Edited by
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PROCEEDINGS OF THE 16TH INTERNATIONAL STUDENT SEMINAR “MICROWAVE AND OPTICAL APPLICATIONS OF NOVEL PHENOMENA AND TECHNOLOGIES”, JUNE 8–9, OULU, FINLAND
Abstract

The present volume contains 9 selected papers from the 16th International Student Seminar “Microwave and optical applications of novel phenomena and technologies”, June 8–9, Oulu, Finland. The authors are young researchers and post-graduate students. Their works reflect the variety of phenomena, models, technologies, and materials currently studied and employed for high-frequency applications in microwave to optical range. Especial emphasis is made on design and miniaturization of microwave components and circuits.

Keywords: design, microwave, technology
Preface

Dear Reader,

In this issue You may see the papers of participants in the 16th International Student Seminar “Microwave and optical applications of novel phenomena and technologies” (ISS-16).

The first International Student Seminar called “Seminar on High-Temperature Superconductors at Microwaves” was held in 1994 at Chalmers University of Technology in the city of Göteborg (Sweden) with support of Professor E. Kollberg. Plenty of exciting discoveries and developments have occurred since then. The content of the annual Seminars was reflecting this progress. Greatly inspired by Prof. O. G. Vendik and Prof. I. B. Vendik, seven of the Seminars were organized in St. Petersburg, Russia. Being internationally acknowledged, the Seminars were arranged in Sweden, France, Germany, UK, and Finland.

In 2009, the scope of ISS-16 covers materials, metamaterials, devices, systems, and technologies for high-frequency applications in microwave to optical spectral range. The Seminar presentations deal with development trends in telecommunication systems, novel design and technological approaches at microwaves, development of advanced tunable components for steerable antenna arrays, metamaterials, ferro- and magneto-electrics for THz applications, and methods of microwave spectroscopy. The selected presentations of the ISS-16 are included in this volume.

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On-orbit Ka-band switch matrix experiment

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Abstract

We have developed a remotely operating switch-matrix Ka-band experiment that will participate in a German satellite mission for on-orbit verification. Estimating the function of the switch matrix during the space mission requires peripheral modules such as detectors, voltage controlled oscillators, and power detectors. The components of the space experiment have been developed using the low-temperature cofired ceramics technology as separate modules, to enable a modular design for future satellite applications. A description of the payload functions and assembly techniques as well as the microwave performance of the components are presented.
1 Introduction

The low-temperature cofired ceramic (LTCC) technology offers three-dimensional design capabilities, hybrid integration, hermetic packaging, low fabrication costs as well as low microwave losses. A wide range of applications takes advantage of these aspects [1], [2]. Ilmenau University of Technology as a part of an industrial-academic consortium [3] have developed a remote operational space experiment that will be launched in 2010 as part of a German test satellite mission. Therefore, all components had to be developed considering requirements such as mechanical and thermal robustness, power consumption and heat load, as well as functional redundancy.
2 LTCC multilayer modules

Independent of the function of the ceramic modules, the size, the materials, and the assembly technologies determine the mechanical and thermal behaviour of the whole system. First qualification tests showed that aluminium as carrier material is unsuitable, in contrast to Kovar [4]. Mainly because of the smaller Young’s modulus, a warpage of the board is caused, potentially causing cracks through the rigid ceramic modules. Still, the remaining differences in thermal expansion coefficient between Kovar and LTCC are critical and would prohibit typical assembly techniques, without the use of an intermediate material.

The ceramic modules are mounted on a four-layer polymer substrate RT/duriod 6002® on which the digital signals and supply voltages are distributed. This soft substrate, serving at the same time as buffer to accommodate the different thermal expansions, is glued onto a gold-plated kovar board [4].

The switch matrix uses eight commercial single-pole four-throw (SP4T) switches resulting in a 4 x 4 topology, i.e., any one of four microwave input ports can be reconfigured to any one of four output ports in the Ka-band. Configuration is obtained by applying 32 carefully adjusted bipolar currents. A small size of the module of 34 mm × 32 mm is achieved by mounting the ICs in cavities on opposite sides of the module [5], [6]. One diode driver module (see Fig. 2) delivers 16 positive or negative currents, depending on the digital inputs, whose magnitude is controlled by appropriately sized resistors and capacitors.
Fig. 2. Diode driver module with the digital inputs on the left-hand side and the current outputs on the right-hand side. The grey and blue shaded patches represent fixed-element resistors and capacitors. The supply voltages are connected via three contact pads on the top of the module. Bondwires are used to connect this module to the carrier board.

The voltage-controlled oscillator (VCO) module depicted in Fig. 3 includes two microwave signal sources covering different frequency ranges, and detectors to monitor the power fed to the switch matrix. The differential analog detector voltage is amplified and converted to single-ended using an operational amplifier. The microwave structures are placed on the top layer, in order to facilitate functional tests and optical inspection.

Fig. 3. VCO-module employing two different materials on the top layer: silver paste for good solderability and gold paste for bond wire adhesion. Coplanar microwave ports are located on the left-hand side, contact pads for supply and tuning voltage on the top side. The coloured patches represent the footprints for components like operational amplifiers, resistors and capacitors.

The outputs of the switch matrix are connected to the detector modules, according to the photograph displayed in Fig. 4. The operational frequency range covers the
output frequencies of the VCOs (see Sec. 3 below). An operational amplifier is used to amplify the detector voltage to a level of about 1 V, to optimise the dynamic range.

Fig. 4. Photograph of the laboratory version of the flight model (FM). The numbers label the FPGA (5) and the LTCC-modules: 1 - switch matrix, 2 - VCO-module, 3 - diode drivers, 4 - detectors. Bond ribbons are used for the microwave connections between the LTCC modules. The inset displays the backside of the switch matrix through an aperture in the board.
3 Microwave Performance

The VCO modules deliver two separate microwave outputs at frequencies between 23.8 - 24.8 GHz and 17.2 - 19.2 GHz with power levels of +12 dBm and +2 dBm, respectively. Measuring the voltage output of the detectors of the VCO-module enables the estimation of the output power. Detectors at the output ports of the switch matrix allow for the calculation of the insertion losses of the paths switched on, as well as a comparison with reference measurements for different paths and environmental conditions. Fig. 5 represents typical S-parameters measured for a complete path incorporating two switch ICs and a total length of connecting transmission lines of about 5 mm.

![S-parameter plot](image)

Fig. 5. Measured scattering parameters of a typical complete I/O path through the switch matrix. The optimised microwave design covers an operational frequency range between 17 and 22 GHz, while the insertion loss remains between 4.5 and 6 dB.

The performance of the switch matrix was found to exhibit a strong dependence on the currents applied to the PIN-diodes. The trade-off between low insertion loss and low power consumption led to a current of 5 mA [7].

During the space experiment, automated measurements of the detector voltages build the basis for data acquisition and subsequent off-line analysis. The estimation of the power delivered to the input ports, and from the output ports, from the detector voltages can be carried out using either detector curves or approximation formulas. To account for the influence of the supply voltage, the
temperature, incident microwave power, and frequency, a formula is suited best for a reliable analysis and reduces the amount of reference measurements [8].

Fig. 6 illustrates first measurement results for the detector voltage as a function of the incident microwave power and frequency. As expected, the detector voltages vary with power with a high sensitivity, while the variation with frequency is negligible.

![Figure 6](image)

**Fig. 6.** Measured detector voltage as a function of incident microwave power and frequency (only a part of the whole characteristic is shown).

During the space mission, two identical boards will be used, to compare the results and to achieve redundancy. Taking advantage of the reconfigurable switch matrix in combination with different microwave sources with tuneable outputs enables to access a wide range of measurement setups. The measured data are transmitted to ground via a satellite communication link and distributed to the authors for evaluation.
4 Conclusions

We have developed a remote operational switch-matrix experiment for space applications. A number of different LTCC modules were successfully developed with special care on robustness, as proven by mechanical and thermal qualification tests. The results are based on theoretical and practical investigations gained during earlier qualification tests and preparations towards the actual payload module. A successful on-orbit verification will pave the way for realistic applications of compact LTCC microwave modules for satellite communications.

Acknowledgement

This work has been funded by the German Federal Ministry of Economics and Technology (BMWi) under the project management by the German Aerospace Center (DLR, no. 50YB0622). We appreciate very much helpful scientific and technical support from A. Schwarz (RHe Microsystems GmbH, Radeberg) and D. Schwanke (Micro Systems Engineering GmbH, Berg) as well as M. Huhn, I. Koch, J. Müller, D. Stöpel, J. Trabert, G. Vogt, and M. Zocher (Ilmenau University of Technology).
References

The Dual-band and Reconfigurable Wilkinson Power Dividers Based on Metamaterial Transmission Lines

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Abstract

An application of metamaterial transmission lines to design novel types of the Wilkinson power dividers with enhanced functionality is described. Designs of the dual-band and reconfigurable Wilkinson power dividers and results of electromagnetic simulation of their characteristics are presented.
1 Introduction

Transmission lines (TLs) with negative dispersion are considered as 1D metamaterials and known as left-handed (LH) TLs [1]. In turn conventional TLs characterizing by positive dispersion are referred to as right-handed (RH) TLs. Using RH and LH TL sections in a combination leads to higher design flexibility. Different microwave devices on a combination of RH and LH TLs with improved performance and enhanced functionality have been recently developed: i) broadband devices [2]-[4]; ii) miniaturized devices [5], [6]; iii) multiband devices [7]-[9].

