MIXED LUMPED AND DISTRIBUTED NETWORK APPLIED TO SUPERCONDUCTING THIN-FILM BROADBAND IMPEDANCE TRANSFORMING

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ABSTRACT: In this paper, a detailed analysis of properties of *mixed lumped and distributed*(MLD) lossless network is first carried out, from which the reason why the MLD network can be used as an extreme impedance transformer between source and load impedances without using extreme impedance values in the network is found. Then, the lossy transformation technique[1], which can be employed for the transformation between MLD lossy or lossless network and lumped lossless one[2,3], is discussed and compared with the method developed by Carlin[4]. Finally, as an example, one quarter-wave and two MLD lossless broadband impedance transformers are synthesized for transforming extremely low input or output impedance of a superconducting device to 50Ω microwave system and the lossy performance of one of the MLD transformers is estimated by means of the lossy transformation technique.

I. Introduction

With the rapid development of microwave integrated circuits and wide applications of many new devices, such as superconducting devices[1-4], in microwave region, a new problem of large impedance mismatch occurs, which restricts the use of these new devices in 50Ω microwave systems, for the input and output impedance levels of many of these devices fabricated with superconducting thin-film, are usually very low. If the conventional methods, such as Chebyshev approximation, are used, the element values of the synthesized network may vary wildly and are often difficult or even impossible to be practically realized.

For purpose of illustration, the element values of two four-step Chebyshev transformers having the same impedance match ratio and fractional bandwidth are shown in Table 1[5,6]. It may be found that i) with the line length reducing from $\lambda_m/4$ to $\lambda_m/16$ for shortening the transformer, the element values vary from a range of about 5.58:1 to 10.54:1; ii) in our particular case, matching a superconducting thin-film device having extremely low input and output impedances, which are typically 2Ω, will result extremely low element values in both transformers, i.e., the lowest element values in both transformers will be 2.6784 and 1.948Ω, respectively, which are unacceptable in a microstrip structure. A simple way to solve this problem is to convert all the lower impedance lines into capacitors[7] or into open-circuited stubs[8]. Nevertheless, these methods and others[9,10] are limited to certain special cases and no general method has been found which could explain why MLD network are superior to lumped or distributed network in certain cases, such as in solving the problem of large impedance mismatch in higher frequency region, and could handle the synthesis and/or design of a generalized MLD network.

II The Properties of Mixed Lumped and Distributed elements

Before detailed description of our method, a difinition is presented first.

Definition: Any lossy or lossless 2-port network, if it is passive, reciprocal and symmetric, will be denoted as a *building block*(BB).

By carefully considering the theorem 1 in [11] and the discussions on symmetric 2-port network and generalized *unit elements*(UEs) in [12,13], it can be found that a BB may be looked as a length of transmission line.

Theorem: Any BB, if its short- and open-circuited impedances, z_1 and z_2, exist, will be equivalent to a UE with its characteristic impedance written as

$$Z_{o,b}(s)=\sqrt{z_1(s)z_2(s)} \quad (1a)$$

and the propagation constant as

$$\gamma_b(s)=\tanh^{-1}(\lambda)/l_b \quad (1b)$$

where

$$\lambda=\sqrt{z_1(s)/z2(s)} \quad (2)$$

may be defined as a new frequency variable, and the real positive constant, l_b, as the equivalent line length of the BB.

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In order to achieve the general properties of a BB, the BB shown in Fig.1(a) is analyzed in detail without loss of generality. It is evident in terms of the theorem that the BB can be looked as a UE shown in Fig.1(b) and its $Z_{o,b}$ and λ as defined in (1) and (2) are given as

$$Z_{o,b}=Z_o(s)\{1+2Y(s)Z_o(s)/\mu(s)+[Y(s)Z_o(s)]^2\}^{-1/2} \tag{3a}$$

$$\lambda=\{1+[\mu^2(s)-1]/[1+Y(s)Z_o(s)\mu(s)]^2\}^{1/2} \tag{3b}$$

with

$$\mu=\tanh[\gamma(s)l] \tag{4a}$$

and the characteristic impedance of the line in the BB being

$$Z_o=Z_{o,t}\delta(s) \tag{4b}$$

as defined in [14], where the real positive multiplicative constant, $Z_{o,t}$, of the frequncy-dependent part, δ, of the Z_o is the static characteristic impedance of the line, γ and l are the propagation constant and the length of the line, respectively.

$$Y(s)=CY_C(s) \tag{5}$$

is the shunt admittance of a lossy or lossless capacitor with C being the capacitance and Y_C, the frequency-dependent part of Y as described in [15] at either end of the line.

