sorption in polymer produces a change in oscillator frequency via a succession of processes: change in vapor concentration $C \rightarrow$ change in polymer mass density $\rho \rightarrow$ change in SAW velocity $v_{SAW} \rightarrow$ change in SAW propagation delay across the feedback device $\tau \rightarrow$ change loop phase $\varphi \rightarrow$ change in oscillator frequency $\Delta\omega$. All the successive changes are linearly related, therefore, the following relation holds for the fractional changes in these quantities

$$\frac{\Delta\omega}{\omega} = \frac{\Delta\varphi}{\varphi} = \frac{\Delta\tau}{\tau} = \frac{\Delta v_{SAW}}{v_{SAW}} = \frac{\Delta\rho}{\rho} . \tag{34}$$

From the second equality in Eq. (34) it can be seen that $\Delta\omega / \omega\Delta\tau = 1/\tau$. Note that for self sustained free running oscillations $\omega\tau = 2\pi$; therefore, $1/\tau = \omega/2\pi = f_0$ (oscillator quiescent frequency). Thus, the sensitivity of the normal SAW oscillator frequency to changes in propagation delay can be obtained under the limit $\Delta\tau \rightarrow 0$ as

$$\frac{1}{\omega}\frac{d\omega}{d\tau} = f_0 . \tag{35}$$

In view of this, writing $\omega_m = 2\pi f_m$ in Eq. (32), one can interpret $\beta' = 2\pi\beta$ as given by Eq. (32b) as the factor by which the sensitivity of the synchronized oscillator sensor enhances relative to the free running oscillator sensor.

3.1. Noise

Uncertainty in signal measurement is characterized by random frequency fluctuations called noise. The noise in a sensing oscillator defines the minimum change in sensor output (signal) that can be measured (or the minimum detection limit of the measurand). The signal to noise ratio is therefore an important figure of merit of the sensor. The frequency fluctuations in SAW oscillators are dominated by phase fluctuations. The standard deviation of a random variable is often taken as measure of uncertainty in its measurement. The standard deviation of frequency measurement however depends on the bandwidth of measuring system or the time interval of measurement. Therefore, in order to specify a noise figure of merit that is independent of measurement condition, a normalized measure referred to 1-Hz bandwidth is used. The most commonly used measure for this purpose is noise power spectral density specified in $(\text{Hz})^2/\text{Hz}$ for frequency fluctuation or in $(\text{radian})^2/\text{Hz}$ for phase fluctuation as a function of position from the carrier (called offset frequency f). The spectral density for frequency fluctuation $S_F(f)$ and phase fluctuation $S_\Phi(f)$ are related as [28]

$$S_\Phi(f) = S_F(f)/f^2 . \tag{36}$$

The spectral density is simply the mean square frequency fluctuation in 1-Hz bandwidth at the offset frequency.

The effect of phase locking or frequency synchronization is known to reduce noise in oscillators (both electronic and non-electronic) [29-32]. It is shown theoretically and experimentally that when N identical oscillators interact on reciprocal basis the phase noise power spectral density in the synchronized state is reduced to $1/N$ of a single oscillator [30, 31]. Recently, in a detailed analysis of coupled neuronal (relaxation) oscillators it was shown that synchronization reduces internal noise and also makes the system more robust against external perturbations [32]. In [2, section 9.2.4] the effect of synchronization on phase noise in two weakly coupled nonlinear limit cycle oscillators has been described. The phase fluctuation of a self sustained oscillator is described as a random walk or diffusion process. The noise power spectral density for a Gaussian frequency fluctuation has been found to be equal to the diffusion constant $D = 2\sigma^2$ where σ denotes the standard deviation of the frequency random walk. Mutual interaction of phase oscillators leading to synchronization allows randomness in their phases also to flow as easily through the coupling network and loop as their unperturbed phases. The noise for the coupled state is calculated by defining a new variable representing the sum of phases of individual oscillators, $\vartheta = \varepsilon(\varphi_1 + \varphi_2)$ whose diffusion coefficient is calculated as $D_\vartheta = 2\varepsilon^2(\sigma_1^2 + \sigma_2^2)$ where $D_1^0 = 2\sigma_1^2$ and $D_2^0 = 2\sigma_2^2$ are phase diffusion coefficients of noninteracting oscillators. The diffusion constant of the individual oscillators in the coupled state is obtained as

