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BROADBAND MICROWAVE MEASUREMENTS OF RELATIVE PERMITTIVITY AND PERMEABILITY OF MATERIALS

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Abstract
A short review of modern techniques for broadband microwave measuring complex permittivity and permeability of materials at broadband microwave frequencies has been presented. Classification of type of the measurement methods to extract the complex permittivity and permeability of bulk materials has been done in this paper. Mathematical models of all the above-mentioned measurement types of methods are reviewed. Advantages and disadvantages of existing measurement devices/systems are described according to the classification made. A new design of test devices for the measurement system operating in broadband microwave frequency range has been proposed. Measurement system on the basis of the proposed device can be used for a measuring the ordinary bulk materials, ferrites (including fully magnetized) and magnetic materials but much cheaper then one to be made. The system is much cheaper than a free space measurement system. The approach to define the permeability tensor of magnetic materials using the test device is proposed as well.

Keywords: Measurement system, permeability tensor, retrieval matter, S-parameters.

INTRODUCTION

Study of the physical phenomena, which determine the behavior of the material subjected to an alternating electric or magnetic field is now undergoing a revival of interest. This is explained by many new applications such as microwave electromagnetic compatibility or high data rate for high density information storage and etc because of dispersion properties of materials especially the magnetic ones. That is why the dispersion properties of materials especially magnetic materials are widely used in many microwave circuits such as circulators, phase shifters and filters. Thus, the development of characterization techniques of materials on microwave frequency range is desperately required at present time for the development of microwave theory and technique, RF applications, metamaterials applications and so on. In the light of the problem, a broadband electromagnetic characterization of materials which requires the development of the measurement techniques for complex permeability and permittivity of materials on microwave frequency range is of big practical interest [Nicolson and Ross 1970, Barry 1986, Ghodgaonkar *et al.* 1989,1990, Belhadj-Tahar *et al.* 1990, Vanzura and Baker-Jarvis 1994]. Moreover the characterization of magnetic materials [Acher 1996, Pain *et al.* 1999, Mallegol *et al.* 2003a] requires the development of an approach to evaluate the permeability tensors [Mallegol *et al.* 2003a,b, Queffelec *et al.* 2002].

In this paper, we present a short review of modern techniques for broadband microwave measuring complex permittivity and permeability of materials including magnetic materials. Classification of type of the measurement methods are made according to the design philosophy and mathematical models of measurement devices/system published in modern literature. Advantages and disadvantages of the existing measurement devices/systems are described based on analysis of a number of publications devoted to the measurements of complex permittivity and permeability of materials including magnetic materials. We also propose the design and mathematical model of quite simple device relative to that of inherent for the free space measurement system. Based on this design there is a possibility to make measurements for ordinary bulk materials, magnetic materials and metamaterials in the microwave frequency range. The procedure to define the permeability tensor of magnetic materials/metamaterials is also proposed.

MEASUREMENT METHODS

Traditional techniques [Nicolson and Ross 1970, Barry 1986, Ghodgaonkar *et al.* 1989,1990, Belhadj-Tahar *et al.* 1990, Vanzura and Baker-Jarvis 1994] for the measurement of the complex permittivity ε^* (or complex dielectric constant ε_r^*) and permeability μ^* (or complex magnetic constant μ_r^*) in the microwave frequency range use the experimentally measured complex S-parameters to utilize them for the evaluation of the complex permittivity and permeability using an adequate mathematical model. According to the mathematical viewpoint of problem a measuring permittivity and permeability are to be defined as complex functions of frequency ω

$$\left.\begin{aligned} \varepsilon^*(\omega) &= \varepsilon'(\omega) + i\varepsilon''(\omega) = \varepsilon'(\omega)\big(1 - i\tan(\delta_\varepsilon(\omega))\big), \\ \mu^*(\omega) &= \mu'(\omega) + i\mu''(\omega) = \mu'(\omega)\big(1 - i\tan(\delta_\mu(\omega))\big) \end{aligned}\right\} \tag{1}$$

where δ_ε is the dielectric loss tangent, δ_μ is the magnetic loss tangent.

As we can see from Archer *et al.* [1996], the evaluation of complex frequency-dependent expressions for permittivity and permeability allows us to investigate absorption properties of materials in pertinent frequency range. Moreover if we know the expressions of above parameters we can evaluate the complex refractive index of materials. Thus, the possibility to measure the complex permittivity and permeability of materials will enable us to characterize the electrical and magnetic absorption and diffraction properties of materials. This problem is of special importance with the viewpoint of development of

metamaterials since the desirable properties of conventional materials are seriously degraded for frequencies above 1 GHz [Kern *et al.* 2005].
Existing experimental techniques for evaluating the complex parameters of materials using measured S-parameters, can be subdivided into three types of the measurement methods/approaches: 1) the free space measurement method (FSM), [Ghodgaonkar *et al.* 1989, 1990)], 2) the waveguide measurement method (WM), [Nicolson and Ross 1970, Belhadj-Tahar *et al.* 1990, Vanzura and Baker-Jarvis 1994)], and 3) the strip line measurement method (SLM), [Barry 1986, Queffelec *et al.* 1994, Salahun *et al.* 2001].

