An Analytical Approach to Computer-Aided Diagnosis and Tuning of Lossy Microwave Coupled Resonator Filters

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***Abstract*—This paper presents a novel analytical approach to extracting of the coupling matrix of a narrow band general Chebyshev bandpass filter with losses. The approach is very useful for computer aided tuning of a microwave bandpass filter. The concept of phase loading is revealed for the first time in the community of computer aided tuning. The analytical approach consists of three elements: 1) a theoretic formula that leads to a practical scheme for determining the phase loading; 2) a theoretic formula for de-embedding the section of unknown transmission lines at the two ports of a filter; and 3) a theory for determining the unloaded Q of a filter if the loss for each resonator is nearly the same. To make the approach easy to use, some practical techniques for reconstructing rational functions of Y-parameters from a set of filter response are also provided in the paper. The proposed diagnosis approach is applicable to a general coupled resonator filter with losses and therefore can be effectively used in computer aided tuning of high order filters with cross couplings.**

***Index Terms*—Chebyshev bandpass filter, computer-aided tuning, coupling matrix, coupling matrix extraction.**

I. Introduction

In order to accurately achieve a desired electric performance of a microwave coupled resonator filter, an effective tuning is a critical and compulsory step in filter production to compensate the manufacturing tolerances and the uncertainties of materials such as the silver plating thickness and the dielectric constant. This situation is particularly true for narrow band general Chebyshev filters in a multiplexer of microwave payload for a satellite.

Since the traditional tuning skill of a human operator is mainly built up by years of tuning experience, the tuning process becomes very labor intensive and expensive. This is true, particularly for high-order filters with multiple cross-couplings as complex fuzzy logic is required. Nevertheless, one of the difficulties associated with the traditional tuning is that it is not a deterministic process. In other words, there is no guarantee that each step of a tuning is always in the right direction. Such predicament is undesirable in tuning a channel filter for space use as a repeated tuning may wear out the plated silver of tuning screws and degrade the unloaded Q of the filter.

Computer-aided tuning (CAT) of a microwave-coupled resonator filter has drawn a great deal of attention in recent years. One major scheme of existing CAT methods is the sequential tuning scheme that is applicable to strongly detuned resonators. For example, the group delay response of each subcircuit is used for the tuning of each coupling element and resonator one by one [1], and a parameter extraction method is applied for tuning of each "sub-filter" using nonlinear optimization [2]. The sequential tuning in time domain [3] is also an effective approach to bring a severely detuned in-line coupled resonator filter to resonance. The major difficulties that come with these sequential tuning schemes are 1) it is difficult to deal with cross coupling in general; 2) it is not always convenient to segregate each resonator or coupling element in a filter structure such as dielectric resonator filters; and 3) there is an accumulated error. On the other hand, a popular CAT approach used in the industry is to extract the entire coupling matrix in one run from the measured filter responses using nonlinear optimization. The extracted coupling matrix is then used to identify and tune the coupling elements that have large discrepancies as compared to the desired coupling matrix [4].

The coupling matrix (CM) of a coupled resonator bandpass filter is a fundamental reference in describing the relationship between a physical realization and the required filter response as each of the coupling elements in a CM uniquely corresponds to a physical tuning element [5]. All the CAT techniques, including the above-mentioned works, require some sort of filter diagnosis or coupling matrix extraction from the measured filter responses. Except the approach used in [1], where the group delay information is utilized directly, most of existing parameter extraction techniques are based on nonlinear optimization [2], [4], [6] and [8]. The optimization procedures are either time consuming (for global optimization) or are very sensitive to the initial value and the number of variables (for gradient based local optimization), and are easily tripped into a local optimum.

In theory, a coupling matrix is determined by the poles and zeros (or the residues of the poles) of a given filter system. It is assumed that the multiple solution is not a concern in the discussion as a CAT process leads the extracted CM to the desired physical solution in practice. For some simple filter configurations, in which each coupling element and resonator can be easily isolated, some analytic formulas have been reported for determining the CM, For example, works in [9] and [10] provided a method that associates the poles and zeros of a filter

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system with the CM for cascaded and symmetrically coupled filters. Special attention must be paid here to the fact that the phase derivative with respect to frequency is used in [9] and [10] for accurately determining the poles and zeros and in [1] for correlating the coupling values in the circuit model with those in a physical model through group delay information.

In filter diagnosis (also called coupling matrix extraction), one must convert filter responses of a physical model (a measured or EM simulated response) into a filter circuit model. This important concept has been ignored in the existing CAT techniques. There are three nonideal effects that need to be removed from raw measured data of a physical filter model before diagnosis: 1) a constant phase loading that is caused by the higher order modes in the vicinity of input/output (I/O) coupling elements; 2) the loss effect associated to each resonator; and 3) a section of embedded transmission line at each port of a filter.

The original concept of a constant phase loading was mentioned in [11] without showing how to determine the constant. Furthermore, the analytical diagnosis approach presented in the work only deals with a lossless case. These two major shortcomings restrict the analytical CAT approach from practical uses. Most recently, the mentioned restrictions are addressed in [12] by the authors, in which the emphasis are only placed on introducing the phase loading concept and the basic procedure of how to extract the unloaded Q.