The Wilkinson power divider is a very popular microwave device used in various power distribution networks. Design of the Wilkinson power dividers with enhanced performance seems very attractive for application in modern communication systems.

This paper presents novel designs of the Wilkinson power dividers with enhanced functionality, which are based on a combination of conventional and metamaterial TLs. A dual-band power divider that can be designed for arbitrary chosen operational frequencies and a reconfigurable power divider providing either 0° or 180° phase difference between the output signals are considered.
2 The dual-band Wilkinson power divider

The classical Wilkinson power divider comprises two quarter-wavelength TL sections (Fig. 1) Due to the periodic resonant properties of these TL sections, the higher harmonic frequencies \( f_i \) of the divider are related to the fundamental frequency \( f_0 \) as \( f_i = (2i-1) f_0 \), where \( i = 1, 2, 3, \ldots \) In order to shift the higher operational frequency band of the Wilkinson power divider to a desired frequency \( f_n \), it is necessary to change the dispersion law of the TL sections. For this purpose a LH TL section is incorporated into each branch of the power divider. The overall phase incursion along a cascade of RH and LH TL sections with the same characteristic impedance \( Z_0 \) is \( \varphi(f) = \varphi_L(f) + \varphi_R(f) \). The values of electrical lengths of RH and LH TL sections have to be adjusted to keep the overall phase incursion along the branch equal to \(+90^\circ\) or \(-90^\circ\) at the frequencies \( f_0 \) and \( f_n > f_0 \), which can be chosen arbitrary. Particularly, the first higher harmonic can be shifted to the frequency \( f_1 = 2f_0 \).

Fig. 1. Schematic diagram of the classical Wilkinson power divider.

The dispersion law is linear for the RH TL: \( \varphi_R(f) = \varphi_R(f_0) \cdot f/f_0 \) and is nonlinear in the case of the LH TL: \( \varphi_L(f) = \varphi_L(f_0) \cdot f_0/f \) as illustrated by Fig. 2.a. Taking into account that the overall phase incursions along the branch have to be equal \( \varphi_R(f_0) = \varphi_L(f_1) = +90^\circ \) or \(-90^\circ\) at the both operational frequencies \( f_0 \) and \( f_1 \) one can easily obtain the electrical length of the RH and LH TL sections: \( \theta_{RH} = 90^\circ\) and \( \theta_{LH} = -180^\circ \).
Fig. 2. a) Phase diagram of the quarter-wavelength RH and LH TL sections and their cascade. b) Equivalent circuit of the dual-band Wilkinson power divider designed for the operational frequencies \( f_0 \) and \( f_1 = 2f_0 \).

An LH TL section is realized in artificial way as a cascade of T- or \( \Pi \)-cells based on lumped LC-elements. At the same time a RH TL section can be implemented as a natural distributed TL or as an artificial one. An equivalent circuit of the dual-band Wilkinson power divider under consideration is shown in Fig. 2.b. The quarter-wavelength RH TL section of each branch has been divided into four sections with the electrical length of \( \theta_{RH} = 22.5^\circ \). The LH TL sections have been designed as three T-cells with the equivalent electrical length of \( \theta_{LH} = -60^\circ \) each. The values of LC-elements have been calculated by comparison of the ABCD matrixes of the TL section and the T-cell.

The demonstrator of the dual-band Wilkinson power divider for operational frequencies \( f_0 = 0.9 \) GHz and \( f_1 = 1.8 \) GHz has been designed. The Rogers RO3010\textsuperscript{TM} (\( \varepsilon_r = 10.2 \)) printed circuit board (PCB) with the thickness of 1.24 mm has been used as substrate material (Fig. 3.a). The RH TL sections have been designed as microstrips. Commercial SMD components have been used to implement the artificial LH TL sections.

The performance of the dual-band Wilkinson power divider obtained by electromagnetic simulation in Sonnet software is presented in Fig. 3.b. At the both chosen operational frequencies (0.9 GHz and 1.8 GHz) the device provides equal division of the input power between two outputs. The insertion loss in the both operational frequency bands is about 0.3 dB. The input reflection coefficients and the isolation of the output ports from each other are better than 20 dB. The experimental investigation of the manufactured demonstrator is in a progress.
Fig. 3. Photograph of the demonstrator (a) and the simulated performance (b) of the dual-band Wilkinson power divider with the operational frequencies $f_0 = 0.9$ GHz and $f_1 = 1.8$ GHz.
3 The reconfigurable Wilkinson power divider

The classical Wilkinson power divider provides zero phase difference between the output signals (0°). The 180° Wilkinson power divider based on LH and RH TLs has been proposed in [10].

We designed the reconfigurable Wilkinson power divider, which allows provision of either 0° or 180° phase difference between the output signals. The schematic diagram of this device is shown in Fig. 4.a. It is based on two composite RH/LH (CRLH) TLs with tunable capacitors. The controllable CRLH TL sections have been designed to behave at the central frequency as the quarter-wavelength RH TL in one state and to exhibit characteristics of the quarter-wavelength LH TL in another state (after tuning the capacitors). While the first CRLH TL is used as the LH TL, the second one is employed as the RH TL and vice versa. As a result, in the initial state the 180° phase difference between the output signals is provided. After tuning the capacitors of the both CRLH TLs, the equivalent electrical length of the TL connected between the resistors becomes equal to 0°, whereas the both main branches behave as the quarter-wavelength RH TLs. Hence, the output signals are in-phase as for the conventional Wilkinson power divider. The equivalent circuit of this reconfigurable Wilkinson power divider is shown in Fig. 4.b.

Fig. 4. Schematic diagram (a) and equivalent circuit using CRLH TLs (b) of the reconfigurable Wilkinson power divider.

The reconfigurable Wilkinson power divider with the central frequency of 2.5 GHz has been designed using commercial SMD components on the Rogers RO3010™ (ε_r=10.2) PCB with the thickness of 0.625 mm (Fig. 5). The quarter-wavelength RH TL sections have been designed in artificial way using two
cascaded lumped-element Π-cells. Commercial semiconductor varactors MV39003 and MV39002 from MDT [11] have been used to implement variable capacitors $C_s$ and $C_p$ in the CRLH TL.

Fig. 5. Top view of the reconfigurable Wilkinson power divider PCB with SMD components.

Characteristics of the reconfigurable Wilkinson power divider obtained by electromagnetic simulation in Sonnet software are presented in Fig. 6. In the both states the device provides equal power division between output ports with the insertion loss is not higher than 0.4 dB at the central frequency. The input reflection coefficients and the isolation of the output ports are better than 20 dB.
Fig. 6. Simulated characteristics of the reconfigurable Wilkinson power divider in two states.
4 Conclusion

Design of the dual-band Wilkinson power divider for applications at $f_0 = 0.9$ GHz and $f_1 = 1.8$ GHz and of the reconfigurable Wilkinson power divider providing either $0^\circ$ or $180^\circ$ phase difference between the output signals have been presented. The results demonstrate a high potential of using metamaterial TLs to develop novel microwave devices with enhanced functionality for future communication systems.

Acknowledgment

Authors highly appreciate the discussions with Irina Vendik. P. K. appreciates very much a scholarship presented by the President of the Russian Federation. Assistance of S. Humbla and M. A. Hein in manufacturing of the dual-band demonstrator is gratefully acknowledged.
References

Miniaturized Broadband LTCC Directional Coupler Using Right/Left-Handed Transmission Lines

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Abstract

Rat-race ring performance improvement due to the use of artificial left- and right-handed transmission line sections is considered. Different rat-race ring designs based on a combination of these lines are compared regarding efficiency of bandwidth enhancement and size reduction.
1 Introduction

A combination of conventional right-handed (RH) transmission lines (TLs) with positive dispersion and metamaterial left-handed (LH) TLs with negative dispersion allows designing microwave devices with reduced size and broadened operational frequency band [1]. Artificial implementation of the RH and LH TL sections as a cascade of lumped-element T- or Π-unit cells is favorably used for device miniaturization. The Low-Temperature Co-fired Ceramics technology (LTCC) suits well for practical realization of microwave devices based on artificial RH/LH TLs [2].

Rat-race ring (RRR) is a well known type of directional couplers. It is a four-port microwave device, which divides the input power equally or unequally between two outputs while the forth port remains isolated. Phase difference between the signals in output ports can be equal to 0° or 180° depending on which port of the device is employed as the input one. RRRs are widely used in balanced mixers, balanced amplifiers, and other microwave circuits.

