Broadband Permittivity and Permeability Extraction of Ferrite Cores up to the GHz Range via Measurements and Simulations

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Abstract **─** In this contribution a method is presented which allows for the characterization of the magnetic (µ'-jµ'') as well as the dielectric (ε'-jε'') properties of ferrites in a broad frequency band (kHz to GHz). In order to determine the material properties at certain frequencies, the simulation model parameters of a particular test setup are tuned via optimization such that the simulated response of a sample to an electromagnetic excitation matches to the measured one. A loop and parallel plate setup support a low frequency parameter extraction of the sample while a coplanar line is used for high frequencies. The extracted material properties are fitted by a broadband causal fit in order to obtain a material model for the whole frequency range. The presented method is verified for a setup with a coil and a ferrite core. The results show that dielectric properties of the ferrite core cannot be neglected for microwave frequencies.

Index Terms **─** Broadband model, EM-simulation, ferrites, material characterization, wireless power.

I. INTRODUCTION

Wireless power transfer applications are emerging in the recent past. In many applications, additional to the power transmission data has to be transferred at the same time [1]. Moreover, often the power coils and data antennas have to be placed in close proximity to each other.

When designing and optimizing such a system containing closely coupled antennas and coils, numerical field simulation techniques are applied widely. For accurate simulation results, a good knowledge of the geometry and the material properties of the components are mandatory in the whole frequency range spanned by the power and data transmission links. The magnetic behavior in the GHz range as well as the dielectric material properties (i.e., the complex permittivity) in the whole frequency range of the used ferrite cores are totally unknown in most cases.

In order to overcome this lack of information, we present an approach which allows for the determination of the complex permeability and the complex permittivity over a broad frequency range (kHz to GHz) for material samples composed of ferrite. In contrast to existing measurement techniques [2-4], the method proposed in this work does not demand for a special preparation or machining of the investigated core sample as long as a closed high permeable path and partially flat and parallel surfaces can be ensured. Examples of feasible samples are a ferrite plate with an arbitrary hole and a pot core investigated in this work.

In Section II, the determination of the material parameters at low frequencies up to a few MHz is focused on. Section III details the extraction of the material parameters at higher frequencies. In order to obtain a material model being feasible for both LF and HF simulations, in Section IV a broadband causal fit is introduced. The determined material parameters are validated in Section V with a typical test structure.

II. LF-PARAMETER EXTRACTION

In order to extract the electromagnetic material properties of simple sample geometries like toroidal cores or brick shaped samples, analytical methods are available [4, 5], which allow for directly computing the material properties from measurement data. Here, measurement equipment with special probe adapters is required. However, for arbitrary geometries, a more general determination procedure is of interest. For this purpose, in this section a material parameter extraction method for low frequency (LF) is described which is based on a comparison between measured and simulated data. LF measurements can be carried out with either impedance analyzers or network analyzers being suited for kHz and MHz frequencies. The procedure is detailed

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with a ferrite pot core [6] with the material "N48" [7] as the material under test (MUT).

As presented in Fig. 1, two measurement setups are used in order to determine the complex permeability $\underline{\mu} = \mu_0(\mu_r' - j\mu_r'')$ and the complex permittivity $\underline{\varepsilon} = \varepsilon_0(\varepsilon_r' - j\varepsilon_r'')$ with μ_0 being the vacuum permeability and ε_0 the vacuum permittivity, respectively. The following iterative procedure similar to [8] was developed which enables the material parameter determination of samples with arbitrary shapes.

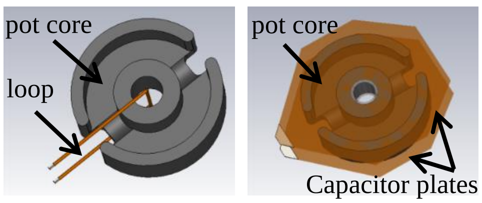

Fig. 1. Model of the LF measurement setups, left: loop setup for determining R and L; right: parallel plate setup for determining G and C.