In this paper, the theory underneath the concept and implementation techniques of the analytical approach are presented for the first time. The theory is apt to lossy resonator filters. To help readers to understand the basic principles, not only a simple-to-operate scheme but also the mathematical model for removing the constant phase loading is presented. In addition, an analytic formula for deterministically de-embedding the unknown transmission lines from the measured data is also provided. To accurately remove the loss effect from measured data, which is crucial in extracting the CM of a lossy filter, four useful mathematic properties of the admittance parameters are revealed as the foundation of the extraction of the unloaded Q. In order to provide a robust diagnosis process, how to deal with the degenerate poles and the measurement noise are also discussed. Furthermore, three practical examples, including a CAT design of an 8–2 circular waveguide dual-mode filter, are given to illustrate the implementation details of the approach and to demonstrate the validation of the analytical approach.

Using the presented approach, there is no need to run nonlinear optimization and no requirement for initial values. The diagnosis and tuning of a practical Chebyshev coupled resonator filter become deterministic and easy to operate and can be running in real time. Because the diagnosis has nearly zero overhead, one can tune either single or multiple coupling elements at each tuning step. It is expected that the proposed analytical approach can significantly accelerate the tuning of high-order Chebyshev coupled resonator filters in a deterministic way and that the approach can become a useful tool in the filter and space microwave payload industries.

II. The Theory

It is well known that the characteristics of a general Chebyshev filter can be represented by a circuit model of intercoupled resonators whose transfer and reflection functions can be characterized by a set of rational polynomials. In the circuit model, the I/O couplings are represented by a simple inverter without any embedded transmission lines at I/O ports, which shift the reference planes. In a physical model, however, the scenario is no longer the same due to the existence of two nonideal effects: 1) a section of transmission line connecting to the I/O inverters, which contributes a frequency dependent phase shift; and 2) a constant phase loading, both of them are caused by the higher order modes at the vicinity of I/O coupling structures. The phase loading is a constant phase shift and can be determined by the phase difference between the phases of the reflection coefficients of the circuit model and those of the physical model. It can be shown that the phase loading is mainly determined by the I/O coupling structure.

A. Removal of the Phase Loading

The existence of a phase loading has been indirectly proved by some previously proposed CAT methods. In [1], [9], and [10], when determining the circuit parameters of a filter from measured data, the effect of a constant phase loading is unintentionally removed by using the derivative of phases with respect to frequency. Nevertheless, a systematic treatment of removing the phase loading from a measured response needs to be developed for a general narrow band coupled resonator filter.

According to the classical filter synthesis theory [5], the reflection coefficient S_{11} of an nth degree lowpass filter prototype can be expressed in terms of rational polynomials as

$$S_{11} = \frac{F(s)/\varepsilon_R}{E(s)} \tag{1}$$

where $E(s)$ is an nth degree polynomial with complex coefficients $e_0, e_1, e_2, \ldots, e_n$, $F(s)$ is an nth degree polynomial with coefficients $f_0, f_1, f_2, \ldots, f_n$, and the coefficient ε_R is determined such that the highest degree coefficients of $E(s)$ and $F(s)$ are normalized to unity. When $s = j\Omega$, where Ω is the normalized frequency in lowpass frequency domain, the phase of the reflection coefficient can be readily obtained as

$$\phi_{S_{11}}(\Omega) = \tan^{-1} \frac{e_{(n-1),r}\Omega^{n-1} + e_{(n-2),i}\Omega^{n-2} + \cdots + e_{0,i}}{\Omega^n + e_{(n-1),i}\Omega^{n-1} + \cdots + e_{0,r}}. \tag{2}$$

Since the coefficients $f_0, f_1, f_2, \ldots, f_n$ are pure real or pure imaginary alternatively, when $s = j\Omega$, $F(s)$ becomes a function that results a real number. Therefore, the phase of S_{11} is independent to the coefficients of polynomial $F(s)$. Therefore, as $\Omega \to \pm\infty$,

$$\phi_{S_{11}} \approx \frac{a_1}{\Omega}. \tag{3}$$

Consequently, for a narrow band filter, when real frequency ω is far away from the passband frequencies, the asymptotic response of the filter group delay can be expressed as

$$\tau_{S_{11}} = \frac{\partial \phi_{S_{11}}}{\partial \omega} \approx -\frac{a_2}{\Omega^2}. \tag{4}$$

In (3) and (4), constants a_1 and a_2 are proportionality coefficients whose values are not important for drawing conclu-

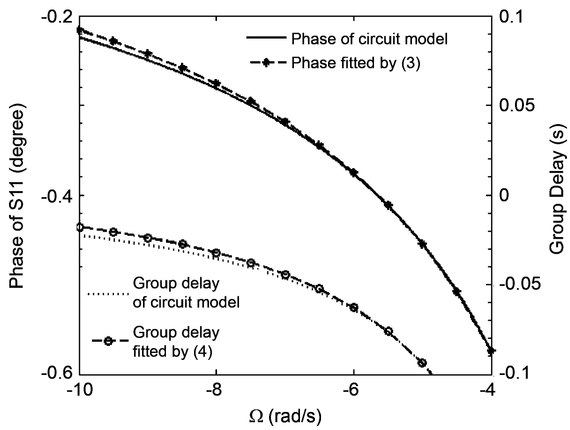

Fig. 1. Phase and group delay of lowpass prototype of a typical four-pole bandpass filter in far low end frequency range and their rational function fitting.