A conventional RRR consists of three TL sections with the electrical length of 90° and one section of 270° (Fig. 1). Using distributed TLs leads to a large size of the RRR, especially in low-frequency applications. Moreover, involving the electrically long 270° TL section results in a strong frequency dependence of characteristics that limits the operational frequency band of the device. Replacement of the 270° RH TL by a LH TL with the equivalent electrical length of -90° improves characteristics and widens the operational bandwidth. At the same time, implementation of all LH and RH TL sections in artificial way makes it possible to reduce the size of the device drastically.

![Fig. 1. Schematic diagram of the conventional RRR.](image)

Design of an RRR based on artificial RH and LH TL sections with a special emphasis to its miniaturization was described in [2]. The size of the planar RRR...
realized in LTCC technology was reduced to one eighth of the guided wavelength ($\lambda_g/8$). Meanwhile the operational bandwidth was about 25 %.

In this paper, a design of miniaturized broadband RRR based on artificial RH and LH TL sections is presented. Three different design approaches, which use distributed RH TL sections as well as artificial LH and RH TLs consisting of different number of unit cells, are compared regarding efficiency of bandwidth enhancement and size reduction.
2 Rat-race ring based on RH and LH TL sections

Due to presence of the electrically long 270° TL section, the operational frequency band of the conventional RRR is limited to 33%, if determined at the 20 dB level of both the return loss and the isolation (Fig. 2.a). Phase difference variation over this bandwidth is ±10°.

![Diagram of Rat-race ring characteristics](image)

**Fig. 2.** Theoretical characteristics of RRRs: (a) conventional RRR using ideal RH TLs, (b) RRR based on ideal RH and LH TL sections (solid lines) and on artificial RH and LH TL sections represented by a single Π-cell (dashed lines) and by two cascaded Π-cells (dash-dot lines).

Improvement of the RRR characteristics after replacement of the 270° RH TL by an ideal -90° LH TL is illustrated by solid lines in Fig. 2.b. The LH TL can be realized in practice as an artificial TL only. Implementation of RH TL sections in artificial way makes it possible to reduce the size of the device drastically. Meanwhile, the device characteristics depend strongly on the number of unit cells used per wavelength in artificial RH and LH TL sections. A more flat frequency dependence of the phase difference with respect to the conventional case is
provided even using only one lumped-element T- or Π-cell for every quarter-wavelength TL section. A remarkable extension of the operational bandwidth (up to 50%) can be theoretically achieved by using two cascaded unit cells per quarter-wavelength that makes the characteristics close to ones of the RRR based on the ideal RH or LH TL sections (see Fig. 2.b). The theoretical evaluations were confirmed by the results of electromagnetic (EM) simulations.

Fig. 3 shows the layout and simulated performance of the conventional RRR with the central frequency of 3 GHz. The device was implemented as a microstrip structure on LTCC substrate consisting of 8 layers of DuPont Green Tape™ 951 with $\varepsilon_r = 7.8$ and the thickness of 95 $\mu$m after co-firing. The substrate size is 28 mm × 17 mm. In the frequency band of about 33% the amplitude unbalance between outputs is ±0.7 dB and the phase difference variation is $+6^\circ/-14^\circ$.

![Fig. 3. Conventional RRR designed as a microstrip structure on LTCC substrate: (a) multilayer structure, (b) characteristics obtained by EM simulation.](image)

As the next step, an RRR consisting of three quarter-wavelength microstrip lines and an artificial LH TL section realized as two cascaded Π-cells was designed for the same frequency and on the same substrate (Fig. 4.a). Due to a very compact size of the quasi-lumped-element artificial LH TL, the entire dimension of the RRR decreased to 17.5 mm × 15.5 mm. According to results of EM simulations the operational bandwidth was broadened to 44%. Over this enlarged bandwidth the phase difference variation becomes even smaller ($+0^\circ/-8^\circ$) while the amplitude unbalance remains nearly the same (Fig. 4.b).
Fig. 4. Miniaturized RRR based on artificial LH and RH TL sections implemented as two Π-cells: (a) equivalent circuit, (b) multilayer LTCC structure.
3 Design of miniaturized broadband rat-race ring

In order to minimize the size of RRR structure, all TL sections have to be implemented as artificial ones. In [2] we presented a design of a miniaturized RRR based on artificial RH and LH TL sections implemented as a single lumped-element Π-cell. The area occupied by the device, which was realized using 8 layers of 95 μm thick DuPont Green Tape™ 951 LTCC, is almost 9 times smaller with respect to the conventional RRR. The miniaturized RRR exhibited a good performance in the operational frequency band of about 25% by the results of EM simulations and experimental investigation.

Two unit cells for each quarter-wavelength section of artificial LH and RH TLs can be used, in order to enhance the operational bandwidth of RRR. The equivalent circuit of the RRR based on two cascaded Π-cells is shown in Fig. 5.a. The novel miniaturized RRR was also accomplished on quasi-lumped elements embedded into 8 layers of 95 μm thick DuPont Green Tape™ 951 LTCC as illustrated in Fig. 5.b. Parallel-plate capacitors and two-turn stacked inductors were used. The size of the RRR designed is 8 mm × 10 mm that corresponds to λg/5×λg/4. Characteristics of this miniaturized RRR obtained by EM simulation of the multilayer structure are presented in Fig. 6. Within the frequency band of 47%, the phase difference variation amounts to +2°/-8°, the amplitude unbalance does not exceed ±0.7 dB. The insertion loss at the central frequency is as low as 0.4 dB.

Fig. 5. Simulated characteristics of the miniaturized RRR based on artificial LH and RH TL sections implemented as two Π-cells (Fig. 5).
Hence, the electrical characteristics of the miniaturized RRR based on two cascaded Π-cells are comparable with those of the mixed lumped/distributed design shown in Fig. 4. However, the area occupied by the miniaturized RRR is more than 3 times smaller. As compared with the conventional RRR design (Fig. 3), the size reduction is equal to 6 times, whereas the operational bandwidth is broadened by the factor of 1.5. With respect to the recently reported miniaturized RRR [2], the novel miniaturized RRR based on two cascaded Π-cells has the almost 2 times larger operational bandwidth that is followed by the increasing of the occupied area in 1.5 times only. A comparison of size of four different RRR designs under consideration is shown in Fig. 7. The essential parameters of different RRR designs are listed in the Table below.

![Fig. 6. Size comparison of different RRR designs: (a) conventional RRR design, (b) mixed RRR design using distributed RH TLs and an artificial LH TL, (c) miniaturized broadband RRR design based on RH/LH TLs implemented as two cascaded Π-cells, (d) miniaturized RRR design based on RH/LH TLs implemented as a single Π-cell [2].](image)

<table>
<thead>
<tr>
<th>RRR design</th>
<th>Operational bandwidth, %</th>
<th>Variation of phase difference between output signals, deg.</th>
<th>Amplitude unbalance between output signals, dB</th>
<th>Factor of area reduction (related to the conventional design)</th>
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<tbody>
<tr>
<td>Conventional (distributed)</td>
<td>33</td>
<td>±10 (+6/-14)</td>
<td>±1</td>
<td>1.0</td>
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<tr>
<td>Conventional with LH TL (mixed)</td>
<td>44</td>
<td>±6 (+0/-8)</td>
<td>±0.6</td>
<td>1.8</td>
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<tr>
<td>Miniaturized broadband (lumped on two Π-cells)</td>
<td>47</td>
<td>±7 (+2/-8)</td>
<td>±0.7</td>
<td>6.0</td>
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<tr>
<td>Miniaturized (lumped on a single Π-cell)</td>
<td>25</td>
<td>±5 (+7/-1)</td>
<td>±1</td>
<td>8.8</td>
</tr>
</tbody>
</table>
4 Conclusion

RRR performance can be significantly improved by using a combination of RH and LH TL sections. Among different designs, the RRR based on RH/LH TLs implemented as two cascaded unit cells has been considered as a reasonable trade-off providing the most efficient band broadening and size reduction.

Acknowledgement

The useful discussions with Irina Vendik are highly appreciated.
References


Modular Design Method for the Efficient Simulation of Multilayer Microwave Circuits based on an Element Library

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Abstract

Although the design of extended microwave circuits becomes more and more relevant, e.g., for multilayer modules, it is still a time-consuming task. At the same time, the available number of operational microwave modules is increasing continuously, opening up the possibility to reuse parts of them as sub-modules for future designs. To address this opportunity, a modular design strategy is investigated. The proposed strategy incorporates a design library which is intended to collect existing model expertise in an appropriate database. Additionally, a cascading technique may be applied to develop new multilayer circuits and to derive their microwave performance. For verification of the results, test structures from an ongoing R&D project have been analysed. To achieve and sustain high precision during the design process, simulation requirements have to be met at the port interfaces between different sub-modules using dedicated empirical studies. The design approach is discussed in detail, together with the promising results achieved so far.
1 Introduction

Low temperature co-fired ceramics (LTCC) are an established technology for the cost-effective integration of truly three-dimensional circuitry at microwave frequencies [1]. Within the German R&D project “KERAMIS” (acronym for “ceramic microwave circuits for satellite communications”) compact and hybrid-integrated microwave modules are developed to serve as future components for Ka-band systems in rough environments, like aboard of a multimedia satellite [2].

