Lecture slides for courses based on the textbook by A. F. Molisch

Wireless Communications

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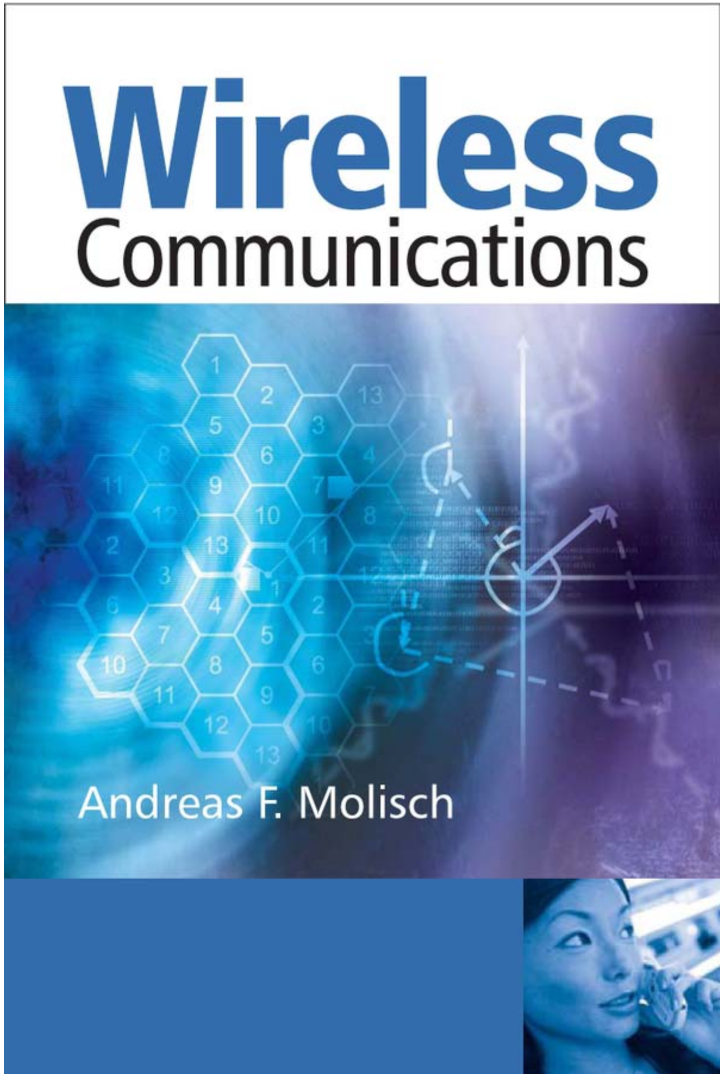
Textbook



Chapter 1

Introduction

Contents

- What are radio systems?
- History of wireless systems
- Classifications
- Requirements for services
- Social and economic aspects

WHAT IS A RADIO SYSTEM?

Radio system?

- From Merriam-Webster Dictionary
  - Radio:

    1 : of, relating to, or operated by radiant energy

    2 : of or relating to electric currents or phenomena (as electromagnetic radiation) of frequencies between about 15 kHz and 100 GHz
  - System:

    1 : a regularly interacting or interdependent group of items forming a unified whole
- ”Radio systems” can be used for many purposes, e.g.
  - Detection and ranging (Radar)
  - Astronomical observation (Radio telescope)
  - Heating food (Microwave oven)
  - Navigation (GPS, etc.)
  - Communication (Cellular telephony, etc.)

Some questions to ask

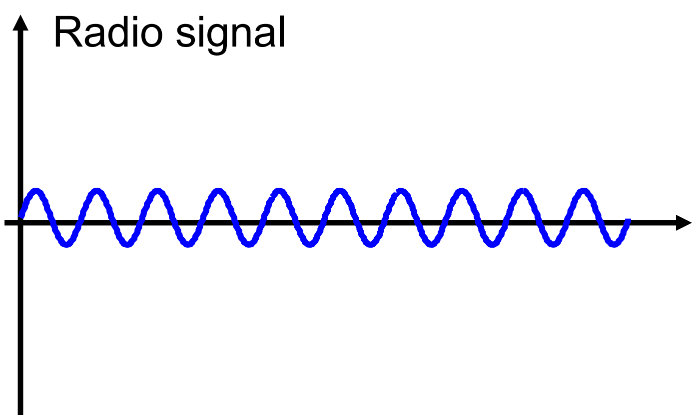
- What do we want to achieve with our system?
  - This gives us design constraints (system requirements)
- What frequency band should we use?
  - Properties of the radio channel changes with frequency
  - Radio spectrum is firmly regulated
- Which technology should we use?
  - Not all technologies can perform the task
  - Cost is important (design, production, deployment, etc.)

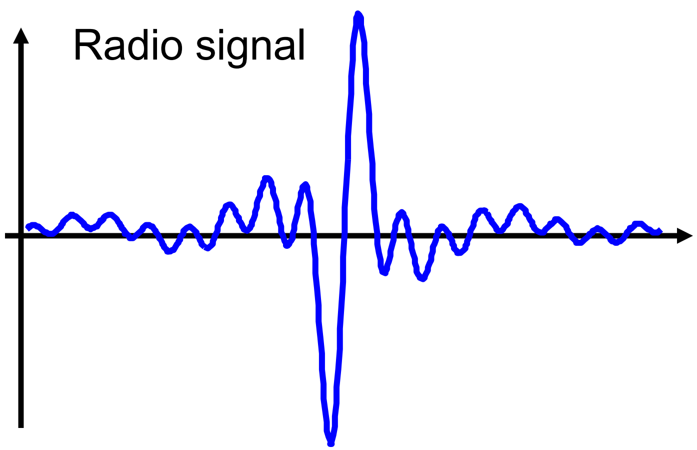
Example: Mobile telephony

Amplifiers with low dynamic range can be made more power efficient than highly linear amplifiers.

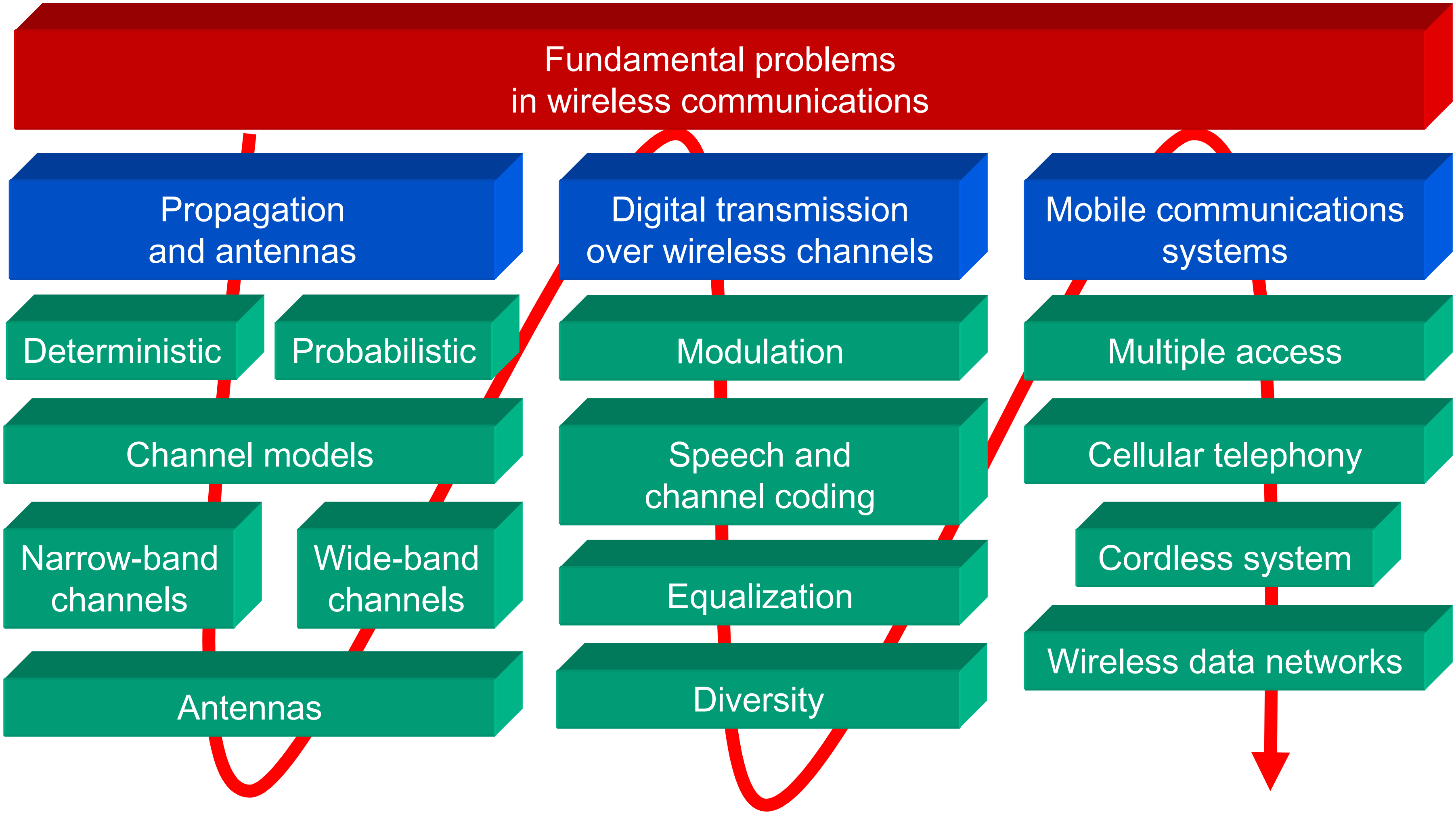
Does this affect the choice of modulation technique?

Copyright: Sony-Ericsson





A rough breakdown into areas



HISTORY OF WIRELESS

History of wireless (1)

- Maxwell: theory
- Hertz: fundamental experiments confirming Maxwell's theory
- 1890-1905: First experiments for wireless information transmission
  - Tesla, Bose, Marconi
- 1905-1946: First systems:
- 1947/1948: fundamental information theory

History of wireless (2)

- Development of cellular telephony: 1950s – 1980s
- Cellular systems
  - GSM (Global System for Mobile communications)
    - First deployment in early 1990s in Europe
    - In 2005, more than 1 billion users
  - IS-95 (cdmaOne)
    - Second-generation system based on different multiple-access scheme
    - Used mainly in US and Korea
  - PDC (Pacific Digital Cellular)
  - Third-generation systems
    - Several different standards
    - All based on CDMA
    - After initial goldrush, now more sedate development

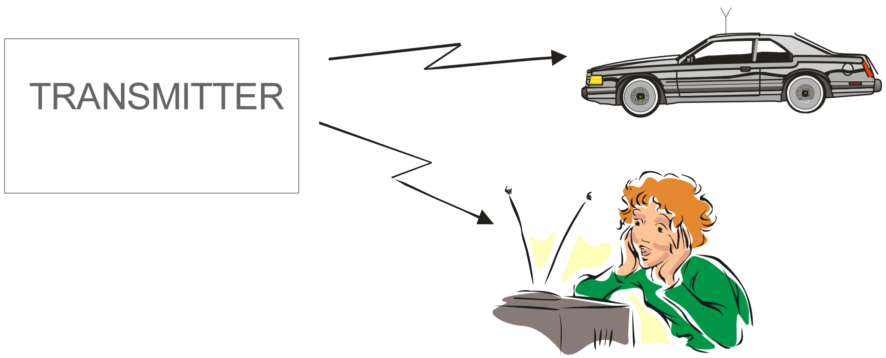
New wireless systems

- Fixed wireless access:
  - for wireless internet
  - Not successful in 1990s
  - In recent years, WiMax was developed, seems more promising
- Cordless phones:
  - DECT: in Europe and most of the world
  - PHS: in Japan
  - CDMA-based: in US
- Wireless LANs
  - Wireless computer networks
  - WiFi
- Personal Area Networks

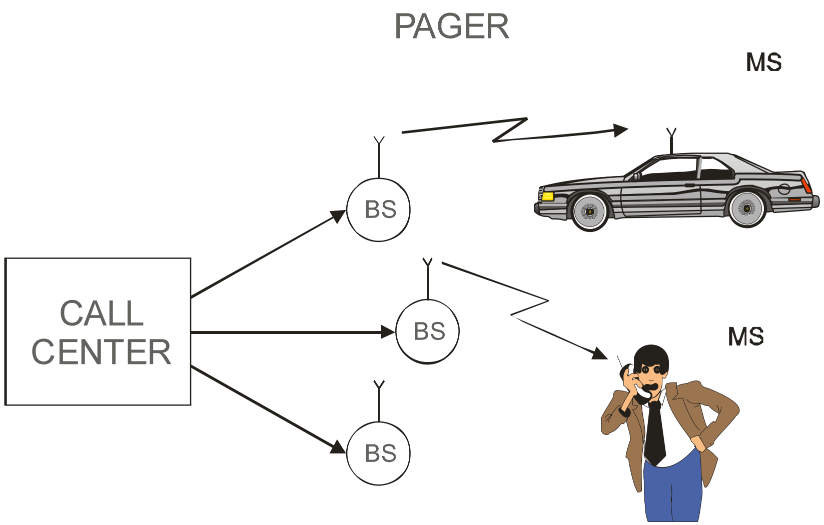
TYPES OF SERVICES

Broadcast

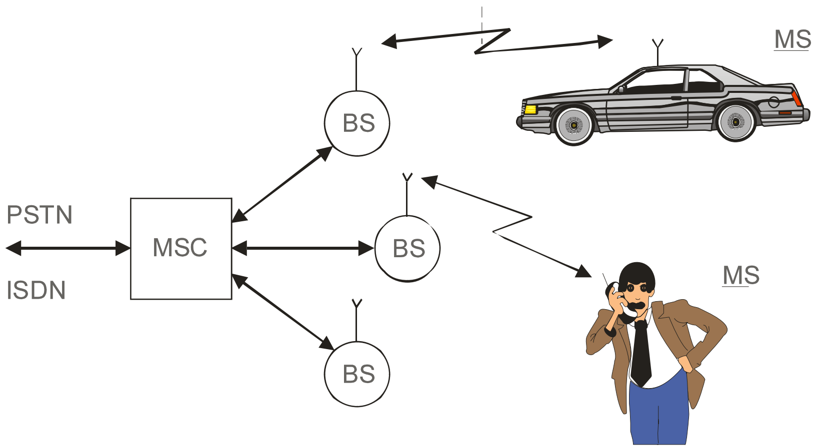
BROADCAST



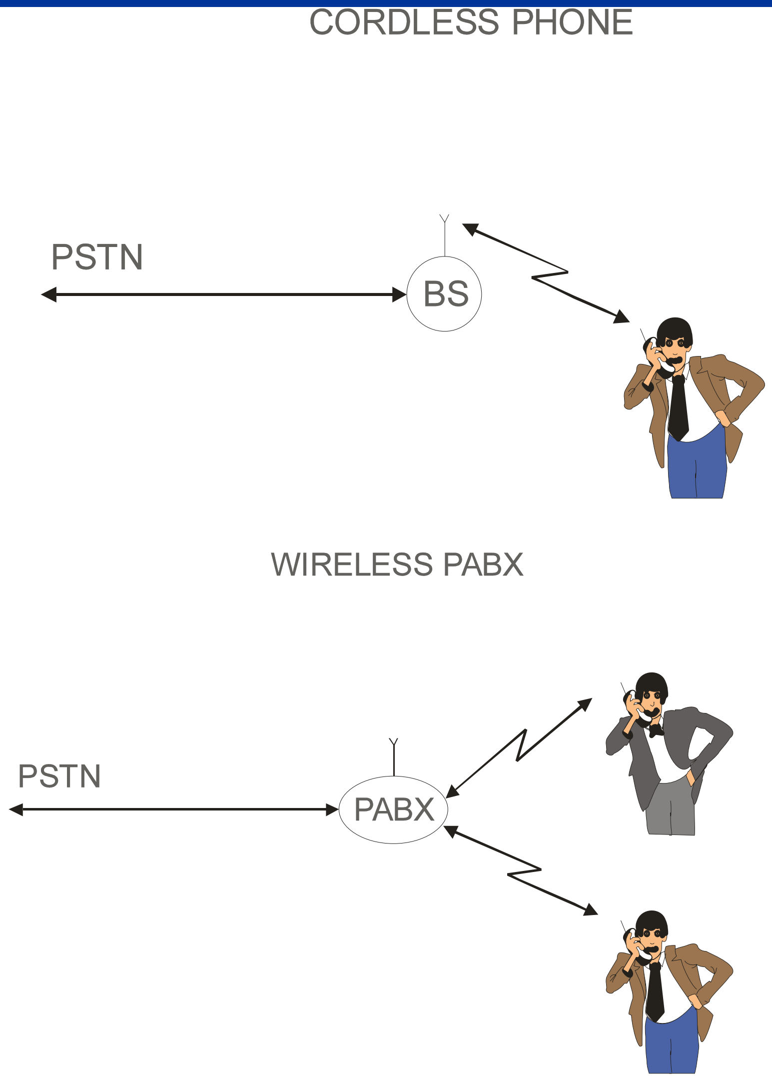
Paging



Cellular phones

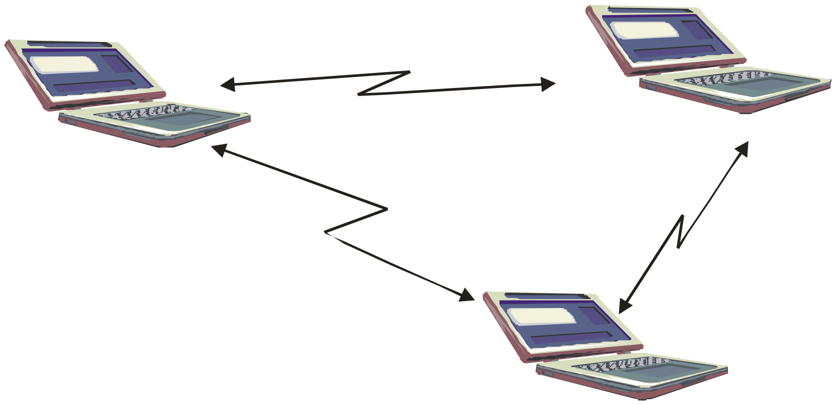


Cordless phones



Wireless LANs and PANs

AD-HOC NETWORK



Fixed wireless and satellite

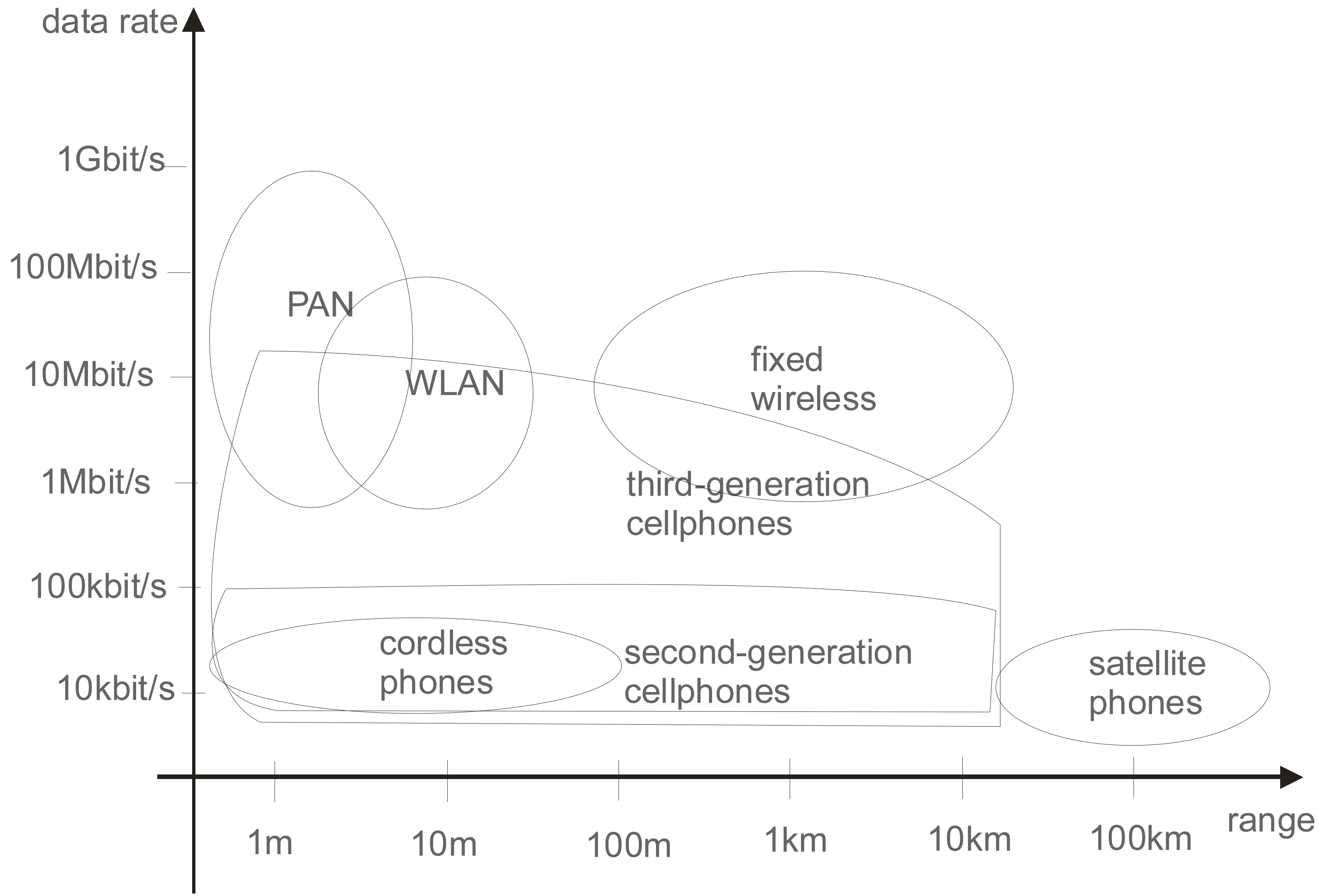
- Fixed wireless systems
  - Long distances between BS and MS
  - No mobility requirements
  - Typically high data rates, but can also be used for voice systems
  - WiMax standard (IEEE 802.16)
- Satellite systems
  - Cover very large area
  - No high density (Erlang/km^2)
  - Iridium system with LEO systems tried to get large user density, but went broke

REQUIREMENTS FOR SERVICES

Data rate

- Sensor networks: <1kbit/s; central nodes need up to 10 Mbit/s
- Speech communications: 5-64 kbit/s, depending in speech coder (vocoder)
- Elementary data services: 10-100 kbit/s
- Communications between computer peripherals: 1 Mbit/s
- Wireless LANs: broadband internet speeds, 1-100 Mbit/s
- Personal Area Networks: >100 Mbit/s

Tradeoff range vs. datarate



Mobility

- Fixed devices: stay in one location; temporal variations due to moving objects in surroundings
- Nomadic devices: MS placed at certain location, stays there for a while (WLANs)
- Low mobility: pedestrian speeds (cordless phones)
- High speed: cellphones in cars
- Extremely high speed: high-speed trains, planes, ….

Spectrum usage

- Spectrum dedicated to specific service and operator
- Spectrum dedicated to specific service
- Free spectrum

ECONOMIC AND SOCIAL IMPACT

Economic requirements

- Systems where mobility is a value by itself
  - Cellphones, etc.
  - Can charge premium for service
- Systems that just are cable replacement
  - E.g., for fixed wireless access
  - Must be cheaper than cabled service
- In either case, quality has to be same than wired
- Systems should contain as many digital components as possible to reduce costs

Behavioral impact

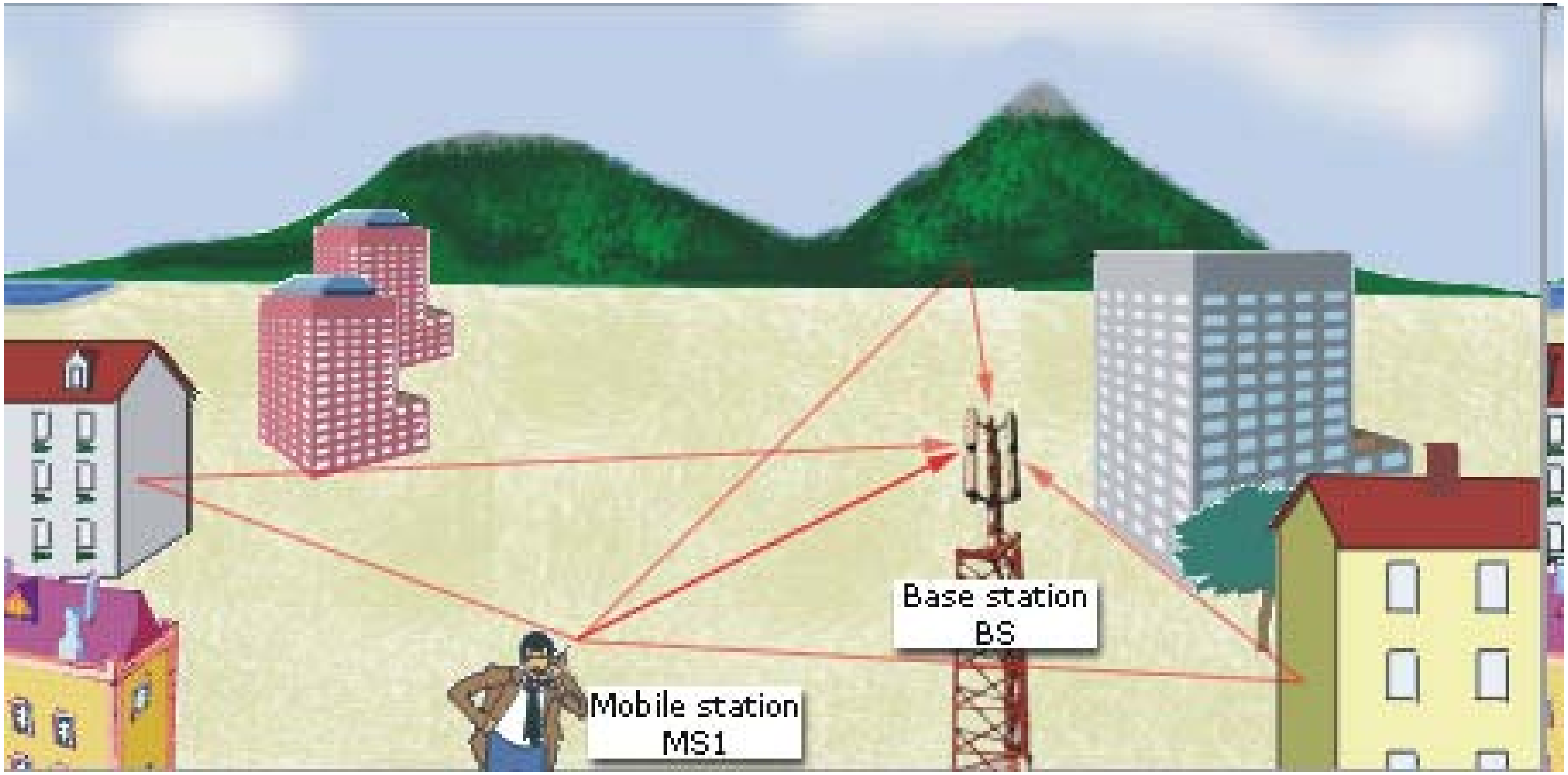
- Communications are now with a person, not with a location
- Allows more flexibility for private/business life, but can also become electronic “ball and chain”
- Cellphone etiquette: generally underdeveloped
- Phoning while driving is dangerous
- Each cellphone has an OFF button

Technical challenges of wireless communications

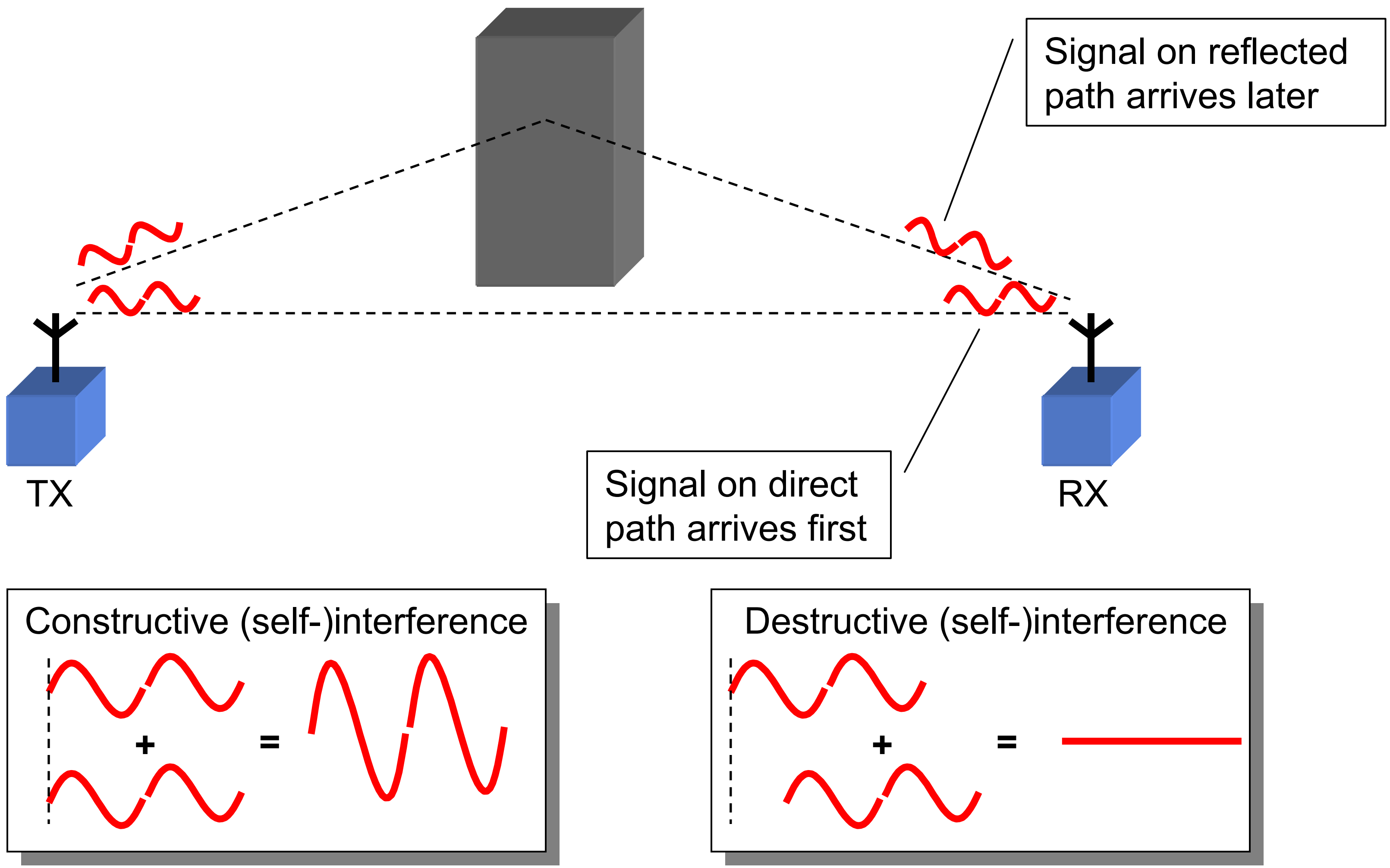
The major challenges

- Multipath propagation
- Spectrum limitations
- Limited energy
- User mobility

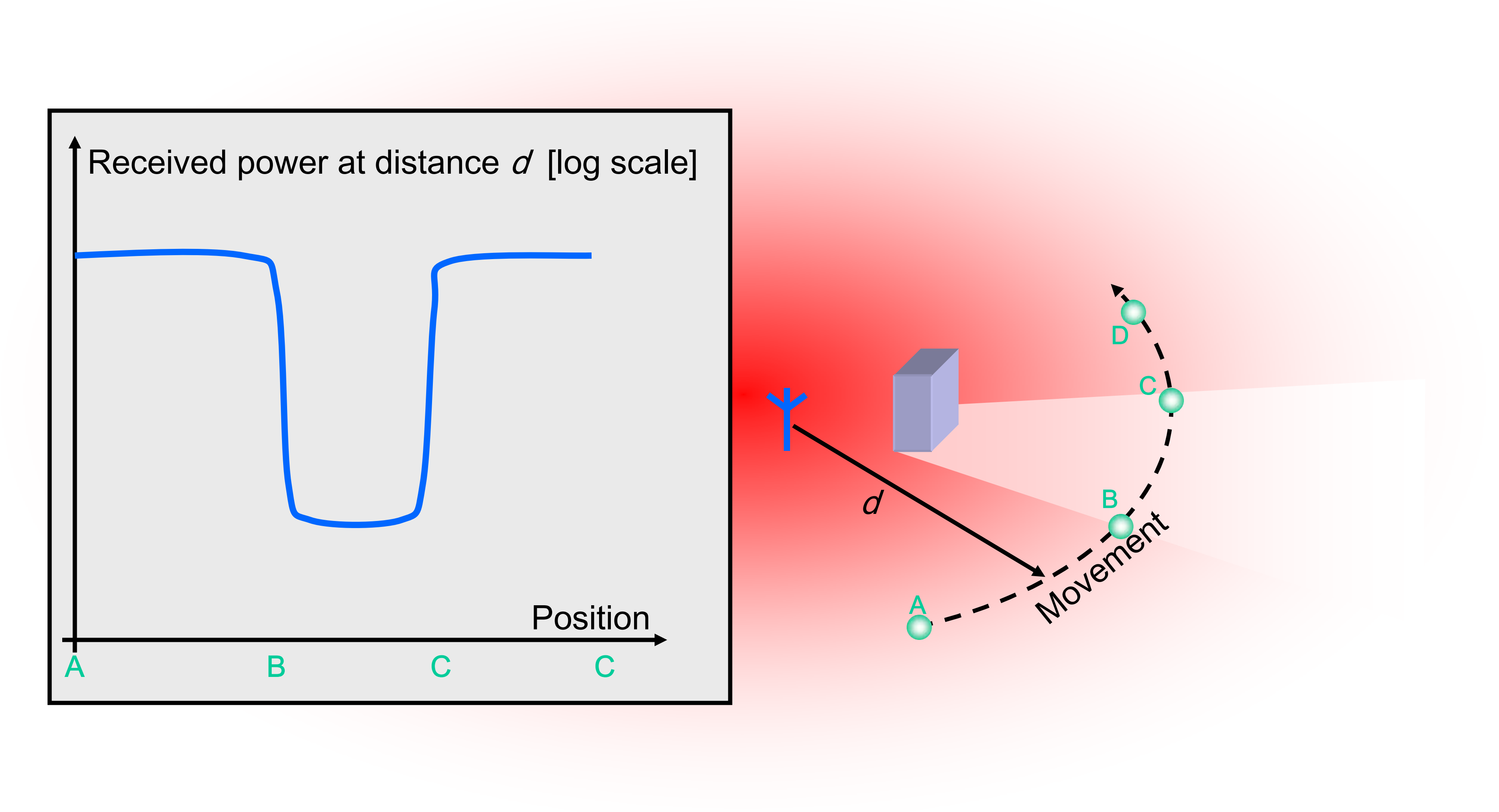
Multipath propagation



Small-scale fading



Large-scale fading

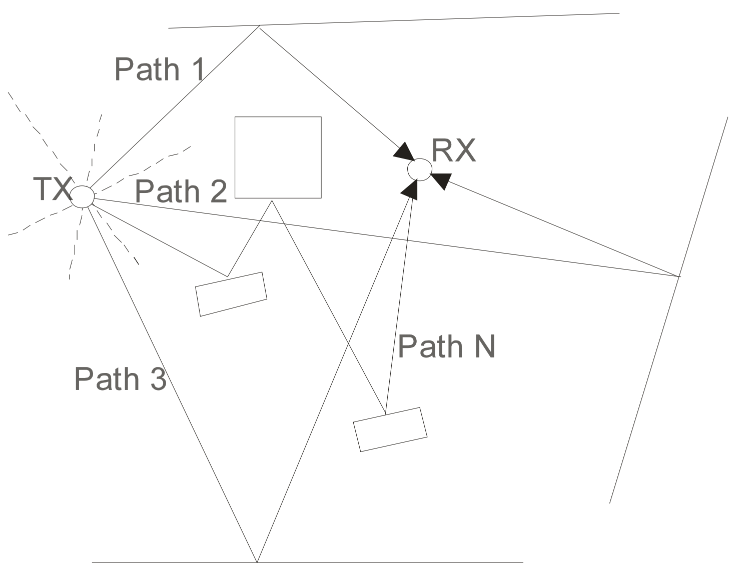


Consequences of fading

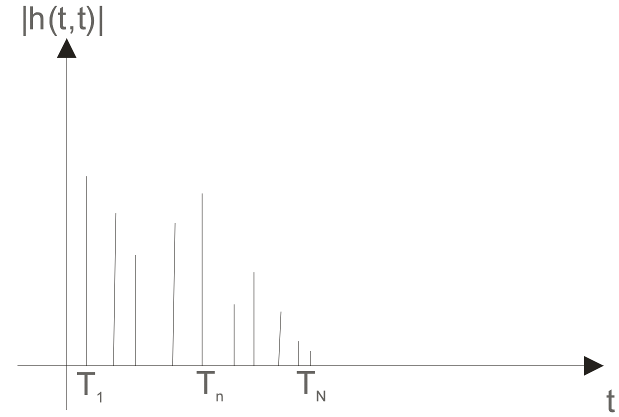
- Error probability is dominated by probability of being in a fading dip
- Error probability decreases only linearly with increasing SNR
- Fighting the effects of fading becomes essential for wireless transceiver design
- Deterministic modeling of channel at each point very difficult
- Statistical modeling of propagation and system behavior

Intersymbol interference (1)

- Channel impulse response is delay-dispersive

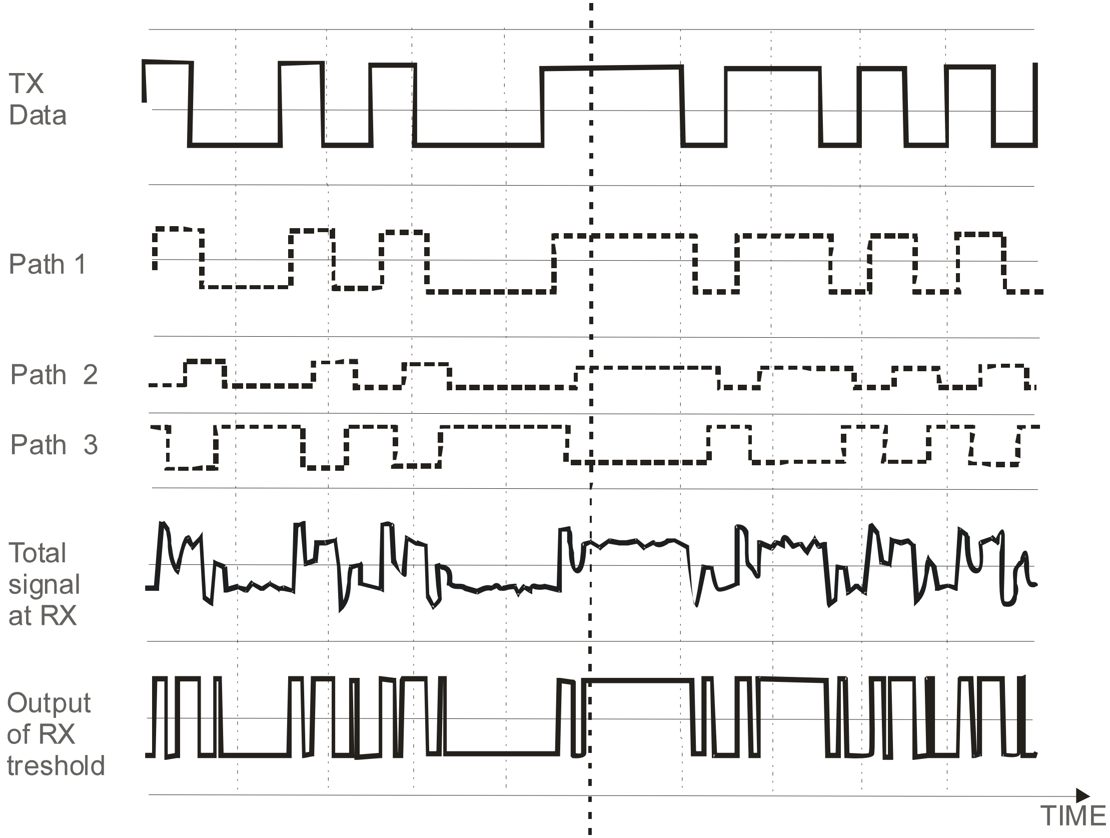


Multipath components with different runtimes



Channel impulse response

Intersymbol interference (2)



Spectrum assignment

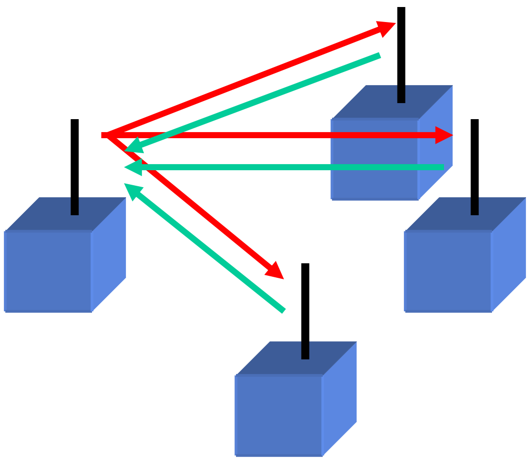
- <100 MHz: CB radio, pagers, and analogue cordless phones.
- 100-800 MHz: broadcast (radio and TV)
- 400-500 MHz: cellular and trunking radio systems
- 800-1000 MHz: cellular systems (analogue and second-generation digital); emergency communications
- 1.8-2.0 GHz: main frequency band for cellular and cordless
- 2.4-2.5 GHz: cordless phones, wireless LANs and wireless PANs (personal area networks); other devices, e.g., microwave ovens.
- 3.3-3.8 GHz: fixed wireless access systems
- 4.8-5.8 GHz: wireless LANs
- 11-15 GHz: satellite TV

Frequency reuse

- Available spectrum is limited
- -> the same frequency (range) has to be used at many different locations
- Regulated spectrum:
  - a single operator owns the spectrum, and can determine where to put TXs
  - cell planning so that interference adheres to certain limits
- Unregulated spectrum:
  - Often only one type of service allowed,
  - Nobody can control location of interferers
  - Power of interferers is limited by regulations

Duplexing and multiple access

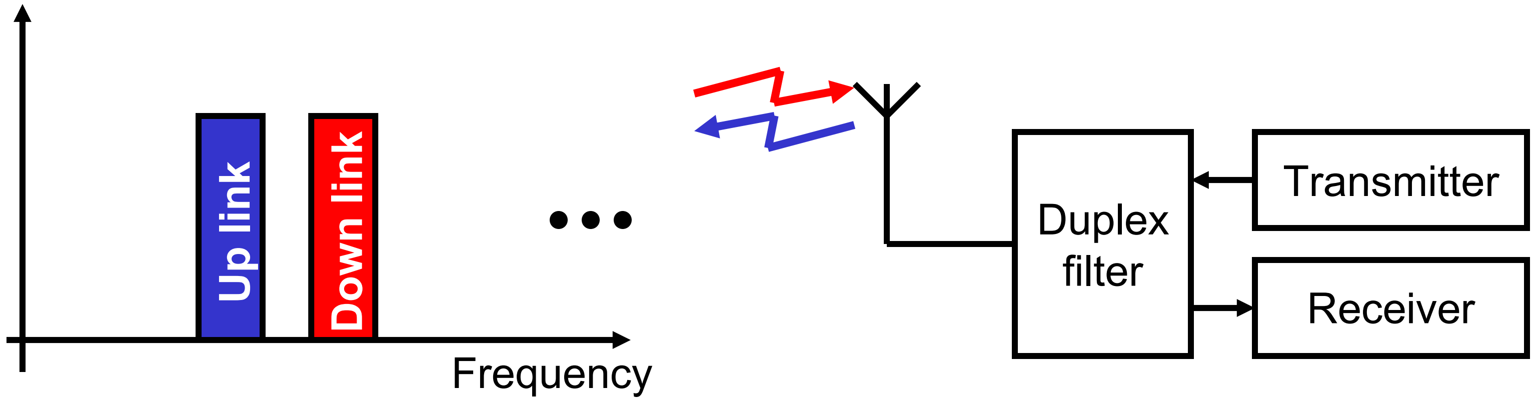
- Within each frequency band, multiple users need to communicate with one BS (multiple access)



Mobile telephony, wireless LAN, ...

- Cellphones have to be able to transmit and receive voice communications (duplexing)

DUPLEX
Frequency-division Duplex (FDD)

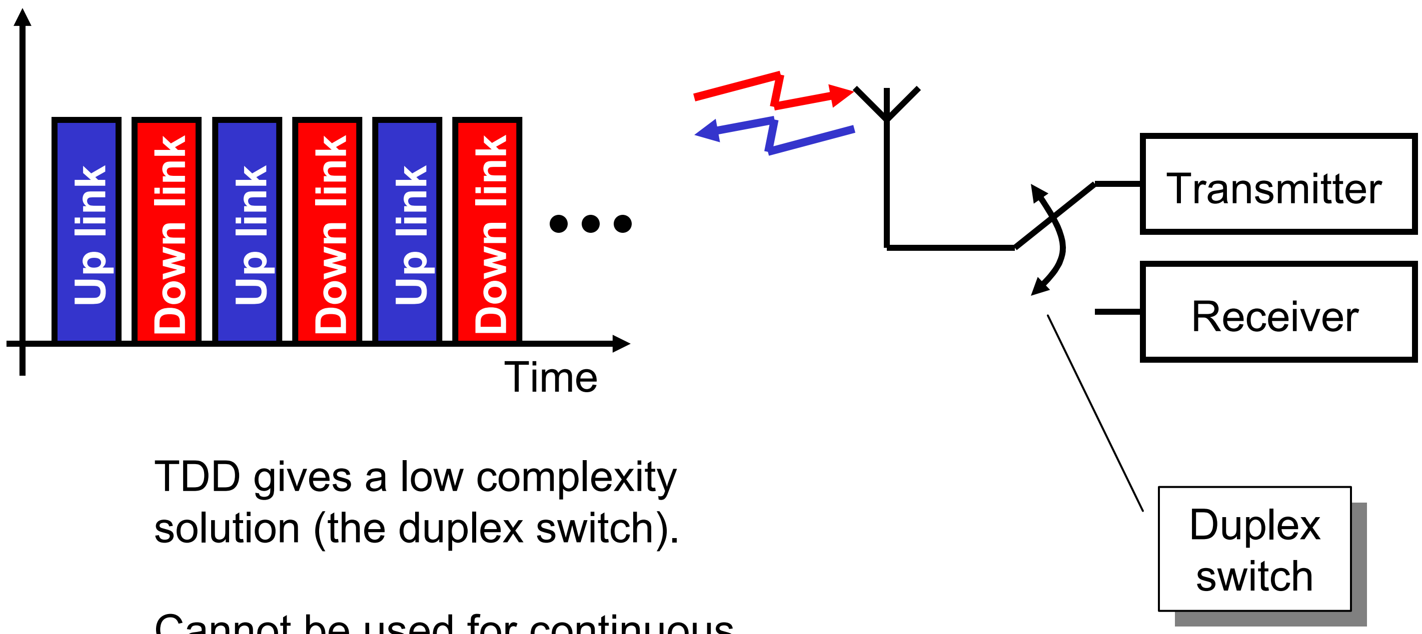


FDD gives a more complex solution (the duplex filter).

Can be used for continuous transmission.

Examples: Nodic Mobile Telephony (NMT), Global System for Mobile communications (GSM), Wideband CDMA (WCDMA)

DUPLEX
Time-division duplex (TDD)

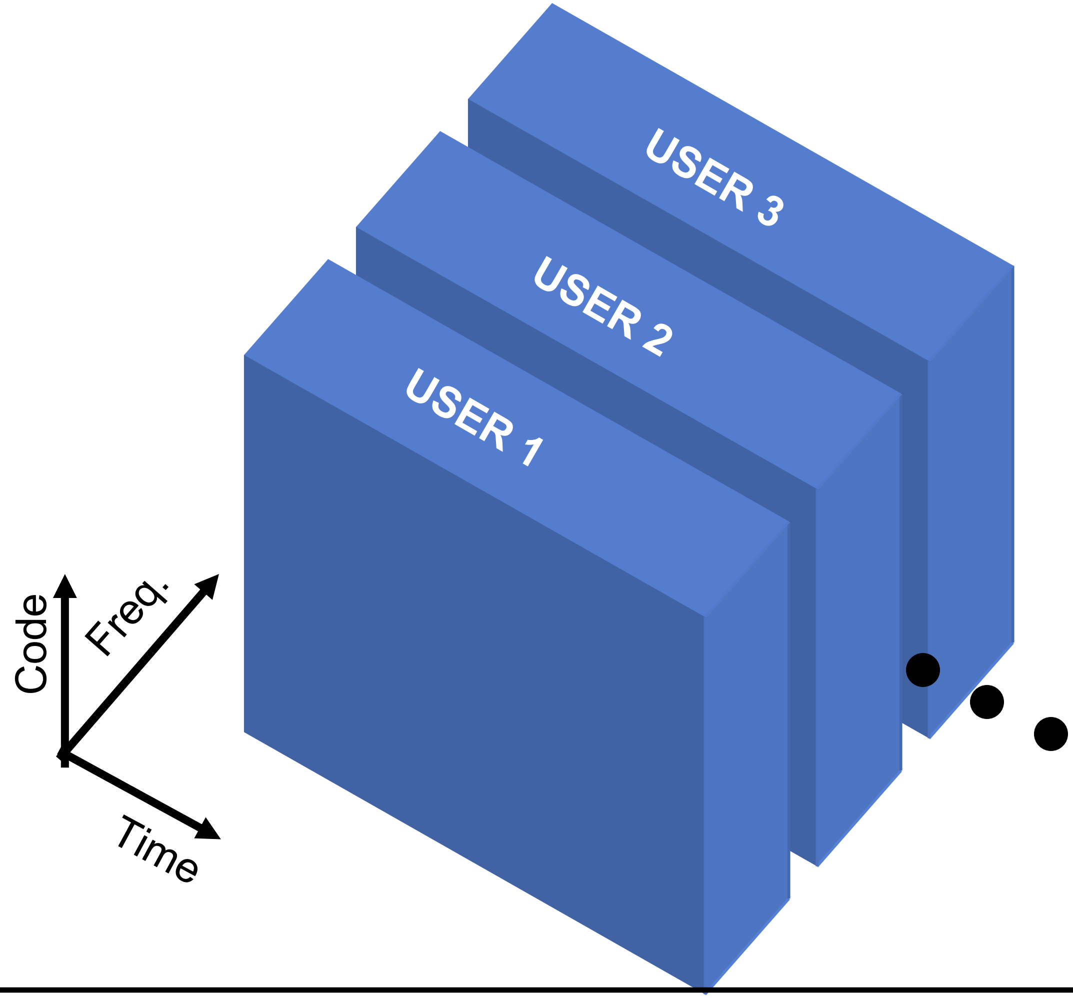


TDD gives a low complexity solution (the duplex switch).

Cannot be used for continuous transmission.

Examples: Global System for Mobile communications (GSM), Wideband CDMA (WCDMA)

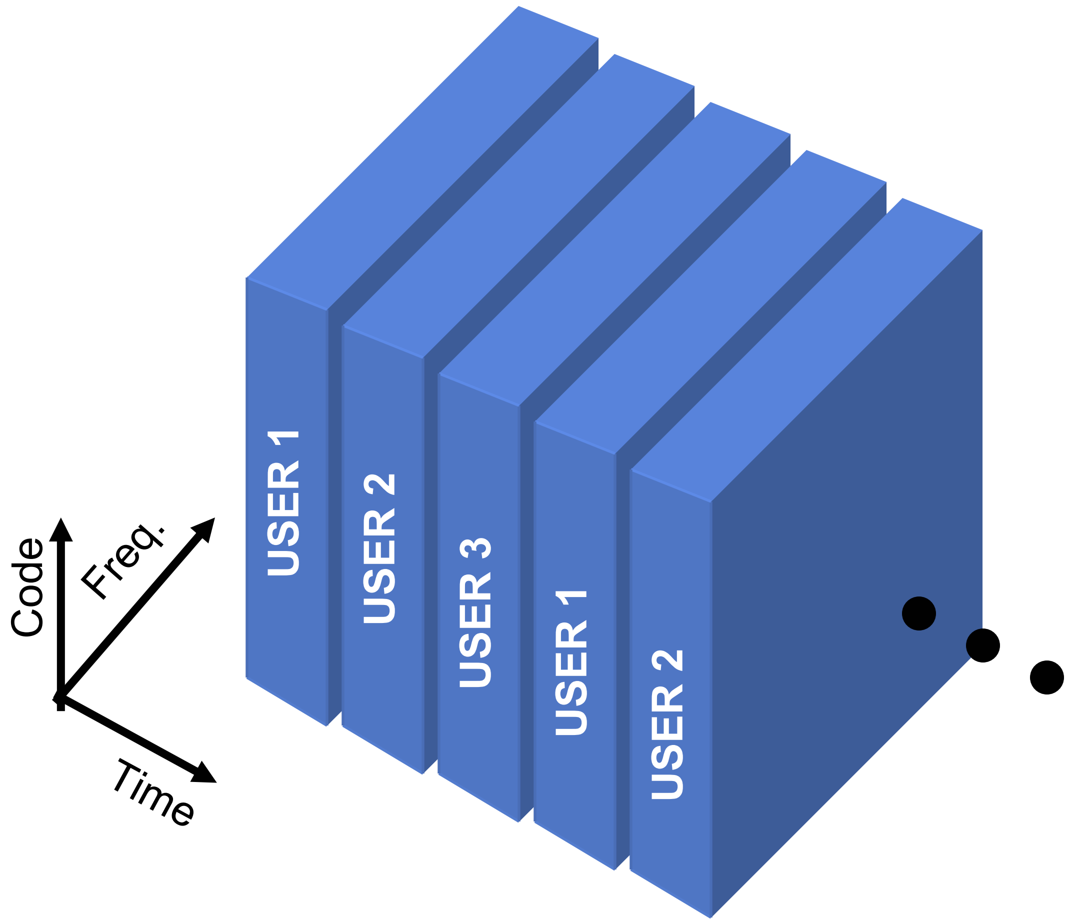
MULTIPLE ACCESS
Freq.-division multiple access (FDMA)



Users are separated in frequency bands.