FSM

FSM utilizes the measurement of S-parameters of a sample suspended between one horn antenna used as a source of irradiating wave and the second horn antenna as receiver of the scattered wave, Fig.1. FSM implies the principle of thin sample (Fig.2): the sample thickness is much smaller than the other dimensions. In this case the mathematical model for obtaining complex expressions for dielectric and magnetic constants using measured S-parameters, is just a solution of appropriative one-dimension problem of Microwave Theory described in paper [Ghodgaonkar *et al.* 1990]. Appropriate formulas for evaluation of complex dielectric and magnetic constants are:

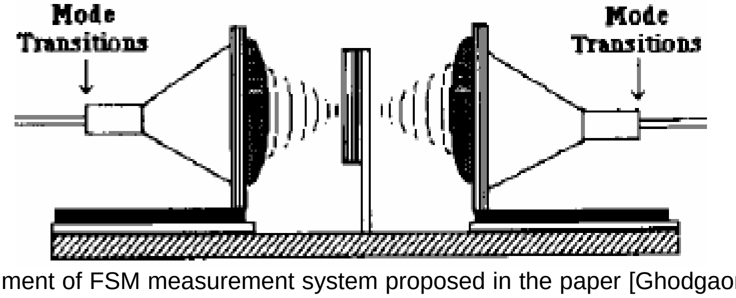

Fig.1: Fragment of FSM measurement system proposed in the paper [Ghodgaonkar *et al.* 1990].

$$\varepsilon_r^*(\omega) = \frac{\gamma^*(\omega)}{\gamma_0(\omega)}\left(\frac{1-\Gamma(\omega)}{1+\Gamma(\omega)}\right), \quad \mu_r^*(\omega) = \frac{\gamma^*(\omega)}{\gamma_0(\omega)}\left(\frac{1+\Gamma(\omega)}{1-\Gamma(\omega)}\right) \qquad (2)$$

where $\gamma^*(\omega) = \gamma_0(\omega)\sqrt{\varepsilon_r^*(\omega)\mu_r^*(\omega)}$ is the complex propagation constant for the sample, $\gamma_0(\omega) = i\omega\sqrt{\varepsilon_0\mu_0} = i2\pi/\lambda_0$ is the propagation constant in free space, λ_0 is the wave length in free space, ε_0 and μ_0 are respectively the permittivity and permeability of free space respectively,

$$\Gamma(\omega) = K(\omega) \pm \sqrt{K^2(\omega)-1}, \quad |\Gamma(\omega)| < 1 \qquad (3)$$

is the Fresnel complex reflection coefficient of sample, the frequency-dependent function $K(\omega)$ is to be calculated by the formula:

$$K(\omega) = \frac{S_{11}^2(\omega) - S_{21}^2(\omega) + 1}{2S_{11}(\omega)}. \qquad (4)$$

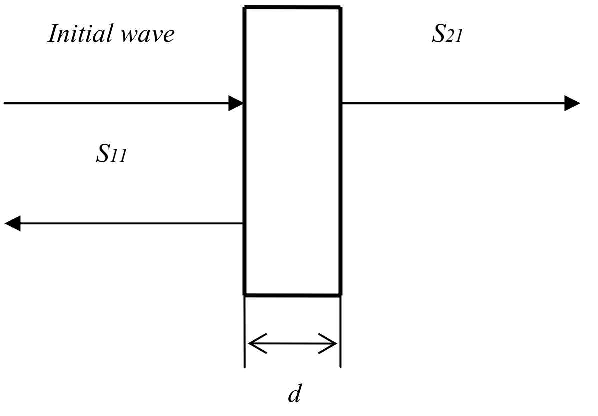

Fig. 2: Schematic diagram of planar sample for FSM.

