DETERMINATION OF THE PERMITTIVITY OF SOME DIELECTRICS IN THE MICROWAVE AND MILLIMETRE WAVE REGION

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Abstract

In the first part of this study, determination of the dielectric properties of the low loss microwave ceramic material, barium nonatitanate ($\text{Ba}_2\text{Ti}_9\text{O}_{20}$), around 1 GHz is discussed. The structures under test were coaxial resonators, the cores of which were made of barium nonatitanate and the metallization was realised by thick film silver. The measured value of the real part of the relative dielectric constant was $\varepsilon_r=37$ and that of the loss tangent was $\tan\delta=0.00014$. The change of the resonance frequency of the coaxial resonators with temperature, in the range -20…+80 °C, was 4 ppm/°C. In addition, realisation of compact interdigital and comb-line bandpass filters was demonstrated for the 900 MHz mobile phone band. Besides $\text{Ba}_2\text{Ti}_9\text{O}_{20}$, $\text{Ba}({\text{Sm},\text{Nd}})_2\text{Ti}_5\text{O}_{14}$ ceramic material with a dielectric constant of $\varepsilon_r=78$ was also employed in order to improve the miniaturisation. The volume of the smallest filter was 2 cm$^3$ and the weight was 9 g.

In the second part of this study, various measurement methods have been demonstrated to determine the real part of the relative permittivity of FR4 fibreglass circuit board. Test structures were straight asymmetric open-circuited strip lines. It was found that the real part of the relative permittivity decreased slightly with frequency. At 0.72 GHz and 4.5 GHz, $\varepsilon_r$ values of 4.3 and 4.1, respectively, were measured. All the characterisation methods used gave consistent values for $\varepsilon_r$, and electromagnetic simulators were used to verify the results.

In the third part of this study, the structures under test were microstrip or strip line transmission lines, the $S$ parameters of which were measured by using on-wafer measurement techniques. It was found that the insertion loss of the 10 mm long etched Cu microstrip was slightly less than that of the Ag microstrips which were manufactured by etching and gravure offset printing techniques, respectively. The performance of the Ag microstrips was, however, similar. In addition, a theoretical basis was established for the determination of $\varepsilon_r$ of the dielectric substrate and the attenuation coefficient of the transmission lines under test. The calculations were based on signal flow diagrams. The method was applied to determine the dielectric and loss properties of a commercial, Kyocera, Japan, LTCC (Low Temperature Cofired Ceramic) material up to 50 GHz. The measured value of the real part of the relative permittivity was $\varepsilon_r=5.7$, and the loss tangent was approximately $\tan\delta=0.003$.

The essential idea behind the studies reported in this thesis was twofold. First, the studies aimed at characterising dielectric media which find usage in the realisation of UHF, microwave or millimetre wave circuits. Second, the studies aimed to obtain data of immediate value in practical design work. For this reason, the structures under test were transmission lines having extensive usage in practical high frequency circuit design.

Keywords: microwave measurement, coaxial resonator, microstrip, strip line
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Espoo, August 1998

Esa Kemppinen
List of original papers

This thesis consists of the following six papers, which will be referred to in the text by their Roman numerals.


In Paper I, determination of the dielectric properties, the real part of the relative permittivity, \( \varepsilon_r \), and loss tangent, \( \tan \delta \), of the low loss microwave ceramic material, barium nonatitanate (Ba\(_2\)Ti\(_9\)O\(_{20}\)), around 1 GHz is discussed. The structures under test were coaxial resonators, the cores of which were made of barium nonatitanate and the metallization was realised by thick film silver. In Paper I, the design of the test structures, the establishment of the test procedures including the theoretical calculations, the performance of the measurements, and the interpretation of the results were the contributions of the author.

In Paper II, the practical realisation of ceramic integrated bandpass filter blocks is presented. This work was based on the determination of the dielectric properties of the ceramic materials and the optimisation of the \( Q \)-values of the quarter wave resonators, as described in Paper I. The design and manufacture of the double tuned resonator pairs, the design of the filters together with the co-authors, the manufacture of the filters, and the theoretical size reduction calculations were the contributions of the author. Determination of the complex permittivity of the coaxial resonators using harmonic resonances and evaluating the size of a ceramic microwave filter in an analytic way were the new contributions provided by Papers I and II.

In Paper III, various measurement methods were demonstrated to determine the real part of the relative permittivity of FR4 fibreglass circuit board. Both frequency and time domain techniques were employed. In addition, two types of electromagnetic simulators were used to verify the measured results. Design of the test structures, measurements using the vector network analyzer and digitizing oscilloscope, theoretical signal flow diagram calculations, electromagnetic analysis, and interpretation of the results were the contributions of the author. The fairly good agreement between the results, \( \varepsilon_r \), and \( Z_c \), from the different measuring and analysing methods was the main new result of this study.

In Papers IV and V, microstrips were employed to determine the dielectric properties of the substrate material as well as the attenuation properties of some Ag and Cu-based microstrip transmission lines up to 50 GHz. In addition, in Paper VI, strip lines implemented in an LTCC (Low Temperature Cofired Ceramic) test piece were also used as test structures. For the first time, attenuation properties of microstrips manufactured by using the gravure offset printing technique, and a special printing technique with light sensitive silver paste, were measured up to 50 GHz, Paper IV. Design of the test structures, conduction of measurements using the vector network analyzer, theoretical flow diagram calculations, and interpretation of the results were the contributions of the author in Papers IV, V, and VI. Evaluation of the dielectric and attenuation properties of the transmission lines under test by using linear regression and \( f_p \) method, based on the
measured magnitude and phase of the $S21$ parameter, and demonstration of the usability of the $f_p$ method up to the millimeter wave region were the new contributions of this study.
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Original papers
1. Introduction

Non-organic dielectric materials are widely used in the realisation of circuitry operating in the microwave and millimetre wave region. Such materials can be employed, for example, as substrate material, resonators, absorbing material, or as the supporting material of high frequency conductors. In this work, the usage of dielectric materials in TEM (Transverse Electromagnetic Field) transmission line structures is addressed. For design purposes, wide-band characteristics of the dielectric materials should be known. The need for appropriate data, for example in microstrip circuit design where the dielectric material is used as a substrate, is clearly stated by Pucel et al. as follows. “For design purposes, it is necessary to know how the characteristic impedance, phase velocity and attenuation constant of the dominant microstrip mode depend on geometrical factors, on the electronic properties of the substrate and conductors, and on the frequency” [1]. Knowledge of these parameters is equally important, not only in the case of microstrips, but also in the case of any transmission line employed in the realisation of microwave circuitry because well characterised circuit elements form the basis for the accurate prediction of the performance of the proposed circuit.

Many of the important properties of the commonly employed planar transmission lines such as microstrips, strip lines and coplanar wave guides, can be calculated theoretically in closed form or by using numerical methods. Recently, electromagnetic simulators have also become much popular in the design work of microwave circuits. However, actual transmission lines fabricated, for example, by screen printing or etching techniques or by laminating fibreglass comprising epoxy and metal, are not in practice ideal. In such cases, electromagnetic simulators can only approximately predict the frequency response of the transmission line. Instead, or in addition, by making well controlled measurements, practical design data characteristic to the fabrication method can be obtained.

In this work, the complex permittivity, $\varepsilon$, of the dielectric material consisting of the real part, $\varepsilon'$, and imaginary part, $\varepsilon''$, has been denoted by $\varepsilon=\varepsilon'-j\varepsilon''$ [2]. The real part of the
relative permittivity, or the real part of the relative dielectric constant, has been denoted by \( \varepsilon_r = \varepsilon'/\varepsilon_0 \); and the loss tangent has been denoted by \( \tan \delta = \varepsilon''/\varepsilon' \); \( \varepsilon_0 = 8.854 \times 10^{-12} \) farad/m.

Attenuation properties of the transmission lines have been denoted by the attenuation coefficient of total loss, \( \alpha_t \), resistive loss, \( \alpha_c \), dielectric loss, \( \alpha_d \), and radiation loss, \( \alpha_r \). In this study, only TEM type transmission lines operating in their dominant modes have been considered because of their great importance and popularity in microwave circuit design.

