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Electromagnetic Investigation of a CMOS MEMS Inductive Microphone

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Abstract: This paper presents a detailed electromagnetic modeling for a new structure of a monolithic CMOS micromachined inductive microphone. We have shown, that the use of an alternative current (AC) in the primary fixed inductor results in a substantially higher induced voltage in the secondary inductor comparing to the case when a direct current (DC) is used. The expected increase of the induced voltage can be expressed by a voltage ratio of AC and DC solutions that is in the range of 3 to 6. A prototype fabrication of this microphone has been realized using a combination of standard CMOS 0.6 μm process with a CMOS-compatible post-process consisting in a bulk micromachining technology. The output voltage of the electrodynamic microphone that achieves the μV range can be increased by the use of the symmetric dual-layer spiral inductor structure. *Copyright © 2009 IFSA.*

Keywords: MEMS, Electromagnetic modeling, inductive microphone, micromachined inductors, induced voltage.

1. Introduction

A microphone is an electroacoustic transducer that converts pressure inputs into an electrical signal and is widely employed in numerous applications from different fields as consumer, industrial [1], health [2], and others [3]. A variety of transduction schemes, such as piezoresistive, piezoelectric,

electrostatic, or optical have been used for acousto-electrical conversion. The vast majority of microphones are designed for audio applications, including a bandwidth less than 20 kHz and a maximum pressure less than 120 dB (ref. to 20 μPa).

Among various designs, piezoresistive microphones have the advantages of simple fabrication processes and integration into semiconductor devices [4]. It uses materials and processes to make gauges that can be integrated on beams or membranes. The electrical resistivity of these gauges changes with the variation of mechanical stress produced by deflection caused by sound waves. Piezoelectric microphones can have a similar design as piezoresistives ones, but the materials used for these devices generate differences in electrical charge at the surface instead of changing resistivity. However, piezoelectric systems suffer from both low sensitivity and the requirement to utilize specific piezo-materials such as ZnO, A1N, PVDF and PZT that need a non standard fabrication process [5].

Another effect, widely used in microphones, is the electrostatic transduction. Microphones employing this effect, called electrostatic or capacitive, are sensitive to variations in an electrostatic field created between two electrodes. This variation is induced by the movement of a microphone membrane due to the acoustic pressure. The electrical field, necessary for the microphone operation can be either supplied by an external source or created by a deposition of a material layer that can quasi-permanently keep the electrical charge between the two electrodes. This material, called electret, brings the potential for the use in portable applications [6]. Capacitive microphones have many advantages such as a relatively high sensitivity, a large bandwidth, inherently low power consumption and a low noise floor. However, in the design of capacitive microphones, specific issues such as electrostatic pull-in instability, output signal attenuation due to the parasitic capacitance and in certain designs, a decreased sensitivity at high frequencies due to the viscous damping of the perforated backplate must be solved [7]. Moreover, if an electret microphone is considered for an integration using MEMS technology, low quality of thin film electrets and charge loss due to humidity are limiting factors. The problem of detecting a small capacitance change has been overcome by clever preamplifier designs. However, narrow mechanical and electronic component tolerances have kept the price of condenser microphones relatively high.

Optical detection is another alternative to detect sound pressure. The idea of an optical microphone is based on the modulation of light rather than the conversion from an existing energy form into light energy [8]. Optical interferometric microphones have, theoretically, a dynamic range and sensitivity comparable to or better than capacitive sensors, depending on the available source power [9]. Optical microphones have the advantage that they can be deployed in harsh environments and are immune from electromagnetic interference; however, they require a stable light source and complex packaging [10].

To overcome problems related to the microphone integration, a more flexible approach is presented in this work. We have shown in our previous work the feasibility of an inductive micromachined single-wafer silicon microphone [11, 12, 16]. The monolithic integration with the electronics using a standard CMOS 0.6μm micromachining process was presented. This monolithic integration will increase performances, miniaturize the system, increase the sensitivity and in particular decrease noise, due to the reduction of interconnections parasitic capacitance. Furthermore, our approach focuses on low-cost MEMS microphone obtained by processing chips issued from an industrial standard CMOS process. In this approach, we use silicon and other layers issued from a CMOS process as basic materials for the mechanical part of the system.

2. Operation Mode of the Microphone

The structure of the electrodynamic microphone, made with a standard CMOS technology, is shown in Fig. 1. The microphone consists mainly of two inductors which occupy separate regions. The first one, a stationary outer inductor L_1, is placed on top of the substrate and an inner inductor L_2 is fabricated on top of a flexible plate suspended over the micromachined cavity. This square plate, attached to the substrate with four arms, one at each corner, has a role of the microphone sensitive part (membrane). The electromagnetic field necessary for the microphone function can be created by an electrical current flowing through one or both inductors. This polarization current can be either direct (DC) or alternative (AC). Thus, a variable magnetic field in space and/or in time will be generated in the vicinity of the inner inductor L_2 that will move due to the movement of the suspended membrane under the incident acoustic energy. According to Faraday's law, the vibration of the inductor L_2 in a magnetic field will generate at its ends an electromotive force (*emf*), proportional to the amplitude of the incident acoustic pressure. This voltage will be processed by the electronic circuitry integrated on the same chip.

The cavity under the suspended membrane is obtained using the front-side bulk micromachining realized after an industrial CMOS process, with no modification in the standard IC fabrication, no additional masking and no negative effects on the associated electronic parts. In this back-end process, a part of the bulk silicon material is etched through openings patterned in dielectric and passivation layers. These openings are easily constructed along with the suspended membrane and the attachments arms through the appropriate placement of open areas in the layout. Nevertheless, design rules of a given technology process must be respected, which defines the minimum open area width (5 µm in our case).