Examples: Nordic Mobile Telephony (NMT), Advanced Mobile Phone System (AMPS)

MULTIPLE ACCESS
Time-division multiple access (TDMA)

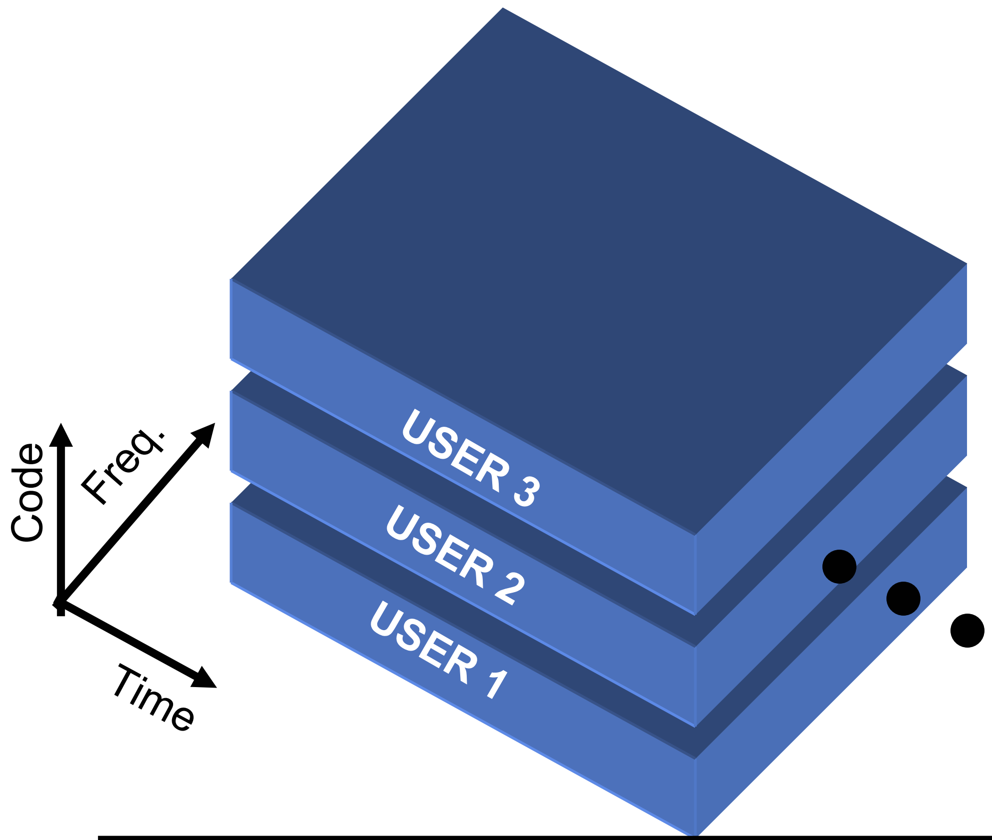


Users are separated in time slots.

Example: Global System for Mobile communications (GSM)

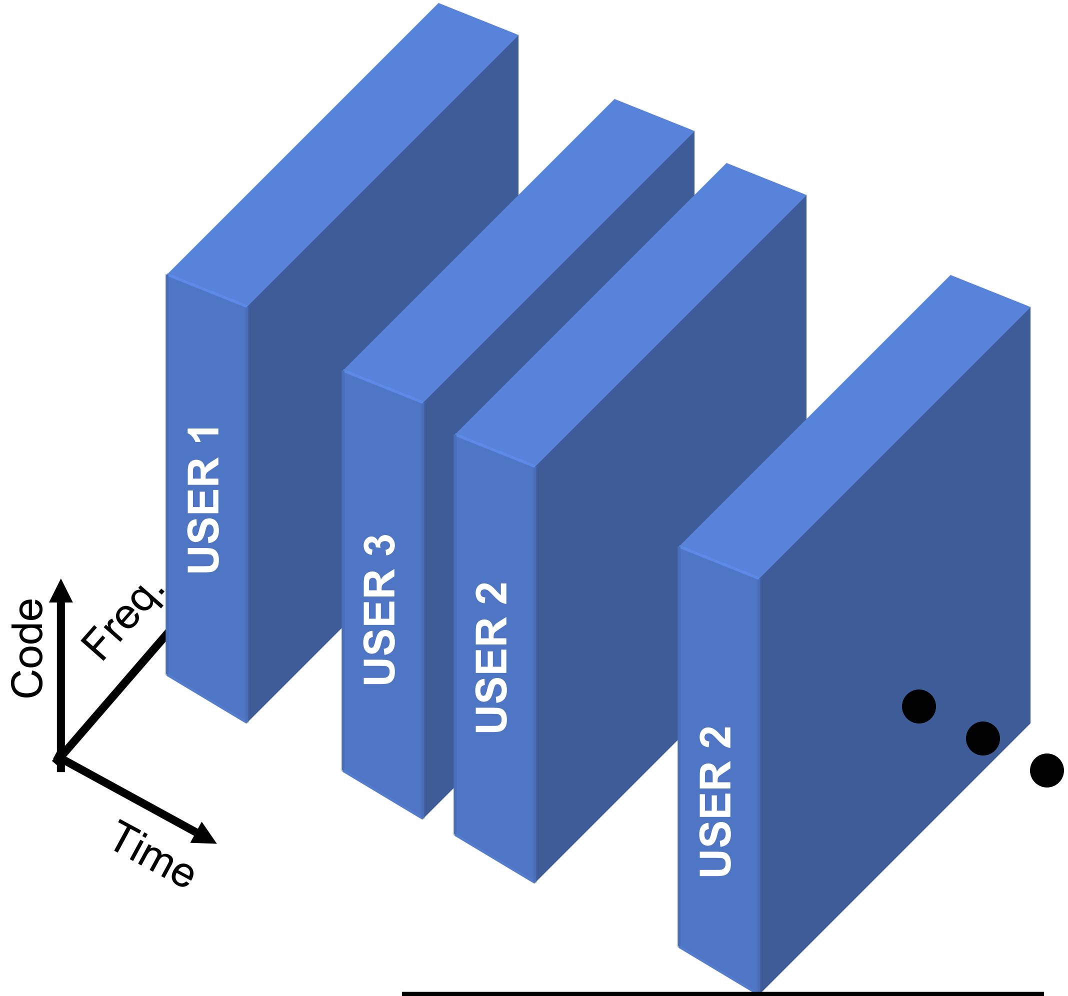
MULTIPLE ACCESS
Code-division multiple access (CDMA)

Users are separated by spreading codes.



Examples: CdmaOne, Wideband CDMA (WCDMA), Cdma2000

MULTIPLE ACCESS
Carrier-sense multiple access (CSMA)



Users are separated in time but not in an organized way. The terminal listens to the channel, and transmits a packet if it's free.

Collissions can occur and data is lost.

Example: IEEE 802.11 (WLAN)

User mobility

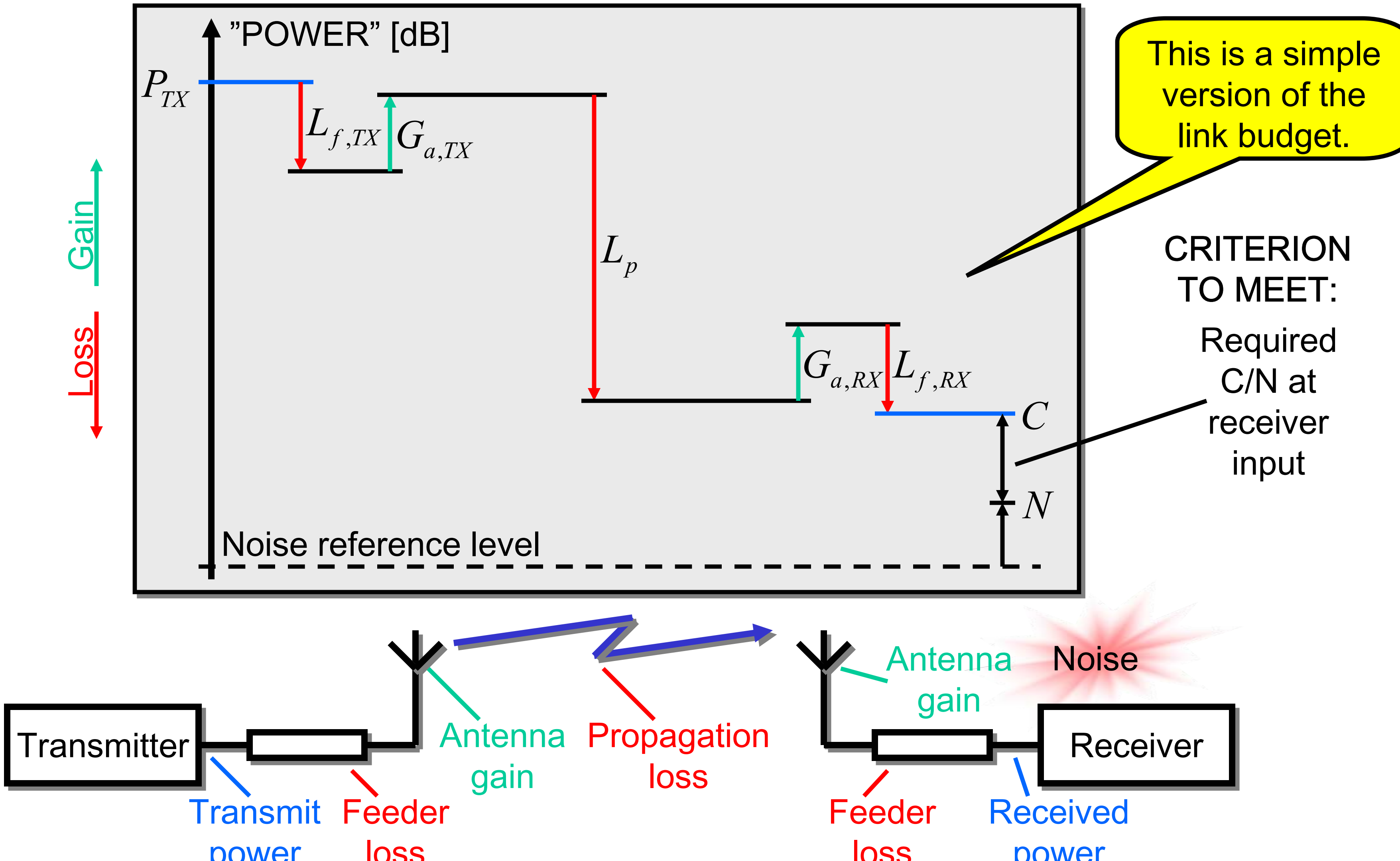
- User can change position
- Mobility within one cell (i.e., maintaining a link to a certain BS): mostly effect on propagation channel (fading)
- Mobility from cell to cell:

Noise- and interference limited systems

Basics of link budgets

- Link budgets show how different components and propagation processes influence the available SNR
- Link budgets can be used to compute, e.g., required transmit power, possible range of a system, or required receiver sensitivity
- Link budgets can be most easily set up using logarithmic power units (dB)

SINGLE LINK
The link budget – a central concept



dB in general

When we convert a measure X into decibel scale, we always divide by a reference value X_{ref}:

$$\frac{X|_{non-dB}}{X_{ref}|_{non-dB}}$$

Independent of the dimension of X (and X_{ref}), this value is always dimension-less.

The corresponding dB value is calculated as:

$$X|_{dB} = 10\log\left(\frac{X|_{non-dB}}{X_{ref}|_{non-dB}}\right)$$

Power

We usually measure power in Watt (W) and milliWatt [mW]

The corresponding dB notations are dB and dBm

	Non-dB	dB
Watt:	$P\|_W$	$P\|_{dB} = 10\log\left(\frac{P\|_W}{1\|_W}\right) = 10\log\left(P\|_W\right)$
milliWatt:	$P\|_{mW}$	$P\|_{dBm} = 10\log\left(\frac{P\|_{mW}}{1\|_{mW}}\right) = 10\log\left(P\|_{mW}\right)$
RELATION:	$P\|_{dBm} = 10\log\left(\frac{P\|_W}{0.001\|_W}\right) = 10\log\left(P\|_W\right) + 30\|_{dB} = P\|_{dB} + 30\|_{dB}$	

Example: Power

Sensitivity level of GSM RX: 6.3×10^{-14} W = -132 dB or -102 dBm

Bluetooth TX: 10 mW = -20 dB or 10 dBm

GSM mobile TX: 1 W = 0 dB or 30 dBm

GSM base station TX: 40 W = 16 dB or 46 dBm

ERP – Effective Radiated Power

Vacuum cleaner: 1600 W = 32 dB or 62 dBm

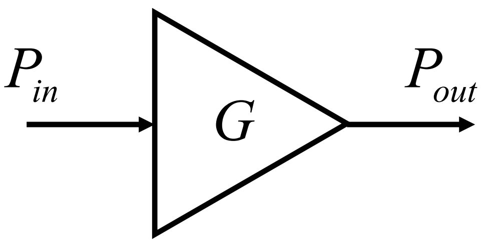
Car engine: 100 kW = 50 dB or 80 dBm

TV transmitter (Hörby, SVT2): 1000 kW ERP = 60 dB or 90 dBm ERP

Nuclear powerplant (Barsebäck): 1200 MW = 91 dB or 121 dBm

Amplification and attenuation

(Power) Amplification:



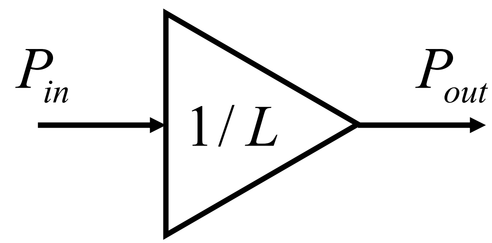
$$P_{out} = GP_{in} \Rightarrow G = \frac{P_{out}}{P_{in}}$$

The amplification is already dimension-less and can be converted directly to dB:

$$G|_{dB} = 10\log_{10} G$$

Note: It doesn't matter if the power is in mW or W. Same result!

(Power) Attenuation:

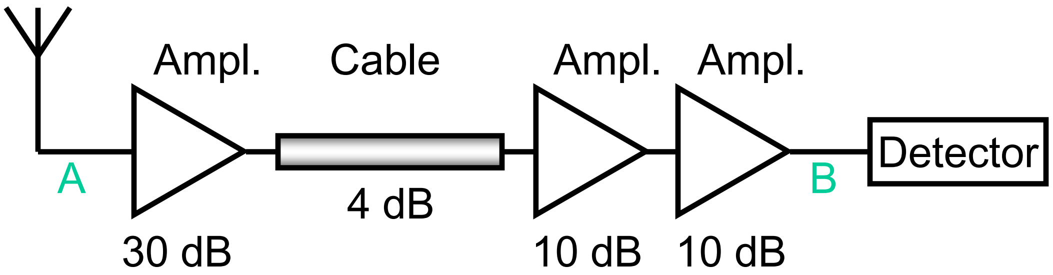


$$P_{out} = \frac{P_{in}}{L} \Rightarrow L = \frac{P_{in}}{P_{out}}$$

The attenuation is already dimension-less and can be converted directly to dB:

$$L|_{dB} = 10\log_{10} L$$

Example: Amplification and attenuation

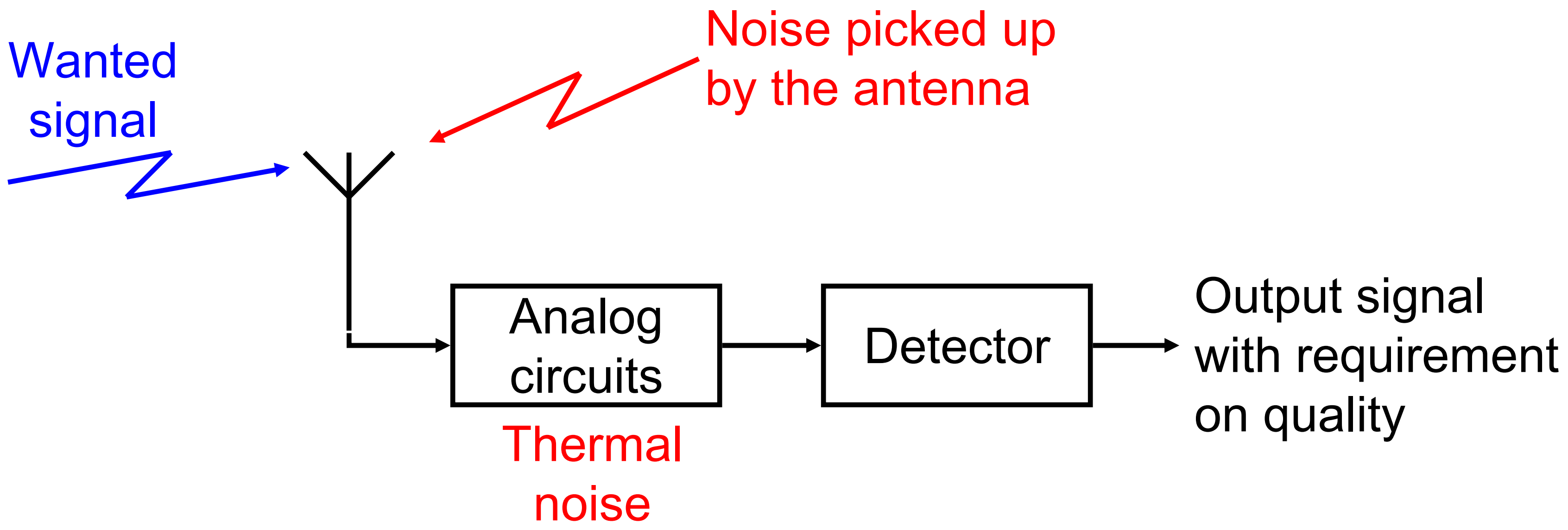


The total amplification of the (simplified) receiver chain (between A and B) is

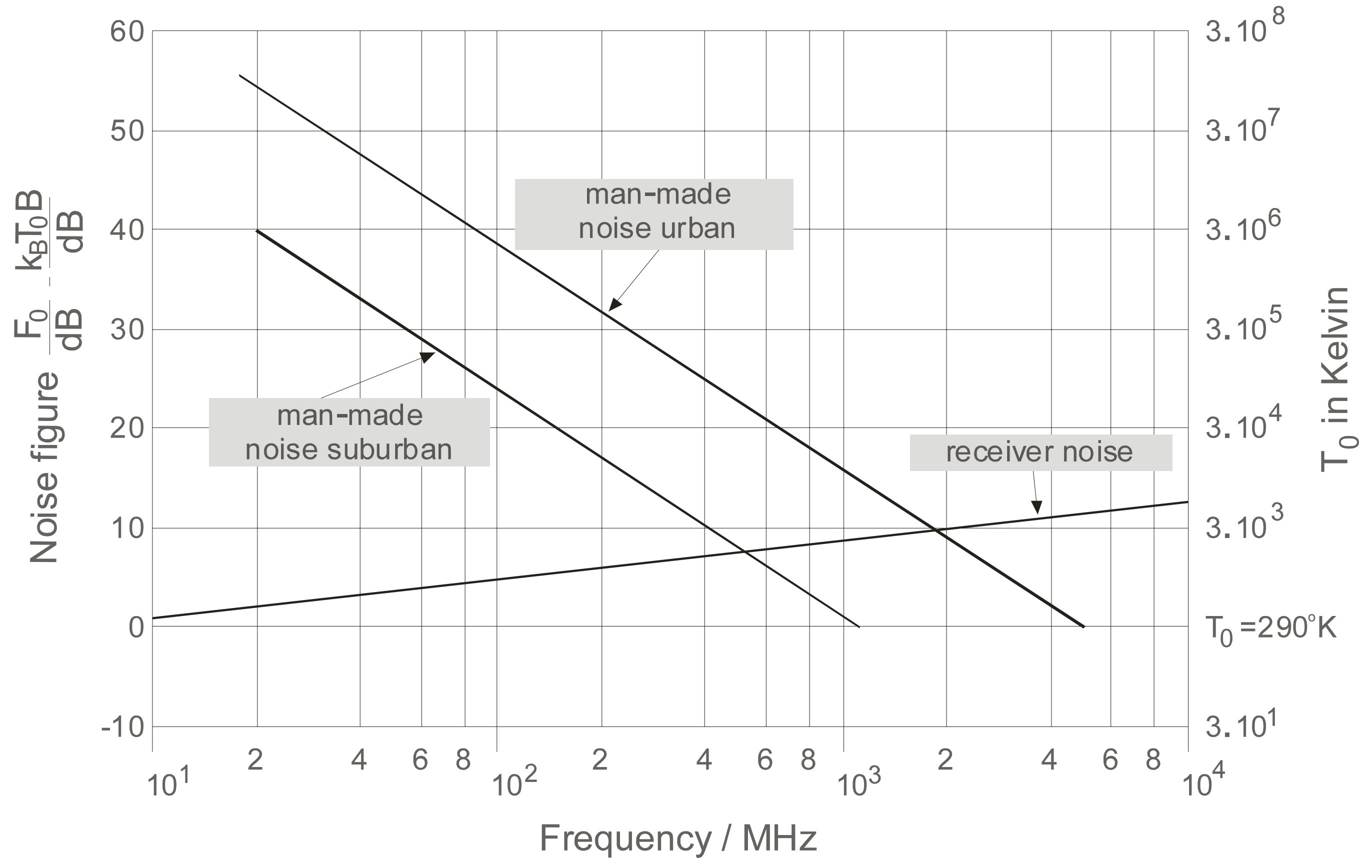
$$G_{A,B}\,|_{dB} = 30 - 4 + 10 + 10 = 46$$

Noise sources

The noise situation in a receiver depends on several noise sources

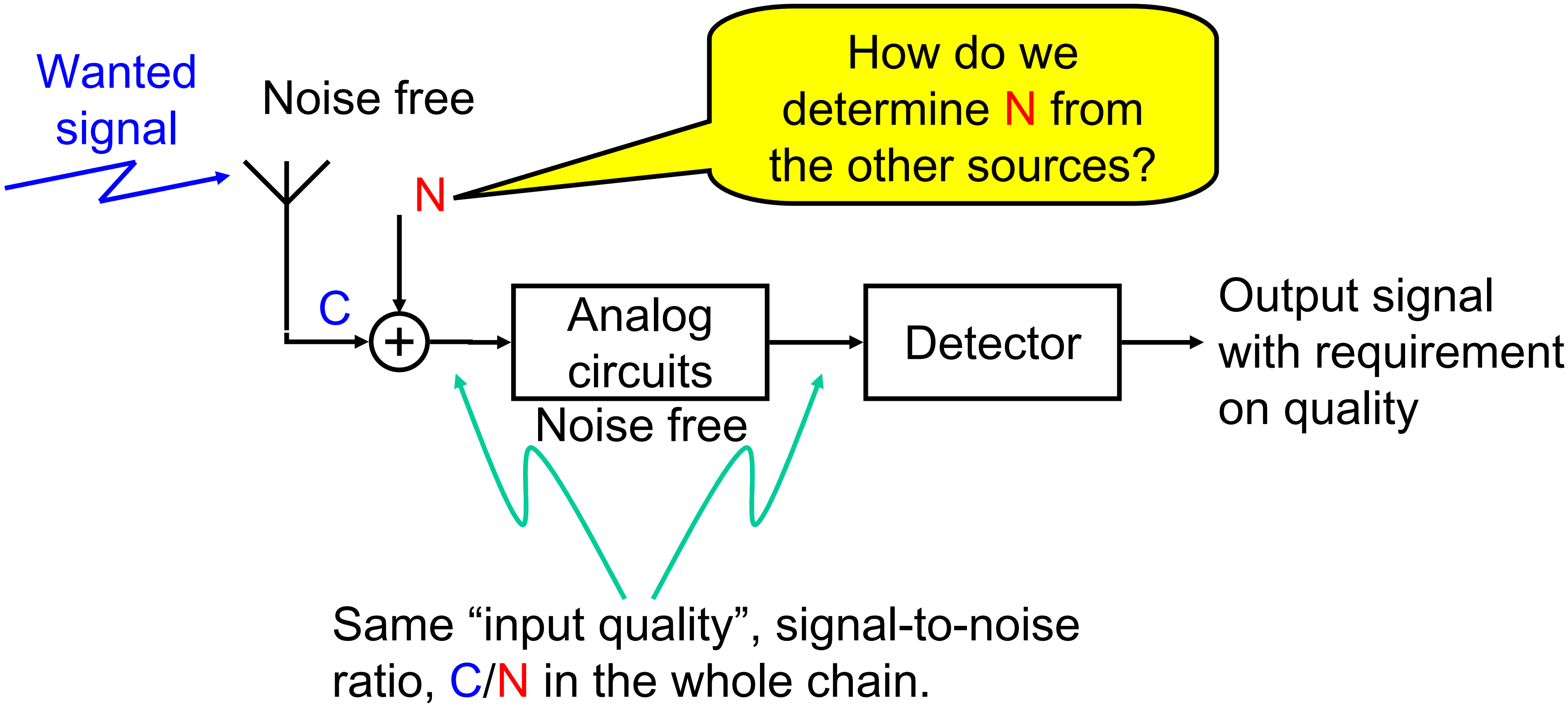


Man-made noise



Receiver noise: Equivalent noise source

To simplify the situation, we replace all noise sources with a single equivalent noise source.



Receiver noise: Noise sources (1)

The power spectral density of a noise source is usually given in one of the following three ways:

1) Directly [W/Hz]: N_s

2) Noise temperature [Kelvin]: T_s

3) Noise factor [1]: F_s

This one is sometimes given i dB and called noise figure.

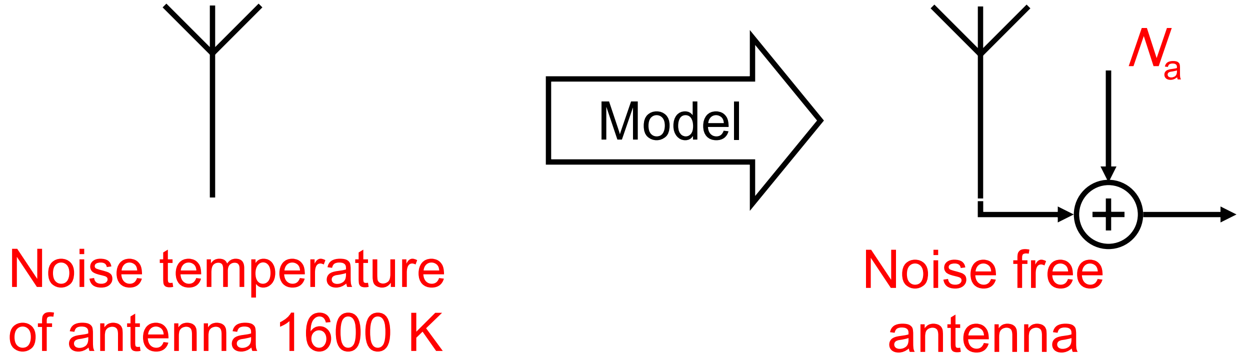
The relation between the tree is

$$N_s = kT_s = kF_sT_0$$

where k is Boltzmann's constant (1.38×10^{-23} W/Hz) and T_0 is the, so called, room temperature of 290 K (17^o C).

Receiver noise: Noise sources (2)

Antenna example



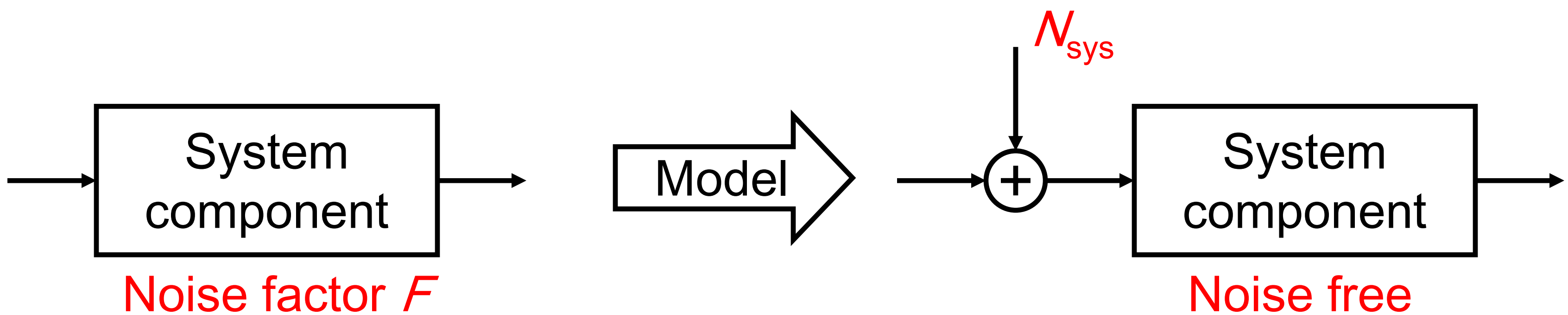
Power spectral density of antenna noise is

$$N_a = 1.38 \times 10^{-23} \times 1600 = 2.21 \times 10^{-20} \text{ W/Hz} = -196.6 \text{ dB[W/Hz]}$$

and its noise factor/noise figure is

$$F_a = 1600 / 290 = 5.52 = 7.42 \text{ dB}$$

Receiver noise: System noise



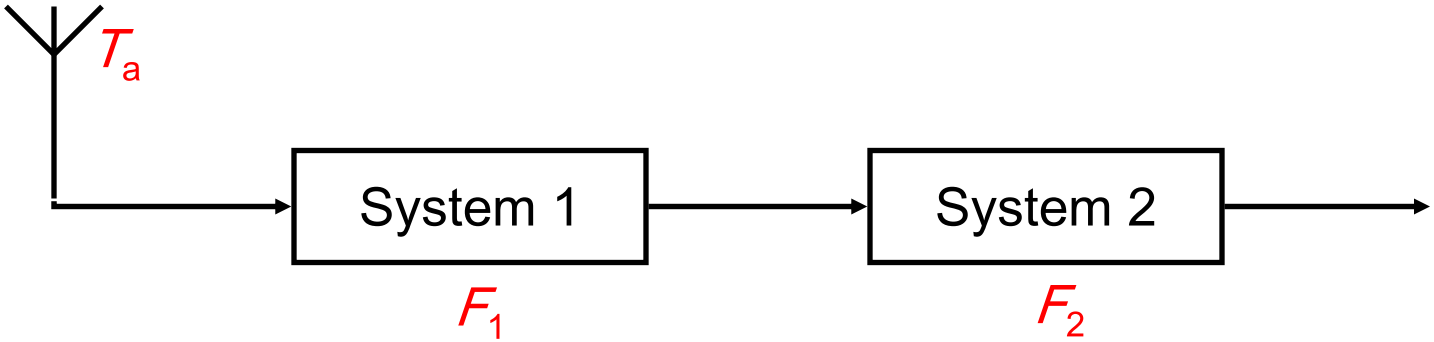
Due to a definition of noise factor (in this case) as the ratio of noise powers on the output versus on the input, when a resistor in room temperature (T_0=290 K) generates the input noise, the PSD of the equivalent noise source (placed at the input) becomes

$$N_{sys} = k \underbrace{\left(F-1\right) T_0}_{\text{Equivalent noise temperature}} \ \mathrm{W/Hz}$$

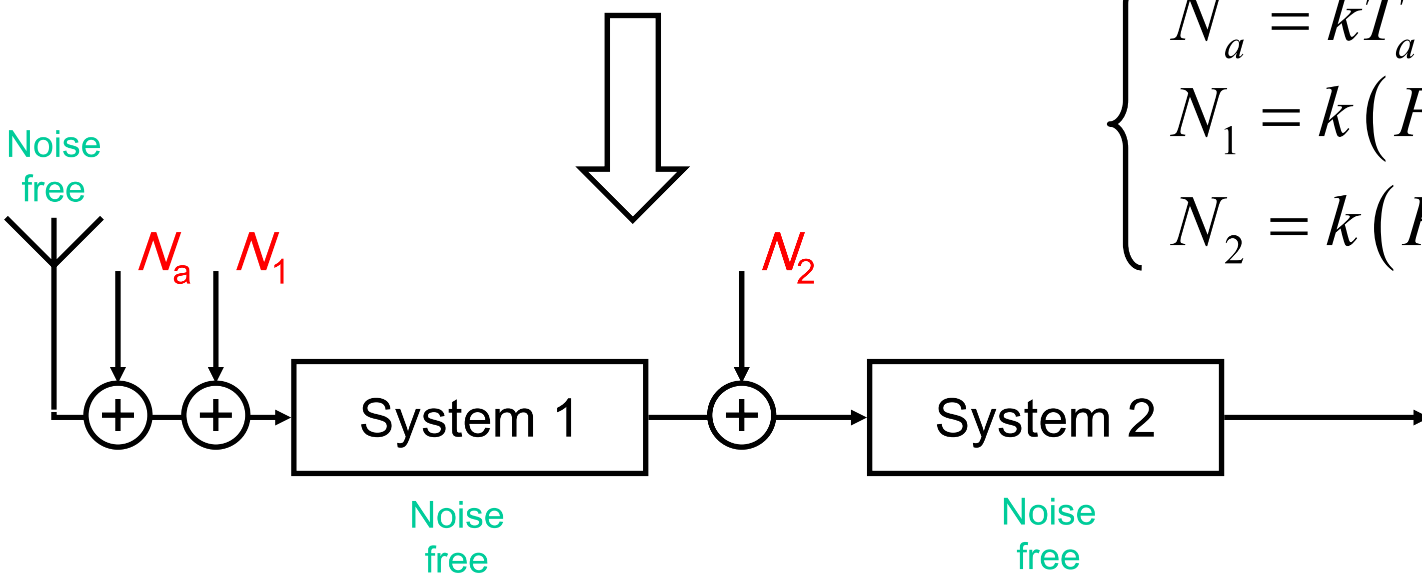
Don't use dB value!

Receiver noise: Sev. noise sources (1)

A simple example



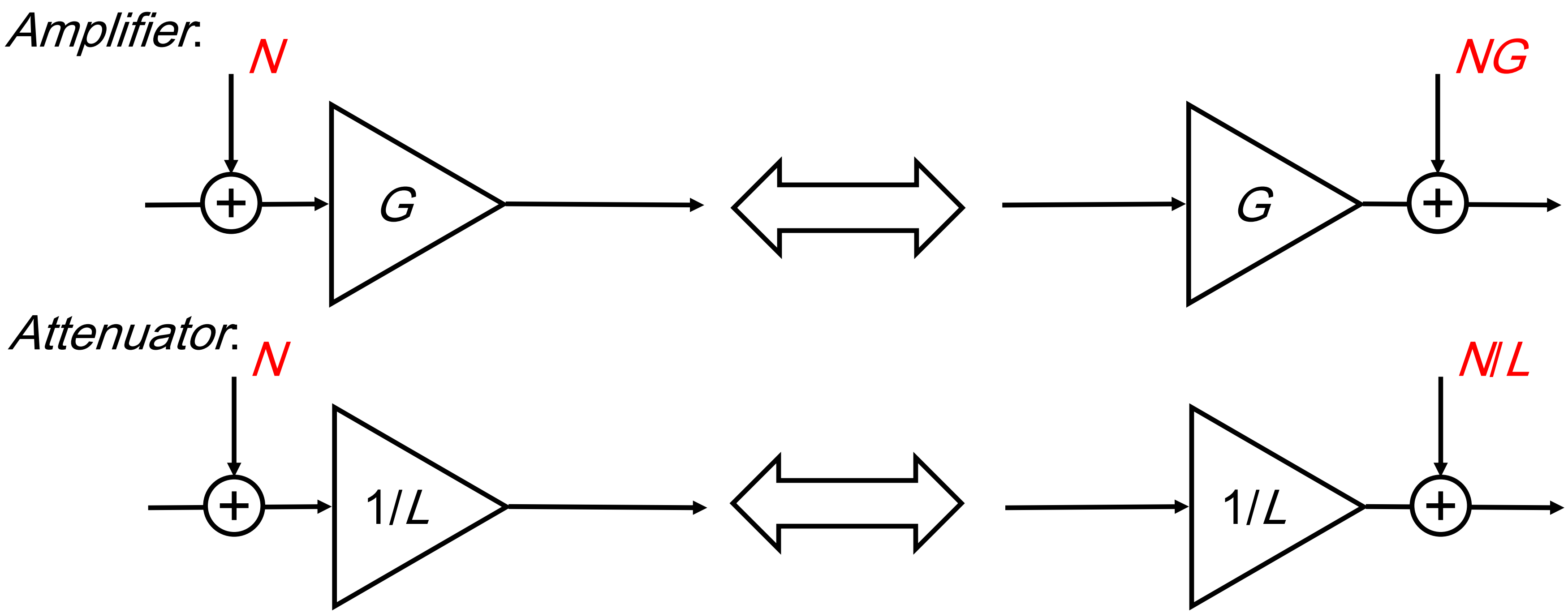
$$
\begin{cases}
N_a = kT_a \\
N_1 = k\left(F_1 - 1\right)T_0 \\
N_2 = k\left(F_2 - 1\right)T_0
\end{cases}
$$



Receiver noise: Sev. noise sources (2)

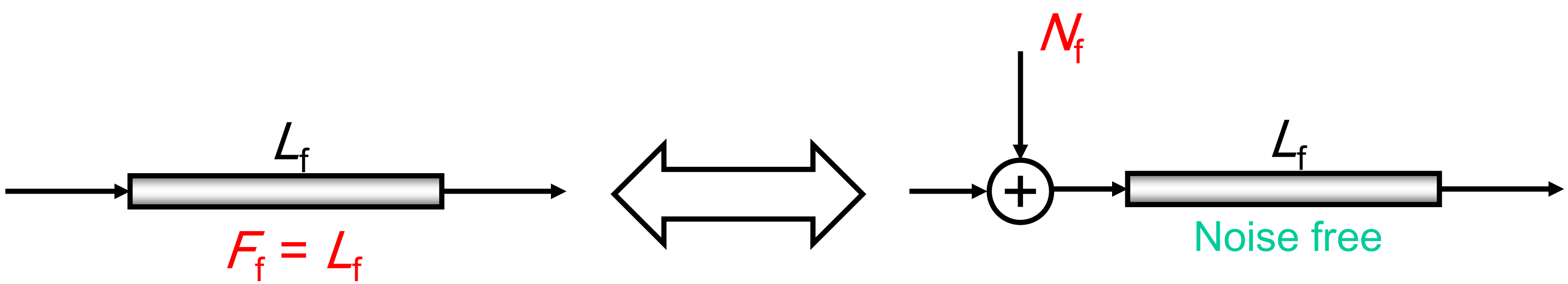
After extraction of the noise sources from each component, we need to move them to one point.

When doing this, we must compensate for amplification and attenuation!



Pierce's rule

A passive attenuator, in this case a feeder, has a noise figure equal to its attenuation.



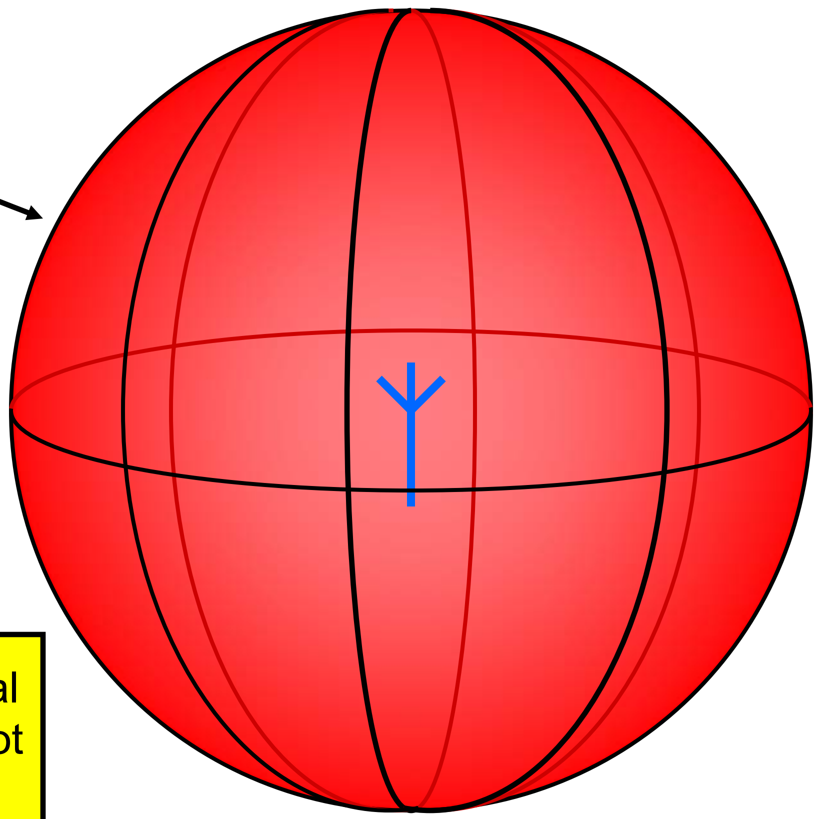
$$N_f = k\left(F_f - 1\right)T_0 = k\left(L_f - 1\right)T_0$$

Remember to convert *from* dB!

The isotropic antenna

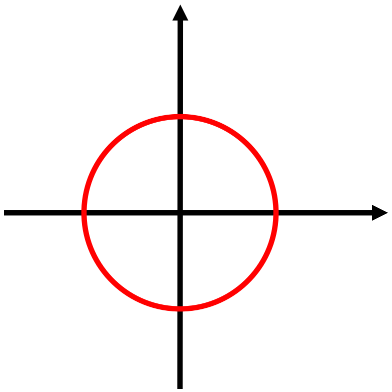
The isotropic antenna radiates equally in all directions

Radiation pattern is spherical

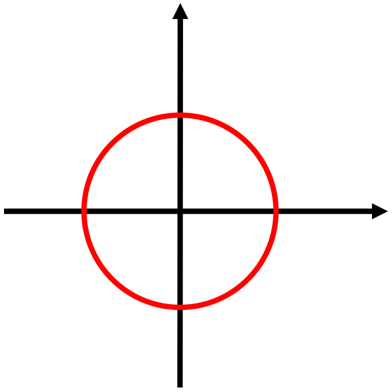


This is a theoretical antenna that cannot be built.

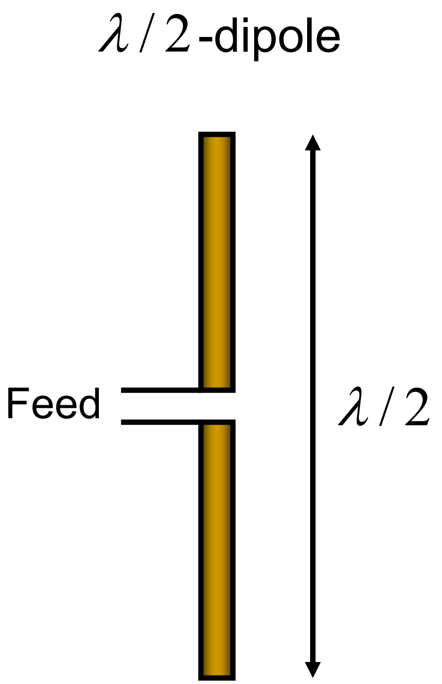
Elevation pattern



Azimuth pattern



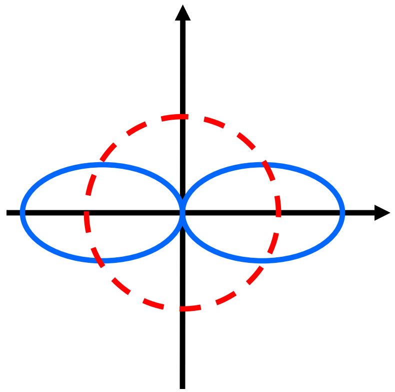
The dipole antenna



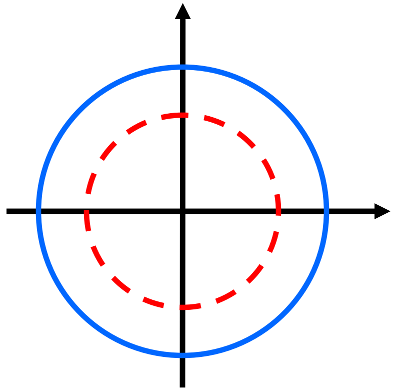
This antenna does not radiate straight up or down. Therefore, more energy is available in other directions.

THIS IS THE PRINCIPLE BEHIND WHAT IS CALLED *ANTENNA GAIN*.

Elevation pattern



Azimuth pattern



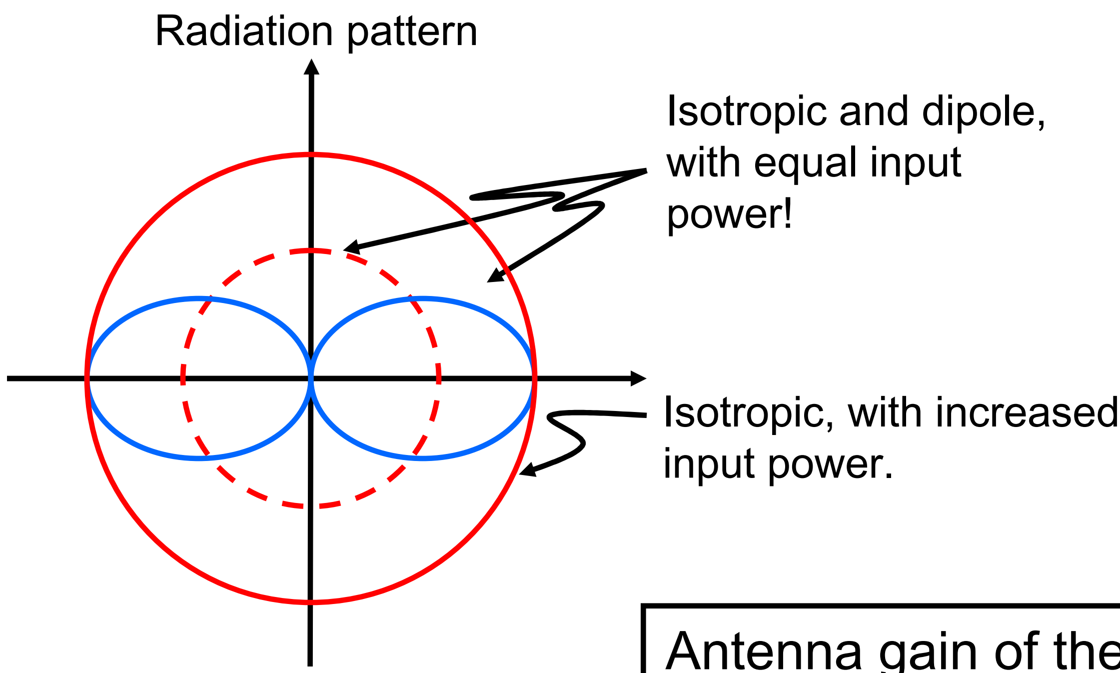
Antenna pattern of isotropic antenna.

A dipole can be of any length, but the antenna patterns shown are only for the $\lambda/2$-dipole.

Antenna gain (principle)

Antenna gain is a relative measure.

We will use the isotropic antenna as the reference.



The amount of increase in input power to the isotropic antenna, to obtain the same maximum radiation is called the antenna gain!

Antenna gain of the $\lambda/2$ dipole is 2.15 dB.

A note on antenna gain

Sometimes the notation dBi is used for antenna gain (instead of dB).

The ”i” indicates that it is the gain relative to the isotropic antenna (which we will use in this course).

Another measure of antenna gain frequently encountered is dBd, which is relative to the λ/2 dipole.

$$G|_{dBi} = G|_{dBd} + 2.15$$

Be careful! Sometimes it is not clear if the antenna gain is given in dBi or dBd.

EIRP: Effective Isotropic Radiated Power

EIRP = Transmit power (fed to the antenna) + antenna gain

$$EIRP|_{dB} = P_{TX|dB} + G_{TX|dB}$$

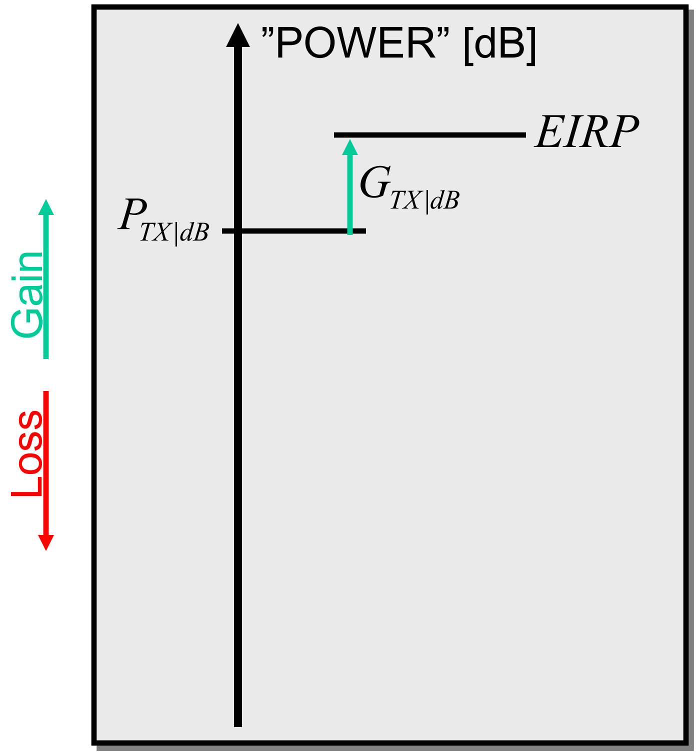
Answers the questions:

How much transmit power would we need to feed an isotropic antenna to obtain the same maximum on the radiated power?

How "strong" is our radiation in the maximal direction of the antenna?

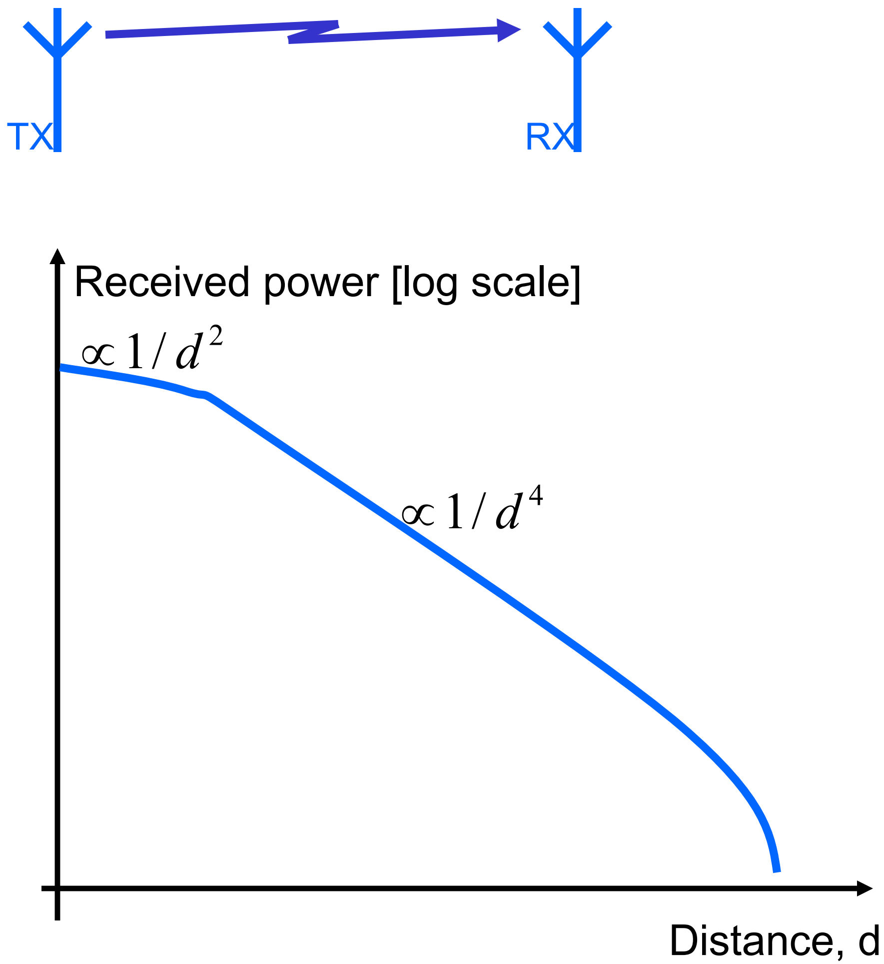
This is the more important one, since a limit on EIRP is a limit on the radiation in the maximal direction.