If the BB is used as an element in a superconducting thin-film transformer, it may be assumed to be lossless approximately with Z_o reducing to $Z_{o,t}$, $\gamma(s)$ to s/C_{vp} and Y_C to s, respectively, where C_{vp} is the velocity of the propagation on the line and $s=j\omega$.

Then, some interesting properties of the BB may be found. i) Proper choice of the ϵ_C (equal to $Z_{o,t}C$) will result in $Z_{o,b}$ being a purely real or imaginary function of the real angular frequency ω. ii) If $Z_{o,b}$ is within real positive region, its value, greater or less than that of $Z_{o,t}$ over the band of interest, will depend on the value of ϵ_C chosen. iii) Similar to $Z_{o,b}$, λ is also purely real or imaginary function of ω. The region of λ being purely real corresponds exactly to that of $Z_{o,b}$ being purely imaginary. The reverse is also true.

For commonly used eight BBs displayed in Fig.2, several regions of ϵ_i (i=L, C, sh, and op) are classfied , which correspond to $Z_{o,b}$ being imaginary (or λ being real), $Z_{o,b}<Z_{o,t}$ and $Z_{o,b}>Z_{o,t}$ (or λ being real), respectively.

It is clear to see from the Table 2(a) that for a fixed ϵ_i, each BB can be refered to as a lossless UE in a certain frequency region. Thus, Richards' correspondence may be employed to treat the synthesis of a MLD commensurate line network consisting of these BBs as was done by Carlin[13]. On the other hand, the difficulty in realizing extremely low or high line impedances in a distributed transformer as mentioned above may be solved by two alternatives. One is to replace the lines having extremely low impedances by certain BBs shown in Fig.2 and by proper choice of ϵ_1, for the characteristic impedance, $Z_{o,b}$, of a BB may be made as low as the that of original line, while the line impedance, $Z_{o,t}$, in the BB is still realizable. For instance, the line having extremely low impedance, Z_2 given in Table 1, may be replaced approximately by the BB shown in Fig.2(D) with $Z_{o,b}$ being still equal to Z_2 at certre frequency f_m(f_m=10GHz), while $Z_{o,t}$ may be made to be 10Ω by letting C=7.9289pF. The other is to convert the lines having extremely high impedances into series-inductors or series- and short-circuited stubs, and extremely low impedances into shunt-capacitors or shunt- and open-circuited stubs. For instance, the line Z_2 may be converted into a shunt-capacitor at the centre frequency with the capacitance be equal to 12.83pF, approximately.

Of course, these two substitutions may result in the insertion loss of the transformer deviating from the original one a little bit at f_m and a great deal over a wider band. Though this problem may be solved by an optimization routine, two design steps will be required. Therefore, it is desired that a general method can be found for synthesizing a generalized MLD network.

It is known from the above discussion that a MLD commensurate line network N, if composed of lossless BBs, may be synthesized by applying the method introduced by Carlin[13] for design of filters with lumped-distributed elements, in which the scattering matrix elements, $\tilde{s}_{i,j}$, (i,j=1,2) of the N normalized to the frequency-dependent part of $Z_{o,b}$, which is defined by

$$\delta_b(s)=Z_{o,b}/Z_{o,bt} \tag{6}$$

are employed. In (6), $Z_{o,bt}$ is the real positive multiplicative constant and known as the static characteristic impedance of a BB in the λ domain.

However, It should be noticed that when and only when $Z_{o,b}$ is real, which corresponds to λ being imaginary, the transducer power gain(TPG) of a impedance transforming system as shown in Fig.3 will keep unchanged before and after the normalization of the scatering matrix of the network N. If $Z_{o,b}$ is complex, which occurs when the BBs are lossless but the λ is within the real region as shown in Table 2(a), or the BBs have losses, then it can be verified easily that normalization to the δ_b (simply called complex normalization) will result in the TPG calculated from the normalized scattering parameters being in the wrong. The reason is due to the source and load impedances, Z_G and Z_L, which will change the ratios between their real and imaginary parts after the complex normalization.Therefore, the method[13] is only suitable for synthesis of MLD lossless N in a certain real frequency range, and will fail if the N is complex normalized.