According to (2-4) the expression for complex propagation constant in the sample is to be evaluated by the formulas [Ghodgaonkar *et al.* 1990]:

$$\gamma^*(\omega) = \frac{\ln(1/T(\omega)) + i(2\pi n - \theta(\omega))}{d} \qquad (5)$$

where $T(\omega)$ is the transmission coefficient between of faces of the sample:

$$T(\omega) = e^{-\gamma^*(\omega)d} = |T(\omega)|e^{i\theta(\omega)} = \frac{S_{11}(\omega) + S_{21}(\omega) - \Gamma(\omega)}{1-(S_{11}(\omega) + S_{21}(\omega))\Gamma(\omega)}, \qquad (6)$$

d is the thickness of sample, $\theta(\omega)$ is the phase constant of the complex phase incursion on sample. It has been obtained in reference [Ghodgaonkar *et al.* 1990] that if the sample thickness d is chosen so that it is less than the wave length in sample material, then expression (5) will give a unique value of the both permittivity and permeability which corresponds to the branch $n = 0$.
As we can see from formulas (2-6) the mathematical model for FSM is quite simple. Moreover, according to the logical design philosophy of measurement system for measuring S-parameters of a flat sample, one can have advantages: 1) the possibility to make the measurements for nonlinear materials, 2) take into account the demagnetizing field for magnetic materials, 3) the possibility to apply the stationary external magnetic field. But the measurement system based on FSM has three serious disadvantages: 1) the high cost of equipment due to the high prices for horn antennas, 2) the poor accuracy to make measurements for thin and flexible materials because of sagging of the sample when mounted on the sample holder, 3) the complicate procedure for calibration. Moreover it should be mentioned that due to the mounting of the sample between antennas the model based on (2-6) will become more complicate since the sample is sandwiched between two fused plates (for instance quartz plates) which are half-wavelength at mid-band. As follows, the actual S-parameters of the sample are calculated from the measured S-parameters of the quartz plate-sample-quartz plate assembly from knowledge of the complex electric constant and thickness of the quartz plates. So the above mentioned disadvantages of the design of FSM measurement system makes it desirable to search principally a new design for S-parameters measurements which is cheaper and simple to be realized.

WM

WM utilizes the same mathematical model as FSM does because a sample is located in a waveguide or coaxial line fully fulfilling the cross section (Fig.3). That is why the waveguide/coaxial line keeps transverse electromagnetic wave (TEM) that is very important in order to use a mathematical model of FSM.
We can see that a WM measurement system realized with a waveguide or coaxial line, has a number of advantages on a FSM measurement system which uses horn antennas: 1) low cost to establish the measurement system, 2) simple procedure for calibration (it is enough to make the matching between fulfilled waveguide/coaxial line and feed connectors), 3) the possibility of measurements for thin and flexible materials, 4) the use of "pure" FSM mathematical model. But any WM measurement system has very serious disadvantages: it cannot be used for measuring S-parameters of magnetic materials because of the necessity to fulfill the cross section of a waveguide/coaxial line fully with the sample. Due to this requirement it is impossible to take in account the magnetizing fields. That is why FSM system is still very useful for characterization of magnetic materials.

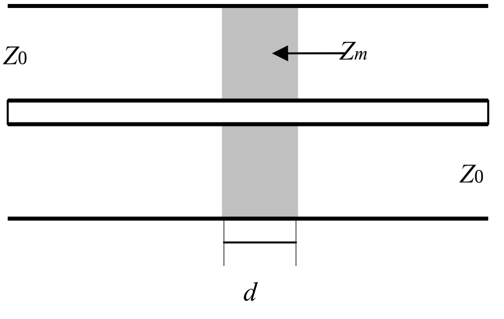

Fig. 3: Schematic diagram of loaded coaxial line of length d, where Z_0 is the characteristic impedance of free space, Z_m is the characteristic impedance of sample material.

SLM

SLM uses special design of device keeping TEM to fulfill easily the cross section of the device with a sample which is parallelepiped. It is very important to perform a series of experiments changing different samples. It is not easy to do this using a waveguide or coaxial line as holder of sample. It is assumed that the device is fed with coaxial connectors.
A quite successful design of device for SLM was proposed by Barry [1986] (Fig. 4). We call a method to analyze the test device as Strip-Line Method due to the fact that strip-line irradiates initial TEM. The strip-line portion of the device was designed for a characteristic impedance of 50 Ω in order to match with the impedance of cables and network analyzer system used to make the S-parameter measurements.
We can see that according to the philosophy of design of the measurement device, it is quite convenient to carry out multi series experiments. Moreover the device keeps TEM mode that simplifies electromagnetic analysis of the device.
In order to obtain the mathematical model for evaluation of complex dielectric and magnetic constants from measured S-parameters, electromagnetic analysis

of the test device through schematic model of Fig. 5 has been made (here γ_0 and Z_0 are the complex propagation constant and impedance of free space respectively, γ^* and Z_m are the complex propagation constant and impedance of resonator space loaded with sample respectively, d is the sample length, l is a distance between connectors and sample, *C1+* and *C1–* are the input terminals, *C2+* and *C2–* are the output terminals). Results of the analysis can be presented by formulas for evaluation of the complex constants in the form of expressions (2) where

