Tenta-kit SSY085

Nils Patriksson

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1 General

Digital signal power: $S = E_b \cdot R_b$
Noise power: $N = n_0 \cdot B$
Shannons formula:

$$\frac{C}{B} = \log_2(1 + SNR)$$

$$SNR = \frac{E_b}{n_0}\frac{R_b}{B}$$

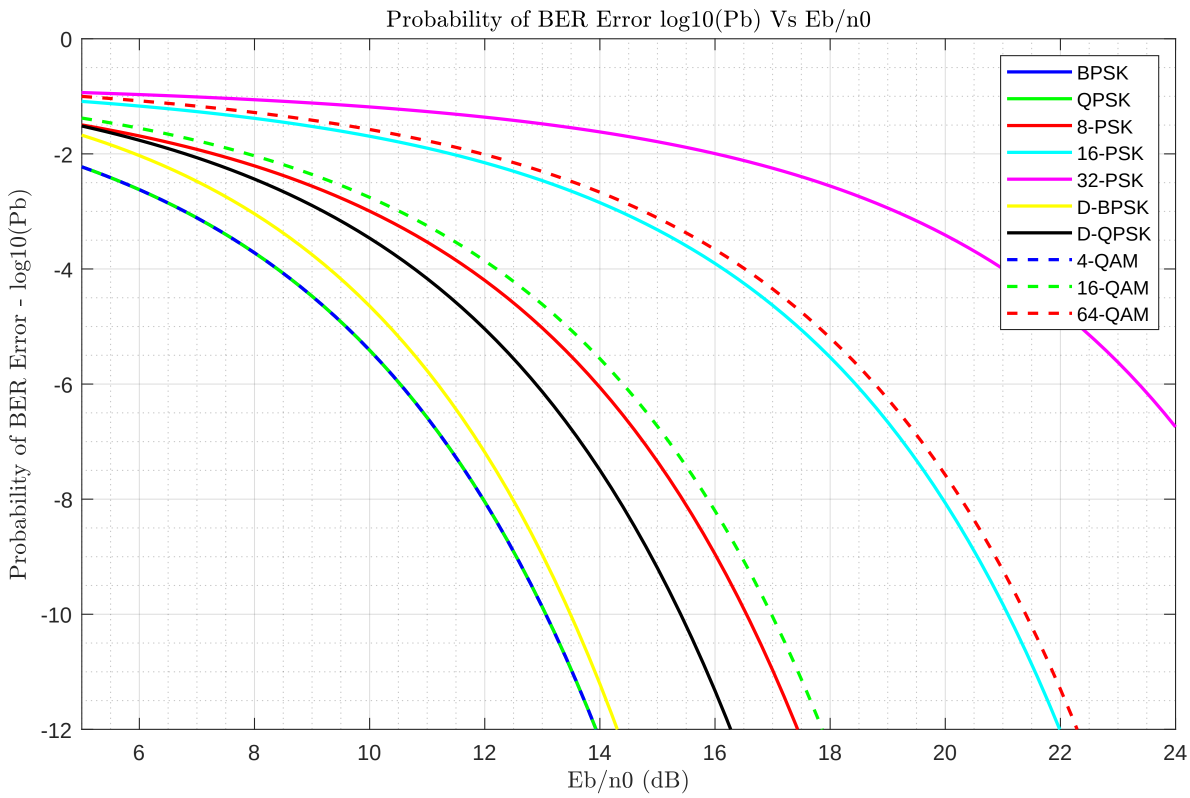
Bit error rates:

$$\text{ASK: } 2 \cdot P_e = \text{erfc}\left(\sqrt{\frac{E_b}{4n_0}}\right)$$

$$\text{FSK: } 2 \cdot P_e = \text{erfc}\left(\sqrt{\frac{E_b}{2n_0}}\right)$$

$$\text{PSK: } 2 \cdot P_e = \text{erfc}\left(\sqrt{\frac{E_b}{n_0}}\right)$$

(see graph)



Noise Figure

- **Noise figure is used to characterize the degradation in SNR for a receiver**

$$F=\frac{\left(S/N\right)_{in}}{\left(S/N\right)_{out}}=\frac{S_{in}\cdot k(T_0+T_e)B}{kT_0B\cdot S_{in}}=1+\frac{T_e}{T_0}\geq 1$$

$$T_e=\left(F-1\right)T_0$$

- **Noise figure depends on T_0**
  - $\boxed{\text{Standard uses } T_0 \equiv 290\text{ K}}$

S/N in digital communication systems

- **Digital signal power, $S = E_b \cdot R_b$**
  - R_b = Bitrate (bps)
- **Noise power, $N = n_0 \cdot B$**
  - B = Bandwidth (Hz)
- **SNR vs. E_b/n_0**

$$\frac{S}{N}=\frac{E_b}{n_0}\frac{R_b}{B}$$

- **Spectral efficiency**
  - Depends on modulation format
  - Typically 1–4 (Lecture 3)

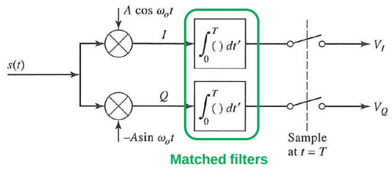
Theoretical bit error probabilities

- **Assuming Gaussian noise, the error probabilities (P_e = BER) can be derived theoretically:**

	Coherent demodulation	Envelope demodulation
ASK	$P_e=\frac{1}{2}\mathrm{erfc}\left(\sqrt{\frac{E_b}{4n_0}}\right)$	$P_e=\frac{1}{2}e^{-r_0^2/2\sigma^2}+\frac{1}{2}\int_0^{r_0}\frac{r}{\sigma^2}e^{-\left(V^2+r^2\right)/2\sigma^2}I_0\left(\frac{Vr}{\sigma^2}\right)dr$
FSK	$P_e=\frac{1}{2}\mathrm{erfc}\left(\sqrt{\frac{E_b}{2n_0}}\right)$	$P_e=\frac{1}{2}e^{-E_b/2n_0}$
PSK	$P_e=\frac{1}{2}\mathrm{erfc}\left(\sqrt{\frac{E_b}{n_0}}\right)$	Not possible

$$\mathrm{erfc}\,x=\frac{2}{\pi}\int_x^{\infty}e^{-x^2}dx$$

QPSK demodulator

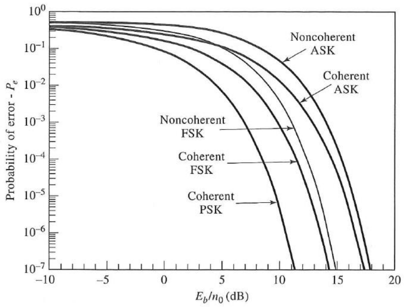


- **Error probability**

$$P_e=\frac{1}{2}\mathrm{erfc}\left(\sqrt{\frac{E_b}{n_0}}\right)$$

- **Same as for BPSK**
  - QPSK has twice the bitrate compared to BPSK
  - QPSK has twice the power (cos + sin modulation)

Comparison of binary modulation techniques

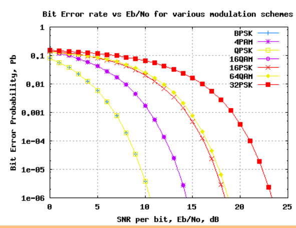


Modulation Method	E_b/n_0 for $P_e = 10^{-5}$
coherent PSK	9.6 dB
coherent FSK	12.6 dB
noncoherent FSK	13.4 dB
coherent ASK	15.6 dB
noncoherent ASK	16.4 dB

- **ASK**
  - Very simple demodulator
  - Common in optical systems
- **FSK**
  - Acceptable performance even with envelope demod
- **PSK**
  - Highest performance
  - Requires synchronized oscillator
  - Used in GPS

Performance comparison for digital modulation formats

Modulation Type	E_b/n_0 (dB) for $P_e = 10^{-5}$	Bandwidth Efficiency
binary ASK	15.6	1
binary FSK	12.6	1
binary PSK	9.6	1
QPSK (4-QAM)	9.6	2
8-PSK	13.0	3
16-PSK	18.7	4
16-QAM	13.4	4
64-QAM	17.8	6

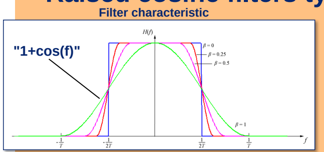


- **There may be other tradeoffs**
  - PSK has constant amplitude → High transmitter efficiency and small linearity problems
  - Receiver complexity
- **Clearly, there is a tradeoff between bandwidth efficiency and E_b/n_0 required. How can this be understood?**

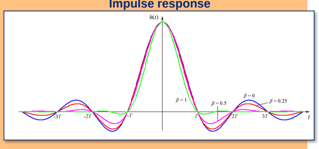
Pulse shaping filters

- **Reduces the spectral widening of a digital pulse train without distorting the message**
  - More efficient spectral usage
  - Designed to have zero inter-symbol-interference
- **Raised cosine filters typically used**

Filter characteristic



Impulse response



  - Roll-off factor parameter (α or β, typ = 0.25)
  - Often split into two root-raised cosine filters (at TX and RX)
- **How is the transmitted signal affected by the filtering?**

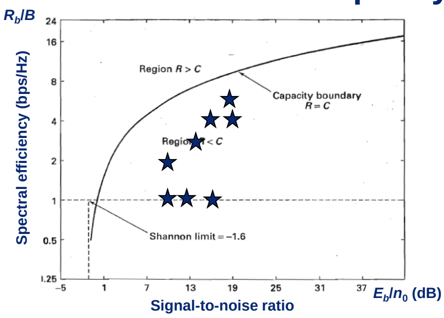
Channel capacity

- **Above a certain E_b/n_0, the bit error probability→0**
- **The corresponding maximum bitrate for error-free transmission is given by:**
- **Shannon's channel capacity theorem**

$$C=B\log_2\left(1+\frac{S}{N}\right)=B\log_2\left(1+\frac{E_b}{n_0}\frac{R_b}{B}\right)$$

- C Max. bitrate for error free transmission (bps)
- B Channel bandwidth (Hz)
- S Signal power (W)
- N Noise power = $n_0 \cdot B$ (W)
- R_b Bitrate (bps)
- E_b Bit energy (J)

Channel capacity



Modulation Type	E_b/n_0 (dB) for $P_e = 10^{-5}$	Bandwidth Efficiency
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binary PSK	9.6	1
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64-QAM	17.8	6

- **Upper limit of spectral efficiency for a given E_b/n_0**
- **Compare w. M-QAM for $P_e = 10^{-5}$**
  - **Higher order modulation schemes utilize spectrum better, but require higher SNR = higher output power**

2 Wireless

2.1 Antennas

Far-field:
Big antennas:

$$R_{ff} = \frac{2 \cdot D}{\lambda}$$

Small antennas:

$$R_{ff} = 2\lambda$$

Other equations:

$$P_r = S_{avg} \cdot A_e$$

(S_{avg} power density, A_e effective aperture area)

$$A_e = \frac{D \cdot \lambda^2}{4\pi}$$

(D directivity)

$$G = e_{rad} \cdot D$$

$$e_{rad} = \frac{P_{rad}}{P_{in}} = 1 - \frac{P_{loss}}{P_{in}}$$

Antenna temperature:

$$T_A = T_b e_{rad} + (1 - e_{rad})T_p$$

Friis equation:

$$P_r = P_t \frac{G_t G_r \lambda^2}{(4\pi R)^2}$$

Modified Friis equation:

$$P_r = P_t \frac{G_t G_r \lambda^2}{(4\pi R_0)^2} \cdot \left(\frac{R_0}{R}\right)^N$$

(R_0: first obstruction, 1-100 m in urban, 1 km in rural, for N, see table)
Reciever SNR:

$$SNR_{in} = \frac{G_r}{T_A} \cdot \frac{P_t G_t \lambda^2}{(4\pi R)^2}$$

Radar equation:

$$P_r = P_t \frac{G_1 G_2 \lambda^2}{(4\pi)^3 R^4} \sigma$$

2.2 Fading

Attenuation: Frequency dependent, see graph.

- Rain
- Atmosphere

Multi-path fading:

- Reyleigh fading (small scale, no LoS):

$$\ln(1 - \mathbf{P}_{outage}) = -\frac{P_{thr}}{P_0}$$

$$-\frac{P_{thr}}{P_0} = \text{Fade depth}$$

(P_0: average power)

- Rician fading (point to point link, direct path and stochastic reflections):

$$K = A_0/2\sigma^2$$

see graph for Fade depth.

- Log-normal fading (large scale, no LoS, multi-path):

$$\mathbf{P_{out}} = \frac{1}{2} + \frac{1}{2} erfc\left(\frac{P_{thr,dBm} - P_{0,dBm}}{\sigma_{dB} \cdot \sqrt{2}}\right)$$

2.3 Cascation

Noise figure:

$$F = F_1 + \frac{F_2 - 1}{G_1} + \frac{F_3 - 1}{G_1 \cdot G_2} + ...$$

Equivalent noise temperature:

$$T = T_1 + \frac{T_2}{G_1} + \frac{T_3}{G_1 \cdot G_2} + ...$$

Output power intercept point 3 (**OIP3** = **P3**!)

$$\frac{1}{P3} = \frac{1}{P3_N} + \frac{1}{P3_{N-1} \cdot G_N} + \frac{1}{P3_{N-2} G_N G_{N-1}} + ...$$

2.4 Power formulas

$$OIP3 = P_{1dB}^{out} - 10\text{dB}$$

$$OIP3 = IIP3 \cdot G$$

Linear dynamic range:

$$DR_l = \frac{P_{1dB}^{out}}{N_{out}}$$

($N_{out} = k(T_A + T_e)BG$, output noise power)
Spurious free dynamic range:

$$DR_{sf} = \left(\frac{OIP3}{N_{out}}\right)^{2/3}$$

2.5 Mixing

$$f_{RF/IM} = f_{LO} \pm f_{IF}$$

$$BW > 5\% f$$

Down or up-conversion need:

$$f_{IF} > \frac{f_{RF}}{40}$$

$$\left[f_{LO}^{(1)}, f_{LO}^{(2)}\right] - f_{IF} \neq \left[f_{IM}^{(1)}, f_{IM}^{(2)}\right] \text{ (no overlap)}$$

(also channels need to fit within filter bandwidths)

$$G_c = \frac{P_{out,RF}}{P_{in,IF}} \text{ or } G_c = \frac{P_{out,in}}{P_{in,RF}}$$

Overall mixers have nasty noise characteristics. Check the slides.

2.6 Oscillators

$$v_0 = V_0 \cos(\omega_0 t + \theta(t)), \;\; \theta(t) = \theta_p \sin(\Delta\omega t)$$

$$L(\Delta\omega) = \frac{P_{noise(SSB)}}{P_{carrier}} = \frac{\theta_p^2}{4} = \frac{\theta_{rms}^2}{2}$$

$$\frac{P_{noise(SSB)}}{P_{carrier}} = \int_0^{B/2} L(\Delta\omega) d\Delta f = L\frac{B}{2} = \frac{\theta_{rms}^2}{2}$$

Multiplier step to increase oscillator frequency affects noise by:

$$L_{VCO\times N} = \left(\frac{N\theta_{rms}}{2}\right)^2 = N^2 L_{VCO}$$

2.7 MIMO

$$E_{tot} = E_{element} \cdot AF$$

$$AF = \sum_{n=0}^{N-1} e^{jn(kd\cos(\theta)+\beta)}$$

$$< P > \propto \frac{E^2}{2} \propto \frac{N^2}{2}$$

2.8 Other

Geo-stationary orbit: 36000 km

Far-field assumption

- E and H fields approximate a planar wave front at far distance
- Far field distance:
  Deviation from planar wave front <22.5°

$$R_{ff} = \frac{2D^2}{\lambda}$$

Antenna "size"

- For small antennas $\boxed{R_{ff} = 2\lambda}$
  - Small dipoles and loops
- All calculations in this course will assume distances $>R_{ff}$

Antenna directivity, *D*

- The ability of an antenna to direct energy

$$D = \frac{\bar{S}_{\max}}{\bar{S}_{\text{avg}}} = \frac{4\pi r^2 \bar{S}_{\max}}{P_{rad}} = \frac{4\pi \bar{S}_{\max}}{\int_{\phi=0}^{2\pi}\int_{\theta=0}^{\pi} \bar{S}(\theta,\phi)\sin\theta \, d\theta \, d\phi}$$

- Directivity of isotropic antenna
  - Assume $S(\theta,\phi) = S_{max} = 1 \Rightarrow D_{isotropic} = 1 = 0$ dB
  - Directivity usually expressed in dBi
- Narrow beam ↔ high directivity
- Approximate expression for narrow beams

$$D \cong \frac{\pi^2}{\theta_1\theta_2}$$

θ_1 and θ_2 in radians

Effective aperture area, A_e

- A receiver antenna captures EM radiation from a certain area, the effective aperture area

$$P_r = S_{avg} A_e$$

- Capture area and directivity are related
  - Natural if we consider reciprocity of antennas

$$A_e = \frac{D\lambda^2}{4\pi}$$

- Effective aperture area is often close to physical aperture area

Radiation efficiency and antenna gain

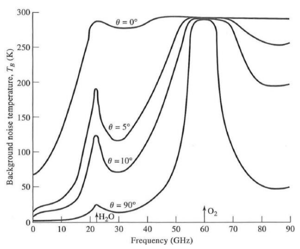
- Radiation efficiency, e_{rad}
  - Amount of power radiated, in relation to the power inserted into the antenna
  - Resistive losses
  - Dielectric losses

$$e_{rad} = \frac{P_{rad}}{P_{in}} = \frac{P_{in} - P_{loss}}{P_{in}} = 1 - \frac{P_{loss}}{P_{in}}$$

- Antenna gain, *G*
  - Directivity, accounting for the losses
  - Suitable for system calculations

$$G = e_{rad} D$$

Background noise temperature, T_B

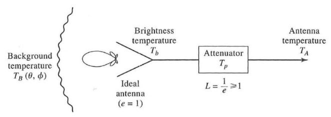


- Noise power

$$\boxed{N_0 = kTB}$$

  - *k*: 1.38 × 10^23
  - *B: Bandwidth*
  - *T: Noise temperature*
- Noise temperature vs. frequency for different antenna angles
  - $\theta = 0°$: Horizon
  - $\theta = 90°$: Towards sky
- How should the figure be interpreted?

