A TIME - DOMAIN TECHNIQUE FOR CHARACTERIZING LEAKY COAXIAL CABLES

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Abstract

A time-domain technique for the measurement of return loss, attenuation and coupling loss of leaky coaxial cables is described. The technique is based on a simple signal flow graph procedure where the leaky cable is represented by a three-port network, with the third port being introduced to represent the leakage from the cable. The measurement procedure is described, while experimental results for three different leaky coaxial cables are presented and compared with frequency-domain data.

Introduction

Because of the inability of free radio propagation to provide radio coverage for underground and certain above ground mobile communications, the leaky feeder technique has been of growing interest. For a successful underground communication the leaky feeders provide guided and also leaky electromagnetic waves so that the field can be picked up by mobile receivers at distances away from and along the leaky line[1]. The performance of such radiating cables is basically determined by measuring the coupling loss between the cable and an antenna, attenuation per unit length and the optimum operating frequency. In addition, the sensitivity of the performance of the cable to mounting position is also of practical importance. It is, therefore, the objective of this paper to present a time-domain technique suitable for describing the performance of leaky feeders. The reason for employing a time-domain method rather than routine frequency-domain measurements is the desire for describing the cable performance over a wide frequency band using a single measurement for each quantity of interest. Hence the behaviour of the line beyond its operating band, and in particular below its emperical optimum frequency, can be studied[2]. Also the advantages of time-domain techniques in reducing errors due to mismatch of the measurement system components and in separating (in time) the so called "end effects" from the overall characteristics of the line are of particular importance. The separation of the "end effects" can also facilitate an isolated study of the periodic effects due to mode convergence at the beginning and end of the line.

The proposed technique employs the equations derived from a signal flow graph procedure where the leaky cable is represented as three-port network with the third port being introduced to represent the leakage. The availability of such a third port in the analysis has also the advantage of describing the cable sensitivity to mounting position, e.g. reflections from the surroundings. It is shown that by sampling only segments of the transient response of the cable to an impulse waveform, the scattering parameters describing the performance of the cable can be obtained. The measurement procedure is described while experimental results for three different leaky coaxial cables are presented and compared with spot check frequency domain measurements.

The Scattering Parameters Measurement In Frequency and Time Domains

A typical measurement system is shown in Fig. 1(a). The leaky cable, which is represented by the network L, is connected to a voltage source and a load by lengths of lossless delay lines ℓ_1 and ℓ_2, respectively. In a practical measurement of the reflection coefficient S_{11}, the transmission coefficient S_{21} and the coupling coefficient S_{31}, it is necessary to measure the voltages E_1^-, E_2^+ and E_3^+, respectively, in addition to the incident voltage E_1^+. Following the system signal flow graph in Fig. 1(b) it is easy to show that

$$E_1^- = \frac{(1-\Gamma_g)}{2\Delta} E\gamma_1^2[S_{11}\{1 - (S_{33}\Gamma_s + S_{22}\gamma_2^2\Gamma_\ell + S_{32}\Gamma_s S_{23}\gamma_2^2\Gamma_\ell) + S_{33}\Gamma_s S_{22}\gamma_2^2\Gamma_\ell\} + \gamma_2^2 S_{21}S_{12}\Gamma_\ell \cdot \{1 - S_{33}\Gamma_s\} + S_{31}\Gamma_s S_{13}\{1 - S_{22}\gamma_2^2\Gamma_\ell\}] \quad (1)$$

$$E_2^+ = \frac{(1-\Gamma_g)}{2\Delta} E\gamma_1[S_{21}\gamma_2(1 - S_{33}\Gamma_s) + S_{31}\Gamma_s S_{23}\gamma_2] \quad (2)$$

$$E_3^+ = \frac{(1-\Gamma_g)}{2\Delta} E\gamma_1[S_{31}k(1 - \gamma_2^2 S_{22}\Gamma_\ell) + S_{21}\gamma_2^2\Gamma_\ell S_{32}k] \quad (3)$$

