Simple and High-Sensitivity Dielectric Constant Measurement Using a High-Directivity Microstrip Coupled-Line Directional Coupler

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Abstract—Simple methods using a microstrip coupled-line directional coupler (CLDC) are presented for dielectric constant measurements. The material under test (MUT) is placed on the coupled-line section of the coupler, and either the coupler’s coupling (|S_{21}|) or its isolation level (|S_{41}|) is considered as the sensor’s response. Putting different MUTs on the microstrip line leads to a change in the effective dielectric constant of the structure and consequently causing a change in the coupling coefficient. In addition, since the isolation level of a microstrip coupled-line coupler depends on the phase velocity difference between the substrate and the medium above the signal strips, putting different MUTs on the line significantly changes the isolation level. This change is significantly greater than the change in |S_{21}| level of a microstrip line when loaded with different MUTs. Validation of the method is presented through measurements for both solid and liquid MUTs.

Index Terms—Coupled-line directional coupler (CLDC), dielectric constant measurement, material characterization, microwave sensors.

I. INTRODUCTION

The response of a material to electrical fields depends on its dielectric constant. Therefore, fast and precise determination of the dielectric constant of materials is of considerable importance, in particular in microwave frequencies, in various areas such as biomedical, materials science, food industry, and circuit board technologies [1]–[3].

Different techniques have so far been proposed for the determination of the dielectric constant of materials at microwave frequencies [4]. These methods can be classified into two categories: resonance and nonresonance methods. The resonance methods are based on the change in the resonance frequency of a resonator either entirely or partially loaded with the materials under test (MUTs) [5]–[14]. In the nonresonance methods, the permittivity of the MUT is fundamentally deduced from the change in the characteristic impedance and/or the phase velocity. In such methods, an electromagnetic wave passes an interface between two media, one is loaded with the MUT, and the other is not. The change in the characteristic wave impedance and wave velocity results in a partial reflection of the electromagnetic wave. Amplitude and phase measurements of reflection [15]–[17], transmission [18], or both [19]–[21] provide enough information to characterize the MUT; however, it is possible to characterize material using amplitude-only measurements [22]–[26].

Different material characterization methods can be classified based on the propagation medium such as the free-space methods [27], [28], metallic waveguides (and cavities) methods [23], [29], [30], coaxial lines methods [24], [25], [31], and planar circuits methods, such as coplanar waveguide structure [32], [33] and microstrip lines [34]–[36]. The microstrip-based methods in which the MUT is placed on a microstrip line or on a planar resonator etched on the signal plane or ground plane have advantages such as inexpensive fabrication (owing to printed circuit board manufacturing technologies), ease of sample preparation, and fast measurement procedure. However, in these devices, the field is mostly concentrated in the substrate (host medium), while the MUT is typically placed on the top or bottom of the substrate. This leads to a weak interaction between the field and the MUT, hence restricting the sensitivity of the sensor to the changes in the MUT.

To enhance the sensitivity of microstrip-based sensors, different designs have been proposed in the literature. In [37] and [38], by building a channel inside the substrate to host the liquid MUTs, the sensitivity is improved at the cost of fabrication complexity. Also, by modifying planar resonators, the sensitivity of resonance-based microstrip sensors has been enhanced. For example, active feedback loops, including a transistor amplifier [39], [40], were introduced to the planar resonators, leading to a significant enhancement in the sensitivity, however, at the cost of increasing the complexity of the circuit. Other examples of planar resonators include shape-optimized resonators [41], [42], multiple coupled resonators [43], using metamaterial coupling [44], and using a...
3-D capacitor within the planar resonator [45] (a nonexhaustive survey of planar resonators is available in [46]).

In nonresonant microstrip-based sensors where the MUT is loaded onto a transmission line (TL), due to the weak interaction between the field and the MUT, the amplitudes of the scattering parameters do not change considerably due to the change in the MUT. Therefore, the phase measurement of the scattering parameters (usually the phase of $|S_{21}|$) is preferred [35], [47]–[49]. In these sensors, the sensitivity of the phase of $|S_{21}|$ with respect to the dielectric constant of the MUT can be enhanced by increasing the length of the line and/or reducing the phase velocity (i.e., increasing the phase constant), where the former represents a penalty in terms of size. The latter can be realized using artificial TLs, such as composite right-/left-handed TLs [47], electro-inductive wave TLs [48], and capacitive-loaded slow wave transmission [35], [49]. To improve the sensitivity of nonresonant microstrip-based sensors, the MUT is placed on a coupled-line section in [50]–[53] and is characterized by extracting and analyzing the odd-mode impedance of the loaded line. Since the odd mode is mainly related to the mutual capacitance between the coupled strips (rather than between the strips and the ground plane), a high sensitivity is achievable.