$$\Gamma(\omega) = \frac{S_{11}(\omega)}{e^{-i2\gamma_0(\omega)l} - S_{12}(\omega)e^{-i\gamma^*(\omega)d}}, \quad |\Gamma(\omega)| < 1, \tag{7}$$

$$\gamma^*(\omega)d = \Theta_G(\omega) + 2\pi n + i\ln(G(\omega)), \quad n = 0, \pm 1, \pm 2, ... \tag{8}$$

and expressions for ε_r^* and μ_r^* are to be defined by the formulas (2). Here

$$G(\omega) = \sqrt{\left(\mathrm{Re}\left(A(\omega) \pm \sqrt{A^2(\omega) - 1}\right)\right)^2 + \left(\mathrm{Im}\left(A(\omega) \pm \sqrt{A^2(\omega) - 1}\right)\right)^2},$$

$$\Theta_G(\omega) = \tan^{-1}\left(\frac{\mathrm{Im}\left(A(\omega) \pm \sqrt{A^2(\omega) - 1}\right)}{\mathrm{Re}\left(A(\omega) \pm \sqrt{A^2(\omega) - 1}\right)}\right),$$

$$A(\omega) = \frac{e^{-i4\gamma_0(\omega)l} + S_{12}^2(\omega) - S_{11}^2(\omega)}{2e^{-i2\gamma_0(\omega)l} S_{12}(\omega)}$$

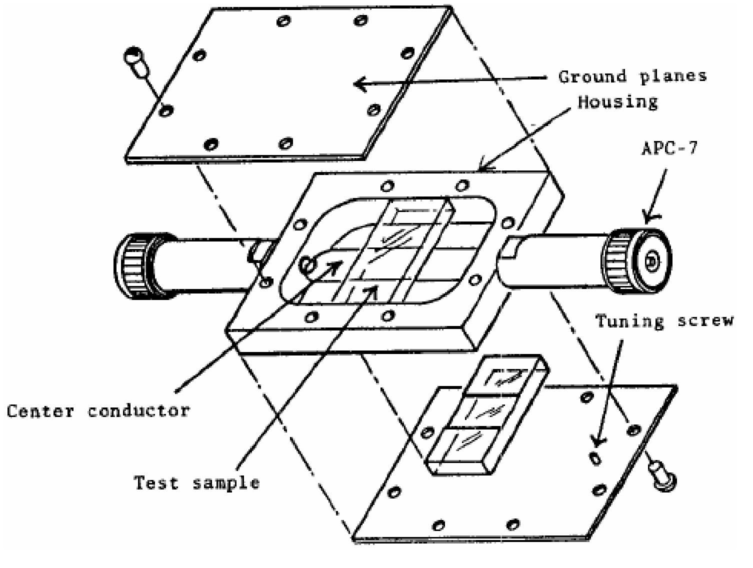

Fig. 4: Strip-line measurement device from reference [Barry 1986]

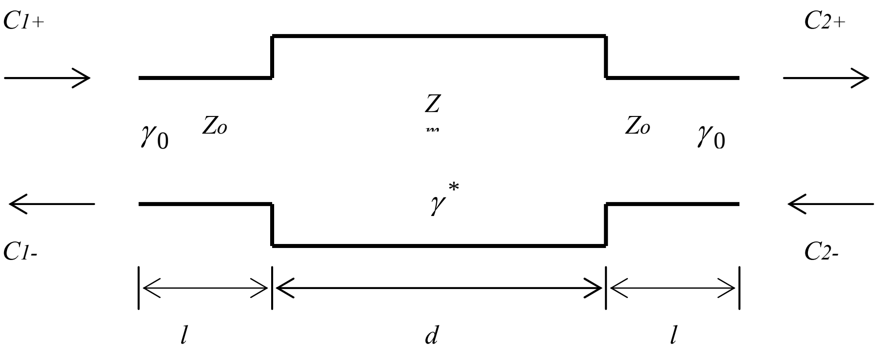

Fig. 5: Schematic diagram for electromagnetic analysis of the test device of Barry [1986].

It has been shown that $n = 0$ if $\lambda_m > d$ (λ_m is the wavelength in sample) and this device is very convenient for measurements in the microwave frequency range.

The proposed device has, in principle, the same advantages device as waveguide/coaxial line test device under WM relative to FSM system. Moreover, the proposed device is much more convenient for multiple measurements than a waveguide/coaxial line test device. Most serious disadvantage of both the devices is impossibility to make measurements for magnetic materials and ones under biasing magnetic fields that is required for ferrites and metaferrites. This disadvantage has been resolved or may be resolved using the so-called Modified Strip Line Method (MSLM) measurement device/system.

