DesignCon 2010

Compliance Testing of Passive Interconnects

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Abstract

Many standardization committees such as SAS, SFF, OIF, IEEE, IBTA, etc. have defined or are in the process of defining informative and/or normative specifications for passive channels operating at multiple gigabit speeds. This paper provides an overview of the different channel performance parameters and compliance testing methods that are applied today in the industry. A new compliance parameter, the total integrated noise is introduced. This parameter includes all interconnect noise sources and makes a separation of the interconnect noise budget in a crosstalk, multiple reflection, inter symbolic interference, transmitter and receiver impedance mismatch and mode conversion budget no longer necessary.

Authors biographies

Stefaan Sercu was born in Ieper, Belgium, on February 6, 1969. He received the degree in electrical engineering from the University of Ghent in 1992. From 1992 to 1998, he worked as a research assistant at the Department of Information Technology (INTEC) of the University of Ghent. His research concentrated on the characterization and modeling of high-speed connectors and interconnections. Since 1998 he works for FCI Electronics in s'-Hertogenbosch, The Netherlands. In 2002 he received the PhD degree in electrical engineering from the University of Ghent. Currently he is leading the Signal Integrity R&D team for the FCI Electronics Division. He has authored and co-authored over 25 technical papers in international journals and international conference proceedings.

Vittal Balasabrumanian received his B.S. in Electrical Engineering from Delhi College of Engineering, University of Delhi in 2001 and his M. Eng. in Electrical Engineering from Penn State University in 2005. He was awarded the Doris Hughes Memorial Award at Penn State in recognition of his outstanding academic achievements and contribution to the college community. He has been working at FCI USA, Inc. since Jan 2005 where his responsibilities include the design and analysis of high-speed connectors and cable assemblies. He also worked for Sapient Corporation Pvt. Ltd. from August 2001 to December 2002 as a technology associate and helped with software consulting. He is a member of the Institute of Electrical and Electronics Engineers since 1998.

Jan De Geest was born in Gent, Belgium on July 30, 1971. He received the degree in electrical engineering from the University of Ghent, Belgium in 1994 and the degree in supplementary studies in aerospace techniques from the University of Brussels, Belgium in 1995. From September 1995 to December 1999 he worked as a research assistant at the Department of Information Technology (INTEC) of the University of Ghent, where he received the PhD degree in electrical engineering in 2000. Since January 2000 he has been working for FCI ELX in 's-Hertogenbosch, The Netherlands. His work focuses on the design, modeling and optimization of high-speed connectors and interconnection links. He has authored and co-authored over 25 technical papers in international journals and international conference proceedings.

Stephen B. Smith's current responsibilities at FCI include primarily customer support in the application of high-speed connectors by working very closely with Marketing and Sales. Prior to coming to FCI in 2000, Smith worked at AMP Incorporated (now Tyco Electronics) for 10 years as a development engineer spending most of that time in the electromagnetics research group where he developed methods of modeling and simulating interconnection systems on projects spanning the frequency spectrum from power-frequency high-current utility connectors to high-speed and r.f. interconnects. Prior to that, he worked in acoustical research at Masland Industries (now Lear Corporation.) He has taught various courses at each of his jobs, and in the last couple of years, he has presented papers at various conferences including the IEEE Holm Conference. Smith has a B.S. in physics and an M.S.E.E., both from Penn State University.

1 Introduction

Today, a lot of standardization committees are in the progress of defining requirements for next generation high speed applications. IEEE802.3ba is specifying requirements for multi-channel 10 Gb/s propagation over both backpanel and cable assemblies. OIF is defining requirements for 25 Gb/s signal propagation over short or long range (backpanel). SAS is moving towards 8 Gb/s. Fibre channel is looking at 16 Gb/s.

The objective set by each standardization committee is to define a set of requirements that needs to guarantee an error free operation of the interconnect application to enable interoperability within an architecture across a wide range of products offered by a host of vendors.

This task is not easy to achieve. Not only a set of performance parameters need to be selected that describes the performance of the passive interconnect, also for each performance parameter a trade-off (Go – No Go) acceptance level must be defined. Only when all acceptance criteria are passed, the passive interconnection link is qualified, otherwise it is rejected.

When testing an interconnect against above set of acceptance criteria, there are 2 cases that must be avoided (see Figure 1)

1) Interconnect fails the compliance testing, however system works without problems: interconnects that could be used in the system are rejected.
2) Interconnect passes the compliance testing, but the system fails: interconnects that cause errors, are accepted.

Therefore sufficient care must be taken that no cost effective solutions are rejected because of failing the criteria but are demonstrated to be working in practice.

As an example this may result in a backplane interconnect with long traces failing the compliance testing when measured with standard PCB material, passes with higher cost HS material, but actually works in practice.

Accepting interconnects that create errors must be avoided. This is typically done by setting good safety margins when defining acceptance criteria. Finding the right balance between the safety margins on one hand and making the system not too complex/expensive is what makes the work difficult.

Another way to deal with interconnects that fail compliance but can be used for considered applications, is by defining the interconnect requirements as informative or guidelines, rather than normative.

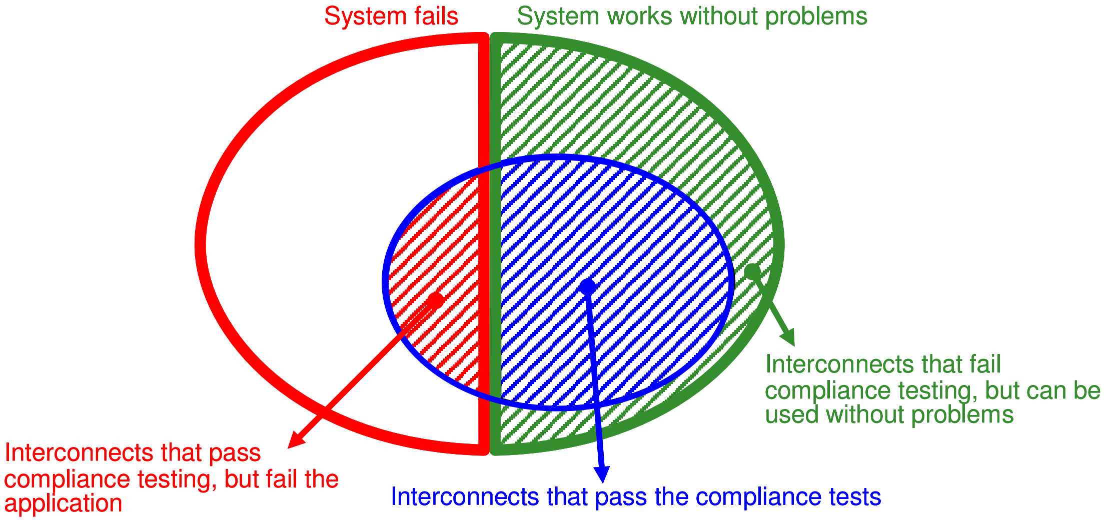


Figure 1: Classification of interconnects

Our objective is to define requirements, that address the above issues, and associated methods for compliance testing.

There is no unique or optimal way to divide the total interconnect loss, jitter or noise budget among the different interconnect performance parameters such as loss, crosstalk, or impedance mismatches. In reality, the different performance parameters are linked with each other. A long link with high losses will generally accept less crosstalk than a short link with lower loss.

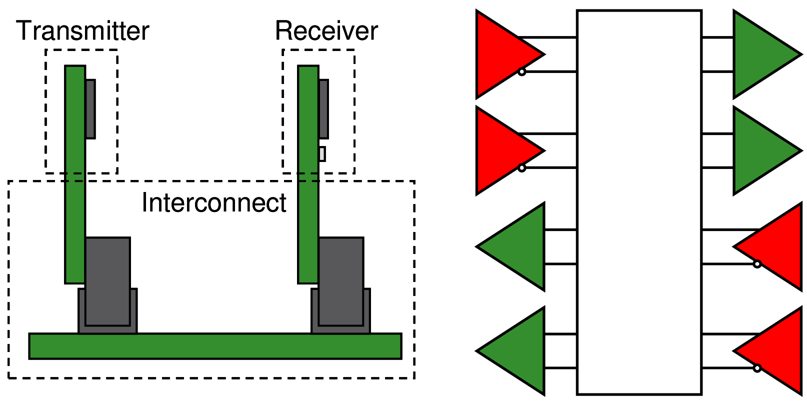
Also the introduction of signal conditioning in transmitters and receivers does not make the problem easier. Traditional test methods such as an eye pattern measurement are becoming no longer adequate, therefore new performance parameters and associated test methods may need to be defined.

Also we must ensure that the test environment represents the real application. Test points need to be defined carefully, compliance boards be specified and better de-embedding methodologies be defined.

An overview of the different passive interconnect parameters that contribute to the total interconnect performance will be presented. An approach is defined to combine the different performance parameters into a new parameter that presents opportunities for a better definition of link qualification criteria: the Total Integrated Noise.

