Impedance Measurements With Second-Order Harmonic Oscillator for Testing Food Sterility

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***Abstract*—In this paper, we present a simple interface circuit for measuring impedance based on a second-order harmonic oscillator. The circuit is intended for testing the sterility of aseptically packed food products, where it must measure the conductivity changes of the packaged food in a nondestructive way. The measured impedance is modeled as a low-ohmic resistor (representing the conductivity of the food) in series with a capacitor (the walls of the food container). The oscillator is built with fast current-feedback operational amplifiers (CFOA). To measure the resistive component of the impedance, an AC-to-DC converter is realized with a wide frequency range analog multiplier. The presented experimental results prove that with this interface circuit an accurate measurement of the impedance components can be achieved in a frequency range up to 10 MHz. The range of the resistive component is from a few ohms up to 200 ohms; the range of the capacitive component is from 50 pF up to 300 pF.**

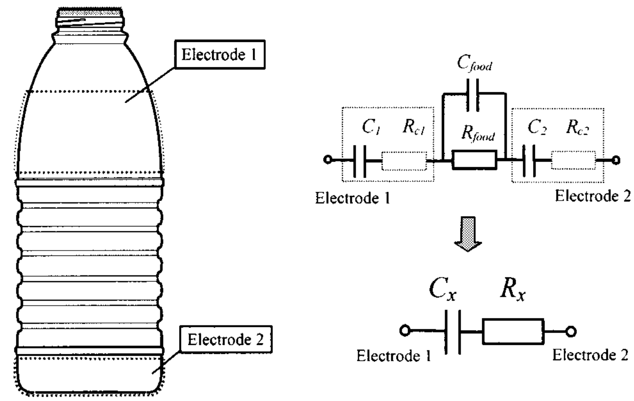

Fig. 1. Plastic bottle with two external electrodes for nondestructive sterility testing and the equivalent electric circuit of the measured impedance.

I. Introduction

CURRENTLY, the most popular way to measure impedance is to apply the auto-balanced bridge technique, also called the vector-voltage-current ratio measurement method [1], [2]. This principle is used in impedance analyzers and RLC meters because of its good accuracy over a wide frequency range [3]. The principle requires at least three sophisticated functional blocks: a tunable harmonic oscillator, an auto-balanced bridge, and a synchronous phase detector. For sensor-interfacing purposes, this solution appears to be quite expensive and complicated. Attempts have been made to optimize and simplify the above-mentioned technique [4] and alternative methods have been sought to interface impedance sensors [5]–[8], but all of them are applicable in a frequency range below 100 kHz.

This paper introduces a novel application-specific impedance-measurement technique for the frequency range up to 10 MHz. This technique uses a second-order harmonic oscillator and is developed for the needs of nondestructive sterility testing of aseptically-packaged, long shelf-life food products [9], [10]. The technique is based on measurement of the conductivity changes that accompany the growth of a large number of bacterial species in food. Because our goal is testing a big part of the production lot and early detection of spoiled food, we are interested in the accurate and fast measurement of the resistive component of the unknown impedance, which is in series with a capacitive component due to the walls of the package. The value of the capacitive component is also informative and is used to test the quality of the food container. The best frequency to sense the conductivity changes of the food vary with the type of the food container, the electrode setting, and the conductivity of the sterile food.

Fig. 1 shows a typical food container for long shelf-life food—a plastic bottle, blow-molded from high-density polyethylene, and the equivalent electric circuit of the measured impedance. Two external electrodes are used to sense the conductivity changes of the food. The values of the coupling capacitors C_1 and C_2, and their loss resistances R_{c1} and R_{c2}, depend on the electrical properties and the thickness of the polyethylene and on the surface of the electrodes. R_{food} and C_{food} represent the electrical properties of the food.

