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Full Length Research Paper

New blend SOC-WCIP strategy for the examination of low pass channel

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The combination of the Wave Concept Iterative Process (WCIP) method combined with the Short-Open Calibration SOC technique to micro strip discontinuity is presented. This SOC technique is directly accommodated in the WCIP method. It is used to remove the unwanted parasitic errors brought by the approximation of the impressed voltage sources and the feed lines. The WCIP is formulated in such a way that the port voltages and currents are explicitly represented through relevant network matrices. The scattering parameters of the filter structure can be obtained through the development and the simulation. This result for micro strip discontinuities is compared with the conventional WCIP method. These instructions provide you guidelines for preparing papers for International Journal. Use this document as a template and as an instruction set.

Key words: Short open calibration (SOC), wave concept iterative process (WCIP), micro strip discontinuity, third-order filter.

INTRODUCTION

Since more than two decades ago, numerous applications of WCIP have been developed to simulate many modern microwave and millimeter-wave planar circuits: Planar active circuits (Sboui et al., 2001), antennas (Sboui et al., 2007a), and filters (Sboui et al., 2007b). The prerequisite of the accurate characterization of planar micro strip discontinuities has been in an extraordinary progress, since the development of monolithic microwave integrated circuits (MMICs) and miniature hybrid microwave integrated circuits (MHMICs) is important.

In our present paper, we present an iterative method based on the wave concept. The principle of this method is based on the solution of the recurrence relationship between the incident and the reflected waves, with respect to the continuity conditions in the spatial domain. Results are approached progressively by successive iterations until their convergence in planar array. This method has been applied since one decade to many kinds of radio frequency integrated circuits (RFIC) and diffraction problems (N'gongo and Baudrand, 1999). However, it is important to examine carefully the discontinuity between the impressed voltage sources and the feed lines.

The short-open calibration (SOC) technique was proposed and developed by Zhu and Wu on the basis of a deterministic MOM algorithm and it has been detailed by Zhu and Wu (2002b) for planar circuit modeling problems. This technique, which was inspired from the real world measurement techniques, differentiates the even from the odd excitations with one section of uniform line to formulate the open/short standards elements and then calculate the error boxes of the structure.

Such SOC calibration techniques have successfully been used in equivalent circuit modeling and applications of various microwave planar structures. The parasitic errors given by the approximation of the impressed voltage sources and also the feed lines be present

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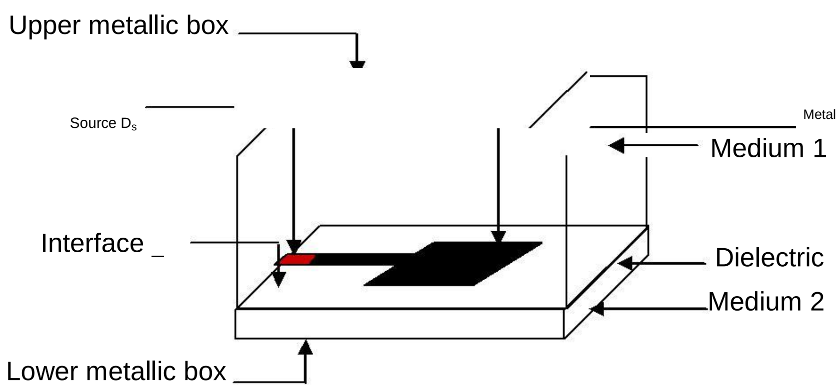

Figure 1. General structure.

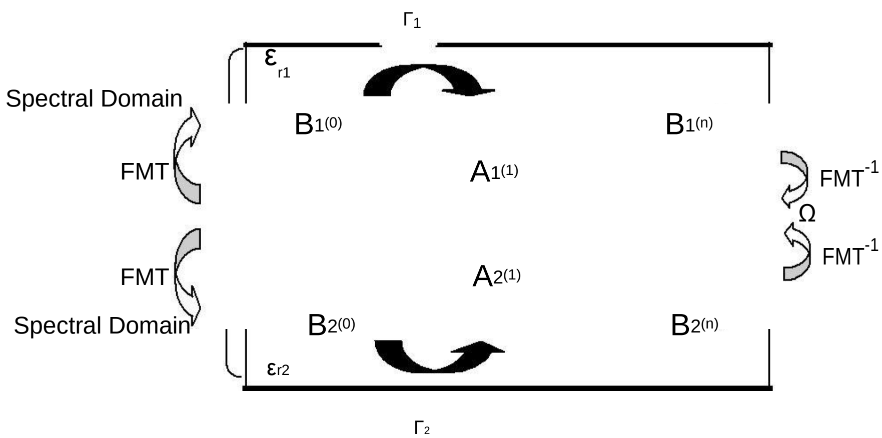

Figure 2. Principle of the wave concept.

