Design of Practical Matching Networks With Lumped Elements Via Modeling

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***Abstract*—It is a common practice to utilize commercially available software tools to design matching networks for wireless communication systems. Most of these tools require a properly selected matching network topology with good initial element values. Therefore, in this paper, a practical method is presented to generate matching networks with initial element values. In the implementation process of the proposed method first, the driving point immitance data for the matching network is obtained in a straight forward manner without optimization. Then, it is modeled as a realizable bounded-real input reflection coefficient which in turn yields the desired matching network with reasonable element values. Eventually, the initial design is improved by optimizing the performance of the matched system employing the commercially available computer-aided design (CAD) packages. An example is given to illustrate the utilization of the proposed method. It is shown that new method provides excellent results as a front-end when utilized together with CAD tools.**

***Index Terms*—Broadband matching, modeling, real frequency techniques.**

I. Introduction

FOR all microwave communication systems, design of wide band 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 process is introduced for matching problems. The new initialization method is based on the reflectance modeling via fixed point iteration (FPI). In the following sections first, the theoretical aspects of the new "initialization method" is introduced. Then, the implementation algorithm is presented. Finally, utilization of the algorithm is exhibited by designing a matching network for a measured passive load.

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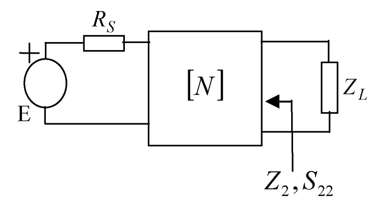

Fig. 1. Single matching arrangement.

II. Generation of Rough Estimate of Driving Point Input Impedance for Matching Network

Let us consider the single matching arrangement as shown in Fig. 1. It is well known that the matching network $[N]$ can completely be specified by the positive real (PR) driving point impedance Z_2 or by the corresponding bounded real (BR) reflectance $S_{22} = (Z_2 - 1)/(Z_2 + 1)$. If one generates $Z_2(j\omega) = R_2(\omega) + jX_2(\omega)$ as a proper data set to optimize the transducer power gain (TPG) of the matched system, then it can be modeled as a PR impedance which in turn yields the desired matching network via synthesis. In fact, Carlin's real frequency line segment technique (RF-LST) is known as the best method to generate the proper or realizable data set for Z_2 [7], [8]. In Carlin's approach, Z_2 is assumed to be minimum reactance function and its real part $R_2(\omega)$ is represented by line segments such that $R_2(\omega) = \sum_{k=1}^{m} a_k(\omega)R_k$, passing through m-selected pairs designated by $\{R_k, \omega_k; k = 1, 2, \ldots, m\}$. In this regard, break points (or break resistances) R_k are considered as the unknowns of the matching problems. Then, these points are determined via nonlinear optimization of TPG, expressed as

$$\text{TPG} = \frac{4R_2R_L}{(R_2 + R_L)^2 + (X_2 + X_L)^2}. \quad (1)$$

In (1) R_L and X_L are the real and the imaginary parts of the measure load data $Z_L(j\omega) = R_L(\omega) + jX_L(\omega)$, respectively, and 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 noted that coefficients $a_k(\omega)$ are known quantities and they are determined in terms of the pre-selected break frequencies ω_k which specify frequency location of the break points R_k. Similarly, coefficients $b_k(\omega)$ are also known and generated by means of Hilbert transformation relation given for minimum reactance functions. In this case, let H{∘} designates the Hilbert transformation operator, then $b_k(\omega) = H\{a_k(\omega)\}$.

In the new technique proposed in this paper, the RF-LST is simply omitted and data for Z_2 are generated without optimization in a straight forward manner as follows.

For a desired shape of TPG $= T(\omega)$ which can even be specified as a set of data points, the ratio defined by $\alpha = R_2/R_L$ can directly be computed under the perfect cancellation condition of the imaginary parts (i.e., $X_2 = -X_L$). Actually, this assumption is a practical one, which maximizes TPG of the matched system over the band of operation.

On the other hand, it is well known that existence of the load network will lower the ideal flat gain from $T(\omega) = 1$, down to a level $T_{\text{flat}} < 1$ in the pass band. Furthermore, TPG must decrease monotonically out side of the band. In this case, one can always select a reasonable-realizable shape for TPG, such as Butterworth or Chybeshev forms, and then, generates the ratio specified by $\alpha = R_2/R_L$ under the perfect cancellation condition. Thus, the data set for the driving point impedance Z_2 given by

$$Z_2(j\omega_i) = R_2(\omega_i) + jX_2(\omega_i) = \alpha(\omega_i)R_L(\omega_i) - jX_L(\omega_i) \quad (2)$$

is computed over the measured frequencies ω_i of the load network.

Let us now derive the ratio $\alpha = R_2/R_L$ when perfect cancellation occurs on the imaginary parts. In this case, TPG is given by

$$\text{TPG} = \frac{4R_2R_L}{(R_2+R_L)^2} \quad (3)$$

or

$$\alpha = \frac{R_2}{R_L} = \frac{(2-\text{TPG}) + 2\mu\sqrt{1-\text{TPG}}}{\text{TPG}} \quad (4)$$

where $\mu = \mp 1$ is a uni-modular constant and lands itself while taking the square-root of $(1 - \text{TPG})$. Obviously, α is derived as a function of the TPG. Hence, for a selected-suitable gain form, the impedance $Z_2 = R_2 + jX_2$ is approximated as

$$R_2 = \alpha R_L \quad (5a)$$

$$X_2 = -X_L. \quad (5b)$$

At this point, it is crucial to choose the form for TPG to describe the matched system performance. In this regard, it may be desirable to have an equal ripple gain shape within the passband as desired in many practical problems. Then, the following low-pass-Chebyshev form may be utilized:

