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DESIGN REQUIREMENTS FOR SENSORS BASED ON AMORPHOUS WIRES RESULTING FROM MEASUREMENTS AND SIMULATION

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Abstract: Amorphous wires are attractive components for sensor and micro-sensor realization. The negative factors that affect the behaviors of these sensor types must be however highlighted and compensated throughout adequate tools. The characteristics of the sensing element can be carried-out starting from the measurement results. A model of a magneto-impedance sensing element is obtained and used for a PSPICE simulation next step. The paper presents the results obtained at radio-frequency, at once with the mathematical modeling of the sensitive part. In addition, there are also the requirements which must be fulfilled in the design of this kind of sensors.

Keywords: amorphous wire, current sensor, magneto-impedance, vectorial measurement.

1. INTRODUCTION

A proposal of non-contact current sensors was investigated through vectorial measurements obtained with instrumentation set-up working, between 1 MHz and 100 MHz [1]. The measurement method is an alternative to the method presented in [2] for evaluating the response of soft magnetic materials from 1 MHz to 6 GHz. As Fig. 1 shows, the sensitive element is made with magnetic amorphous wires. These wires have a diameter and a length of 120 μm and of 20 mm respectively, while the chemical composition is $(Co_{94}Fe_6)_{72.5}Si_{12.5}B_{15}$. Because the element is axially mounted inside a solenoid [1], the magnetic field H_{ex} created by the measurand, the dc-current, becomes intermediary quantity acting on the magneto-impedance (MI) components [3]. The impedance Z seen by an RF current passing through the MI element in presence of a strong skin effect ($\delta << a$, skin depth δ much smaller than the wire radius a) is:

$$Z = \left(a/2\sqrt{2\rho}\right)R_{dc}\left(1+j\frac{\rho\ell}{\pi a^2}\right)\sqrt{\omega\cdot\mu_\theta(f,H_{ex})} \qquad (1)$$

where R_{dc} is the dc resistance of the element, ρ is the resistivity, ω is the angular frequency of the alternating current flowing through the element, ℓ is the amorphous wire length and μ_θ is the circumferential maximum differential permeability.

We propose a current meter based on a quantity conversion inside a coil fed by the same unknown current and by means of a MI sensor [1]. The device scheme is depicted in Fig. 1.

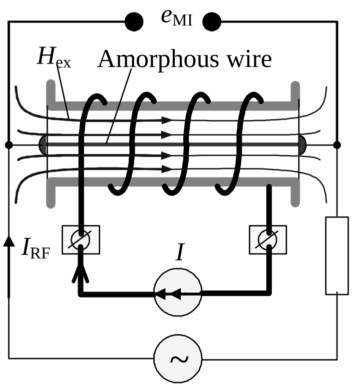

Fig. 1. The schematic arrangement of the current sensor

The coil of the current sensor is designed for obtaining an inner magnetic field H_{ex} - expressed in Oe – ten times the value of the input current expressed in A. From measurement data [1], the dependence was obtained between the sensor input current I and the MI components of the same the sensor, i.e. resistance R, inductance L and impedance magnitude $|Z|$. The following two diagrams show this dependence for a test current of 1 mA at 36 MHz. Although data were acquired in a large range of the magnetic field (its upper limit was over 6 Oe), only a reduced range, limited by the skin conditions, was selected for modeling before and then for simulating the sensor behaviors.