EIRP and the link budget

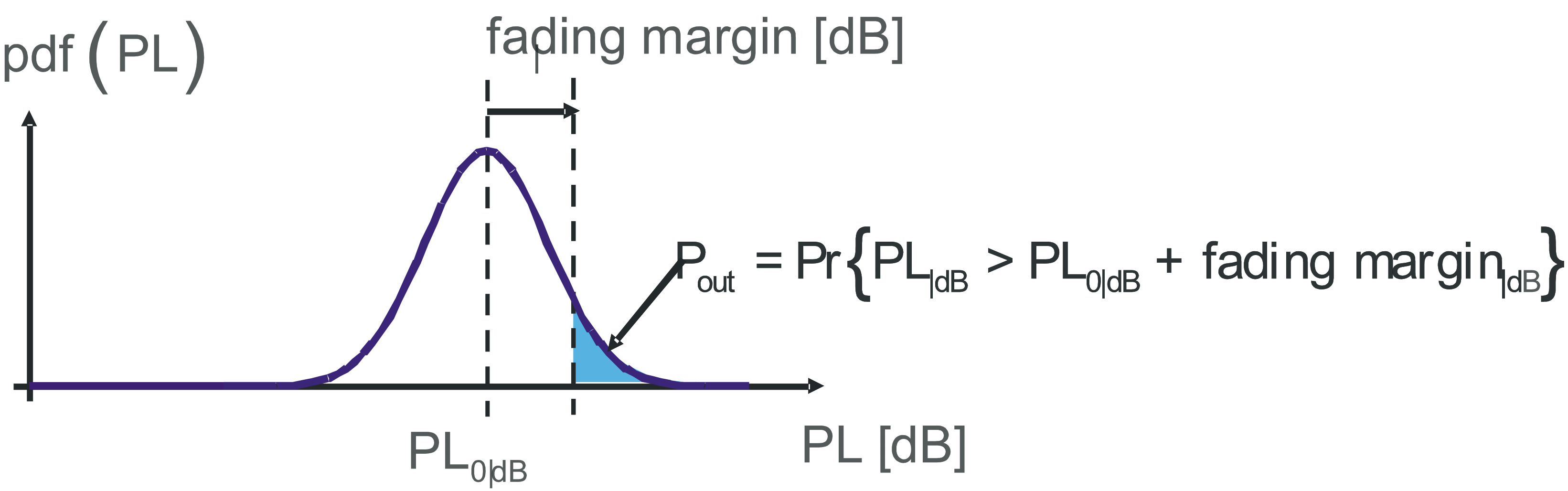


$$EIRP\,|_{dB} = P_{TX|dB} + G_{TX|dB}$$

Path loss

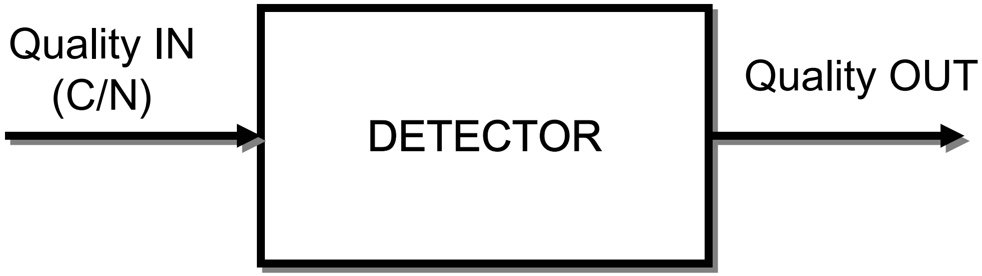

TX
RX
Received power [log scale]
$\propto 1/d^2$
$\propto 1/d^4$
Distance, d

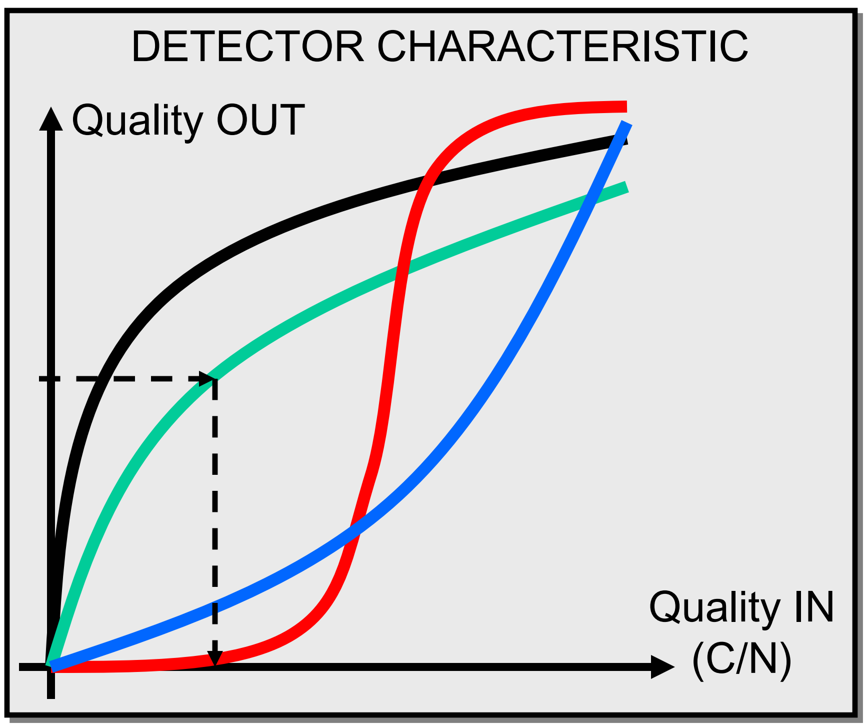
Fading margin



$$P_{out} = \Pr\left\{PL_{|dB} > PL_{0|dB} + \text{fading margin}_{|dB}\right\}$$

Required C/N – another central concept



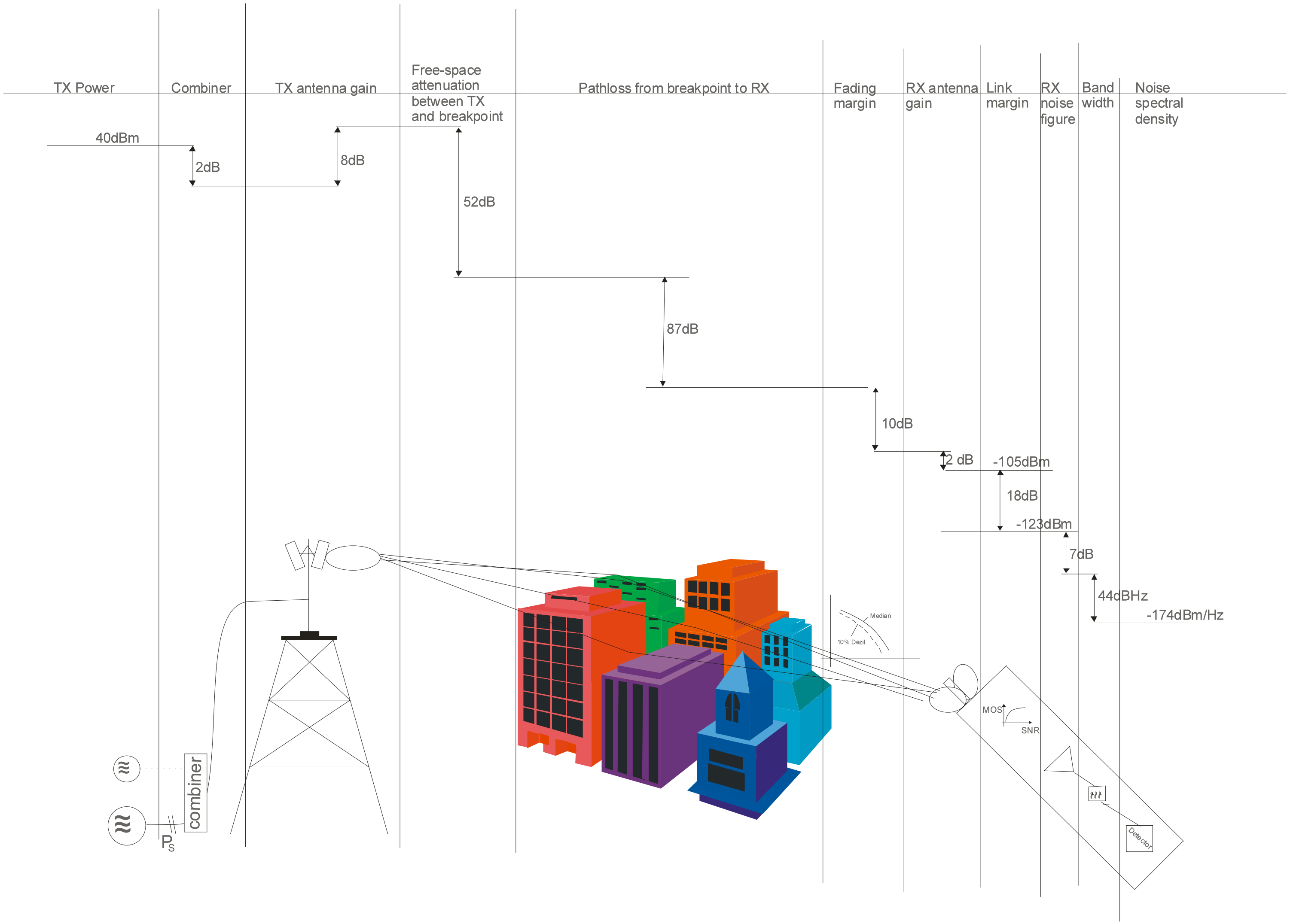


The detector characteristic is different for different system design choices.

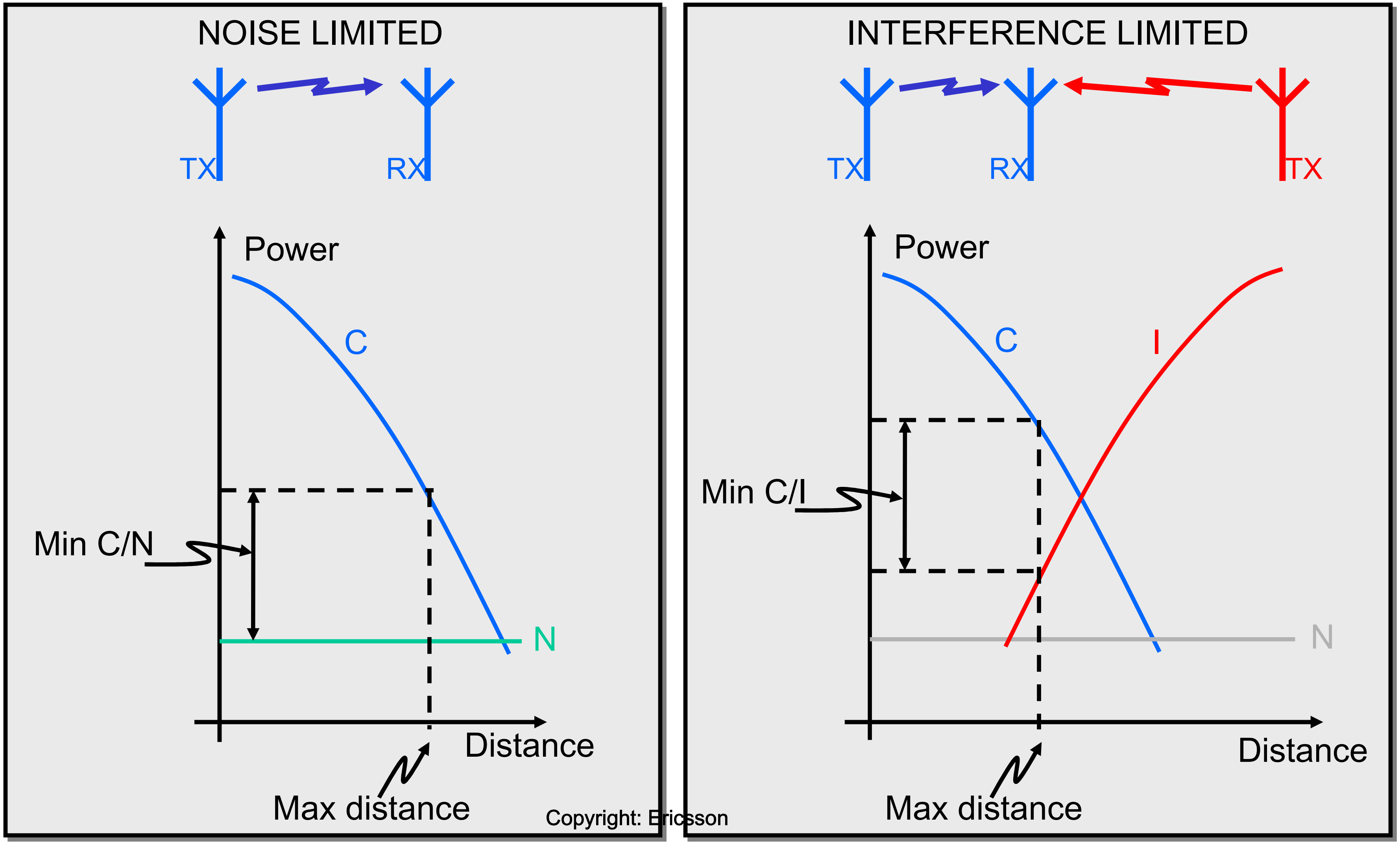
REQUIRED QUALITY OUT:

Audio SNR
Perceptive audio quality
Bit-error rate
Packet-error rate
etc.

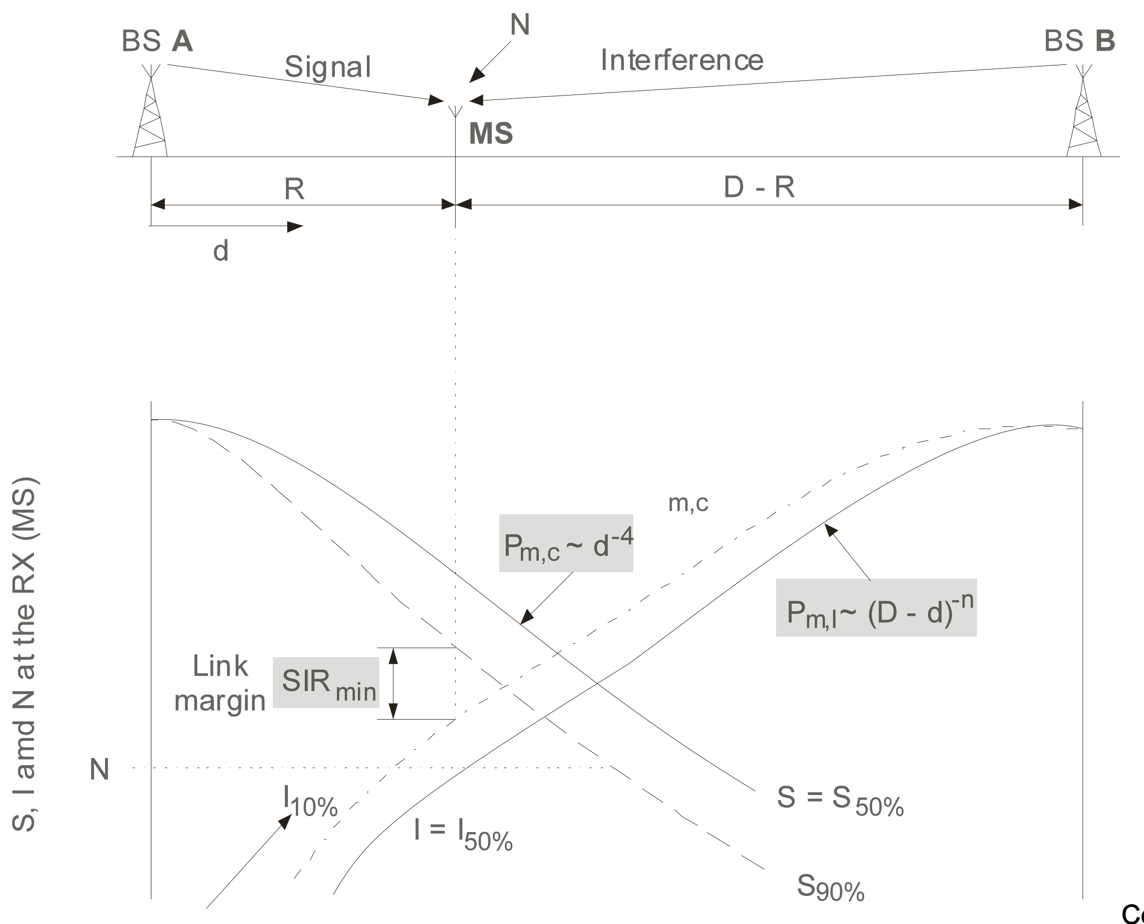
Example for link budget



Noise and interference limited links



What is required distance between BSs?



Chapter 4

Propagation effects

Why channel modelling?

- The performance of a radio system is ultimately determined by the radio channel
- The channel models basis for
  - system design
  - algorithm design
  - antenna design etc.
- Trend towards more interaction system-channel
  - MIMO
  - UWB

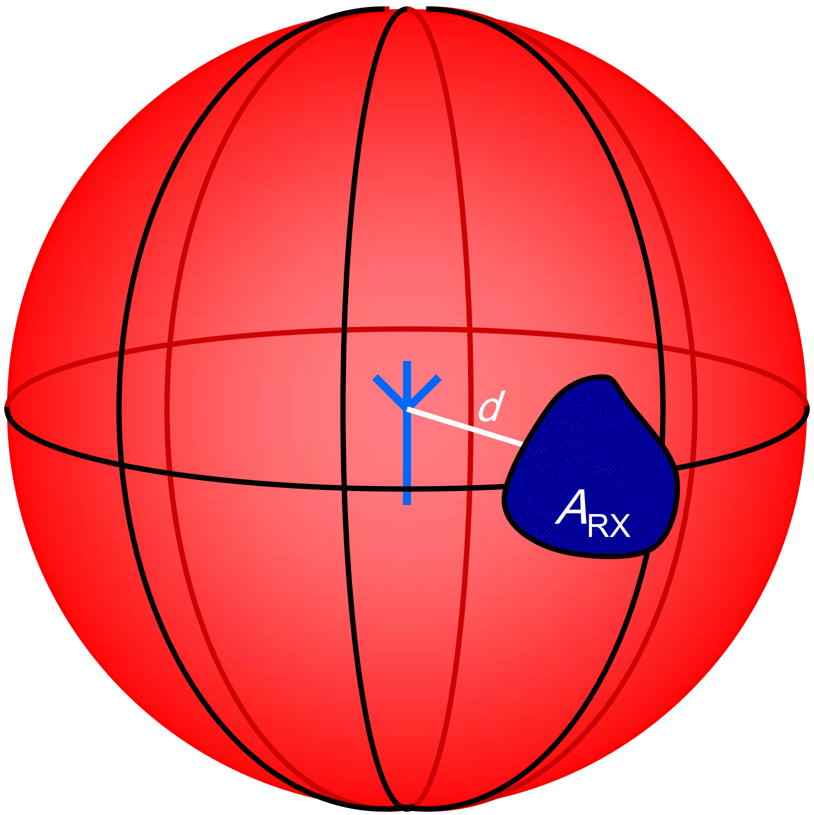
Without reliable channel models, it is hard to design radio systems that work well in *real* environments.

THE RADIO CHANNEL
It is more than just a loss

- Some examples:
  - behavior in time/place?
  - behavior in frequency?
  - directional properties?
  - bandwidth dependency?
  - behavior in delay?

BASIC PROPAGATION MECHANSISMS

Free-space loss



If we assume RX antenna to be isotropic:

$$P_{RX} = \left(\frac{\lambda}{4\pi d} \right)^2 P_{TX}$$

Attenuation between two isotropic antennas in free space is (free-space loss):

$$L_{free}(d) = \left(\frac{4\pi d}{\lambda} \right)^2$$

Free-space loss
Friis' law

Received power, with antenna gains G_{TX} and G_{RX}:

$$P_{RX}(d) = \frac{G_{RX} G_{TX}}{L_{free}(d)} P_{TX} = P_{TX} \left(\frac{\lambda}{4\pi d} \right)^2 G_{RX} G_{TX}$$

Valid in the far field only

$$P_{RX|dB}(d) = P_{TX|dB} + G_{TX|dB} - L_{free|dB}(d) + G_{RX|dB}$$

$$= P_{TX|dB} + G_{TX|dB} - 10\log_{10}\left(\frac{4\pi d}{\lambda} \right)^2 + G_{RX|dB}$$

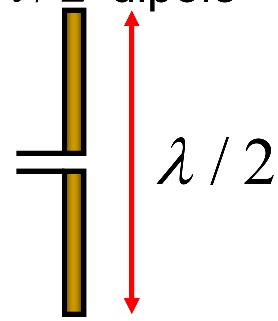
Free-space loss
What is far field?

Rayleigh distance:

$$d_R = \frac{2L_a^{\;2}}{\lambda}$$

where L_a is the largest dimesion of the antenna.

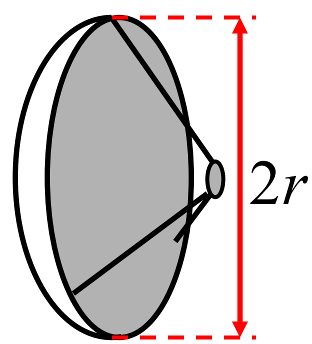
$\lambda/2$-dipole



$L_a = \lambda/2$

$d_R = \lambda/2$

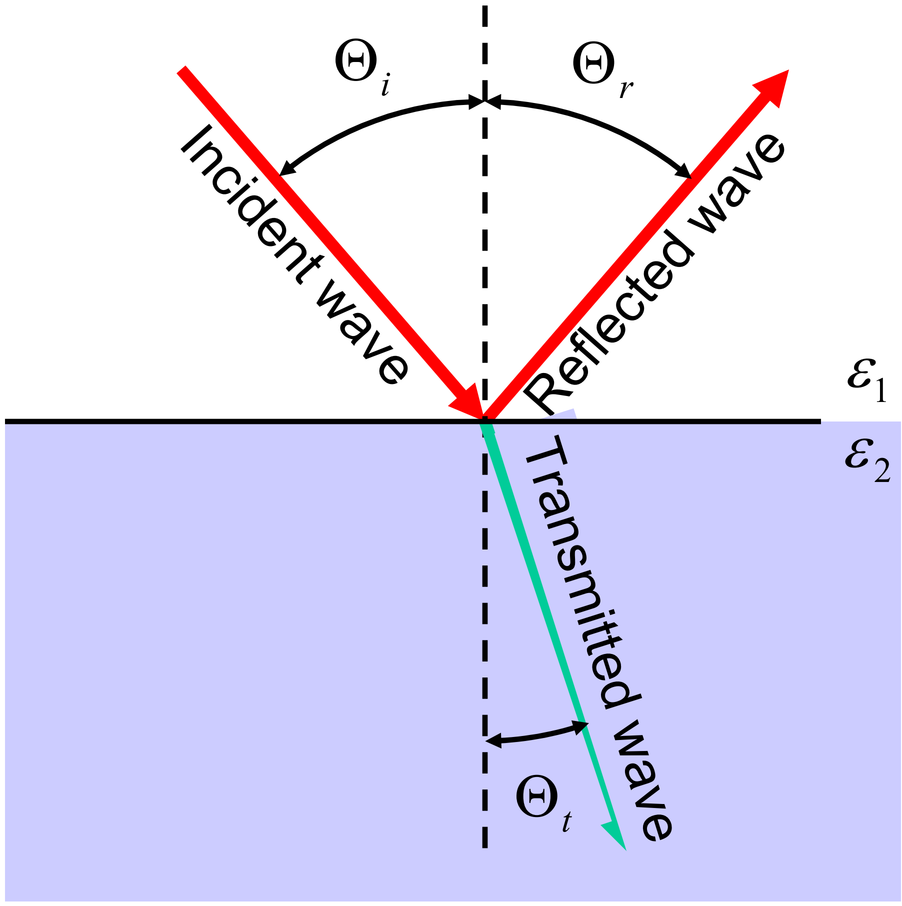
Parabolic



$L_a = 2r$

$$d_R = \frac{8r^2}{\lambda}$$

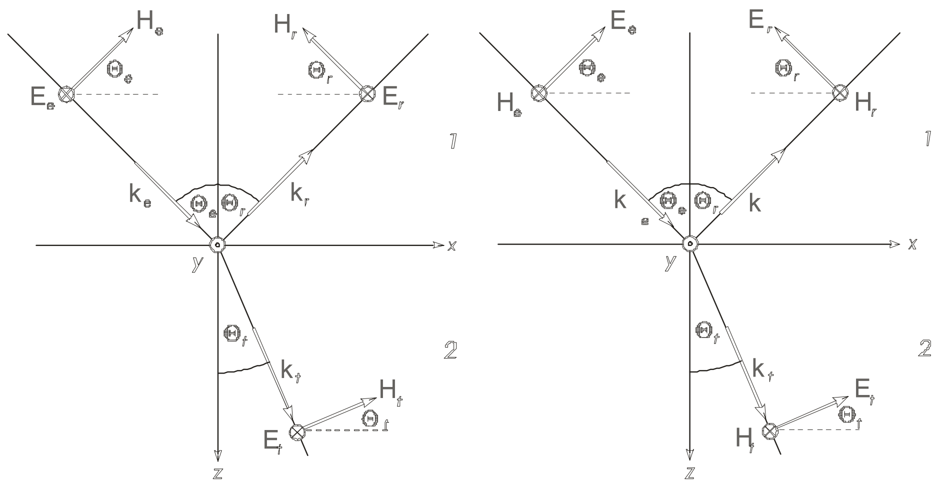
Reflection and transmission (1)



Reflection and transmission (2)

- Snell's law
  - Reflection angle $\Theta_r = \Theta_e$
  - Transmission angle $\dfrac{\sin\Theta_t}{\sin\Theta_e} = \dfrac{\sqrt{\varepsilon_1}}{\sqrt{\varepsilon_2}}$
- Transmission and reflection: distinguish TE and TM waves

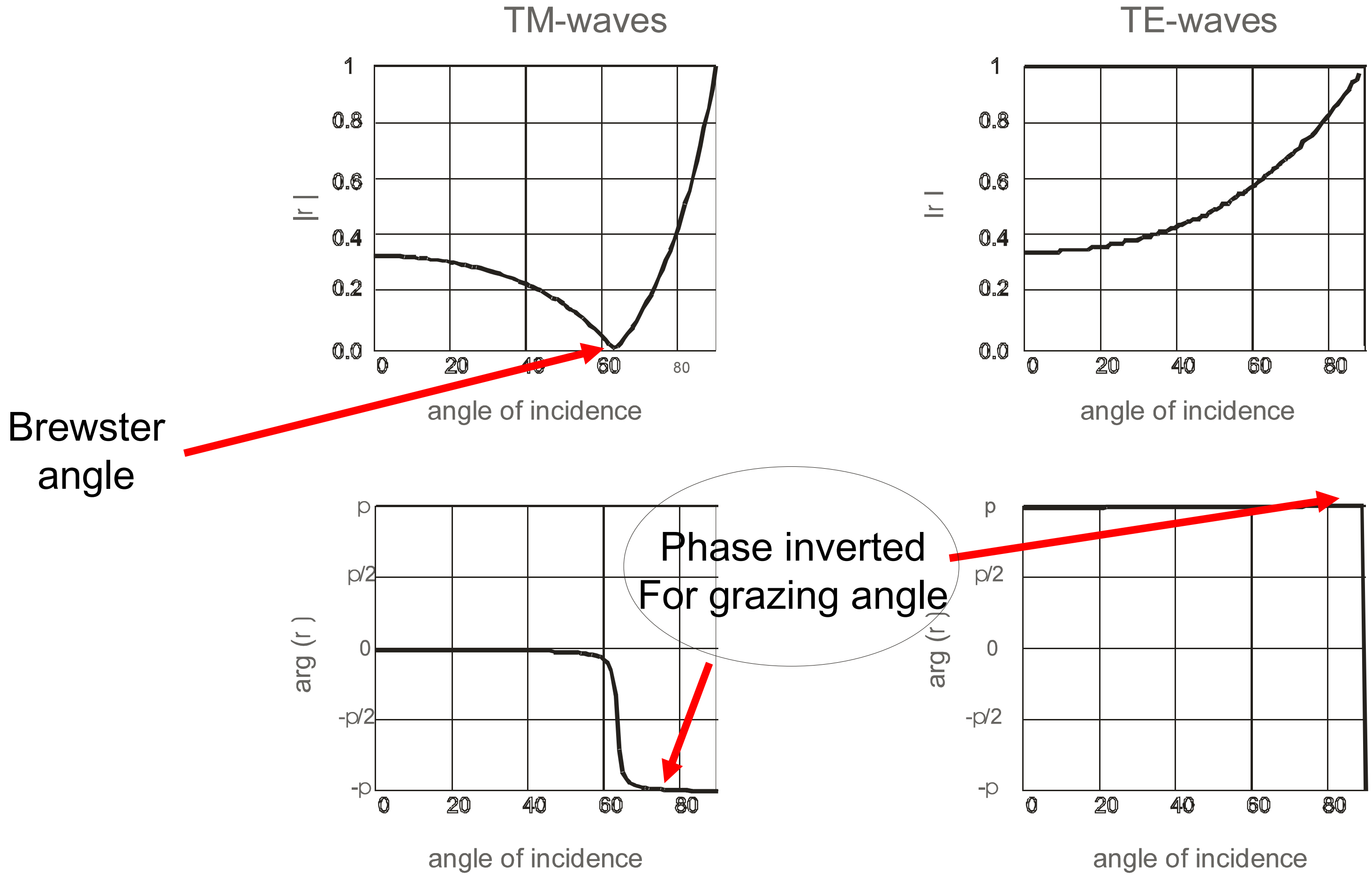
TE TM



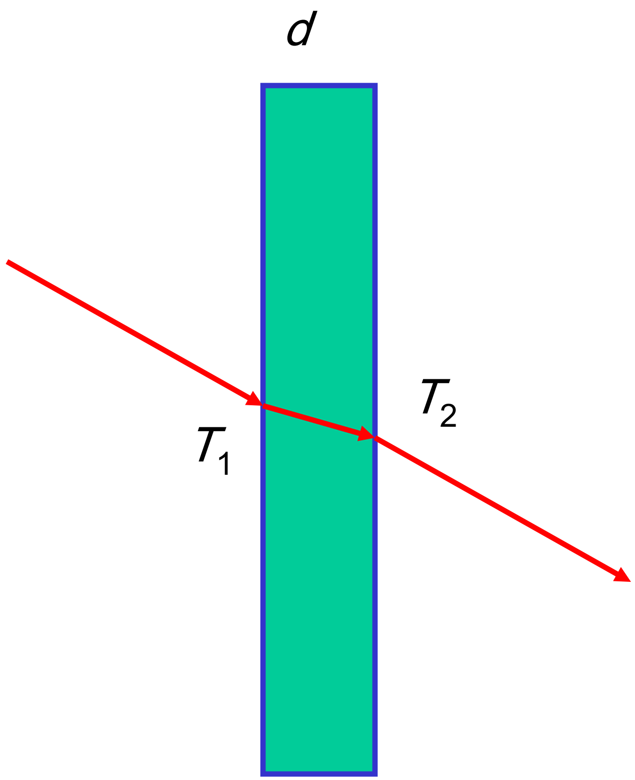
Reflection and transmission (3)

$$\rho_{TM} = \frac{\sqrt{\varepsilon_2}\cos\Theta_e - \sqrt{\varepsilon_1}\cos(\Theta_t)}{\sqrt{\varepsilon_2}\cos\Theta_e + \sqrt{\varepsilon_1}\cos(\Theta_t)}$$

$$\rho_{TE} = \frac{\sqrt{\varepsilon_1}\cos(\Theta_e) - \sqrt{\varepsilon_2}\cos(\Theta_t)}{\sqrt{\varepsilon_1}\cos(\Theta_e) + \sqrt{\varepsilon_2}\cos(\Theta_t)}$$



Transmission through a wall – layered structures



Total transmission coefficient

$$T = \frac{T_1 T_2 e^{-j\alpha}}{1+R_1 R_2 e^{-2j\alpha}}$$

total reflection coefficient

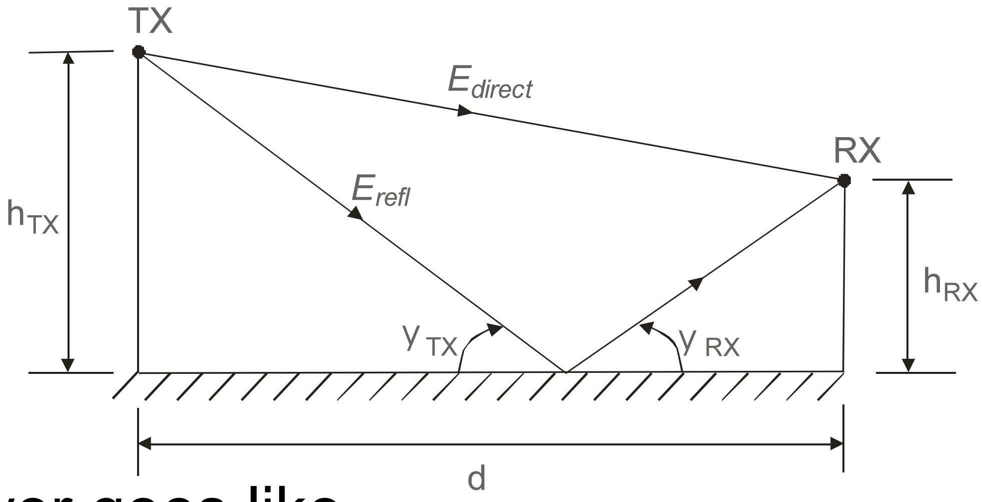
$$\rho = \frac{\rho_1 + \rho_2 e^{-j2\alpha}}{1+\rho_1 \rho_2 e^{-2j\alpha}}$$

with the electrical length in the wall

$$\alpha = \frac{2\pi}{\lambda}\sqrt{\varepsilon_1}\, d_{\mathrm{layer}} \cos(\Theta_{\mathrm{t}})$$

The d^{-4} law (1)

- For the following scenario



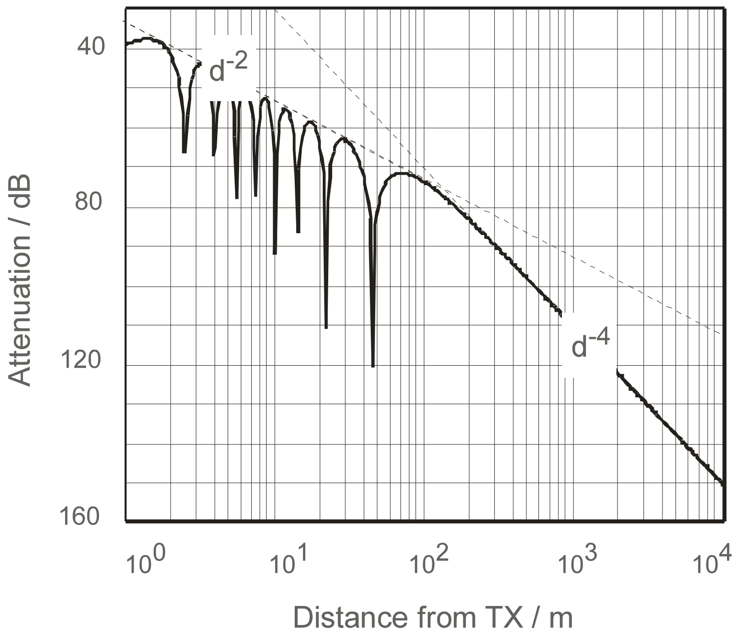
- the power goes like

$$P_{RX}(d) \approx P_{TX} G_{TX} G_{RX} \left(\frac{h_{TX} h_{RX}}{d^2} \right)^2 .$$

- for distances greater than

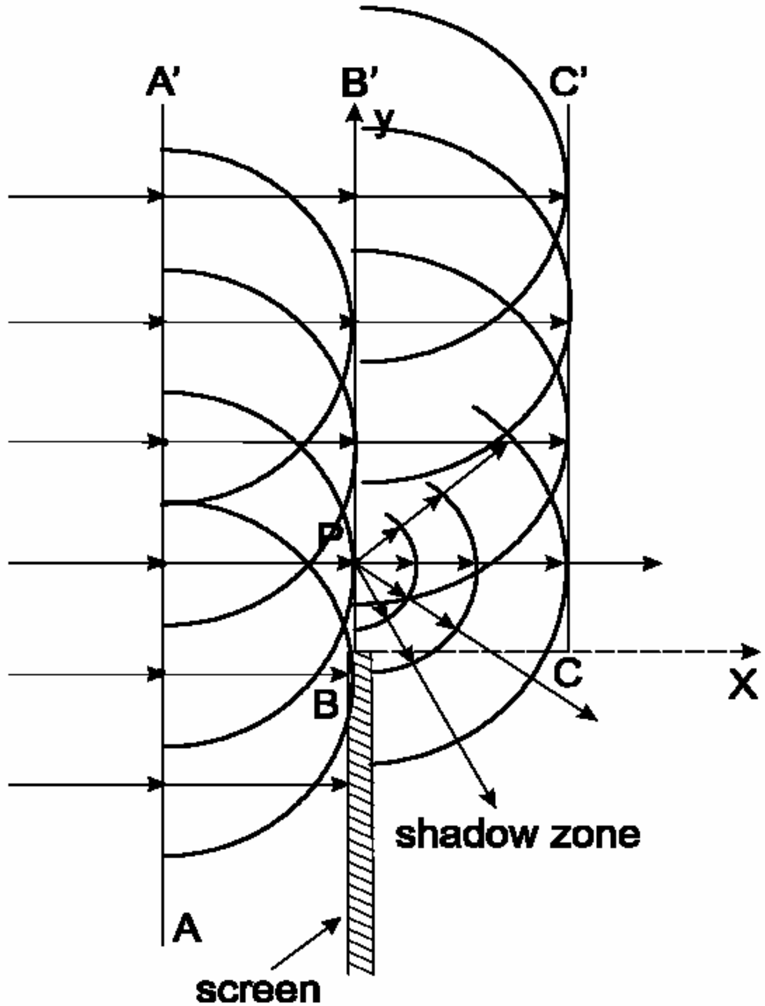
$$d_{\text{break}} \gtrsim 4 h_{TX} h_{RX} / \lambda$$

The d^{-4} law (2)

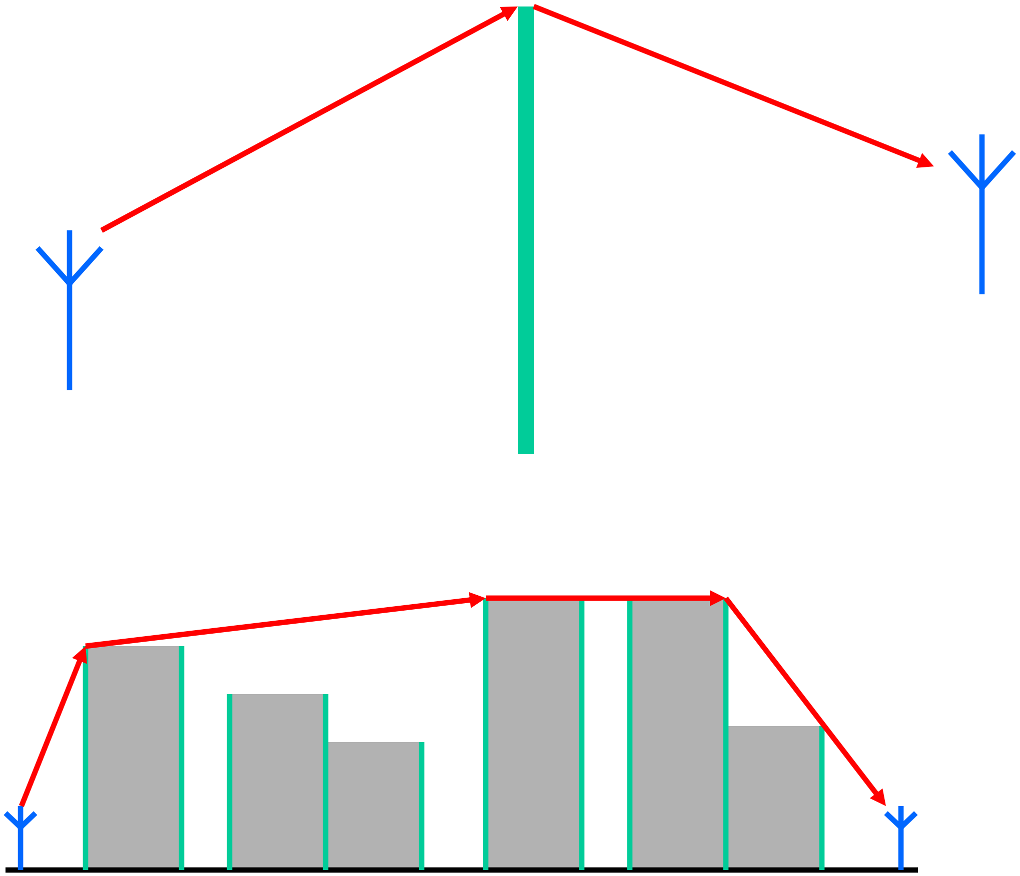


40
80
120
160
d^{-2}
d^{-4}
Attenuation / dB
10^0
10^1
10^2
10^3
10^4
Distance from TX / m

Diffraction, Huygen's principle

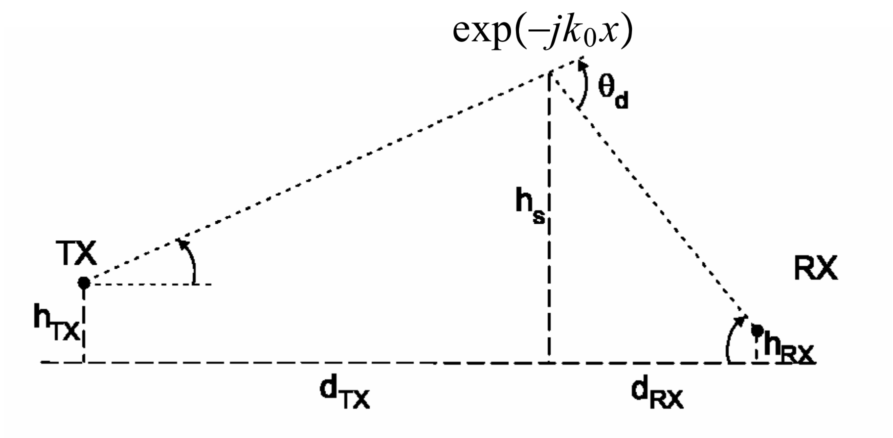


Diffraction



- Single or multiple edges
- makes it possible to go behind corners
- less pronounced when the wavelength is small compared to objects

Diffraction coefficient



The Fresnel integral is defined

$$F(\nu_F) = \int_0^{\nu_F} \exp\left(-j\pi \frac{t^2}{2}\right) dt.$$

with the Fresnel parameter

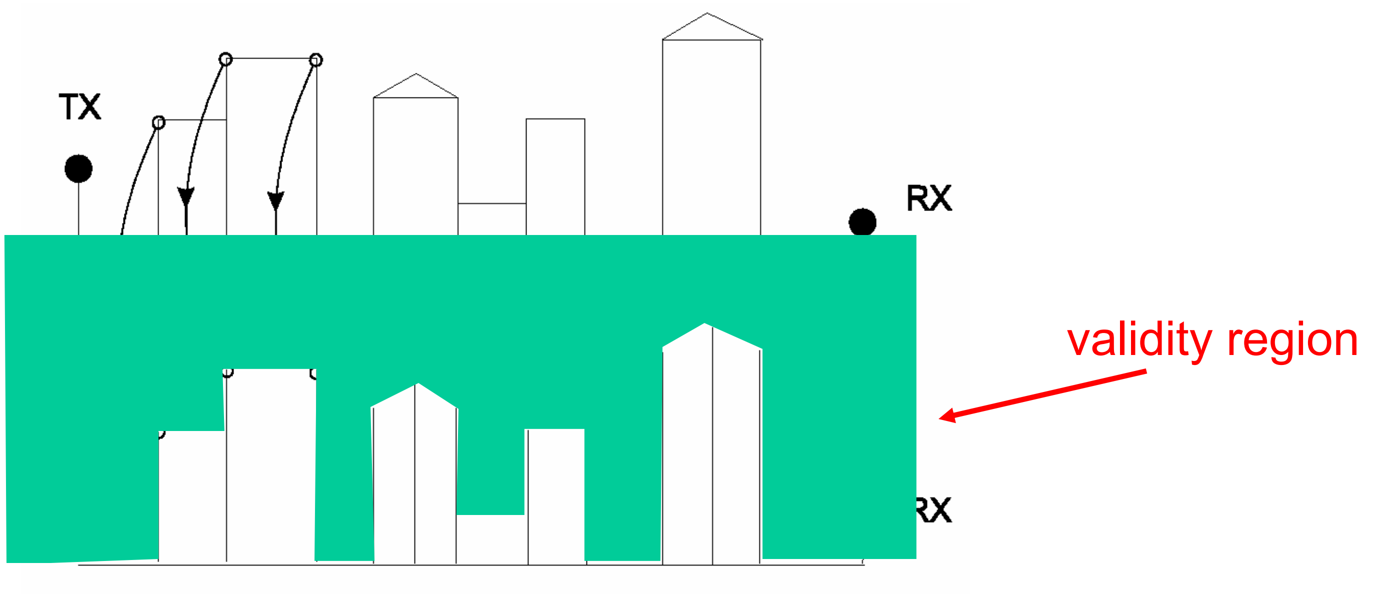
$$\nu_F = \alpha_k \sqrt{\frac{2d_1 d_2}{\lambda(d_1+d_2)}}$$

Total field

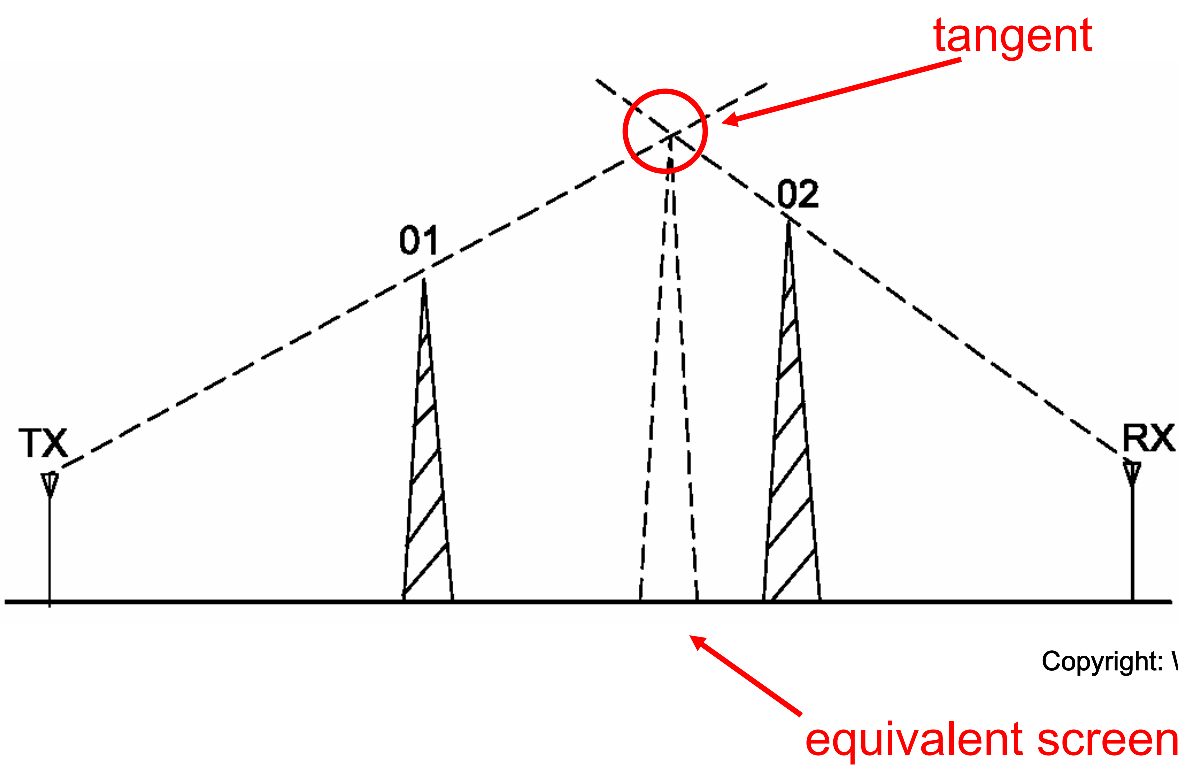
$$E_{\text{total}} = \exp(-jk_0 x)\left(\frac{1}{2} - \frac{\exp(-j\pi/4)}{\sqrt{2}} F(\nu_F)\right)$$

Fresnel integral

Diffraction in real environments

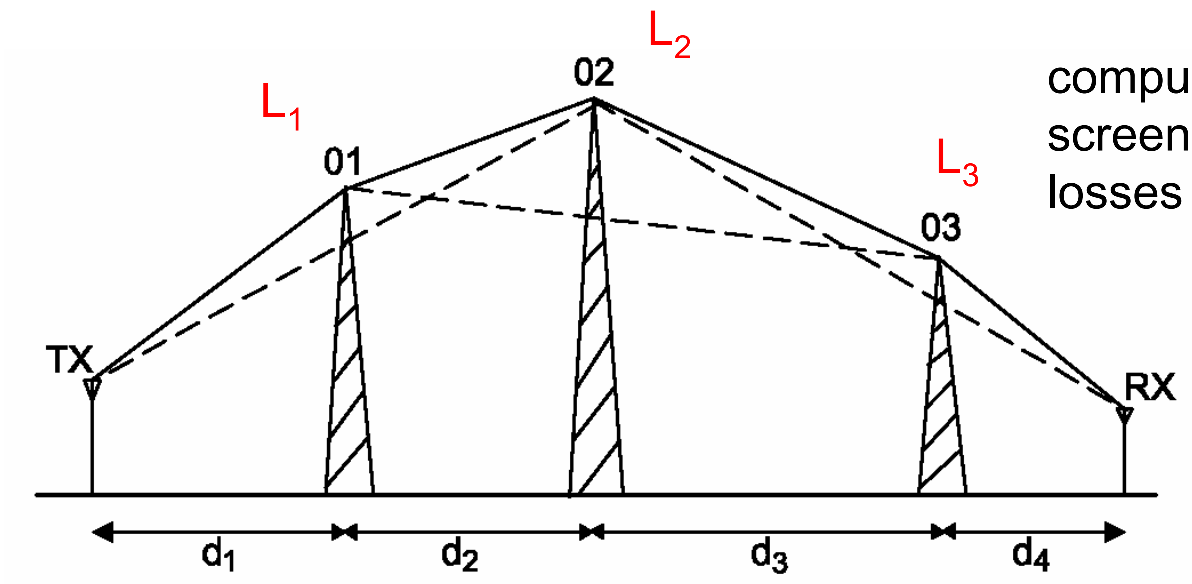


Diffraction – Bullington's method



$$E_{\text{total}} = \exp(-jk_0x)\left(\frac{1}{2} - \frac{\exp(-j\pi/4)}{\sqrt{2}}F\left(\nu_{\mathrm{F}}\right)\right) \qquad \nu_{\mathrm{F}} = \alpha_k\sqrt{\frac{2d_1d_2}{\lambda(d_1+d_2)}}$$