One of the components designed at Ilmenau University of Technology is a 4x4 reconfigurable switch matrix which incorporates four PIN-diode switches on both sides, as well as highly integrated waveguide structures distributed across six layers of DuPont’s 951 tape substrate [3]. A photograph of the module is shown in Fig. 1. Each switched transmission line path is made up of two PIN-diode switches together with appropriate top-layer matching networks and buried stripline connections.

![Photograph of the switch matrix module](image)

Fig. 1. Photograph of the switch matrix module [4]. A Kovar frame is used for a subsequent hermetic packaging of the bare switch dies. The arrows indicate one of the transmission line paths connecting two switch dies on opposite faces of the module diagonally.

The design of such complex microwave structures is quite challenging from the geometrical as well as from the electrical point of view. In particular for finite-difference time-domain (FDTD) solvers, a nonlinear increase of the computational efforts, like simulation time and memory (RAM), is observed when transmission line structures of increasing length and spatial extent are being investigated [5]. Therefore, the main goal of this work is to make the design of future modules faster and more efficient while remaining flexible enough to fulfil
different specifications, by starting from the models and expertise already accumulated within and beyond the afore described project.
2 Modular Design Methodology

The applied design strategy shown in Fig. 2 incorporates three major parts. At first, models and relevant simulation or measurement data of existing modules are collected and arranged in an easy-to-use database (blue shapes in Fig. 2). Then, a geometrical layout procedure based on three-dimensional CAD (computer aided design) tools is performed to develop new structures (yellow shapes in Fig. 2). By using prepared component models from the database, this can be done very quickly and flexibly, in particular for modules which represent enhancements or modifications of earlier architectures.

Fig. 2. Flow chart of the described LTCC design approach. Main features are the model database (blue shapes), the geometrical circuit layout and optimisation (yellow shapes), and the element-wise electromagnetic simulation together with de-embedding and cascading strategies (red shapes).

To achieve the desired microwave performance is the third aspect of the design process (red shapes in Fig. 2), which normally runs in parallel with the CAD layout optimisation. Continuing the modular concept, the electromagnetic simulation and optimisation shall be carried out element-wise, and thus more efficiently [6]. A simple way to perform the final cascading of the individual
microwave responses is given through the computation of transmission parameters [7]. It turns out that the simulation of expanded elements and a subsequent de-embedding are effective approaches to maintain high precision [5].
3 Verification Example and Results

After successful initial tests with basic coplanar waveguide transitions, the modular design strategy was used to investigate a transmission line path from the switch matrix as indicated by the arrows in Fig. 1. The CAD model of the path containing matching networks at both ends and some via transitions in between, together with the modular decomposition procedure is depicted on the left-hand side of Fig. 3.

Fig. 3. Left-hand panel: Geometrical decomposition of the switch matrix path. At first, distributed transition elements (red) and transmission line elements (green) are identified. Additional electrically short line steps (blue) are inserted at the interfaces if necessary. Right-hand panel: Resulting scattering parameters of the cascaded element-wise simulation (dashed curves) compared with the electromagnetic simulation of the entire path (solid curves).

The path was simulated element-wise by applying the expansion and de-embedding techniques according to the described method. The results of the subsequent cascading process are displayed on the right-hand side of Fig. 3. For comparison, the entire path was simulated separately. With a remaining average error between both simulations of about 5 %, the modular design approach reproduced the microwave performance of the switch matrix path quite well [5]. Furthermore, the cascading technique permitted a reduction of the computation time by a factor of six to ten, depending on the additional usage of parallel computing facilities.
4 Empirical Design Studies and Conclusions

To further study the expansion and de-embedding approach, an empirical analysis of two basic transition elements was carried out. A discrete impedance step and a double via transition were simulated and de-embedded, respectively, using different lengths of feeding lines, as shown on the left-hand side of Fig. 4. The mean deviations between the scattering parameters of structures with increasing lengths of their feeding lines were computed across the entire range of frequencies, with the results depicted on the right-hand side of Fig. 4.

For a double via (red curves) with up to three feeding line elements (each 500µm long), a considerable decrease of the simulation error due to field distortions at the interface ports is observed. Beyond three feeding line elements only numerical variations remained. In case of a line step (blue curves), the results are less clear. Obviously, numerical errors seem to dominate the performance. But due to the short electrical length, the transmission of the step is less significant, and thus a limit of about four feeding line elements is proposed.

Even though their lengths may be limited to comparatively small values, the influence of the feeding lines on the simulation results is not negligible.
Additional studies on error propagation have to be carried out to evaluate the contribution of feeding lines of sub-modules to the performance of cascaded structures. Although promising results have been achieved using the investigated method, further empirical results and derived guidelines are important for the correct application of the cascading technique for future design projects, in particular for automated design approaches which require efficient and reliable design strategies.

Acknowledgment

This work has been funded by the German Federal Ministry of Economics and Technology (BMWi) under the project management by the German Aerospace Center (DLR, 50YB0622).
References

Planar Leaky-Wave Antennas Based on Networks of Loaded Transmission Lines

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Abstract

A two-dimensional (2-D) backward-wave (BW) transmission-line network (TLN) antenna is designed analytically by tuning the dispersion and Bloch impedance to obtain the desired operation of the TLN in a certain frequency range. Functionality of the antenna is inspected with analytical calculations and full-wave simulations. A method based on transmission (ABCD) and scattering parameters is developed to tune the load values of the full-wave model so that the dispersion region with zero phase-propagation coefficient is reasonably smooth. The feeding of a 2-D TLN structure with \( N \times N \) unit cells is studied with one feed at the corner. The input impedance can be estimated well for one-dimensional (1-D) and 2-D antennas in most cases using a circuit simulator.
1 Introduction

Loaded transmission lines have been used to make 1-D periodical leaky-wave antennas (LWAs) by using the $n = -1$ spatial harmonic as the leaky mode [1]. Recently, the unit-cell structure widely used in BW TLNs [2] was found to be beneficial for 1-D LWAs. Namely, with a proper choice of loading elements the so-called balanced condition can be realized, which makes it possible to design LWAs with $n = 0$ spatial harmonic leaky mode, enabling broadside radiation [3].

A planar TLN with the balanced condition is an attractive concept given the possibility of zero phase-propagation coefficient $\beta$ (infinite phase velocity) with a non-zero group velocity. This would, in a sense, produce an aperture the size of the TLN with a uniform phase distribution and broadside radiation. A 2-D TLN antenna was designed in [4] with the feed realized to one edge with power dividers. However, this solution takes space and makes the feeding complicated.

Due to the in-plane isotropy of the 2-D TLN, the aperture can be perhaps fed with just one feed. However, the design of even 1-D balanced transmission line (TL) is extremely difficult in the vicinity of $\beta = 0$ [5]: even small deviations from the optimal series impedance $Z$ and shunt admittance $Y$ lead to reduced matching and disturbed radiation pattern. Because of this, we must somehow relate $Z$ and $Y$ of the dispersion to the input impedance of the network.
2 Theory

The traditional way to analyze unit cells of loaded TLs is the Floquet principle [6]. We also begin the analysis from the T-type unit circuit of the periodic 1-D TL structure, seen in Fig. 1, where $Z_0$ is the characteristic impedance and $\theta$ is the electric length of the unit cell. The transmission matrix (ABCD matrix) of the unit cell is easily solved by multiplication of the discrete matrices [6], and the elements of the ABCD matrix read

$$A = \cos^2 \theta \left[ 1 + ZY / 2 \right] - \sin^2 \theta + j \sin \theta \cos \theta \left[ Z_0 Y + ZY_0 \right],$$  \hspace{1cm} (1)

$$B = \cos^2 \theta \left[ Z + Z^2 Y / 4 \right] - \sin^2 \theta \left[ Z + Z^2 Y / 4 \right]$$
\[+ j \sin \theta \cos \theta \left[ Z^2 Y_0 / 2 + Z_0 ZY + 2Z_0 \right], \hspace{1cm} (2)

$$C = Y \cos^2 \theta + j 2Y_0 \sin \theta \cos \theta,$$  \hspace{1cm} (3)

$$D = A, \hspace{1cm} (4)$$

where $Y_0 = 1/Z_0$. The balanced condition with nonzero group velocity and $\beta = 0$ for a 1-D TL is achieved when $Z_0 = \sqrt{Z/Y}$.

The dispersion and propagation in a 2-D TLN look however different in different directions. If we consider the different propagation directions separately, we can reduce the complex circuit model to the 1-D T-model of Fig. 1 [7]. The axial propagation direction is presented in Fig. 2. The unit cell can be seen to be composed of two 1-D unit cells with two admittances in parallel, so $Y_{2-D}$ is half the original value. Also, with the axial propagation direction (from bottom left to top right in Fig. 2) the two side stubs can be reduced to additional shunt admittances, $Y_0 = 1/Z_0 = jY_0 \tan(\theta/2)$ just like matching stubs with open-circuit termination. The diagonal propagation seen in Fig. 3 can be solved by splitting the unit cell into two parts, since there is no interaction between the halves. In fact, this same reasoning for axial and diagonal propagation directions can be easily extended to three dimensions as well. Table 1 shows the values of $Y$ with different propagation directions.

