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NEW WAVEGUIDE METHOD FOR DIELECTRIC PARAMETER MEASUREMENT

Abstract. *Perfect knowledge of dielectric parameters is necessary for its application in various devices. In spite of the whole range of measurement techniques, their practical implementation in the microwave frequency band runs into some difficulties. This article describes a new method for nonmagnetic dielectrics permittivity and loss tangent measurement in the microwave frequency band. A dielectric specimen slab is placed in the short-circuited waveguide section normal to its axis and fills the whole cross-section of the waveguide at approximately quarter wavelength from its short-circuited endpoint. By means of the vector network analyzer the waveguide section reflection factor is measured. Objective function is determined as difference between calculated and measured module and phase of the reflection factor. Specific code for objective function calculation and its minimization is worked out. Minimization of this function by varying dielectric parameters makes it possible to find real values of these parameters. The method needs no de-embedding and can be used with non-calibrated waveguide-to-coax transitions. Also it is less sensitive to the noise component of reflected signal. The testing results show that new method's error does not exceed 0.2 % for relative permittivity and 1% for dielectric loss tangent.*

Key words: dielectric parameters, microwaves, waveguides, optimization methods

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НОВЫЙ ВОЛНОВОДНЫЙ МЕТОД ИЗМЕРЕНИЯ ПАРАМЕТРОВ ДИЭЛЕКТРИКОВ

Аннотация. *Точное знание параметров диэлектрика необходимо при его применении в самых различных устройствах. Несмотря на наличие целого ряда известных методов измерения этих параметров, практическое их применение в микроволновом диапазоне частот наталкивается на ряд трудностей. В данной статье описан новый волноводный метод измерения диэлектрической проницаемости и тангенса угла потерь немагнитных диэлектриков в микроволновом диапазоне. Пластина диэлектрика помещается в короткозамкнутый отрезок волновода перпендикулярно его оси, заполняя все поперечное сечение на расстоянии примерно четверти длины волны от короткозамкнутого конца отрезка. С помощью векторного анализатора цепей измеряется коэффициент отражения от входа волновода. Для определения параметров диэлектрика по этим данным составлена программа вычисления и минимизации целевой функции, которая определяется как разность между вычисленными значениями модуля и фазы коэффициента отражения на входе волновода и измеренными значениями этого коэффициента. Минимизация этой функции при варьировании параметров диэлектрика позволяет определить указанные параметры. По сравнению с известными, представленный в настоящей статье метод не требует переноса плоскостей отсчета векторного анализатора цепей к поверхностям образца и менее чувствителен к шумовой составляющей измерительного сигнала. Это позволяет использовать при измерении некалиброванные коаксиально-волноводные переходы. По результатам тестирования метода погрешность измерения относительной диэлектрической проницаемости не превышает 0,2 %, а тангенса угла диэлектрических потерь – 1 %.*

Ключевые слова: параметры диэлектриков, микроволны, волноводы, методы оптимизации

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Introduction. Dielectrics permittivity and loss tangent measurement in the microwave frequency band has always been the subject of research and development [1], [2]. Lately these measurements have gained in particular importance due to appearance and use of new materials in microwave engineering, such as SiC absorbers, meta-materials, etc. Although there is a lot of well-known measuring techniques, not all of them are suitable for the solution of this particular problem. For example, cavity methods are hardly suitable for broadband measuring; methods based on coaxial transmission line require sample pieces in the form of the disk with a hole that are difficult to produce from fragile solid materials; methods using open-circuit systems often fail to provide adequate accuracy.

Waveguide methods are the most suitable for broadband frequency measuring. For instance, the well-known Nicolson-Ross-Weir (NRW) method [3], [4] is widely used for parameters measuring of both nonmagnetic and magnetic dielectrics. The method is based on the measurement of the scattering matrix (S-matrix) of the waveguide section with a test material slab. To evaluate the sample parameters the method uses scattering matrix defined between the front and back planes of the slab.

Since a vector network analyzer measures S-parameters with respect to reference planes, the method requires de-embedding input and output port reference planes to the sample surface. For de-embedding uniqueness, group velocity is calculated $d\omega/d\beta$, where ω represents angular frequency; β is constant phase in the waveguide. Based on this method in particular, a measuring unit was designed and the software for processing measurement results was elaborated [5]. However, during the operation it was found that the noise in measuring signal of Rohde&Shcwarz ZVL-13 vector network analyzer [6] results in different signs of group velocity values obtained at different frequencies. This leads to wrong measurement results. The attempts to smooth over the measured S-parameters dependencies did not give the desired effect. Due to this, the present article proposes a new waveguide method for measuring parameters.