Antenna noise temperature, T_A



- Contributions from losses and background
- Antenna sees background temperature, T_B
  - Weighted with antenna pattern, yields brightness temp, T_b
- Antenna losses
  - Expressed by radiation efficiency, e_{rad}, where $G = e_{rad}D$
  - Antenna held at physical temperature, T_p
- Overall equivalent antenna temperature, T_A

$$T_A = e_{rad} T_b + (1 - e_{rad}) T_p$$

Properties of Friis equation

$$\boxed{P_r = P_t \frac{G_t G_r \lambda^2}{(4\pi R)^2}}$$

- Assumes free space propagation
  - Applicable for point-to-point links, satellite communication etc.
- Received power is proportional to $1/R^2$
  - In practice, $1/R^N$, where $1.5 < N < 6$
    - When could $N < 2$ be possible?
  - Still better than cable losses for long distance
    - Loss proportional to $e^{\alpha R}$
  - More detailed and realistic models presented next lecture

Modified Friis equation

- Modified Friis equation for generic path loss exponent, *N*

$$\boxed{P_r = P_t \frac{G_t G_r \lambda^2}{(4\pi R_0)^2} \frac{1}{(R/R_0)^N}}$$

- Free space propagation until a distance R_0
  - R_0 = (representative) distance to first obstruction
    - Urban areas and indoors: R_0 = 1m - 100m
    - Rural areas: R_0 = 1km is a common assumption
- R_0 must be in far-field region

Realistic path loss

- **Path loss exponent depends on environment and assumptions**

$$P_r = P_t \frac{G_t G_r \lambda^2}{(4\pi R)^2}$$

Environment	Path Loss Exponent
Free space	2
Urban	2.7–3.5
Shadowed urban	3–5
In-building LOS	1.6–1.8
In-building shadowed	4–6
Factory shadowed	2–3
Retail store	2.2
Office—soft partitions	2.4

- **Path loss exponent need often to be characterized experimentally in practical situations**

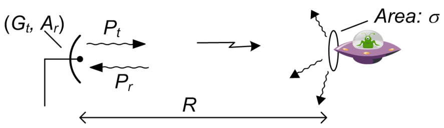
G/T

$$\frac{S_i}{N_i} = \frac{G_r G_t P_t \lambda^2}{kT_A B (4\pi R)^2} = \boxed{\frac{G_r}{T_A}} \frac{G_t P_t \lambda^2}{kB(4\pi R)^2}$$

Note:
Error in Pozar, eq (4.30)

- **G_r/T_A is the only parameter that can be controlled by the receiver design**
  - **Important figure-of-merit for receivers**
- **G/T usually expressed in units "dB/K"**
  - **G/T (dB/K) = $10 \cdot \log_{10}(G_r/T_A)$**
  - **For receivers: T_A may include the receiver noise temperature**
- **Improved by high increasing antenna gain**
  - **May not be allowed if an omni-directional coverage is desired**

The RADAR equation



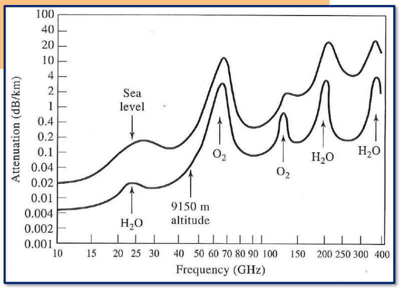
- **Finally, assuming identical transmit and receive antennas (using relation between gain and area):**

$$P_r = P_t \frac{G_t^2 \lambda^2}{(4\pi)^3 R^4} \sigma$$

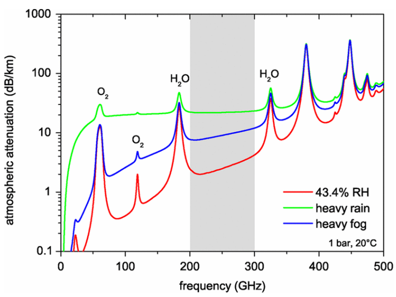
- **Very strong dependence on distance, *R***

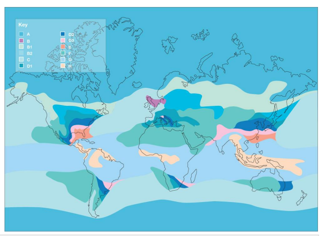
Attenuation effects

- **Loss may also be caused by losses in the propagation path**
  - **Atmospheric losses: O_2 and H_2O resonances**
  - **Precipitation**
  - **Walls**
  - **Ceilings**



Rain attenuation

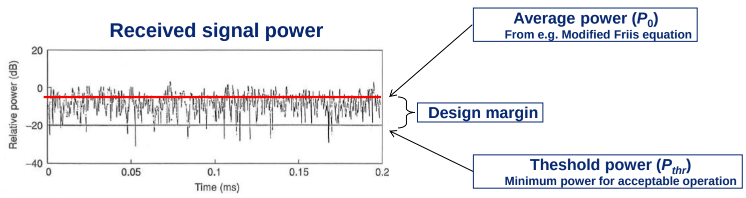




Rainfall Rate (R)

Percentage of Year (%)	A (mm/hr)	B (mm/hr)	B1 (mm/hr)	B2 (mm/hr)	C (mm/hr)	D1 (mm/hr)	D2 (mm/hr)	D3 (mm/hr)	E (mm/hr)	F (mm/hr)	G (mm/hr)	H (mm/hr)
1	0.2	1.2	0.8	1.4	1.8	2.2	3.0	4.6	7.0	0.6	8.4	12.4
0.5	0.5	2.0	1.5	2.4	2.9	3.8	5.3	8.2	12.6	1.4	13.2	22.6
0.1	2.5	5.7	4.5	6.8	7.7	10.3	15.1	22.4	36.2	5.3	31.3	66.5
0.05	4.0	8.6	6.8	10.3	11.5	15.3	22.2	31.6	50.4	8.5	43.8	97.2
0.01	9.9	21.1	16.1	25.8	29.5	36.2	46.8	61.6	91.5	22.2	90.2	209.3
0.005	13.8	29.2	22.3	35.7	41.4	49.2	62.1	78.7	112.0	31.9	118.0	283.4
0.001	28.1	52.1	42.6	63.8	71.6	86.6	114.1	133.2	176.0	70.7	197.0	542.6

Rayleigh fading: Received power vs. time



Rayleigh fading: Outage probability

- **Practical systems have a minimum acceptable power threshold level for proper operation, P_{thr}**
- **The outage probability, p_{out}, of the received power being $< P_{thr}$ is then given by**

$$p_{out} = \Pr(P < P_{thr}) = \int_0^{P_{thr}} f_P(P)\, dP$$
$$= \int_0^{P_{thr}} \frac{1}{P_0} e^{-P/P_0} dP = 1 - e^{-P_{thr}/P_0}$$

Rician fading: Received signal power

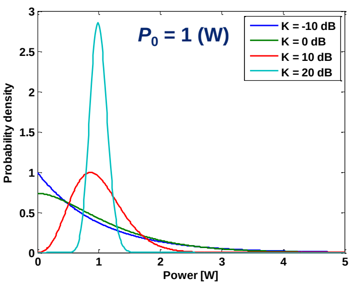
- **The direct path results in the Gaussian variables *X* and *Y* in Rayleigh fading having non-zero mean**
  - **Received power follows in this case a "non-central chi-square" distribution:**

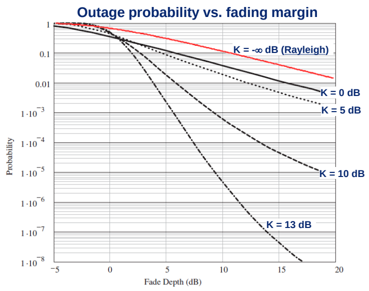
$$f_P(P) = \frac{(1+K)e^{-K}}{P_0} e^{-(1+K)P/P_0} I_0\left(\sqrt{\frac{4K(1+K)P}{P_0}}\right)$$

  - **I_0 is the modified Bessel function of first kind**
  - **P_0 is the average, total power received**
  - ***K* is a parameter relating the power of the line of sight component to the stochastic components:** $K = \frac{A_0^2}{2\sigma^2}$

Rician fading: Received signal power

- Probability density function for received power
- Converges to Rayleigh when $K \to 0$ (= $-\infty$ dB)
  - Explain figure for K → ∞!





Log-Normal fading: Outage probability

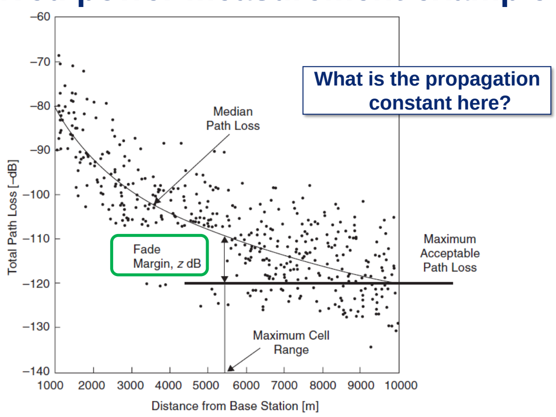
- The outage probability, p_{out}, of the received power being $<P_{thr,dBm}$ is then given by

$$p_{out} = \Pr\left(P_{r,dBm} < P_{thr,dBm}\right) = \int_{-\infty}^{P_{thr,dBm}} f\left(P_{r,dBm}\right) dP_{r,dBm}$$

$$= \int_{-\infty}^{P_{thr,dBm}} \frac{1}{\sigma_{dB}\sqrt{2\pi}} \exp\left(-\frac{\left(P_{r,dBm} - P_{0,dBm}\right)^2}{2\sigma_{dB}^2}\right) dP_{r,dBm}$$

$$= 0.5 + 0.5 erf\left(\frac{P_{thr,dBm} - P_{0,dBm}}{\sigma_{dB}\sqrt{2}}\right)$$

Log-Normal fading: Received power measurement example



Source: Saunders - "Antennas and propagation for wireless communication systems", Wiley & Sons, 1999.

Noise figure of cascaded components

- Generalized expression for noise temperature

$$T_e = T_{e1} + \frac{T_{e2}}{G_1} + \frac{T_{e3}}{G_1 G_2} + \cdots$$

- Generalized expression for noise figure

$$F = F_1 + \frac{F_2 - 1}{G_1} + \frac{F_3 - 1}{G_1 G_2} + \cdots$$

Intercept point of cascaded components

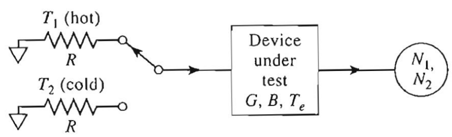
- General expression for output intercept point of cascaded stages

$$P_3 = \left(\frac{1}{G^{(2)}G^{(3)}\cdots G^{(N)}P_3^{(1)}} + \cdots + \frac{1}{G^{(N-1)}G^{(N)}P_3^{(N-2)}} + \frac{1}{G^{(N)}P_3^{(N-1)}} + \frac{1}{P_3^{(N)}}\right)^{-1}$$

$G^{(i)}$: Gain of i^{th} stage
$P_3^{(i)}$: Output intercept point of i^{th} stage

- Compare with expression for cascaded noise
- Influence scales inversely with gain in front of it
  - In transmitters, the last stage typically dominates ($P_3 \approx P_3^{(N)}$)

Noise temperature measurements The Y-factor method



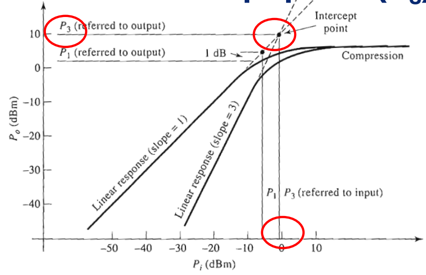
- An Y-factor is defined as

$$Y = \frac{N_1}{N_2} = \frac{T_1 + T_e}{T_2 + T_e}$$

- The unknown noise temperature is then

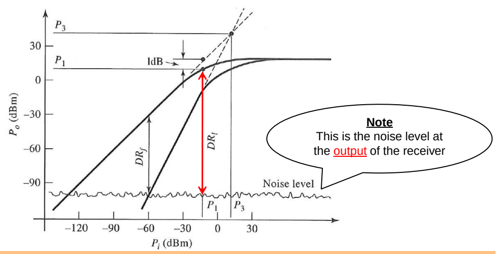
$$T_e = \frac{T_1 - YT_2}{Y - 1}$$

3rd order intercept point (P_3)



- Extrapolation of linear response and IM3 signals yield an intercept point (IP3 = P_3 in Pozar)
  - Referenced to input power: IIP3
  - Referenced to output power: OIP3
- Approximate relation: $P_3 \approx P_1$ + 10 dB

Linear dynamic range, DR_l

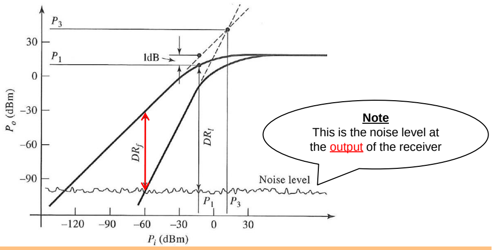


- Linear dynamic range
  - Power range between noise at low power and 1-dB compression point at high power
  - Referenced to output:

$$DR_l = \frac{P_1}{N_o}$$

P_1: 1-dB compression point (output)
N_o: Noise power at output, typ. $k(T_e+T_0)BG$

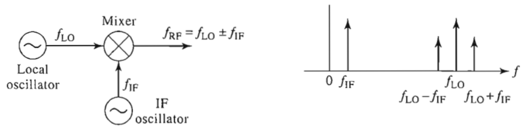
Spurious free dynamic range, DR_f



- **Spurious free dynamic range**
  - Power range where signal is higher than noise floor and distortion levels are below noise floor

$$DR_f = \left.\frac{P_{\omega 1}}{P_{2\omega 1-\omega 2}}\right|_{P_{2\omega 1-\omega 2}=N_0} = \left\{P_{2\omega_1-\omega_2} = \left(P_{\omega_1}\right)^3 / \left(P_3\right)^2 = N_0\right\} = \left(\frac{P_3}{N_0}\right)^{2/3}$$

Up conversion mixer



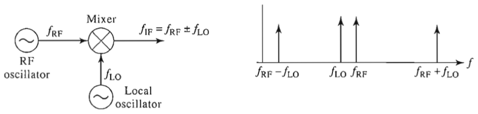
- **RF output signal**
  - Assuming that the mixer is a multiplying device

$$v_{RF}(t) = K \cdot v_{IF}(t) \cdot v_{LO}(t) = K\cos(2\pi f_{LO} t)\cos(2\pi f_{IF} t)$$
$$= \frac{K}{2}\left[\cos(2\pi(f_{LO}-f_{IF})t) + \cos(2\pi(f_{LO}+f_{IF})t)\right]$$

- **Output signal has spectral components at**

$$f_{RF} = f_{LO} \pm f_{IF}$$

Down conversion mixer



- **IF output signal**
  - Assuming that the mixer is a multiplying device

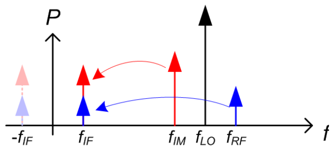
$$v_{IF}(t) = K \cdot v_{RF}(t) \cdot v_{LO}(t) = K\cos(2\pi f_{RF} t)\cos(2\pi f_{LO} t)$$
$$= \frac{K}{2}\left[\cos(2\pi(f_{LO}-f_{RF})t) + \cos(2\pi(f_{LO}+f_{RF})t)\right]$$

- **Desired output frequency is:**

$$f_{IF} = f_{RF} - f_{LO}$$

The other component (f_{RF} + f_{LO}) is easy to filter away (lowpass filter)

Image frequency in receivers

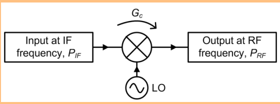


- **Two RF frequencies produce the same IF output frequency in down-conversion mixers**
  - The desired frequency: $f_{RF} = f_{LO} + f_{IF}$
  - The image frequency: $f_{IM} = f_{LO} - f_{IF}$

Mixer conversion gain (or loss)

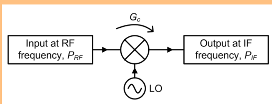
- **Conversion gain (or loss) specifies the ratio of output signal power to the power available at the input**
- **For up conversion (transmitter):**

$$G_c = \frac{P_{out,RF}}{P_{in,IF}}$$

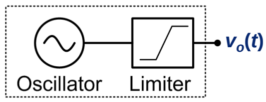


- **For down conversion (receivers):**

$$G_c = \frac{P_{out,IF}}{P_{in,RF}}$$



Phase noise representation



- **General oscillator output signal**

$$v_o(t) = V_0\left[1 + A(t)\right]\cos\left[\omega_0 t + \theta(t)\right] \approx V_0\cos\left[\omega_0 t + \theta(t)\right]$$

  - $A(t)$: Amplitude modulation (AM) noise
  - AM noise can be suppressed by amplitude limiter
  - $\theta(t)$: Phase modulation (PM) noise = phase noise
- **Phase noise at offset $\Delta\omega$**
  - Represented as a frequency modulation ($\Delta\omega$) of carrier frequency

$\theta(t) = \theta_p \sin \Delta\omega t$ θ_p = Peak phase deviation

Phase noise representation

- **Resulting output signal, assuming small θ_p**

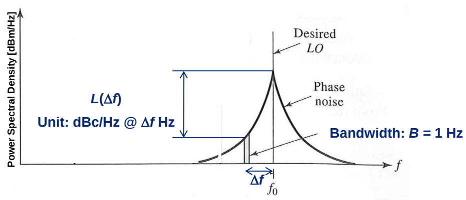
$$v_o(t) = V_0\left\{\cos\omega_0 t - \frac{\theta_p}{2}\underbrace{\left[\cos(\omega_0+\Delta\omega)t - \cos(\omega_0-\Delta\omega)t\right]}_{\text{Phase noise sidebands at } \omega_0 \pm \Delta\omega}\right\}$$

- **Phase noise at $\Delta\omega$: $L(\Delta\omega)$:**

$$L(\Delta\omega) = \frac{P_{noise(SSB)}}{P_{carrier}} = \frac{\frac{1}{2}\left(V_0\theta_p/2\right)^2}{V_0^2/2} = \frac{\theta_p^2}{4} = \frac{\theta_{rms}^2}{2}$$

  - Only one sideband is used in the definition
- **Note that the phase noise power scales with the received signal power**
  - Compare with thermal noise and shot noise...

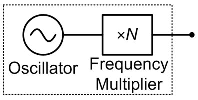
Oscillator output spectrum



- **Noise power spectral density (W/Hz) relative to total oscillator power (W)**
  - $L(\Delta f) = (P_{noise(SSB)}/B) / P_{carrier}$
- **Phase noise within channel bandwidth (B) is added at receiver output**

$$P_{noise,SSB}/P_{carrier} = \int_0^{B/2} L(\Delta f)\, d\Delta f = L\,B/2 = \frac{\theta_{rms}^2}{2}$$

Frequency multiplication



- **Frequency multiplier = Phase multiplier**

$$v_o(t) = V_0 \cos\left[\omega_0 t + \theta(t)\right] \rightarrow v_o(t) = V_0 \cos\left[N\omega_0 t + N\theta(t)\right]$$

... and therefore a phase noise multiplier

$$\theta_{rms} \rightarrow N\theta_{rms}$$

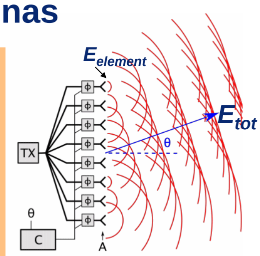
- **Phase noise (*L*) after multiplier**

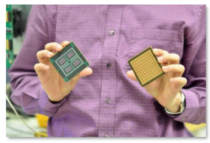
$$L_{VCO \times N} = \frac{\left(N\theta_{rms}\right)^2}{2} = N^2 L_{VCO}$$

  - **Degradation of phase noise with $20\log_{10}(N)$ dB**

Phased array antennas

- **An array of antennas transmitting phase-delayed copies**
  - **Main beam direction controlled by phase shift**
  - **Electrical phase-shifters → fast scanning**
- **Applications**
  - **Electrically scanned radars**
  - **Directed communication (5G)**
  - **Satellite communication**
- **$E_{tot}(\theta,\varphi) = E_{element}(\theta,\varphi)$*AF$(\theta,\varphi)$**
- **AF = array factor**



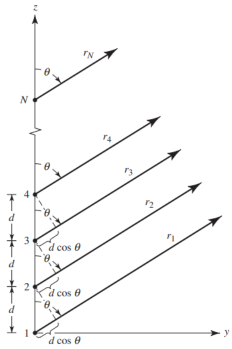


Phased array antennas

- **Example: Linear array**

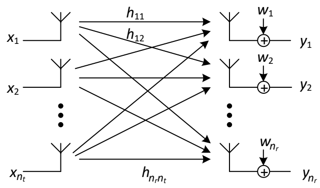
$$AF = \sum_{n=0}^{N-1} e^{jn(kd\cos\theta+\beta)}$$

  - ***N* = Number of antenna elements**
  - ***k* = wave number**
  - ***d* = antenna separation**
  - ***β* = Phase progression between antennas**
- **max(AF) = *N* → $E_{tot,max} = E_{element}$**N***
- **Multiplication of antenna gain with N^2**



[Balanis]

MIMO system model



$$\begin{bmatrix} y_1 \\ \vdots \\ y_{n_r} \end{bmatrix} = \begin{bmatrix} h_{11} & \cdots & h_{1n_t} \\ \vdots & \ddots & \vdots \\ h_{n_r 1} & \cdots & h_{n_r n_t} \end{bmatrix} \begin{bmatrix} x_1 \\ \vdots \\ x_{n_t} \end{bmatrix} + \begin{bmatrix} w_1 \\ \vdots \\ w_{n_r} \end{bmatrix}$$

- **Generic model of MIMO system**

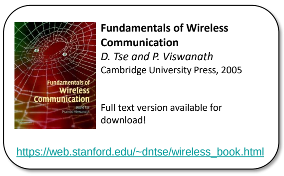
$$\mathbf{y} = \mathbf{Hx} + \mathbf{w}$$

x = Transmitted signals

H = Channel matrix

w = Receiver noise

y = Received signals



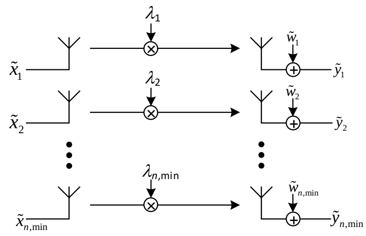
Singular value decomposition (SVD) of H

$$\mathbf{H} = \mathbf{U}\mathbf{\Lambda}\mathbf{V}^*$$

- **$U^{(nr,nr)}$ = Receiver rotation matrix ; $UU^* = I$**
- **$V^{(nt,nt)}$ = Transmitter rotation matrix ; $VV^* = I$**
- **$\Lambda^{(nmin,nmin)} = \mathrm{diag}(\lambda_1, \lambda_2, \ldots, \lambda_{n,min})$**
  - **diagonal matrix of eigenvalues**
  - **$n_{min} = \mathrm{rank}(H) \le \min(n_t, n_r)$**

Data multiplexing

$$\mathbf{y} = \mathbf{U}\mathbf{\Lambda}\mathbf{V}^*\mathbf{x} + \mathbf{w}$$
$$\mathbf{U}^*\mathbf{y} = \mathbf{\Lambda}\mathbf{V}^*\mathbf{x} + \mathbf{U}^*\mathbf{w}$$
$$\left.\begin{aligned} \tilde{\mathbf{x}} &= \mathbf{V}^*\mathbf{x} \\ \tilde{\mathbf{y}} &= \mathbf{U}^*\mathbf{y} \\ \tilde{\mathbf{w}} &= \mathbf{U}^*\mathbf{w} \end{aligned}\right\} \Rightarrow \tilde{\mathbf{y}} = \mathbf{\Lambda}\tilde{\mathbf{x}} + \tilde{\mathbf{w}}$$



- **A transformed system with n_{min} independent paths**
- **Power in each path scales with the eigenvalue**

3 Photonics

(n_1:core, n_2:cladding)

$$\Delta = \left(\frac{n_{core} - n_{clad}}{n_{core}}\right)$$

$$NA = \sin(\theta_{i,max} n_0 = \sqrt{n_1^2 - n_2^2} \approx n_1\sqrt{2\Delta}$$

Typical fiber has: $n_1 = 1.5$, $\Delta = 0.1\%$.