$$\text{TPG} = \frac{T_{\max}}{1 + \varepsilon^2 T_n^2(\omega)} \quad (6a)$$

where ε is the ripple factor and $T_n(\omega)$ is the n^{th} order Chebyshev polynomial. The degree n specifies the total number of reactive elements in the equalizer topology. TPG takes its maximum value $T_{\max}$ at the zeros of the Chebyshev polynomial $T_n(\omega)$. It is minimum (TPG $= T_{\min}$) when $T_n(\omega) = 1$. Obviously, ε is specified by

$$\varepsilon^2 = \frac{T_{\max} - T_{\min}}{T_{\min}} \quad (6b)$$

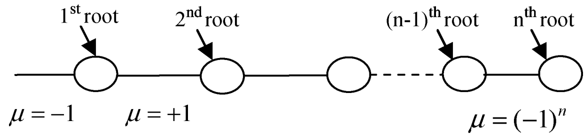

Fig. 2. Selection of the sign of μ.

and the average flat gain level is determined as

$$T_{\text{flat}} = \frac{T_{\max} + T_{\min}}{2}. \quad (6c)$$

Let us point out that beyond simple matching problems, it is almost impossible to determine the ideal value of T_{flat} analytically. Selection of the sign of the uni-modular constant μ of (4) is important to end up with realizable driving point impedance Z_2. In this regard, it is appropriate to flip the sign of μ along the frequency axis as TPG fluctuates around its mean value T_{flat} within the passband. For example, when working with Chebyshev forms of (6), it is known that TPG changes its direction of movement up and down at the roots of the Chebyshev polynomial $T_n(\omega)$. Starting with $\mu = -1$, the sign of μ is flipped as the frequency ω_i of $\alpha(\omega_i)$ moves between the roots of $T_n(\omega)$ of (6a) as shown by Fig. 2.

Once, the data for the driving point impedance $Z_2(j\omega)$ is generated, then it is modeled employing the reflectance based method presented in the following section. Finally, the reflectance model 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.

III. Reflectance Based Data Modelling via Fixed-Point Interpolation

In this section, the reflectance data specified by $S_{22}(j\omega) = (Z_2 - 1)/(Z_2 + 1)$ are considered as the input reflection coefficient of a lossless equalizer $[N]$ and it is modeled as a rational-bounded real scattering coefficient in Belevitch form as

$$S_{22}(j\omega) = \frac{h(j\omega)}{g(j\omega)} = S_R(\omega) + jS_X(\omega) \quad (7a)$$

where ω represents the normalized angular frequency. In the classical literature however, it is referred as the "real frequency."

On the real frequency axis, let the numerator polynomial be

$$h(j\omega) = h_R(\omega) + jh_x(\omega) \quad (7b)$$

and the denominator polynomial be

$$g(j\omega) = g_R(\omega) + jg_x(\omega). \quad (7c)$$

Then, at selected points ω_i, the real part $S_R(\omega_i)$ and the imaginary part $S_X(\omega_i)$ are determined by manipulation as follows:

$$S_R(\omega_i) = \frac{h_R g_R + h_x g_x}{g_R^2 + g_x^2} \quad (7d)$$

and

$$S_X(\omega_i) = \frac{h_x g_R - h_R g_x}{g_R^2 + g_x^2}. \tag{7e}$$

From (7), one can readily obtain h_R and h_x as

$$h_R = [g_R S_R - g_x S_X] \tag{8a}$$

and

$$h_x = -[g_x S_R + g_R S_X]. \tag{8b}$$

The above equations indicate that, if the denominator polynomial $g(j\omega) = g_R(\omega) + jg_X(\omega)$ is known, then the numerator polynomial $h(j\omega) = h_R(\omega) + jh_X(\omega)$ can readily be obtained. In fact, this way of thinking constitutes the crux of the method in the following manner.

At the first glance, the equalizer topology may be constructed with lumped elements, namely by inductors and capacitors. In this case, the model built for the BR reflection coefficient S_{22} will be expressed as a function of the classical complex variable $p = \sigma + j\omega$. Thus, one can write

$$h(p) = S_{22}(p) \cdot g(p) \tag{9a}$$

or employing the concept of interpolation, at a given single frequency ω_i, the following equation must be satisfied:

$$h(j\omega_i) = S_{22}(j\omega_i) \cdot g(j\omega_i). \tag{9b}$$

Since $S_{22}(p) = h(p)/g(p)$ belongs to a lossless-reciprocal two-port and is specified by the given data, then rest of the scattering parameters of $[N]$ are also represented in Belevitch form as

$$S_{12}(p) = S_{21}(p) = \frac{f(p)}{g(p)} \text{ and } S_{11} = -\frac{f}{f_*}\frac{h_*}{g} \tag{10a}$$

satisfying the losslessness condition of

$$gg_* = hh_* + ff_* \tag{10b}$$

where "$*$" designates the complex conjugate (or para conjugate) of the given complex valued quantity. Thus, on the real frequency axis $|S_{21}|^2$ is given by

$$|S_{21}|^2 = 1 - \rho^2 = \frac{|f|^2}{|g|^2} = \frac{|f|^2}{g_R^2 + g_x^2}. \tag{10c}$$

It should be noted that the numerator polynomial $f(p)$ of $S_{21}(p)$ includes transmission zeros of the matching network to be designed. At the frequencies where $f(j\omega)$ vanishes, the reflectance data $|S_{22}(j\omega)|^2$ becomes unity. Hence, the real frequency zeros of $f(j\omega)$ is dictated by the given reflectance data. Furthermore, some practical considerations shape the polynomial form of $f(p)$ which in turn specifies the strictly Hurwitz denominator polynomial $g(p)$ as described in the following paragraph.

