



# OMEGA link model

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# Introduction and definitions

# Link model — Introduction

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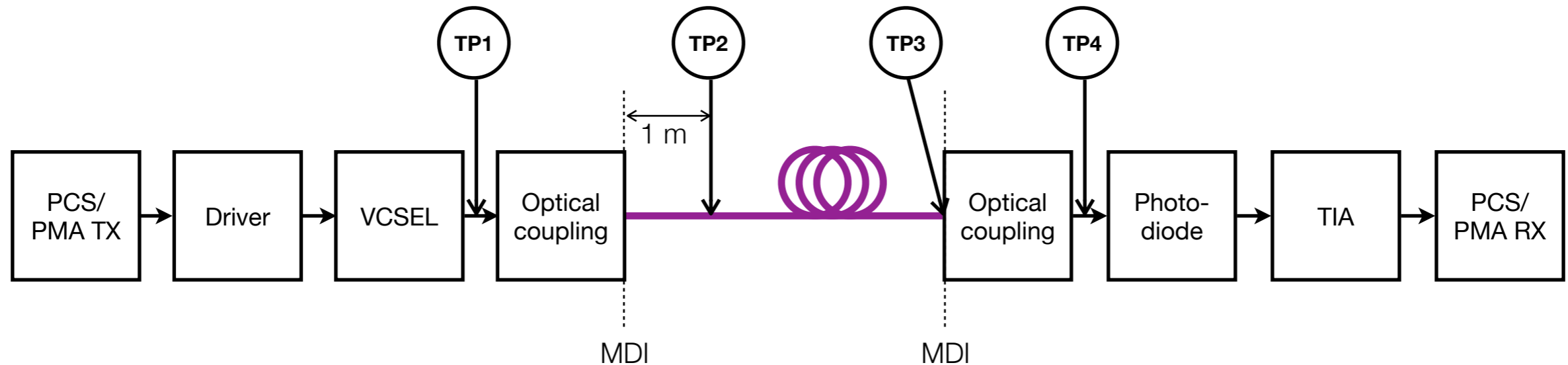
- The link model is based on Shannon's capacity, which takes into account every noise source with impact in the communications link as well as every transfer function that distorts the communications signal
- The receiver is accurately modeled including optimum equalizer to approach the channel communications capacity
- Effects of FEC (Forward Error Correction) code-rate and net coding-gain are included
- The model is built in the electrical domain (vs optical domain), because the communications system under consideration is IM/DD (Intensity Modulation / Direct Detect) and the main noise sources are electrical
- The optical link is short ( $\leq 40$  m per objectives): CD (Chromatic Dispersion) and MPN (Mode Partition Noise) are negligible
- MN (Modal Noise) is negligible: hypothesis that needs to be proven for the optical connections under consideration suitable for automotive applications, i.e. no MSL (Mode Selective Loss) is produced, cable vibrations produce negligible effect, etc.
- Considered noise sources: VCSEL RIN (Relative Intensity Noise), photodiode, TIA (Trans-Impedance Amplifier), TX/RX clock jitter, ADC (Analog-to-Digital Converter) quantization