Initially, the magnetic and electric material parameters are guesses. A frequency f_1 is selected from a given vector containing M frequencies at which the material parameters are sought. The series inductance L and resistance R of a loop enclosing the MUT as shown in Fig. 1 are computed at the selected frequency by using the finite element method simulation tool CST Microwave Studio. The simulation model contains 24.4k and 100k tetrahedron mesh cells for the inductive and capacitive setup respectively utilizing third order basis functions. Comparing the measured and computed impedances, a goal function can be determined:

$$\Delta = \left|(\Delta A_S - \Delta A_M)/\Delta A_M\right| + \left|(\Delta B_S - \Delta B_M)/\Delta B_M\right|. \qquad (1)$$

In the above equation ΔA_S and ΔB_S represent the simulated series resistance and inductance change of the loop with respect to the setup without the MUT, whereas ΔA_M and ΔB_M are the measured counterparts.

In order to minimize (1) a trust region optimization alters the material parameters. If the goal function is lower than a predefined boundary, the optimization loop is aborted. The electric parameters are extracted in a similar way from the parallel plate setup. Equivalently to above, the goal function (1) is computed in this case with ΔA_M and ΔB_M representing the simulated change of the conductance and capacitance of the parallel plated setup respectively when the MUT is inserted into the parallel plate setup, and ΔA_M as well as ΔB_M are the measured quantities. As the permittivity of the MUT might influence the loop impedance, the magnetic parameters are optimized again with the found dielectric parameters. The whole procedure is repeated N times. In our investigation three iterations proved feasible.

If for all given frequencies the material parameters are determined the task is finished. The material parameters of the previous frequency samples are used as starting values for optimizations at the next frequency. The number of simulations involved in the optimization can be reduced by this if the material parameters do not change significantly at two adjacent elements of the frequency vector. In [8] a sketch of a similar optimization sequence can be found.

It should be mentioned that the presented method can be applied to many other ferrite sample geometries like U-cores [8], E-cores, disks, and alike.

III. HF-PARAMETER EXTRACTION

In this section, the extraction method of the material parameters at high frequencies (HF) is focused on since the previous setups (closed loop and parallel plate capacitor) are motivated by mainly independent responses of the setups to magnetic and dielectric properties of the MUT which is no longer valid at high frequencies. Here, all electromagnetic properties must be taken into account simultaneously. For this reason, analog to [9] the MUT is placed on top of a coplanar transmission line and the two port scattering parameters are measured over a broad frequency range.

In a similar way as in the previous section simulations and measurements are compared in order to determine the material parameters. However, as presented in Fig. 2, in this case one single measurement and simulation setup is used in order to determine all four material parameters. Here, the FEM model consists of 41k mesh cells using third order basis functions.

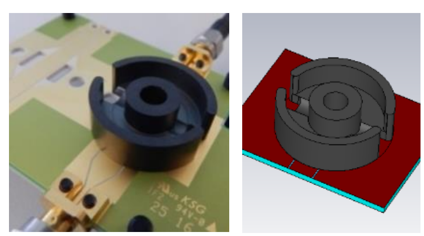

Fig. 2. Ferrite pot core on top of a transmission line, left: measurement setup; right: simulation model.

The complex scattering parameters of the coplanar line with the MUT are first measured and repeatedly simulated afterwards. As the structure is symmetric $\underline{S}_{11} = \underline{S}_{22}$ and reciprocal $\underline{S}_{12} = \underline{S}_{21}$ it is sufficient to consider only two of the four scattering parameters. A goal function is defined as:

$$\Delta_S = \left|(\underline{S}_{11,S} - \underline{S}_{11,M})/\underline{S}_{11,M}\right| + \left|\underline{S}_{12,S} - \underline{S}_{12,M}/\underline{S}_{12,M}\right|. \qquad (2)$$

In order to minimize the goal function (2) the complex permittivity and the complex permeability are perturbed at the same time for each frequency $\{f_1, f_2, \ldots f_M\}$ given by the frequency vector. Again, a trust region optimization is applied for the minimization task. The influence of the connectors as well as the tapering section of the transmission line is removed as suggested in [10] by application of a through-reflect-line calibration.