For many practical problems, it is customary to work with low-pass *LC* ladders with all transmission zeros located at infinity demanding $f(p) = 1$. This means that for a given BR reflection coefficient $S_{22} = S_R + jS_X$, one can readily compute the amplitude square of the denominator polynomial g, by selecting a proper form of f. Thus,

$$|g|^2 = g_R^2 + g_x^2 = \frac{|f|^2}{1 - \rho^2}. \tag{11}$$

Hence, (11) describes a known quantity over the specified frequencies with pre-selected f. In this case, the Hurwitz polynomial $g(p)$ can be constructed by means of well established numerical methods [9].

Briefly, data points given by (11) for $|g(j\omega)|^2$, describe an even polynomial such that

$$G(\omega^2) = G_0 + G_1\omega^2 + \ldots + G_n\omega^{2n} > 0 \qquad \forall\omega. \tag{12}$$

Coefficients $\{G_0, G_1, G_2, \ldots G_n\}$ can easily be found by linear or non linear interpolation or curve fitting methods. Then, replacing ω^2 by $-p^2$, one can extract $g(p)$ from $G(-p^2) = g(p)g(-p)$ using explicit factorization techniques. At this point, the roots of $G(-p^2)$ may be computed and then $g(p)$ is constructed on the left half plane (LHP) roots of $G(-p^2)$ as a strictly Hurwitz polynomial.

Once $g(p)$ is generated, then $g_R = \mathrm{Re}\{g(j\omega)\}$ and $g_x = \mathrm{Im}\{g(j\omega\}$ are computed which in turn yields the numerical pair of $\{h_R, h_x\}$ by means of (8). Let $h(p) = \sum_{k=0}^{n} h_k p^k$ designate the numerator polynomial of $S_{22}(p) = h(p)/g(p)$. In this representation $\{h_k; k = 0, 1, 2, \ldots, n\}$ are the arbitrary real coefficients, and n specifies the total number of elements in the matching network.

Thus, data points corresponding to the real and the imaginary parts of $h(j\omega)$ are given by

$$h_R(\beta) = \sum_{k=0}^{m} (-1)^k h_{2k} \beta^{2k} \tag{13a}$$

where $m = n/2$ if n is even. $m = (n-1)/2$ if n is odd

$$h_x(\beta) = \sum_{k=1}^{m} (-1)^{k-1} h_{2k-1} \beta^{2k-1} \tag{13b}$$

where $m = n/2$ if n is even. $m = (n+1)/2$ if n is odd.

Then, one can immediately determine the unknown real coefficients $\{h_k; k = 0, 1, 2, \ldots, n\}$ by means of straight linear interpolation over the selected frequencies.

At this point it is crucial to point out that polynomials $g(p)$ and $h(p)$ must satisfy the losslessness condition of $g(p)g(-p) = h(p)h(-p) + f(p)f(-p)$ rather than on the frequencies selected for interpolation. Therefore, herewith, an iterative approach which is named as the "interpolation via fixed-point interpolation" is introduced which yields the consistent triple of $\{h(p), g(p), f(p)\}$ satisfying the losslessness condition.

A. Fixed-Point Interpolation of $h(p)$

In this section, let us first briefly review the technique, as it is described in classical numerical analysis text books such as [10].

Zeros of a nonlinear function $G(X) = F(X) - X$ can be determined using the iterative loop described by

$$X_r = F(X_{r-1}). \tag{14a}$$

It is straight forward to prove that for any initial guess X_0, (14a) converges to one of the real root $X_{root} = \lim_{r\to\infty} F(X_r)$ if and only if $|dF/dX| < 1; \forall X$.

For the problem under consideration, in fact, the numerator polynomial $h(j\omega)$ can be determined point by point by means of an iterative process which may be described employing (9b) over the selected frequencies ω_i such that

$$h_r(j\omega) = S_{22}(j\omega) \cdot g_{r-1}(j\omega). \tag{14b}$$

In this case, one has to show that the term $S_{22} \cdot g$ describes a function $h = F(h)$ for which $|dF/dh| < 1; \forall h$.

In the following, first the iterative process of (14b) is described, then its convergence is proven.

After selecting $f(p)$, in (14b), $g_0(p)$ is generated solely in terms of the given data $S_{22}(j\omega)$ employing the explicit factorization of (12) as described above. Then, the first loop is initiated by computing $h_1(j\omega_i) = h_R(\omega_i) + jh_x(\omega_i)$ over the chosen set of frequencies $\{\omega_i; i = 01, 2 \ldots n\}$ and using (13), analytic form of $h_1(p)$ is obtained by means of a linear interpolation algorithm.

Employing the losslessness equation

$$\begin{aligned} G_1(-p^2) &= g_1(p)g_1(-p) = h_1(p)h_1(-p) + f(p)f(-p) \\ &= G_{10} - G_{11}s^2 + \ldots + (-1)^{n_s}p^{2n}. \end{aligned} \tag{15a}$$

$g_1(p)$ is generated on the LHP roots of $G_1(-p^2)$. Hence, the second iteration loop starts on the computed $g_1(j\omega_i)$ which yields $h_2(j\omega_i)$. Then, g_2 is constructed yielding h_3 etc. Iterative loops continue until $\|h_r - h_{r-1}\| \le \delta$. Here, δ is selected as a negligibly small positive number to terminate the iterations.

The above process describes the interpolation of $h(p)$ via over the selected frequencies. As a matter of fact, the denominator polynomial g can be described in terms of the numerator polynomial h by using losslessness condition

$$\begin{aligned} g(j\omega) &= h(j\omega)\frac{h(-j\omega)}{g(-j\omega)} + f(j\omega)\frac{f(-j\omega)}{g(-j\omega)} \\ &= h(j\omega)S_{22}(-j\omega) + f(j\omega)S_{21}(-j\omega). \end{aligned} \tag{15b}$$

Using (15) in (14) one obtains

$$h(j\omega) = h(j\omega)\rho^2 + S_{22}(j\omega)f(j\omega)S_{21}(-j\omega) \tag{16}$$

where $\rho = |S_{22}(j\omega)|$ and it is specified by the given data.

