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Modeling based real frequency technique

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Abstract

Usually commercially available software tools are used, to design matching networks for wireless communication systems. But a properly selected matching network topology with good initial element values must be supplied to these tools. Therefore, in this paper a modeling-based real frequency technique (M-RFT) is presented, to generate matching networks with initial element values. In the proposed method, output impedance data of the matching network are obtained in terms of *ABCD*-parameters of the load model. Then, they are modeled which in turn yields the desired matching network with initial element values. It is not needed to select a circuit topology for the matching network, which is the natural consequence of the matching processes. Also, there is no need to select the desired transducer power gain level; the proposed technique naturally provides a gain curve fluctuating around a flat level. Eventually, the initial design is improved by optimizing the performance of the matched system employing the commercially available computer aided design (CAD) packages. An algorithm and example are given, to illustrate the utilization of the proposed technique.

Keywords: Broadband matching; Modeling; Real frequency technique

1. Introduction

For all microwave communication systems, design of broadband matching networks or so-called equalizers have been considered as an essential problem for engineers [1]. In this regard, analytic theory of broadband matching [2,3], and computer aided design (CAD) methods are available for the designers [4–6]. It is well known that analytic theory is difficult to utilize. Therefore, it is always preferable to employ CAD techniques, to design matching networks. All the CAD techniques optimize the matched system performance. As the result of this process, element values of the matching network are obtained. It should be mentioned that performance optimization is highly nonlinear with respect to element values and requires very good initials. In this respect, selection of initial element values is crucial for successful optimization. Therefore, in this paper, a well-established initialization technique is introduced for broadband matching problems. The new technique is based on the generation of the output impedance data of the matching network, and modeling them. In the proposed method, there is no need to select matching network topology which is the natural consequence of the modeling processes, and no need to guess the available transducer power gain level, the algorithm realizes the optimization to obtain nearly flat transducer power gain in the passband as opposed to the existing methods [7–9].

In the following section, a lumped-element two-port is described in terms of its scattering parameters, and the rationale of the new matching technique is explained. Subsequently, the algorithm of the proposed technique is presented. Finally, utilization of the algorithm is exhibited by designing a matching network for a selected passive load.

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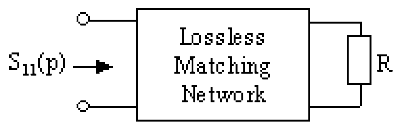

Fig. 1. A lossless two-port with reflection coefficient $S_{11}(p)$.

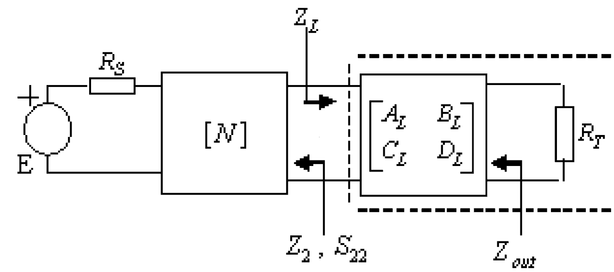

Fig. 2. Single matching arrangement.

2. Mathematical framework

2.1. Scattering parameters

For a lossless two-port like the one depicted in Fig. 1, the canonic form of the scattering matrix is given by [10]

$$S(p) = \begin{bmatrix} S_{11}(p) & S_{12}(p) \\ S_{21}(p) & S_{22}(p) \end{bmatrix} = \frac{1}{g(p)} \begin{bmatrix} h(p) & \mu f(-p) \\ f(p) & -\mu h(-p) \end{bmatrix}, \tag{1}$$

where $p = \sigma + j\omega$ is the complex frequency, and $\mu = \pm 1$ is a unimodular constant. If the two-port is reciprocal as well, then the polynomial $f(p)$ is either even or odd. In this case, $\mu = +1$ if $f(p)$ is even, and $\mu = -1$ if $f(p)$ is odd. Thus, for a lossless, reciprocal two-port

