Compensation of high frequency effects in dielectric spectroscopy with the coaxial transmission line structure fixture

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Abstract. The application of complex impedance detection shows a trend of multi-direction development. The impedance detection technology at home and abroad has expanded from the traditional fields of physics, electricity and chemistry or other fields. Therefore, under the condition of satisfying precision, the design of equipment needs miniaturization and simplicity more and more. Aiming at this trend, a fixture with coaxial structure is calculated and processed based on theoretical knowledge, and its availability is verified by actual measurement and HFSS simulation. On this basis, the physical phenomena of multiple reflection and electrical delay that cause the measurement error are analyzed, and the analytical compensation algorithm is given.

1. Overall scheme
Complex impedance detection technology was initially applied to measure the impedance value of components in physical research. With the increasing demand of all aspects of society, this technology has been widely used in other fields. Nowadays, diversified complex impedance detection technology is gradually applied in different fields such as electrical, electronic, chemical, biological and medical, etc. People not only pursue high precision, but also have various requirements on the object, method and process of detection technology\cite{1-2}.

Due to the increasing demand, more and more scientific research institutions and researchers join in the development of measuring instruments. This paper mainly studies the influence of the physical phenomena produced in the measurement of high frequency dielectric spectrum when using the measurement fixture based on the theory of coaxial transmission line structure, and puts forward the method that can improve the measurement results according to the theoretical analysis and formula derivation, and gives the corresponding theoretical derivation\cite{3-4}.

2. Design and processing of fixture
The fixture processing. Under the condition of low frequency, the size of the tested sample can be larger, and the lumped parameters can be calculated simply to obtain each parameter. As the frequency
increases, the wavelength will become shorter and shorter. In this case, the sample size needs to be smaller to ensure the reliability of measurement. To measure dielectric value by means of plate capacitance, a transmission line is required to access the signal. The frequency of APC-7 can reach 18 GHz, \( \phi = 7 \) mm in diameter, cross section is opposite bigger, meet the requirements of our design of fixture structure. Therefore, we processed the APC-7 coaxial cable structure fixture. In the actual circuit, it is equivalent to a transmission line, as shown in FIG. 1 (a). The network analyzer characteristic impedance circuit for 50 \( \Omega \) and the test fixture design also need to consider the impedance matching requirements. At the same time, because the structure of APC-7 coaxial cable, so the fixture of the diameter of the inner conductor \( \phi = 3.04 \) mm, outer conductor diameter \( \phi = 7 \) mm. As shown in figure 2 (b). Coaxial transmission line characteristic impedance calculation formula is:

\[
Z_c = \frac{60}{\sqrt{\varepsilon_r}} \ln \frac{b}{a}.
\]

In the formula \( \varepsilon_r \) represents the relative dielectric constant of the filled medium between the inner and outer conductors of the coaxial line. When the medium is air \( \varepsilon_r = 1 \). As shown in figure 1 (c), \( a \) represents the diameter (mm) of the outer conductor of the coaxial line; \( b \) represents the diameter (mm) of the inner conductor of the coaxial line. Using the Eq.1 calculate the test fixture when no-load characteristic impedance is about 50.04 \( \Omega \).

The coaxial adapter of APC-7 is selected to be connected with both ends of the test device. Since the inner conductor of the APC-7 connector has internal thread, the external thread needs to be processed for the fixture to facilitate the connection between the two. As shown in figure 1 to the actual machining of finished products, two is the application of the two same frequency connector, can fit into a large-area (\( \phi = 7 \) mm) of the tested samples, form the structure of capacitance, measured by capacitance method[5-6].

![Fig. 1 Design and processing of fixture physical map](image)

working principle. Network analyzer with the characteristic impedance of 50 \( \Omega \) cable circuits, the design of the transmission line also need to consider 50 \( \Omega \) impedance matching requirements. The transmission line adopts all-metal structure. According to Eq. (1), it can be seen that when heating, the characteristics of metal expansion and contraction should be considered. Stainless steel has the property of metal expansion and contraction, and the temperature increases by 1\( ^\circ \)C per meter to enlarge 0.016mm, but the change is very small9. In addition, according to the basic theory of transmission line, the characteristic impedance formula of transmission line is

\[
Z_c = \frac{R + j\omega L}{\sqrt{G + j\omega C}},
\]

