Interference between voice and data channels sharing the same unshielded twisted pair cable's sheath in structured cabling systems

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Abstract

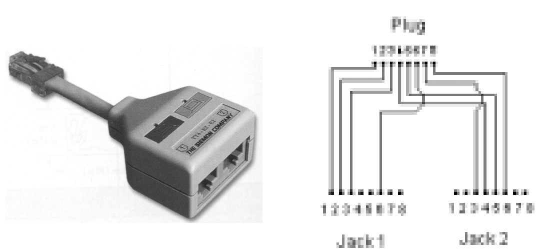
The goal of this paper is to study the interference between data and voice channels using different pairs inside the same four-pair UTP cable's sheath in structured cabling systems. Interference effects are evaluated theoretically from the channel model developed for this study which has been validated through field measurements in regular used structured cabling systems. The interference of the data channel over the voice channel is evaluated by means of the induced psophometric voltage through field measurements. Likewise the interference of the voice channel over the data channel is evaluated through the bit error rate (BER) in the data channel when the voice channel is active. More than just analyze the interference between data and voice channels this paper shows some interesting interference relations between analog and digital signals as well.

Keywords

Structured cabling systems, electromagnetic interference, psophometric voltage noise, bit error rate, crosstalk noise.

INTRODUCTION

The most popular adapter used in structured cabling systems is known as "Y" adapter. This adapter allows two different applications or services to be implemented in only one telecommunications outlet of a certain work area making possible to users to take more advantage of the system's flexibility as shown in Figure 1.



(a) (b)

Figure 1. 1(a) "Y" adapter, 1(b) "Y" adapter wiring scheme.

Some level of interference is then expected by using these adapters[1]. The most significant interference that can be found in cabling systems is the crosstalk in its two most concerning types: NEXT (Near End Crosstalk) and FEXT (Far End Crosstalk). Crosstalk occurs due to the inductive and capacitive couplings which is the most significant impairment that causes interference because it appears as errors on data and voice communications by introducing noise on the disturbed channel.

Impulse noise is also a significant transmission impairment specially in analog voice channels due to the ring signal used by analog telephony. When coupled by the data channel, impulse noise can cause ISI (Intersymbol Interference) by changing the phase of the digital signal and consequently changing the binary symbols (i.e. zero for one and vice-versa).

Thus, digital and analog signals are also corrupted by noise due to crosstalk, impulse noise, external voltages, phase distortion, return loss, and other types of interference that are also present in the input of the active equipment receiver. In the case of data communications the receiver must decide at each sampling interval if there is or there is not a valid pulse in its input. In digital communication systems, a small part of the digital data is not detected, or noise may be detected as valid data when data is not present. The ratio between the received bits in error and the actual bits sent in a communication system in a measurement interval is called bit error rate (BER). Several types of interference that are present in communications channels can contribute to increase the probability of bit error of the system (Figure 2).

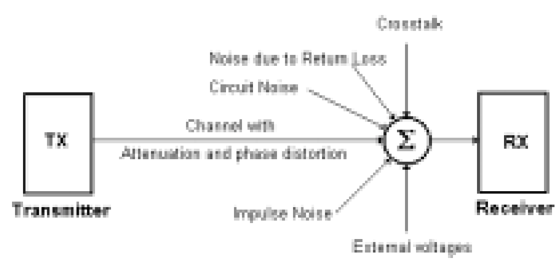


Figure 2. Sources of interference present in a telecommunication channel.

BER is the best indicator of the data communication systems performance. It can be reduced in a specific system by controlling the signal to noise ratio (SNR) at the input of the decision circuit of the receiver or by controlling the

[1] This is due to the use of unshielded cable for the construction of these adapters.

SNR of the channel itself. It means that the received signal level, the line attenuation, as well as the noise due to crosstalk coupling and external voltages must be kept within certain limits established by technical standards.

The focus of this paper is on the analysis of interference between data and voice signals in structured cabling systems. The noise voltage induced on voice channel (disturbed circuit) due to data propagation over an adjacent pair (disturbing circuit) appears as induced psophometric voltage that degrades the quality of voice communications. On the other hand, noise voltages induced on data channel (disturbed circuit) due to voice signal propagation over an adjacent pair (disturbing circuit) appear as BER that may impair data transmission on the data channel by the relations previously mentioned.