$$\mu = \frac{f(-p)}{f(p)} = \pm 1. \tag{2}$$

For a lossless two-port with resistive termination, energy conversation requires that

$$S(p)S^{\mathrm{T}}(-p) = I, \tag{3a}$$

where I is the identity matrix, and T represents the transpose operation. The explicit form of Eq. (3a) is known as the Feldtkeller equation and given as

$$g(p)g(-p) = h(p)h(-p) + f(p)f(-p). \tag{3b}$$

In Eqs. (1) and (3b), $g(p)$ is the strictly Hurwitz polynomial of nth degree with real coefficients, and $h(p)$ is a polynomial of nth degree with real coefficients. The polynomial function $f(p)$ includes all transmission zeros of the two-port.

2.2. Rationale of the matching technique

Let us consider the single matching arrangement as shown in Fig. 2, where $[N]$ represents the matching network, and Z_L is the load. It is well known that the impedance (Z_{out}) seen in Fig. 2 can be calculated as [11]

$$Z_{\text{out}} = \frac{D_L Z_2 + B_L}{C_L Z_2 + A_L}, \tag{4}$$

where $\{A_L, B_L, C_L, D_L\}$ are the $ABCD$-parameters of the load model, and Z_2 is the output impedance of the matching network. For maximum power transfer from the output of the load model (Z_{out}) to the load termination resistance (R_{T}), the output impedance must be equal to the termination impedance, i.e. $Z_{\text{out}} = R_{\mathrm{T}}$. So from Eq. (4), the output impedance of the matching network can be written as

$$Z_2 = \frac{B_L - Z_{\text{out}} A_L}{C_L Z_{\text{out}} - D_L} = \frac{B_L - R_{\mathrm{T}} A_L}{C_L R_{\mathrm{T}} - D_L}. \tag{5}$$

As a result, once the data of the output impedance of the matching network (Z_2) are generated, then they can be modeled [12]. Finally, the model function is synthesized yielding the desired equalizer topology with initial element values. Eventually, performance of the matched system is optimized utilizing the commercially available CAD packages.

By using the proposed technique, matching networks with lumped, distributed or mixed (lumped and distributed) elements can be designed. The element type depends on the selected modeling tool.

3. Algorithm

In this section, the algorithm is presented, to design broadband matching networks via obtaining and modeling the output impedance data of the matching network.

3.1. Inputs

- ω_i; $i = 1, 2, \ldots, N_\omega$: Sample frequencies.
- N_ω: Total number of sample frequencies.
- $Z_L(j\omega_i) = R_L(\omega_i) + jX_L(\omega_i)$; $i = 1, 2, \ldots, N_\omega$: Given load impedance data.
- n: Desired number of elements in the matching network.
- $f(p)$: A monic polynomial constructed on the transmission zeros of the matching network. A practical form is $f(p) = p^k$, where k is the total number of transmission zeros at DC ($k \leqslant n$).
- $h_0, h_1, h_2, \ldots, h_n$: Initialized coefficients of the polynomial $h(p)$.
- δ: The stopping criteria for the sum of the squared errors. For many practical problems, it is sufficient to choose $\delta = 10^{-3}$.

3.2. Computational steps

Step 1: Construct the load model from the given load impedance data, which can be obtained by using the methods explained in [12]. In the reflectance-based modeling methods, three canonic polynomials $h_L(p)$, $g_L(p)$, $f_L(p)$ of the load are formed. Then, obtain the values of the scattering parameters $S_{11,L}(\mathrm{j}\omega_i)$, $S_{12,L}(\mathrm{j}\omega_i)$, $S_{21,L}(\mathrm{j}\omega_i)$, $S_{22,L}(\mathrm{j}\omega_i)$ by substituting $p=\mathrm{j}\omega_i$ into Eq. (1). By using S- to $ABCD$-parameters conversion formulae [11], calculate the values of A_L, B_L, C_L, D_L at the sample frequencies. Termination resistor (R_T) is obtained by synthesizing the reflection coefficient of the load $S_{11,L}(p)=h_L(p)/g_L(p)$.

