Design of Transformerless Quasi-Broad-Band Matching Networks for Lumped Complex Loads Using Nonuniform Transmission Lines

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Abstract—**A simple design method for transformerless and lossless quasi-broad-band matching of a lumped *RC* load is presented by use of a parabolic nonuniform transmission line. The key idea in removing an ideal transformer from the matching network is based on impedance transformation of the nonuniform transmission line, whose mixed lumped and distributed equivalent circuit contains an ideal transformer. Also, illustrative examples and some design curves are presented.**

I. Introduction

THE CLASSIC MATCHING problem was initiated by Bode [1] for complex *RC* and *RL* loads and solved by Fano [2] for an arbitrary load impedance. Youla [3] refined the new theory of broad-band matching based on the principle of complex normalization. Fano's and Youla's techniques have been extended and elaborated by many authors [4]–[10]. Recently, Carlin *et al.* have developed a CAD method called the real frequency method [11]–[14].

In the field of microwave engineering, many design methods also have been established by using distributed parameter elements such as quarter-wave transformers, noncommensurate lines, nonuniform transmission lines, and so on [15]–[22].

A passive broad-band matching network, according to Youla [3], may be designed by starting from a calculation of the gain–bandwidth restriction imposed by the prescribed load. The resulting matching network will often contain an ideal transformer to satisfy the gain–bandwidth restriction except in special cases. In practice, the realization of the ideal transformer is difficult, and should be avoided unless it can be realized as a coupled-line transformer.

In this paper, we present a simple design method using parabolic tapered nonuniform transmission line (PTL) to achieve transformerless and lossless quasi-broad-band matching between a resistive generator and a complex load. By using PTL, we can eliminate an ideal transformer from the matching network. However, this will cause a decrease of theoretical maximum gain prescribed by the broad-band matching theory [3]–[5]. Since the broad-band matching theory requires the best gain performance in the passband, our proposed matching technique may be termed quasi-broad-band matching. A PTL, whose equivalent circuit is represented by mixed lumped and distributed elements, can transform the lumped series *RC* loads into a different lumped impedance. This transformation of the load results in a new impedance, mainly dependent on the taper ratio of the PTL prior to calculation of the gain–bandwidth restriction. Recalculation of the restriction to the transformed load impedance gives a modified restriction equation which contains the taper ratio of the PTL explicitly. We are able to find this taper ratio by solving a simple nonlinear equation so as to avoid the ideal transformer in the matching network structure. The maximum obtainable gain for this transformed impedance will be lower than that of the original one.

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II. Impedance Transformation with Parabolic Tapered Transmission Line

The characteristic impedance distribution of the parabolic tapered transmission line (PTL) is given by [23], [24]

$$W(x)=W_0\left(1+\frac{1}{K_1}\frac{x}{l}\right)^2 \qquad (1)$$

where W_0 is the front-end $(x=0)$ characteristic impedance, K_1 is a positive constant, and l is the line length of the PTL. A PTL loaded by a lumped series *RC* impedance Z_L',

$$Z_L'=R_L+\frac{1}{j\omega C_L} \qquad (2)$$

is shown in Fig. 1(a) and its equivalent circuit is shown in Fig. 1(b) [23]. In the equivalent circuit, element values are given as follows:

$$k=1+\frac{1}{K_1}>1 \qquad (3)$$

$$\mathbb{W}_0=k^2W_0 \qquad (4)$$

$$\mathbb{C}_0=(1+K_1)l/(k^2W_0v) \qquad (5)$$

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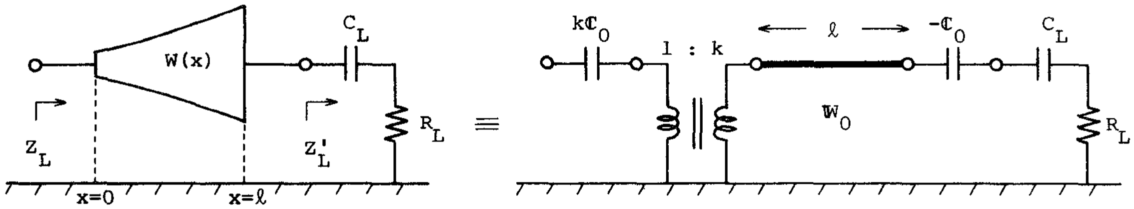

Fig. 1. (a) The parabolic tapered transmission line loaded by lumped series RC impedance and (b) its equivalent circuit.

where v denotes the velocity of the wave. If we set $\mathbb{C}_0 = C_L$ and $\mathbb{W}_0 = R_L$, then the driving-point impedance Z_L becomes [24]

$$Z_L = \frac{R_L}{k^2} + \frac{1}{j\omega k C_L}. \tag{6}$$

A PTL having a parameter value k larger than unity decreases the real part of the original load impedance and must be determined prior to the design of a quasi-optimum impedance matching network.

