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MEASUREMENT OF COMPLEX PERMEABILITY IN THE RF BAND

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The short circuit (coaxial transmission line) method has been used to determine the complex permeability of NiZnCu spinel ferrite and its composite material based on this ferrite (in the powder form) and a non-magnetic PVC polymer matrix over the frequency range 1 MHz to 1 GHz. We also report here on the design and verification of a simple high frequency coaxial line sample holder. The measurement principle as well as calibration and compensation procedures are discussed in detail. Impedance measurements were optimised by the partial elimination of undesirable influences. The measured values of impedances were recalculated to the equivalent values of complex relative permeability according to proposed formulas and the obtained results have been discussed. A user-friendly program written in VEE environment (Agilent technology) is provided for computer control and measurement process.

Keywords: complex permeability, short circuit method, sintered ferrite, ferrite polymer composite, impedance measurement

1 INTRODUCTION

Magnetic permeability describes the interaction of a material with a magnetic field and it is the ratio of magnetic flux density B to the applied magnetizing field H. Complex relative permeability $\tilde{\mu}_r = \mu_r' - j\mu_r''$ consists of the real part μ_r' that represents the energy storage term and the imaginary part μ_r'' that represents the power dissipation term. In order to determine the complex relative permeability of toroidal–shaped samples, impedance measurements have been performed using the HP 4191A impedance analyser. This instrument uses network analysis method, Fig. 1.

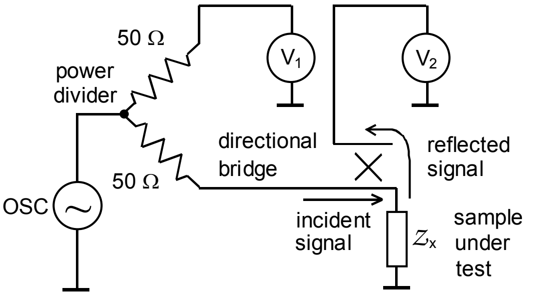

Fig. 1. Measurement principle of the HP 4191A analyser.

The complex reflection coefficient $\tilde{\Gamma} = \Gamma_1 + j\Gamma_2$ is obtained by measuring the ratio of a reflected signal to the incident signal. A directional bridge or coupler is used to detect the reflected signal and the analyser equipped with this bridge is used to supply and measure the signals. Since this method makes it possible to measure reflection at the sample under test, it is usable in the RF band. The measured values of reflection coefficient are automatically converted into the corresponding values of impedance in the frequency range concerned. The high frequency permeability of test samples can be obtained by measuring the input impedance differences between the coaxial sample holder loaded with a toroidal sample and the holder without the sample.

A user–friendly program written in the VEE environment (Agilent technology) is provided for computer control and measurement process. For this purpose, the HP 4191A analyser is connected with a host PC computer via a GP–IB interface. The developed program allows to:

- define the measurement conditions: frequency range (1MHz to 1 GHz), dc bias field (–40 to +40 V), frequency step (linear or logarithmic),
- perform the electrical length compensation,
- perform the residual/stray impedance compensation,
- recalculate the measured values of complex impedance to the corresponding values of complex permeability and view the curves,
- store the measured data on a hard or floppy disk and export them in a format convenient for another processing.

The aim of this paper is to present the experimental technique for measuring the complex permeability of spinel ferrites, ferrite polymer composites and related materials over a broad frequency range concerned. Impedance measurements were optimised by the partial elimination of undesirable influences. A simple holder for high frequency measurement of impedance of toroidal magnetic samples has been designed and verified. The frequency dependences of the complex relative permeability of two samples (NiZnCu ceramic ferrite and ferrite polymer composite) have been obtained by means of this technique and the results are discussed.

2 DESCRIPTION AND MEASUREMENT PRINCIPLE OF THE COAXIAL SAMPLE HOLDER

A sample holder, shown in Fig. 2, is a conductive (brass) shield surrounding the central conductor, which terminates in a short circuit. The short circuit produces a maximum magnetic field and a minimum electric field near the sample, thus making the “short circuit” technique

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particularly suited for the measurement of the magnetic properties of test sample such as permeability.