OTHER TECHNIQUES

Last two measurement techniques have the same serious disadvantages because of impossibility to take into account the magnetizing fields. That is why the both techniques are not useful for characterizing magnetic materials. It follows to find out new design of measurement device which can be used for characterizing magnetic materials and can be alternative to FSM system. Such kind of device has been proposed in [Queffelec *et al.* 1994]. Cross section principle scheme can be found at Fig. 6. In the chamber of test device, the alumina substrate is placed on the lower ground plane. The housing is made of aluminum. An air gap appears between the sample and the alumina substrate, near the central conductor, because the central conductor thickness is not equal to zero. The air gap can increase on account of a bad sample surface. The sample, which is placed directly and symmetrically on the line (Fig. 7) is a parallelepiped. The test device must be matched with feed coaxial connectors through the strip-line.

Retrieval of complex dielectric and magnetic constants from the measured S-parameter requires a rigorous electromagnetic analysis of the test device (direct problem) together with an optimization program (inverse problem).

The direct problem can partially be solved using the analysis of test device of SLM considered by Barry [1986] because of pertinent design philosophy (Fig. 7) which enables to keep the hypothesis of a quasi-TEM mode valid since the low material volume (bringing about a low perturbation of the electromagnetic symmetry of the microstrip test device).

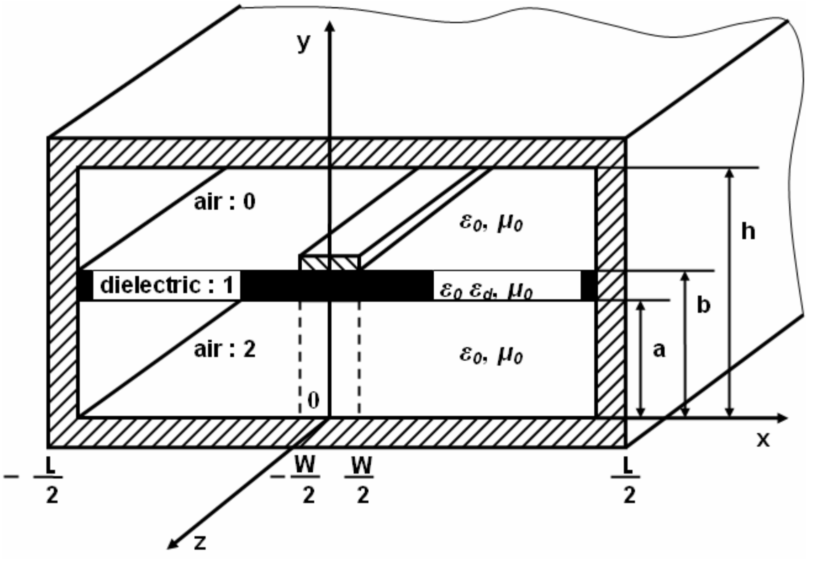

Fig. 6: Microstrip device cross section of reference [Queffelec *et al.* 1994]

So using terminology of the models (2-8), we can write formulas of retrieval of S-parameters

$$S_{11}(\omega) = S_{22}(\omega) = i\,\frac{2\Gamma(\omega)\cdot\sin(\gamma^{*}(\omega)d)\cdot e^{-2i\gamma_0 l}}{e^{i\gamma^{*}(\omega)l} - \Gamma^2(\omega)\cdot e^{-i\gamma^{*}(\omega)l}}, \tag{9}$$

$$S_{12}(\omega) = S_{21}(\omega) = \frac{(1-\Gamma^2(\omega))\cdot e^{-2i\gamma_0 l}}{e^{i\gamma^{*}(\omega)l} - \Gamma^2(\omega)\cdot e^{-i\gamma^{*}(\omega)l}} \tag{10}$$

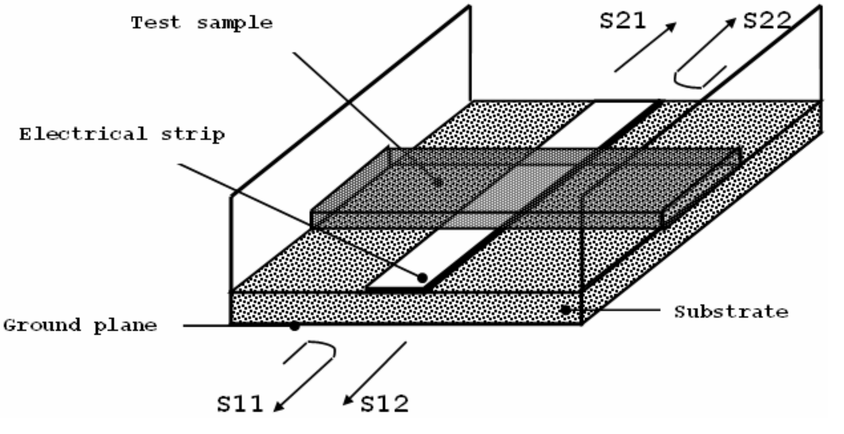

Fig. 7: The test device of reference [Queffelec et al. 1994] loaded with the sample.

