Cable measurements supporting xDSL technologies

The ongoing developments in the field of high-speed data transmission over twisted-pair cables (DSL – Digital Subscriber Line) leads to methods that can exploit the transmission media with increased efficiency. MIMO (Multiple-Input Multiple-Output) transmission, i.e., the coordinated use of many pairs within a bundle, is such an emerging DSL technology. A further goal is to maintain a certain transmission quality in the presence of non-DSL noise. For example, cancellation techniques to combat narrowband radio interference, which is generated by radio amateur transmitters and may be picked up by the wire close to the subscriber or by the in-house wiring, become important.

The efficiency of these techniques strongly depends on cable properties that have not been the focus of cable measurements so far. This paper gives an overview of special cable measurements and shows their importance for future DSL technologies.

Keywords: DSL (Digital Subscriber Line); cable measurements; MIMO (Multiple-Input Multiple-Output); RFI (Radio Frequency Interference)

Kabelmessungen für zukünftige DSL-Technologien.

Die stetig voranschreitende Entwicklung im Bereich der hochratigen Datenübertragung über die Kupferzweidrahtleitung (DSL – Digital Subscriber Line) führt zu immer komplexeren Verfahren mit dem Ziel, das Medium so gut wie möglich auszunutzen. Ein zentrales Thema ist die koordinierte Nutzung vieler Paare eines Kabelbündels, was unter dem Begriff MIMO(Multiple-Input Multiple-Output)-Übertragung zusammengefasst wird. Ein weiteres Ziel ist die Garantie einer bestimmten Übertragungsqualität bei Gegenwart von nicht DSL-artigem Rauschen. Verfahren, z. B. zur Kompensation von schmalbandigen Störungen, welche durch Amateurfunker verursacht werden und für die vor allem der Leitungsabschnitt nahe am Teilnehmer sowie die Hausverkabelung wie eine Antenne wirken, gewinnen an Bedeutung.

Die Effektivität dieser Verfahren hängt stark von Leitungseigenschaften ab, welche bisher nur unzureichend untersucht wurden. Dieser Beitrag behandelt spezielle Kabelmessverfahren und zeigt ihre Bedeutung für die DSL-Technik von morgen.

Schlüsselwörter: DSL (Digital Subscriber Line); Kabelmessungen; MIMO (Multiple-Input Multiple-Output); RFI (Radio Frequency Interference)

T. MAGESACHER, W. HENKEL, G. TAUBÖCK, T. NORDSTRÖM

1. Introduction

Broadband access technologies over copper twisted pairs, summarized in the keyword DSL (Digital Subscriber Line), have experienced significant progress in recent years. Field trials with prototypes have been carried out successfully and operators have already started the deployment of several DSL types.

Concerning research there are two goals: First, the twisted-pair channel should be used as efficiently as possible in terms of achievable throughput under the current operating conditions. The performance of today's DSL systems is often limited by crosstalk between neighboring loops. If it is possible to use all the loops of a bundle, these crosstalk paths can be utilized for data transmission. Thus, the drawback of crosstalk can actually be turned into an advantage. Systems that consider the whole bundle as channel employ MIMO (Multiple-Input Multiple-Output) techniques, which are also very important in wireless communications. Second, it is vital to provide reliable transmission in the presence of various other disturbances. One challenging phenomenon is the undesirable reception of radio signals by cables. In particular, narrowband transmitters can generate interference that is stronger than the far-end signal at the receiver. Decreasing radio frequency interference (RFI) is thus vital in terms of transmission quality.

Extensive research on MIMO and interference mitigation has lead to good theoretical results. In order to evaluate the performance of the algorithms and techniques in real systems, knowledge about special cable properties is necessary. In MIMO systems, e. g., the crosstalk behavior between all combinations of pairs is essential. For assessment of RFI mitigation techniques, information about coupling properties between differential-mode and common-mode is vital.

This paper deals with measurement techniques yielding the cables' properties required for future generations of DSL. Chapter 2 provides an overview on the MIMO technique in wireline communications. Chapter 3 deals with mitigation of interference from radio transmitters. Cable measurement techniques supporting these techniques are described in Chap. 4. A summary and concluding remarks follow in Chap. 5.

MAGESACHER, Thomas, Dipl.-Ing., HENKEL, Werner, Dr., TAUBÖCK, Georg, Dipl.-Ing., NORDSTRÖM, Tomas, Dr., ftw. Forschungszentrum Telekommunikation Wien, Techgate, Donau-City-Straße 1/3, A-1220 Wien.

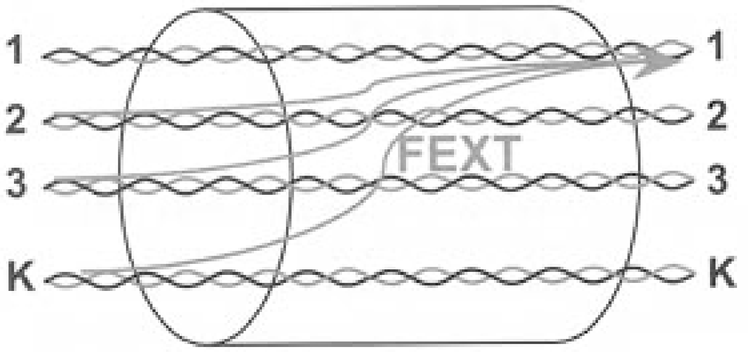

Fig. 1. MIMO cable channel

2. Multiple-Input Multiple-Output transmission

In order to exploit the copper cable, the interaction between individual pairs of a bundle has to be taken into account. A first step towards a more cooperative treatment was made in frequency planning and in the design of power back-off algorithms for VDSL. In the wireless domain, we are used to applying multiuser detection. In wireline, the cable is also a multiple transmit channel. Thus, a whole bundle of pairs can easily be considered as one transmit channel and a so-called MIMO transmission is obtained. This MIMO channel consists not only of the direct loop responses, but also of the so-called FEXT (Far-End Crosstalk) responses, which describe the coupling from one pair to another on the opposite side (Fig. 1).