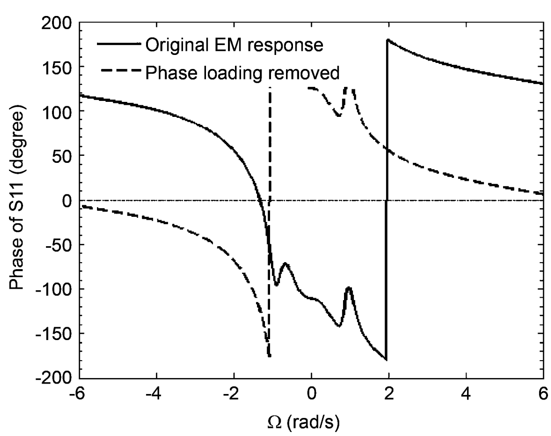

Fig. 2. Phases of a 4–2 filter with and without removing the phase loading.

sions. Fig. 1 shows the phase and the group delay responses of a four-pole bandpass filter using the circuit model at the far low end frequencies. It can be seen that the phase and the group delay can be fitted very well by (3) and (4), respectively.

Relation (3) reveals that the asymptote of the S_{11} phase outside of the pass band approaches to zero counter- symmetrically from negative and positive half planes. This fact suggests that the phase loading can be determined by finding the phase difference at the two symmetric frequency points that are a few fold of bandwidth away from the center frequency, say $\Omega = \pm 4\ (\mathrm{rad/s})$, in the lowpass frequency domain.

The concept of the phase loading can be illustrated by a simple filter example. Fig. 2 shows the phase responses of a typical 4–2 waveguide dual-mode filter obtained by an EM simulation. The original phase and the one with 127° phase loading removed are shown, illustrating the asymptotic behavior of the S_{11} phase of a physical model after removing the phase loading. The frequency variable has been transformed to its lowpass domain in order to view the respective limits.

A phase loading is originated from the higher order modes in the vicinity of the I/O coupling element, whose circuit model can be represented by a reactive T-network with a phase offset φ connected to each port [13]. The reactive T-network is equivalent to an admittance inverter and the phase offset can be very well approximated by the following function in a wide frequency range:

$$\varphi = \varphi_0 + \beta \Delta l \tag{5}$$

where φ_0 is constant term, β is the propagation constant of the interfacing transmission line and Δl is an equivalent length of transmission line to be de-embedded. The frequency invariant term φ_0 is called phase loading.

B. De-Embedding of Transmission Line

In a physical filter model, there is always an unwanted length of transmission line at a filter port between the physical reference plane and the port of the corresponding inverter in the circuit model. The length of this transmission line is difficult to measure because of the higher order mode effect at the port of the physical filter model.

For a typical transmission line, the wave number β can be approximated, when $k \gg k_c$, by

$$\beta \approx k - \frac{k_c^2}{2k}. \tag{6a}$$

As $k \gg k_c$, term $k_c^2/2k$ approaches to a very small constant, and β can be further approximated by

$$\beta \rightarrow \omega\sqrt{\mu\varepsilon} - b \tag{6b}$$

where the coefficient of the first term is a known constant associated to the dielectric constant of the transmission line and b is a very small constant and has no effect in this approach. Therefore, in conjunction with (4), when the working frequency is much higher than the cutoff frequency, the group delay of S_{11} of a physical filter model with an assumed length of transmission line can be expressed by

$$\tau_{S_{11}} \approx -\frac{a_2}{\Omega^2} - 2\sqrt{\mu\varepsilon}\Delta l \tag{7}$$

where Δl is the length of a transmission line to be de-embedded. The constant term in (7) can be found by least square fitting.

It can be seen from (7) that a length of transmission line contributes to a constant shift of the group delay. The constant shift can be determined by curve-fitting the group delay of a physical model at frequency points far below or above the center frequency by the function form of (7). Having found the parameter Δl the following phase change should be removed from the measured phase:

$$\Delta\phi_{S_{11}} = -2\beta\Delta l \approx -2\Delta l\sqrt{\mu\varepsilon}\omega. \tag{8}$$

Failing to de-embed the unwanted transmission line will also cause the displacement of the poles of Y-parameters, and thus leads to an incorrect solution.

