Modified Transmission and Reflection Coefficients of Nonuniform Transmission Lines and Their Applications

Te-Wen Pan and Ching-Wen Hsue, *Senior Member, IEEE*

***Abstract*—By employing the $ABCD$ transmission matrix of a transmission line, we formulate the reflection and transmission coefficients of a nonuniform line as polynomial ratios in Z-transforms. Such formulations reveal explicit relationship between transmission and reflection coefficients of a nonuniform line. These formulations, in conjunction with a reconstruction method, lead to the realization of nonuniform lines from either the reflection or transmission coefficients. Several examples are presented to illustrate the applications of modified transmission and reflection coefficients in practical circuits.**

***Index Terms*—Filter, inverse scattering, transmission line.**

I. Introduction

NONUNIFORM transmission lines (NTL's) have been studied by many authors for decades. Most of those investigations laid great stress on wave interaction with an NTL. They analyzed scattering characteristics of a known structure of NTL's in both frequency and time domains [1]-[8]. Few papers [9]-[12] were concerned with inverse scattering problems in which the structures of transmission lines are obtained from given scattering parameters. As far as the direct scattering is concerned, both computation efficiency and computation accuracy become major focuses. However, from the point-of-view of inverse scattering, the format of scattering parameters plays an important role in facilitating the inverse problem. Therefore, we may formulate the scattering parameters of nonuniform line in various forms to fit specific considerations.

In many practical applications, we need to build an NTL that meets certain scattering parameters in either frequency or time domain. We typically use a lumped *LC* circuit to approximate the prescribed characteristics. After obtaining the values of *LC* elements, we then use the Richard's transformation [13] or others to convert the equivalent *LC* values into an appropriate structure of nonuniform lines. The motivation of this paper is to study the scattering characteristics of an NTL and deduce a scheme that constructs the physical structure of nonuniform lines from reflection or transmission coefficients. The method shown here is different from the conventional scheme, which includes the concept of the *LC* circuit.

Manuscript received December 12, 1997; revised July 7, 1998. This work was supported by the National Science Council, R.O.C., under Grant NSC87-2218-E011-012.

The authors are with the Department of Electronic Engineering, National Taiwan University of Science and Technology, Taipei, Taiwan, R.O.C.

Publisher Item Identifier S 0018-9480(98)09052-8.

In this paper, we employ the $ABCD$ transmission matrix of transmission lines to develop reflection and transmission coefficients of a nonuniform line. By treating an NTL as a cascaded commensurate finite number of sections of signal lines, we express the scattering parameters of a nonuniform line as polynomial ratios in Z-transforms. We show that explicit relationship exists between the reflection and transmission coefficients. In particular, a given reflection coefficient of an NTL will lead to an explicit unique transmission coefficient. However, a given transmission coefficient may yield multiple corresponding reflection coefficients. These formulations, in conjunction with a reconstruction scheme [12], help the realization of an NTL from either a reflection or transmission coefficient. We present several examples to illustrate the applications of such formulated coefficients in practical circuits.

II. Formulations of Reflection and Transmission Coefficients

The $ABCD$ transmission matrix of a uniform transmission line in frequency domain can be written as [13]

$$\begin{bmatrix} A & B \\ C & D \end{bmatrix} = \begin{bmatrix} \cos\beta l & jZ\sin\beta l \\ jY\sin\beta l & \cos\beta l \end{bmatrix} \tag{1}$$

where β is the phase constant, Z is the characteristic impedance, Y is the characteristic admittance, and l is the physical length of transmission line.

The $ABCD$ matrix in frequency domain can be converted into the $ABCD$ matrix in Z domain by using $z = e^{j\omega}$, where ω is the angular frequency. If the propagation delay time of uniform line is τ, i.e., $\beta l = \omega\tau$, the $ABCD$ matrix of the signal line can be cast into the form of

$$\begin{bmatrix} A & B \\ C & D \end{bmatrix} = \begin{bmatrix} \frac{1}{2}\left(z^{\tau}+z^{-\tau}\right) & \frac{Z}{2}\left(z^{\tau}-z^{-\tau}\right) \\ \frac{Y}{2}\left(z^{\tau}-z^{-\tau}\right) & \frac{1}{2}\left(z^{\tau}+z^{-\tau}\right) \end{bmatrix}. \tag{2}$$