Nevertheless, it is easy to see that after the complex normalization, the BBs and the short- and open-circuited "stubs" in a MLD lossy or lossless N will become their corresponding "lumped" lossless UEs, lossless inductors and capacitors in the λ domain and $\tilde{S}_{i,j}$ will have a general expressions written as

$$\tilde{S}_{11}(\lambda)=\frac{h(\lambda)}{g(\lambda)}=\frac{h_1+h_2\lambda+h_3\lambda^2+\ldots+h_n\lambda^{n-1}+h_{n+1}\lambda^n}{g_1+g_2\lambda+g_3\lambda^2+\ldots+g_n\lambda^{n-1}+g_{n+1}\lambda^n} \quad (7a)$$

$$\tilde{S}_{12}(\lambda)=\tilde{S}_{21}(\lambda)=(+/-)f(\lambda)/g(\lambda)=(+/-)\lambda^k(1-\lambda^2)^{m/2}/g(\lambda) \quad (7b)$$

$$\tilde{S}_{22}(\lambda)=(-1)^{k+1}h(-\lambda)/g(\lambda) \quad (7c)$$

where h_i and g_i, $(i=1,2,\ldots,n+1)$ are the real coefficients of the numerator and denominator polynomials, $h(\lambda)$ and $g(\lambda)$, of $\tilde{S}_{11}$; $f(\lambda)$ is the numerator polynomial of $\tilde{S}_{12}$ with the order k of the first term being the number of highpass stubs and the order m of the second term being the number of UEs.

By an appropriate renormalization of the scattering matrix $\tilde{S}$ to a real positive 1Ω resistor (simply called complex renormalization), the unit normalized scattering matrix S of the network N can then be found from $\tilde{S}$, via

$$S=[\delta_b(I+\tilde{S})-(I-\tilde{S})][\delta_b(I+\tilde{S})+(I-\tilde{S})]^{-1} \quad (8)$$

where I is the identity matrix. Thus, the TPG of the impedance transforming system can be computed by S, the source and the load impedances in terms of the expression defined by [15] and an optimal performance can be approached by an optimization routine with the h_i as variables.

An interesting thing can be easily found in comparison with the lossy transformation formula (11b) in [15] that equation (8) can also be obtained by replacing $\sqrt{Z_1Z_2}$ in (11b) by δ_b. Hence, it can be said that the lossy transformation can be considered not only as a kind of mapping between the unit normalized scattering matrices of the N in the real frequency domain and a supposed corresponding lumped lossless network $\tilde{N}$ in the λ domain as described in [15] and [11], but also as the complex normalization and renormalization of the scattering matrix of the N from the view point of normalization.

III. Design Examples

As a demostration of the new synthesis approach mentioned above, several broadband impedance transformers between an extremely low input or output impedance of a superconducting device and a 50Ω microwave system over the band of 6 to 15GHz is synthesized. Their topologies are displayed in Fig.4(a), (b), and (c) with the element values being given in Table 2. It can be found from the TPGs shown in Fig.5 and obtained by [9] for transforming the same impedance matching ratio that even if the TPG of the $\lambda_m/16$ transformer is smaller than that of the quarter-wave transformer, it is better than -1.5dB achieved by [9]. The main advantage of the $\lambda_m/16$ transformer is that its lowest line impedance is about three times as large as that of the quarter-wave transformer, which will be more easy to be practically realized. Besides, the former is much shorter in length, so that the chip area will be greatly saved for integration.

Since the size of the three-step $\lambda_m/16$ transformer is quite small in comparison with that of the three-step quarter-wave transformer, a four-step $\lambda_m/16$ transformer displayed in Fig.4(c) is then synthesized. It can be seen from Fig.6 that its lossless TPG is as small as that of the three-step quarter-wave transformer, while its lowest line impedance is about twice as that of the later.