2 Interconnect performance parameters

A system consists typically of three parts: one or multiple transmitters, the passive interconnect, and one or multiple receivers. The transmitter or receiver package and decoupling capacitors that might be needed in the system are typically considered to be part of the transmitter and/or receiver and are not part of the passive interconnect. The passive interconnect typically consists of stripline and/or microstrip traces on PCB's, via holes, connectors and cable assemblies. A typical backpanel system is shown on Figure 2.



(a) *(b)*

Figure 2: a) Typical backpanel system, b) schematic representation

The performance of the interconnect is determined by the signal that arrives at the receiver. This signal consists of two parts: the useful signal and the noise.

Received signal = useful signal + noise

The useful signal is the signal that contains the information that will be detected/decoded by the receiver. Noise is all unwanted signals that are superimposed on the useful signal and that makes the detection of the useful signal difficult. Noise should be as small as possible.

Depending on the physical source of the noise, the noise can be classified into

- Multiple reflections noise
- Inter symbolic interference noise
- Differential near end crosstalk or noise
- Differential far end crosstalk or noise
- Transmitter and receiver mismatch noise
- Common mode to differential mode transmission noise
- Common mode to differential mode crosstalk or noise

A differential signal that propagates through a channel of an interconnect will be attenuated by this interconnect before it arrives at the receiver and due to dispersion, a single bit will be spread over multiple bits. The signal that arrives at the receiver can be divided into *the useful signal* and *inter symbolic interference noise*. The propagated signal will not only be attenuated by the interconnect. Each time an impedance mismatch or discontinuity is encountered, part of the signal will be reflected. This reflected signal on his turn will be re-reflected and will finally arrive as differential noise at the receiver. This noise is called the *multiple reflection noise*. Part of the signal that is reflected by the interconnect arrives at the receiver. If the receiver is not perfectly matched, this signal will be reflected at the receiver, propagates over the interconnect and arrives as *transmitter mismatch noise* at the receiver. The useful signal that arrives at the receiver will also be reflected by the receiver if the receiver is not perfectly matched. This reflected signal will be re-reflected by the interconnect and arrives as *receiver mismatch* *noise* at the receiver. As transmitters are not perfect components, a transmitter will never transmit a perfect differential signal. Due to differences in rise time, amplitude, delay between positive and negative part of the differential signal, the transmitted signal also contains a common mode component. This common mode component propagates over the interconnect and due to asymmetries in the interconnect, this common mode component will be transformed into a differential signal. This differential signal is the *common mode to differential mode transmission noise.*

Near end and far end crosstalk are well known. Assume that a second channel is in close proximity of the first channel and that a signal propagates over this second channel. The signal that propagates over this second channel has a differential and a common mode component. Due to the close proximity of this second channel, a signal that propagates over this second channel will couple to the first channel. This coupled signal arrives as crosstalk noise at the receiver. The differential noise coming from the differential component is called *the differential near* end or far end crosstalk or noise. The differential noise coming from the common mode component is the *common mode to differential mode crosstalk or noise*.

A compliance specification has typically one or more performance parameters that describe and control the noise and useful signal of the interconnect , making sure that the receiver is capable of detecting the transmitted signal without any problem.

In the next paragraph these performance parameters are described more in detail.

3 Interconnect characterization methods

3.1 Transmission or useful signal parameters

3.1.1 Insertion loss and fitted attenuation

A first well known parameter that is used to characterize interconnects is the differential insertion loss or the S_{DD21} parameter expressed in dB. For uniform interconnects, insertion loss is identical to the attenuation of the interconnect. For non uniform interconnect, insertion loss is equal to the attenuation + multiple reflections.

$$S_{DD21}(f) = A(f) + MR(f)$$

$$IL(f) = 20.\log_{10}|S_{DD21}(f)|$$

with A(f) = attenuation and MR(f) = multiple reflections of the interconnect

Since IL contains multiple reflections, it is not the most optimal performance parameter for the useful signal. It might be that IL of an interconnect fails marginally a predefined limit due to the multiple reflections but that the interconnect can be used without any problem in the application. To overcome this problem, in some standards, insertion loss has been replaced by a fitted attenuation and insertion loss deviation requirement.

Fitted attenuation is a good approximation for the attenuation of an interconnect. It is calculated as the least mean square line fit to the insertion loss computed over a predefined frequency range.

$$\text{Attenuation}_{\text{fitted}} = Af + b$$
$$= IL_{avg} + \frac{\sum_n (f_n - f_{avg})(IL(f_n) - IL_{avg})}{\sum_n (f_n - f_{avg})^2}(f - f_{avg})$$

With $IL_{avg} = \frac{1}{n}\sum_n IL(f_n)$ and $f_{avg} = \frac{1}{n}\sum_n f_n$

Figure 3 shows the IL, and fitted attenuation of a typical backpanel link.

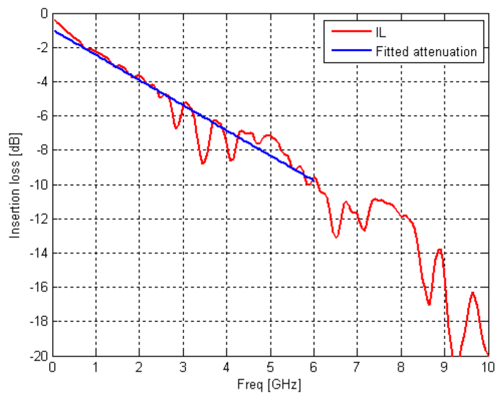


Figure 3: IL, and fitted attenuation of a typical backpanel link

As the losses of an interconnect consist of conductor ($\sim\sqrt{f}$) and dielectric losses (~f), in some standards the fitted attenuation is defined as

$$\text{Attenuation}_{\text{fitted}} = a + b\sqrt{f} + cf$$

or even as

$$\text{Attenuation}_{\text{fitted}} = b\sqrt{f} + cf + df^2$$

If IL fails marginally the compliance limits, then quite often the fitted attenuation does pass the compliance limits.

3.1.2 Eye Pattern, Deterministic Jitter

A second transmitter parameter that is often used in standards is the eye pattern. The measurement set-up of Figure 4 is used to measure the eye pattern. A differential pulse train generated by a pattern generator is transmitted through an interconnect. At receive side a sampling scope detects all received bits and puts them on top of each other. As a result, an eye pattern is obtained. Compliance testing with an eye pattern is quite simple. A mask is defined that must fit in the eye. If the mask doesn't fit then the interconnect fails. Important is that the impact of the test environment is minimized or taken into account in the specification. Especially the jitter of the pattern generator must be considered. Therefore not only a receiver mask is specified, but quite often also a transmitter mask is specified. As for IL, impact of multiple reflections and inter symbolic interference noise are included in the measurement results.

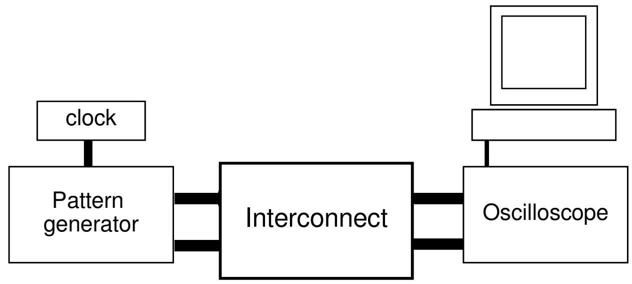
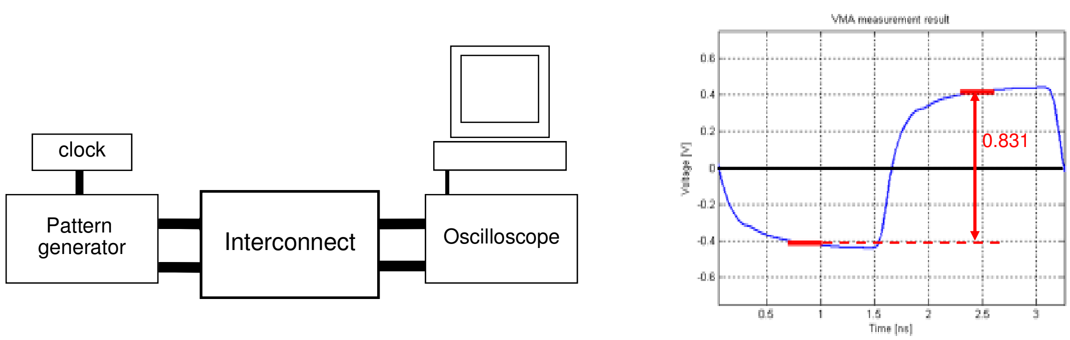


Figure 4: (a) Measurement set-up to measure eye pattern

Problem with an eye pattern is that if losses are too high the eye pattern will be closed. This is especially a problem if receivers with signal conditioning are used to compensate for the losses of the interconnect.