C. Admittance Parameter Extraction

After removing the phase loading and the embedded transmission line, one can convert the modified S-parameters to the numerical admittance parameters. For an nth degree filter, the

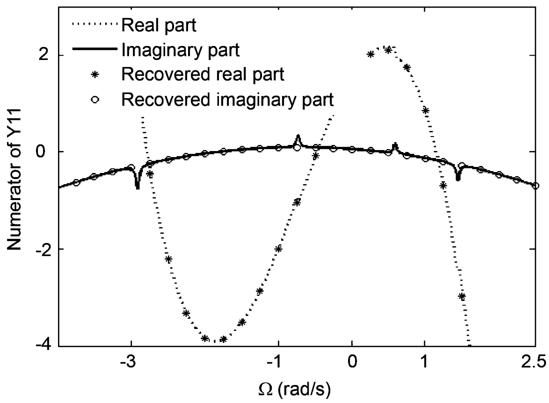

Fig. 3. Fitting of the numerator of Y_{11} of a fourth degree circuit filter model.

admittance parameters containing n single-pole terms can be written as [5]

$$[Y] = \begin{bmatrix} y_{11}(s) & y_{12}(s) \\ y_{21}(s) & y_{22}(s) \end{bmatrix} = \sum_{k=1}^{n} \frac{1}{s - j\lambda_k} \begin{bmatrix} r_{11k} & r_{12k} \\ r_{21k} & r_{22k} \end{bmatrix}. \quad (9)$$

The "peaks" of the admittance parameters indicate the locations of the poles. With the poles founded from the peaks, the explicit polynomial expression for the denominator can be found easily. The numerical values of the numerator of Y_{11} can be determined by multiplying the numerical values of Y_{11} by the product of the denominator of each partial fraction such that

$$Y_{n11} = Y_{11} \prod_{i=1}^{n} \big(\Omega - \lambda(i)\big) = \sum_{i=1}^{n} \left[r_{11}(i) \prod_{\substack{j=1 \\ j \neq i}}^{n} \big(\Omega - \lambda(j)\big) \right]. \quad (10)$$

Since the curve of Y_{n11} is a smooth function the polynomial expression of the numerator can be found accurately by a simple least square fitting. Fig. 3 shows typical curves of the numerator of a typical fourth degree circuit filter model.

Having determined the polynomial expressions of the numerator and the denominator of Y_{11}, when the frequency is equal to the value of a pole, the corresponding residue can be easily found by

$$r_{11}(i) = \frac{Y_{n11}\big(\lambda(i)\big)}{\prod\limits_{\substack{j=1 \\ j \neq i}}^{n} \big(\lambda(i) - \lambda(j)\big)}. \quad (11)$$

The residues of Y_{22} and Y_{21} can be found by the same token. This approach works accurately for lossless case. For lossy cases, the residues calculated by this approach will be complex numbers, which lead to an undesirable complex coupling matrix.

D. Removal of Loss Effect and Recovery of Coupling Matrix

When losses are presented, Y parameters will exhibit a real part. If poles are treated as purely imaginary numbers, the residues will be complex, which represent nonphysical coupling elements of ordinary filters. To describe a coupled resonator filter using a CM, it is desirable to have real and positive residues of Y_{11} and Y_{22}.

In the synthesis of a predistortion filter, by rightward shifting the poles on the left half of the complex plane, a real part of the complex frequency in denominator is introduced to compensate the effect of loss [5]. Being inspired by this treatment, the real part of the frequency can be included in the poles as described by (12) while the residues remaining to be positive real numbers.

$$\begin{aligned} Y_{11} &= \sum_{i=1}^{n} \frac{r_{11}(i)}{(j\Omega + \sigma(i)) - j\lambda(i)} \\ &= \sum_{i=1}^{n} \frac{r_{11}(i)}{j\Omega - (j\lambda(i) - \sigma(i))}. \end{aligned} \quad (12)$$

Therefore, the real and the imaginary parts of the admittance parameters of a lossy filter can be expressed as

$$\mathrm{Re}(Y_{11}) = \sum_{i=1}^{n} \frac{r_{11}(i)\sigma(i)}{(\Omega - \lambda(i))^2 + \sigma(i)^2} \quad (13)$$

$$\mathrm{Im}(Y_{11}) = j \sum_{i=1}^{n} \frac{r_{11}(i)(\omega - \lambda(i))}{(\Omega - \lambda(i))^2 + \sigma(i)^2}. \quad (14)$$

Four important conclusions can be drawn from above two equations: 1) a local maximum of the real part of Y_{11} is achieved when the frequency equals to the imaginary part of a pole; 2) when the frequency equals to the imaginary part of a pole the corresponding term in the imaginary part of Y_{11} goes to zero; 3) the location of the local maximum in the imaginary part at the ith pole is attained when the frequency is shifted away from the imaginary axis by amount of $\sigma(i)$; and consequently, 4) the maximum value of the imaginary part at the ith pole is half of that of the corresponding real part.