Since an NTL can be treated as a cascaded finite number of sections of signal lines, the $ABCD$ matrix of an NTL is then represented by a sequential multiplication of the corresponding $ABCD$ matrices of uniform transmission lines. We then have

$$\begin{bmatrix} A & B \\ C & D \end{bmatrix} = \prod_{i=1}^{N} \begin{bmatrix} A_i & B_i \\ C_i & D_i \end{bmatrix} \tag{3}$$

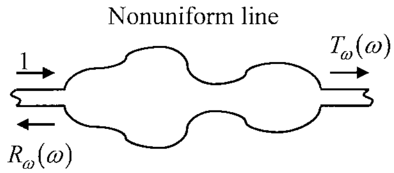


Fig. 1. A nonuniform line and its frequency-domain reflection and transmission coefficient.

where N is the number of finite lines, and A_i, B_i, C_i, and D_i are the matrix elements that represent the ith $(i \leq N)$ line.

Assuming that the propagation delay of each divided subsection is τ, from (2) and (3), we obtain the $ABCD$ matrix of an NTL in Z-transforms

$$A = \sum_{\substack{k=-N \\ \text{increment by 2}}}^{N} \alpha_k z^{k\tau} \tag{4a}$$

$$B = \sum_{\substack{k=-N \\ \text{increment by 2}}}^{N} \beta_k z^{k\tau} \tag{4b}$$

$$C = \sum_{\substack{k=-N \\ \text{increment by 2}}}^{N} \gamma_k z^{k\tau} \tag{4c}$$

$$D = \sum_{\substack{k=-N \\ \text{increment by 2}}}^{N} , \eta_k z^{k\tau} \tag{4d}$$

where α_k, β_k, γ_k, and η_k are real coefficients. The determinant of the $ABCD$ matrix is one because the determinant of each $ABCD$ matrix representing each subline is one.

If we set $z = e^{j\omega}$ in (4), we obtain the $ABCD$ transmission matrix of nonuniform line in frequency domain. The frequency-domain scattering parameters are related to $ABCD$ matrix elements in frequency domain with the following relationships [13]:

$$R_\omega(\omega) = \frac{A + B/Z_0 - CZ_0 - D}{A + B/Z_0 + CZ_0 + D} \tag{5a}$$

$$T_\omega(\omega) = \frac{2}{A + B/Z_0 + CZ_0 + D} \tag{5b}$$

where Z_0 is the reference characteristic impedance, $R_\omega(\omega)$ is the reflection coefficient, and $T_\omega(\omega)$ is the transmission coefficient, as shown in Fig. 1.

By substituting A, B, C, and D in Z domain into (5a) and (5b), we obtain

$$R_z(z) = \frac{\displaystyle\sum_{\substack{k=-N \\ \text{increment by 2}}}^{N} q_k z^{k\tau}}{\displaystyle\sum_{\substack{k=-N \\ \text{increment by 2}}}^{N} p_k z^{k\tau}} \tag{6a}$$

and

$$T_z(z) = \frac{2}{\displaystyle\sum_{\substack{k=-N \\ \text{increment by 2}}}^{N} p_k z^{k\tau}} \tag{6b}$$

where p_k and q_k are real coefficients. Note that $R_\omega(\omega)$ and $T_\omega(\omega)$ are related to $R_z(z)$ and $T_z(z)$, respectively, as

$$R_\omega(\omega) = R_z(z)|_{z=e^{j\omega}} \tag{7a}$$

and

$$T_\omega(\omega) = T_z(z)|_{z=e^{j\omega}}. \tag{7b}$$

If we divide the denominators and numerators of (6a) and (6b) with a common factor $p_N z^{N\tau}$, we obtain the general forms of reflection and transmission coefficients in Z domain

$$R_z(z) = \frac{\displaystyle\sum_{n=0}^{N} b_n z^{-2n\tau}}{1 + \displaystyle\sum_{n=1}^{N} a_n z^{-2n\tau}} \tag{8a}$$

$$T_z(z) = \frac{b z^{-N\tau}}{1 + \displaystyle\sum_{n=1}^{N} a_n z^{-2n\tau}} \tag{8b}$$

where b_n, a_n, and b are real coefficients.

