

Description

Method for determining the dielectric constant and the loss factor of the individual insulant levels of multilayer printed-circuit boards.

5 The invention relates to a method for determining the dielectric constant and/or the loss factor of insulator layers of printed-circuit boards by using test structures fitted onto or into the printed-circuit board at measuring frequencies in the MHz to GHz range, as well
10 as to a printed-circuit board having at least one test structure for determining the dielectric constant and/or the loss factor of insulator layers of printed-circuit boards at measuring frequencies in the MHz to GHz range.

 For the electrical properties of printed circuits
15 on printed-circuit boards, further to the physical thickness of the insulant layers and the physical width of the conductive tracks, in the case of a radiofrequency circuits, in particular the relative permittivity and the loss factor of the insulants are of great importance.
20 This being the case, multilayer printed-circuit boards generally have a particular pairing of the values of the dielectric constant and of the loss factor for each insulant level, depending on the materials used and on the production process.

25 It is known to determine the dielectric constant, for example, by means of capacitance measurements with capacitor arrangements which use the insulant of the printed-circuit board as dielectric. In order to make it possible to take measurements even on the printed-circuit
30 board, it is known to produce narrow-band filter arrangements on the printed-circuit board, in which it is possible to determine the dielectric constant as well as the loss factor of the insulant by means of a transmission measurement, through the transmission characteristics.
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 The object of the invention is to provide a method and a test structure with which the dielectric



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constant and/or the loss factor can be determined on prefabricated or finished printed-circuit boards with the aid of standard measuring instruments and using solder-free connections. The intention is for this to be possible, even on a commercial scale, in the shortest possible time for mass-produced articles which already have a printed circuit.

On the basis of a method of the type mentioned at the start, this object can be achieved in that, as the test structure, at least one test resonator is fitted into or onto the printed-circuit board, this resonator being designed as a planar microwave line with predetermined dimensions and predetermined position of at least one measuring point, in that, for each test resonator, its resonant frequency is determined in a frequency band in which this resonant frequency is to be expected, and in that the impedance at the measuring point is then determined at this resonant frequency, and lastly the dielectric constant and/or the loss factor are determined by computer from the predetermined dimensions of each test resonator and from the determined values of the resonant frequency and of the real part of the impedance by using the line equations.

The object can also be achieved with a printed-circuit board of the type mentioned above, in which according to the invention at least one test structure is a test resonator, designed as a planar microwave line and having predetermined dimensions and predetermined position of at least one measuring point, fitted into or onto the printed-circuit board.

The invention permits quick and simple determination of the dielectric constant and the loss factor, even on already finished printed-circuit boards, with the aid of standard measuring instruments, in particular network analysers, in the MHz to GHz range, and has been proven, for example, in measurements on printed-circuit boards for mobile radio at frequencies of, for example, 900 MHz.

Further features of the invention are characterized in the dependent Claims 2 to 8 and 10 to 17.



The invention, together with other advantages, is explained in more detail below with reference to embodiments as examples, these being illustrated in the drawing. In the drawing, Fig. 1 shows a plan view of a section of a printed-circuit board according to the invention having a planar microwave line, Fig. 2 shows a section on the line II-II in Fig. 1, Fig. 3 shows a section on the line III-III in Fig. 1, Fig. 4 shows a schematic cross-section through a triplate line having two dielectrics, Fig. 5 shows a diagram of the field-line pattern associated with a triplate line and a representation of the capacitances per unit length, Fig. 6 shows the equivalent circuit diagram of a test resonator used in the invention, and Fig. 7 shows a schematic representation of a printed-circuit board according to the invention, having three insulant layers and three test resonators.

According to Figs 1 to 3, a printed-circuit board 1 consists of an insulant 2, for example made of a glass-fibre epoxy resin support material, in which a conductor 3, designed as a metal strip, is embedded. Facing the conductor 3, there are earth layers 4 on the two surfaces of the printed-circuit board 1. A planar microwave line of this type is referred to as a "triplate" line and is represented later in further detail with its dimensions in Fig. 4.

On one surface, the upper surface in Fig. 3, a coaxial measuring point 5 is designed which consists of a circular recess 6 in the upper earth layer 4 and of a contact pin 7 which extends as a through-contact from the conductor 3 to the plane of the earth layer 4. At one end, the left-hand end in Fig. 3, through-contacts 8 are provided which extend from the upper earth layer 4 to the lower earth layer 5 and also touch the conductor 3. At the other end, the right-hand end in Fig. 3, through-contacts of this type are not provided but instead, as indicated at one position in Fig. 1, lateral through-contacts 9 can be provided which screen the conductor 3 on both sides in the manner of a cage. This is expedient

particularly when the breadth of the earth layers 4 cannot be chosen to be very large.