(Zhu and Wu, 1999, 2002a, b; Chen et al., 2000, 2003; Li et al., 2000; Li and Wu, 2004). In fact, with SOC-WCIP method, the proposed problem is resolved; this is not possible with other numerical tools. In this paper, we proposed the new combination SOC-WCIP method. Here, a pair of impressed electrical fields having identical magnitude and reverse orientation is used to excite the two ports in order to create the “short” element, and a pair of impressed electrical fields having identical magnitude and orientation is used to excite the two ports in order to create the “open” element.

THEORY

Preliminary, we present a short description of the iterative so called (W.C.I.P.). Hence, we consider the general structure of Figure 1, representing an example of planar circuit printed on a dielectric substrate. The interface Ω between the two homogeneous mediums (1) and (2) can support circuit design and includes three sub-domains, which are the source, the metal and the dielectric. The electromagnetic field is known on all point of the plane interface. The iterative process is introduced in Figure 2. The formulation of this method is based on the wave concept, an incident wave and a reflected wave. These waves are defined by linear combination of the tangential currents and voltages. The passage of them from the spectral and spatial domain is ensured by the iterative process using the fast modal transformation (FMT).

$$\begin{aligned} E_{T1} &= E_{T2} = E_0 - (H_T \wedge n).z_0 && Source \quad (D_s) \\ E_{T1} &= E_{T2} = 0 && Metal \quad (D_m) \qquad (1) \\ J_{Tot} &= J_1 + J_2 && Dielectric \ (D_d) \end{aligned}$$

In this last equation, E_0 is the field source, it is defined on the source domain D_S and it is used to initialize the iterative process. E_{T1} and E_{T2} are the tangential components of the electric field at the medium ‘i’ (i=1 or 2) and it is continuous. H_{T1} and H_{T2} are the corresponding magnetic field, their sum is equal to zero because the density of current is null.

In the two sides of circuit (mediums 1 and 2), the higher-order modes are shunted by their reactive admittance $\hat{Y}_i$, which relates the tangential electric field E_T to the total current density J; hence, we obtain:

$$J_i = \hat{Y}_i E_{Ti} \tag{2}$$

In this last equation, we have associated the vector current density to the magnetic field, $\vec{J} = H \wedge n$. The spectral development of the admittance operator $\hat{Y}_i$ is defined in Equation 3.

$$\hat{Y}_i = \sum_{mn} \left| \psi^{TE}_{mn} \right\rangle Y^{TE}_{mn,i} \left\langle \psi^{TE}_{mn} \right| + \sum_{mn} \left| \psi^{TM}_{mn} \right\rangle Y^{TM}_{mn,i} \left\langle \psi^{TM}_{mn} \right| \tag{3}$$

With $Y^{TE}_{imn} = \dfrac{jw\varepsilon_0 \varepsilon_{ri}}{\gamma_{imn}}$ and $Y^{TM}_{imn} = \dfrac{\gamma_{imn}}{jw\propto_0}$

Where γ_{imn} is the propagation constant of the medium 'i' and it is given by:

$$\gamma_{imn} = \sqrt{\left(\frac{m\pi}{a}\right)^2 + \left(\frac{n\pi}{b}\right)^2 - k_0^2 \varepsilon_{ri}}$$

Where a and b are the box dimensions and ε_0, ε_i and $\propto_0$ are respectively the permittivity of the vacuum, the relative permittivity of the medium 'i' and the permeability of the vacuum.

Therefore, the expression of the electromagnetic fields from waves is given from the Maxwell theory. As an application of the theory, we attempt to analyze a little cube by the transverse line matrix (T.L.M.) as a basis structure of empty space in which there are the incident and reflected waves at each face of the cube. Its scattering matrix is computed with a finite difference scheme applied to Maxwell equations and an appropriate definition of waves (Krumpholz and Russer, 1994).

$$\begin{vmatrix} A \\ B \end{vmatrix} = \frac{1}{2\sqrt{Z_{0i}}} \begin{vmatrix} 1 & Z_{0i} \\ 1 & -Z_{0i} \end{vmatrix} \begin{vmatrix} E_i \\ H_i \wedge n_i \end{vmatrix} \tag{4}$$

Z_{0i} is an intrinsic impedance characterizing the medium, it is often chosen as $\sqrt{\dfrac{\propto_0}{\varepsilon_0 \varepsilon_i}}$ and n_i is the unit vector orthogonal to Ω, giving the direction of the incident and reflected waves. This last equation allows us to calculate the boundary conditions given by the Equation (1) as the following general spatial equation.

$$\begin{vmatrix} B_1 \\ B_2 \end{vmatrix} = \underbrace{\{S_m + S_d + S_s\}}_{S_\Omega} \begin{vmatrix} A_1 \\ A_2 \end{vmatrix} + \begin{vmatrix} B_{01} \\ B_{02} \end{vmatrix} \tag{5}$$