In this article, two methods using a microstrip coupled-line directional coupler (CLDC) for dielectric constant measurement are presented. In both the methods, the MUT is placed on the coupled-line section of the coupler. The first method, which is a resonant method, is based on the change in the coupling coefficient of the coupler, while in the second method, which is a nonresonant method, the sensing is based on the isolation coefficient of the coupler. We show that for a unique determination of the dielectric constant of the MUT from the isolation level, a conventional microstrip CLDC which has a low level of isolation in its unloaded state does not work, and the coupler needs modifications for a high level of isolation (i.e., a high directivity). While in the works of Piekarz et al. [50]–[52] and Sorocki et al. [53] (in which coupled-line sensors were used), measuring the amplitude and phase of all the transmission ($S_{ij}$) and reflection ($S_{ii}$) coefficients is required for odd-mode impedance extraction, our proposed methods need only the amplitude of one transmission coefficient (not its phase). Therefore, there is no need for a vector network analyzer (VNA), but instead either a scalar network analyzer (SNA) or a frequency synthesizer and a power detector, which makes the measurements simpler and lower cost. To close this section, the advantages of this work are the simplicity and low cost of the measurement setup, low cost and ease of sample preparation, and high sensitivity to the changes in MUT’s dielectric constant.

II. DIELECTRIC CHARACTERIZATION MECHANISM

Fig. 1 shows a microstrip CLDC where the MUT is placed on top of the coupled-line section. For the sake of simplifying the analysis, coupled microstrip lines are sometimes assumed to operate in the transverse electromagnetic (TEM) mode, which is approximately (not rigorously) valid [54]. By such an assumption and applying an even–odd mode analysis, $|S_{31}|$ is given as [54]

$$|S_{31}| = \frac{C|\tan \theta|}{\sqrt{1 - C^2 + j \tan \theta}}$$

where $\theta = \beta_{\text{eff}} L = \beta_0 (\varepsilon_r^{\text{eff}})^{1/2} L$ is the electrical length of the coupled lines, $\beta_{\text{eff}}$ is the effective propagation constant in the microstrip medium, $L$ is the physical length of the coupled lines, $\beta_0$ is the propagation constant in free space, $\varepsilon_r^{\text{eff}}$ is the effective dielectric constant of the microstrip medium, and $C$ is the maximum coupling coefficient (which is equal to $|S_{31}|$ for $\theta = \pi/2$). $|S_{31}|$ of a coupler (with $C = -10$ dB) in terms of $\theta$ is plotted in Fig. 2 showing that $S_{31} = 0$ for $\theta = \pi$ and for all integers that are multiples of $\pi$. For deriving (1), it was assumed that the even and odd modes of the coupled line structure have the same velocities of propagation (i.e., the same effective propagation constant), so that the line has the same electrical length for both the modes. For coupled microstrip lines not supporting a pure TEM mode, however, this condition is not satisfied exactly, and the propagation velocity (and consequently the propagation constant and electrical length) in even and odd modes is slightly different. In fact, in (1), $\theta$ (and $\beta_{\text{eff}}$) is the average of the electrical lengths (and effective propagation constants) of the two modes.