The best sensitivity for R_{food} can be achieved at the frequency at which the influence of the loss resistance in the coupling capacitors (R_{c1} and R_{c2}) and the dielectric properties of the food (C_{food}) are the smallest and we can neglect them. At these frequencies, the equivalent electric circuit (see Fig. 1) can be simplified to one resistor $R_x \approx R_{\mathrm{food}}$ and one capacitor $C_x \approx C_1C_2/(C_1+C_2)$, in series. For example, with the setting in Fig. 1, the coupling capacitors C_1 and C_2 have values of approximately 250 pF, R_{food} for sterile UHT-milk is about 50 Ω, and the best testing frequency is between 1 MHz and 2 MHz. The R_{food} of fish sauce is about 20 Ω, and the best testing frequency is between 3 MHz and 5 MHz.

Section II discusses the principle of operation of the interface circuit. In Section III, the measurement setup is presented. In

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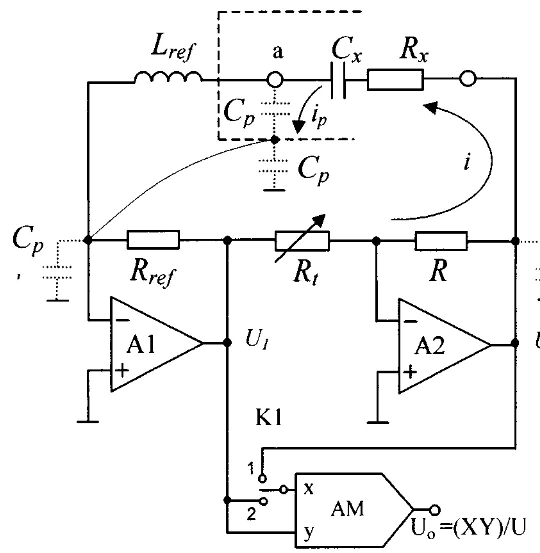

Fig. 2. Simplified impedance interface circuit.

Section IV, the experimental results are presented and discussed. Finally, the conclusions are given in Section V.

II. Principle of Operation

To build a second-order oscillator, we need two basic blocks—a frequency-selective network (resonator) and an undamping circuit that compensates for the energy loss in the resonator and ensures continuous oscillations. Fig. 2 shows a simplified circuit, diagram of the impedance sensor interface. A natural choice for the resonator is a serial RLC circuit in which R_x and C_x are the components of the unknown impedance and L_{ref} is a reference inductor. As an undamping circuit, two inverting stages are used. In this way, low-ohmic input and output are ensured, which eliminates the influence of the parasitic capacitors C_{p1} and C_{p2} at both ends of the resonator. Assuming that the amplifiers A1 and A2 from Fig. 2 are ideal (infinite gain and unlimited bandwidth), the circuit will start oscillating at a frequency that is equal to the resonance frequency of the resonator

$$\omega_o = 1/\sqrt{C_x L_{\mathrm{ref}}} \tag{1}$$

if

$$(R_{\mathrm{ref}} R_1)/(R_x R_t) \geq 1. \tag{2}$$

To ensure stable sine oscillations for an arbitrary value of R_x, we have to tune one of the other three resistors automatically (in Fig. 2 this is R_t) in such a way that it is always true that

$$(R_{\mathrm{ref}} R_1)/(R_x R_t) = 1. \tag{3}$$

The existence of a parasitic capacitance C_p (not shown in Fig. 2) between point $\mathbf{a}$, at which L_{ref} and the measured impedance are connected, and ground will cause a part (i_p) of the loop current i to be directed to ground. This will result in a change of the oscillating frequency

$$\omega_o = 1/\sqrt{L_{\mathrm{ref}}(C_x + C_p)} \tag{4}$$

and will result in a relative error for R_x of

$$\varepsilon_{R_x} = -\frac{C_p}{C_p + C_x}. \tag{5}$$

If the value of this parasitic capacitance is too high, one can use an electric-field shield that is connected to the inverting input of A1 (see Fig. 2); in this way, the initial parasitic capacitance C_p is split in two parts—C_{p3} and C_{p4}. C_{p4} will appear in parallel to C_{p1} and its influence will be reduced significantly, just like that of C_{p1}. The leakage current $\boldsymbol{i}_p$ through C_{p3} will be added to the loop current $\boldsymbol{i}$ and the error for R_x (5) will be eliminated. What remains is an offset error for C_x, which can be eliminated by applying, for example, a multi-signal measurement technique.

