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Feasibility and problems of DSL loop topology identification via single-ended line tests

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Abstract- Digital subscriber lines (DSL) offer the possibility to deliver broadband services over the existing telephone network. Yet, before deploying DSL, the subscriber loops must be tested to see whether they can support high-speed data services, and at what rate. Single-ended automatic qualification is essential for achieving low-cost deployment of DSL, since it allows loops to be qualified in bulk without human intervention at the customer's location. An even more ambitious challenge is to fully characterize the loop, i.e. to identify its loop topology (number of loop sections, length and cable type of each section). This paper discusses the feasibility and the problems of loop topology identification via single-ended measurements.

I. Introduction

Nowadays Digital Subscriber Line (DSL) technology is one of the possible ways to offer broadband access. It exploits the telephone network of the subscriber to transport high-speed data. Yet, before deploying DSL to a new customer, the subscriber loop must be tested to see if it can support the considered DSL service. In Europe, a subscriber loop most often consists of a cascade of several line sections with increasing diameter (gauges) from the central office towards the customer. Identifying the topology of a subscriber loop (number of sections, length and gauge of each section) is important for a realistic capacity estimation, but can also benefit network planning and maintenance in general. Several measurement techniques exist, however single-ended testing is often preferred over dual-ended line testing because it eliminates the necessity of dispatching a technician to the customer's site. The tests can then be run periodically from the central office, to test the loops in bulk without human intervention.

The underlying physical phenomenon that allows estimation of the loop topology through single-ended line tests is that gauge changes and the end of the loop represent a discontinuity in the characteristic impedance along the loop. As a result, when a signal is injected into the loop, every discontinuity will generate a reflection. Analysis of all these echoes allows (at least in theory) inducing the complete loop topology.

There are few papers in literature addressing the problem of subscriber loop topology estimation through single-ended tests. Galli et al [1,2] use time domain reflectometry (TDR) with a square pulse as excitation signal. Consequently the measured reflections are dependent on the shape of the injected pulse. Boets et al therefore propose to use the one-port scattering parameter S_{11}, which is the ratio of the reflected to the injected voltage wave, as an indicator of the loop topology [3,4]. Here, the excitation signals are discrete multitones (DMT) placed on the ADSL grid. Another measurement technique proposed by Dodds et al [5], energizes the loop with a sinusoid that increases in frequency in discrete steps and measures the received reflections through coherent detection. Both latter measurement techniques can be catalogued as frequency domain reflectometry (FDR). Nevertheless, both TDR and FDR encounter difficulties when attempting to identify the loop topology. This paper discusses the main problems in both domains. As the paper will show, even for very simple cases, the evaluation of the loop topology is challenging.

The remainder of this paper is structured as follows. Section II gives some theoretical background and states the problems, which are then further detailed in Sections III and IV. Section V summarizes the most important conclusions.

II. Problem statement

With single-ended line testing all measurements are performed from the central office and by consequence the subscriber loop should be considered as a one-port. All the information that can be acquired about the loop through this port is contained in the one-port scattering parameter $S_{11}(f)$, which is the ratio of the reflected voltage wave to the incident voltage wave. The one-port scattering parameter can be measured by means of a network analyser in the frequency domain. If the subscriber loop consists of a single line, then this transmission line can be modelled as derived in [4]:

$$S_{11}(f) = \frac{-\rho_g + \rho_L e^{-2\gamma l}}{1 - \rho_g \rho_L e^{-2\gamma l}} \tag{1}$$

l is the unknown length of the line, γ the propagation constant of the line, ρ_L the reflection due to the line end and ρ_g the reflection at the measurement device, given by:

$$\rho_g = \frac{Z_g - Z_c}{Z_g + Z_c}; \quad \rho_L = \frac{Z_L - Z_c}{Z_L + Z_c} \tag{2}$$

with Z_g the impedance of the measurement device, Z_c the characteristic impedance of the transmission line and Z_L the load at the line end as given in Figure 1. If we linearize formula (1) according to $\frac{1}{1+x} = 1 - x + x^2 - \cdots$, we obtain a structure which allows easy interpretation.