Equation (8a) and (8b) represents modified reflection and transmission coefficients of nonuniform signal lines in Z-transforms. These equations have been widely used in digital signal processing (DSP) studies [14]. In fact, (8a) and (8b) are the general forms of discrete-time system functions. For a given system specification in either frequency or time domain, we can easily find the coefficients a_n, b_n, and b, shown in (8), by using DSP techniques. This reveals that many application tools developed in DSP are applicable to the study of NTL's. For example, in order to get an NTL filter that has the desired transmission coefficient $T_z(z)$, we may use digital filter-design techniques[1] [14] or system-idenitification techniques [15] to obtain the coefficients a_n and b in $T_z(z)$. However, these are all DSP techniques and have been well developed.

In many practical applications, we know only the transmission coefficient $T_z(z)$, but not the reflection coefficient $R_z(z)$. It is a rather simple process to obtain the impedance profile $Z(x)$ of an NTL from the reflection coefficient by using a reconstruction method [12], where x is a space variable. However, it is quite difficult to obtain the structure of a nonuniform line from a given transmission coefficient. This, in turn, indicates that it is useful to study the method of obtaining the reflection coefficient from the transmission coefficient.

[1] T. P. Krauss, L. Shure, and J. N. Little, *Signal Processing Toolbox User's Guide*, The Math Works Inc., Natick, MA, 1994.

III. Relationship Between $R_z(z)$ and $T_z(z)$

According to the conservation of energy, we have

$$|T_z(z)|^2 + |R_z(z)|^2|_{z=e^{j\omega}} = 1. \tag{9}$$

Equation (9) indicates that a transmission coefficient $T_z(z)$ may not yield a corresponding unique reflection coefficient and vice versa. For example, in a single-section line, two configurations $50\ \Omega - 100\ \Omega - 50\ \Omega$ and $50\ \Omega - 25\ \Omega - 50\ \Omega$ have the same transmission coefficients $T_z(z)$, but these two configurations produce different reflection coefficients $R_z(z)$.

A. To Obtain $T_z(z)$ from $R_z(z)$

For a dc condition, we have $\omega = 0$, $z = 1$, and $T_z(z) = 1$. As shown in (8b), this leads to the following condition:

$$b = \left(1 + \sum_{n=1}^{N} a_n\right). \tag{10}$$

Substituting b into (8b), we obtain $T_z(z)$. Therefore, it is a straightforward procedure to obtain $T_z(z)$ from a given $R_z(z)$. In particular, we show that a given $R_z(z)$ will lead to a unique $T_z(z)$.

B. To Obtain $R_z(z)$ from $T_z(z)$

If we set $X = z^{2\tau}$, we then can cast (8b) in the form of

$$\begin{aligned} T_z(z = X^{1/2\tau}) &\equiv T(X) \\ &= \frac{bX^{-(N/2)}}{1 + \sum_{n=1}^{N} a_n X^{-n}} \\ &= \frac{bX^{-(N/2)}}{\prod_{k=1}^{N}(1 - d_k X^{-1})}. \end{aligned} \tag{11}$$

Note that both a_n and b are real values. However, d_k could be a complex value and d_k is the pole of $T(X)$. By taking the square of absolute value $|T(x)|$, we obtain

$$|T(X)|^2 = T(X)T^*\left(\frac{1}{X^*}\right) = \frac{b^2}{\prod_{k=1}^{N}(1 - d_k X^{-1})(1 - d_k^* X)}. \tag{12}$$

Equation (12) indicates that, for each pole d_k in $T(X)$, there exists two poles d_k and $(d_k^*)^{-1}$ in $|T(X)|^2$. Substituting (12) into (9), we get

$$\begin{aligned} |R(X)|^2 &= 1 - |T(X)|^2 \\ &= \frac{\left[\prod_{k=1}^{N}(1 - d_k X^{-1})(1 - d_k^* X)\right] - b^2}{\prod_{k=1}^{N}(1 - d_k X^{-1})(1 - d_k^* X)}. \end{aligned} \tag{13}$$

If c is a solution to $|R(X)|^2 = 0$, we then have

$$\left[\prod_{k=1}^{N}(1 - d_k c^{-1})(1 - d_k^* c)\right] - b^2 = 0. \tag{14a}$$