The value of R_x can be obtained in several ways. When resistors R_{ref} and R_1 are known, the only value we need to calculate R_x is the value of R_t. At lower frequencies, we could use a digital potentiometer as R_t. In this way, the circuit will automatically generate a digital code at the control input of the potentiometer, which can be used as a measure for R_x. Unfortunately, the big parasitic capacitance at the signal inputs of existing digital potentiometers would pose a serious problem at higher frequencies. Another alternative is to use a MOSFET as a controlled resistor. The dc voltage at the gate of the MOSFET can also be used as a measure for R_x, but then we have to deal with nonlinearity problems caused by the nonlinear behavior of the MOSFET as a voltage-controlled resistor and the temperature dependence of its parameters.

Our choice is to use the MOSFET-transistor as a voltage-controlled resistor, but to measure the value of R_x with a high-frequency analog multiplier (AM in Fig. 2). The measurement is carried out in two steps. First, with switch K1 in position 1 (see Fig. 2), we measure the dc component of the multiple of U_1 and U_2

$$U_{o1} = \frac{U_1 U_2}{U} \tag{6}$$

and then, with switch K1 in position 2, the dc component of the square value of U_1

$$U_{o2} = \frac{U_1^2}{U}. \tag{7}$$

In (6) and (7), the $U_{o1(2)}$ are the dc components of the output voltage of the multiplier and U is a reference dc voltage. The ratio between (6) and (7) equals the ratio between U_2 and U_1, which in turn is equal to R_x/R_{ref}

$$\frac{U_{o1}}{U_{o2}} = \frac{U_2}{U_1} = \frac{R_x}{R_{\mathrm{ref}}}. \tag{8}$$

From (8), the value of R_x can be determined by

$$R_x = R_{\mathrm{ref}} \frac{U_{o1}}{U_{o2}}. \tag{9}$$

There are three basic advantages to measure R_x, as shown above. First of all, by using the ratio of the two output voltages of the analog multiplier we eliminate its multiplicative error, which could be significant at high frequencies. Secondly, the effect of the harmonic distortion caused mainly by the voltage-