In Fig. 3 it can be seen that the permeability curve gained with the HF-extraction technique in the range from 70 MHz to 100 MHz is in good agreement with the LF-extraction results. For lower frequencies the LF data with the closed magnetic path provide the most accurate results. Here, the extracted HF material parameters diverge due to a low sensitivity of the coplanar line setup at low frequencies. The relative permeability approaches the properties of vacuum at frequencies in the GHz range.

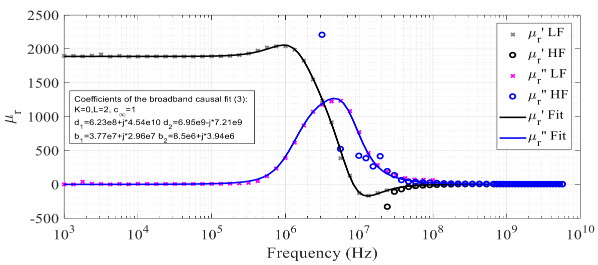

Fig. 3. Extracted relative permeability of the pot core from the LF-extraction (x markers), the HF-extraction method (circles) and the broadband causal model response (solid lines).

In Fig. 4 a steep drop in the real part of the relative permittivity gained from the LF approach is not present in the HF-data. Both, the real and the imaginary parts show material properties significant different from those of vacuum in the whole frequency range.

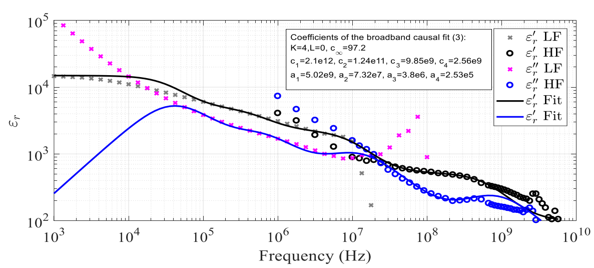

Fig. 4. Extracted relative permittivity of the pot core from the LF-extraction using Fig. 3 marker and line definitions.

IV. BROADBAND CAUSAL MODEL

Until now, two different approaches (LF and HF) with three different test scenarios (loop, capacitor, transmission line) have been set up in order to determine the material properties of the MUT at certain individual frequency samples. However, for numerical field simulation in time domain like the FDTD method and transient models derived from field simulation data (e.g., SPICE models), a causal response of the material data [11] is important. Such a model is advantageous also for simulations in the frequency domain as it guarantees a smooth behavior across the frequency spectrum.

From the vector fitting technique it is known that any linear time invariant system response can be approximated by poles and residues of a rational function [12] which is causal by definition. Therefore in order to describe the material parameters with a causal model it is necessary to find the poles a_p, residues c_∞, c_p and the complex conjugated pairs b_q, b_q^* and d_q, d_q^* giving the complex frequency response:

$$\{\underline{\mu}, \underline{\varepsilon}\} = c_\infty + \sum_{p=1}^{K} \frac{c_p}{a_p + j\omega} + \sum_{q=1}^{L} \left(\frac{d_q}{b_q + j\omega} + \frac{d_q^*}{b_q^* + j\omega} \right). \quad (3)$$

In (3), K represents the order of single poles and residues and L represents the complex conjugated order of the model. In Fig. 3 and Fig. 4 the determined parameters of the model of the investigated pot core model as well as the frequency response are displayed.

V. VALIDATION

In order to validate the determined material model for the MUT, a loop was designed. The loop is a preferable structure for wireless power and data transmission as it provides a rotation invariant coupling between a transmitter and a receiver. The loop structure realized on a printed circuit board (PCB) is inserted into the core as depicted in Fig. 5 and the impedance of the loop is measured in a broad frequency range via a network analyzer.

