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Maxwell-Wien bridge with vector voltmeter system for measurement small and rapid changes in inductive-loop sensor impedance components

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ABSTRACT

The paper presents a low-cost electronics system for impedance measurement of an inductive-loop sensor as well as small, rapid changes in impedance components. The idea of the system is based on the adaptation of a vector voltmeter instead of the Maxwell-Wien bridge balance indicator and supporting the system with a microcontroller. The proposed procedures, based on the complex voltage of bridge unbalance, allow to calculate the resistance that can balance the bridge in a short time. This, in turn, allows to measure the sensor impedance components, identically, as in the case of a well-known balanced bridge method. However, what is original in this work, is that measurements can be taken in a state of bridge imbalance and with high relative accuracy to small changes in sensor impedance.

1. Introduction

Metal objects, such as motor vehicles, passing through the magnetic field of the inductive-loop (IL), cause fluctuations to the sensor impendence both real and imaginary parts [1]. These fluctuations, which reach a few percent of nominal impedance values, provide useful information for vehicle traffic detection and measurement systems [2]. On the other hand, the narrow IL sensor used for vehicle axle detection requires more sensitive impedance measurement systems, since relative changes in these parameters are smaller than 1% of their nominal values [3].

Changes in the mentioned values represented as a function of time travelled by the vehicle passing over the IL are called the magnetic profile or magnetic signature of the vehicle. In many applications, the magnetic profile is essential for vehicle classification [4].

The Maxwell-Wien bridge is characterized by a simple construction and high sensitivity [5,6]. Based on the output voltage, the bridge allows to detect very small changes in the sensor impedance components. Obtaining a precise result (resistance and inductance values) using the traditional Maxwell-Wien bridge measurement method requires high-resolution adjustable resistors, time-consuming balancing and constant sensor impedance during the measurement. This is generally considered a disadvantage of bridge methods.

The next section presents a proposition of measurement method free of these drawbacks. The numerical circuit models were presented, followed by system hardware implementations and prototype verification, supported by experimental results and conclusions.

2. The proposition of the measurement method

In the layout shown in Fig. 1, the Maxwell-Wien bridge constructed from R_{1h},R_{2h}, C, R_4 with known reference values, and Sensor with unknown values L_3,R_3 is powered by the sinusoidal generator with a known voltage $\underline{E}$ and a known constant frequency f.

Microcontroller (μC) can change the values R_{1h},R_{2h} in hardware-determined resolution and range. This is possible thanks to digitally controlled potentiometers. The numerical *RRC* model of a circuit consisting of the R_{1h}, R_{2h}, C elements which is clearly marked in Fig. 1, allows to calculate the $\underline{U}_2$ complex voltage value

$$\underline{U}_2 = RRC(R_{1h},R_{2h},C,\underline{E},f) \tag{1}$$

The numerical model of the RRC circuit and its inverse model are presented separately in the next section. Alternatively, the $\underline{U}_2$ can be measured by VVM. However, this method would require the use of additional switches. In order to avoid this, the *RRC* model was used. Based on the measured voltage $\underline{V}$, the voltage $\underline{U}_4$ can be simply calculated as:

$$\underline{U}_4 = \underline{U}_2 + \underline{V} \tag{2}$$

The Maxwell-Wien bridge is balanced when $\underline{V} = (0 + i0)$V, i.e. when $\underline{U}_4 = \underline{U}_2$. The theoretical R_1 and R_2, which can balance the bridge much more accurately than R_{1h} and R_{2h}, are calculated using the *invRRC* function:

$$[R_1,R_2] = invRRC(\underline{U}_4,C,\underline{E},f) \tag{3}$$

Numerically calculated values of R_1 and R_2, which in theory

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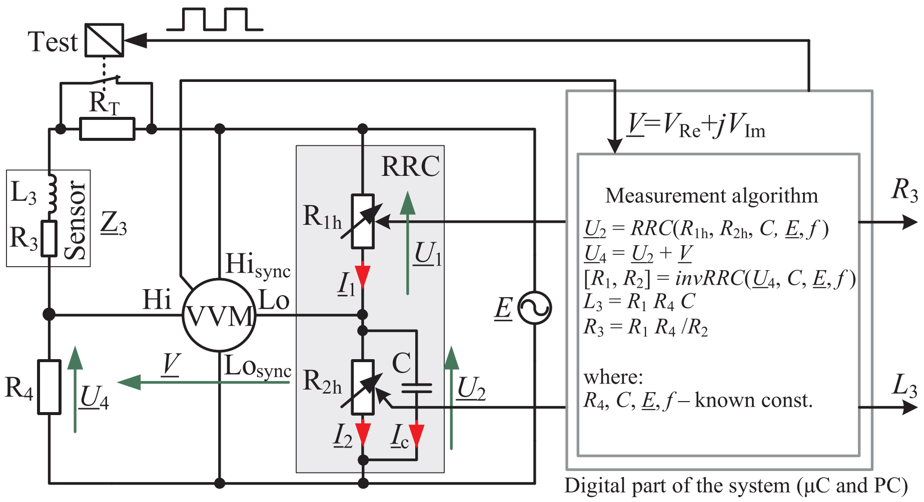

Fig. 1. The Maxwell-Wien bridge system with vector voltmeter (VVM), measurement algorithm, and with extra R_T to test phase error of VVM.

perfectly balance the Maxwell-Wien bridge are used to calculate the R_3 and L_3 values of the Sensor impedance parameters from:

$$\begin{aligned} R_3 &= R_1 R_4 / R_2 \\ L_3 &= R_1 R_4 C \\ \underline{Z}_3 &= R_3 + i\,2\pi f L_3 \end{aligned} \tag{4}$$

It is important that the numerical R_1 and R_2 values allow the bridge to be balanced in one cycle by setting up hardware-capable (near-available) resistance R_{1h} and R_{2h}. This avoids the time-consuming, gradual balancing, as is the case of the standard Maxwell-Wien bridge measurement method.