# Link model — Introduction



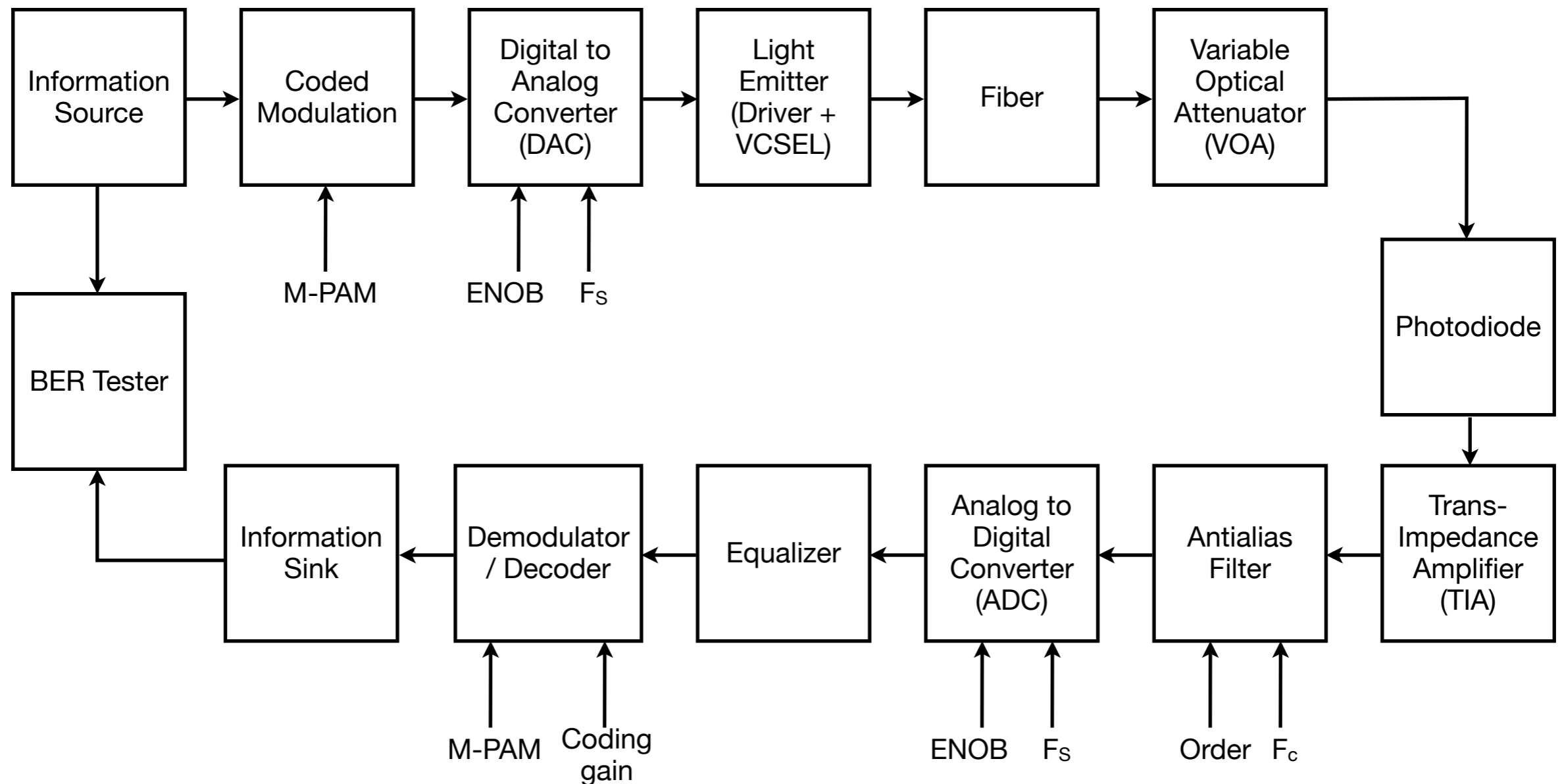
- Sensitivity obtained in electrical domain is translated into optical power terms through the photodiode responsivity
- Two signal paths are considered, DC and AC, both closely related by the Extinction Ratio (ER) and the modulation scheme
- Noise models:
  - Noise sources coming from photodiode and TIA are very well modeled as additive gaussian colored noise, equivalent to AWGN (Additive White Gaussian Noise) filtered by a linear system
  - Jitter effect is artificially modeled as gaussian colored additive noise with PSD (Power Spectral Density) dependent of sampled signal PSD
  - Quantization noises are artificially modeled as additive white gaussian noises derived from the ENOB (Equivalent Number Of Bits) specification; non-linearities of DAC (Digital-to-Analog Converter) and ADC are not explicitly considered and they are accounted within the ENOB terms
  - No impulse noise and RFI noises are considered; the transmission media is immune to any EM interference surrounding it

# Link model — test points definition



- These test (specification) points are defined for link model definition and link budget analysis
- TP1: VCSEL optical transmit signal
  - Optical power characterized per optical coupling and LIV method described in [perezaranda\\_OMEGA\\_01b\\_0720\\_VCSEL\\_test\\_methods.pdf](#)
- TP2: Optical transmit signal at the output end of 1 meter of optical fiber consistent with the link segment type connected to MDI (Media Dependent Interface)
- TP3: Optical receive signal at the output of the fiber optic cabling, which in a link segment is connected to the receiver (i.e. MDI)
- TP4: Optical receive signal coupled into the photodiode device
- The link model is valid regardless the MDI implementation; no special assumptions about the MDI have been considered in the link model development