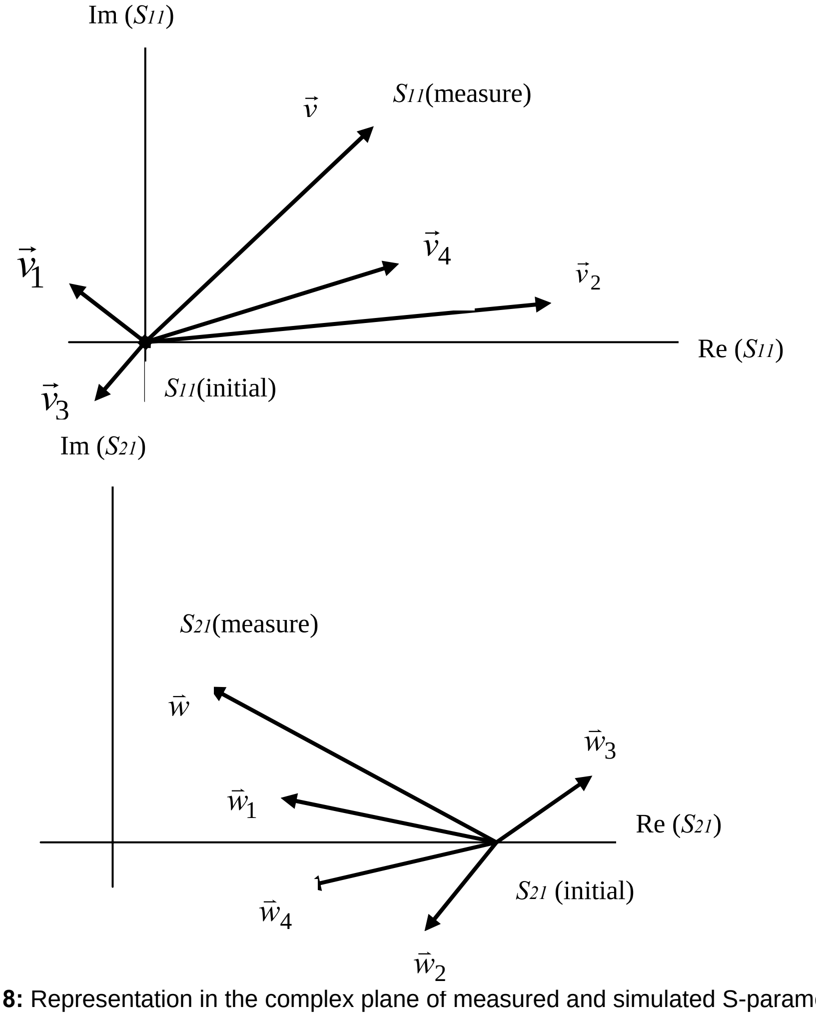

Fig. 8: Representation in the complex plane of measured and simulated S-parameters.

The relationship between the S-parameters and the complex dielectric and magnetic constants is a matter of the inverse problem which requires a numerical optimization method together with the electromagnetic analysis of the test device. The optimization procedure used is based on the gradient method [Rosloniec 1990]. According to this method the initial values of the problem correspond to a permittivity and permeability equal to one and loss equal to zero. Theoretical S-parameters $S_{11}(initial)$ and $S_{21}(initial)$ are computed from these characteristics (direct problem, formulas (9-10)). Then four new states are defined by iteration of ε', ε'', μ' and μ'' successively. The representation in the complex plane of these states is shown in Fig. 8. The vectors $\vec{v}$ and $\vec{w}$ join the initial state to the measured value of S_{11} and S_{21}, and the vectors $\vec{v}_i$ and $\vec{w}_i$ $(i = 1,2,3,4)$ join the initial state to the previously described states and can be expressed by the formulas:

$$\vec{V} = a\vec{V}_1 + b\vec{V}_2 + c\vec{V}_3 + g\vec{V}_4 , \tag{11}$$

$$\vec{W} = a\vec{W}_1 + b\vec{W}_2 + c\vec{W}_3 + g\vec{W}_4 \tag{12}$$

A set of four equations with four unknowns is obtained from the separation real and imaginary components:

$$\begin{pmatrix} a \\ b \\ c \\ g \end{pmatrix} = \begin{pmatrix} \mathrm{Re}(V_1) & \mathrm{Re}(V_2) & \mathrm{Re}(V_3) & \mathrm{Re}(V_4) \\ \mathrm{Im}(V_1) & \mathrm{Im}(V_2) & \mathrm{Im}(V_3) & \mathrm{Im}(V_4) \\ \mathrm{Re}(W_1) & \mathrm{Re}(W_2) & \mathrm{Re}(W_3) & \mathrm{Re}(W_4) \\ \mathrm{Im}(W_1) & \mathrm{Im}(W_2) & \mathrm{Im}(W_3) & \mathrm{Im}(W_4) \end{pmatrix}^{-1} \begin{pmatrix} \mathrm{Re}(V) \\ \mathrm{Im}(V) \\ \mathrm{Re}(W) \\ \mathrm{Im}(W) \end{pmatrix} \tag{13}$$

After the solution of the set, a new state is defined as follows:

$$\left.\begin{aligned} \varepsilon_2^{'} = 1 + a, \quad \varepsilon_2^{''} = c, \\ \mu_2^{'} = 1 + b, \quad \mu_2^{''} = g \end{aligned}\right\} \tag{14}$$

where the index 2 means a second step of iterative procedure of the gradient method ($\varepsilon_1^{'} = 1$, $\varepsilon^{''}{}_1 = 0$, $\mu_1^{'} = 1$, $\mu_1^{''} = 0$). And more generally for any *i*-th state beginning from third one:

$$\left.\begin{aligned} \varepsilon_{i+1}^{'} = \varepsilon_i^{'} + a \cdot \Delta, \quad \varepsilon_{i+1}^{''} = \varepsilon_i^{''} + b \cdot \Delta, \\ \mu_{i+1}^{'} = \mu_i^{'} + c \cdot \Delta, \quad \mu_{i+1}^{''} = \mu_i^{''} + g \cdot \Delta \end{aligned}\right\} \tag{15}$$

where Δ is the correction increment. This new state takes place out of the previous one, and so on, until convergence of the simulated S-parameters to the measured one is achieved.

The considered measurement device has an essential advantage relative to SLM's device, namely, one takes in account the magnetizing fields because a sample does not fill the cross section of the test device fully. It means that this device can be used for characterization of magnetic materials, even thin film samples. But the device has also essential disadvantage relative to the device of FSM system: it does not enable to apply a biasing magnetic field to the test sample. Proposed in reference [Salahun *et al.* 2001] the test device avoids this disadvantage due to its design, Fig. 9. The measurement cell is an asymmetrical strip-line ended by two tapers. These discontinuities, taken into account in the calibration procedure, are made to avoid capacitances between the conducting strip and conductor planes where strip connectors are fixed. The non equal distance between the strip and the ground planes is made to approach microstrip configuration. The electromagnetic energy is mainly confined in the space between the strip and the nearest ground plane. No laterally conductive walls are necessary because the ground plane width is wider than three times the separation between the strip and the nearest ground plane.

The direct and inverse problems should also be resolved here in order to extract the expressions of complex dielectric and magnetic constants from measured S-parameters.

The direct problem is to determine the constants of the cross section as a function of the frequency. The hypothesis of a quasi-TEM mode is valid for microwave frequency range because for low frequencies, longitudinal components of the microwave fields can be neglected compared with transversal ones. Since the dominant mode is quasi-TEM, the effective electromagnetic parameters are determined from parameters using the Nicolson-Ross procedure [Nicolson and Ross 1970].

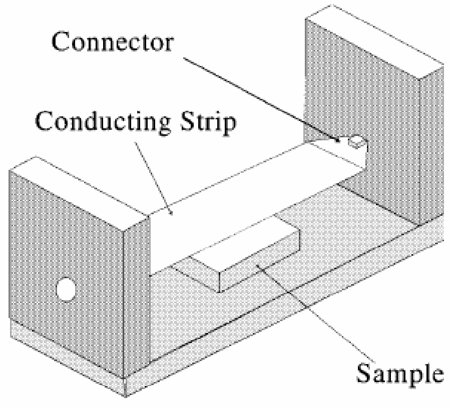

Fig. 9: Schematic diagram for the test device of reference [Salahun *et al.* 2001] without upper ground plane