Since in microstrip circuits the electric field exists in both the substrate and the medium above the signal strip (namely superstrate), $\varepsilon_r^{\text{eff}}$ depends on both the dielectric constants of the substrate and superstrate (while having a value between them) [54]. Therefore, by placing different MUTs on the
coupled-line section, $\varepsilon_r^{\text{eff}}$ and consequently $\theta$ vary. As a result, the frequency at which $S_{11} = 0$ (i.e., $\theta = \pi$) changes. Hence, this frequency namely zero-coupling frequency ($f_0$) can be used to retrieve the dielectric constant of the MUT. It should be noted the microstrip line is a dispersive TL above a specific frequency which is typically in the 3–10 GHz range. For frequencies higher than this specific frequency, the effective dielectric constant starts approaching the substrate’s dielectric constant in a way that in high frequencies, the medium above the microstrip line is uniform and the operating mode will be TEM (not quasi-TEM), so in that case, $S_{11}$ will be very low (close to zero). In Fig. 4, $|S_{11}|$ at 2 GHz is plotted versus $\varepsilon_r^{\text{MUT}}$. In Fig. 4, it is clear that $|S_{11}|$ is not a monotonic function, making the unique extraction of $\varepsilon_r^{\text{MUT}}$ impossible, because every value of $|S_{11}|$ between $-43$ and $-22$ dB is mapped to two values of $\varepsilon_r^{\text{MUT}}$. To resolve this problem, the design of the coupler needs modifications so that $S_{41} \approx 0$ occurs for $\varepsilon_r^{\text{MUT}} = 1$ instead of $\varepsilon_r^{\text{MUT}} = \varepsilon_r^{\text{Sub}}$. In other words, a coupler with a high isolation ($S_{41} \approx 0$) and consequently a high directivity in the unloaded state is required.

It should be noted that by increasing the dielectric constant of the MUT, the change in the isolation level ($|S_{41}|$) of the coupler would be significantly greater than the change in the transmission coefficient ($|S_{21}|$) of a microstrip line when loaded with the same MUT. In Fig. 5, the simulated transmission coefficients ($|S_{21}|$) of a 50-$\Omega$ microstrip line on the same substrate are given for different $\varepsilon_r^{\text{MUT}}$. The sample size of the MUT is assumed to be the same as the sample size in the coupler. By comparing Fig. 3(b) with Fig. 5, it can be observed that by increasing $\varepsilon_r^{\text{MUT}}$ from 1 to 10, the change in the coupler’s isolation level ($|S_{41}|$) is a few tens of dB, while the maximum change in the microstrip line’s transmission

![Fig. 3](image-url) Simulated (a) $|S_{11}|$ and (b) $|S_{41}|$ of the microstrip CLDC shown in Fig. 1 when loaded with different MUTs.

![Fig. 4](image-url) Isolation level ($|S_{41}|$) of the coupler shown in Fig. 1 at 2 GHz as a function of $\varepsilon_r^{\text{MUT}}$. Fig. 3(b) shows that the change in $\varepsilon_r^{\text{MUT}}$ which results in the change in the phase velocity in the medium above the coupled microstrip lines has a significant impact on the isolation level ($|S_{41}|$) of the coupler. It should be noted that by assuming a TEM mode and applying an even–odd mode analysis, $|S_{41}|$ of a coupled-line coupler is analytically calculated as zero (for every $\theta$). However, in microstrip CLDC operating in a quasi-TEM mode (not pure TEM), the difference between the propagation velocity (and consequently the propagation constant and electrical length) in the even and odd modes results in a nonzero $|S_{41}|$.