III. Gain–Bandwidth Restriction for Quasi-Broad-Band Matching Network Design Imposed by the Transformed Load

Consider the scheme of broad-band matching shown in Fig. 2. Our objective is to match the transformed load Z_L in (6) to a resistive generator with a lossless and transformerless two-port network. Since the series RC load is a high-pass circuit, we consider an nth order high-pass Butterworth transducer power gain characteristic between the generator and the load,

$$G(\omega^2) = \frac{K(\omega/\omega_c)^{2n}}{1+(\omega/\omega_c)^{2n}}, \qquad 0 \leqslant K \leqslant 1 \tag{7}$$

where ω_c is the 3-dB radian frequency and K is the maximum gain at $\omega \to \infty$.

To obtain a gain–bandwidth restriction imposed on the matching network N by the load, we employ the well-known broad-band matching theory [3], [4]. The series RC load Z_L possesses a class IV zero of transmission of order 1 at the origin. To apply the theory, consider the following steps.

Let $s(s)$ be the minimum-phase reflection function determined from factorization of the equation

$$s(s)s(-s) = 1 - G(-s^2). \tag{8}$$

A real regular all-pass function $b(s)$ is defined by the poles of $Z_L(-s)$ in $\mathrm{Re}\{s\} > 0$. For the series RC load, this is given by

$$b(s) = 1 \tag{9}$$

since the only such pole is at the origin.

Let $f(s)$ be

$$f(s) = 2b(s)\,\mathrm{Ev}\{Z_L(s)\} = 2R_L/k^2 \tag{10}$$

where $\mathrm{Ev}\{Z_L(s)\}$ is the even part of the load impedance.

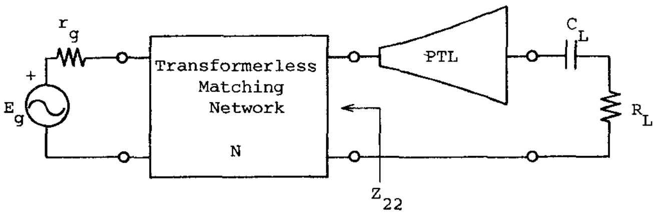

Fig. 2 The scheme of the broad-band matching problem with PTL.

For a class IV zero of transmission of order 1 at the origin, the basic constraint imposed by the load becomes

$$1-(1-K)^{1/2n} \leqslant 2\omega_c \frac{C_L R_L}{k} \cdot \sin\left(\frac{\pi}{2n}\right). \tag{11a}$$

And this equation may be solved for the gain K as

$$K \leqslant 1 - \left\{1 - 2\omega_c \frac{C_L R_L}{k} \cdot \sin\left(\frac{\pi}{2n}\right)\right\}^{2n}. \tag{11b}$$

Fulfillment of this gain–bandwidth restriction ensures the positive-reality of the back-end impedance $Z_{22}(s)$, which is given by

$$Z_{22}(s) = \frac{f(s)}{b(s)-s(s)} - Z_L(s). \tag{12}$$

Equation (11a) or (11b) shows that the use of a PTL (constrained to $k > 1$) causes the decrease of maximum gain K. In compensation for this decreased gain, the ideal transformer is removed from the matching network. In the limit as $n \to \infty$,

$$K \leqslant \lim_{n\to\infty}\left[1 - \left\{1 - 2\omega_c \frac{C_L R_L}{k} \cdot \sin\left(\frac{\pi}{2n}\right)\right\}^{2n}\right] = 1 - \exp(-2\pi\omega_c C_L R_L / k). \tag{13}$$

Equation (13) shows that the reflection attenuation is proportional to $C_L R_L / k$.