3.1 Dispersion

$$\sigma_{\tau,step} \simeq \frac{L\Delta n_1}{2c}$$

$$\sigma_{\tau,grin} \simeq \frac{L\Delta^2 n_1}{4c}$$

$$\sigma_{\tau,step} \simeq D\sigma_\lambda L$$

$$\sigma_\lambda = \Delta\lambda = \Delta f \frac{\lambda^2}{c}$$

$$D = 16 \text{ ps/nm/km @1.55µm}$$

$$\sigma_{sys}^2 = \sigma_{TX}^2 + \sigma_\tau^2 + \sigma_{RX}^2$$

$$\sigma_{TX/RX}^2 = \frac{1}{B^2}, \text{BW for device!}$$

We need to have for example:

$$\sigma_{sys} < 70\% \cdot T_{bit} = 70\% \cdot \frac{1}{R_{baud}}$$

$$\sigma_\tau < \frac{T}{4} = \frac{1}{4B_{signal}}$$

3.2 Noise, SNR, Q-value

$$SNR = \frac{I_p^2}{\sigma^2} = \frac{(R \cdot P_{in})^2}{\sigma_T^2 + \sigma_S^2 + \sigma_{s-sp}^2 + \sigma_{sp-sp}^2}$$

$$Q = \frac{I_1 - I_0}{\sigma_1 + \sigma_0}$$

$$Q > 6 \Leftrightarrow BER < 10^{-9}, Q > 7 \Leftrightarrow BER < 10^{-12}$$

(for OOK)
Shot noise:

$$\sigma_S^2 = 2qI_pB = 2qRP_{in}\Delta f$$

(P_{in} on photo diode,$P_{in} = PLO$ for coherent)

(R responsivity)

$$R = \eta\frac{q}{h\nu} = \eta\frac{\lambda[\text{µm}]}{1.24}$$

Thermal noise:

$$\sigma_T^2 = \frac{4kT}{R_L}B = \frac{4kT}{R_L}\Delta f$$

($\sigma_T^2 \approx \sigma_S^2$@$P_{in} = 0.83$mW@1.55µm)
Signal spontaneous noise:

$$\sigma_{s-sp}^2 = 4R^2 P_{in} S_{sp}\Delta f$$

(S_{sp} spectral noise density)

$$S_{sp} = \frac{F_n}{2} Gh\nu \cdot N$$

(balanced detector, for coherent $P_{in} = 2\sqrt{P_s P_{LO}}$)
Spontaneous spontaneous beat noise:

$$\sigma_{sp-sp}^2 = 4R^2 S_{sp}^2 \Delta\nu\Delta f$$

$$OSNR = \frac{P_{sig}}{P_{ASE}} = \frac{P_{sig}}{S_{sp}\Delta\nu \cdot N} = \frac{P_{sig}}{\frac{F_n}{2}Gh\nu\Delta\nu \cdot N}$$

Coherent receiver power:

$$P(t) = P_s + P_{LO} + 2\sqrt{P_s P_{LO}} \text{@ photo diode}$$

Pre-amplified reciever (consists of a optical amplifier followed by a optical pass band filter followed by a photo diode detector for example):

$$P_{in} = h\nu F_n B\left(Q^2 + Q\sqrt{\frac{\Delta\nu}{\Delta f}}\right)$$

(or just use dominating noise)

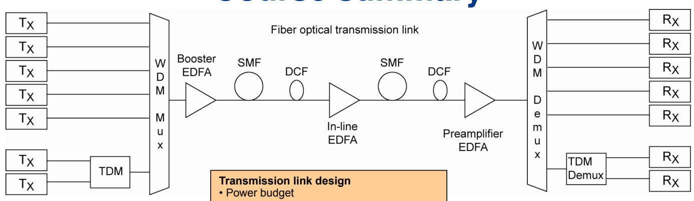
3.3 Other

Bandwidth:

$$\Delta\lambda = \Delta f\frac{\lambda^2}{c}$$

C-band: $\Delta\lambda = 35$ nm, $\Delta f = 4{,}37$ THz.

Course summary



Transmission link design
- Power budget
  - Launched power, link attenuation, power margin, receiver sensitivity ➔ system limitations
- Rise time budget
  - Fiber types, modal dispersion, chromatic dispersion (GVD), receiver and transmitter bandwidths, dispersion compensation ➔ system limitations

Transmitters
- Laser sources
- Direct modulation
- External modulation
- Modulation formats
  - OOK
  - Differentially coded signals (DPSK)
  - Coherent modulation

Receiver design
- Detectors (PIN and APD)
- Detector performance (quantum efficiency, responsivity, bandwidth)
- Receiver noise (shot noise, thermal noise, ASE)
- Receiver sensitivity, BER, Q-value

Multiplexing
- Time division multiplexing (TDM) ➔ high channel bit rate
- Wavelength division multiplexing (WDM) ➔ many independent wavelength channels

➔ **Very high capacity possible**

Optical amplification (EDFA)
- Booster amplifier, in-line amplifier, preamplifier
- EDFA design (Erbium ions, pumping, stimulated emission)
- EDFA performance (gain, bandwidth, noise figure)
- ASE noise, spontaneous emission, beat noise
- Cascaded EDFAs, effective noise figure, SNR degradation, SNR limited transmission

Ray-optics description of step-index fiber

Apply Snell's law at the input interface:

$$n_0 \sin(\theta_i) = n_1 \sin(\theta_r)$$

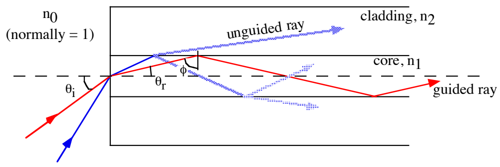
For total internal reflection at the core/cladding interface we have a critical, minimum, angle:

$$n_1 \sin(\phi_c) = n_2 \sin(90°) \rightarrow \sin(\phi_c) = n_2/n_1$$

Relate to maximum entrance angle:

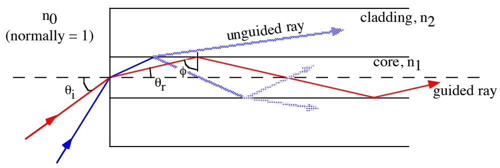
$$n_0 \sin(\theta_{i,\max}) = n_1 \sin(\theta_{r,\max}) = n_1 \sin(90 - \theta_c) =$$

$$n_1 \cos(\theta_c) = n_1 \sqrt{1 - \sin(\theta_c)^2} = \sqrt{n_1^2 - n_2^2}$$



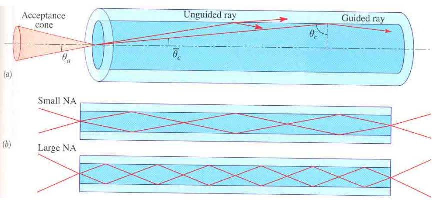
Δ is the index difference
$\Delta = (n_1 - n_2)/n_1 \ll 1$
$\Delta \approx$ 1-3% for MM fibers,
$\Delta \approx$ 0.1-1% for SM fibers

Numerical aperture



- The numerical aperture, NA, is a measure of the light gathering power of an optical system, originating from microscopy.
- For fibers it is defined as

$$NA = n_0 \sin\theta_{i,\max} = \sqrt{n_1^2 - n_2^2} \approx n_1\sqrt{2\Delta}$$



Transmitters - overview

- Directly driven LED
  - Low cost
  - Incoherent (high linewidth) source
  - High coupling losses due to weak directionality
  - MMF
  - Mb/s links, OOK

- Directly modulated Semiconductor Laser
  - Higher cost
  - Coherent (low linewidth) source
  - Less coupling losses due to stronger directionality of radiation
  - MMF or SMF
  - ≤ 10-100 Gb/s link, OOK

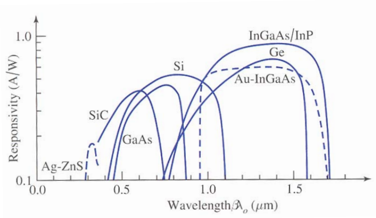
- Externally modulated Semiconductor Laser
  - High cost
  - Coherent (low linewidth) source
  - Less coupling losses due to stronger directionality of radiation
  - SMF
  - ≥ 100 Gb/s links, QPSK

Affordable - Expensive

Photodetectors for fiber systems - overview

- Responsivity increases linearly with wavelength...
- ...until the photon energy is below the bandgap energy of the semiconductor.
- 870 nm range
  - Si p-i-n or APD
    - low cost
    - low noise
- 1300-1600 nm range
  - InGaAs, Ge, InP
    - InGaAs APD common in commercial systems (10 Gb/s per WDM channel)
    - InGaAs/InP p-i-n common in high bandwidth systems (> 40 Gb/s per channel)

$$R = \frac{\eta q}{h\nu} = \frac{\eta q \lambda}{hc} \text{ if } h\nu > E_g$$



Material	hc/E_g $[\mu m]$
Si	1.15
Ge	1.88
GaAs	0.87
InAs	3.5

Responsivity

A requirement is that the detector material bandgap energy (E_g) must be smaller than the photon energy ($h\nu$).

The detector responsivity R is defined as the photo-current/incident optical power:

$$R = \eta \frac{q}{h\nu} = \eta \frac{\lambda}{1.24}$$

η = the quantum efficiency = the number of e-h pairs generated per photon (ideally, η = 100%), λ is wavelength in µm.

Material used: InGaAs (sometimes Ge), with typical data:

$\eta \approx 0.8$, $R \approx 1$ A/W, BW ≈ 10 GHz into 50 Ω, dark current ≈ 0.1 nA

The p-i-n characteristics

A photo diode p-n junction has the following characteristics

$$i = i_s(\exp(\frac{eV}{k_B T}) - 1) - RP_{\text{in}}$$

i_s is the "dark current"
$\Phi = RP_{in}$ denotes photoelectron flux

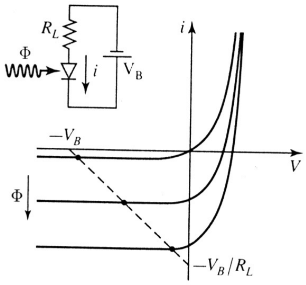
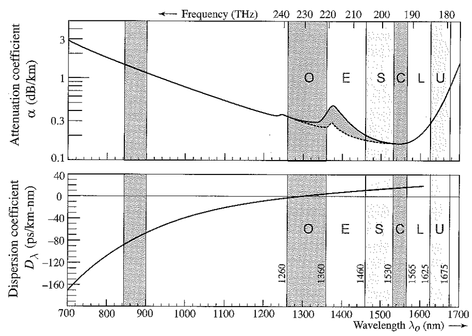


Photo diodes are usually operated strongly reverse-biased:

- reduces the carrier transit time - increases the speed (bandwidth)
- increases the linearity

The operating point is indicated by a dashed line

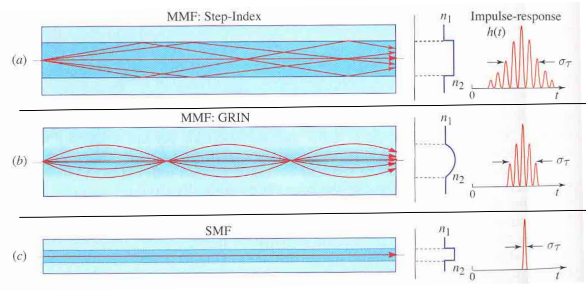
What wavelengths are used in fiber optics?



C: 'conventional', the conventional EDFA band
O: 'original', historically the first band used
S: 'short' wave EDFA band
L: 'long' wave EDFA band
U: 'ultra long' wave EDFA band
E: 'extended' wave band

- 850 nm: common in short MMF-based links
- 1550nm: most commonly used SMF wavelength
- 1300-1600: also used in SMF but less frequently

Dispersive broadening summary



$$\sigma_{\tau,\text{step}} \cong \frac{L\Delta n_1}{2c}$$

$$\sigma_{\tau,\text{GRIN}} \cong \frac{L\Delta^2 n_1}{4c}$$

$$\sigma_\tau = D\sigma_\lambda L$$

Rise time budget

- A time (or rise time) budget accounts for the dispersive broadening of the fiber, as well as the response times of the transmitter and receiver, σ_{tx} and σ_{rx}.
- Assuming Gaussian responses, a quadratic summation must be used.

Arbitrary but common criterion

$$\sigma_{sys}^2 = \sigma_{tx}^2 + \sigma_{rx}^2 + \sigma_\tau^2 \boxed{< (0.7\,T)^2 = \frac{0.49}{B^2}}$$

Often the fiber dispersion-induced rise time is designed to be less than:

$$\sigma_\tau = \frac{T}{4}$$

- How large is this in various fibers?

Signal-to-Noise Ratio (SNR)

The total noise variance is the sum of the noise contributions (since the noise sources are independent):

$$\sigma^2 = \sigma_s^2 + \sigma_T^2$$

The SNR of an electrical signal is simply defined as:

$$SNR = \frac{signal \;\; power}{noise \;\; power} = \frac{I_P^2}{\sigma^2}$$

Q-value

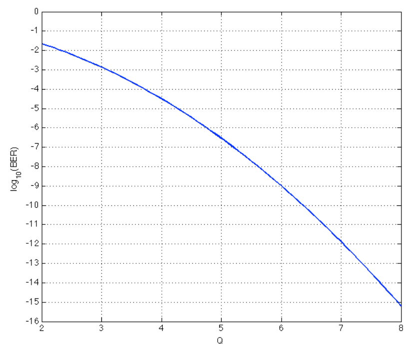
To calculate the BER in optical systems with OOK it is common to introduce the ***Q-value.*** Note that Q^2 is (approx.) proportional to the SNR.

$$\boxed{Q \equiv \frac{I_1 - I_0}{\sigma_1 + \sigma_0}}$$

Note: σ_1 and σ_0 may not be equal, for example due to shot noise! With optimum setting of the decision point, and Gaussian noise PDFs:

$$\boxed{BER \approx \frac{\exp\left(-Q^2/2\right)}{Q\sqrt{2\pi}}}$$

Q > 6 is needed for BER < 10^{-9}
Q > 7 is needed for BER < 10^{-12}



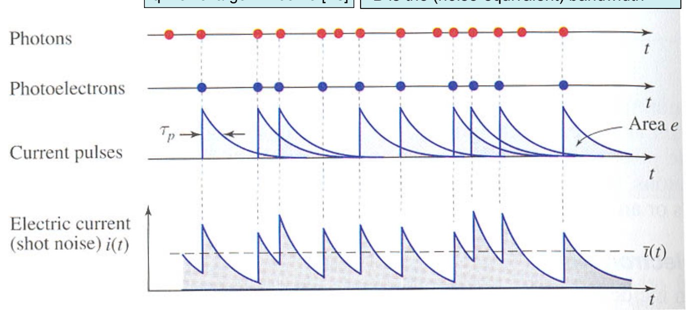
Shot noise

Shot noise arises from the particle nature of light (photons) and charges (electrons). The variance of the photo current I_p is (in units of [A²]):

$$\sigma_s^2 = 2qI_pB$$

q = el. charge = 1.6e-19 [As]

B is the (noise-equivalent) bandwidth



Thermal noise

(circuit noise, Johnson/Nyquist noise)

Thermal noise arises from random movements of conducting electrons due to the temperature. The noise current variance is (in units of [A²]):

$$\sigma_T^2 = \frac{4k_BT}{R_L}B$$

(Boltzmann's const. x Temp) [J], [Ws]

Load resistance [Ohm]

To account for electrical amplifiers generating excess thermal noise the above expression is sometimes modified to:

$$\sigma_T^2 = \frac{4k_BTF_n}{R_L}B$$

F_n is the electric amplifier noise figure (>1)

The optical beat noise terms

- The signal-spontaneous (s-sp) beat noise (signal beats with the noise) is proportional to the electrical bandwidth:

$$\sigma_{s-sp}^2 = 4R^2P_{in}S_{sp}\Delta f$$

ASE noise power spectral density [W/Hz]

el. BW [Hz]

- The spontaneous-spontaneous (sp-sp) beat noise (noise beats with itself) is proportional to also the optical noise bandwidth

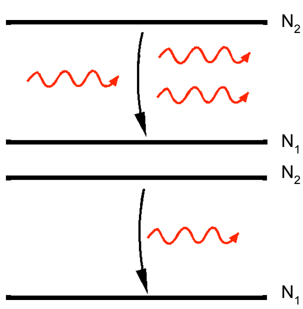
$$\sigma_{sp-sp}^2 = 4R^2S_{sp}^2\Delta\nu\Delta f$$

opt. BW [Hz]

- An optical bandpass filter can reduce the sp-sp part significantly.

Gain and noise in optical amplifiers

- Most amplifiers operate by the principle of stimulated emission
- Noise added through spontaneous emission is amplified along with the signal:
Amplified spontaneous emission (ASE)
- The population-inversion factor, n_{sp} depends on the population in the ground state (N_1) and in the excited state (N_2)



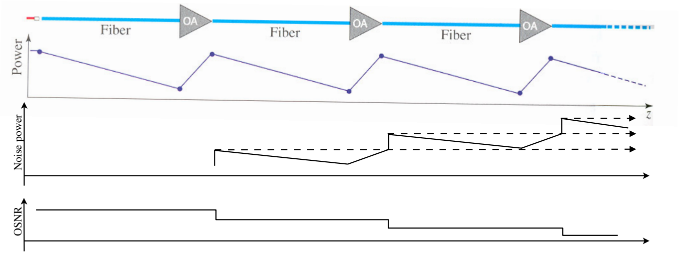
Optical power spectral density of ASE in an amplifier:

$$S_{sp}(\nu) = (G-1)n_{sp}h\nu = \frac{F_n}{2}Gh\nu \quad \text{[W / Hz]}$$

$$n_{sp} = \frac{N_2}{N_2 - N_1}$$

$$F_n = 2n_{sp}\frac{G-1}{G}$$

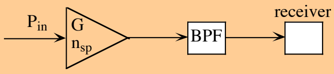
Optically amplified chains



- Each amplifier produces optical noise, *amplified spontaneous emission*, ASE, which propagates through the chain.
- The noise power adds incoherently and increases linearly with the number of amplifiers. $P_{ASE} = P_{ASE,1} + P_{ASE,2} + \ldots = Nh\nu \frac{F_n}{2} G\Delta\nu$
- The optical SNR, the OSNR, will decrease linearly. $OSNR = \frac{P_{sig}}{P_{ASE}}$

The preamplified receiver

One of the main applications for EDFAs is as an amplifier in front of the receiver, in order to overcome thermal noise.



The amplified signal entering the detector is $P_{amp} = GP_{in} + P_{sp}$

where $P_{sp} = S_{sp}\Delta\nu_{opt} = (G-1)n_{sp}h\nu\Delta\nu_{opt}$

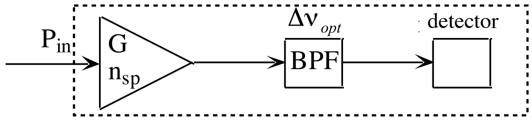
is the spontaneous emission noise power, and $\Delta\nu_{opt}$ is the bandwidth of the optical bandpass filter.