$S_m = \hat{H}_m \begin{pmatrix} -1 & 0 \\ 0 & -1 \end{pmatrix}$, $S_d = \hat{H}_d \begin{pmatrix} I_1 & I_1 \\ I_2 & -I_1 \end{pmatrix}$

and $S_S = \hat{H}_s \begin{pmatrix} ds_1 & ds_2 \\ ds_2 & ds_3 \end{pmatrix}$

Where: $\hat{H}_\delta$ is the Heaviside unit step defined as the following:

$$\hat{H}_\delta = \begin{cases} 1 & \text{if } \delta = m, d, \text{ or } s \\ 0 & \text{elsewhere} \end{cases}$$

And: $ds_1 = \dfrac{1 - n_1 + n_2}{1 + n_1 + n_2}$, $ds_2 = \dfrac{2n}{1 + n_1 + n_2}$, $ds_3 = \dfrac{1 + n_1 - n_2}{1 + n_1 + n_2}$

$I_1 = \dfrac{1 - n_{12}}{1 + n_{12}}$, $I_2 = \dfrac{2n_{12}}{1 + n_{12}}$, $n_1 = \dfrac{y_0}{y_{01}}$, $n_2 = \dfrac{y_0}{y_{02}}$, $n_{12} = \dfrac{y_{01}}{y_{02}}$, $n = \dfrac{\sqrt{y_{01} y_{02}}}{y_0}$

B_{01} and B_{02} are the components of the source wave; they defined by Equation 6.

$$\begin{vmatrix} B_{01} \\ B_{02} \end{vmatrix} = \frac{\hat{H}_s}{2y} \begin{vmatrix} \sqrt{y_{01}} \\ \sqrt{y_{02}} \end{vmatrix} \tag{6}$$

By combining Equations (2) and (3), a spectral equation can be defined in the modes domain (7).

$$\begin{vmatrix} A_1 \\ A_2 \end{vmatrix} = \begin{vmatrix} \Gamma_1 & 0 \\ 0 & \Gamma_2 \end{vmatrix} \begin{vmatrix} B_1 \\ B_2 \end{vmatrix}_\alpha$$

α is the TE or TM mode (7).

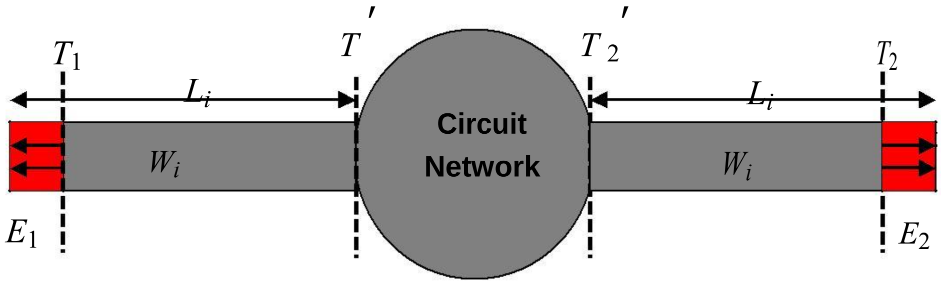

Figure 3. Physical model.

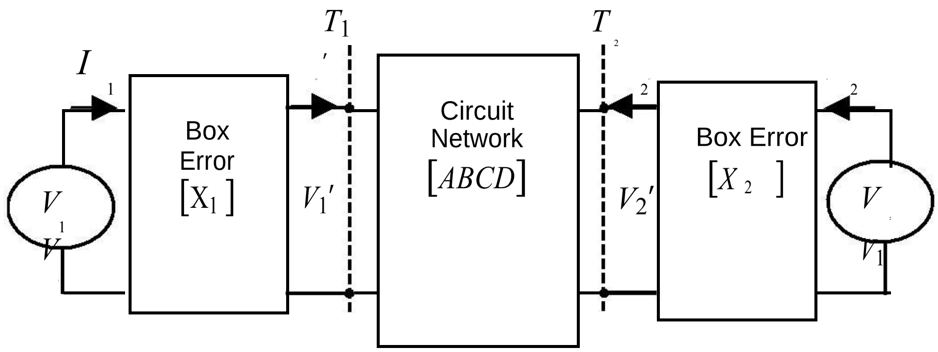

Figure 4. Process of the calibration SOC.