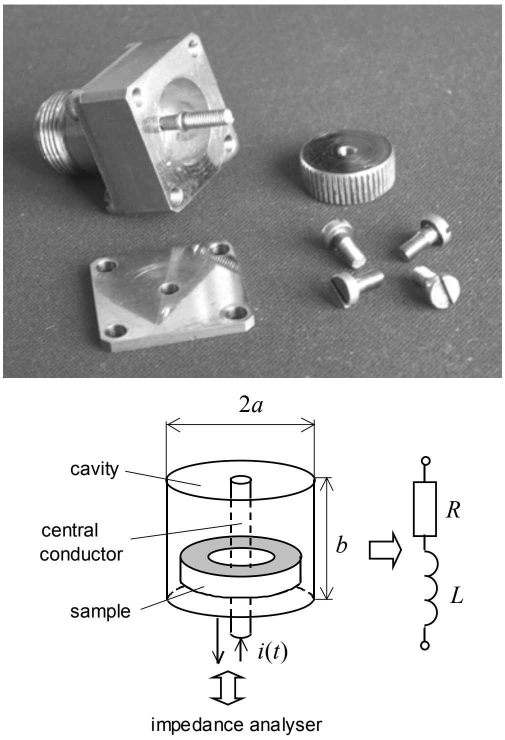

Fig. 2. View of the sample holder and its measurement principle.

The medium between the inner and outer conductors is air. The dimensions of the sample holder cavity are inner length $b = 10$ mm and inner diameter $2a = 15$ mm. The length b obeys the condition $b < \lambda/4$ for $f \leq 1$ GHz in order to avoid the dimensional resonance effect.

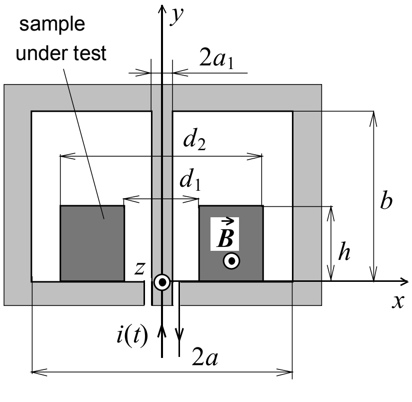

Fig. 3. Cross-section of the sample holder.

When the sample is inserted into the holder, the whole system is completely closed and then connected through the N to APC–7 adapter to the previously calibrated impedance analyser. The analyser supplies an electromagnetic wave propagating in a TEM mode. The reflection coefficient is measured, permitting the determination of the input impedance $\mathcal{Z}_{\rm in}$ of the holder with the sample. In this section we will derive the equation for the determination of complex permeability of the holder equipped with the test sample, Fig. 3. Since the construction of this holder creates one turn around the toroid (with no magnetic flux leakage), the complex magnetic flux of the measurement circuit including the ring core is given by the equation:

$$\tilde{\Phi} = \iint_S \vec{\mathcal{B}} \cdot d\vec{S} = \int_0^a \int_0^b \frac{\mu_{\rm o} \tilde{\mu}_{\rm r} \mathcal{J}}{2\pi \cdot x} dxdy \,. \tag{1}$$

By unfolding Eq. (1) with $\vec{\mathcal{B}}$ as complex phasor–vector of magnetic flux density, $\mu_{\rm o}$ as permeability of free space, and $\mathcal{J}$ as complex phasor of harmonic time–dependent electric current $i(t)$, the following formula can be written:

$$\tilde{\Phi} = \int_{\frac{d_2}{2}}^{a} \int_0^b \frac{\mu_{\rm o} \mathcal{J}}{2\pi \cdot x} dxdy + \int_{\frac{d_1}{2}}^{\frac{d_2}{2}} \int_0^h \frac{\mu_{\rm o} \tilde{\mu}_{\rm r} \mathcal{J}}{2\pi \cdot x} dxdy + \int_{\frac{d_1}{2}}^{\frac{d_2}{2}} \int_h^b \frac{\mu_{\rm o} \mathcal{J}}{2\pi \cdot x} dxdy +$$

$$+ \int_0^{\frac{d_1}{2}} \int_0^b \frac{\mu_{\rm o} \mathcal{J}}{2\pi \cdot x} dxdy - \int_0^{a_1} \int_0^b \frac{\mu_{\rm o} \mathcal{J}}{2\pi \cdot x} dxdy \,. \tag{2}$$