The corresponding photocurrent is

$$I = R|\sqrt{G}E_s + E_{sp}|^2 + i_s + i_T = R(GP_{in} + 2\sqrt{G}E_sE_{sp} + |E_{sp}|^2) + i_s + i_T$$

Signal field | ASE noise field | shot/ thermal noise | signal-spontaneous beating | spontaneous-spontaneous beating

The optically preamplified lightwave receiver



The power entering the detector is: $GP_{in} + S_{sp}\Delta\nu_{opt} =$

$= GP_{in} + (G-1)n_{sp}h\nu\Delta\nu_{opt}$

The input average power needed to reach a certain Q is:

$$P_{in} = h\nu F_n B(Q^2 + Q\sqrt{\frac{\Delta\nu_{opt}}{B}})$$

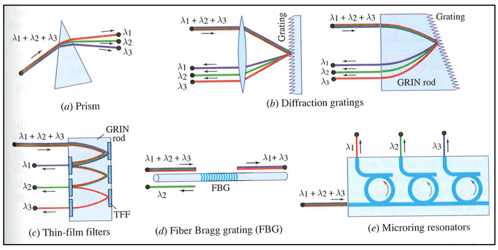
Derived from:
$$Q = \frac{RP_1 - RP_0}{\sigma_1 + \sigma_0} = \frac{RP_1}{\sqrt{\sigma^2_{s-sp} + \sigma^2_{sp-sp}} + \sigma_{sp-sp}}$$

Use a narrow optical filter (but not too narrow!) and a low-noise EDFA

At 10 Gb/s, BER = 10[-9] (Q=6) can be achieved with P_{in} = -40 dBm (100 nW)

Wavelength MUX/DEMUX

- A number of photonic structures can provide wavelength routing capability.
- The most popular is the Arrayed Waveguide Grating - AWG (fig 23.2-8).



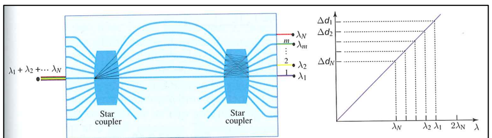
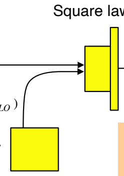


Figure 23.2-8 Wavelength-division demultiplexing by use of a wave-grating router (WGR).

Coherent detection - Basic Concept



$E_s = A_s e^{-j(\omega_0 t + \phi_s)}$ Input optical signal

$E_{LO} = A_{LO} e^{-j(\omega_{LO} t + \phi_{LO})}$ Local oscillator laser

$(E_s + E_{LO})^2 = A_s^2 + A_{LO}^2 + 2A_sA_{LO}\cos((\omega_0 - \omega_{LO})t + \phi_s - \phi_{LO})$

Corresponding optical quantities:

$P(t) = P_s + P_{LO} + 2\sqrt{P_sP_{LO}}\cos(\omega_{IF} t + \phi_s - \phi_{LO})$

$\omega_{IF} = \omega_0 - \omega_{LO}$

- Interference takes place in the photodiode
- Traditional square law detector, pin or APD
- Three cases,
  - homodyne detection ($\omega_{IF} \approx 0$)
  - heterodyne detection ($\Delta f_{det} > \omega_{IF} > \Delta f_{sig}$)
  - intradyne detection ($\omega_{IF} \approx 0$, tracked via digital signal processing)

Homodyne / Intradyne receiver

Detected signal current (I-quadrature)

$I \approx R(P_{LO} + 2Re(E_sE^*_{LO}))$

Dominating noise (if no optical amps):

LO shot noise: $\sigma_s^2 = 2qRP_{LO}\Delta f$

$$SNR = \frac{4R^2P_sP_{LO}}{2qRP_{LO}\Delta f} = 2\frac{P_s}{h\nu\Delta f}$$

Example: BPSK

$$Q = \frac{\mu_1 - \mu_0}{\sigma_1 + \sigma_0} = 2\sqrt{\frac{P_s}{2h\nu\Delta f}}$$

$\mu_1 = \mu_0 = 2R\sqrt{P_sP_{LO}}$

$\sigma_1 = \sigma_0 = 2qRP_{LO}\Delta f$

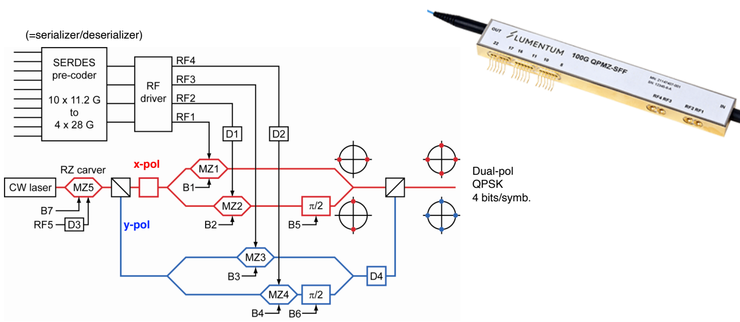
- High receiver gain is possible by using high-power LO
- Phase locking of LO required!
  - Usually done via DSP!
- Best possible receiver!

SNR for Homodyne and direct detection

	Homodyne / Intradyne	Direct Detection
Signal power:	$I(t)^2 = 4R^2P_sP_{LO}$	$I(t)^2 = R^2P_s^2$
Shot noise:	$\sigma_s^2 = 2qRP_{LO}\Delta f$ (LO-limited)	$\sigma_s^2 = 2qRP_s\Delta f$
	$SNR = \frac{I^2}{\sigma^2} = \frac{4R^2P_sP_{LO}}{\sigma_T^2 + 2qRP_{LO}\Delta f}$	$SNR = \frac{I^2}{\sigma^2} = \frac{R^2P_s^2}{\sigma_T^2 + 2qRP_s\Delta f}$
	For sufficiently large LO power the thermal noise can be neglected, and one finds (use R=q/hv):	For sufficiently large signal power the thermal noise can be neglected, and one finds (use R=q/hv):
	$SNR = \frac{I^2}{\sigma_s^2} = \frac{2P_s}{h\nu\Delta f} = 4\bar{m}$	$SNR = \frac{I^2}{\sigma_s^2} = \frac{P_s}{2h\nu\Delta f} = \bar{m}$

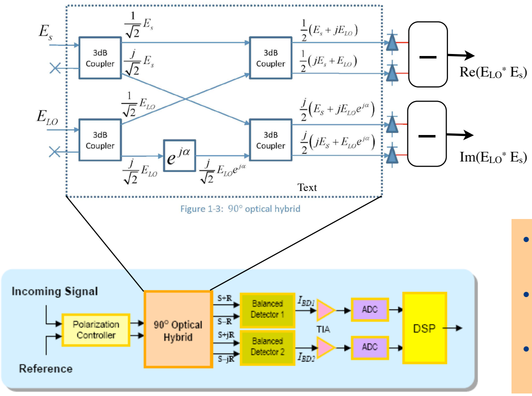
the average photon number per bit, using a Nyquist-limited bitrate B=Δf/2

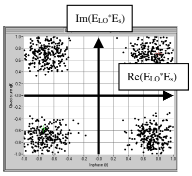
Coherent Transmitter



- A CW laser is split into 2 polarizations…
- …each modulated with dual-arm Mach-Zehnder modulators.
- 4 binary data signals, 6 bias voltages

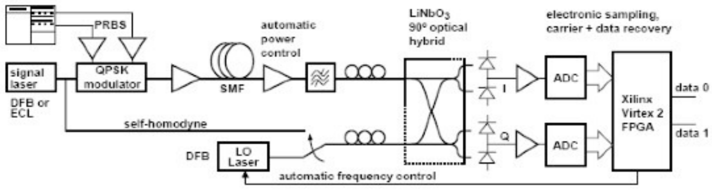
Quadrature Detection - 90-deg hybrid





- Enables coding in phase and/or frequency
- Fundamentally different from direct detection
- Well known from electrical communication

Example of Intradyne detection



After Pfau et al, Univ of Paderborn 2006

- The phase locking is done via electronic processing, in FPGA:s
- Similar performance as homodyne
- Complex electric signal is mapped to the electric domain
- Used in all modern coherent receivers (ASIC-based).

Digital Demodulation in optical receivers

Before the decisions, "undo" the effects of the channel.

For details: Seb J. Savory, IEEE J. Select. Topics Quantum Electron., 16, 5, 2010

This is a simplified list of all signal processing done by the DSP in a modern optical coherent receiver.

Subsystem	Functionality
ADC	Sample and convert data to digital domain
Static filtering	FIR filter to compensate for the chromatic dispersion
Dynamic filtering	Compensate polarization mixing and track polarization
Frequency estimation	Frequency offset compensation
Phase estimation	Phase drift compensation
Symbol decisions	Data recovery
FEC decoder	Decode words and correct errors

Nonlinear limits on transmitted power

The Kerr effect (cause of FWM, and other NL distortions)

- refractive index slightly power dependent
- induced nonlinear phase shift from power P over a fiber length L is

$$\phi = \gamma P L$$

- where $\gamma \approx 1.5$ [W^{-1} km^{-1}] in SMF is the nonlinear coefficient.
- the nonlinear phase shift arises during the first loss length, $1/\alpha \approx 20$ km after each amplifier.
- After N amplifiers accumulated nonlinear phase shift is

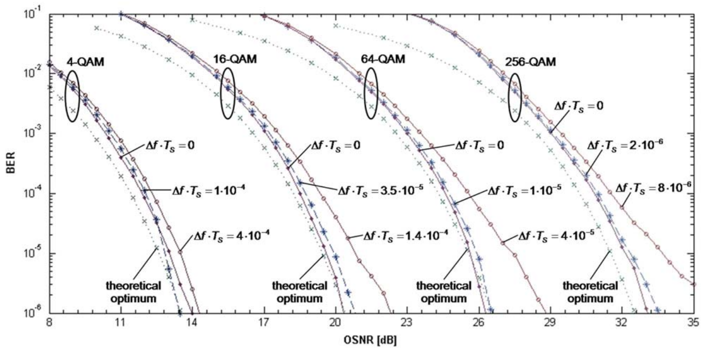
$$\phi_{NL} = \int_0^L P_0 \exp(-\alpha z)\gamma dz \approx \frac{N\gamma P_0}{\alpha} < \pi$$

total peak power

Nbr of amplifier spans

- This limits the total transmitted power to be < 0.1-10 mW in most systems.

Linewidth Requirements



T. Pfau et al., J. Lightwave Technol. 27, pp.989 (2009)

Department of Microtechnology and Nanoscience

Wireless and Photonic System Engineering SSY085

2015-10-27, 14:00-18.00

Teachers in charge:
Magnus Karlsson, phone 772 1590, email: magnus.karlsson@chalmers.se
Christian Fager, phone 772 5047, email: christian.fager@chalmers.se

Aids: Open book examination. Any printed material and calculator of choice is allowed. Communication devices (computers, mobile phones etc.) are *not* allowed.

Examination checking: Tue. Nov. 17th, 12-13 in room A703 "Donatorn", MC2 floor 7.

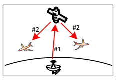
Convince yourself that you have understood the problem before you get started. Constructive and valuable gambits will also give points. If information is lacking in the description of the task, you must yourself introduce technical plausible assumptions. Make sure you clearly state such assumptions.
Grades: 3: ≥ 24, 4: ≥ 36, 5: ≥ 48

1. A future Stockholm-Göteborg link
Since the demand for broadband access keeps increasing, so does the demand for trunk network capacity. The Swedish University Network (SUNET):s next generation Stockholm-Gothenburg link (600 km) will likely use coherent transponders at 100 Gb/s (with 28 GHz of signal bandwidth) per wavelength channel and operate at an accumulated data rate of 10 Tb/s in the EDFA C-band. Thanks to the use of forward error correction and other digital signal processing algorithms, the required Q for such a channel is reduced from the conventional 6 to 3.

Design the link including necessary hardware, and consider as a design goal that you want to minimize the number of in-line amplifiers for cost reasons. Describe the link with block diagrams for the hardware and show with supporting calculations that it will work. Make reasonable and realistic assumptions for all ingoing system parameters.

(30 points)

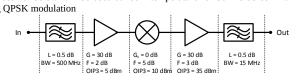
2. Earth-to-satellite-to-aircraft link
A telecom operator uses a geostationary satellite (G/T = 3 dB/K), supporting the 10.35 – 10.45 GHz frequency band, to provide on-board internet to aircraft passengers. The earth station uses a 5m diameter parabolic antenna and BPSK to reach a total capacity of 20 Mbit/s in the #1 link.



a) Calculate the minimum output power from the earth station, assuming BER<10^{-5}.
b) Describe which possibilities the operator have to increase the #1 link capacity.
c) Present a complete RF block diagram for a homodyne earth station transmitter (<u>no numbers needed</u>) and describe in detail <u>which</u> and <u>how</u> the components would have to be replaced to increase the link capacity.

(10 points)

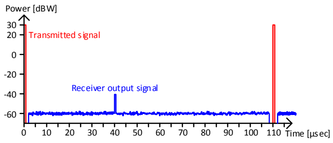
3. Receiver dynamic range
The receiver below will be used at room temperature for 30 Mbit/s communication using QPSK modulation



a) What is the linear dynamic range of the receiver?
b) We are going to use this amplifier in an urban area where there is no direct path between the transmitter and the receiver. The outage probability should be <10 %. What is ratio between the maximum and the minimum distance (R_{max}/R_{min}) if the required BER = 10^{-4}.

(10 points)

4. Pulsed radar system
Consider an ground-based aircraft surveillance radar (free space conditions). The radar transmits a pulsed 10 GHz RF signal after which it switches to receive mode to listen for possible radar echoes. The figure below shows an example of the transmitted and measured receiver output power versus time. The radar uses a parabolic antenna with 0.4m diameter (e_{rad} = 80%) and a receiver with 80 dB of gain and 1 MHz bandwidth.



a) What is the noise figure of the radar receiver?
b) What is the distance to the object in front of the radar?
c) What is the radar cross section of the object?
d) Assuming an object with radar cross section 1m², what is the maximum range for this radar?

(10 points)

Solutions

1.

2. Earth-to-satellite-to-aircraft link

a) Output power

```
% Matlab solution
f = 10.4e9; % Frequency in the middle of the band
lambda = 3e8/f;
G_T = 10^(3/10);
d = 5;
Ap = pi*d^2/4; % Physical antenna area
Gt = 4*pi*Ap/lambda^2; % Physical area and antenna area is similar
k = 1.38e-23; % Boltzmann constant
R = 36000e3; % Geostationary satellite = 36000km
Rb = 20e6;
B = Rb; % Spectral efficiency = 1 for BPSK
S_N = 10^(9.6/10); % Lecture Modulation, slide 29

% Non-fading lecture, slide 22
Pt = S_N/(G_T*Gt*lambda^2)*(k*B*(4*pi*R)^2);
Pt_dBm = 10*log10(Pt/1e-3);
```

<u>Minimum transmit power = 1 W = 30 dBm</u>

b) Improvement of capacity
- Increase the bandwidth utilization with the same modulation (it is not fully utilized now)
- Increase the modulation order
- (More antennas / but MIMO is probably not an option in satellite links)
- More satellites, so that each can serve fewer aircraft.

c)
- Bandwidth: Filters need to be replaced, A/D and D/A converters may need to be changed. Transmit power need to be increased kTB, or antennas replaced.
- Higher modulation order, same bandwidth: Transmit power increased (higher SNR requirement) (larger antenna or larger PA), new algorithms.
- More satellites…

3. Receiver dynamic range

a) Two equations are used:

(1) $T = T_{antenna} + T_{amplifier}$

(2) $\frac{1}{P3} = \frac{1}{Gc.Gb.P3a} + \frac{1}{Gc.P3b} + \frac{1}{P3c}$

DR_L (dB) = (P3 (dBm)-10-No (dBm)), No=G*k*B*T
G = -0.5+30+0+30-0.5=59 dB
$T_{amplifier}$ (input) ≈ $(10^{0.05}$-1) x290+ $(10^{0.2}$-1)x290/$10^{-0.05}$ = 225 K →
T_{sys} = 290+225=515 K (mixer noise can be neglected since the LNA gain is high)
→ No=$10^{5.9}$ x1.38x10^{-23}x15x10^6x515=8.47e-8 = -40.7 dBm
P_3 from (2) → P_3 = 31.35 dBm
DR_L (dB) = 31.35-10+40.7 = 62.05 dB, where it has been used that
P_{1dB} ~ P_3 dBm – 10 dB

b) SNR = Eb/n0 x Rb/B
For BER = 1e-4 (QPSK) {Lecture Modulation, curves slide 29) → Eb/n0 @ BER = 10^{-4} = ca 8 dB → SNR = 11 dB

Since there is no direct path between the transmitter and the receiver, we have a Rayleigh fading situation. From the figures in lecture notes (Lecture 8, p.32), Fading margin, Fd = 11 dB is needed for a 10% outage probability.

So the new DR_f can be find as
DR_L = (P3 (dBm)-10 dB -No (dBm)- SNR (dB) - Fd (dB)) = (31.35-10+40.7-11-11) dB =40.05 dB

So $(R_{max}/R_{min})^n = 10^{4.005}$ and for the urban area n = 3 → (R_{max}/R_{min}) = 21.63

As a remark, in real systems e.g. mobile communication, power control is used so that the base station or mobile phone adjusts its transmit power depending on the distance. Variable gain amplifiers are also used in the receiver to keep a constant power level to the demodulator / analog-to-digital converter.

4. Pulsed radar system

```
%% Matlab solution
clear;
clc
f = 10e9;
lambda = 3e8/f;
```

a) Receiver noise figure

```
erad = 0.8; % Radiation efficiency
B = 1e6;
k = 1.38e-23;
Tb = 50; % Brightness temperature, approximately from Lecture Non-fading systems, slide 18.
TA = erad*Tb + (1-erad)*290; % Same lecture slide 20
Nout = 10^(-60/10); % The output noise is read to be -60 dBW (note, not dBm) in the graph in the problem.
GRX = 10^(80/10); % The receiver gain is 80 dB
Nin = Nout/GRX; % Referring from output to input noise

Te = Nin/(B*k)- TA; % Equivalent receiver noise, subtracting the antenna noise
F = 1+Te/290; % Noise figure
F_dB = 10*log10(F)
```

<u>Total receiver noise figure = 5 dB.</u>

b) Distance

```
% According to the figure, the received pulse arrives 40 usec after the transmit pulse. Having traveled twice the distance yields
t = 40e-6;
R = t*3e8/2
```

<u>Distance = 6 km.</u>

c) Radar cross section

```
d = .4; % Antenna diameter
Ae = pi*d^2/4; % Antenna area
D = 4*pi*Ae/lambda^2; % Directivity
Gt = erad*D; % Antenna gain
Pt = 10^(30/10); % Transmit power, from graph
Pr_out_dBW = -40; % Radar echo power, dBW from graph
Pr = 10^((Pr_out_dBW-80)/10); % Convert to W, subtract receiver gain
RCS = Pr*(4*pi)^3*R^4/(Pt*Gt^2*lambda^2) % Radar equation
```

<u>Radar cross section (RCS) = 1.45 m².</u>

d) Max range

```
Pr = Nin; % Max distance is when the radar echo level is the same as the noise floor (Nin, see problem 4a)
RCS = 1; % According to problem
Rmax_noise = (RCS*Pt*Gt^2*lambda^2/(Pr*(4*pi)^3))^(1/4) % Radar equation, another way around
```

<u>Max noise-limited range (Rmax_noise) = 17.3 km.</u>

But, we also have to check if the echo arrives before the next pulse is transmitted.
t_{max} = 108µsec → $R_{max,time}$ = t_{max}*3e8/2 = <u>16.2 km < Rmax_noise.</u>

The maximum range is limited by the time between the pulses in this case.

<u>The maximum range is therefore 16.2 km for an object with 1m² radar cross section.</u>

Department of Microtechnology and Nanoscience

Wireless and Photonic System Engineering SSY085

2016-01-07, 08:30-12.30

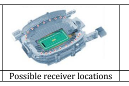
Teachers in charge:
Magnus Karlsson, phone 772 1590, email: magnus.karlsson@chalmers.se
Christian Fager, phone 772 5047, email: christian.fager@chalmers.se

Aids: Open book examination. Any printed material and calculator of choice is allowed. Communication devices (computers, mobile phones etc.) are *not* allowed.

Examination checking: Contact Christian or Magnus.