Where:

$$(\ _i)_\alpha = \frac{y_{0i} - (Y_i)_\alpha}{y_{0i} + (Y_i)_\alpha}$$

Hence, an iterative resolution of problem is established by considering a recursive relationship between the two spectral and spatial Equation (8).

$$B = S_\Omega A + B_0 \quad \text{on } \Omega$$
$$\ \ \hat{=}\ B \quad \text{on the two sides} \qquad (8)$$

Where:

$$B = \begin{vmatrix} B_1 \\ B_2 \end{vmatrix}, A = \begin{vmatrix} A_1 \\ A_2 \end{vmatrix}, B_0 = \begin{vmatrix} B_{01} \\ B_{02} \end{vmatrix}, \text{ and } \hat{\ } = \begin{vmatrix} \ _1 & 0 \\ 0 & \ _2 \end{vmatrix}$$

The two equations are defined respectively in the spatial and spectral domain. At each iteration, it is necessary to change the type of domain. In this paper such as in "Electromagnétisme Et Interconnexions", Charruau (1997), we combine the wave concept with the 2D-FFT algorithm to change the type of domain. This technique is called fast mode transformation F.M.T. The use of FMT is required to mesh the 2D circuit into small rectangular cells. Hence, the boundary conditions are satisfied at each cell.

The mechanism of the iterative procedure is summarized in Figure 2. Finally, it is possible to determine the admittance matrix, from that, the scattering matrix S_{ij} for a two port can be obtained by the following equation:

$$S_{ij} = [1 - Y][1 + Y]^{-1} \quad (9)$$

[Y] is the admittance matrix.

After recovering the parameters of the structure, we applied the process of calibration which was inspired from the real world measurement techniques, makes use of the even/odd excitation with one section of uniform line to formulate the open/short standards and then calculate the error boxes of the structure (Zhu and Wu, 2002b). The corresponding network model of the structure as shown in Figure (3) can also be arranged via three distinct circuit topologies, namely, the error box [X1], the circuit network [ABCD], and the error box [X2]. They are sequentially cascaded, as shown in Figure (4). In this way, all of the parasitic effects brought by the feed lines and port discontinuities in the simulation model are considered through network parameters of the two error blocks, which may be considered as "lumped-element" arrangements of the parasitic effects. Thus, the network parameters of two error blocks contain all of the parasitic effects brought by the feed lines and the port discontinuities. These two error

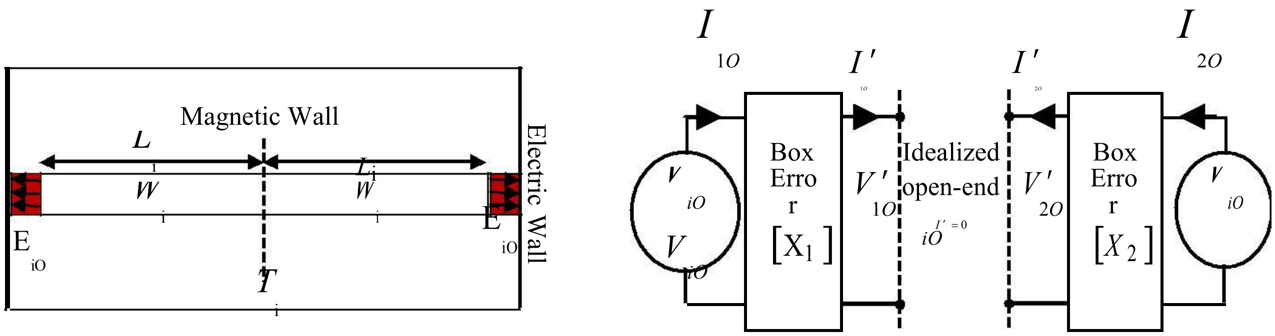

Figure 5. Open-end case, ideal open-end element, matrix standard open-end.

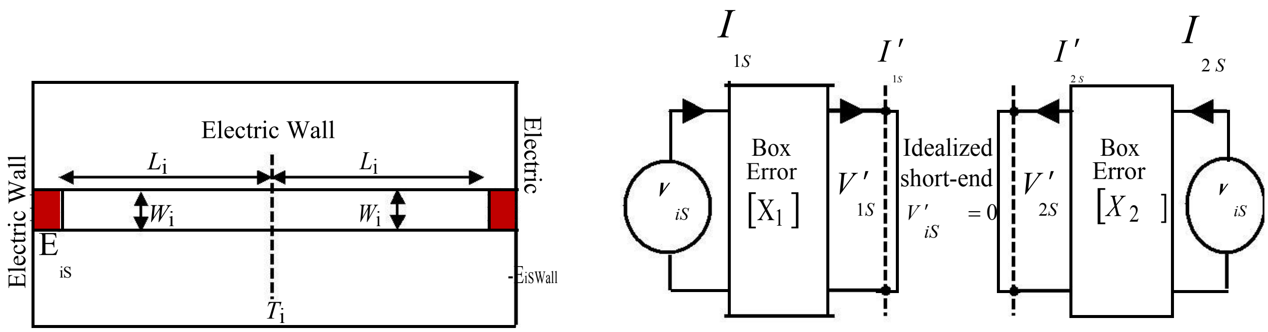

Figure 6. Short-end case, ideal short-end element, matrix standard short-end.

blocks make it easy to analytically deduce the equivalent voltages and currents (V_1', I_1', V_2', I_2') at the reference planes I_1', I_2' from those (V_1, I_1) and (V_2, I_2) defined at the port planes T_1 and T_2, which can be obtained by the WCIP method.