The objective of this work is to introduce and demonstrate measuring methods, some of which have been established during the work, to obtain data of immediate value in practical design work with dielectrics employed in the microwave and millimetre wave circuit designs corresponding to a certain manufacturing method. In addition, the aim is to determine the dielectric properties, the real part of the relative permittivity of the materials, and in some cases also the loss tangent. In the case of asymmetric strip lines in FR4 material, efforts have also been made to demonstrate methods to determine the characteristic impedance of the transmission lines. Coaxial resonators have been used to characterise barium nonatitanate (Ba_2 Ti_9 O_20) material at frequencies around 1 GHz. Asymmetric strip lines have been used to characterise an FR4 environment up to about 5 GHz by employing different kinds of techniques, including frequency and time domain methods. Microstrips on the alumina substrate, and microstrips and strip lines on or in an LTCC substrate, have been used to characterise the dielectric properties of the ceramic substrates from 45 MHz to 50 GHz. Attenuation properties of Cu-based and Ag-based microstrips have been compared. The Cu-based test circuits were manufactured using etching techniques whereas the Ag-based test circuits were manufactured by using gravure offset printing and by a special printing method employing light sensitive paste. Measurements of the electrical properties of the transmission lines under test were mainly conducted using a vector network analyzer. In some TDR measurement, a digitizing oscilloscope was used.

In the literature, there are many reports dealing with measuring methods of the dielectric properties of electronic materials to be employed in microwave or millimetre wave circuitry. For example, in references [3] and [4], there are listed more than 60 papers about measurements employing microstrips or strip lines as test structures. This list of papers is not, however, repeated in this work. It is clear that the measuring methods can be divided into two main categories. The methods in the first category employ resonators in the determination of the dielectric and attenuation properties of the substrate material. The methods in the second category employ non-resonating structures, and the determination of the dielectric and attenuation properties of the materials are based on the measured propagation constant of the transmission lines [5]. Typically, dielectric properties of the substrate material are addressed through the real part of the relative permittivity and attenuation properties are addressed through the loss tangent. In
addition, attenuation coefficients of the resistive, dielectric, and radiation loss are typically indicated in dB per unit length [6]. This latter method is very practical because, at its best, it gives an insight into the actual properties of the transmission line under test. In addition to these, there are several methods aimed at determining the dielectric properties of the material only, rather than those seen by the electromagnetic signal propagating along a transmission line. In these techniques, for example, a dielectric specimen made of the substrate material is measured in a wave guide [7], or if the specimen is in the form of a cylindrical puck, between two large round metal plates [8]. Moreover, a whole substrate can be metallised to form a multimode resonator and the dielectric properties of the specimen can be determined from the actual resonance frequencies [3].
2. Determination of the permittivity of $\text{Ba}_2\text{Ti}_9\text{O}_{20}$ and the use of the material in a filter application

2.1. Test elements

From the ceramist’s standpoint, it is important to have a reliable measurement method for both the development of microwave ceramics and for quality control in their manufacture. Different methods have been used to determine the dielectric properties of low loss high frequency ceramics. The important electrical parameters that are measured are the complex relative dielectric constant, $\epsilon' - j\epsilon''$, which yields the loss tangent, $\tan\delta = \epsilon''/\epsilon'$, and the variation of complex relative dielectric constant with temperature. Typically, non-metallised cylindrical specimens have been used as test structures in the rod resonator method [8], the wave guide method [7] and the microstrip transmission line method [9]. In all three methods, the ceramic samples are in the form of cylinders with aspect ratios such that they resonate with an external electromagnetic field. Knowing the geometry of the sample and the electromagnetic field configuration, it is possible to calculate the complex dielectric constant from the measured resonance curves. In practice, however, the calculation can be inaccurate if the configuration is prone to radiation.

Cylindrical resonators are most practically employed at microwave frequencies because of their manageable size. For example, in the case of the rod resonator method, at a frequency of 1 GHz, the approximate dimensions of a barium nonatitanate, $\varepsilon_r=37$, dielectric cylindrical resonator would be 65 mm in diameter and 36 mm in height. Such a large resonator is difficult to manufacture reliably using the manual cold pressing method. In contrast, at UHF frequencies, coaxial resonators are well suited to determine the merits of the electrical performance of the ceramic material employed. Compared to dielectric cylinders, coaxial resonators need metallisation of relevant surfaces in order to be able to support the TEM propagation mode. The metallizations can be used to yield
well defined boundary conditions for the TEM fields, especially in the case of a half guided wave length resonator, allowing reliable determination of the complex dielectric constant of the ceramic core. In the case of barium nonatitanate, the length of a guided quarter wave length resonator is around 14 mm at 900 MHz. The cross sectional area is dictated by the desired unloaded $Q$-value, provided that higher order modes are to be avoided [Paper I].

The use of coaxial resonators as test structures is extremely practical when it is desired simultaneously to obtain data that can also be used for practical design purposes. For example, design of a 900 MHz bandpass filter incorporating a ceramic core is one such practical application. With such an application in mind, together with the need to characterise barium nonatitanate ceramic materials around 1 GHz, we employed coaxial resonators to determine the dielectric properties of barium nonatitanate at 900 MHz.

The selection of starting materials to manufacture barium nonatitanate and the fabrication routes used are beyond the scope of this study, and they have been detailed elsewhere [10]. A short description of the manufacture of the ceramic coaxial resonators is, however, given as background information. Ceramic resonator cores with a central cylindrical hole were cold-pressed from calcined powder in a steel mould and then sintered to a peak temperature of 1350 °C with an appropriate temperature vs. time profile. In the sintering process, the shrinkage was about 20%. Thick film silver paste was brushed onto the surface of the sintered resonator and fired on at temperatures in the range 900-930 °C, depending on the ceramic and silver, to give a high conductivity coating. Firing at a lower temperatures resulted in a lower conduction value of the metallization and thus in a higher resistive loss. Two silver coatings were applied and separately fired on to ensure that the thickness of the layer was more than ten times the skin depth, that is more than about 20 μm at 1 GHz.

2.2. Determination of the permittivity and results

Because of the high dielectric constant of the microwave ceramics, three basic TEM coaxial resonator types were employed. These are:

- $\lambda g/4$ resonator - where all the surfaces, with the exception of one end, are covered with metal

- $\lambda g/2$ resonator - where all the surfaces are covered with metal but where, in the middle of the outer conductor, there is a small opening for coupling
- $\lambda g/2$ resonator - where all the surfaces, except for both ends are covered with metal.

Here, $\lambda g$ is the guided wavelength in the coaxial resonator. Theoretically, the second method is the most accurate of the three because of the well defined boundary conditions achieved by covering both the ends with silver. However, the first method yielded nearly the same result as the second one because of the high dielectric constant of barium nonatitanate and the corresponding well defined boundary conditions of the resonator. Additionally, the third method also yielded a quite accurate estimate of the dielectric constant which deviated by less than 1 per cent from that given by the first method, Table 1. The calculated real part of the relative permittivity and the standard deviation are based on measurements of 5-15 test resonators.

**Table 1. Real part of relative dielectric constant of Ba$_2$Ti$_9$O$_{20}$ as determined by various resonator methods, [Paper I].**

<table>
<thead>
<tr>
<th>Resonator type</th>
<th>Frequency range [MHz]</th>
<th>Dielectric constant</th>
<th>Standard deviation</th>
</tr>
</thead>
<tbody>
<tr>
<td>$\lambda g/4$ resonator</td>
<td>900</td>
<td>36.86</td>
<td>0.14</td>
</tr>
<tr>
<td>$\lambda g/2$ short-circuited resonator</td>
<td>1800</td>
<td>36.63</td>
<td>0.02</td>
</tr>
<tr>
<td>$\lambda g/2$ open-circuited resonator</td>
<td>1800</td>
<td>37.12</td>
<td>0.25</td>
</tr>
<tr>
<td>Rod resonator</td>
<td>8300</td>
<td>37.41</td>
<td>0.20</td>
</tr>
</tbody>
</table>

For comparison, the relative dielectric constant obtained from the rod resonator measurement at 8300 MHz is also given. This is very consistent with the values obtained with the coaxial resonator method.

