IMPROVED THE MULTIPLE SIGNAL CLASSIFICATION ALGORITHM TO ESTIMATE THE COMPLEX RELATIVE PERMITTIVITY OF MATERIAL BASED ON THE REFLECTION MEASUREMENT IN FREE-SPACE AT X-BAND

CẢI TIẾN THUẬT TOÁN PHÂN LOẠI ĐA TÍN HIỆU ĐỂ ƯỚC LƯỢNG ĐIỆN MÔI TƯƠNG ĐỐI PHỨC CỦA VẬT LIỆU DỰA TRÊN PHÉP ĐO PHẢN XẠ TRONG KHÔNG GIAN TỰ DO Ở BĂNG TẦN X

Ho Manh Cuong, Le Trong Hieu

Electric Power University

Ngày nhận bài: 23/03/2020, Ngày chấp nhận đăng: 14/07/2020, Phản biện: TS. Hoàng Phương Chi

Abstract:

This paper aims to improve the multiple signal classification (MUSIC) algorithm to estimate the complex relative permittivity of a metal-backed planar material sample placed in a free-space based on reflection measurement at X-band. The measurement system consists of a pyramidal horn antena operating at X-band and the material sample with the thickness is changed. From the measured values of the reflection coefficients and a known thickness of a planar slab of the material samples, the complex relative permittivity of the material sample is estimated by the proposed algorithm. The proposed algorithm is verified with different thickness Teflon-PTFE materials at X-band. The estimation results show that the complex relative permittivity of a large thickness sample is more accurate than that of a small thickness one.

Keywords:

Complex relative permittivity, dielectric constant, dielectric loss tangent, MUSIC (Multiple Signal Classification), CST (Computer Simulation Technology).

Tóm tắt:

Mục đích của bài báo này nhằm cải tiến thuật toán phân loại đa tín hiệu (MUSIC) để ước lượng điện môi tương đối phức của mẫu vật liệu phẳng với mặt sau tráng kim loại được đặt trong không gian tự do dựa trên phép đo phản xạ ở băng tần X. Hệ thống đo lường bao gồm một anten loa tháp hoạt động ở băng tần X và mẫu vật liệu với độ dày được thay đổi. Từ các giá trị đo được của các hệ số phản xạ và độ dày đã biết của mẫu vật liệu, điện môi tương đối phức của mẫu vật liệu được ước lượng bởi thuật toán đề xuất. Thuật toán đề xuất được kiểm chứng với vật liệu Teflon-PTFE có độ dày khác nhau ở băng tần X. Kết quả ước lượng chỉ ra rằng điện môi tương đối phức của mẫu vật liệu có độ dày lớn chính xác hơn so với mẫu vật liệu có độ dày nhỏ.

Từ khóa:

Điện môi tương đối phức, hằng số điện môi, tổn hao điện môi, MUSIC (phân loại đa tín hiệu), CST (công nghệ mô phỏng máy tính).

1. INTRODUCTION

The reflection method is a type of nonresonant method, the properties of a sample are obtained from the reflection due to the impedance discontinuity caused by the presence of the sample in a transmission structure. In the measurement of the effective permittivity of composite materials, it is required that the sample dimensions should be much larger than the sizes of the inclusions. For composites with inclusions whose sizes are comparable with the wavelength of the microwave signal, for example, fiber composites, the conventional coaxial line, and waveguide methods cannot be used. In these cases, free-space methods are often used. Besides, the free-space methods for determining the parameters of material are nondestructive, contactless, and sample preparation requirements are minimal. Therefore, they are especially suitable for the measurement of the parameters of material under high-temperature conditions [1, 2]. The free-space methods are based on the measurements of the phase of the reflection (S_{11}) and transmission (S_{21}) coefficient through a known thickness of the material samples. The reflection and transmission coefficients are determined by measuring the attenuation and phase shift introduced by a sample placed between two antenas. However, when the wavelength in the material sample is smaller than the thickness of the material under test, a phase problem is encountered [3-10]. Because phase-angle measurements are only possible between -180^o and $+180^o$. The total phase shift is measured using the vector network analyzer, shifted by n times 360^o, where n is an integer to be determined. This problem was solved by making measurements on samples of different thickness [11] or by the delay time if the wave is non-dispersive in the observed frequency range [12] or by the measurement technique, based on reflection, for thickness and permittivity determination [13] or by the based on selecting a sample thickness that allows the phase to remain within the measurement limits of the instrumentation at a given frequency or two frequencies [14].

