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Design of practical broadband matching networks with commensurate transmission lines

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ABSTRACT

To design broadband matching networks for microwave communication systems, commercially available computer aided design (CAD) tools are always preferred. But these tools need proper matching network topology and element values. Therefore, in this paper, a practical method is proposed to generate distributed-element matching networks with good initial element values. Then, the gain performance of the designed matching network can be optimized employing these tools. The utilization of the proposed method is illustrated by means of the given example. It is shown that proposed method provides very good initials for CAD tools.

1. Introduction

In the design of high frequency communication systems, if the wavelength of the operation frequency is comparable with physical size of the lumped circuit elements, usage of distributed elements is inevitable. Therefore, at Radio Frequencies (RF), design of broadband matching networks with distributed elements or commensurate transmission lines have been considered as a vital problem for engineers [1].

Although analytic theory of broadband matching may be employed for simple problems [2,3], it is well known that this theory is inaccessible except for simple problems. Therefore, for practical applications, it is always preferable to utilize CAD tools, to design matching networks with distributed elements [4–6]. Matched system performance is optimized by all the commercially available CAD tools. At the end of this process, characteristic impedances and the delay lengths of the transmission lines are obtained. But performance optimization is highly nonlinear with respect to characteristic impedances and delay lengths, and requires proper initials [7]. Furthermore, selection of initial values is vital for successful optimization, since the convergence of the optimization depends on the selected initial values.

Therefore, in this paper, a well-established process is proposed, to design broadband matching networks with equal length or commensurate transmission lines. These lines are also called as unit elements (UEs).

2. Broadband matching problem

The broadband matching problem can be considered as the design of a lossless two-port network between a generator and complex load, in such a way that power transfer from the source to the load is maximized over a frequency band. The power transfer capability of the lossless matching network is best measured by means of the transducer power gain which can be defined as the ratio of power delivered to the load to the available power from the generator.

The matching problems can be grouped basically as single matching and double matching problems. In the single matching problems, the generator impedance is purely resistive and the load impedance is complex. On the other hand, if both terminating impedances are complex, then the problem is called as the double matching problem.

Let us consider the classical double matching problem depicted in Fig. 1. Transducer power gain (*TPG*) can be written in terms of the real and imaginary parts of the load impedance $Z_L = R_L + jX_L$ and those of the back-end impedance $Z_2 = R_2 + jX_2$, or in terms of the real and imaginary parts of the generator impedance $Z_G = R_G + jX_G$ and those of the front-end impedance $Z_1 = R_1 + jX_1$ of the matching network as follows:

$$TPG(\omega) = \frac{4R_\alpha R_\beta}{(R_\alpha + R_\beta)^2 + (X_\alpha + X_\beta)^2}. \quad (1)$$

Here if $\alpha = 1$, $\beta = G$, and if $\alpha = 2$, $\beta = L$.

The objective in broadband matching problems is to design the lossless matching network in such a way that *TPG* given by (1) is maximized inside a frequency band. So the matching problem in this formalism can be regarded as the determination of a realizable

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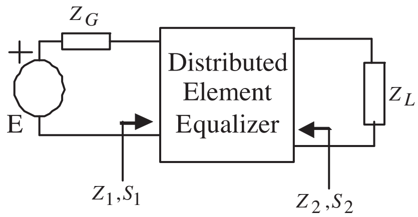

Fig. 1. Double matching arrangement.

impedance function Z_1 or Z_2. Once Z_1 or Z_2 is obtained properly, the lossless matching network can be synthesized easily.

Real frequency line segment technique proposed by Carlin (RF-LST) is one of the best techniques to determine a realizable data set for Z_2 [8,9]. In this method, Z_2 is realized as a minimum reactance function and its real part $R_2(\omega)$ is resembled by line segments in such a way that $R_2(\omega)=\sum_{k=1}^{m}a_k(\omega)R_k$, passing through m-selected pairs designated by $\left\{R_k, \omega_k; k=1,2,\ldots,m\right\}$. Here, break points (or break resistances) R_k are considered as the unknowns of the problem. Then, these points are obtained via nonlinear optimization of *TPG*.

The imaginary part $X_2(\omega)=\sum_{k=1}^{m}b_k(\omega)R_k$ of Z_2 is also expressed by means of the same break points R_k. It is important to note that the coefficients $a_k(\omega)$ are known quantities and they are calculated in terms of the pre-selected break frequencies ω_k. The coefficients $b_k(\omega)$ are obtained by means of Hilbert transformation relation given for minimum reactance functions. If $H\{\circ\}$ represents the Hilbert transformation operator, then $b_k(\omega)=H\left\{a_k(\omega)\right\}$.

In RF-LST, two independent approximation steps seem to be disadvantages of the method. Although it is possible to extend the method to solve double matching problems, the computational efficiency applies only for single matching problems.