The real part of the relative dielectric constant shown in Table 1 has been calculated from the measured length, $L$, and resonance frequency, $f_r$, of the resonator by assuming that the length of the cavity is a guided quarter wave length (method 1), equation (2-1) or a guided half wave length (methods 2 and 3), equation (2-2).
\varepsilon_r = \left[ \frac{c}{4L_f} \right]^2 \quad \text{(2-1)}

\varepsilon_r = \left[ \frac{c}{2L_f} \right]^2 \quad \text{(2-2)}

where \( c \) is the velocity of light in vacuum. The dielectric constant can also be calculated at the harmonic frequencies, especially by using method 2, because of the well defined boundary conditions at the ends of the resonator. The only sources of error in equations (2-1) and (2-2) affecting the accuracy of determination of \( \varepsilon_r \) are \( L \) and \( f_r \). Because both \( L \) and \( f_r \) can be determined quite accurately within approximately 0.1 per cent, the dielectric constant can be determined, at least, within the accuracy of better than 1 per cent that can be obtained by using the method of total difference.

The resonance frequency of the coaxial resonators, \( f_r \), was measured with an HP8410 vector network analyzer. In addition, the lower and upper -3 dB frequencies, \( f_1 \) and \( f_2 \) respectively, of the \( S_{21} \) response were determined in order to allow calculation of the unloaded \( Q \)-value of the resonators, equation (2-3)

\[ Q = \frac{f_r}{f_2 - f_1} \left( 1 + \frac{K}{K} \right) \quad \text{(2-3)} \]

where \( K \) is the correction factor, \( K=10^{-S_{21}/20} - 1 \) [11]. This correction factor, where \( S_{21}<0 \), takes into account the strength of coupling between the resonator and the measurement system. In practice, transmission type measurements were used where swept frequency power was introduced to the open end of the resonator with a short, flexible open-ended coaxial cable probe [Paper I]. Power was also picked up at the second port of the network analyzer with a similar flexible coaxial cable probe. Loose coupling values, \( S_{21}=-20 \text{ dB} \) to \(-25 \text{ dB} \), were used to make the measurement less sensitive to the correction factor \( K \). Looser coupling values would have resulted in noisier resonance curves on the network analyzer impairing the accuracy of determination of the \( f_r, f_1 \), and \( f_2 \) frequencies. In this study, the measured resonance frequency was used to determine the unloaded \( Q \)-value of the resonator, and the effect of the high but non-infinite unloaded \( Q \)-value of the resonator on the resonance frequency was not considered [12].
2.3. Measuring method of the loss tangent and results

In this study, knowledge of \( \tan \delta \) was gained from measurements of two coaxial resonators of equal cross sections but of different length. If the lengths of the resonators are \( L \) and \( 3L \), then the resonance frequency of the first resonator resonating in the \( \lambda g/4 \) mode would be expected to be the same as that of the second resonator resonating in the \( 3\lambda g/4 \) mode. In practice, the equality of the resonance frequencies was ensured by properly lapping the lengths of the resonators. It is possible to derive equations for the unloaded \( Q \)-values of both \( \lambda g/4 \) and \( 3\lambda g/4 \) resonance modes from the basic definition of \( Q \) and the electromagnetic field inside the resonator, [Paper I]. Because there are two unknown factors in the equations, namely \( \tan \delta \) and the skin depth, \( \delta \), of the metal at the resonance frequency, it is possible to solve both these factors in closed form from two equations. By using this procedure, the value of \( \tan \delta = 0.00014 \) at about 1 GHz was obtained. This also indicated that the value of the skin depth was 2.2 \( \mu m \) which is 0.2 \( \mu m \) greater than that calculated for bulk silver. It can therefore be deduced that the conductivity of the paste is 83 % of that of silver, and by using it, approximately 90 % of the theoretical unloaded \( Q \)-value can be achieved, provided that the ceramic is totally loss-less. When this method is used to determine the value of \( \tan \delta \) it is essential that the quality of the ceramic and the metal paste are the same for both resonators. The low measured value of \( \tan \delta \) indicates that, at 1 GHz, resistive loss is the dominating loss mechanism in ceramic coaxial resonators.

2.4. Temperature coefficient of the resonator

The temperature coefficient of the resonance frequency of the coaxial resonators was measured in an environmental chamber with a network analyzer in the temperature range -20 \(^\circ\)C to +80 \(^\circ\)C. The variation in resonance frequency with temperature of a typical barium nonatitanate coaxial \( \lambda g/4 \) resonator is shown in Fig 1.

Although the curve in Fig. 1 is not strictly linear, an estimation of the slope can be calculated by using the extreme temperature vs. resonance frequency points yielding 4 kHz/\(^\circ\)C. Because the resonance was around 1 GHz the slope value can also be indicated as about 4 ppm/\(^\circ\)C, which is a suitable value for many practical applications.
Fig. 1. The variation in resonance frequency of a barium nonatitanate coaxial \( \lambda g/4 \) resonator with temperature, [Paper I].

The value of the slope is characteristic for undoped barium nonatitanate and can be, if needed, increased by adding small amounts of a suitable dopant, for example Neodymium, to the starting materials. This small amount of Neodymium does not affect the value of \( \tan \delta \), nor the unloaded \( Q \)-value at around 1 GHz.

2.5. UHF-filters employing a ceramic core

By employing a core made of low loss microwave ceramic material with a high permittivity, instead of air, it is possible to reduce the size of, for example, interdigital and comb line filters. Because the basic resonators in such filters are of the TEM type, the length of each cavity can be reduced proportionally to the square root of the relative permittivity if radiation from the open end is neglected. Neglecting radiation of the resonator is, in practice, well justified because of the very high permittivity of the ceramic core. The filter can be realised by using discrete resonators, or it can be an integrated body. An integrated filter bar is better suited for high volume production because the core of the filter can be manufactured by a simple uniaxial pressing with subsequent metallization of the relevant surfaces by silver paste. In the design of the filter, a basic knowledge of the dielectric properties of the material is necessary. Also needed is a knowledge of how the unloaded \( Q \)-value is related to the mechanical dimensions of the resonator, and how metallization affects the \( Q \)-value. The basic design data was established in Paper I.
As a practical example which utilises design data obtained in the basic measurements of the discrete resonators, four different bandpass filters operating in the 900 MHz band mobile phone band, pass band 890-915 MHz, were realised and measured [Paper II]. In general, such small-sized UHF filters can be employed, for example, in wireless communication. The objective of the study was to demonstrate how the use of ceramic materials can result in smaller filter size. Further decrease of the volume of the filter bar was possible by using irregular structures, such as grooves and cuts, between the resonators. These grooves and cuts reduced the weight of the filter but they also made the manufacturing process more demanding. The filters were made of two different ceramic materials and were of interdigital and comb-line types with round central conductors. The basic resonator of the filter bar is thus of strip line type rather than coaxial type. The filters comprised four poles, but in three structures there was one extra resonator, making it possible to realise a notch in the stop band, Table 2.

Table 2. Four different bandpass filter structures, ceramic material employed, relative dielectric constant, volume, and weight. The heights of the filters are about 9 mm. [Paper II].

<table>
<thead>
<tr>
<th>Filter type and Structure</th>
<th>Ceramic material</th>
<th>Dielectric constant</th>
<th>Volume [cm³]</th>
<th>Weight [g]</th>
</tr>
</thead>
<tbody>
<tr>
<td>4 band-pass resonators, Interdigital</td>
<td>Ba₃Ti₉O₂₀</td>
<td>37</td>
<td>5.8</td>
<td>26</td>
</tr>
<tr>
<td>4 band-pass resonators + 1 band-stop resonator, Interdigital</td>
<td>Ba₃Ti₉O₂₀</td>
<td>37</td>
<td>7.2</td>
<td>31</td>
</tr>
<tr>
<td>4 band-pass resonators + 1 band-stop resonator, Comb-Line</td>
<td>Ba₃Ti₉O₂₀</td>
<td>37</td>
<td>4.6</td>
<td>17</td>
</tr>
<tr>
<td>3 band-pass resonators + 1 band-stop resonator, Interdigital</td>
<td>Ba(Sm,Nd)₂Ti₅O₁₄</td>
<td>78</td>
<td>2</td>
<td>9</td>
</tr>
</tbody>
</table>

The familiar Tchebyscheff synthesis was applied in the design of the filters [13]. In the case of the irregular shapes, the coupling coefficients had to be determined experimentally because no closed form formulae were available.
It can be seen that the smallest volume was achieved when the ceramic material with a dielectric constant of $\varepsilon_r = 78$ was used. However, in this design there was one resonator less than in the other designs. The use of the grooves also contributed to the miniaturisation of the structure. Comb-line structures resulted in a smaller volume than interdigital structures because of the inherently weaker coupling between the resonator elements. In the comb-line structure, the central conductors had to be placed closer together than in the interdigital filter, which resulted in a smaller volume. Further reduction of the volume of the filter bar in this application is possible by using a different type of filter synthesis approach instead of the pure Tchebyscheff synthesis.