![Fig. 1. Circuit model for a symmetrical 1-D unit cell.](image)
By freely choosing the series impedance $Z$ and then calculating $Y$ from the table, the values can be used in Equations (1) to (4). Then, the dispersion can be calculated from [6]

$$\cosh(\gamma d) = A$$

where $\gamma = \alpha + j\beta$ is the complex propagation coefficient with $\alpha$ being the attenuation coefficient and $d$ is the period of the network. The Bloch impedance $Z_B$ can be calculated from

$$Z_B = \frac{\pm B}{\sqrt{A^2 - 1}}$$

where the plus-minus sign should be chosen according to the sign of $\beta$. 

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Table 1. Values for $Y$ in 1-D, 2-D and 3-D cases in axial, in-plane diagonal (2-D diagonal) and diagonal (3-D diagonal) propagation directions with chosen constant $Z$ and $f_0$ [7].

<table>
<thead>
<tr>
<th></th>
<th>1-D TLN</th>
<th>2-D TLN</th>
<th>3-D TLN</th>
</tr>
</thead>
<tbody>
<tr>
<td>Intrinsic shunt $Y$</td>
<td>$\frac{f_0}{Z}$</td>
<td>$\frac{f_0}{Z} - 2\theta$</td>
<td>$\frac{f_0}{Z} - 3\theta$</td>
</tr>
<tr>
<td>Axial propagation $Y$</td>
<td>$\frac{f_0}{Z}$</td>
<td>$\frac{f_0}{Z} + j2\tan(\theta/2)$</td>
<td>$\frac{f_0}{Z} + j4\tan(\theta/2)$</td>
</tr>
<tr>
<td>2-D diagonal propagation $Y$</td>
<td>$\frac{f_0}{Z}$</td>
<td>$\frac{f_0}{Z} + j2\tan(\theta/2)$</td>
<td>$(\frac{f_0}{Z} + j2\tan(\theta/2))^2$</td>
</tr>
<tr>
<td>3-D diagonal propagation $Y$</td>
<td>$\frac{f_0}{Z}$</td>
<td>$\frac{f_0}{Z} + j2\tan(\theta/2)$</td>
<td>$\frac{f_0}{Z}$</td>
</tr>
</tbody>
</table>


3 Simulations

The method of Section 2 or the equations found in [8] give the same results for the dispersion and the Bloch impedance. If we choose $L = 1$ nH and $C = 0.80$ pF according to equation $Y_{2-D} = Z/((2Z_0^2)$ found in Table 1, set $d = 20$ mm and $Z_0 = 50$ Ω, we can solve the dispersion and $Z_B$ from (5) and (6). The curves for the axial and diagonal directions are shown in Fig. 4 and Fig. 5. The leaky mode is $n = 0$ spatial harmonic in this kind of BW TLN, and leaky-wave radiation is produced in the frequency region where $|\beta| \leq k = \omega \sqrt{\mu_0 \varepsilon_0}$ and seen in Fig. 4 as the region where $\beta$ is inside the so-called light line. We can see that the leaky mode has in-plane isotropy in the 2-D TLN, and the frequency with $\beta = 0$ at $f_0 = 2.38$ GHz. The Bloch impedance is not equal to $Z_0$ at this frequency, but $L$ and $C$ can be tuned so that this holds at $f_0$.

This analytical formulation for 2-D BW TLNs is helpful when we want to design a TLN using lumped or distributed elements. A $6 \times 6$ TLN with aforementioned values was designed and simulated in both Ansoft HFSS full-wave simulator and ADS circuit simulator. The finite TLN was bounded by edge terminations of $Z_0$ at the end of each waveguide. The lumped series elements used in HFSS were 2-mm long ideal reactive sheets and shunt elements were $h = 0.787$-mm tall inductive sheets. The substrate in the model had the relative permittivity $\varepsilon_r = 2.33(1 - j0.0012)$.

It was seen that despite of ideal models of $L$ and $C$ in the full-wave simulator, the realized load values did not produce the balanced dispersion, but led to a stop band in the dispersion diagram manifested as a stop band in $S_{11}$ curve. With the help of a 1-D TLN of 2-D unit cells (Fig. 2), the load elements were tuned to produce a reasonably smooth $S_{11}$ as seen in Fig. 6. To tune the load values, one must use many unit cells so that the mutual coupling between the unit cells is taken into account. The dispersion, or $\beta$, of a finite sized TLN can be smooth even with a stop band, but the effect is more easily seen in either $\alpha$ or in $S_{11}$ as used in this paper, and finally in the radiation pattern.
Fig. 4. Axial and diagonal dispersion.

Fig. 5. The Bloch impedance in axial and diagonal directions.

Fig. 6. $S_{11}$ for axial propagation for $N=6$ long 2-D TLN after some tuning.
Fig. 7. The reflection coefficient $S_{11}$ from a 6×6 TLN fed from one corner.

The 6×6 TLN was fed from one of the corners (the shunt admittance was replaced by a feed port) and the input impedance and $S_{11}$ seen in Fig. 7 were studied. It was seen that even with the analytical load values in ADS simulation we see a region of bad matching near $f_0$. It turns out that this is an inherent property of an edge-terminated 2-D TLN. Since $\beta = 0$ near $f_0$, the effect of the TLs and the load elements disappears and the input impedance $Z_{in}$ looks as a parallel connection of edge terminations. In our case, the relation can be easily calculated to be ($N = 6$ in our simulations)

$$Z_{in} = \frac{Z_0}{2(2N-1)}$$

since we do not use any terminations at the unit cell with the feed. The circuit simulator gives similar results near $f_0$ to the full-wave simulator. Because of this, the TLN with a single feed requires some matching and cannot be used at a wide frequency range with the frequency scanning properties of 1-D LWAs. The radiation was indeed broadside near $\beta = 0$, and the directivity proportional to the aperture size [58].
4 Conclusions

A 2-D BW TLN with a balanced condition was studied for broadside antenna applications. A method for calculating the ABCD parameters for axial and diagonal directions in 2-D and 3-D structures was developed and used to match the simulated load elements to the balanced condition. Regardless of the balanced condition, an inherent impedance mismatch was seen to rise in edge-terminated TLNs when the phase-propagation coefficient is near zero.
References

Design of Subwavelength Tunable and Steerable Fabry-Perot Antennas

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Abstract

A Fabry-Perot antenna can be realized by placing a highly reflective Frequency Selective Surface (FSS) at a proper distance from a ground plane. The cavity, excited by a low gain antenna, transforms an omnidirectional field distribution into a highly directive one [1].

In the case of a metallic ground plane, the cavity height equals \( \lambda /2 \). However, when the metallic plate is replaced by an High Impedance Surface (HIS), the reflection phase of this ground plane can be properly chosen leading to a subwavelength design [2]. By varying, at the same time, the cavity height and the return loss matching of an active patch antenna, employed to feed the cavity, a multi-frequency design can be also obtained [3]. Changing of the cavity height, keeping constant the working frequency, leads to a beam steering. However, a mechanical tuning of the cavity is not a good practical design. Instead, by employing an active HIS ground plane [4,5] the phase \( \psi_R \) can be electronically controlled leading to a very compact tunable and steerable antenna.
1 Formulation

The layout of the analyzed structure and its transmission line equivalent circuit are shown in Fig. 1. The antenna is analyzed by a transmission line approach where the source is assumed to be an horizontal infinitesimal electric dipole inside the resonant Fabry-Perot cavity [6,7]. The original problem of analyzing the far field radiation of the Hertzian dipole inside the cavity is transformed into a scattering problem. Thus, the calculation reduces to a calculation of the field in the position of the source dipole due to an incident plane-wave. In the equivalent TL approach this corresponds to determine the voltage $V_s$ in correspondence of the dipole position. The voltage $V_s$ is computed by solving the following system of equations [6]

$$
\begin{bmatrix}
V_s & I_N \n
\end{bmatrix} =
\begin{bmatrix}
A & B \\
C & D
\end{bmatrix}
\begin{bmatrix}
I_0 \n
\end{bmatrix}Z_{HIS} =
\begin{bmatrix}
V_s \\
I_s
\end{bmatrix} =
\begin{bmatrix}
A' & B' \\
C' & D'
\end{bmatrix}
\begin{bmatrix}
I_0 \n
\end{bmatrix}Z_{HIS}
$$

where $I_N$ is the current in the N-th transmission line and $A$, $B$, $C$, $D$ constitute the chain matrix of the whole system. Such matrix can be computed as product of the three matrix composing the system

$$
\begin{bmatrix}
A & B \\
C & D
\end{bmatrix} = [M_{FSS}] [M_2] [M_1]
$$

with

$$
[M_n] = \begin{bmatrix}
\cos(k_{z_n}d_n) & jZ_{zn} \sin(k_{z_n}d_n) \\
\frac{j \sin(k_{z_n}d_n)}{Z_{zn}} & \cos(k_{z_n}d_n)
\end{bmatrix}
$$

$$
[M_{FSS}] = \begin{bmatrix}
1 & 0 \\
1/Z_{FSS} & 1
\end{bmatrix}
$$

where $Z_{FSS}$ represents the lumped impedance of the frequency selective surface (its expression is given later). The matrix $A'$, $B'$, $C'$, $D'$, representing the chain matrix of the transmission line between the source and the load impedance $Z_{HIS}$, is equal to $[M_1]$. The voltage can be computed both for TE and TM polarization by employing the following relations.
By solving the system in (1) the following equation for the voltage is provided [6]

\[ V_s = V_x - \frac{A'Z_{HSS} + B'}{CZ_{HSS}R_S + DR_S + A'Z_{HSS} + B} \]  \hspace{1cm} (5)