3.1.3 Voltage Modulation Amplitude (VMA) and VMA loss

Last transmission parameters that are considered are VMA and VMA Loss. Voltage Modulation Amplitude or VMA is the difference between the nominal one and zero of an electrical signal. To measure the VMA, measurement set-up of Figure 5 is used. A bitrate is selected and a pulse train consisting of 8 consecutive ones and zeros is transmitted through an interconnect and measured on a scope. The measured pulse train is divided into two equal time intervals (8UI long) aligned to the average time of both edges. The average voltage level in the central 20% of each time interval is measured. The difference between the one level and the zero level is the VMA. The VMA measurement is very useful to determine the risetime of a signal. To determine the 10-90% or 20-80% risetime of a pulse, the amplitude needs to be known. The VMA is a unique way to do this.



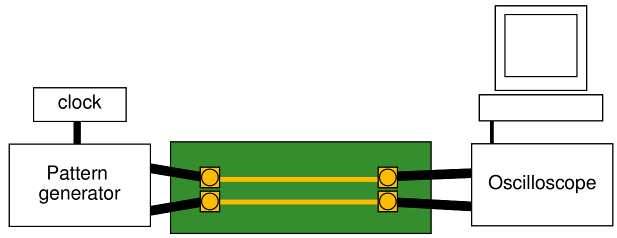
(a) *(b)*

Figure 5: (a) VMA measurement set-up, (b) example of a VMA measurement

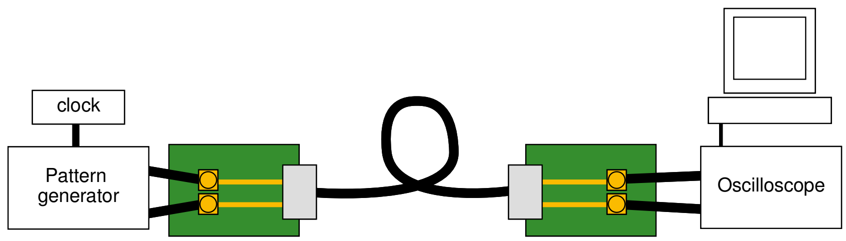
VMA loss of an interconnect is the difference between the VMA at the output of a calibration trace (VMA_i) and at the output of an interconnect (VMA_o). The calibration trace can be a perfect thru connection.

$$L = 20\log_{10}\left(\frac{VMA_i}{VMA_o}\right)$$

Figure 6 shows a typical VMA measurement set-up and measurement results. Figure (a) shows the calibration trace measurement set-up. Figure (b) shows the measurement set-up for a cable assembly. Figures (c) and (d) show the VMA measurement results for a calibration trace and a 5m SFP+ cable assembly. The selected bitrate is 10 Gb/s. The VMA loss of the cable assembly for a 10 Gb/s bitrate is 2.87 dB.



(a)



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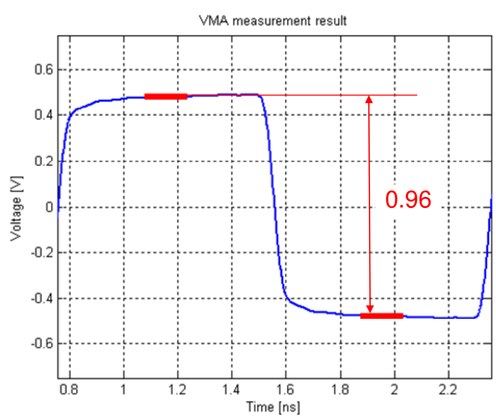

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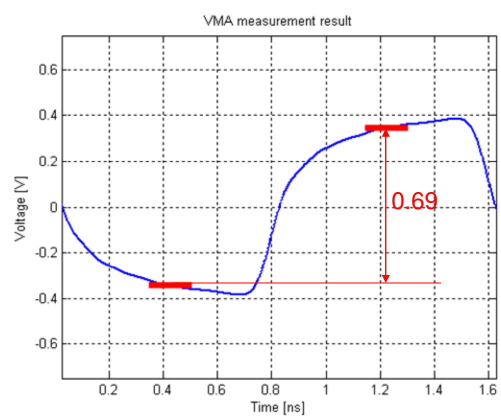


(d)

Figure 6: (a) and (b) Measurement set-up for VMA loss measurements; example of VMA loss measurement for a calibration trace (c) and of a 5m SFP+ cable assembly (d)

To show the correlation between the IL and VMA Loss, the VMA loss has been calculated for two links: a 5 m SFP+ cable assembly and a 15 cm long backpanel link. The VMA loss has been calculated as a function of bitrate. The calibration trace that is used is a perfect thru. Figure 7 compares the IL and the VMA loss. IL is plotted vs frequency. VMA loss is plotted against bitrate/20. From the figure it can be concluded that the VMA loss is approximately equal to the IL at a frequency equal to the selected bitrate/20. The factor 20 is empirically determined.

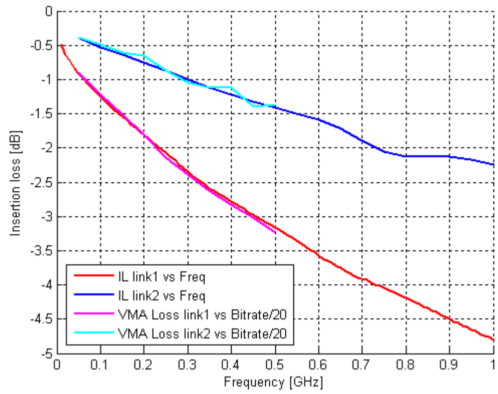


Figure 7: IL and VMA loss vs bitrate for 2 different links

3.2 Multiple Reflections

3.2.1 Introduction

As is shown in previous paragraph, multiple reflections cause noise at the receiver and are unwanted. To make sure that these multiple reflections don't cause system failure, a parameter such as impedance or RL or insertion loss deviation is defined in standards to limit the noise caused by multiple reflections.

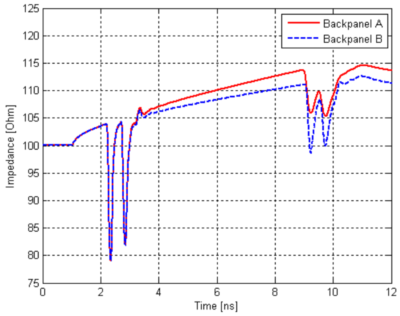
3.2.2 Impedance

As multiple reflections are caused by impedance mismatches it is clear that the impedance profile is an indication for the multiple reflection noise.

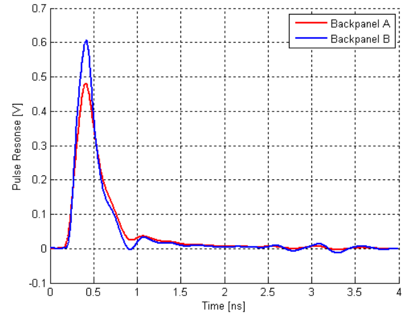
If you look to the history of standards then you will notice that impedance is the most used parameter for the description of a passive interconnect. For a long time, it was the most important parameter. Not only because it is a measure for the reflections and mismatches that occur in channel and due to the physical correlation that exist between an impedance profile and the physical structure that is tested, but also because it can easily be measured with the first available test sytem: a time domain reflectometer (TDR).

Although impedance is the most used parameter, it is not the most important parameter as it doesn't say anything about the signal that arrives at the receiver of an interconnect.

Consider 2 backpanel systems that consist of a backpanel, 2 connectors and 2 component cards. Both backpanels are identical apart from the PCB material of the backpanel. Backpanel A is made of a PCB material with dissipation factor 0.022, while the backpanel B is made of a PCB material with dissipation factor 0.007. Figure 8 a) shows the impedance profile of the 2 backpanels. The min and max impedance of both backpanels are more or less identical. Figure b) shows the pulse response of both backpanels. Figures c) shows the multiple reflection noise. From the Figures can be concluded that although impedance is more or less identical, backpanel B performs significant better than backpanel A.



(a)



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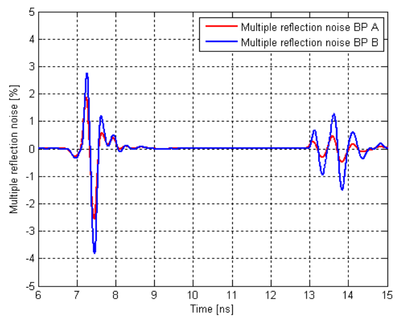
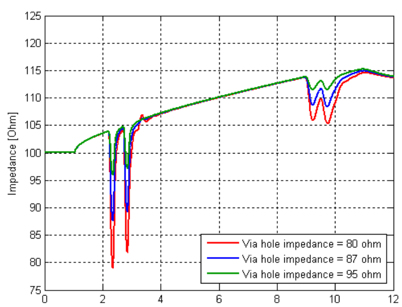


(c)

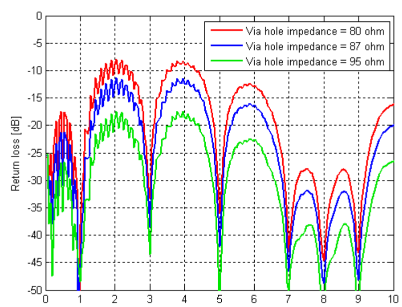
Figure 8 a) Impedance profile, b) pulse response and c)multiple reflection noise of two backpanel systems.