By taking the complex conjugate on both sides of (14a), we obtain

$$\left[\prod_{k=1}^{N}(1 - d_k^*(c^*)^{-1})(1 - d_k c^*)\right] - b^2 = 0. \tag{14b}$$

A close examination on (13), (14a), and (14b) reveals that, if there is a complex number c that makes $|R(X)|^2 = 0$, then there exists another zero $(c^*)^{-1}$ that makes $|R(X)|^2 = 0$. Therefore, (13) can be cast in the form of

$$|R(X)|^2 = \frac{K^2 \prod_{k=1}^{N}(1 - c_k X^{-1})(1 - c_k^* X)}{\prod_{k=1}^{N}(1 - d_k X^{-1})(1 - d_k^* X)} \tag{15}$$

where c_k, d_k are complex numbers, and K is a real number. To obtain c_k from d_k, we first expand the numerator of (13) into a polynomial form with X as a variable. Obviously, c_k are roots of that polynomial. We may use the built-in function in MATLAB software tools to facilitate the procedure of finding c_k from d_k.[1]

Equation (8a) and (8b) reveals that both $R(X)$ and $T(X)$ have the same poles. While no zero is found in $T(X)$, there exists N zeros in $R(X)$. In order to obtain $R(X)$ from $T(X)$, we need to compute all the zeros of $R(X)$. Comparing (8a) with (15), we find that, for each zero c_k in $R(X)$, there exist two zeros c_k and $(c_k^*)^{-1}$ in $|R(X)|^2$. A similar situation holds for the poles. If we want to obtain $R(X)$ from $|R(X)|^2$, we choose one zero from each pair of zeros and one pole from each pair of poles. Due to causality, the poles selected must be within a unit circle. Therefore, to obtain $R(X)$ from $|R(X)|^2$, we have only one choice for each pair of poles. However, we have two choices for the selection of each pair of zeros in $|R(X)|^2$. This shows that we may have 2^N choices for the selection of all proper zeros. Note that the zero at $X = 1$ in $|R(X)|^2$ is a second-order zero.

It is pertinent to point out that, although we have many choices to determine $R(X)$, there is no guarantee that every choice would eventually lead to a physical configuration that is practically realizable. For example, a nonuniform line consisting of extremely large or small values of characteristic impedances cannot be practically implemented.

IV. Applications

In this section, we present some examples to illustrate the applications of transmission and reflection formulations in (8) to practical circuits. In particular, we address the procedure that converts a given frequency-domain transfer (or reflection) coefficient into the form shown in (8).

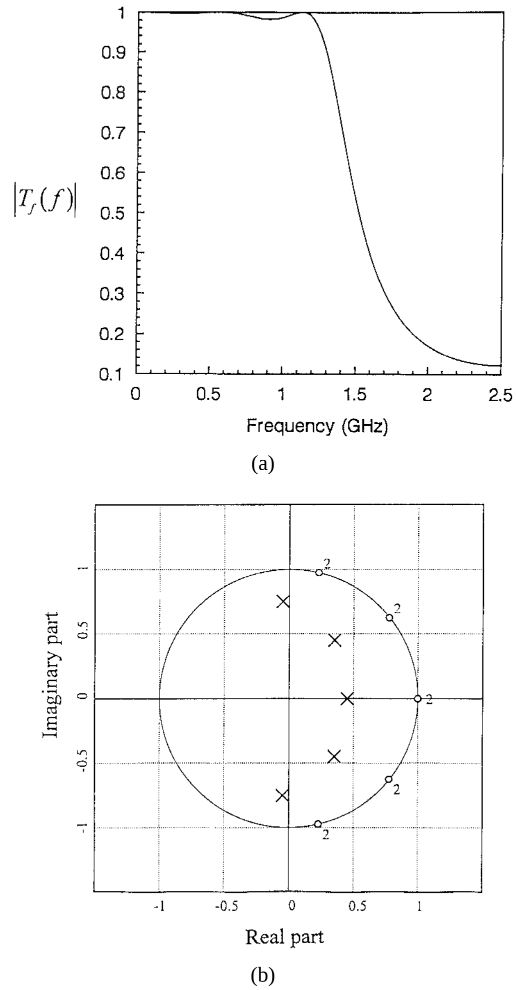


Fig. 2. (a) The transmission function of a postulated low filter having −3-dB point at 1.4 GHz. (b) The poles (x) for both $T(X)$ and $R(X)$, and the zeros (o) for $|R(X)|^2$ in Z plane.