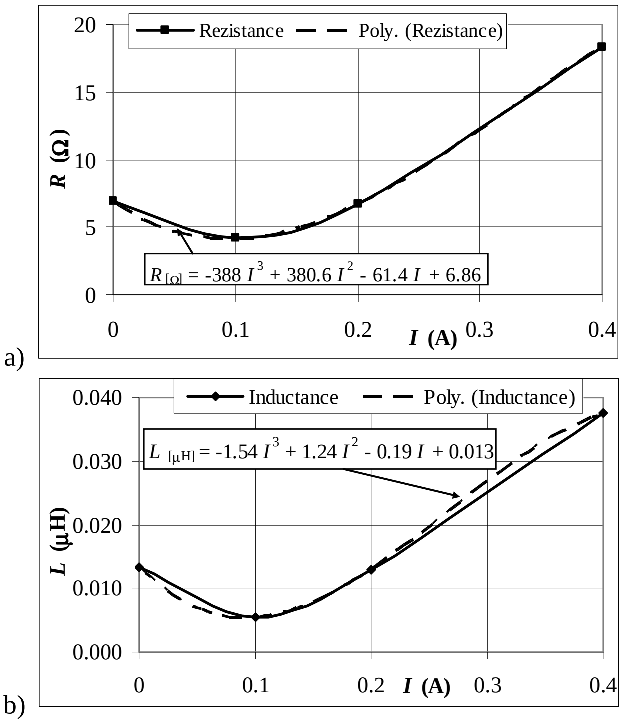

Fig. 2: The experimental results at 32 MHz for resistance - a), and inductance - b), of MI together with the corresponding trend lines used in simulation.

In the range 0 - 0.4 A of the input quantity I, the relevant quantities of the MI are described by the fitting 3rd order polynomial functions:

$$R_{[\Omega]} = -388\, I^3 + 380.6\, I^2 - 61.4\, I + 6.86 \qquad (2)$$

$$L_{[\mu H]} = -1.54\, I^3 + 1.24\, I^2 - 0.19\, I + 0.013 \qquad (3)$$

These two laws were used in the simulation of a Collpits oscillator, adequate to implement the sensing capabilities of the amorphous wires. The main requirements for the current sensor taken in account in the simulation step are: low RF test current, bias in an adequate quiescent point for the sensing element and by using of the negative feedback for restrain the nonlinearity in the transfer curve.

2. SIMULATION STAGE

A model for these sensors is neither simple to find and not unique, due to their nonlinearity. A general model for the investigated sensor is given in Fig. 3. This is in accordance with the informational and energetic models of the sensors [4] implementing the relation between the input quantity – current– and the output quantity – impedance. The first part of this model has a linear dependence law, but the second one is completely non-linear. The magnetic quantity is an intermediary one but it cannot be hidden because this type of quantity is used also for biasing and for feedback. The ambient temperature, T_a, is also considered, but it is difficult to take into account its real contribution [3]. Both magnetic fields, for bias and for feedback, are created passing electric currents through known coils. In this case, the model of the sensor becomes a converter [4] instead of two linked transducers. In the simulation step we used this last model.

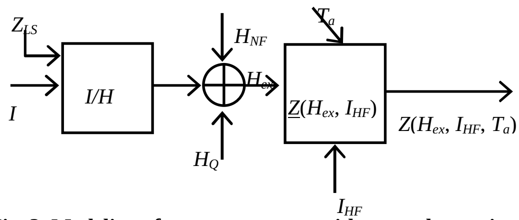

Fig. 3. Modeling of a current sensor with amorphous wire by two transducers; H_Q - biasing magnetic field, H_{NF} - negative feedback field

2.1. PSPICE simulation circuit

The current sensor was considered as part of a Collpits oscillator, Fig. 4, realized with a dual-gate MOSFET M1, the inductance L1, and the capacitors C1 and C4. The MI effect of the sensing part was considered through the transfer function of an E-type multiplier [5], E1, having as input quantities the RF test current, converted at voltage by H1, and modeling (2) with an H-type dual port, H2. The MI inductance variation, small in comparison with L1 value, is neglecting in this PSPICE simulation.