$$S_{11}(f) = -\rho_g + \rho_L\left(1 - \rho_g^2\right)e^{-2\gamma l} + multiples \tag{3}$$

The first term represents the reflection at the measurement device, due to the impedance mismatch between the measurement device and the characteristic impedance of the line. The second term is the reflection caused by the line end, which is the actual meaningful reflection containing information about the line length. We will also receive multiples of this reflection because the signal bounces (in theory) infinitely back and forth on the line. In practice, however, the attenuation strongly increases with every multiple and as such only a few multiples (if any) will be visible. We can rewrite equation (3) with the complex propagation constant $\gamma = \alpha + j\beta$ and the following assumptions to keep the example simple without loss of generality: a) we ignore the multiple reflections, b) we neglect ρ_g^2, since a good measurement device will be approximately matched to the line under test, leading to a small ρ_g and a negligible ρ_g^2, c) we assume the line end to be open (as is the case for an on-hooked telephone), resulting in $\rho_L = 1$. Therefore $S_{11}(f)$ for a single line simplifies to:

$$S_{11}(f) = -\rho_g + e^{-2\alpha l} e^{-2j\beta l} \tag{4}$$

Since in general ρ_g is a complex function, we can write the real and imaginary part of $S_{11}(f)$ as:

$$\Re\{S_{11}(f)\} = -\Re\{\rho_g\} + e^{-2\alpha l}\cos(-2\beta l) \tag{5}$$

$$\Im\{S_{11}(f)\} = -\Im\{\rho_g\} + e^{-2\alpha l}\sin(-2\beta l) \tag{6}$$

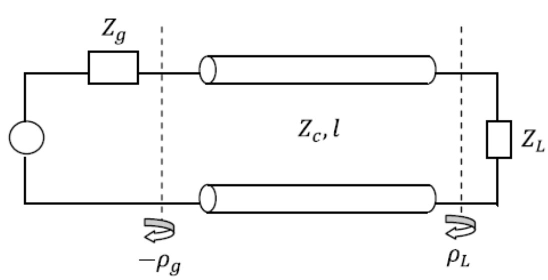


Figure 1. Reflections for a subscriber loop consisting of a single line

Although formulas (5) and (6) do not look very complicated, it is still hard to identify the line length l from $S_{11}(f)$-measurements. The first problem we encounter is the unwanted reflection ρ_g at the measurement device. Section III will describe the influence of this term. Secondly, the unknown parameter l is present in the exponent as well as in the sinusoidal function. Since the attenuation coefficient α and the phase shift β are frequency dependent, classical signal parameter estimation is impossible. Moreover, if the gauge type is not known a priori, then α en β are also unknowns which have to be estimated as well. Section IV will focus on the estimation of the line length l.

III. Near-end reflection

In general, the measurement device is not matched to the loop under test. The measured scattering parameter of the subscriber loop is expressed in reference impedance Z_g, which is typically 50 Ω, while the characteristic impedance Z_c of a twisted pair varies around 115 Ω. This impedance mismatch will generate a reflection at the connection of the measurement device, called the "near-end reflection" (first term in (5) and (6)), potentially masking the real reflections in the frequency domain as well as the time domain ([1],[3]). Figure 2 illustrates this problem for a single line of 1800 m. Several approaches are possible to tackle this problem.
In [3] the impedance in which the measured scattering parameter is expressed is changed, as to match the characteristic impedance of the line as well as possible ($Z_{new} \approx Z_c$). This is done through post-processing with (7). For this, the characteristic impedance $Z_c(f)$ of the line needs to be estimated first.

$$S_{11}\big|_{Z_{new}} = \frac{Z_g\left(\dfrac{1+S_{11,meas}}{1-S_{11,meas}}\right) - Z_{new}}{Z_g\left(\dfrac{1+S_{11,meas}}{1-S_{11,meas}}\right) + Z_{new}} \tag{7}$$

In [1] and [5] the near-end reflection ρ_g is estimated by measuring a very long line, in the time and frequency domain respectively. The near-end reflection is then subtracted from the measured one-port scattering parameter. However, this assumes that a very long line of similar type is available. Another possibility would be to fit the near-end reflection. In [6] a rational function of second order over first order was found to give satisfactory results.