The key idea in our method is to use $|S_{41}|$ to detect $\varepsilon_r^{\text{MUT}}$. But as shown in Fig. 3(b), by increasing $\varepsilon_r^{\text{MUT}}$, the trend of $|S_{41}|$ is not monotonically increasing or decreasing. It can be seen that by increasing $\varepsilon_r^{\text{MUT}}$ from 1 to 4 (which is close to the dielectric constant of the substrate), $|S_{11}|$ decreases. However, by increasing $\varepsilon_r^{\text{MUT}}$ from 4 to 10, $|S_{41}|$ increases. In fact, for $\varepsilon_r^{\text{MUT}}$ equal to $\varepsilon_r^{\text{Sub}} = 3.55$, the medium below and above the microstrip line is uniform and the operating mode will be TEM (not quasi-TEM), so in that case, $|S_{41}|$ will be very low (close to zero). In Fig. 4, $|S_{11}|$ at 2 GHz is plotted versus $\varepsilon_r^{\text{MUT}}$. In Fig. 4, it is clear that $|S_{41}|$ is not a monotonic function, making the unique extraction of $\varepsilon_r^{\text{MUT}}$ impossible, because every value of $|S_{41}|$ between $-43$ and $-22$ dB is mapped to two values of $\varepsilon_r^{\text{MUT}}$. To resolve this problem, the design of the coupler needs modifications so that $S_{41} \approx 0$ occurs for $\varepsilon_r^{\text{MUT}} = 1$ instead of $\varepsilon_r^{\text{MUT}} = \varepsilon_r^{\text{Sub}}$. In other words, a coupler with a high isolation ($S_{41} \approx 0$) and consequently a high directivity in the unloaded state is required.
level ($|S_{21}|$) is about 0.2 dB. This comparison shows the high sensitivity of using the isolation of a coupler for sensing $\varepsilon_r$. While in previous simulations the MUTs were assumed lossless, the effect of MUT losses on the sensor’s responses ($|S_{31}|$ and $|S_{41}|$) should be investigated. If the change in the loss tangent (which is equivalent to the conductivity) of the MUT (with the same dielectric constant) results in a considerable change in $|S_{31}|$ or $|S_{41}|$, it means an undesired cross-sensitivity in the sensor response which may lead to an inaccurate retrieval of $\varepsilon_r$. Fig. 6 shows the simulated $|S_{31}|$ and $|S_{41}|$ of the coupler for different MUTs having different values of conductivity between $\sigma = 0$ (loss-less material) and $\sigma = 6.7 \times 10^{-2}$ S/m (equivalent to a loss tangent of 0.1 at 2 GHz), with the same dielectric constant of $\varepsilon_r = 6$. It can be observed that the dielectric loss of the MUT does not have considerable impact on the zero-coupling frequency ($f_0$) and $|S_{41}|$. It should be noted that similar results were observed for other values of $\varepsilon_r$. Therefore, the proposed sensor has negligible cross-sensitivity to MUT’s loss. However, the insensitivity of the responses to MUT’s loss prevents characterizing the complex permittivity of the MUT.

### III. PROOF OF CONCEPT

As mentioned in Section II, to use $|S_{41}|$ to detect $\varepsilon_r$, a high-directivity coupler (with near zero $|S_{31}|$ in the unloaded state) is required, while a conventional microstrip CLDC, as shown in Section II, has a low directivity (and isolation) due to the inequality of the phase velocities in the substrate and the medium above the strips. One way to improve the directivity of a CLDC is to load two capacitors between the coupled strips at their ends [56], [57]. In the low gigahertz range, the two stubs at the ends of the coupled-line section, as shown in Fig. 7(a), can play the role of the capacitors between the strips. Fig. 7(a), the substrate is 1.6-mm-thick Rogers RO4003C laminate, and the length and width of the lines and the distance between them were adjusted, as given in the caption of Fig. 7(a), so that the ports are well-matched to 50 $\Omega$ and the maximum value of $|S_{31}|$ equals $C \approx -18$ dB at around 1.4 GHz. This value of the coupling coefficient was chosen merely as an example, and couplers with more tightly or more loosely coupled lines are applicable as well.

#### A. Application to Solid MUTs

The fabricated coupler was measured using an SNA. Then, as shown in Fig. 7(b), by putting different standard laminates, including Rogers RT/duroid 5880, Rogers RO4003C, FR4, and Rogers AD1000 having dielectric constants of 2.2, 3.55, 4.4, and 10.2, respectively, as MUT on the coupled-line section [sensing area in Fig. 7(a)], the coupler’s $|S_{31}|$ and $|S_{41}|$ were measured. The results are shown in Fig. 8 and are compared with the full-wave simulation results. Strong agreement between the simulations and the measurements is observed except for $|S_{41}|$ in the unloaded state. In fact, the discrepancy between the simulated and the measured $|S_{41}|$, in this case, is due to a very small $|S_{41}|$, which, in practice, can be affected by the reflection from nonideal ports. Nevertheless, the measured (as well as simulated) $|S_{41}|$ demonstrates that
Fig. 8. Simulated and measured results of the proposed sensor (high isolation CLDC shown in Fig. 7) for different values of \( \varepsilon_{\text{MUT}} \): (a) coupling coefficient (|\( S_{31} \)|) and (b) isolation coefficient (|\( S_{41} \)|).

using capacitive stubs led to a high level of isolation (around −40 dB) in the unloaded state.