$$E_1^+ = \frac{(1-\Gamma_g)}{2\Delta} E\{1 - (\gamma_2^2 S_{22}\Gamma_\ell + \Gamma_s S_{33} + S_{32}\Gamma_s S_{23}\gamma_2^2\Gamma_\ell) + (S_{33}\Gamma_s S_{22}\gamma_2^2\Gamma_\ell) \quad (4)$$

where the network determinant Δ is given by

$$\Delta = 1 - (\gamma_1^2\Gamma_g S_{11} + S_{22}\gamma_2^2\Gamma_\ell + \gamma_1^2\gamma_2^2\Gamma_g\Gamma_\ell S_{21}S_{12} + \Gamma_s S_{33} + \Gamma_g\gamma_1^2 S_{31}\Gamma_s S_{13} + S_{32}\Gamma_s S_{23}\gamma_2^2\Gamma_\ell) + (\Gamma_g\Gamma_\ell\gamma_1^2\gamma_2^2 S_{11}S_{22} + \Gamma_g S_{11}\gamma_1^2 S_{33}\Gamma_s + S_{22}\gamma_2^2\Gamma_\ell S_{33}\Gamma_s + \Gamma_g\gamma_1^2 S_{31}\Gamma_s S_{13}S_{22}\gamma_2^2\Gamma_\ell + S_{32}\Gamma_s S_{23}\gamma_2^2\Gamma_\ell S_{11}\Gamma_g\gamma_1^2 + S_{33}\Gamma_s\Gamma_g\Gamma_\ell\gamma_1^2\gamma_2^2 S_{21}S_{12}) - (\Gamma_g\gamma_1^2 S_{11}S_{22}\gamma_2^2\Gamma_\ell S_{33}\Gamma_s) \quad (5)$$

and the scattering coefficients are normalized with respect to the characteristic impedance of the delay transmission lines.

To simplify the analysis, the third port will be assumed matched and hence no reflections are received from this port (i.e. $\Gamma_s = 0$). This condition, however, is only valid if the cable is appropriately mounted away from any interfering objects. By employing the above condition and expanding $1/\Delta$ it is easy to show that

$$E_1^- = V(\omega)\ \Gamma'_g\ \gamma_1^2\ S_{11}[1 + f_1(\gamma_1,\gamma_2)] \quad (6)$$

$$E_2^+ = V(\omega)\ \Gamma'_g\ \gamma_1\gamma_2\ S_{21}[1 + f_2(\gamma_1,\gamma_2)] \quad (7)$$

$$E_3^+ = V(\omega)\ \Gamma'_g\ \gamma_1\ S_{31}[1 + f_3(\gamma_1,\gamma_2)] \quad (8)$$

$$E_1^+ = V(\omega)\ \Gamma'_g\ [1 + f_4(\gamma_1,\gamma_2)] \quad (9)$$

where $f_{1,2,3,4}$ are functions containing at least γ_1^2, γ_2^2 or $\gamma_1\gamma_2$, $V(\omega)$ is the Fourier transform of $\nu(t)$ which is the impulse type waveform of E and $\Gamma'_g = (1 - \Gamma_g)/2$. Taking the inverse Fourier transform and neglecting f_1, f_2, f_3 and f_4 since they will transform with a time delay of at least $2\tau_1$, $2\tau_2$ or $\tau_1 + \tau_2$, we obtain $g_1^-(t) = \nu(t)*\Gamma'_g(t)*s_{11}(t - 2\tau_1)$, $g_2^+(t) = \nu(t)*\Gamma'_g(t)*s_{21}(t - \tau_1 + \tau_2)$, $g_3^+(t) = \nu(t)*\Gamma'_g(t)*s_{31}(t - \tau_1)$ and $g_1^+(k) = \nu(t)*\Gamma'_g(k)$ where s(t) is the Fourier transform of any scattering coefficient $S(\omega)$ and * denotes a convolution[3]. Hence if the delays (τ_1 and τ_2) introduced by the delay transmission lines are made to be longer than the period of time required for the transient responses of S_{11}, S_{21} and S_{31} to the incident waveform, then by performing a Fourier transform on these waveform segments, the scattering parameters $S_{11}(\omega)$, $S_{21}(\omega)$ and $S_{31}(\omega)$ can be obtained by combining (6) to (9). It should be noted that, by neglecting f_3 in (8) the main coupling path E to S_p which involves the coupling coefficient S_{31} is only considered.