In this paper, a high-resolution algorithm is proposed based on the improved MUSIC algorithm and exploiting the relationship of the permittivity and the refractive index of the material to solve the phase ambiguity problem. Therefore, the proposed algorithm can be estimated accurately the complex relative permittivity of the sample when the thickness is greater than the wavelength at X-Band without determining an integer time the wavelength in the sample under test.

The rest of this paper is organized as follows. Section 2 describes the reflection while the estimation procedure is presented in section 3. Section 4 is the implementation modeling by CST software. The estimation results are shown and discussed in section 5. Finally, a brief conclusion is given in section 6.

2. REFLECTION MEASUREMENT METHOD

Figure1 shows a typical setup for a free-space reflection measurement.

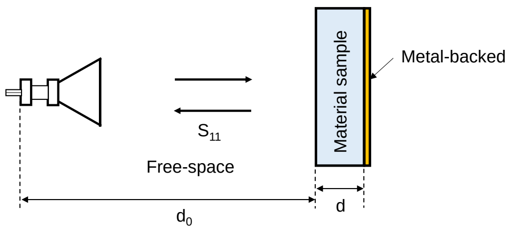

Figure 1. Free-space reflection method

Reflection measurement of the system is only considered on *D* discrete reflection points. If the multiple reflections are negligibly small, the number of signals is equal to *D*. The following formulation must also be applied to this measurement. It is assumed that the antena has *D* signal sources changing from f_1 to f_M (bandwidth equivalent to $f_M - f_1$) and radiating through the free-space and material sample to the metal-back then reflected back. The data obtained by the system above are the reflection coefficients at *M* uniformly sampled frequency points f_i ($i = 1,2,...,M : M > D$) with the same different frequency Δf between adjacent points. Delay time is represented corresponding to the k^{th} reflection point by t_k ($k = 1,2,...,D$). Then, the measurement value of reflected signals at the frequency f_i is given by

$$x_i(t) = \sum_{k=1}^{D} s_k(t)e^{-j2\pi f_i t_k} + w_i(t) \quad (1)$$

where $s_k(t)$ is the reflection coefficient of the k^{th} reflection point at the frequency f_i, $w_i(t)$ is the additive noise with zero mean and variance σ^2.

Figure 1 shows a planar sample of thickness *d* placed in free space. The complex permittivity, relative to free space, is defined as:

$$\varepsilon^* = \varepsilon' - j\varepsilon'' \quad (2)$$

where ε' and ε'' are the real and imaginary parts of the complex relative permittivity.

When the signal passes through the material sample, some part of it is always attenuated. To take this into account, an imaginary part is added to the refractive index. The complex refractive index (η^*) is defined as:

$$\eta^* = \eta' - j\eta'' \quad (3)$$

where η' and η'' are the real and imaginary parts of the refractive index.

The complex relative permeability of non-magnetic materials is equal 1. The complex refractive index is related to the complex permittivity by the following equation:

$$\eta^* = \sqrt{\varepsilon' - j\varepsilon''} \quad (4)$$

The time of arrival is also equal to delay time ($\tau = t_k$) and It is determined by

$$\tau = 2\left(d\frac{\sqrt{\varepsilon' - j\varepsilon''}}{c} + d_0\frac{1}{c} \right) \quad (5)$$

where d_0 is the distance between the port of antena and material sample, *c* is the light velocity in free-space.