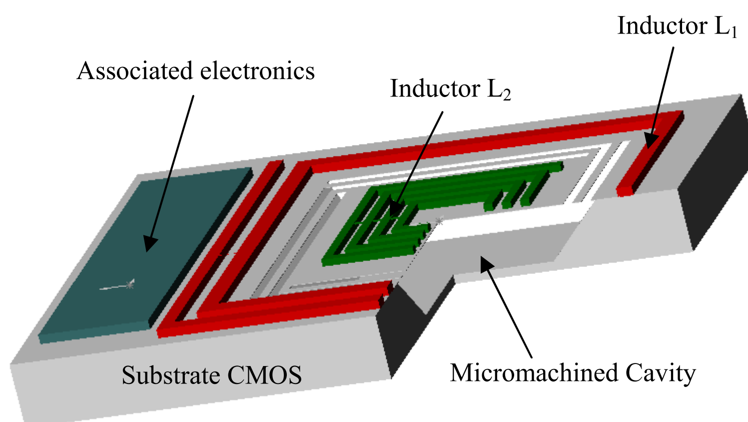

Fig. 1. Structure of the electrodynamic microphone.

3. Induced Electromotive Force (emf)

The magnetic field (B-field), at any point P in the space, produced by straight wire carrying current I is given by Biot-Savart law as shown in the following equation:

$$B_z = \frac{\mu_0}{4\pi}\frac{I}{r}\left(\sin\theta_1 + \sin\theta_2\right), \tag{1}$$

where μ_0 is the magnetic permeability of the air, r is the perpendicular distance separating P from the wire, and θ_1, θ_2 are the angles shown in Fig. 2 (a).

Circuits having a rectangular shape are made of straight segments and the B-field of such circuits can be obtained by adding their separate contributions. Thus, for a squared loop of side *a,* placed in the x-y-plane (Fig. 2 (b)), flown by a uniform current density on its four segments, will produce three magnetic field components, constant in time and nonuniform in space as given by the following set of equations [12]:

$$B_x(P) = \frac{\mu_0 I\, z}{4\pi}\left[\frac{1}{(\frac{a}{2}-x)^2+z^2}\left(\frac{\frac{a}{2}-y}{A_1}+\frac{\frac{a}{2}+y}{A_2}\right)-\frac{1}{(\frac{a}{2}+x)^2+z^2}\left(\frac{\frac{a}{2}-y}{A_3}+\frac{\frac{a}{2}+y}{A_4}\right)\right]$$

$$B_y(P) = \frac{\mu_0 I\, z}{4\pi}\left[\frac{1}{(\frac{a}{2}-y)^2+z^2}\left(\frac{\frac{a}{2}-x}{A_1}+\frac{\frac{a}{2}+x}{A_3}\right)-\frac{1}{(\frac{a}{2}+y)^2+z^2}\left(\frac{\frac{a}{2}-x}{A_2}+\frac{\frac{a}{2}+x}{A_4}\right)\right]$$

$$B_z(P) = \frac{\mu_0 I}{4\pi}\left[\frac{\frac{a}{2}+x}{(\frac{a}{2}+x)^2+z^2}\left(\frac{\frac{a}{2}-y}{A_1}+\frac{\frac{a}{2}+y}{A_2}\right)+\frac{(\frac{a}{2}-x)}{(\frac{a}{2}-x)^2+z^2}\left(\frac{\frac{a}{2}-y}{A_3}+\frac{\frac{a}{2}+y}{A_4}\right)\right.$$
$$\left.+\frac{\frac{a}{2}-y}{(\frac{a}{2}-y)^2+z^2}\left(\frac{\frac{a}{2}-x}{A_1}+\frac{\frac{a}{2}+x}{A_3}\right)+\frac{\frac{a}{2}+y}{(\frac{a}{2}+y)^2+z^2}\left(\frac{\frac{a}{2}-x}{A_2}+\frac{\frac{a}{2}+x}{A_4}\right)\right] \quad (2)$$

where, $A_1 = \sqrt{(\frac{a}{2}-x)^2+z^2+(\frac{a}{2}-y)^2}$, $A_2 = \sqrt{(\frac{a}{2}-x)^2+z^2+(\frac{a}{2}+y)^2}$,

$A_3 = \sqrt{(\frac{a}{2}+x)^2+z^2+(\frac{a}{2}-y)^2}$, $A_4 = \sqrt{(\frac{a}{2}+x)^2+z^2+(\frac{a}{2}+y)^2}$