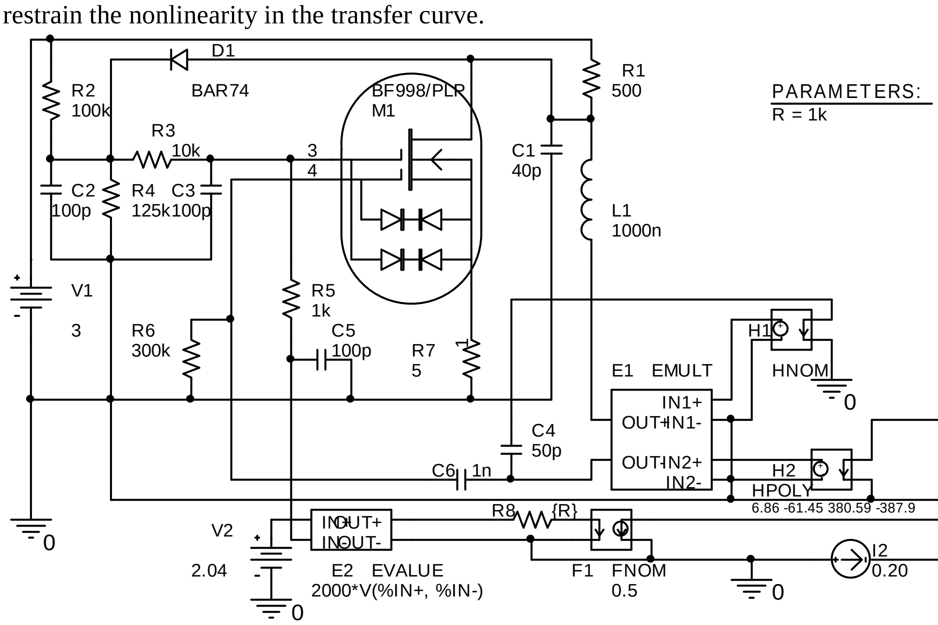

Fig. 4. Electrical diagram for PSPICE simulation

The behavior of the modeling part for MI is justified by the following equation:

$$
\begin{aligned}
Z_{E1_{\mathrm{OUT}}}(I_1) &= \frac{V_{E1_{\mathrm{OUT+}}} - V_{E1_{\mathrm{OUT-}}}}{I_{E1_{\mathrm{OUT}}}} = \frac{V_{E1_{\mathrm{IN1}}} \times V_{E1_{\mathrm{IN2}}}}{I_{E1_{\mathrm{OUT}}}} \\
&= \left(-388\, I_{H2}^{3} + 380.6\, I_{H2}^{2} - 61.4\, I_{H2} + 6.86\right) \frac{I_{H1}}{I_{E1_{\mathrm{OUT}}}} \quad [\Omega]
\end{aligned} \tag{4}
$$

It can be considered the same quantity for both factors of the last ratio in (4) because the current through C6 is insignificant. Therefore, the expression for the output impedance of E1 expressed in ohms becomes similar with (2) for resistive part of MI. The input quantity for the sensor is a sum between the current of the generator I1, the biasing current I_2 and the compensation current I_{F1}. The influence of MI parameters in the output oscillation signal is in both frequency and amplitude:

$$
f = \frac{\sqrt{\dfrac{1}{C_1} + \left(1 + \dfrac{R_{MI}}{Ri_{G1_{M1}}}\right)\dfrac{1}{C_4}}}{2\pi\sqrt{L_1 + L_{MI}}} \tag{5}
$$

where $Ri_{G1_{M1}}$ is the small signal input impedance at the first gate of M1. The changes in MI parameters are sensed by a peak detection of the oscillation, realized by a serial detector constituted from D1-C2. A double filtration realized by R2-C2 and R3-C3 assures a good DC signal for both negative feedback loops. One of it acts through the second gate of M1 and keeps the signal amplitude in a close range. The other loop, constituted by E-type dual port E2, conversion resistor R8 and F-type dual port F1, make possible a compensation of the input quantity I_1. The current of F1 will be, at the balance, a measure of the input current I_1. Two thresholds are used: one for the peak detector, realized by R2-R4 divider, and the other for the compensation loop – realized by the voltage source V2. Their role is complex: in settling time value and in linearity of the transfer function $I_{F1}(I_1)$.