Thus, the analysis of the interference from the data channel over the voice channel in structured cabling systems is obtained by the relationship between crosstalk voltages and currents, and the psophometric voltage induced on the voice channel.

Similarly, the analysis of the interference from the voice channel over the data channel in cabling systems is obtained by the relation between voltages and currents induced on the data channel, and its bit error rate that is related here to the channel's ACR which is the parameter whose measurement performed for certification of structured cabling systems is closer to the SNR (Signal to Noise Ratio).

For a better understanding of the noise effects on a given digital communication system it is important to know some of its characteristics such as signal encoding as well as signal detection and decision mechanisms used by the receptor. The focus of this study is on the analysis of PCM (Pulse Code Modulation) signals - Manchester encoding which is the most common code used for Ethernet applications (e.g. 10BASE-T, 100BASE-TX). Even so, this analysis can also be applied to applications based on PAM (Pulse Amplitude Modulation) as in the case of Gigabit Ethernet (e.g. 1000BASE-T, 1000BASE-TX).

PCM communication systems performance is commonly analyzed through its BER which is the term frequently used in the technical jargon as mentioned earlier in this paper. The existence of a relation between the system bit error rate and its signal to noise ratio is depicted in Figure 3 [1].

However, signal to noise ratio is not a parameter whose measurement is specified in technical standards applicable to structured cabling systems certification. The equivalent parameter that can be observed in structured cabling systems that represents its signal to noise ratio is the ACR (Attenuation to Crosstalk Ratio). For this reason the analysis of data channel response for BER, considering PCM Manchester, was made by obtaining the relation between ACR and system bit error rate in structured cabling. In order to determine such a relation a theoretical channel model was developed using MATLAB and some testing setups were arranged for the measurement of the psophometric noise voltage induced on the voice channel as well as the ACR response of the data channel under the conditions of interest in this study.

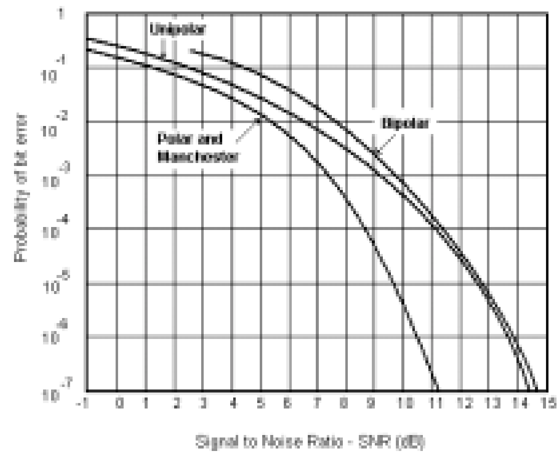


Figure 3. Probability of bit error as function of SNR

The theoretical model was then validated trough the field and laboratory measurements and safe limits were established for operation of voice and data applications in the presence of noise in structured cabling systems.

CHANNEL MODELING

Transmission lines or channels can be represented by distributed network parameters such as resistance, inductance, capacitance, and conductance per unit length. In this study a model, known as "T" model, is adopted to represent an equivalent circuit for a length, Δl, of the channel due to its demonstrated accuracy in a previous study [4]. The network parameters (R,L,G and C) are assumed to be uniformly distributed over the channel length. The model becomes as accurate as Δl approaches zero. In practice a differential model is adopted [5].

Copper cable channels can be approached as a cascade of symmetrical T sections mathematically represented by a two-port transmission matrix whose analytical parameters are classical and well-known or by a set of differential equations obtained by assuming a differential length of line, calculating the voltage and the current at the input and output of the cable section for Δl approaching zero. Classical results obtained from this analysis can be seen below.

$$V_l = C_1 e^{\gamma l} + C_2 e^{-\gamma l} \quad (1)$$

$$I_l = C_1 Y_0 e^{\gamma l} - C_2 Y_0 e^{-\gamma l} \quad (2)$$

Where V_l and I_l are the voltage and the current along the transmission line respectively. C_1 and C_2 are determined by boundary conditions at the source and line's termination. Y_0 is the characteristic admittance and γ is the propagation constant of the channel related as follows

$$Y_0 = \sqrt{\frac{Z}{Y}} \quad (3)$$

$$\gamma = \sqrt{ZY} = \alpha + j\beta \quad (4)$$

where α is the attenuation constant and β is the phase constant, with

$$Z = R + j\omega L \quad (5)$$

$$Y = G + j\omega C \quad (6)$$

The T-model adopted for this study can be seen in Figure 4.