In short, right-hand side of (16) describes a function F in h such that $F(h) = h.\rho^2 + S_{22}.f.S_{21}^*$

Therefore, $h = F(h)$ describes a convergent process provided that $|dF/dh| < 1$. In fact, $S_{21}(j\omega)$ is also specified by means of $S_{22}(j\omega)$ and pre-selected $f(j\omega)$. Then, practically, $dF/dh = \rho^2 < 1$ over the entire frequencies by bounded realness; except at isolated points where ρ hits unity. Thus, for the given reflection coefficient data, the polynomial form of $h(p)$ is readily obtained via of $h = F(h)$ which in turn results in a realizable driving point reflectance $S_{22}(p) = h(p)/g(p)$.

The above results can be collected under the following theorem to generate the reflectance based circuit model.

B. Theorem: Modeling via Fixed-Point Interpolation

Referring to Fig. 1, let $S_{22}(j\omega_i) = R(\omega_i) + jX(\omega_i)$ be the input reflectance coefficient data of the lossless matching network $[N]$ specified over the real frequency points ω_i such that $|S_{22}(j\omega)| < 1$ for all frequencies. Let $\{S_{ij}; i, j = 1, 2\}$ be the real normalized bounded real scattering parameters of the lossless matching network $[N]$ described in Belevitch sense. Once, the polynomial f of $S_{21} = f/g$ is selected properly, then, the iterative process $h_r = F(h_{r-1})$ described by (14b) is always convergent and yields the numerator polynomial h of $S_{22} = h/g$ satisfying the losslessness condition of $gg_* = hh_* + ff_*$.

Obviously, proof of this theorem follows as in above.

Depending on the modeling problem under consideration, numerical implementation of the method may require some care. Therefore, in the following section some practical issues are covered.

IV. Numerical Aspects

In order to end up with a successful equalizer design, the fit between the generated reflectance data and the model must be as good as possible. In this regard, it has been experienced that the following numerical implementation which is called the "Foster Approach" improves the quality of data fitting [9].

For the sake of completeness of the paper herewith, the Foster Approach is summarized.

A. Foster Approach to Model the Given Data

In this approach, $Z_2 = R_2 + jX_2 = \alpha R_L - jX_L$ is decomposed into its minimum $Z_M = R_M + jX_M = R_2 + jX_M$ and Foster parts as in [9]. Thus, it is expressed as $Z_2 = Z_M + Z_F$. In this representation, Z_M is a minimum reactance function and its imaginary part is computed using the Hilbert transformation relation such that

$$X_M(\omega) = H\{R_2(\omega)\} = \frac{2\omega}{\pi}\int_0^\infty \frac{R_2(y)}{y^2 - \omega^2}dy. \tag{17}$$

This integral can easily be computed numerically since $R_2(\omega)$ rolls off within couple of octaves outside the band of operation depending on the shape of TPG. In practice, one octave beyond the upper edge of the pass band is even sufficient to assume $R_2(\omega) \approx 0$.

In this case, $S_M = (Z_M - 1)/(Z_M + 1)$ is modeled using the method. The Foster part Z_F is specified point by point over the real frequencies such that $Z_F(j\omega) = jX_F(\omega)$ with $X_F(\omega) = -(X_L + X_M)$ and it is modeled employing the Foster form given by

$$X_F(\omega) = k_\infty \omega - \frac{k_0}{\omega} + \sum_{p=1}^{N_p} \frac{k_p \omega}{\omega_p^2 - \omega^2}. \tag{18}$$

In (18), residues k_∞, k_0, and k_p are determined via linear interpolation or curve fitting techniques provided that the poles ω_p are fixed properly in advance outside the passband [9]. In many daily life problems however, it is sufficient to employ only the first term or the second term of X_F. Rarely, the first two terms or some times one or two finite poles may be required in the foster part to improve fitting quality.

B. Selection of the Transmission Zeros

One of the crucial issues of the design process is the selection of the transmission zeros of the equalizer which are included in the numerator polynomial $f(p)$ of the transfer scattering parameter $S_{21}(p) = f(p)/g(p)$. When working with lumped elements, it is well known by classical theory that $f(p)$ has the following general form [10]:

$$f(p) = p^{k_1} \prod_{j=1}^{k_2} \left(p^2 + \omega_{zj}^2\right) \prod_{i=1}^{k_3} (p - \sigma_i) \times \prod_{l=1}^{k_4} \left[(p + \sigma_l)^2 + \omega_{zl}^2\right] \left[(p - \sigma_l)^2 + \omega_{zl}^2\right] \quad (\sigma_i, \sigma_l) > 0 \tag{19}$$

such that $k_1 + 2k_2 + k_3 + 4k_4 \le n$ where n is the degree of the denominator polynomial $g(p) = g_0 + g_1 p + g_2 p^2 + \ldots + g_n p^n$ and specifies the total number of elements in the equalizer to be designed.

General form given by (19) is almost impossible to be manufactured physically. Therefore, in many matching problems $k_1 = k_2 = k_3 = k_4 = 0$ is selected when appropriate. In other words, designers prefer to employ the simplest form of $f(p) = 1$. Depending on the matching problem, rarely the bandpass forms $f(p) = p^{k_1}$ may be utilized.

On the other hand, if $1 - |\rho|^2$ hits the real frequency axis at some points ω_{zj} then in (19), k_2 will be the count of these hits and the real frequency zeros are easily placed in $f(p)$ as the multiplicative terms of $(p^2 + \omega_{zj}^2)$.

C. Normalizations

In the course of design process, numerical stability is maintained by means of frequency and impedance normalizations. In other words, all the computations must be carried out in the normalized domain. Eventually, de-normalization is performed on the final element values of the matching network. In this regard, it may be appropriate to normalize the frequencies at the upper edge of the frequency band. For the impedance normalization, standard $R_0 = 50\ \mathrm{ohm}$ termination may be utilized.