If a system has access to both sides of a number of cable pairs, the FEXT paths can be seen as useful transmit channels. Examples would be the transmission between the central office and a street cabinet or transmission from a cabinet to a bigger company or a mobile station where a whole cable bundle is used. Nevertheless, there exist important cases where only one-sided processing of multiple signals is possible. This is typically the case when wire pairs branch out from a cabinet (or central office) to customers at different locations. In this paper we will focus on the two-sided case. A study of the alternative one-sided case can be found, e.g., in (*Ginis, Cioffi, 2001*).

In time domain, the direct and crosstalk channels can in principle be described by convolutions with the corresponding impulse responses h_{kj}, where h_{kk} represent the direct paths and h_{kj}, $k \neq j$ are the FEXT paths. The NEXT (Near-End Crosstalk) responses are denoted by g_{kj}. The MIMO channel output samples c_k can thus be written as

$$c_k = h_{kk} \cdot a_k + \underbrace{\sum_{j=1,\, j\neq k}^{K} h_{kj} \cdot a_j}_{\text{FEXT}} + \underbrace{\sum_{m=1}^{L} g_{km} \cdot b_m}_{\text{NEXT}}, \tag{1}$$

where a_j and b_m are the far-end and near-end transmit samples, respectively. Hereinafter, we assume that NEXT is avoided by, e.g., a suitable duplexing method. We proceed with a frequency domain description of multicarrier transmission with DMT (Discrete MultiTone) over the MIMO channel. Being aware that a cyclic convolution (assuming a sufficiently long cyclic prefix) in time domain maps into a multiplication by a factor in DFT domain, we obtain a channel matrix for every single frequency:

$$\mathbf{y}(n) = \mathbf{H}(n) \cdot \mathbf{x}(n), \quad n = 1, \ldots, N. \tag{2}$$

The channel matrix $\mathbf{H}(n)$ thus contains the direct paths in its main diagonal and the FEXT responses elsewhere. In ideal case all FEXT components would vanish and a purely diagonal matrix would remain. This would lead to separate transmit channels without any coupling. Even with nonzero FEXT components, we can easily obtain a diagonalization of the matrix $\mathbf{H}(n)$ by applying the singular value decomposition, resulting in

$$\mathbf{H}(n) = \mathbf{Q}(n) \cdot \mathbf{\Lambda}(n) \cdot \mathbf{P}^H(n), \quad n = 1, \ldots, N, \tag{3}$$

where $\mathbf{Q}(n)$ and $\mathbf{P}(n)$ are unitary matrices with the property that inverse and Hermitian transpose are identical and $\mathbf{\Lambda}(n)$ is a diagonal matrix.

If we now perform a pre-processing by $\mathbf{P}(n)$ in the transmitter and a post-processing by $\mathbf{Q}^H(n)$ in the receiver, we obtain

$$\mathbf{H}(n) = \mathbf{Q}^H(n) \cdot \mathbf{Q}(n) \cdot \mathbf{\Lambda}(n) \cdot \mathbf{P}^H(n) \cdot \mathbf{P}(n) = \mathbf{\Lambda}(n). \tag{4}$$

We obtain a diagonal channel matrix for every carrier frequency, which actually means decoupling without a loss in channel capacity. We can thus write the transfer relation between the input and the output vectors $\mathbf{t}(n)$ and $\mathbf{r}(n)$, respectively, for each frequency as

$$\mathbf{r}(n) = \mathbf{\Lambda}(n) \cdot \mathbf{t}(n), \quad n = 1, \ldots, N. \tag{5}$$

Note that the input and output vectors are across the K channels. The transmission structure with pre- and post-processing for every carrier frequency is depicted in Fig. 2. This MIMO approach is described in more detail in (*Tauböck, Henkel, 2000*).

When comparing such a MIMO system with regular systems that would have FEXT as a disturbance, we have to note that (assuming no NEXT) for long cables FEXT does not play a major role any more, since it experiences the same attenuation as the direct data signal itself. Thus, with increasing length, usual non-MIMO systems starting from a FEXT-limited system become more and more limited by the background noise. If we estimate the performance of such a system (*Nordström, Bengtsson, 2001*), assuming a very high number of 25 pairs for a MIMO system, which is currently not realizable, we obtain rate-length dependencies depicted in Figs. 3 and 4 for background-noise levels of − 140 dBm/Hz and − 120 dBm/Hz, respectively. Due to the mentioned length dependency of FEXT, such MIMO systems are not very suitable for longer cables (> 800 m) and