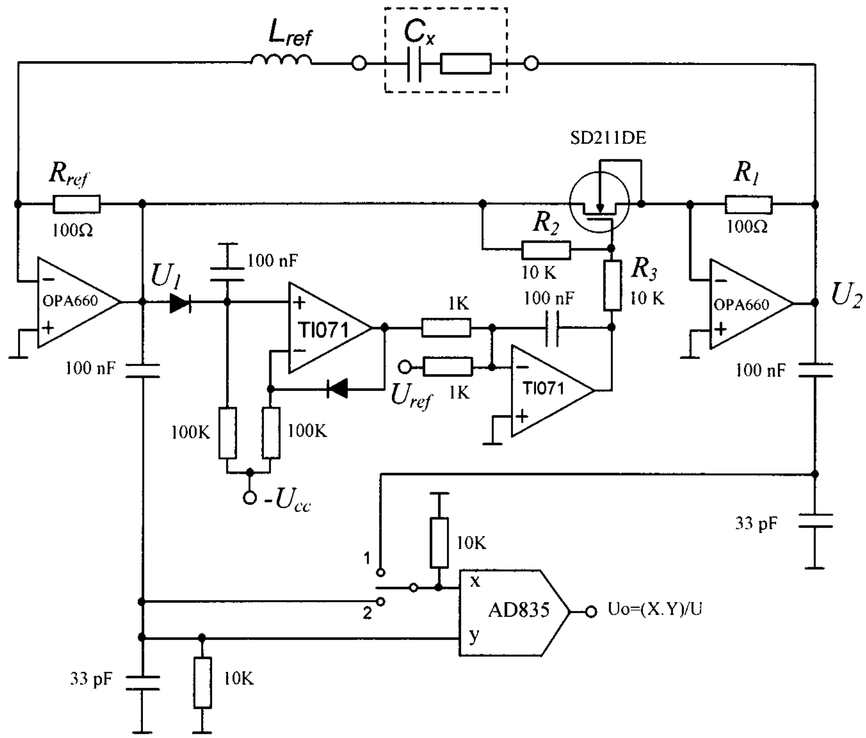

Fig. 3. Circuit diagram of the impedance sensor interface with which the experimental results were obtained.

controlled resistor R_t is considerably reduced. The harmonic distortion of the signal at the output of A2 (U_2) is highest at the highest value of R_t. At the same time, the resonator acts as a filter for the high-level harmonics and its best performance is when R_x has its smallest value and R_t has its largest value. When U_1 is multiplied with U_2, only the first harmonic of U_2 has an impact on the output dc signal.

A third advantage is related to a certain compensation of the error due to the phase shift $\varphi_A = \varphi_{A1} + \varphi_{A2}$ introduced by the amplifiers A1 and A2. This phase shift causes the oscillator to oscillate at frequency ω_g, which differs from the resonance frequency of the resonator ω_o with

$$\omega_g \approx \omega_o \left(1 + \frac{\varphi_A}{2Q}\right) \tag{10}$$

where Q is the quality factor of the resonator. At the same time, when a phase shift of φ_{A2} between U_1 and U_2 occurs, (6) becomes

$$U_{o1} = \frac{U_1 U_2 \cos(\varphi_{A2})}{U}. \tag{11}$$

Taking into account the phase shift introduced by the amplifiers, we can show that (9) changes into

$$R_x = R_{\text{ref}} \frac{U_{o1}}{U_{o2}} \cos(\varphi_{A2}) \sqrt{1 + \tan^2(\varphi_{A1} + \varphi_{A2})}. \tag{12}$$

From (12), it is easily seen that the full compensation effect is achieved when $\varphi_{A1} \ll \varphi_{A2}$.

The value of R_x directly affects the quality factor Q. To ensure that the oscillating frequency ω_g, from which we calculate C_x, remains independent of the value of R_x, we have to use amplifiers with a very small phase shift in the measurement frequency range, as shown in (10).

III. Measurement Setup

Experiments were carried out with the circuit shown in Fig. 3. For a reference inductance L_{ref}, we used molded inductors IM-6-38 (Dale Electronics, Inc.) with the features: small package, epoxy molded construction with superior moisture protection, high Q and self-resonant frequency, low DC resistance, and an insulation resistance higher than 1000 MΩ.

The undumping circuit is realized with two wide frequency bandwidth ($f_{-3\,\text{dB}} \approx 850$ MHz) current-feedback operational amplifiers OPA 660 (Burr Brown). A current-feedback operational amplifier (CFOA) is the best choice in our application for two reasons. First, it has a high-frequency performance that is superior to the performance of the voltage-feedback amplifiers (a higher cut-off frequency, $f_{-3\,\text{dB}}$). This means we have a lower phase shift φ_A in our measurement frequency range. Secondly, the cut-off frequency and the phase shift φ_A change only slightly for different values of R_x. Although a multiplicative error is introduced, the transfer characteristic stays linear and the phase-shift error in (11) can be cancelled by using the multi-signal measurement technique. The relatively high input current noise of a CFOA (a few tens of pA/$\sqrt{}$Hz) is not a serious problem in this application because the loop current we use is in the milliampere range and, besides, the output noise of A2 is filtered by the resonance circuit.