3.2.3 Return loss

As return loss is the frequency domain representation of impedance, return loss is also an indirect measure for the noise caused by multiple reflections. The lower the return loss, the less reflections that occur and the lower the noise, caused by multiple reflections, will be. Figure 9. a) shows the impedance profile of 3 backpanel links. All links are identical apart from the via hole impedance. The via hole impedance varies between 80 and 95 ohm. Figure b) shows the corresponding return loss. When impedance and return loss are specified in one specification, one has to make sure that one specification is not more stringent than the other one.



(a)



(b)

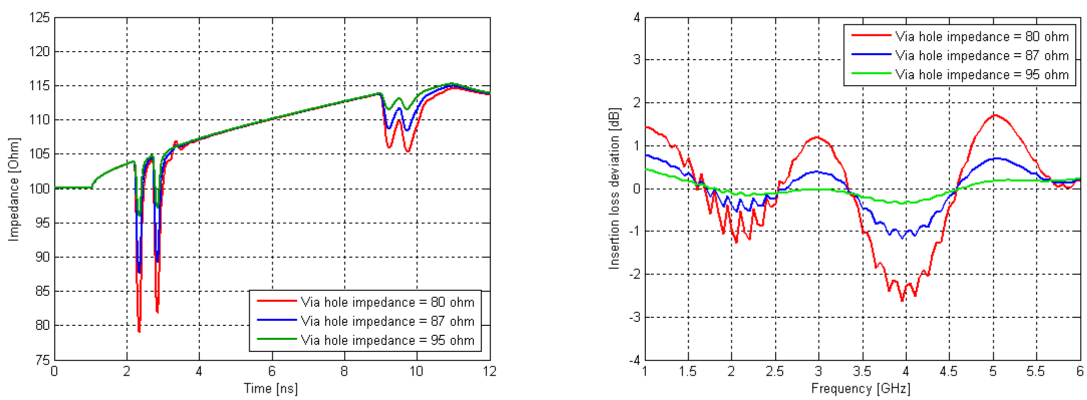
Figure 9 a) Impedance and b) Return loss of a backpanel link with via hole impedance varying between 70 ohm and 90 ohm

3.2.4 Insertion loss deviation (ILD)

Insertion loss deviation is the difference between the IL (in dB) and the fitted Attenuation (in dB).

$$ILD = IL - Attenuation_{fittted}$$

As is illustrated on Figure 10, insertion loss deviation is a measure for the impedance mismatches and as such for the noise caused by the multiple reflections of an interconnect. Figure 10 shows how the ILD of 3 nearly identical backpanel links. Backpanel links are identical apart from the connector footprint. As is shown on figure (a) impedance of the connector footprint varies between 80 and 95 ohm. From the Figure can be concluded that as impedance mismatch increases, ILD also increases.



(a) *(b)*

Figure 10: (a) Impedance and (b) ILD for 3 different backpanel links.

3.2.3 Multiple reflections (MR)

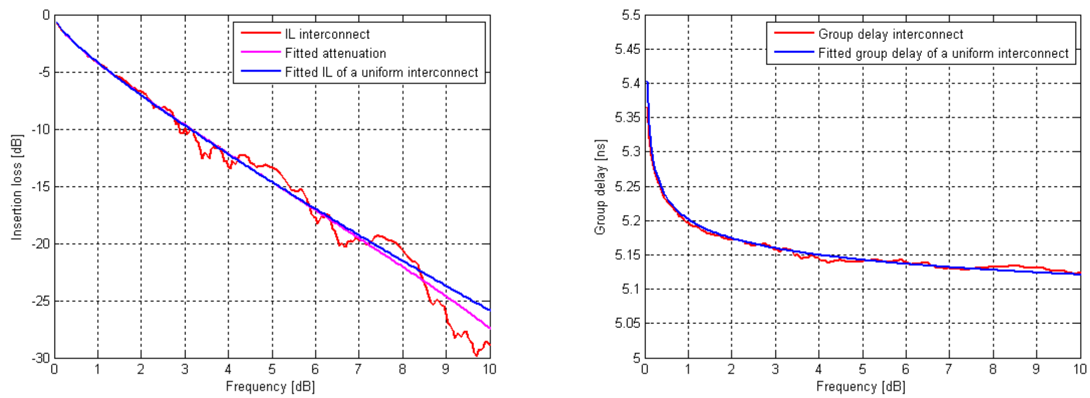
Main disadvantage of impedance, RL and ILD is that they are indirect parameters. They don't say anything about the actual multiple reflection noise that arrives at the receiver and it is not possible to compare the multiple reflection noise to the other noise parameters that arrive at the receiver. To overcome this problem, in this paragraph a new parameter is introduced: the multiple reflections noise (MR).

MR is currently not used by specifications but is a logical extension for the ILD. MR is defined as

$$MR = 20\log_{10}\left|S_{DD21} - S_{fitted,21}^{uniform}\right|$$

and is expressed in dB.

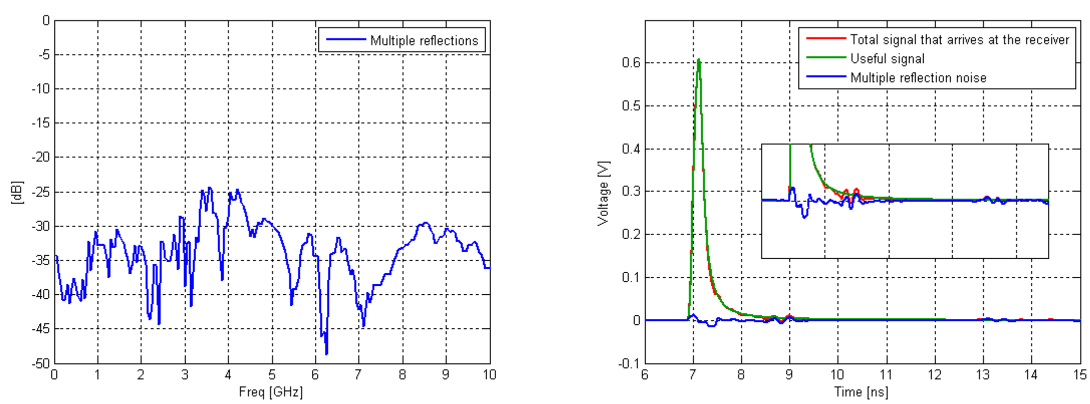
The $S_{fitted,21}^{uniform}$ is the S_{21}-parameter of a uniform interconnect (with no discontinuities) and with loss and group delay fitted to the loss and group delay of the interconnect. Figure 11 shows an example of such a fit for a typical backpanel link. Figure (a) compares IL while Figure (b) compares the group delay.



(a) *(b)*

Figure 11: (a) Loss and (b) group delay of a uniform interconnect fitted to the loss and group delay of a typical backpanel link.

Figure 12 shows the multiple reflections of a typical interconnect. Figure (a) shows the frequency domain performance (MR transfer function) while figure (b) shows the time domain multiple reflection noise that arrives at the receiver when a pulse with risetime 50 ps and pulsewidth 100 ps is launch at the transmitter site of the interconnect.. Also shown on the figure are the complete signal that arrives at the receiver and the useful signal.



(a) (b)

Figure 12: (a) Multiple reflections of a typical backpanel link, and (b) multiple reflection noise that arrives at the receiver.

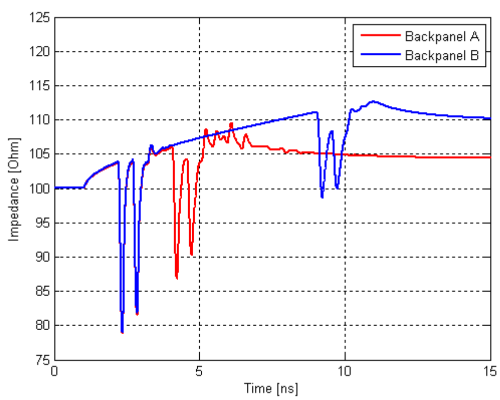
3.3 Transmitter and receiver mismatch noise

3.3.1 Introduction

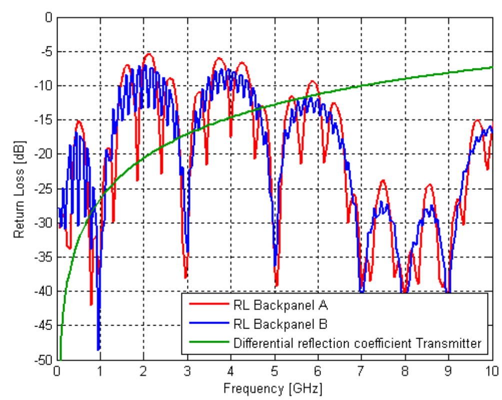
Consider an arbitrary interconnect. Unless the interconnect is perfectly matched to the transceiver, a differential signal launched from a transceiver will partially be reflected at the interconnect. This signal propagates back to the transceiver. If the transceiver is not perfectly matched, the reflected signal will be re-reflected and will propagate over the interconnect and will finally arrive at the receiver. This signal that arrives at the receiver is unwanted and is the transmitter mismatch noise (TMN). At the same time the launched differential signal propagates over the interconnect and arrives at the receiver. If the receiver is not perfectly matched, part of the signal reflects, and returns to the interconnect. At the interconnect, this signal is re-reflected and finally arrives at the receiver as receiver mismatch noise (RMN)

The parameters that are used in a specification to control the transmitter and receiver mismatch noise are the interconnect return loss and the transmitter and receiver differential reflection coefficient. Limiting these parameters will limit the transmitter and receiver mismatch noise.