2.2. Simulation results

The simulation results presented in Fig. 5 were obtained with a negative feedback loop having 10^3 gain, biasing the amorphous wire in a 0.2 A quiescent point and for 0.1 A step variation in a range of 0.4 A for the input current I_1. Fig. 5 shows how the control loop acts in transient regime, the settling time *vs.* input current and attest a linear transfer characteristic at the balance; traced

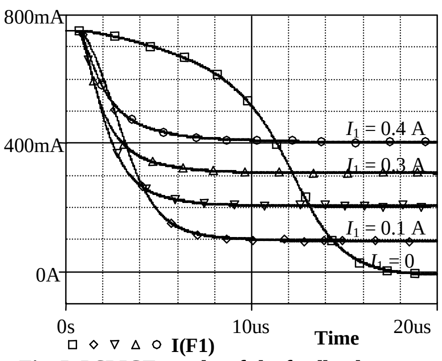

Fig. 5. PSPICE results of the feedback current for various input currents.

for long-standing regime, this transfer characteristic shows like in Fig. 6. This line has less than one percent deviation from the ideal unitary value in a twice range than necessary one.

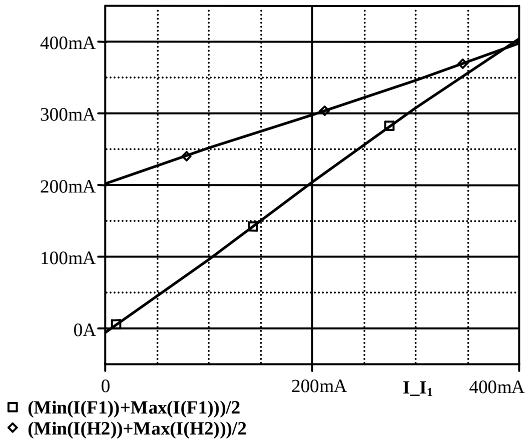

Fig. 6. Transfer characteristic of the feedback current I_{F1} *vs.* I_1.

Due to increasing of the settling time at the full balance by negative feedback, a partial compensation was used. A compromise between the dynamic range and the settling time was obtained in a quiescent point imposed by the bias current I_2 = 0.2 A and a gain of the voltage to current conversion of 10^3. In this manner, the slope of the transfer characteristic for the resulting current I(H2) is 0.5, two times lower than without negative feedback. Moreover, the corresponding magnetic field in the amorphous wire will have a half dynamic range too. Consequently, the input quantity can be in a wide variation interval or the linearity will be better.

Because of this incomplete compensation method applied, many other parameters do not remain constantly. One of them is the resistive part of the amorphous wire impedance; its variation is following a straight line limited between 7 Ω and 18 Ω, show in Fig. 7. This variation will affect the parameters of the oscillation signal.

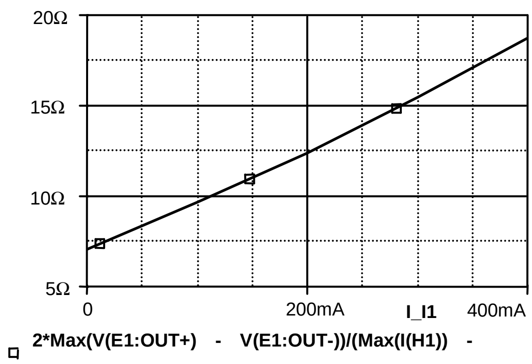

Fig. 7. Amorphous wire resistance variation *vs.* I_1.

First of all, it is important to know the variation of the frequency. The variation of the test frequency generated by Collpits oscillator can modify the value of the amorphous wire parameters because there is a small dependence. The signal period variation shows in Fig. 8 proves frequency change under ±0.8% and do not justify a complicated PSPICE model of the amorphous wire.