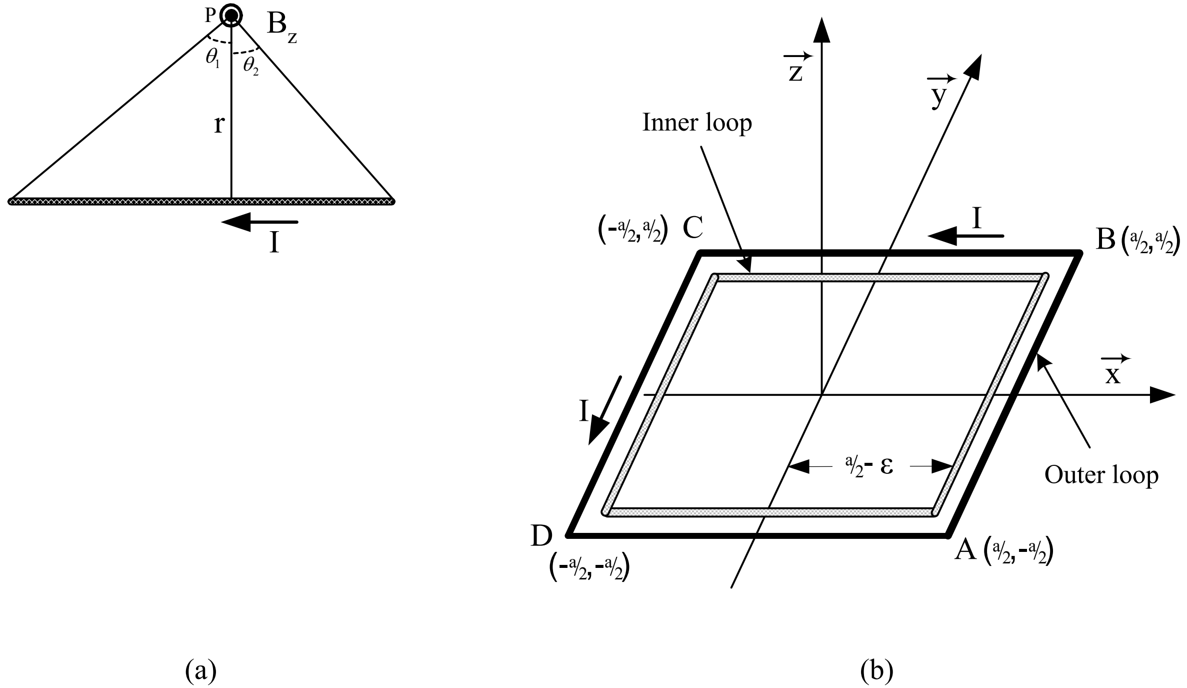

Fig. 2. Geometric relations between magnetic field and conductors in the case of: straight wire (a), and closed loops (b).

In general, the total electromotive force, as given by Faraday's law, is a voltage that arises in a closed conducting path from the time rate of change of the flux (*emf=-dΦ/dt*). A nonzero value of the *emf* may result from a relative motion between a steady flux and the closed path, Φ_c, and/or a time-changing flux linking the stationary closed path like its show by equation (3) [21].

$$emf = -\frac{d\Phi}{dt} = -\frac{d}{dt}\iint_S \vec{B}.d\vec{S} - \frac{d\Phi_c}{dt} \tag{3}$$

In the case of two inductively coupled coils, we can identify the leakage flux Φ_{22} produced by the current i_2 and the flux Φ_{21} that links both currents i_1 and i_2 like its show by equation (4)

$$\Phi = \iint_S \vec{B}.d\vec{S} = \Phi_{22} + \Phi_{21} = L_2 i_2 + M i_1 \tag{4}$$

In the following sub-sections, we will derive expressions for the induced voltage in different possible cases of polarization, depending on currents flowing in one or both inductors.

3.1. Case 1: Outer Inductor Carrying a DC Bias Current

We consider two squared loops, the outer is stationary and the inner one moves up and down along the z-axis with the deflection of the membrane caused by incident acoustic pressure. In a general case, the incident acoustic pressure is a broad-band signal. For simplicity, we will take into account, in this scalable study, only one frequency. Supposing that the membrane exercise a piston-like movement, the instantaneous inductor displacement is given by $\xi_m = h(\omega_p)\ sin(\omega_p t)$, where ω_p is the angular velocity of the incident acoustic pressure and h is the corresponding amplitude of the membrane deflection. According to Faraday's law, an induced electromotive force (*emf*) in the inner closed-loop circuit will be produced. This electromotive force, called motional *emf*, is generated by the movement of the closed conducting loop through a time-independent but spatially varying magnetic field. The expression of the emf_m is proportional to the variation of the magnetic flux, Φ_c, as given by the following equation [13]:

$$emf_m = -\frac{d\Phi_c}{dt} = \oint_{loop} \vec{E_m}.\vec{dl} = \oint_{loop} (\vec{v} \wedge \vec{B}).\vec{dl}\ , \tag{5}$$

where E_m is called the motional electric field, dl represents the elementary displacement of the loop and v is the conductor velocity given by deriving the displacement ξ_m with respect to the time. The value of the emf_m is equal to the algebraic sum of the voltage drops across each segment forming the loop. For time-constant B-fields, the emf_m is given by equation (6):

$$\begin{aligned} emf_m = h\,\omega_p\, cos(\omega_p t) \times \Bigg[& \int_{AB} \left(\vec{z} \wedge B_{x/(x=a/2-\varepsilon)}.\vec{x}\right).\,dy\vec{y} + \int_{BC} \left(\vec{z} \wedge B_{y/(y=a/2-\varepsilon)}.\vec{y}\right).\,dx\vec{x} \\ & - \int_{CD} \left(\vec{z} \wedge B_{x/(x=-a/2+\varepsilon)}.\vec{x}\right).\,dy\vec{y} - \int_{DA} \left(\vec{z} \wedge B_{y/(y=-a/2+\varepsilon)}.\vec{y}\right).\,dx\vec{x} \Bigg], \end{aligned} \tag{6}$$

where ε is the average distance separating the inner closed-loop circuit from the outer. B_x and B_y are odd functions relative to the x and y variables because of the symmetry properties of the square loop geometry. The resulting simplified expression of the emf_m is given by equation (7):

$$emf_m = 4h\omega_p\, cos(\omega_p t) \int_{-\frac{a}{2}+\varepsilon}^{\frac{a}{2}-\varepsilon} B_{x/(x=a/2-\varepsilon)}.\,dy \tag{7}$$

The accurate expression of the emf_m as a function of the z-axis displacement, ξ_m, is given by equation (8):

$$emf_m = \frac{2\mu_0 I}{\pi} \frac{\xi_m \left(\sqrt{\varepsilon^2 + \xi_m^2 + (a-\varepsilon)^2} - \sqrt{2\varepsilon^2 + \xi_m^2} \right)}{\varepsilon^2 + \xi_m^2} h\omega_p \cos(\omega_p t) = A_1 h\omega_p \cos(\omega_p t) \quad (8)$$

The A_1 factor in the expression of the emf_m given by the Equation (8) depends on the displacement ξ_m and its value is shown for given dimensions in Fig. 3. This factor can be considered as linear if displacements are sufficiently small (less than 20µm in our case (Fig. 4)). When we make a series expansion for ξ_m neighboring zero, the simplified emf_m expression, for a linear range, can be approximated by the equation (9). Accordingly, the total emf_m will remain sinusoidal if the amplitude of the displacement is restricted to this limited linear range.