The magnetic flux of measurement circuit is then:

$$\tilde{\Phi} = \frac{\mu_{\rm o} \mathcal{J}}{2\pi} \left\{ \left(\tilde{\mu}_{\rm r} - 1 \right) \cdot h \cdot \ln\left(\frac{d_2}{d_1} \right) + b \cdot \ln\left(\frac{a}{a_1} \right) \right\} \tag{3}$$

and the complex susceptibility of a sample under test is given by:

$$\tilde{\chi} = \frac{2\pi \cdot \left(\tilde{\Phi} - \tilde{\Phi}_{\rm air} \right)}{h \cdot \mathcal{J} \cdot \mu_{\rm o} \cdot \ln\left(\frac{d_2}{d_1} \right)} \tag{4}$$

where $\tilde{\Phi}_{\rm air} = \left(b\mu_{\rm o}\mathcal{J}/2\pi \right) \cdot \ln\left(a/a_1 \right)$ is the magnetic flux when ferrite core is not mounted in the holder. The measured complex impedance $\mathcal{Z}$ of an equivalent electric circuit of the holder loaded with the ferrite core shown in Fig. 2 is given by $\mathcal{Z} = R + {\rm j}\omega L = {\rm j}\omega \tilde{\Phi}/{\rm I}$. Instead of fluxes $\tilde{\Phi}$ and $\tilde{\Phi}_{\rm air}$ in Eq. (4) we can use corresponding complex impedances $\mathcal{Z}$ (measured on a holder with ferrite core) and $\mathcal{Z}_{\rm air}$ (measured on a holder without ferrite core):

$$\tilde{\mu}_{\rm r} = 1 + \tilde{\chi} = 1 + \frac{\left(\mathcal{Z} - \mathcal{Z}_{\rm air} \right)}{{\rm j}h \cdot \mu_{\rm o} \cdot f \cdot \ln\left(\frac{d_2}{d_1} \right)} \tag{5}$$

where d_1 and d_2 are the inner and outer diameters of the toroid, respectively, h is the height of the toroid, and f is the frequency of applied ac electromagnetic field. Complex (relative) permeability is therefore calculated from difference between the impedance $\mathcal{Z}$ of the holder loaded with the toroidal sample and the impedance $\mathcal{Z}_{\rm air}$ of the empty holder.

The impedance analyser measures the complex reflection coefficient $\tilde{\Gamma} = \Gamma_1 + {\rm j}\Gamma_2$, which is recalculated to the input impedance of the holder $\mathcal{Z}_{\rm in}$ (with or without sample) according to the following equation:

$$Z_{\mathrm{in}} = Z_{\mathrm{o}} \frac{1+\widetilde{\Gamma}}{1-\widetilde{\Gamma}} \qquad (6)$$

where $Z_{\mathrm{o}} = 50\ \Omega$ is the characteristic impedance of the 7 mm (APC–7) test port. Resistance (R_{in}) and reactance (X_{in}) values of the input impedance of the holder $Z_{\mathrm{in}} = R_{\mathrm{in}} + \mathrm{j}X_{\mathrm{in}}$ (with or without sample) are then derived as:

$$R_{\mathrm{in}} = Z_{\mathrm{o}} \frac{1-\Gamma_1^2-\Gamma_2^2}{(1-\Gamma_1)^2+\Gamma_2^2}\,, \quad X_{\mathrm{in}} = Z_{\mathrm{o}} \frac{2\Gamma_2}{(1-\Gamma_1)^2+\Gamma_2^2}\,. \qquad (7,8)$$