Conclusion 4) can be easily derived from the following inequality:

$$\frac{r_{11}(i)}{(\Omega - \lambda(i)) + \frac{\sigma(i)^2}{(\Omega - \lambda(i))}} \leq \left(\frac{r_{11}(i)}{2\sigma(i)} \right). \quad (15)$$

Making use of the above mentioned properties, a two-step approach for accurately determining the complex poles is proposed. For a lossy bandpass filter, the scheme for determining the real positive residues discussed earlier is still applicable. The first step of the approach provides an approximated values of the complex poles, and the second step is to fine tune the complex poles to satisfy the four properties.

Since the loss factor $\sigma(i)$ are usually small as compared to $\lambda(i)$, when Ω is near a pole, the ith term in the partial fraction expression will dominate. Using (13) and (14), an approximated loss factor $\sigma(i)$ can be found by the derivative of the ratio of the real and the imaginary parts of Y_{11} near the ith pole:

$$\sigma(i) = \left(\frac{-d\big(\mathrm{Re}(Y_{11})/\mathrm{Im}(Y_{11})\big)}{d\Omega} \right)^{-1}. \quad (16)$$

As the magnitude of the Y parameters at a pole is mainly determined by the value of corresponding loss factor σ according to (11), if the imaginary part of the pole is accurately determined, the loss factor $\sigma(i)$ calculated by (16) can be fine tuned by matching the finite "peak" values of the Y parameters from the measured data and the recovered ones.

In numerical process, the residues found using the complex poles may be complex numbers. However, their imaginary parts are a few orders of magnitude smaller than that of the real parts. It can be observed in practice that when an appropriate loss factor is found, the absolute value of the imaginary parts of the residue reaches to its minimum.

An $(n+2)$ coupling matrix of a given filter responses can be constructed by the imaginary parts of poles and corresponding real residues [5]. With complex poles, the CM can also be presented as a complex matrix. The imaginary part of the CM represents the coupling between coupled resonators for lossless case. The real part of the CM represents the loss of the filter and can be represented by a diagonal matrix as

$$\mathrm{Re}(CM) = \begin{bmatrix} 0 & 0 & 0 & \dots & 0 & 0 \\ 0 & \sigma(1) & & & & 0 \\ 0 & & \sigma(2) & & & 0 \\ \dots & & & \dots & & \dots \\ 0 & & & & \sigma(n) & 0 \\ 0 & 0 & 0 & \dots & 0 & 0 \end{bmatrix}. \quad (17)$$

For a given filter coupling topology, a sequence of similarity transformations need to be applied to the real and imaginary CM separately. As the result, the imaginary part of the coupling matrix reflects the actual couplings of a given filter response. When the loss factors are the same, the real part of CM in (17) will not be changed by the transformations. However, if the loss factors are not the same, the transformations will result in a full matrix. The more uniform the loss factors are, the more dominant the diagonal elements in the transformed loss matrix will be. In practice, because the loss among the resonators in a filter does not vary too much, the resultant off-diagonal elements in the transformed loss matrix can be ignored, and the effective unloaded Q for the ith resonator can be found by

$$Q_u(i) = \frac{f_0}{BW} \cdot \frac{1}{\sigma(i)} \quad (18)$$

where $\sigma(i)$ is the ith diagonal element of the transformed loss matrix. It is worth mentioning that if the loss factors are not uniform, the solution of the effective unloaded Q's is not unique. As a side product, the effective Q values calculated by (18) gives a good estimation of the unloaded Q of the filter.

At the end of this section, two practical issues need to be addressed. The first issue is the removal of the measurement noise in the original data acquired from a vector network analyzer. A straightforward method to remove the noise is to represent the raw measured data by a rational function using an adaptive frequency sampling algorithm [14]. The second issue is the degenerate pole problem. For a filter of a high-order degree, it is frequently seen that two or more poles of the Y parameters are located so close to each other such that only one "peak" is seen. This problem is more severe in diagnosing a high order lossy filter. To deal with this problem, the Thieles' continued fraction method [15] can be utilized to find the rational expression of the Y-parameters. More accurate pole locations can be found by solving for the roots of the fitted denominator polynomial rather than finding the poles from the magnitude.

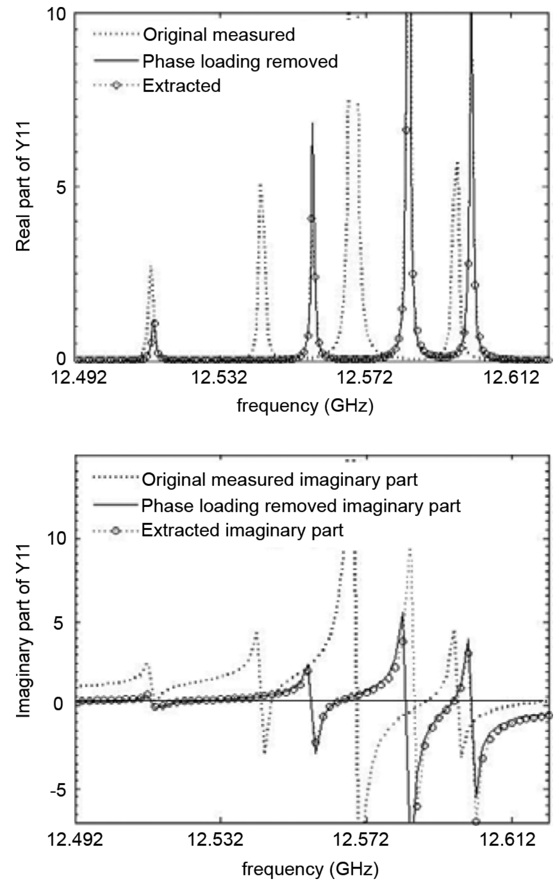

Fig. 4. (a) Extraction of real part of Y_{11} of the 4–2 filter. (b) Extraction of imaginary part of Y_{11} of the 4–2 filter.