The volume of the filter bar can be reduced by making it lower. In such a case, of course, the distance between the central conductors has to be decreased resulting in a smaller filter width. However, reduction of the height of the filter makes the unloaded $Q$-value of each resonator smaller. This reduction in turn increases the insertion loss of the filter because the unloaded $Q$-value and mid band insertion loss are inversely related [13]. Insertion loss of the duplex filter in the transceiver parts of wireless telecommunications equipment is of great concern. For system design purposes, the relationship between the insertion loss and the volume of the interdigital four-pole filter described above is evaluated. The reasoning here is as follows. It was found experimentally that the unloaded $Q$-value of a single strip line resonator was close to that of a coaxial resonator when the height of the strip line element was equal to the outer diameter of the coaxial resonator [14]. Thus, the formula relating the dimensions of the optimum unloaded $Q$-value of the coaxial resonator to the height of the filter bar could be employed as a good approximation. For the basic regular filter structure with round central conductors, a closed form equation is available which allows the coupling coefficients to be calculated as determined by the filter synthesis [15]. This, combined with the relationship between the insertion loss and the unloaded $Q$-value, and by assuming that the length of each resonator is a guided quarter wave, allows the volume of the filter as a function of insertion loss to be calculated, Fig. 2.

It can be seen that there is a strong relationship between the volume of the filter and the insertion loss. The requirement of small insertion loss results in a large volume, and thus weight, of the filter. On the other hand, if a moderate insertion loss of about 2 dB in the mid band can be tolerated, the volume of the filter decreases drastically. However, a large reduction in the size of the filter can cause manufacturing problems. With small values of the insertion loss, the permittivity of the ceramic material has an important effect on the volume.
Fig. 2. The calculated volume in cubic centimetres of the four-pole Tchebyscheff interdigital filter bar with 0.1 dB ripple as a function of mid-band insertion loss. Relative permittivity of the ceramic core is $\varepsilon_r=37$ and $\varepsilon_r=78$, respectively, [Paper II].

However, with moderate insertion loss values, when the dielectric material with $\varepsilon_r=78$ is used instead of barium nonatitanate, $\varepsilon_r=37$, the effect on the volume of the filter is not significant. Using the above assumptions, the volume of filter bars with different pole numbers can be evaluated in a straightforward manner.
3. Determination of the permittivity of FR4 substrates and the characteristic impedance of asymmetric strip lines

3.1. Test elements

Laminated fibreglass epoxy (FR4) has been commonly used in the electronic industry to produce printed circuit boards (PCBs). PCB structures vary from simple single sided substrates to complicated multilayer structures. The trend has been to use FR4 substrates in applications where higher frequencies, for example a few gigahertz, are encountered because the price of FR4 is considerably lower than that of dedicated microwave substrate materials. Typical high frequency analog applications of FR4 include, for example, PCBs for portable cellular telephones and other equipment used in wireless transmission products. Also, in high speed digital applications, the microwave properties of the PCB should be known if the harmonic frequencies of the clock signal extend up to the microwave region.

Regardless of the type of application, analog or digital, successful circuit design in the microwave frequency region necessitates a knowledge of the basic electrical properties of the transmission lines employed, namely the propagation factor, $\gamma$ and the characteristic impedance $Z_c$. Typically, some TEM transmission line types are well suited to be designed inside the PCB, such as strip line, coplanar waveguide (CPW), and coplanar strips, whereas some types are applicable on the PCB, such as microstrip, CPW, and coplanar strips. Basic characteristics of the most important lines are well documented in the literature, for example [3, 4, 6, 16].

In this study, the basic aim was to create a quickly performed quality test to enable a PCB manufacturer to assess the quality of the PCBs in terms of the dielectric properties of the PCB and the characteristic impedance of the asymmetric strip lines manufactured inside the PCB. The characteristic impedance of the TEM line and the permittivity $\varepsilon$ are
related through the static equation (3-1) where $Z$ is the intrinsic impedance, $Z=(\mu/\varepsilon)^{\frac{1}{2}}$, of the medium surrounding the conductors [2].

$$Z_C = \varepsilon Z.$$ \hspace{1cm} (3-1)

where $C$ is the electrostatic capacitance of the center and outer conductors per unit length. In spite of the straightforward relationship between $Z_C$ and $\varepsilon$, knowledge of both these quantities forms a good basis for the quality check. To keep the measurements sufficiently simple, open-circuited strip lines were used as the basic test structures. Thus, all the measurements in the frequency and time domain were conducted as 1-port measurements.

The cross sectional geometry of an asymmetric strip line is illustrated in Fig. 3. The asymmetric geometry is related to the manufacturing process of the PCB. The characteristic impedance of the strip line is a function of the distances between the strip and the reference ground planes ($H_1$, $H_2$), the strip cross section and thickness ($W$, $W_{\text{min}}$, $t$) and the relative permittivity, $\varepsilon_r$, of the dielectric medium [4]. The height $H_3$ has no effect on $Z_C$.

![Cross section of an asymmetric strip line](image)

**Fig. 3. Cross section of an asymmetric strip line. Typical measured dimensions are $H_1=274 \ \mu m$, $H_2=518 \ \mu m$, $H_3=274 \ \mu m$, $W=254 \ \mu m$, $W_{\text{min}}=244 \mu m$, and $t=32 \ \mu m$.**

The first five parameters $H_1$, $H_2$, $W$, $W_{\text{min}}$ and $t$ are physical dimensions which can be measured, if necessary, from the cross section of the strip line. On the other hand, $\varepsilon_r$ can only be measured by an electrical method. Thus, if the physical dimensions of the
transmission line are within the specifications but the measured $Z_c$ is not, then the dielectric constant is not correct, perhaps due to a failure in the fabrication process. In this way, the measurement of the dielectric constant and its possible dependence on frequency can give information about the quality of the fabrication process.

### 3.2. Measuring methods of the permittivity and results

Four different methods were used to measure the real part of the relative permittivity of the asymmetric strip lines, [Paper III]:

- TDR method, direct measurement in the time domain

- Frequency/time domain transformation by using the software of the HP8510C network analyzer

- Frequency/time domain transformation using our own software

- $f_p$ method, direct measurement in the frequency domain.

In the TDR method, the time difference between the reflections from the open ends of two strip lines of different lengths was measured. From this time difference, the real part of the relative permittivity of the dielectric medium was calculated. The basic idea in the second and third method was similar to the TDR method, but first the measured frequency domain data, $S_{11}$, was transformed into the time domain by using an inverse Fourier transformation. Finally, in the fourth method, measured data only in the frequency domain was used. In this method, the real part of the relative permittivity of the dielectric medium was calculated at the frequencies, $f_p$, at which the length difference of the two test strip lines corresponded to a guided quarter wave length or its multiples. However, the use of the $f_p$ frequencies made the calculation of $\varepsilon_r$ independent of the discontinuity of the connecting adapter when the length difference corresponded to a guided half wave length or its multiples. The advantage of the $f_p$ method, although most laborious of all the four, is that it gives the frequency dependence of $\varepsilon_r$, whereas the other three methods are not capable of yielding this dependence. For this reason, the candidate for the real part of the relative permittivity obtained by using the first three methods, which rely on the time domain technique, is regarded as the apparent relative permittivity. However, if the frequency dependence of $\varepsilon_r$ is small, a reasonable estimation
for the dielectric constant can also be obtained by using these three methods. The \( \varepsilon_r \) values and error margins obtained by the four methods are summarised in Table 3. Two pairs of test lines of lengths 100 mm and 200 mm, respectively, have been used as test structures.