Finally, the lossy performance of the four-step $\lambda_m/16$ transformer is estimated under the assumption that the elements in the transformer are not the ideal superconducting ones with QL and QC being the conductor and dielectric losses of the transmission lines and Qc and Qd being the corresponding losses of the capacitors as defined in [11].

In conclusion, our method can handle the synthesis of MLD lossy or lossless networks not only for impedance transforming, but also for filtering, double broadband matching, and so on.

Helpful discussions with Prof. Yunyi Wang are greatfully acknowledged!

IV. References

(1) K.K.Lihkarev,IEEE Trans.Magn.,vol.MAG-15,no.1,pp.420-423,1979.
(2) T.V.Rajeevakumar,Appl.Phys.Lett.,vol. 39,pp.439-441,1981.
(3) B.J.Van Zeghbroeck,Appl.Phys.Lett.,vol. 42,pp.736-739,1983.
(4) D.P.McGinnis,et al.,J.Appl.Phys.,vol. 59,pp.3917,1986.
(5) G.L.Matthaei,et al.,New York:McGraw-Hill,1964.
(6) G.L.Matthaei,IEEE Trans.MTT,pp.372-383, Aug.1966.
(7) R.Levy, IEEE Trans.MTT,pp.223-233,Mar. 1972.
(8) Pieter W.Van Der Walt,IEEE Trans.MTT, pp.863-868,Aug.1986.
(9) A.Ghiasi,et al.,IEEE Trans.MTT,pp.673-675,May 1990.
(10) D.P.McGinnis,et al.,IEEE Trans.MTT,pp. 1521-1525,Nov.1988.
(11) Lizhong Zhu,"Computer-aided synthesis of mixed lumped and distributed lossy broadband matching networks in MMIC's," submitted for publication.
(12) H.J.Carlin,Proc.IEEE,pp.1059-1081,July 1971.
(13) H.J.Carlin,et al.,IEEE Trans.MTT,pp. 598-604,Aug.1969.
(14) Lizhong Zhu,"Computer-aided synthesis of a lossy commensurate line network and its application in MMIC's," to be published in IEEE Trans.MTT,April 1991.
(15) Lizhong Zhu,et al.,IEEE Trans.MTT,pp. 1614-1620,Dec.1988.