III. Examples

The theory discussed above can be easily implemented for a generic coupled resonator bandpass filter. In this section, the validation of the approach will be demonstrated by three examples, including the measured data of practical waveguide filters and the data from a full wave electromagnetic simulation.

1) Example 1: Diagnosis of a 4–2 Waveguide Filter: The diagnosis of a 4–2 Ku band dual-mode circular waveguide filter with a center frequency of 12.572 GHz and a bandwidth of 0.04 GHz is chosen as the first example. To demonstrate the applicability of the proposed approach to a severely detuned filter, one of the frequency tuning screws was pushed in. Fig. 4 shows the recovery of the admittance parameters after removing the phase loading (79.5° and 86.4° separately for input and output ports) and the embedded transmission lines (both are zero in this case).

It can be seen that the locations of the poles are greatly affected by the existence of the phase loadings. The pole that is far away from the pass band corresponds to the severely detuned resonator. After removing the phase loading, the measured Y-parameters should satisfy the four conclusions drawn from (13) and (14).

In Fig. 5, the original measured S-parameters are compared with those calculated by the extracted coupling matrix. Very good agreement can be observed. The extracted coupling matrix

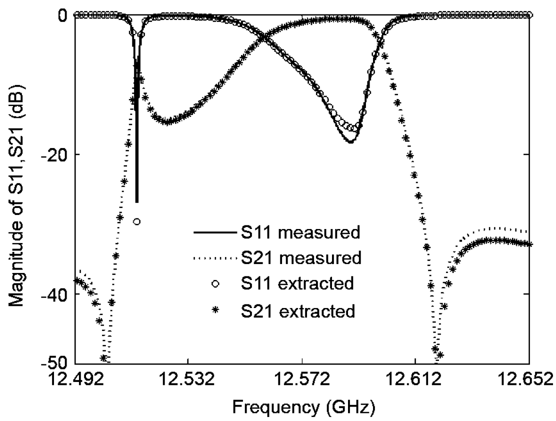

Fig. 5. S-parameters of the 4–2 filter: the measured and the recovered.

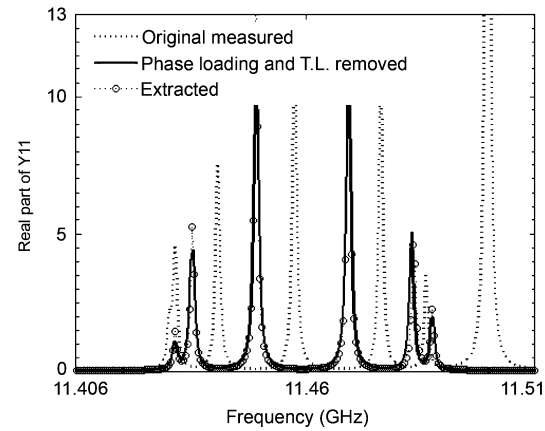

Fig. 6. Real part of Y_{11} of the 8–4 dual-mode waveguide filter.

TABLE I
EXTRACTED $N+2$ COUPLING MATRIX OF THE 4–2 FILTER

0.0000	0.9992	0.0000	0.0000	0.0000	0.0000
0.9992	-0.5090	-0.9536	0.0000	-0.2140	-0.0145
0.0000	-0.9536	1.0609	1.7595	0.2807	0.0000
0.0000	0.0000	1.7595	0.9368	-0.6976	0.0000
0.0000	-0.2140	0.2807	-0.6976	0.1119	1.0106
0.0000	-0.0145	0.0000	0.0000	1.0106	0.0000

TABLE II
EXTRACTED LOSS MATRIX OF THE 4–2 FILTER

0.00	0.00	0.00	0.00	0.00	0.00
0.00	**-2.30E-02**	8.14E-04	-1.97E-03	-1.69E-04	0.00
0.00	8.14E-04	**-2.83E-02**	4.40E-03	-9.23E-04	0.00
0.00	-1.97E-03	4.40E-03	**-2.72E-02**	2.56E-04	0.00
0.00	-1.69E-04	-9.23E-04	2.56E-04	**-2.22E-02**	0.00
0.00	0.00	0.00	0.00	0.00	0.00

is listed in Table I, in which a large M_{22} represents the severe detuning of the second resonator.