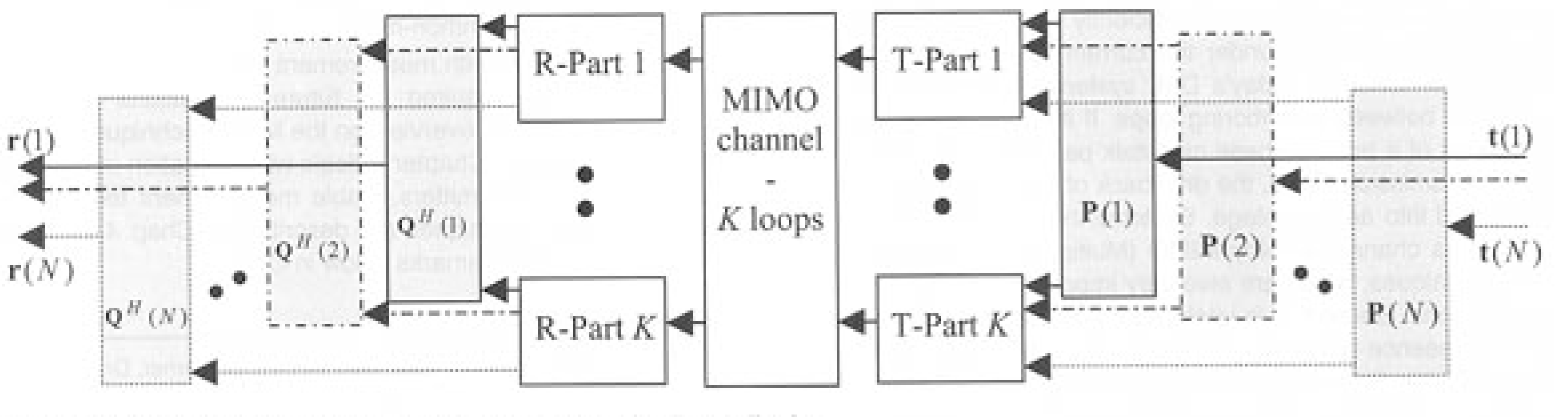

Fig. 2. MIMO-DMT system with pre- and post-processing for diagonalization

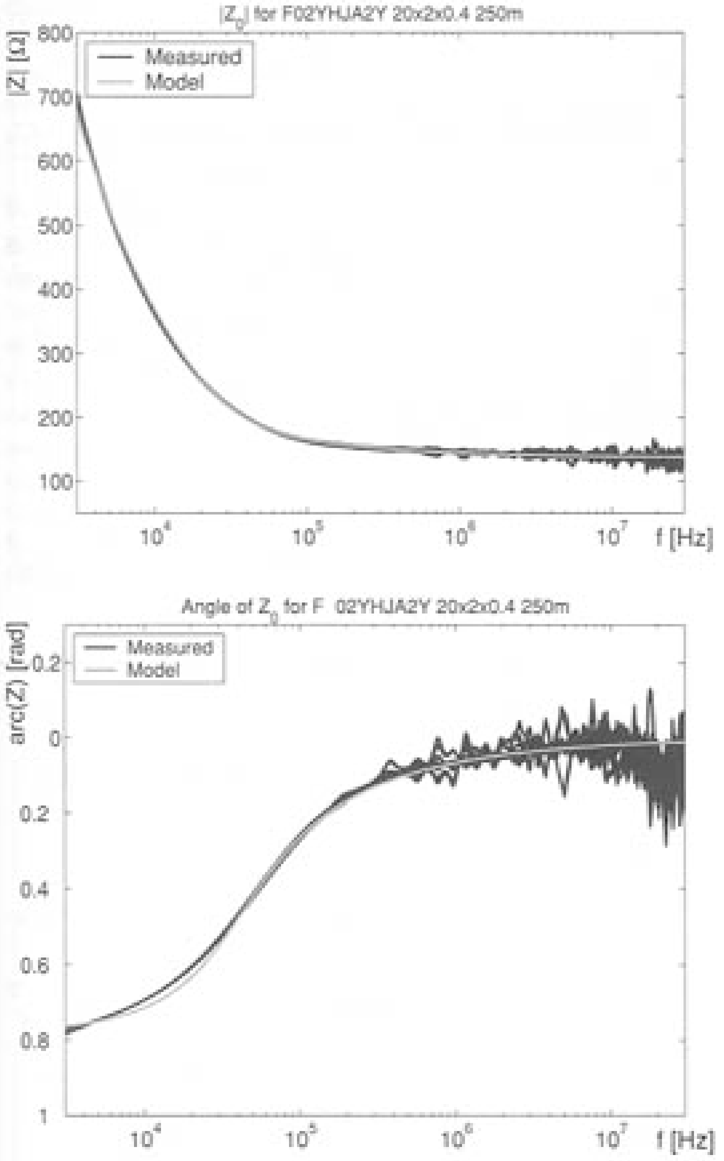

Fig. 9. Characteristic impedances of the layered cable

We will not focus on NEXT, since our MIMO approach assumes that NEXT had been taken care of by other means, i.e., by using FDD (frequency division duplexing) or TDD (time-division duplexing) transmission. However, NEXT can easily be measured with almost every network analyzer and two baluns. It is simply the transfer function between two pairs on one side of the cable (when terminating the opposite side). The cable length should be at least around 100 m in order to be able to neglect length dependencies. The NEXT responses of the layered cable are shown in Fig. 10. As a comparison with the 99 % worst-case model proposed by ETSI shows, the layered cable differs significantly from other European cables.

Far-end crosstalk, in contrast to its near-end counterpart, is essential for a MIMO system, since these paths will be considered as useful transmission channels. Since FEXT is the transfer function between opposite ends of different pairs, the measurement is very similar to the one for NEXT. There is, however, one major difference: The length dependency cannot be neglected, since the coupled signal travels along the loops (in parts on both loops) and is thus attenuated. Usually, the loop response is eliminated from the FEXT response leading to the so-called EL-FEXT (Equal-Level FEXT)

$$N_{\text{EL-FEXT}}(f) = \underbrace{N_{\text{FEXT}}(f)}_{\text{measured}} \cdot \sqrt{1\,\text{km}/l} \cdot 10^{+\alpha(f)\,l/20} \cdot e^{+j\beta(f)\,l} \qquad (9)$$