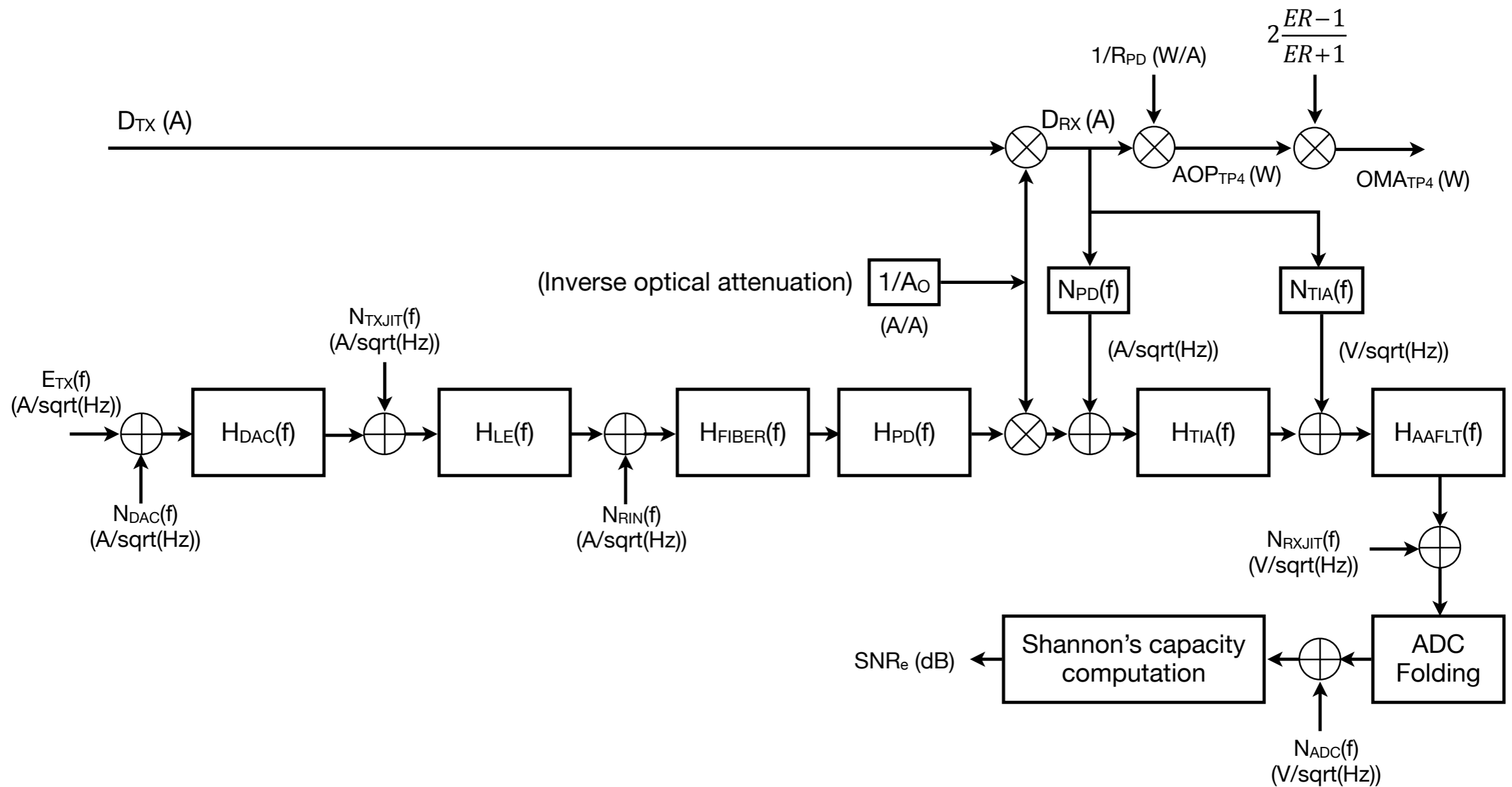
# Link model – Conceptual block diagram





# Information Theory Model (ITM)

# Information theory model





# Variables



- $E_{TX}(f)$ : TX electrical current spectral density referred to photodiode output
- $D_{TX}$ : average TX electrical current referred to photodiode output
- $D_{RX}$ : average RX electrical current at the output of photodiode
- $R_{PD}$ : Photodiode responsivity (A/W)
- $AOP_{TP4}$ : Average Optical Power in TP4
- $OMA_{TP4}$ : Optical Modulation Amplitude in TP4
- $A_O$ : optical attenuation
- $N_{DAC}(f)$ : ENOB dependent DAC noise spectral density in terms of electrical current referred to photodiode output
- $N_{ADC}(f)$ : ENOB dependent ADC noise spectral density in terms of electrical voltage
- $N_{PD}(f)$ : photodiode shot noise spectral density as electrical current at the photodiode output
- $N_{TIA}(f)$ : TIA output voltage noise spectral density
- $N_{TXJIT}(f) / N_{RXJIT}(f)$ : noise spectral density produced by TX and RX clock jitter, respectively
- $N_{RIN}(f)$ : VCSEL relative intensity noise spectral density
- $H_{DAC}(f)$ ,  $H_{LE}(f)$ ,  $H_{FIBER}(f)$ ,  $H_{PD}(f)$  and  $H_{AAFLT}(f)$ : transfer functions of DAC, light emitter, fiber, photodiode and anti-alias filter, respectively (DC normalized)
- $H_{TIA}(f)$ : transfer function of the trans-impedance amplifier (no DC normalized)
- $SNR_e$  (dB): effective SNR (Signal-to-Noise Ratio) calculated in the output of the channel

# Transmit signal model

- Let us assume the TX communication signal be a uniformly distributed M levels PAM, that take values from the set  $\{-M+1, -M+3, \dots, +M-3, +M-1\}$ , where M is a positive integer  $\geq 2$
- The crest-factor of TX signal is given by:

$$CF_{TX} = \sqrt{3 \frac{M-1}{M+1}}$$

- Under non-negative optical power constraint, the optical channel input is limited in power peak
- Based on that, the TX electrical current spectral density is given by:

$$E_{TX}(f) = D_{TX} \frac{ER-1}{ER+1} \frac{1}{CF_{TX}} \frac{1}{\sqrt{F_S/2}} \quad (\text{A/sqrt(Hz)})$$

where ER is the extinction ratio and  $F_S$  is the symbol rate.