Convince yourself that you have understood the problem before you get started. Constructive and valuable gambits will also give points. If information is lacking in the description of the task, you must yourself introduce technical plausible assumptions. Make sure you clearly state such assumptions.
Grades: 3: ≥ 24, 4: ≥ 36, 5: ≥ 48

1. Real-time on-field football player tracking
Wireless technology is now being investigated to enhance the live experience in various sports. One idea is to equip the football players with <u>small battery powered</u> wireless transmitters. By receiving these signals from multiple positions around the arena, all player locations and movements can be accurately monitored and analyzed.

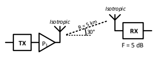
		
Transmitter placement	Possible receiver locations	Real-time tracking

Your task is to design a wireless player tracking system for American Football based on the idea above. Your system should be able to simultaneously track up to 40 players with at least 25 position updates per second.

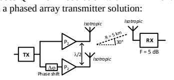
Your solution should include the design of the battery powered transmitters at the players and receivers around the arena, including choice of frequencies, modulation format, block diagrams, antennas, output power, etc. Use realistic assumptions where applicable.

(30 points)

2. Phased array transmitter
The following 5.2 GHz wireless system should transfer 100 Mbit/s of data to a receiver which is located line-of-sight in a direction of 30° and 5 km distance compared to the transmitter. The system uses QAM-64 modulation with BER < 10^{-3}. Your task is to evaluate a traditional, and a phased array transmitter solution:



Case 1: Traditional transmitter



Case 2: Phased array transmitter

First, consider Case 1 with one isotropic transmitter.
a) How much transmit power (P_1) is needed to fulfil the specifications?
Now consider Case 2 with two phase controlled transmitters, separated by λ/2.
b) How should the phase shift, Δφ, be set to maximize the received power?
c) With this setting, how much transmit power is needed from each of the two amplifiers (P_2) to fulfil the specifications?
d) Compare the total transmit power: P_1 vs. 2×P_2. Why are they different/same?

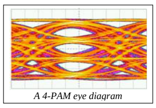
(10 points)

3. Combined amplifiers
You have three EDFAs, with gains G_1=10 dB, G_2=12 dB and G_3=15 dB. The corresponding noise figures are F_{n1}=5 dB, F_{n2}=7 dB and F_{n3}=8 dB. If the amplifiers are cascaded (used in series) the total gain is obviously 37 dB. Calculate the maximum and minimum effective noise figure such a cascade can have.

(10 points)

4. An MMF link
Within a datacenter a 50 m link is needed. The used wavelength is 850 nm, at which the fiber modal dispersion is around 1 ns/km and attenuation 2 dB/km. Assign 6dB for connector losses and splices, and another 5 dB in margin.
Directly modulated lasers with a bandwidth of 25 GHz and photodetectors with the same bandwidth are used. An average output power of 0 dBm is possible.

Calculate the highest datarate that can be achieved to reach a BER=10^{-9}, and whether the link is attenuation or dispersion limited. Make reasonable assumptions for the rise times. Consider two cases:
a) On-off-keying is used
b) 4-PAM is used.
Hint: 4-PAM needs approximately 3 times higher (optical) power than OOK for the same BER, since it can be seen as 3 OOK eye-diagrams stacked on top of each other (see figure).



A 4-PAM eye diagram

(10 points)

Solutions

1.

2. Phased array transmitter

a) *Output power, traditional system*

```
% Basic data given in problem
f = 5.2e9; lambda = 3e8/f; R = 5000; Rb = 100e6;
F = 10^(5/10); % Receiver noise figure
Gt = 1; Gr = 1; % Isotropic transmit and receive antennas

% SNR requirement given by modulation and BER requirement
EbN0= 10^(15/10); % 15 dB Eb/n0 required for BER = 1e-3 with QAM64 (see Modulation lecture, slide 29)
RbB = 6; % Spectral efficiency for QAM64
SNR = EbN0*RbB;

% Calculate minimum power needed
B = Rb/RbB; T = 290; k = 1.38e-23;
Nin = F*k*T*B; % Noise power at receiver input
Pr = Nin*SNR; % Add SNR margin

% Calculate transmit power needed using Friis formula
P1 = Pr/(Gt*Gr*lambda^2)*(4*pi*R^2);
```

<u>Minimum transmit power, P_1 = 3.78 W = 35.7 dBm</u>

b) *Phase difference*
With some simple trigonometric calculations, it is clear that the signal from the lower antenna has to travel a distance $d = \lambda/2\times\sin(30°)$ longer than the upper one, to reach the receiver. This corresponds to a phase difference of $\varphi = k\times d$, where k is the wave number: $k = 2\pi/\lambda$. When combined, this means that the $\Delta\varphi = +(2\pi/\lambda)\times\lambda/2\times\sin(30°) = \pi\times\sin(30°) = \pi/2 = +90°$ would be needed to make the signals align in the 30° direction.

c) *With this setting, how much transmit power is needed from each of the two amplifiers (P_2) to fulfil the specifications?*
When the signals are phase aligned, then their voltage contributions (E-field components) add in the receiver. Assuming a 50W reference impedance (will not change the end result), power and voltage are related by: $P = |V|^2/(2\times50)$.
Since the received power and the distance is the same in both cases we have:
$P_r = x\times|V_1|^2 = x\times|V_2+V_2|^2 \rightarrow |V_2|^2 = ¼|V_1|^2$, where V1 and V2 are the voltages produced by the transmitters. This corresponds to $P_2 = ¼P_1$ = 3.78/4 = 0.945 W.
<u>Each of the two amplifiers need to provide 0.945W.</u>

d) *Compare the total transmit power: P_1 vs. 2×P_2. Why are they different/same?*
From the results in b) and c) clearly P_1 is twice as large as 2×P_2. So, the phased array solution allows the total power to be reduced, with no effect on the performance. The explanation is the fact that the phase alignment only works in the desired direction. In other directions the signals may cancel. So, effectively, the combined effect of the two transmit antennas is to direct the energy to the receiver. This is identical of replacing the isotropic antenna in Case 1 with a directive antenna. For two elements, there is a factor 2 (3dB) gain in directivity. This factor is called the array factor, and is added on top of the antenna gain characteristics of each antenna.

3. **Cascaded amplifiers**
If the amplifiers are used in order a,b,c, the total emitted noise power is proportional to
$((G_a F_{na}+F_{nb})G_b+F_{nc})G_c=G_aG_bG_c F_{n,eff}$
so that
$F_{n,eff}= F_{na}+F_{nb}/G_a+F_{nc}/(G_b G_a)$.

Thus the first amplifiers noise figure will dominate. Setting the amps in the order 123 will thus be best and have a lowest NF of
10^(5/10)+10^[(7-10)/10]+10^[(8-22)/10]=3.70=5.69 dB
and the order 321 will get the highest NF of
10^(8/10)+10^[(7-15)/10]+10^[(5-27)/10]=6.47= 8.11 dB

All other combinations will lie between these values (8.08 dB, 7.34 dB, 7.20 dB, 5.81 dB) for the respective (312, 231, 213,132)-orderings.

4. **MMF link max data rate**

Make a rise time budget first. Assume Tx/Rx rise times are inverse bandwidths, i.e. 40 ps. The fiber rise time is 1000 ps/km*0.05 km=50ps. Total rise time is then sqrt(50^2+2*40^2)=75.5 ps, and if that amounts of 70 % of the symbol time slot, a rise-time-limited symbol rate of 0.7/0.0755=9.3 GHz. For OOK that is 9.3 Gb/s, for 4-PAM it is twice this, i.e. 18.6 Gb/s.

Now the power budget. Consider OOK first.
We require Q=6, for BER=10^{-9}, where $Q=I_1/(\sigma_0+\sigma_1)$, $I_1=2 R P_{rec}$, P_{rec} is the *average* received power (which is equally splitted between 1s and 0s, hence the factor of 2), and R the responsivity. The thermal noise has the variance
$\sigma_1=\sigma_0=\sigma_T$=sqrt($4k_BTB/R_L$). Hence $Q=RP_{rec}/\sigma_T$. The allowed thermal noise variance is then $\sigma_T^2=(R P_{rec}/Q)^2$ with the corresponding bandwidth given by
$B=(R P_{rec}/Q)^2R_L/(4k_BT)$.

The received optical power is 0 dBm minus 5 dB in margin, 6 dB in coupling losses and 2 dB/km•0.05 km=0.1 dB in fiber losses, giving -11.1 dBm=77.6 µW. With load resistance R_L=50 Ohm, a responsivity at 850 nm of R=0.68 A/W, and $k_BT=4.14\cdot10^{-21}$ J, we get
$B=(0.68\cdot77.6\cdot10^{-6}/6)^2\cdot50/(4\cdot4.14\cdot10^{-21})$=233 GHz
This is well beyond the bandwidth limited by the dispersion, so OOK is definitely dispersion limited.

For 4-PAM we use the same calculation, but insert 3 times lower optical power, i.e. 77.6/3=25.9 µW. This gives nine time lower bandwidth since the optical power scales with the square, i.e. 25.9 GHz. This is closer to, but still above, the dispersion limited bandwidth.

Thus the dispersion limited datarate is 9.3 Gb/s for OOK and twice that value, 28.6 Gb/s for 4-PAM

Department of Microtechnology and Nanoscience

Wireless and Photonic System Engineering SSY085

2016-10-25, 14.00-18.00

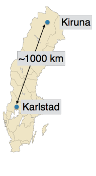
Teachers in charge:
Magnus Karlsson, phone 772 1590, email: magnus.karlsson@chalmers.se
Christian Fager, phone 772 5047, email: christian.fager@chalmers.se

Aids: Open book examination. Any printed material and calculator of choice is allowed. Communication devices (computers, mobile phones etc.) are *not* allowed.

Examination checking: On Monday Nov 14, during the lunch break 12-13 in room A402 'Gaten' (next to Café Canyon).

Convince yourself that you have understood the problem before you get started. Constructive and valuable gambits will also give points. If information is lacking in the description of the task, you must yourself introduce technical plausible assumptions. Make sure you clearly state such assumptions.
Grades: 3: ≥24, 4: ≥36, 5: ≥48

1. The Karlstad-Kiruna link



The longest span of the Swedish university network (SUNET) is the1000 km fiber optic link connecting Karlstad in the south with Kiruna in northern Sweden. The link should be able to provide 1 Tb/s initially, but be possible to upgrade to 10 Tb/s in (a not-so-distant) future. For cost reasons the number of amplifiers and other expensive components over this long distance should be minimized.
The link is based on standard SMF hung aerially along power electric transmission lines.
Two reconfigurable optical add-drop multiplexers (ROADM:s) are inserted along the link, one in Östersund, 500 km north of Karlstad and one in Arvidsjaur, 300 km south of Kiruna. Each ROADM adds around 3 dB insertion losses to the link budgets.

Design the link including selection of the necessary hardware, modulation format, data rates etc. Make reasonable parameter estimates for all components and verify that the link will work according to a reasonable specification that you clearly state. It is enough to design for one direction only.

(30 points)

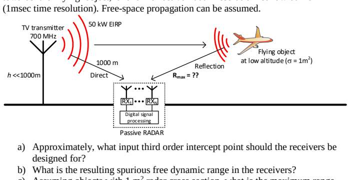
2. Path loss and fading margins
You are designing a 3.5 GHz cellular wireless system with 5km range. The system is operating in a dense urban area.
a) How much path loss, in dB, should the system be designed for?
b) How much margin must be added to account for heavy rain losses?
c) How much fading margin do we need to add if 5% outage probability is desired?

(10 points)

3. Receiver block diagram
Consider a terrestrial 30 GHz point-to-point radio link system, designed for QPSK modulation at a maximum bitrate of 200 Mbit/s (BER <10^{-5}). Your task is to present a super heterodyne receiver block diagram (including filter, amplifier, oscillator, and mixer specifications) for this system. The receiver should have a sensitivity better than -75 dBm and a total gain of >80 dB. All component values must be carefully motivated.

(10 points)

4. Passive RADAR
Passive radar systems detect and track objects by processing the difference between reflections and direct signals from non-cooperative sources of illumination, such as commercial TV transmitters.
Consider the following passive radar system, which uses a 700 MHz TV transmitter with EIRP = 50 kW at 8 MHz signal bandwidth. The radar receivers (RX_1-RX_N) each have low gain antennas (3 dBi) and 3 dB noise figure. 8 MHz signals from several receiving antennas are combined digitally to give an array antenna gain of 20 dBi towards the flying object, and an effective radar resolution bandwidth of 1 kHz (1msec time resolution). Free-space propagation can be assumed.



a) Approximately, what input third order intercept point should the receivers be designed for?
b) What is the resulting spurious free dynamic range in the receivers?
c) Assuming objects with 1 m² radar cross section, what is the maximum range of the radar (R_{max})?

(10 points)

Solutions

<u>**Problem 1:**</u>

<u>**Problem 2:**</u>
a) Path loss
Modified Friis equation will be used, with path loss exponent of N = 3 to account for dense urban area. The distance to the nearest obstacle is assumed to be in the order of R_0 = 100m. The path loss in Friis equation becomes:
PathLoss = $\lambda^2/(4*\pi*R_0)^2*(R_0/R)^N = (3e8/3.5e9)^2/(4*\pi*100)^2*(100/5000)^3$ = -134 dB

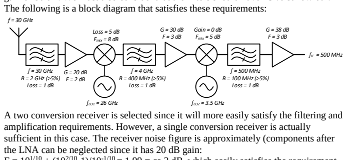
b) Rain margin
According to slide 17, Lecture 8, heavy rain has very small influence at low frequencies. Reading the graph, the rain loss at 3.5 GHz is in the order of 0.1 dB/km, which means that only 0.5 dB need to be added to account for the rain. Rain can be neglected.

c) Fading margin
Dense urban → we can assume Rayleigh fading. For this case, the relation between outage probability and fading margin is given by:
$p_{out} = 1\text{-}e^{margin}$ → margin =ln(1-p_{out}) = -0.0513 = -12.9 dB. We therefore need to add ca 13 dB margin to account for fading in this scenario.

<u>**Problem 3:**</u>
From the problem, we need a sensitivity of at least -80 dBm. The sensitivity is the minimum detectable power level.
Assuming ambient temperature = 290 K, we have P_{min} = k*T*B*F*SNR

where T = 290K, B = bitrate/spectral efficiency = 200e6/2 = 100 MHz.
SNR is the required signal-to-noise ratio for QPSK at BER = 10^{-5}. From Table 9.5 in Pozar, we have SNR = Eb/n0*(Rb/B) = 12.6 dB = 18.2.

With Pmin = -75 dBm we can solve for the required receiver noise figure → F < 4.3 = 6.3 dB. Our receiver block diagram must therefore satisfy this noise figure and a total gain of 80 dB. We must also consider that the filters should have >5 % bandwidth. The following is a block diagram that satisfies these requirements:



A two conversion receiver is selected since it will more easily satisfy the filtering and amplification requirements. However, a single conversion receiver is actually sufficient in this case. The receiver noise figure is approximately (components after the LNA can be neglected since it has 20 dB gain:
F = $10^{1/10}$ + $(10^{2/10}$-1)/$10^{-1/10}$ = 1.99 = ca 3 dB, which easily satisfies the requirement above.

Problem 4:

```
%% Input data
freq = 500e6;
lambda = 3e8/freq;
EIRP = 50e3;
RCS = 1; % Radar cross section
R_TV = 1000;
R_object = 1000:1000:400000; % This is unknown - different values will be tested
k = 1.38e-23;
T_a = 50; % Approximately, from slide 8, Lecture 7
B_RX = 8e6;
B_RX_array = 1e3;
F = 2; % 3 dB
G_array = 100; % Antenna array gain, 20 dBi
G_rx = 2; % Individual receiver antenna, 3 dBi

%% Received power from direct TV signal
% In order to avoid compression of the receivers, we ensure that the power from the
% direct path is reaching the 1-dB input compression point. This is then translated to
% IIP3 by the approximate formula
Pr_TV = EIRP*G_rx*lambda^2./((4*pi*R_TV).^2);
Pr_TV_dBm = 10*log10(Pr_TV/1e-3);
RX_P1dB = Pr_TV_dBm;
RX_IIP3_dBm = RX_P1dB + 10;
RX_IIP3 = 1e-3*10^(RX_IIP3_dBm/10);
```

a) What input third order intercept point should the receivers be designed for? RX_IIP3_dBm = 6.7 dBm

```
%% Noise power at receiver inputs
P_noise_in = k*B_RX*((F-1)*290 + T_a);

%% Spurious free dynamic range
% The spurious free dynamic range is defined at the output of the receivers
% DR_f = (OIP3/P_noise_out)^(2/3)
% Referring both noise and IP3 to the input instead means scaling with the
% gain. The gain will, however, cancel:
DR_f = (RX_IIP3/P_noise_in)^(2/3);
DR_f_dB = 10*log10(DR_f);
```

b) What is the resulting spurious free dynamic range in the receivers? DR_f_dB = 72 dB

```
%% Solution
% Power reaching the flying object
Pr_object = EIRP*RCS./(4*pi*(R_TV+R_object).^2);
Pt_object = Pr_object; % Isotropic re-radiation

% Received power from radar reflection. Here we can account for the antenna
% array gain
Pr_reflection = Pt_object*1*G_array*lambda^2./((4*pi*R_object).^2);

% For the radar echo processing, we can consider the effective resolution
% bandwidth.
P_noise_array = k*B_RX_array*((F-1)*290 + T_a);

% Signal-to-noise ratio.
SNR = 10*log10(Pr_reflection./P_noise_array);

% Max range, R_max, is found for the R_object setting where the SNR is = 0 dB.
% There is no analytical solution to this if we use the expressions above.
% With some easy numerical testing, we will find that Rmax = ca 160 km
% It is, however, rather easy to conclude that the range is much longer
% than the 1km range from the TV station. This simplifies the derivation,
% and the following expression can be used to analytically find the max
% range:

R_max = ((EIRP*RCS*G_array*lambda^2)/(k*T_a*B_RX*F*(4*pi)^3))^(1/4);
```

c) Assuming objects with 1 m2 radar cross section, what is the maximum range of the radar (R_{max})? R_max = 1.69e5 m. In practice, the range will be reduced by scattering, non-line of sight propagation, etc. Please look at https://en.wikipedia.org/wiki/Passive_radar for more information.

Problem 3:

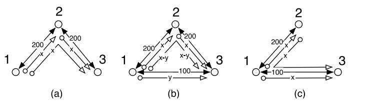
a)
This will be a thermal-noise-limited link, and the receiver sensitivity P_{rec} (average power=$P_1/2$) is obtained from Q=6, where $Q=I_1/(2\sigma_T)=2R\,P_{rec}/(2\,\sigma_T)=6$. In this case, $\sigma_T^2 = 4k_B T\,\Delta f/R_L$ = 1.66e-11 A^2, from which we get $P_{rec}=6\,\sigma_T\,(4k_B T\,\Delta f/R_L)^{1/2}/R$=-17.1 dBm (where we used responsivity R=1.25 A/W). The allowed losses are thus 3-3-1-1-(-17.1)=15.1 dB which corresponds to a link length of 15.1/0.2=76 km.

b)
Here signal-spontaneous noise will likely limit the link, and using P_{rec} to denote the average power in to the amplifier, the Q becomes $Q=I_1/\sigma_{s-sp}=2RGP_{rec}/(4R^2GP_{rec}S_{sp}\Delta f)^{1/2}=(P_{rec}/(F_n h\nu\,\Delta f/2))^{1/2}=6$, by using $S_{sp}=F_n h\nu G/2$. From this we find the sensitivity to be $P_{rec}=18\,F_n\,h\nu\,\Delta f$.
By using F_n=4, $h\nu\Delta f$=6.4e-9 W, we get the receiver sensitivity as P_{rec}=-33.4 dBm, and the allowed losses are 3-3-1-1-33.4=31.4 dB. The corresponding fiber length is 31.4/0.2=157 km.
Note that the EDFA gain does not enter the expressions, but it is assumed that it is so large that the s-sp noise is much larger than the thermal noise. We can check if this is reasonable via $\sigma_{s-sp}^2=\sigma_T^2$ which occurs for an EDFA gain of 13.9 dB. So if the gain is 20 dB or higher we can safely neglect the thermal noise.

Problem 4:

Refer to the below figure.