The real part of the extracted coupling matrix in its canonical form is founded by applying the same similarity transformations as applied to the imaginary part and is listed in Table II. It can be seen that the diagonal elements are at least one order of magnitude larger than the off-diagonal elements indicating that the losses contributed by each individual resonator is quite uniform. Applying (18) to the diagonal elements leads to the effective unloaded Q of 13 900, 11 200, 11 700 and 14 000, for each resonator, respectively.

2) Example 2: Diagnosis of an 8–4 Waveguide Dual Mode Filter: A set of measured responses of an eight-pole Ku band dual-mode waveguide filter with four transmission zeros are tested to validate the proposed approach. In this case, the center frequency is 11.46 GHz, and the bandwidth is 0.054 GHz.

The phase loadings of the input and output ports are extracted as -72.1° and -76.5° , respectively. The length of the embedded transmission lines are extracted to be 99.1 and 94.0 mm for ports 1 and 2, respectively. Fig. 6 shows the real part of Y_{11} from the raw measured data, the measured data after removing the phase loading and transmission lines (T.L.) and the filter responses from the extracted CM.

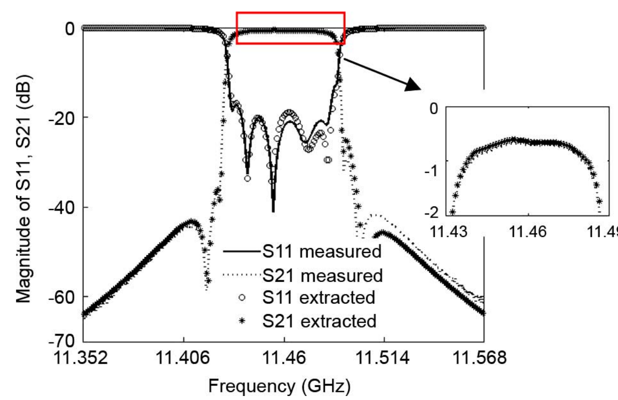

Fig. 7. S-parameters of the 8–4 dual-mode waveguide filter.

Obviously, only the six poles seen in the plot can not fully describe this eight-pole filter. By interpolating the denominator of the Y-parameters using an eighth-order polynomial all the eight poles can be revealed accurately. In this case, each of the two outer most "peaks" consists of two poles. The values of poles in the lowpass frequency domain are -1.2874, -1.2822, -0.9555, -0.3878, 0.3983, 0.9636, 1.2126, and 1.2807.

The measured S-parameters are superposed with those from the extracted CM (real and imaginary parts) in Fig. 7. The filter responses from the extracted circuit model fit the original measured filter responses very well. The detailed comparison of the insertion losses in the pass band is zoomed in, demonstrating the accuracy of the extracted effective unloaded Q values.

3) Example 3: CAT for EM Design of an Eight-Pole Filter: The last example concerns with the design of an equal-ripple eight-pole dual-mode circular waveguide filter. The filter structure is shown in Fig. 8. The filter response by an in-house full-wave mode-matching-based electromagnetic simulation software is used as the physical model. The center frequency of the filter is 12.0 GHz, and the bandwidth is 0.05 GHz. In order to study the proposed approach for a lossy filter, the conductivity of the metal is set to 4.0×10^7 Siemens in the simulation.

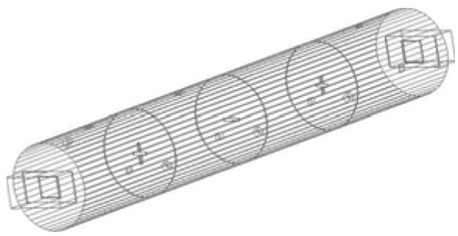

Fig. 8. Eight-pole circular waveguide dual-mode filter for Example 3.

TABLE III
COUPLING ELEMENTS IN SELECTED TUNING STEPS

	Initial	Step 3	Step7	Step 20	Desired
M11	-0.0389	0.0281	0.0076	0.0104	0.0000
M22	-0.1235	0.0006	-0.0011	0.0018	0.0000
M33	-0.1984	-0.0227	-0.0081	-0.0009	0.0000
M44	-0.0343	-0.0305	0.0012	-0.0067	0.0000
M55	-0.1929	-0.0234	-0.0074	0.0076	0.0000
M66	0.3568	-0.0306	0.0005	-0.0039	0.0000
M77	0.1441	-0.0431	0.0097	0.0101	0.0000
M88	0.2756	0.0298	-0.0100	0.0050	0.0000
Ms1	1.0400	1.0347	1.0363	1.0405	1.0452
M12	-0.8287	0.8325	-0.8317	-0.8349	0.8390
M23	-0.3131	0.4763	-0.4712	-0.4630	0.4602
M34	-0.5734	0.5638	-0.5615	-0.5616	0.5640
M45	-0.5474	0.5470	-0.5467	-0.5500	0.5543
M56	-0.4493	0.4608	-0.4582	-0.4595	0.4587
M67	-1.0048	0.9061	-0.8352	-0.8293	0.8235
M78	-0.8696	-0.7846	-0.7780	-0.7885	0.7870
M8L	1.0960	1.0425	1.0358	1.0460	1.0452
M14	-0.1780	0.2256	-0.2239	-0.2193	0.2164
M58	0.4437	0.4041	0.3714	0.3669	-0.3623

In each tuning step, the characteristics of the filter are interpreted by a coupling matrix. Then, the difference between the extracted CM and the desired golden template can guide the tuning process. For most of the cases, adjustments should be made to the tuning of the elements with large differences. When the return loss level is lower than -17 dB and the maximum difference between the extracted CM and the golden template is around 0.01, a sensitivity analysis of all the coupling elements in the extracted CM should be conducted first in order to find the proper elements to be tuned in the next step. In this way, the tuning process utilizing the CM extraction can be iterated until the desired response is achieved.