$$D_1 = D_2 = D = \frac{\sigma_1^2 + \sigma_2^2}{2} . \tag{37}$$

If both the oscillators in their free running states have the same amount of noise power spectral density (that is, if $\sigma_1 = \sigma_2 = \sigma$), then from Eq. (34) we see that in the coupled state $D = \sigma^2$, which is ½ the noise power spectral density of the individual oscillators in free running state. This result is the same as that mentioned above for the noise reduction by $1/N$ due to synchronization. Otherwise, it can be noted that the noise in coupled state is always lower than the noisiest oscillator in free running state.

4. Results and Discussion

The relations for synchronization frequency Ω, phase difference at synchronization Ψ_d and sensitivity factor β given by Eqs. (24), (25) and (32b) are functions of 2π-periodic sinusoidal functions. All these relations depend on the phase shift across the coupling delay $\omega_m\tau_c$, frequency detuning parameter Δ, difference between feedback delays of the two oscillators τ_d, and the total delay (both loops + coupling) τ_s. In general, they represent complex oscillatory behaviour. However, to make the coupling SAW device as sensing platform we must seek simplified conditions under which high sensitivity linear response could be obtained. Let us examine the following three conditions:

(i) $A_1 = (2n+1)\pi/2 + z$ such that $\sin A_1 \approx \pm 1$ and $\cos A_1 \approx \pm z$,
(ii) $A_1 = n\pi + z$ such that $\sin A_1 \approx \pm z$ and $\cos A_1 \approx \pm 1$, and
(iii) $A_1 \approx A_2$,

where n denotes an integer and z denotes some small phase deviation from half-integer or integer multiples of π. In writing these conditions, approximations $\sin z \approx z$ and $\cos z \approx 1$ have been used.

Condition 1:
$\sin A_2 = \Delta/2K \sin A_1 \approx \pm\Delta/2K$, and $\cos A_2 = \sqrt{1-(\Delta/2K\sin A_1)^2} \approx 1$ by noting that $\Delta << 2K$. Hence, from Eq. (25) and Eq. (32b)

$$\Omega \approx \omega_m \pm zK \tag{38a}$$

$$\beta \approx \pm 1/z \tag{38b}$$

Condition 2:
$\sin A_2 \approx \pm\Delta/2zK$, and $\cos A_2 \approx 0$. Hence, from Eq. (25) and Eq. (32b)

$$\Omega \approx \omega_m \tag{39a}$$

$$\beta \approx z \tag{39b}$$

Condition 3:
$\sin A_1 = \sin A_2 \approx \Delta/2K$, and $\cos A_1 \cos A_2 \approx 1 - \Delta/2K$. Hence, from Eq. (25) and Eq. (32b)

$$\Omega \approx \omega_m \pm K(1-\Delta/2K) \tag{40a}$$

$$\beta \approx -1. \tag{40b}$$

The condition 2 implies that the coupled system is synchronized at mean frequencies ω_m, and the sensitivity to perturbation can be made arbitrarily small by maintaining the condition close to $A_1 = n\pi$ so that $z \to 0$. This condition is suitable for making stable phase locked oscillator system. The condition 3 tells that the frequency locking can occur at $(\omega_m \pm K(1-\Delta/2K)$ with sensitivity factor close to the uncoupled oscillator sensitivity. Under this condition there does not seem to accrue any advantage either from stability or sensing point of view. The condition 1, however, provides the basis for making the coupled oscillator system a sensor whose sensitivity can be adjusted to any desired level (limited only by the practical constraints) by keeping z low enough to maintain the operation point close to $A_1 = n\pi + \pi/2$. The synchronization frequency varies linearly with z (that is τ_c) and sensitivity can be adjusted by fixing the phase shift across the coupling delay line somewhere between integer to half-integer multiple of $\pi/2$.