The difference in phase ($\Delta\varphi_i$) of the signal when it goes through the medium (free space and material sample) at the first frequency and the i^{th} frequency is:

$$\Delta\varphi_{\mathrm{i}} = \frac{4\pi(i-1)\Delta f}{c}\left(d\sqrt{\varepsilon' - j\varepsilon''} + d_0\right) \tag{6}$$

Combine this with the phase difference, the equation (1) can be written using the vector notation as follows:

$$X = AS + W \tag{7}$$

where:

$X = \left[x_1(\mathrm{t}), x_2(\mathrm{t}), \ldots x_{\mathrm{M}}(\mathrm{t})\right]^T$, X is the $M\times1$ output vector measured at receiver, T denotes transpose.

$S = \left[s_1(\mathrm{t}), s_2(\mathrm{t}), \ldots, s_{\mathrm{D}}(\mathrm{t})\right]^T$, S is the vector of the k arriving signals.

$W = \left[w_1(\mathrm{t}), w_2(\mathrm{t}), \ldots, w_{\mathrm{M}}(\mathrm{t})\right]^T$, W is the noise vector:

$A = \left[a(\varepsilon'_1, \varepsilon''_1), a(\varepsilon'_2, \varepsilon''_2), \ldots, a((\varepsilon'_{\mathrm{D}}, \varepsilon''_{\mathrm{D}})\right]$, A is an MxD "parameter" matrix with $a(\varepsilon'_{\mathrm{k}}, \varepsilon''_{\mathrm{k}}) = \left[\mathrm{e}^{-j\Delta\varphi_1}, \mathrm{e}^{-j\Delta\varphi_2}, \ldots, \mathrm{e}^{-j\Delta\varphi_{\mathrm{i}}}\right]^T$ is a "parameter" vector of each signal.

3. ESTIMATION PROCEDURE

The multiple signal classification algorithm was proposed by R. Schmidt [15]. The basic approach of this algorithm is that from the received signal, the covariance matrix is calculated and then eigenvectors decomposition is carried out. The signal subspace and noise subspace are determined based on eigenvectors and eigenvalues. The results showed that the $M - D$ dimensional subspace spanned by the $M - D$ noise eigenvectors as the noise subspace and the D dimensional subspace spanned by the incident signal parameter vectors as the signal subspace; they are disjoint. The signal and the noise subspaces are calculated by matrix algebra and they are found to be orthogonal to each other. Therefore, the signal and noise subspaces are isolated by the orthogonal property of this algorithm. Thus, the complex relative permittivity of the material sample is estimated by combining the autocorrelation and MUSIC function of the received signal.

The corresponding data covariance matrix in equation (7) is given by

$$R_{\mathrm{XX}} = E\left\{XX^H\right\} = AS_{\mathrm{XX}}A^H + \sigma^2 I \tag{8}$$

where $S_{\mathrm{XX}} = E\left\{SS^H\right\}$ denotes the signal covariance matrix, I is the identity matrix, H denotes complex conjugate transpose.

The eigenvalues of R_{XX} are $\lambda_1, \lambda_2, \ldots, \lambda_{\mathrm{D}}$ such that:

$$det\left(R_{\mathrm{XX}} - \lambda_{\mathrm{i}} I\right) = 0 \tag{9}$$

Substituting (8) to (9):

$$det\left(AS_{\mathrm{XX}}A^H - (\lambda_{\mathrm{i}} - \sigma^2)I\right) = 0 \tag{10}$$

The eigenvalues ξ_{i} of $AS_{\mathrm{XX}}A^H$ are:

$$\xi_{\mathrm{i}} = \lambda_{\mathrm{i}} - \sigma^2 \tag{11}$$

If the eigenvalues ξ_{i} of $AS_{\mathrm{XX}}A^H$ are zero, $AS_{\mathrm{XX}}A^H$ is singular. This means that the number of incident wave fronts D is less than the number of frequency elements M. Thus, The minimum eigenvalue of R_{XX} is equivalent to σ^2 with multiplicity M - D. Therefore:

$$\lambda_1 \geq \lambda_2 \geq \cdots \geq \lambda_{\mathrm{D}} \succ \lambda_{\mathrm{D+1}} = \lambda_{\mathrm{D+2}} = \cdots = \lambda_{\mathrm{M}} = \sigma^2 \tag{12}$$