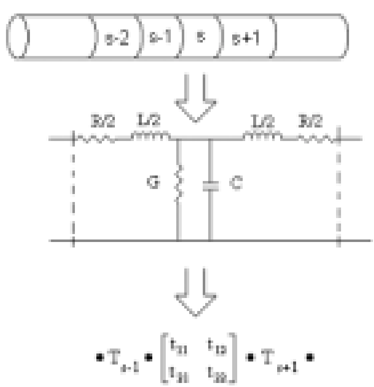


Figure 4. T model adopted in this study

It were included in the model the skin effect, capacitance unbalance (pair-to-pair and pair-to-ground), additive white gaussian noise (thermal noise), psophometric noise voltage, attenuation, near end crosstalk, far end crosstalk, residual NEXT, attenuation to crosstalk ratio, and bit error rate for PCM codes[4].

CROSSTALK MODELING

In this study crosstalk coupling mechanisms (near end crosstalk and far end crosstalk) were modeled according to the classical Campbell's formula [2]. This formula was initially developed for the analysis of electrically short parallel conductors' elements with equal lengths and terminated by loads with equal characteristic impedances. The scheme depicted in Figure 5 can be used for deducting the Campbell's formula.

The voltage source, V_1, creates a current $2i_c$ through the capacitance unbalance C_{ub} (pair-to-pair) and a current i_c flows to both ends of the circuit 2 (disturbed pair). Current I_1 is responsible for inducing a voltage in the disturbed pair due to the mutual inductance L_M and a loop current i_l flows through circuit 2 (notice the direction of i_l in both ends), that can be expressed as

$$i_l = \frac{v_N}{2Z_0} \quad (7)$$

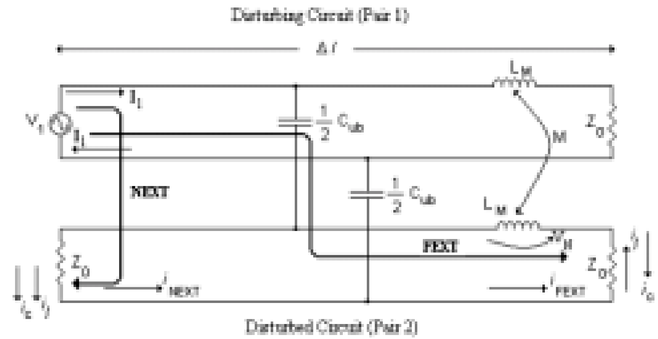


Figure 5. Crosstalk coupling mechanisms (NEXT and FEXT)

V_N, which is the noise voltage induced in the disturbed pair due to I_1 (in the disturbing pair) can be defined as follows

$$V_N = j\omega L_M I_1 \quad (8)$$

Thus the crosstalk current due to the inductive coupling can be re-written as

$$i_l = j\frac{\omega L_M J_1}{2Z_0} \quad (9)$$

By measuring the induced signal at the near-end, one can notice that the polarity of the inductive and capacitive couplings are coincident. Actually the phases of i_c and i_l can oppose to each other or not. This behavior shows that if the effects of the inductive coupling are added to the effects of the capacitive coupling for NEXT, they will be subtracted for FEXT or vice-versa. Thus crosstalk currents i_c and i_l tend to add in the near-end (taking the end where the noise source, V_1, is placed as reference) and subtract in the far-end.

Then, Campbell's formula can be written as

$$\frac{I_2}{I_1} = \left[\frac{j\omega C_{ub} Z_0}{8} \pm \frac{j\omega L_M}{2Z_0}\right] \quad (10)$$

Where,

$\omega = 2\pi f$

f is the operation frequency (Hz)

L_M is the mutual inductance (H)

Z_0 is the characteristic impedance (Ω).