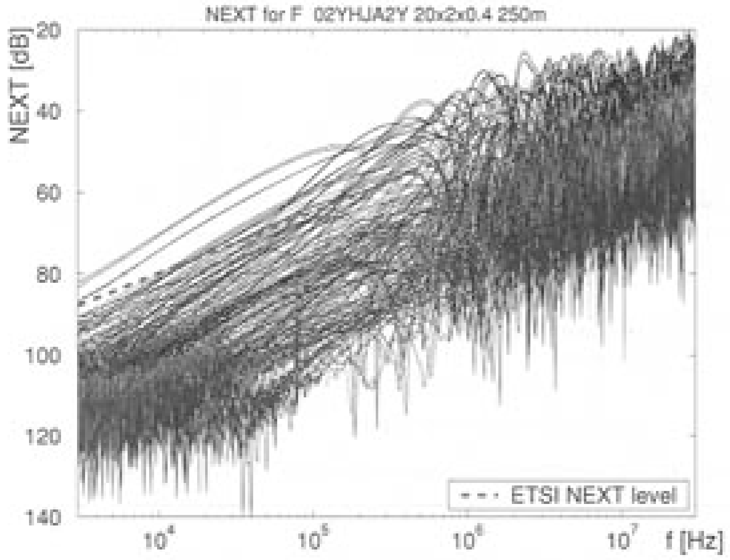

Fig. 10. NEXT responses of the layered cable

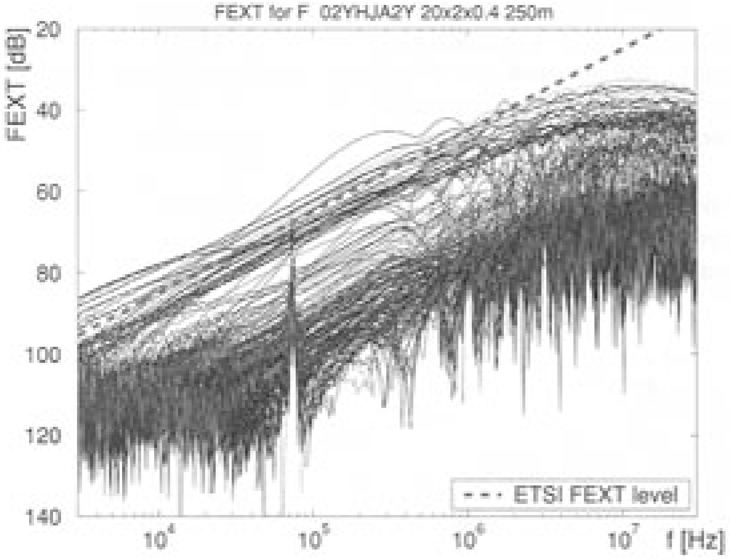

Fig. 11. Equal-level FEXT responses of the layered cable

Since the coupled signal travels along both loops, average values for $\alpha(f)$ and $\beta(f)$ of the corresponding pairs are used. Note that power coupling is considered to be linearly dependent on the length. Therefore, the square-root term appears in the formula for the EL-FEXT. The EL-FEXT responses of the layered cable are shown in Fig. 11, again together with an ETSI reference curve. Also here we see that the layered cable differs significantly from the ETSI reference.

4.2 Symmetry properties

The symmetry parameters defined above can only characterize the cable itself. The properties of the radio channel between the disturber and the cable are not described. However, models for the cable's symmetry form the basis for ingress and egress calculations. LCL and TCL measure the degree of unbalance at one end of the cable and may therefore be relevant in case of RFI ingress close to the modem. LCTL and TCTL assess the unwanted modes attenuated by the line and should characterize situations where ingress and egress mainly happen at the other end of the line.

Two different types of cables have been measured:

(1) Layered cable (see previous section for details): representing a typical layered cable between central office and customer premises, measured on a drum;

Figure 7 shows a block diagram of the principle (*Magesacher et al., 2001*). The common-mode signal is obtained at the center-tap of the primary side of the transformer. An adaptive algorithm adjusts its amplitude and phase to yield a counter-interferer, which is then subtracted from the received signal before the AD converter thus removing the interference.

To assess the performance and detailed implementation requirements of methods like the one described above, it is vital to model the coupling as accurately as possible. The following parameters to assess the cable's symmetry are based on the recommendations provided in (*ITU-T 1996, ITU-T 1999*). The frequency dependent loss measure is the ratio of a driving signal and the resulting unwanted signal:

$$\text{loss\_measure}\,(f) = \frac{\text{driving signal}}{\text{unwanted signal}}. \tag{6}$$

Table 1 summarizes the different loss measures resulting from the combination of common-mode and differential-mode signals as driving and unwanted signals at both ends of the cable (Fig. 12). Since the cable is a passive reciprocal element, both LCL and TCL, as well as LCTL and TCTL, are identical in theory. In practice, this is true for the cable itself, however, in real systems the circuitry interfacing the cable may not be exactly the same at both ends, possibly degrading the symmetry.

Table 1. Summary of symmetry measures and their definition with regard to (6); C = common-mode, D = differential-mode; 1 and 2 denote different ends of the cable

Name of symmetry measure	driving signal	unwanted signal
Longitudinal Conversion Loss (LCL)	C_1	D_1
Transverse Conversion Loss (TCL)	D_1	C_1
Longitudinal Conv. Transfer Loss (LCTL)	C_1	D_2
Transverse Conv. Transfer Loss (TCTL)	D_1	C_2

4. Cable measurements

In this paper, we concentrate on two aspects of DSL: MIMO systems and RFI mitigation. Both require special measurements to collect the basic data for performance studies. For the MIMO case, we need exact knowledge of the transmit characteristics of the cable, i.e., the transfer and the crosstalk functions in amplitude and phase. RFI investigations, however, require knowledge about the symmetry properties of the cable, i.e., the coupling behavior between differential-mode and common-mode. The following two subsections describe the underlying measurements.