Once computed the voltage for TE and TM polarization, the radiation patterns can be obtained on both E plane and H plane, by normalizing the absolute value of the field magnitude to its maximum. The expression of the field components in the two main planes are

\[ H - \text{plane} \ (\phi = 90^\circ): |E_y| = |V_x^\text{TE}|; \quad E - \text{plane} \ (\phi = 0^\circ): |E_x| = |V_x^\text{TM}| \]  \hspace{1cm} (6)

The impedance of the FSS, \( Z_{FSS} \), is not assumed constant with the incident angle as in [6] but it is defined according to the averaged boundary condition theory [9]. In this approach the impedance is derived assuming an average current on the plane of the wires comprising the frequency selective surface. By applying the Babinet theorem, the surface impedance of an array of patches is obtained as well. These expressions are valid both for normal and oblique incidence under the hypothesis of homogeneous array \( (D < \lambda/3) \). In the present work the FSS is composed by an inductive grid and its impedance reads [10]:

\[
Z_{FSS,\text{grid}}^{\text{TE}} = j \frac{\omega \mu_0 D}{2\pi} \ln \left( \frac{1}{\sin \left( \frac{\pi r}{2D} \right)} \right)
\]

\[
Z_{FSS,\text{grid}}^{\text{TM}} = j \frac{\omega \mu_0 D}{2\pi} \left( 1 - \frac{k_0}{k_{\text{eff}}} \frac{\sin^2 (\vartheta)}{2} \right) \ln \left( \frac{1}{\sin \left( \frac{\pi r}{2D} \right)} \right)
\]  \hspace{1cm} (7)

where \( \varepsilon_{\text{eff}} = \left( \varepsilon_w + \varepsilon_{\text{down}} \right) / 2 \) represents the effective permittivity of an equivalent uniform medium composed by the two dielectrics surrounding the grid, D is the grid period, w is the strip width, \( \mu_0, k_0 \) are, respectively, the permeability and the wave
number in free space and $k_{\text{eff}} = k_0 \sqrt{\varepsilon_{\text{eff}}}$ is the wave number of the incident wave vector in the effective host medium.

The load impedance of the antenna equivalent circuit is the surface impedance of the high impedance surface. This impedance is computed by the parallel connection between the impedance of the grounded dielectric slab and the grid impedance of a patch array [10]

$$
Z_{\text{HIS\-patch}}^{\text{TE}} = \frac{j \omega \mu_0}{k_r} \frac{\tan(k_d d)}{k_r} \left( 1 - k_0^2 (\varepsilon_{r1} + \varepsilon_{\text{HIS}}) \frac{D}{\pi} \ln \left( \frac{1}{\sin \left( \frac{\pi}{2D} \right)} \right) \left( 1 - \sin^2 \left( \theta_i \right) \right) \varepsilon_{\text{HIS}} \right)
$$

$$
Z_{\text{HIS\-patch}}^{\text{TM}} = \frac{j \omega \mu_0}{k_r} \frac{\tan(k_d d)}{k_r} \left( 1 - k_0^2 (\varepsilon_{r1} + \varepsilon_{\text{HIS}}) \frac{D}{\pi} \ln \left( \frac{1}{\sin \left( \frac{\pi}{2D} \right)} \right) \left( 1 - \sin^2 \left( \theta_i \right) \right) \varepsilon_{\text{HIS}} \right)
$$

where $D$ is the periodicity of the patch array, $w$ is the gap width between patches, $\varepsilon_{\text{HIS}}$ is the permittivity of the dielectric slab composing the HIS, $\mu_0$, $k_0$ are the permeability and the wave number in free space, respectively.

The analysis of the active high impedance surface is performed by adding an additional lumped capacitor, in parallel with the capacitance created by the patch array [5]. In the full-wave simulations, this lumped varactors need to be connected between all the neighbouring patches.

![Fig. 1. (a) Layout of the analyzed structure and (b) equivalent circuit.](image)
2 Numerical Results

The analytical model previously presented has been used in order to show the main ideas of the paper.

The FSS has been designed as a planar inductive grid with a periodicity of 22 mm and a strip with of 8 mm. In order to allow a more reliable practical design, a supporting dielectric for the FSS has been considered. The dielectric has a thickness of 3.198 mm with a permittivity equal to 2.55. The height of the cavity is 13 mm that corresponds to roughly $\lambda_0/6$ at the chosen operating frequency of 3.4 GHz. The high impedance surface is composed by a patch array printed on a grounded dielectric slab with the same characteristic of the one of the FSS. The patch array is characterized by a periodicity of 15 mm and a gap between the patches of 1 mm. In Fig 2 the radiation patterns of the structure are shown at the operating frequency of 3.42 GHz, varying the diode capacitance. As expected, this electronic variation leads to a fictitious variation of the cavity high and a consequent steering of the beam. Moreover, this active configuration can be used to change the operating frequency of the antenna because, when the diode polarizations are changed, the broadside operating frequency shifts towards lower values. In Fig 3 the directivity in the broadside direction of the antenna is shown for different polarizations of the active elements. The figure shows how as the increasing of the capacitance value brings to a shift of the maximum directivity. This happens because, when the capacitance value is increased, the reflection phase value of the high impedance surface at a given frequency decreases leading to an equivalent higher cavity height. In Fig 4 the radiation patterns in correspondence of the directivities maxima are reported in order to show that, at these frequencies, the antenna radiates in broadside direction.

Fig. 2. Radiation pattern at 3.42 GHz obtained by varying diode capacitances.
Fig. 3. Directivity in the broadside direction by varying diode capacitances.

Fig. 4. Radiation patterns in correspondence of the peaks of the directivities for different capacitances.
References

Nonlinear Composite with Ferroelectric Nanoparticles

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Abstract

A composite formed by ferroelectric spheroids embedded in a linear dielectric matrix is considered. The nonlinearity of the ferroelectric material was taken into account. The goal of the work is to discuss nonlinear properties of the composite considered. The effective dielectric permittivity, the tuneability, and the loss factor of the composite will be estimated.
1 Introduction

The aim of this paper is to estimate the property of a composite material containing ferroelectric nonlinear inclusion embedded in a linear dielectric matrix. Ferroelectric composites were described earlier [1]. The general aim of developing the ferroelectric composites is the following:

1. Obtain smaller dielectric permittivity as compared with a pure ferroelectric,
2. Obtain tuneability of the material comparable with the tuneability of a pure ferroelectric
3. Decrease the microwave loss as compared with a pure ferroelectric.

It is suggested to consider a composite, in which the ferroelectric inclusions are in form of thin disks. The disk faces are parallel to the applied electric field. In this case the external field can effectively penetrate inside the embedded particles.

The inclusions will be considered as oblate spheroids (Fig.1).

![Fig. 1. Schematic presentation of the composite structure. Notations: a and c are semiaxes of ellipsoidal particle.](image)

The relation between external and internal fields in a spheroid can be easily described [2]. The quantitative spheroid parameters of the averaged field must not have a big discrepancy with the disk of the same size.

As usual any ferroelectric particle is surrounded by a thin non-ferroelectric layer with a small dielectric permittivity, which is referred to as a dead layer. In this consideration, the dead layer between ferroelectric material of the spheroid and surrounding non-ferroelectric media is not taken into account. The influence of the dead layer on properties of a composite was considered previously [3]. There was shown that in the case of the rather small filling factor the influence of the dead layer on the dielectric properties of the composite can be neglected.
2 Effective dielectric permittivity, tuneability and loss factor of the composite

The relation between the field inside the spheroid $E_{\text{in}}$ and the external field $E_{\text{ext}}$ is determined by the following formula

$$E_{\text{in}} = E_{\text{ext}} \left[ 1 + \left( \frac{\varepsilon_{\text{in}}(E_{\text{in}})}{\varepsilon_{\text{ext}}(E_{\text{in}})} - 1 \right) \cdot n_s(a,c) \right]^{-1},$$

(1)

where $\varepsilon_{\text{in}}(E_{\text{in}})$ and $\varepsilon_{\text{ext}}$ are dielectric permittivity of the ferroelectric inside the spheroid and dielectric permittivity of the host material; $n_s(a,c)$ is depolarization coefficient [2].