$$emf_m \approx \frac{2\mu_0 I}{\pi} \frac{\sqrt{\varepsilon^2 + (a-\varepsilon)^2} - \sqrt{2\varepsilon^2}}{\varepsilon^2} \xi_m h\omega_p \cos(\omega_p t) = A_2 h\omega_p \cos(\omega_p t) \quad (9)$$

In order to extend the obtained results for simple conducting loops towards planar inductors having n_1 and n_2 turns, we need to apply the following approximations. The sum of all electromotive forces will be approximated by the equivalent average effect of n_1 segments of L_1 on n_2 segments of the inner inductor L_2. Each inductor will be represented by its average length and both inductors will be placed at the average distance. Let l_a and m_a be the length of the average spire in the outer and the inner inductor, respectively. ε_a is the average distance separating the two medium spires of the two inductors ($\varepsilon_a=(l_a-m_a)/2$). When we substitute ξ_m by $h\sin(\omega_p t)$ in equation (9) the total open-circuit output voltage is given by the following equation

$$emf_m \approx n_1 n_2 \frac{\mu_0 I}{\pi} \frac{\sqrt{\varepsilon_a^2 + (l_a-\varepsilon_a)^2} - \sqrt{2\varepsilon_a^2}}{\varepsilon_a^2} h^2 \omega_p \sin(2\omega_p t) \quad (10)$$

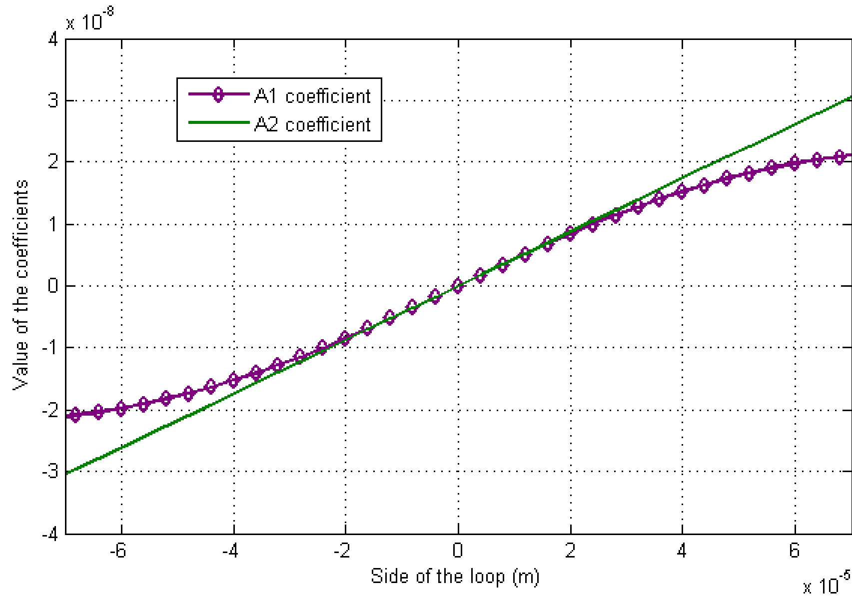

Fig. 3. Variation of the two voltage factors A_1 and A_2 of the emf_m as a function of the displacement ξ_m (for a =1.5 mm, ε=107 µm and I=5 mA).

For a medium frequency of $f_p=\omega_p/2\pi=1\ kHz$ and a maximal displacement equal to $h=10\ \mu m$, we can obtain a sinusoidal emf_m voltage in the μV range. In the final equation (10), the produced emf_m is nearly inversely proportional to the distance ε_a^2. Then, the two inductors L_1 and L_2 are required to be placed as closest as possible in order to obtain the maximal output voltage. The minimal distance is limited by the chosen technology and geometry. We can also notice that, the frequency of the output signal is doubled, compared to the incident pressure wave. This is due to the induction B_x that is an odd function of the z variable. This effect is not desirable in the microphone operation. We can reach a high voltage value if we have a large displacement but the signal includes higher harmonics. Then, we need to use filtering to recover the original signal.

3.2. Case 2: Both Inductors Carrying a DC Bias Current

In this case we consider the two inductors, inner and outer, carrying a DC bias current. This will form a tapped monolithic planar transformer illustrated in Fig. 4. The tapped transformer is actually as an inductor with a middle tap that occupies a separate region. Although it is not a symmetric design, it allows a large range of turn ratios to be realized [14, 15]. Some circuits (e.g., differential circuits) prefer transformers with a symmetrical layout. Depending on whether the lateral or vertical magnetic coupling is used, transformer structures have two categories: planar or stacked. Tapped transformer relies only on lateral magnetic coupling only. All windings can be implemented with the top metal layer, thereby minimizing terminal-to-substrate parasitic capacitances. The relationship between the secondary voltage and the primary and secondary currents is given by the following equation.

$$v_2 = \frac{dL_2 i_2}{dt} + \frac{dM i_1}{dt}, \tag{11}$$

where L_2 is the total inductance of the inner (secondary) spiral inductor and M is the mutual inductance between the segments of the inner inductor and the outer (primary) one, L_1. For a DC biasing of both inductors, equation (11) can be simplified to [16]:

$$v_2 = i_1 \frac{\partial M}{\partial t} \tag{12}$$

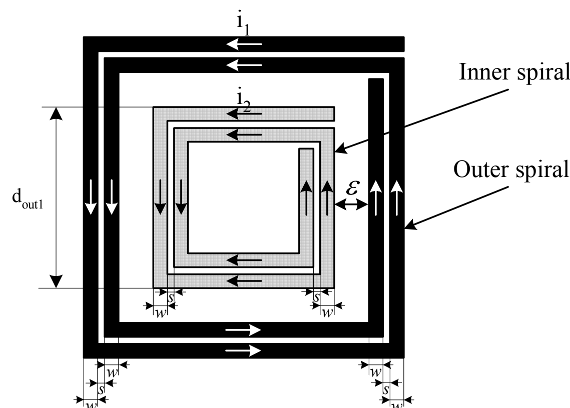

Fig. 4. Planar tapped transformer structure.