Diffraction – Epstein-Petersen Method



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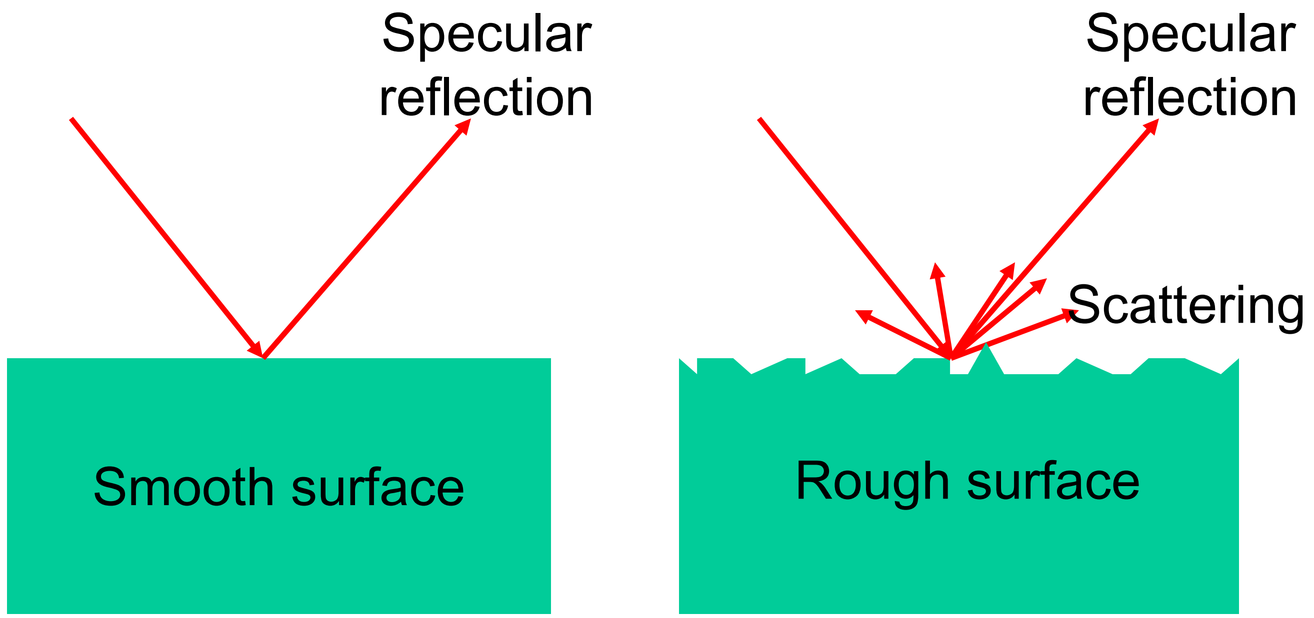
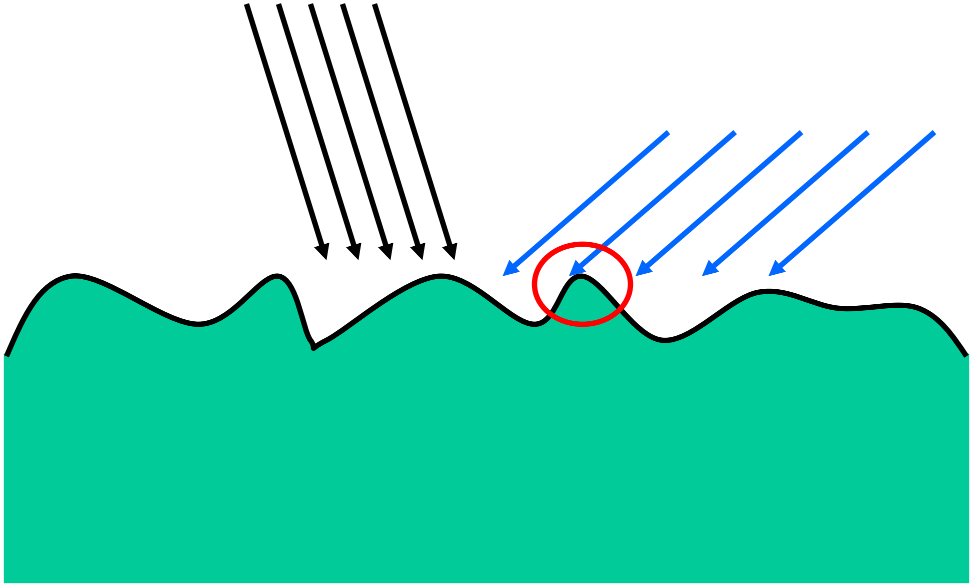
compute diffraction loss for each screen separately and add the losses

$$L_{tot}=L_1+L_2+L_3$$

Scattering


Specular
reflection
Smooth surface
Specular
reflection
Scattering
Rough surface

Kirchhoff theory – scattering by rough surfaces



for Gaussian surface distribution

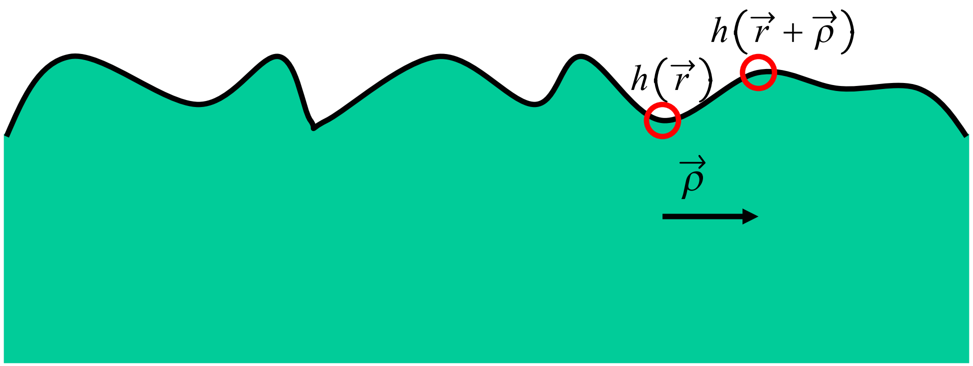
$$\rho_{\text{rough}} = \rho_{\text{smooth}} \exp\left[-2\left(k_0 \sigma_{\text{h}} \sin\psi\right)^2\right]$$

angle of incidence

standard deviation of height

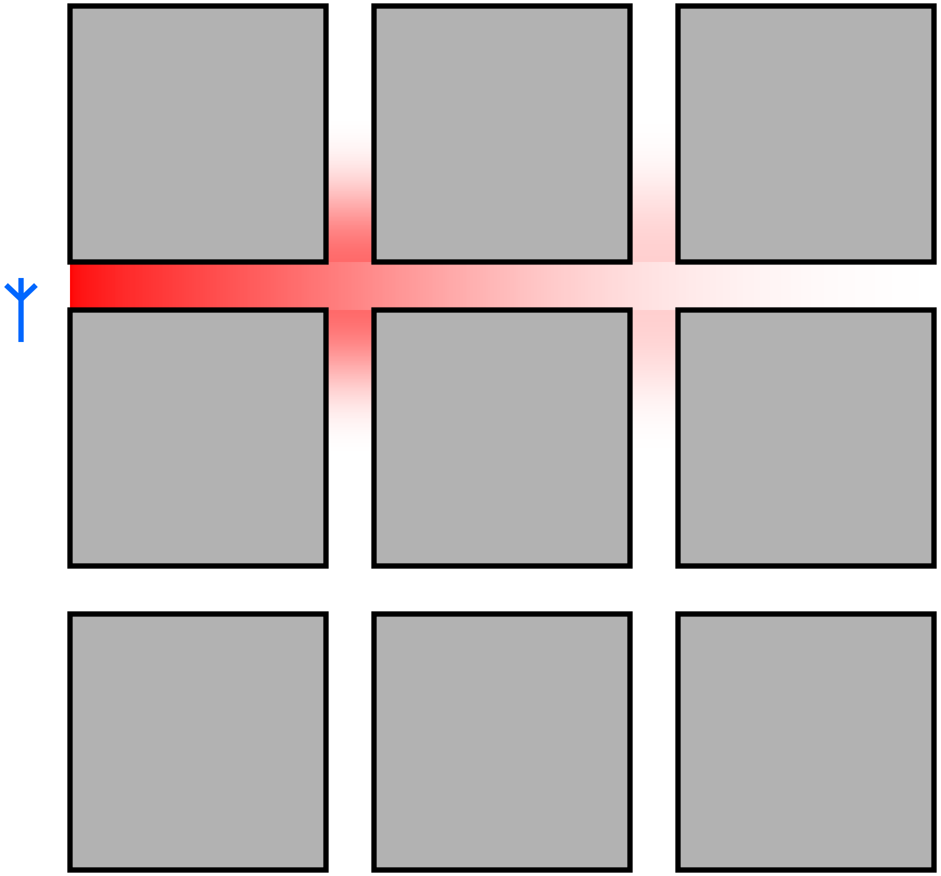
Pertubation theory – scattering by rough surfaces

$$\sigma_h^2 W(\vec{\rho}) = E_{\vec{r}}\{h(\vec{r})h(\vec{r}+\vec{\rho})\}$$



More accurate than Krichhoff theory, especially for large angles of incidence and “rougher” surfaces

Waveguiding



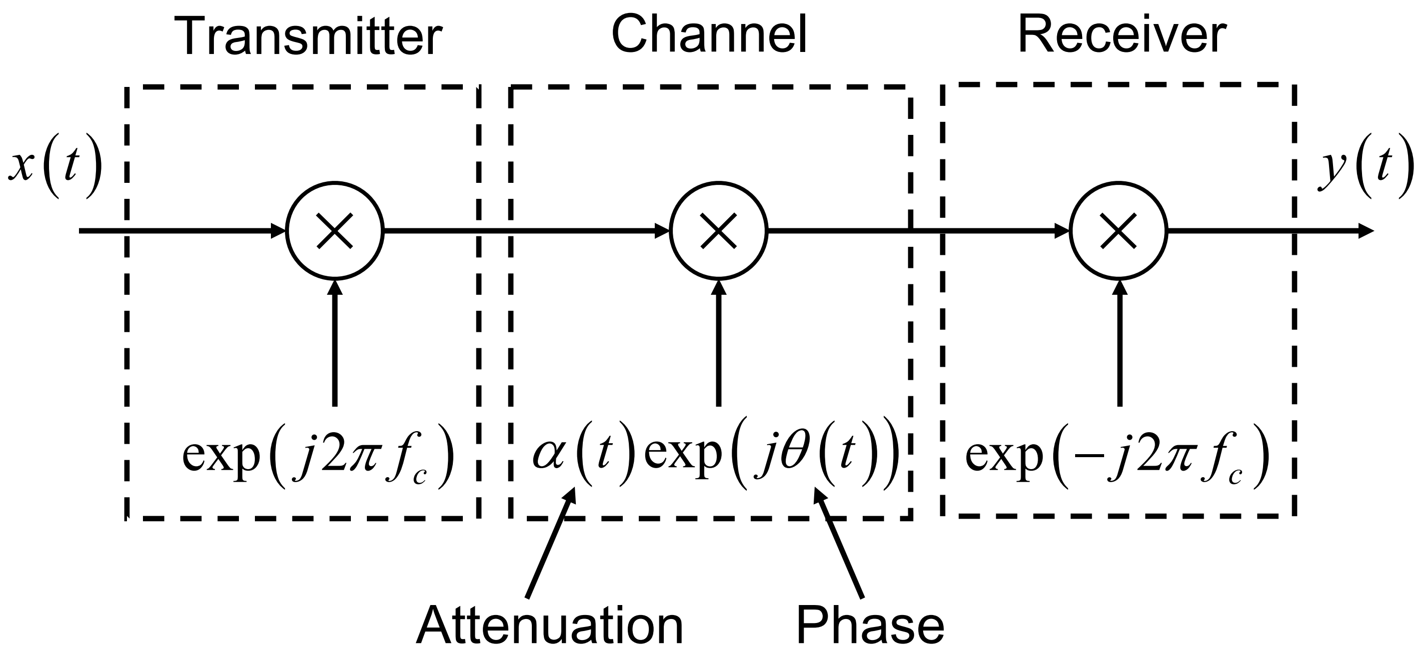
Waveguiding effects often result in lower propagation exponents

n=1.5-5

This means lower path loss along certain street corridors

Statistical modeling

A narrowband system described in complex notation (noise free)



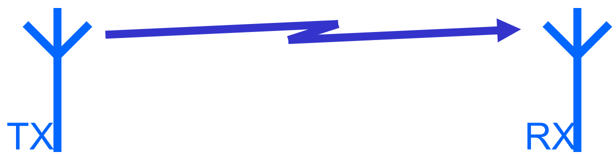
In: $x(t) = A(t)\exp(j\phi(t))$

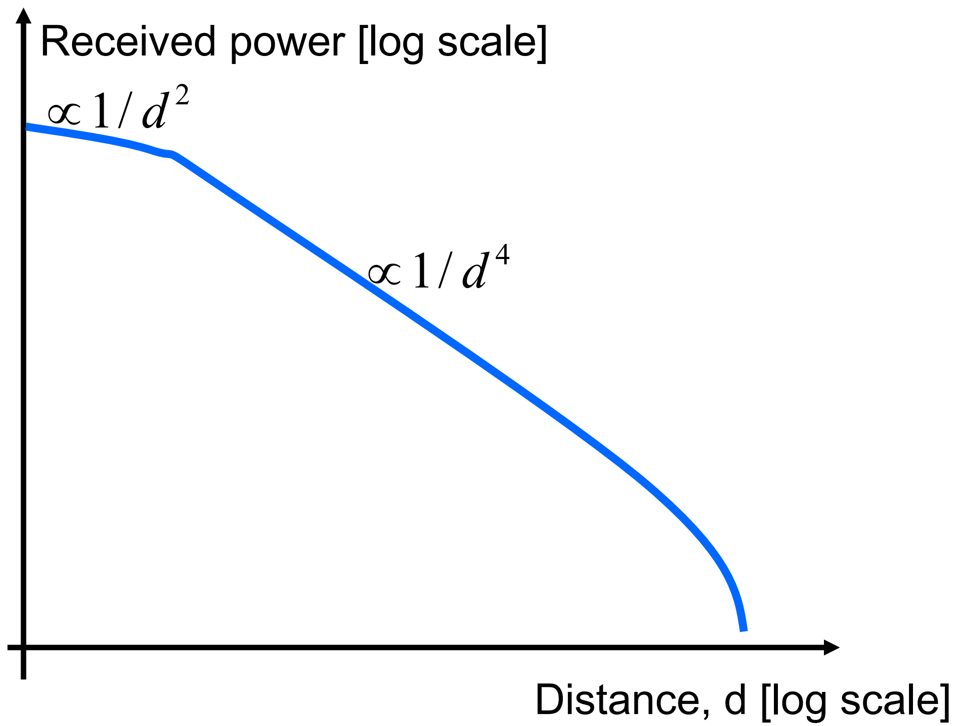
Out: $y(t) = A(t)\exp(j\phi(t))\cancel{\exp(j2\pi f_c t)}\alpha(t)\exp(j\theta(t))\cancel{\exp(-j2\pi f_c t)}$

$$= A(t)\alpha(t)\exp(j(\phi(t)+\theta(t)))$$

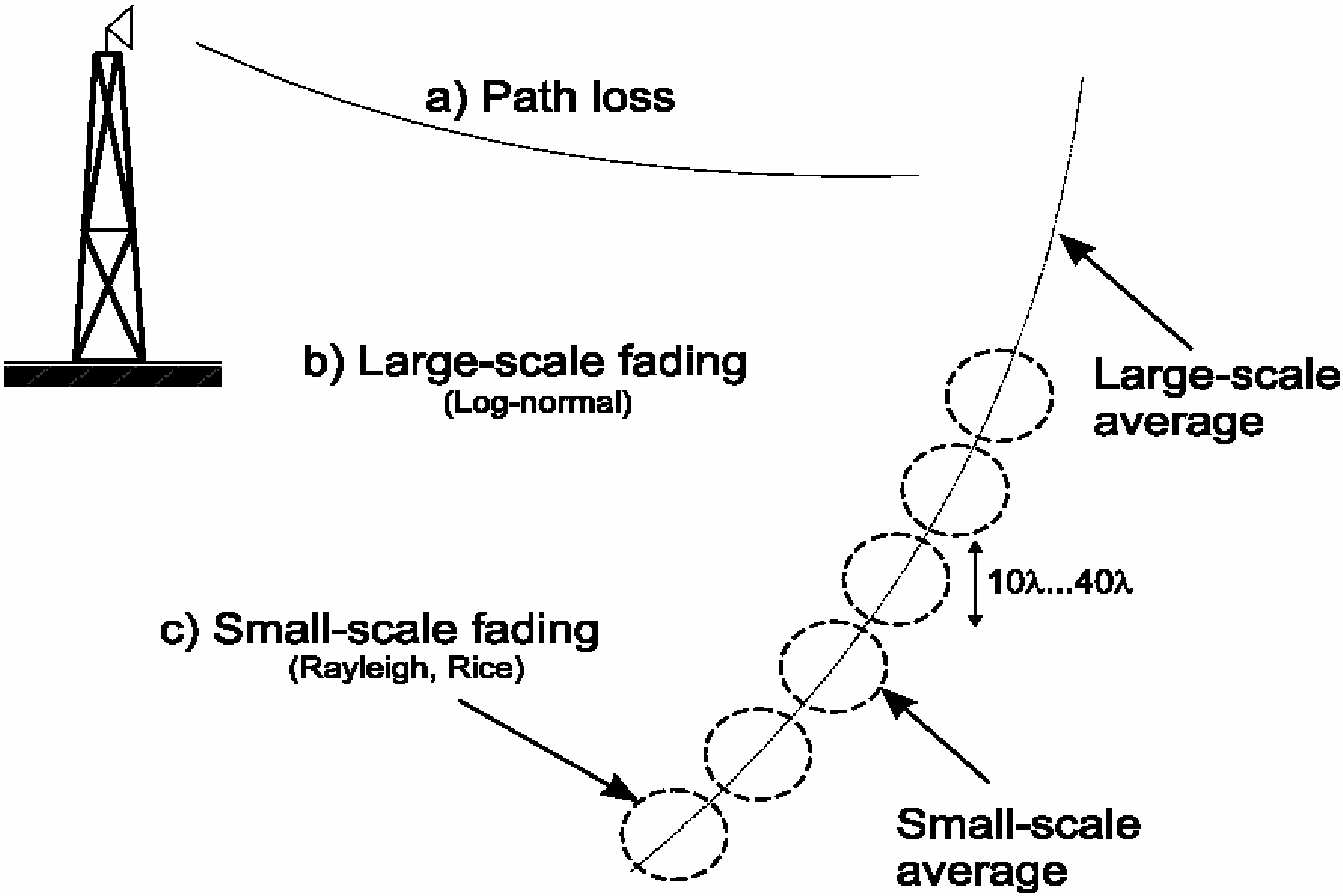
It is the behavior of the channel attenuation and phase we are going to model.

THE RADIO CHANNEL
Path loss

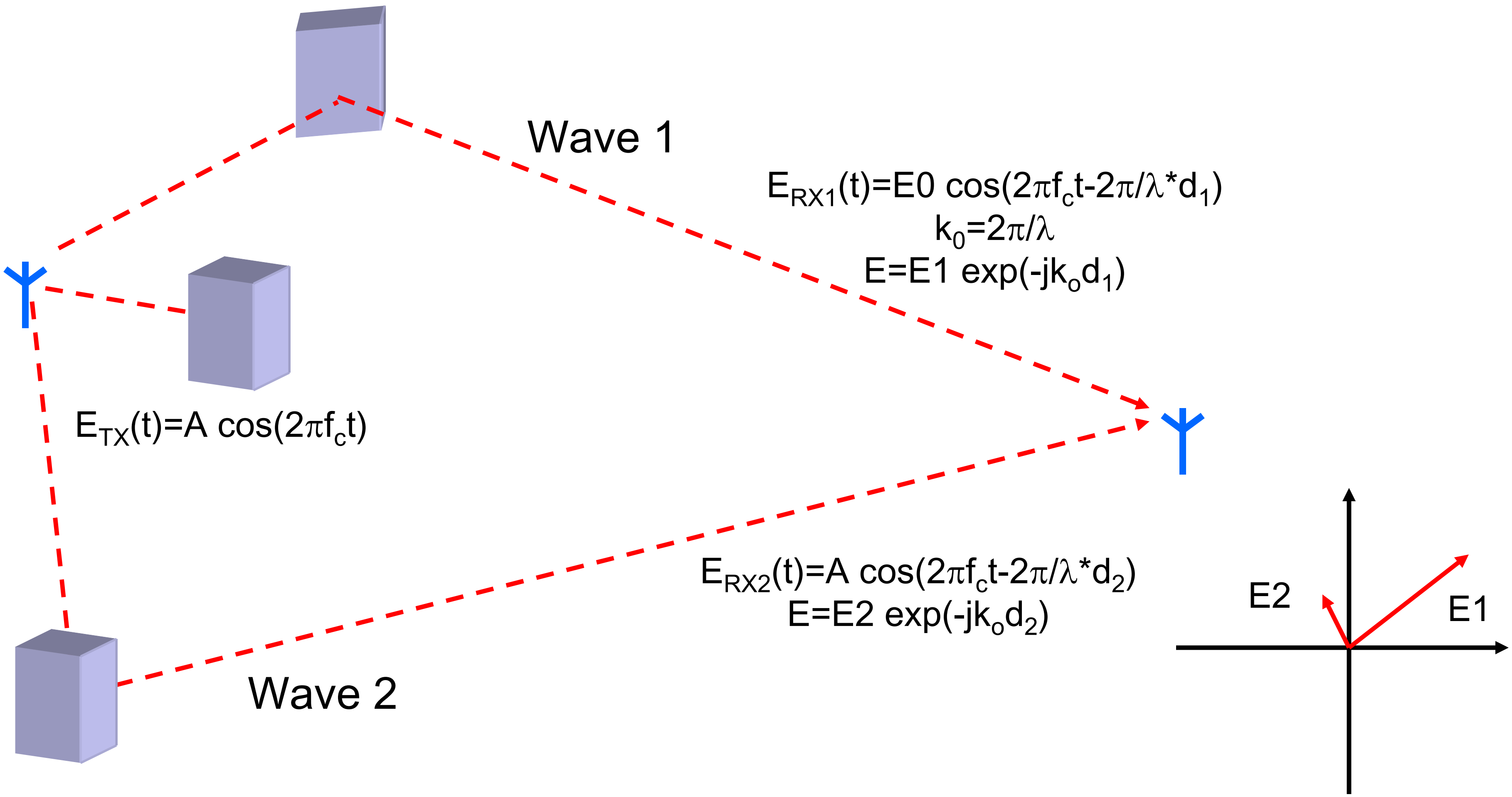




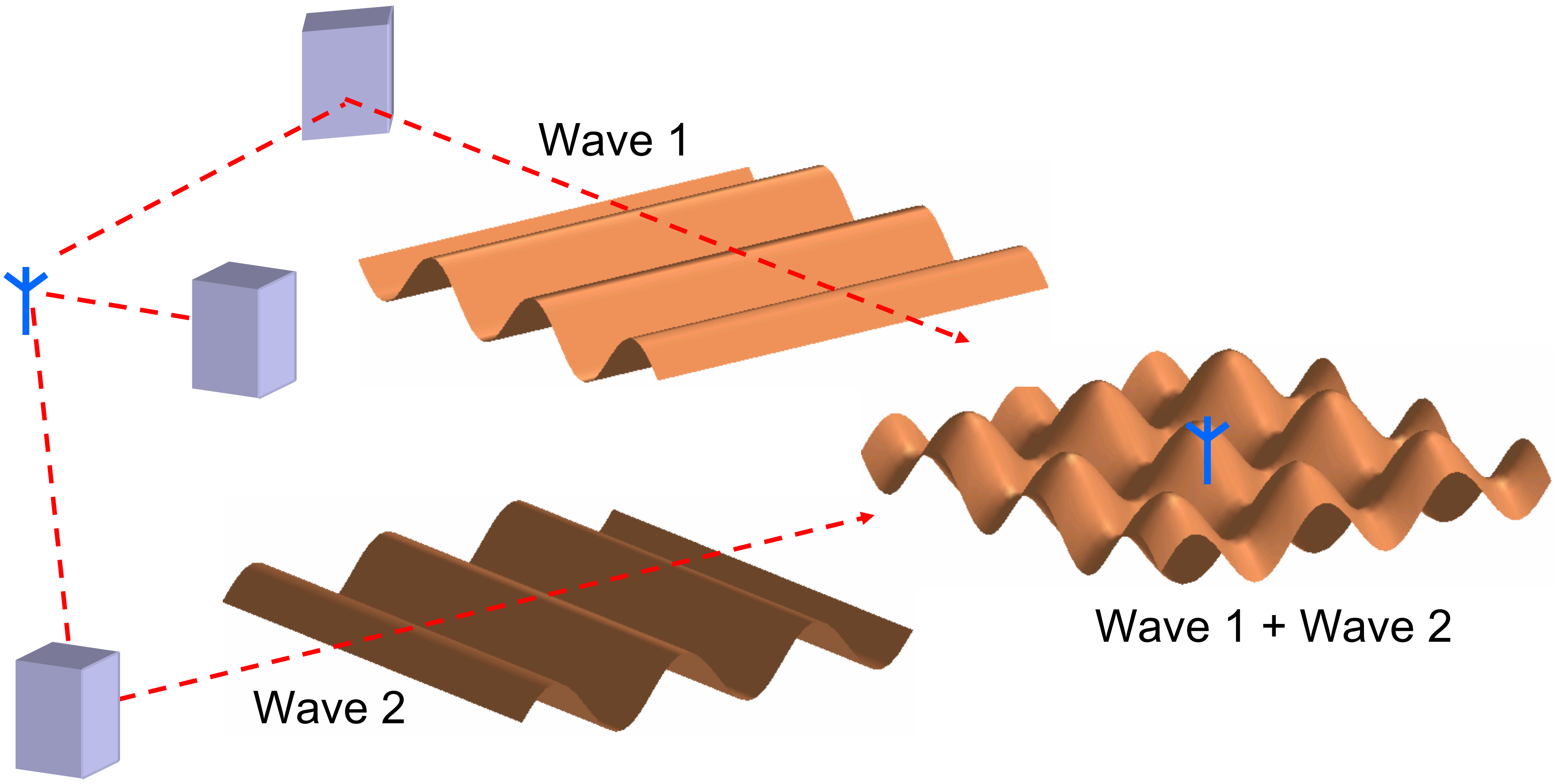
What is large scale and small scale?



Small-scale fading
Two waves



Small-scale fading
Two waves



Small-scale fading (cont.)

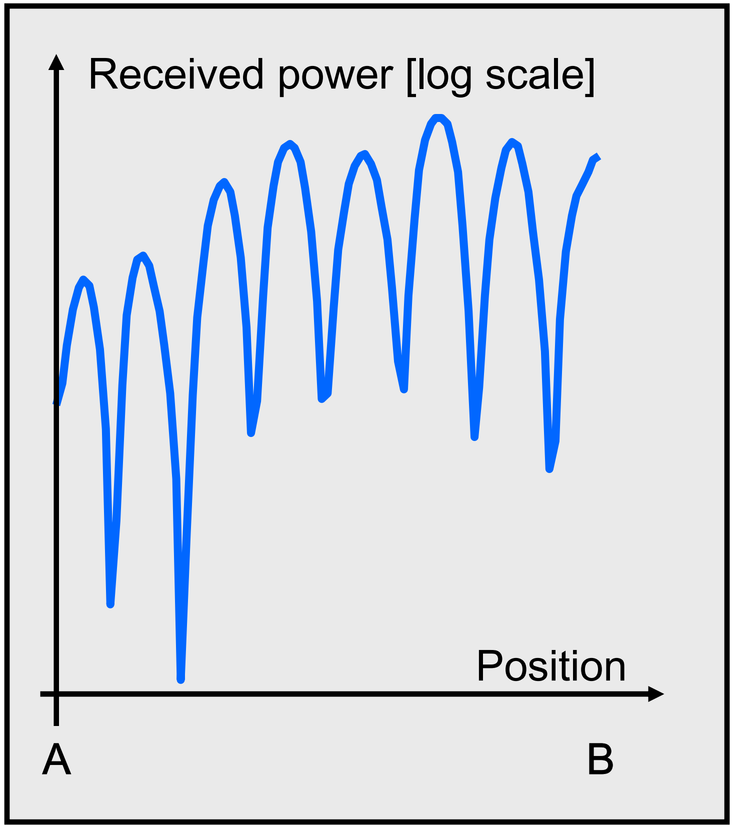
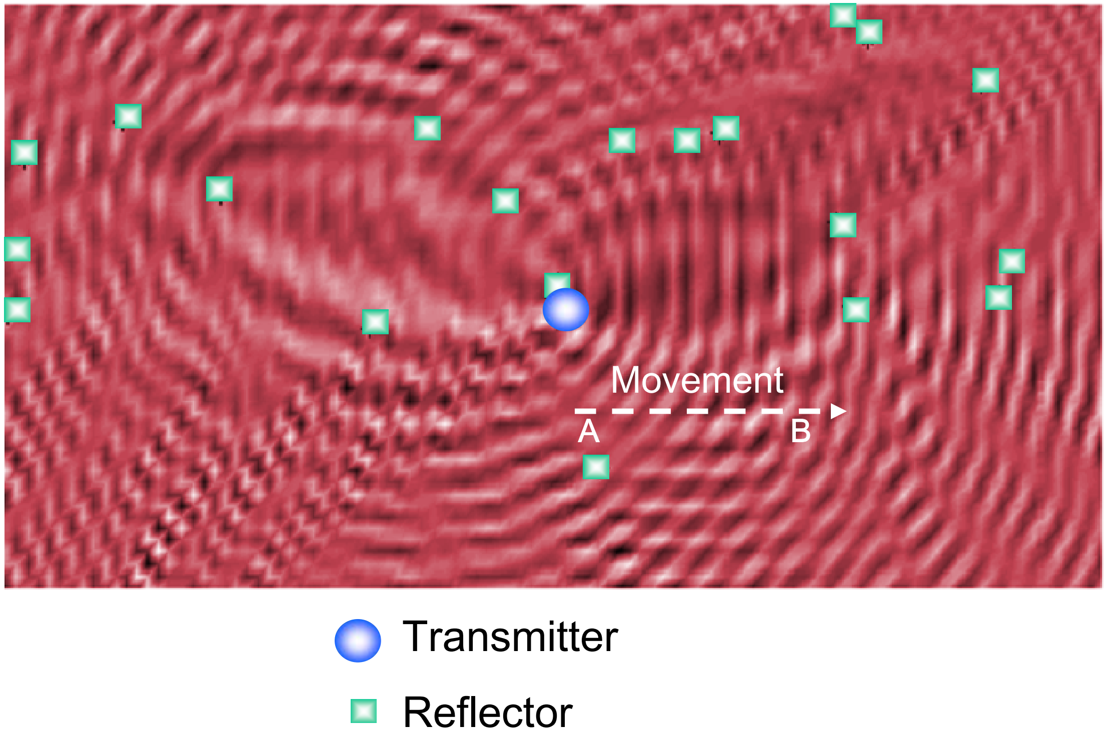
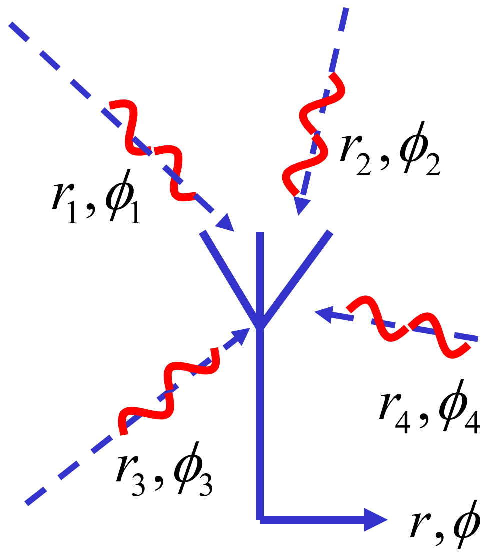


Illustration of interference pattern from above

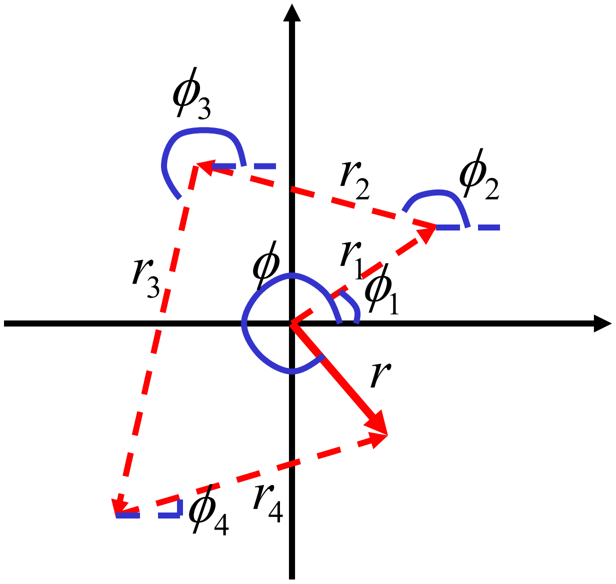


Small-scale fading
Many incoming waves

Many incoming waves with independent amplitudes and phases



Add them up as phasors



$$r\exp(j\phi) = r_1\exp(j\phi_1) + r_2\exp(j\phi_2) + r_3\exp(j\phi_3) + r_4\exp(j\phi_4)$$

$\mathrm{Re}(r) \in N(0,\sigma^2)$

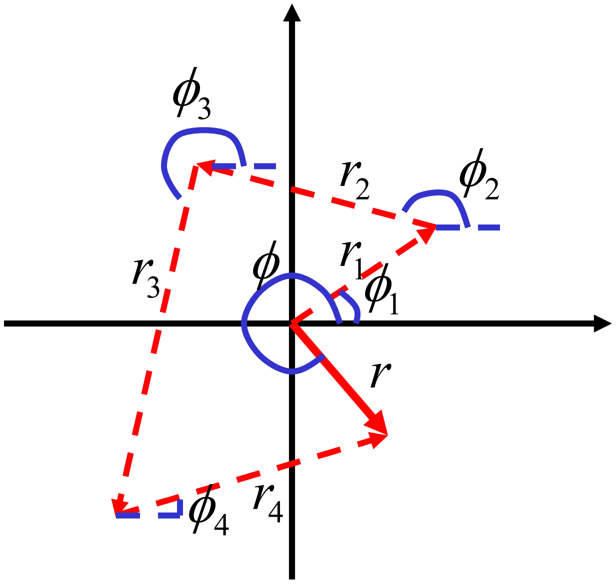
Small-scale fading
Many incoming waves

Re and Im components are sums of many independent equally distributed components

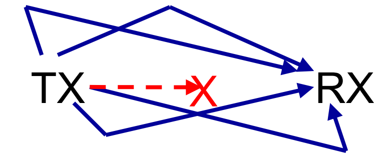
$$\mathrm{Re}(r) \in N(0,\sigma^2)$$

Re(r) and Im(r) are independent

The phase of r has a uniform distribution

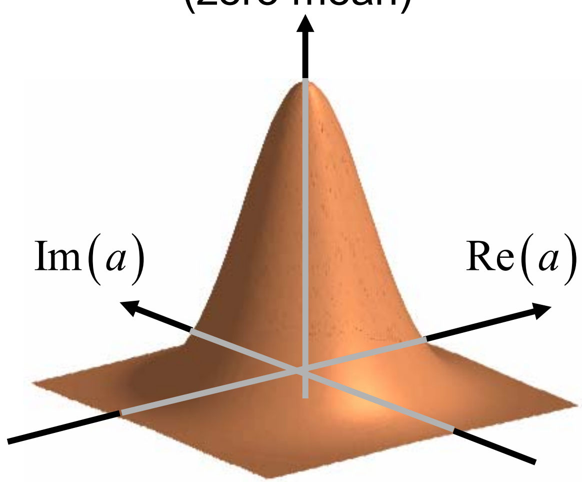


Small-scale fading
Rayleigh fading



No dominant component
(no line-of-sight)

Tap distribution
2D Gaussian
(zero mean)

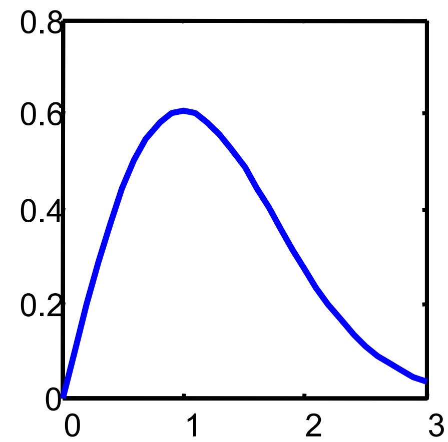


No line-of-sight component

$$r = |a|$$

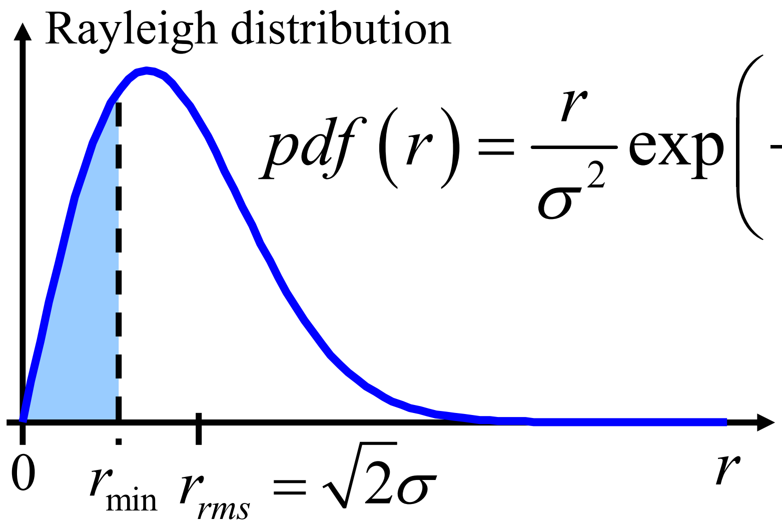
Amplitude distribution

Rayleigh



$$pdf(r) = \frac{r}{\sigma^2}\exp\left(-\frac{r^2}{2\sigma^2}\right)$$

Small-scale fading
Rayleigh fading



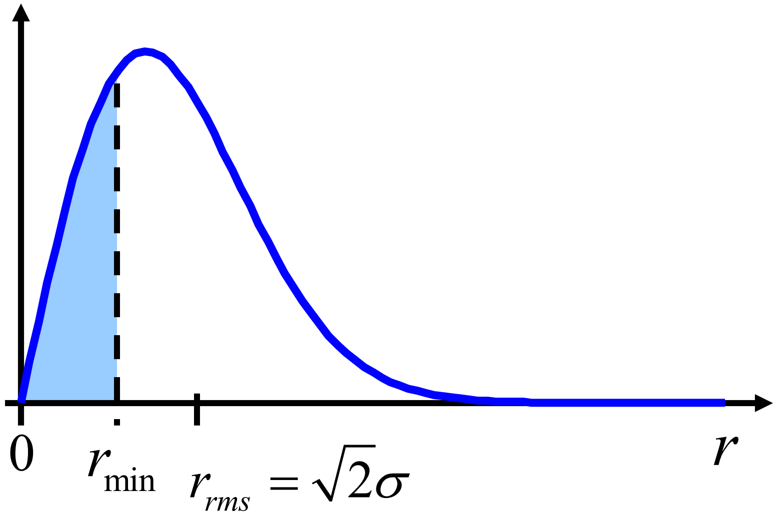
$$pdf(r) = \frac{r}{\sigma^2}\exp\left(-\frac{r^2}{2\sigma^2}\right)$$

$$\Pr\left(r < r_{\min}\right) = \int_0^{r_{\min}} pdf(r)\,dr = 1 - \exp\left(-\frac{r_{\min}^{\;2}}{r_{rms}^{\;2}}\right)$$

Small-scale fading
Rayleigh fading – fading margin

$$M = \frac{r_{rms}{}^2}{r_{\min}{}^2}$$

$$M_{|dB} = 10\log_{10}\left(\frac{r_{rms}{}^2}{r_{\min}{}^2}\right)$$



Small-scale fading
Rayleigh fading – fading margin

How many dB fading margin, against Rayleigh fading, do we need to obtain an outage probability of 1%?

$$\Pr\left(r < r_{\min}\right) = 1 - \exp\left(-\frac{r_{\min}^2}{r_{rms}^2}\right) = 1\% = 0.01$$

Some manipulation gives

$$1 - 0.01 = \exp\left(-\frac{r_{\min}^2}{r_{rms}^2}\right) \Rightarrow \ln\left(0.99\right) = -\frac{r_{\min}^2}{r_{rms}^2}$$

$$\Rightarrow \frac{r_{\min}^2}{r_{rms}^2} = -\ln\left(0.99\right) = 0.01 \quad \Rightarrow M = \frac{r_{rms}^2}{r_{\min}^2} = 1/0.01 = 100$$

$$\Rightarrow \boxed{M_{|dB} = 20}$$

Small-scale fading
Rayleigh fading – signal and interference

- What is the probability that the instantaneous SIR will be below 0 dB if the mean SIR is 10 dB when both the desired signal and the interferer experience Rayleigh fading?

$$\Pr\left(r < r_{\min}\right) = 1 - \frac{\overline{\sigma}^2 r_{\min}}{(\overline{\sigma}^2 + r_{\min}^2)} = 1 - \frac{10}{(10+1)} \approx 0.09$$

Small-scale fading
one dominating component

In case of Line-of-Sight (LOS) one component dominates.

- Assume it is aligned with the real axis

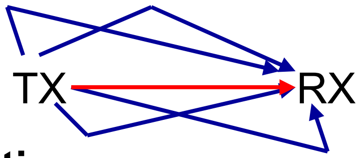
$$\mathrm{Re}(r) \in N(A, \sigma^2) \quad \mathrm{Im}(r) \in N(0, \sigma^2)$$

- The received amplitude has now a Ricean distribution instead of a Rayleigh

- The ratio between the power of the LOS component and the diffuse components is called Ricean K-factor

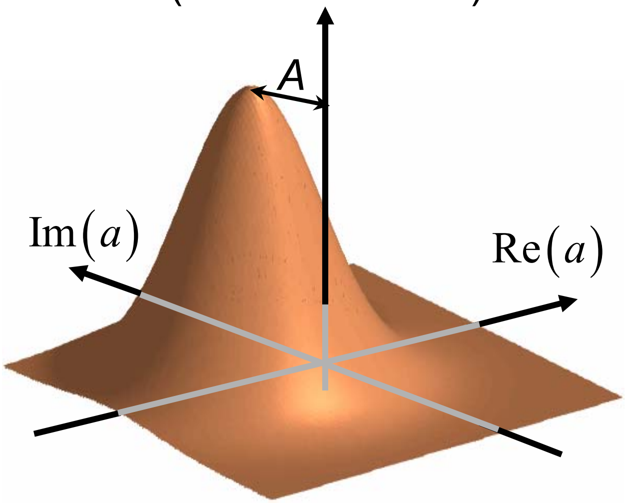
$$k = \frac{\text{Power in LOS component}}{\text{Power in random components}} = \frac{A^2}{2\sigma^2}$$

Small-scale fading
Rice fading

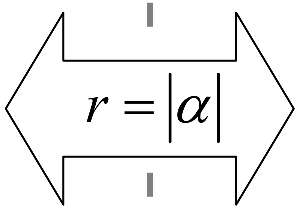


A dominant component (line of sight)

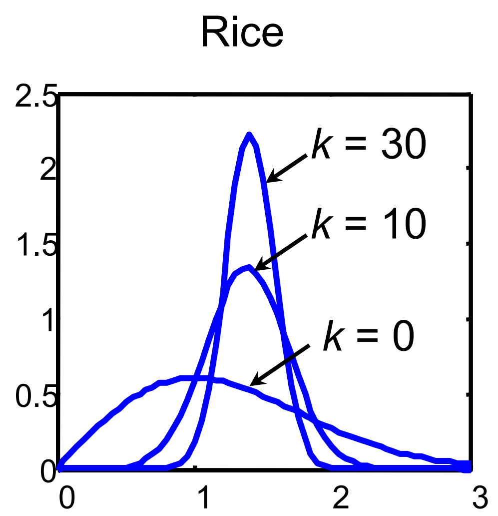
Tap distribution
2D Gaussian
(non-zero mean)



Line-of-sight (LOS) component with amplitude *A*.

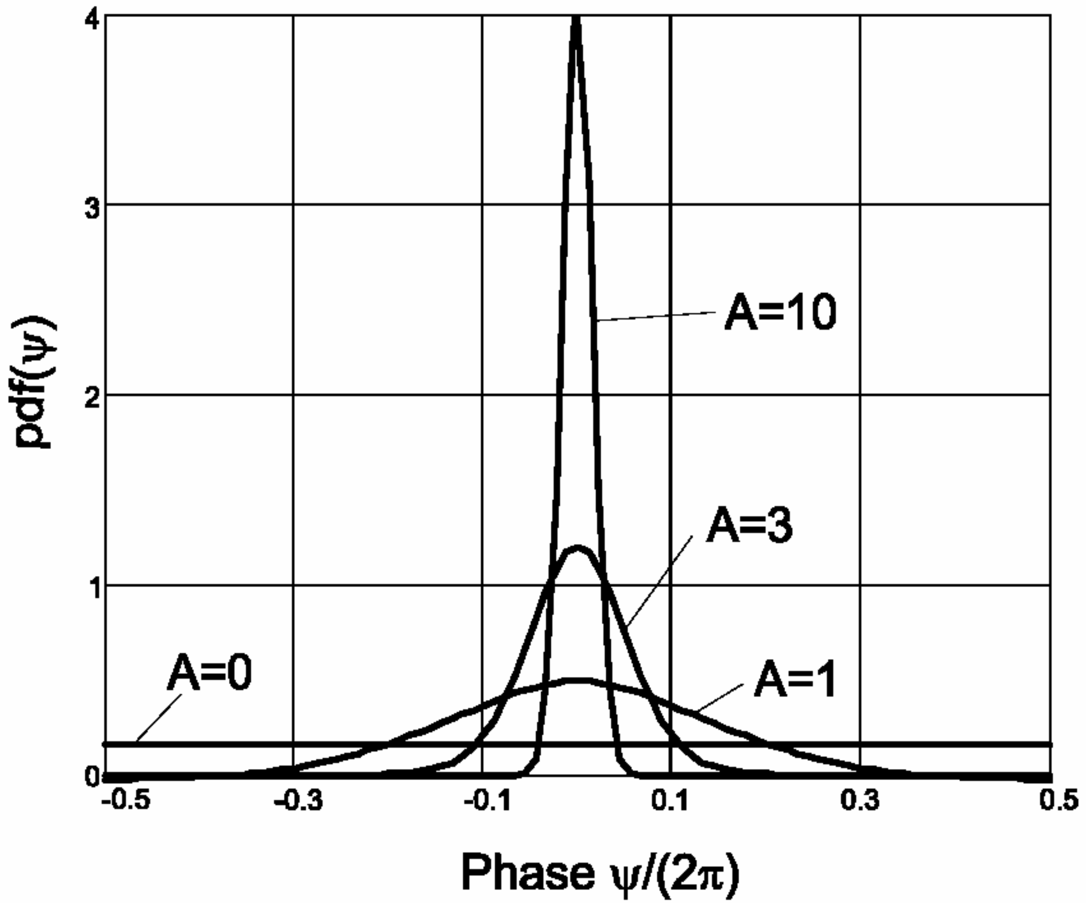


Amplitude distribution



$$pdf(r) = \frac{r}{\sigma^2} \exp\left(-\frac{r^2 + A^2}{2\sigma^2}\right) I_0\left(\frac{rA}{\sigma^2}\right)$$

Small-scale fading
Rice fading, phase distribution



Small-scale fading
Nakagami distribution

- In many cases the received signal can not be described as a pure LOS + diffuse components
- The Nakagami distribution is often used in such cases

$$pdf(r) = \frac{2}{\Gamma(m)}\left(\frac{m}{\Omega}\right)^m r^{2m-1} \exp\left(-\frac{m}{\Omega} r^2\right)$$

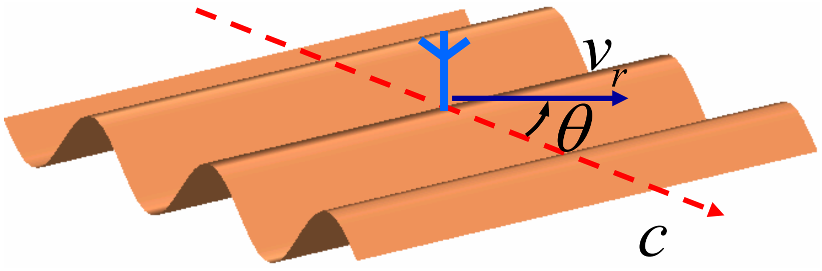
$\Gamma(m)$ is the gamma function

$$\Omega = \overline{r^2}$$

$$m = \frac{\Omega^2}{\overline{(r^2 - \Omega)^2}}$$

with m it is possible to adjust the dominating power

Small-scale fading
Doppler shifts



Receiving antenna moves with speed v_r at an angle θ relative to the propagation direction of the incoming wave, which has frequency f_0.

Frequency of received signal:

$$f = f_0 + \nu$$

where the Doppler shift is

$$\nu = -f_0 \frac{v_r}{c} \cos(\theta)$$

The maximal Doppler shift is

$$\nu_{\max} = f_0 \frac{v}{c}$$

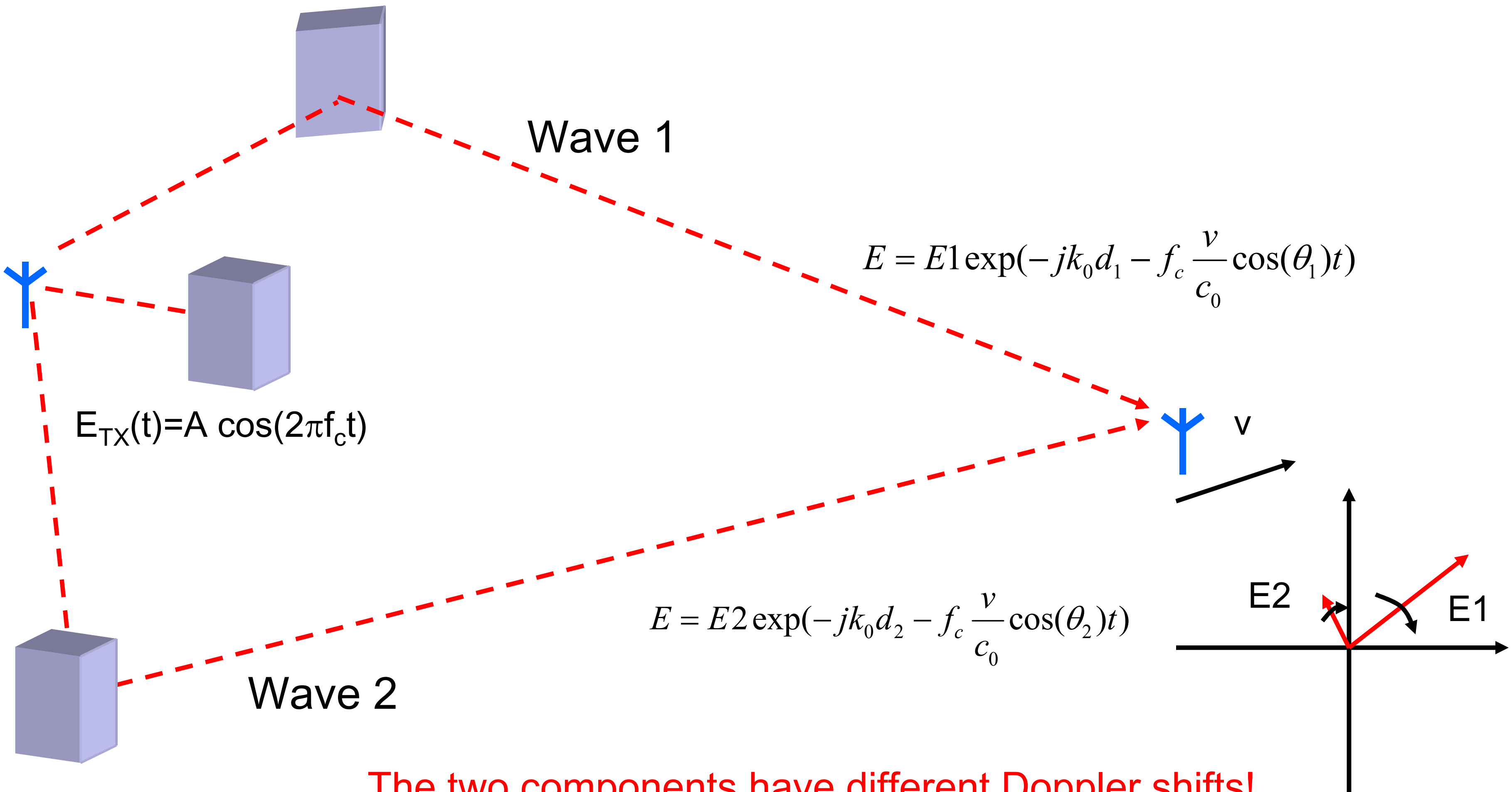
Small-scale fading
Doppler shifts

How large is the maximum Doppler frequency at pedestrian speeds for 5.2 GHz WLAN and at highway speeds using GSM 900?

$$\nu_{\max} = f_0 \frac{v}{c}$$

- f_0=5.2 10^9 Hz, v=5 km/h, (1.4 m/s) ⇨ 24 Hz
- f_0=900 10^6 Hz, v=110 km/h, (30.6 m/s) ⇨ 92 Hz

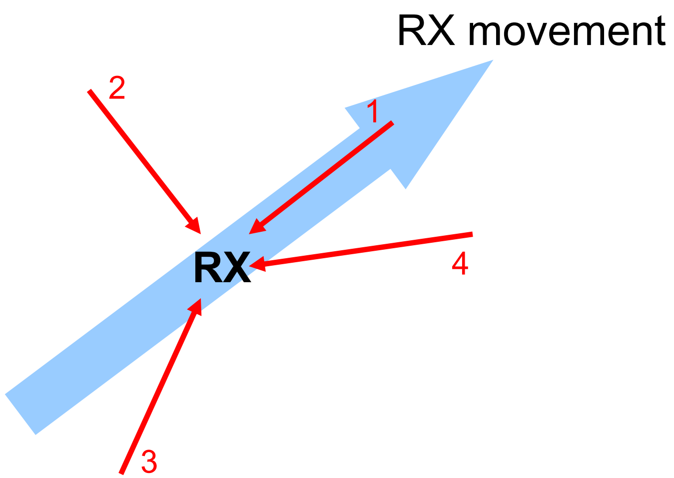
Small-scale fading
Doppler spectra



The two components have different Doppler shifts!
The Doppler shifts will cause a random frequency modulation

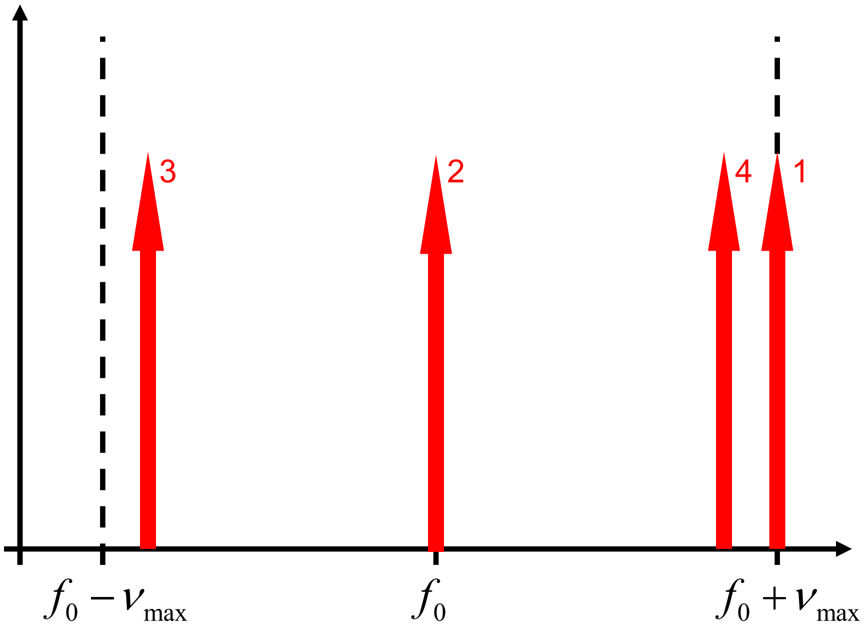
Small-scale fading
Doppler spectrum

Incoming waves from several directions
(relative to movement or RX)



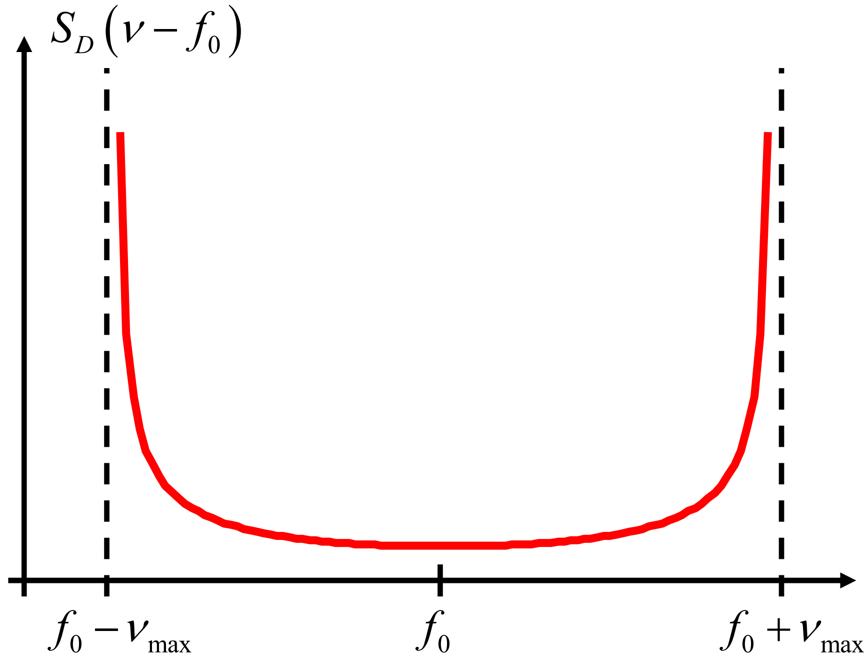
All waves of equal strength in this example, for simplicity.