Under the assumption that the two-port network N in Fig. 2 contains only the the high-pass transformerless ladder structure, the maximum gain G obtained at $\omega \to \infty$ is given by

$$G(\infty) = \frac{4 r_g R_L k^{-2}}{\left(R_L k^{-2} + r_g\right)^2}. \tag{14}$$

From the high-pass transducer power gain characteristic of (7), the maximum attainable gain K in (11b) is achieved at $\omega \to \infty$. When the gain K equals the infinite-frequency transducer power gain $G(\infty)$, the two-port network N does

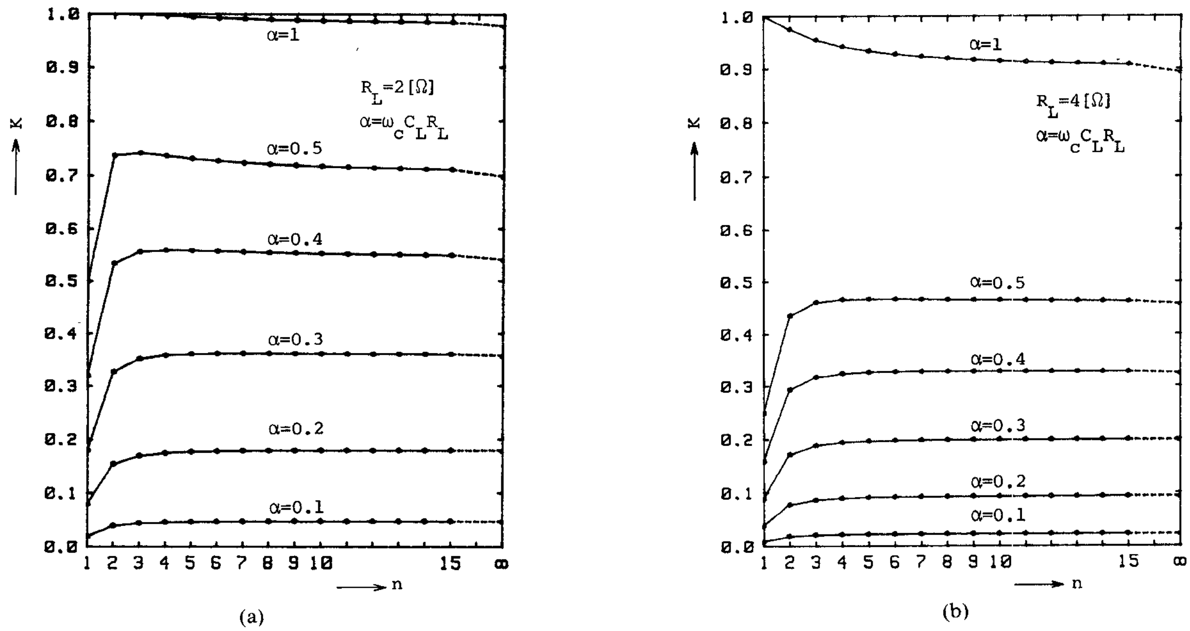

Fig. 3. The maximum attainable gain K as a function of order n.

not contain an ideal transformer. Namely,

$$1-\left\{1-2\omega_c\frac{C_LR_L}{k}\cdot\sin\left(\frac{\pi}{2n}\right)\right\}^{2n}=\frac{4r_gR_Lk^{-2}}{\left(R_Lk^{-2}+r_g\right)^2}. \quad (15)$$

This nonlinear equation may be solved for a parameter k by numerical methods such as bisection.

Before solving (15) it is necessary to consider the case achieving unity gain. If the inequality

$$2\omega_cC_LR_Lk^{-1}\cdot\sin(\pi/2n)>1$$

is established under the condition $k=\sqrt{R_L/r_g}$ we can set $K=1$ as the maximum gain. A PTL having $k=\sqrt{R_L/r_g}$ transforms the resistive part of the load R_L into r_g, and there is no need to use an ideal transformer in the matching network structure. In this case the PTL method may be able to stand comparison with an ordinary broad-band matching. It is interesting to note that using an equivalent representation of the PTL, the resulting matching network structure will appear to be a typical $R-R$ high-pass filter with Butterworth response.

Except for the case of $K=1$ mentioned above, it is necessary to solve (15) numerically for the parameter k such that

$$1\leqslant\sqrt{R_L/r_g}<k. \quad (16)$$

The available gain is then given by (11b) using the equals sign. It is obvious that the maximum obtainable gain K will be decreased by the taper ratio k of the PTL. If we accept the decreased gain and select a maximum gain with optimum k in (11b), the quasi-broad-band matching will be achieved without an ideal transformer.