Figs. 3 and 4 show the results for the phase difference, synchronization frequency and sensitivity factor corresponding to condition 1 and condition 2 respectively. The following values of parameters were used: $f_1 = \omega_1/2\pi = 199.9$ MHz, $f_2 = \omega_2/2\pi = 200.1$ MHz, $\tau_1 = 3\,\mu\mathrm{s}$, $\tau_2 = 3.1\,\mu\mathrm{s}$, $H_c = 0.005$, $n = 100$. Note the hyperbolic sensitivity $(\sim 1/z)$ in Fig. 3(a) for the coupling delay variations close to $A_1 = (2n+1)\pi/2$ as given by Eq. (38b). The synchronization frequency varies linearly with delay variation within region of high sensitivity. Therefore, this can condition is appropriate for developing sensors based on synchronization. The sensitivity factor however changes sign as one crosses this point even though the oscillator remain synchronized in phase, see Fig. 3b. This means that accurate adherence to odd multiple of $\pi/2$ condition will make the sensor highly unstable. The operating point should therefore be set on one side of this point at a distance such that it can cater for the range of frequency variations for a given sensing application without crossing this point.

The results in Fig. 4 are as expected from the condition 2. The system is synchronized at mean frequency with sensitivity being close to zero, Fig. 4a. The two coupled oscillators however oscillate in phase on one side, and in opposite phase on the other side, see Fig. 4b. This tells that if the system is locked accurately at $A_1 = n\pi$, it maintains the locked state in which both the oscillators manoeuvre their phases to go from in-phase or out of phase around this point. This state is ideal for achieving high quality phase locked oscillator system.

Further note from these figures (Figs. 3 and 4) that the range of τ_c variation over which either frequency locked or sensitive regions are maintained is very small, $\Delta\tau_c/\tau_c \approx 0.0007$ or 700 ppm. However, most vapor sensing applications involve ppm to sub-ppm variations of stimulant. Hence, the present configuration provides adequate stability for making sensors.

Figs. 5 and 6 show the variation of synchronization characteristics over wider ranges of τ_c variation for higher values of detuning parameters. The parameters are: $f_1 = \omega_1/2\pi = 199.75$ MHz, $f_2 = \omega_2/2\pi = 200.25$ MHz, $\tau_1 = 3\,\mu\mathrm{s}$, $\tau_2 = 3.3\,\mu\mathrm{s}$, $H_c = 0.005$, and $\Delta\tau_c = 0.004\,\mu\mathrm{s}$ in Fig. 5, and $\Delta\tau_c = 0.1\,\mu\mathrm{s}$ in Fig. 6. Several frequency-locked and sensitive regions can be easily identified in these figures. As the range of τ_c is increased the number of such regions also increase. This suggests that if the coupling path incorporates a phase noise source of enough strength such that it can make the system hop over different synchronization frequencies it may be useful for communication systems based on frequency hopping.

From the results shown above and from the considerations on noise in synchronized oscillator system described in Section 3.1 it is clear that by inserting coupling delay line with proper delay time one can realize much higher signal-to-noise ratio SAW sensors if the coupling SAW device is made the sensing platform. For practical implementation, perhaps, it may be more prudent to incorporate an adjustable phase shifter in the coupling path so that the operating condition for high signal-to-noise could be fine tuned. Synchronization sensor may provide other advantage also like immunity against spurious electromagnetic interferences by keeping both the oscillators in synchronized state.