The key parameters in inductors design involve the outer dimensions, width and spacing of the metal tracks, thickness of the metal, number of turns of the spiral and the substrate material. The total

inductance of a planar spiral square inductor can be calculated according to the Greenhouse theory [17]. Basically, the spiral square inductor is split up into sections consisting of straight conductors. Then the self-inductance L_i of each section is calculated and summed up. Beside the consideration of the self-inductance of each straight conductor, the mutual inductance interaction $M_{i,j}$ (positive or negative) between the ith and j^{th} parallel segments needs to be included for the calculation of the overall inductance. Hence, supposing that each conductor includes 4 straight segments, the accurate expression of the total inductance L_1 and L_2 can be calculated using the following equation [18]:

$$L_T = \sum_{i=1}^{n} L_i + \sum_{i=1}^{n} \sum_{j=1, j \neq i}^{n} M_{i,j} \tag{13}$$

A practical approximate expression of the expression (13) can be found in [19]. The mutual inductance of two parallel conductors having the same length shown in Fig. 5-(a) is a function of this length l and of the geometric mean distance (GMD), between them. The mutual inductance parameter, M_l, is given by the following equation [17]:

$$M_l = \frac{\mu_0 l}{2\pi} \left\{ ln\left[\frac{l}{GMD} + \sqrt{1 + \frac{l^2}{GMD^2}} \right] - \sqrt{1 + \frac{GMD^2}{l^2}} + \frac{GMD}{l} \right\}, \tag{14}$$

where the *GMD* between the two conductors is approximately equal to the distance between the track center. The exact value of the GMD can be obtained from Grover's tables [29] and is given by the following equation:

$$ln(GMD) = ln(d) - \left\{ \frac{1}{12\left(\frac{d}{w}\right)^2} + \frac{1}{60\left(\frac{d}{w}\right)^4} + \frac{1}{168\left(\frac{d}{w}\right)^6} + \frac{1}{360\left(\frac{d}{w}\right)^8} + \frac{1}{660\left(\frac{d}{w}\right)^{10}} + \dots \right\}, \tag{15}$$

where d is the center to center separation between the conductors, and w is the width of the conductors. If we extend our reasoning to the configuration of the spirals in a microphone, we must take into account the instant vertical distance of both coils. The mutual inductance between both inductors spirals depends on the distance between the two surfaces of L_1 and L_2. For an inner inductor moving sinusoidally with an amplitude equal to ξ_m, the maximal distance d becomes $d=\sqrt{(\varepsilon_a^2+\xi_m^2)}$ (Fig. 5-(b)). In a planar inductor corresponding to our microphone design, we have to take into account different segment lengths as shown in Fig. 5-(c). The mutual inductance of two parallel conductors of lengths l and m, can be expressed as a difference of inductances of equal length segments, given by equation (16) [18]. Since each segments of length m are situated symmetrically with respect to the middle of length l, the expression for the mutual inductance can be written in a simplified form as:

$$M_{l,m} = M_{m+p} - M_p \tag{16}$$

To evaluate the total voltage induced in the secondary inductor due to its vertical movement, we will apply the relations obtained for the case of two rectangular conductors to the two inductors shown in Fig. 4. For the segments of the two inductors, parallel current flowing in same phase produces positive mutual inductance components, to which contribute interaction between parallel segments on the same side of a square. If the current is out-of-phase the mutual components contribution is negative, such as current in one side and the other. In our case, the negative mutual inductance can be neglected because of the large diameter of the inner inductance (range of mm). Using symmetry and noting that segments perpendicular to one another have zero mutual inductance, the total mutual inductance interactions can be approximated by an equivalent of $8n_1 \times n_2$ average interactions between segments of average length at an average distance. Taking into account these considerations and combining equations. (14), (15)

and (16), the normalized mutual inductance between two parallel concentric loops of different lengths is given by equation (17):

$$M = M_0 + \Delta M\left(\xi_m\right), \tag{17}$$

where M_0 is the normalized steady mutual inductance corresponding to the length separation between the loops of the two in-plane inductors ε_a, and $\Delta M(\xi_m)$ is the additional value of the normalized mutual inductance resulting from a distance between the two surfaces of the two inductors (see Figure 5-(b)). For values used in our design, an accurate representation of equation (17) is given in Fig. 6. The simplified terms of the mutual inductance components for small displacement can be approximated by the use of a series expansion of ξ_m, and results in:

$$M_0 \approx n_1 n_2 \frac{4\mu_0}{\pi}\left[\left(l_a - \varepsilon_a\right) ln\left(\frac{\left(l_a - \varepsilon_a\right) + \sqrt{\varepsilon_a^2 + \left(l_a - \varepsilon_a\right)^2}}{\varepsilon_a}\right) - \sqrt{\varepsilon_a^2 + \left(l_a - \varepsilon_a\right)^2} - \varepsilon_a\left(ln\left(1+\sqrt{2}\right) - \sqrt{2}\right)\right] \tag{18}$$

$$\Delta M\left(\xi_m\right) \approx n_1 n_2 \frac{2\mu_0}{\pi}\left(\frac{\sqrt{\varepsilon_a^2 + \left(l_a - \varepsilon_a\right)^2} - \sqrt{2\varepsilon_a^2}}{\varepsilon_a^2}\right)\xi_m^2 \tag{19}$$