Still in terms of relations between currents, crosstalk coupling can be written through the Campbell's formula, in dB, as

$$C = 20\log\frac{I_2}{I_1} = 20\log\left(\frac{j\omega C_{ub} Z_0}{8} \pm \frac{j\omega L_M}{2Z_0}\right)(\text{dB}) \quad (11)$$

Thus, from the expression (11) we can express NEXT and FEXT couplings as follows

$$NEXT = 20\log\left(\frac{j\omega C_{ub} Z_0}{8} + \frac{j\omega L_M}{2Z_0}\right)(\text{dB}) \quad (12)$$

$$FEXT = 20\log\left(\frac{j\omega C_{ub}Z_0}{8} - \frac{j\omega L_M}{2Z_0}\right)(\text{dB}) \qquad (13)$$

Equation (13) can be applied to pairs inside the same cable's sheath whose separations are minimal. For pairs whose separations are higher (e.g. between pairs of different cables in a bundle) the capacitive coupling (pair-to-pair) is negligible and the first term of equation (13) can be neglected. Now the equation becomes

$$FEXT_d = 20\log\left(\frac{j\omega L_M}{2Z_0}\right)(\text{dB}) \qquad (14)$$

where the index d is used to indicate "distant pairs".

As the Campbell's formula was developed to analyze cables electrically short, it's necessary to adjust it so the formula can be applied to long cable lengths. The following expression can be used to determine the NEXT coupling under this condition based on Campbell's formula

$$NEXT = 20\log\left(\frac{j\omega C_{ub}Z_0}{8} + \frac{j\omega L_M}{2Z_0}\right)\cdot \frac{1}{\sqrt{n}}\cdot\sqrt{1-\frac{e^{-4\alpha l}}{4\alpha N}}\ (\text{dB}) \qquad (15)$$

where,

n is the length of a T-section, in meters (m)

α is the propagation constant, in dB/m

l is the total cable length, in meters (m)

N is the number of T-sections cascaded to model a cable of length l. By following the same behaviours, far end crosstalk coupling for a cable with length l will be

$$FEXT = 20\log\left(\frac{j\omega C_{ub}Z_0}{8} - \frac{j\omega L_M}{2Z_0}\right)\cdot \frac{1}{\sqrt{n}}\cdot\sqrt{1-\frac{e^{-4\alpha l}}{4\alpha N}}\ (\text{dB}) \qquad (16)$$

Different performance levels for NEXT and FEXT in structured cabling systems occur due to the nature of FEXT, which is the addition of crosstalk along the total cable run. For NEXT the dependence of the cable run length is negligible. The best strategy to achieve closer performance levels for NEXT and FEXT is by obtaining better possible inductive coupling characteristics in the manufacturing process of UTP cables as well as connecting hardware. The capacitive coupling can be easily kept within stable values for the whole frequency operation range.

TESTING METHODOLOGY AND RESULTS

For the measurements of the interference from the data channel over the voice channel the test setup depicted in Figure 6 was used.

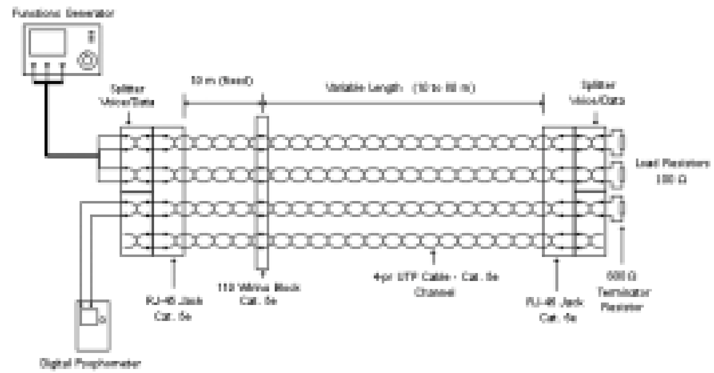


Figure 6. Testing setup for the measurement of the interference from the data channel over the voice channel

The channel was implemented to allow the variation of its length from 10 to 100 meters in steps of 10 m. To facilitate this behavior a 110-style wiring and connecting block was used in the telecommunications outlet-end where the work area splitter was installed. In a sequence, a function generator and a personal computer connected to a Fast Ethernet switch were used as the interference sources. A digital psophometer was used to measure the coupled noise on the voice channel while the data channel was active. The voice channel was terminated at its far-end by a 600-ohm resistor to make the psophometric measurements feasible. The functions generator was used to transmit random sequences of pulses from 100 kHz to 100MHz through the data channel that under this condition was terminated by a 100-ohm resistor as load.

Figure 7 shows the response curve of the voltage coupled by the voice channel due to the data channel active.