4.1 Transmission properties

The transmission properties are on the one hand determined by the characteristics of the direct channel, i.e., the transfer function of a single copper pair. On the other hand, there are also NEXT and FEXT functions that specify the coupling between pairs on the same side of the cable and between pairs on the opposite sides, respectively. Usually, NEXT and FEXT paths disturb other DSL systems and are thus considered to be an unwanted effect of non-ideal cable construction. In a MIMO system, however, especially FEXT is treated as a useful transmit signal. Thus, detailed knowledge of these channels is required. Let us first consider the direct transmission over one pair, described by the transfer function

$$H(f) = e^{-\gamma(f)l} = e^{-(\alpha(f)+j\beta(f))l}, \tag{7}$$

assuming the cable of length l (in km) to be terminated by the characteristic impedance Z_w. The transmission constant γ consists of the real part $\alpha(f)$, the attenuation per length in dB/km, and the imaginary part $\beta(f)$, which is the corresponding phase per length in rad/km. The transmission constant $\gamma(f)$ and the characteristic impedance Z_w are the so-called secondary cable parameters.

The primary cable parameters are the resistance, conductance, capacitance, and inductance per length. There is a direct correspondence between these parameter sets. For determining the secondary parameters, we chose a method that is based on measuring the input impedances Z_{oc} and Z_{sc} for open and shorted output of the cable, respectively. From these we obtain

$$\begin{aligned} \gamma l &= \operatorname{atanh}\left(\sqrt{Z_{sc}/Z_{oc}}\cdot e^{jn\pi}\right) + jm\pi, \\ Z_w &= \sqrt{Z_{sc}\cdot Z_{oc}}\cdot e^{jn\pi} \end{aligned} \tag{8}$$

The n and m multiples of π represent the multiple solutions of the square-root and atanh functions. Choosing the right one is actually one of the critical aspects of cable measurements.

Practically, the measurement is carried out by using either an impedance analyzer or a network analyzer together with a reflection bridge. The use of a balun (*bal*anced/*un*balanced, basically a transformer for measurement purposes) is always recommended. For all the measurements regarding MIMO, a shielded layered cable of 250 m length with 20 twisted-pairs (type: F-02YHJA2Y / 20 × 2 × 0.4) was used. We will refer to this cable as layered cable from this point on. The secondary cable parameters of the layered cable are shown in Figs. 8 and 9 together with the so-called MAR model (Marconi) (*Musson, 1998*).