We emphasize that an air gap between the coupler and the MUT is inevitable. To minimize this gap, as shown in Fig. 7(b), four Teflon screws were used. However, the air gap can never be completely eliminated, so that the simulation and measurement results can be further matched together by assuming an appropriate amount of a gap in the simulations. The dotted lines in Fig. 8 show the simulation results by considering a 15-μm air gap between the coupler and the MUT. The slight difference between the simulated and measured results, even when assuming a gap in the simulations, turns out to be due to the fact that we assumed the same gap for different MUTs in the simulations. But in practical implementation, we cannot achieve the same gap when changing the samples because of the surface roughness of the sample under test and the main board in the measurement setup. These discrepancies can also be attributed to the reflection from nonideal ports, fabrication inaccuracies, and measurement errors. In addition, our simulations and measurements show that if the thickness of the MUT is greater than about 1 mm, the change in MUT’s thickness will not affect the responses due to the negligible strength of the fields far away from the coupler.

As shown in Fig. 8(a), by increasing \( \varepsilon_{\text{MUT}} \), the zero-coupling frequency (\( f_0 \) at which |\( S_{31} \)| is minimum) decreases. Moreover and importantly, while in a conventional CLDC, as presented in Section II, the trend of |\( S_{41} \)| versus \( \varepsilon_{\text{MUT}} \) was not monotonically increasing or decreasing, Fig. 8(b) demonstrates that using the high-directivity coupler shown in Fig. 7, |\( S_{41} \)| increases monotonically by increasing \( \varepsilon_{\text{MUT}} \), making the unique retrieval of \( \varepsilon_{\text{MUT}} \) possible. The measurement results of \( f_0 \) and |\( S_{41} \)| at 2 GHz, for different MUTs, are listed in Table I. It is seen that by increasing \( \varepsilon_{\text{MUT}} \) from 1 (unloaded) to 10.2 (AD1000), \( f_0 \) and |\( S_{41} \)| decrease and increase by about 42% and 30 dB, respectively. Therefore, each of these two parameters can be used to retrieve \( \varepsilon_{\text{MUT}} \) with a high sensitivity.

To derive closed-form expressions of \( \varepsilon_{\text{MUT}} \) in terms of \( f_0 \) and |\( S_{41} \)|, by setting the data of RO4003C, which is considered as the test data, aside, a curve-fitting method was applied to the data of the other four MUTs (i.e., air, RT5880, FR4, and AD1000) given in Table I. The curve-fitting

| MUT             | \( \varepsilon_{\text{MUT}} \) | \( f_0 \) (GHz) | \( |S_{41}| \) (dB) at 2GHz |
|-----------------|-----------------|-----------------|-----------------|
| Air (Unloaded)  | 1               | 2.77            | -37.9           |
| RT/duroid 5880 | 2.2             | 2.39            | -28.7           |
| RO4003C        | 3.55            | 2.17            | -20.4           |
| FR4            | 4.6             | 2.04            | -16.9           |
| AD1000         | 10.2            | 1.62            | -8.0            |

| MUT     | Reference value | Retrieved value using \( f_0 \) | Retrieved value using |\( S_{41} \)| |
|---------|-----------------|-------------------------------|-----------------------|
| RO4003C | 3.55            | 3.51 (1.1 %)*                  | 3.50 (1.4 %)*         |

*The numbers in the parentheses show the relative errors.
resulted in

\[ \varepsilon_r^{\text{MUT}} = a_3(f_0)^3 + a_2(f_0)^2 + a_1(f_0) + a_0 \] (2)
\[ \varepsilon_r^{\text{MUT}} = b_3(S_{41})^3 + b_2(S_{41})^2 + b_1(S_{41}) + b_0 \] (3)

where \( a_3 = -2.91, a_2 = 2.60 \times 10^4, a_1 = -7.92 \times 10^3, a_0 = 8.26 \times 10^3, b_3 = 5.72 \times 10^{-4}, b_2 = 5.12 \times 10^{-2}, b_1 = 1.63, b_0 = 2.02 \times 10^3, \) and \( f_0 \) and \( |S_{41}| \) are in GHz and dB scales. Fig. 9(a) and (b) shows \( \varepsilon_r^{\text{MUT}} \) as functions of the zero-coupling frequency \( (f_0) \) and the isolation level \( (|S_{41}|) \), respectively. In Fig. 9(a) and (b), a good agreement between the measured results and the fitted curves is observed. It should be noted that either (2) [i.e., Fig. 9(a)] or (3) [i.e., Fig. 9(b)] can be used for the retrieval of \( \varepsilon_r^{\text{MUT}} \) alternatively; in other words, measuring one of \( |S_{31}| \) and \( |S_{41}| \) is sufficient. It is noticeable that although for retrieving \( \varepsilon_r^{\text{MUT}} \) using the isolation level we used \( |S_{41}| \) at 2 GHz, its value at other frequencies can be used. The reason that we used 2 GHz is that with varying \( \varepsilon_r^{\text{MUT}} \) from 1 to 10.2, the change in \( |S_{41}| \) at 2 GHz is more extensive than other frequencies; thus, the highest possible sensitivity is achieved.

