An Efficient Systematic Approach to Model Extraction for Passive Microwave Circuits

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***Abstract*—This paper introduces a new systematic approach to equivalent circuit model extraction for linear microwave passive circuits directly from full-wave frequency domain simulation. The devices being modeled may be either lossless or lossy. Adaptive frequency sampling is used to minimize the computational effort of EM simulation while critically assisting in determining the pole locations of an RF circuit. A simple circuit model for lossy RF circuits along with a determined starting point of optimization of lumped element component values is also presented in detail. The overall result is an efficient and accurate means to produce a complete equivalent lumped element model for RF circuits that is suitable for use in conventional SPICE-like simulation software.**

***Index Terms*—Adaptive frequency sampling, equivalent model, microstrip components, microstrip discontinuities, parameter extraction, RF circuit.**

I. Introduction

WITH the increasing advent of high-speed mixed signal circuits, maintaining signal integrity has become one of the major issues facing modern circuit design. In order to accommodate this trend and perform overall system simulation accurately requires one mixed signal analysis of conventional digital circuits with RF/microwave distributed circuits. This may be accomplished through one of two approaches. The first is to integrate the digital circuit segment into a full-wave electromagnetic (EM) analysis. Preliminary attempts have been made [1]. The second approach to overall system analysis is to extract an equivalent circuit model from a full-wave EM simulation of the distributed circuit section and embed the model in a conventional circuit simulation [2], such as SPICE. In order to preserve computational effort, the most efficient means of system simulation is to minimize the amount of full-wave EM simulation. Typical conventional circuit simulation time is insignificant in comparison to full-wave EM methods. The obvious avenue is to extract an accurate equivalent model from the full-wave response and implement the resulting equivalent circuit via SPICE-like simulation techniques.

Although many equivalent circuit models exist for commonly used RF circuits, most are either derived from empirical formulas or constructed based on physical intuition [2], [3]. The common deficiency in many microwave circuit models is that they are only valid within a limited bandwidth. The major reason is that it is often difficult to determine the number of poles and their locations for an RF circuit in a given frequency band simply by physical intuition. This limitation restricts many accepted models usage in high-speed applications where wide spectra of information is needed. Therefore, in order to perform efficient system-level simulation of high-speed digital/RF mixed signal circuits, it is crucial to develop a more general, accurate, and systematic approach for model extraction of arbitrary passive RF circuits.

Recently, a general approach to extract lumped element equivalent circuits of RF linear circuits for an arbitrary but finite frequency band has been proposed [4]. The significance of the work is that it constructs the topology of the model by matching the properties of the poles instead of by physical intuition. Although this approach has demonstrated its potential in relevant applications, several serious drawbacks have to be amended to enable the approach more efficient, practical, and easy to use: 1) the approach requires the use of time domain EM simulation techniques such as FD-TD or TLM. Therefore, commonly used EM simulators developed with frequency domain techniques are not easily used; 2) the model for lossy RF circuits is complicated and cumbersome such that the positive real (PR) [5] condition for a realizable network is difficult to be reinforced; and 3) the approach does not provide a systematic means to determine the initial starting point of the circuit parameters so that extensive optimization is required to grope the solution.

In this work it will be shown that a systematic approach can be implemented upon results obtained directly from frequency domain EM simulations to extract an equivalent lumped element circuit model for lossless and lossy linear passive structures. This approach is based on the Cauer network synthesis technique [6] using full-wave EM frequency domain analysis results and an adaptive frequency sampling (AFS) algorithm designed for wide band applications to minimize computation. To improve the efficiency, a simple circuit model for lossy RF circuits is developed. The model is applicable across a sufficient wide frequency bandwidth so that it dictates the major characterization of an RF circuit in the time domain. The model is obtained quickly and accurately without requiring extensive computations from either EM simulation or optimization techniques.

The extraction approach uses a general circuit topology that is fundamentally fixed in form but capable of representing structures with an arbitrary number of poles. The resultant equivalent circuit can be used in any circuit simulator for digital system applications. The overview of this paper begins with the Cauer expansion for lossless circuits and an outline of the equivalent circuit topology, which are both extended to the lossy case. The wide band adaptive sampling algorithm is then presented, and an

Manuscript received June 16, 1999.
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Publisher Item Identifier S 0018-9480(00)07402-0.

effective means of obtaining the starting point of optimization for the lumped element parameter values is illustrated. The validation of the proposed approach is then demonstrated through three numerical examples.

II. Cauer Expansion for Lossless Circuits

The proposed approach uses Cauer's network synthesis technique [7]. To describe the concept, only two port passive circuits will be illustrated. The concept can easily be extended to a multiple-port network [4]. Synthesis begins with S-parameters of the device being modeled. These can be obtained from either measurements or directly from full-wave frequency domain EM simulation. To represent the acquired information in a physically meaningful format, the S-parameters need to be converted to the impedance parameters.