The maximum attainable infinite-frequency gain K, which may be called the quasi-optimum gain, can be calculated by finding the optimum k as a function of n, $\alpha=\omega_cC_LR_L$, and R_L, the resistive part of the load Z'_L. The relation between order n and the quasi-optimum gain K is presented in Fig. 3(a) for $R_L=2\Omega$ and in (b) for $R_L=4\ \Omega$. The plots of gain K as a function of $\alpha=\omega_cC_LR_L$ are shown in Fig. 4(a) and (b). Plots of gain K as a function of R_L, for $\alpha=1$, are shown in Fig. 5. In these figures, the generator resistance is normalized to unity, namely $r_g=1\ \Omega$.

The parameter α is the normalized cutoff angular frequency, and a large α means high cutoff frequency and narrow bandwidth. An inspection of these curves shows that

1) the gain K is almost identical throughout a large variation of order n, while there exists an optimum gain for a specified α that is attained at a low value of n;
2) K increases as R_L approaches unity; and
3) K increases as α is increased.

For practical applications, it is useful to note that the characteristic impedance of a PTL will be equal to R_L at the far end ($x=l$) and will be close to r_g at the front end ($x=0$) when the gain K is close to unity. The line length l of the PTL is dependent on R_L, C_L, and k and is given by

$$l=(1-k^{-1})vC_LR_L. \quad (17)$$

IV. Examples

Example 1

As an example, consider the series load impedance of a 100 Ω resistor and a 0.75 pF capacitor. It is desired to match this load to a resistive generator of internal resistance 50 Ω, and to achieve fifth-order high-pass

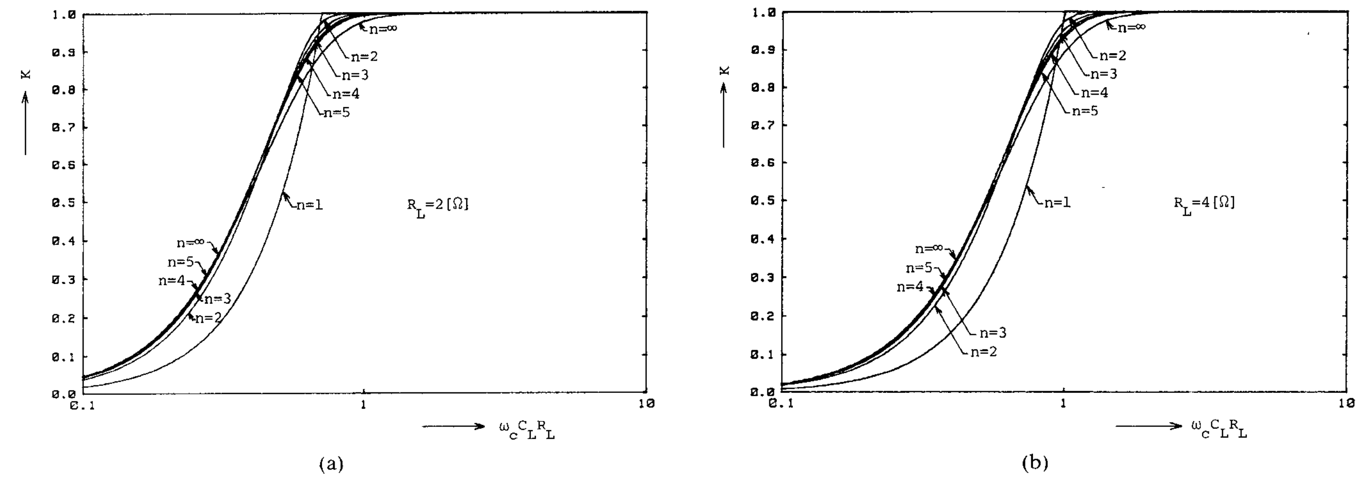

Fig. 4. The maximum attainable gain K as a function of $\alpha = \omega_c C_L R_L$.

Butterworth transducer power gain. The cutoff radian frequency is 10^{10} rad/s.