The eigenvector u_i associated with the eigenvalue λ_i satisfies the following equation:

$$(R_{XX} - \lambda_i I)u_i = 0 \tag{13}$$

For eigenvectors associated with the minimum eigenvalue, the (14) is suggested by substituting (8) and (12) into (13).

$$AS_{XX}A^H u_i = 0 \tag{14}$$

Since A has full rank and S_{XX} is non-singular, thus:

$$A^H u_i = 0 \tag{15}$$

This means that the eigenvectors corresponding to the minimum eigenvalue are orthogonal to the columns of the matrix A. Namely, they are orthogonal to the "parameter" vector of the signals:

$$\{u_{D+1},...,u_M\} \perp \{a(\varepsilon_1', \varepsilon_1''),\dots,a(\varepsilon_D', \varepsilon_D'')\} \tag{16}$$

It implies that the squared norm of $A^H u_i$ is zero

$$\|A^H u_i\|^2 = a^H(\varepsilon', \varepsilon'')U_M U_M^H a(\varepsilon', \varepsilon'') = 0 \tag{17}$$

where $U_M = [u_{D+1}, u_{D+2}, ..., u_M]$ represents the eigenvectors associated with the noise subspace of the covariance matrix R_{XX}.

The pseudo-spectrum of the MUSIC function as (18) is given by combining the autocorrelation function of signal subspace:

$$P_{MUSIC}(\varepsilon', \varepsilon'') = \frac{a^H(\varepsilon', \varepsilon'')a(\varepsilon', \varepsilon'')}{a^H(\varepsilon', \varepsilon'')U_M U_M{}^H a(\varepsilon', \varepsilon'')} \tag{18}$$

The values of ε' and ε'' that make P_{MUSIC} reach a peak that are chosen from the result of the estimation.

4. IMPLEMENTATION MODELING

In order to make the modeling determining the reflection coefficients (S_{11}) for the free-space reflection method presented in section 2. In this part, we have implemented modeling by CST software to determine parameter S_{11} as shown in Figure 2.

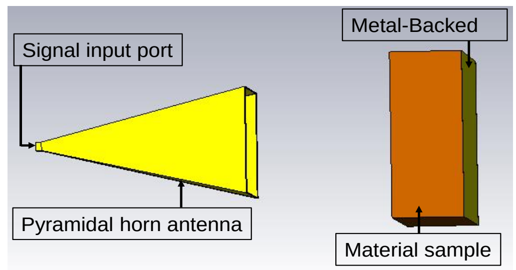

Figure 2. Modeling determining the parameters (S_{11}) of material sample using CST

In Figure 2, one pyramidal horn antena is designed to operate well in the frequency range of 8.0 - 12.0 GHz [16]. The gain and voltage standing wave ratio of the pyramidal horn antena are 20 dBi and 1.15 at the center frequency. In this model, the distance between the port of the antena and the material sample is 1082.5 mm. Losses due to the spacing of the free-space are removed through calibration by calculating for an air material sample with the same condition.

The selected material sample is a Teflon-PTFE nonmagnetic material. The Teflon-PTFE is widely used in communication devices, electronic devices, aerospace, and military equipment. In these devices and equipment, this material plays a vital role in many components, such as power divider, combiner, power amplifier, line amplifier, base station, RF antena, etc. The sample has parameters as follows: the width and length of the sample are similar in size of 150 mm, the complex relative permittivity of the sample at 10.0 GHz is $\varepsilon^* = 2.1 - j0.0002$.

5. RESULTS

The Teflon-PTFE samples are set up to measureat 801 different frequencies from 8.0 to 12.0 GHz with the scale of 5 MHz. From Figure 3 to Figure 7 show the pseudo-spectrum for the permittivity of Teflon-PTFE samples with the thickness of 10 mm, 30 mm, 50 mm, 70 mm, and 90 mm, respectively.