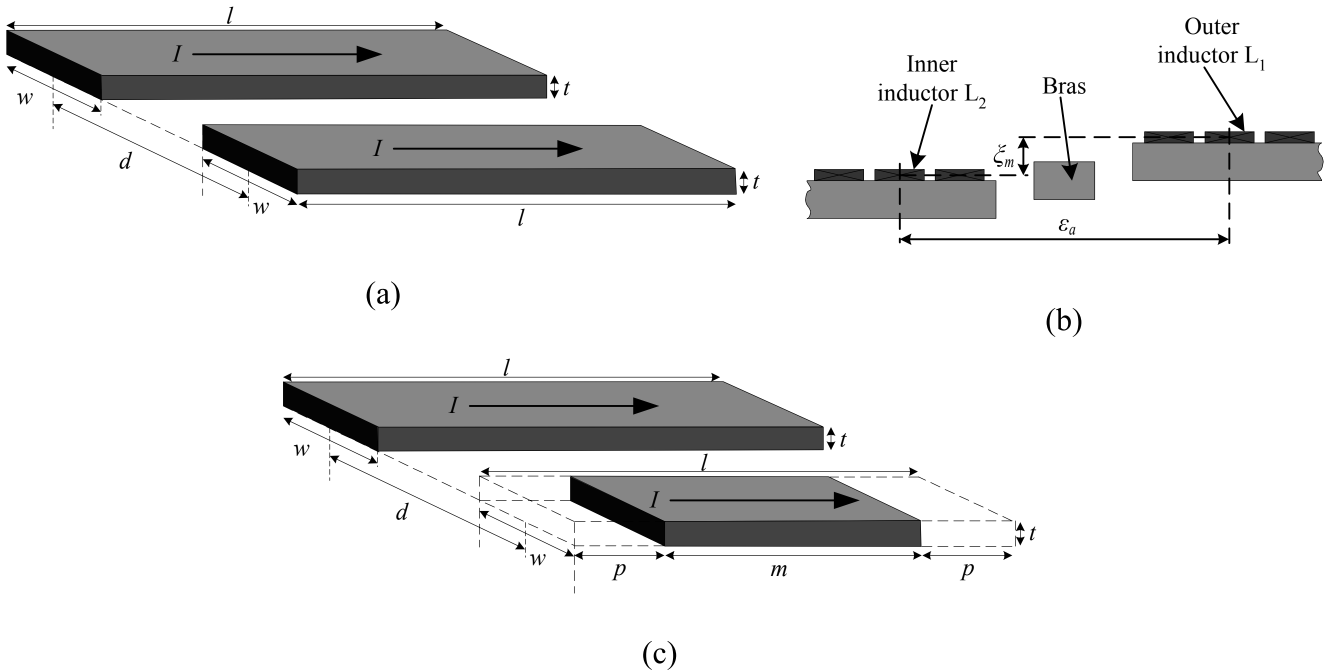

Fig. 5. Layout of two parallel-filament geometry: (a) the basic configuration (b) the situation respecting the conductor movement in a vertical direction, (c) symmetrical configuration of two segments of a different length.

The estimated expression of mutual inductance in equations (18) and (19) represent a parabolic relationship between the motion, ξ_m, of the inner inductor and the resulting mutual inductance. They widely coincide with the one in equation (17) for amplitude displacement less than 50µm (Fig. 6). We notice that, in our case, the mutual coupling coefficient M is quite small (range of µT) because of the high spatial separation, ε_a, between the two inductors.

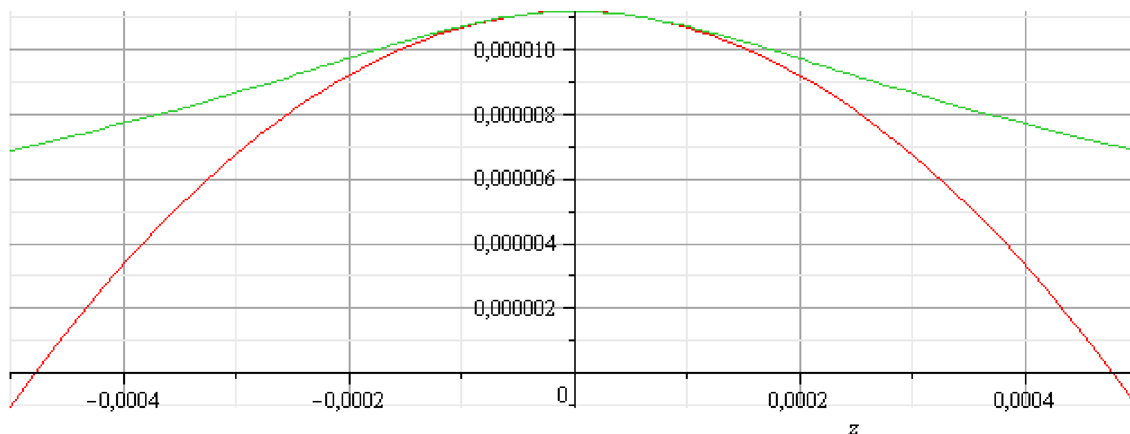

Fig. 6. Variation of the normalized mutual inductance *M* between the two inductors as a function of displacement ξ_m, with length l_a=1.5 mm, ε_a = 107 µm, thickness *w* = 0.8 µm and distance *d* = 1.0 µm.

Thus, Combining equations (12) and (17), the final expression for the normalized open-circuit output voltage of the electrodynamic microphone in the linear range can be obtained by deriving by time $\Delta M(\xi_m)$ as shown in the following equations:

$$v_2 = i_1 \frac{d\Delta M(\xi_m)}{dt} \tag{20}$$

$$v_2 \approx 2 i_1 n_1 n_2 \frac{4\mu_0}{\pi} \left(\frac{\sqrt{\varepsilon_a^2 + (l_a - \varepsilon_a)^2} - \sqrt{2\varepsilon_a^2}}{\varepsilon_a^2} \right) \xi_m \xi_m' \tag{21}$$

$$v_2 \approx 2 n_1 n_2 \frac{\mu_0 i_1}{\pi} \frac{\sqrt{\varepsilon_a^2 + (l_a - \varepsilon_a)^2} - \sqrt{2\varepsilon_a^2}}{\varepsilon_a^2} h^2 \omega_p \sin(2\omega_p t) \tag{22}$$

We can notice that the evaluated expression for the induced voltage v_2 given by the Equation (22), includes and depends on the same variable as the emf_m found in the Section 1, then, the same consequences are applied to v_2. Although, its amplitude represents the double of the emf_m, this can be explained by the mutual effect of the conductors forming the two inductors each one on the other. We can also gain more in linearity that reaches 50µm compared to the emf_m which is linear for only a *20 µm* range. For different transformer structures the mutual coupling coefficient, *k*, equal to $M/\sqrt{(L_1L_2)}$ in the tapped transformer is relatively small *(< 0.4)* [15]. Stacked transformers have a higher mutual coupling, because of the low distance between layers, but we are constrained by the operating principle and the use of standard technology in our electrodynamic microphone.