Method description. The method uses a waveguide short-circuited section as a measuring chamber containing a specimen of dielectric in the form of the slab entirely filling the waveguide cross section (Fig. 1). The dielectric is supposed to be nonmagnetic (relative magnetic permeability $\mu_r = 1$).

Let us denote the distance from the waveguide short circuit to the specimen by l_1, the distance from the specimen to the input port by l_2 and slab thickness by l. It is assumed that only H_{10}-mode propagates in the air-filled waveguide sections.

Boundary conditions on the air-dielectric interface require continuous of tangential electric and magnetic fields. It follows therefrom that H_{10}-mode also propagates in dielectric-filled waveguide section and there is no higher mode excitation even though their propagation in this section is possible.

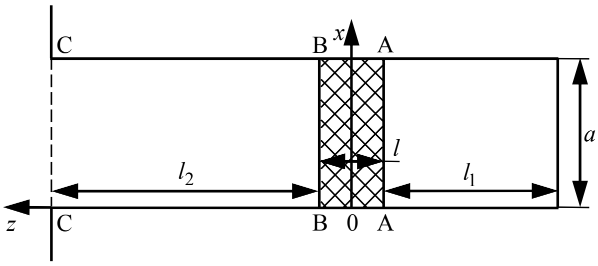

Fig. 1

Let us find the waveguide section input impedance. The short-circuited waveguide section input impedance in A–A plane is

$$Z_{AA} = iZ_{g0}\tan(\beta_0 l_1),$$

where $Z_{g0} = (2b/a)\eta_0\left[1-(\lambda/\lambda_c)^2\right]^{-1/2}$ is wave impedance of free-space waveguide; $\beta_0 = 2\pi/\lambda_g$ is its phase constant with b, a being the dimensions of the waveguide narrow and wide sides, respectively; $\eta_0 = 120\pi$ Ом is free space wave impedance, λ is free space wavelength; $\lambda_c = 2a$ is the cut-off wavelength in the free-space waveguide, $\lambda_g = \lambda\Big/\sqrt{1-(\lambda/\lambda_c)^2}$ is the wavelength in the free-space waveguide.

This impedance serves as the load impedance for the dielectric-filled waveguide section (crosshatched in Fig. 1). Its input impedance is

$$Z_{BB} = Z_{g1}\frac{Z_{AA} + iZ_{g1}\tan(\gamma_1 l)}{Z_{g1} + iZ_{AA}\tan(\gamma_1 l)},$$

where

$$Z_{g1} = \frac{(2b/a)\dot{\varepsilon}_r^{-1/2}}{\sqrt{1-(\lambda/\lambda_c)^2/\dot{\varepsilon}_r}};\quad \gamma_1 = \frac{2\pi\dot{\varepsilon}_r^{1/2}}{\lambda}\sqrt{1-(\lambda/\lambda_c)^2/\dot{\varepsilon}_r}$$

are wave impedance and propagation constant of waveguide section with the sample piece with $\dot{\varepsilon}_r = \varepsilon_r (1 - i \tan \delta_\varepsilon)$ being complex permittivity of the sample ($\tan \delta_\varepsilon$ is dielectric loss tangent).

Input impedance in the C–C plane can be found in a similar way:

$$Z_{CC} = Z_{g0} \frac{Z_{BB} + iZ_{g0} \tan(\beta_0 l_2)}{Z_{g0} + iZ_{BB} \tan(\beta_0 l_2)}.$$

Reflection factor of the measuring chamber is:

$$\Gamma = \frac{Z_{CC} - Z_{g0}}{Z_{CC} + Z_{g0}}.$$

It is obvious that with properly selected values of ε_r and $\tan \delta_\varepsilon$ the calculated and measured reflection factors must be the same. Hence, we can derive objective function:

$$F = (|\Gamma| - |\Gamma_m|)^2 + \alpha(\varphi - \varphi_m)^2,$$

where $|\Gamma_m|$, φ_m present modulus and phase of the measured reflection factor; α is a weight factor defined experimentally. Minimization of this function when using ε_r and $\tan \delta_\varepsilon$ as varying parameters makes it possible to find parameters of the measured material.

In order to calculate objective function and to minimize it, the special program "EPS" was written in MATLAB. Since the objective function is not unimodal, the program uses the genetic algorithm [7] for the global minimum search.

Measuring unit and measurement results. Measuring unit (Fig. 2) consists of the vector network analyzer Rohde&Shcwarz ZVL-13 *1*, coaxial cable *2*, waveguide-to-coax transition (WTC) *3* and measuring chamber *4* with a specimen. The analyzer transmits measurement results to the computer *5*. The unit allows measurement in a given frequency band with a fixed frequency step.