To assess the accuracy of the proposed retrieval methods, the measured data of RO4003C, which was omitted from the curve-fitting process, were used. The retrieved dielectric constant from (2) and (3) using zero-coupling frequency and isolation level, respectively, is given in Table II. Table II also shows the relative error of each method, calculated as

\[ \text{Error} = \frac{|\varepsilon_{\text{retrieved}} - \varepsilon_{\text{reference}}|}{\varepsilon_{\text{reference}}} \] (4)

where \( \varepsilon_{\text{retrieved}} \) and \( \varepsilon_{\text{reference}} \) are the retrieved and true (reference) values of the dielectric constant, respectively. It is seen that the relative retrieval error in both the methods is very small, around 1%.

It is worth mentioning that we do not rely on the simulation data in the retrieval process. But we first calibrate the sensor with some known samples and extract the governing formula [i.e., (2) or (3)]. Then, the sensor can be used to retrieve the dielectric constant of unknown MUTs.

**B. Application to Liquid MUTs**

To equip the sensor with the means of measuring the dielectric constant of liquids, a set of walls, built from Rogers RO4003C with a height of about 4 mm, were added to the structure to form a small pool around the sensing area as shown in Fig. 10(a). Before testing liquids, to consider the effect of the dielectric walls on the coupler’s response, the measured \( |S_{31}| \) and \( |S_{41}| \) of the coupler when loaded with the Rogers RT/duroid 5880 with and without walls are compared, and the results are shown in Fig. 10(b). It can be observed that the dielectric walls do not have a considerable impact on the sensor’s responses.
To show the applicability of the proposed methods to liquids, diesel fuel and chloroform were used as the MUT. The simulated and measured \( |S_{31}| \) and \( |S_{41}| \) of the coupler loaded with diesel fuel and chloroform are shown in Fig. 11(a) and (b), respectively. In the simulations, the dielectric constant of diesel fuel and chloroform was set to 2.05 and 4.75, respectively, as reference values, obtained by performing a set of VNA measurements and applying the well-known Nicolson–Ross–Weir (NRW) method [58], [59]. A good agreement is observed between the measured and simulated results demonstrating that both the proposed methods (which use coupler’s zero-coupling frequency and isolation level) are applicable to retrieve the dielectric constants of liquids. However, it is essential to note that the formulas obtained earlier [i.e., (2) and (3)] are valid only for solids and should not be relied upon for liquids since a liquid MUT does not leave a gap between the sensor and itself. In other words, to retrieve the dielectric constant of unknown liquids, a set of new calibrating measurements with some known liquid samples whose reference dielectric constants are at hand are necessary. Then, equations similar to (2) and (3) can be derived for retrieving the dielectric constants of liquids. It should also be noted that due to the direct contact between the liquid samples and the coupler, a cleaning procedure is required after each measurement.

C. Comparison to the Literature

A detailed comparison between the characteristics and performance of the proposed methods and other prior published works is provided in Tables III and IV. It should be noted that two different methods, both based on a microstrip CLDC, were presented in this article. The method, which uses zero-coupling frequency, is a resonant method, while the one using the isolation level is a nonresonant method. So, the former is compared with other resonant methods, whereas the latter is compared with other similar nonresonant methods which use only the amplitude of a transmission coefficient, as in our method using only the amplitude of \( S_{41} \).