However these parameters do not take into account that the signal also propagates over the interconnect. If the interconnect has high losses, then the propagated signal will be attenuated and the TMN and RMN will be low, even if RL and/or transmitter and receiver mismatch are high. This is illustrated in Figure 13. Figure (a) shows the impedance of two backpanel links. Both links are identical apart from the length of the backpanel. The backpanel of link A is 10 cm long, while the backpanel of link B is 50 cm long. Figure (b) shows the return loss of both links and the reflection coefficient of the transmitter. Figure (c) shows the transmitter mismatch noise at the receiver. It is clear that the noise is lower for the long link due to the attenuation over the backpanel.



(a)



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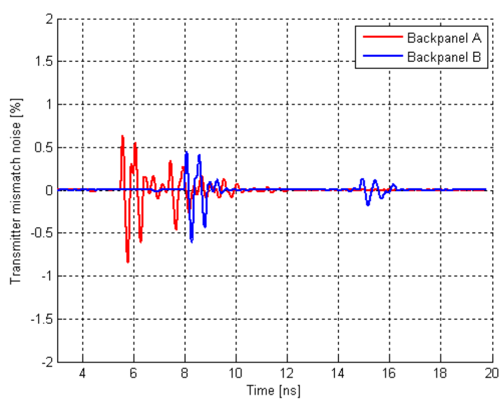


(c)

Figure 13: (a) Impedance profile of two backpanel links, (b) corresponding return loss and transmitter reflection coefficient, and (c) transmitter mismatch noise at the receiver.

Further more RL, transmitter and receiver differential reflection coefficient are indirect parameters, they do not say anything about the noise that actually arrives at the receiver and it is not possible to compare the contribution of these noise sources with other noise sources. Therefore in this paragraph two new performance parameters are introduced: transmitter mismatch noise (TM) and receiver mismatch noise (RM).

3.3.2 Transmitter and receiver mismatch (TM and RM)

Rather then using the RL and transceiver and receiver differential reflection coefficient, it is better to use the transmitter mismatch or receiver mismatch as both parameters take into account the loss of the interconnect. Transmitter and receiver mismatch noise are defined as

$$\begin{aligned} TM &= \frac{S_{11}^{channel}.S_{11}^{transmitter}.S_{21}^{channel}}{1 - S_{11}^{channel}.S_{11}^{transmitter}} \\ &\approx S_{11}^{channel}.S_{11}^{transmitter}.S_{21}^{channel} \end{aligned}$$

$$\begin{aligned} RM &= \frac{S_{21}^{channel}.S_{11}^{receiver}.S_{22}^{channel}}{1 - S_{22}^{channel}.S_{11}^{receiver}} \\ &\approx S_{21}^{channel}.S_{11}^{receiver}.S_{22}^{channel} \end{aligned}$$

Figure 15 shows the TM and RM for a typical backpanel link under the assumption that the reflection coefficient of the transmitter and receiver can be modeled by the circuit shown on Figure 14 with C = 0.3 pF and Z_0 = 50 ohm (single ended).

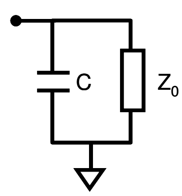
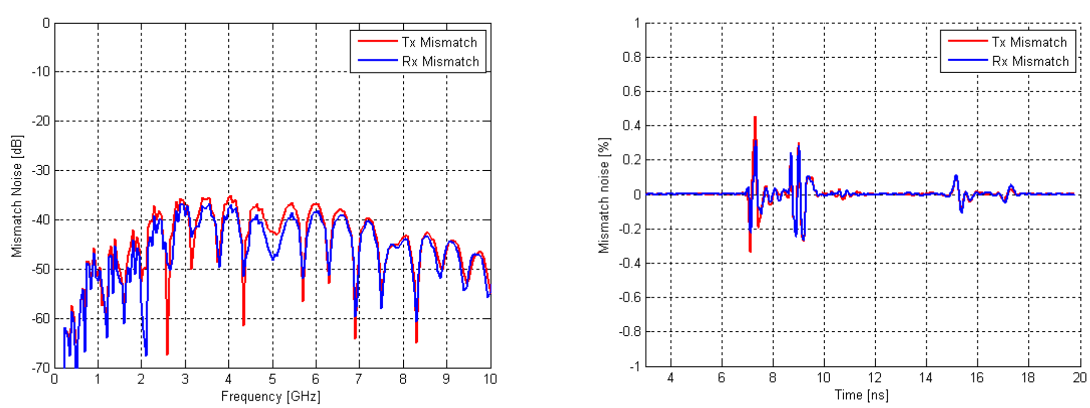


Figure 14: Circuit model of the transmitter and/or receiver input impedance



(a) *(b)*

Figure 15: Transmitter and receiver mismatch noise of a typical backpanel, (a) frequency domain results, (b) corresponding time domain results (excitation = pulse, risetime = 35 ps, pulsewidth = 100 ps)

3.4 Differential near end and far end crosstalk

3.4.1 Introduction

An important parameter as it is one of the primary reasons why systems don't seem to work is differential crosstalk. As is well known, crosstalk is a measure for the coupling between channels within a system. A distinguish is made between near end and far end crosstalk. Crosstalk can be expressed in time domain (in %) or frequency domain (in dB). As pair to pair crosstalk is not a measure for the total differential pair to pair crosstalk, it is not often used for compliance testing.

For a long time, worst case synchronous crosstalk, the sum of the peak values of the pair to pair crosstalk, was used in specifications. However this description does not include crosstalk duration and crosstalk ripple, therefore in high speed standards, the worst case synchronous crosstalk is replaced by the frequency domain power sum crosstalk.

3.4.4 Power sum crosstalk

Power sum crosstalk is a frequency domain measure for the total crosstalk that arrives at the receiver. Consider a passive interconnect that consists of N+F+1 channels with N the number of channels with a near end contribution (near end channels) at the receiver and F the number of channels with a far end contribution (far end channels). Let $NEXT_i$ (i=1,…,N) be the pair to pair near end crosstalk at the receiver of near end channel i (expressed in dB) and let $FEXT_i$ be the far end crosstalk at the receiver of far end channel i (expressed in dB) then the power sum crosstalk is defined as

$$PSXT = -10 \log_{10}\left(10^{-PSNEXT(f)/10} + 10^{-PSFEXT(f)/10}\right)$$

with

$$PSNEXT = -10 \log_{10}\left(\sum_{i=1}^{N} 10^{-NEXT_i(f)/10}\right)$$

and

$$PSFEXT = -10 \log_{10}\left(\sum_{i=1}^{F} 10^{-FEXT_i(f)/10}\right)$$

Figure 16 shows a PSXT measurement result of a typical backpanel link.

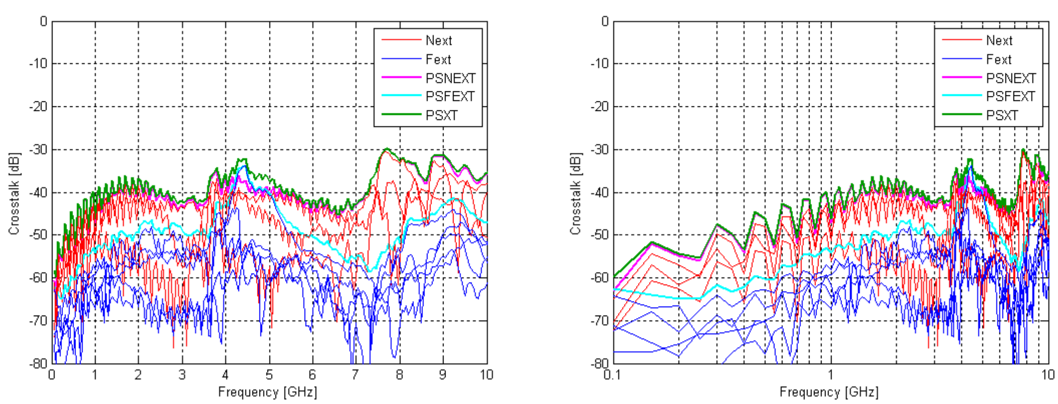


Figure 16: Pair to pair next, fext and power sum crosstalk of a typical backpanel link.