(a) In fig (a): this case is indicated. We show only connections C_{12}, C_{23}, C_{13} with open arrows, which should all be x, and the other three in the opposite directions will be the same due to symmetry when using the bidirectional transmission links in the other direction. The total capacity in link 1-3 (and 2-3) is thus 2x ≤ 200, which gives a maximum x of 100 Gb/s.
(b) This case is shown in fig (b). The additional path 1-3 is assumed to be used at a rate y Gb/s, so the capacity C_13 is shared between path 1-3 and path 1-2-3 as shown in (b). We thus have that y≤100 for path 1-3, and that 2x-y≤200 for path 1-2. The maximum value of x in this case is 150 Gb/s, for y=100.
(c) This case is shown in (c), which is similar to (a) but path 1-3 now limits the rate via 2x≤100, so the maximum x is now 50 Gb/s.



(a) (b) (c)

Problem 4: Solution.

Problem 2:

2a. Max link distance.
Point-to-point link with high gain antennas -> Friis equation. At 145 GHz we need to consider the atmospheric losses. According to Lecture 10, slide 16, the loss is ca α = 2 dB/km. Combined with Friis equation this gives:

$P_r = P_t G_t G_r \lambda^2/(4\pi R)^2 \cdot 10^{-\alpha(R/1000)/10}$

Similar as in Problem 1, $P_{r,min} = SNR_{out}(T_A + T_e)kB$ = 9.12*(290 + 290*($10^{10/10}$-1))*1.38e-23*10e9 = 3.6*10^{-9} W. With Pt, Gt, Gr given in the problem, we can solve for the range. Since *R* appears in both exponentially with a and quadratically, a numerical solution is preferred. This gives: R_{max} = 1100 m.

2b. How much transmit power, if BPSK → QAM64
Since the range (and antennas and frequency) is fixed, we have the following relationship from 2a:

$P_t = P_r G_t G_r \lambda^2/(4\pi R_{max})^2 \cdot 10^{\alpha(R_{max}/1000)/10} = P_r \cdot 3.7\cdot10^{-7}$ ←→
P_t [dBm] = P_r [dBm] - 64.3 dB → P_r [dBm] = P_t [dBm] + 64.3 dB

When changing from BPSK to QAM64, the SNR_{out} requirement will change, and the bandwidth will reduce by a factor of 6 due to the spectral efficiency. $SNR_{out,QAM64}$ = {Lecture 3, slide 29} = Eb/n0*(R_b/B) = $10^{17.8/10}$*6 = 361.5. Hence, $P_{r,min}$ = 361.5*(290 + 290*($10^{10/10}$-1))*1.38e-23*10e9/6 = 2.4*10^{-8} W = -46.2 dBm → P_t = 18.1 dBm. This is 8 dB, or 6.3 times, higher power than for BPSK.

2c. What is the theoretically minimum transmit power, according to Shannon
Shannon equation:

$R_{b,max} = B\log_2(1+S/N)$ → $R_b/B = \log_2(1+P_{r,min}/(kB(T_e+T_A)))$ →
$P_{r,min} = kB(T_e+T_A)(2^{R_b/B}-1)$, where R_b/B = spectral efficiency, η. This will, for the two cases η = 1 and η = 6 result in:
$P_{r,min,\eta=1}$ = 4.0*10^{-10} W = -64.0 dBm → $P_{t,\eta=1}$ = 0.3 dBm
$P_{r,min,\eta=6}$ = 4.2*10^{-9} W = -53.8 dBm → $P_{t,\eta=6}$ = 10.5 dBm

Clearly, with advanced coding techniques (for example forward error correction), the capacity can be increased for given output power.

Department of Microtechnology and Nanoscience

Wireless and Photonic System Engineering SSY085
2016-12-22, 08.00-12.00

Teachers in charge:
Magnus Karlsson, phone 772 1590, email: magnus.karlsson@chalmers.se
Christian Fager, phone 772 5047, email: christian.fager@chalmers.se

Aids: Open book examination. Any printed material and calculator of choice is allowed. Communication devices (computers, mobile phones etc.) are *not* allowed.

Examination checking: Contact Magnus or Christian via mail or phone above.

Convince yourself that you have understood the problem before you get started. Constructive and valuable gambits will also give points. If information is lacking in the description of the task, you must yourself introduce technical plausible assumptions. Make sure you clearly state such assumptions.
Grades: 3: ≥24, 4: ≥36, 5: ≥48

1. Balloon internet

Up to 50% of the world population still lack internet access. One of the solutions investigated to expand coverage in rural areas is to replace traditional base station infrastructure with air balloons traveling in the stratosphere at ca 20 km altitude. Each balloon provides ca 5000 km^2 land area with a total of 100 Mbit/s downlink capacity directly to standard 4G devices (phones etc). The balloons are interconnected between themselves in a wireless network (see illustration). The balloons are driven by solar panels producing 100W of electric power.

Your task is to design the balloon downlink transmitter, and also the bi-directional wireless link between two balloons. This includes choice of frequencies, modulation format, block diagrams, antennas, output power, etc. Use realistic assumptions where applicable.

(30 points)

Wireless and Photonic System Engineering SSY085
2017-10-24, 14.00-18.00

Teachers in charge:
Magnus Karlsson, phone 772 1590, email: magnus.karlsson@chalmers.se
Christian Fager, phone 772 5047, email: christian.fager@chalmers.se

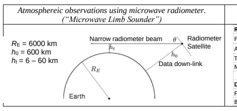
Aids: Open book examination. Any printed material and calculator of choice is allowed. Communication devices (computers, mobile phones etc.) are *not* allowed.

Examination checking: On Monday Nov 13th, during the lunch break 12-13 in room A604

Convince yourself that you have understood the problem before you get started. Constructive and valuable gambits will also give points. If information is lacking in the description of the task, you must yourself introduce technically plausible assumptions. Make sure you clearly state such assumptions.
Grades: 3: ≥24, 4: ≥36, 5: ≥48

1. Climate observations using microwave radiometry
The molecular composition in upper atmospheric layers can be characterized by their specific microwave thermal emissions. The noise temperature of these emissions can be measured by pointing a highly directive antenna tangentially towards the atmosphere, see illustration below.

Atmosphereic observations using microwave radiometer. ("Microwave Limb Sounder")



System specifications

Radiometer specifications	
Frequency, f_{O3}	320 GHz
Altitude resolution, Δh_t	3 km
Time per measurement	0.2 s
Measurement accuracy	± 1 K
Downlink	
Frequency	10 GHz
Bitrate	100 kbit/s

Repeated noise temperature measurements are used to improve the sensitivity, giving rise to the "radiometer equation": $S/N = T_{src}/\left(T_{RX}/\sqrt{B\cdot\Delta t}\right)$, where *B* is the receiver bandwidth and Δ*t* is the averaging time. T_{RX} and T_{src} are the noise temperatures of the receiver and the atmosphere within the antenna beam, respectively. The vertical profile, h_t, is found by scanning the antenna angle (*θ*).

The first task is to design a microwave radiometer for ozone emissions at 320 GHz. The **second task** is to design a downlink for the scientific data, which is sent once per day to a fixed ground station. The **overall system specifications** are given in the table above. Your solutions to both tasks should include block diagrams, choice of frequencies, component specifications, antennas, power levels, modulation format (for the downlink), etc. Use realistic assumptions where applicable.

(30 points)

A simple choice of the dispersion limit condition is 25 % of the bit period. The bit period for 5 Gb/s is 200 ps. Thus the allowed dispersive broadening is σ_t=50 ps. We next use the expressions for dispersive broadening given on slide 22 in Lect 7, "Dispersion and attenuation".

a) For a step index MMF we have $\sigma_t = L\Delta n_1/(2c)$, or $L= 2c\sigma_t/(\Delta\, n_1)$, and with an index difference of Δ=0.01 and n_1=1.5, we get L=2 meter. Not much…
b) For a graded index MMF we have $\sigma_t = L\Delta^2 n_1/(4c)$, or $L= 4c\sigma_t/(\Delta^2 n_1)$, or a factor 2/Δ=200 better than the above step index fiber, i.e. L=400 meter. That is better.
c) For the SMF we have the formula $\sigma_t = D\,L\sigma_\lambda$ we need the spectral width σ_λ, and 5 GHz corresponds to 0.04 nm. With D=16 ps/nm km, we find L=50/(16*0.04)=78 km. This is clearly the best.

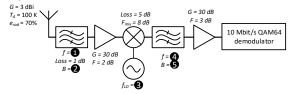
Problem 4:

a)
The total losses of the link is 270 km x 0.20 dB/km=54 dB. Hence G_1G_2=54 dB, but for the A, B and C options these are divided as $G_1\,G_2$ = 20+34, 24+30 and 32+22 dB, respectively. Which if these produce the lowest noise? Each amplifier produces a noise power spectral density of $S_{sp}=GF_n\,h\nu/2$, and at the receiver this is $S_{sp1}+S_{sp2}=(G_1+G_2)F_n\,h\nu/2$. Thus the position that minimizes (G_1+G_2) is the best. We can calculate the three cases
A: $G_1+G_2=10^{2.0}+10^{3.4}$=2612
B: $G_1+G_2=10^{2.4}+10^{3.0}$=1251
C: $G_1+G_2=10^{3.2}+10^{2.2}$=1744

Clearly option B is the best one, in the sense that it give the least noise.

b)
Which noise source dominates? Since it is an amplified system, it is signal-spontaneous noise, which is given by $\sigma_{s-sp}^2 = 4R^2 S_{sp} P_{rec}\,\Delta f$
The SNR at the receiver is given by SNR=(RP_{rec})^2/σ_{s-sp}^2, and we obtain SNR=P_{rec}/[2 (G_1+G_2) F_n hν Δf] =39.0=16 dB

2. Receiver design
The following block diagram is implementing a 5 GHz receiver for 100 Mbit/s QAM64 signals.



a) Which receiver architecture is this?
b) Assign suitable values for the missing parameters in the diagram (1) – (5).
c) What is the minimum received signal power that gives BER<10^{-4}?

(10 points)

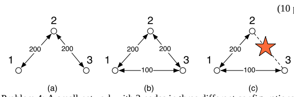
3. Transmission reach
Calculate the attenuation-limited reach for an OOK system using SMF (loss 0.2 dB/km) that operates at 50 Gb/s. The transmit average power is 3 dBm, the coupling losses are 1 dB each in the transmitter and receiver, and one would like 3 dB of power margin in the receiver at a BER of 10^{-9}. The receive is a p-i-n diode with quantum efficiency=1. How far can you go if you have
a) a direct detection receiver (i.e. no optical amplifier)?
b) a preamplified optical receiver using an EDFA with noise figure of 6 dB?

(10points)

4. A small network
A simple optical network consists of 3 nodes, numbered 1,2,3. It is supposed to sustain a *uniform* traffic load matrix C where the elements C_{jk} equals the data rate from node j to node k. In other words we want the same traffic between all node pairs, so that C=[0 x x; x 0 x; x x 0], where x is the data rate between any pair of nodes. We seek the highest value of x for three different scenarios as outlined below.

(a) At first install, 200 Gb/s of bidirectional capacity (i.e. 200 Gb/s can be transmitted in both directions simultaneously) was installed between nodes 1-2 and 2-3, as shown in Fig 4. (a). What is the maximum x in this case?
(b) The network was then upgraded with a third bidirectional path between nodes 1-3, but as the distance between these nodes are larger this link could only sustain 100 Gb/s of data rate, see (b). What is the maximum x in this case?
(c) A cable break in path 2-3 occurs, as indicated in (c). What is the maximum rate x in this case?

(10 points)



(a) (b) (c)

Problem 4: A small network with 3 nodes in three different configurations. Bidirectional data rates in Gb/s are indicated on the arrows.

You are part of a research team that should demonstrate 10 Gbit/s point-to-point communication at 145 GHz (D-band). You have a receiver with 10 dB noise figure, and transmit and receive antennas with 35 dBi gain available.
a) Calculate the maximum link distance if the transmitter has 10 dBm output power and uses BPSK modulation (spectral eff. = 1 bit/s/Hz, BER = 10^{-5}).
b) Keeping the same distance as in (a), how much transmit power is needed if the modulation is changed to QAM-64 (spectral eff. = 6 bit/s/Hz, BER = 10^{-5})?
c) According to Shannon's theory, what is the theoretically minimum transmit power needed for error-free communication at spectral efficiency of 1 bit/s/Hz and 6 bit/s/Hz, respectively?

(10 points)

3. Impact of fiber types
You are to transmit 5 Gb/s of OOK modulated data at the wavelength of 1550 nm. What is the dispersion-limited distance if you use
a) a step-index multi-mode fiber?
b) a graded-index multi-mode fiber?
c) a standard single-mode fiber?
Make your own reasonable assumptions with respect to fiber parameters and the definition of "dispersion-limited distance", and state them clearly.

(10 points)

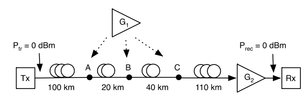
4. An amplified terrestrial link
You are designing a fiber optic link between two cities, 270 km apart, that transmits 20 Gb/s using on-off-keying on a single wavelength channel. The transmitter (average) power is 0 dBm, and to avoid nonlinearities this power must not exceed 0 dBm anywhere along the link. The receiver site has an amplifier with gain G_2 that amplifies the signal back to 0 dBm before it hits the photoreceiver. You are allowed to insert another amplifier with gain G_1 in the link. There are 3 possible options for the placement of this amplifier, marked A, B and C in the figure, where the link passes suitable hub stations with access to electricity. Assume an effective fiber loss of 0.20 dB/km, which accounts also for other sources of attenuation such as connectors, splices, filters and dispersion compensating elements. The amplifier gains G_1, G_2 can be adjusted to compensate for the losses as needed, and the amplifier noise figures are 6 dB, irrespective of the gain. The noise-noise beating is optically filtered and can be neglected.

a) Which site (A, B, or C) would you select for inline amplifier in this system, and what would the gains G_1, G_2 be? Motivate!

(5 points)

b) What will be the received SNR of the signal with your choice?

(5 points)



Problem 4: Choose among three possible positions for the inline optical amplifier.

Department of Microtechnology and Nanoscience

Wireless and Photonic System Engineering SSY085
2018-11-02, 08.30-12.30

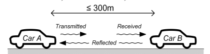
Teachers in charge:
Magnus Karlsson, phone 772 1590, email: magnus.karlsson@chalmers.se
Christian Fager, phone 772 5047, email: christian.fager@chalmers.se

Aids: Open book examination. Any printed material and calculator of choice is allowed. Communication devices (computers, mobile phones etc) are *not* allowed.

Examination checking: Wednesday Nov 21st, 12-12:30, Room A604, MC2.

Convince yourself that you have understood the problem before you get started. Constructive and valuable gambits will also give points. If information is lacking in the description of the task, you must yourself introduce technical plausible assumptions. Make sure you clearly state such assumptions.
Grades: 3: ≥24, 4: ≥36, 5: ≥48

1. Automotive radar communication
Most automotive radars operate at 77 GHz and are presently used for collision avoidance, and intelligent cruise-control. In the near future, these radars will play an increasingly important role as drive-automation progresses. It is also proposed to use radar signals for communication between cars. Multiple cars can then share information about traffic conditions, internet access, etc.



You should design such a 77 GHz combined radar/communication system, so-called RadCom. The system should be able to detect cars up to 300m ahead, while simultaneously provide a data transfer rate of at least 10 Mbit/s.

To get full marks you must present block diagrams with specifications for all RF components and antennas for the transmitter and receiver installed on Car A,and the receiver installed on Car B. The component parameters, antenna sizes, bandwidths, modulation formats, etc., should be motivated considering signal and noise power levels, power consumption, propagation effects, BER, etc.

(30 points)

Solutions

Problem 1:

Problem 2:
a) This is a (super) heterodyne receiver architecture
b) Values:
1. **5 GHz** (= center frequency)
2. Max 5% of center frequency, i.e. ca **300 MHz**
3. The LO frequency has to be set so that the image frequency and LO leakage is eliminated by the bandpass filter: $f_{IF} > ½BW_{RF}$ → f_{IF} > 150 MHz. Adding also the 16 MHz signal bandwidth and some margins, I select f_{IF} = 200 MHz. fLO = $f_{RF} \pm f_{IF}$ = 5 GHz ± 200 MHz.
I choose f_{LO} = **5.2 GHz**.
4. f_{IF} = **200 MHz**
5. B = signal bandwidth = bitrate / spectral efficiency for QAM64 = 100e6/6 = 16 MHz. I choose **20 MHz** to include some margin for filter rolloff and manufacturing tolerances.

c) Matlab solution

```
Ffilt = 10^(1/10);
Gfilt = 10^(1/10);
FLNA = 10^(2/10);
GLNA = 10^(30/10);

F = Ffilt + (FLNA - 1)/Gfilt
Te = 290*(F-1); % This is the noise temperature at the receiver input (at
the filter/antenna interface)
TA = 100; % This includes the brightness temperature, and the antenna
efficiency.

Trx = Te + TA; % The total receiver input noise (at the filter/antenna
interface).
Ebno = 10^(16.5/10); % read for QAM64 curve, Modulation lecture slide 29
spec_eff = 6; % For QAM64, 2^6 = 64
SNR_out = Ebno*spec_eff;

Rb = 100e6;%Bitrate
B = Rb/spec_eff; % Bandwidth
k = 1.38e-23;
Nin = k*Trx*B;
Sin = Nin*SNR_out;

e_rad = 0.75; % Antenna radiation efficiency

Sin_antenna = Sin/e_rad;% Total receiver input signal must also overcome
antenna losses
Sin_antenna_dBm = 10*log10(Sin_antenna/1e-3)
```

Minimum received signal power = -75.94 dBm.

Problem 1:
This problem is about 1) Design of a receiver for measurement of the atmospheric temperature (a *radiometer*) 2) Radiometer antenna, and 3) Design of a 100kbit/s downlink.

1) Radio receiver
As the radiometer equation shows (and as I mentioned at the exam), the signal corresponds to T_{src} (in this case the atmosphere temperature), while the noise corresponds to T_{RX}/sqrt(B*Δt). Here B = receiver bandwidth and Δt is the averaging time. According to information in the table, the receiver should have an accuracy better than 1K. Hence, the radiometer noise must be less than 1K. Moreover, the specifications mention that the time per measurement is 0.2s:
T_{RX}/sqrt(B*0.2) < 1 → T_{RX}/sqrt(B) < sqrt(0.2) = 0.44

Let's assume that the receiver noise figure is 10 dB (I will be very generous about your assumptions here) → T_{RX} = 290*(10-1) = 2610. We can then solve for the receiver bandwidth: B = (2610/(0.44))^2 = 35 MHz.

So, in summary, the first task is to design a 320 GHz receiver with 35 MHz bandwidth. This cannot be done in one conversion step ("the 5% rule"). The following intermediate frequencies could, e.g. be used: 320 GHz → 10 GHz → 250 MHz. The 35 MHz bandwidth can be filtered out from the 250 MHz IF and processed to estimate the atmospheric temperature.

2) Radiometer antenna design
It is given that the altitude resolution Δh_t should be 3km. Looking at the figure, this corresponds to the following angular resolution from satellite perspective:
θ = arctan($d/\Delta h_t$) = 1.15*10^{-3} rad, where d = sqrt($(6000+600)^2$-$(6000+60)^2$) ≈ 2600 km is the distance from the satellite to the atmosphere. From Lecture 5, we have the relationship between directivity (*D*) and beam width θ: D = π^2/θ^2 → D = 7.4970e+06 = 68.7 dBi (which is a very high number indeed). With some losses a 70 dBi antenna at 320 GHz would have the following area: $A_e = D\lambda^2/4\pi$ = 0.52 m^2. Assuming a parabolic antenna, this corresponds to a diameter of 0.8m.

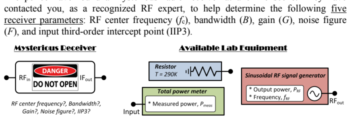
3) 10 GHz downlink
We can assume QPSK modulation and BER = 1e-5, which gives us a bandwidth of 50 kHz, and a receiver output SNR requirement: SNR_{out} = 18.24. According to Lecture 6 (slide 18) the sky temperature is only <10 K at 10 GHz, wielding T_A ≈ 10. Assuming also a receiver noise figure of F = 5 dB yields the following expression for the minimum permissible received power at the Earth station:

$P_{r,min}$ =
$SNR_{out}(T_A + T_e)kB$ = 18.24*(10 + 290*($10^{5/10}$-1))*1.38e-23*50e3 = 8*10^{-15} W

Assuming free space propagation, we can safely use Friis equation. If we assume antenna gains of 20 dBi at satellite and 40 dBi at Earth station (smaller at satellite due to weight), then Friis equation gives us the transmit power needed

$P_t = P_{r,min}(4\pi h_0)^2/(G_t G_r \lambda^2)$ = 5.1*10^{-4} W = -3 dBm, which is easy to generate. In reality, we will need to add margin for atmospheric losses due to rain etc. We could also reduce the Earth station antenna gain to make it easier to track the satellite motion.