Measurement System

A measurement system for both the transmission S_{21} and reflection S_{11} coefficients is straightforward as has been reported before[3,4]. The system simply consists of a pulse generator, delay lines and sampling oscilloscope. The network L is placed either after the pulse generator or the sampling head depending on whether the transmission or the reflection coefficients are being measured, respectively[3]. It should be noted that ℓ_1 and ℓ_2 are required to produce a delay longer that the transient response being measured so as to eliminate any overlap from multiple reflections.

The coupling coefficient S_{31}, on the other hand, is calculated from the Fourier transform of the signal received by a standard dipole antenna at a distance which is sufficient to ensure far field calculations[5]. Such measurements, although neglect the unavoidable effect of the environment[6], give results which are characteristic of a typical line[7]. The effect of a practical environment (i.e. metallic obstacles, tunnels or uneven ground), however, can be taken into account by evaluating Γ_s (see Fig. 1(b)) in a typical situation by sampling a larger transient segment which includes the effect of Γ_s. Also, since the measurement of the received signal will depend on the characteristics of the receiving dipole, a calibration procedure similar to that reported by Bates et al. is required[5].

Such procedure basically involves the calculation of the transmit $T_s(f)$ and receive $R_s(f)$ transfer functions of the standard dipole in terms of its input impedance Z(f) and effective height $h_e(f)$ using simple transmission and reception equivalent circuits[5]. It should be noted that Z(f) and $h_e(f)$ are known rigorously for a standard dipole antenna and can also be spot checked using standard frequency-domain methods. The transmit transfer function T(f) of the test leaky coaxial cable can then be determined in terms of $T_s(f)$ and $R_s(f)$ leading to the field intensity at a distance from the cable[5]. It should also be noted that the coupling loss can simply be calculated by multiplying the field intensity incident upon a receiving antenna by the receive transfer function R(f).

Experimental Measurements and Results

To test the feasibility of the method, comparative measurements are made on three leaky coaxial cables (set of three Radiax cables RX 4-1, RX 4-3A and RX 5-1) in both time and frequency domains. For time-domain measurements, the transmitted pulse was generated from 150 ps step of an HP-1415A TDR system by using a 25Ω short circuited stub as a pulse forming network[8]. The appropriate segments from the analog signals are sampled and stored in a buffer of a PDP-11/40 computer system, while their Fourier transforms are calculated using FFT-subroutine in the same computer[9]. Although discrete frequency values will be only available from the FFT, intermediate values could also be obtained from interpolation formulae[3]. It should also be noted that for the coupling loss measurement, the cable is mounted in an anechoic chamber to eliminate any interference from the surroundings.

Experimental results for the return, insertion and coupling losses are presented graphically as a function of frequency and compared with spot check frequency domain measurements using a network analyzer. The uncertainty of the time-domain results, based on the combined effect of possible errors, is also estimated[4,10]. The advantages and disadvantages of using the time-domain technique as well as the limitation on the length of the leaky cable are discussed.

Acknowledgments

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References

1. R.A. Farmer and N.H. Shepard, "Guided radiation ... the key to tunnel talking", IEEE Trans. on Vehicular Communication, vol. VC-14, pp. 93-102, 1965.

2. D.J.R. Martin, "A general study of the leaky feeder principle", The Radio and Elect. Engin., vol. 45, pp. 205-214, 1975.

3. A.M. Nicolson, "Broad-band microwave transmission characteristics from a single measurement of the transient response", IEEE Trans. Inst. Measurement, vol. IM-17, pp. 395-402, 1968.

4. M.F. Iskander and S.S. Stuchly, "A Time-domain technique for measurement of dielectric properties of biological substances", IEEE Trans. Inst. Measurement, vol. IM-21, pp. 425-429, 1972.

5. R.H.T. Bates and G.A. Burrell, "Towards faithful radio transmission of very wide bandwidth signals", IEEE Trans. Antennas and Propagation, vol. AP-20, pp. 684-690, 1972.