In order to find the internal field in the spheroid, we have to solve the nonlinear equation with respect to $E_{\text{in}}$ using the known function $\varepsilon_{\text{in}}(E_{\text{in}})$.

The nonlinearity of the ferroelectric material was taken into account. The known model of the dielectric response $\varepsilon_f(T,E,\xi)$ of incipient ferroelectrics was used [4].

Let us suppose that $\varepsilon_{\text{in}}(E_{\text{in}})$ in (1) is equal to $\varepsilon_f(T,E,\xi)$, where $E = E_{\text{in}}$. Thus, one obtains nonlinear equation with respect to internal electric field inside the spheroid:

$$\left( 2 + \frac{\varepsilon_f(T,E,\xi)}{\varepsilon_{\text{ext}}} \right) \cdot E_{\text{in}} - 3E_{\text{ext}} = 0.$$

(2)

After solving the equation (2), the relative permittivity of the composite with respect to the low level microwave field can be found as a function of the high level external $dc$ field. For that, the following formula can be used:

$$\varepsilon_{\text{eff}}(a,c,E_{\text{in}},E_{\text{ext}},\xi) = \frac{\varepsilon_{\text{ext}}}{\gamma} \left[ 1 + \left( \frac{\varepsilon_f(T,E_{\text{in}}(a,c,E_{\text{in}},E_{\text{ext}}),\xi)}{\varepsilon_{\text{ext}}(E_{\text{in}})} - 1 \right) \cdot n_s(a,c) \right] + \varepsilon_{\text{ext}}$$

(3)

where $\gamma$ is the inversed filling factor.

Relative permittivity of the composite as a function of the external field is presented in Fig. 2.
Fig. 2. Relative permittivity of the composite as a function of the external field.

The tuneability of the composite is determined as follows:

\[ n_{eff}(T, E_{ext}, \xi) = \frac{\varepsilon_{eff}(T, 0, \xi)}{\varepsilon_{eff}(T, E_{ext}, \xi)} \]  (4)

The tuneability of the composite \((a = 500\,nm, c = 50\,nm, \gamma = 3)\) is shown in Fig.3.

Fig. 3. Tuneability of the composite as a function of the external field.

In order to take into account the inherent loss of the ferroelectric material, we replace \(\varepsilon_f(T, E_{in}(a, c, E_{ext}, \xi))\) in formula (3) by \(\varepsilon_f(T, E_{in}(a, c, E_{ext}, \xi)) \cdot (1 - i \tan \delta)\). (Here \(i\) is the imaginary unit, \(\tan \delta\) is the
loss factor of the material). For the original bulk ferroelectric material, it was taken: \( \tan \delta = 0.02 \). Thus we find the real and imaginary part of the effective permittivity of the composite and consequently the composite loss factor. The result of the simulation is presented in Fig. 4.

\[
\tan \delta_{\text{eff}}(E_{\text{ext}})
\]

**Fig. 4. Loss factor of the composite as a function of the external field.**

**Acknowledgment**

The author is grateful to Professor Orest G. Vendik for the problem formulation and support of the problem solving.
References

EM simulation for the material electrical properties measurement in microwave range

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Abstract

One of the basic problems in measurements of the electrical characteristics of materials is finding relations between measured electrical values and characteristics of the material. Especially when the shape of a material is complex.

The EM simulation software calculates electrical parameters of defined structure. Alas, at measurements, it is necessary to solve a reverse problem, for what the software is not adapted. Using some properties of the program it is possible to achieve the solution of a reverse problem. Also it allows to take into account the spurious factors arising at connection of a measured sample to a measuring circuit.

Thus, it is possible to consider the EM simulation software as a powerful and universal tool for the measurements of material electrical properties.
1 Introduction

The measurements of the permittivity and permeability consist of three main components: measurement equipment, investigated material in the device under test and mathematical model which relates measured by equipment electrical values and physical values of the material properties, what is the goal of the measurement (Fig. 1).

**Fig. 1.** Measurement scheme.

For example, we measure complex reflection coefficient by vector network analyzer and using the mathematical model, we can calculate dielectric permittivity.

The mathematical model is simple only in very few cases: like the capacitor with dielectrics between plates, or the transmission line in the dielectric media. Alas, when frequency is high or permittivity is relatively high, there are difficulties to make device under test in small dimensions and various spurious effects appear, thus simplest models are not acceptable. Other models can be very complex and difficult. For example, the complex reflection coefficient $\Gamma$ of the dielectric rod in the rectangular waveguide (Fig. 2) is defined by equation [1]:

\[ \text{Equation [1]} \]
\[ \Gamma = \frac{4(e-1)J_1(\beta)}{\pi \sqrt{\left(\frac{2a}{A}\right)^2 - 1}} \]

\[ \Delta = e J_1(\beta) \left[ H_0(\alpha) + \sum_{n=1}^{\infty} 2(-1)^n H_0(nka) \right] - \sqrt{e} J_0(\beta) H_i(\alpha), \]

\[ \beta = kr \sqrt{e}, \]

\[ \alpha = kr \]

where \( r \) is the radius of the rod, \( e \) - complex dielectric permittivity, \( a \) – the wide wall of the waveguide, \( k \) – is the wave number, \( \lambda \) - the wavelength in the free space.

Fig. 2. Dielectric rod in the rectangular waveguide.

Thus, the development of the measurement’s mathematical model may be serious problem. Especially, when the geometrical shape of the sample is complicated, no analytical solution can be obtained from Maxwell equations, as it was shown in the example above. In such cases, only numerical solution can be helpful.
2 EM simulation software application for the material electrical properties calculation

Recently, the calculation power of the computers significantly increased. The commercial software, based on various numerical Maxwell equations solution methods appeared which allows EM simulating complex 3D structures. For example, it is possible to calculate S parameters of the device, when dimensions and shape of this one and electrical properties of the material are known. Doing dielectric properties measurements it is necessary to solve the reverse task: to find electrical properties using measured S parameters. Alas, such software are not adapted for this task.

Some of the EM simulation software has optimization option. It means that if the optimization goal as S parameters is defined, the program will select device dimensions and (or) electrical properties, in our case the sample’s which is a part of the device under test electrical properties. Software Ansoft HFSS, version 11 was used for our calculations.

In Fig. 3 optimisation setting sample is shown, when device under test is already mentioned dielectric rod in the waveguide.

![Fig. 3. Optimisation setting sample of dielectric rod in the waveguide.](image)
Solid curves are measured reflection and transmission coefficients of cylindrical rod in the rectangular waveguide. We must mark, that the same reflection and transmission may be with different permittivity. In such case is much better define electrical parameters at least at two different frequencies, which allows solving the problem unambiguously.

In Fig. 4 the comparison of S parameters calculated by using simulation software (solid lines) and solving equation (1) (dotted line) are shown. The mesh generated by program is shown on the waveguide. We obtain good agreement between these two different methods.

Another example is the rectangular dielectric rod with both opposite sides metallization, as shown in Fig. 5. We took values of permittivity: $\varepsilon''=140$ $\varepsilon'''=8$. Lumped ports connected between conductive surfaces at the narrow ends.
Fig. 5. Rectangular dielectric rod with both opposite sides metallization with the electromagnetic wave inside it. Arrows show magnetic field component at 12 GHz.

At the low frequencies this device can be considered as a microstrip line. At the high frequencies as shown in Fig. 5, longitudinal fields’ components with in-line dispersion appear and microstrip model is not more reasonably accepted for the calculations.

Using EM simulation software with optimisation option it is possible to calculate dielectric permittivity from measured S parameters, both on the low and on the high frequencies limits.

The transmission line shown in the Fig. 5 is idealisation, because it is connected to ideal ports, sample is in the free space, conductors are perfect. Real sample is connected to analyzer by coaxial line, sample is placed on conductive plate, metallisation has finite thickness and conductivity. Such more realistic device is shown in the Fig. 6.
Fig. 6. Sample connected to coaxial lines.

It is very complicated problem to develop and to solve such model by usual analytical methods. By using numerical methods, there is no difference what structure is under simulation - complex or simple. Difference will be only in computation time. On the Fig. 7 calculated S parameters of the real and ideal samples (doted lines) of the geometries presented in the Figs. 5 and 6 are shown.
Fig. 7. Calculated S parameters of the real (solid lines) and ideal samples (doted lines) of the geometries presented in the Figs. 5 and 6.