D. Selected Forms of TPG

It has been experienced that utilization of monotone roll-off Chebyshev transfer functions are useful to generate matching networks with initial element values. For low-pass proto-types, TPG is given by

$$\mathrm{TPG} = \frac{T_{\max}}{1 + \varepsilon^2 \left[\cos\left(n \cos^{-1}(\omega)\right)\right]^2}. \tag{20}$$

The above form results in an equal ripple monotone-roll-off transfer function over the frequency band $-1 \le B(\omega) \le +1$.

For bandpass problems described by $\omega_1 \le B(\omega) \le \omega_2$, first, the frequency band dictated by (20) must be normalized to yield the desired band width over $-\omega_c \le B(\omega) \le +\omega_c$ such that $\omega_c = B/2 = (\omega_2 - \omega_1)/2$ and then it is shifted by an amount of $(B/2) + \omega_1$ to obtain the required shape of the TPG in the frequency interval specified by $\omega_1 \le B(\omega) \le \omega_2$. This process replaces the frequency ω of (20) by

$$\omega \Rightarrow \left[\omega - \left(\frac{B}{2} + \omega_1\right)\right] \Big/ \omega_c. \tag{21}$$

E. Equalizer Design Based on Driving Point Admittance

So far, the lossless matching network to be designed has been described in terms of its PR driving point impedance $Z_2 = R_2 + jX_2$ or equivalently by the corresponding BR reflection coefficient $S_{22} = (Z_2 - 1)/(Z_2 + 1)$. Certainly, the above description can as well be made by means of PR admittance function $Y_2 = 1/Z_2 = G_2 + jB_2$ which in turn results in $S_{22} = (1 - Y_2)/(1 + Y_2)$. In this case, TPG given by (1) expressed in terms of admittances as $\mathrm{TPG} = 4G_2G_L/((G_2 + G_L)^2 + (B_2 + B_L)^2)$; where load $Y_L(j\omega_i) = G_L(\omega_i) + jB_L(\omega_i)$ is the load admittance. Similarly, (4) becomes $\alpha = G_2/G_L$. Then, rest of the notation is preserved through out the modelling process described in this paper.

Eventually, the lossless equalizer is built by synthesizing either the driving point admittance or the impedance function which ever is preferred.

For the sake of clear understanding, let us now summarize the details of the proposed design procedure in the following algorithm.

V. Algorithm: Construction of Lossless Matching Networks Without Optimization

This algorithm outlines the procedure to construct lossless equalizers for single matching problems without optimization.

Inputs:

- Measured load data in the form of impedance or admittance: $\{R_L(\omega_i), X_L(\omega_i); i = 1, 2, \ldots, N\}$ where N designates the total sample points.
- Desired form of the TPG TPG $= T(\omega)$ over the entire frequency band: It should be noted that this form can be input either in closed form as in (6) or as sample points. In this manner, monotone-roll off Chebyshev forms is recommended as in Section IV-E.
- Realizable gain levels $T_{\max}$ and $T_{\min}$ over the pass band: In this regard, $T_{\max}$ and $T_{\min}$ are selected with practical

considerations. For example, a low-pass matching network which is free of ideal transformer, demands $T_{\max} = 1$. On the other hand, $T_{\min}$ may be selected as the allowable minimum gain level in the passband.

- Lower (f_{LE} or f_1) and the upper (f_{UE} or f_2) edges of the passband.
- Normalization frequency f_{Norm} (or f_c), Impedance Normalization Number R_0 in ohms.
- n: Desired number of elements in the equalizer.
- Selected form of the numerator polynomial $f(p)$ of the transfer scattering parameter S_{21} (see Section IV-B).
- δ: Stopping criteria selected to terminate the fixed-point interpolations. Note that if the computations are run on PC, δ is usually selected as $10^{-5} \leq \delta \leq 10^{-3}$.

Outputs:

- Analytic form of the input reflection coefficient of the lossless equalizer given in Belevitch form of $S_{22}(p) = h(p)/g(p) = (h_0 + h_1 p + \ldots + h_n p^n)/(g_0 + g_1 p + \ldots + g_n p^n)$. It is noted that this algorithm determines the coefficients $\{h_0, h_1, \ldots, h_n\}$ and $\{g_0, g_1, \ldots, g_n\}$, which in turn optimizes the matched system performance.
- Circuit topology of the lossless equalizer with element values: The circuit topology with element values is obtained as the result of the synthesis of $S_{22}(p)$ in series with the foster section. Synthesis is carried out in Darlington sense. That is, $S_{22}(p)$ is synthesized as a lossless two-port which is the desired equalizer.

Computational Steps:

Step 1:

(a) Normalize the measured frequencies with respect to f_{Norm} and set all the normalized angular frequencies

$$\omega_i = f_{i_{(\text{measured})}}/f_{\text{Norm}}$$

(b) Normalize the measured load impedance with respect to normalization number R_0 and $R_L = R_{L\text{measured}}/R_0$; $X_L = X_{L\text{measured}}/R_0$ over the entire frequency band. It should be noted that if the load is specified as the measured admittance data then, the normalization resistance R_0 multiplies the measured real and the imaginary parts of the admittance data (i.e., $G_L = G_{L\text{measured}} * R_0$; $B_L = B_{L\text{measured}} * R_0$).

Step 2: Employing $T_{\max}$ and $T_{\min}$, compute the ripple factor $\varepsilon^2 = (T_{\max} - T_{\min})/T_{\min}$ as in (6b).

Step 3: Compute the real roots of the Chebyshev polynomial in ascending order $-\omega_{Rn} < \ldots < -\omega_{R1} < \ldots < +\omega_{R1} < \ldots < +\omega_{Rn}$ for the given degree n.