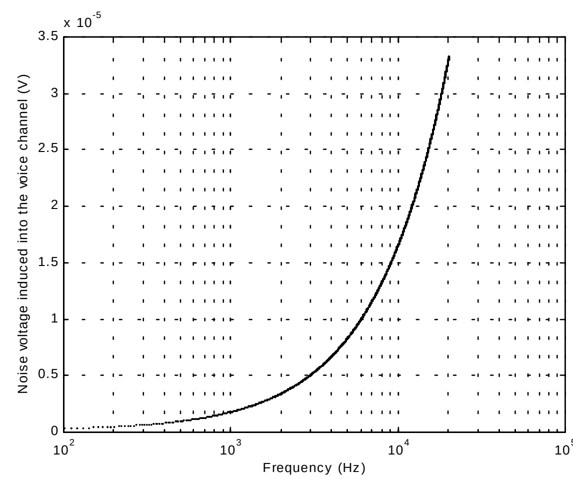


Figure 7. Noise voltage induced into the voice channel due to the data channel active for a length of 100m

Table 1 shows the psophometric noise levels coupled by the voice channel due to the data channel active.

Table 1. Psophometric noise level measured in the voice channel

Channel length (m)	Induced psophometric noise voltage (dBm)/(μV rms)
10	- 100.1 / 7.63
20	- 97.5 / 10.33
30	- 95.9 / 12.40
40	- 94.8 / 14.14
50	- 93.8 / 15.67
60	- 93.1 / 17.06
70	- 92.5 / 18.34
80	- 92.0 / 19.53
90	- 91.5 / 20.65
100	- 91.0 / 21.71

For the measurements of the interference from the voice channel over the data channel the test setup depicted in the Figure 8 was deployed.

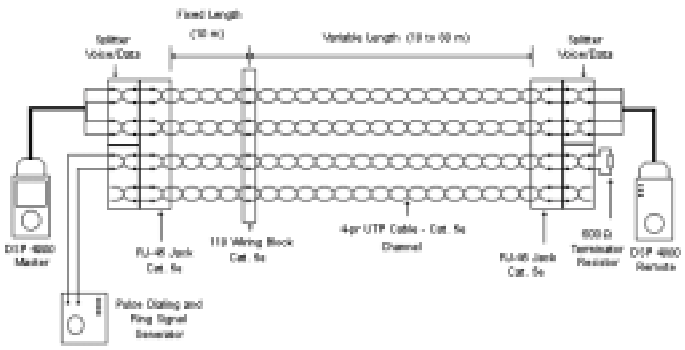


Figure 8. Testing setup for the measurement of the interference from the voice channel over the data channel

In the test setup shown in Figure 8 a pulse and ring generator was used to simulate activity in the analog voice channel (pulse dialing and ring signal as well). The ring signal has a peak voltage of approximately 150 V with a frequency of 25 Hz. The pulse created in a process of pulse dialing has a duration of 33.33 ms (30 Hz) to place the pulses on the line with a separation of 66.66 ms between each pulse. The peak voltage of those pulses are approximately 70 V.

A cablemeter was connected to the data channel in order to measure the variation of the ACR on it due to the activity of the voice channel as described above. The tests were performed for channel lengths varying from 20 to 100 m in steps of 20 m.

Table 2 presents the testing results for ACR and the corresponding calculated bit error rates for PCM symbols, Manchester encoding.

Table 3 presents a comparison between the values of ACR from the model and those obtained by the measurements.

Table 4 shows the bit error rate as function of attenuation to crosstalk ratio for PCM codes.

Table 2. Measured ACR values as function of the activity in the voice channel and the corresponding BER (calculated)

Channel Length (m)	ACR for Pulse Dialing (dB)	BER (Pulse Dialing)	ACR for Signal Ring (dB)	BER (Signal Ring)	Freq. (MHz)
20	46.5	2.10^{-42}	41.4	7.10^{-38}	96.75
40	41.3	8.10^{-38}	41.3	8.10^{-38}	91.75
60	38.1	5.10^{-35}	38.0	6.10^{-35}	99.25
80	28.1	3.10^{-26}	28.1	3.10^{-26}	96.75
100	26.8	2.10^{-25}	23.9	1.10^{-22}	92.75

Table 3. Comparison between ACR values obtained from the model and measured values with voice channel active (for pulse dialing and ring signal)