3.3. Case 3: Outer Inductor Carrying AC Bias Current

To increase the *emf* voltage produced by the acoustic pressure, we can generate a time-varying magnetic field, which will add a voltage component in the inner closed circuit. An electromotive force, as given by Faraday's law, is merely a voltage that arises from conductors moving in a constant magnetic field or a conductor submerged in a time-varying magnetic field or a combination of the two. An AC bias current applied to the outer inductor will generate a time-varying magnetic field, which

will add a voltage component in the inner closed circuit. The voltage produced by the time-varying magnetic field is called transformer action and is defined by [13]:

$$emf_t = -\frac{d\Phi}{dt} = \oint_{loop} \vec{E}.\vec{dl} = -\frac{d}{dt}\iint_S \vec{B}.d\vec{S} \ , \tag{23}$$

where S is the surface bounded by the loop formed by the spires of the inner inductor and E is the electric field. When the outer inductor is driven by an AC current, the total emf_T is given by the superposition of the two contributions defined by equations (5) and (23). The motional *emf* is a function of the velocity and the magnetic flux density, while the electromotive force induced in a stationary closed circuit is equal to the negative varying rate of the magnetic flux.

$$emf_T = \oint_{loop}\left(\vec{v}\wedge\vec{B}+\vec{E}\right)\vec{dl} = \underbrace{\oint_{loop}(\vec{v}\wedge\vec{B}).\vec{dl}}_{\text{Motional induction (generator)}} - \underbrace{\frac{d}{dt}\iint_S \vec{B}.d\vec{S}}_{\text{Transformer induction}} \tag{24}$$

In the AC configuration, the value of the current flowing through the outer inductor will be $i_1=I_0.sin(\omega_c t)$. Where, ω_c is the angular velocity. Thus, the new value of the motional *emf* will be a signal modulated in frequency and given by the following equation:

$$emf_m \approx n_1 n_2 \frac{4(\sqrt{2}-1)\mu_0 a}{\pi \varepsilon^2} I_0 \omega_p .h^2(\omega_p).sin(2\omega_p t).sin(\omega_c t) \tag{25}$$

This also can be expressed by the equation below:

$$emf_m \approx n_1 n_2 \frac{2(\sqrt{2}-1)\mu_0 a}{\pi \varepsilon^2} I_0 \omega_p .h^2.\left[cos(\omega_c t - 2\omega_p t) - cos(\omega_c t + 2\omega_p t)\right] \tag{26}$$

For the transformer induction, the corresponding *emf* will be evaluated by the following equation:

$$emf_t = -\frac{d}{dt}\iint_S B_z \, dx\, dy \tag{27}$$

To simplify the equation (27), for low amplitude displacement, supposing that B_z varies only in time and not in space, we can find a sinusoidal induced voltage of the same frequency as the has the polarization current. A series expansion neighboring zero of the *emf* can be evaluated by the following equation:

$$emf_t = -n_1 n_2 \frac{\mu_0 a}{16\pi}\left(ln\tfrac{a}{\varepsilon}+2\right) I_0 \omega_c \, cos(\omega_c t) \tag{28}$$

Depending on the angular velocity of the current in the outer inductor, we can achieve a total emf_t value that is higher or lower than the value of a motional emf_m. However, our goal is to increase the induced voltage produced by the displacement of the membrane. Then, we can choose ω_c that will result at a multiplication factor greater than 2. In the case shown by Fig. 7 we can reach an emf_T voltage 3 times higher than the emf_m with two harmonic signals, the first have an angular velocity at ω_c-$2\omega_p$ and the second at ω_c+$2\omega_p$. This was done for a displacement of 20 µm, f_p=*1 kHz* and f_c=*10 kHz*. We can pass the resulting wave through a band-stop filter, to recuperate the original signal

proportional to the membrane deflection, permitting only the angular velocity of ω_c-2ω_p and/or ω_c+2ω_p and rejecting the others.

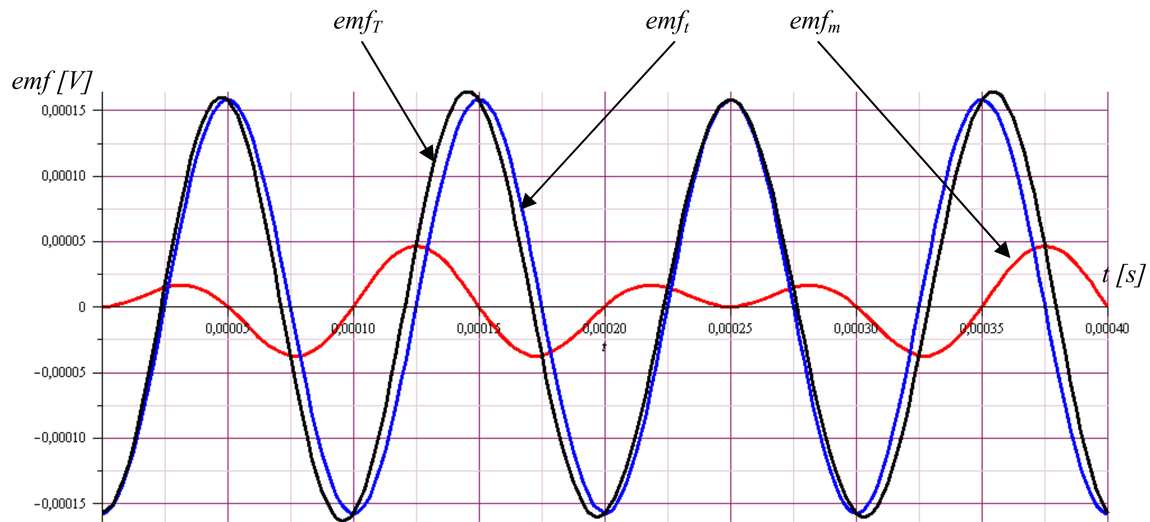

Fig. 7. Voltage form result, emf_T, of the sum of the emf_t and the emf_m.