Spectrum of received signal
when a f_0 Hz signal is transmitted.



Small-scale fading
The Doppler spectrum

Doppler spectrum
at center frequency f_0.



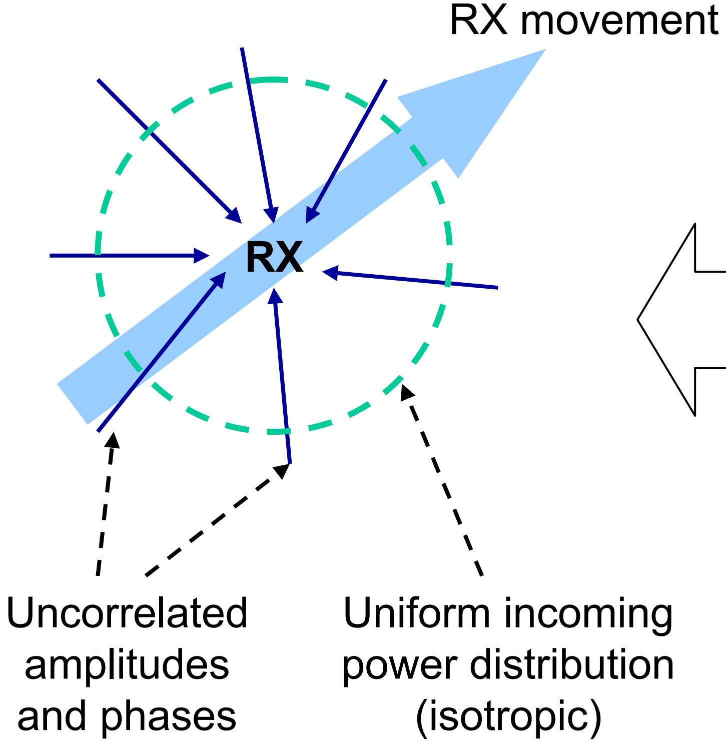
$$S_D(\nu) = \int \rho(\Delta\tau) e^{-j2\pi\nu\Delta\tau} d\Delta\tau$$

$$\propto \frac{1}{\pi\sqrt{\nu_{\max}{}^2 - \nu^2}}$$

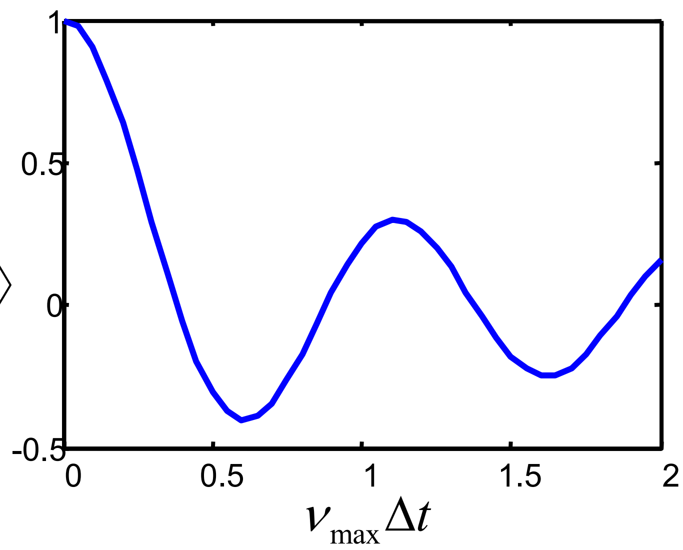
for $-\nu_{\max} < \nu < \nu_{\max}$

Small-scale fading
Doppler spectrum

Isotropic uncorrelated scattering



Time correlation



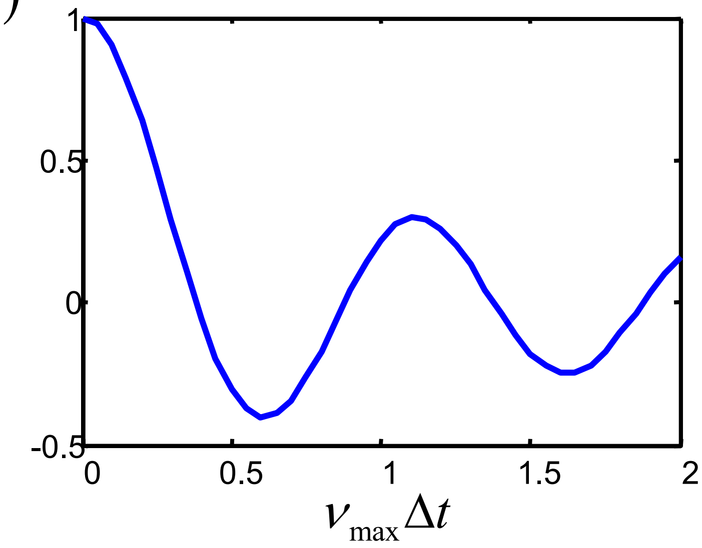
Small-scale fading
Doppler spectrum

- Time correlation – how static is the channel?

$$\rho\left(\Delta t\right) = E\left\{a\left(t\right)a^*\left(t+\Delta t\right)\right\} \propto J_0\left(2\pi\nu_{\max}\Delta t\right)$$

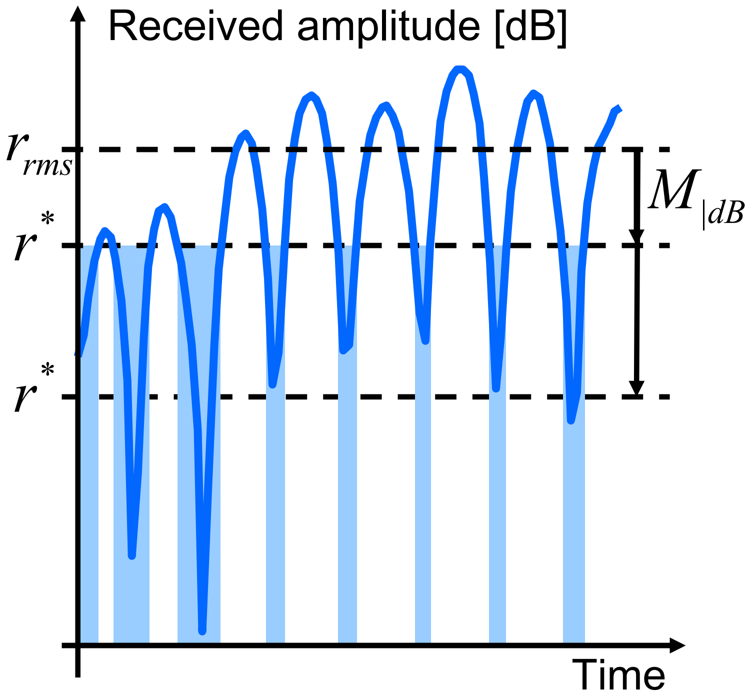
- The time correlation for the amplitude is

$$\rho\left(\Delta t\right) \propto J_0^2\left(2\pi\nu_{\max}\Delta t\right)$$



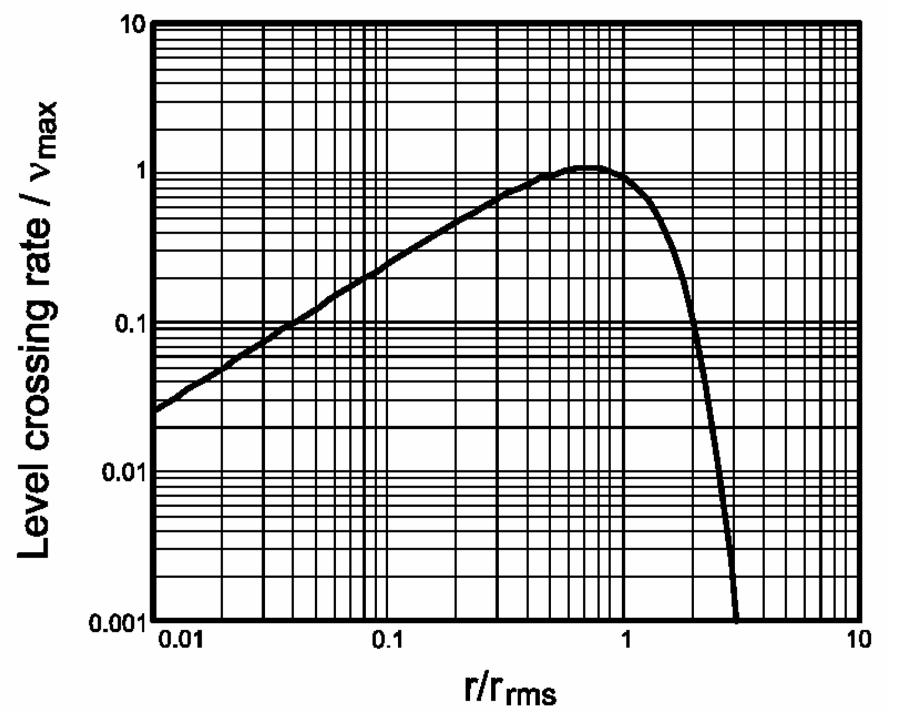
Small-scale fading
Fading dips

What about the length and the frequency of fading dips ?

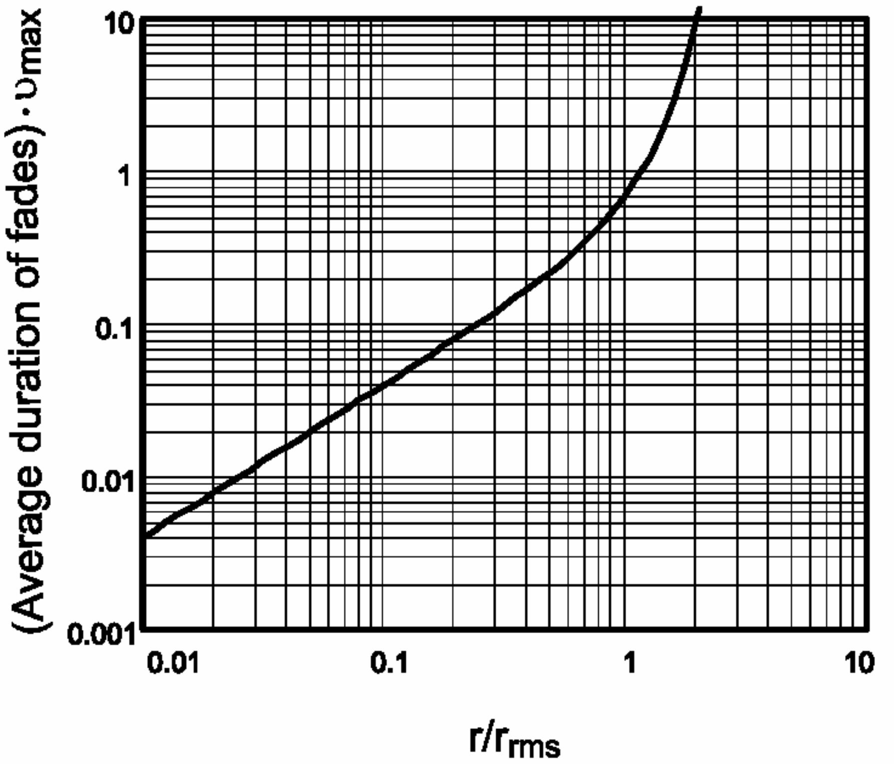


Small-scale fading
Statistics of fading dips

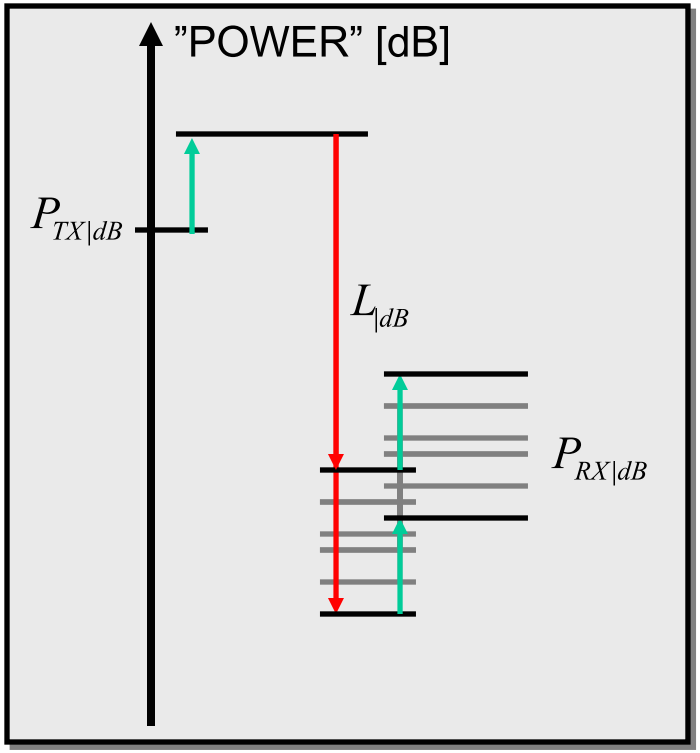
Frequency of the fading dips
(normalized dips/second)

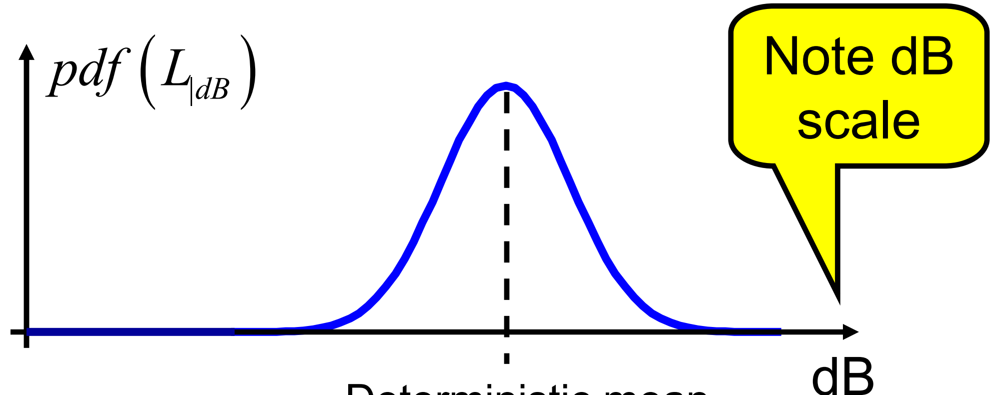


Length of fading dips
(normalized dip-length)



Large-scale fading
Log-normal distribution



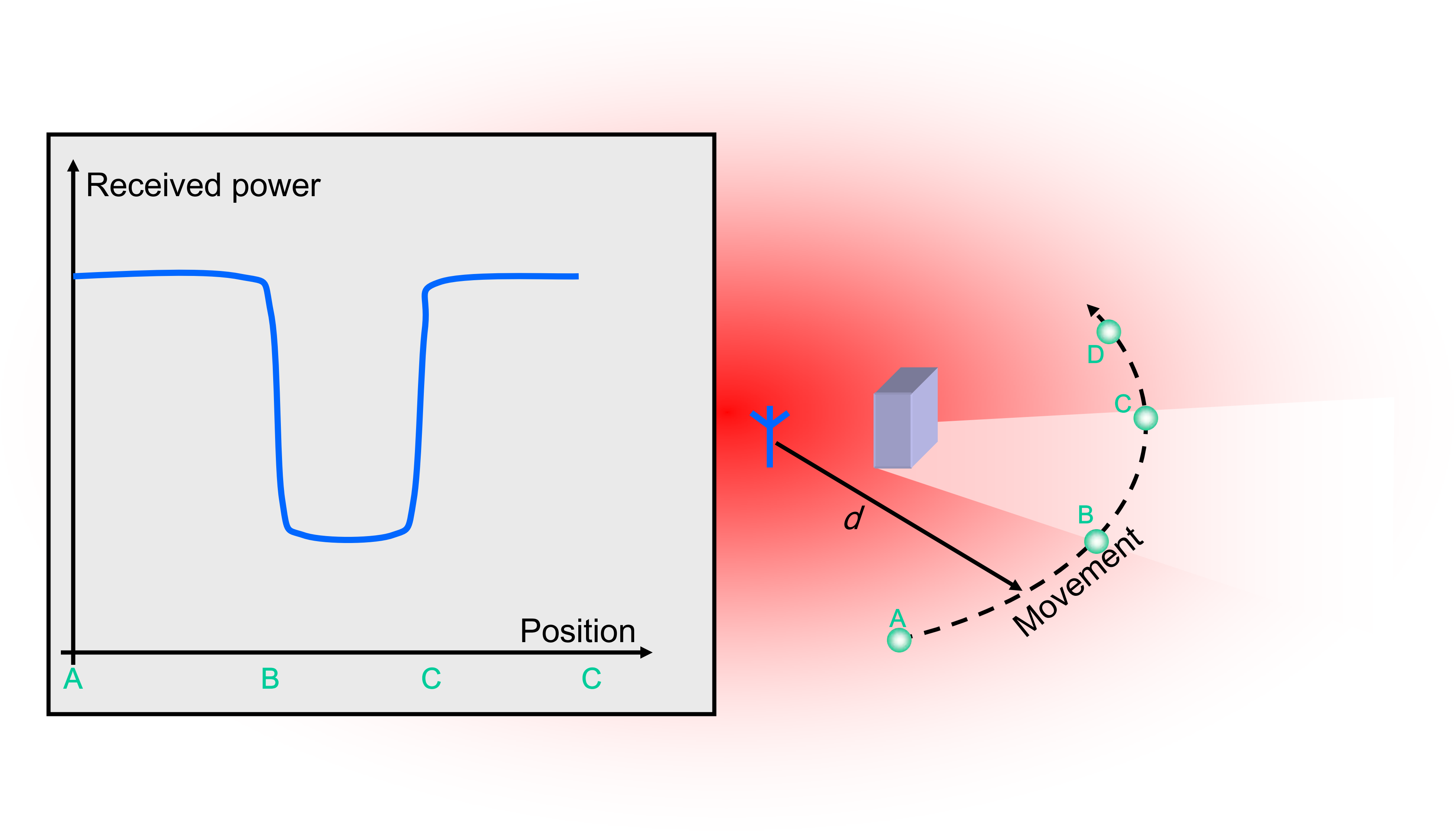


Deterministic mean value of path loss, $L_{0|dB}$

$$pdf\left(L_{|dB}\right) = \frac{1}{\sqrt{2\pi}\sigma_{F|dB}} \exp\left(-\frac{\left(L_{|dB} - L_{0|dB}\right)^2}{2\sigma_{F|dB}^{\;2}}\right)$$

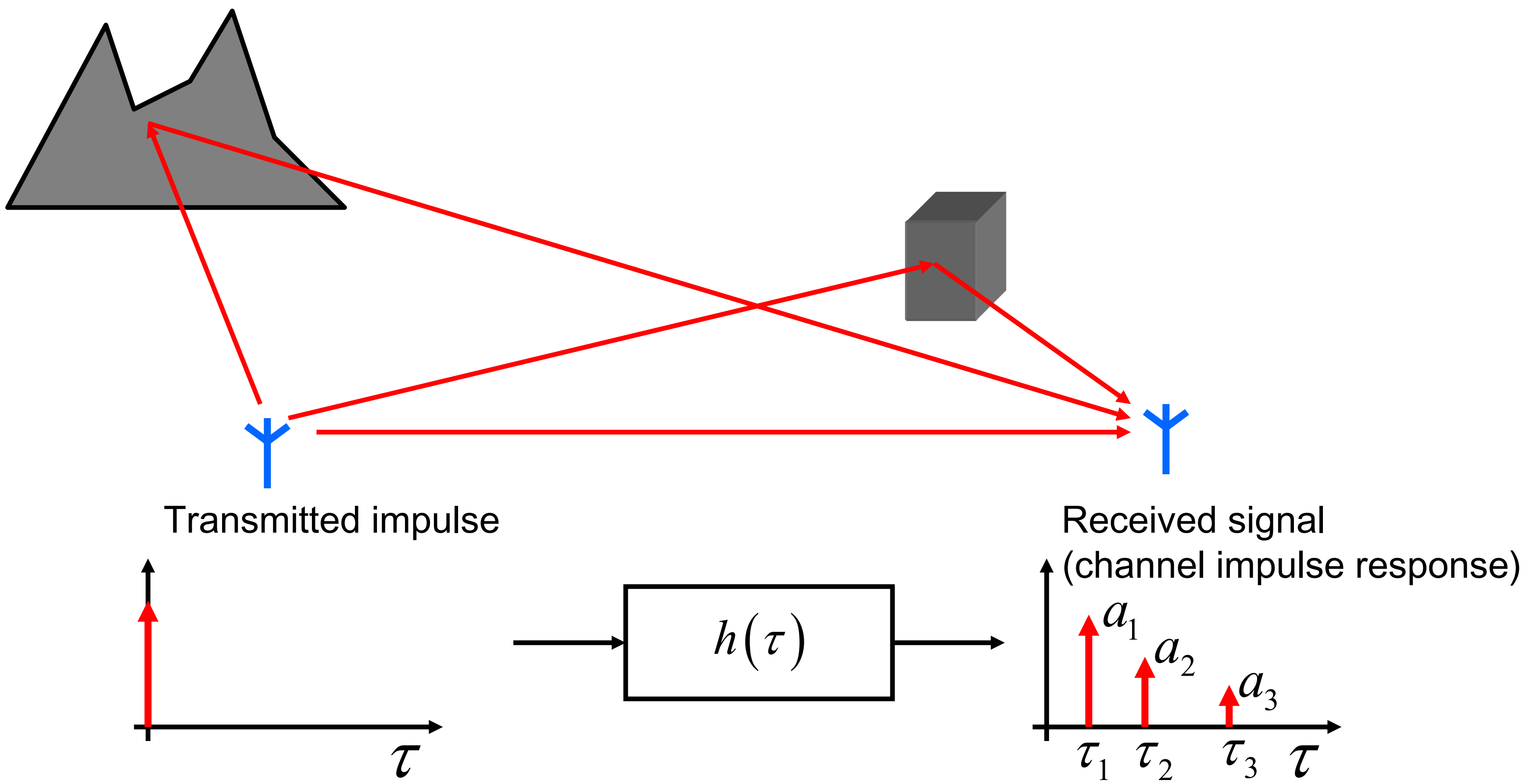
Standard deviation $\sigma_{F|dB} \approx 4\ldots10 \text{ dB}$

Large-scale fading
Basic principle



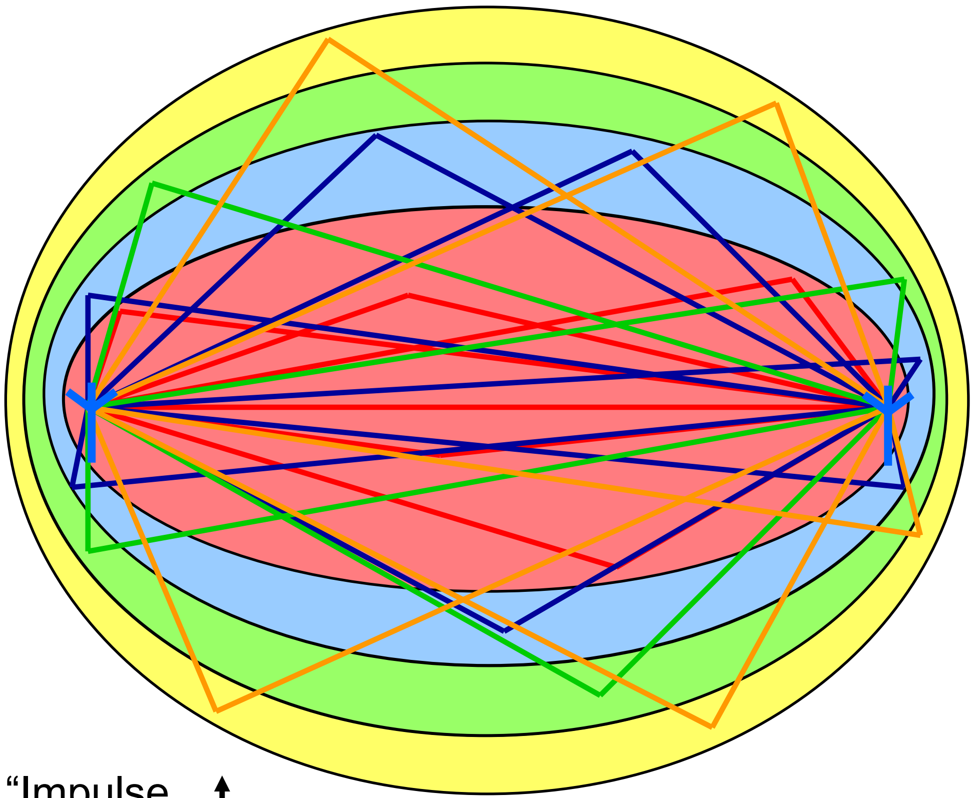
Wideband channels

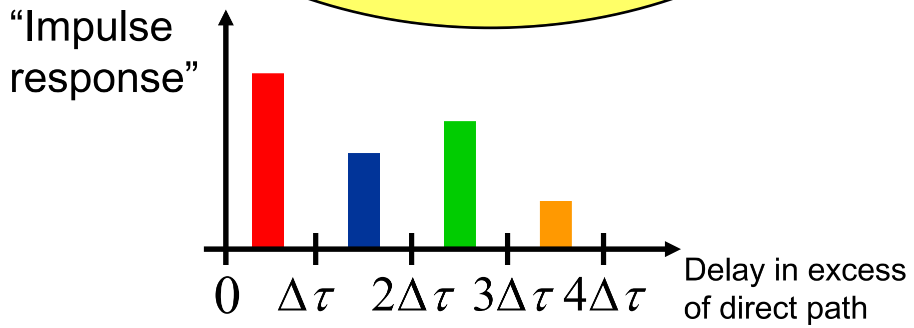
Delay (time) dispersion
A simple case



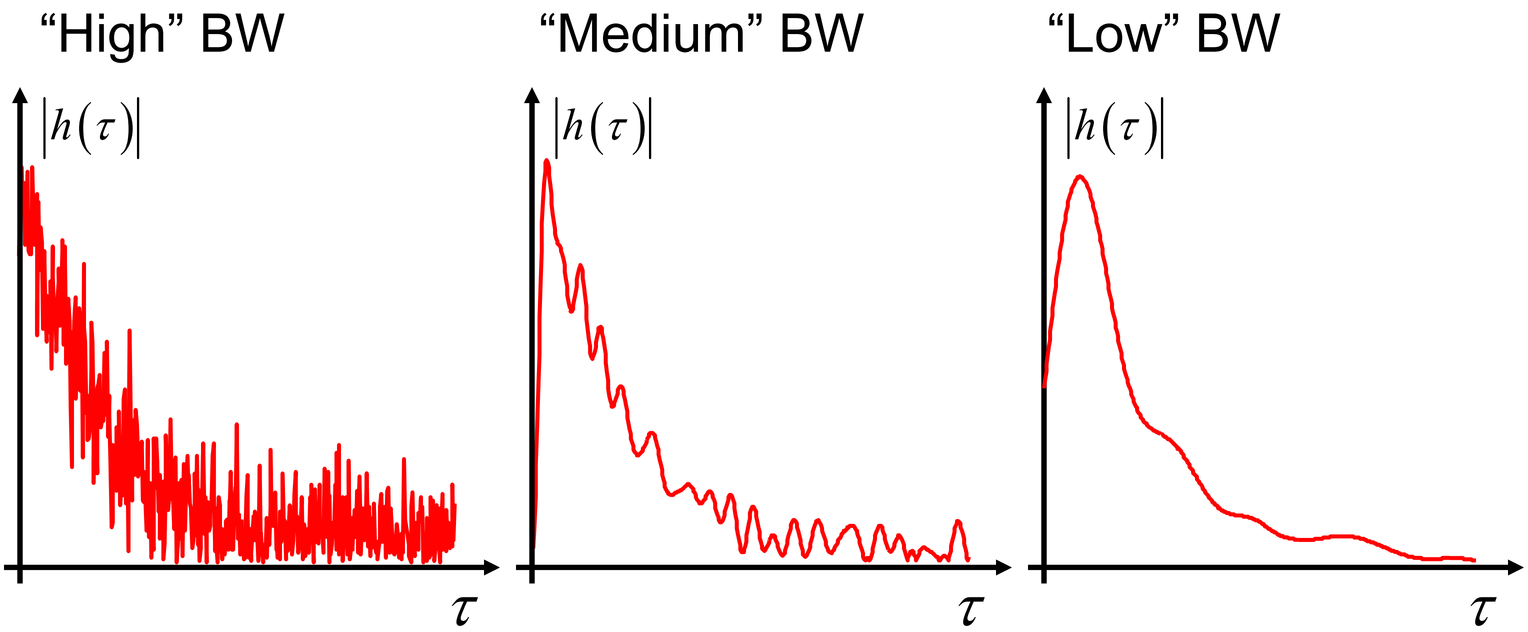
$$h(\tau) = a_1\delta(\tau-\tau_1) + a_2\delta(\tau-\tau_2) + a_3\delta(\tau-\tau_3)$$

Delay (time) dispersion
One reflection/path, many paths

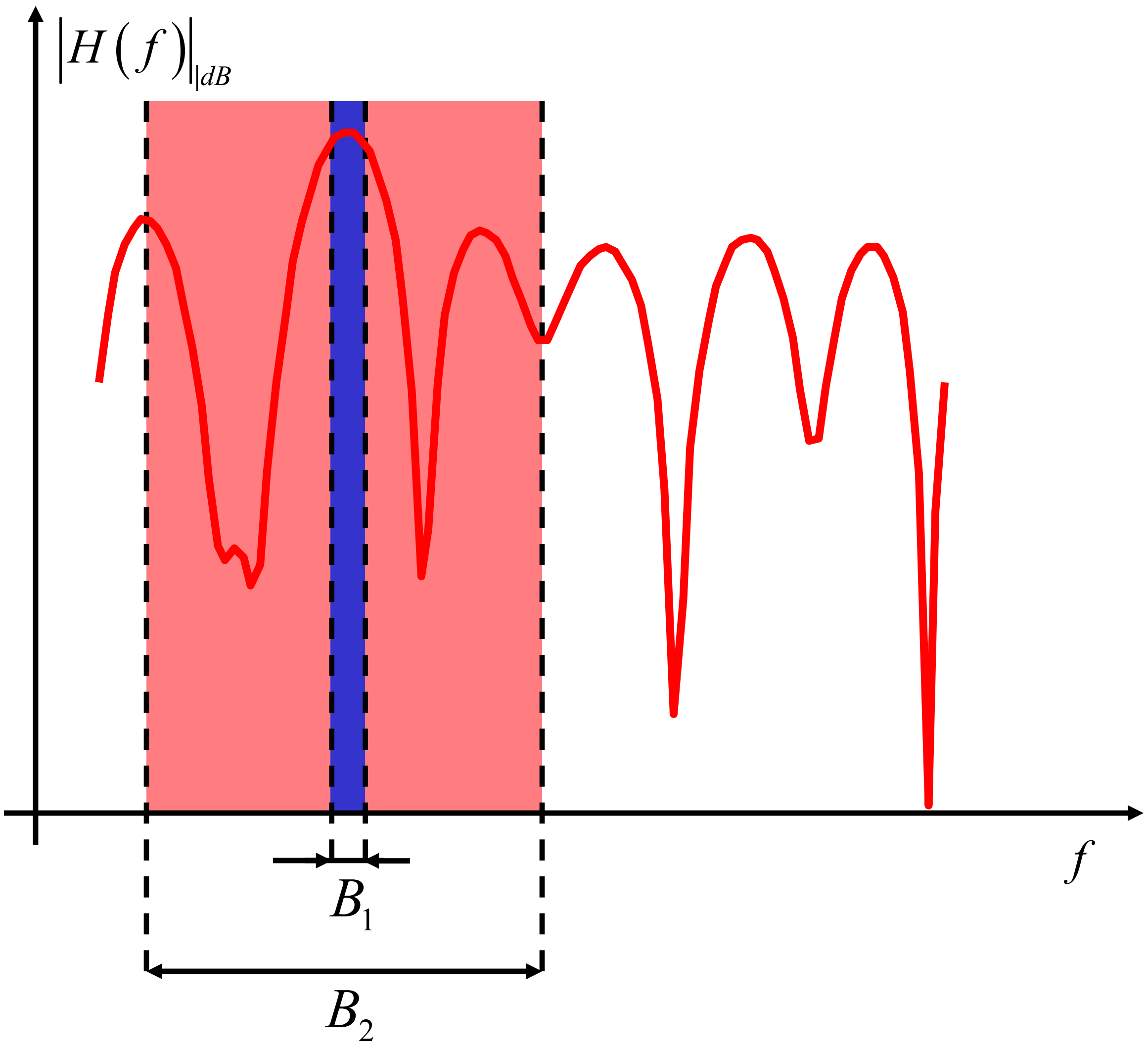




Narrow- versus wide-band
Channel impulse response



Narrow- versus wide-band
Channel frequency response



System functions (1)

- Time-variant impulse response $h(t,\tau)$
  - Due to movement, impulse response changes with time
  - Input-output relationship:

$$y(t) = \int_{-\infty}^{\infty} x(t-\tau)h(t,\tau)d\tau$$

- Time-variant transfer function *H(t,f)*
  - Perform Fourier transformation with respect to τ

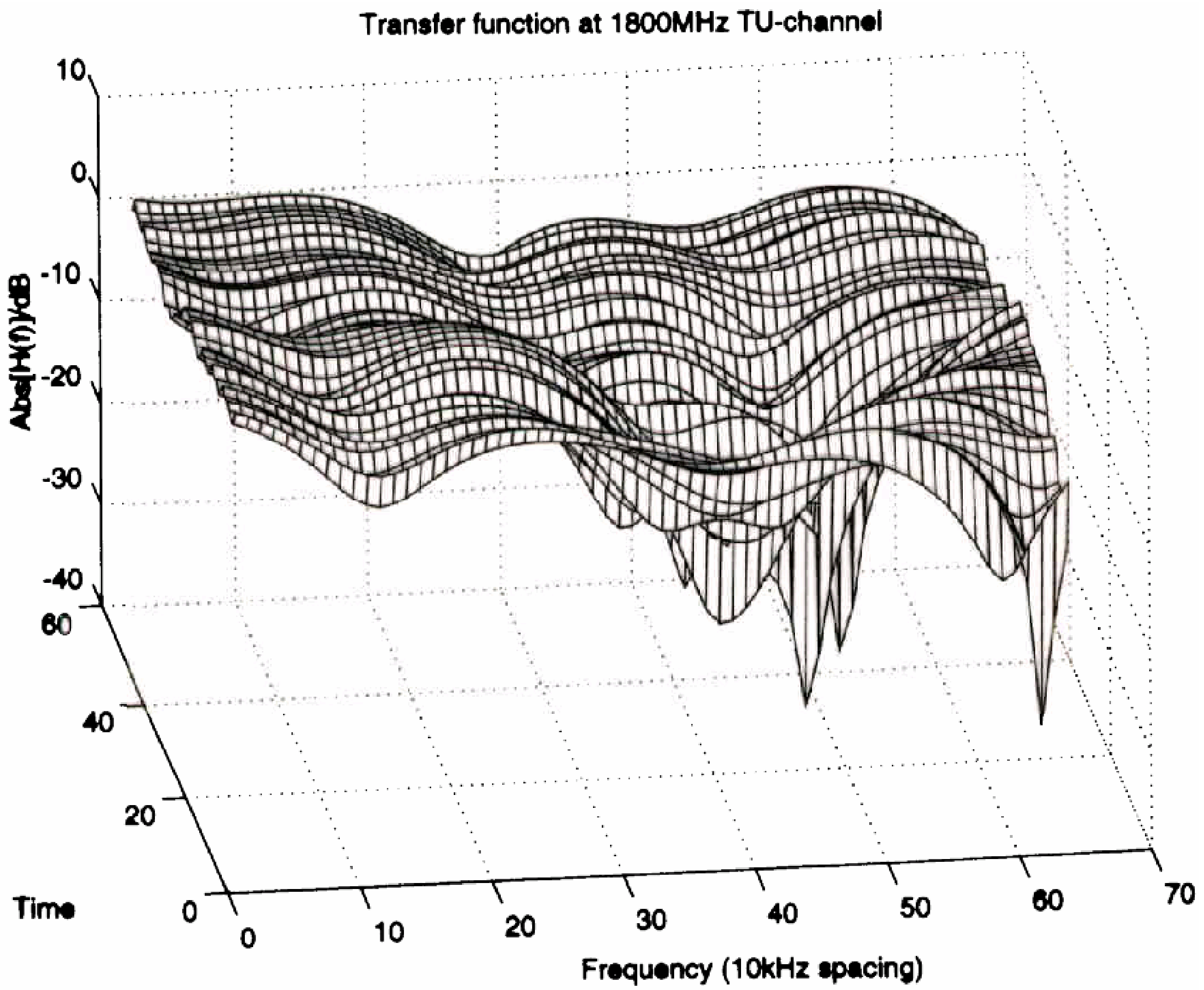
$$H(t,f) = \int_{-\infty}^{\infty} h(t,\tau)\exp(-j2\pi f\tau)d\tau$$

  - Input-output relationship

$$Y(\tilde{f}) = \int_{-\infty}^{\infty}\int_{-\infty}^{\infty} X(f)H(t,f)\exp(j2\pi ft)\exp(-j2\pi\tilde{f}t)dfdt$$

becomes *Y(f)=X(f)H(f)* only in *slowly* time-varying channels

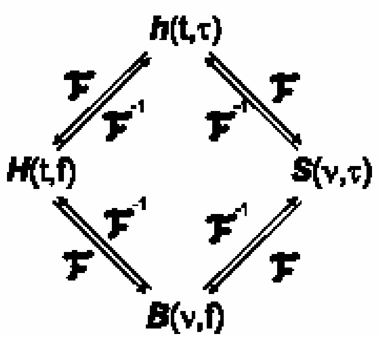
Transfer function, Typical urban



System functions (2)

- Further equivalent system functions:
  – Since impulse response depends on two variables, Fourier transformation can be done w.r.t. each of them

  -> *four* equivalent system descriptions are possible:
    - Impulse response
    - Time-variant transfer function
    - Spreading function $S(\nu,\tau) = \int_{-\infty}^{\infty} h(t,\tau)\exp(-j2\pi\nu t)dt$
    - Doppler-variant spreading function $B(\nu,f) = \int_{-\infty}^{\infty} S(\nu,\tau)\exp(-j2\pi f\tau)d\tau$



Stochastic system functions

- autocorrelation function (second-order statistics)

$$R_h(t,t',\tau,\tau') = E\{h^*(t,\tau)h(t',\tau')\}$$

- Input-output relationship:

$$R_{yy}(t,t') = \int_{-\infty}^{\infty}\int_{-\infty}^{\infty} R_{xx}(t-\tau,t'-\tau')R_h(t,t',\tau,\tau')d\tau d\tau'$$

The WSSUS model: mathematics

- If WSSUS is valid, ACF depends only on two variables (instead of four)
- ACF of impulse response becomes

$$R_h(t, t+\Delta t, \tau, \tau') = \delta(\tau - \tau') P_h(\Delta t, \tau)$$

$$P_h(\Delta t, \tau) \text{....} \textit{Delay cross power spectral density}$$

- ACF of transfer function

$$R_H(t+\Delta t, f+\Delta f) = R_H(\Delta t, \Delta f)$$

- ACF of spreading function

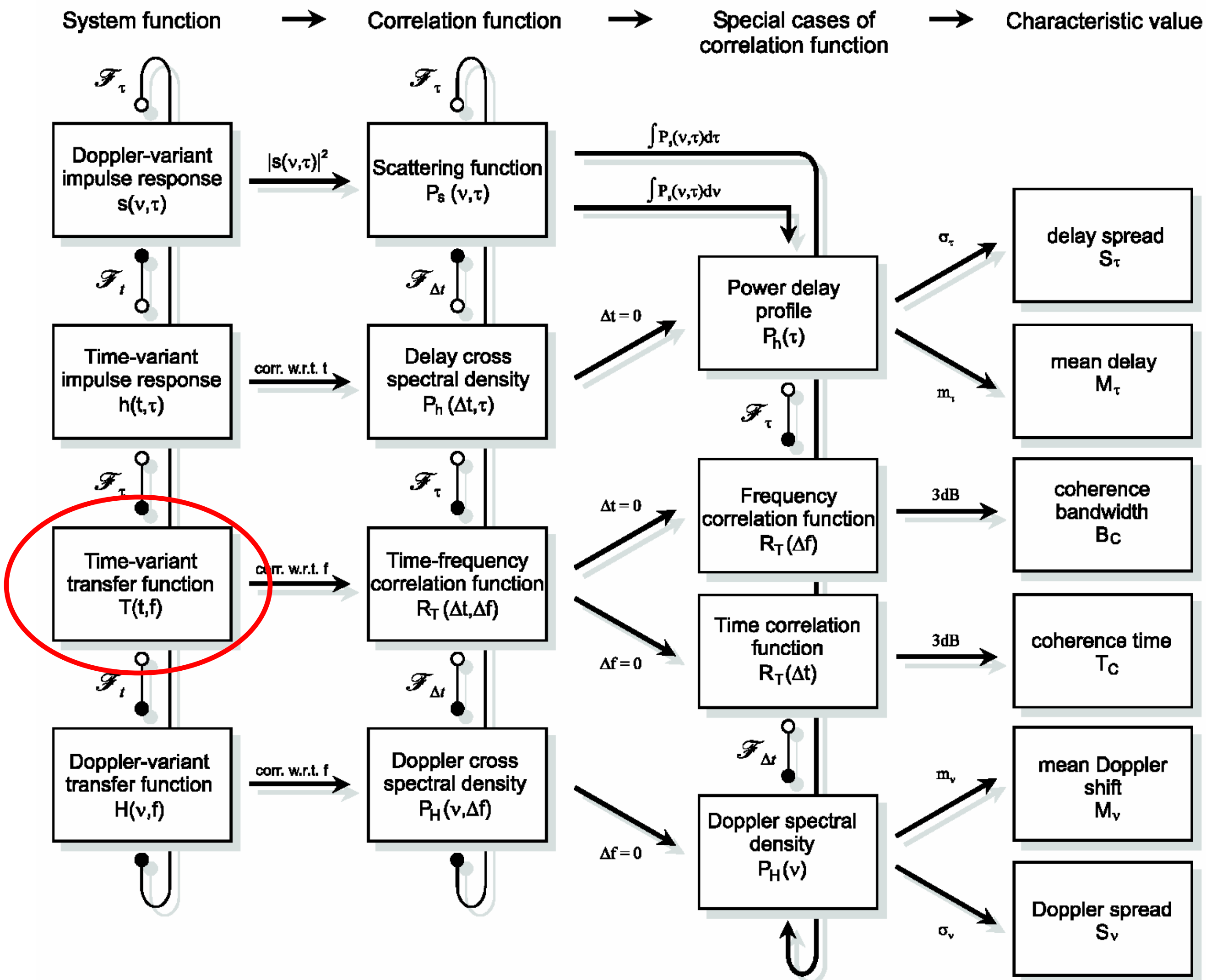
$$R_s(\nu, \nu', \tau, \tau') = \delta(\nu - \nu')\delta(\tau - \tau') P_s(\nu, \tau)$$

$$P_s(\nu, \tau) \text{.......} \textit{Scattering function}$$

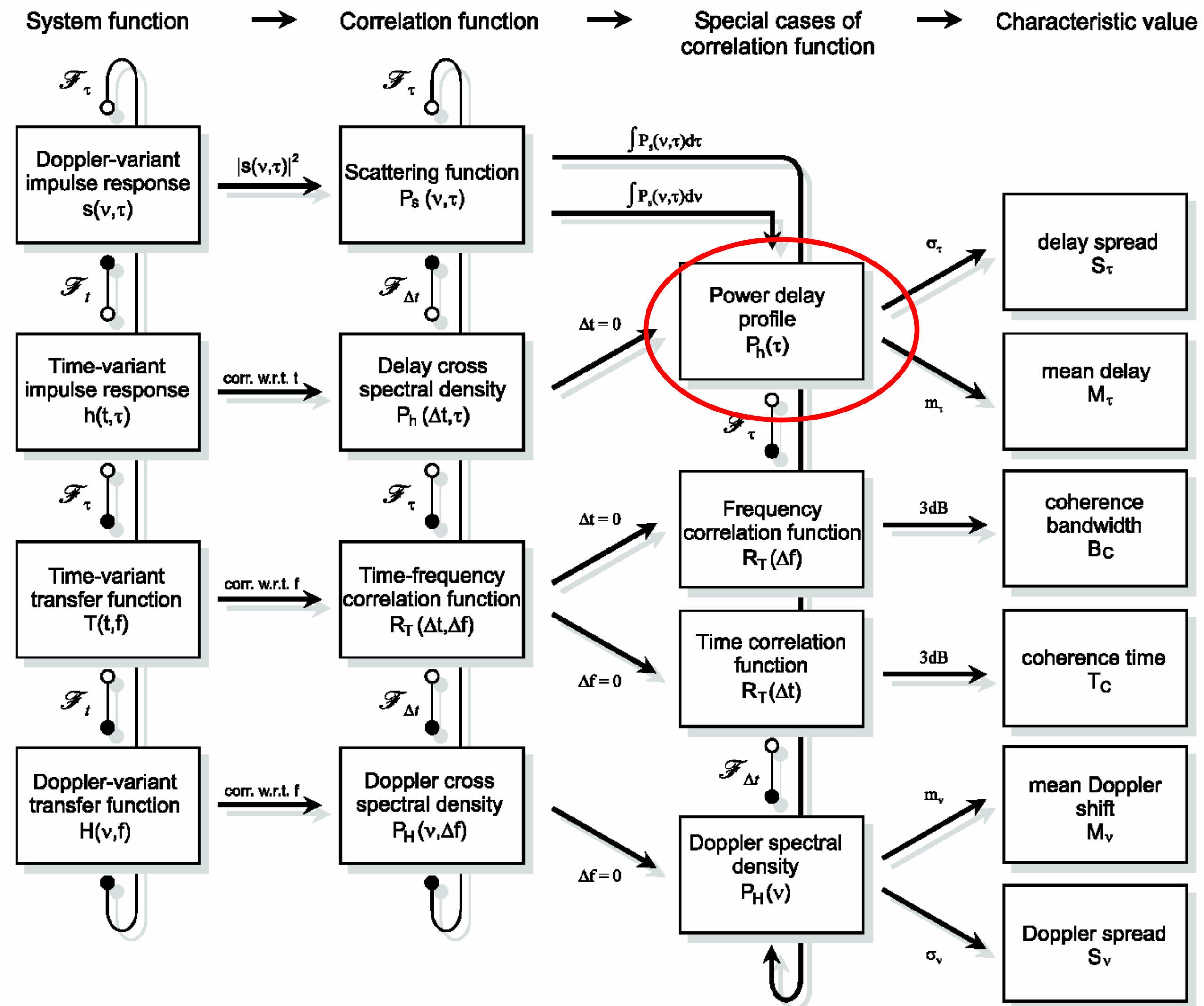
Condensed parameters

- Correlation functions depend on two variables
- For concise characterization of channel, we desire
  - A function depending on one variable or
  - A single (scalar) parameter
- Most common condensed parameters
  - Power delay profile
  - Rms delay spread
  - Coherence bandwidth
  - Doppler spread
  - Coherence time

Channel measures



Channel measures



Condensed parameters
Power-delay profile

One interesting channel property is the power-delay profile (PDP), which is the expected value of the received power at a certain delay:

$$P(\tau) = \mathrm{E}_t\left[\left|h(t,\tau)\right|^2\right]$$

E_t denotes expectation over time.

For our tapped-delay line we get:

$$P(\tau) = \mathrm{E}_t\left[\left|\sum_{i=1}^{N}\alpha_i(t)\exp\left(j\theta_i(t)\right)\delta(\tau-\tau_i)\right|^2\right]$$

$$= \sum_{i=1}^{N}\mathrm{E}_t\left[\alpha_i^2(t)\right]\delta(\tau-\tau_i) = \sum_{i=1}^{N}2\sigma_i^2\delta(\tau-\tau_i)$$

Average power of tap i.

Condensed parameters
Power-delay profile (cont.)