Channel Length (m)	ACR for Pulse Dialing – Model (dB)	ACR for Pulse Dialing – Measured (dB)	ACR for Ring Signal – Model (dB)	ACR for Ring Signal– Measured (dB)	Freq. (MHz)
20	49.16	46.5	42.75	41.4	96.75
40	42.96	41.3	36.59	39.3	91.75
60	37.57	38.1	31.22	33.9	99.25
80	33.17	28.1	26.79	27.9	96.75
100	29.23	26.8	22.86	23.9	92.75

Table 4. Bit error rate as function of the attenuation to crosstalk ratio for PCM codes

Line Code (Signal Encoding)	Bit Error Rate (Probability of Bit Error)
NRZ or Unipolar RZ	$erfc\left(\sqrt{ACR}\right)$
NRZ or Polar RZ	$erfc\left(\sqrt{2ACR}\right)$
NRZ or Bipolar RZ	$\frac{3}{2}erfc\left(\sqrt{ACR}\right)$
NRZ Manchester or RZ Manchester	$erfc\left(\sqrt{2ACR}\right)$

Figure 9 shows the measured ACR response curve in the data channel with the voice channel active (for ring signal and pulse dialing as well).

All results presented on Tables 1 through 3 are based on the worst case values, which are, the very worst values obtained for the worst pair of the four pairs of the UTP category 5e channel under test. The frequencies corresponding to each of those values were also registered and reported.

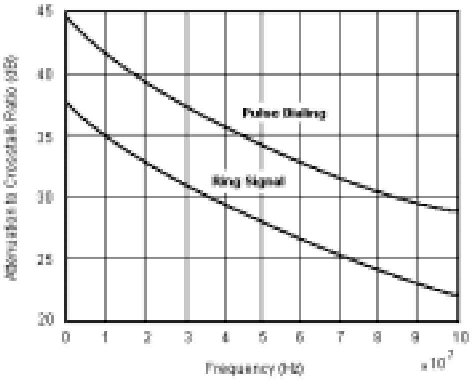


Figure 9. Measured ACR in the data channel due to the voice channel active for the conditions of pulse dialing and ring signal

OTHER RESULTS

Table 5 shows the ACR data channel response (worst case) for some impulse noise generated and applied in an adjacent pair inside the same cable's sheath, and the associated BER for each impulse magnitude.

Table 5. ACR and BER as function of the amplitude of impulse noise applied in an adjacent pair

Impulse Noise Amplitude (V_{peak})	ACR (dB)	Bit Error Rate
36	35	10^{-32}
75	29	10^{-27}
150	22	10^{-21}
300	15	10^{-15}
700	9	10^{-9}
980	6	10^{-6}
1500	2	10^{-3}

Table 6 shows a comparison between ACR for category 5e and category 6 channels according to the ANSI/TIA/EIA-568-B.2 and '568-B.2-1[2].

Table 6. Comparison between ACR response for categories 5e and 6 channels

Frequency (MHz)	ACR (dB) Cat. 5e channel config.	ACR (dB) Cat. 6 channel config.
1.0	-	62.9
16.0	40.5	45.2
25.0	34.8	39.9
31.25	32.3	37.0
62.5	26.4	26.9
100.0	12.9	18.6
200.0	-	3.3
250.0	-	-2.8

CONCLUSIONS

Testing setup and configuration adopted in this study have demonstrated satisfactory levels of accuracy and allowed validating the theoretical model as well considering the results shown in the previous sections.

Through the mentioned results (practical and theoretical) we can conclude that the level of interference between analog and digital signals propagating over different pairs inside the same four-pair UTP cable's sheath (under the conditions considered here) is not important and can be neglected for the majority of applications currently in use in structured cabling systems. Thus the operation of analog telephony, low-voltage circuits and devices used in building automation systems, Ethernet, Fast Ethernet, and even Gigabit Ethernet applications can be implemented in conjunction (sharing the same cable or adjacent cables in a bundle) in structured cabling systems successfully.

In addition, by analyzing Table 6 we can also guarantee lower interference levels for Category 6 cabling systems (given their ACR response) which offer positive margins allowing the implementation of future applications under the conditions considered in this study . Also the frequency range is much higher than that for Category 5e cabling systems.

So, as Category 6 cabling system transmission characteristics are better than those of Category 5e systems, the interference levels will be lower and its ACR and BER responses will still be better.

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