3. The RRC and invRRC numerical models

It is assumed that the distribution of currents and voltages, in the RRC circuit, in steady state, is as shown in Fig. 2. It is assumed that the current flowing through the VVM is zero.

Consequently, for known values of $R_{1h}, R_{2h}, C, \underline{E}, f$, voltage $\underline{U}_2$ can be easily calculated numerically, implementing Algorithm 1 in program language that supports complex numbers (e.g. Matlab).

Algorithm 1. Model of RRC circuit

```
function  $\underline{U}_2 = RRC(R_{1h}, R_{2h}, C, \underline{E}, f)$
    $X_C = 1/(i\,2\pi f\ C)$;
    $\underline{Z}_2 = (R_{2h}\,X_C)/(R_{2h} + X_C)$;
    $\underline{U}_2 = \underline{E}\ \underline{Z}_2/(R_{1h} + \underline{Z}_2)$;
end
```

Inversely, in order to obtain the desired $\underline{U}_2$ voltage at the RRC circuit output, a numerical model for calculating R_1 and R_2, represented by Algorithm 2, was derived and implemented.

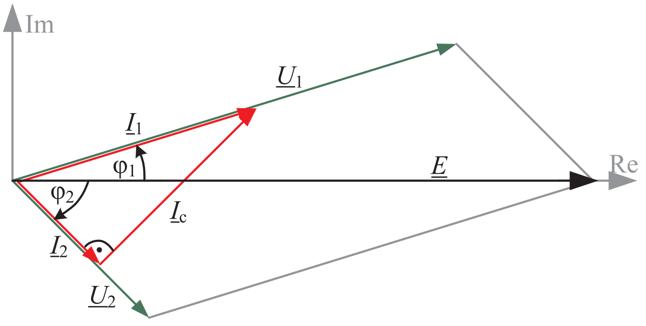

Fig. 2. Phasor diagram of voltages and currents in the RRC circuit (marked in Fig. 1).

Algorithm 2. Solver of inverse RRC problem

```
function $[R_1, R_2] = invRRC(\underline{U}_2, C, \underline{E}, f)$
    $\underline{U}_1 = \underline{E} - \underline{U}_2$;
    $\varphi_1 = \mathrm{atan}(imag(\underline{U}_1)/real(\underline{U}_1))$;
    $\varphi_2 = \mathrm{atan}(imag(\underline{U}_2)/real(\underline{U}_2))$;
    $U_{1m} = \mathrm{abs}(\underline{U}_1)$;
    $U_{2m} = \mathrm{abs}(\underline{U}_2)$;
    $X_{Cm} = 1/(2\pi f\ C)$;
    $R_1 = \sin(\varphi_1 + \mathrm{abs}(\varphi_2))\,X_{Cm}\,(U_{1m}/U_{2m})$;
    $R_2 = \tan(\varphi_1 + \mathrm{abs}(\varphi_2))\,X_{Cm}$;
end
```

4. Hardware implementation

The R_T, located in the sensor branch (Fig. 1), is non-inductive, has a low resistance value, and is bypassed by normally closed, Test switch. This circuit is used to determine the correction value of the VVM phase error.

The bridge uses 10-bit, digitally controlled (via SPI) potentiometers (AD5293) with a nominal resistance of 20 kΩ and 100 kΩ, for R_{1h} and R_{2h} respectively. Given the dedicated operating frequency ($f = 10$ kHz) and the nominal sensor parameters $\underline{Z}_{3N} = (152.4 + i431.5)\,\Omega$, $R_4 = 590\,\Omega$, $C = 963$ pF (pre-measured at 25 °C, $C_N = 1$ nF, ±5%, Polyester) were used.

The VVM [7], shown in Fig. 3, was implemented using two synchronous demodulators (AD698). Lowpass filters were configured to operate at 100 Hz, which limited the fluctuation measurement bandwidth of rapid changes in inductive-loop impedance components. The AD698 also includes a sinusoidal voltage generator that was used to power the bridge. The VVM contains a C-S block, which supplies two signals precisely shifted by 90°. The technique of generating a signal with a frequency 4-times greater than the input frequency, properly dividing it twice by two and synchronizing it with the input signal was used. For this purpose, PLL (4046) and two JK-MS flip-flops (4027) as the frequency dividers, were adopted.

The gain of the input amplifier (Amp) is 2 V/V, the gain in synchronous demodulator channels is 5 V/V, the gain of Amp1 and Amp2 is 100 V/V. The prototype of the device uses 12-bit ADC. The ADC1 and ADC3 measuring ranges are ±10 V, and are used for large bridge unbalances, while ADC2 and ADC4 allow measurement in the smaller ±10 mV range. Due to the limited resolution of R_{1h} and R_{2h}, synchronous demodulators tend to generate a DC voltage offset (even tens of mV) at the outputs. Therefore, in order to avoid saturation of Amp1 and