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Design of broadband single matching networks

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Abstract

In general commercially available software tools are preferred, to design broadband matching networks for wireless communication systems. But they need a properly selected matching network topology with good initial element values. Therefore, in this paper a new real frequency technique is presented, to generate broadband single matching networks with suitable initial element values. In the proposed method, load impedance is written in terms of *ABCD*-parameters of the desired matching network and the source resistor. Then, free parameters are optimized 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 the final available 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 two examples are given, to illustrate the utilization of the proposed technique.
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1. Introduction

For microwave communication systems, design of broadband matching networks is the essential problem for engineers [1]. Analytic broadband matching theory [2,3], and computer-aided design (CAD) methods are available for the designers [4–6]. But it is well known that analytic theory is difficult to implement. Therefore, CAD techniques have been preferred, to design matching networks. All the CAD techniques optimize the matched system performance. As a result of this process, element values of the matching network are obtained. It is important to notice that performance optimization is highly nonlinear with respect to element values and requires suitable initials. Namely, selection of initial element values is very important for successful

optimization. Therefore, in this paper, a well-established initialization technique is introduced for broadband single matching problems.

In [7], a similar technique (modeling-based real frequency technique (M-RFT)) has been proposed. But in M-RFT, circuit model of the load must be formed from the given numerical load data. But now, it is not necessary to obtain the model of the load. So the proposed method is simpler than M-RFT, but gives the same performance.

In the proposed technique, load impedance is written in terms of the parameters of the matching network, then an optimization algorithm has been used to get the best parameter values until the error between the given and calculated load data is smaller than a user-defined acceptable level. Like M-RFT, there is no need to define a desired gain level, the algorithm realizes the optimization to obtain transducer power gain fluctuating around the final available level in the passband as opposed to the existing other methods [8–11].

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In the following section, the scattering parameters of a lumped-element two-port is described, and the rationale of the new matching technique is explained. Then, the algorithm of the proposed technique is presented. Finally, utilization of the algorithm is exhibited by designing matching networks for a selected passive load and for a monopole antenna.

2. Mathematical framework

2.1. Scattering parameters

The canonic form of the scattering matrix of a lossless two-port like the one depicted in Fig. 1 is given by [12],

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

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

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

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

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

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

$$g(p)g(-p) = h(p)h(-p) + f(p)f(-p). \quad (3b)$$

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, which is usually given in numerical data measured at several frequencies.
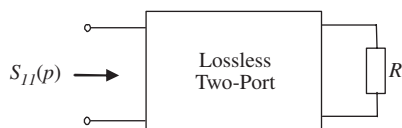

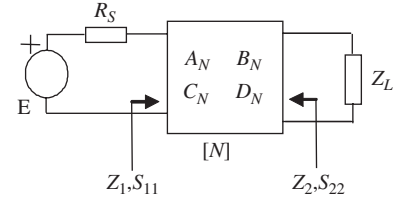**Fig. 1.** A lossless two-port with reflection coefficient $S_{11}(p)$.

Fig. 2. Single matching arrangement.

It is well known that the impedance (Z_1) seen in Fig. 2 can be calculated as [13]

$$Z_1 = \frac{A_N Z_L + B_N}{C_N Z_L + D_N}, \quad (4)$$

where $\{A_N, B_N, C_N, D_N\}$ are the *ABCD*-parameters and Z_1 is the input impedance of the matching network, respectively. For maximum power transfer from the source resistor (R_S) to the input of the matching network (Z_1), the input impedance must be equal to the source resistor, i.e. $Z_1 = R_S$. So from Eq. (4), the load impedance can be written as

$$Z_L = \frac{B_N - Z_1 D_N}{C_N Z_1 - A_N} = \frac{B_N - R_S D_N}{C_N R_S - A_N}. \quad (5)$$

As a result, after initializing the optimization parameters, *ABCD*-parameters of the matching network and the resistor (R_S) are calculated. Then, optimization is run until the error between the given load data and the values calculated via Eq. (5) is smaller than a user-defined acceptable level. Finally, obtained 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.

If an equation is written for the load impedance in terms of the network parameters (impedance or admittance) and the input impedance of the network (like Eq. (5)), then this equation can be used in the design process.

It can be seen that the proposed method is very simple, and there is no need to obtain a proper model of the given load data, as opposed to the method (M-RFT) described in [7].

By using the proposed technique, matching networks with lumped, distributed or mixed (lumped and distributed) elements can be designed.

3. Algorithm

In this section, the proposed algorithm is presented, to design broadband single matching networks.