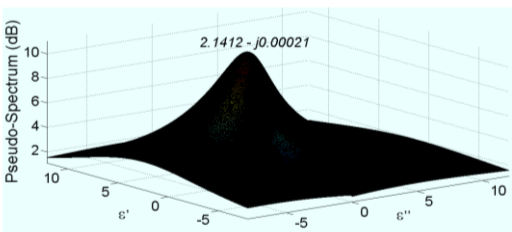

Figure 3. Pseudo-spectrum of Teflon-PTFE sample at 801 frequencies, frequency range of 4.0 GHz and thickness of 10 mm

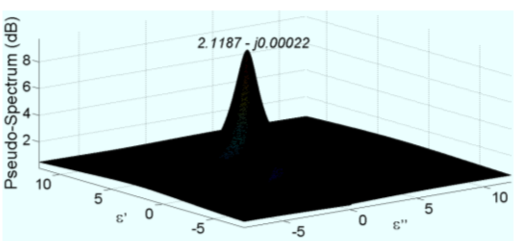

Figure 4. Pseudo-spectrum of Teflon-PTFE sample at 801 frequencies, frequency range of 4.0 GHz and thickness of 30 mm

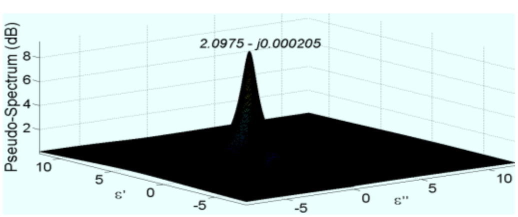

Figure 5. Pseudo-spectrum of Teflon-PTFE sample at 801 frequencies, frequency range of 4.0 GHz and thickness of 50 mm

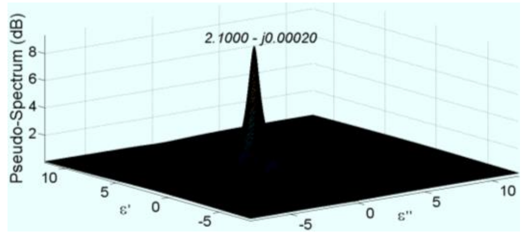

Figure 6. Pseudo-spectrum of Teflon-PTFE sample at 801 frequencies, frequency range of 4.0 GHz and thickness of 70 mm

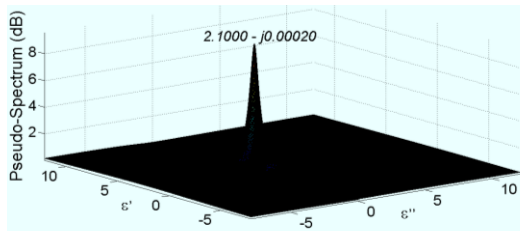

Figure 7. Pseudo-spectrum of Teflon-PTFE sample at 801 frequencies, frequency range of 4.0 GHz and thickness of 90 mm

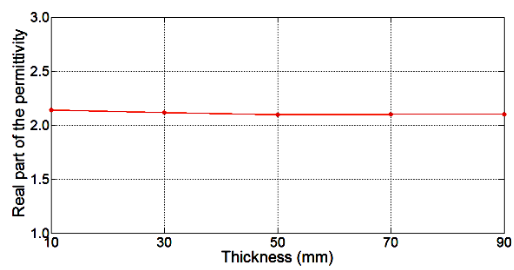

Figure 8. The real part of the complex relative permittivity of Teflon-PTFE sample is estimated by the MUSIS algorithm

It can be seen from Figures 3, 4, 5, 6, and 7 that the change in the thickness of the sample affects both the sharpness and the position of the peak of pseudo-spectrum. The change in the position of the peak from the expected value means that the estimation is not accurate for samples with small thickness. Furthermore, the drop in the sharpness of the peak makes the determination of the point of the greatest spectrum more difficult because the points around the peak can become equal to or greater than the supposed peak. These changes combined to create a sharp decrease in the accuracy of the results as the thickness of the samples decreases.