Table III provides a comparison between the resonant sensors based on a microstrip resonator. While the earlier works used the zero-transmission frequency (i.e., the frequency at which the transmission coefficient of a two-port circuit is almost zero), in our work, zero-coupling frequency is used. In such resonant methods, since the change in the dielectric constant of the MUT is sensed by the change in the resonance frequency, when loaded with a specific sample, the higher the resonance frequency shift the loaded sensor exhibits with respect to the unloaded one, the higher the sensitivity of the sensor [41]. Therefore, the normalized resonance frequency shift of the sensor when loaded with a specific sample (with respect to the unloaded state) can be a measure of the sensitivity. Since FR4 has been commonly used as MUT in the literature, in Table III, the works giving the results for this material are listed to perform a comparison between the shift in the resonant frequency when the sensor is loaded with this material with respect to the unloaded sensor. The measured unloaded \( (f_{0}^{\text{unloaded}}) \) and loaded \( (f_{0}^{\text{FR4}}) \) resonant frequencies are given in Table III, and the normalized resonant frequency shift is calculated as

\[
\Delta f_0 = \frac{f_{0}^{\text{unloaded}} - f_{0}^{\text{FR4}}}{f_{0}^{\text{unloaded}}}. \tag{5}
\]

It is observed from Table III that the normalized resonant frequency shift, equivalently the sensitivity, in this work is considerably higher than previously reported works.

The proposed method based on the isolation level is a kind of (nonresonant) amplitude-only transmission method since it only requires the amplitude of \( S_{41} \) and the transmission coefficient between the input and isolated ports. In Table IV, a qualitative comparison between different transmission media that can be applied in the amplitude-only transmission methods is provided. Among them, microstrip TL and microstrip CLDC used in our work, with a planar structure, have the advantage
of low-cost fabrication. Also, when using these microstrip-based structures, the preparation of MUT is quite easy, while in the case of metallic waveguide and coaxial line, the sample needs to be cut and drilled precisely in a specific manner.

In the amplitude-only transmission methods, since the change in the dielectric constant of the MUT is measured by the change in the amplitude of a transmission coefficient, the higher the change in this parameter, the higher the sensitivity. Therefore, to quantify the comparison between the sensitivity of different approaches, the change in the transmission/isolation level when the sensor is loaded with two specific MUTs (with dielectric constants of 3 and 4) is considered as the measure of sensitivity as

\[ \Delta T = S_{ij} (\varepsilon_r^{\text{MUT}} = 4) - S_{ij} (\varepsilon_r^{\text{MUT}} = 3) \]  

(6)

where \( S_{ij} \) is the transmission/isolation level in dB scale. By modeling a 50-\( \Omega \) microstrip line on a Rogers RO4003C substrate with a thickness of 1.6 mm, a coaxial line with an inner and outer conductor with diameters of 2.5 and 5.8 mm, respectively (having a characteristic impedance of 50 \( \Omega \) in the unloaded state), and a WR430 rectangular waveguide with an operating frequency of 1.7–2.6 GHz in the full-wave simulator, \( \Delta T \) were obtained for different structures indicated in Table IV which is shown in Fig. 12. It should be noted that a gap of 15 \( \mu \)m was considered between the MUT and the substrate when simulating the microstrip TL and CLDC.

As can be seen in Fig. 12, when using a microstrip CLDC, \( \Delta T \) (which is the change in the isolation/transmission level) is noticeably higher than when a microstrip TL, coaxial line, and waveguide are applied, which is identical to the higher sensitivity of the proposed retrieval method based on CLDC.

### IV. CONCLUSION

Two alternative methods using a microstrip CLDC were proposed to retrieve the dielectric constant of the MUT, one using the zero-coupling frequency (frequency at which \( S_{31} \approx 0 \)) and the other using the isolation level (\( |S_{41}| \)) at a specific frequency. The proposed methods were based on the change in the effective dielectric constant of the microstrip medium and the phase velocity in the medium above the microstrip line when placing different MUTs on the coupled-line section which leads to significant changes in coupler’s coupling and isolation levels.

It was shown that for unique retrieval of MUT’s dielectric constant, the coupler should provide a high directivity (\( S_{41} \approx 0 \)) in the unloaded state, while a conventional microstrip coupled-line coupler does not. By loading two capacitive stubs to the coupled lines, a high-isolation coupler was designed and fabricated. \( |S_{11}| \) and \( |S_{41}| \) of the coupler were measured when different solid and liquid samples were placed on it. The measurement results, which were in strong agreement with full-wave simulations, validated the proposed method and demonstrated the high sensitivity of the proposed methods to the change in the dielectric constant of the MUT in such a way that by increasing \( \varepsilon_r^{\text{MUT}} \) from 1 to 10.2, \( f_0 \) and \( |S_{41}| \) change by approximately 42% and 30 dB, respectively. Comparison with similar previous works shows a superior sensitivity in the proposed methods.

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