It is usually a plural and is related to the operating frequency. The characteristic impedance is determined by the distribution parameters of the transmission line itself, independent of the load and the signal source. For a uniform lossless transmission line, \( R = 0, G = 0, \ Z_c = \frac{\omega L}{\sqrt{\omega C}} \). It is important to note that the high frequency, the skin effect increase the effective resistance of conductor \( R \). The higher the frequency, skin effect is more significant, \( R \) can be ignored in low frequency approximation for nondestructive line, when the conductor by high frequency alternating current, the current distribution on the conductor section will appear uneven status, current flows only in the conductor surface, the
surface current density, the largest conductor deep current density is small, the alternating current table phenomenon, called the current skin effect. Therefore, at high frequencies, an approximately lossless transmission line may become lossy transmission line, due to this phenomenon. Based on the transmission line theory, the length $l$, characteristic impedance $Z_c$ and propagation constant $\gamma$ of the fixture can be obtained after measuring and calculating each parameter of the fixture. According to the $Z_m = \frac{Z_r+Z_c \text{th}(\gamma l)}{Z_r+Z_c \text{th}(\gamma l)}$ after measuring the overall impedance $Z_m$ of the fixture after placing the sample accurately, the impedance $Z_r$ of the sample itself can be deduced. Therefore, to verify the reliability of fixture, it is necessary to measure at different temperatures and different frequencies to verify whether $Z_c$ is unchanged.

Feasibility analysis. Firstly, it analyzes the open circuit and short circuit status of the fixture in the condition of high frequency. This part adopts the actual measurement system for measurement, and HFSS simulation software simulates the frequency response of open circuit, short circuit and matching in the condition of high frequency according to the actual size of the fixture. The physical connection diagram of coaxial transmission line in the actual built measurement system is shown in figure 2.

Fig. 2 A physical connection diagram of a coaxial transmission line in measurement

Through actual measurement, the short circuit and open circuit of transmission line are measured respectively. In the case of short circuit, the left end of Smith's impedance circle diagram is clockwise along the outer ring of the circle, which is equivalent to pure inductance. In the case of open circuit, the right end of Smith's impedance circle diagram is clockwise along the outer ring of the circle, which is equivalent to pure capacitance.

Secondly, HFSS graph parameters were set according to the actual size of the fixture for modeling and simulation. Its simulation model is shown in figure 4 below.
Through parameter setting of the model, we get the frequency response of S parameter. The result of S parameter is shown in the Smith circle diagram as the short-circuit simulation result, as shown in figure 5, and the open-circuit simulation result, as shown in figure 6:
Compared with the above actual measurement results and simulation results, we can see that the fixture has usability.

3. Error Analysis and Improvement

In the high frequency condition, due to the direct propagation of waves in different media, reflection phenomenon will be generated, which will greatly affect our final measurement results3-6. In addition, due to the different thickness of the measured sample put in the fixture, the wave will produce a certain phase change when propagating in the object, which is called electrical delay phenomenon. This phenomenon will also have a great impact on the measurement results.

In order to solve the problem of measuring impedance compensation of the above coaxial fixture, we first need to combine the physical phenomenon of microwave reflection on the cross section between two different characteristic impedances and the electrical delay generated when microwave signal propagates in the sample.

Multiple reflection. Test under normal temperature condition, as the chip structure of sample material part is considered to be a medium of transmission line under test frequency range is limited in several hundred MHZ, especially in the case of a higher dielectric constant measurement of material, in this case the measured results will have certain deviation with real data, the reasons for the deviation because through this party when testing formula for an ideal situation, and it does not take into account a wave travels on the sample section, in other words without considering happens many times, and not to be added to the measurement results.

A dielectric sample with thickness \(d\) and diameter \(a\) and \(\varepsilon^* = \varepsilon - j\varepsilon_r\) a relative complex dielectric constant of are placed at the end of the central conductor of the coaxial waveguide. At this time, the position of the sample is on the same central axis as the central conductor. The sample is completely wrapped by the outer tube of the fixture, forming a circular parallel plate capacitor. The experimental model after the fixture fills the tested sample is shown in figure 7. This structure is the same as the traditional lumped capacitor model.

![Fig. 7 Experimental model after the clamp is filled with the tested sample](image)

The structure can be equivalent to a circuit network, where the dielectric - filled capacitor part can be expressed as a function of transmission line \(\varepsilon^*(\omega)\). Here we assume that the type of wave propagated is TEM mode formula wave. There are no electric and magnetic fields in the direction of propagation, which is called transverse electromagnetic wave.
FIG. 8 Propagation diagram of multiple reflections in a thin parallel-plate capacitor terminating coaxial waveguide.