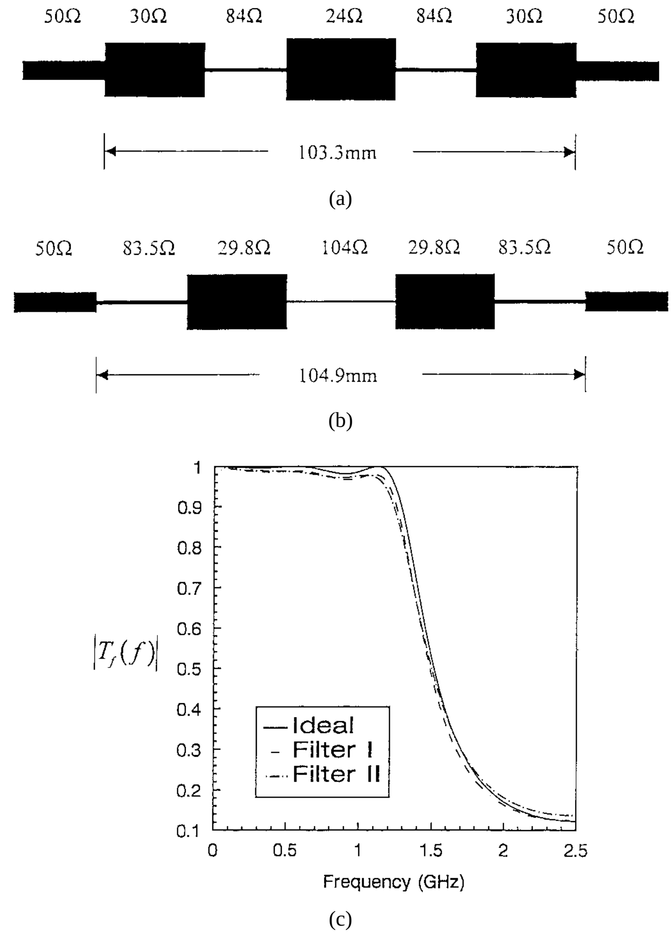
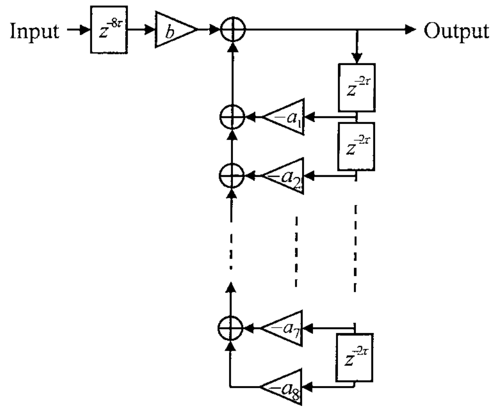


Fig. 3. (a) The physical layouts of two NTL's low-pass filters. (b) The measured responses of two low-pass filters shown in (a).

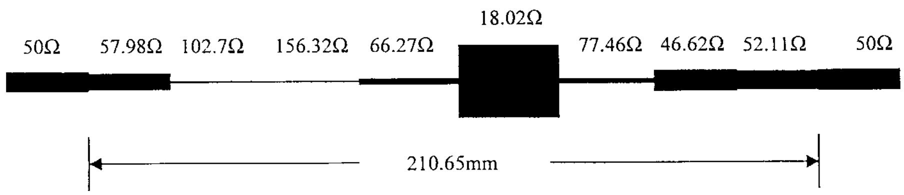
A. Filter Design

We assume that we need to construct a nonuniform line filter from a given transmission function in frequency domain. As shown in (8b), the transmission function is well represented by an all-pole function. This indicates that we essentially should design an all-pole filter that satisfies the transmission function. Here, we employ digital filter-design techniques which have been developed for many years in DSP studies. For a given transmission function that shows the magnitude responses as a function of normalized frequency, we use a finite impulse response (FIR) technique to obtain an all-zero filter [14].[1] We then convert the FIR filter into an Nth-order all-pole filter.[1] By employing these procedures, we obtain N poles of the transmission coefficient. All N poles are located within a unit circle in Z plane. Upon the substitution of poles into (11), we obtain the transmission function $T(X)$. We then obtain the corresponding reflection coefficient $R(X)$ by using the method described in Section III. The conversion from transmission function to both $T_z(z)$ and $R_z(z)$ is facilitated by the aid of a MATLAB software tool.[1] The physical structure that yield the prescribed reflection coefficient $R_z(z)$ or $R_\omega(\omega)$ can be obtained by using the reconstruction technique in [12].