Since the holder together with the N connector is connected to the test port via an N to APC–7 adapter, special attention must be paid to the errors caused by these additional parts – one is residual impedance error and the other one is electrical length error, Fig. 4. The residual impedance, which includes distributed parameters such as residual resistance, residual inductance, stray conductance and stray capacitance, makes the measurement values inaccurate. The electrical length error occurs due to a phase shift of the test signal along the 50 Ω transmission line between the calibration plane and the holder. This leads to an impedance measurement error when measuring phase. Phase shift error can be compensated with electrical length compensation function, which the HP 4191A analyser provides automatically. Residual impedance error can be compensated with short compensation function. The HP 4191A instrument does not have such a possibility. Therefore we can remove the residual impedance error from the measurement results by computing the short compensation values using the developed software (written in the VEE environment) through an external computer.

The electrical length l ($l = l_1 + l_2$, l_1 is the length of N connector and l_2 is the length of the N to APC–7 adapter) establishes the plane "A" (Fig.4), from which the instrument takes the holder into account. Because the reflection coefficient $\widetilde{\Gamma}$ is a function of the propagation line length l according to the equation $\widetilde{\Gamma}(l) = \widetilde{\Gamma}(0) \cdot \mathrm{e}^{-\mathrm{j}2\gamma l}$ where $\gamma = \beta + \mathrm{j}\alpha$ is the propagation coefficient (β is the attenuation coefficient and α is the phase coefficient of the transmission line), all measurement results automatically compensate for the electrical length effect of the holder using correction calculations based on the equation for $\widetilde{\Gamma}(l)$. It should be noted that in reality the electrical length l is given by the relation $l = l'\lambda_{\mathrm{o}}/\lambda$ with λ_{o} the wavelength in vacuum, λ the actual wavelength in the coaxial transmission line (i.e. N connector together with N/APC–7 adapter), and l' the actual transmission line length. The above discussion assumed that $\lambda \approx \lambda_{\mathrm{o}}$, *ie* $l \approx l'$ because the medium between inner and outer conductors of the additional parts is air.

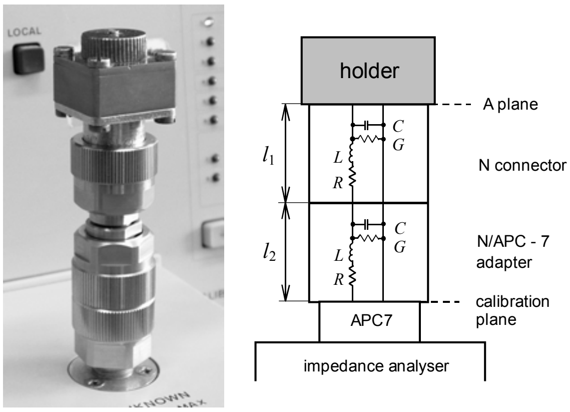

Fig. 4. View of additional parts: N connector and N/APC–7 adapter, and their equivalent circuit model.

After the calibration of the analyser, the electrical length l as well as the residual impedances are compensated. Without these compensations spurious geometrical resonances might be present in the experimental data. It should be noted that one factor still remains. This factor is regarded as the time fluctuation (or instability) of compensated values. We should also not forget about other physical factors that affect measurement results include temperature, humidity, and external magnetic fields. Once Z_{in} of the holder with and without toroidal sample is known, the complex (relative) permeability $\widetilde{\mu}_{\mathrm{r}}$ is obtained by (5).