Now the so-called "short" and "open" elements can naturally be formulated in the WCIP to accurately simulate the ideal open-end and short-end. It is accomplished by exciting a pair of odd and even impressed voltages along a uniform feed line, as shown in Figures (5b) and (6b), whose symmetrical plane T_i becomes short (electric wall) and open (magnetic wall), respectively. The uniform feed line is suitably selected with a length of twice the distance L_i, which is defined between the port and reference plane of the *i*th (*i*=1, 2) feed line. As shown in Figures (5a) and (6a), applying "short" and "open" procedures via the impressed voltages results in two sets of equivalent networks with an idealized open-end and short-end, respectively. A pair of impressed electrical fields having identical magnitude and reverse orientation E_i is used to excite the two ports in order to realize "short" element in the proposed circuit model. Here Gaussian pulse *Vs* is used. Then the values of voltages at the planes T_1, T_1', T_2', T_2 are calculated by WCIP method. A pair of impressed electrical fields having identical magnitude and orientation E_i is used to excite the two ports in order to realize "open" element in the proposed circuit model.

For the short-end element, the voltage at the symmetrical plane is ideally equal to zero so that a perfect electric wall can be simulated. This electric wall divides this line into two independent parts, one of which can be treated as the *i*th (*i* = 1, 2) transition block loaded by an idealized short-end in the equivalent circuit network of the *i*th feed line, shown as Figure (6a). Based on the network topology, two equations for calculating the [*ABCD*] transmission matrix of the feed line is obtained in terms of the calculated port voltage and current:

$$a_i I_{iS} + b_i V_{iS} = 0 \tag{10}$$

$$c_i I_{iS} + a_i V_{iS} = 0 \tag{11}$$

As for the open-end element, a perfect magnetic wall is at the symmetrical plane, and the current at the plane is equal to zero. Therefore, another equation toward the solution of the [*ABCD*] matrix can be established on the basis of network theory with the port voltage and port current

$$a_i I_{iO} + b_i V_{iO} = 0 \tag{12}$$

Use of the reciprocity of the transition networks leads to an additional equation as:

$$a_i d_i - b_i c_i = 1 \tag{13}$$

In order to simplify the description that follows, all equivalent currents above, respectively, are normalized by tensions applied to the ports for the two kinds of excitation arrangements: "short circuit" and "open circuit":

$$\begin{aligned} \bar{I}_{iO} &= I_{iO}/V_{iO} \\ \bar{I}_{iS} &= I_{iS}/V_{iS} \\ \bar{I}'_{iS} &= I'_{iS}/V_{iS} \end{aligned} \tag{14}$$

In this normalization the letter “O” signified the "open circuit" and the letter “S” signified the "short circuit". Through the solution of Equations (10), (11), (12), (13), and (14) the [*ABCD*] transmission matrix for the error block can be formulated by:

$$X_i = \begin{bmatrix} a_i & b_i \\ c_i & d_i \end{bmatrix} = \begin{bmatrix} \bar{I}_{iO}/\bar{I}'_{iS} & -1/\bar{I}'_{iS} \\ -\bar{I}_{iS}\bar{I}_{iO}/(\bar{I}_{iS}-\bar{I}_{iO}) & \bar{I}'_{iS}/(\bar{I}_{iS}-\bar{I}_{iO}) \end{bmatrix} \tag{15}$$

This matrix chains, corresponding to the error box, correct all parasitic effects brought by the two discontinuities of the port of non-ideal excitations and the two lines of excitation in the dynamic model of the equivalent circuit; and could be determined precisely by the three numerically measured currents while using the iterative method: I_{iO}, I_{iS} and I'_{iS}. With this matrix calculated meadow, the equivalent current I'_i and the equivalent tension V'_i, define on the ports of excitations of the circuit to study that is characterized by the matrix[ABCD], express themselves according to tensions and currents, applied to the whole circuit I_i and V_i with (i=1,2), and the error matrix by the following relation:

$$\begin{bmatrix} V_i' \\ I_i' \end{bmatrix} = [X_i] \begin{bmatrix} V_i \\ I_i \end{bmatrix} \tag{16}$$

After giving the matrixes values of the error box, we can now extract the real matrix of the circuit to study it in term of ABCD matrix; indeed the setting in cascade of the three matrixes gives by the following equation:

$$\begin{bmatrix} A_{circuit} & B_{circuit} \\ C_{circuit} & D_{cicuit} \end{bmatrix} = [X_1] \begin{bmatrix} A & B \\ C & D \end{bmatrix} [X_2] \tag{17}$$

With:

$\begin{bmatrix} A_{circuit} & B_{circuit} \\ C_{circuit} & D_{circuit} \end{bmatrix}$: The relative matrix of the whole circuit