The measuring chamber is a section of standard rectangular waveguide with the length of 40 mm and cross section $23 \times 10\ \text{mm}^2$. We used 2 mm thick specimen placed at a distance of 19 mm from the shorted end. The chamber is driven by the vector network analyzer Rohde&Shcwarz ZVL-13 by means of WTC. The distance between the short circuit plane and the specimen makes approximately a quarter-wavelength in the waveguide. In this case the electric field near the specimen and hence the method sensitivity is maximal.

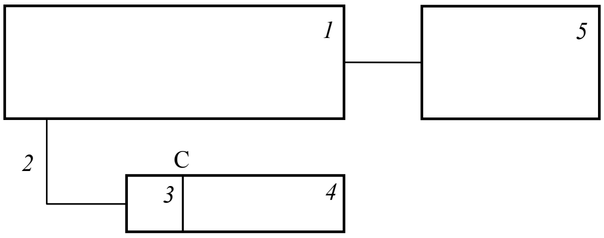

Fig. 2

The method was tested by means of mathematical simulation of measuring chamber using the RFS program [8]. Dependence of the electric field on the y-coordinate (Fig. 1) in different cross sections is shown on Fig. 3. The input power is 1 W. The cross section $x = 0$ is in the middle of the specimen, the cross section $x = -18$ mm is close to the shorted end of the section, and the cross section $x = 18$ mm is close to the input port (Fig.1). Field distribution corresponding to H_{10} mode and free from higher modes was observed in all the cross sections.

Reflection factor s_{11} of the measuring chamber with the specimen was calculated in RFS program and then entered in EPS program. The program calculation results were compared with the specimen preset parameters in RFS. The weighting factor α in the objective function was taken to be equal to one.

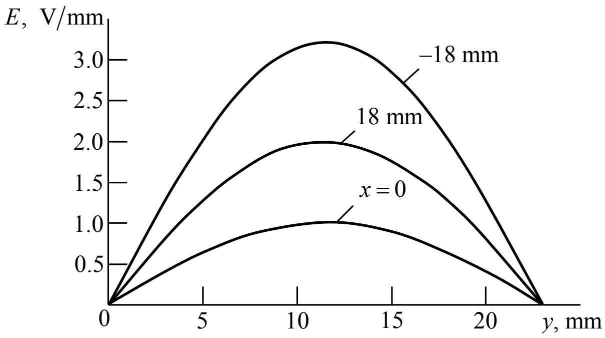

Fig. 3

Table 1

Frequency, GHz	Parameters		Test results		Relative error, %	
	ε_r	$\tan \delta_\varepsilon$	ε_r	$\tan \delta_\varepsilon$	ε_r	$\tan \delta_\varepsilon$
8	2	0.0003	2.000471	0.000303	0.023	1.0
	14	0.3	14.00618	0.300226	0.065	0.07
10	2	0.0003	2.000544	0.000301	0.0272	0.33
	14	0.3	14.02879	0.300171	0.205	0.057
12	2	0.0003	2.001372	0.000325	0.0686	0.83
	14	0.3	13.99038	0.299225	–0.068	–0.25

Table 1 provides data for PTFE (Teflon) having low ε_r and low losses, as well as for polycrystalline SiC-based absorbing material $(\varepsilon_r = 14;\ \tan\delta_\varepsilon = 0.3)$. As we can see, error of dielectric permittivity does not exceed 0.1 %, and 1 % for dissipation factor. According to the measurements, these results do not depend on the length of the measuring chamber, as well as on the position of the specimen in it, which is an apparent advantage of the proposed technique. The similar results are obtained for the specimens with different thickness and made from different materials. Note that the use of this method avoids the necessity for reference planes de-embedding.

Estimation of the method errors. Main error sources of the method are: inexact thickness of the specimen, its gapping with the walls of the waveguide and inaccurate sizing of l_1 and l_2.

Since analytical calculation of the error in this case is impossible, mathematical simulation of the chamber was performed with the size and position of the sample deviating from the values set in the result processing program.

The simulation was performed by means of the RFS program, with the same size of the measuring chamber, as it was set during the testing. As a result, the error sensitivity was found for a typical 2 mm thick specimen with $\varepsilon_r = 10$ and $\tan\delta_\varepsilon = 0.3$. Simulation was performed on 10 GHz frequency. Relative sensitivity of q parameter to resizing of p argument was calculated by the following formula

$$\delta(q)_p = \left(\frac{q - q_0}{q_0}\right) \Big/ \left(\frac{p - p_0}{p_0}\right),$$

where q_0, p_0 are the reference values of the function and the argument respectively.