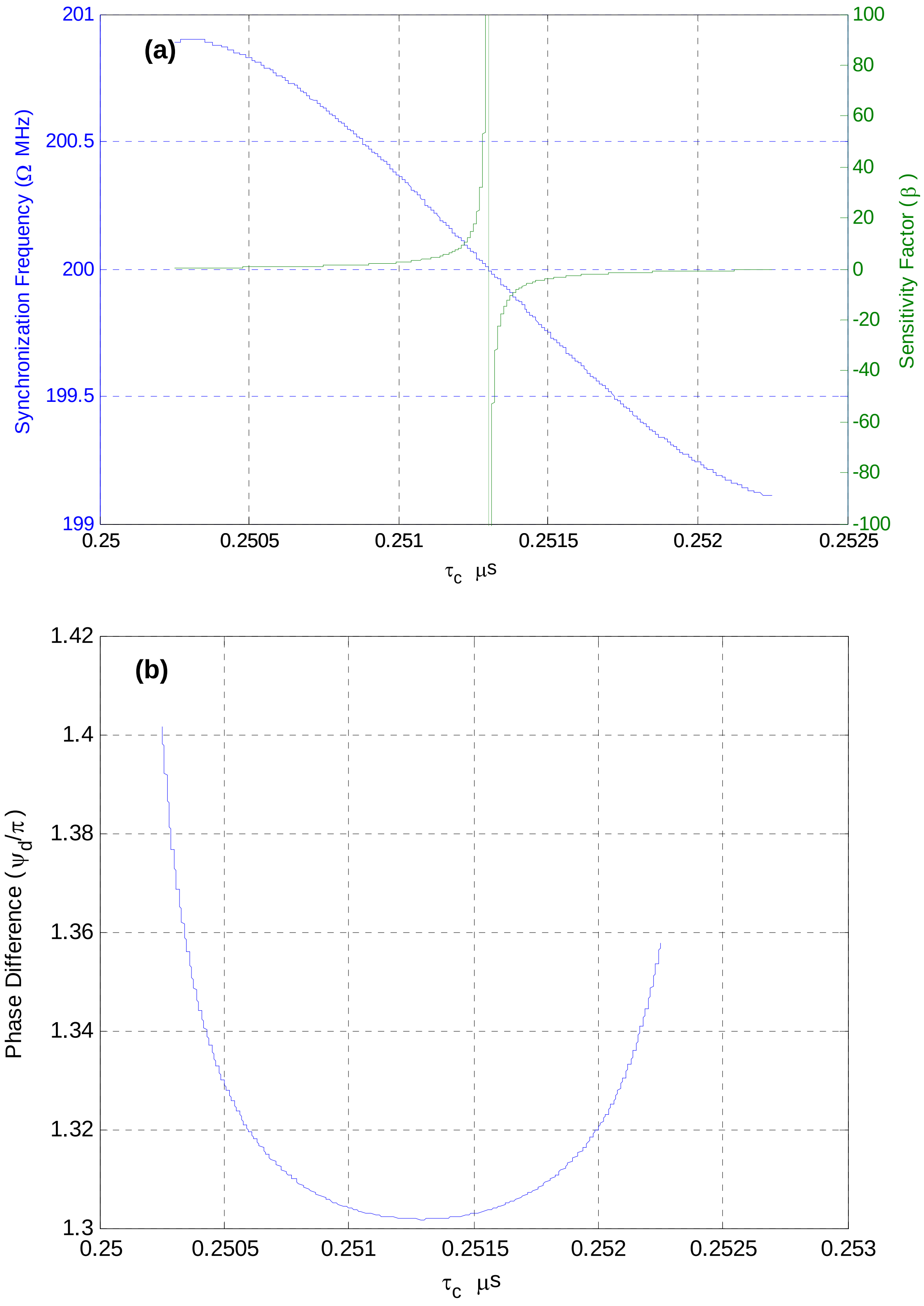

Fig. 3. State of synchronization for the phase change across the coupling SAW delay line being an odd multiple of $\pi/2$.