2. The receiver detective
The police have found a mysterious radio receiver at a crime scene. They have contacted you, as a recognized RF expert, to help determine the following five receiver parameters: RF center frequency (f_c), bandwidth (*B*), gain (*G*), noise figure (*F*), and input third-order intercept point (IIP3).



Explain, in words, illustrations, and equations, how you will use the available lab equipment (see Figure above) to determine each of these five receiver parameters.

(10 points)

3. The MOST link
Inside automobiles the MOST (media oriented system transport) standard specifies the use of plastic optical fibers for data transmission. The plastic fibers have the following parameters: diameter 1 mm, core index 1.492, cladding index 1.402, and loss 140 dB/km at the wavelength of 650 nm which is used in these links. The transmitter is an LED with output power 1 mW and bandwidth 200 MHz, and the receiver is a pin-diode with 300 MHz bandwidth. The transmission distance is 20 m.
a) Calculate the receiver sensitivity using this pin-diode, assuming OOK modulation and a required BER of 10^{-9}.

(3 points)

b) Make a rise-time budget and use a suitable condition to calculate the dispersion-limited data rate.

(5 points)

c) Using the data rate found from (b), how much extra attenuation (in addition to the fiber loss) can be allowed between transmitter and receiver?

(2 points)

4. Choice of modulation

A long amplified WDM fiber link consists of 8 amplifiers with gain 18 dB noise figure of 5 dB. The WDM multiplexers and demultiplexers use a channel spacing of 50 GHz, of which 30 GHz is a "flat passband" which is available as useable signal bandwidth. The total output power from each EDFA can be no higher than 15 dBm, to avoid nonlinear distortions in the transmission.

You have two choices of modulation: (i) 16-QAM data that requires an SNR of 20 dB, or (ii) QPSK that requires 14 dB SNR at the receiver for it to work properly. Calculate for these 2 cases: the required transmitted power per channel, the number of channels that can be transmitted, and the resulting total data rate.

(10 points)

Solutions
Problem 1:
Car A assumptions:
- Transmit power = P_{TX} = 100 mW (<<10 W limitation, and assuming a low transmitter efficiency at the high frequency used)
- Antenna gain = G_A = 35 dB (High so that good separate cars can be tracked/identified. Can be implemented with small horn antenna at these frequencies). Separate, but identical antennas are used for transmit and radar receiver. This will automatically give isolation to prevent strong leakage between the transmitted and the reflected received signal in Car A. Note that filtering cannot be used to separate RX and TX in radars since the reflected and transmitted signals are at the same frequency.
- Receiver noise figure = F_A = 5 dB (Rather low, but a critical parameter – could be obtained with good receiver LNA)
- Modulation format = QPSK → Spectral efficiency = R_b/B = 2 → {Rb = 10 Mbit/s} → B = 5 MHz

Radar calculations:
- I assume that Car B has a radar cross section of RCS = 1m²
- I will use a ground reflection propagation model. Other models could be used e.g. using Modified Friis-equation with N>2; see Fading Lecture. Free-space propagation is not a realistic choice.
- Applying the ground reflection model with the radar equation gives:
  `Prx_A = Ptx*G_A^2*4*pi*RCS*(h_A^2*h_B^2)^2/(R^8*lambda^2)`
  where R = 300 m and h_A and h_B are the antenna heights of the cars, assumed to be h_A = h_B = 0.4m.
  Inserting numbers yields Prx_A = 8.3e-12 W = -80 dBm.
- The car A receiver noise power, referred to the input, is
  Nrx_A = k*T0*B*F_A = 1.38e-23*290*5e6*10^0.5 = 6.3e-14 W = -102 dBm
- Comparing the radar signal power Prx_A with the noise power Nrx_A, the signal is higher meaning that the other car can be detected at this range.
- The ground reflection model is valid for R > 20*h_A*h_B/lambda = 820 m, which indicates it could be needed to compare the results also with other propagation models in this case to guarantee the operation. The ground reflection model probably gives a pessimistic result and the margin is probably higher in practice.

Car A → Car B communication calculations:
- Assume car B receiver noise figure = F_B = 10 dB (a cheap noisier receiver is used)
- Assume car B receiver antenna gain = G_B = 3 dB (a small, low gain antenna is used to collect signal)
- Again using ground propagation model:
  `Prx_B = Ptx*G_A*G_B*h_A^2*h_B^2/(R)^4`
  Inserting numbers yields Prx_B = 2.0e-9 W = -57 dBm
- The car B receiver noise power, referred to the input, is
  Nrx_B = k*T0*B*F_B = 1.38e-23*290*5e6*10 = 2.0e-13 W = -97 dBm
- The signal to noise ratio is 40 dB, which is more than enough for error free data transmission between the cars

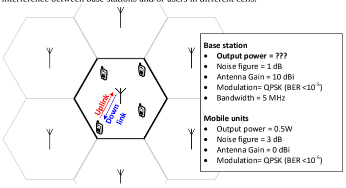
2. Receiver design
Please consider a tunable 9-12 GHz super-heterodyne receiver, for 50 Mbit/s QAM16 modulated signal at BER < 10^{-5} and with a sensitivity better than -90 dBm and a spurious free dynamic range of at least 40 dB. The receiver output is interfacing a low frequency IF A/D-converter, which you can assume to be ideal but with -30 dBm minimum input power. Your task is to present complete RF block diagram of the receiver above. All component parameters must be specified so that the receiver will meet the requirements above.

(10 points)

3. Point-to-point link
Your task is to design a 3 km, 70 GHz point-to-point communication link. The link should support 100 Mbit/s, QAM64, BER < 10^{-5}. The receiver has a noise figure of 3 dB and 20 cm diameter parabolic antennas (e_{rad} = 80%) on both TX and RX. Please calculate the required transmit power under the following conditions:
a) Free-space propagation
b) Rice fading (K = 10 dB, outage probability <10^{-3}) + clear sky
c) Rice fading (K = 10 dB, outage probability <10^{-3}) + heavy rain attenuation.

(10 points)

4. Cellular network planning
You should design a traditional cellular mobile network at 2 GHz (downlink ≈ 1.9 GHz, uplink ≈ 2.1 GHz) for a dense urban area. There are max 100 simultaneously active users per square km. A minimum of 100 kbit/s downlink capacity per user, at an outage probability of <1% is required. Base station and mobile phone characteristics are given in the figure below. You do not need to worry about interference between base stations and/or users in different cells.



a) Calculate the minimum base station output power
b) How many base stations are needed to cover a city with an area of 10 km × 10 km?
c) Calculate the minimum up-link data rate for the mobile users.

(10 points)

```
% We have the following equations
% From receiver sensitivity
% Pr_min_UL = k*T0*Rb_UL/spec_eff*F_mobile*SNR;

% Received power from the transmitter
% Pr_min_UL = Pt_mobile*G_basestation*G_mobile*path_gain*
Rayleigh_fading;
%
% Here, Rb_UL is the only unknown --> it can be solved for:

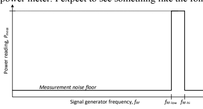
Rb_UL = Pt_mobile*G_basestation*G_mobile*path_gain*Rayleigh_fading *
spec_eff/ (F_mobile*SNR*k*T0)

Rb_UL = 190e3
```

However, the total base station receiver bandwidth is limited to 5 MHz. So, as calculated in problem 4a, the 100 users will anyway be limited to 100 kbit/s.

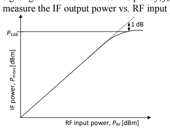
Answer: The users will have an uplink capacity of 100 kbit/s in this example.

Problem 2
I will first connect the RF signal generator to the RX input, set a low power (e.g. P_{RF-set} = -50 dBm) sweep the frequency (f_{RF}) and measure the IF output power with the power meter. I expect to see something like the following result:



The receiver will only pass the RF signal from the signal generator when the frequency is within its pass band. Hence, from the measurements and the sketch, the **RF center frequency** will be in the middle of the band: $f_c = (f_{RF-low} + f_{RF-hi})/2$.
The **bandwidth** is given by: $B = f_{RF-hi} – f_{RF-low}$.
The **receiver gain** is given by the dB-difference between the power set on the generator, and the power measured in-band: G [dB] = $P_{IF-inband}$ [dBm] – P_{RF-set} [dBm].

To find the IIP3, I will use the same experimental setup as above. I will tune the signal generator to the center frequency, f_c and sweep the RF signal power, P_{RF-set}. I measure the IF output power vs. RF input power. I expect to see something like this:



From the 1-dB compression point P_{1dB}, I estimate OIP3 [dBm] = P_{1dB} [dBm] +10 dB. With the receiver gain we can then get the **input third-order intercept point**:
IIP3 [dBm] = OIP3[dBm] – G[dB]

The receiver noise figure is found by connecting the 290K resistor as a noise source at the receiver input and measuring the IF noise output power. Similar to the top figure above, the IF power will be dominated by the in-band receiver noise power. The expression for the noise output power is: $N_{out} = k*B*(T_e+290)*G$, where N_{out} is the total power meter reading, B and G the bandwidth and gain from above, and T_e the receiver equivalent noise temperature, and 290 being the noise temperature of the resistor at the input. This yields: $T_e = N_{out}/(k*B*G) – 290$. From this, the **receiver noise figure** is given by $F = 1 + T_e/290 = N_{out}/(k*B*G*290)$

Solutions
Problem 1:
Total capacity is C=MxNxRxS
M=nbr of fibers (given =4), N=nbr of wavelength channels, R=symbol rate per channel, S=spectral efficiency in bits/symbol of the chosen modulation format.

First make assumptions on fiber attenuation (e.g. 0.18 dB/km), amplifier spacing (e.g. 50 km) dispersion compensation and its attenuation influence.

Assume baudrate and channel spacing, with some guard band margin, calculate from this the total SNR per channel.

Consider the C-band and how many channels will fit.

Check that the total output power is reasonable to avoid NL penalties (e.g. < 10 dBm), select a format compatible with this SNR, and a receiver (direct detection or coherent) based on the chosen format.

Sketch system and specify choices of components in the Tx and Rx. Sketch Rx/Tx layouts.

Reasonable guesstimate based on experience without any calculations: 32 Gbaud, coherent PM-QPSK (=100 Gb/s per channel, deducting FEC OH) , 50 GHz channel spacing (80 channels in the C-band), gives 0.1*80*=8 Tb/s per fiber and 32 Tb/s in total for the cable.

Problem 2
Two conversion steps are needed; the first IF must be high enough to avoid the image frequency to fall within the 9-12 GHz band. One good example could be to use 3.5 GHz, corresponding to an LO range of 5.5 – 8.5 GHz. A second conversion step is needed to facilitate the filtering of the modulated signal to the desired 15 MHz (50Mbits/s / 4 bit/s/Hz for QAM16).

We need to calculate the receiver noise floor under realistic noise figure assumptions, to see if the sensitivity can be met.

```
Rb = 50e6; % Given in problem
spec_eff = 4; % QAM16
B = Rb/spec_eff;
Eb_no = 10^(13.4/10); % BER <1e-5 for QAM16
Pr = 1e-3*10(-90/10); % Sensitivity given in the problem
Frx = 10^(3/10);
```

Frx = 0.22 dB, which is very challenging. The total receiver gain should be the difference between minimum input power and ADC power level, which is -30 dBm. This yields a gain of >60 dB, which should be divided between the different conversion stages.
The output noise power is given by: `No = k*T0*B*Frx*Grx.` The spurious free dynamic range can be used to determine the receiver output intercept point:
`OIP3 = SFDR^(3/2)*No = 10.5 dBm`
Use IP3 cascade formula to verify that the receiver components provide sufficient OIP3.

Wireless and Photonic System Engineering SSY085
2019-11-01, 08.30-12.30

Teachers in charge:
Dhecha Nopchinda, phone 076-651 3144, email: dhecha@chalmers.se
Christian Fager, phone: 031-772 5047, email: christian.fager@chalmers.se

Aids: Open book examination. Any printed material and calculator of choice is allowed. Communication devices (computers, mobile phones etc.) are *not* allowed.

Examination checking: Wednesday Nov. 20th, 12:00 – 12:30, room A604.

Convince yourself that you have understood the problem before you get started. Constructive and valuable gambits will also give points. If information is lacking in the description of the task, you must yourself introduce technically plausible assumptions. Make sure you clearly state such assumptions.
Grades: 3: ≥24, 4: ≥36, 5: ≥48

1. Mars – Earth communication
Your task is to design the next generation Mars – Earth communication system. The system should support up to 50 terminals anywhere on Mars surface (vehicles, bases, astronauts, etc.), each with an up- and down-link capacity of 200 kbit/s.

The Mars terminals communicate with Earth through relay satellites in Mars-synchronous orbits, see Fig. 2 below. The satellites share data using ideal free space optical links (not part of your design).

The relay satellites communicate with Earth through the NASA Deep Space Network (DSN) whereby at least one of three 34m-diameter antennas sites, located in different parts of the Earth, provide line-of-sight connection to the Mars satellites. DSN Earth transmitters: f = 34.45 GHz, P_{max} = 500W; DSN Earth station receivers: f = 32.05 GHz, T_{sys} = 28K.

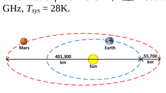


Fig. 1: Earth and Mars solar orbits.

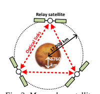


Fig. 2: Mars relay satellite configuration.

Fig. 3: DSN site in Goldstone/US.

Your solution should include the design of transmitters and receivers for both the Mars terminals and the relay-satellites, including frequencies, modulation format, block diagrams, antennas, output power, etc. Use realistic assumptions where applicable.

(30 points)

Problem 3
a) The receiver sensitivity for BER=10e9 can be obtained from the condition Q=(I1-I0)/2σT = RP1/2σT=RPav/σT=6, from which we get Pav=(6σT)/R (assuming thermal noise dominates). The bandwidth of the diode is 300 MHz, from which we find the thermal noise σT=sqrt[4kTΔf/RL]=3.15e-7 [A] (using 50 ohms and 300 K). The responsivity at 650 nm (ν = 461 THz) is R= η/hν=0.524 [A/W] (assuming unit quantum efficiency). The receiver sensitivity then becomes Pav=3.61e-6 W= – 24.4 dBm. If a signal uses less bandwidth than 300 MHz we can filter this noise to the used bandwidth and increase the sensitivity.
b) We thus need to make a risetime budget: $\sigma_{sys}^2=\sigma_{tx}^2+\sigma_{rx}^2+\sigma_{fib}^2$ where we use the inverse bandwidths as risetimes for the transmitter and receiver, i.e. σ_{tx} =1/0.2 GHz=5 ns, σ_{rx} =1/0.3 GHz=3.33 ns. The fiber dispersion gives a risetime according to σ_{fib} = LΔn1/(2c)=L(n1-n2)/2c = 20*(1.492-1.402)/(2*3e8) =3.0 ns. This give a system rise time of σ_{sys}=6.72 ns, and if we use as a design condition that this is to be 70 % of the bit period, we get a bit period of 6.72/0.7=9.6 ns, which corresponds to a bit rate of B=104 Mb/s.
c) If we filter the signal from the pin-diode with 104 MHz, we increase the sensitivity from problem a) a factor sqrt(300/104)=sqrt(3)=2.4 dB, which gives a receiver sensitivity of -24.4-2.4=-26.8 dBm. With a transmit power of 0 dBm, we can thus allow for 26.8 dB of losses. The fiber losses amounts to 20m*0.14 dB/m=2.8 dB, which leaves 24 dB for other losses such as coupling, bend losses etc.

Problem 4
Case 1 - 16-QAM: Both formats indicate that coherent receivers are used, and in that case the SNR is the same as in the optical domain, i.e. SNR=P_{sig}/P_{ase} with $P_{ase}=S_{ase}$ Δf where the used bandwidth is Δf=30 GHz and the power spectral density of the ASE noise is S_{ase}=N F G hν/2, where N=8 is the number of amplifiers, F=$10^{0.5}$=3.1 is the noise figure, G=$10^{1.8}$ = 63.1 is the gain and hν = 1.28e-19 J is the photon energy. We thus get the SNR = $2P_s$/(NFG hν Δf). The required signal power for 16 QAM is then P_{16-QAM}=100NFG hν Δf/4=0.385 mW. With an available total signal power of $10^{1.5}$=31.6 mW we can sustain 31.6/0.385 =82.1, i.e., 82 channels. The total data rate becomes 30 Gsymb/s*4 bits/symb*82 channels=9.84 Tbit/s.

Case 2 - QPSK: As above we find P_{QPSK}=($10^{1.4}$) NFG hν Δf/2=0.096 mW. We can sustain $10^{1.5}$/0.096 =328 channels. Note however that C-band is not sufficient for this, since the total available channel count is 4.4THz/50 GHz=88 channels. We will thus be limited by the available bandwidth to 88 channels. The total data rate is then 30 Gsymb/s*2 bits/symb*88 channels=5.28 Tbit/s.

Thus 16-QAM is a better choice in this case.

Problem 3
Matlab script for solution

```
k = 1.38e-23;
T0 = 290;
lambda = 3e8/70e9;
Rb = 100e6;
spec_eff = 6; % QAM64
B = Rb/spec_eff;

% Calculate the antenna gain
d_ant = 0.2;
erad = 0.8;
A = pi*d_ant^2/4;
D = 4*pi*A/(lambda^2);
G = D*erad;

% Calculate the minimum receive power
R = 3e3; % 3km link distance
F = 10^(3/10);
Eb_no = 10^(17.8/10); % BER <1e-5 för QAM64
SNR = Eb_no*spec_eff;
Pr_min = k*T0*B*F*SNR;

%% a) Calculate the necessary transmit power under free space
operating conditions
Pt = Pr_min*(4*pi*R)^2/(G*G*lambda^2);
Pt_dBm = 10*log10(Pt/1e-3)
Pt_dBm = -19 dBm
Answer: Transmit power under free space conditions = -19 dBm

%% b) Rice fading (K = 10 dB, outage probability <10^-3) + clear sky
% From Fading Systems lecture, slide 31, we can read that outage
probability for Rice fading (K = 10dB) corresponds to a fading dip of
ca 10dB. The output power must therefore be increased by 10dB to
guarantee the link performance
Pt_dBm_wRice = -19 + 10 = -9 dBm
Answer: Transmit power under Rice fading conditions = -9 dBm

%% c) Rice fading (K = 10 dB, outage probability <10^-3) + heavy rain
% Again, from Fading Systems lecture, slide 17, we can read that the
atmospheric attenuation at 70 GHz and heavy rain is in the order of
20dB/km. For a link distance of 3km, this gives 3*20 = 60 dB
additional attenuation. The output power must therefore be increased
by an additional 60 dB to guarantee the link performance


Pt_dBm_wRain_wRice = -19 + 10 + 60 = 51 dBm
Answer: Transmit power under Rice fading conditions and heavy rain =
51 dBm = 125 W. This is unrealistic.
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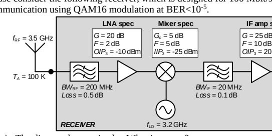
Please consider the following receiver, which is designed for 100 Mbit/s communication using QAM16 modulation at BER<10^{-5}.



a) The diagram has a mistake. What is wrong?
b) Calculate the receiver sensitivity when T_A = 100K.
c) You wish to increase the linear dynamic range by 10 dB, while maintaining the overall receiver gain. Which component do you need to replace, and what would the new component specifications be?

(10 points)

3. Dispersive broadening
Consider a standard single mode fiber link transmitting OOK data at 1550 nm. Dispersion compensating elements are not available. Formulate your own design requirement and assume reasonable parameter values to answer the following questions:
What is the dispersion-limited transmission distance for OOK-modulated data at
a) a data rate of 10 Gb/s, and
b) a data rate of 40 Gb/s ?
c) Generally, how does the dispersion-limited distance scale with the data rate in a standard single mode fiber?

(10 points)

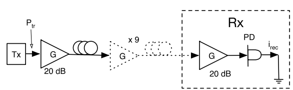
4. A preamplified receiver
A wavelength-division multiplexed system consists of 10 EDFA spans with span losses of 20 dB, as shown in the figure. The receiver has an optical preamplifier with gain 20 dB as shown in the figure. All 11 amplifiers are similar with a noise figure of 6 dB. The channel spacing is 200 GHz and each channel transmits 40 Gb/s OOK. The photodetector quantum efficiency is 1.

a) Calculate all noise variances after the photodetection at a transmit power of P_{tr} = -10 dBm. Which noise source dominates?

(7 points)

b) What transmit power P_{tr} is required to get a BER < 10^{-9}?

(3 points)



Problem 4: A preamplified photonic receiver

Department of Microtechnology and Nanoscience

Wireless and Photonic System Engineering SSY085
2019-01-09, 08.30-12.30

Teachers in charge:
Magnus Karlsson, phone 772 1590, email: magnus.karlsson@chalmers.se
Christian Fager, phone 772 5047, email: christian.fager@chalmers.se

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Examination checking: by request from Christian or Magnus

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Grades: 3: ≥24, 4: ≥36, 5: ≥48

1. The Curie cable
The below press release describes Google's new "Curie cable" which will go live 2019. Notably, it does not mention the total data capacity of the cable. Your challenge is to use your engineering skills to estimate what the total data rate this cable can supply (in Tb/s). Since no data other than the transmission distance and number of fibers used is given, you are expected to make many assumptions of your own, which need to be clearly specified.
The "branching unit" halfways (outside Panama) can be assumed to be a *regenerator site* where the data is detected electrically and retransmitted, making the longest transmission distance you need to design for 5000 km.

(30 points)

The Curie submarine cable system is a four-fibre-pair and 10,000km cable connecting Los Angeles, California, and Valparaíso, Chile, with a branching unit for future connectivity to Panama.

Named after physicist and chemist Marie Curie, the Curie cable system will make Google the first major non-telecom company to build a private intercontinental cable. Google claims it will be the first new cable to land in Chile in almost 20 years, and will become the largest single data pipe connecting the country. The Curie cable system is supplied by TE SubCom and is set to go live in 2019.

Source: TE connectivity press release, Jan 16, 2018.

Problem 4