- For NRZ scheme, i.e.  $M = 2$ :

$$E_{TX}(f) = D_{TX} \frac{ER-1}{ER+1} \frac{1}{\sqrt{F_S/2}} \quad (\text{A/sqrt(Hz)})$$

# DAC model



- DAC magnitude transfer function:

$$H_{DAC}(f) = \left| \text{sinc}\left(\frac{f}{F_s}\right) \right|; \quad \text{where} \quad \text{sinc}(x) = \frac{\sin(\pi x)}{\pi x}$$

- DAC noise spectral density:

$$N_{DAC}(f) = D_{TX} \frac{ER-1}{ER+1} \frac{1}{\sqrt{3}} \frac{1}{2^{ENOB_{DAC}}} \frac{1}{\sqrt{F_s/2}} \quad (\text{A}/\sqrt{\text{Hz}})$$

- Bandwidth limitations of DAC implementation are negligible compared with VCSEL response
- For NRZ scheme, DAC only generates 2 levels and  $N_{DAC}(f) = 0$ 
  - $N_{DAC}(f)$  can be either considered included into RIN or negligible compared to RIN

# TX clock jitter noise model

- The effect of TX clock jitter in the symbols generated by the DAC is modeled as an additive noise with a PSD given by:

$$N_{TXJIT}(f) = E_{TX}(f) \cdot 2\pi \cdot \sqrt{2} \cdot t_{JTX} \cdot f \cdot H_{DAC}(f) \quad (\text{A}/\sqrt{\text{Hz}})$$

where  $t_{JTX}$  is the phase jitter of the TX clock, defined as the standard deviation of the time error of actual clock with respect to an un-jittered clock reference of the same frequency

- The factor of  $\sqrt{2}$  has been included to account the long term jitter, making the effect of the specification pessimistic and independent of the receiver CDR (Clock Data Recovery) implementation

# VCSEL driver models



- Two types of simple models can be considered:
  - Current driver model: for an specified  $I_{BIAS}$  it transforms amplitude information into current levels to produce an specified ER taking into account the  $I_{TH}$  of VCSEL
  - Voltage driver model: for an specified  $I_{BIAS}$  it transforms amplitude information into voltage levels to produce an specified ER taking into account the  $I_{TH}$ ,  $R_J$ ,  $R_S$  and  $f_p$  of VCSEL and a source impedance  $Z_o$

• Current driver frequency response is given by  $H_{DRV-CURRENT}(f) = 1$

• Voltage driver frequency response is given by:

$$H_{DRV-VOLTAGE}(f) = \left| \frac{(R_J + R_S + Z_o) \left(1 + \frac{f}{f_p} j\right)}{(R_J + R_S + Z_o) + (R_S + Z_o) \frac{f}{f_p} j} \right|$$

where  $R_S$  is the VCSEL series resistance,  $R_J$  is the VCSEL junction resistance,  $f_p$  is the VCSEL extrinsic response pole, and  $Z_o$  is the driver source impedance

- Bandwidth limitations of the drivers have been considered negligible compared with VCSEL, specially true in high temperature, low bias current conditions
- Physical limitations of current injection from the voltage driver model are not considered

# VCSEL driver models, cont'd



- The band-limited response of the VCSEL can be improved using symbol-rate feed-forward equalization (FFE) in the driver

- Z-transform of FFE is given by  $F(z) = \sum_{k=0}^{N_F} b_k z^{-k}$ , where  $b_k$  are the FIR filter coefficients

- The frequency response of this filter is calculated as:

$$H_{FFE}(f) = \left| \sum_{k=0}^{N_F} b_k \exp \left\{ -2\pi k \frac{f}{F_s} j \right\} \right|$$

- The frequency response of the driver is therefore calculated as:

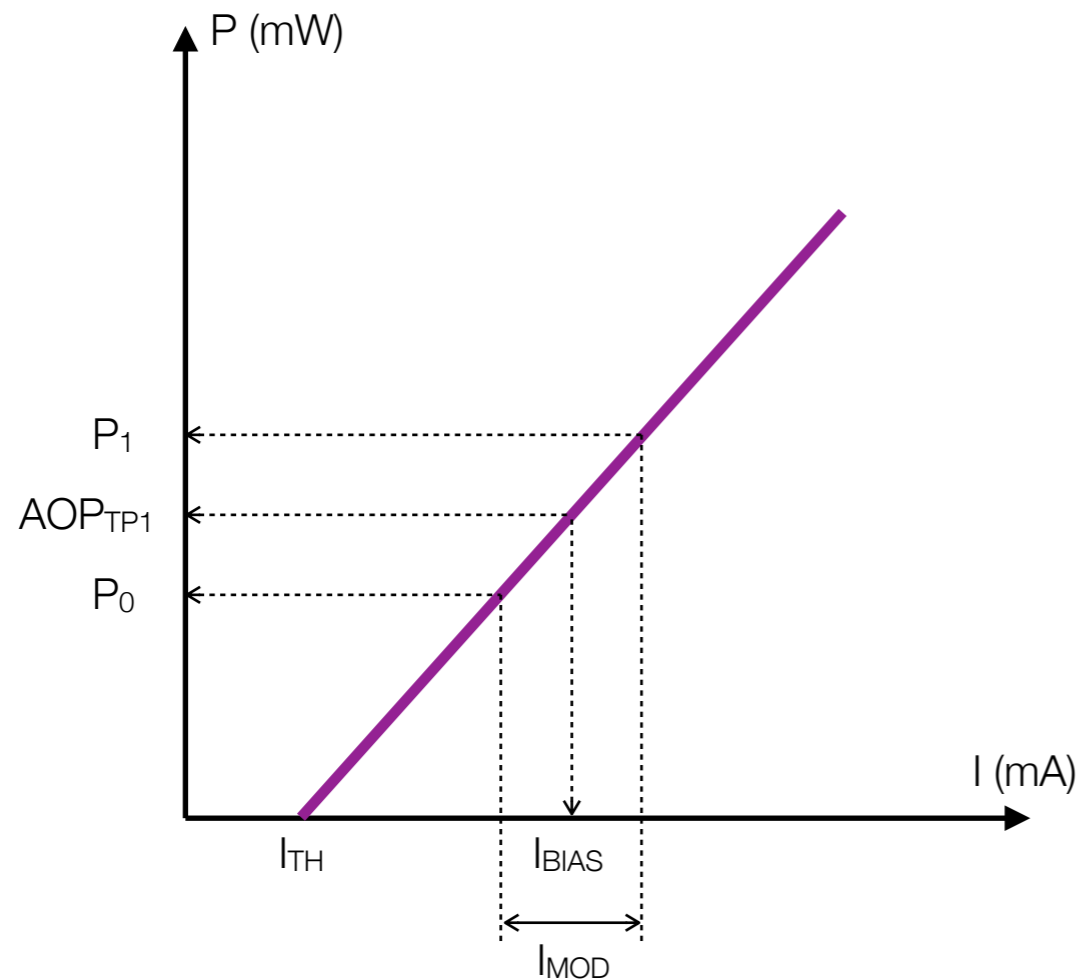
$$H_{DRV}(f) = H_{FFE}(f) \cdot H_{DRV-\{type\}}(f)$$

where  $H_{DRV-\{type\}}$  is  $H_{DRV-CURRENT}$  or  $H_{DRV-VOLTAGE}$ , per provided definitions in previous slide.

# VCSEL model — DC



- From L-I-V characterization, the model assumes affine transformation between driving current and optical power generated by the VCSEL



$$ER = \frac{P_1}{P_0}$$

$$OMA_{TP1} = P_1 - P_0$$

$$AOP_{TP1} = \frac{1}{2}(P_1 + P_0)$$

$$OMA_{TP1} = 2 \frac{ER - 1}{ER + 1} AOP_{TP1}$$

# VCSEL model — AC



- The considered linear model of VCSEL is consistent with the characterizations for several VCSEL parts in July and August meetings
- Signal transfer function (DC normalized):

$$H_{VCSEL-INT}(f) = \left| \frac{f_r^2}{f_r^2 - f^2 + j \frac{f}{2\pi} \gamma} \right| \quad H_{VCSEL-EXT}(f) = \left| \frac{1}{1 + j \frac{f}{f_p}} \right|$$

$$H_{VCSEL}(f) = H_{VCSEL-INT}(f) \cdot H_{VCSEL-EXT}(f)$$

where the response is split in intrinsic part (i.e. VCSEL-INT) and extrinsic part (i.e. VCSEL-EXT).

- $f_r$  is the resonance frequency,  $\gamma$  is the damping rate and  $f_p$  is the extrinsic cut-off frequency
- Full transfer function of light emitter (LE) is given by:

$$H_{LE}(f) = H_{DRV}(f) \cdot H_{VCSEL}(f)$$



# VCSEL model – RIN



- RIN specification is given in terms of OMA, which can be directly measured in a Digital Communications Analyzer (DCA). Test method was described in perezaranda\_OMEGA\_01b\_0720\_VCSEL\_test\_methods.pdf.