After the direct tunings of the first seven steps, a sensitivity analysis of the extracted CM is needed for the rest of fine-tunings. The number of the rest of tunings strongly depends on the knowledge of the mapping relation between the couplings in the circuit model and those in the physical model. Fig. 9(a)–(d) shows the S-parameters of the EM simulated and the circuit model of the extracted CM at the initial step and steps 3, 7, and 20, respectively. The process eventually ends at an equal-ripple response. The coupling elements in the CM at some selected tuning steps are list in Table III, illustrating how the analytical diagnosis guides the filter tuning.

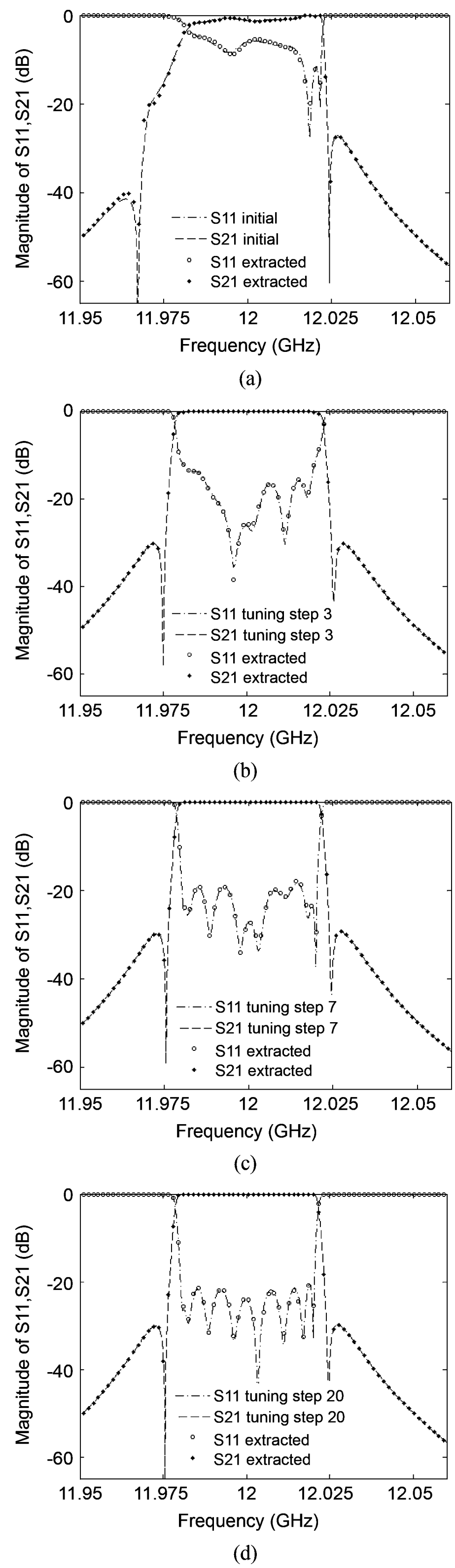

Fig. 9. S-parameters of the EM model and the circuit model of the extracted coupling matrix of the eight-pole circular waveguide dual-mode filter at (a) initial step; (b) step 3; (c) step 7; and (d) step 20.

IV. Conclusion

A novel systematic approach to analytically extracting of the coupling matrix and the equivalent unloaded Q of a practical narrow band coupled resonator filter is presented. This method is applicable to any measured filter response as long as the poles of the filter fall within the range of measurement. The filter diagnosis is a very important step in computer aided tuning of a microwave bandpass filter. In this paper, the concept of phase loading is proposed and stressed for the first time in the field. To convert a physical model to a circuit model, the theoretic formulas for conveniently removing the phase loading effect and the embedded transmission line from a given filter response are developed. An analytical formula for calculating the loss factor in the system poles is also developed in the proposed approach. Being able to accurately determine the complex poles of a filter within the frequency range of measurement and consequently the effective unloaded Q values of resonators, the analytical approach can be effectively used for diagnosing a lossy filter.

Three practical examples, including two measured filter responses and one electromagnetic design case study, are provided to show the details in the implementation and the effectiveness of the approach for practical filter tuning.

In addition to its simplicity, the proposed approach is robust and efficient warranted by its deterministic attribute without a need for an initial value. This analytical tool will find many practical applications for the computer aided tuning of microwave filters in the industry.

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