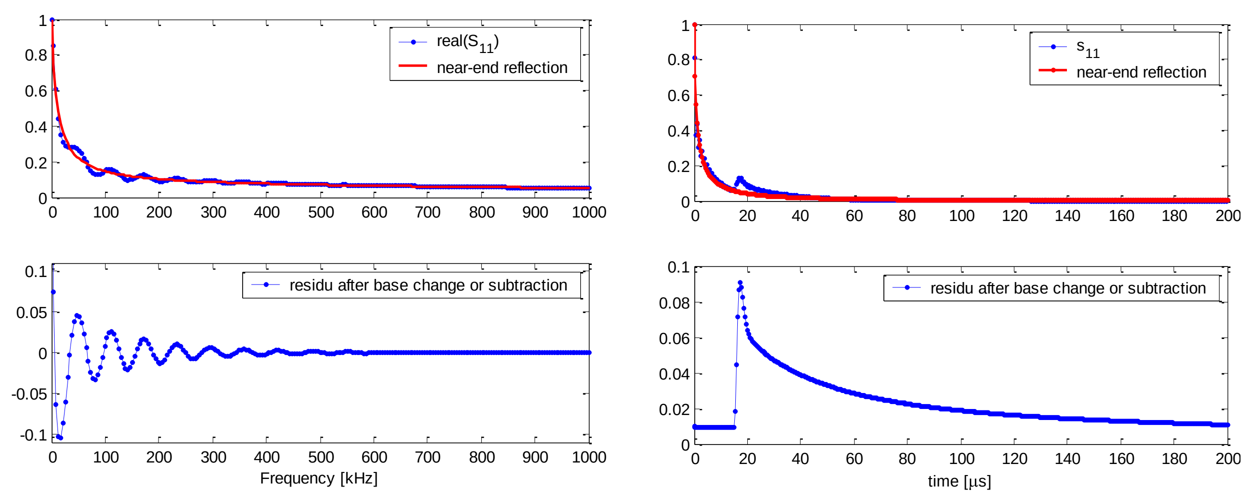


Figure 2. Near-end reflection partially masking the line end reflection of a 1800 m line (top) in the frequency domain (left) and in the time domain (right); after base change or subtraction of near-end reflection, the end reflection becomes more clearly visible (bottom).

In any case, it is important to remove the near-end reflection ρ_g or at least reduce it as much as possible to ease the identification of the loop topology.

IV. Estimation of line length l

When considering a subscriber loop made up of a cascade of several line segments, it is important to keep the reflections as narrow as possible. This is necessary to be able to resolve two closely separated reflections and to be able to distinguish a small reflection in the proximity of a large reflection. Therefore, we return to the example of the single line, to examine the factors that define the width of a reflection.

Once the near-end reflection is removed, formulas (5) and (6) further simplify to:

$$\Re\{S_{11}(f)\} = e^{-2\alpha(f)l}\cos(-2\beta(f)l) \quad (8)$$

$$\Im\{S_{11}(f)\} = e^{-2\alpha(f)l}\sin(-2\beta(f)l) \quad (9)$$

This means that for a single line, the one-port scattering parameter is an exponentially damped sine. The exponential term reflects the attenuation of the line and depends on the attenuation constant α and the line length l. The sinusoidal term represents the standing waves created on the line due to the excitation. It depends on the line length l, as well as on the phase constant β. Note that α and β are frequency dependent, and not constants as their name would suggest.