The stripline according to Figs 1 to 3 therefore has one short-circuited and one open end, the length
5 between the short-circuited end and the measuring point being denoted l_x and the length between the measuring point and the open end being denoted l_0 . In general, the line is designed as an $1/4$ resonator, although the invention is not restricted thereto, and there could, for
10 example, also be a $3\ 1/4$ or an $1/2$ resonator.

A concise theoretical explanation of the measuring method according to the invention will firstly be given below with reference to Figs 4, 5 and 6, the literature listed below also being relevant:

15 n Integrierte Mikrowellenschaltungen [Integrated microwave circuits], Reinmut K. Hoffmann, Springer Verlag 1983
n Wheeler, H.A., Formulas for the Skin Effect, proc. IRE, Vol. 39, Sept. 1942, pp 412-424
n Meinke-Gundlach, Taschenbuch der Hochfrequenztechnik
20 [Guide to radiofrequency technology], Springer Verlag 1992

The input impedance Z_1 of a line which has length l and is terminated by an impedance Z_2 is equal to:

$$Z_1 = Z_0 \frac{Z_2 \cdot \cosh(\gamma l) + Z_0 \sinh(\gamma l)}{Z_0 \cosh(\gamma l) + Z_2 \sinh(\gamma l)}$$

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(1)

25 with the line impedance Z_0 and the propagation constant γ . For the sake of simplicity, only "wide" offset strip-line arrangements will be considered, for which $W / (D_u - T) \geq 0.35$ and $W / (D_o - T) \geq 0.35$. W is the physical line width, D_u , D_o are the substrate thicknesses
30 and T is the conductor thickness (Fig. 4).

For the line impedance Z_0 of a wide stripline of this type:

$$Z_1 \cong \sqrt{\frac{\epsilon_0 \mu_0}{\epsilon_{r,eff}}} \frac{1}{C_{p0} + C_{p0} + 2C_{s0} + 2C_{s0}}, \quad (2)$$

with the relevant capacitances per unit length

$$C_{p0} = \frac{\epsilon_0 W}{D_s} \quad \text{and} \quad C_{s0} = \frac{\epsilon_0 W}{D_s}, \quad (3)$$

$$C_{s0} \cong \frac{\epsilon_0}{\pi} \left[\left(2 + \frac{T}{D_s} \right) \ln \left(2 + \frac{T}{2D_s} \right) - \frac{T}{2D_s} \ln \left(\frac{T}{D_s} + \left(\frac{T}{2D_s} \right)^2 \right) \right], \quad (4)$$

$$C_{s0} \cong \frac{\epsilon_0}{\pi} \left[\left(2 + \frac{T}{D_s} \right) \ln \left(2 + \frac{T}{2D_s} \right) - \frac{T}{2D_s} \ln \left(\frac{T}{D_s} + \left(\frac{T}{2D_s} \right)^2 \right) \right].$$

5 Fig. 5 shows the meaning of the capacitances per unit length modelled thereby.

The effective dielectric constant $\epsilon_{r, eff}$ represents an equivalent to the dielectric constants ϵ_{ru} , ϵ_{ro} of the different insulants present in the configuration:

$$10 \quad \epsilon_{r,eff} = \frac{\epsilon_{ru}(C_{p0} + 2C_{s0}) + \epsilon_{ro}(C_{p0} + 2C_{s0})}{C_{p0} + C_{p0} + 2C_{s0} + 2C_{s0}} \quad (5)$$

It is only meaningful to specify an effective dielectric constant if the electromagnetic field structure in the existing configuration approximately corresponds to the field of the uniform equivalent structure.
 15 This is true so long as the dielectric constants ϵ_{ru} , ϵ_{ro} differ from one another only slightly, as is the case in conventional multilayer printed-circuit boards.