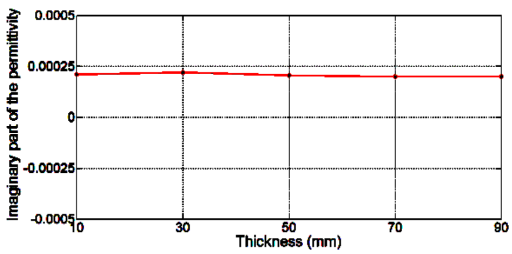

Figure 9. The imaginary part of the complex relative permittivity of Teflon-PTFE sample is estimated by the MUSIS algorithm

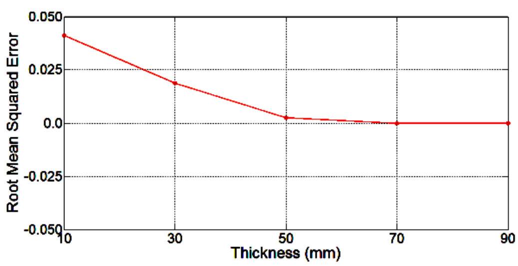

Figure 10. Root mean squared error versus thickness graph for ε'

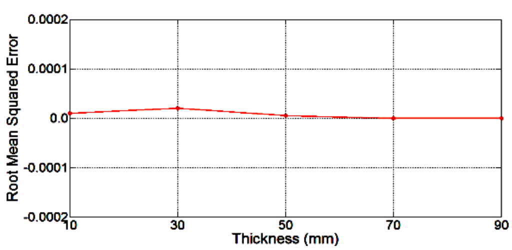

Figure 11. Root mean squared error versus thickness graph for ε''

The estimation results show that the complex relative permittivity of the Teflon-PTFE is accurate when the thickness changes as Figures 8 and 9. Figures 10 and 11 show the root mean squared error (RMSE) versus the thickness graph calculated from the simulation results for the samples at different thicknesses. From the results, the algorithm can solve for the unique value of the permittivity regardless of the thickness *d* but is more accurate with samples of larger thickness. The thickness of the sample affects the accuracy of the measurement of both ε' and ε''. To get the accurate value of the complex relative permittivity, the thickness *d* needs to be approximately 50mm or higher.

6. CONCLUSION

To propose a super high-resolution algorithm to accurately estimate the complex relative permittivity of the planar material samples using the reflection method in free-space. The system consists of a pyramidal horn antena and a metal-backed Teflon-PTFE placed in a free-space. The parameter vectors of the improved MUSIC algorithm describe the difference in phase, which indicates the difference in frequencies and arrival time of the simulated signals. These parameter vectors are calculated by using the relation between the permittivity and the refractive index. The performance of the proposed algorithm is verified for all scenarios in the simulation. Through the results, the ability of the proposed algorithm to solve the problem of ambiguity of the conventional method is also validated. The estimation results of the complex relative permittivity using proposed algorithm are accurate when the thickness of the sample is at least 50 mm. The proposed algorithm has great benefits in determining the characteristic parameters of new materials.

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Biography:

Ho Manh Cuong was born in Ha Noi, Vietnam, in 1977. He received the Bachelor degree in Radio Physics and Electronics at VNU University of Science in 1999 and the Master degree in Electronic Engineering at Le Quy Don University in 2006. In 2019 he received a Ph.D. degree in Electronics Engineering at Le Quy Don University. Now, he is a lecturer in Electric Power University, Vietnam. He has published many national as well as international papers. His current research interests are microwave engineering, antena, electromagnetic theory.

Le Trong Hieu was born in Hanoi, Vietnam in 1986. He graduated at Le Quy Don Technical University in Electronics and Telecommunications, in June 2009. He received the M.Sc. and Ph.D. degrees in Electromagnetic Field and Microwave Technology from the State Key Laboratory of Millimeter Waves, School of Information Science and Engineering, Southeast University, Nanjing, China, in 2013 and 2018, respectively. Now, he is a lecturer in the Faculty of Electronics and Telecommunications, Electric Power University, Hanoi, Vietnam. His fields of research are RF/Microwave and Millimeter-waves circuits such as filters, amplifiers, antenas for wireless communication applications.

(ISSN: 1859 – 4557)