Due various spurious factors described above, there is no difference between realistic and ideal model only at low frequencies. This method has been applied for the microwave measurements of BZN pyrochlore ceramics microwave properties, as shown in the Fig. 8 (green points).
Fig. 8. Summary of the methods, used for the investigation of BZN ceramics in a wide frequency range.
3 Conclusions

Using optimisation option in the commercial EM simulation software it is possible to find the electrical material properties of the samples in very complex shape. Using numerical methods allow us to take into account the spurious factors arising at connection of a measured sample to a metering circuit. Thus, it is possible to consider the EM simulation software as a powerful and universal tool for the material properties measurements.
References

Miniaturized Low-Loss LTCC Bandpass Filters on Capacitively Loaded Cavities

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Abstract

Miniaturization of high-Q LTCC cavity resonators by using a capacitive loading is considered as applied to design of low-loss and small-size narrowband filters for low-frequency applications. It is demonstrated by the results of 3D electromagnetic simulations that the size of a highly loaded LTCC cavity resonator can be as low as one eighth of the guided wavelength while the unloaded Q-factor still remains much higher than for a quasi-lumped-element LTCC resonator. Design of two LTCC bandpass filters for telecommunications, which are based on miniaturized capacitively loaded cavities, is presented.
1 Introduction

The multilayer Low-Temperature Co-fired Ceramics (LTCC) technology allows designing very compact microwave devices as 3D highly integrated modules. A lot of miniature LTCC filters for L- and S-bands were designed up to date based on quasi-lumped elements. The main problem of those designs deals with a rather high insertion loss due to insufficient Q-factor of quasi-lumped-element resonators.

At the same time low-loss LTCC filters based on high-Q cavity resonators were designed for millimeter-wave applications [1]-[3]. Furthermore, it is well known that by loading a cavity resonator with a capacitive post, one can reduce the resonant frequency of the cavity while still maintaining a relatively high unloaded Q-factor [4]. Various designs of compact microwave filters based on highly loaded evanescent-mode cavities, which use different fabrication technologies (LTCC, PCB, silicon micromachining process), were recently presented [4]-[6].

In this paper, possibilities of a further miniaturization of capacitively loaded cavity resonators to design small-size and low-loss narrowband LTCC filters for S-band are discussed. It is demonstrated that the size of a highly loaded LTCC cavity can be as low as one eighth of the guided wavelength ($\lambda_g/8$) while the Q-factor of such a miniaturized resonator still remains much higher as compared to a quasi-lumped-element LTCC resonator. The two-pole and four-pole low-loss LTCC bandpass filters for UMTS/LTE-2100 applications (2110–2170 MHz) are presented as the design examples illustrating the successful proof of principle.
2  Miniaturized capacitively loaded cavity resonators

A schematic representation of a capacitively-loaded cavity resonator is shown in Fig. 1-a. The resonator consists of a cavity with a conductive post, which is short-circuited at one end and capacitively-loaded at another end. The higher capacitive loading, the lower is the resonant frequency. For a given resonant frequency, dimensions of the cavity decrease with increasing the capacitive loading. On the other hand, it is followed by decreasing the unloaded Q-factor of the resonator.

![Diagram of capacitively loaded cavity resonator](image)

**Fig. 1.** Capacitively loaded cavity resonator: (a) schematic representation, (b) LTCC structure, (c) equivalent circuit, (d) characteristics obtained by EM simulation.

The highly loaded cavity can be considered as a quasi-lumped-element resonator since the electric field is concentrated on the bottom of the capacitive post but the magnetic field distribution remains relatively unchanged. Thus, the resonant frequency is reduced significantly when the bottom of the capacitive post is at close proximity to the bottom of the cavity due to the increased capacitance. However, the metal loss which is associated with the tangential magnetic field on
metal surfaces does not change significantly. As a result the unloaded Q-factor of the resonator does not drop as quickly as the resonant frequency does [4].

Fig. 1-b shows a possible LTCC implementation of capacitively loaded square-shape cavity resonator. On the top and the bottom the resonator is covered by two ground planes. In the figure, the middle part of the top ground plane is not shown to demonstrate the internal structure of the resonator. Rows of stacked vias are used as sidewalls. The square post is also formed by rows of stacked vias situated along the post perimeter and connected with the ground plate on the top and with the capacitive plate on the bottom. The resonator is connected with external circuits by inductive coupling elements. Input and output coplanar feed lines are situated on the top of the LTCC structure. Equivalent circuit of the capacitively loaded LTCC cavity with inductive coupling elements is presented in Fig. 1-c.

Due to technological reasons, the number of layers in LTCC multilayer structures is basically limited to 10-15. Hence, neither the height of LTCC cavity resonator nor the post length can be larger than 2–3 mm approximately. Such a cavity resonator operates on TM_{110} mode [1]. The electrical length of the post is equal to a few degrees and consequently the post inductance $L$ is rather low. It facilitates a high capacitive loading of the resonator that leads to an effective size reduction. The area of the capacitive plate can be decreased by using the lower LTCC layer as thin as possible.

Fig. 1-d presents simulated characteristics of the LTCC capacitively loaded cavity resonator accomplished using six layers of DuPont Green Tape™ 951 LTCC ($\varepsilon_r = 7.8$) with the thickness of 210 $\mu$m and one layer with the thickness of 95 $\mu$m. Thus, the height of LTCC structure is about 1.36 mm. The resonator was developed for the resonant frequency of 2140 MHz. The internal size of the cavity was reduced to $6 \text{ mm} \times 6 \text{ mm} \approx \lambda_g/8$. The size of the capacitive post was chosen as $2 \text{ mm} \times 2 \text{ mm}$ with the capacitive plate of $4.6 \text{ mm} \times 4.6 \text{ mm}$. Electromagnetic (EM) simulation of the multilayer LTCC structure of the resonator was carried out using Ansoft HFSS 3D electromagnetic field solver. The dielectric loss tangent specified by DuPont ($\tan(\delta) = 0.002$) and a typical value of the conductive paste surface resistance ($R_{\text{sur}} = 5 \text{ mOhm/square}$) were taken into account during the simulation. The estimated unloaded Q-factor of this resonator is about 220 that is at least three times higher than for a quasi-lumped-element LTCC resonator.
3 Design of bandpass LTCC filters on capacitively loaded cavity resonators

Two small-size and low-loss narrowband LTCC filters were designed on the base of miniaturized highly loaded cavity resonator mentioned above.

Fig. 2-a illustrates a LTCC structure of two iris-coupled capacitively loaded cavities. Though such a coupling is generally mixed, the main contribution is made by the electric field. Hence, the coupling can be considered as mainly capacitive one. Equivalent circuit of these coupled resonators is presented in Fig. 2-b. The value of the coupling capacitance $C_c$ depends on the iris size $b$ as shown in Fig. 2-c. The larger iris size, the higher is the coupling capacitance. Since $b < a$ (see Fig. 2-a), the cavity size $a$ cannot be chosen arbitrary. It should be sufficient to provide a required coupling that is weak for a narrowband filter design whereas a tight coupling is required to provide relatively wide passband.

![LTCC structure](image1)

![Equivalent circuit](image2)

![Coupling capacitance as a function of normalized iris size](image3)

![Characteristics of the 2 pole bandpass filter](image4)

**Fig. 2.** Two iris-coupled capacitively loaded cavity resonators: (a) LTCC structure, (b) equivalent circuit, (c) coupling capacitance as a function of normalized iris size, (d) characteristics of the 2 pole bandpass filter obtained by EM simulations of lossless (dashed lines) and lossy (solid lines) LTCC structure.
A two-pole bandpass LTCC filter with the Chebyshev characteristic and the fractional bandwidth of 2.8% was designed for UMTS/LTE-2100 application on the base of the structure shown in Fig. 2-a. The entire multilayer LTCC structure of the two-pole filter has the size of 13 mm × 7 mm. Characteristics of the filter obtained by EM simulations of lossless and lossy filter structure with the aid of Ansoft HFSS software are presented in comparison in Fig. 2-d. The simulation revealed that the filter midband insertion loss is as low as 1.2 dB.

As the next step we designed the four-pole Chebyshev bandpass filter for the same frequency band. Equivalent circuit and multilayer LTCC structure of this filter are shown in Fig. 3-a and Fig. 3-b, correspondingly. The filter size is 13 mm × 13 mm. Simulated characteristics of the filter with and without loss, are presented in Fig. 3-c. The midband insertion loss is equal to 2.9 dB.

Fig. 3. The four-pole bandpass filter: (a) equivalent circuit, (b) LTCC structure, (c) characteristics obtained by EM simulations of lossless (dashed lines) and lossy (solid lines) filter structure.
4 Conclusions

A capacitive loading allows a significant decreasing of the size of LTCC cavity resonator. It was illustrated by the results of 3D electromagnetic simulations that the size of such a resonator can be reduced to $\lambda_g/8$ while the unloaded Q-factor remains much higher with respect to a quasi-lumped-element LTCC resonator. This gives possibilities to design compact and low-loss narrowband LTCC filters for low-frequency applications.

The first results of designing low-loss and small-size LTCC bandpass filters on capacitively-loaded cavities for UMTS/LTE-2100 application demonstrate a successful proof of principle and a high potential for further innovations.

Acknowledgement

Helpful discussions with Irina Vendik and Orest Vendik are gratefully acknowledged.
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University of Oulu, Finland

Distributed by OULU UNIVERSITY LIBRARY
P.O. Box 7500, FI-90014
University of Oulu, Finland
Edited by
Marina Tyunina & Orest Vendik

PROCEEDINGS OF THE 16TH INTERNATIONAL STUDENT SEMINAR “MICROWAVE AND OPTICAL APPLICATIONS OF NOVEL PHENOMENA AND TECHNOLOGIES”, JUNE 8–9, OULU, FINLAND