The complex propagation constant γ can be written as

$$20 \quad \gamma = a + jb,$$

the attenuation constant a is in turn composed of a component a_r for the resistive losses and a component a_d for the dielectric losses:

$$a = a_r + a_d.$$

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In lines which are filled with a substantially uniform insulant, losses in the dielectric can be described by:

$$\alpha_s \cong \pi \sqrt{\epsilon_{r,eff}} \frac{f \tan \delta_{eff}}{c_0}, \quad (8)$$

- 5 with the frequency f and the speed of light in vacuo c_0 . The effective loss factor $\tan \delta_{eff}$ for wide offset strip-lines is given, by considering the distribution of the electric field energy, from the loss factors $\tan \delta_u$, $\tan \delta_0$ of the dielectrics involved

$$\tan \delta_{eff} = \frac{\tan \delta_u \epsilon_u (C_{pu0} + 2C_{pu})}{\epsilon_u (C_{pu0} + 2C_{pu}) + \epsilon_n (C_{pu0} + 2C_{pu})} + \frac{\tan \delta_n \epsilon_n (C_{pu0} + 2C_{pu})}{\epsilon_n (C_{pu0} + 2C_{pu}) + \epsilon_u (C_{pu0} + 2C_{pu})}. \quad (9)$$

10

If the skin depth of the current in the material of the conductor is substantially less than the thickness of the conductor, then a good approximation of the resistive losses per unit length is:

$$\alpha_p = \frac{\sqrt{\pi f \rho \epsilon_0 \epsilon_{r,eff}}}{Z_1} \frac{\delta Z_1}{\delta n} \quad \text{with} \quad \frac{\delta Z_1}{\delta n} = \frac{\partial Z_1}{\partial D_s} + \frac{\partial Z_1}{\partial D_o} - \frac{\partial Z_1}{\partial W} - \frac{\partial Z_1}{\partial T} \quad (10)$$

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with the conductivity ρ of the conductor material. The phase constant b is

$$\beta = 2\pi f \sqrt{\epsilon_{r,eff} \epsilon_0 \mu_0} \quad (11)$$

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- 20 Substituting Eqns (2)...(11) in Eq.(1) gives the impedance Z_1 occurring for arbitrary line terminations Z_2 . The two important special cases of open-circuit and short-circuit at the end of the line should be considered separately: In the case of a short-circuited line of length l_k , the input impedance is calculated according to

Eq. (1) as

$$Z_k = Z_1 \tanh(\gamma l_k). \quad (12)$$

The impedance of an open line of effective length l_0 is

$$5 \quad Z_0 = Z_1 \tanh^{-1}(\gamma l_0). \quad (13)$$

In this case, it should be borne in mind that the effective length l_0 of the open line is not identical to the physical length of the open line. The cause resides in the additional capacitance C_{ende} of the open line end, which can be approximately modelled by a corresponding extension l_{ende} of the open line. If a wide line is assumed, than

$$C_{\text{ende}} \cong W(\epsilon_n C_{\text{po}} + \epsilon_n C_{\text{so}}) \quad (14)$$

and therefore

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$$15 \quad l_{\text{ende}} \cong \frac{C_{\text{ende}}}{\epsilon_n (C_{\text{po}} + 2C_{\text{so}}) + \epsilon_n (C_{\text{po}} + 2C_{\text{so}})} \quad (15)$$

In a preferred embodiment of the invention, each test resonator consists of an $l/4$ line having one short-circuited and one open end, which is represented in the equivalent circuit diagram according to Fig. 6.

20 At the measuring point of the modelled test resonator, the impedances Z_k , Z_0 form a parallel circuit:

$$Z_{\text{meas}} = Z_1 [\tanh^{-1}(\gamma l_k) + \tanh(\gamma l_0)]^{-1} \quad (16)$$

Close to $l/4$ resonance, the impedance at the measuring point reaches a maximum. The value of this impedance is determined by the line impedance, the losses and the position of the measuring point. For predetermined losses, the impedance measured in the case of resonance can be influenced by the selection of the ratio in length between the short-circuited line and the open line, since the dissipative impedances along the line change.

A network analyser is suitable for measuring the



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impedance, it being necessary to bear in mind that its greatest accuracy in impedance measurements can be achieved close to the reference impedance (e.g. $Z_0 = 50\Omega$). If the measured impedance is close to this reference impedance, then the measurement errors occurring as a result of contact resistances and parasitic effects (stray capacitances or inductances) are also minimized. In particular, the use of conventional coaxial test pins and therefore automated measurement with tolerable measurement errors is made possible. The aim of the dimensioning of the test resonators is therefore, for typical values of $e_{r,eff}$, $\tan\delta_{eff}$ to be measured, to bring the measured impedance at $1/4$ resonance close to the reference impedance $Z_0 = 50\Omega$.

The values $e_{r,eff}$, $\tan\delta_{eff}$ are measured in two steps:

- The frequency f_R at which the maximum return loss occurs is determined
- the real part $\text{Re}(Z_R)$ of the impedance at the measuring point at the frequency f_R is measured

With corresponding adjustment, a network analyser can automatically search for the frequency f_R and simultaneously determine the resistance $\text{Re}(Z_R)$.