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 $h(p)$. Since there is no restriction of these coefficients, the topology of the matching network is completely free.
- δ: The stopping criteria for the sum of the square errors. For many practical problems, it is sufficient to choose $\delta = 10^{-3}$.

Computational steps

Step 1: Obtain the polynomial $g(p)$ by using the initial and constructed polynomials $h(p)$ and $f(p)$ via Eq. (3b), respectively. Briefly, $G(\omega^2) = g(j\omega)g(-j\omega) = h(j\omega)h(-j\omega) + f(j\omega)f(-j\omega)$ is an even polynomial in ω. Therefore, the strictly Hurwitz polynomial $g(p)$ can be constructed by means of well-established numerical methods [14]. The data points describe a polynomial such that $G(\omega^2) = G_0 + G_1\omega^2 + \cdots + G_n\omega^{2n} > 0; \forall\omega$. 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. Alternatively, the proposed method in [15] can be utilized, to form the polynomial $g(p)$. Then, obtain the values of the scattering parameters $S_{11,L}(j\omega_i)$, $S_{12,L}(j\omega_i)$, $S_{21,L}(j\omega_i)$, $S_{22,L}(j\omega_i)$ by substituting $p = j\omega_i$ into Eq. (1). By using S- to *ABCD*-parameters conversion formulae [13], calculate the values of A_N, B_N, C_N, D_N at the sample frequencies. Termination resistor (R_S) is obtained by synthesizing the reflection coefficient of $S_{22}(p) = -\mu h(-p)/g_L(p)$.

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

Step 3: Optimize initialized coefficients of the polynomial $h(p)$ until the sum of the squared errors between Z_L and Z_L^C is smaller than the stopping criteria.

Step 4: Synthesize the reflection coefficient $S_{22}(p)$, and obtain the matching network and the termination resistor R_S.

4. Examples

Example 1. In the following, an example is given, to illustrate the implementation of the algorithm defined above.

The same load used in [7] has been chosen. Namely, 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 .

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

Table 1. Given normalized load impedance data

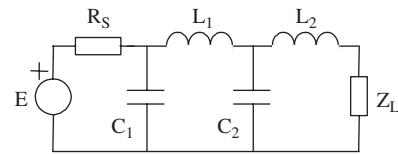| ω_i	$R_L(j\omega_i)$	$X_L(j\omega_i)$
0.1	0.962	−0.092
0.2	0.862	−0.145
0.3	0.735	−0.141
0.4	0.610	−0.088
0.5	0.500	0.000
0.6	0.410	0.108
0.7	0.338	0.227
0.8	0.281	0.351
0.9	0.236	0.476
1.0	0.200	0.600

Fig. 3. Designed lumped-element matching network, proposed and 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: $L_1 = 0.8861$, $L_2 = 0.03621$, $C_1 = 1.106$, $C_2 = 2.678$, $R_S = 0.6321$.

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 has been solved by M-RFT in [7] and the simplified real frequency technique (SRFT). Finally, the obtained element values via the proposed method 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 after final optimization 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 like M-RFT, and as opposed to SRFT. Also a load model has not been utilized in this method, but in M-RFT, a load model must be formed from the given numerical load data. The performance of the matched network has been improved by using a CAD tool.

Example 2. In this example, a matching network is designed to transfer maximum power to a short monopole antenna. Data for the antenna are provided over 20–100 MHz.
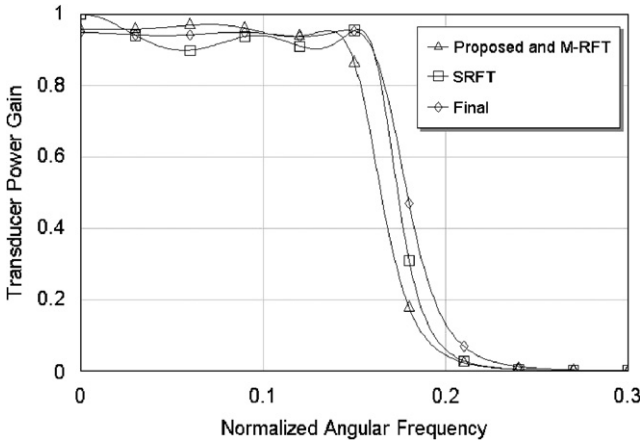
Fig. 4. Performance of the matched system.