It is postulated that for a passive linear circuit, the impedance parameter Z_{ij} can be characterized by the following partial fraction expansion [6]:

$$Z_{ij}(s) = \frac{k_{ij}^{(0)}}{s} + \sum_{m=1}^{M} \frac{2k_{ij}^{(m)} s}{s^2 + \omega_m^2} + \cdots + k_{ij}^{(\infty)} s \tag{1}$$

where

s	complex frequency;
$k_{ij}^{(0)}$ and $k_{ij}^{(\infty)}$	residues of poles at zero and infinity, respectively;
the mth term in the summation	contribution from the intermediate pole of the complex frequency ω_m.

Since the approximation of a lossless circuit is made upon fitting these partial fraction forms to the obtained EM simulation results, all the residues k_{ij} are real and poles are approximated to lie along the imaginary axis. For a realizable network, functions of Z_{11} and Z_{22} must be positive real (PR). Thus the residues for these expansions are anticipated to be positive. Z_{12}, however, need not be a PR function and therefore its' residues can be either positive or negative. Regardless, for a realizable network, the residue condition must be satisfied

$$k_{11}^{(v)} k_{22}^{(v)} - (k_{12}^{(v)})^2 \geq 0. \tag{2}$$

III. The Lossless Equivalent Network

Once the Cauer expansions have been obtained for each impedance parameter, we may use the residues and poles to determine the equivalent circuit component values. For each term in the Cauer expansion, there will be a corresponding sub T-network. Thus for a one-pole linear circuit, the equivalent circuit may contain three sub T-networks, one for the residue at zero frequency, one for the residue at infinite frequency, and one for the intermediate pole occurring at the frequency ω_1. Such an equivalent network would take the form as shown in Fig. 1.

In Fig. 1, the T-network composed of capacitors corresponds to the pole at zero frequency, the inductor T to the pole at infinite frequency, and the LC tanks of T-network represent the contribution of an intermediate poles. Each LC tank T-network

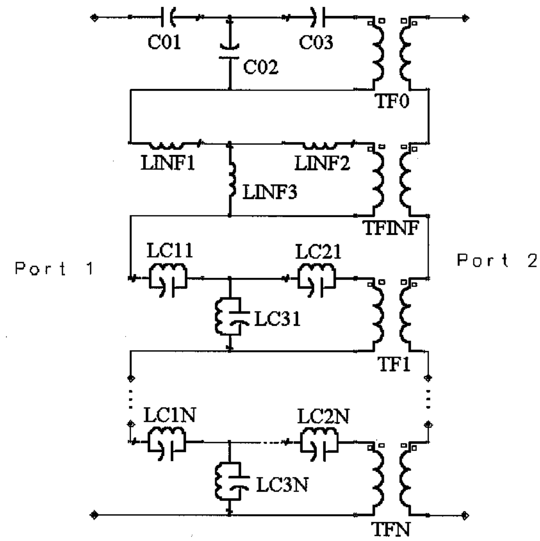


Fig. 1. The fixed network topology of Cauer expansion for lossless circuits.

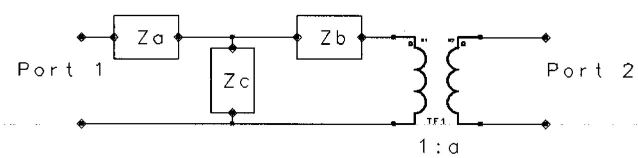


Fig. 2. The general T-network with transformer turns ratio 1 : a.

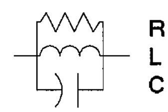


Fig. 3. The proposed Cauer T-network component for an intermediate pole of a lossy structure.

corresponds to each intermediate pole in the band of interest. Effects of each branch of the sub T-network on Z_{11}, Z_{12} and Z_{22} can be represented using the T-network illustrated in Fig. 2.

Using the general two-port arrangement, we can illustrate how each branch contributes to the input and transmission parameters. This form may be expanded to an arbitrary number of ports [6]. However, the two-port form may be expressed in reference to the notation illustrated in Fig. 2

$$Z_{11} = Z_a + Z_c \tag{3a}$$

$$Z_{12} = aZ_c \tag{3b}$$

$$Z_{22} = a^2 Z_b + a^2 Z_c \tag{3c}$$

where Z_a, Z_b, and Z_c represent either an inductor, capacitor, or an LC tank. By using this form, it is possible for equivalent models to be produced which represent symmetrical or nonsymmetrical devices. For the process of model extraction, we are interested in determining a set of branch components that produce the same response obtained from full-wave EM analysis. Once the Cauer expansion has been obtained, we may use the residue values of each Z_{11}, Z_{12}, and Z_{22} function for each pole to calculate the equivalent lumped element circuit parameters. In order for the equivalent network to be realizable, we start with the fact that the transformer turns ratio, $a^{(m)}$, for an arbitrary pole m must be contained within a certain range [7]

$$\frac{\left|k_{12}^{(m)}\right|}{k_{11}^{(m)}} \le a^{(m)} \le \frac{k_{22}^{(m)}}{\left|k_{12}^{(m)}\right|} \tag{4}$$

where k_{ij} are the residues of pole m. Furthermore, for realizability the residue condition also dictates that the turns ratio must comply with the following equation [7]:

$$\frac{k_{12}^{(m)}}{a^{(m)}} \ge 0. \tag{5}$$

Once an acceptable turns ratio $a^{(m)}$ that meets the requirements imposed by (4) and (5) for each pole m has been chosen, the impedance contribution of each branch in the equivalent T network corresponding to pole m may be obtained from the following relations:

$$Z_a^{(m)} = \left[k_{11}^{(m)} - \frac{k_{12}^{(m)}}{a^{(m)}}\right] f_a(s) = A_a^{(m)} f(s) \tag{6a}$$

$$Z_b^{(m)} = \left[\frac{k_{22}^{(m)}}{\left(a^{(m)}\right)^2} - \frac{k_{12}^{(m)}}{a^{(m)}}\right] f_a(s) = A_b^{(m)} f(s) \tag{6b}$$

$$Z_c^{(m)} = \left[\frac{k_{12}^{(m)}}{a^{(m)}}\right] f_a(s) = A_c^{(m)} f(s) \tag{6c}$$

where

$f(s)$	of the form $1/s$ for a pole at zero;
s	for a pole at infinity;
$2s/(s^2+\omega_m^2)$	for an intermediate pole, as outlined in the Cauer expansion.

Once the T-network branch impedance coefficients have been determined, it is possible to directly calculate the values of the lumped elements. For a pole at zero, the capacitance of branch x is given by

$$C_x^{(0)} = \frac{1}{2\pi A_x^{(0)}}. \tag{7}$$

Similarly for a pole at infinity, the inductance of branch x is given by

$$L_x^{(\infty)} = \frac{A_x^{(\infty)}}{2\pi}. \tag{8}$$

For the intermediate pole at frequency ω_m, the values of the LC tank in branch x are given by

$$C_x^{(m)} = \frac{1}{4\pi A_x^{(m)}} \tag{9a}$$

$$L_x^{(m)} = \frac{A_x^{(m)}}{\pi\omega_m^2}. \tag{9b}$$

Via the above procedure, it is demonstrated that the general T-network can be used to produce a lossless lumped element equivalent model for a passive linear circuit.

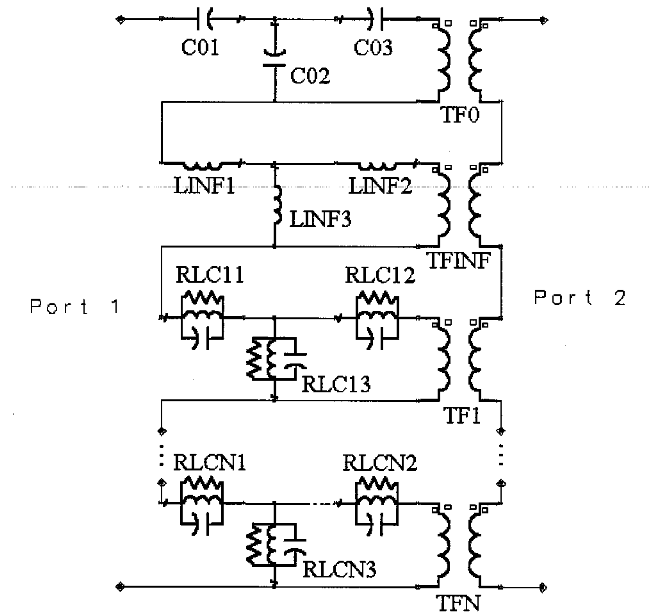


Fig. 4. The proposed fixed network topology of Cauer expansion for lossy circuits.

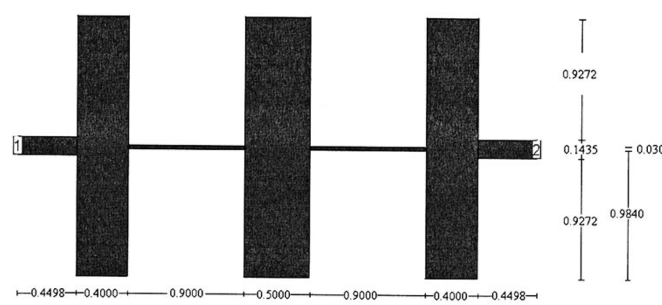


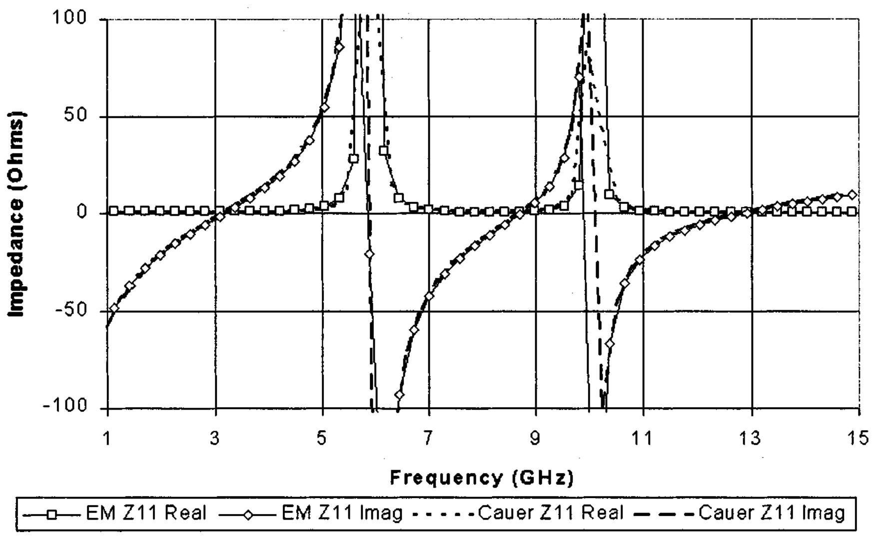
Fig. 5. A low pass microstrip filter. The dielectric constant $\varepsilon_r = 10$ and the height of the dielectric plate $h = 0.2000$ mm.