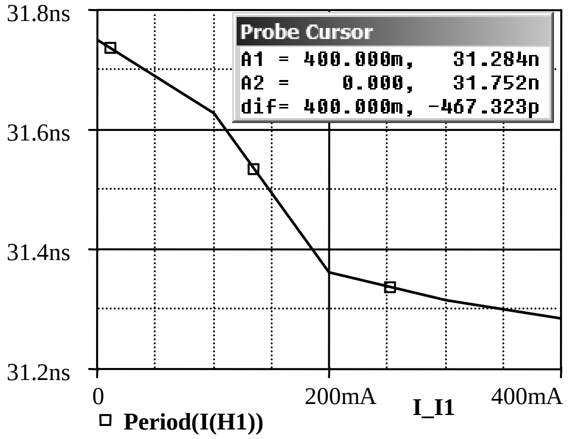

Fig. 8. Period of oscillation of the test current I_{H1} *vs.* I_1.

The negative feedback loop is not a selective one and the test signal purity can become important. A FFT analysis of the test current, represented in Fig. 9, shows the first harmonic of this signal below -20 dB than the fundamental. A global parameter for the test signal purity, like distortion coefficient, can be computed:

$$\delta = \sqrt{\left(\frac{A_c}{A_2}\right)^2 + \left(\frac{A_c}{A_3}\right)^2 + \ldots} = \sqrt{10^{-\frac{(A_c - A_2)_{dB}}{10}} + 10^{-\frac{(A_c - A_3)_{dB}}{10}} + \ldots} \quad (6)$$

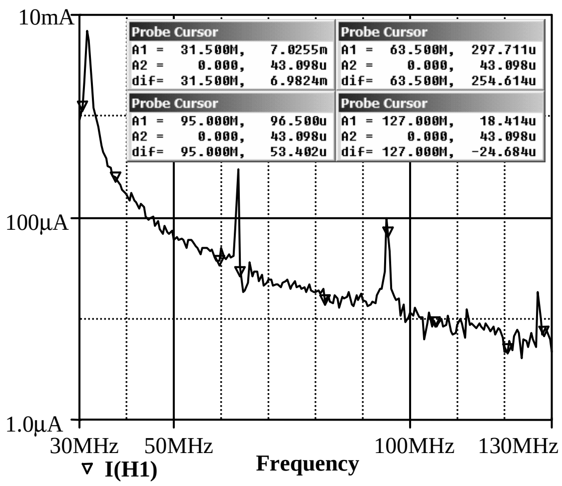

Fig. 9. Power spectral density of the test current I_{H1}; parameter - I_1.

In the middle of the considered input range, I_1 = 0.2A, this coefficient had the value of 4.5% and can be considered well enough; the same coefficient of the drop voltage across amorphous wire was a little bit greater, under 4.8%. These values it allows to take in account only the fundamental component of the test signal and justify the use of non-selective leveling and negative feedback loops. The dependence of the test current effective value, Fig. 10, shows a decreasing of 1.8 mA for the wide input range but values situated under the proposed limit of 10 mA.

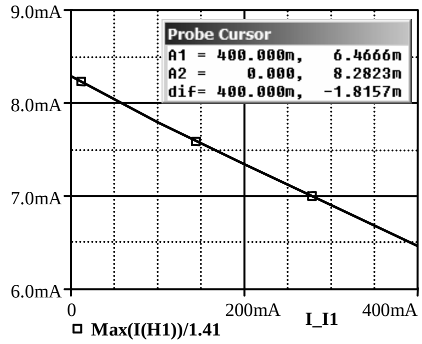

Fig. 10. The effective value of the current I_{H1} through the amorphous wire *vs.* I_1.

3. CONCLUSION

The characteristics of the amorphous wires can be theoretically modeled, but vectorial measurements are important for obtaining the fitting laws in the real conditions. Starting from the experimental data [1], the mathematical, informational and energetic models are developed as a first step in PSPICE modeling for a current sensor based on amorphous wires. In the simulation phase, amorphous wires based sensors nearly complete behaviors were obtained. The resulted fast balancing loop, obtained by applying an incomplete compensation method, has good behaviors proved by many quality parameters. Moreover, not easy to do in current measurements are obtained as function of the dc input quantity. A compact non-contact current sensor and its electronic processing unit, designed on Collpits oscillator structure, can be developed starting from the presented simulation diagram. The aim for the future is to extend the study of this current sensor in ac and to find the requirements for signal processing in this case.

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