Table 3. Measured real part of the relative permittivity, \( \varepsilon_r \), of the FR4 material using asymmetric strip lines as test structures. Results from four different measuring methods are shown together with the measurement equipment and measuring conditions, [Paper III].

<table>
<thead>
<tr>
<th>Method</th>
<th>( \varepsilon_r ) Line A100</th>
<th>( \varepsilon_r ) Line A200</th>
<th>( \varepsilon_r ) Line B100</th>
<th>( \varepsilon_r ) Line B200</th>
</tr>
</thead>
<tbody>
<tr>
<td>TDR method, HP54120A</td>
<td>4.22( \pm )0.04</td>
<td>4.22( \pm )0.04</td>
<td>(apparent ( \varepsilon_r ))</td>
<td>(apparent ( \varepsilon_r ))</td>
</tr>
<tr>
<td>Bandwidth is 12.4 GHz</td>
<td></td>
<td></td>
<td></td>
<td></td>
</tr>
<tr>
<td>f/t transform, HP8510B</td>
<td>4.21( \pm )0.05</td>
<td>4.31( \pm )0.05</td>
<td>(apparent ( \varepsilon_r ))</td>
<td>(apparent ( \varepsilon_r ))</td>
</tr>
<tr>
<td>45-4545 MHz, 101 points</td>
<td></td>
<td></td>
<td></td>
<td></td>
</tr>
<tr>
<td>f/t transform, Our program</td>
<td>4.24( \pm )0.05</td>
<td>4.35( \pm )0.05</td>
<td>(apparent ( \varepsilon_r ))</td>
<td>(apparent ( \varepsilon_r ))</td>
</tr>
<tr>
<td>45-4545 MHz, 101 points</td>
<td></td>
<td></td>
<td></td>
<td></td>
</tr>
<tr>
<td>Frequency domain measurement, ( f_p )</td>
<td>4.30( \pm )0.08, ( f_p )=723 MHz</td>
<td>4.29( \pm )0.08, ( f_p )=724 MHz</td>
<td></td>
<td></td>
</tr>
<tr>
<td></td>
<td></td>
<td></td>
<td></td>
<td></td>
</tr>
<tr>
<td></td>
<td>4.11( \pm )0.02, ( f_p )=3699 MHz</td>
<td>4.17( \pm )0.02, ( f_p )=3671 MHz</td>
<td></td>
<td></td>
</tr>
</tbody>
</table>

It can be seen that all the measured values of \( \varepsilon_r \) are quite close to each other. For the pair B100, B200 the frequency to time domain transforms seemed to give slightly higher values than for the pair A100, A200. On the other hand, the result from the TDR measurements and those from the analysis performed in the frequency domain were consistent. The deviations from the average value of \( \varepsilon_r \) given in Table 3 are based only on the two test circuit pairs measured because the purpose was to demonstrate different measurement methods of the relative dielectric constant for quality check purposes. By measuring a larger number of test circuits, a more realistic insight into the variations caused by the manufacturing method can be obtained. According to the measurements in the frequency domain, the frequency dependence of \( \varepsilon_r \) of the FR4 material was not strong at VHF frequencies and beyond but should be taken into consideration in critical designs, for example, in filter circuits. At frequencies lower than 700 MHz, \( \varepsilon_r \) can be obtained with the \( f_p \) method by manufacturing test structures whose length difference is larger than the 100 mm used in this study.
3.3. Measuring methods of the characteristic impedance and results

Three different methods were used to measure the characteristic impedance of the asymmetric strip lines. In addition, two electromagnetic simulators were employed to calculate $Z_c$ based on the measured physical dimensions of the test structure. Finally, two formulae, available in the literature, were used to calculate the $Z_c$ values [Paper III]:

- TDR method, direct measurement in the time domain
- Frequency/time domain transformation by using the software of the HP8510C network analyzer
- Frequency/time domain transformation using our own software
- HFSS, electromagnetic simulator
- Momentum, electromagnetic simulator
- Formulae available in the literature.

The TDR measuring method was simply conducted by using the digitizing oscilloscope, HP54120A. Both of the frequency to time domain transformation methods were quick to use. In our own software, the vertical axis was directly scaled to ohms which was not possible in the case of the network analyzer. Typically, HFSS is used to simulate real three-dimensional structures and, as such, it was suitable for calculating the characteristic impedance of the asymmetric strip line. On the other hand, Momentum is used merely to analyse planar structures. For this reason, in the Momentum analysis, we used a central conductor with zero thickness. The $Z_c$ values obtained by the three measuring methods, electromagnetic simulations and closed form equations are summarised in Table 4. Estimated error margins for the measurements and HFSS analysis are also shown. The same test strip lines which were used to assess the dielectric constant have been used here.

All the values for the characteristic impedance determined by the first three methods were reasonably close to each other. The amount of deviation in the characteristic impedance is consistent with the manufacturing tolerances of the strip lines. HFSS tended to give slightly smaller values than the three measuring methods. In the Momentum analysis, the average dimensions of the cross section were used and the thickness of the
line had to be set to zero. This resulted in too high a value for the characteristic impedance. Also, in calculating the characteristic impedance from the formulae available in the literature, average dimensions for the cross-section were employed.

Table 4. Comparison between the characteristic impedances determined by three different measuring methods of four test strip lines. Also shown are the results obtained from the HFSS analysis, Momentum analysis and formulae available in the literature are shown, [Paper III].

<table>
<thead>
<tr>
<th>Method</th>
<th>Line A100</th>
<th></th>
<th>Line A200</th>
<th></th>
<th>Line B100</th>
<th></th>
<th>Line B200</th>
<th></th>
</tr>
</thead>
<tbody>
<tr>
<td>TDR method, HP54120A Bandwidth is 12.4 GHz</td>
<td>54.6\pm0.5</td>
<td>55.2\pm0.5</td>
<td>54.5\pm0.5</td>
<td>53.9\pm0.5</td>
<td></td>
<td></td>
<td></td>
<td></td>
</tr>
<tr>
<td>f/t transformation, HP8510B, 45-4545 MHz</td>
<td>55.0\pm0.6</td>
<td>56.1\pm0.6</td>
<td>55.2\pm0.5</td>
<td>55.3\pm0.7</td>
<td></td>
<td></td>
<td></td>
<td></td>
</tr>
<tr>
<td>f/t transform, Our software, 45-4545 MHz</td>
<td>54.7\pm0.6</td>
<td>55.1\pm0.7</td>
<td>55.0\pm0.5</td>
<td>54.6\pm0.6</td>
<td></td>
<td></td>
<td></td>
<td></td>
</tr>
<tr>
<td>HFSS, ( f=100 \text{ MHz} )</td>
<td>53.3\pm0.8</td>
<td>53.6\pm0.8</td>
<td>52.4\pm0.8</td>
<td>52.4\pm0.8</td>
<td></td>
<td></td>
<td></td>
<td></td>
</tr>
<tr>
<td>Momentum, ( f=100 \text{ MHz} )</td>
<td>58.3, ( t=0 \text{ ( \mu )m} )</td>
<td></td>
<td></td>
<td></td>
<td></td>
<td></td>
<td></td>
<td></td>
</tr>
<tr>
<td>Formula [4]</td>
<td>57.1, ( t=0 \text{ ( \mu )m} )</td>
<td></td>
<td>52.0, ( t=32 \text{ ( \mu )m} )</td>
<td></td>
<td></td>
<td></td>
<td></td>
<td></td>
</tr>
<tr>
<td>Formula [17]</td>
<td>55.0, ( t=32 \text{ ( \mu )m} )</td>
<td></td>
<td></td>
<td></td>
<td></td>
<td></td>
<td></td>
<td></td>
</tr>
</tbody>
</table>