$[X_i]$: The relative matrix of the error box

$\begin{bmatrix} A & B \\ C & D \end{bmatrix}$: The real matrix of the circuit to study

Therefore,

$$\begin{bmatrix} A & B \\ C & D \end{bmatrix} = [X_2]^{-1} \begin{bmatrix} A_{circuit} & B_{circuit} \\ C_{circuit} & D_{circuit} \end{bmatrix} [X_1]^{-1} \tag{18}$$

For the reason that the elements of matrix $[X_1]$ and $[X_2]$ are normalized by the current I'_{iS}, then the determinant of these matrix is equal to 1. While replacing $[X_1]$ and $[X_2]$ by their expressions in the equation (15), we gets:

$$\begin{bmatrix} A & B \\ C & D \end{bmatrix} = \begin{bmatrix} \frac{\bar{I}'_{2S}}{(\bar{I}_{2S}-\bar{I}_{2O})} & \frac{1}{\bar{I}'_{2S}} \\ \frac{\bar{I}'_{2S}\bar{I}_{2O}}{(\bar{I}_{2S}-\bar{I}_{2O})} & \frac{\bar{I}_{2S}}{\bar{I}'_{2S}} \end{bmatrix} \begin{bmatrix} A_{circuit} & B_{circuit} \\ C_{circuit} & D_{circuit} \end{bmatrix} \begin{bmatrix} \frac{\bar{I}'_{1S}}{(\bar{I}_{1S}-\bar{I}_{1O})} & \frac{1}{\bar{I}'_{1S}} \\ \frac{\bar{I}'_{1S}\bar{I}_{1O}}{(\bar{I}_{1S}-\bar{I}_{1O})} & \frac{\bar{I}_{1S}}{\bar{I}'_{1S}} \end{bmatrix} \tag{19}$$

In Figure (7) we proposed the algorithm result of this SOC-WCIP method; where we present a schematic description which summarizes the main operation of this method.

APPLICATION

To evaluate the numerical complexity of the procedure described in paragraph III, we considered the structure of the low pass filter is presented in the Figure (8) and we have the relative dielectric constant of 4.23 and thickness of 1.20 mm. Two coupled micro strip lines, that have a characteristic impedance of 50 _, are used as an input and output signal. In Figure (9), the amplitude of the S-parameters is illustrated as a function of the iteration number that validates this new SOC-WCIP method. As seen the convergence is achieved in 200 iterations. The results of the WCIP method combined with SOC technique are compared to those obtained only with the WCIP method. Figure (10) and (11) show respectively the amplitude and the phase of the *S*11 parameter as function of frequency. They demonstrate that the cutoff frequency is translated until 4.2 GHz in case of hybrid approach