We can “reduce” the PDP into more compact descriptions of the channel:

Total power (time integrated):

$$P_m = \int_{-\infty}^{\infty} P(\tau)\, d\tau$$

Average mean delay:

$$T_m = \frac{\int_{-\infty}^{\infty} \tau P(\tau)\, d\tau}{P_m}$$

Average rms delay spread:

$$S = \sqrt{\frac{\int_{-\infty}^{\infty} \tau^2 P(\tau)\, d\tau}{P_m} - T_m}$$

For our tapped-delay line channel:

$$P_m = \sum_{i=1}^{N} 2\sigma_i^{\,2}$$

$$T_m = \frac{\sum_{i=1}^{N} \tau_i 2\sigma_i^{\,2}}{P_m}$$

$$S = \sqrt{\frac{\sum_{i=1}^{N} \tau_i^{\,2} 2\sigma_i^{\,2}}{P_m} - T_m}$$

Condensed parameters
Frequency correlation

A property closely related to the power-delay profile (PDP) is the frequency correlation of the channel. It is in fact the Fourier transform of the PDP:

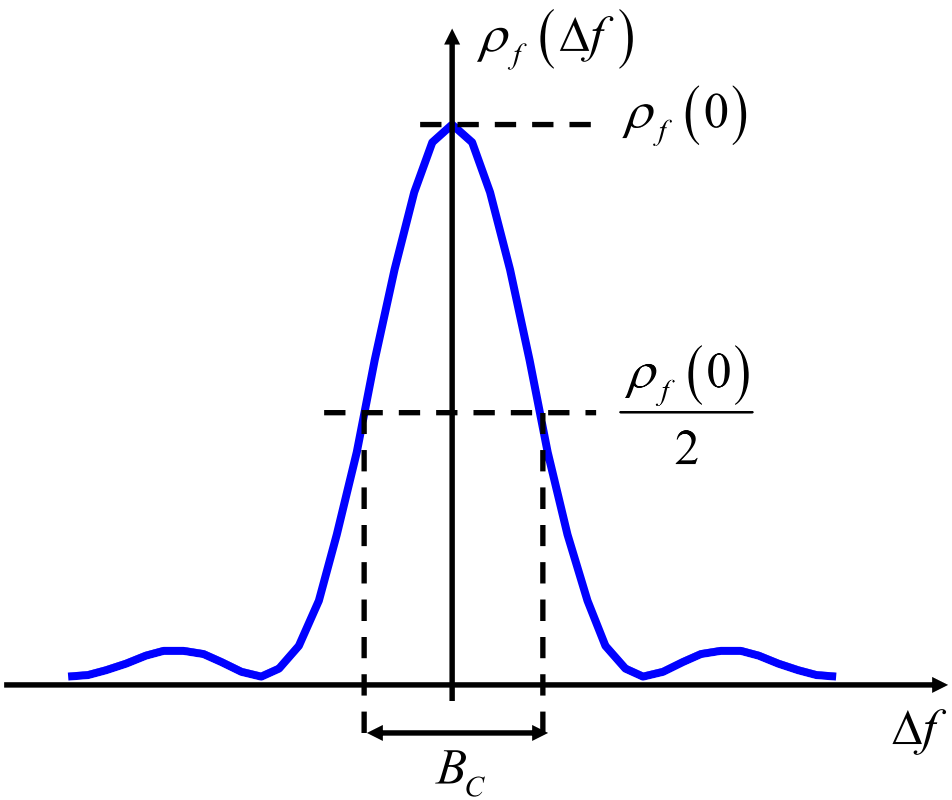
$$\rho_f\left(\Delta f\right) = \int_{-\infty}^{\infty} P\left(\tau\right) \exp\left(-j2\pi\Delta f\,\tau\right) d\tau$$

For our tapped delay-line channel we get:

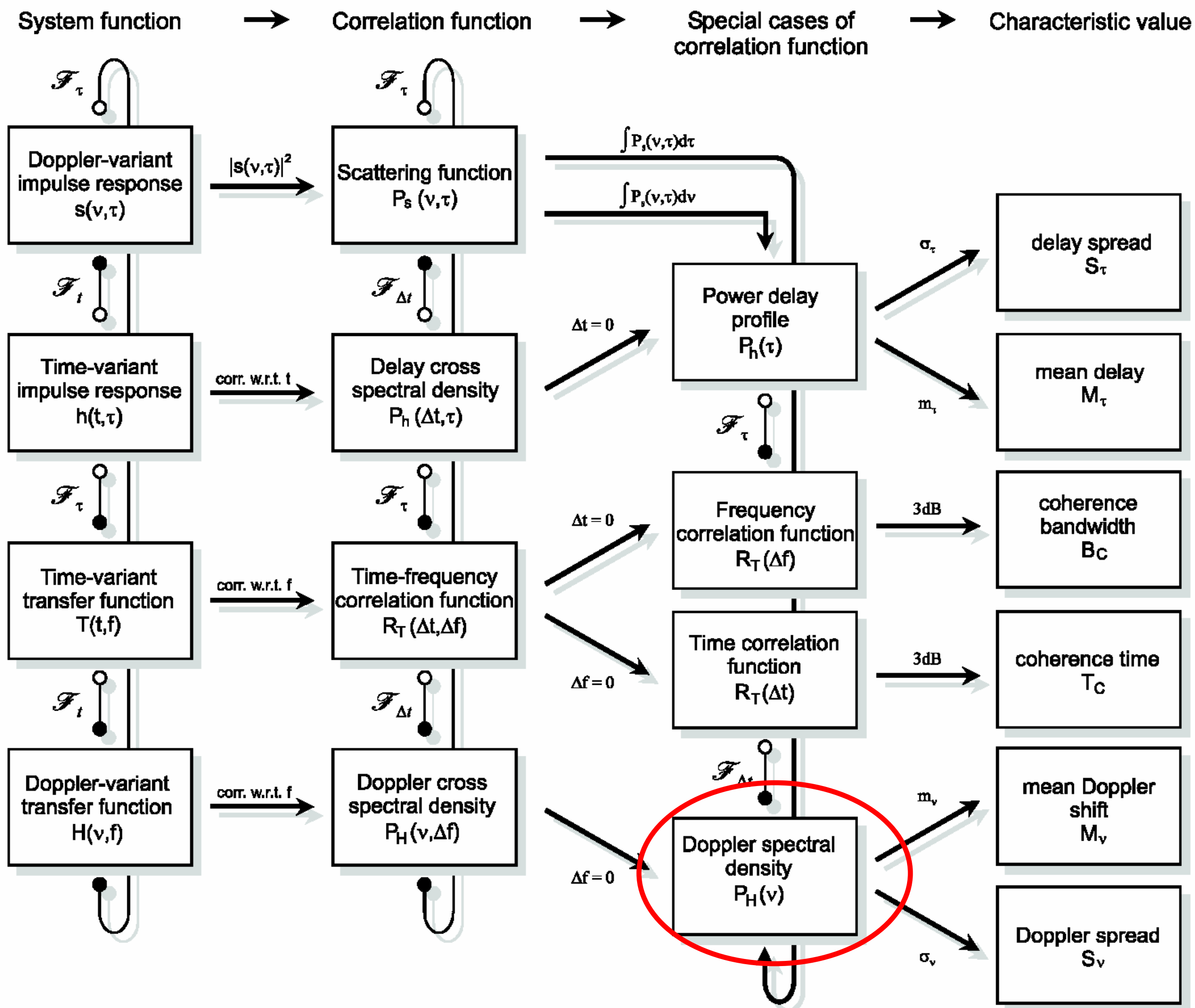
$$\rho_f\left(\Delta f\right) = \int_{-\infty}^{\infty} \left(\sum_{i=1}^{N} 2\sigma_i^{\,2} \delta\left(\tau - \tau_i\right) \right) \exp\left(-j2\pi\Delta f\,\tau\right) d\tau$$

$$= \sum_{i=1}^{N} 2\sigma_i^{\,2} \exp\left(-j2\pi\Delta f\,\tau_i\right)$$

Condensed parameters
Coherence bandwidth



Channel measures



Condensed parameters
The Doppler spectrum

Given the scattering function P_s (doppler spectrum as function of delay) we can calculate a total Doppler spectrum of the channel as:

$$P_B(\nu) = \int P_S(\nu, \tau)\, d\tau$$

For our tapped delay-line channel, we have:

Doppler spectrum of tap *i*.

$$P_S(\nu, \tau) = \frac{2\sigma_i^{\ 2}}{\pi\sqrt{\nu_{i,\max}^{\ \ 2} - \nu^2}} \delta(\tau - \tau_i)$$

$$P_B(\nu) = \int_{-\infty}^{\infty} \frac{2\sigma_i^{\ 2}}{\pi\sqrt{\nu_{i,\max}^{\ \ 2} - \nu^2}} \delta(\tau - \tau_i)\, d\tau$$

$$= \sum_{i=1}^{N} \frac{2\sigma_i^{\ 2}}{\pi\sqrt{\nu_{i,\max}^{\ \ 2} - \nu^2}}$$

Condensed parameters
The Doppler spectrum (cont.)

We can "reduce" the Doppler spectrum into more compact descriptions of the channel:

Total power (frequency integrated):

$$P_{B,m} = \int_{-\infty}^{\infty} P_B(\nu)\, d\nu$$

Average mean Doppler shift:

$$T_{B,m} = \frac{\int_{-\infty}^{\infty} \nu P_B(\nu)\, d\nu}{P_{B,m}}$$

Average rms Doppler spread:

$$S_B = \sqrt{\frac{\int_{-\infty}^{\infty} \nu^2 P(\nu)\, d\nu}{P_{B,m}} - T_{B,m}}$$

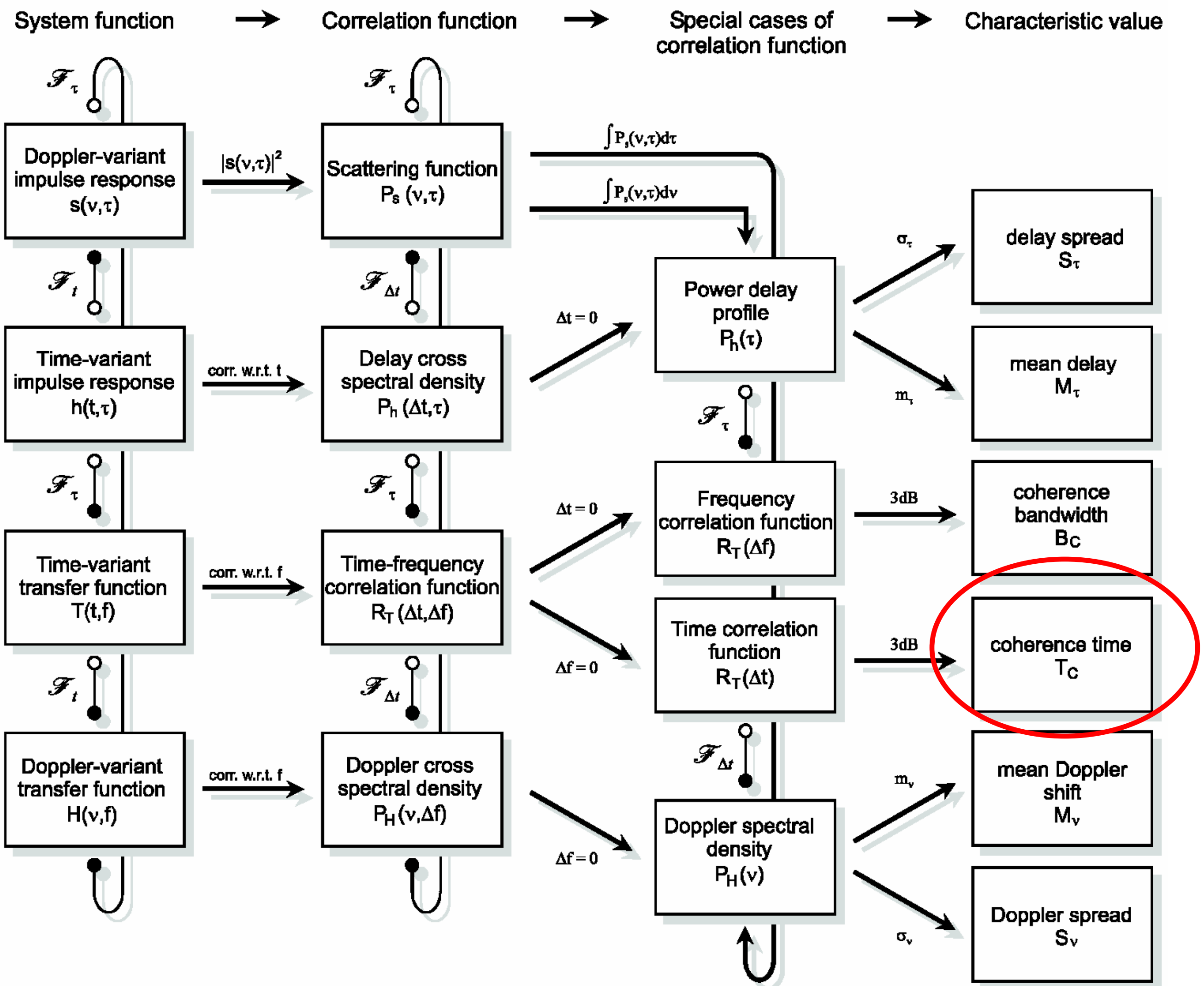
For our tapped-delay line channel:

$$P_{B,m} = \sum_{i=1}^{N} 2\sigma_i^{\ 2}$$

$$T_{B,m} = 0$$

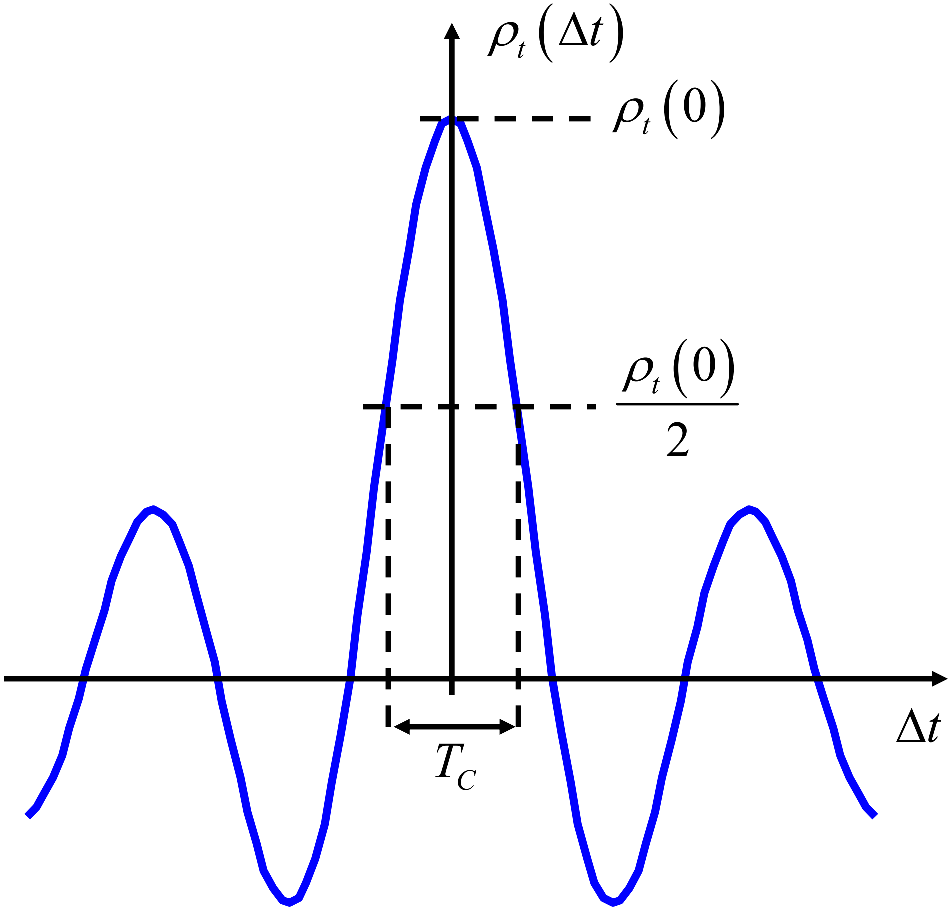
$$S_B = \sqrt{\frac{\sum_{i=1}^{N} \sigma_i^{\ 2} \nu_{i,\max}^{\ \ 2}}{P_{B,m}}}$$

Channel measures



Condensed parameters
Coherence time

Given the time correlation of a channel, we can define the coherence time T_C:



Condensed parameters
The time correlation

A property closely related to the Doppler spectrun is the time correlation of the channel. It is in fact the inverse Fourier transform of the Doppler spectrum:

$$\rho_t(\Delta t) = \int_{-\infty}^{\infty} P_B(\nu) \exp(j2\pi\nu\Delta t)\, d\nu$$

For our tapped-delay line channel we get

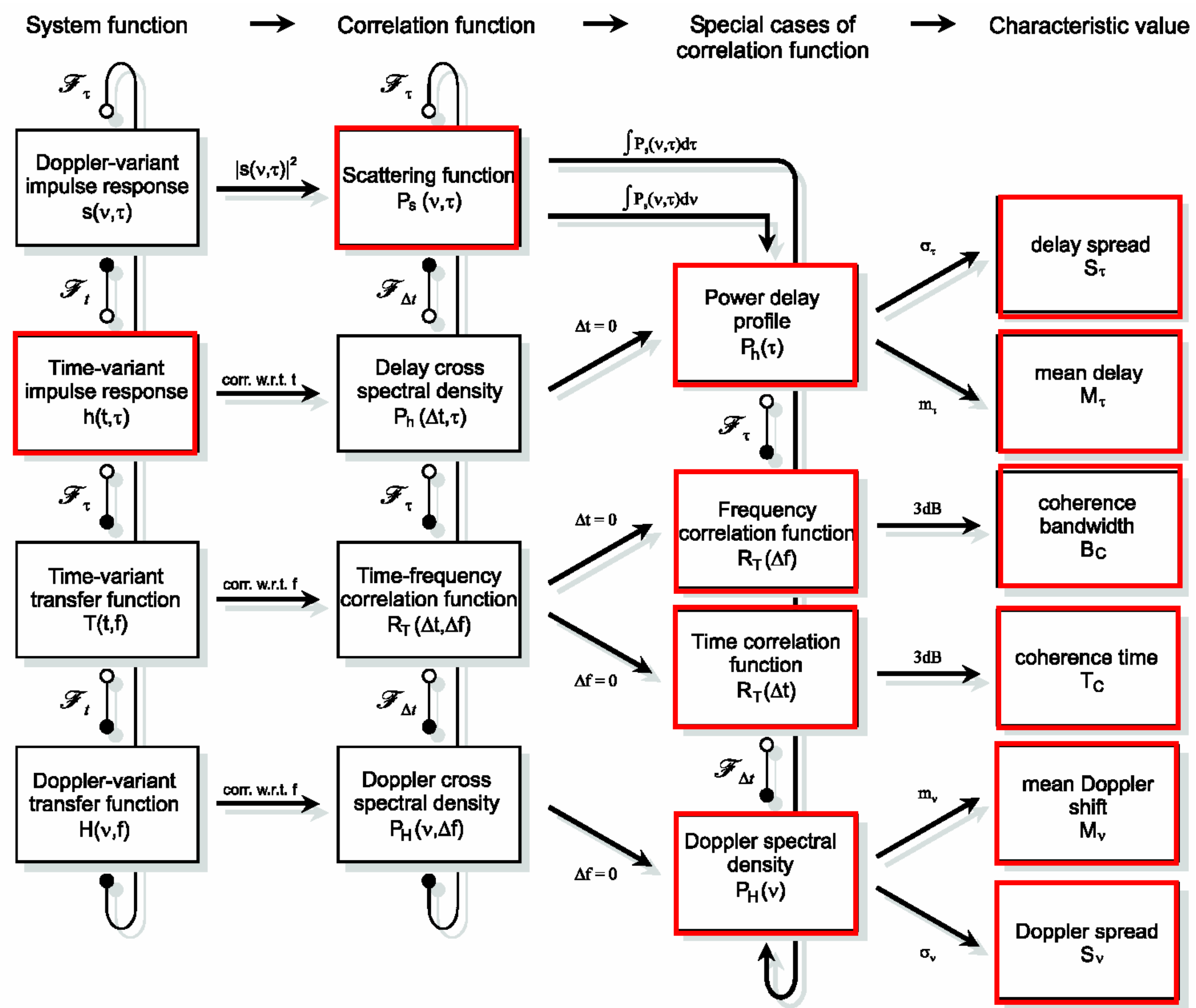
$$\rho_t(\Delta t) = \int_{-\infty}^{\infty} \sum_{i=1}^{N} \frac{2\sigma_i^2}{\pi\sqrt{\nu_{i,\max}^2 - \nu^2}} \exp(j2\pi\nu\Delta t)\, d\nu$$

$$= \sum_{i=1}^{N} \int_{-\infty}^{\infty} \frac{2\sigma_i^2}{\pi\sqrt{\nu_{i,\max}^2 - \nu^2}} \exp(j2\pi\nu\Delta t)\, d\nu$$

$$= \sum_{i=1}^{N} 2\sigma_i^2 J_0\left(2\pi\nu_{i,\max}\Delta t\right)$$

Sum of time correlations for each tap.

It's much more complicated than what we have discussed!



Double directional impulse response

TX position

RX position

number of multipath components for these positions

$$h(t, \vec{r}_{\mathrm{TX}}, \vec{r}_{\mathrm{RX}}, \tau, \Omega, \Psi) = \sum_{\ell=1}^{N(\vec{r})} h_\ell(t, \vec{r}_{\mathrm{TX}}, \vec{r}_{\mathrm{RX}}, \tau, \Omega, \Psi)$$

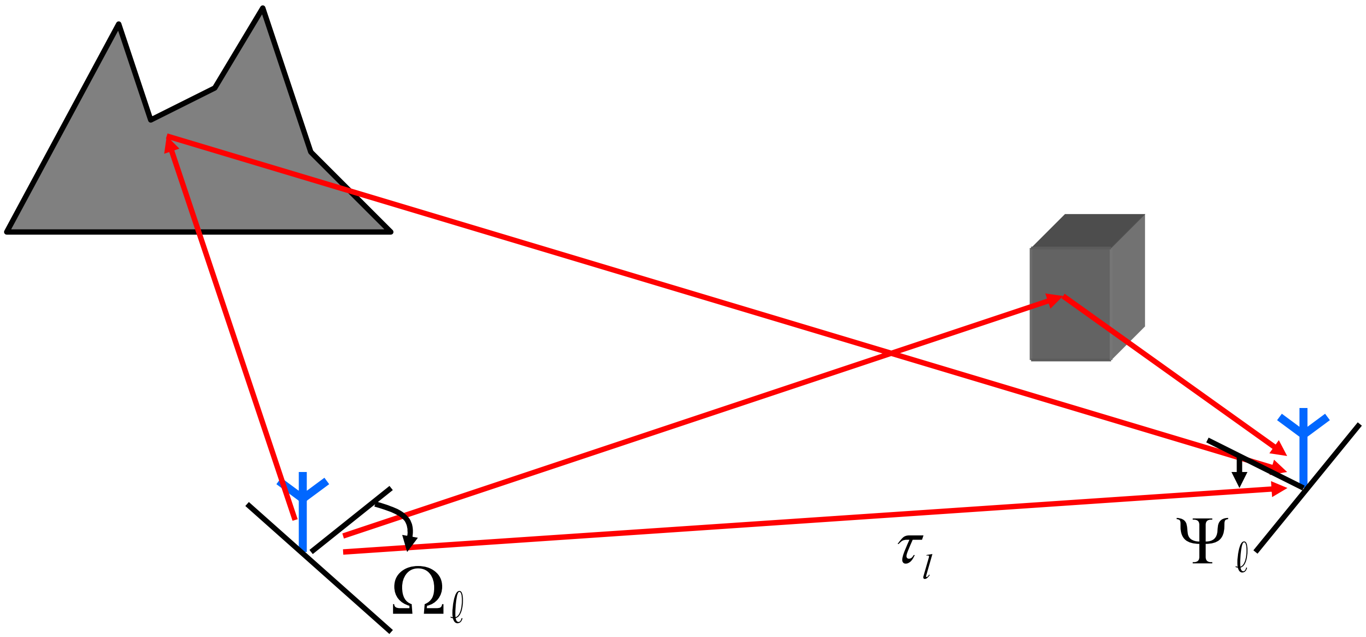
direction-of-arrival

delay

direction-of-departure

$$h_\ell(t, \vec{r}_{\mathrm{TX}}, \vec{r}_{\mathrm{RX}}, \tau, \Omega, \Psi) = |a_\ell| e^{j\varphi_\ell} \delta(\tau - \tau_\ell) \delta(\Omega - \Omega_\ell) \delta(\Psi - \Psi_\ell)$$

Physical interpretation



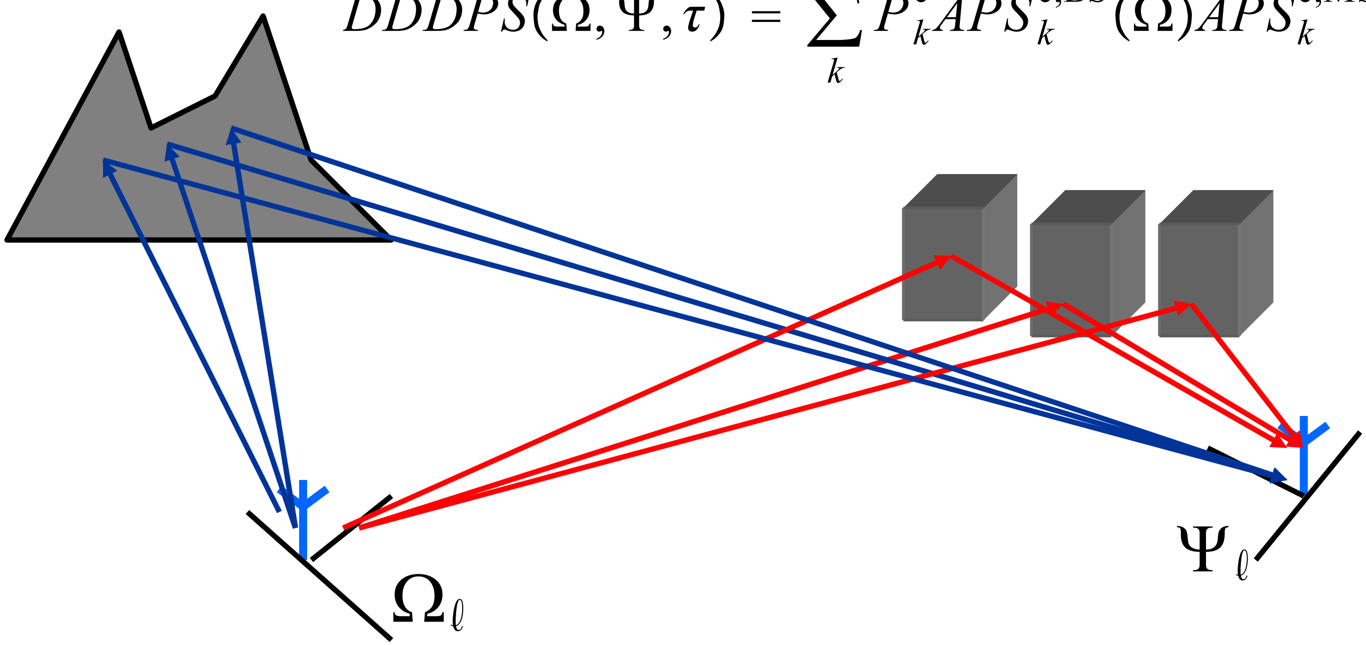
Directional models

- The double directional delay power spectrum is sometimes factorized w.r.t. DoD, DoA and delay.

$$DDDPS(\Omega,\Psi,\tau) = APS^{\mathrm{BS}}(\Omega)APS^{\mathrm{MS}}(\Psi)PDP(\tau)$$

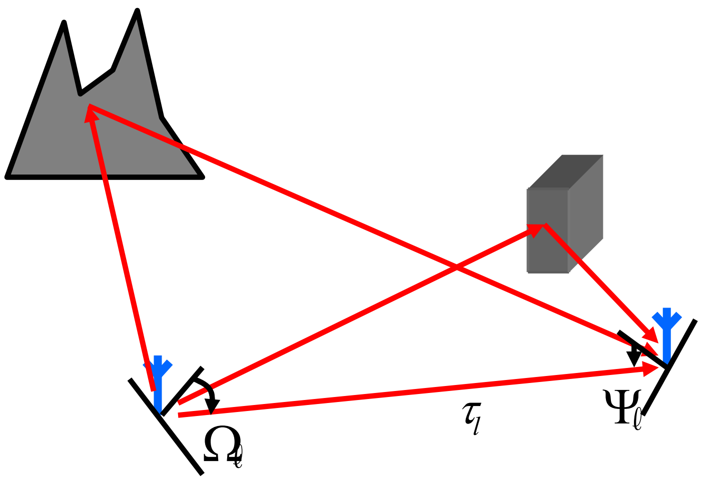
- Often in reality there are groups of scatterers with similar DoD and DoA – clusters

$$DDDPS(\Omega,\Psi,\tau) = \sum_k P_k^{\mathrm{c}} APS_k^{\mathrm{c,BS}}(\Omega)APS_k^{\mathrm{c,MS}}(\Psi)PDP_k^{\mathrm{c}}(\tau)$$



Angular spread

$$E\{s^*(\Omega,\Psi,\tau,\nu)s(\Omega',\Psi',\tau',\nu')\} = P_s(\Omega,\Psi,\tau,\nu)\delta(\Omega-\Omega')\delta(\Psi-\Psi')\delta(\tau-\tau')\delta(\nu-\nu')$$



double directional delay power spectrum

$$DDDPS(\Omega,\Psi,\tau) = \int \mathrm{P_s}(\Psi,\Omega,\tau,\nu)\mathrm{d}\nu$$

angular delay power spectrum

$$ADPS(\Omega,\tau) = \int DDDPS(\Psi,\Omega,\tau)G_{\mathrm{MS}}(\Psi)d\Psi$$

angular power spectrum

$$APS(\Omega) = \int \mathrm{APDS}(\Omega,\tau)\mathrm{d}\tau$$

power

$$P = \int \mathrm{APS}(\Omega)\mathrm{d}\Omega$$

Chapter 7

Channel modeling

Modelling methods

- Stored channel impulse responses
  - realistic
  - reproducible
  - hard to cover all scenarios
- Deterministic channel models
  - based on Maxwell's equations
  - site specific
  - computationally demanding
- Stochastic channel models
  - describes the distribution of the field strength etc
  - mainly used for design and system comparisons

Narrowband models
Review of properties

Narrowband models contain ”only one” attenuation, which is modeled as a propagation loss, plus large- and small-scale fading.

Path loss: Often proportional to $1/d^n$, where n is the propagation exponent. (n may be different at different distances)

Large-scale fading: Log-normal distribution (normal distr. in dB scale)

Small-scale fading: Rayleig, Rice, Nakagami distributions ... (not in dB-scale)

Okumura's measurements

Extensive measurement campaign in Japan in the 1960's.

Parameters varied during measurements:

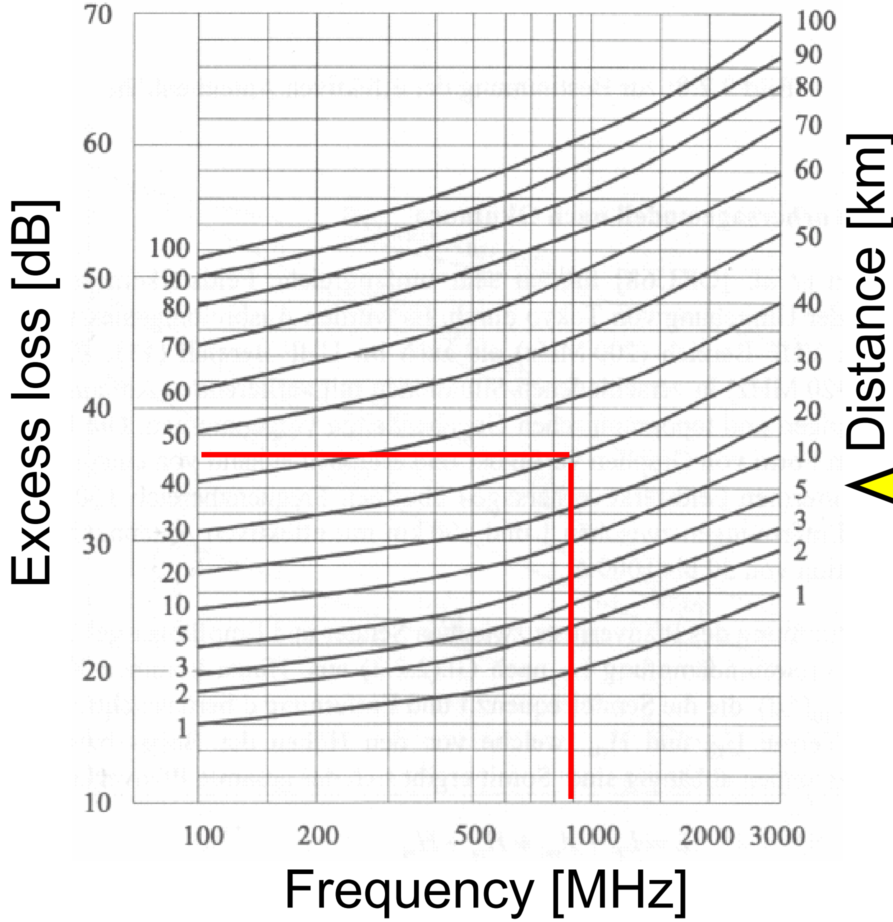
Frequency	100 – 3000 MHz
Distance	1 – 100 km
Mobile station height	1 – 10 m
Base station height	20 – 1000 m
Environment	medium-size city, large city, etc.

Propagation loss is given as median values (50% of the time and 50% of the area).

Okumura's measurements excess loss

Example

FIGURE 7.12



These curves are only for h_b=200 m and h_m=3 m

900 MHz and
30 km distance

From [Okumura et al.]

The Okumura-Hata model
How to calculate prop. loss

$$L_{O-H} = A + B\log\left(d_{|km}\right) + C$$

$$A = 69.55 + 26.16\log\left(f_{0|MHz}\right) - 13.82\log\left(h_b\right) - a\left(h_m\right)$$

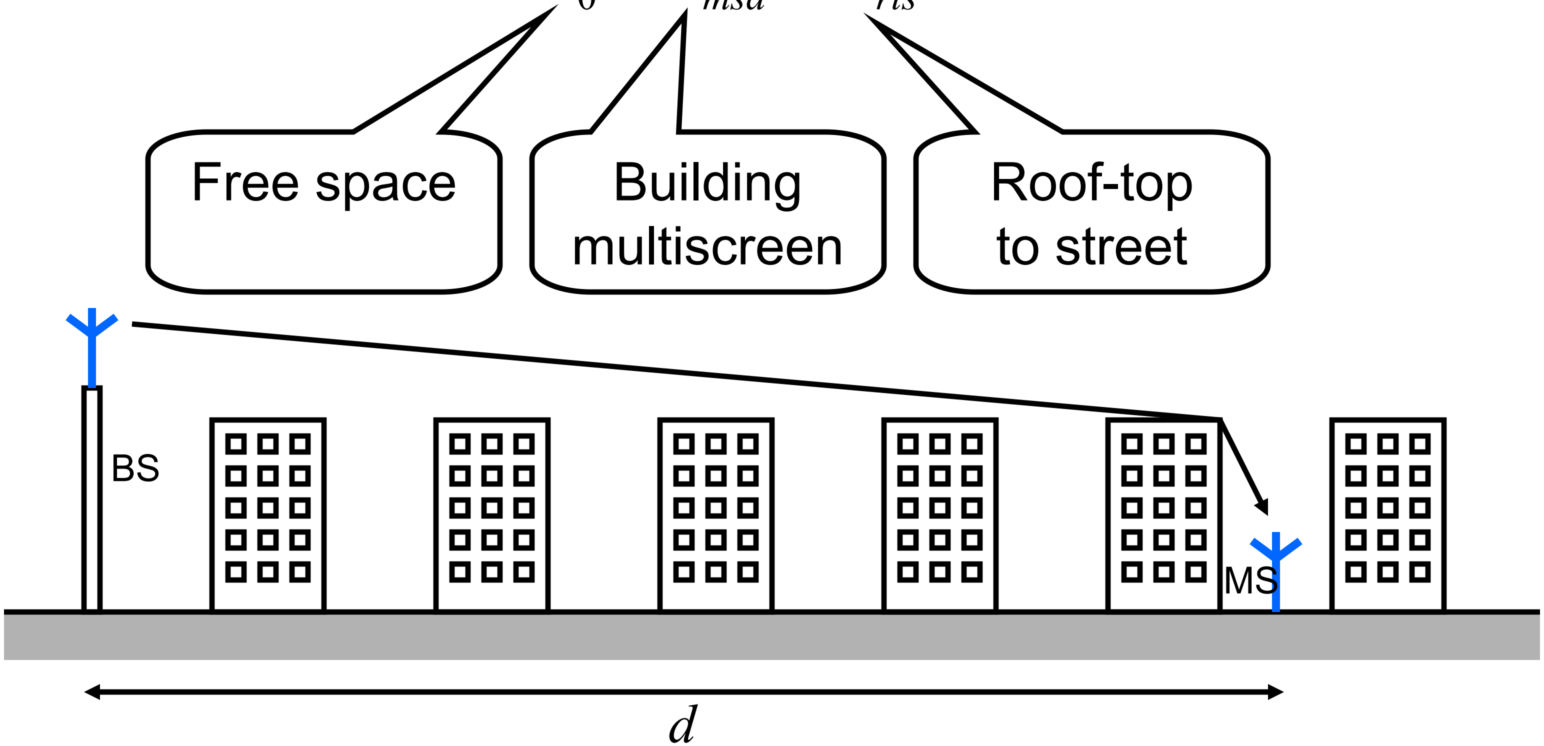
$$B = 44.9 - 6.55\log\left(h_b\right)$$

h_b and h_m in meter

	$a\left(h_m\right) =$	$C =$
Metropolitan areas	$8.29\left(\log\left(1.54h_m\right)\right)^2 - 1.1$ for $f_0 \leq 200$ MHz $3.2\left(\log\left(11.75h_m\right)\right)^2 - 4.97$ for $f_0 \geq 400$ MHz	0
Small/medium-size cities	$\left(1.1\log\left(f_{0\|MHz}\right) - 0.7\right)h_m - \left(1.56\log\left(f_{0\|MHz}\right) - 0.8\right)$	0
Suburban environments	$\left(1.1\log\left(f_{0\|MHz}\right) - 0.7\right)h_m - \left(1.56\log\left(f_{0\|MHz}\right) - 0.8\right)$	$-2\left[\log\left(f_{0\|MHz}/28\right)\right]^2 - 5.4$
Rural areas	$\left(1.1\log\left(f_{0\|MHz}\right) - 0.7\right)h_m - \left(1.56\log\left(f_{0\|MHz}\right) - 0.8\right)$	$-4.78\left[\log\left(f_{0\|MHz}\right)\right]^2 + 18.33\log\left(f_{0\|MHz}\right) - 40.94$

The COST 231-Walfish-Ikegami model
How to calculate prop. loss

$$L = L_0 + L_{msd} + L_{rts}$$



Details about calculations can be found in the textbook, Section 7.6.2.

Motley-Keenan indoor model

For indoor environments, the attenuation is heavily affected by the building structure, walls and floors play an important rule

$$PL = PL_0 + 10n\log(d/d_0) + F_{\mathrm{wall}} + F_{\mathrm{floor}}$$

distance dependent path loss

sum of attenuations from walls, 1-20 dB/wall

sum of attenuation from the floors (often larger than wall attenuation)

site specific, since it is valid for a particular case

Wideband models

- Tapped delay line model often used

$$h(t,\tau) = \sum_{i=1}^{N} \alpha_i(t) \exp\left(j\theta_i(t)\right) \delta(\tau - \tau_i)$$

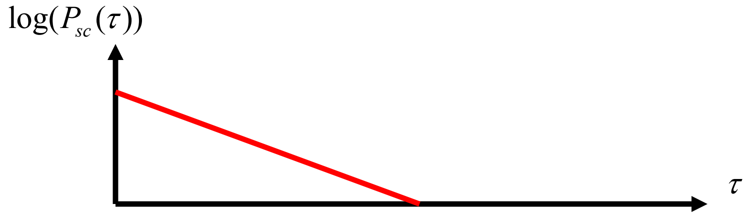
- Often Rayleigh-distributed taps, but might include LOS and different distributions of the tap values
- Mean tap power determined by the power delay profile

Power delay profile

- Often described by a single exponential decay

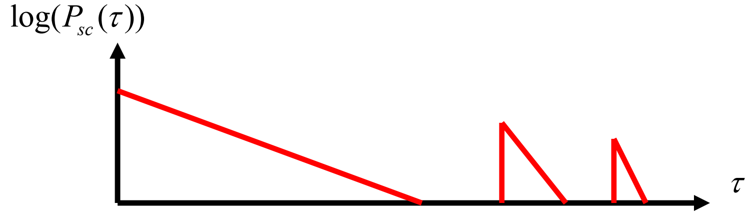
$$P_{sc}(\tau) = \begin{cases} exp(-\tau / S_\tau) & \tau \geq 0 \\ 0 & otherwise \end{cases}$$

delay spread



- though often there is more than one "cluster"

$$P(\tau) = \begin{cases} \sum_k \frac{P_k^c}{S_{\tau,k}^c} P_{sc}(\tau - \tau_{0,k}^c) & \tau \geq 0 \\ 0 & otherwise \end{cases}$$



arrival time

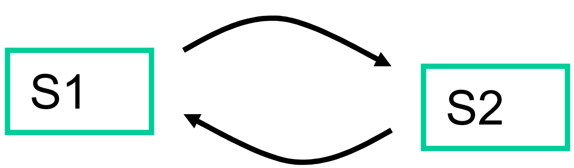
- If the bandwidth is high, the time resolution is large so we might resolve the different multipath components
- Need to model arrival time
- The Saleh-Valenzuela model:

ray arrival time (Poisson)

$$h(\tau) = \sum_{l=0}^{L} \sum_{k=0}^{K} \alpha_{k,l}(\tau)\delta(\tau - T_l - \tau_{k,l})$$

cluster arrival time (Poisson)

- The Δ-K-model:



arrival rate: $\lambda_0(t)$ $K\lambda_0(t)$

Wideband models
COST 207 model for GSM

The COST 207 model specifies:

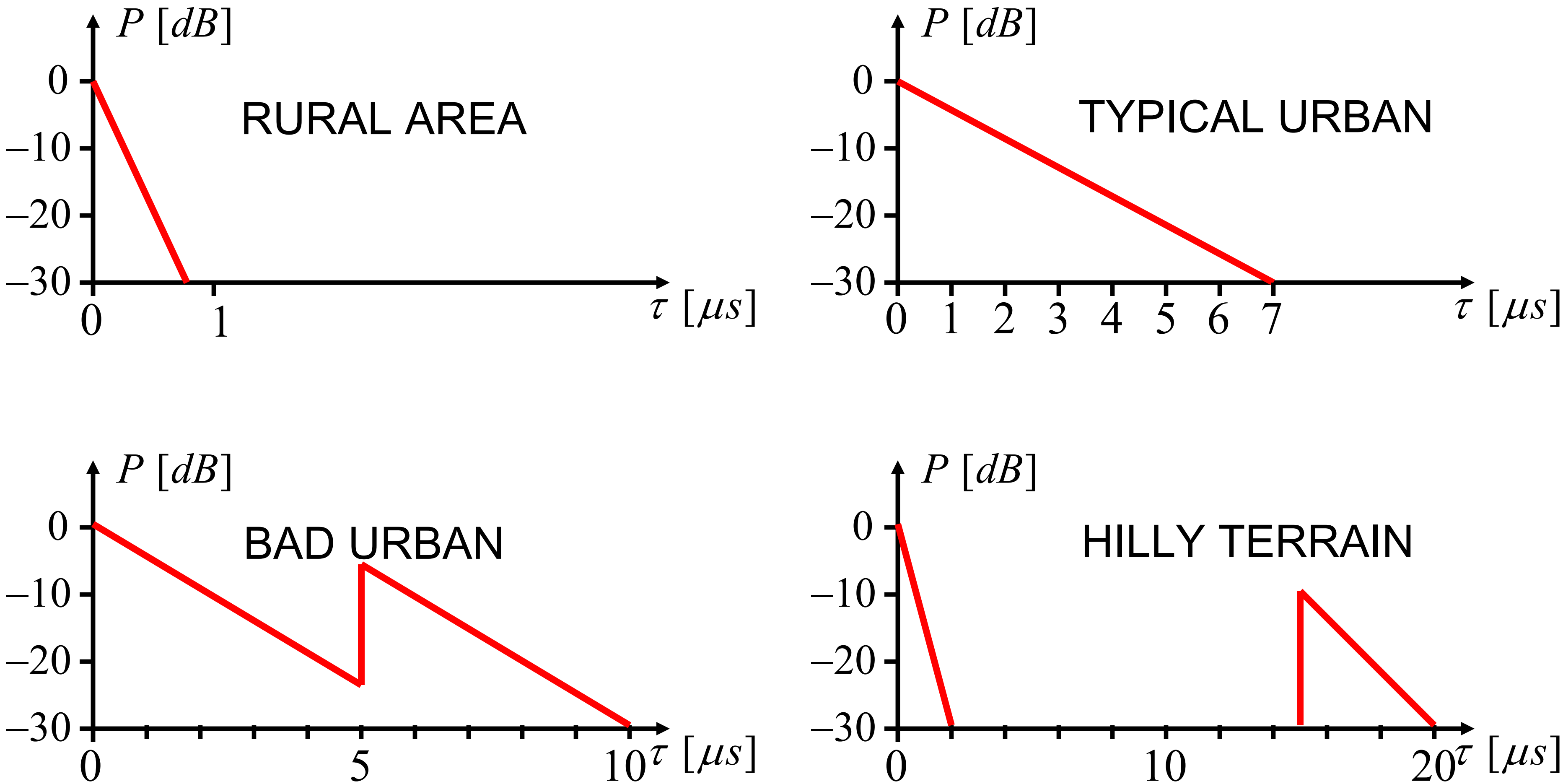
FOUR power-delay profiles for different environments.

FOUR Doppler spectra used for different delays.

IT DOES NOT SPECIFY PROAGATION LOSSES FOR THE DIFFERENT ENVIRONMENTS!

Wideband models
COST 207 model for GSM

Four specified power-delay profiles

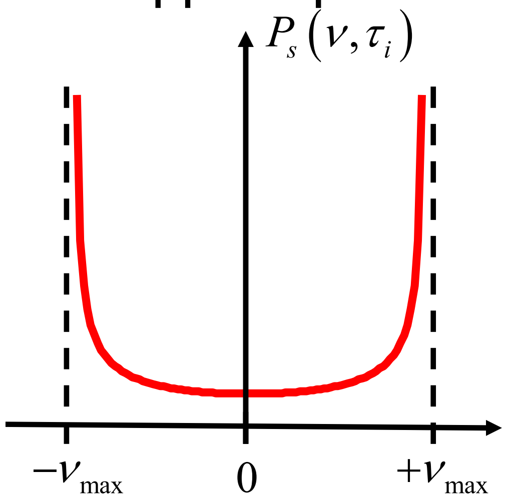


Wideband models
COST 207 model for GSM

Four specified Doppler spectra

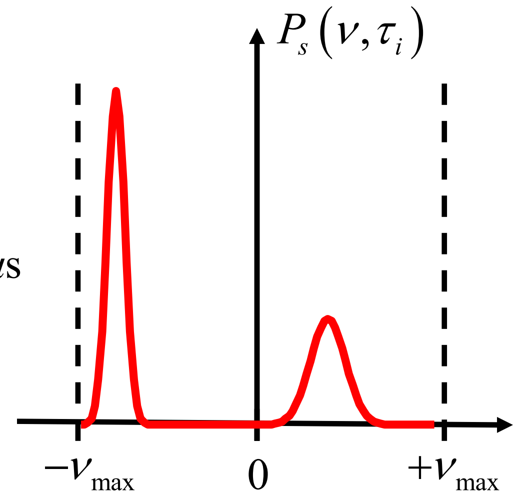
CLASS

$\tau_i \leq 0.5\ \mu s$



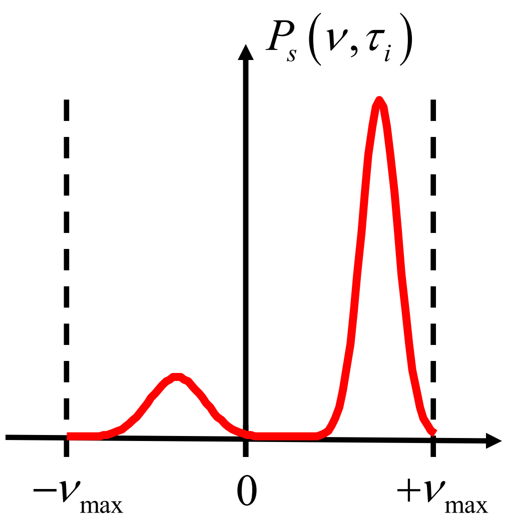
GAUS1

$0.5\ \mu s < \tau_i \leq 2\ \mu s$



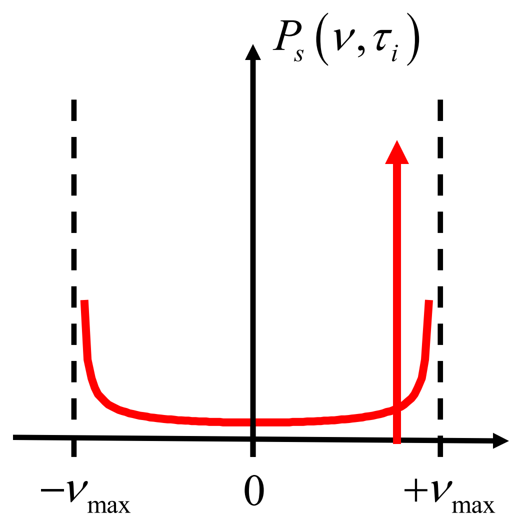
GAUS2

$\tau_i > 2\ \mu s$

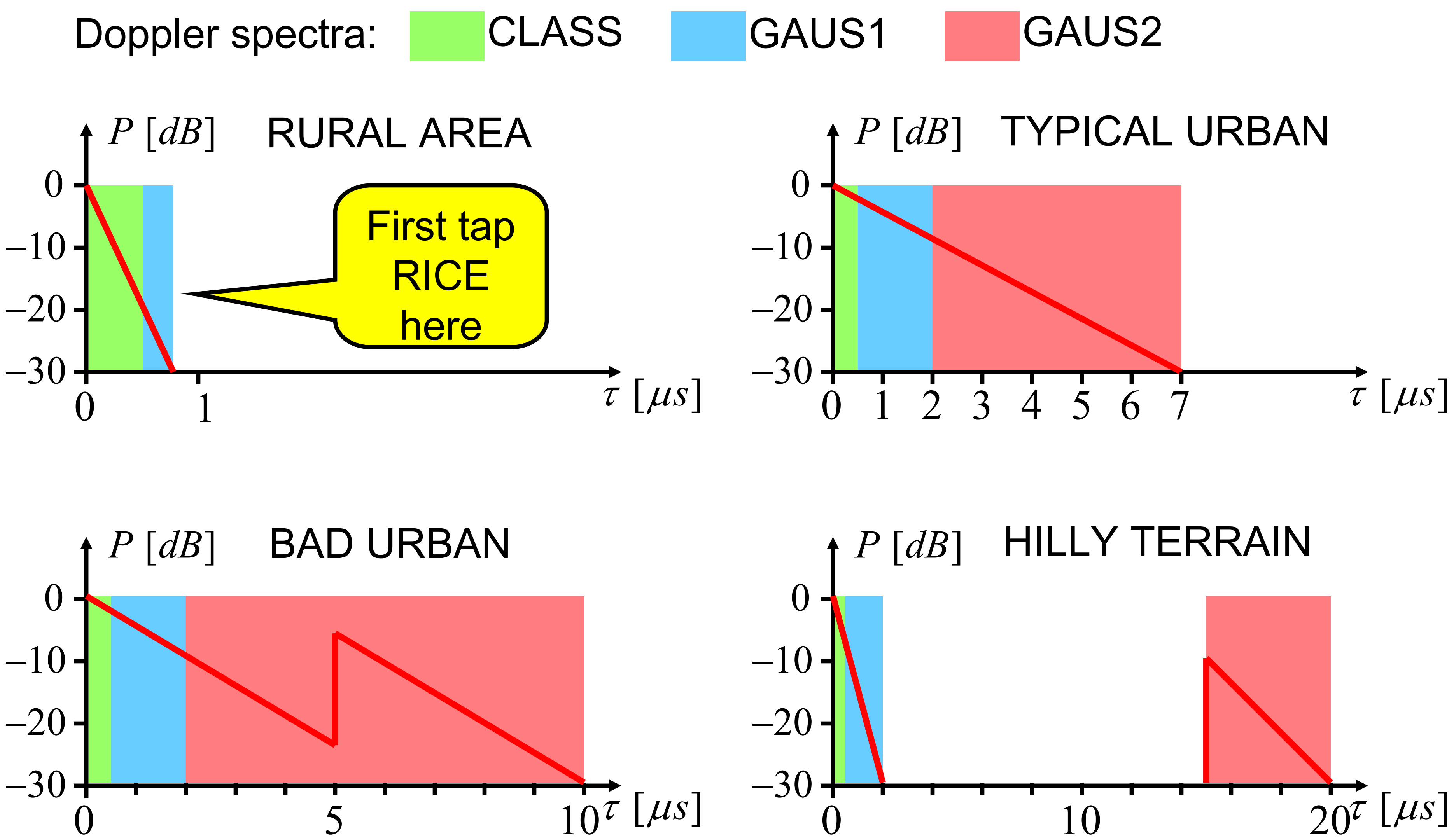


RICE

Shortest path in rural areas



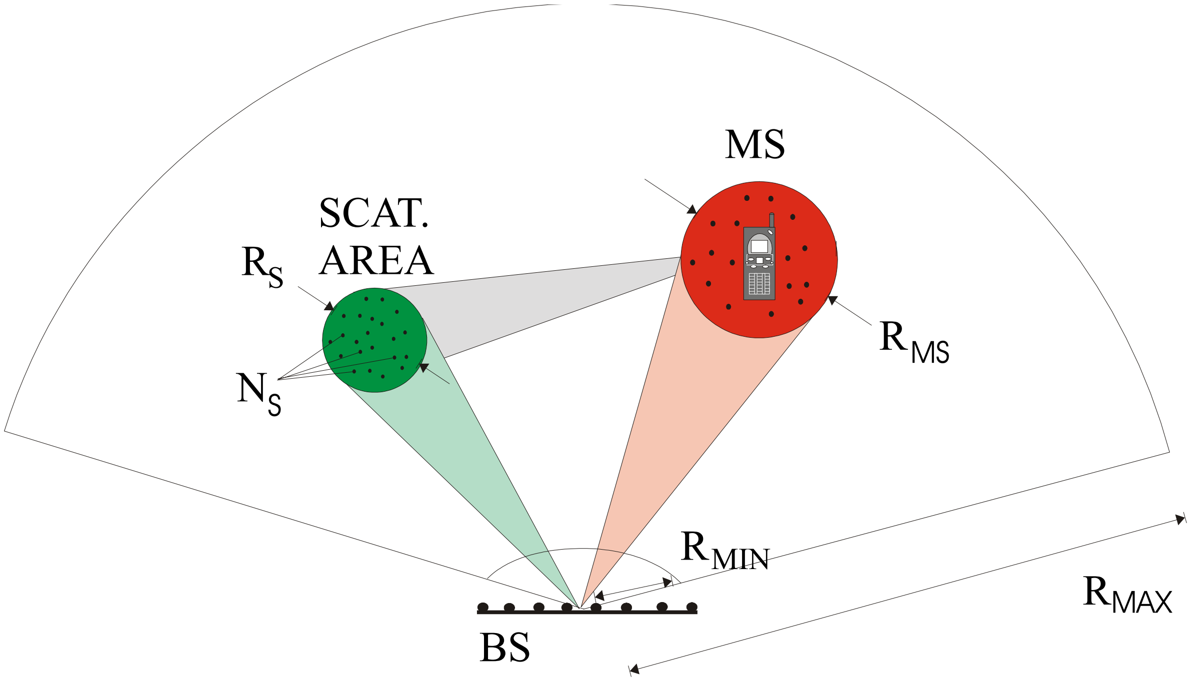
Wideband models
COST 207 model for GSM



Wideband models
ITU-R model for 3G

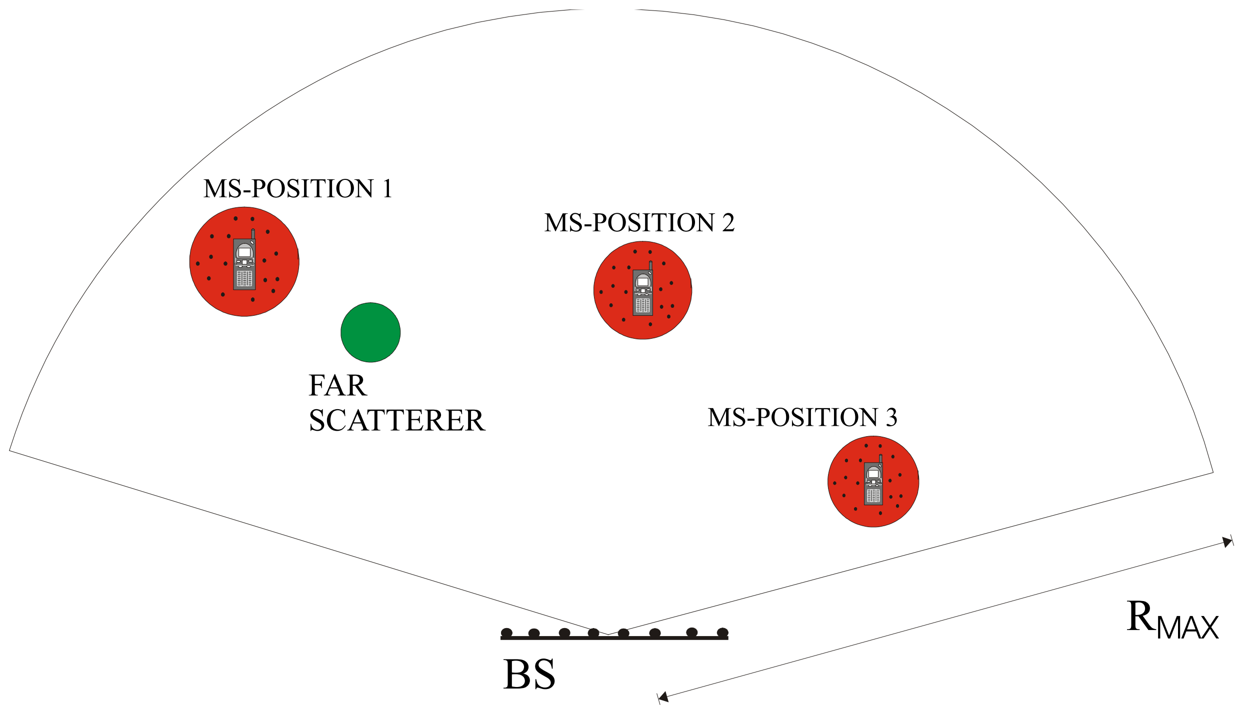
Tap No.	delay/ns	power/dB	delay/ns	power/dB
INDOOR	**CHANNEL A** (50%)		**CHANNEL B** (45%)	
1	0	0	0	0
2	50	-3	100	-3.6
3	110	-10	200	-7.2
4	170	-18	300	-10.8
5	290	-26	500	-18.0
6	310	-32	700	-25.2
PEDESTRIAN	**CHANNEL A** (40%)		**CHANNEL B** (55%)	
1	0	0	0	0
2	110	-9.7	200	-0.9
3	190	-19.2	800	-4.9
4	410	-22.8	1200	-8.0
5			2300	-7.8
6			3700	-23.9
VEHICULAR	**CHANNEL A** (40%)		**CHANNEL B** (55%)	
1	0	0	0	-2.5
2	310	-1	300	0
3	710	-9	8900	-12.8
4	1090	-10	12900	-10.0
5	1730	-15	17100	-25.2
6	2510	-20	20000	-16.0