For simplicity, the load impedance is normalized to $R_0 = 50\ \Omega$ and to an angular frequency $\omega_0 = 10^{10}$ rad/s. This results in the normalized quantities $r_g = 1\ \Omega$, $R_L = 2\ \Omega$, $C_L = 0.375$ F, $\omega_c = 1$ rad/s, and $\alpha = 0.75$.

Applying an ordinary broad-band matching network design procedure [3], we can achieve almost unity maximum attainable gain

$$K = 1 - \{1 - 2\omega_c C_L R_L \cdot \sin(-\pi/2n)\}^{2n}$$
$$= 0.9980253$$

but achieving the above gain between the resistor $r_g = 1\ \Omega$ and $R_L = 2\ \Omega$ at infinite frequency requires an ideal transformer with a transformation ratio of 1:1.478518.

To design a transformerless quasi-broad-band matching network employing the proposed technique, we first must determine k, the optimum taper ratio of the PTL. For the given specification and the trial value of $k = \sqrt{R_L/r_g} = \sqrt{2}$, we compute

$$2\omega_c C_L R_L k^{-1} \cdot \sin(\pi/2n) = 0.3277620 < 1.$$

In this case, it is not possible to obtain unity gain as described in Section III, so we have to solve (15) for a suitable value of k, giving

$$k = 1.764933$$

and the corresponding maximum gain K is obtained as

$$K = 0.9524828.$$

This is smaller than unity, and is 0.2 dB attenuation in passband.

For constructing the back-end impedance $Z_{22}(s)$, the needed minimum-phase reflection function $s(s)$ may be obtained from (7) and (8) as

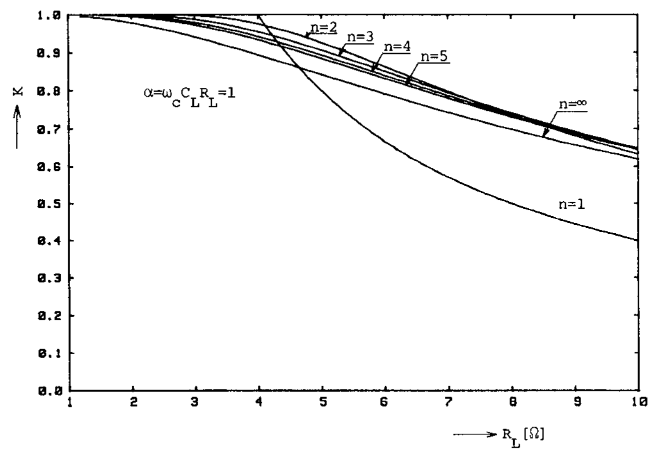

Fig. 5. The maximum attainable gain K as a function of R_L.

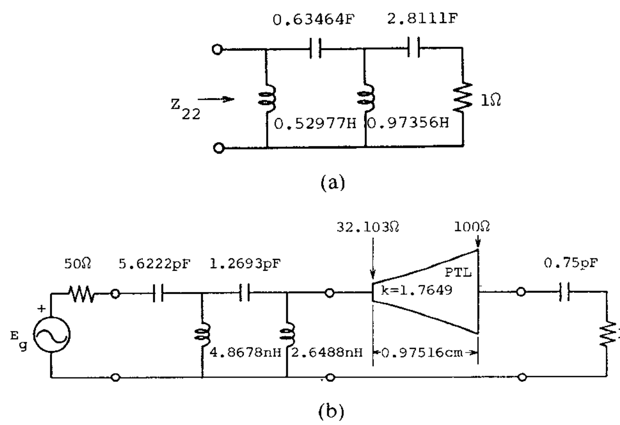

Fig. 6. (a) The normalized back-end driving point impedance Z_{22} and (b) the denormalized matching network structure without an ideal transformer.

$$s(s) = \frac{1 + 2.386177s + 2.846921s^2 + 2.099232s^3 + 0.9566604s^4 + 0.2179843s^5}{1 + 3.236068s + 5.236068s^2 + 5.236068s^3 + 3.236068s^4 + s^5}. \tag{18}$$

The back-end impedance $Z_{22}(s)$ in (12) is given by

$$Z_{22}(s) = \frac{f(s)}{b(s) - s(s)} - Z_L(s) = \frac{0.2979929s + 0.8376945s^2 + 0.999665s^3 + 0.5175769s^4}{0.5625000 + 1.581256s + 2.076114s^2 + 1.508625s^3 + 0.5175769s^4}. \tag{19}$$