Step 2: Calculate the values of the impedance Z_2 at sample frequencies via Eq. (5).

Step 3: By using initialized polynomial $h(p)$ and selected form of $f(p)$, obtain polynomial $g(p)$ via the Feldtkeller equation, Eq. (3b). Briefly, $G(\omega^2) = g(\mathrm{j}\omega)g(-\mathrm{j}\omega) = h(\mathrm{j}\omega)h(-\mathrm{j}\omega) + f(\mathrm{j}\omega)f(-\mathrm{j}\omega)$ is an even polynomial in ω. Therefore, the strictly Hurwitz polynomial $g(p)$ can be constructed by means of well-established numerical methods [12]. The data points describe a polynomial such that $G(\omega^2)=G_0+G_1\omega^2+\ldots+G_n\omega^{2n}>0;\ \forall\omega$. The coefficients $G_0, G_1, G_2, \ldots, G_n$ can be determined easily by linear or nonlinear interpolation or curve fitting methods. Then, replacing ω^2 by $-p^2$, the roots of $G(-p^2)=g(p)g(-p)$ can be extracted using explicit factorization techniques, and $g(p)$ constructed from the left half-plane (LHP) roots of $G(-p^2)$ as a strictly Hurwitz polynomial. Then, form S-parameters of the matching network, and calculate the output impedance values via $Z_2(\mathrm{j}\omega) = (1 + S_{22}(\mathrm{j}\omega))/(1 - S_{22}(\mathrm{j}\omega))$.

Step 4: Optimize initialized coefficients of the polynomial $h(p)$ until the sum of the squared errors between Z_2 values calculated at Step 2 and 3 are smaller than the stopping criteria.

Step 5: Synthesize the reflection coefficient $S_{22}(p)$, and obtain the matching network.

4. Example

In the following, an example is given, to illustrate the implementation of the algorithm defined above.

Load was selected as an inductor $L=1$ in series with the parallel combination of a capacitor $C=2$ and a resistor $R=1$ (i.e. $L+C//R$). Calculated load impedance data are given in Table 1.

Three canonic polynomials and termination resistor of the load were obtained by modeling the load impedance data given in Table 1 as

$$h_L(p)=p^2-0.5p,\quad g_L(p)=p^2+1.5p+1, \tag{6}$$
$$f_L(p)=1,\quad R_\mathrm{T}=1. \tag{7}$$

Then, these polynomials were used in the algorithm to calculate the values of the $ABCD$-parameters of the load.

Desired number of elements in the matching network and polynomial $f(p)$ were selected as $n=4$ and $f(p)=1$, respectively. Then, by ad hoc choice of the initial coefficients of the polynomial $h(p)$, the proposed matching algorithm was run.

Table 1. Given normalized load impedance data

ω_i	$R_L(\mathrm{j}\omega_i)$	$X_L(\mathrm{j}\omega_i)$
0.1	0.962	−0.092
0.2	0.862	−0.145
0.3	0.735	−0.141
0.4	0.610	−0.088
0.5	0.500	0.000
0.6	0.410	0.108
0.7	0.338	0.227
0.8	0.281	0.351
0.9	0.236	0.476
1.0	0.200	0.600

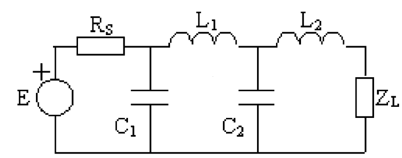

Fig. 3. Designed lumped-element matching network, M-RTF:$L_1=1.0437$, $L_2=0.23483$, $C_1=1.1442$, $C_2=2.6497$, $R_S=0.66091$, SRFT:$L_1=1.304$, $L_2=0.1621$, $C_1=1.124$, $C_2=2.322$, $R_S=1$, Final:$L_1=0.8861$, $L_2=0.03621$, $C_1=1.106$, $C_2=2.678$, $R_S=0.6321$.