IV. The Lossy Equivalent Network

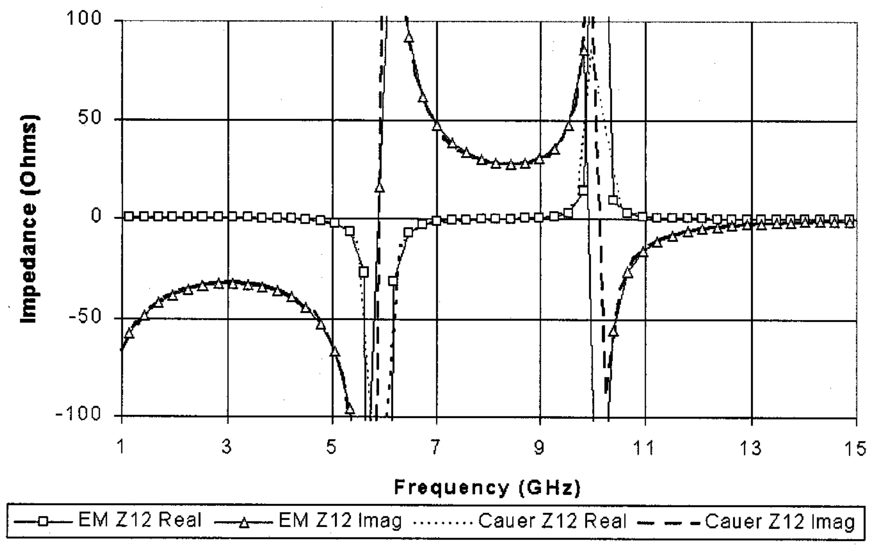
Since the technique outlined above is for lossless networks, it becomes critical for the general case to accommodate circuits that exhibit lossy characteristics. It is desirable to prescribe the lossy case in the same fashion as the lossless, using sub T-networks and transformers of fixed turns ratio between ports. This allows effects such as radiation loss of a microstrip circuit at intermediate poles ω_m to be modeled effectively. We assume that the Cauer expansion of an impedance matrix element Z_{ij} near a lossy pole ω_m can be characterized primarily by the following expression:

$$Z_{ij}(s) \approx \frac{2k_{ij}^{(m)} s}{s^2 + (\omega_m')^2} \tag{10}$$

where ω_m' is a complex pole. A complex pole location is characteristic of a lossy device. A simple modification to the lossless LC tank for a resistive property could take the form of Fig. 3. Consequently, the complete equivalent network can be represented by the structure shown in Fig. 4.



(a)



(b)

Fig. 6. (a) and (b): Z_{11} and Z_{12} parameters for the low pass filter of Fig. 4 obtained from full-wave EM simulation and their respective Cauer expansions.

Returning to first principles, the above structure can be characterized near an arbitrary pole location for any Z_{ij} by

$$Z_{ij}(s) = \frac{\frac{s}{C}}{\frac{1}{LC} + \frac{s}{RC} - s^2} \tag{11}$$

which modifies the lossless Cauer expansion term of (1) to be

$$Z_{ij}(s) \approx \frac{2k_{ij}^{(m)} s}{s^2 + \omega_m^2 + \frac{js}{B^{(m)}}}. \tag{12}$$

To obtain the tank resistor values for a lossy circuit, it is assumed the transformer turns ratio is the same for the real and imaginary $Z_{ij}(s)$. In other words, the turns ratio for the lossy nature of the device between ports exists in approximately the same proportion as the lossless case. Thus the lossy case is simply an extension of the lossless. Once a lossless approximation is made through fitting the Cauer expansion to the imaginary part of impedance, simply adding a linear term to the denominator will account for a lossy structure. For each branch of the equivalent sub T-network the resistor value is easily calculated by

$$R_x^{(m)} = \frac{B^{(m)}}{C_x^{(m)}} \tag{13}$$

in which $C_x^{(m)}$ for each T-network branch is already determined from the lossless Cauer expansion. $B^{(m)}$ is determined directly from the real part of impedance obtained from EM simulation.