Fig. 2(a) shows a transmission function $T_f(f)$ of a low-pass filter having −3-dB point at 1.4 GHz. The frequency of interest extends from dc to 2.5 GHz. We normalize the frequency so that the uppermost frequency 2.5 GHz becomes 1 Hz and the −3-dB point is at $1.4/2.5 = 0.56$ Hz. We assume to use a five-section line, i.e., $N = 5$, to implement such a low-pass filter. Fig. 2(b) shows the locations of five poles that represent the transmission function shown in Fig. 2(a). In addition, Fig. 2(b) also shows the locations of zeros that occur in the corresponding reflection coefficient $R(X)$. Note that all five zeros are located on the contour of a unit circle. The symbol "2" at each zero location represents a second-order zero. The reflection coefficient is then converted into the impedance profile of an NTL by using the reconstruction method [12]. We obtain two different nonuniform lines having impedance profiles $30\ \Omega - 84\ \Omega - 24\ \Omega - 84\ \Omega - 30\ \Omega$ and $83.5\ \Omega - 29.8\ \Omega - 104\ \Omega - 29.8\ \Omega - 83.5\ \Omega$, respectively. The physical length of each section is determined by the highest frequency of interest. If the propagation delay of each section is τ, the interval between adjacent impulses in both $R_t(t)$ and $T_t(t)$ is 2τ, where $R_t(t)$ is the reflection coefficient and $T_t(t)$ is the transmission coefficient in time domain. According to discrete Fourier transform theory, the repetitive period of both $R_\omega(\omega)$ and $T_\omega(\omega)$ in frequency domain is $1/2\tau$ and the highest operating frequency of the filter is $1/4\tau$. Therefore, if



(a)



(b)

Fig. 4. (a) An AR processor. (b) An eight-section nonuniform line that resembles the performance of an AR shown in (a).

the highest frequency of interest is f_h, the propagation delay of each section is $1/4f_h$. This, in turn, indicates that the physical length of each section is $\lambda_h/4$, where λ_h is the wavelength of highest frequency signal in the respective section. Note that, because of the variation of effective dielectric constants, the physical length of each section may have a different value.

The low-pass filters are built on Duroid substrate having thickness of 31 mil and relative dielectric constant of 2.5, which are shown in Fig. 3(a) and (b). The width and length of each uniform line are calculated by using microstrip formulations [13]. We use an HP8510C network analyzer to measure the transmission coefficients of these two nonuniform microstrip lines. Fig. 3(b) shows the measurement results of two low-pass filters. For convenience, we also show the original filter specification, which is shown in solid line. The slight discrepancy between the measurement results and original specification is due to the loss factor and big impedance discontinuity occurring at the junction of two uniform lines. The effects of both loss factor and big impedance discontinuity on an NTL are not taken into account for the present consideration.

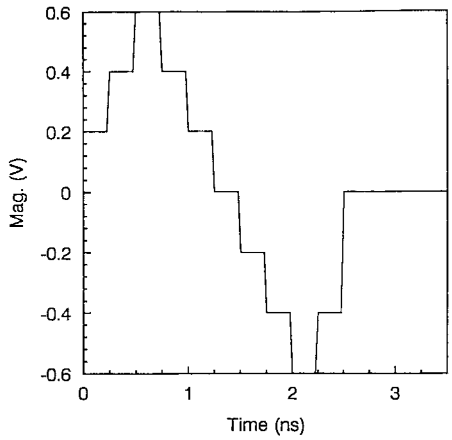
Although the physical structures of low-pass filters are similar to those of conventional low-pass filters, it is pertinent to point out that the method presented here is quite different from conventional equivalent *LC* circuit approach [13]. In particular, two structures having the same measurement results basically agree with our prediction. Furthermore, each uniform line in either filter configuration has the same propagation delay. This property is not found in conventional *LC* approach filters.