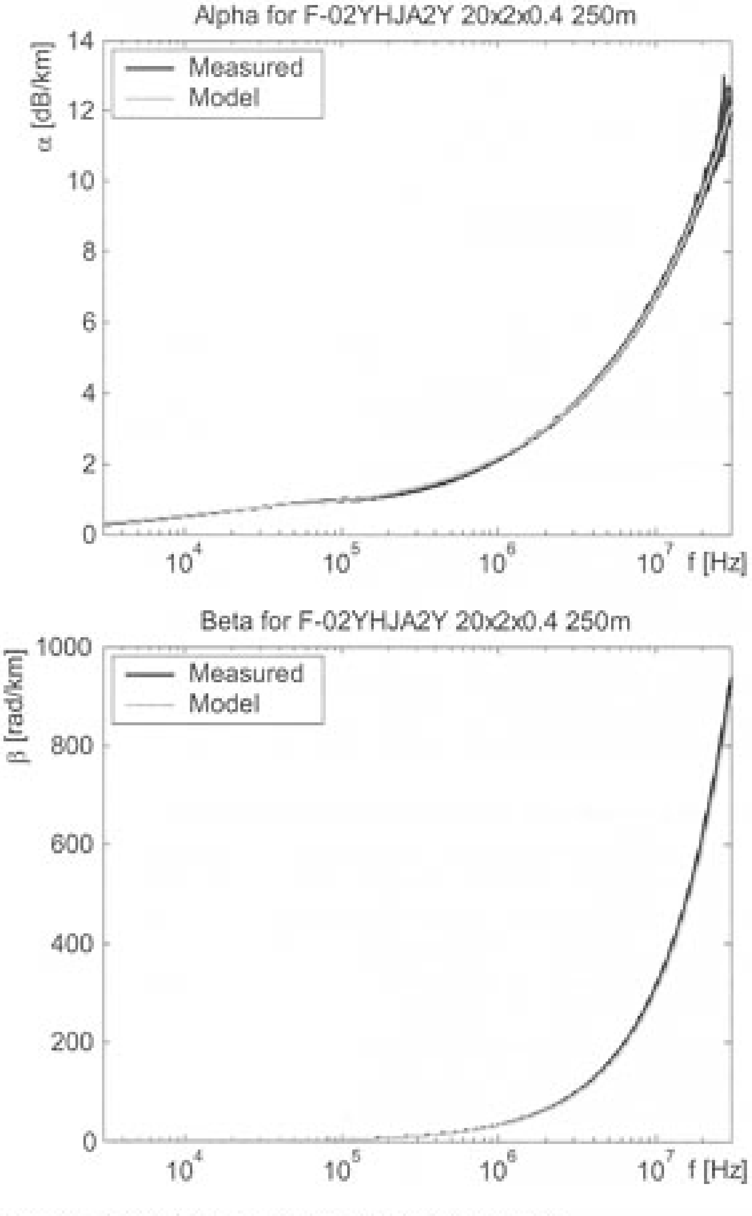

Fig. 8. Transmission constants of the layered cable

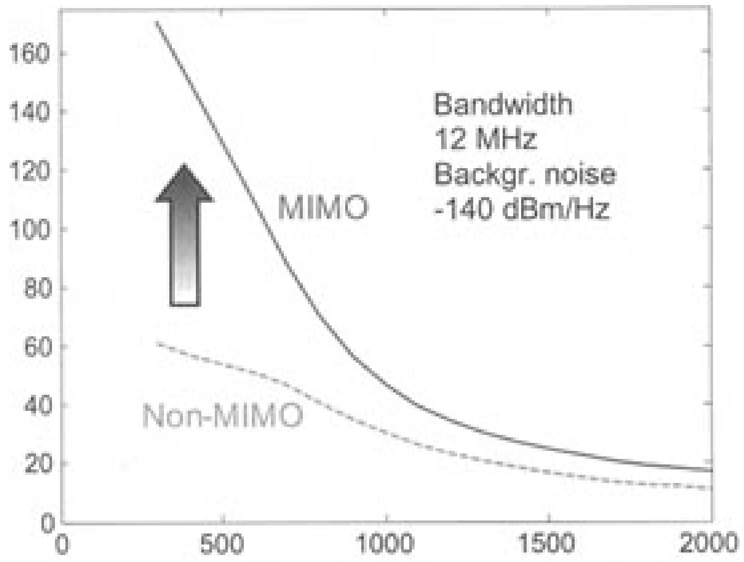

Fig. 3. Comparison of MIMO system with ordinary transmission (25 pairs; performance per pair; background noise − 140 dBm/Hz)

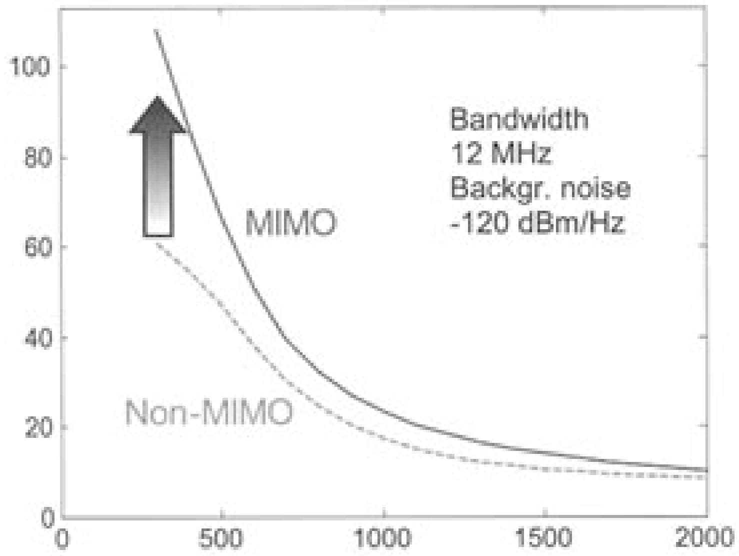

Fig. 4. Comparison of MIMO system with ordinary transmission (25 pairs; performance per pair; background noise − 120 dBm/Hz)

the performance and possible gains depend strongly on the actual background noise level. In the standardization of VDSL (Very High Bit-Rate DSL), a level of − 140 dBm/Hz had been assumed, but following the discussions in the standardization bodies, this level may be questionable and is not easy to measure in practice. For shorter loops, it is possible to achieve gains in transmission rate in the order of a factor of 2.

3. Narrowband interference mitigation

DSL systems transmit data in frequency bands of up to 12 MHz. Theoretically, a perfectly twisted cable does neither emit nor pick up radio waves. However, in reality, perfect twisting is not possible, thus the cable starts to act like an antenna. This unwanted behavior has two consequences as illustrated in Fig. 5: On the one hand, the cable receives radio signals (RFI ingress). The two main disturbers in that respect are broadcast radio stations and amateur radio (HAM) transmitters, which can produce strong and intermittent disturbance. On the other hand, the cable emits radio signals (RFI egress), which disturb wireless services like HAM transmitters. This led to severe restrictions concerning the modems' allowed transmit power in certain frequency bands in standardization (*ETSI TMS, ANSI T1E1.4*).

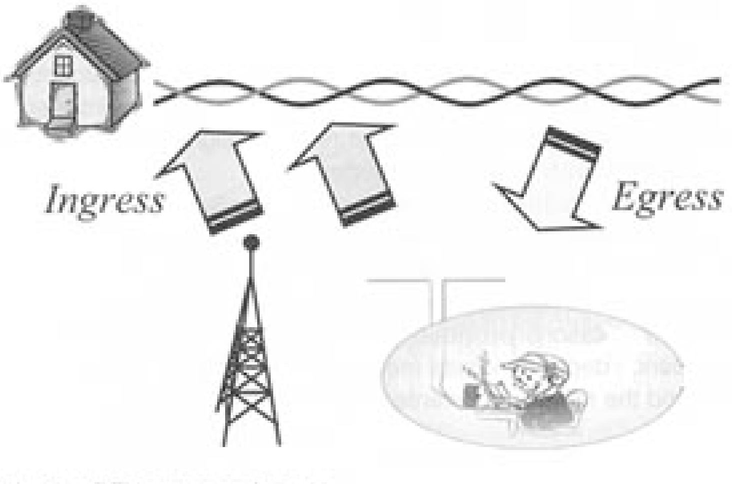

Fig. 5. RFI ingress and egress

The actual impact of RFI strongly depends on the distance from the disturber and the installation of the cable. The last meters to the customer are often aerial cables and in-house wiring is typically unshielded. Thus, this part of the access network is highly susceptible to RFI ingress. RFI may strongly affect transmission quality and could be a severe problem for cable infrastructures with a large percentage of overhead or above-ground cables, like in the United States or in the United Kingdom.