The formula from [4], narrow strip assumption, resulted in \( Z_c=52.0 \text{ \( \Omega \)\), which is lower than the measured values. This value is, however, close to that obtained with HFSS
because the formula, as well as HFSS analysis, assumed that the dielectric medium was homogenous and isotropic. However, the result from the formula [17], $Z_c=55.0 \ \Omega$, compared well with the measured values. The validity of the formula [4] was compared with the Momentum analysis when the thickness of the strip was set to zero. The calculated value of 57.1 $\Omega$ compared satisfactorily with the result from Momentum analysis, $Z_c=58.3 \ \Omega$. 
4. Determination of the permittivity of alumina and LTCC materials and attenuation properties of microstrips and strip lines

4.1. Test elements

The ceramic substrates under study were made of alumina and LTCC, and the test elements were microstrips and strip lines. On the conventional alumina substrates, only microstrips were manufactured. They were, however, made of copper and silver by using the same pattern file and identical substrates so that direct comparison between the attenuation coefficients was possible [Paper IV]. The copper strips were manufactured by the direct copper bond technique with subsequent etching. Some of the silver conductors were manufactured by screen printing photosensitive silver paste with subsequent etching, whereas some silver conductors were manufactured by gravure offset printing. The LTCC substrate was a laminated four-layer structure. It was thus possible to design both microstrip and strip line test structures in such a way that the lengths of the transmission lines were equal. The strip lines were, of course, narrower than the microstrips because a characteristic impedance of 50 Ω was pursued. In the LTCC structure, copper metallization was used. However the surface Cu traces were covered with electrolytic Ni/Au overplate.

The layout of the alumina substrate with Cu microstrips and that of the LTCC test substrate is shown in Fig. 4. The actual area of the alumina substrate was 2” x 2”, and the nominal height was 0.25 mm. The area of the LTCC structure was 30 mm x 50 mm, and the nominal height was 1 mm consisting of four 250 μm thick ceramic sheets, the two lowest of which were only to make the structure more robust. In this study, only straight microstrips and strip lines were used as test circuits but, for comparison purposes, straight, stub, and ring resonators were designed on the alumina substrate. Also, straight and ring resonators were designed in the LTCC test piece.
The quality of the physical properties of the transmission lines under test were evaluated in terms of strip width, strip height, profile of the cross section and edge smoothness of the strip. The physical dimensions were measured by using a profilometer and tooling microscope. The characteristic impedance of the transmission lines was assessed through calculations and measurements. The calculations were based on the measured dimensions whereas, in the measurement, the frequency to time domain transformation option of the HP8510C network analyzer was used. These main parameters, which are based on measurements of several strips under test, are summarised in Table 5.

The profiles of all the microstrips looked approximately rectangular with the exception of those made using the gravure offset method. Because of the relatively viscous silver paste used, the corners of the cross sections of the strips were rounded. However, the quality of the microstrips in this study was appropriate for measurements up to 50 GHz.

The nominal width of the strip lines in the LTCC structure was 180 μm, and the measured and calculated values of the characteristic impedance were 47.5 Ω and 49.2 Ω, respectively. Again, a Zc of 50 Ω was pursued. The cross sections of the strip lines did not look like proper rectangles because the borders of the central conductors were quite sharp. The maximum thickness of the central conductor was 8 μm.
Table 5. Strip width and thickness, edge smoothness, measured and calculated Zc of the microstrips fabricated by different methods, [Paper IV].

<table>
<thead>
<tr>
<th></th>
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<th></th>
<th></th>
<th></th>
<th></th>
</tr>
</thead>
<tbody>
<tr>
<td>Cu strip on Alumina, etched</td>
<td>293 ± 1.5</td>
<td>24</td>
<td>3</td>
<td>45.8</td>
<td>44.6</td>
</tr>
<tr>
<td>Ag strip on Alumina, etched</td>
<td>239 ± 4</td>
<td>10</td>
<td>5</td>
<td>50.0</td>
<td>50.1</td>
</tr>
<tr>
<td>Ag strip on Alumina, Gravure offset</td>
<td>269 ± 12</td>
<td>9</td>
<td>26</td>
<td>48.4</td>
<td>47.4</td>
</tr>
<tr>
<td>Cu/Ni/Au strip on LTCC, Screen printed</td>
<td>400 ± 5</td>
<td>8</td>
<td>3</td>
<td>47.5</td>
<td>49.2</td>
</tr>
</tbody>
</table>

To make on-wafer measurements possible, there must be a special connecting structure at the ends of the transmission lines under test. In the case of microstrips, the connecting structure comprised a signal line and two vias connected to the ground plane of the substrate. In practice, this structure consisted of a short piece of coplanar waveguide completing the transition to the microstrip line. In the case of the strip lines there is an opening in the upper ground plane for the signal line, and the central conductor of the strip line is accessed through a via. Additionally, both the upper and lower ground planes are connected together with a multiplicity of vias. Typical connecting structures of strip lines and microstrips of the LTCC test structure are illustrated in Fig. 5.

Although in the case of microstrips the length of the connecting structure is short compared to the total length of the transmission line, the design of the connecting structure has an effect on the measured S parameters of the line under test. Ideally, the characteristic impedances of the coplanar wave guide sections should be 50 Ω. Deviation from 50 Ω impairs the measured values of S11 and S22, especially in the millimetre wave region whereas at low microwave frequencies the connecting structure can be regarded as almost ideal. To achieve 50 Ω coplanar sections, the minimum line width of the manufacturing technique should be approximately 50 μm. This requirement was best met with the LTCC test structure. As a consequence, the measured S parameters of the microstrips of the LTCC structure were slightly better behaved than those of the
microstrips on alumina. However, all the measured $S$ parameters of the microstrips could be used to determine the dielectric properties of the ceramic substrates.

![Fig. 5. Connecting structure of the strip line a), and micro strip b), of the LTCC substrate. The width of the microstrip is approximately 400 $\mu$m and the via diameter in both photographs is nominally 100 $\mu$m, [Paper VI].](image)

The frequency response of the connecting structure of the strip lines was valid only up to about 10 GHz. The obvious reason for this was that the gaps between the signal pad and the ground planes were so narrow, Fig. 5a, that they caused coupling with increasing frequency. This harmful coupling did not, however, degrade the phase of $S_{21}$ too much so that determination of the permittivity of the LTCC material up to 50 GHz was possible.

### 4.2. Measuring set-up

In the measurement of the microstrips and strip lines, on-wafer measurement techniques were used. The measuring set-up consisted of Cascade’s manual probe station, Summit 9651, and a vector network analyzer, HP8510C. The microwave probe heads were of ground-signal-ground type, Cascade ACP50, with a pitch size of 200 $\mu$m. All the coaxial connectors of the microwave probe heads and connecting cables were of 2.4 mm standard in order properly to cover the frequency range up to 50 GHz. Measurements, calibrations, and the storing of the measured data were controlled by Cascade Microtech’s WinCal software, version 2.1. The main equipment of the measuring set-up is illustrated in Fig. 6.
The on-wafer measurement approach adopted has the advantage over the traditional approach employing coaxial connectors that many more test circuits can be accommodated on one substrate. Also, the on-wafer type connecting structure is easily repeatable if the tolerances of the manufacturing method of the circuits can be kept tight. In addition, the calibration of the network analyzer can be performed conveniently by using planar calibration standards.

All the frequency domain measurements were made from 45 MHz to 50 GHz. The number of frequency points was 801, the largest value the HP8510C can accept, in order to obtain experimental data at as many frequency points as possible. Several calibration procedures of the WinCal software were tested, and the LRRM Port 1 calibration method was obtained to give the most repeatable results up to 50 GHz. This procedure was also suggested by Cascade Microtech based on [18].

The calibration standards used were on Cascade’s Impedance Standard Substrate (ISS) made of alumina. Thus the match between the permittivities of the test circuits and the ISS was good. In the LRRM Port 1 calibration, line, reflection and matching standards were employed. The match elements were laser trimmed to 50 Ω within an accuracy of 0.3 per cent. The length of the line element, as expressed in terms of propagation delay, was 1 ps. The validity of the calibration was checked from time to time by raising the probe of Port 1 well up and measuring the $S_{11}$ response and making comparisons with earlier measurements, a property of the WinCal software. Repeatability of the measurements was verified to be good.
4.3. Measuring methods of the permittivity and results

Two methods were used to determine the real part of the frequency dependent relative effective permittivity, \( \varepsilon_{re}(f) \), of the microstrips. Both methods were based on 2-port phase measurement, and they were also used to determine the real part of the relative permittivity, \( \varepsilon_r \), of the strip lines. [Paper IV, Paper V, Paper VI]. In this study, the methods are called

- Linear regression method
- \( f_p \) method.