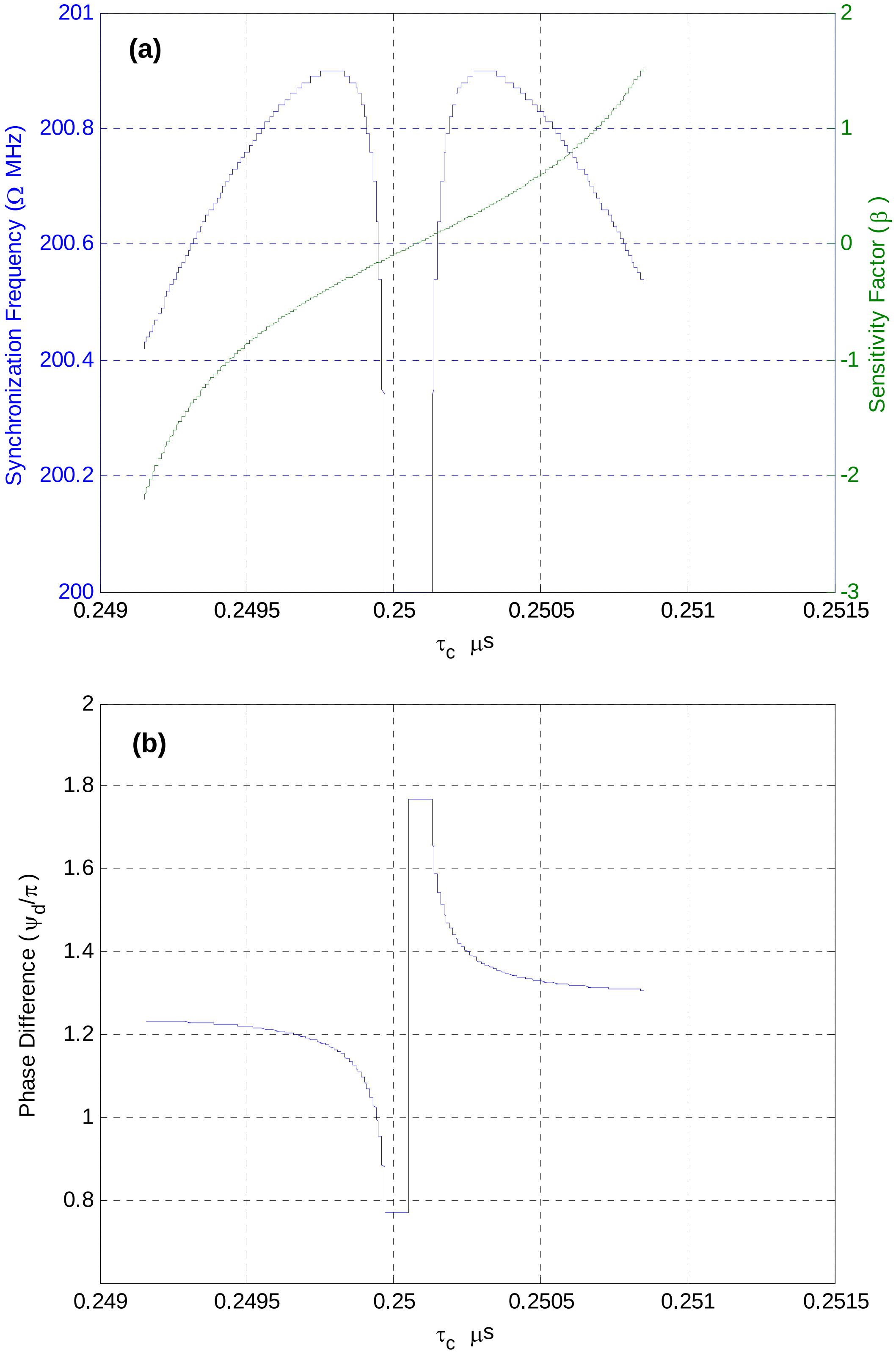

Fig. 4. State of synchronization for the phase change across the coupling SAW delay line being an integer multiple of $\pi/2$.