- RIN<sub>OMA</sub> is defined  $RIN_{OMA} = 10 \cdot \log \left( \frac{(RN_{one} + RN_{zero})^2}{OMA^2 \cdot BW_N} \right) \text{ dB/Hz}$ , where:

- RN<sub>one/zero</sub>: standard deviation of noise measured in level 1/0 (in Watts)
- OMA: optical modulation amplitude (in Watts)
- BW<sub>N</sub>: noise equivalent bandwidth (in Hz). BW<sub>N</sub> = 1.04 · BW<sub>-3dB</sub>, for a 4th order Bessel filter

- RIN model assumes noise is additive and independent of the light power level

- RIN power density is given by  $N_{RIN}(f) = \frac{A \cdot f^2}{(f_r^2 - f^2)^2 + (\gamma/2\pi)^2 f^2} \text{ (dB/Hz)}$ ,

where constant A has to be calculated to meet  $RIN_{OMA} = \int_0^{BW_N} N_{RIN}(f) df$ .

- N<sub>RIN</sub>(f) is consistent with the one reported in [1].
  - Here we consider B=0 to make possible unique solution based on RIN<sub>OMA</sub> specification
  - B=0 approximation is consistent with the observation of RIN (f > f<sub>r</sub>) is higher than RIN (f ≈ 0)

# Fiber model

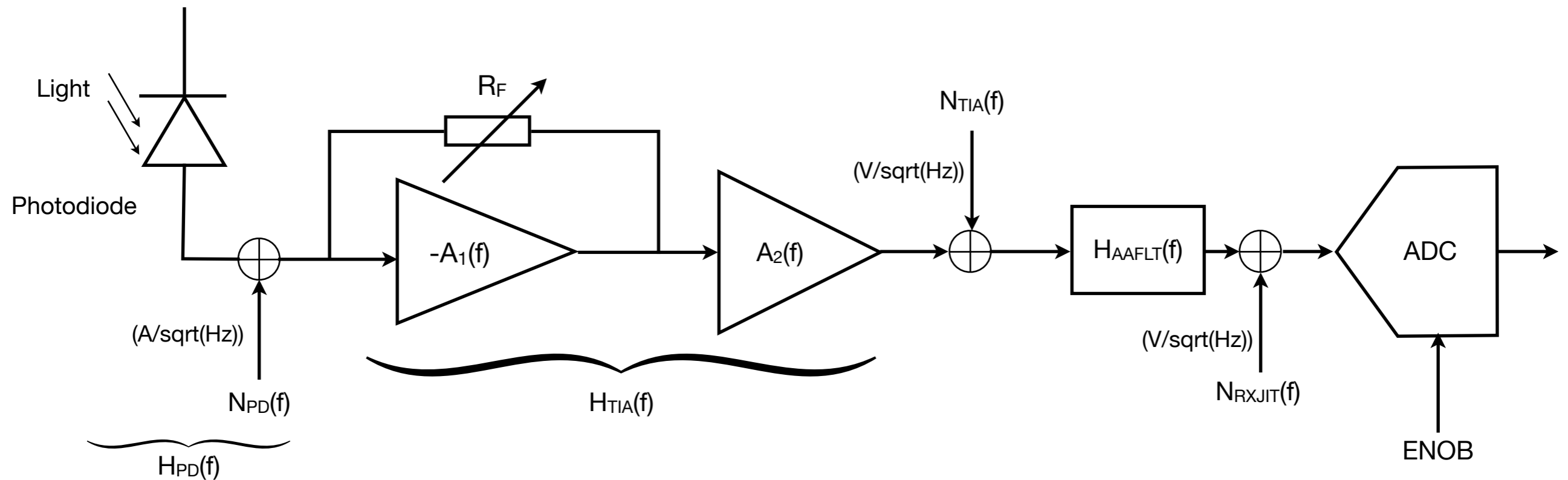


- The fiber frequency response is modeled as a Gaussian filter that meets the BLP (bandwidth length product) specification
- The magnitude of the transfer function is given by:

$$H_{FIBER}(f) = \exp\left\{-\ln(2)\left(L\frac{f}{BLP}\right)^2\right\}$$

where  $L$  is the fiber length (m), BLP is given in (Hz × m).

# RX general model



- Resistive feedback TIA followed by a voltage gain stage is implemented
- $R_F$  feedback is controlled by AGC to accommodate the gain to input optical power conditions
- Antialias filter is implemented before ADC sampling to reduce the TIA out of band noise folding inside the Nyquist band
- ADC operates at symbol rate with CDR circuit providing the optimum sampling phase for max signal to noise ratio
- Electrical simulations of transistor level implementation are used in co-simulation with communications system implemented in MATLAB®
- Receiver is simulated worst production corner conditions in order to get worst case sensitivity results

# RX clock jitter noise model

- The effect of RX clock jitter in the symbols sampled by the ADC is modeled as an additive noise with a PSD given by:

$$S_{RX}(f) = E_{TX}(f) \cdot H_{DAC}(f) \cdot H_{LE}(f) \cdot H_{FIBER}(f) \cdot H_{PD}(f) \cdot \frac{1}{A_0} \cdot H_{TIA}(f) \cdot H_{AAFLT}(f)$$

$$N_{RXJIT}(f) = S_{RX}(f) \cdot 2\pi \cdot \sqrt{2} \cdot t_{JRX} \cdot f \quad (\text{A/sqrt(Hz)})$$

where  $t_{JRX}$  is the phase jitter of the RX clock, defined as the standard deviation of the time error of actual clock with respect to an un-jittered clock reference of the same frequency