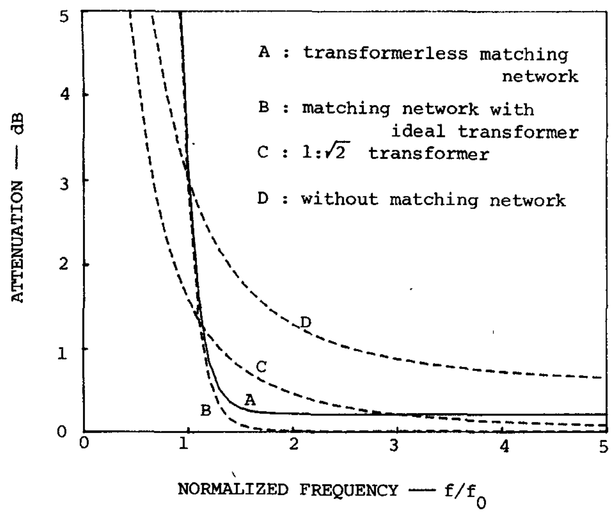

Fig. 7. The frequency response of the attenuation.

This driving-point impedance is realized by an LC ladder as shown in Fig. 6(a) with no ideal transformer. The denormalized matching network is also shown in Fig. 6(b). The free-space line length of the PTL needed is $l = 0.9751637$ cm for this case.

The frequency response of the attenuation between generator and load is shown in Fig. 7. In this figure, line A indicates the attenuation response employing the transformerless matching network designed by the method proposed in this paper, line B indicates the response employing the optimum broad-band matching network with an ideal transformer ($K = 0.9980258$), line C indicates the response employing a simple $1:\sqrt{2}$ ideal transformer, and line D indicates the response without matching network.

Example 2

For the series RC load of $R_L = 100\ \Omega$ and $C_L = 20$ pF, design a third-order Butterworth response matching network without an ideal transformer. The internal resistance of the generator is $r_g = 50\ \Omega$ and the cutoff radian frequency is $\omega_c = 10^9$ rad/s.

In this case, we can achieve unity gain because the inequality

$$2\omega_c C_L R_L k^{-1} \cdot \sin(\pi/2n) = \sqrt{2} > 1$$

is satisfied when we set

$$k = \sqrt{R_L/r_g} = \sqrt{2}.$$

The maximum attainable gain K is unity and the needed back-end impedance $Z_{22}(s)$ is

$$Z_{22}(s) = \frac{0.8284271 + 0.8284271s + 1.828427s^2 + 1.414214s^3}{2.828427s + 2.828427s^2 + 1.414214s^3 + 0.7898546s^4}. \tag{20}$$

The line length of the PTL is calculated to be $l = 17.57359$ cm, considerably shorter than 47.12389 cm, the quarter-wave length of the cutoff frequency.

The ordinary design method also allows unity gain, but the resulting matching network will contain an ideal transformer with a transformation ratio of $1:\sqrt{2}$ unless we settle for a reduced gain of

$$K = \frac{4r_g R_L}{(R_L + r_g)^2} = \frac{8}{9} = 0.8888889. \tag{21}$$

Equation (21) gives the maximum gain when the generator is connected to the load directly.

V. Conclusions

In this paper we have presented a simple technique for designing a transformerless high-pass impedance matching network having a Butterworth response of arbitrary order using a parabolic tapered transmission line. The PTL operates as an ideal transformer, as given by its equivalent circuit. We have also shown design curves calculated from the gain–bandwidth restriction.

Parallel RL loads are dual to series RC loads and may be treated in a dual manner by using the reciprocal parabolic tapered transmission line. Other transducer power gain response, such as Chebyshev, may be treated by similar techniques.