3 EXPERIMENTS AND RESULTS

As an example, Fig. 5 shows the frequency dependences of real μ_{r}' and imaginary μ_{r}'' parts of complex relative permeability $\widetilde{\mu}_{\mathrm{r}} = \mu_{\mathrm{r}}' - \mathrm{j}\mu_{\mathrm{r}}''$ calculated using (5) from measured impedance values for sintered NiZnCu ferrite of composition $Ni_{0.27}Zn_{0.63}Cu_{0.1}Fe_2O_4$ (Fig. 5a) and composite material made of this ferrite in the powder form (with particle size 0–250 μm) and a non–magnetic polymer matrix (PVC) with ferrite volume concentration 73 vol% (Fig. 5b), [1]. We also checked the level of ac current passing through the holder, *ie* the amplitude of ac magnetic field intensity H_{ac} that should not be higher than about 0.5 A/m in order to fulfil the initial permeability condition. The dimensions of both toroidal samples (ferrite and composite) used here were 12x6x3 mm. The thickness $h = 3$ mm of samples satisfies the condition, at which dimensional resonance effect cannot occur in the measured results: $h < \lambda/4$. The value of H_{ac} was investigated because the HP 4191A analyser does not have the constant–current mode.

The change of real part of complex permeability with frequency close to the critical frequency f_{c} is called the dispersion of permeability and the change of imaginary part with frequency is known as absorption.

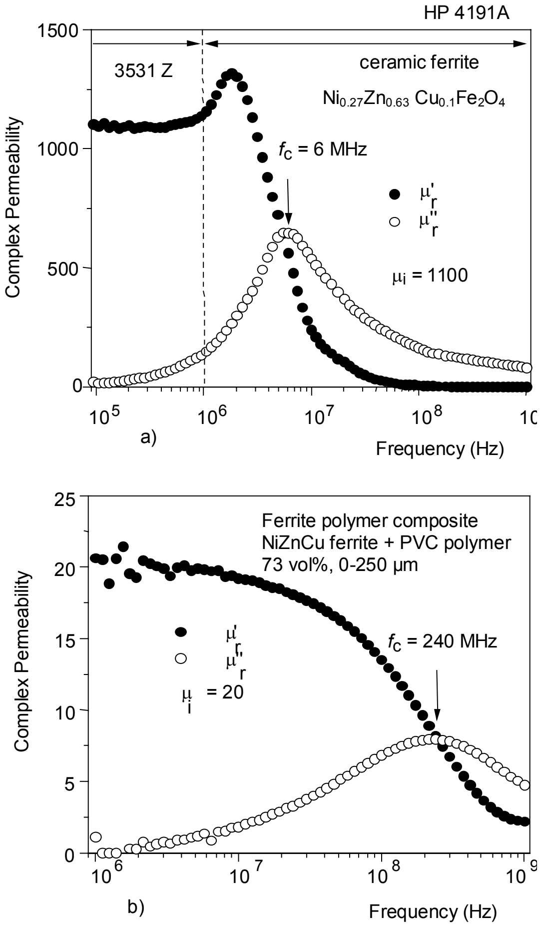

Fig. 5. Complex permeability spectra for a) sintered ferrite, and b) its composite material.

NiZnCu ceramic ferrite showed a resonance type of frequency dispersion of permeability and ferrite polymer composite material showed a relaxation one. In the case of composite structure, the low frequency real part of complex permeability is lower (20) in contrary to sintered ferrite (1100) but the critical resonance/relaxation frequency (at which the imaginary part has a maximum) is considerably higher (240 MHz) in comparison with the ceramic ferrite (6 MHz). These changes are due to: a) existence of Snoek's limit given by the product $\mu_r' \cdot f_c$ (proportional to the saturation magnetisation M_s of ferrite filler), b) arising demagnetisation fields of ferrite particles incorporated into the polymer matrix, [2].

4 CONCLUSIONS

We have described an optimised sample holder for the measurement of complex permeability of toroidal magnetic samples in RF band (1–1000 MHz) and convenient calibration/compensation procedures have been discussed in detail. The described equipment gives us the ability to perform measurements over a wide range of frequencies in order to achieve better understanding of resonance/relaxation processes particularly in spinel ferrites and related materials. The data treatment is discussed, paying special attention to the proper conversion of the impedance measurement to permeability spectra. The designed sample holder can be connected with the test port via a precision coaxial cable in order to provide measurements in the dependence on temperature (in a convenient furnace) and/or dc bias field (in a suitable electromagnet). All system parameters are computer controlled by the VEE software.

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