Step 4: Using the positive roots, constitute frequency intervals I_k such that

$$I_1 = \{\omega_1 \leq \omega \langle \omega_{R1}\}, I_2 = \{\omega_{R1} \leq \omega \langle \omega_{R2}\}$$
$$I_3 = \{\omega_{R2} \leq \omega \langle \omega_{R3}\}, \ldots, I_{n+1} = \{\omega_{Rn} \leq \omega \leq \omega_2.\}.$$

Step 5: Compute $\alpha(\omega_i)$ using (4) over the frequencies for which the load data is measured. In the course of computations set $\mu = (-1)^i$ when $\omega \in I_i$.

Step 6: Compute the real part $\{R_2(\omega_i) = \alpha(\omega_i) R_L(\omega); i = 1, 2, \ldots, N\}$ point by point and using line segment approach, extrapolate it beyond the measured frequencies. At this step, it may be suitable to fix $R_2 = 1$ at DC (i.e., $\omega = 0$) and $R_2 = 0$ for $\omega > 1.5\omega_{UE}$ for low-pass designs (i.e., when $f(p) = 1$).

Step 7: Generate the minimum reactance function $X_M(\omega_i)$ point by point using (17) and compute the Foster data $X_F = -X_L - X_M$ over the measured frequencies ω_i.

Step 8: Generate the reflection coefficient $S_M = ((R_2 + jX_M) - 1)/((R_2 + jX_M) + 1) = S_R + jS_X$ over the measured frequencies.

Step 9: Employing the method, model the reflection coefficient as $S_M = (h_0 + h_1 p + \ldots + h_n p^n)/(g_0 + g_1 p + \ldots + g_n p^n)$. Note that the process described by (7)–(16) stops when $\|h_r - h_{r-1}\| \leq \delta$.

Step 10: Using the Foster approach, model the generated data given for X_F as in (18). It should be remarked that in practice, the Foster part must be as simple as possible. The simplest situation is no Foster part. However, a simple Foster topology may be described with a single inductor and perhaps in series with a capacitor or it may be just a single series capacitor. On the other hand, step 7 and this step can simply be omitted if one directly generates and models the reflectance data given by $S_{22} = ((R_2 - jX_L) - 1)/((R_2 - jX_L) + 1)$ as in step 8.

Step 11: Synthesize the modeled reflectance as a lossless two-port terminated in R_0. In this step, decomposition technique of Fettweis [11], zero shifting method or simple continuous fraction expansion can be used to end up with equalizer topology with normalized-initial element values. Then, actual element values are obtained by de-normalization. In this case, actual element values are given by

$$\begin{aligned}&\text{Actual Capacitor}\\&\quad = (\text{Normalized Capacitor}/2\pi f_{\text{Norm}})/R_0\\&\text{Actual Inductor}\\&\quad = (\text{Normalized Inductor}/2\pi f_{\text{Norm}})R_0\\&\text{Actual Line Impedance}\\&\quad = (\text{Normalized Line Impedance})R_0.\end{aligned}$$

Eventually, the above algorithm can be integrated with a commercially available CAD package to further improve the performance of the matched system via optimization [4]–[6].

VI. Comments on the Nonlinearity of TPG Function

In order to appreciate the usage of the newly proposed FPI technique presented above, let us comment on the nature of the matching problem as far as its nonlinear behavior is concerned.

Assume that we try to solve the matching problem using the well commercialized CAD procedures where the designer first selects the circuit topology and then, determines its element values to optimize the TPG of the matched structure. In this regard, consider a simple case where we start with a low-pass *LC* ladder which consists of n-sec*tions*. Let us designate the elements values of this ladder by X_i. In other words, X_i either designates a series inductance L_i or a shunt capacitor C_i. Let us now derive the driving point impedance $Z_2(p)$ in terms of the elements values X_i when the source end is terminated in unit resistance.

Thus, $Z_2(p) = N(p)/D(p)$ is given by

$$Z_2(p) = X_1 p + \frac{1}{X_2 p + \frac{1}{X_3 p + \frac{1}{\cdots\cdots\cdots\cdots\cdots\cdots + \frac{1}{X_n p + 1}}}}.$$

For example if $n = 1$, $Z_2(p) = X_1 p + 1$. For $n = 2$, $Z_2(p) = (X_1 X_2 p^2 + X_2 p + 1)/(X_1 p + 1)$; or $n = 5$ yields

$$Z_2(p) = \frac{N^{(5)}(p)}{D^{(4)}(p)} = \frac{a_5 p^5 + a_4 p^4 + a_3 p^3 + a_2 p^2 + a_1 p + 1}{b_4 p^4 + b_3 p^3 + b_2 p^2 + b_1 p + 1}$$

where

$$\begin{aligned}
a_5 &= X_1 X_2 X_3 X_4 X_5\\
a_4 &= X_2 X_3 X_4 X_5\\
a_3 &= X_1 X_4 X_5 + X_2 X_4 X_5 + X_1 X_2 X_5\\
a_2 &= X_2 X_5 + X_4 X_5\\
a_1 &= X_1 + X_2 + X_3\\
b_4 &= X_1 X_2 X_3 X_4\\
b_3 &= X_1 X_4 + X_2 X_4 + X_1 X_2\\
b_2 &= X_1 X_4 + X_2 X_4 + X_1 X_2\\
b_1 &= X_2 + X_4.
\end{aligned}$$

As far as the measure of the nonlinearity is concerned, the polynomial $N^{(1)}(p) = X_1 p + 1$ is said to be linear in variable X_1, the polynomial $N^{(2)}(p) = X_1 X_2 p^2 + X_2 p + 1$ is quadratic in X_1 and X_2. Similarly, polynomials $N^{(5)}(p)$ and $D^{(4)}(p)$ have degree of nonlinearity $d_{\text{non}} = 5$ and $d_{\text{non}} = 4$ in variables X_i; (i = 1,2,3,4,5) respectively.