TABLE I
ELEMENT VALUES OF THE EQUIVALENT CIRCUITS IN NUMERICAL EXAMPLES, WHERE CAPACITANCE IS IN pF, RESISTANCE IS IN OHMS AND INDUCTANCE IN NANOHENRYS

	EXAMPLE 1	**EXAMPLE 2**	**EXAMPLE 3**
C_{01}	4439.36	131.7109	87.41037
C_{02}	4436.83	143.9383	90.88498
C_{03}	2.53298	12.22744	3.474608
TF_0	1.0000	1.004065	.9888998
$L_{\infty 1}$	.291125	0.86368	.5105067
$L_{\infty 2}$	.302425E-7	0.03553157	.08372643
$L_{\infty 3}$	.302425E-7	0.037056	.1001520
TF_∞	-2194.097	3.519946	-1.412827
L_{11}	0.004574	0.0090648	.002112553
R_{11}	8.913788	75.28294	3.978740
C_{11}	159.1113	205.6096	1786.067
L_{12}	0.004529	0.0089954	.002115179
R_{12}	8.82662	74.7065	3.983686
C_{12}	160.6825	207.196	1783.850
L_{13}	.463201	1.174845	1.701795
R_{13}	902.6826	9757.029	3205.127
C_{13}	1.571188	1.586435	2.217166
TF_1	-0.999953	.9353968	-1.122033
L_{21}	0.00103827	0.00423324	.006328008
R_{21}	3.486293	38.6098	25.33494
C_{21}	240.6048	366.2818	134.7907
L_{22}	0.0010209	0.00421866	.006201302
R_{22}	3.42804	38.47676	24.82765
C_{22}	244.6933	367.5482	137.5448
L_{23}	0.061101	1.224336	.3097070
R_{23}	205.1645	11166.69	1239.949
C_{23}	4.088518	1.266451	2.754076
TF_2	.9989993	-1.052925	1.505813
L_{31}			.001645911
R_{31}			8.892608
C_{31}			288.4778
L_{32}			.001660050
R_{32}			8.969002
C_{32}			286.020
L_{33}			.1932377
R_{33}			1044.034
C_{33}			2.457123
TF_3			-1.412688
L_{41}			.0009977455
R_{41}			7.082305
C_{41}			276.9706
L_{42}			.0009891258
R_{42}			7.021120
C_{42}			279.3842
L_{43}			.1144938
R_{43}			812.7126
C_{43}			2.413633
TF_4			.8747841

The lossless approximation of the Cauer expansion is not affected by this linear term in the denominator. The case for which loss occurs near the intermediate pole location ω_m can be evaluated after the residue $k_{ij}^{(m)}$ has been determined.

V. ADAPTIVE FREQUENCY SAMPLING (AFS)

To implement the Cauer synthesis technique, it is essential that the pole location along the imaginary frequency axis be accurately determined. Thus, a large number of frequency samples are required across the frequency band of interest. To enhance the efficiency, a multisectional adaptive frequency sampling (AFS) [8] is developed so that a minimal amount of EM simulation is required to characterize an RF circuit. The concept of AFS is that through sampling a system in the frequency domain adaptively one models the frequency response with a rational function. Once this function is determined, it can be used to generate an arbitrary number of frequency domain responses, to locate the poles of the system described by the Z-parameters. The rational function used takes the following form:

$$S_{ij}(f) = \frac{\sum_{k=0}^{n} c_k f^k}{1 + \sum_{l=1}^{m} d_l f^l}. \tag{14}$$

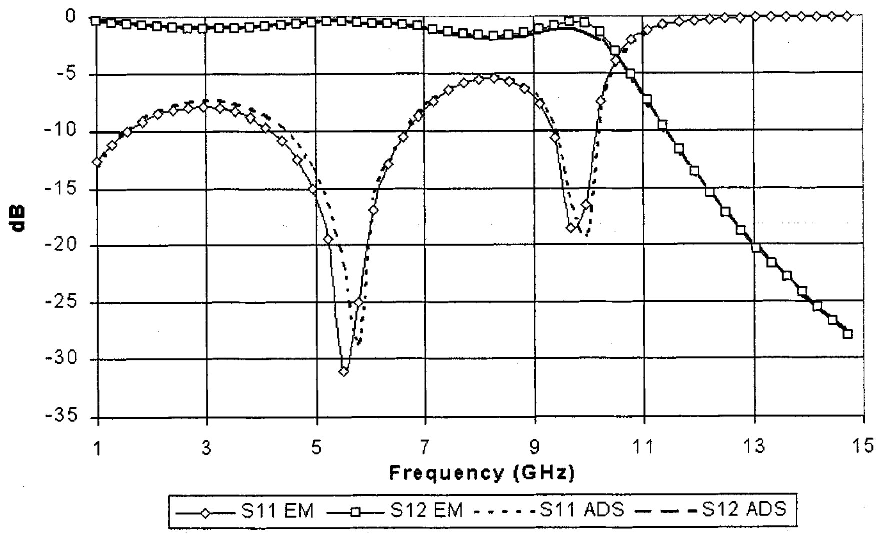
The frequency is denoted by f, and c_k and d_l are the coefficients of the polynomials of the numerator and denominator, respectively. Modeling is accomplished by designating a preliminary number of samples, perhaps at equal intervals, across the full band of interest. Two models are generated whereby the numerator and denominator of one differ in degree by one in comparison to the other. The S-parameters of the models are then compared piecewise. The point of greatest difference between the two rational functions constitutes the next sampling point for EM simulation. By definition, this point will be contained in the next pair of functions for $S_{ij}(f)$, of which both would be of one degree higher than the previous set. The numerator and denominator are increased in degree in an alternating fashion. Ideally, the two models will converge to a common result.