The automatic gain-control circuit (AGC) is realized with two OpAmp TL071 (TI) stages, and the MOSFET SD211DE (Siliconix) is used as a voltage-controlled resistor. By keeping the amplitude of U_1 constant, the AGC circuit fixes the loop current

TABLE I
MEASURED RELATIVE POWER OF THE SECOND AND THE THRID HARMONIC OF U_1 AND U_2

		Harmonics of U_1		Harmonics of U_2	
F, MHz	Rx, Ω	2nd, dBm	3rd, dBm	2nd, dBm	3rd, dBm
2	50	-42	-39	-26	-30
	100	-44	-45	-30	-40
6	50	-32	-43	-19	-30
	100	-35	-49	-30	-46
9	50	-30	-42	-20	-38
	100	-31	-47	-27	-45

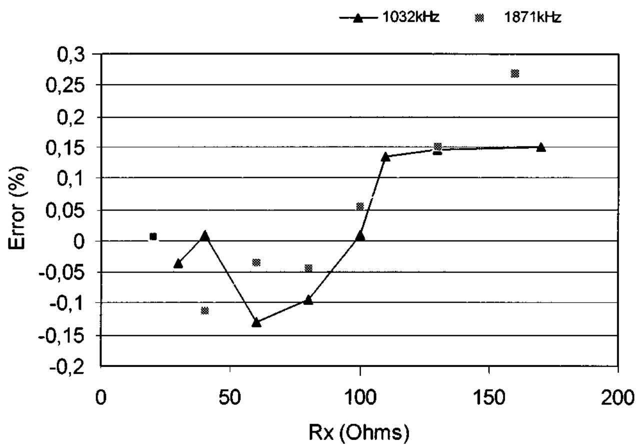

Fig. 4. Relative error of the measured resistive component for lower frequencies.

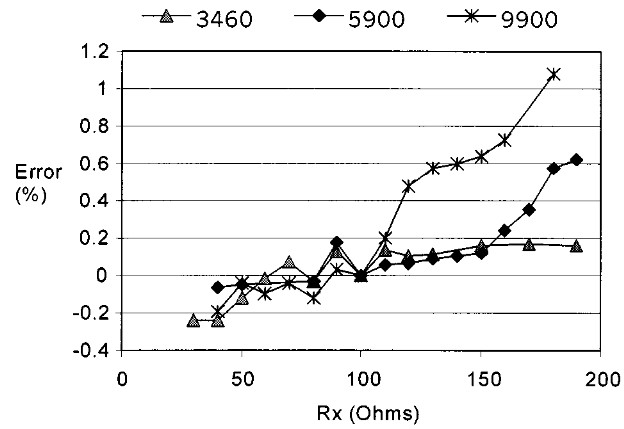

Fig. 5. Relative error of the measured resistive component R_x at higher frequencies.

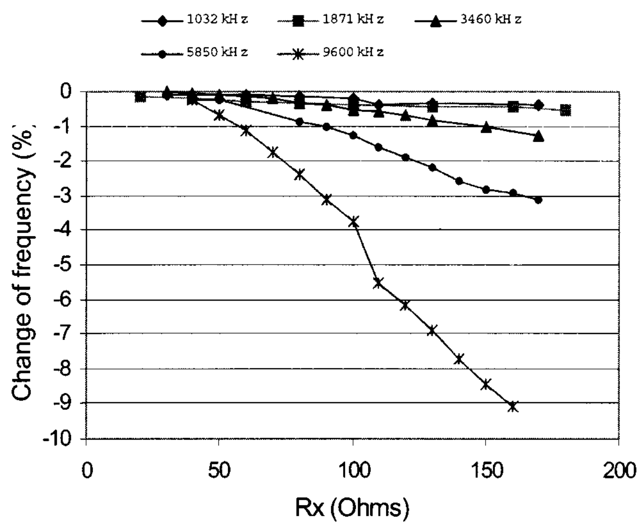

Fig. 6. Dependence of the oscillating frequency on the value of R_x.

at $\boldsymbol{i} \approx U_1/R_{\text{ref}}$ so it is independent of the value of R_x. With resistor R_2 and R_3, the linear operating region of the MOSFET is extended [11].