The obtained values of sensitivity factors are shown in Table 2. According to the table, with the reference dimensions the proposed method induces dielectric permittivity error of 0.08 % and loss tangent error of 0.07 %. The specimen thickness provides the most significant effect on the results of calculation. 0.1 mm error in its setting (with the reference value of 2 mm) results in changes ε_r and loss tangent error by 13 and 3 % respectively. Moreover, the specimen gapping with the walls of the waveguide is of critical importance. 0.1 mm gap along the wide wall gives 6.48 % error in ε_r, and 5.63 % error in loss tangent.

Before making measurements, it is necessary to calibrate the circuit analyzer in the specified frequency range with respect to C–C reference plane lying at the junction of WTC and measuring chamber (see Fig. 2). For this purpose, the circuit analyzer is configured with a special procedure. As it is mentioned in the circuit analyzer manual [9], this is a narrowband type of calibration, thus the measurements are made in comparatively narrow band of frequencies.

Fig. 4 demonstrates the A Caprolone ε_r and $\tan\delta_\varepsilon$ measurement results on the thickness of the specimen. They are within the range of reference values [10] for this type of material.

Table 3 shows measurement results for some dielectrics at the frequency of 10 GHz. Comparison with the reference parameters [8] shows that the measured parameter values are within the range of reference values. Comparatively big difference of ceramics 22XC loss tangent from the reference values may be caused by nonideal termination during calibration of the equipment.

It is necessary to mention the importance of proper calibration of the circuit analyzer in the plane

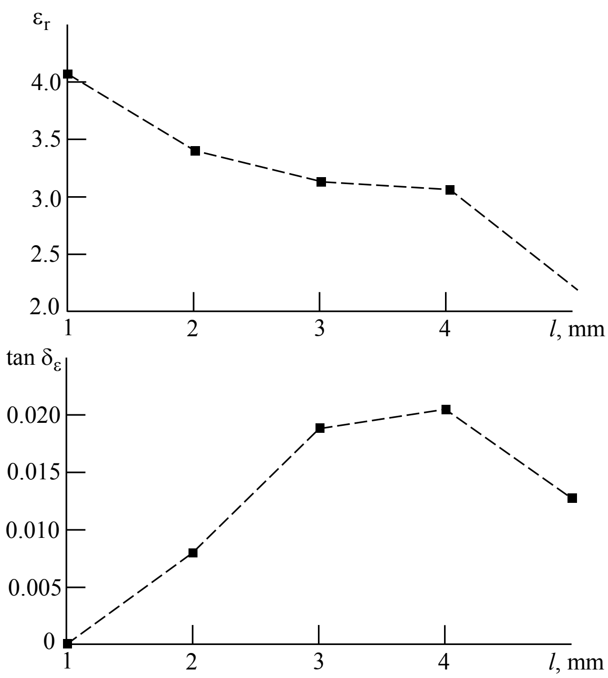

Fig. 4

Table 2

l_1, mm	l_2, mm	l, mm	b_1, mm	ε_r	$\tan\delta_\varepsilon$	$\delta(\varepsilon_r)$	$\delta(\tan\delta_e)$
10.0	20.0	2.0	10.0	10.0082	0.3002	–	–
9,9	20.0	2.0	10.0	9.9325	0.3089	0.13	–0.58
10.0	19.9	2.0	10.0	9.2793	0.2881	0.4	–0.37
10.0	20.0	1.9	10.0	9.3523	0.2831	13.2	–3.2
10.0	20.0	2.0	9.9	8.7949	0.2623	6.48	5.63

Table 3

Material	*t*, mm	Reference parameters [9]		Measured parameters	
		ε_r	$\tan\delta_\varepsilon$	ε_r	$\tan\delta_\varepsilon$
Caprolone A	2	3…4	0.025	3.38	0.027
Caprolone B	2	–	–	2.28	0.11
Abrasive SiC	2	–	–	16.17	0.08
Ceramic 22XC	1	9.3	0.0015	9.48	0.0063

of the WTC waveguide flange. Calibration involves subsequently attaching short circuiting plate, section of short-circuited waveguide of specified length and termination to the WTC. Here, it is very important to provide fail-safe connection and aligning of the WTC and calibration components. VSWR of operating frequency termination have not to exceed 1.05. Not meeting these requirements leads to law accuracy of measurements.

Conclusion. The method described allows measuring dielectric permittivity and loss tangent of nonmagnetic solid dielectrics in broad frequency band with sufficient accuracy. The specimen is to be in the form of a slab sized according to the waveguide cross sectional dimensions. The method requires accurate specification of the specimen thickness, the measuring waveguide length and the specimen placing in it. The specimen is to be fixed without any waveguide walls-to-specimen gaps and the network analyzer with WTC is to be carefully calibrated. To process the measurement results a special software written in MATLAB is used.

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