Geometry-based stochastic channel model (GSCM)



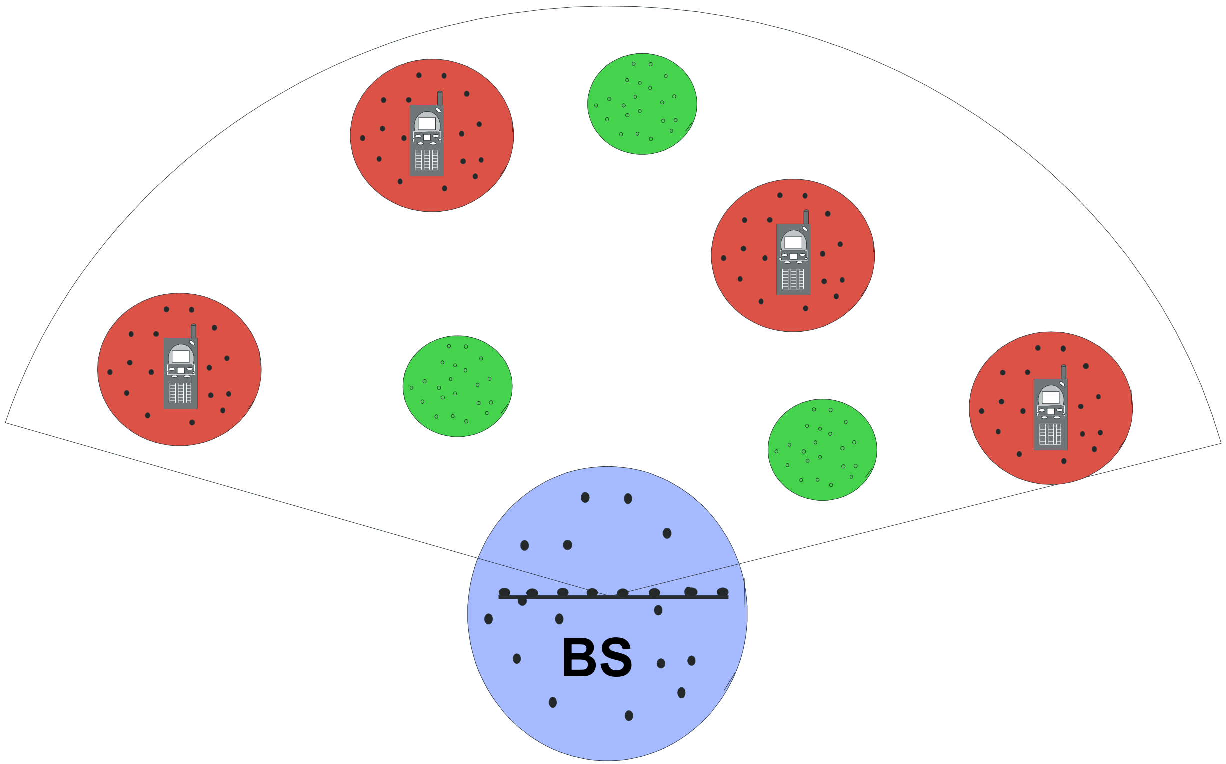
Temporal evolution - GSCM

- Temporal evolution of channel easily implemented



Modeling interference with GSCM

- Spatial correlation between interfering mobiles



MIMO channel

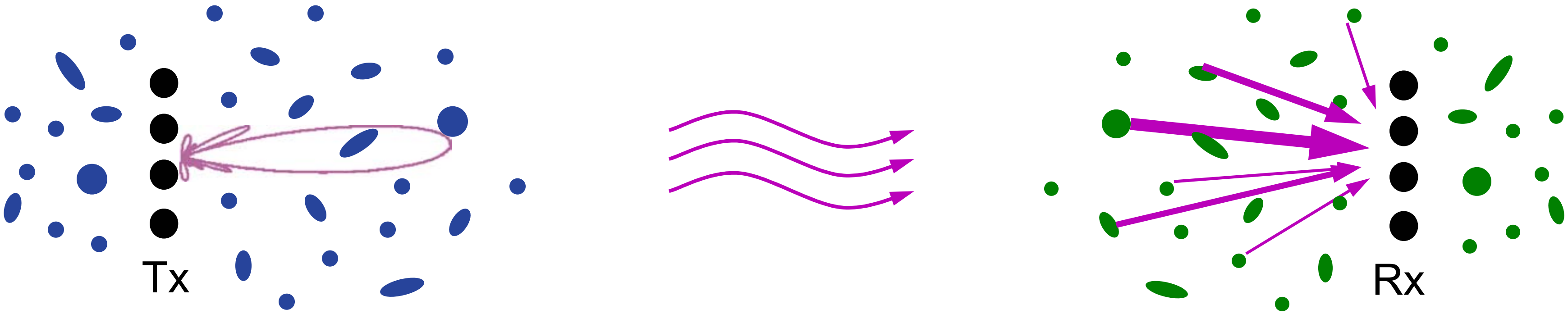
- channel matrix
$$\boldsymbol{H}(\tau)=\begin{bmatrix} h_{11}(\tau) & h_{12}(\tau) & \cdots & h_{1M_{\mathrm{Tx}}}(\tau) \\ h_{21}(\tau) & h_{22}(\tau) & \cdots & h_{2M_{\mathrm{Tx}}}(\tau) \\ \vdots & \vdots & \ddots & \vdots \\ h_{M_{\mathrm{Rx}}1}(\tau) & h_{M_{\mathrm{Rx}}2}(\tau) & \cdots & h_{M_{\mathrm{Rx}}M_{\mathrm{Tx}}}(\tau) \end{bmatrix}$$

- signal model
$$\boldsymbol{y}(t)=\sum_{\tau=0}^{D-1}\boldsymbol{H}(\tau)\cdot\boldsymbol{x}(t-\tau)$$

- mean channel
$$\overline{\boldsymbol{H}}(\tau)=\mathrm{E}\{\boldsymbol{H}(\tau)\}$$

- correlation *tensor* of order four
$$R_{mp}^{nq}(\tau)=\mathrm{E}\{h_n^m(\tau)\cdot h_p^{q^*}(\tau)\}$$

Kronecker model



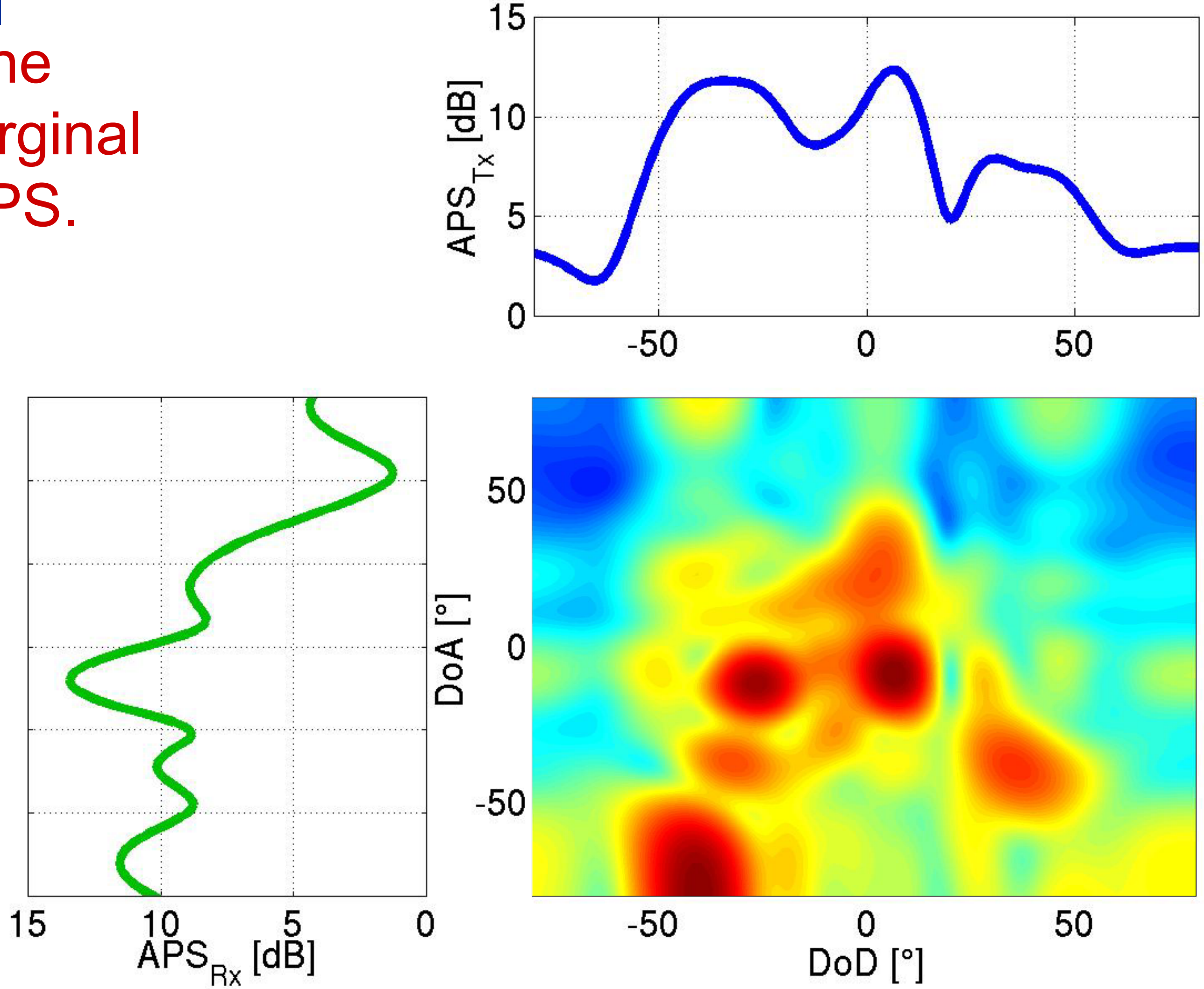
- The spatial structure of the MIMO channel is neglected.
- The MIMO channel is described by separated link ends:

$$\boldsymbol{R}_{\boldsymbol{H}} = c \cdot \boldsymbol{R}_{\mathrm{Tx}} \otimes \boldsymbol{R}_{\mathrm{Rx}} \qquad \boldsymbol{H} = \boldsymbol{R}_{\mathrm{Rx}}^{1/2} \boldsymbol{G} \boldsymbol{R}_{\mathrm{Tx}}^{\mathrm{T}/2}$$

Any **transmit signal results in one and the same receive correlation!**

Kronecker model (cont.)

Joint APS is the product of marginal Rx- and Tx-APS.

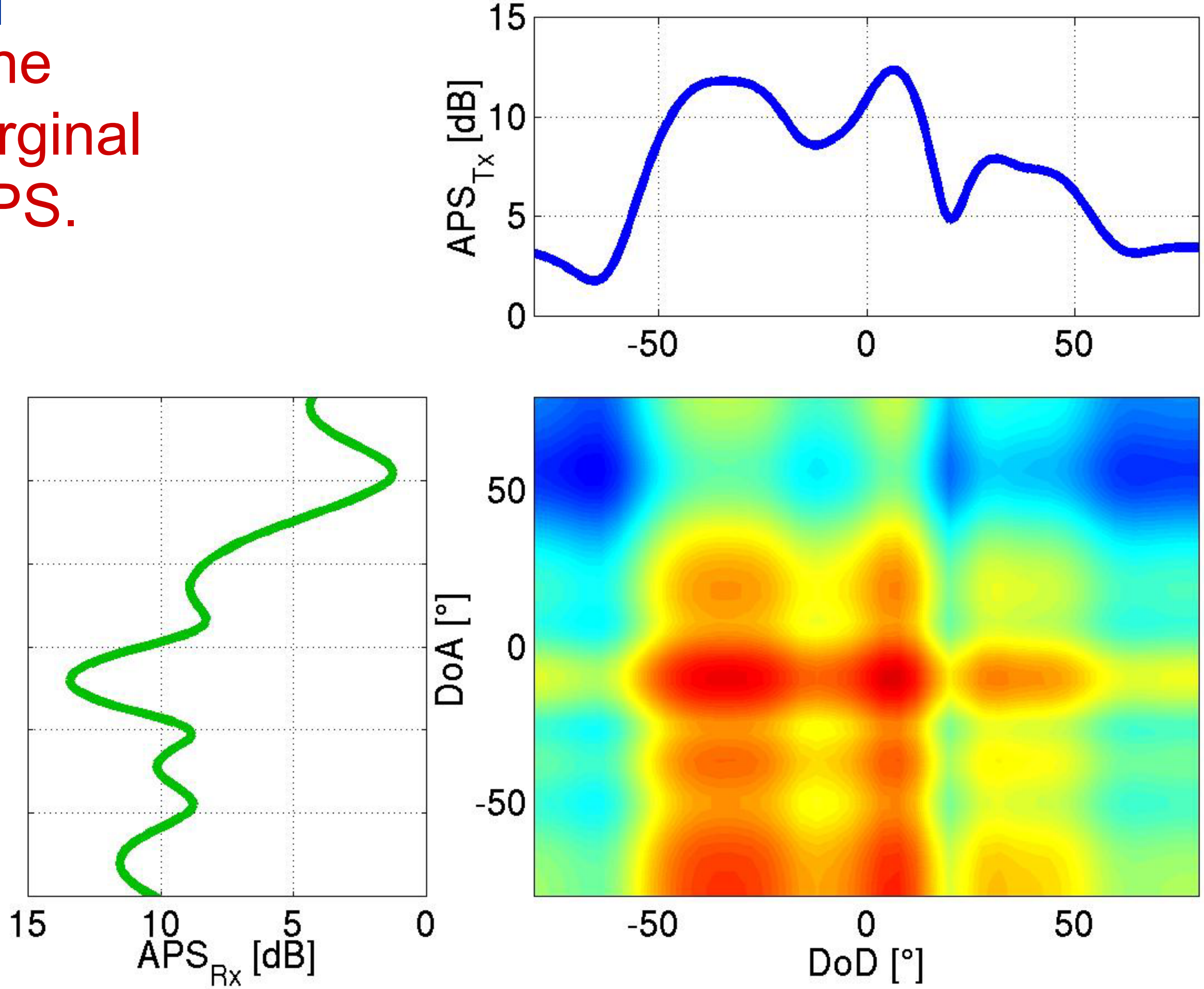


measurement

Kronecker model (cont.)

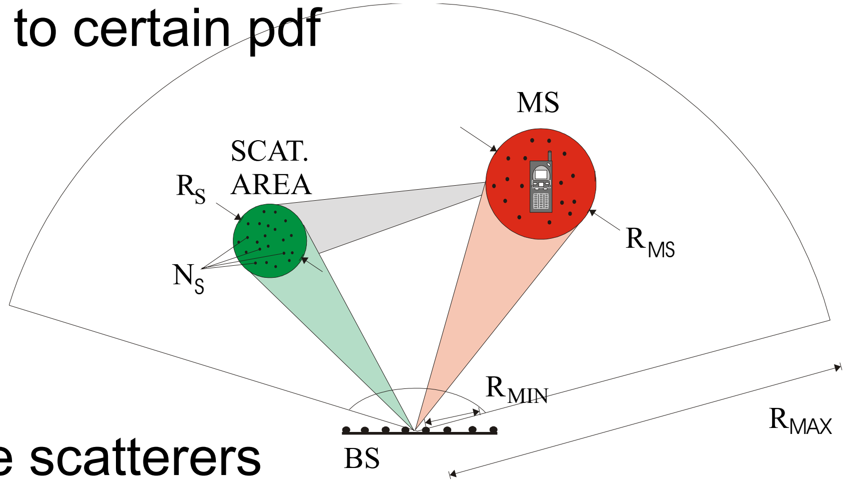
Joint APS is the product of marginal Rx- and Tx-APS.

Kronecker approximation



GSCM for MIMO

- GSCM – original version:
  - Locate scatterers according to certain pdf
  - only single scattering



- MIMO version:
  - model all effects that involve scatterers
  - Relative strength of propagation processes by weighting
  - Single scattering is not sufficient for MIMO!
  - MIMO capacity strongly depends on the angular spread.
  - Double- (multi-) Scattering increases angular spread.

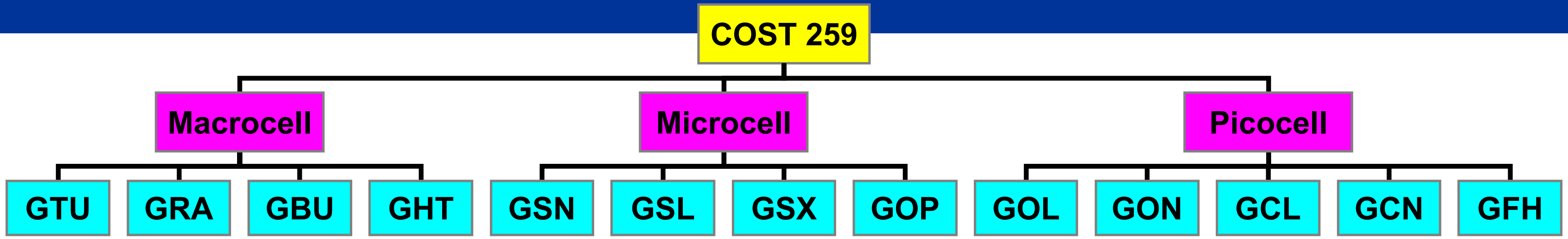
The COST 259 DCM

- COST 259 “Flexible Personalized Wireless Communication” Subgroup 2.1 Directional Channel Model
- European research initiative
- Includes operators, manufacturers, universities
- Close cooperation with other European programs
- Model widely used for smart antenna simulations
- Now also used for MIMO

COST 259 DCM - Philosophy

- Parametric approach, WSSUS not required
- No statement about implementation method (stochastic or GSCM)
- Based on clustering approach
- Multi-layer approach:
  - Radio environments
  - Large-scale effects
  - Small-scale effects

Radio environments



GTU	Generalized Typical Urban
GRA	Generalized Rural Area
GBU	Generalized Bad Urban
GHT	Generalized Hilly Terrain

GSN	Generalized Street NLOS
GSL	Generalized Street Canyon LOS
GSX	Generalized Street Crossing
GOP	Generalized Open Place

GOL	Generalized Office LOS
GON	Generalized Office NLOS
GCL	Generalized Corridor LOS
GCN	Generalized Corridor NLOS
GFH	Generalized Factory Hall

COST 259 DCM - Simulation procedure

Simulation steps:

1) select scenario

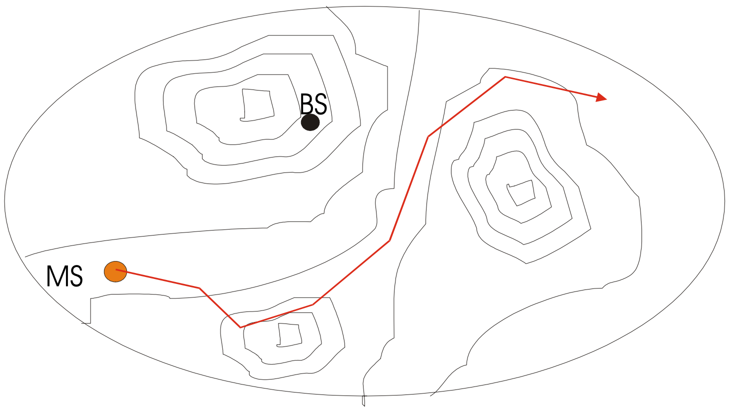
2) select global parameters (number of clusters, mean Rice factor,....)

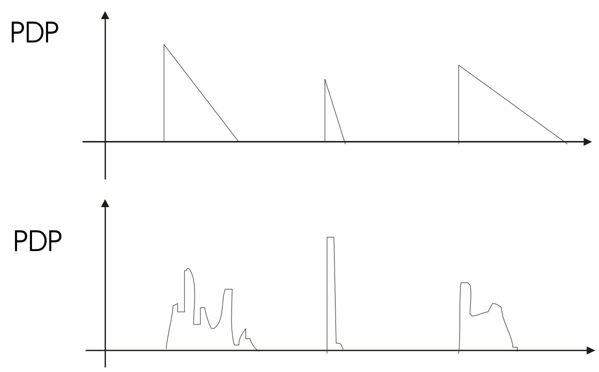
3) REPEAT

compute one realization of global parameters. This realization prescribes small-scale averaged power profiles (ADPS)

create many instantaneous complex impulse responses from this average ADPS

Generalized Hilly Terrain (GHT)





realization of
global parameters:
3 clusters, Rice factor of first: 3.2,
Rice factor of later clusters: 0
delay spread of clusters: 1,0.3,2s μ

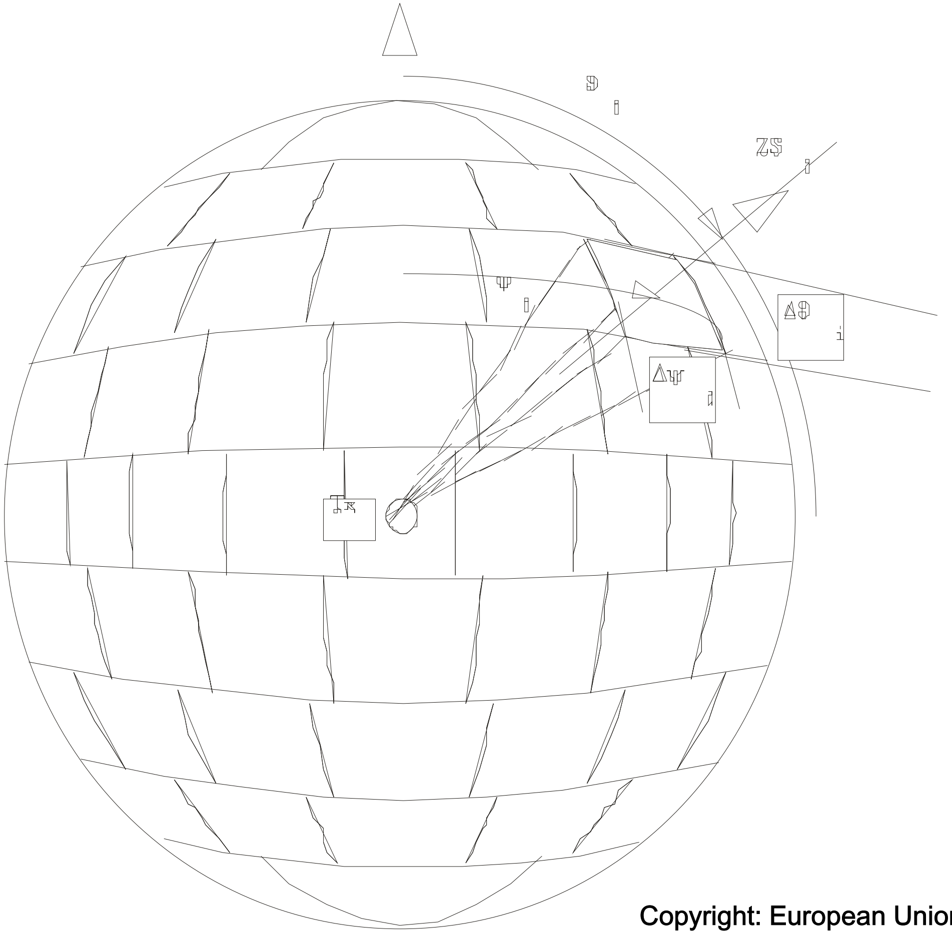
COST 259 DCM - Important features

- Very realistic !
- Distinguishes 13 different radio environments
- Treats large-scale and small-scale variations
- Far scatterer clusters included, with birth/death process
- Delay spread and angular spread treated as (correlated) random variables
- Angular spectra are functions of delay
- Azimuth and elevation

Deterministic modeling methods

- Solve Maxwell's equations with boundary conditions
- Problems:
  - Data base for environment
  - Computation time
- "Exact" solutions
  - Method of moments
  - Finite element method
  - Finite-difference time domain (FDTD)
- High frequency approximation
  - All waves modeled as rays that behave as in geometrical optics
  - Refinements include approximation to diffraction, diffuse scattering, etc.

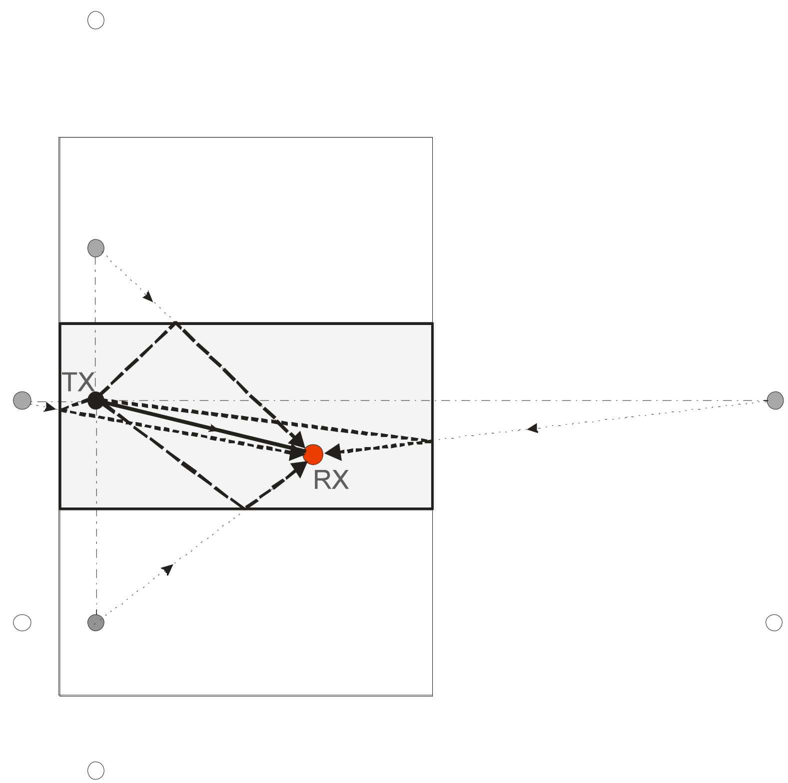
Ray launching



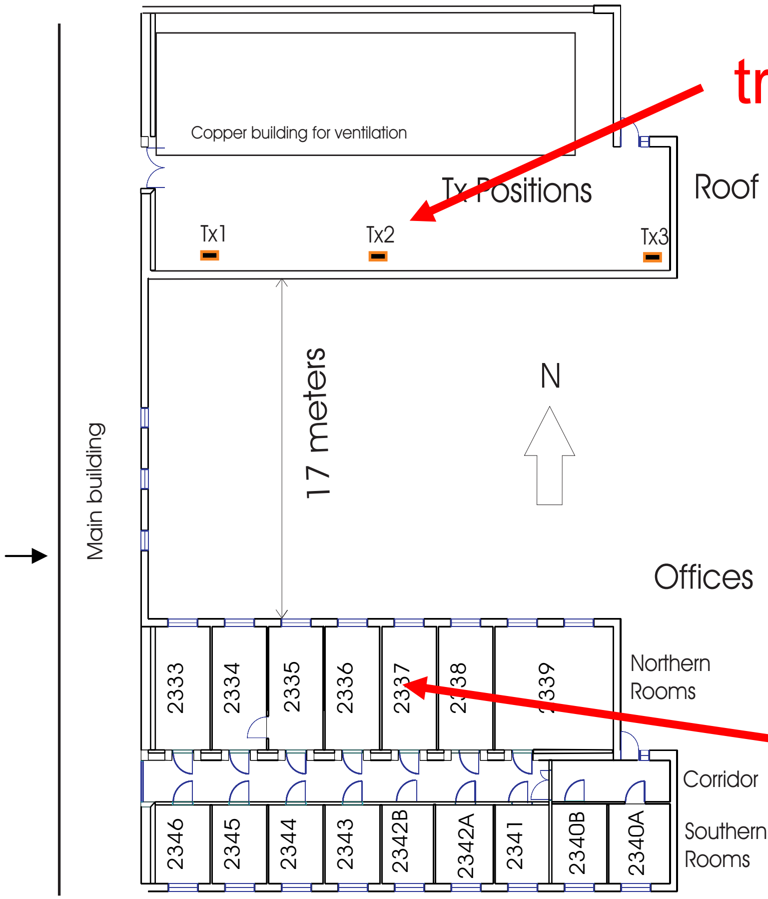
Ray tracing

- Determines rays that can go from one TX position to one RX position
  - Uses imagining principle
  - Similar to techniques known from computer science

- Then determine attenuation of all those possible paths



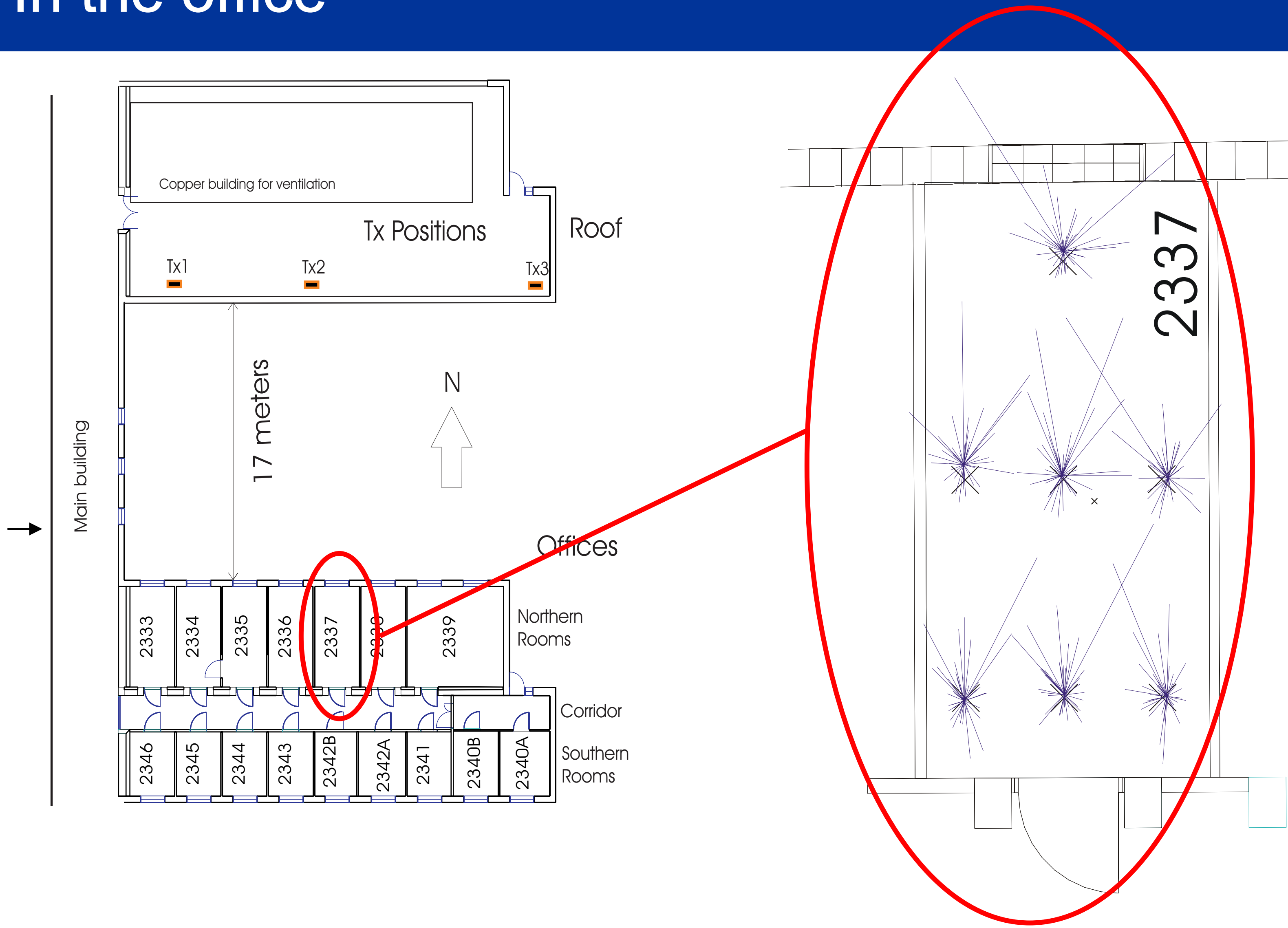
How does the signal reach the receiver Outdoor-to-indoor



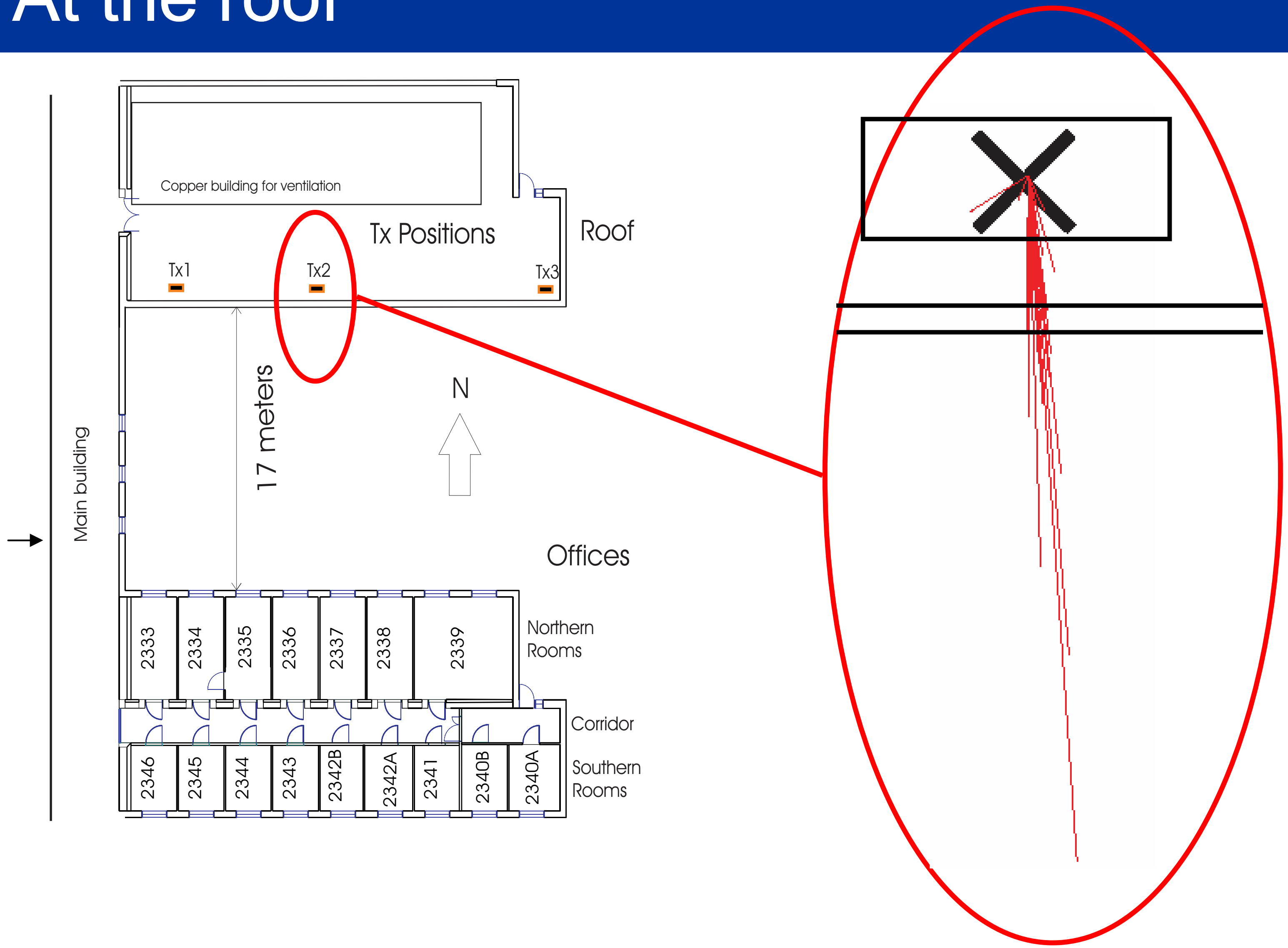
transmitter

receiver

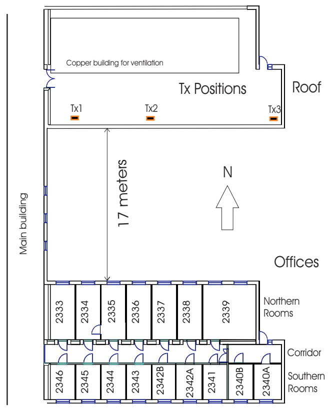
How does the signal reach the receiver In the office

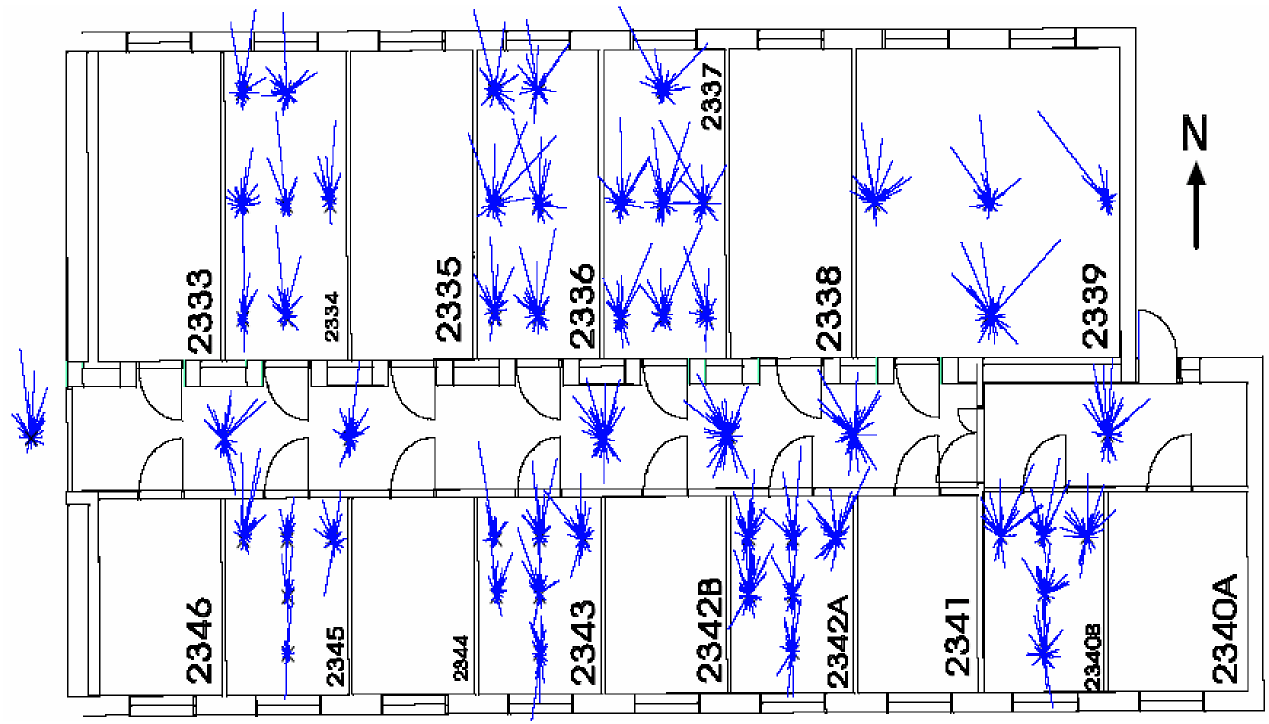


How does the signal leave the transmitter
At the roof



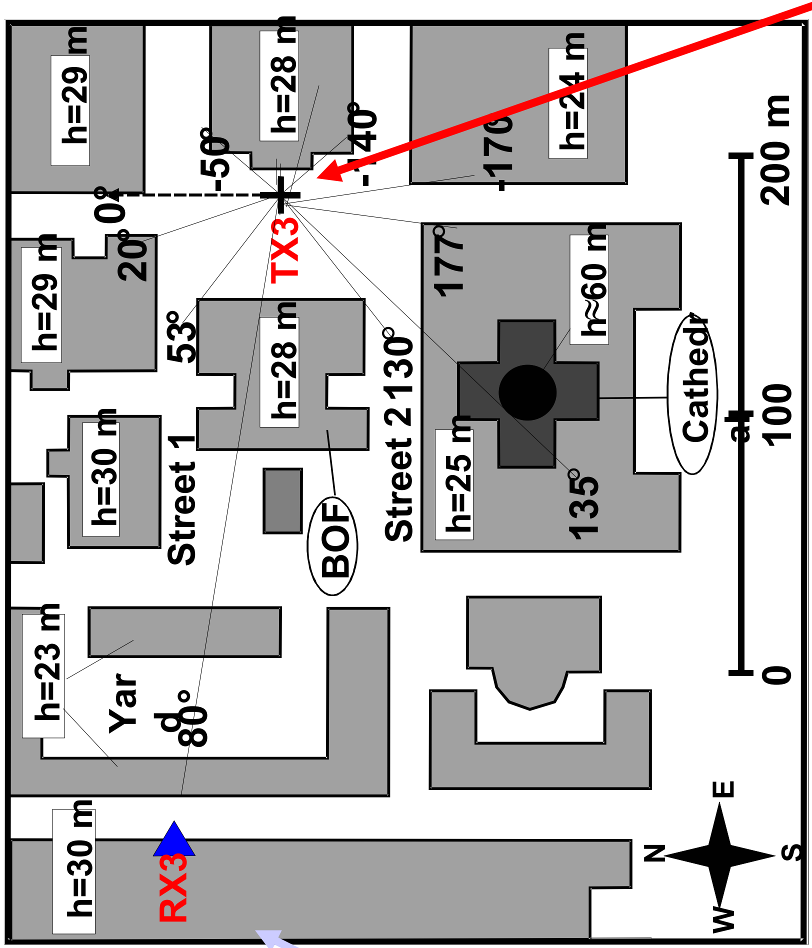
In all offices

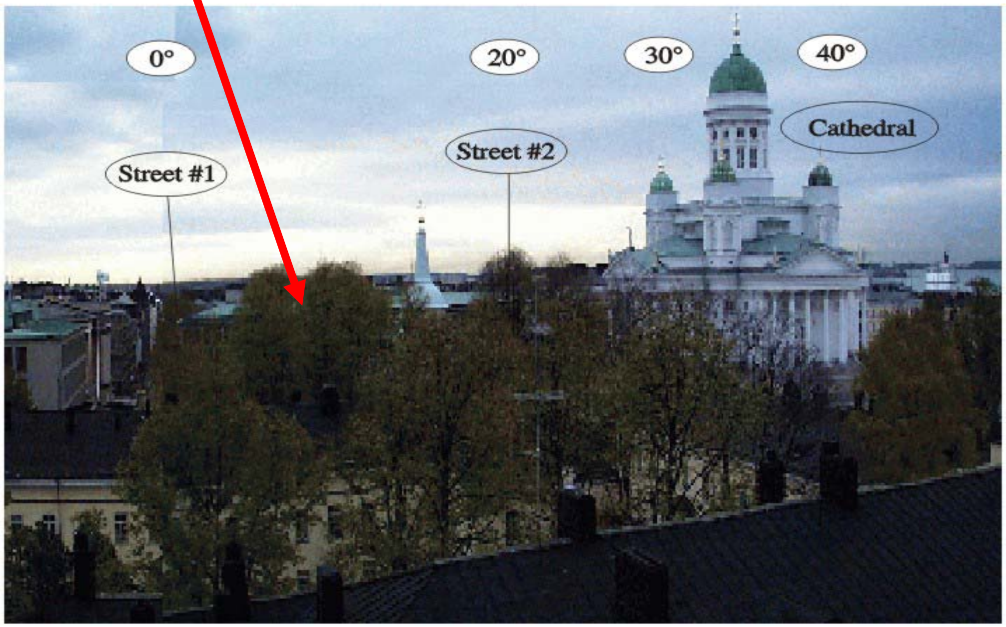




How does the signal reach the receiver outdoor urban

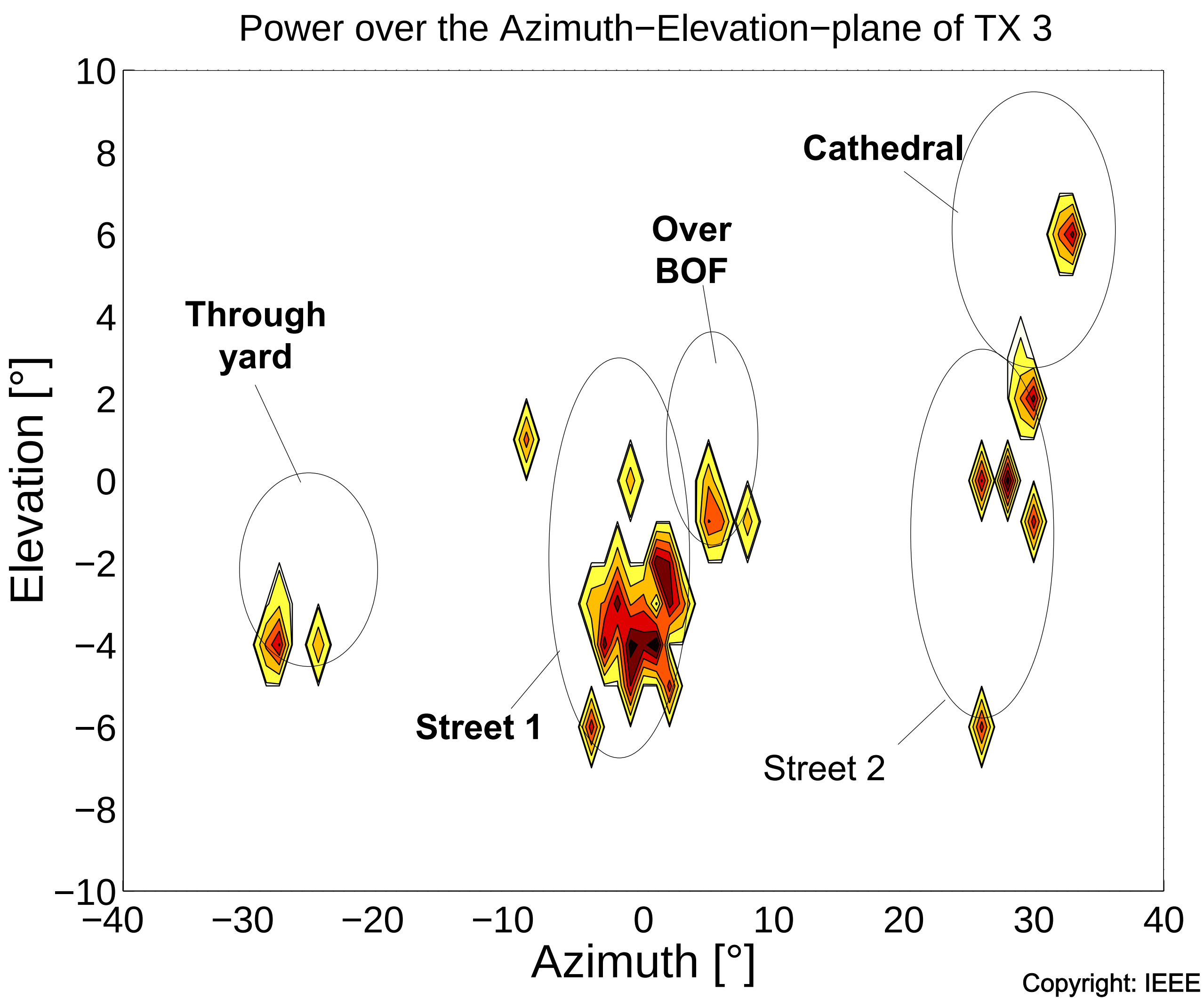
Transmitter



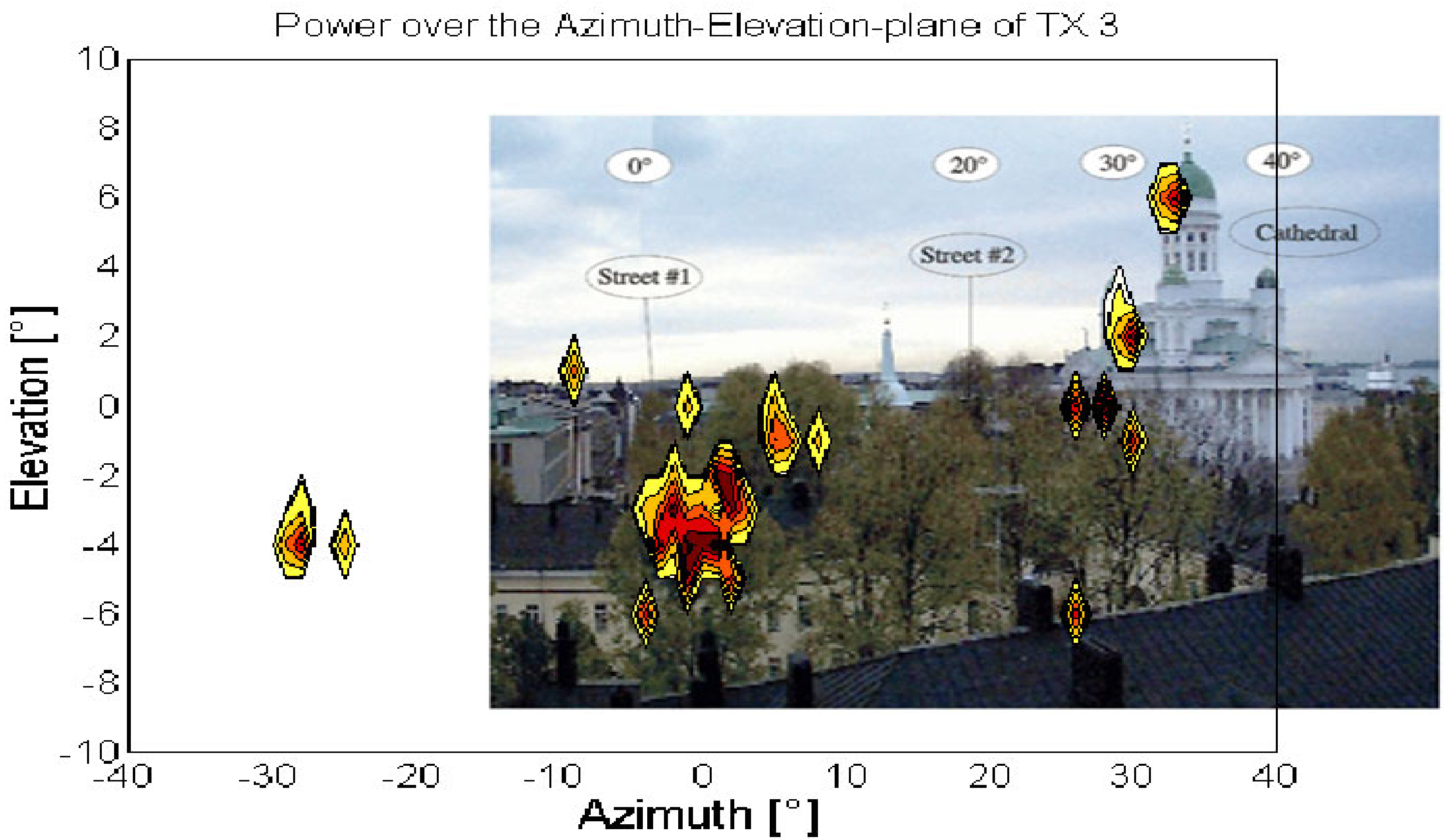


Mottagare

Signal arrives from some specific areas



Diffraction, reflection, scattering, transmission



Chapter 8

Channel sounding

Channel measurements

In order to model the channel behavior we need to measure its properties

- Time domain measurements
  - impulse sounder
  - correlative sounder
- Frequency domain measurements
  - Vector network analyzer