Therefore, considering multiple reflections is an indispensable factor in this study. Considering multiple reflections, the scattering coefficient $S_{11}$ of the capacitor edge of the parallel plate can be expressed as Eq.1.

$$S_{11} = \rho + (1 - \rho^2)e^{-\gamma a}\sum_{n=1}^{\infty}(-\rho e^{-\gamma a})^{n-1} = \frac{\rho + e^{-\gamma a}}{1 + \rho e^{-\gamma a}}$$  \hspace{1cm} (1)

In Eq.1, $\rho = \left(1 - \sqrt{\varepsilon^*}\right)/\left(1 + \sqrt{\varepsilon^*}\right)$, $\gamma$ is the propagation constant (constant value), $\omega$ is the angular frequency, and $C$ is the velocity of light in air ($3*10^8$m/s). On the other hand, the input admittance of the sample part is $Y_{in}$ with characteristic conductance $G_s$, which can be expressed as the function of $S_{11}$. The formula of its representation is shown in Eq.2:

$$Y_{in} = G_s\frac{1-S_{11}}{1+S_{11}}$$  \hspace{1cm} (2)

Combining Eq.1 and Eq.2, the following equation can be simplified:

$$Y_{in} = G_s\frac{(1-\rho)(1-e^{-\gamma a})}{(1+\rho)(1+e^{-\gamma a})} = G_s\sqrt{\varepsilon^*}\tanh\left(\frac{\gamma a}{2}\right)$$  \hspace{1cm} (3)

If $x = \frac{\omega a\sqrt{\varepsilon^*}}{2c}$ and $h(jx) = \frac{1}{\cot(x)}$, Eq.3 can be simplified to the form of Eq.4 as follows:

$$Y_{in} = G_s\frac{j\omega a}{2c}\varepsilon^* \frac{1}{\cot(x)}$$  \hspace{1cm} (4)

Eq.4 can be used to describe the input admittance of a partial network, including the complex dielectric constant of the parallel plate capacitor $\varepsilon^*$ of the network, which is $C^*$ to represent the part of the complex capacitor. Eq.4 can be expressed as Eq.5:

$$Y_{in} = j\omega C^* \frac{1}{\cot(x)}$$  \hspace{1cm} (5)

According to Eq.4 and Eq.5, the dielectric sample can be expressed as a transmission line whose electric length (electric length: the ratio of the physical length of the micro-strip transmission line to the wavelength of the transmitted electromagnetic wave is electric length) is 1/2, and $\frac{1}{\cot(x)}$ represents the multiple reflection phenomenon of the sample part. If the electrical length is less than the wavelength, the value of $\frac{1}{\cot(x)}$ approaches unity. At this point, Eq.5 can be simplified to the formula of the input admittance of the conventional lumped capacitance model, as shown in Eq.6 below:

$$Y_{in} = j\omega C^*$$  \hspace{1cm} (6)

In the actual measurement process, the end of the coaxial waveguide fixture is selected as the reference plane, and the scattering coefficient is represented by $S_{11}$. The network parameters above the terminal termination interface (reference surface) of the sample are shown in figure 2. It should be
noted that $S_{11}^m$ here is different from the parameter $S_{11}$ used in Eq.1, and specific values can be given by measurement. By imitating Eq.2, we can obtain the new formula of input admittance, as shown in Eq.7 below:

$$Y_{in} = G_1 \frac{1-S_{11}^m}{1+S_{11}^m}$$

(7)

Where $G_1$ represents the characteristic conductance of coaxial waveguide. By combining Eq.5 and Eq.7, we can obtain the quantifiable complex permittivity $\varepsilon^*$, which can be expressed in Eq.8 as follows:

$$\varepsilon^* = \frac{2\varepsilon}{\text{j} \mu \omega G_1} \frac{1-S_{11}^m}{1+S_{11}^m} \text{xcotx}$$

(8)

The parameter $g$ in Eq.8 is $g = \frac{G_1}{G_2}$, which represents the battery constant (the ratio between the distance between the two electrodes in the battery and the cross-sectional area of the two electrodes). Usually, a solution which has been accurately measured in the conductivity is used to measure the conductivity in the battery, and then the battery constant is obtained. The value can be obtained by measuring the material with known complex dielectric constant. Since the shape of the material selected in each test is different, it is more practical to use the complex capacitance as a parameter than the battery constant[7-8]. Since the multiple reflections of the wave at the sample are considered, the capacitance of the sample part is given by the combination of parallel plate capacitors, which is divided into two parts: one is a parallel plate capacitor formed by filling the dielectric sample, and the other is a capacitor of the edge field after considering the multiple reflections of the edge[9]. In this case, the complex capacitance can be expressed as the sum of the two parts, as shown in Eq.9 below:

$$\varepsilon^* = \frac{C_{p} \varepsilon^* + C_{f}}{\text{j} \omega \varepsilon \varepsilon_0 + 2\varepsilon_0 \ln \left(\frac{b-a}{2d}\right)}$$