The basic principle of the direct computational technique (DCT) is similar to that of the real frequency line segment technique [10]. In this method, the real part of the unknown matching network impedance R_2 is written as a real even rational function. Then the unknown coefficients of this function are optimized to get the best gain performance.

In DCT, the unknown coefficients of R_2 must be determined so that R_2 is a nonnegative even rational function, which in turn ensures the realizability of the resulting impedance function Z_2. So in order to guarantee the realizability, an auxiliary polynomial is utilized for constructing an intrinsically nonnegative real part function R_2. By the introduction of this polynomial, although the realizability is simply ensured, the computational effort and the nonlinearity of the transducer power gain with respect to the optimization parameters are increased.

In Fettweis's method, parametric representation of the positive real back-end driving point impedance Z_2 is utilized [11]. Namely, the positive real impedance Z_2 is expressed in a partial fraction expansion, and then the poles of Z_2 are optimized to get the best gain performance of the system in the frequency band.

The parametric method constitutes an efficient approach for solving single matching problems. The only problem is the initialization of the location of poles, which may be critical.

In all the methods explained briefly above, the lossless matching network is described in terms of a set of free parameters by means of back-end driving point impedance Z_2. But, the matching problem can also be described by means of any other set of parameters. In the real frequency scattering approach which is referred to as the Simplified Real Frequency Technique (SRFT), the canonic polynomial representation of the scattering matrix is employed to describe the lossless matching network [12,13].

In another method proposed in [7,14], the back-end driving point impedance of the matching network Z_2 is modeled as a minimum reactance function, then, if necessary, a Foster impedance is connected in series.

As the result of the explanation above, it is desired to express the back-end impedance Z_2 of the matching network in terms of any set of free parameters. Then gain performance of the matching network is optimized via (1). But the determination of the back-end impedance expression is complicated. There is a very simple and obvious way to determine the back-end impedance Z_2 or front-end impedance Z_1 of the matching network. This is the crux of the proposed method.

In the proposed method, these driving point impedances (Z_2 or Z_1) are determined utilizing the scattering parameters of the lossless matching network, source and load reflection coefficients. So in the next section, canonic polynomial representation of a distributed-element two-port network is briefly summarized, and then rationale of the proposed method is given.

3. Canonic polynomial representation of a distributed element two-port network

Most of the design methods for microwave networks incorporate finite homogenous transmission lines of commensurable lengths as ideal UEs [15]. By commensurate, it must be understood that all line lengths in a network are multiples of the UE length. Richards has shown that the distributed-element networks composed of commensurate transmission lines (UEs) can be proceeded in analysis or synthesis as lumped element networks under the transformation

$$\lambda = \tanh p\tau,$$

where τ is the commensurate delay of the transmission lines, p is the usual complex frequency variable ($p=\sigma+j\omega$) and λ is the so called Richards variable, $\lambda=\Sigma+j\Omega$. Specifically, on the imaginary axis, the transformation takes the form $\lambda=j\Omega=j\tan\omega\tau$.

Referring to the double matching configuration shown in Fig. 1, the scattering parameters of the lossless matching network can be written in terms of three real polynomials by using the well known Belevitch representation as follows:

$$S_{11}(\lambda)=\frac{h(\lambda)}{g(\lambda)}, \qquad S_{12}(\lambda)=\frac{\mu f(-\lambda)}{g(\lambda)},$$
$$S_{21}(\lambda)=\frac{f(\lambda)}{g(\lambda)}, \qquad S_{22}(\lambda)=-\frac{\mu h(-\lambda)}{g(\lambda)}, \tag{2}$$

where g is a strictly Hurwitz polynomial, f is a real polynomial which is constructed on the transmission zeros of the matching network and μ is a unimodular constant ($\mu=\pm1$). If the two-port is reciprocal, then the polynomial f is either even or odd and $\mu=f(-\lambda)/f(\lambda)$.

The polynomials $\left\{f,g,h\right\}$ are related by the Feldtkeller equation

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

It can be concluded from (3) that the Hurwitz polynomial $g(\lambda)$ is a function of $h(\lambda)$ and $f(\lambda)$. If the polynomials $f(\lambda)$ and $h(\lambda)$ are known, then the scattering parameters of the two-port network, and then the network itself can completely be defined.

In almost all practical applications, the designer has an idea about transmission zero locations of the matching network. Hence the polynomial $f(\lambda)$ is usually constructed by the designer. For practical problems, the designer may use the following form of $f(\lambda)$

$$f(\lambda)=f_0(\lambda)(1-\lambda^2)^{n_\lambda/2} \tag{4}$$

where n_λ specifies the number of equal-length transmission lines in cascade, and $f_0(\lambda)$ is an arbitrary real polynomial. A powerful class of networks contains series or shunt stubs and equal-length