- The factor of  $\sqrt{2}$  has been included to account the long term jitter, making the effect of the specification pessimistic and independent of the receiver CDR implementation

# ADC model



- ADC model is composed by two steps: folding + quantization
- Folding: signal PSD in the input of the ADC is folded into the 1<sup>st</sup> Nyquist band with power addition. The same performed for the noise PSD
  - After folding operation we have  $S_{RX-ADC}(f)$  and  $N_{RX-ADC}(f)$ , defined between 0 and  $F_s/2$
- Quantization is modeled by additive noise with power spectral density that is calculated as:

$$N_{ADC}(f) = ADC_{FS-PP} \frac{1}{\sqrt{3}} \frac{1}{2^{ENOB_{ADC}}} \frac{1}{\sqrt{F_s/2}} \quad (\text{V}/\sqrt{\text{Hz}})$$

where  $ADC_{FS-PP}$  is the ADC full scale in peak-to-peak volts

# Effective signal-to-noise ratio ( $SNR_e$ )

- In the output of the ADC, after folding and quantization, the effective SNR ( $SNR_e$ ) is measured
- $SNR_e$  is defined in the output of the ADC as the SNR that provides the same capacity of the channel defined by  $S_{RX-ADC}(f)$  and  $N_{RX-ADC}(f)$ , computed in the first Nyquist band

$$C \triangleq \frac{F_s}{2} \log_2(1 + SNR_e) \text{ bits/s}$$

$$SNR_e = \exp \left\{ \frac{2}{F_s} \int_0^{\frac{F_s}{2}} \ln \left( 1 + \frac{S_{RX-ADC}(f)}{N_{RX-ADC}(f)} \right) df - 1 \right\}$$

- $SNR_e$ , evaluated in an ISI (Inter-Symbol Interference) channel, is equal to the  $SNR_d$  (SNR in detector) provided by an ideal infinite length MMSE (Minimum Mean Square Error) DFE (Decision Feedback Equalizer)
  - Error propagation and residual ISI due to finite length implementation are not considered



# Receiver sensitivity & link budget

# Required $SNR_d$ for transmission scheme



- The minimum  $SNR_d$  required for a coded M-PAM scheme is approximated by:

$$SNR_d = \frac{2 M^{2C_R} - 1}{3 C_G} \left( \operatorname{erfc}^{-1} \left( \frac{M \log_2(M)}{M-1} P_b \right) \right)^2$$

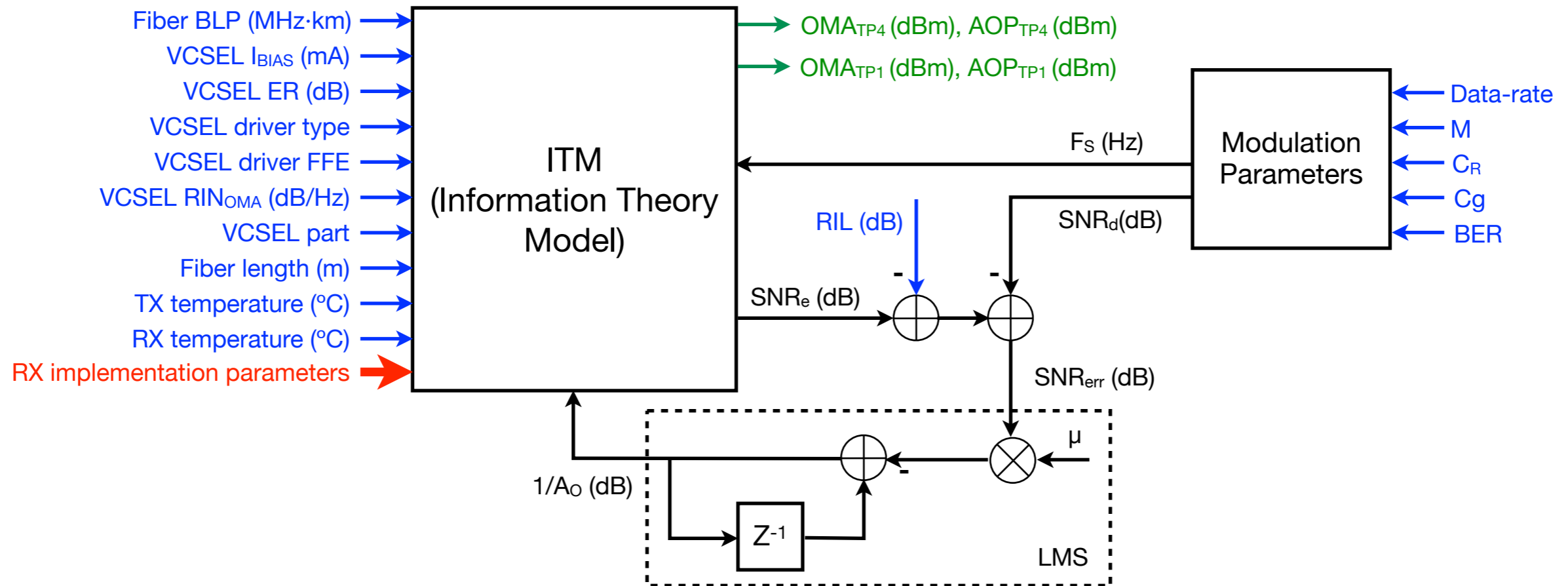
where inverse of complementary error function is defined as:

$$\operatorname{erfc}(\operatorname{erfc}^{-1}(x)) = x; \quad \operatorname{erfc}(x) = \frac{2}{\sqrt{\pi}} \int_x^{\infty} e^{-t^2} dt$$

- AWGN channel is assumed in the detector after noise whitening and ISI compensation by MMSE DFE
- Parameters:
  - M: number of levels of PAM scheme ( $M \geq 2$ )
  - $C_R$ : FEC code-rate;  $C_R < 1$  for coded schemes,  $C_R = 1$  for uncoded scheme
  - $C_G$ : net coding gain respect to an uncoded scheme;  $C_G > 1$  for coded schemes,  $C_G = 1$  for uncoded scheme. Coding gain is defined for a given  $P_b$  in non-asymptotic codes
  - $P_b$ : bit error probability (i.e. BER, Bit-Error Rate) after demodulation and decoding



# Receiver sensitivity



- The ITM is evaluated in several iterations in a LMS (Least Mean Squares) loop to find the minimum  $AOP_{TP4}$  (and  $OMA_{TP4}$ ), i.e. optical receiver sensitivity, that meets the BER requirement for the coded modulation scheme ( $F_s$ ,  $M$ ,  $C_R$ ,  $C_G$ ),
- The simulation stops when  $|SNR_{err}| < 0.01$  dB
- RIL (Resultant Implementation Losses)  $> 0$  dB accounts for the real PHY implementation losses where finite size circuits and non-linearities are considered
- RIL is not known a priori. To get a confident value for RIL, it is needed to implement a time-domain simulation of the complete system including e.g. non-linear VCSEL model, real CDR and adaptive equalization loops, as well as finite size and resolution circuitry, etc.

# Link budget



- After ITM simulation we get as results:
  - $OMA_{TP1}$  (dBm): optical modulation amplitude at TP1 for specified TX conditions and parameters
  - $OMA_{TP4}$  (dBm): optical receiver sensitivity in terms of OMA that meets data-rate and BER requirements
- If we know the different insertion losses that affect the optical signal, we can calculate link budget and link margin:
  - $OMA_{TP2}$  (dBm) =  $OMA_{TP1}$  (dBm) -  $IL_{TP1-to-TP2}$  (dB)
  - $OMA_{TP3}$  (dBm) =  $OMA_{TP4}$  (dBm) +  $IL_{TP3-to-TP4}$  (dB)
  - Link budget  $LB_{TP1-TP4}$  (dB) =  $OMA_{TP1}$ (dBm) -  $OMA_{TP4}$ (dBm)
  - Link budget  $LB_{TP2-TP3}$  (dB) =  $OMA_{TP2}$ (dBm) -  $OMA_{TP3}$ (dBm)
  - Link margin LM (dB) =  $LB_{TP1-TP4}$  -  $IL_{TP1-to-TP2}$  (dB) -  $IL_{TP3-to-TP4}$  (dB) -  $IL_{FIBER}$  (dB) -  $N_{IC} \times IL_{IC}$  (dB) -  $IL_{BEND}$ (dB)
  - Link margin LM (dB) =  $LB_{TP2-TP3}$  -  $IL_{FIBER}$  (dB) -  $N_{IC} \times IL_{IC}$  (dB) -  $IL_{BEND}$ (dB)
- Insertion losses:
  - $IL_{TP1-to-TP2}$ : TP1 to TP2 insertion loss. It shall include aging and worst case conditions.
  - $IL_{TP3-to-TP4}$ : TP3 to TP4 insertion loss. It shall include aging and worst case conditions.
  - $IL_{FIBER}$ : fiber insertion loss
  - $IL_{IC}$ : insertion loss per inline connection
  - $N_{IC}$ : number of inline connections
  - $IL_{BEND}$ : insertion loss due to macro-bending and micro-bending
- $LM > 0$  dB: the system is feasible

# References

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- [1] Seyed Ehsan Hashemi, “Relative Intensity Noise (RIN) in High-Speed VCSELs for Short Reach Communication”, Master of Science Thesis in Photonics Engineering, Chalmers University of Technology



Thank you!