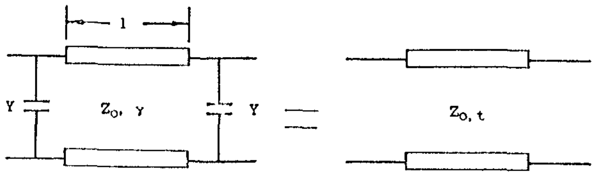

Fig 1 Circuit equivalence

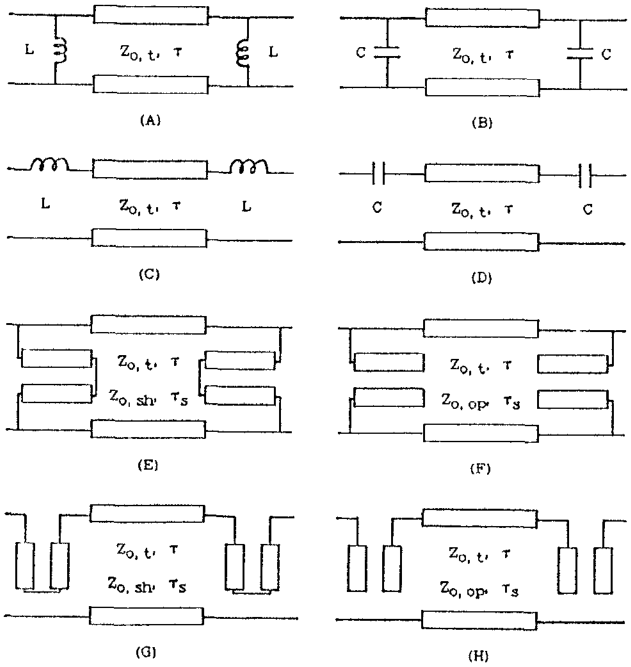

Fig. 2 Commonly used BBs

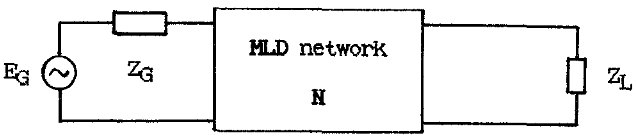

Fig. 3 Impedance transforming system

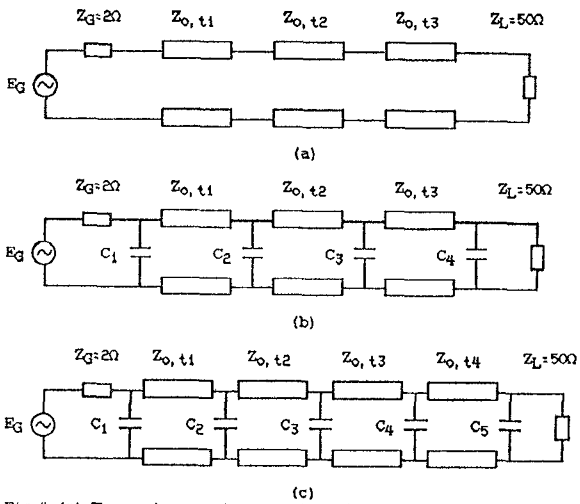

Fig. 4 (a) Three-step quarter-wave transformer; (b) Three-step $\lambda_m/16$ transformer; (c) Four-step $\lambda_m/16$ transformer

Table 1 Element values of the transformers providing an impedance match between resistances 1 and 10Ω and fractional bandwidth B_f=1

Line impedance \ Transformer Type	Quarter-wave	Short-stub
	l=$\lambda_m/4$	l=$\lambda_m/16$
Z_1	1. 3392Ω	5. 2485Ω
Z_2	2. 2840Ω	0. 9740Ω
Z_3	4. 3783Ω	10. 2669Ω
Z_4	7. 4671Ω	1. 9053Ω

Table 2 The element values of the transformers in Fig. 4

Parameter \ Transformer type	Three-step	Three-step	Four-step
	$\lambda_m/4$	$\lambda_m/16$	$\lambda_m/16$
$Z_{o,t1}$ (Ω)	3. 770	11. 044	7. 747
$Z_{o,t2}$ (Ω)	10. 028	27. 368	16. 637
$Z_{o,t3}$ (Ω)	26. 584	63. 080	39. 514
$Z_{o,t4}$ (Ω)			77. 834
C_1 (pF)		3. 6645	5. 4236
C_2 (pF)		5. 1433	7. 9491
C_3 (pF)		2. 1246	3. 5888
C_4 (pF)		0. 6456	1. 6031
C_5 (pF)			0. 5398

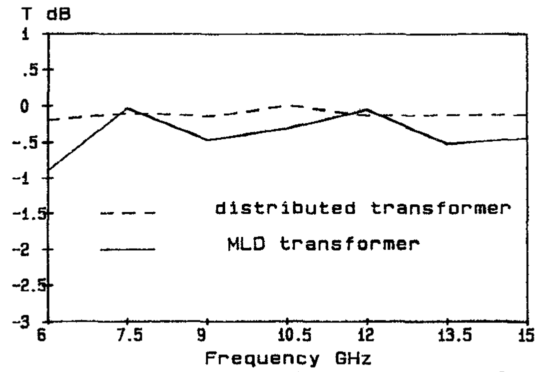

Fig. 5 The TPGs of the two three-step transformers in Fig. 4(a) and (b)

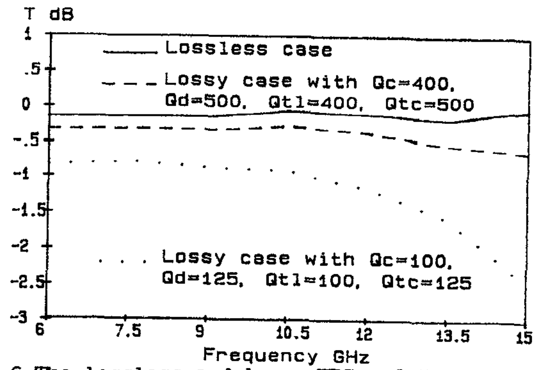

Fig. 6 The lossless and lossy TPGs of the transformer in Fig. 4(c)