Obviously, nonlinearities double when we work with the norms of the above polynomials.

For example, the even polynomial

$$\begin{aligned}
\hat{N}^{(10)}(\omega^2) &= N^{(5)}(j\omega) N^{(5)}(-j\omega)\\
&= \hat{a}_5 \omega^{10} + \hat{a}_4 \omega^8 + \hat{a}_3 \omega^6 + \hat{a}_2 \omega^4 + \hat{a}_1 \omega^2 + 1
\end{aligned}$$

has degree of nonlinearity $d_{\text{non}} = 10$ in variables X_i since its leading coefficient is specified by $\hat{a_5} = -(X_1 X_2 X_3 X_4 X_5)^2$ doubling the nonlinearity.

In short, we say that for an *n-element* ladder network, degree of nonlinearity of the norm function $\hat{N}(\omega^2)$ is $d_{\text{non}} = 2n$ in element values X_i. Now, let us consider the scattering parameters $S_{22}(p)$ and $S_{21}(p)$ of the *LC* ladder under consideration.

By proper normalization, $S_{22}(p)$ is given by

$$S_{22}(p) = \frac{Z_2(p) - 1}{Z_2(p) + 1} = \frac{N(p) - D(p)}{N(p) + D(p)} = \frac{h(p)}{g(p)}$$

where $h(p) = N(p) - D(p) = h_0 + h_1 p + \ldots + h_n p^n$ and $g(p) = N(p) + D(p)$.

The transfer scattering parameter $S_{21}(p)$ is specified as $S_{21}(p) = 1/g(p)$. It should be noted that losslessness condition requires $g(p)g(-p) = h(p)h(-p) + 1$.

Now, let us consider the simplest hypothetical matching problem where the source and the load networks are purely resistive.[1] In this case, the TPG is given by

$$T(\omega) = |S_{21}(j\omega)|^2 = \frac{1}{g(j\omega)g(-j\omega)} = \frac{1}{h(j\omega)h(-j\omega) + 1}$$

or equivalently the denominator polynomial

$$\begin{aligned}
P(\omega^2) &= \frac{1}{T(\omega)} = g(j\omega)g(-j\omega)\\
&= h(j\omega)h(-j\omega) + 1.
\end{aligned}$$

As explained above, $P(\omega^2)$ has the nonlinearity degree of $d_{\text{non}} = 2n$ in terms of the element values X_i since the nonlinearity is specified by the norm function

$$g(j\omega)g(-j\omega) = [D(j\omega) - N(j\omega)]\,[D(-j\omega) - N(-j\omega)]\,.$$

On the other hand, nonlinearity is always quadratic in terms of the real coefficients $\{h_0, h_1, \ldots, h_n\}$ of the polynomial $h(p)$ no matter what the total number of circuit elements are.

For the matching problem under consideration, maximization of the gain function is equivalent to minimization of the polynomial $P(\omega^2)$ in terms of the selected unknowns over the specified bandwidth. In the theory of optimization, it is well known that, if the degree of the nonlinearity of the objective function, which is subject to minimization, goes beyond 2 (in this case $P(\omega^2)$) then, one may easily be trapped in local minima and perhaps ultimate convergence becomes impossible.

Therefore, the designer, who starts the matching problem with the selection of appropriate circuit topology, must have excellent initial element values to end up with a successful optimization.

On the other hand, if the designer initiates the matching network design on the real coefficients of the polynomial $h(p)$, for sure, the optimization is quadratic and the convergence is guaranteed. This is the situation for the hypothetical problem stated above.

If the load network is complex then, optimization becomes even harder on the element values. However, the newly proposed FPI method is always convergent as proven above and results in optimum matching network topology with element values.

It should be noted that for narrow bandwidth problems, effect of the high degree nonlinearities beyond 2 may be neglected for

[1]This is actually the well-established filter design problem. We can always understand matching problem as a special filter design problem where resistive terminations are degenerated gradually to converge to the given complex source and the load impedances [1].

TABLE I
GIVEN NORMALIZED IMPEDANCE DATA

ω	R_L	X_L
0.0	1.00	0.00
0.1	0.86	-0.34
0.2	0.60	-0.49
0.3	0.41	-0.49
0.4	0.28	-0.45
0.5	0.20	-0.40
0.6	0.14	-0.35
0.7	0.11	-0.32
0.8	0.09	-0.28
0.9	0.07	-0.26
1.0	0.06	-0.23

the small frequency values. In this case, one may wish to directly start with a simple circuit topology with one or two elements in the matching network then, proceeds with optimization on the element values. However, for wideband matching problems, this approach usually does not work due to nonlinear behavior of the TPG function unless one starts with good initial element values. Therefore, it is highly recommended to generate the matching network topology with element values using the design procedure presented in this paper. Having obtained the matching network topology with excellent element values, one can always carry out further simulations and re-optimize the physical dimensions of the circuit layout employing the commercially available CAD packages.

Let us now present an example to design a practical matching network for a physical one port device described in the following section.

VII. EXAMPLE

In this section, an example is presented to design a practical matching network for a physical one port device for which the normalized impedance data is given by Table I.