(9)

Where, $C_{p}$ is the geometric capacitance of the empty parallel-plate, $C_{f}$ is the edge capacitance, and $\varepsilon_0$ is the complex dielectric constant of the air. It should be noted that Eq.9 is only applicable to the case of $d \ll b-a$ and $\lambda \gg 2\pi d$. The study in this paper is aimed at thin lamella with a diameter of less than 7mm, so Eq.9 is applicable to the study in this paper. Thus, we finally get the expression of the complex dielectric constant of the sample, as shown in Eq.10 below:

$$\varepsilon^* = \frac{G_1 \frac{1-S_{11}^m}{1+S_{11}^m} \text{xcotx} - C_{f}}{C_{p}}$$

(10)

The $\text{xcotx}$ part in Eq.10 represents multiple reflections of the sample part. As we mentioned before, at a low frequency, $x$ in $\text{xcotx}$ tends to 0, so $\text{xcotx}$ tends to 1. Therefore, Eq.10 can be simplified as follows:

$$\varepsilon^* = \frac{G_1 \frac{1-S_{11}^m}{1+S_{11}^m} - C_{f}}{C_{p}}$$

(11)

It corresponds to the traditional model, so multiple launches are not required. The complex dielectric constant is used as a function of the frequency of the calculation and is used in Eq.1 to include an undesired influence factor equivalent to $\frac{1}{\text{xcotx}}$, which has a given discrete point

$$x = \left(n - \frac{1}{2}\right)\pi \quad (n = 1, 2, 3, ... ... \)$$

(12)

From the relation $x = \frac{\omega a \sqrt{\varepsilon_0}}{2\varepsilon}$, the initial frequency is expressed as follows when $n = 1$:

$$f_r = \frac{c}{2a \sqrt{\varepsilon_0}}$$

(13)

Thus, the resonant dispersion of the lumped capacitance model is found to be $1/\text{xcotx}$. Thus the real part experiences a numerical oscillation of 0 at the resonant frequency, and the imaginary part experiences a peak at that frequency. The expression of Eq.10 can eliminate the resonance dispersion
and more accurate the complex dielectric constant of the sample. Electrical delay. Based on the above theoretical analysis, in order to get more accurate measurement results, we need to consider the fixture configuration. As shown in figure 8, the dielectric sample is placed on the reference plane at the end of the coaxial waveguide during the actual measurement. Therefore, due to the different thickness of the sample in the transmission process of TEM mode microwave samples, the phase change of the microwave in the transmission process of the sample results in the electric delay phenomenon. We need to modify the parameters $t_{ed}$ in Eq.10, which represents the electric delay part, so as to better estimate the complex dielectric constant. By introducing electrical delay parameters, the following equation can be obtained:

$$e^{j\omega t} \rightarrow e^{j\omega (t+t_{ed})}$$

(14)

Considering the influence factors of electrical delay, $S_{11}$ in the above formula (1) can be redefined as the form given by Eq.15:

$$S'_{11} = \rho e^{j\omega t_{ed}} + (1 - \rho^2)e^{-j\omega a} \sum_{n=1}^{\infty} (-\rho e^{-j\omega a})^{n-1} e^{j\omega t_{ed}}$$

(15)

According to Eq.15, we can deform the formula to get the expression of the complex dielectric constant $\varepsilon^*$, and finally get the related parameters of the measured sample[10].

4. Conclusion
The application of complex impedance detection is developing in many directions. Impedance detection technology has been extended from the traditional fields of physics, electricity and chemistry to other fields. Examples include bioelectrical impedance spectroscopy for biomedical tissue detection, electrochemical impedance spectroscopy for chemical reactions, and biochip laboratories for reagent, contamination, or disease detection. Therefore, under the condition of satisfying precision, the design of equipment needs miniaturization and simplicity more and more.

In this paper, the physical phenomenon of the measuring fixture based on the coaxial transmission line structure is analyzed under the measurement of high frequency dielectric spectrum. Firstly, the relevant technical parameters of coaxial fixture are obtained through theoretical calculation. Secondly, through the actual connection of vector network analyzer and the use of HFSS simulation software, according to the actual size of the fixture, the simulation is carried out at high frequency to obtain the frequency response when the fixture is open, short and matched. The results of both operations indicate the usability of the fixture. On this basis, this paper analyzes the phenomenon of multiple reflection and electrical delay that may occur when the fixture is used for measurement at high frequency, and gives the corresponding improved compensation algorithm based on theoretical knowledge.

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