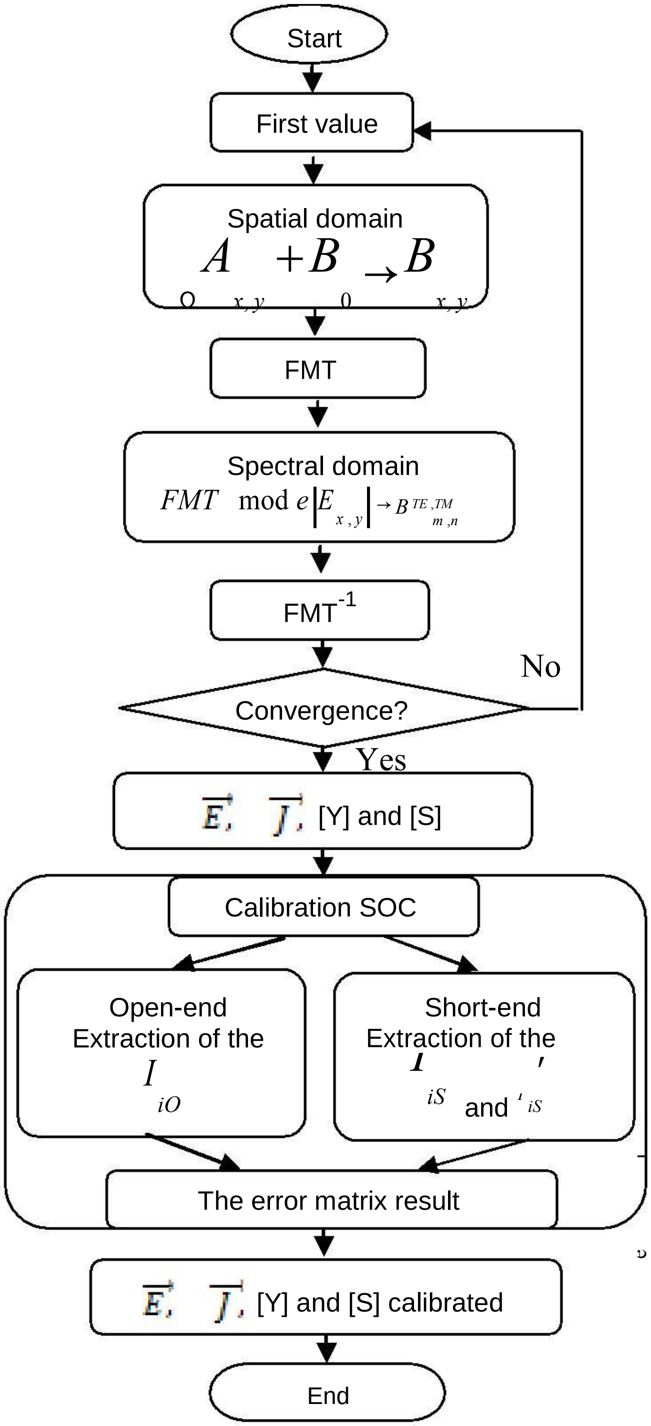

Figure 7. Algorithm result.

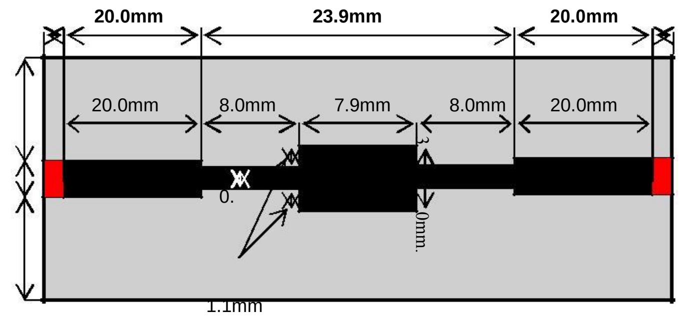

Figure 8. Geometry of the low pass filter.

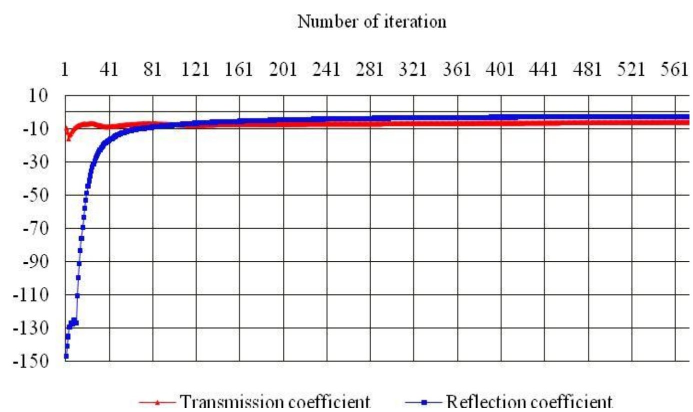

Figure 9. Convergence of the S-parameters according to the iterations for f = 2 Ghz.

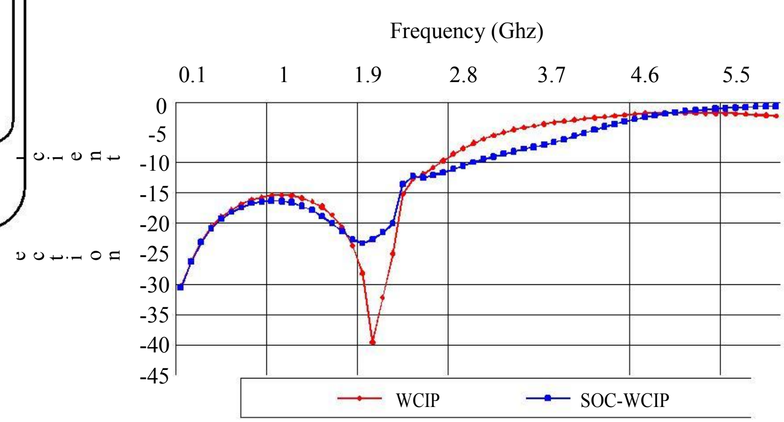

Figure 10. Amplitude of the reflection coefficient S_{11} for the low pass filter.

(WCIP-SOC) since the iterative only shows 3.3 GHz as cutoff frequency.