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References

[1] H. W. Bode, *Network Analysis and Feedback Amplifier Design*. New York: van Nostrand, 1945.
[2] R. M. Fano, "Theoretical limitations on broadband matching of arbitrary impedances," *J. Franklin Inst.*, vol. 249, nos. 1 and 2, pp. 57–83 and 139–154, 1950.
[3] D. C. Youla, "A new theory of broad-band matching," *IEEE Trans. Circuit Theory*, vol. CT-11, pp. 30–50, Mar. 1964.
[4] W. K. Chen, *Theory and Design of Broadband Matching Networks*. New York: Pergamon Press, 1976.
[5] W. K. Chen, "Unified theory of broadband matching," *J. Franklin Inst.*, vol. 310, no. 6, pp. 287–301, Dec. 1980.
[6] W. K. Chen, "Explicit formulas for the synthesis of optimum broad-band impedance-matching networks," *IEEE Trans. Circuits Syst.*, vol. CAS-24, pp. 157–169, Apr. 1977.
[7] W. K. Chen and K. G. Kourounis, "Explicit formulas for the synthesis of optimum broad-band impedance-matching networks II," *IEEE Trans. Circuits Syst.*, vol. CAS-25, pp. 609–620, Aug. 1978.
[8] W. K. Chen and T. Chaisrakeo, "Explicit formulas for the synthesis of optimum bandpass Butterworth and Chebyshev impedance-matching networks," *IEEE Trans. Circuits Syst.*, vol. CAS-27, pp. 928–942, Oct. 1980.
[9] Y. T. Chan and E. S. Kuh, "A general matching theory and its application to tunnel diode amplifier," *IEEE Trans. Circuit Theory*, vol. CT-13, pp. 6–18, 1966.
[10] T. M. Chien, "A theory of broadband matching of a frequency-dependent generator and load—Part I: Theory," *J. Franklin Inst.*, vol. 298, pp. 181–199, 1974.
[11] H. J. Carlin, "A new approach to gain-bandwidth problems," *IEEE Trans. Circuits Syst.*, vol. CAS-24, pp. 170–175, Apr. 1977.

[12] H. J. Carlin and P. Amstutz, "On optimum broad-band matching," *IEEE Trans. Circuits Syst.*, vol. CAS-28, pp. 401–405, May 1981.

[13] H. J. Carlin and B. S. Yarman, "The double matching problem: Analytic and real frequency solutions," *IEEE Trans. Circuits Syst.*, vol. CAS-30, pp. 15–28, Jan. 1983.

[14] H. J. Carlin and P. P. Civalleri, "On flat gain with frequency-dependent terminations," *IEEE Trans. Circuits Syst.*, vol. CAS-32, pp. 827–839, Aug. 1985.

[15] G. L. Matthaei, L. Young, and E. M. T. Jones, *Microwave Filters, Impedance Matching Networks and Coupling Structures*. New York: McGraw-Hill, 1964.

[16] R. E. Collin, *Foundation for Microwave Engineering*. New York: McGraw-Hill, 1966.

[17] H. J. Carlin and J. J. Kohler, "Direct synthesis of band-pass transmission line structures," *IEEE Trans. Microwave Theory Tech.*, vol. MTT-13, pp. 283–297, May 1965.

[18] H. J. Carlin and J. J. Komiak, "A new method of broad-band equalization applied to microwave amplifiers," *IEEE Trans. Microwave Theory Tech.*, vol. MTT-27, pp. 93–99, Feb. 1979.

[19] R. Levy and J. Helszajn, "Specific equation for one and two section quarter-wave matching networks for stub-resistor loads," *IEEE Trans. Microwave Theory Tech.*, vol. MTT-30, pp. 55–63, Jan. 1982.

[20] R. M. Cottee and W. T. Joines, "Synthesis of lumped and distributed networks for impedance matching of complex loads," *IEEE Trans. Circuits Syst.*, vol. CAS-26, pp. 316–329, May 1979.

[21] D. A. E. Blomfield, A. G. Williamson, and B. Egan, "A design method for noncommensurate broad-band matching networks," *IEEE Trans. Microwave Theory Tech.*, vol. MTT-31, no. 9, Sept. 1983.

[22] S. C. Dutta Roy, "Low frequency wide-band impedance matching by exponential transmission line," *Proc. IEEE*, vol. 67, Aug. 1979.

[23] K. Kobayashi, Y. Nemoto, and R. Sato, "Kuroda's identity for mixed lumped and distributed circuit and their application to nonuniform transmission lines," *IEEE Trans. Microwave Theory Tech.*, vol. MTT-29, pp. 81–86, Feb. 1981.

[24] I. Endo, Y. Nemoto, and R. Sato, "Impedance transformation and matching for lumped complex load with nonuniform transmission line," *IEEE Trans. Microwave Theory Tech.*, vol. MTT-33, pp. 2–8, Jan. 1985.

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