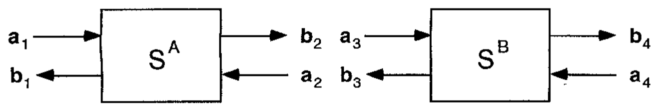

Fig. 2. Signal flow graph of cascaded two-ports.

and

$$a_3 = b_2. \tag{4}$$

In the classical waveguide circuit theory, these conditions arise directly from the continuity of the voltage and current. They are so fundamental as to be intuitive, and they form the basis of signal flow graph analysis and indeed of circuit modeling in general. However, when a and b are Youla's waves, the boundary conditions (3) and (4) do *not* apply. In other words, Youla's waves are not subject to signal flow graph analysis. A corollary is that Frickey's defined transmission matrices, formed from the scattering parameters using his Table VI, do not *function* as transmission matrices. In other words, let us denote the transmission matrix of A by T^A, that of B by T^B, and that of the circuit AB by T^{AB}. A functional transmission matrix must satisfy the condition that $T^A T^B = T^{AB}$. However, algebraic manipulation of Frickey's expressions for the transmission matrix in terms of voltage-current parameters confirms that, for his definitions

$$T^A T^B \neq T^{AB}. \tag{5}$$

Equality in (5) holds true only when the reference impedances on adjoining ports are complex conjugates, a restriction with numerous negative implications. This result of the above paper demonstrates that the counterintuitive nature of Youla's waves can easily lead to serious errors.

In the above paper, Frickey compares his results to those of a commercial simulator. From that comparison, it appears that the simulator also defines scattering parameters in terms of Youla's parameters. This suggests caution in the use of scattering parameters based on a complex reference impedance.

An alternative to Youla's theory is the general waveguide circuit theory of [4], which preserves the essential features of the classical theory while allowing for complex characteristic and reference impedances.

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Reply to Comments on "Conversions Between S, Z, Y, h, $ABCD$, and T Parameters which are Valid for Complex Source and Load Impedances"

D. A. Frickey

I would like to thank Mr. Marks and Mr. Williams for pointing out the error in using the definition of a_j and b_j in the above paper[1] as I was unaware of the implications involved. Also, I would like to thank the authors for bringing to my attention their work in [1].

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Comments on "An Equivalent Transformation for the Mixed Lumped Lossless Two-port and Distributed Transmission Line"

R. Finkler and R. Unbehauen

Stimulated by previous articles [1]–[4] by the authors of the above paper,[1] we have done related research. In doing so, we have found additional results and synthesis applications ([5], parts also in [6]) that we would like to communicate here briefly.

In [6] and (more conveniently in [5]) we gave formulas for the transformation of the D section with l'Hospital's rule already incorporated, so that no indefinite expressions such as 0/0 (cf. p. 277, text between (80) and (81)) occur. According formulas for the other sections are also given in [5], [6]. These formulas seem to be more suited for the use in the synthesis applications described below.

The equivalent transformation treated in the Theorem in Section V of the above paper, which we in accordance to the idiomatic usage in [1], [2] and due to [7] called extended Levy transformation, can also be performed numerically. This can be done by solving a system of ordinary differential equations, where the line length l is the independent variable and the coefficients of the numerators of the lumped lossless two-port chain matrix elements are the functions to be determined. Reference [5] contains some additional theorems on the asymptotic behavior of this transformation for $l \to \infty$.

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[1] I. Endo, Y. Nemoto, and R. Sato, *IEEE Trans. Microwave Theory Tech.,* vol. 42, pp. 272–282, Feb. 1994.

There exists a related extension of the Richards transformation [8], leading to the following statement: An impedance $Z(j\omega)$ $(Z(\infty) \neq 0, \infty)$, which can be realized as a lumped passive one-port, can also be realized as a distributed transmission line of arbitrary length l terminated in an impedance $Z'(j\omega)$ $(Z'(\infty) \neq 0, \infty)$ that is also realizable as a lumped passive one-port. If $Z(0) \neq 0, \infty$, $Z'(j\omega)$ tends to a constant ohmic resistance of value $Z(0)$ as $l \rightarrow \infty$ (otherwise $Z'(j\omega)/l^k \rightarrow$ const with some integer k). Under certain conditions, the extended Richards transformation can be generalized for lossy lines.

Both of these extended transformations can be used for the synthesis of 1) distributed transmission lines and 2) cascades, in which such lines and lumped lossless two-ports follow one another alternately. In both cases, the synthesis starts from a lumped reference network. Case 1 is based upon the fact that in Fig. 4 of the paper[1] with $W_0(x) = R_2 \equiv$ const the properties of the transformed distributed transmission line with characteristic impedance $W(x)$ tend to those of the lumped lossless two-port N_R (reference network) in some sense as $l \rightarrow \infty$.

For case 2, consider the lumped reference network terminated in the ohmic resistance R_2. Now insert a cascade of uniform transmission lines all with the same characteristic impedance R_2 between the reference network and its terminating resistance. This does not affect the input impedance and changes the transfer properties (from the input port of the reference network to the terminals of the resistance) only by a constant time delay. Then represent the lumped reference network by a cascade of Darlington sections (and maybe an ideal transformer).

The final cascade can then be constructed by a multiple application of the extended Levy transformation. It can be shown that if the zeros of transmission all occur at real frequencies, and if the lengths of the lines between the lumped lossless two-ports in the final cascade are chosen appropriately, transformers, which in general appear in the lumped reference network, can be avoided.

In both cases, instead of using the extended Levy transformation, we can also apply the extended Richards transformation (in case 2) together with the extraction cycles [9] of lumped network synthesis) to the input impedance of the reference network terminated in R_2.

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Corrections to "TE and TM Modes in Circularly Shielded Slot Waveguides"

J. L. Tsalamengas, I. O. Vardiambasis, and J. G. Fikioris

In the above paper[1], the following misprints should be corrected:

1) Just below (15) and just before (28), $x_2 = -h + wt$ should read: $x_2 = -h + wt''$.
2) In (20), $\begin{matrix} H_0^{(2)} \\ \cdot(k_c|x - x'|) \end{matrix}$ should read: $H_0^{(2)}(k_c|x - x'|)$.
3) In (27), $\begin{matrix} B_m \\ \cdot(\alpha) \end{matrix}$ should read: $B_m(\alpha)$.
4) In (28), $H_0^{(2)}[k_c|2h + w(t - t)|]$ should read: $H_0^{(2)}[k_c|2h + w(t - t'')|]$.

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[1] J. L. Tsalamengas, I. O. Vardiambasis, and J. G. Fikioris, *IEEE Trans. Microwave Theory Tech.*, vol. 41, no. 6/7, pp. 966–972, June/July 1993.

Corrections to "Experimental Proof-of-Principle Results on a Mode-Selective Input Coupler"

Jeffrey P. Tate

Upon careful review of the above paper,[1] two errors were found. The cutoff frequency for the TE_{01} coaxial mode was incorrectly shown as 13.81 GHz in Fig. 4. The results in Fig. 9(a) and 9(b), which compare theory and experiment, are also incorrect. The new figures that should replace them are shown below as Fig. 1(a) and 1(b). The figure captions used for Fig. 9(a) and 9(b) are unchanged. These new graphs correctly illustrate the effect discussed in the text

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