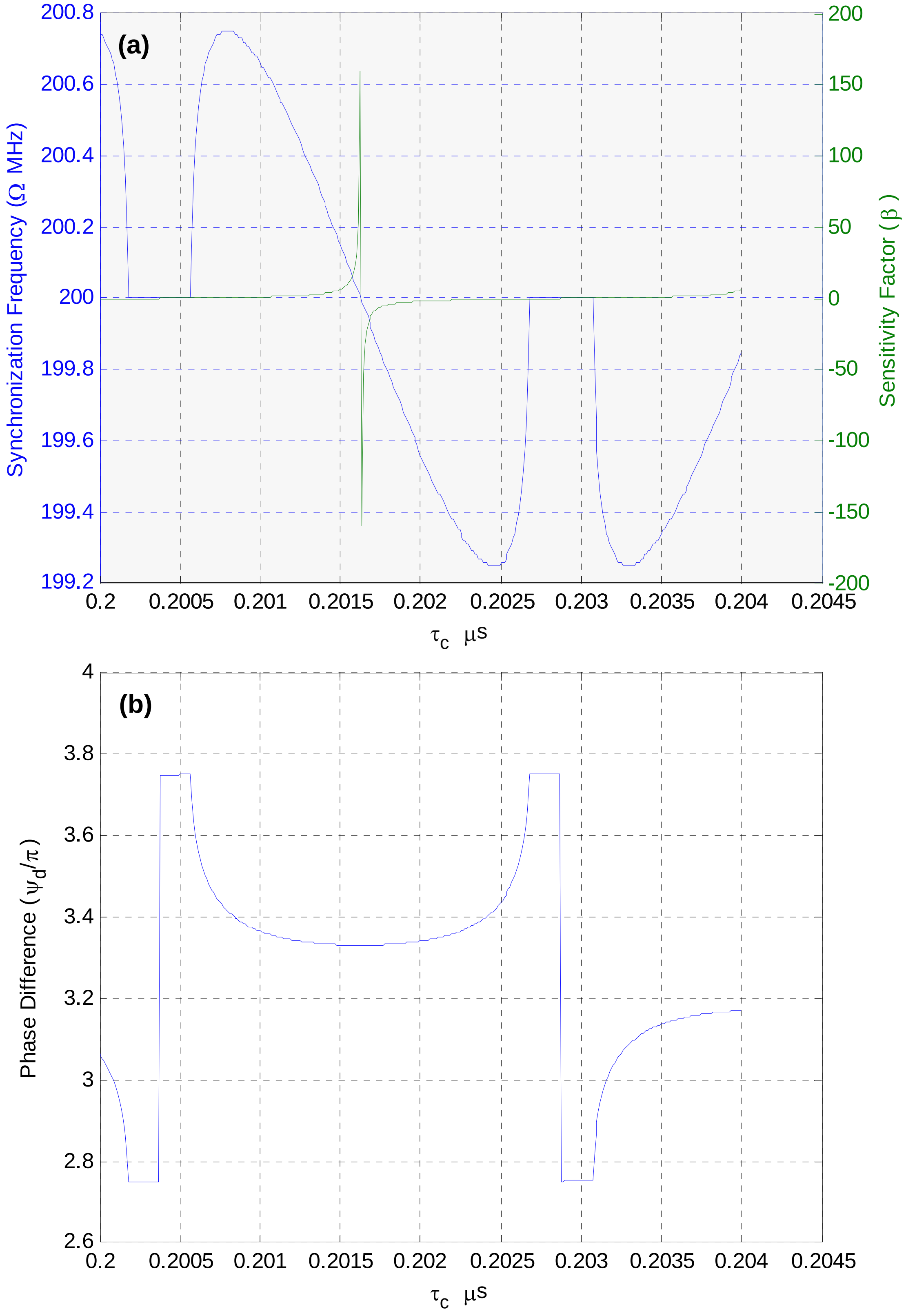

Fig. 5. State of synchronization as a function of coupling delay over $\Delta\tau_c = 0.004$ μs.