This technique was originally proposed for narrow band modeling. When used for wide band characterization, over sampling may produce an ill-conditioned matrix. To prevent this, the original model is divided into a number of separate models, each of which is developed individually across their respective frequency band. EM samples of previous wider band models are recycled into the next generation of models, thus no computational effort is wasted.

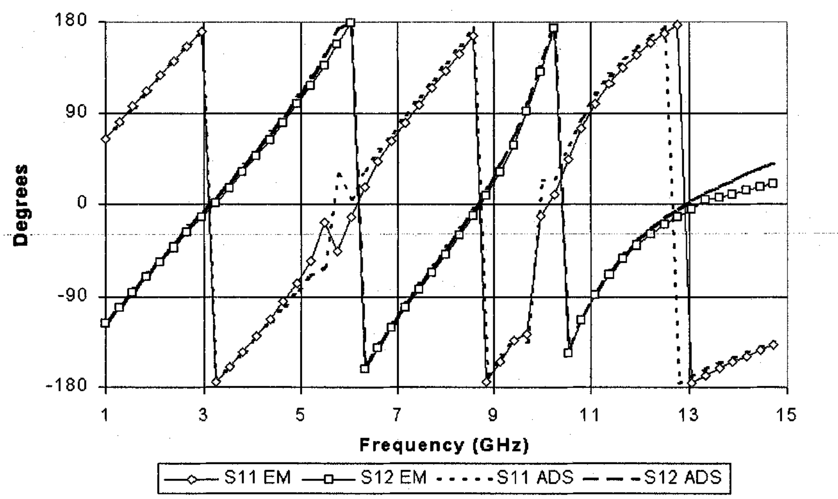
The most important advantage of using the rational function for EM full-wave frequency domain response is that poles of the Z-parameters are easily located accurately and with a minimal amount of full-wave simulation. This is most apparent when the modeled S-parameters are converted to impedance parameters. A large number of sampling points near a pole location finds extreme values of impedance as the magnitude of the function near a pole approaches infinity.

VI. INITIAL APPROXIMATION OF THE CAUER EXPANSION AND OPTIMIZATION

Since AFS allows generation of a large number of samples, we may use this ability to assist in the location of intermediate



(a)



(b)

Fig. 7. (a) and (b): Equivalent circuit S-parameters from HP ADS RF circuit simulator compared to the full-wave EM simulated S-parameters for S_{11} and S_{12} for the low pass filter of Fig. 5.

poles along the imaginary frequency axis. This is a very good approximation for the lossy case, since the real component of the complex pole ω_m for most applications is quite small. Once the approximate pole locations along the imaginary axis are found, a set of linear equations can be established for the lossless Cauer expansion to give initial residue values. Re-examining the Cauer expansion

$$Z_{ij}(s) = \frac{k_{ij}^{(0)}}{s} + \sum^{M} \frac{2k_{ij}^{(m)} s}{D_m(s)} + \cdots + k_{ij}^{(\infty)} s. \qquad (15)$$

Once ω_m for all m are located accurately the denominator $D_m(s)$ at a point near the pole can be effectively made a constant. Using the original $Z_{ij}(s)$ parameters obtained from full-wave EM analysis, we may create a set of linear equations and solve for initial residue values directly. Afterward, a minimal amount of optimization is anticipated to adjust the residues of the expansion to best match the original EM simulation results.

The scheme for optimization is simple and straightforward. The starting point for the partial fraction expansion provided resembles strongly the impedance function obtained from the full-wave analysis S-parameter results. A gradient-based scheme is implemented such that the residues of all impedance functions and the corresponding linear terms in each dominator are adjusted such that the Cauer expansion resembles the original EM simulation results as closely as possible. The error function is evaluated by converting the impedance parameters of the Cauer expansion to scattering parameters for each

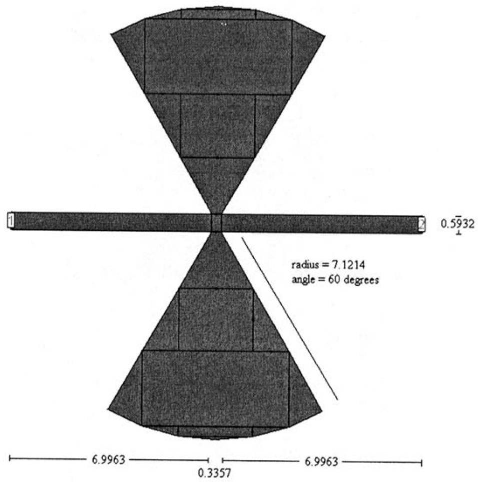


Fig. 8. A radial stub microstrip filter. All dimensions are in millimeters. $\varepsilon_r = 10$ and $h = 0.6350$ mm.