B. High-Speed Infinite Impulse Response (IIR) Circuit

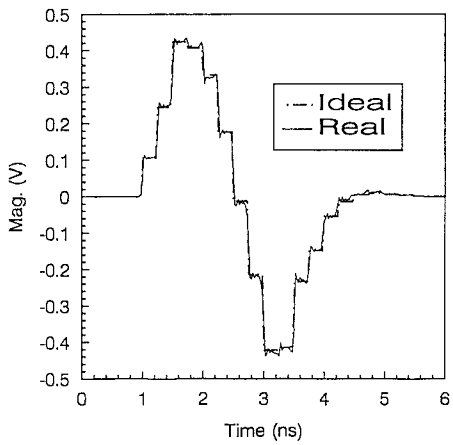
From the viewpoint of DSP, (8a) and (8b) indicate that an NTL can also be treated as an IIR circuit. While the reflection coefficient can be regarded as an autoregressive moving average (ARMA) process, the transmission coefficient can be treated as an autoregressive (AR) process [15]. A DSP generally can be implemented by using a microprocessor or special integrated circuit (IC). However, when we implement a DSP with a microprocessor, we cannot obtain a speed that exceeds several hundred megahertz. However, if we implement an IIR circuit by using an NTL, its speed is able to exceed gigahertz fairly easily.

Fig. 4(a) shows an AR process whose performance resembles the transmission coefficient in (8b). Although we may arbitrarily select the values of $a_i (i = 1, 2, \cdots, 8)$, it is pertinent to point out that the choices of a_i must satisfy $|T_z(z = e^{j\omega})| < 1$. For the present consideration, we set $a_1 = -0.3$, $a_2 = -0.3$, $a_3 = 0.056$, $a_4 = 0.0897$, $a_5 = -0.0054$, $a_6 = -0.0058$, $a_7 = -0.0012$, and $a_8 = -0.0015$. The value of b is obtained via (10). We show here that an NTL can resemble the performance of an AR processor shown in Fig. 4(a).

By using the same procedures addressed in previous sections, we obtain an eight-section NTL having impedance profile $57.98\ \Omega - 102.70\ \Omega - 156.32\ \Omega - 66.27\ \Omega - 18.02\ \Omega - 77.46\ \Omega - 46.62\ \Omega - 52.11\ \Omega$. Note that this impedance profile is just one selection among many choices. The operation speed of the AR process is determined by the delay time of each section of transmission line. We assume that the operation speed is 4 GHz, i.e., the output signal could change state every 0.25 ns. The delay time of each section must be 0.125 ns because the output signal will be changed in the interval of signal round-trip time. This nonuniform line is built on a Duroid substrate having 31-mil thickness and relative dielectric constant 2.5, as shown in Fig. 4(b). When we apply an input signal in Fig. 5(a) to the eight-section lines,



(a)



(b)

Fig. 5. (a) Input signal to the AR shown in Fig. 4(b). (b) Comparison of AR output to ideal AR output.

we obtain the output of the nonuniform line, which is shown in Fig. 5(b). Notice that the time-domain response in Fig. 5(b) is obtained by taking the inverse Fourier transform of its frequency-domain result, which extends from 50 MHz to 20 GHz. For convenience, we also show in Fig. 5(b) the output of an ideal AR processor. Fig. 5(b) shows that slight discrepancy exists between measurement result and ideal value.

V. Conclusion

We have derived reflection and transmission coefficients of NTL's in Z-transform forms. In particular, we show that the transmission parameter can be obtained from the reflection parameter and vice versa. The AR format of scattering parameters of an NTL reveals physical insights and deduces applications to practical circuits.

References

[1] F. Y. Chang, "Waveform relaxation analysis of nonuniform lossy transmission lines characterized with frequency-dependent parameters," *IEEE Trans. Circuits Syst.*, vol. 38, pp. 1484–1500, Dec. 1991.

[2] Q. Gu and J. A. Kong, "Transient analysis of single and coupled lines with capacitively loaded junctions," *IEEE Trans. Microwave Theory Tech.*, vol. MTT-34, pp. 952–964, Sept. 1986.