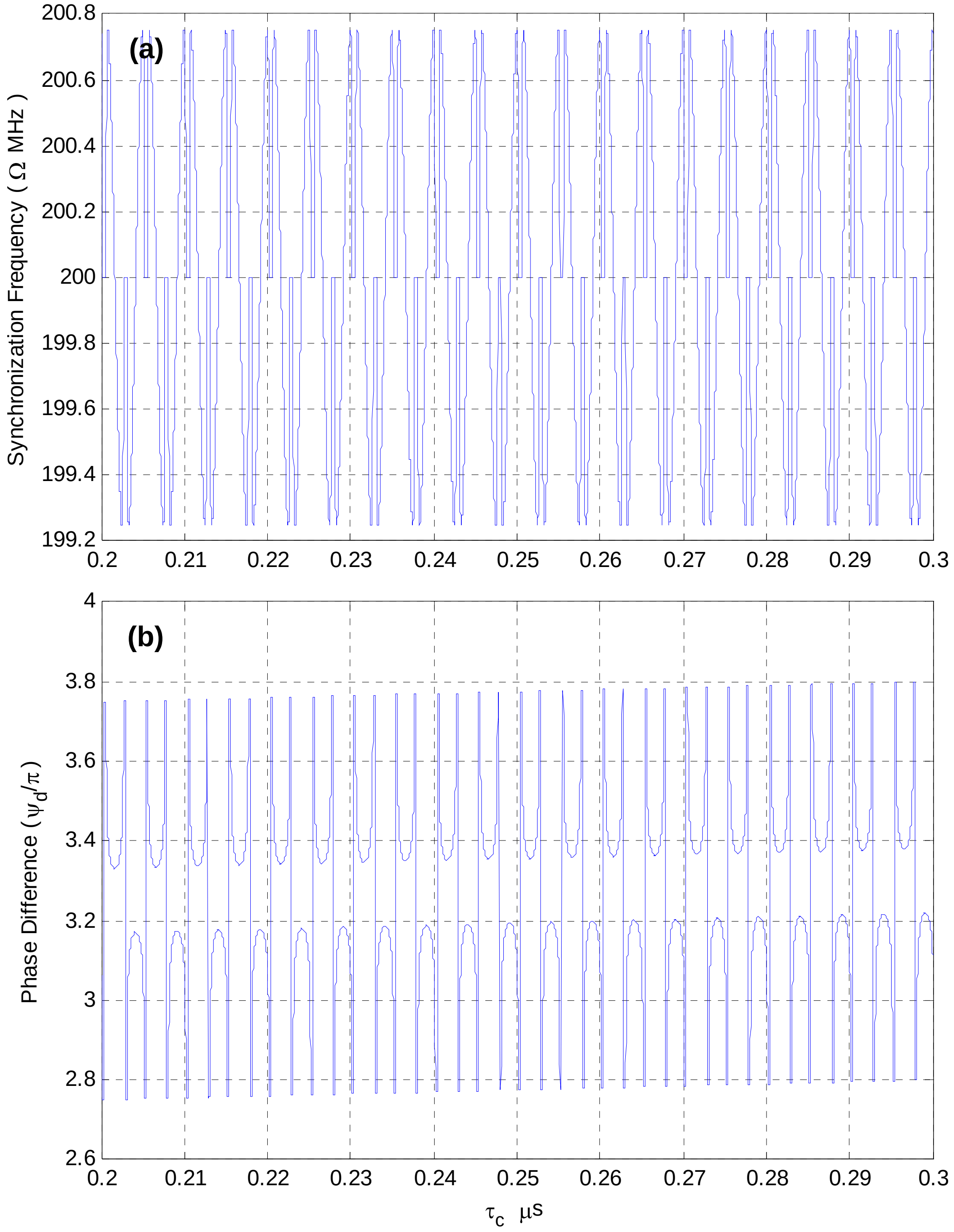

Fig. 6. Synchronization frequency and phase difference for $\Delta\tau_c = 0.1$ μs .

5. Conclusions

Two limit cycle SAW feedback oscillators with cubic nonlinearity amplifiers in loops can be synchronized by coupling through a simple SAW delay line. The state of synchronization defined by time-invariance of the phase difference between coupled oscillators can be controlled by the time delay in coupling. Two types of phase locked conditions are noticed. In one, the phase locking occurs at a fixed frequency such that the synchronization states become insensitive to perturbations. In second, the synchronization frequency varies linearly with the amount of perturbation. The second types of synchronization states are suitable for making the coupling SAW delay line as sensor. The synchronized SAW oscillator system provides flexible conditions for adjustment of sensitivity and robustness. The synchronized states are less noisy compared to free running states. This adds to the potentiality of synchronization as method for developing high performance SAW sensors.

Acknowledgements

The author SSJ acknowledges the JRF support by UGC, Govt. of India, New Delhi. The authors are thankful to Mr. Prashant Singh for his valuable help in MatLab programming.

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