As there are no analytical expressions to determine the electromagnetic constants (ε_r^*, μ_r^*) of the sample due to the heterogeneity of the loaded cross section of the test device, the inverse problem of the data processing procedure consist in extracting these parameters from the measured effective constants. Complex electromagnetic parameters of the material are calculated by matching theoretical and measured effective values where theoretical values can be evaluated by the formulas:

$$\left.\varepsilon_r^{th} = \frac{L}{L_0}, \quad \mu_r^{th} = \frac{C}{C_0}\right\} \tag{16}$$

L and C are the inductance and capacitance of per unit length of strip-line with sample under/on it, L_0 and C_0 are the inductance and capacitance per unit length of strip-line without sample under/on it. L, C, L_0, C_0 are calculated on basis of the quasi-static theory based on the Green's potential functions and on the transverse transmission line method, allows, for multilayered dielectrics, the calculation of the characteristic impedance and of the propagation constant of a transmission line [Crampagne *et al.* 1978]. Based on the Kaneki's relations, this approach can be used for magnetic media [Kaneki 1969].

Errors equations

$$\left.\begin{array}{l} G(\varepsilon', \varepsilon'') = \left|\varepsilon_r^* - \varepsilon_r^{th}\right|^2, \\ F(\mu', \mu'') = \left|\mu_r^* - \mu_r^{th}\right|^2 \end{array}\right\} \tag{17}$$

for the complex dielectric and magnetic constants of the material are solved using a dichotomous procedure in the complex plane.

Last considered test device has a very important advantage in comparison with the test device proposed in reference [Queffelec *et al.* 1994]: it allows to apply a biasing magnetic field to magnetic materials under measurement (if it requires) due to the absence of lateral walls, so a measurement system built on the basis of this device can be alternative relative to FSM measurement system. Moreover, the device can be adapted for measuring thin films which is very difficult in case of FSM measurement system. But as proposed in reference [Pozar 1998], the measurement system on basis of this test device exhibits essential fluctuations of measured parameters that is a serious disadvantage especially in the case of measuring thin films. These fluctuations are due to approximate mathematical model based on the principle of quasi-TEM and quasi-static theory. Thus the comparison and analysis of all of the considered methods and devices allow us to assume that for optimal design of a measurement system can be reached if mathematical model of the device of reference [Barry 1986] will be adapted to a device considered in reference [Salahun *et al.* 2001] since this model has shown quite smooth dependencies of the complex constants versus frequency in microwave region.

RESTORATION OF PERMEABILITY TENSOR

Characterization of magnetic materials especially ferrites requires to restore the permeability tensor. Exciting methods [Mallegol 2003 a,b, Queffelec *et al.* 2002] are directed to specific use. It means that one has to obtain the permeability tensor of magnetic materials without having a possibility to measure complex permeability/magnetic constant and permittivity/dielectric constant. We show for the case of ferrites that preferred design of test device enables us to obtain the complex permeability tensor due to a possibility to make measurement of complex dielectric and magnetic constants simultaneously.

Let us apply a biasing magnetic field directed parallel to the direction of propagation of an initial microwave signal. Let us suppose also that initial microwave signal is a circularity polarized plane TEM wave. If permeability tensor of ferrite is to be defined by formulas [Pozar 1998]:

$$\hat{\mu}^* = \begin{pmatrix} \mu & ik & 0 \\ -ik & \mu & 0 \\ 0 & 0 & \mu_0 \end{pmatrix} \tag{18}$$

then there is a relation between the complex constants of test ferrite and the permeability tensor elements [Pozar 1998]:

$$\left(\frac{\gamma_\pm^*}{\omega}\right) = \varepsilon_0 \varepsilon_r^* (\mu \pm k) \tag{19}$$

where “+” stands for the case of a right-hand circularity polarized plane TEM wave and “-” stands for the case of a left-hand circularity polarized plane TEM wave. In the case of a biasing field directed perpendicular to the propagation of an initial microwave signal relation between the complex constants of test ferrite and the permeability tensor elements is to be defined by the formula [Pozar 1998]:

$$\mu_0 \mu_r^* = \frac{\mu^2 - k^2}{\mu}. \qquad (20)$$

Solving together the equations (19-20) gives the permeability tensor elements μ and k.

CONCLUSIONS

A review of measurement methods and devices to retrieval the expressions of complex dielectric and magnetic constants of materials in microwave frequency range were presented. The review has been done in order to find out optimal design of a test device with a viewpoint of cost effectiveness of the device and suitability of mathematical model to extract the complex expressions of constants Emphasis has been made on a possibility for characterizing materials including magnetic materials and thin films. A classification of measurement methods for characterizing materials was made preliminary. The cost-effective and simple design for measuring test device was found together with the way to obtain an expression for permeability tensor of ferrites like materials using measured complex dielectric and magnetic constants of materials.

Practical realization of the measurement system on the basis of proposed design of the test device is currently under consideration.

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