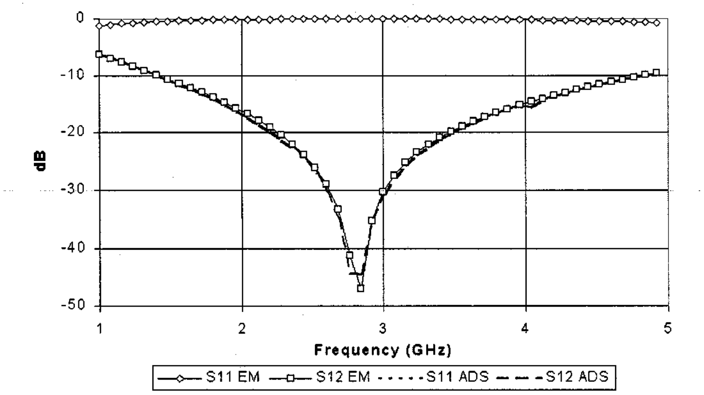
iteration. This approach is best suited as the extreme values of impedance near each intermediate pole ω_m contribute overwhelmingly when evaluating the error function through impedance parameters.

VII. Numerical Results

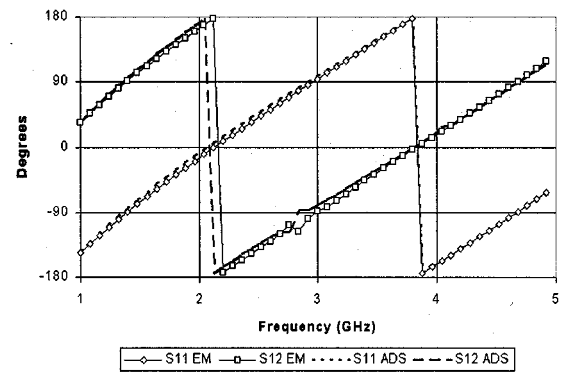
The following three two-port examples are given to illustrate the extraction process outlined above. The full wave analysis results were obtained from the moment method simulation package IE3D from Zeland Software. AFS was used to sample the S-parameters across the frequency band of interest, and the model extraction process implemented directly upon these results. No user intervention was required. To obtain the equivalent circuit, it is essential that the impedance parameters of the full-wave analysis be de-embedded.

The first example is a microstrip low pass filter as illustrated in Fig. 5. Using AFS for this example, an arbitrary number of S-parameter results can be obtained from 1 to 15 GHz with only ten full-wave EM simulation samples in the frequency domain. The S-parameters are sufficient for not only determining the locations of poles but also the extraction of the residues. The AFS process in this case began with seven initial points spread equally from 1 to 15 GHz. Converting the S to Z-parameters with 50 Ω line impedance, one can easily obtain the Cauer expansion. Fig. 6 compares the synthesized Cauer expansion to the results obtained directly from EM simulation. The pole locations are easily identified once the S-parameters are converted to Z-parameter form. Two intermediate poles within the band of interest are observed at 5.9011 and 10.0021 GHz. The effects of the pole at zero can be seen at lower frequencies in all Z-parameters, as well as the effects of the pole at infinity, observable at high frequencies. Synthesized element values are listed in Table I.

Using HP ADS the extracted circuit is simulated for the purpose of comparison to the original full-wave EM analysis S-parameters. The results are presented in Fig. 7. Good agreement



(a)



(b)

Fig. 9. (a) and (b): S_{11} and S_{12} comparisons of extracted model HP ADS S-parameter simulation compared to the original full-wave EM simulation for the radial stub filter depicted in Fig. 8.

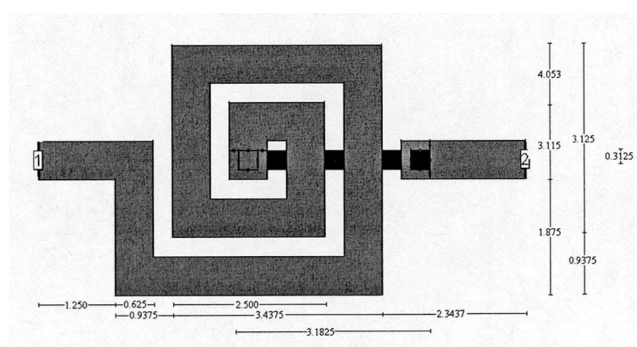
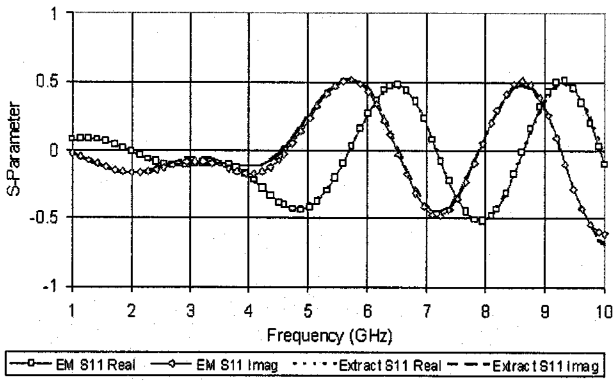


Fig. 10. A spiral inductor. All dimensions are in millimeters. $\varepsilon_r = 9.8$ and $h = 0.6350$ mm.

can be observed for the S-parameter comparison of the equivalent circuit model to the original full-wave EM simulation.