4. Use of the Dual-layer Spiral Inductor

To improve the *emf* produced by the internal inductor and also to enhance the B-field we can use the symmetric dual-layer spiral inductor, structure shown in Fig. 8. Mutual inductance between adjacent spirals increases the total inductance of the multi-layer spiral inductor. In comparison with the conventional single-layer spiral inductors, theory and measurements show that, for a given silicon area, the dual-layer inductor provides the inductance value nearly three times higher than that of the single-layer inductor [22, 23]. The use of this kind of inductance structure in our microphone structure can lead to an *emf* which is four times higher than that given in the Equation 10. This can easily be seen since the number of inductor segments has been multiplied by two for both inductors.

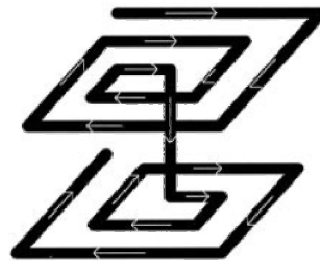

Fig. 8. Symmetric dual-layer spiral inductor structure.

5. Conclusion

We have shown a new structure of a monolithic CMOS integrated inductive microphone. It is based on the variation of the mutual inductance between an outer fixed inductor and an inner suspended inductor. This inner inductor is designed on a suspended membrane attached with four arms. Detailed electromagnetic modeling and calculation was done to find solutions leading to the increase of the

induced voltage gained in the inner inductor. A prototype chip fabrication of this inductive microphone has been done in a CMOS compatible process using a 0.6 μm technology. The electromagnetic study shows that the use of an AC current in the primary fixed inductor gives a larger induced voltage in the inner inductor. The induced voltage evaluated, in the μV range using a DC current solution, is found to be multiplied by a factor of 3 to 6 when using the alternative current. The output voltage of the electrodynamic microphone that achieves the μV range can be increased by the use of the symmetric dual-layer spiral inductor structure.

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Guide for Contributors

Aims and Scope

Sensors & Transducers Journal (ISSN 1726-5479) provides an advanced forum for the science and technology of physical, chemical sensors and biosensors. It publishes state-of-the-art reviews, regular research and application specific papers, short notes, letters to Editor and sensors related books reviews as well as academic, practical and commercial information of interest to its readership. Because it is an open access, peer review international journal, papers rapidly published in *Sensors & Transducers Journal* will receive a very high publicity. The journal is published monthly as twelve issues per annual by International Frequency Association (IFSA). In additional, some special sponsored and conference issues published annually. *Sensors & Transducers Journal* is indexed and abstracted very quickly by Chemical Abstracts, IndexCopernicus Journals Master List, Open J-Gate, Google Scholar, etc.

Topics Covered

Contributions are invited on all aspects of research, development and application of the science and technology of sensors, transducers and sensor instrumentations. Topics include, but are not restricted to:

- Physical, chemical and biosensors;
- Digital, frequency, period, duty-cycle, time interval, PWM, pulse number output sensors and transducers;
- Theory, principles, effects, design, standardization and modeling;
- Smart sensors and systems;
- Sensor instrumentation;
- Virtual instruments;
- Sensors interfaces, buses and networks;
- Signal processing;
- Frequency (period, duty-cycle)-to-digital converters, ADC;
- Technologies and materials;
- Nanosensors;
- Microsystems;
- Applications.

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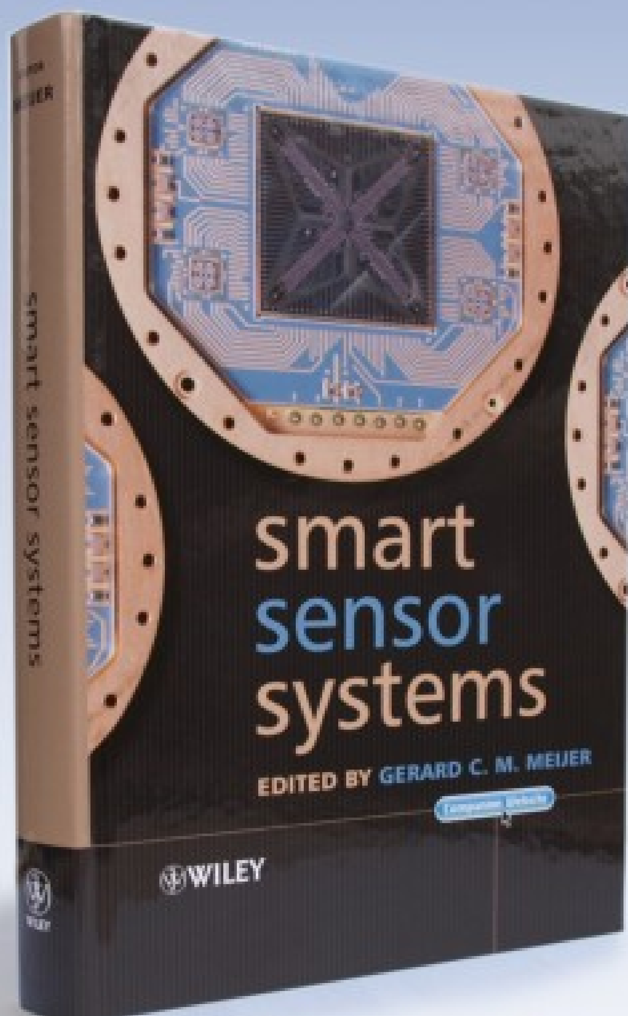

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