It should be noted that the above data can easily be modeled using the FPI technique as a capacitor $C_L = 4$ in parallel with a resistance $R_L = 1$ (i.e., $R_L//C_L$ type of load). In this case, using Fano's or Youla's relations [1]–[3], the ideal flat gain level T_{flat} is computed as

$$T_{\text{flat}} = 1 - \exp(-2\pi/R_L C_L \omega_c) = 1 - \exp(-2\pi/1 \cdot 4 \cdot 1) = 0.7921.$$

Let us design the equalizer over the normalized pass band of $0 \le B \le 1$. Thus, a low-pass Chebyshev transfer function of (6) can be utilized. In this manner, let us choose $T_{\max} = 1$ and $T_{\min} = 0.792$, then the ripple factor ε^2 is found as

$$\varepsilon^2 = \frac{T_{\max} - T_{\min}}{T_{\min}} = \frac{1 - 0.7921}{0.7921} = 0.2625.$$

TABLE II
CALCULATED TPG, α, R_2, X_M AND X_F DATA

ω	*TPG*	α	R_2	X_M	X_F
0.0	0.7921	0.3737	0.3737	0.00	0.00
0.1	0.8690	0.4685	0.4039	-0.036	0.3809
0.2	0.9496	0.6334	0.3862	-0.094	0.5821
0.3	0.9953	0.8712	0.3570	-0.134	0.6256
0.4	0.9930	1.1830	0.3323	-0.162	0.6114
0.5	0.9532	1.5520	0.3104	-0.188	0.5881
0.6	0.8978	1.9401	0.2870	-0.215	0.5698
0.7	0.8457	2.2937	0.2595	-0.241	0.5578
0.8	0.8087	2.5548	0.2273	-0.266	0.5503
0.9	0.7925	2.6730	0.1915	-0.289	0.5470
1.0	0.8003	2.6161	0.1539	-0.334	0.5694

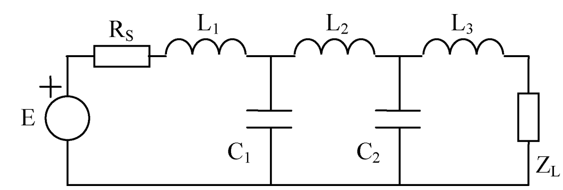

Fig. 3. Lumped element equalizer topology with initial element values: $L_1 = 0.3311, L_2 = 0.6550, L_3 = 0.6078, C_1 = 3.8438, C_2 = 4.8705, R_S = 0.3796$.

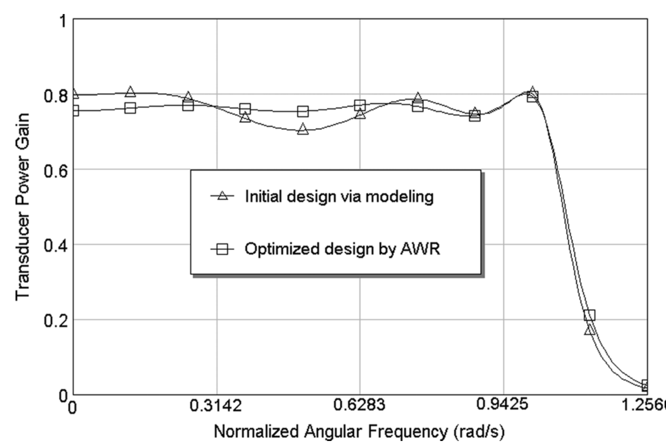

Fig. 4. Performance of the matched system designed with lumped elements.

To ease the physical implementation, let us employ only four elements in the equalizer topology. Thus, selecting $n = 4$, preselected form of TPG is found as $\text{TPG} = 1/(1+\varepsilon^2 T_4^2(\omega))$ with $T_4(\omega) = 8\omega^4 - 8\omega^2 + 1$.

Using the above inputs, the proposed design algorithm is run. Thus, the quantities α, minimum reactive impedance $Z_M = R_M + jX_M$ and Foster part $X_F = -(X_L + X_M)$ are computed point by point as listed in Table II.

Eventually, the driving point reflectance is modeled and synthesized as follows. First the reflectance $S_M(p) = h(p)/g(p)$ which corresponds to the minimum reactance driving point impedance is modeled utilizing the FPI technique. Thus, selecting $f(p) = 1$, $h(p)$ and $g(p)$ polynomials are found as

$$h_{MR}(p) = -3.2944p^4 - 3.1010p^3 - 4.1546p^2 - 1.8843p - 0.5035$$
$$g_{MR}(p) = 3.2944p^4 + 4.4539p^3 + 5.7057p^2 + 3.4847p + 1.1196.$$

Then, the Foster data is modeled as a simple series inductor $L_3 = 0.6078$.

Finally, $S_M(p) = h(p)/g(p)$ is synthesized and it is connected in series with the Foster part yielding the equalizer topology with initial element values as shown in Fig. 3.

As it is seen from Fig. 4, initial performance of the matched system looks pretty good. However, it is further improved via optimization utilizing the commercially available design package called Microwave Office of Applied Wave Research Inc. (AWR) [4]. Thus, the final normalized elements values are given as $\mathrm{L}_1 = 0.2433$, $\mathrm{L}_2 = 0.579$, $\mathrm{L}_3 = 0.5937$, $\mathrm{C}_1 = 4.197$, $\mathrm{C}_2 = 5.061$, $\mathrm{R}_\mathrm{S} = 0.3382$. For comparison purpose, both initial and the final performances of the matched system are depicted in Fig. 4.

VIII. 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 (CAD-Tools) are utilized. Once the matching network topology is provided, these packages are excellent tools to optimize system performance by working on the physical dimensions of the circuit elements. From the practical point of view, the designer prefers to select a proper topology suitable for production. 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, an "Easy to Use" initialization procedure is proposed to construct lossless equalizers for matching problems. The new procedure consists of three major steps. In the first step, for a pre-selected TPG form, optimum input reflectance of the equalizer is generated as data set. Then, this data is modeled as a Bounded-Real reflectance function via method. Finally, it is synthesized as a lossless two-port in resistive termination yielding the desired equalizer topology with initial element values. Eventually, the actual performance of the matched system is improved utilizing a commercially available CAD tool which in turn results in the physical layout of the matching network to be manufactured as a microwave monolithic integrated circuit. An example is 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 procedure 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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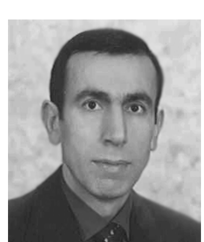

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