[3] K. N. S. Rao, V. Mahadevan, and S. P. Kosta, "Analysis of straight tapered microstrip lines—ASTMIC," *IEEE Trans. Microwave Theory Tech.*, vol. MTT-25, pp. 164–164, Feb. 1977.

[4] O. P. Rustogi, "Linearly tapered transmission line and its applications in microwave," *IEEE Trans. Microwave Theory Tech.*, vol. MTT-17, pp. 166–168, Mar. 1969.

[5] J. E. Schutt-Aine, "Transient analysis of nonuniform transmission lines," *IEEE Trans. Circuits Syst. I*, vol. 39, pp. 378–385, May 1992.

[6] K. Lu, "An efficient method for analysis of arbitrary nonuniform transmission lines," *IEEE Trans. Microwave Theory Tech.*, vol. 45, pp. 9–14, Jan. 1997.

[7] J. P. Mahon and R. S. Elliott, "Tapered transmission lines with a controlled ripple response," *IEEE Trans. Microwave Theory Tech.*, vol. 38, pp. 1415–1420, Oct. 1990.

[8] R. W. Klopfenstein, "A transmission line taper of improved design," *Proc. IRE*, vol. 44, pp. 31–35, Jan. 1956.

[9] S. C. Burkhart and R. B. Wilcox, "Arbitrary pulse synthesis via nonuniform transmission lines," *IEEE Trans. Microwave Theory Tech.*, vol. 38, pp. 1514–1518, Oct. 1990.

[10] L. A. Hayden and V. K. Tripathi, "Characterization and modeling of multiple line interconnections from time-domain measurements," *IEEE Trans. Microwave Theory Tech.*, vol. 42, pp. 1737–1743, Sept. 1994.

[11] S. D. Corey and A. T. Yang, "Interconnect characterization using time domain reflectometry," *IEEE Trans. Microwave Theory Tech.*, vol. 43, pp. 2151–2156, Sept. 1995.

[12] C.-W. Hsue and T.-W. Pan, "Reconstruction of nonuniform transmission lines from time-domain reflectometry," *IEEE Trans. Microwave Theory Tech.*, vol. 45, pp. 32–38, Jan. 1997.

[13] D. M. Pozar, *Microwave Engineering*. Reading, MA: Addison-Wesley, 1990.

[14] A. V. Oppenheim and R.-W. Schafer, *Discrete-Time Signal Processing*. Englewood Cliffs, NJ: Prentice-Hall, 1989.

[15] A. Papoulis, *Probability, Random Variable and Stochastic Processes*, 3rd ed. New York: McGraw-Hill 1991.

Te-Wen Pan was born on August 28, 1969, in Taipei, Taiwan, R.O.C. He received the B.S. and the M.S. degrees in electronic engineering from the National Taiwan University of Science and Technology (formerly, the National Taiwan Institute of Technology), Taiwan, R.O.C., in 1994 and 1996, respectively, and is currently working toward the Ph.D. degree.

From July 1996 to February 1997, he was with the Electronic Testing Center, Taoyuan, Taiwan, R.O.C., where he worked on electromagnetic compatibility testing and research. His current research interests are in the areas of electromagnetic compatibility, microwave circuit design, and transmission-line modeling and applications.

Ching-Wen Hsue (SM'91) was born in Tainan, Taiwan, R.O.C.. He received the B.S. and M.S. degrees in electrophysics and electronic from National Chiao-Tung University, Hsin-Chu, Taiwan, R.O.C., in 1973 and 1975, respectively, and the Ph.D. degree from Polytechnic University (formerly, the Polytechnic Institute of Brooklyn), Brooklyn, NY, in 1985.

From 1975 to 1980, he was a Research Engineer at the Telecommunication Laboratories, Ministry of Communication, Taiwan, R.O.C. From 1985 to 1993, he was with Bell Laboratories, Princeton, NJ, as a Member of Technical Staff. In 1993, he joined the Department of Electronic Engineering, National Taiwan University of Science and Technology, Taipei, Taiwan, R.O.C., as a Professor, and since August 1997, he has served as the department Chairman. His current interests are in pulse-signal propagation in lossless and lossy transmission media, wave interactions between nonlinear elements and transmission lines, photonics, high-power amplifiers and electromagnetic inverse scattering.