3.4.5 Insertion loss to crosstalk ratio

PSXT is a measure for the total crosstalk that arrives at the receiver. Specifying a PSXT limit in a specification is limiting the total crosstalk at the receiver. However in applications, the total noise that a system can handle is function of the loss of the interconnect. Long links with higher losses can have less noise than short links with lower losses. Part of the loss budget can be taken by the crosstalk. To take this into account in a specification, insertion loss to crosstalk ratio (ICR) has been introduced:

$$\mathrm{ICR} = -\mathrm{IL} + \mathrm{PSXT}$$

Also here it turned out that a number of interconnects marginally failed, at discrete frequencies, the ICR limit in specifications. But BER simulations showed a good performance with these interconnects. Therefore, ICR in specifications was replaced by fitted ICR: $\mathrm{ICR}_{\mathrm{fit}}$. Figure 17 shows a typical ICR and $\mathrm{ICR}_{\mathrm{fit}}$ measurement result.

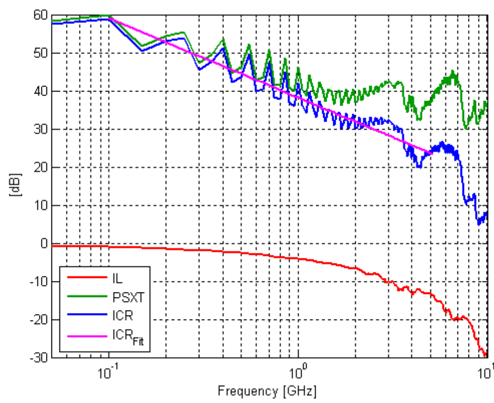


Figure 17: ICR and ICR_{fit} measurement of a typical backpanel link

3.4.6 Integrated crosstalk noise

3.4.6.1 Introduction

Although the fitted ICR already is an improvement over ICR and PSXT, recently in some standards, it was replaced by a new parameter: integrated crosstalk noise (ICN). Main advantage of this new parameter is that it takes into account the spectrum of the excitation signal. The next paragraphs describe what ICN is and how it is calculated.

3.4.6.2 Energy and power transfer

Consider a linear time invariant system with transfer function H(f). Let $v_{in}(t)$ be a time limited signal with duration T at the input of the system and $v_{out}(t)$ is the signal at the output of the system then

$$V_{out}(f) = H(f).V_{in}(f)$$

with $V_{out}(f)$ and $V_{in}(f)$ the frequency domain representations of $v_{in}(t)$ and $v_{out}(t)$.

The energy at receive side is given by

$$E_{receiver} = \int_{-\infty}^{\infty} v_{out}^2(t)dt$$

and the power at the receiver is

$$P_{receiver} = \frac{E_{receiver}}{T} = \frac{1}{T}\int_{-\infty}^{\infty} v_{out}^2(t)dt = \frac{1}{T}\int_{-\infty}^{\infty} |V_{out}(f)|^2 df$$

taking into account Parceval's theorem.

3.4.6.3 Integrated crosstalk noise

Consider an interconnect which consists of N+1 channels: one transmission channel or channel under test and N crosstalk channels. Let $XT_i(f)$ be the crosstalk transfer function of the i^{th} channel. Let H_t be the transmitter transfer function that shapes the transmitter output pulse. Assume that this is a second order butterworth filter:

$$H_t(f) = \frac{1}{1+(f.T_r/0.2365)^2}$$

T_r is the 20-80% risetime of the output pulse.

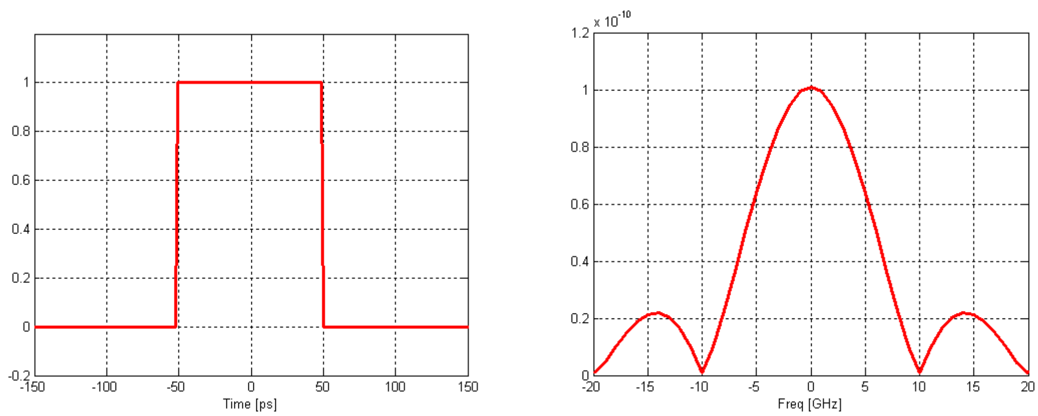
Assume that the receiver filter $H_r(f)$ is a 4the order butterworth filter with 3-dB bandwidth f_r.

$$H_r(f) = \frac{1}{1+(f/f_r)^4}$$

Assume that the crosstalk channels are excited with a single square pulse (Figure 18) with pulsewidth τ and amplitude A. Spectrum of this pulse is given by

$$V_{dif,in}(f) = A.\tau.\text{sinc}(f/f_b)$$

with $f_b = 1/\tau$.



(a) *(b)*

Figure 18: (a) Square pulse, (b) spectrum of the square pulse

The noise contribution of the i^{th} crosstalk channel at the output of the receiver filter can be written as

$$N_i(f) = H_t(f).XT_i(f).H_r(f).V_{dif,in}(f)$$
$$= \frac{1}{1+(f.T_r/0.2365)^2}.XT_i(f).\frac{A.\tau.\text{sinc}(f/f_b)}{1+(f/f_r)^4}$$

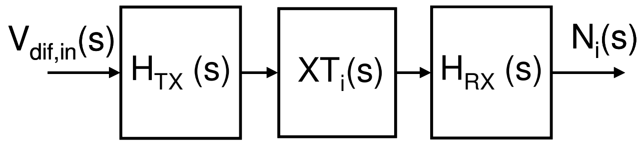


Figure 19: Schematic representation of crosstalk contribution of the i^{th} crosstalk channel at the receiver

The total average crosstalk noise power that arrives at the receiver during period T=τ is given by

$$
\begin{aligned}
N_{receiver} &= \sum_{i=1}^{N} \frac{1}{\tau} \int_{-\infty}^{\infty} n_i^2(t)dt \\
&= \sum_{i=1}^{N} \frac{1}{\tau} \int_{-\infty}^{\infty} \left|N_i(f)\right|^2 df \\
&= \sum_{i=1}^{N} \frac{1}{\tau} \int_{-\infty}^{\infty} \left|H_t(f).XT_i.H_r(f).V_{dif,in}(f)\right|^2 df \\
&= \int_{-\infty}^{\infty} f_b.\left|H_t(f)\right|^2.\left|H_r(f)\right|^2.\left|V_{dif,in}(f)\right|^2.\sum_{i=1}^{N}\left|XT_i\right|^2 df \\
&= \int_{-\infty}^{\infty} W(f).\sum_{i=1}^{N}\left|XT_i\right|^2 df
\end{aligned}
$$

with

$$
\begin{aligned}
W(f) &= f_b.\left|H_t(f)\right|^2.\left|H_r(f)\right|^2.\left|V_{dif,in}(f)\right|^2 \\
&= \frac{f_b}{1+\left(f.T_r/0.2365\right)^4}.\frac{A^2.\tau^2.\mathrm{sinc}^2(f/f_b)}{1+\left(f/f_r\right)^8}
\end{aligned}
$$

Notice that

$$\sum_{i=1}^{N}\left|XT_i\right|^2 = PSXT_{abs}^2$$

with

$$PSXT_{abs}^2 = 10^{PSXT/10}$$

and thus

$$N_{receiver} = \int_{-\infty}^{\infty} W(f).PSXT_{abs}^2 df$$

The integral can be approximated by a discrete sum:

$$
\begin{aligned}
N_{receiver} &= \int_{-\infty}^{\infty} W(f).PSXT_{abs}^2\, df \\
&= 2.\int_{0}^{\infty} W(f).PSXT_{abs}^2 df \\
&\approx 2.\sum_{k=1}^{K} W(f).PSXT_{abs}^2.\Delta f
\end{aligned}
$$

The integrated crosstalk noise σ is equal

$$\sigma = \sqrt{N_{receiver}}$$

As for ICR, the integrated crosstalk noise can be higher if insertion loss is lower: links with less loss can have more crosstalk. To take this into account, in specifications ICN is plotted against the IL@fundamental frequency (f_0) of the transmission channel and must fall in a compliance region which is function of frequency. As can be concluded from Figure 20, the higher the IL@fundamental frequency, the less the ICN can be to fall within the compliance region.