Finally, $S_{22}(p)$ was generated as $S_{22}(p)=-\mu h(-p)/g(p)$, where

$$h(p)= -0.3020p^4-0.8867p^3+0.8534p^2 -0.7560p+0.2085$$

and

$$g(p)=0.3020p^4+1.6854p^3+2.5483p^2 +2.3285p+1.0215.$$

Then, the reflection coefficient $S_{22}(p)$ was synthesized, and the equalizer topology with element values seen in Fig. 3 was obtained. The same matching problem was solved by the simplified real frequency technique (SRFT, [7]). Finally, the obtained element values via M-RFT were optimized by using a CAD tool [4].

Transducer power gain (TPG) curves of the matching network designed by the proposed method (M-RFT), SRFT and obtained by a CAD tool are given in Fig. 4. Close examination of Fig. 4 reveals that in the passband, a nearly flat gain curve is obtained without selecting any TPG level as input. The performance of the matched network has been improved by using a CAD tool.

In broadband matching designs, it is desired to get maximum possible, flat transducer power gain in the passband. In the existing broadband matching methods [7–9], the designer must guess and supply this gain level to the algorithm, and free parameters are optimized until reaching to this level. But by using analytic theory of broadband matching, this level can be calculated only for simple loads.

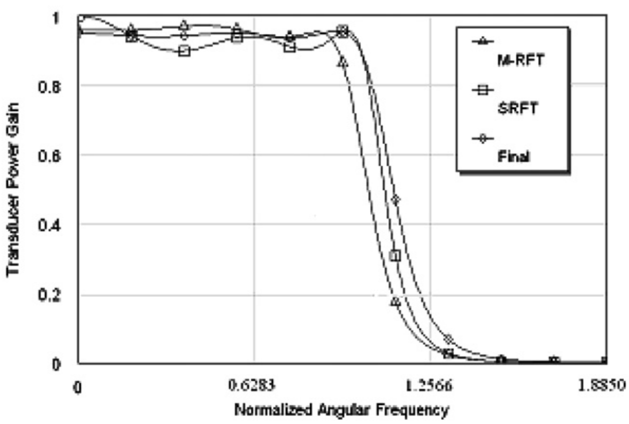

Fig. 4. Performance of the matched system.

So designer has to try lots of levels, to find the best one. But in the proposed technique, there is no need to choose any gain level. Algorithm optimizes the free parameters, to fluctuate the gain around a nearly flat level in the passband.

5. Conclusion

Design of practical matching networks is one of the essential problems of the microwave engineers. In this regard, commercially available computer aided design tools are utilized. Once the matching network topology is provided, these packages are excellent tools, to optimize system performance. At this point, initialization process becomes very crucial, since the system performance is highly nonlinear in terms of the element values of the chosen circuit topology. Therefore, in this paper, a modeling-based real frequency technique (M-RFT) is proposed, to construct lossless equalizers for broadband matching problems. In the method, output impedance data of the matching network are calculated without optimization in terms of $ABCD$-parameters and the termination resistor of the load model. Then, the calculated impedance data are modeled, and after synthesizing the obtained network function, desired matching network is formed. In the method, there is no need to select matching network topology which is the natural consequence of the matching processes, and no need to guess available transducer power gain level, the algorithm realizes the optimization to obtain nearly flat transducer power gain in the passband. Eventually, the actual performance of the matched system may be improved utilizing a commercially available CAD tool. An example is presented, to construct a matching network with lumped elements.

It is exhibited that the proposed method provides very good initials, to further improve the matched system performance by working on the element values. Therefore, it is expected that the proposed design technique is used as a front-end for the commercially available CAD packages, to design practical matching networks for wireless or in general microwave communication systems.

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