The second example presented is a microstrip radial stub filter. It is illustrated in Fig. 8. Performing the model extraction in the same manner as that of the low pass filter, two intermediate poles are observed in the 1 to 7 GHz frequency band. The



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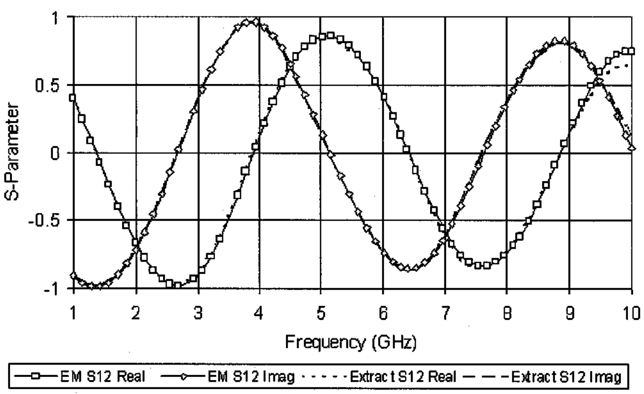

(b)

Fig. 11. (a) and (b): S_{11} and S_{12} parameters of the HP ADS equivalent circuit simulation compared to the full-wave EM simulation results for the spiral inductor illustrated in Fig. 10.

poles occur in the band of interest at 3.6867 and 4.0408 GHz. The S-parameters comparing the full-wave simulation results to the equivalent model are depicted in Fig. 9. Good agreement is obtained. Synthesized element values for this example are also listed in Table I.

The final example of a spiral inductor as illustrated in Fig. 10 is presented to demonstrate the proposed extraction method. The device produces four poles over the frequency range of 1–10 GHz. They are located at 2.5910, 5.4495, 7.3040, and 9.5470 GHz. Performing the same model extraction process, we obtain an equivalent circuit model and element values that are listed in Table I. S-parameter comparisons for the HP ADS equivalent model simulation and full-wave EM simulation are presented in Fig. 11. Since the fourth pole of the spiral inductor is near the upper end of the band of interest, it is difficult to fit an accurate pole at infinity residue for the Cauer expansion. As can be observed from Fig. 11, this effect becomes more pronounced as frequency increases. For this case modeling response beyond the range of the 9.5470 GHz pole would result in a more accurate equivalent circuit model.

VIII. Conclusion and Discussions

The Cauer network synthesis technique proves to be effective in providing an efficient approach for equivalent model extraction from frequency domain simulation results. The adaptive frequency sampling and the linear equation method for the initial residue values are essential in producing an effective and accurate model directly from full-wave EM frequency domain simulation results. This has been demonstrated by several typical two port microstrip circuit structures. Since this technique is applicable to the lossless and lossy cases, the approach can easily be implemented on any commercial EM simulation software, efficiently and systematically producing accurate equivalent circuit models for passive linear RF circuits.

It is worth mentioning that for a very high-frequency application, in which cross couplings occur, the Cauer's network may not be able to represent the characteristics of the RF circuit pertinently. In this case, some appropriate lumped coupling elements need to be added to accommodate the effect.

Acknowledgment

The authors would like to thank Dr. T. Mangold of Technische Universitat München, Germany, for his detailed explanation of his published works.

References

[1] M. Piket-May, A. Taflove, and J. Baron, "FD-TD modeling of digital signal propagation in 3-D circuits with passive and active loads," *IEEE Trans. Microwave Theory Tech.*, vol. 42, pp. 1514–1523, 1994.

[2] J.-G. Yook, L. P. B. Katehi, K. A. Sakallah, R. S. Martin, L. Huang, and T. A. Schreyer, "Application of system-level EM modeling to high-speed digital IC packages and PCB's," *IEEE Trans. Microwave Theory Tech.*, vol. 45, pp. 1847–1856, 1997.

[3] P. L. Werner, R. Mittra, and D. H. Werner, "Extraction of equivalent circuits for microstrip components and discontinuities using the genetic algorithm," *IEEE Microwave Guided Wave Lett.*, vol. 8, pp. 333–335, 1998.

[4] T. Mangold and P. Russer, "Full-wave modeling and automatic equivalent-circuit generation of millimeter-wave planar and multilayer structures," *IEEE Trans. Microwave Theory Tech.*, vol. 47, pp. 851–858, 1999.

[5] F. F. Kuo, *Network Analysis and Synthesis*, New York: Wiley, 1962, pp. 255–276.

[6] W. Cauer, *Synthesis of Linear Communication Networks*: McGraw-Hill, 1958, vol. 1 and 2, pp. 181–220.

[7] M. E. Van Valkenburg, *Introduction to Modern Network Synthesis*, New York: Wiley, 1960, pp. 297–322.

[8] T. Dhaene, J. Ureel, N. Fache, and D. De Zutter, "Adaptive sampling algorithm for fast and accurate S-parameter modeling of general planar structures," in *IEEE MTT-S Int. Microwave Symp. Dig.*, 1995, pp. 1427–1430.

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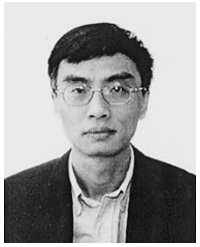

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