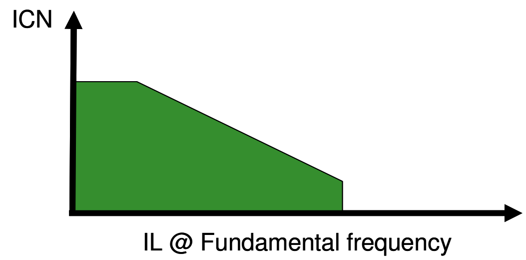


Figure 20: ICN compliance region

One problem however is that multiple reflections can have an impact on the result. As is shown on Figure 21. There fore rather than plotting ICN vs IL@fundamentel frequency, it probably is better to plot ICN vs IL_{FIT}@fundamental frequency.

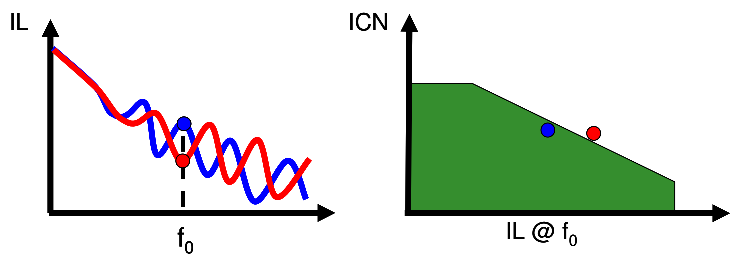


Figure 21: Ripple on the IL can have an impact on the ICN result

3.5 Common mode to differential mode transmission noise

Signals generated by differential transceivers are not perfect differential. Due to risetime differences, amplitude differences and transmitter within pair skew, the excitation signal has a common mode component. This common mode component also propagates over the passive interconnect. Due to asymmetries in the interconnect, part of this common mode signal is conversed to differential mode signal and appears as noise at the transceiver. A measure for this noise is the common mode to differential mode S-parameter. This mode conversion can easily be calculated out of the single ended S-parameters of an interconnect. Let port 1 and 2 be the single ended ports of the excitation pair and let port 3 and 4 be the single ended ports of the receiver pair then the mode conversion is given by

$$DCT = S_{DC,21} = 0.5 * (S_{3,1} + S_{3,2} - S_{4,1} - S_{4,2})$$

Controlling the noise caused by mode conversion is done by defining a limit for $S_{DC,21}$.

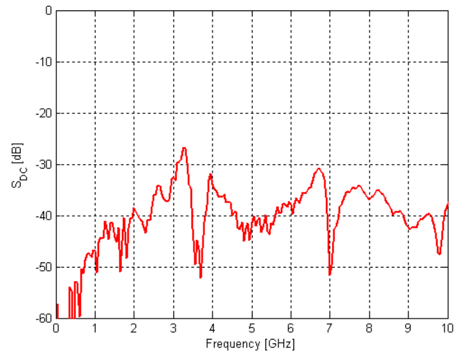


Figure 22: Typical mode conversion measurement result.

3.6 Common mode to differential mode crosstalk noise

Not only the transmission channel exhibits a common mode component, also the signals propagating over the crosstalk channels contain a common mode component. This common mode component can, due to mode conversion, appear as unwanted differential crosstalk at the receiver. This common mode to differential mode crosstalk can be calculated out of the single ended S-parameters of the interconnect.

Assume that port 5 and 6 the single ended ports are of a crosstalk channel and that port 3 and 4 are connected with the receiver then the common mode to differential mode crosstalk is calculated as

$$S_{DC,31} = 0.5 * (S_{3,5} + S_{3,6} - S_{4,5} - S_{4,6})$$

or

$$DCXT = 20 \log_{10} |S_{DC,31}|$$

If multiple crosstalk channels are involved then the common mode to differential mode powersum crosstalk needs to be considered:

$$DCPSXT = -10 \log_{10}\left(\sum_{i=1}^{X} 10^{-DCXT_i(f)/10}\right)$$

with X the total crosstalk channels (near end and far end).

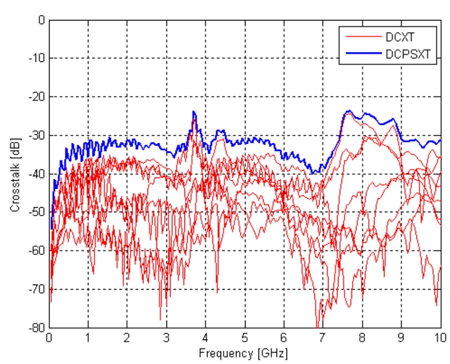


Figure 23: Example of a DCPSXT measurement result

4 Total Integrated Noise

4.1 Total integrated noise

Previous paragraphs describe a number of parameters that can be used for controlling the signals that arrive at the receiver. Each parameter typically controls/limits the contribution of one specific noise source. In standards you will typically find back for each noise parameter that is important, one of these parameters. Challenge is finding the right budget balance between the different noise sources: how much budget must be assigned to inter symbolic interference, how much to multiple reflections, how much to crosstalk, how much to mode conversion …? For every choice made, there will be links that fail the compliance criteria but will work without any problem in an application, simply because there is no perfect choice. Main problem is that it is not possible or very complex to link the different performance parameters with each other. ILD should be function of ICR: if multiple reflections are low, more crosstalk could be allowed, etc.

In this paragraph we introduce the integrated total noise. The integrated total noise combines all noise parameters into one parameter. The noise at the receiver is controlled by controlling this single parameter. It is no longer needed to specify a separate performance parameter for each different noise source.

Starting point is the integrated crosstalk noise. The integrated crosstalk noise makes use of the crosstalk power at the receiver:

$$N_{\text{differential crosstalk @ receiver}} = N_{PXSXT} \approx 2.\sum_{k=1}^{K} W_{DD}(f).PSXT_{abs}^2.\Delta f$$

Notice that the weight function $W_{DD}(f)$ is determined by the transmitter and receiver characteristics and by the differential excitation source. The passive interconnect defines the $PSXT_{abs}^2$.

In an identical way we can calculate the integrated multiple reflection noise, the integrated transmitter and receiver mismatch noise. In stead of using the power sum crosstalk, the multiple reflection noise (MR), the transmitter mismatch noise (TM) and receiver mismatch noise (RM) functions must be used:

$$N_{\text{multiple reflections @ receiver}} = N_{MR} \approx 2.\sum_{k=1}^{K} W_{DD}(f).MR_{abs}^2.\Delta f$$

$$N_{\text{transmitter mismatch @ receiver}} = N_{TM} \approx 2.\sum_{k=1}^{K} W_{DD}(f).TM_{abs}^2.\Delta f$$

$$N_{\text{receiver mismatch @ receiver}} = N_{RM} \approx 2.\sum_{k=1}^{K} W_{DD}(f).RM_{abs}^2.\Delta f$$

Also the common mode to differential mode transmission noise and common mode to differential mode crosstalk noise can be calculated in a similar way. In stead of using the spectrum of the differential component of the excitation function in the weight function, the spectrum of the common mode component of the excitation function must be used.

$$N_{\text{common mode to differential transmission@receiver}} = N_{DCT} \approx 2.\sum_{k=1}^{K} W_{CD}(f).DCT_{abs}^{2}.\Delta f$$

$$N_{\text{common mode to differential crosstalk@receiver}} = N_{DCPSXT} \approx 2.\sum_{k=1}^{K} W_{CD}(f).DCPSXT_{abs}^{2}.\Delta f$$

Main advantage of this approach is that all noise parameters can be compared with each other and that the noise source with the highest contribution can be identified. Also the total integrated noise can then be calculated as

$$\sigma_{total} = \sqrt{N_{PSXT} + N_{MR} + N_{TM} + N_{RM} + N_{DCT} + N_{DCPSXT}}$$

As the total allowed noise in an interconnect will also be function of the loss of the interconnect, σ_{total} can be plotted against the IL_{FIT}@fundamental frequency and a compliance region can be defined as function of the IL_{FIT}@fundamental frequency.

4.2 Inter symbolic interference noise

In the previous paragraph we did not look at the transmitted signal. In this paragraph we will analyze the transmitted signal (excluding multiple reflections) and we will calculate the transmitted power. This power can be divided into the integrated received power P and the integrated inter symbolic interference noise N_{ISI}. This latter is the noise that can be reduced by signal conditioning.