In the first method, the phase of the \( S_{21} \) parameter of microstrips which were similar but of different lengths was measured. Because of the multiplicity of microstrips under test, in order to decrease the effect of measurement errors, linear regression was used as the mathematical tool to determine the dielectric properties of the substrate. The second method was also based on phase measurement but now only two microstrips were employed. The frequencies, denoted by \( f_p \), at which the length difference of the microstrips corresponded to a multiple of a half guided wave length, were determined from the measured \( S_{21} \) parameters. From these \( f_p \) frequencies, \( \varepsilon_{re}(f) \) of the substrate was determined. The \( f_p \) method applied to the ceramic structures is an extension of the method applied to the FR4-structures because now 2-ports, instead of 1-ports, were used as test elements to avoid radiation from the open ends of the transmission lines. The theoretical background of the extended \( f_p \) method is given in Paper V where the relevant equations have been derived using signal flow diagram analysis. Once the real part of the effective relative permittivity was obtained, the actual \( \varepsilon_r \) was calculated using formulae available in the literature, for example [6]. The dispersion in the case of microstrips was taken into account by using the model suggested by Kirschning and Jansen [19]. The use of strip lines in the LTCC structure as test circuits directly yielded the actual \( \varepsilon_r \) value of the substrate. Results of both measuring methods are summarised in Fig. 7 when alumina and LTCC substrates were used as the test vehicles.

In the case of the alumina substrate, Fig. 7a and Fig. 7b, both measured and calculated curves agree well. The agreement is better than 2 per cent in the case of the linear regression method, Fig. 7a, and in the case of the \( f_p \) method, the curves are virtually overlapping. The slightly higher \( \varepsilon_{re}(f) \) in Fig. 7a is explained by the fact that the characteristic impedance of the copper microstrips was 44.6 \( \Omega \), as determined by Linacalc software from the actual physical dimensions, whereas \( Zc \) of the silver microstrips, Fig. 7b, was approximately 50 \( \Omega \).
The measured low frequency value of $\varepsilon_r(f)$ was about 6.5, Fig. 7a. Based on this value and the physical dimensions of the microstrip, the dielectric constant of the substrate was calculated to be $\varepsilon_r = 9.5$, which value was also given by the manufacturer.

The values of $\varepsilon_r(f)$ and $\varepsilon_s$ of the LTCC material were determined up to 50 GHz by using the linear regression and $f_p$ method, Fig. 7c and Fig. 7d, respectively. Now, all the test microstrips used to determine $\varepsilon_r(f)$, and strip lines used to determine $\varepsilon_s$ were within the same test piece, so that direct comparison of the results is justified. It can be seen that
the measured and calculated $\varepsilon_r(\omega)$ curves agree well, as do the measured $\varepsilon_r$ curves, Fig. 7c and Fig. 7d. In the measured $\varepsilon_r$ curves, there is a small undulation present which was attributed to the connection structures of the strip lines. Basically, microstrips were better behaved test structures than strip lines. It can be concluded, however, that the measured value of the real part of the relative permittivity, $\varepsilon_r=5.7$, of the LTCC material agreed well with that of the LTCC manufacturer which was also $\varepsilon_r=5.7$.

4.4. Determination of the attenuation properties and results

Two methods were used to determine the attenuation coefficient of the total loss, $\alpha$, of the microstrips and strip lines employing ceramic substrates. Both methods were based on the measurement of the magnitude of the $S_{21}$ parameter of the transmission lines under test, Paper V, and Paper VI. The methods were the same as in the case of the determination of the dielectric properties of the ceramic substrates but now the focus was on the attenuation properties:

- Linear regression method
- $f_p$ method.

In the first method, the magnitude of the $S_{21}$ parameter of similar transmission lines of different lengths was measured. Again, because of the multiplicity of transmission lines under test, linear regression was used as the mathematical tool to determine the attenuation coefficient. The second method was also based on the measured magnitude of the $S_{21}$ parameter but now only two transmission lines were employed. The frequencies, denoted by $f_p$, at which the length difference of the transmission lines corresponded to a multiple of a half guided wave length in the case of 2-port measurements, were determined from the measured $S_{21}$ parameters. At these $f_p$ frequencies, $\alpha$ of the transmission lines under test was determined. It was found, however, that because of radiation losses from the open end of the microstrips under test, the 2-port measuring method resulted in more reliable results than the 1-port method. For this reason, 1-port measurements are not dealt with in this study.

The measured attenuation coefficients of the microstrips and strip lines of the LTCC substrate are shown in Fig. 8. For comparison, attenuation coefficients calculated from closed form formulae, are also shown for both microstrips [6] and strip lines [20]. In the calculations, measured values of $\varepsilon_r(\omega)$, $\varepsilon_r$, $Z_c$, and physical dimensions of the test circuits were used. The small increase of the characteristic impedance of the microstrips with
The conductivity of copper, $\sigma_{Cu}=5.85\times10^7$ S/m [21], and loss tangent given by the manufacturer, $\tan\delta=8.5\times10^{-4}$, were employed in the calculations.

Fig. 8. Measured and calculated attenuation coefficient of microstrips on LTCC substrate using linear regression method a), and $f_p$ method b). Measured and calculated attenuation coefficient of strip lines in LTCC substrate using linear regression method c), and $f_p$ method d), [Paper VI].
It can be seen that the measured and calculated values of $\alpha_t$ of the microstrips are in reasonable agreement up to 20 GHz in the case of the linear regression method and up to 35 GHz in the case of the $f_p$ method. The slightly larger measured values of the attenuation constant can most likely be explained by loss contributions from surface roughness and non-ideal edge definition of the Cu conductors. It is also possible that the actual conductivity of copper is less than the bulk value [22]. At 50 GHz, the calculated value of $\alpha_t$ is about 2 to 3 times the measured value.

The agreement between the measured and calculated values in the case of strip lines is worse than in the case of microstrips. Attenuation coefficient measurements of the strip lines could be extended up to nearly 10 GHz by using the linear regression method and up to about 20 GHz by using the $f_p$ method. The measured attenuation values are about 1.5 to 2 times the calculated values in both cases. The most significant reason for the difference is presumably due to the cross sectional geometry of the strip line. Calculation assumes that the cross section of the strip line is rectangular but, in reality, the edges or the central conductor are sharp, resulting in a higher resistive loss and degraded characteristic impedance.

Attenuation of strip lines is larger than that of microstrips, as was also predicted by the formulae. For example, the measured $\alpha_t$ of strip line was 0.03 dB/mm at 10 GHz, whereas that of microstrip was about one third of this value, $\alpha_t$ = 0.012 dB/mm. Because of the similarity of the results, both measuring methods can be used to assess the approximate attenuation coefficients of the microstrips fabricated on LTCC material. The same applies to strip lines, at least up to 10 GHz, and after redesign of the connecting structure, most likely up to the millimetre wave region.

### 4.5. Calculation of the loss tangent

In this study, no special sets of test circuits [23] were designed in order to evaluate the value of tan$\delta$. However, to determine the approximate order of tan$\delta$ of the LTCC material, we used the approach employed in [24]. The approach is also valid with other substrate materials. In this approach, the amount of dielectric loss of the microstrip line was obtained by subtracting the theoretically calculated resistive loss from the measured total loss. The amount of resistive loss can also be calculated by using numerical methods [25]. The loss tangent was readily calculated from the dielectric loss contribution by using a closed form equation [6]. In this approach, possible loss contributions from radiation, leakage of energy, surface roughness and non-ideal edge smoothness of the metallisation were omitted. This is better justified at low frequencies since in the millimetre wave region these factors typically increase the coefficient of total loss. To
determine $\tan\delta$, we used $\alpha_t$ values of the microstrips obtained by both the linear regression method and the $f_p$ method. Fig. 9 shows the values of $\tan\delta$ obtained by this approach.