Several techniques exist to mitigate RFI ingress (*de Clercq et al., 2000*). For low ingress levels, counter-measures in the digital part of the modem are sufficient. If the interference levels are high enough to overload the analog-to-digital (AD) converter of the modem, mitigation techniques in the analog domain have to be employed. One method is cancellation based on the common-mode signal. Figure 6 shows the underlying coupling model. The differential-mode signal, which is the difference of the two voltages between each wire and ground, consists of the desired signal and an RFI component. The common-mode signal, i.e., the arithmetic mean of the two voltages between each wire and ground, also consists of both an RFI and a signal component. The cancellation principle is based on the fact that the RFI component in common-mode is much stronger than the signal. Thus, the common-mode signal may serve as a reference.

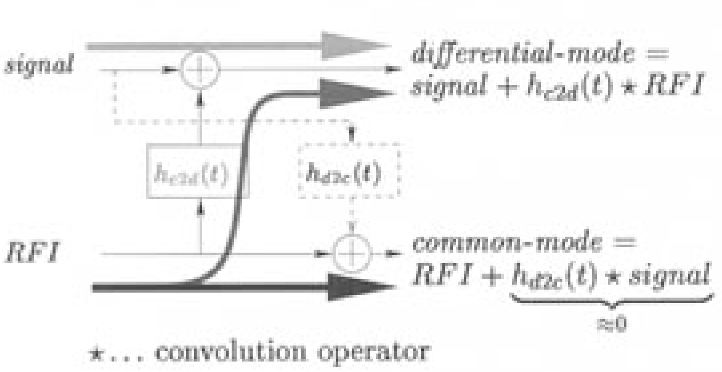

Fig. 6. Coupling model for reference-based RFI cancellation

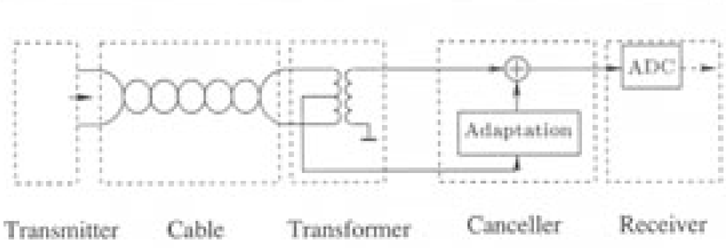

Fig. 7. Principle of reference-based RFI cancellation

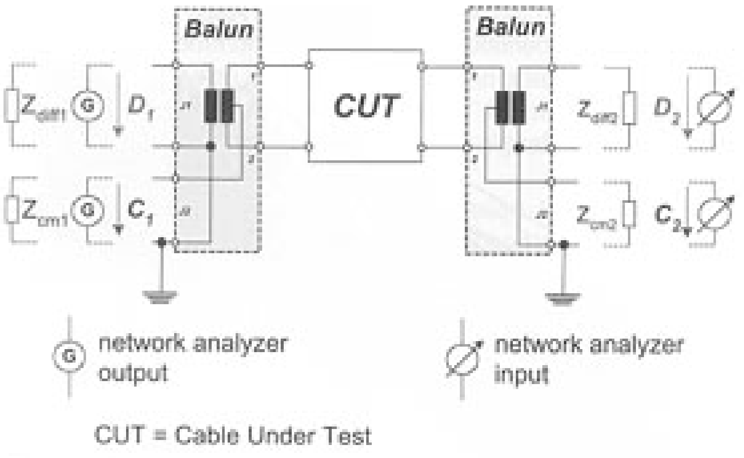

Fig. 12. Symmetry measurement setup

(2) In-house cable: unshielded, not twisted cable with 5 wires (DFYY / 5 × 0.5 EL), typically used for in-house wiring in Austria, length = 30 m, cable was arranged straightly to resemble the in-house installation case.

Figure 12 shows the measurement setup. Depending on the parameter that is actually measured, the four ports of the two baluns are connected to the output/input of the network analyzer or terminated by appropriate impedances. It is vital for balance measurements that all the ports are terminated properly.

Figure 13 shows the LCL curves for the 10 pairs of the in-house cable resulting from all combinations of the 5 wires. The symmetry performance of the measurement equipment itself plays an important role. Auxiliary connection equipment, like short pieces of wire, plugs and sockets, etc. must be avoided so that the residual leakage from one mode to the other is determined by the baluns only. The results depicted in Fig. 13 are valid for all the frequencies where the LCL of the two baluns exceeds the LCL of the cable, i.e., when the asymmetry is determined by the cable.