The total power received at the receiver is defined as

$$\begin{aligned} P_{receiver} &= \frac{1}{T}\int_{-\infty}^{+\infty} v_{out}^{2}(t)dt \\ &= \frac{1}{T}\int_{-\infty}^{t_2} v_{out}^{2}(t)dt + \frac{1}{T}\int_{t_1}^{t_2} v_{out}^{2}(t)dt + \frac{1}{T}\int_{t_2}^{+\infty} v_{out}^{2}(t)dt \\ &= P + N_{ISI} \end{aligned}$$

with

$$P = \frac{1}{T}\int_{t_1}^{t_2} v_{out}^{2}(t)dt$$

and

$$\begin{aligned} N_{ISI} &= \frac{1}{T}\int_{-\infty}^{t_2} v_{out}^{2}(t)dt + \frac{1}{T}\int_{t_2}^{+\infty} v_{out}^{2}(t)dt \\ &= P + N_{ISI} \end{aligned}$$

With t_2-t_1= τ and t_1 and t_2 chosen so that the P is maximized. Calculation of transmitted power and inter symbolic interference is illustrated on Figure 24. The integrated inter symbolic interference noise can be compared with the other noise sources and can be included in the calculation of the total noise:

$$\sigma_{total} = \sqrt{N_{PSXT} + N_{MR} + N_{TM} + N_{RM} + N_{DCT} + N_{DCPSXT} + N_{ISI}}$$

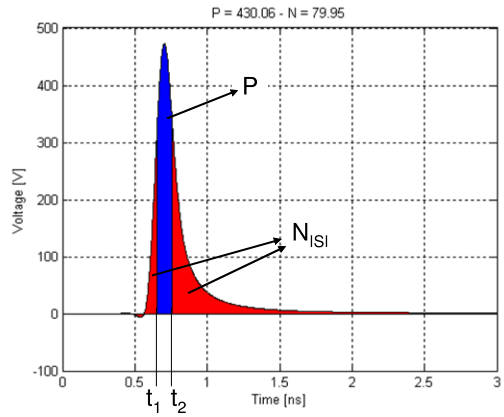


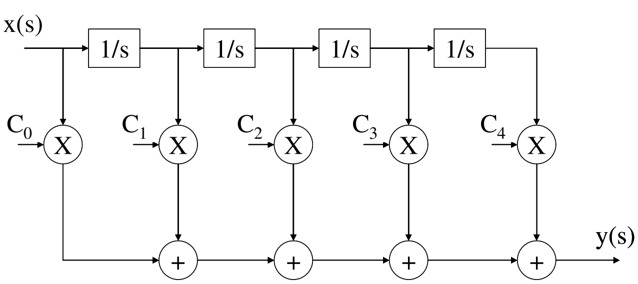
Figure 24: Calculation of transmitted power P and intersymbolic interference noise N_{ISI}

Since the integrated transmitted power at the receiver is calculated the signal to noise ratio can also be defined as an overall performance parameter for an passive interconnect:

$$SNR = \frac{P}{\sigma_{total}}$$

5 Taking into account Signal conditioning

Linear signal conditioning such transmitter de-emphasis (DE), receiver feed forward equalization (FFE), and continuous time linear equalizer (CTLE) can easily be taken into account as the weight function can easily be extended with the equalization transfer functions ($H_{DE}(s)$, $H_{CTLE}(s)$, $H_{FFE}(s)$).



$$\begin{aligned} y(s) &= H_{FFE}(s).x(s) \\ &= \left(C_0 + \frac{C_0}{s} + \frac{C_1}{s^2} + \frac{C_2}{s^3} + \frac{C_3}{s^4} \right).x(s) \end{aligned}$$

(a) *(b)*

Figure 25: 5 taps FIR filter (DE or FFE) and (b) associated transfer function

Also decision feedback equalization (DFE) can be taken into account if we assume that we can linearize the equalization. Once linearized, a transfer function exists that can be included in the weight function. In stead of using the schematic of Figure 19 for the calculation of the integrated noise, the schematic of Figure 26 should be used.

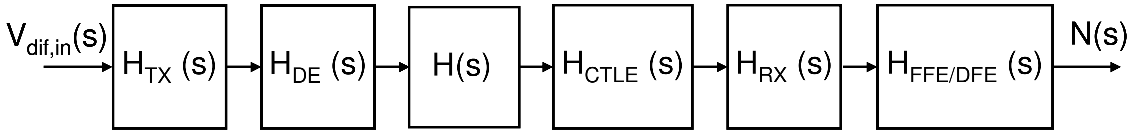
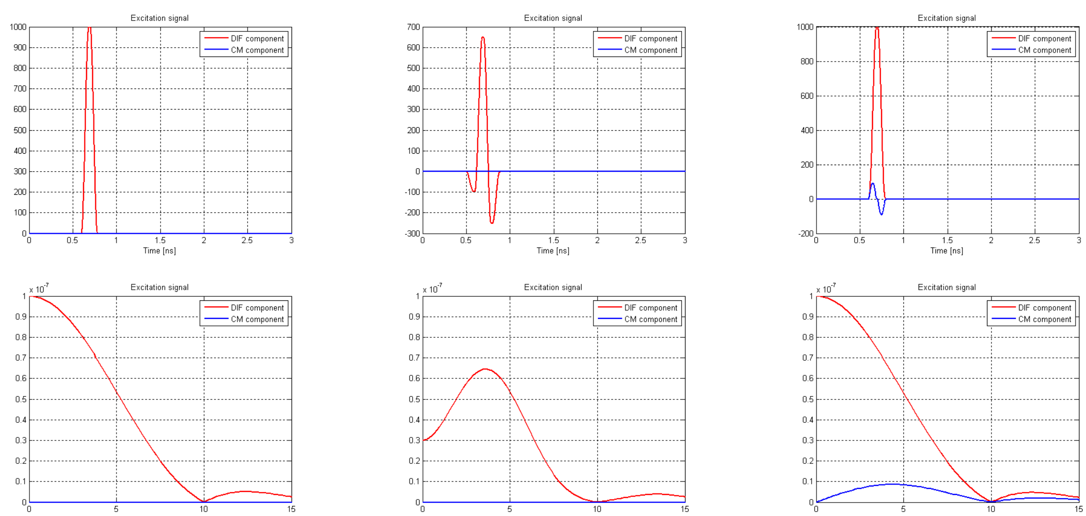


Figure 26: Schematic representation for the calculation of the noise at the receiver including signal conditioning.

6 Example

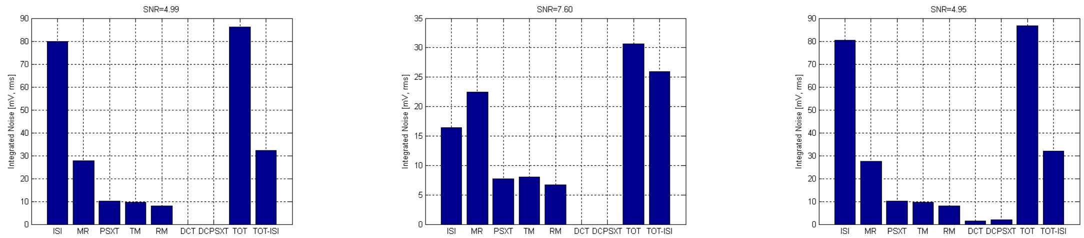
In this paragraph we will calculate the total integrated noise of a typical backpanel link, with and without signal conditioning. Also the impact of a common mode component in the excitation signal on the total noise will be illustrated.

Consider the backpanel link of Figure 2 a) with a total length of 50 cm (20 inch). The backpanel link contains 2 connectors and is build out of Nelco 4000-12 board material. In the results presented are 8 crosstalk channels included: 3 near end channels and 5 far end channels. The link is excited three times. First time with a differential pulse with bitrate 10 Gb/s, no signal conditioning is applied. Second time with the same pulse but this time the transmitter has a 3 taps FIR filter implemented. Third time again without signal conditioning but this time the + and – signal of the differential pulse have 10 ps skew between them. Figure 27 shows the excitation signals and associated spectrums. **Error! Reference source not found.** shows the contribution of the different noise sources (MR, PSXT, TM, RM, DCT, DCPSXT, ISI) and compares the total noise of the different excitations (TOT). As ISI can be minimized using signal conditioning the total noise without signal conditioning is also shown on the figures (TOT-ISI). The SNR is 4.99 in case no signal conditioning is applied and 7.60 with signal conditioning. Including a skew of 10ps does not very much adversely effect the SNR (4.95 iso 4.99 in case of no signal conditioning). From figure (a) can be concluded that ISI is the main contributor to the total integrated noise. This was expected as no signal conditioning is applied. Also notice that noise caused by multiple reflections is more important than crosstalk. When a 3 taps FIR filter is applied then the ISI noise and total noise reduces spectacular (Figure (b)). Figure (c) finally shows that when common mode is excited, the mode conversion noise is very small and has nearly no impact on the total noise.



(a) (b) (c)

Figure 27: Signals used for the excitation of the backpanel link:a) no signal conditioning, no skew; b) 3 taps FIR filter, no skew; c) no signal conditioning, 10 ps signal skew; top figures: time domain representation, bottom figures: spectrum.



(a) *(b)* *(c)*

Figure 28: Integrated noise at the receiver for the different excitations, a) no signal conditioning, no skew, b) 3 taps FIR filter, no skew, c) no signal conditioning, 10 ps signal skew.

7 Conclusion

In this paper an overview is given of the different performance parameters that are used for describing the different noise sources at the receive side of a passive interconnect and that are used for compliance testing of passive interconnects.

A new parameter, the total integrated noise has been introduced. This parameter is a measure for all possible noise sources at the receiver and allows a comparison between the different noise sources.

By making use of this parameter for compliance testing, it is no longer needed to specify a compliance limit for the individual noise sources. An additional advantage of the proposed approach is that it possible to determine the noise source with the highest contribution to the total noise. Finding a solution to minimize this noise will have the highest effect on the total interconnect performance.