For the ac-to-dc conversion, we use the four-quadrant, voltage-output multiplier AD835 (Analog Devices). It has an internal reference voltage U with a nominal value of 1.05 V and -3 dB output bandwidth of 250 MHz. The inputs of the analog multiplier are capacitively coupled to the outputs of the amplifiers A1 and A2. For the biasing of AD835, two resistors of 10 kΩ are used in parallel with capacitors of 33 pF, which produces an output offset voltage of about 52 mV. The capacitors are used to prevent the circuit from oscillating at frequencies higher than 10 MHz.

IV. EXPERIMENTAL RESULTS

With a spectrum analyzer, the harmonic distortion levels of U_1 and U_2 (see Fig. 3) were measured for two values of R_x: 50 Ω and 100 Ω, and for an amplitude of U_1 of 0.75 V. Table I presents the results for the measured relative power of the second and the third harmonic of U_1 and U_2 with respect to the power of the first harmonic. As one can see, the distortion level of U_2 is quite low, and it is further reduced by the filtering effect of the resonator. Fig. 4 shows the relative error of the measured resistive component R_x of the impedance versus its value for two frequencies: 1032 kHz and 1871 kHz. The measurement range of the resistive component was from 30 Ω to 170 Ω.

As a reference, impedance analyzer (HP4194A) was used, with an error specification for this frequency range that does not exceed 0.125%. The measurement results showed a maximum error of $\pm 0.15\%$ for a frequency of 1032 kHz and a maximum error of $+0.3\%$ and -0.2% for a frequency of 1871 kHz.

For higher frequencies, the impedance analyzer was not suitable as a reference measurement instrument since its error specification at 10 MHz is already above 1%. For this reason, we took as reference our result for $R_x = 100\ \Omega$ and by changing the DC value of R_x by 10 Ω, we covered the range from 30 Ω to 200 Ω. Fig. 5 shows the relative error that we measured with the relative changes of R_x. For values of R_x below 100 Ω, which is the range that is most interesting for us, the error is within $\pm 0.2\%$. Above 100 Ω, the error starts to grow. The highest value is 1.1% for a frequency of 9.9 MHz and $R_x = 180\ \Omega$. The errors displayed in Figs. 4 and 5 have a clear systematic component, which can be explained with (11). For higher values of R_x and lower values of R_t, the phase shift φ_{A1} of the stage A1 is slightly growing, and the phase shift φ_{A2} of the stage A2 is decreasing, which ensures better frequency compensation. The systematic error, especially for values of R_x above 100 Ω, can be further reduced by using, for example, an appropriate read-out table.

Fig. 6 shows the dependence of the oscillating frequency on the value of the resistive component R_x. This dependence means that the measured value of C_x is a function of the value of R_x. For frequencies up to 2 MHz, the frequency changes only $\pm 0.25\%$ for R_x from 30 Ω to 170 Ω. With an increase in frequency, the same small variation of the frequency is achieved for a reduced range of R_x. In other words, if the value of R_x is higher, the measurement frequency must be lower in order to keep the error in determining C_x small. This is, of course, a natural choice for many applications, including the one presented in this article. An alternative solution is to compensate the error in determining C_x by applying a multi-signal or statistical error-correction measurement technique.

V. Conclusion

A simple, self-oscillating circuit for impedance measurement is presented. It is intended to test the sterility of aseptically packed food products by measuring the conductivity changes of the packaged food in a nondestructive way. The oscillator has been built with fast current-feedback operational amplifiers. To measure the resistive component of the impedance, an AC-to-DC converter with a wide frequency range analog multiplier is used.

The advantages of this interface circuit are: i) an external oscillator is not needed, ii) synchronous detection is not necessary, and iii) at resonance frequency, the reactive components compensate each other to ensure the highest possible sensitivity for the resistive component. The experimental results presented prove that with this interface circuit, an accurate measurement of the impedance components can be achieved in a frequency range up to 10 MHz.

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