Table 2. Reflectance data for the monopole antenna

f_i	Re(S)	Im(S)
0.20	0.9170	−0.3113
0.30	0.5545	−0.5446
0.40	−0.1111	0.0000
0.45	0.3289	0.4698
0.50	0.6437	0.3325
0.55	0.7828	0.2270
0.60	0.8686	0.1248
0.65	0.9040	0.0528
0.70	0.9189	−0.0187
0.75	0.9233	−0.0831
0.80	0.9214	−0.1415
0.90	0.9019	−0.2566
1.00	0.8288	−0.3767

The reflectance data for the antenna are given in Table 2 over a normalized frequency band of [0.2–1].

Desired number of elements in the matching network and polynomial $f(p)$ were selected as $n = 2$ and $f(p) = p$, respectively. Then, by ad-hoc choice of the initial coefficients of the polynomial $h(p)$, the proposed matching algorithm was run. Finally, $S_{22}(p)$ was generated as $S_{22}(p) = -\mu h(-p)/g(p)$, where

$$h(p) = -0.8392p^2 + 0.9605p - 0.0211$$

and

$$g(p) = 0.8392p^2 + 1.3865p + 0.0211.$$

Then, the reflection coefficient $S_{22}(p)$ was synthesized, and the equalizer topology with element values seen in Fig. 5 was obtained. The same matching problem has been solved by using SRFT. Finally, the obtained element values via the proposed method were optimized by using a CAD tool [4].

TPG curves of the matching network designed by the proposed method, SRFT and obtained by a CAD tool after final optimization are given in Fig. 6. Close examination of
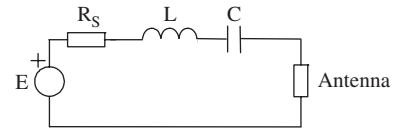

Fig. 5. Designed lumped-element matching network for the monopole antenna: proposed: $L = 3.93991$, $C = 10.0948$, $R_S = 5.5094$; SRFT: $L = 0.58968$, $C = 3.6535$, $R_S = 6.0362$; and final: $L = 3.6260$, $C = 20.3205$, $R_S = 4.167$.
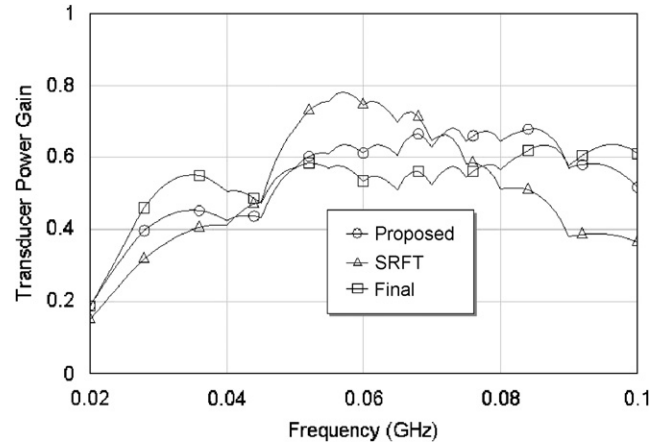

Fig. 6. Performance of the matched system with the monopole antenna.

Fig. 6 reveals that the final available gain curve is nearly obtained without selecting any TPG level as an input as opposed to SRFT. 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. Almost in the all existing broadband matching methods [8–11], the designer must guess and supply this gain level to the algorithm, and free parameters are optimized until reaching to this level. On the other hand, by using analytic theory of broadband matching, this level can be calculated only for simple loads. So designer has to try lots of levels, to find the best one. In M-RFT [7], there is no need to select desired gain level, but the designer must form the model of the load described by numerical data. But in the proposed technique, there is no need to choose any gain level, and no need to form a load model. Algorithm optimizes the free parameters, to fluctuate the gain around the final available level in the passband.

5. Conclusion

For microwave engineers, design of practical matching networks is one of the essential problems. Once the matching network topology is provided, commercially available computer-aided design (CAD) tools are excellent, to optimize system performance. At this point,

initialization process becomes very important, since the system performance is highly nonlinear in terms of the element values of the chosen circuit topology. Therefore, in this paper, a new real frequency technique is proposed, to construct broadband single matching networks. In the method, load impedance is written in terms of *ABCD*-parameters of the matching network and the source resistor. Then, the free parameters are optimized until the error is smaller than an acceptable level, 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, no need to form a model for the load, and no need to guess desired transducer power gain level, the algorithm realizes the optimization, to obtain transducer power gain fluctuating around the final available level in the passband. Eventually, the actual performance of the matched system is improved utilizing a commercially available CAD tool. Two examples are presented, to construct matching networks 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 broadband single matching networks for microwave communication systems.

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