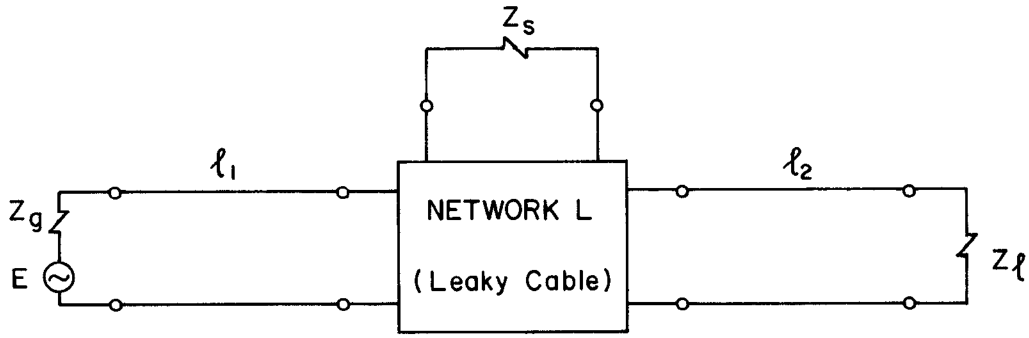
6. S. Okada, et al, "Leaky coaxial cable for communication in high speed railway transportation", Radio and Elect. Engin., vol. 45, pp. 224-228, 1975.

7. D.J. Cree and L.G. Giles, "Practical performance of radiating cables", Radio and Elect. Eng., vol. 45, pp. 215-223, 1975.

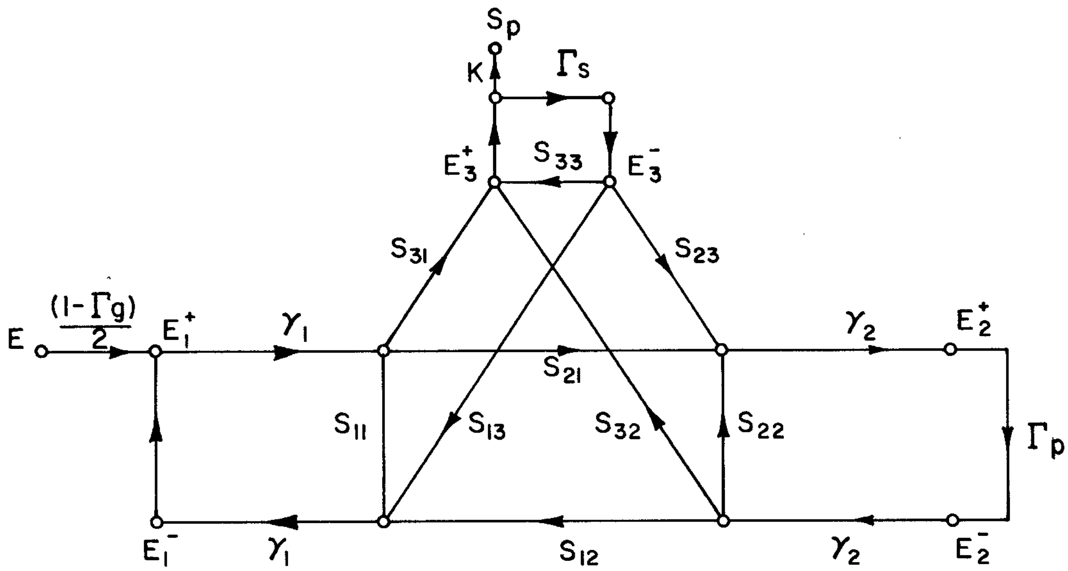
8. G.F. Ross, "The synthetic generation of phase-coherent microwave signals for transient behaviour measurements", IEEE Trans. Microwave Theory and Techniques, vol. MTT-17, pp. 704-706, 1965.

9. Digital Equipment Corporation, Maynard, Mass., Lab Applications - 11.

10. G.F. Ross, A.M. Nicolson and R.S. Smith, et al, "Transient behaviour of microwave networks", RADC-TR-68-92, 1968.



(a) Schematic diagram of the three-port leaky cable network



(b) Signal flow graph for the leakv cable svstem

Fig. 1 Schematic Diagram of a Typical System Characterization