![Graphs showing Loss tangent of the LTCC material](image)

**Fig. 9.** Loss tangent of the LTCC material obtained from microstrip measurements by using linear regression a), and $f_p$ method b), [Paper VI].

The value of the loss tangent given by the LTCC manufacturer was $\tan\delta=8.5\times10^{-4}$ at 9.8 GHz. It can be seen from Fig. 9 that $\tan\delta$ is, at its largest, several times this value. The error reflects the sensitivity of the subtraction method to errors in the measured $\alpha_t$ value and to the simplifying assumptions made in the calculations. Also, the conductivity of the metal used in the theoretical calculation of the resistive loss can deviate from the bulk metal’s conductivity [22] resulting in erroneous value of $\tan\delta$. By averaging the values of $\tan\delta$ over the measured band the approximate values, $\tan\delta=0.0032$ and $\tan\delta=0.0025$, were obtained using the linear regression and $f_p$ methods, respectively. Both of these values were about three to four times that given by the manufacturer.
5. Conclusions

The objective of this work was to introduce and demonstrate measuring methods to obtain data of dielectric materials with immediate value in practical design work for microwave and millimetre wave circuits. An additional aim of this work was to determine the dielectric properties, the real part of the relative permittivity of the materials, and in some cases also the loss tangent. Dielectrics under test were barium nonatitanate, FR4 fibreglass composite in a multilayer printed circuit board, alumina, and low temperature cofired ceramic. All the test structures were supposed to operate in their dominant TEM mode. In the case of barium nonatitanate, the test structures were coaxial resonators, and the dielectric properties were determined at approximately 1 GHz by using coaxial cable probes. The FR4 test structures were asymmetric strip lines, the properties of which were measured through coaxial SMB connectors up to about 5 GHz. In the remaining cases, alumina and LTCC, the test structures were microstrips and strip lines which were measured using on-wafer techniques from 45 MHz up to 50 GHz. Such analyses as the determination of the complex permittivity of the coaxial resonators using harmonic resonances, evaluating the size of a ceramic microwave filter in an analytic way, and evaluation of the dielectric and attenuation properties of the transmission lines under test by using linear regression and $f_r$ method were the new contributions provided by this work. A new contribution was also the demonstration of the usability of the $f_r$ method to determine dielectric and attenuation properties of the transmission lines under test.

The real part of the permittivity of barium nonatitanate was measured to be $\varepsilon_r=36.63-37.12$ at around 1 GHz by using three different resonator structures. The loss tangent of the material was determined, by using the method of harmonic resonances, to be $\tan\delta=0.00014$ at about 1 GHz. The change of the resonance frequency of the coaxial resonators with temperature, in the range of -20…+80 °C, was only 4 ppm/°C. Finally, as a practical consequence of the resonator study, it was demonstrated how low loss microwave ceramic materials can be used to realise compact bandpass filters with small size and yet good performance for the 900 MHz mobile phone band. Besides $\text{Ba}_3\text{Ti}_5\text{O}_{20}$, $\text{Ba(}\text{Sm,Nd})_2\text{Ti}_5\text{O}_{14}$ ceramic material with a dielectric constant of $\varepsilon_r=78$ was also
employed in order to improve the miniaturisation. Both interdigital and comb line filters were realised. The volume of the smallest filter was only 2 cm$^3$, and the weight was 9 g.

Four different methods were used to determine the permittivity of FR4 material employed in multilayer printed circuit boards. Three of the methods were based on measurement in the time domain, and the real part of the permittivity obtained was 4.21...4.24, as measured from one pair of test structures. The disadvantage of the time domain methods was that they could not reveal the possible frequency dependence of the real part of the relative permittivity. For this reason, the values obtained were called the real part of the apparent relative permittivity. However, at low microwave frequencies, say above 1 GHz, the change of $\varepsilon_r$ was found to be so small that the time domain methods were able to give a reasonably accurate estimate for $\varepsilon_r$. Employment of the frequency domain measurement, $f_p$ method, showed that the real part of the relative permittivity decreased slightly with frequency. At 723 MHz and 3699 MHz, $\varepsilon_r$ was measured to be 4.30 and 4.11, respectively. The characteristic impedance of the structures under test was also measured using time domain methods. Typically, values of the order from $Z_c=54.6\pm0.5$ to $55.0\pm0.6$ $\Omega$ were obtained, measured from a single test structure, showing that the characteristic impedance was slightly above the reference impedance level of 50 $\Omega$.

Wide-banded dielectric properties of the alumina substrate from 45 MHz to 50 GHz were determined by linear regression and the $f_p$ method using microstrips as test structures. A known substrate was employed to validate both methods and linear regression and the $f_p$ method resulted in the same value of the real part of the relative permittivity, $\varepsilon_r =9.5$. This was also the value given by the substrate manufacturer. Moreover, the same dispersion behaviour of the microstrips was obtained with both measuring methods. The dispersion was also in very good agreement with the Kirschning Jansen dispersion model. The attenuation properties of Cu-based and Ag-based microstrips manufactured on alumina were compared up to 50 GHz. The Cu-based test circuits were manufactured using an etching technique whereas the Ag-based test circuits were manufactured using gravure offset printing and a special printing method with light sensitive paste. Although the bulk conductivity of silver is higher than that of copper, the attenuation of the Cu-based microstrip, $S_{21}$ parameter, was slightly smaller than that of the Ag-based microstrip. The measured attenuation coefficient of the total loss of the Cu-based microstrip was $\alpha =0.033$ dB/mm at 40 GHz and that of the Ag-based microstrips was $\alpha =0.038-0.0043$ dB/mm. This finding is best explained by the higher quality of the direct bond manufacturing process of the copper microstrips. Typically, the measured $\alpha$ values were 1.4-1.9 times the calculated values.

Wide-banded dielectric properties of the LTCC material were also determined from 45 MHz to 50 GHz by linear regression and the $f_p$ method. Both microstrips and strip
lines were used as test structures and both methods resulted in the same value of $\varepsilon_r = 5.7$, regardless of the test structure type. This $\varepsilon_r$ value was also given by the manufacturer, Kyocera, Japan. However, microstrips were found to be more practical test structures than strip lines because of the smaller discontinuity in the connection structure of the microwave measuring probe. Because of this discontinuity, the attenuation coefficient of the strip lines could be measured only up to about 10 GHz, whereas $\alpha_t$ of microstrips could be determined up to 50 GHz. The attenuation coefficient obtained with the linear regression method was about 20 per cent larger than the calculated value up to 20 GHz. Similar agreement extended up to 35 GHz in the case of the $f_p$ method. At 20 GHz, the measured attenuation coefficient of microstrips obtained using both methods was about $\alpha_t = 0.02$ dB/mm. The difference between the measured and calculated $\alpha_t$ values of the strip lines increased with frequency, independently of the measuring method. While being in reasonable agreement at low microwave frequencies, the measured value at 10 GHz was about $\alpha_t = 0.03$ dB/mm and the calculated value $\alpha_t = 0.016$ dB/mm. Clearly, the attenuation of the strip line was higher than that of microstrip.

In order to determine an approximate value for the loss tangent of the LTCC material, the amount of dielectric loss of the microstrip line was obtained by subtracting the theoretically calculated resistive loss from the measured total loss. The loss tangent was readily calculated from the dielectric loss contribution by using a closed form equation. In this approach, only resistive loss and dielectric loss were taken into account. The value of the loss tangent given by the LTCC manufacturer was $\tan \delta = 8.5 \times 10^{-4}$ at 9.8 GHz which was obtained using a cavity resonator method. The measured loss tangents were not flat as a function of frequency. However, by averaging the values over the measuring band, the approximate values, $\tan \delta = 0.0032$ and $\tan \delta = 0.0025$, were obtained using linear regression and $f_p$ methods, respectively. These values were about three to four times the value given by the manufacturer. The error reflects the sensitivity of the subtraction method to errors due to the measured $\alpha_t$ value and the simplifying assumptions made in the calculations. Also, the conductivity of the metal used in the theoretical calculation of the resistive loss can deviate from the bulk metal’s conductivity, resulting in erroneous values for $\tan \delta$. 
References


Errata

Paper I: p. 144 the complex dielectric constant should read in the correct form as:

\[ \varepsilon' - j\varepsilon'' \]