```
%% a) Calculate the minimum base station output power
lambda_DL = 3e8/1.9e9;
spec_eff = 2; % QPSK
B_tot_DL = 5e6;
Rb_tot_DL = spec_eff*B_tot_DL;
Rb_per_user_DL = 100e3; % According to system spec
Users_per_basestation = Rb_tot_DL/Rb_per_user_DL;
Users_per_km2 = 100; % User density --> area per base station
Area_per_base_station = Users_per_basestation / Users_per_km2;
base_station_radius = 1000*sqrt(Area_per_base_station/pi); % Assume
circular coverage area

% Mobile phone sensitivity
k = 1.38e-23;
T0 = 290;
F_mobile = 10^(3/10); % Specified noise figure
Eb_no = 10^(9.6/10); % BER <1e-5 for QPSK
SNR = Eb_no*spec_eff;
Pr_min_DL = k*T0*Rb_per_user_DL/spec_eff*F_mobile*SNR;

% Rayleigh fading_margin
p_outage = 0.01;
Rayleigh_fading = -log(1-p_outage); % p_outage = 1-exp(-fade_margin)
Pr_min_fading_DL = Pr_min_DL / Rayleigh_fading;

% Path gain (= 1/path loss). Dense urban area --> Modified Friis (N =
3, R0 = 5m)
N = 3;
R0 = 5;
path_gain = lambda_DL^2/(4*pi*R0)^2/(base_station_radius/R0)^N;

% Antenna gains
G_basestation = 10^(5/10);
G_mobile = 10^(0/10);

% Combine to calculate the required transmit power
Pt_basestation =
Users_per_basestation*Pr_min_fading_DL/(G_basestation*G_mobile)/path_
gain

Pt_basestation = 5.2 W
Answer: Base station transmit power = 5.2W

%% b) How many base stations are needed to cover a city with an area
of 10 km × 10 km?

Area_tot = 10*10; % Given in problem
Base_stations_tot = Area_tot/Area_per_base_station

Base_stations_tot = 100
Answer: 100 base stations are needed

%% c)Calculate the minimum up-link data rate for the mobile users
% Similar calculations as for the downlink. I will assume the same
path loss as for the downlink (the frequency is almost the same)
Pt_mobile = 0.1;
F_basestation = 10^(1/10);
```

Free space.
$R = 17000\,km - 6760\,km = 10240\,km$
Take $R = 15000\,km$ to be safe

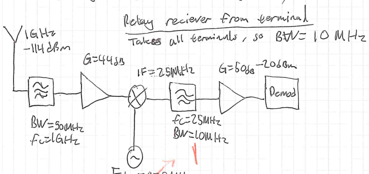
$P_r = P_t \frac{G_t G_r \lambda^2}{(4\pi R)^2}$

$P_{t\,dBm} = P_{r\,dBm} - G_t - G_r - 10\lg \frac{\lambda^2}{(4\pi R)^2}$

Choose $f_c = 10\,GHz$
$\therefore \lambda = \frac{c}{f_c} = 0.03m$

$P_t = -114\,dBm - 3\,dBi - 30\,dBi - (-176\,dB)$
$= 29\,dBm$ (Pretty much... Test lower frequency. higher antenna gain.)

Relay reciever from terminal
Takes all terminals, so BW = 10 MHz



Gtot = 94 dB
Very high gain on each step...

1 $R_B = 200$ kbit/s

$f_u = 34.45$ GHz
$P_{tu} = 500$ W
$f_d = 32.05$ GHz
$T_{sys} = 28$ K

What we need:
Relay satellite:
1 Recieve from earth
2 Transmit to earth
3 Recieve from terminal
4 Transmit to terminal
Terminal
5 Recieve from relay
6 Transmit to relay

Part 1 Recieve from Earth
$d_{earth} = 34$ m
$A_{e_1} = \frac{\pi d^2}{4} = 908\ m^2$

$D_1 = \frac{4\pi A_{e1}}{\lambda_u^2} = \left[\lambda_u = 8.7\cdot 10^{-3} m\right] = 150.7\cdot 10^6$

$= 81.8$ dBi

Assume $G_1 \approx D_1$

$P_{max} = 500$ W
$P_{channel} = \frac{500 W}{50} = 10$ W/channel

Recieved at relay
$P_r = \frac{P_t G_1 G_2 \lambda_u^2}{(4\pi R)^2}$ (Assume point to point link)

Satellite antenna size = 2m
So $A_{e2} = \pi\ m^2$

$\therefore D_2 = \frac{4\pi A_{e2}}{\lambda_u^2} = 521.6\cdot 10^3 = 57.2$ dBi

Assume $G_2 \approx D_2$

$R \in (55700\ km, 401300\ km)$
Recieved from earth:
$P_{r_1} \in (1.2\cdot 10^{-7} W,\ 2.34\cdot 10^{-9} W) = (-39.2 dBm, -56.3 dB$

Assume QPSK, $BER = 10^{-5}$
$S_{eff} = 2 \Rightarrow BW = 100$ KHz
$\frac{E_b}{N_0} = 9.5$ dB

$\therefore SNR_{out} = 10^{0.95}\cdot 3 = 12.5$ dB
Antenna temp assumed
$T_A = 50$ K

Add losses from rain:
20 dB/km · height of clouds ≈ 2 km ++

$P_{r_1} = (-79.2$ dBm, -96.3 dBm$)$

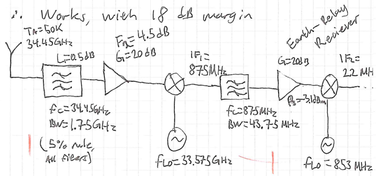
Reciever $F_{n_1} = 5$ dB

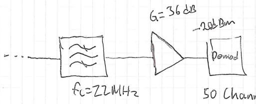
$SNR_{in} = 17.5$ dB

$N_{in} = kT_A B = 6.9\cdot 10^{-17} = -131.6$ dBm

Sensitivity $= -131.6$ dBm $+ 17.5$ dB
$= -114$ dBm

∴ Works, with 18 dB margin





50 Channels of 200 KHz spacing ⇒ 10 MHz and Bandwidth

Need total gain $G = -20$ dBm $- (-96$ dBm$)$
$= 76$

$F_{N_1} = 5$ dB $= F_1 + \frac{F_2 - 1}{G_1} + \dots$

$\Leftrightarrow F_2 = G_1\cdot(10^{0.5} - F_1) + 1 = 2.8 \approx 4.5$ dB

Dynamic range = $P_{rmax} - P_{sensitivity} = -79$ dBm $- (-114$ dBm$)$
$= 35$ dB

Assume Last amplifier most important P_3

$DR_f = \left(\frac{P_3}{N_0}\right)^{2/3}$

$\Leftrightarrow P_3 = DR_f^{3/2}\cdot N_{in}\cdot G = 4.88\cdot 10^{-4}$ W
$= -3.1$ dBm

Part 2 transmit to earth

Interpret T_{sys} as equivalent temp of reciever at earth.
Assume $T_A = 50K$ again
QPSK, $BER = 10^{-5}$

$F_n = 1 + \frac{24 K}{290K} = 1.083 = 0.35$ dB
(Very good, but maybe possible at NASA?)

$SNR_{in} = 12.5$ dB $+ 0.35 \approx 13$ dB

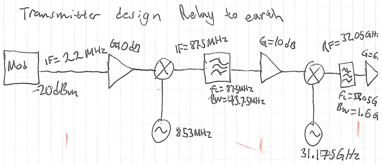
$N_{in} = (T_A + T_{sys}) kB = 1.1\cdot 10^{-17} W = -140$ dBm

$P_{sensitivity} = -130$ dBm
Assume downlink same path loss as uplink.

Path loss from part 1 = 136 dB at worst

∴ $P_t = -130$ dBm $+ 136$ dB $= 6$ dBm per channel, so in total
$P_t = 0.2$ W $= 23$ dBm.
High but reasonable. — not very high

Transmitter design Relay to earth



total Gain = 26 dB
Assume linear enough amps.

Part 3 Recieve from terminal

Antenna Gain relay = 30 dBi
Antenna Gain on terminal = 3 dBi
QPSK $BER = 10^{-5}$
$SNR_{out} = 12.5$ dB
$F_n = 5$ dB
$SNR_{in} = 17.5$ dB
Antenna temp when pointing at mars?
Maybe 50 K?
So $N_{in} = -131.6$ dBm
$P_{sensitivity} = -114$ dBm

Part 4 Transmit to terminal

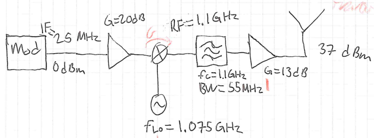
Assume same antennas, same path loss
Same modulation, very close frequency

$SNR_{in} = 17.5$ dB
Antenna temperature on mars with low gain antenna ≈ 290 K

$N_{in} = 4\cdot 10^{-16}$ W $= -124$ dBm
$P_{in} = N_{in}\cdot SNR = -124$ dBm $+ 17.5$ dB
$= -106$ dBm

∴ $P_t = -106$ dBm $+ 143$ dB $= 37$ dBm
All terminals means
$P_{tot} = 50 P_t = 54$ dBm

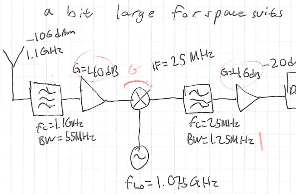
Very high for a satellite, must have very good solar panels or something.



Part 5 Recieve from relay

Antenna was $G = 3$ dBi $\approx D$
so $A_e = \frac{D\lambda^2}{4\pi} = 0.014\ m^2$

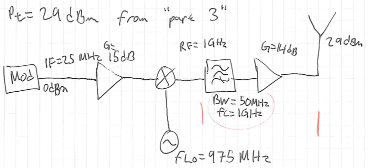
Parabolic antennas
$d = \sqrt{\frac{4A}{\pi}} = 13$ cm. Size ok, a bit large for space suits maybe.



G = 86 dB

Part 6 Transmit to relay from terminal

$P_t = 29$ dBm from "part 3"



Final words.
Shortness of time, i have left out noise figure and dynamic range calculations in part 3 and 5. They are done in the same way as in "part 1".

Very good!

Problem 1:
Solution not available yet…

Problem 2:
a) Design error
100 Mbit/s communication with QAM16 modulation requires a bandwidth of at least 25 MHz. However, in the diagram there is a design error since the IF filter has a bandwidth of only 20 MHz.

b) Receiver sensitivity
From Table 9.5 in Pozar, we have that QAM16 at BER<10^{-5} requires Eb/n0 = 13.4 dB. With a spectral efficiency of 4, this means that the output SNR must be at least SNR = 13.4 + 6 dB = 19.4 dB = 87.5.
The total receiver noise figure, F_{tot}, can be calculated using the cascade formula.

$F_{tot} = F_{filter} + (F_{LNA}-1)/G_{filter} + \varepsilon$

where F_{filter} = 0.5 dB is the filter loss, and G_{filter} = -0.5 dB. ε represents noise contributions from later stages, which become negligible since the gain of the LNA is very high. Inserting numbers, we get F_{tot} = 1.78 = 2.5 dB, which corresponds to an equivalent noise temperature of T_e = 290(F-1) = 226.2 K.

Referring all receiver noise sources to the input, we have $N_{in} = k*(T_A+T_e)*B$ = 1.38e-23*326.2*25e6 = 1.15e-13 W = -99.5 dBm. Together with the SNR requirement set by the modulation, we get the receiver sensitivity, i.e. the minimum input power required:
$P_{r,min}$ = SNR*N_{in} = 87.5*1.15e-13 = **1.0e-11 W** = **-80.0 dBm**

c) Dynamic range improvement by 10 dB
Linear dynamic range is defined by $DR_l = (OIP_3/N_{out})$, where OIP3 is the output third order intercept point of the receiver. The output noise is not easy to improve since the gain should be the same, and the noise is determined by the noise figure of the first stage. It is already 2 dB, so we can't improve it by 10 dB. First, we therefore need to determine which component is limiting the third order intercept point. This is easiest if we consider $OIP_3 = P_{1dB}$ + 10dB, i.e. which component has the most limiting output power.
The IF amp has $OIP_{3,IF}$ = 20 dBm and, therefore, $P_{1dB,IF}$ = 10 dBm. Referred to its input, and using its 25 dB gain, this corresponds to an input compression power of -15 dBm. The mixer, on the other hand, has an input intercept point, $IIP_{3,mix}$ = -25 dBm, which together with its 5 dB gain corresponds to $OIP_{3,mix}$ = -20 dBm or an output 1-dB compression power of $P_{1dB,mix}$ = -30 dBm, which is significantly lower than the -15 dBm input power where the IF amplifier will compress. Using a similar reasoning, the LNA compression point is also clearly higher than the input compression to the mixer. Thus, in this case, the mixer IP3 therefore limit the dynamic range.
In order to find how much we need to improve the mixer IP3, we set up the expression for the total OIP3:

$$\frac{1}{OIP_{3,tot}} = \frac{1}{OIP_{3,LNA}G_{mix}G_{filt}G_{IF}} + \frac{1}{OIP_{3,mix}G_{filt}G_{IF}} + \frac{1}{OIP_{3,IF}}$$

Thus, to improve OIP3tot by a factor of 10 (10dB), we need to change $OIP_{3,mix}$:

$$\frac{1}{10OIP_{3,tot}} = \frac{1}{OIP_{3,LNA}G_{mix}G_{filt}G_{IF}} + \frac{1}{OIP_{3,mix,new}G_{filt}G_{IF}} + \frac{1}{OIP_{3,IF}}$$

We can eliminate $OIP_{3,tot}$ by combining these expressions. This gives, after some manipulations, the following expression for $OIP_{3,mix,new}$

$$OIP_{3,mix,new} = \frac{10}{\frac{1}{OIP_{3,mix}} - 9\left(\frac{1}{OIP_{3,LNA}G_{mix}} + \frac{G_{filt}G_{IF}}{OIP_{3,IF}}\right)}$$

Inserting the given numbers yields $OIP_{3,mix,new}$ = 0.22 mW = -6.4 dBm. Referred to the input, this is equivalent to **$IIP_{3,mix,new}$ = -11.4 dBm**.

Thus, in summary, the mixer has to be replaced with another one having IIP_3 > -11.4 dBm, in order to improve the linear dynamic range of the receiver by 10 dB.

Problem 3:
A typical design criterion could be that the dispersive broadening can be only 25 % of the bit period T. The dispersive broadening is Δt=D Δλ L, where D=16 ps/nm km and Δλ denotes the signal bandwidth in [nm] units. We need to relate bandwidth in nm to GHz, and using Δλ/λ=Δf/f we find that Δλ =1 nm corresponds to Δf =125 GHz, or 1 GHz is 0.008 nm. Thus 10 GHz corresponds to 0.08 nm, and 0.32 nm corresponds to 40 GHz.

The allowed Δt for 10 GHz is (given our 25 % criterion) 100/4=25 ps and for 40 GHz it is 25/4=6.25 ps.
This gives for 10 Gb/s: L_d= 25/(16*0.08)=19.5 km
and for 40 Gb/s: L_d= 6.25/(16*0.32)=1.22 km

Finally the scaling: $L_d \sim \Delta t/(D\,\Delta\lambda)$ and since Δt scales as ~ 1/Δf and Δλ ~ Δf we have that $L_d \sim \Delta f^{-2}$. That 19.5/1.22=16 should be a hint as well.

Problem 4:
We have 4 noise sources to calculate; thermal shot, signal-spontaneous and spontaneous-spontaneous beat noise.

We need some data to do this. The received optical power is P_r = 10 dBm (note that the preamp makes the receive optical power higher than the transmit power). The received amplifier noise PSD is
S_{sp}=N G hν NF/2=2.78e-16 [W/Hz]
where N is 11, G=100, the photon energy hν =1.28e-19 J, and the noise figure is NF=$10^{0.6}$=3.98. The photodetector has a responsivity of R=q/hν=1.25 A/W.
The bandwidth Δf=40 GHz and the optical bandwidth is given by the channel spacing to be Δν=200 GHz.

This gives:
σ_t^2= 4kT Δf/R_L= 1.33e-11 [A^2]
σ_s^2= 2q(RP_r)Δf= 1.60e-11 [A^2]
$\sigma_{s\text{-}sp}^2$= 4RS_{sp}(RP_r) Δf= 6.95e-8 [A^2]
$\sigma_{sp\text{-}sp}^2$= 4$(RS_{sp})^2$ Δν Δf = 3.86e-9 [A^2]

We see that s-sp noise is largest, as expected. We can neglect the others and solve prob b) with only s-sp noise.

$I_1=2RP_r$ (assuming P_r is the average received power)
$I_0= \sigma_0=0$
$\sigma_1= \sigma_{s\text{-}sp} =\sqrt{4RS_{sp}(RP_r)\,\Delta f}$.

This gives $Q^2=P_r/(S_{sp}\,\Delta f)$=36 from which we find P_r=36 S_{sp} Δf=4.0 e-4 W=0.4 mW= - 4 dBm.
The transmit power is 20 dB lower than this and thus -24 dBm.
(Note: A more accurate but also more complicated result to derive is if the sp-sp noise is not neglected. This gives a 2.5 dB higher power of -21.5 dBm.)

4 To think about designing systems

4.1 Wireless

- Frequency planning
- Explain all assumtions
- Parameter values:
  Antennas: e_{rad}, G, T_A, f
  Filters: f_c, B, L
  Mixers: P_3, P_1, F, G
  Local oscillators: f_{LO},
- RX sensitivity
- Discussion on parameter values
- Dynamic range
- Duplexing

4.2 Photonics

- Component choices, for example: external/directly modulated, laser type, detectors, amplifiers, dispersion compensation, SMF/MMF/GRIN, etc.
- Total data capacity:

  $$C_{tot} = N_{fibers} \times N_{\lambda-channels} \times R_{symbol} \times Sp.Eff.$$

- Guard bands

Design flow for long-haul fiber system (excerpt from Magnus tips):

1. Assume reasonable channel spacing, modulation format and symbol rate to get the desired bitrate. This gives the required bandwidth (=symbol rate) per channel which is needed for noise calculations.
2. Compensate the dispersion, using reasonable assumptions for DCF (or fiber Bragg gratings) and SMF including parameters. This leads to the total system losses. If you have a coherent receiver the dispersion can be compensated electronincally in the receiver. The drawback is the coherent system is more complex, and you must mention this fact as one part of the coherent DSP
3. Assume a reasonable amplifier spacing. Reasonable span losses is around 10-20 dB for ultralong (usually submarine) links and 20-30 dB for shorter terrestrial links. This info together with the total system losses leads to the required number of spans.
4. Use the required Q (for OOK) or SNR (for phase modulation or QAM) to get a transmit power per channel (which you want not much larger than 0 dBm). If too high (e.g. total signal power summed over all channels $>$ 20 dBm) consider either: reducing amplifier span loss (more amplifiers), reducing channel bandwidth, or using FEC and design with a lower Q/SNR.
5. Check all noise sources and verify that s-sp (or LO-sp) dominates.