Whereas, the amplitude and phase of the *S*21 parameter are shown respectively in Figures (12) and (13) and demonstrate the response of our proposed filter in case of iterative approach (WCIP) and hybrid method WCIP-SOC. Here, results prove the theoretical approach given by Zhenwang et al. (2005) but novelty is shown in pass band what is located between 0.1 and 3.6 GHz in case of iterative approach and that will be translated until the 4.4 GHz frequency in case of combined method WCIP-SOC. These later result proves the effectiveness of this new hybrid approach what ameliorate the pass band

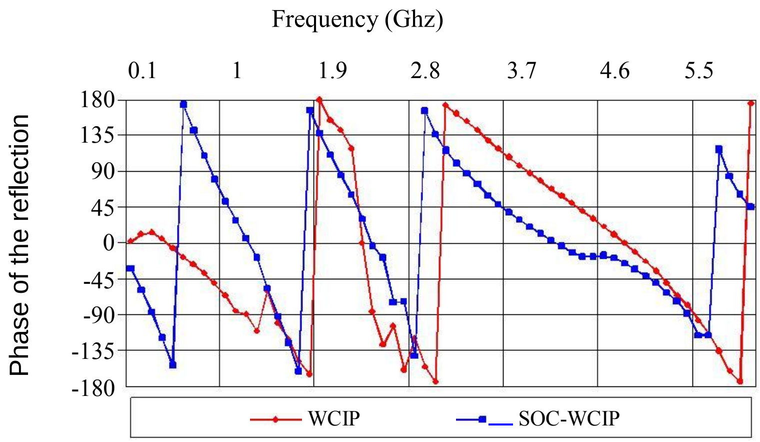

Figure 11. Phase of the reflection coefficient S_{11} for the low pass filter.

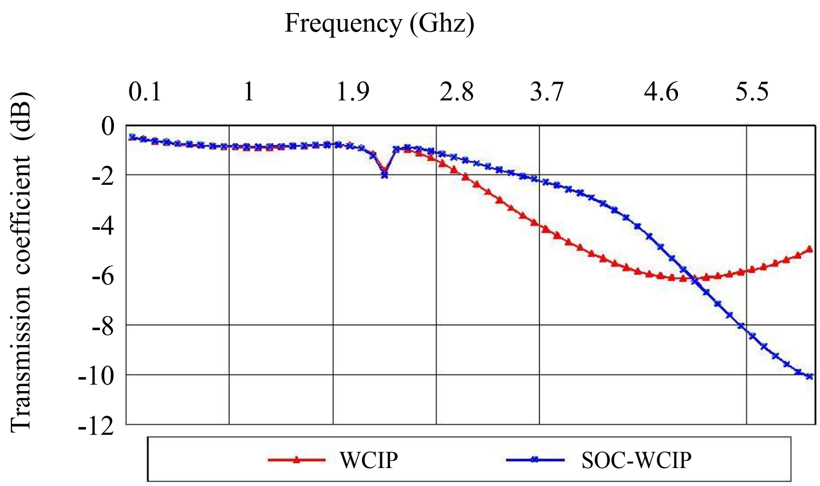

Figure 12. Amplitude of the transmission coefficient S_{11} for the low pass filter.

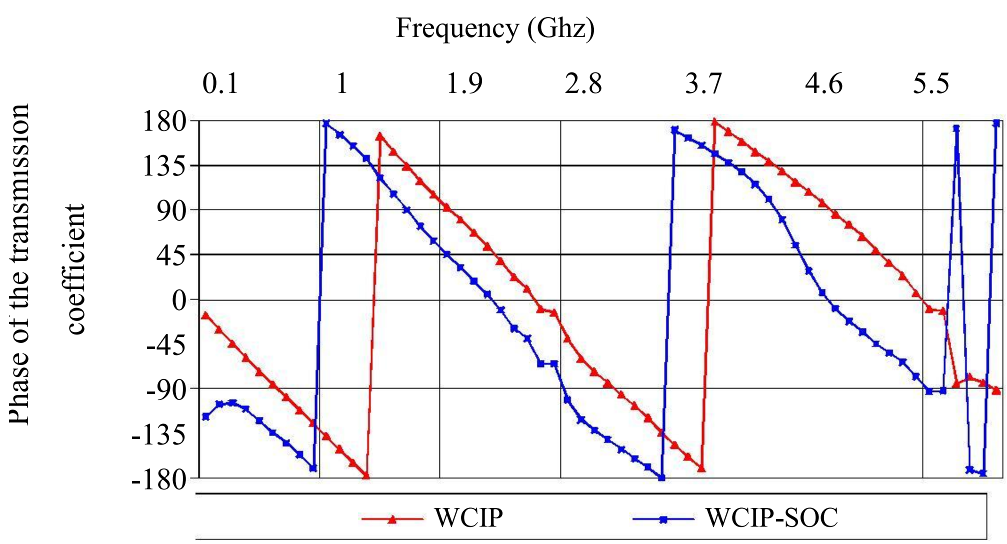

Figure 13. Phase of the transmission coefficient S_{11} for the low pass filter.

spectrum. Other advantage of this new technical approach is the peak of frequency what is increasing.

CONCLUSION

In this paper, we have overviewed the new combined method WCIP-SOC; indeed we have presented the state-of-the-art of the iterative method (WCIP) and the basic element of the numerical calibration (SOC).The future investigations concerning this numerical calibration are applied this combined method for other structure and extraction of each equivalent schema exact without the limitation of the excitation part and the discontinuities between port of excitation and the real structure.

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