A. Phase constant β

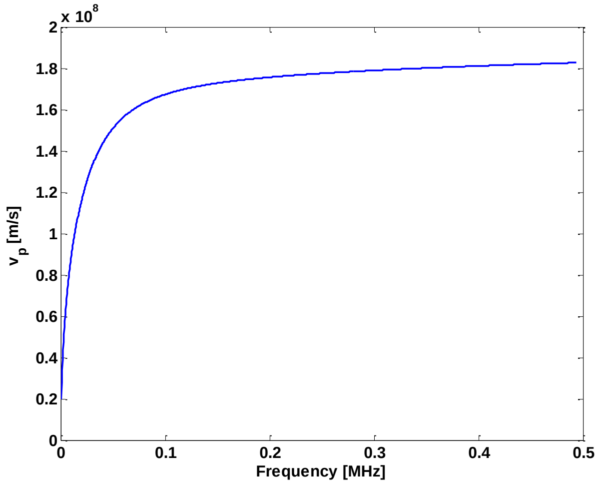
If the phase constant β would be perfectly linear ($\beta = m \cdot f$ with $m \in \mathbb{R}$), an inverse Fourier transform of the sinusoidal term would yield a Dirac pulse at time instant ml/π. This can then be related to the unknown line length l through the propagation speed v_p. In practice, β is not a linear function of frequency and as a consequence we have a lobe around the exact time value instead of a Dirac pulse. In the time domain, this phenomenon manifests itself as dispersion due to the fact that the propagation velocity is frequency dependent. The propagation speed v_p is related to β as follows:

$$v_p(f) = \frac{2\pi f}{\beta(f)} \quad (10)$$

Measuring the one-port scattering parameter in the frequency domain and performing an inverse Fourier transform corresponds to measuring the impulse response in the time domain, which is strongly influenced by the dispersion, thus causing broad reflections. However, if $S_{11}(f)$ is multiplied with a whitener in the frequency domain, the reflection can be narrowed. This corresponds with the use of an equalizer in the time domain, as used in DSL modems to counter the channel dispersion and avoid intersymbol interference.

B. Attenuation constant α

Moreover, the exponential attenuation term in (8) and (9) will also widen the lobe. The fact that the attenuation constant α is not constant, as its name would suggest, but rather frequency dependent, causes supplementary dispersion in the time domain. In theory, the attenuation could be compensated by multiplying the signal with a factor $e^{2\alpha(f)l}$. However, this can only be done when the unknown line length has already been approximated and the attenuation constant α is known. Compensation results in an improvement in resolution, from which the identification can benefit. An alternative approach would be to estimate the envelope of $S_{11}(f)$ and to compensate by dividing it by this estimation. In this way, the exponential decay is also countered.



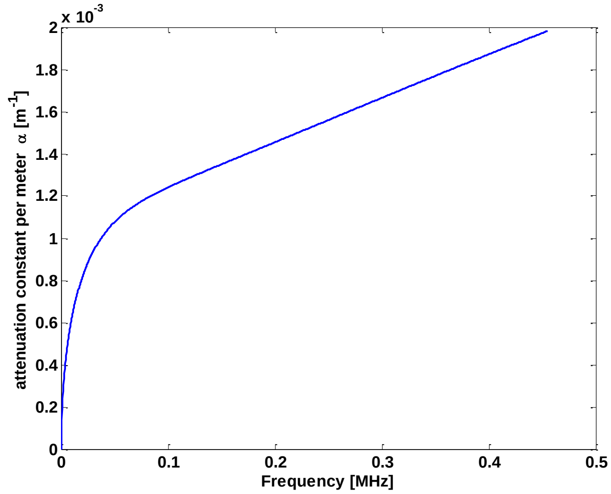


Figure 3. The frequency dependent propagation speed (left) and attenuation constant (right) cause dispersion on the subscriber loop.

V. Conclusions

This paper addressed some of the major problems when attempting to identify the topology of a subscriber loop with single-ended measurements. Even in the most simple case, namely a single line, the near-end reflection might mask the reflection from the loop end. Moreover, when performing an inverse Fourier transform on the measured one-port scattering parameter $S_{11}(f)$, we do not obtain a Dirac pulse corresponding to the reflection, but rather a lobe around this location. As the paper explained, there are two main reasons for this widening.

1. The phase characteristic is not linear.
2. The line attenuation $e^{-2\alpha(f)l}$ is frequency dependent.

This is equivalent to the concept of dispersion in the time domain. These drawbacks are not a major problem for a single line. On the contrary, if we consider cascades of lines, as is mostly the case in subscriber loops, more than one reflection is received. It is then of uttermost importance to keep the lobe as small as possible to attain a high resolvability, especially when the reflections are in each others proximity.

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