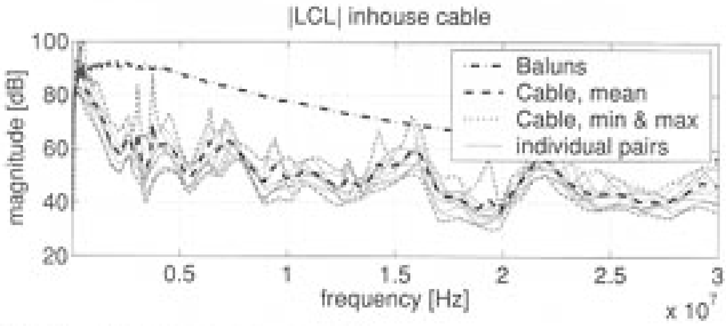

Fig. 13. LCL of the in-house cable

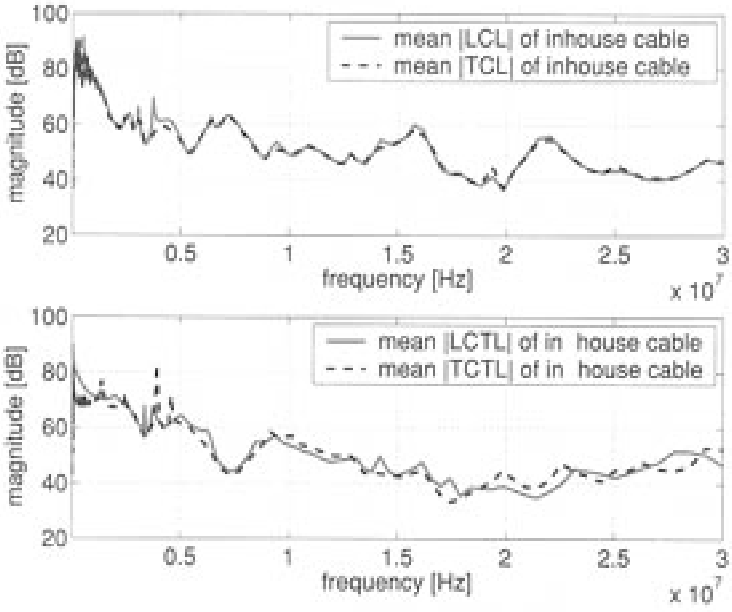

Fig. 14. Reciprocity of LCL/TCL and LCTL/TCTL

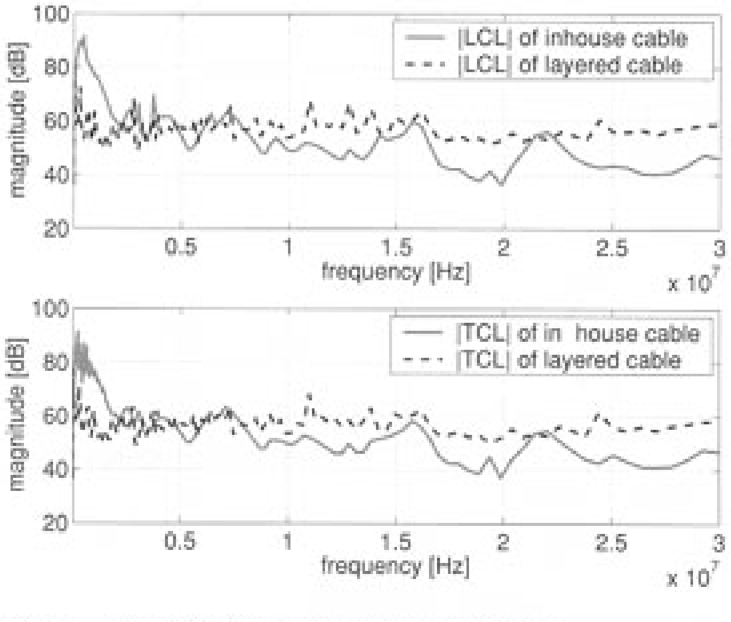

Fig. 15. LCL and TCL of in-house and layered cable

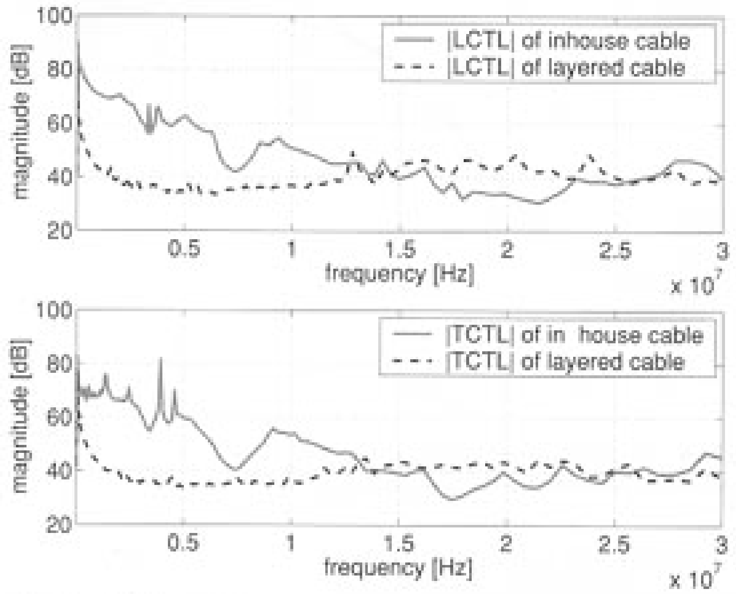

Fig. 16. LCTL and TCTL of in-house and layered cable

Figure 14 shows that the reciprocity assumption for LCL and TCL as well as LCTL and TCTL is reasonable.

The mean values of the four balance parameters of the two cables under consideration are compared in Figs. 15 and 16. As already mentioned before, TCTL and LCTL implicitly contain the loop attenuation. Thus, a normalization of these measures is necessary in order to compare cables of different length:

$$\{L,T\}CTL(f) = \{L,T\}CTL_{\text{measured}}(f)\cdot H_{\text{loop}}(f). \qquad (10)$$

Note that the symmetry measures describe the ratio of input signal to output signal, which is the inverse of a transfer function. Thus, the normalization in (10) is performed by a multiplication with the loop's transfer function. The layered cable shows worse symmetry properties compared to the in-house cable for low frequencies. This is a surprising result since the in-house cable is not twisted. However, it is much shorter and the wires are not squeezed as tightly compared to the layered cable, which preserves the symmetry a little more. For high frequencies the symmetry of the layered cable is slightly better, however, its shield makes it much less susceptible to RFI, anyway.

5. Summary and conclusions

Reliable performance assessment of techniques for future DSL systems requires detailed knowledge of the cable properties. Two topics have been discussed in here: MIMO transmission and RFI mitigation.

For MIMO transmission, the crosstalk behavior of the whole cable is of importance. Regarding RFI, the symmetry parameters of the cable are essential for coupling models. Measurement results of a layered cable and an in-house cable have been presented. The measurement results are the basis for studying emerging technologies, like MIMO transmission and RFI mitigation.

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