The relationship between the measured impedance Z_{modell} and the values $e_{r,eff}$, $\tan\delta_{eff}$, f is known through Eq. (16). With an iterative method, it is then possible to find a pair of values $e_{r,eff}$, $\tan\delta_{eff}$ such that, at the frequency $f = f_R$, the calculated reflection factor reaches a minimum and, at the same time, $\text{Re}(Z_{modell}) = \text{Re}(Z_R)$ is satisfied for the calculated measured impedance. The numerical solution method is simplified by the fact that the parameters $e_{r,eff}$, $\text{Re}(Z_R)$ and $\tan\delta_{eff}$, f_R are only weakly coupled if the measured impedance Z_R is close to the reference impedance Z_0 .

In order to make it possible to determine the material properties of a number N of different insulant layers individually, it is necessary to measure at least N different test resonators. This being the case, the test resonators need to be chosen in such a way that all



the insulant layers to be measured are involved. Fig. 7 shows a group of three resonators for measuring a four-layer printed-circuit board ($N=3$) having three insulant layers 2a, 2b, 2c with three conductors 3a, b, c and with
5 corresponding earth layers 4a, b, c.

The effective physical constants of the stacked insulant layers are given from the individual physical constants and the layer thicknesses in similar fashion to a series circuit of capacitances. Suitable design of the
10 test structure gives rise to a system of equations which can be used to isolate the physical constants $\epsilon_{r(1..N)}$, $\tan\delta_{(1..N)}$ from the effective physical constants $\epsilon_{r,eff}$, $\tan\delta_{eff}$. The system of equations is generally non-linear, but can be solved numerically, for example by a random-
15 ized optimization method, especially if very accurate initial values for the physical constants are known. If more resonators than necessary are available and the system of equations is therefore overdefined, then the least squares method is a suitable solution strategy.

Even though the method according to the invention, and the associated printed-circuit board, are suitable for determining both the dielectric constant and the loss factor, in many cases it is sufficient to determine merely the dielectric constant $\epsilon_{r,eff}$. Because
25 of the weak coupling between the parameters $\epsilon_{r,eff}$ and $\text{Re}(Z_R)$, it is in this case sufficient to determine the resonant frequency f_R , and a simple scalar reflection-factor measuring instrument suffices for the measurement. Fundamentally, the weak coupling between $\text{Re}(Z_R)$ and $\epsilon_{r,eff}$
30 is based on the fact that the changes in impedance of a test arrangement which are due to the loss factor and to the dielectric constant are orthogonal to one another (this is clear from a Smith chart). The calculation of the material property ϵ_r takes place for the simplified
35 case in similar fashion to the full calculation, except that the measured value $\text{Re}(Z_R)$ is replaced by the estimate Z_0 .

The invention has been practically implemented on printed-circuit boards of mobile radio sets, the test



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resonators being arranged at the edge of the blank.

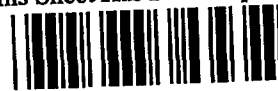
An embodiment of the test resonator, as represented in Figs 1 to 3, is particularly expedient in so far as selecting the position of the measuring point 5 makes it easy to match the impedance Z_R which occurs at resonance to a reference impedance Z_0 of the measuring instrument which is used. It is clear that the physical dimensions are determined by the desired measurement frequency and by the actual dielectric constant, and the position of the measuring point dictates the input impedance at the measuring point.

In arrangements which correspond to Fig. 7, that is to say when a plurality of different insulant layers are present, the measuring points of the individual resonators can be shifted by suitable through-connection to one or other surface of the printed-circuit board, or some to one surface and others to the other surface. However, it is practical to put all the measuring points on the same side of the printed-circuit board. In the context of the invention, the term "measuring point" should not be taken too literally, since it could, for example, be necessary to arrange the measuring point in a different place from the actual resonator and to make the connection to the measuring instrument via a lead. In such a case, the impedance of the simplified model, specified in equation (1), would need to be modified by the addition of a lead. It is also, for example, possible for additional capacitances or inductances to occur at the measuring point, which can then easily be accounted for in the model.