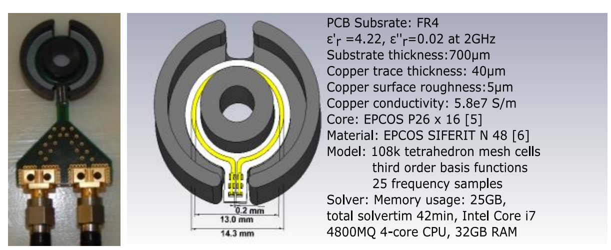

Fig. 5. Photo and model of the test loop antenna.

To determine the loop impedance, as shown in Fig. 5 the scattering parameters at the two coaxial terminals are measured. The impedance is computed from the differential scattering parameters which are gained from the nodal scattering parameters via:

$$\underline{S}_{dd} = 0.5\left(\underline{S}_{11} - \underline{S}_{21} - \underline{S}_{12} + \underline{S}_{22}\right). \quad (4)$$

The influence of the connectors as well as the feeding structure is removed with a differential open-short-load calibration.

In Fig. 6 and Fig. 7 it can be seen that measurement and simulation of the loop resistance as well as the loop inductance is in a good match when the causal model of the extracted material data is used. In the frequency range between 100 MHz and 1 GHz, a deviation between measurement and simulation with the extracted material data is visible which can be explained by an underestimated air gap in the HF-material parameter extraction [10] and an imperfect calibration of the loop. Nevertheless, the simulated loop impedance using the provided manufacturer data shows much larger deviations in the loop resistance as well as the loop inductance in the hole investigated frequency range showing that the permittivity of the ferrite core strongly influences the loop impedance in the whole investigated frequency range. Since no permittivity information is given by the manufacturer [7], $\varepsilon_r = 1 - j0$ was assumed in this case.

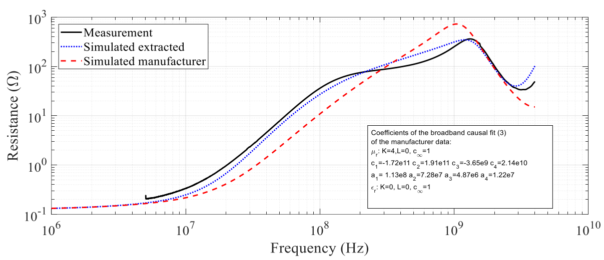

Fig. 6. Comparison of the loop resistance: Measured (solid line), simulated with causal fit of the extracted data (dotted line) and simulated with causal fit of the manufacturer provided data (dashed line).

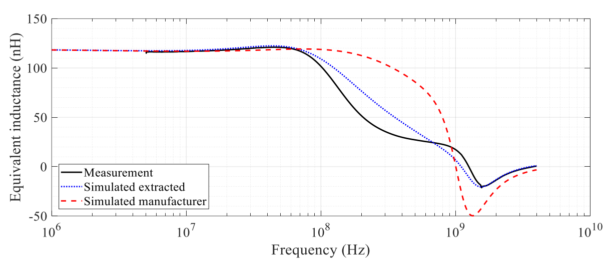

Fig. 7. Comparison of the loop inductance. The same line definitions as introduced in Fig. 6 are used.

VI. CONCLUSION

In this work a method was presented allowing for the broadband determination of the permeability and permittivity of ferrite cores typically applied in wireless power transfer systems. Two measurement setups for the low frequency parameter extraction as well as one setup for high frequency characterization were presented. A causal material model is described allowing for stable time domain simulations. The method is validated with a loop antenna inserted into the core showing a good agreement between the measurement and simulation when the extracted material data is used. The ferrite core permittivity, in general not provided by the manufacturer, strongly influences the impedance of structures close to the investigated pot core.

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