In order to make it possible to use a network analyser and the possibility of problem-free connection, for example with coaxial test pins, it is merely necessary for the measured impedance resulting at the measuring point or at the measuring instrument to be close to the reference impedance Z_0 of the measuring instrument. As already mentioned, a customary network analyser can be used for the measurement, which can automatically determine the frequency f_R of the maximum return loss and the

real part $\text{Re}Z_R$ of the impedance occurring at this frequency, for example with a single-port measurement. Since the impedance to be measured is advantageously close to the reference impedance of the network analyser, its measurement accuracy can be used to the full. With this choice of the measured impedance, even possible contact resistances between the test pins and measuring point do not introduce substantive errors into the determination of the desired values. Although a triplate line has especially been tried and tested for the purposes of the invention, according to practical requirements it is also possible to employ other planar microwave lines, for example screened or unscreened micro-striplines, micro-striplines with earthing slot, slotted lines, etc.

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Patent claims

1. Method for determining the dielectric constant and/or the loss factor of insulator layers of printed-circuit boards by using test structures fitted onto or
5 into the printed-circuit board at measuring frequencies in the MHz to GHz range, characterized in that, as the test structure, at least one test resonator is fitted into or onto the printed-circuit board, this resonator being designed as a planar microwave line with pre-
10 determined dimensions and predetermined position of at least one measuring point, in that, for each test resonator, its resonant frequency (f_R) is determined in a frequency band in which this resonant frequency is to be expected, and in that the impedance (Z_R) at the
15 measuring point is then determined at this resonant frequency, and lastly the dielectric constant and/or the loss factor are determined by computer from the predetermined dimensions of each test resonator and from the determined values of the resonant frequency and of the
20 real part of the impedance by using the line equations.
2. Method according to Claim 1, characterized in that the line resonator is an $1/4$ resonator.
3. Method according to Claim 1 or 2, characterized in that the line resonator has one short-circuited and
25 one open end with, lying between these ends, a measuring point which is at a distance (l_R) from the short-circuited end and at a distance (l_O) from the open end.
4. Method according to one of Claims 1 to 3, characterized in that the resonant frequency (f_R) is determined
30 as that frequency at which the return loss of the test resonator reaches its maximum or the reflection factor reaches its minimum.
5. Method according to one of Claims 1 to 4, characterized in that the real part $\text{Re}(Z_R)$ of the impedance
35 (Z_R) is determined.
6. Method according to one of Claims 1 to 5, characterized in that the test resonator is designed as a triplate line with predetermined distances (D_O , D_U) of





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the conductor from the earth layers.

7. Method according to Claim 6, characterized in that, for a printed-circuit board having N insulator layers, at least N test resonators are used.

5 8. Method according to one of Claims 1 to 7, characterized in that the physical dimensions of each test resonator are chosen in such a way that its impedance (Z_R) occurring at the measuring point on resonance lies in the region of the reference impedance (Z_0) of the
10 measuring instrument which is used, for example a network analyser.

9. Printed-circuit board (1) having at least one test structure for determining the dielectric constant and/or the loss factor of insulator layers of printed-
15 circuit boards at measuring frequencies in the MHz to GHz range, characterized in that at least one test structure is a test resonator (R; R1; R2, R3), designed as a planar microwave line and having predetermined dimensions and predetermined position of at least one measuring point
20 (5), fitted into or onto the printed-circuit board (1).

10. Printed-circuit board (1) according to Claim 9, characterized in that the microwave line has one short-circuited and one open end.

11. Printed-circuit board (1) according to Claim 10,
25 characterized in that a measuring point (5) is provided at predetermined distances (l_K , l_0) from the short-circuited end and from the open end.

12. Printed-circuit board (1) according to one of Claims 9 to 11, characterized in that the test resonator
30 (R) is an $1/4$ line.

13. Printed-circuit board (1) according to one of Claims 9 to 12, characterized in that the test resonator (R) is designed as a triplate line with predetermined distances (D_0 , D_u) of the conductor (3) from the earth
35 layers (4).

14. Printed-circuit board (1) according to one of Claims 9 to 13, characterized in that the physical dimensions of the test resonator (R) are chosen in such a way that its impedance (Z_R) occurring at the measuring



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point (5) on resonance lies in the region of the reference impedance (Z_0) of customary measuring instruments, for example network analysers, and is in particular about 50 ohms.

- 5 15. Printed-circuit board (1) having N insulator layers, according to one of Claims 9 to 14, characterized in that it has at least N test resonators (R1, R2, R3), at least one insulant layer (2a, 2b, 2c) being allocated as dielectric to each test resonator.
- 10 16. Printed-circuit board (1) according to one of Claims 9 to 15, characterized in that each test resonator (R) is arranged at the edge of the board blank.
- 15 17. Printed-circuit board (1) according to one of Claims 9 to 16, characterized in that the measuring points (5) are designed as coaxial measuring points for coaxial test pins.

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