- Directional measurements
  - directional antennas
  - real antenna arrays
  - multiplexed arrays
  - virtual arrays

Basic identifiability of the channel

- The channel can be measured uniquely only if
  - sampling theorem

$$f_{\mathrm{rep}} \geq 2\nu_{\mathrm{max}}$$

$$\frac{1}{f_{\mathrm{rep}}} \geq \tau_{\mathrm{max}}$$

- Therefore, a channel can only be measured uniquely if it is *underspread*

$$2\tau_{\mathrm{max}}\nu_{\mathrm{max}} \leq 1$$

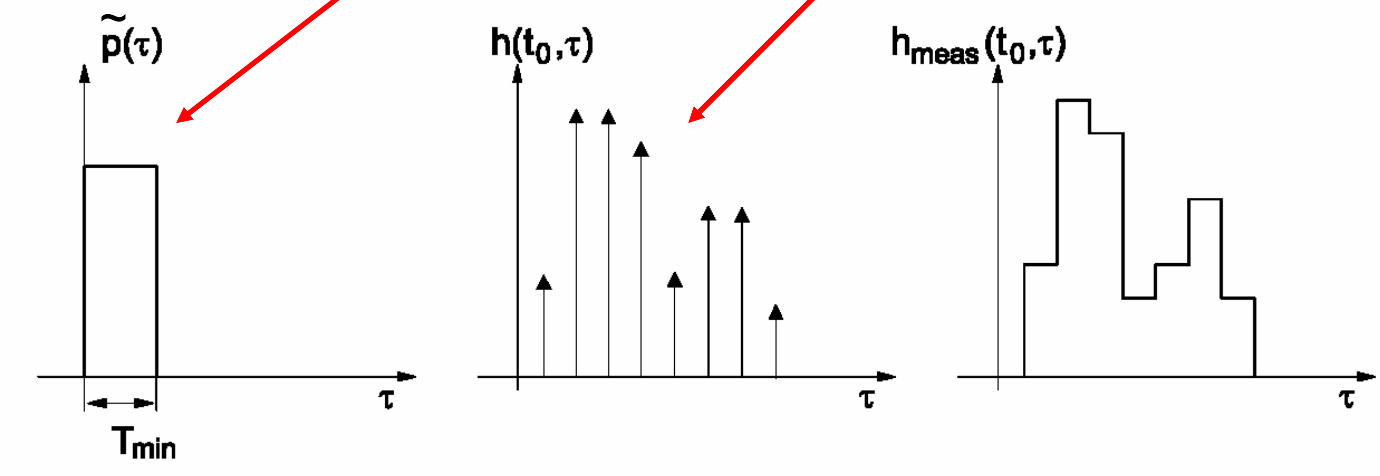
- This condition is fulfilled in all practical wireless applications

Impulse sounder

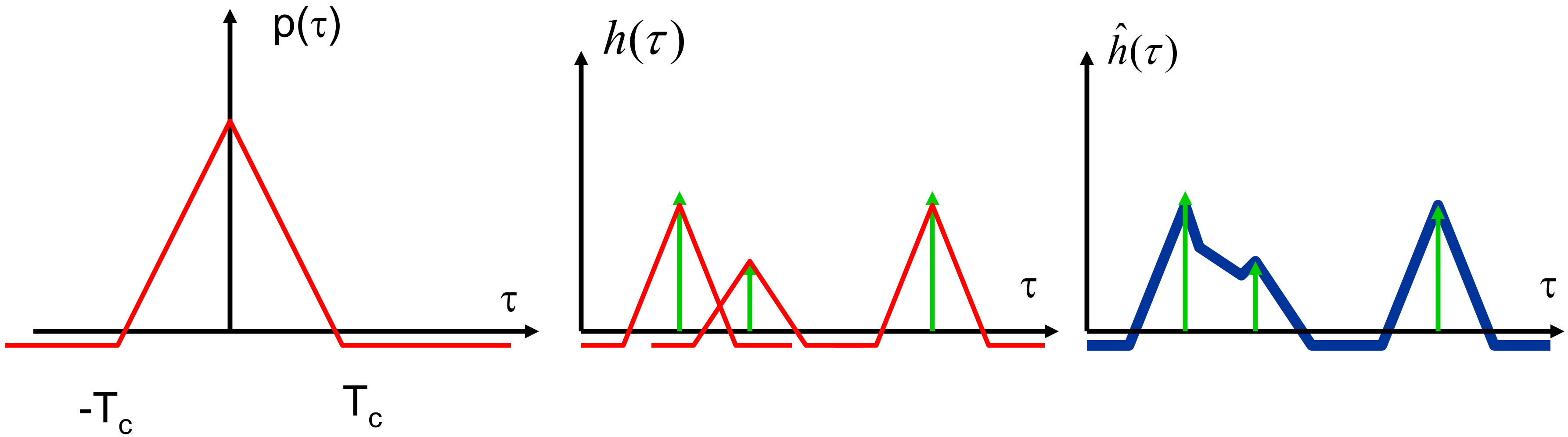
$$h_{\mathrm{meas}}(t_i,\tau) = \tilde{p}(\tau) * \boxed{h(t_i,\tau)}$$

impulse response
of sounder

impulse response
of channel



Correlative sounder



correlation peak impulse response measured impulse response

Frequency domain measurements

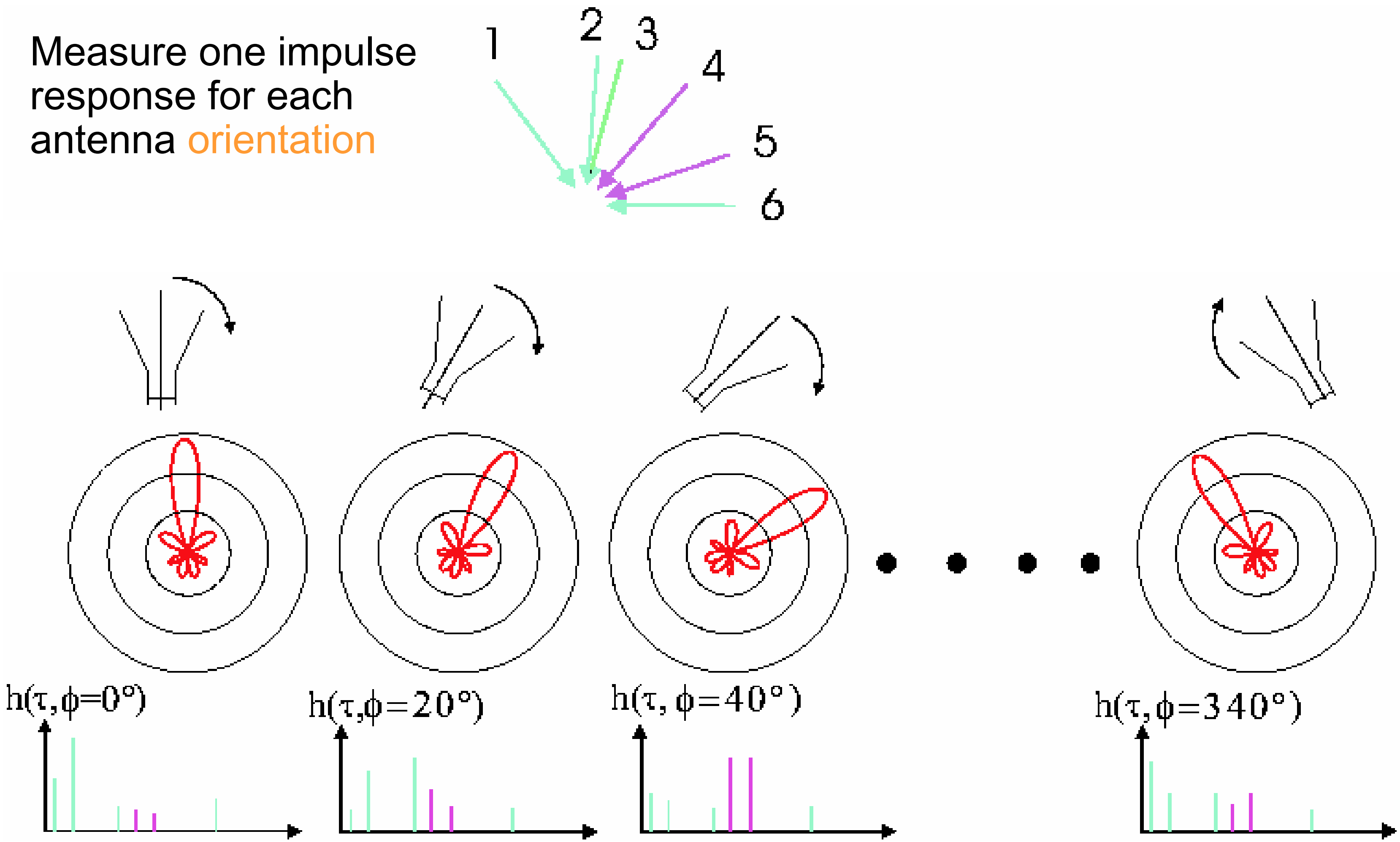
- Use a vector network analyzer or similar to determine the transfer function of the channel

$$H_{meas}(f) = H_{TXantenna}(f) * \boxed{H_{channel}(f)} * H_{RXantenna}(f)$$

- Need to know the influence of the measurement system

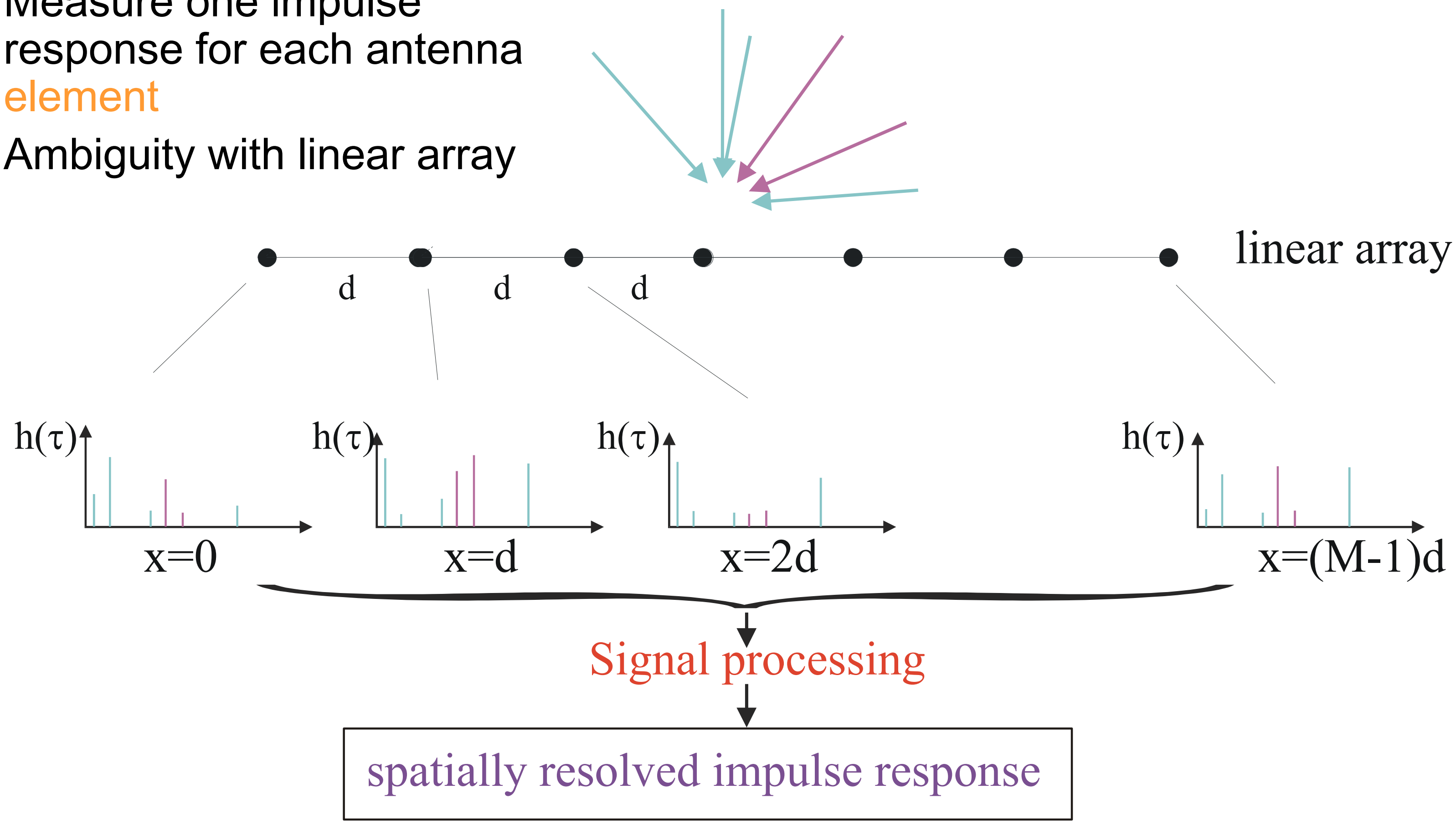
Channel sounding – directional antenna

- Measure one impulse response for each antenna orientation



Channel sounding – antenna array

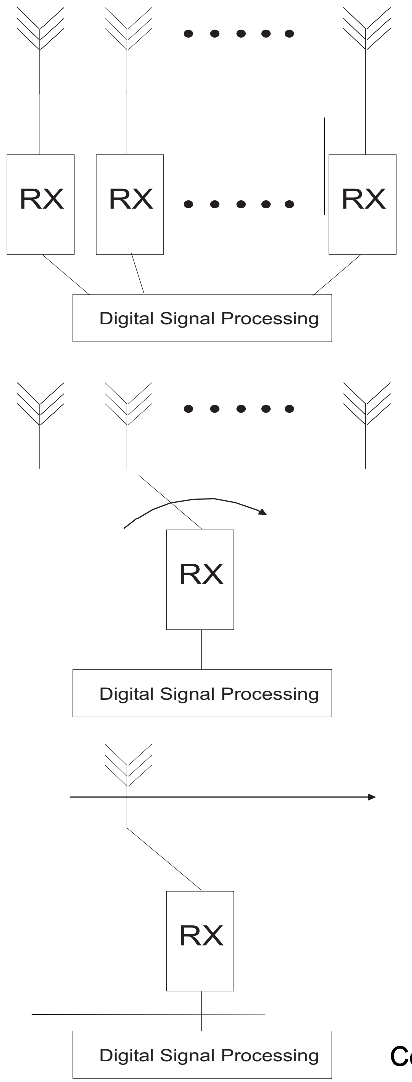
- Measure one impulse response for each antenna element
- Ambiguity with linear array



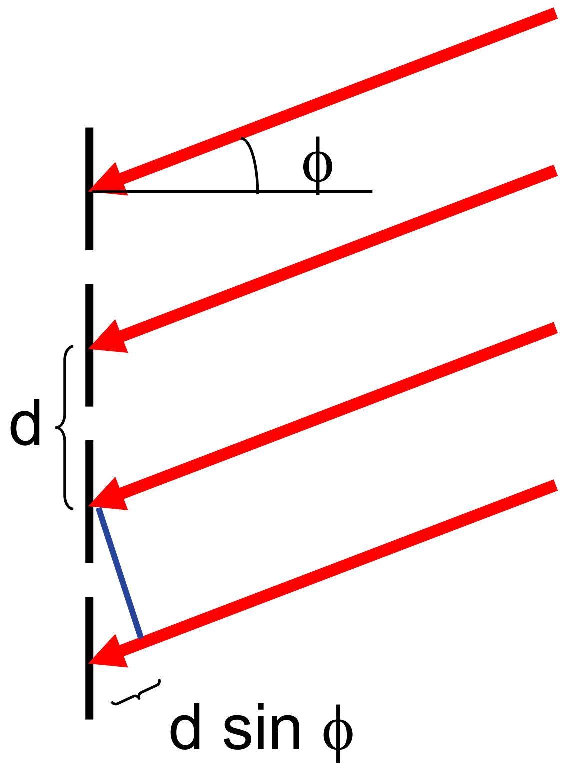
Real, multiplexed, and virtual arrays

- Real array: simultaneous measurement at all antenna elements
- Multiplexed array: short time intervals between measurements at different elements
- Virtual array: long delay

  no problem with mutual coupling



Directional analysis



- The DoA can, e.g., be estimated by correlating the received signals with steering vectors.

$$\vec{a}(\phi) = \begin{pmatrix} 1 \\ \exp(-jk_0 d\cos(\phi)) \\ \exp(-j2k_0 d\cos(\phi)) \\ \vdots \\ \exp(-j(M-1)k_0 d\cos(\phi)) \end{pmatrix}$$

- An element spacing of d=5.8 cm and an angle of arrival of ϕ =20 degrees gives a time delay of $6.6 \cdot 10^{-11}$ s between neighboring elements

High resolution algorithms

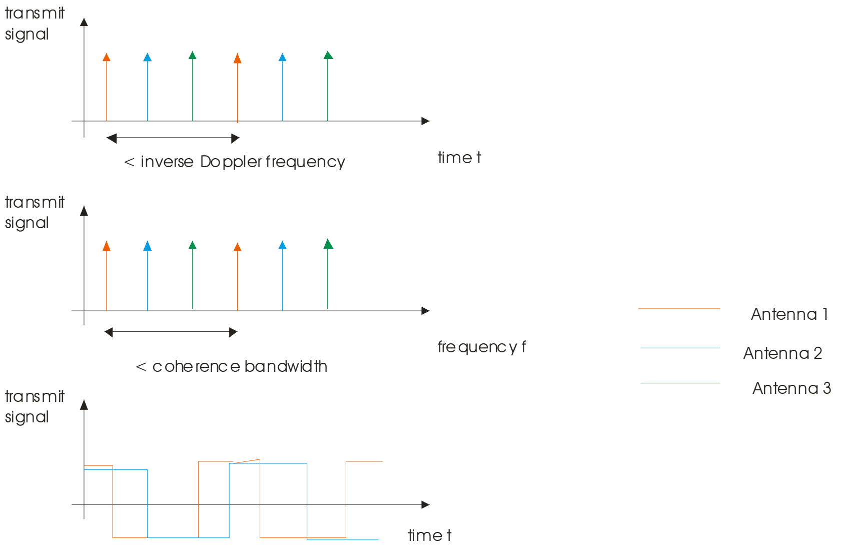
- In order to get better angular resolution, other techniques for estimating the angles are used, e.g.:
  - MUSIC, subspace method using spectral search
  - ESPRIT, subspace method
  - MVM (Capon’s beamformer), rather easy spectral search method
  - SAGE, iterative maximum likelihood method
- Based on models for the propagation
- Rather complex, one measurement point may take 15 minutes on a decent computer

Antenna array TX

- Transmission must be done so that RX can distinguish signals from different TX receivers

  →Transmit signals should be orthogonal

- Orthogonality in time
- Orthogonality in frequency
- Orthogonality in code



Chapter 9

Antennas

Antennas in real channels

- One important aspect is how the channel and antenna interact
  - The antenna pattern determines what the system sees
  - Delay spread and angular spread affected by the antenna pattern

- The user may have a large influence on the behavior of the antenna
  - Change in antenna pattern
  - Change in efficiency – miss-match

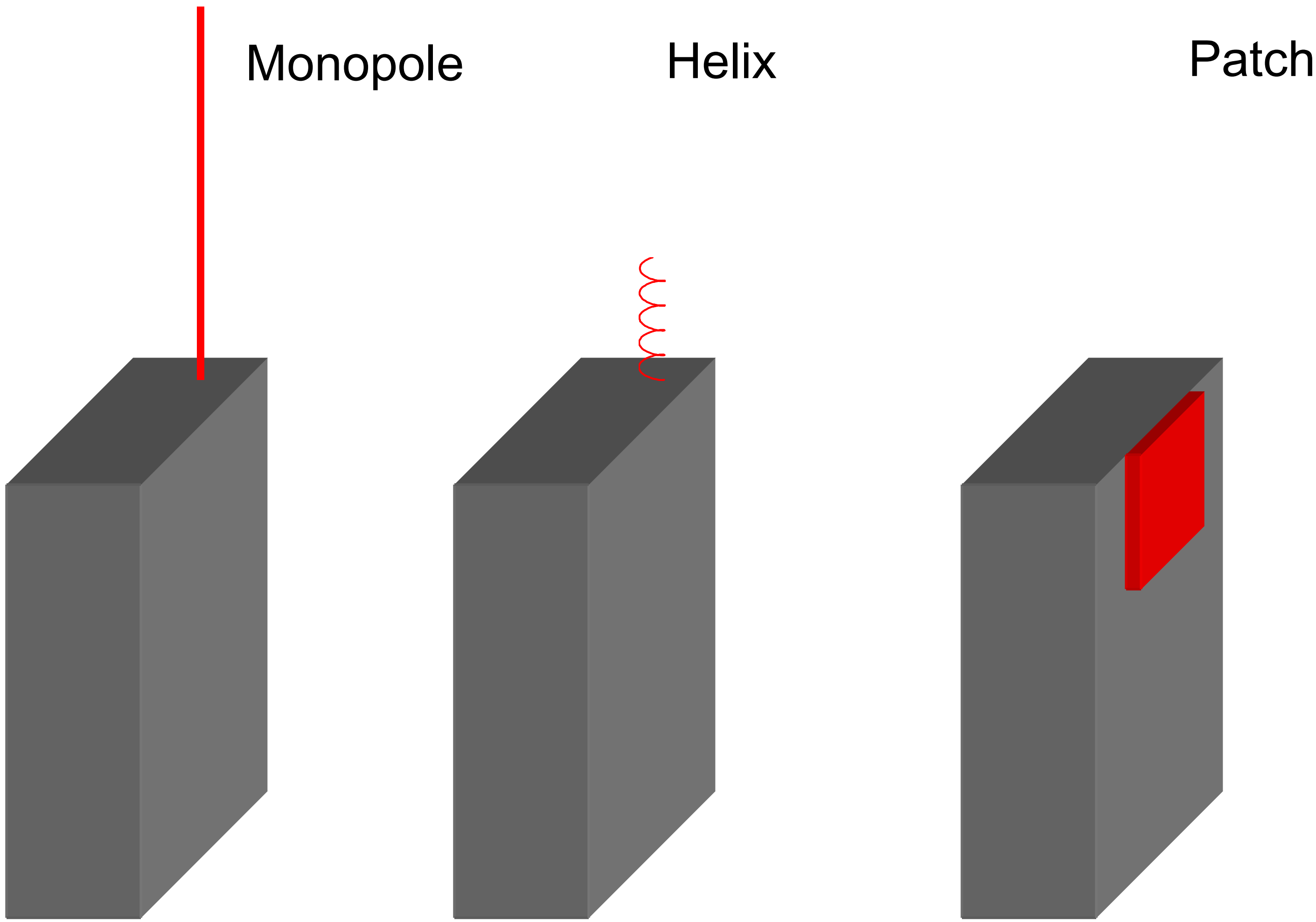
Important antenna parameters

- Directivitiy
  - Total power in a certain direction compared to total transmitted power
- Efficiency

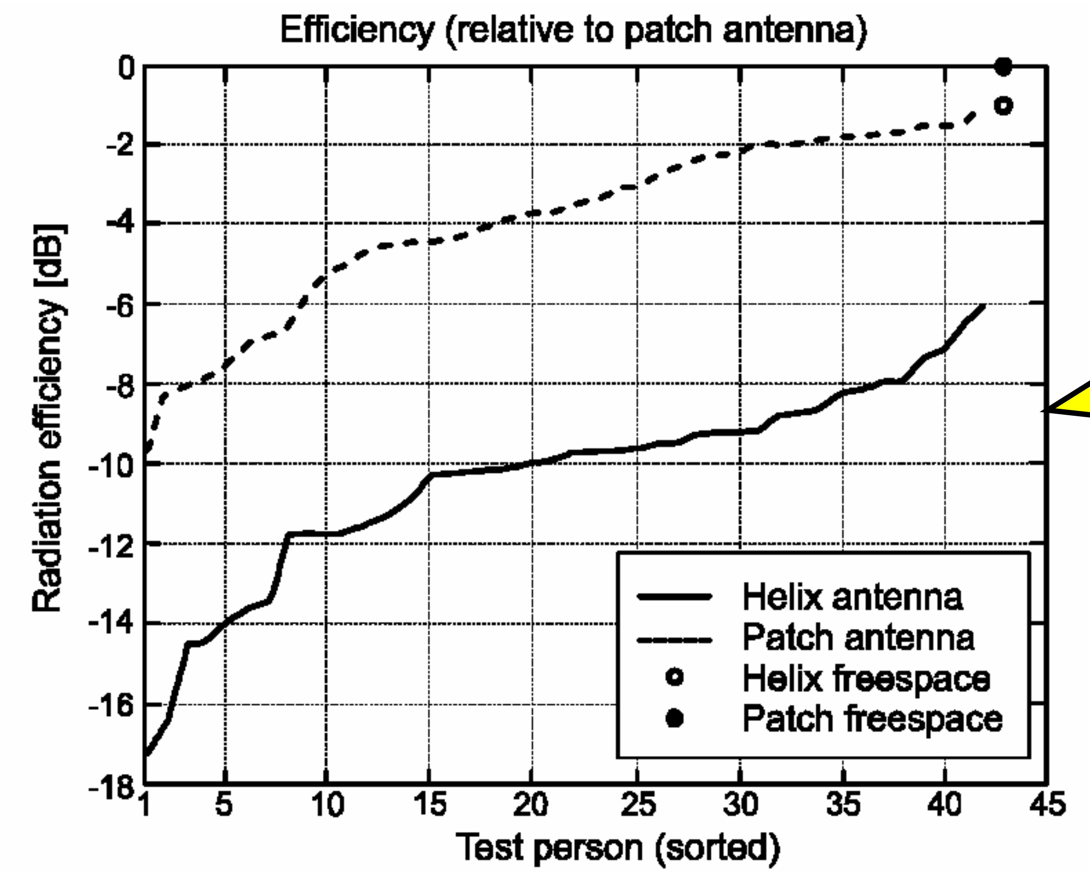
$$\eta = \frac{R_{rad}}{R_{rad} + R_{ohmic} + R_{match}}$$

- Q-factor
  - Stored energy compared to dissipated energy
- Mean effective gain
  - Include influence of random channel
  - Average received power compared to average received power by isotropic antenna in real environment
- Polarization
- Bandwidth

Mobile station antennas



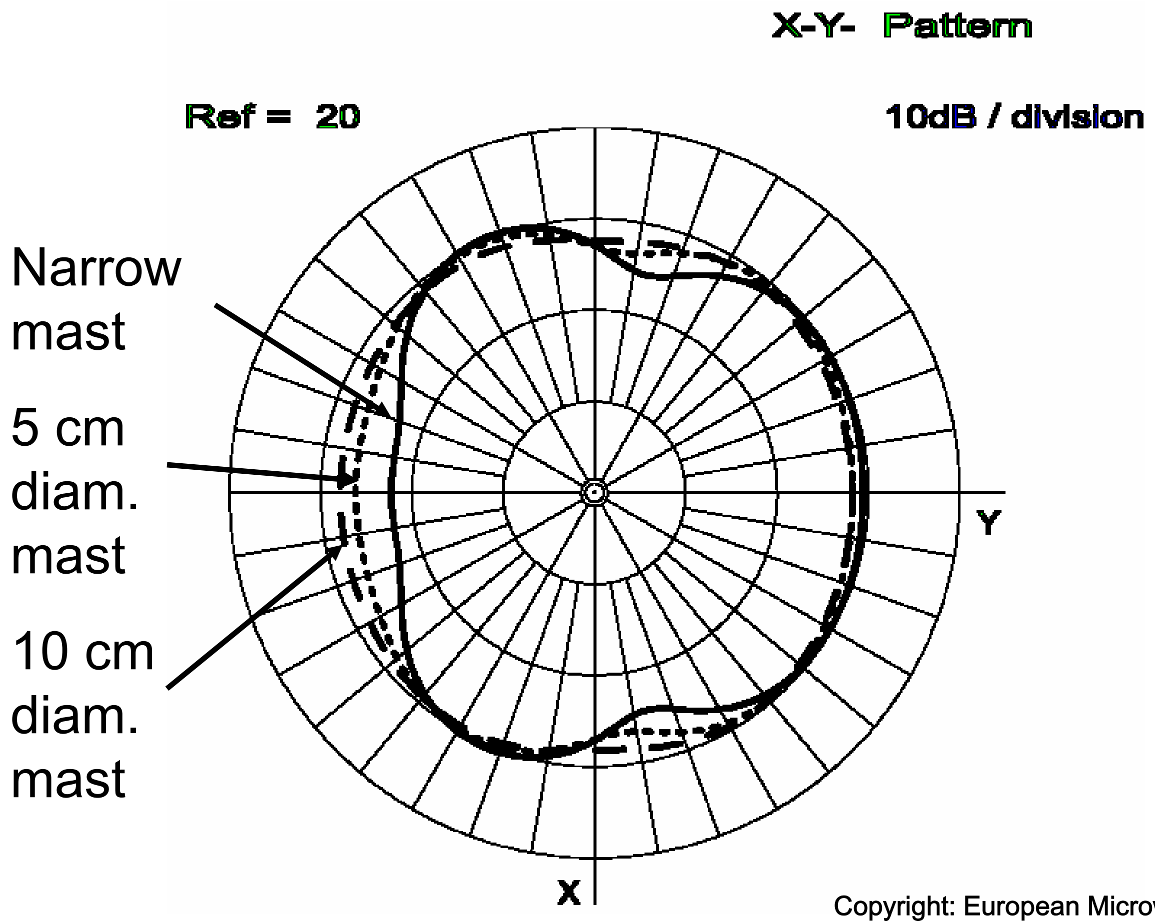
Impact of user on MS antenna



Up to around 10 dB difference, depending on person.

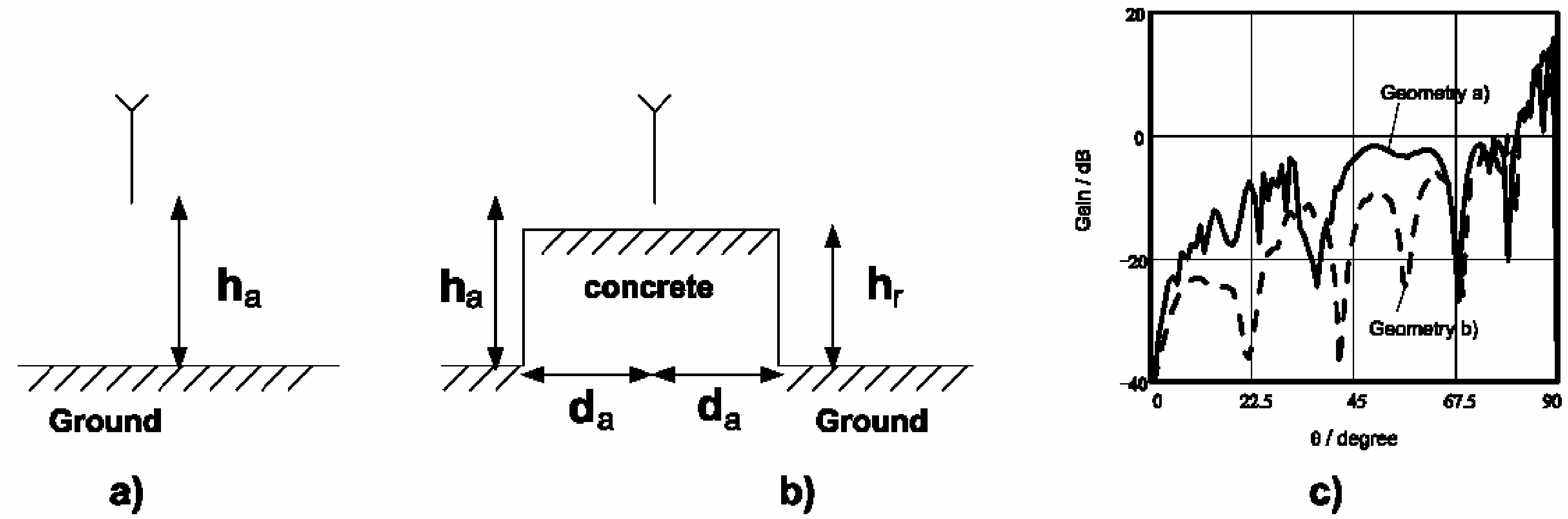
Base station antennas

Base station antenna pattern affected by the mast (30 cm from antenna).



Base station antennas

Base station antenna pattern affected by a concrete foundation.



Common antenna types

- Linear antennas (dipole, monopole)
- Helical antennas
- Microstrip antennas
- PIFA and RCDLA antennas

Linear antennas (1)

- Hertzian dipole (short dipole)
  - Antenna pattern:

    $$\widetilde{G}(\varphi,\theta) \propto \sin(\theta)$$

  - Gain

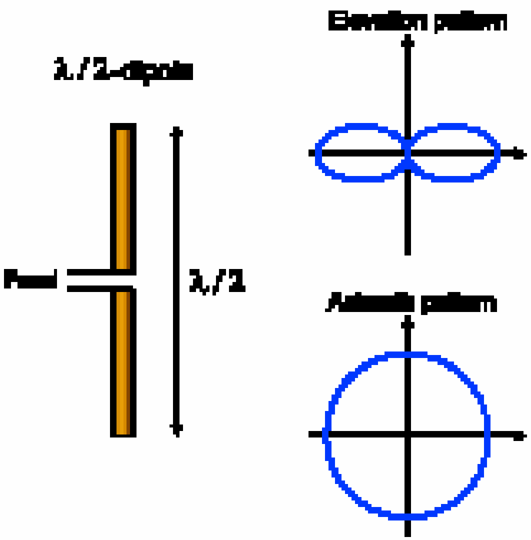
    $$G_{\max} = 1.5$$

- $\lambda/2$ dipole
  - Pattern

    $$\widetilde{G}(\varphi,\theta) \propto \frac{\cos\left(\frac{\pi}{2}\cos(\theta)\right)}{\sin(\theta)}$$

  - Gain

    $$G_{\max} = 1.64$$



Linear antennas (2)

- Radiation resistance of dipoles
  - Uniform current distribution

$$R_{\text{rad}}^{\text{uniform}} = 80\pi^2 (L_{\text{a}}/\lambda)^2$$

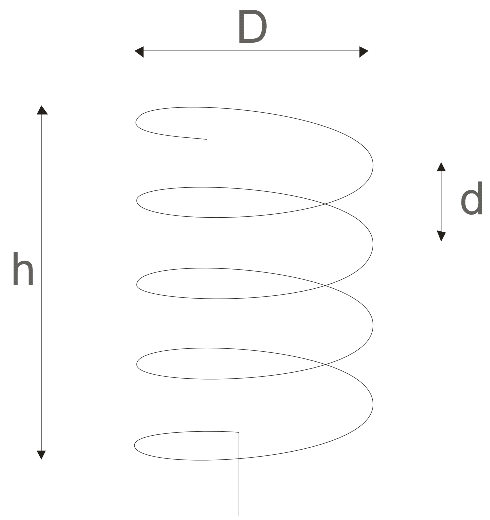
  - Tapered current distribution

$$R_{\text{rad}}^{\text{tapered}} = 0.25 R_{\text{rad}}^{\text{uniform}}$$

- Monopole over groundplane
  - Twice the gain of dipole
  - Half the radiation resistance of dipole

Helical antenna

- Combination of loop antenna and linear antenna
  - If dimensions much smaller than wavelength, behaves like linear antenna
  - Bandwidth, efficiency, and radiation resistance increase with increasing *h*



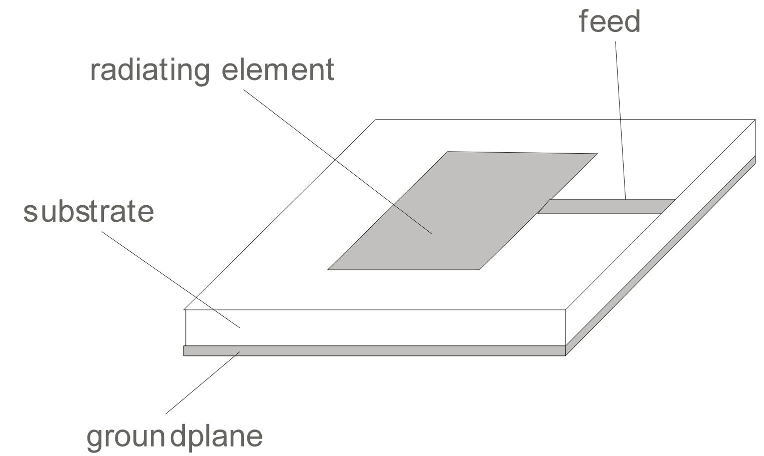
Microstrip antennas

- Dielectric substrate with ground plane on one side, and metallic patch on the other
- Properties determined by
  - Shape of patch: size must be at least

    $$L = 0.5\lambda_{\mathrm{substrate}}$$

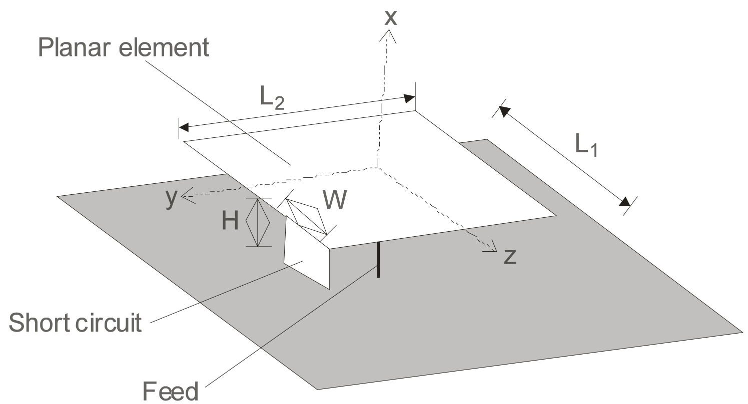
  - Dielectric properties of substrate

    $$\lambda_{\mathrm{substrate}} = \lambda_0/\sqrt{\varepsilon_{\mathrm{r}}}$$

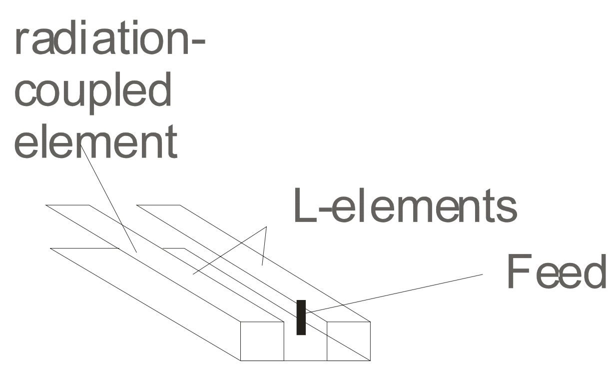


PIFA and RCDLA

- PIFA (Planar inverted F antenna)

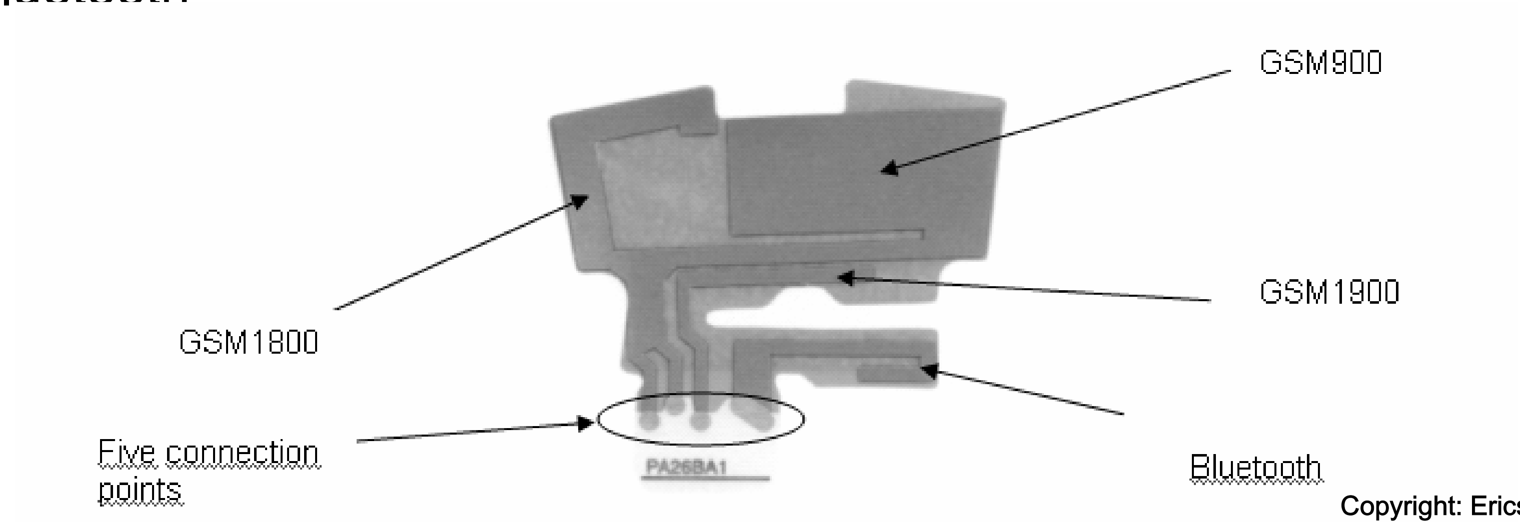


- RCDLA (Radiation-coupled dual-L antenna



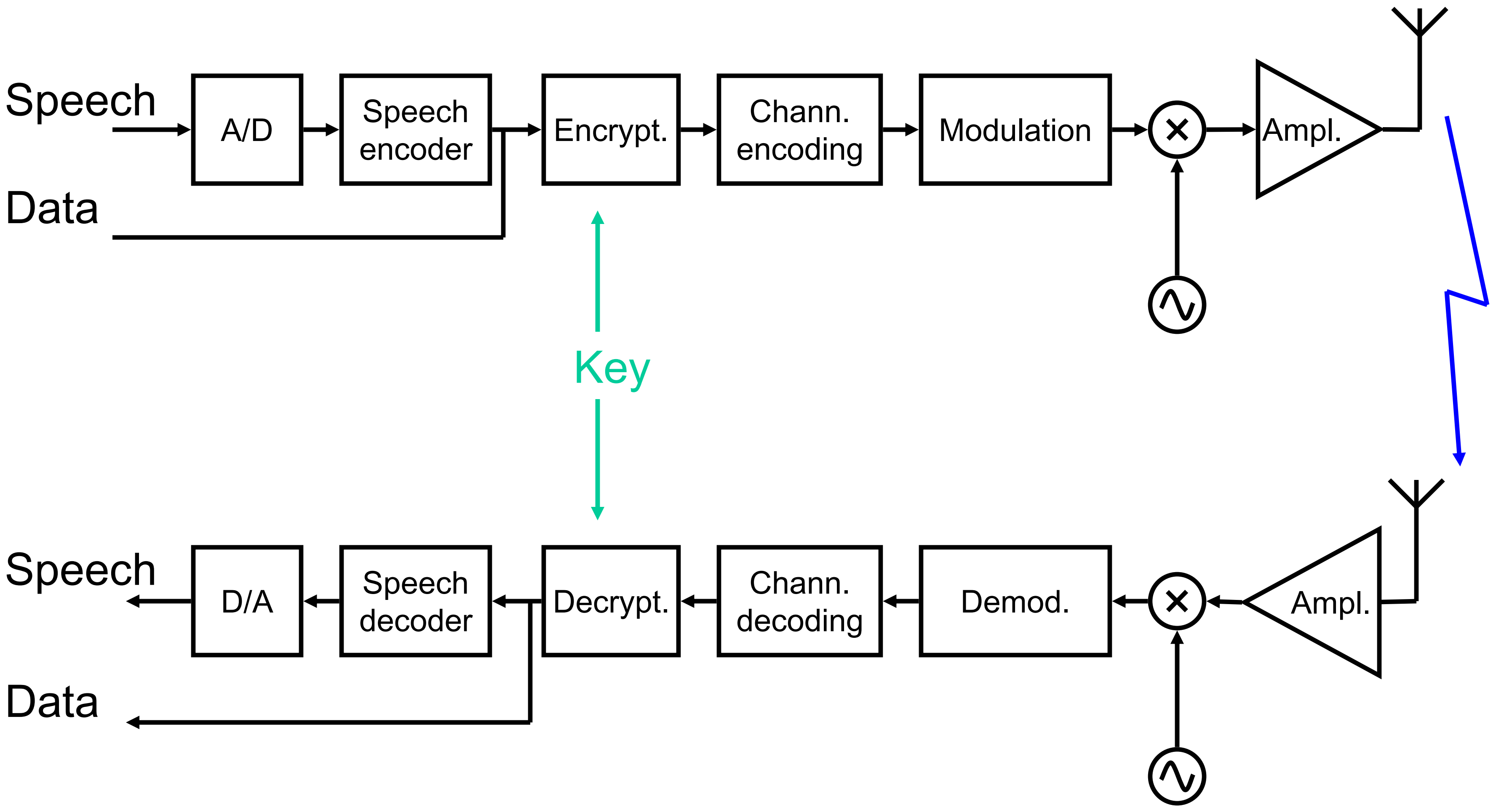
Multiband antennas

- For many applications, different wireless services need to be covered
- Example: cellular handset
  - GSM 900
  - GSM 1800
  - GSM 1900
  - Bluetooth

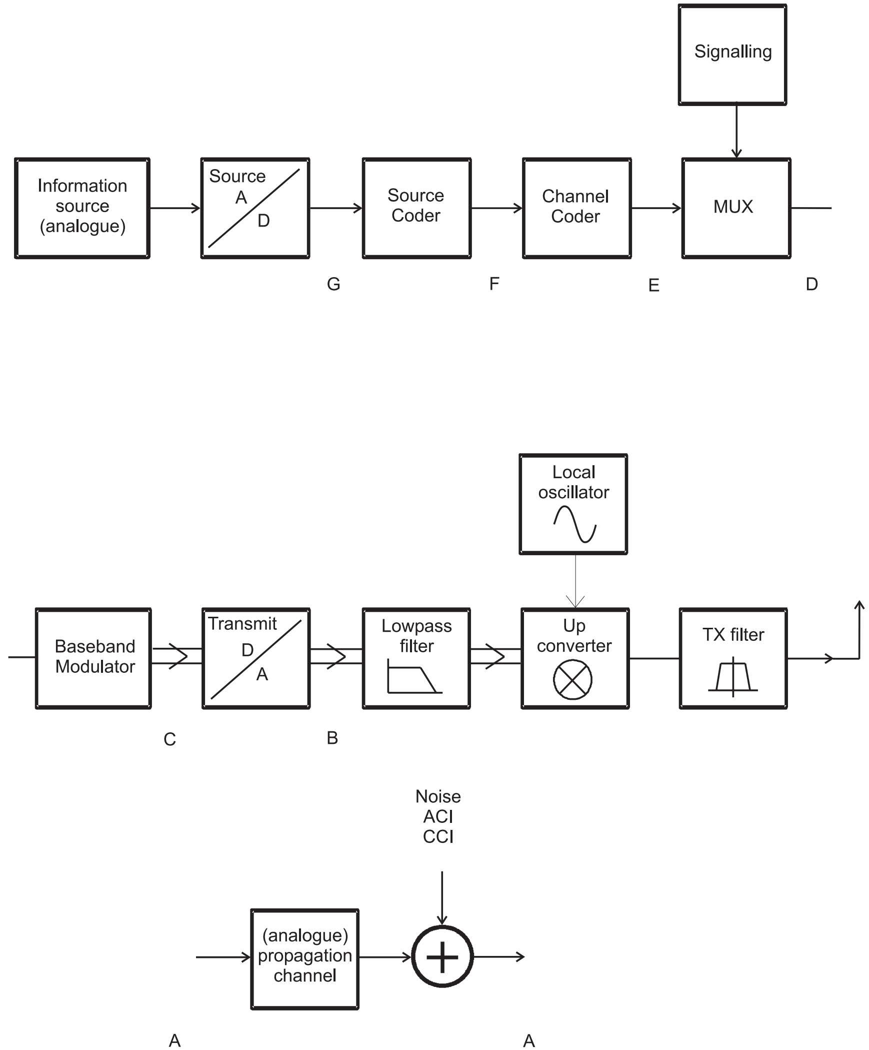


Structure of a wireless communications link

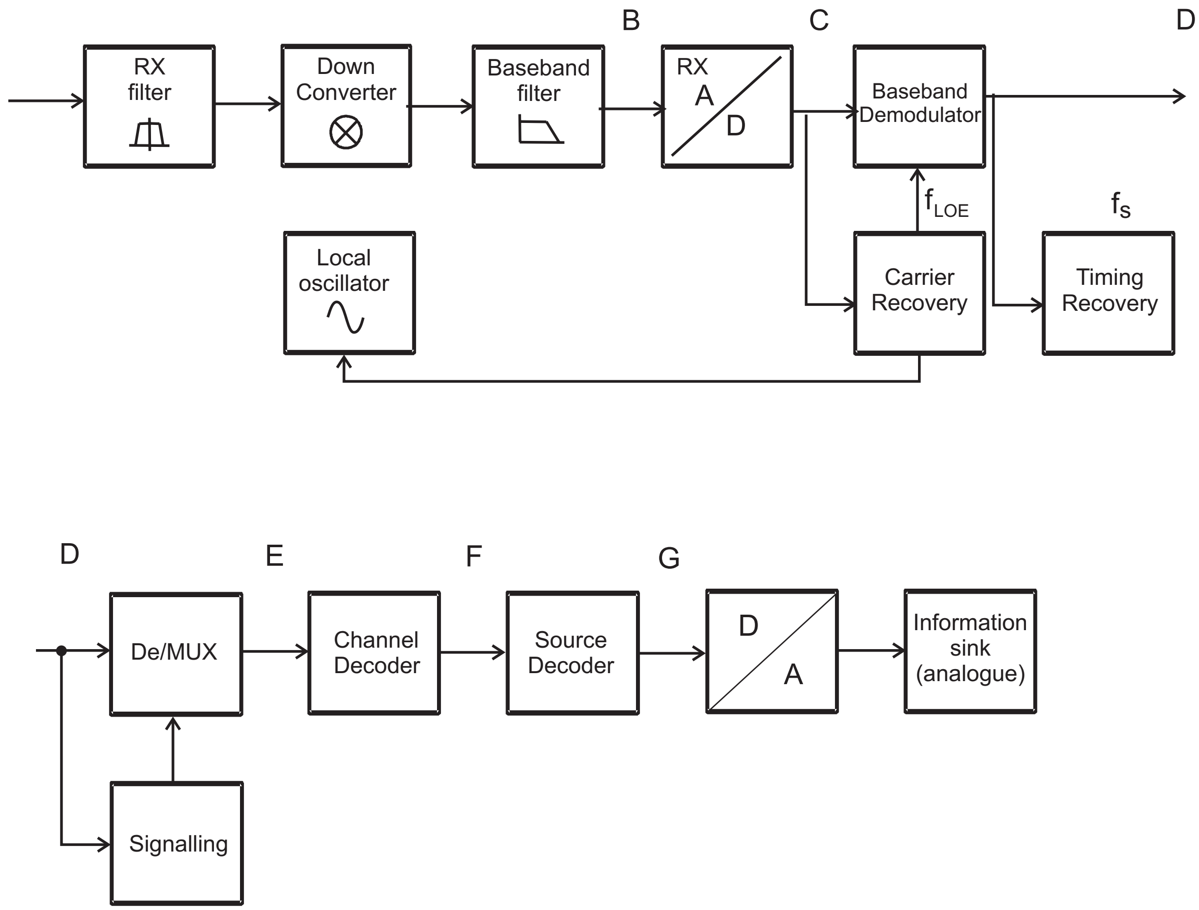
Block diagram



Block diagram transmitter



Block diagram receiver



Chapter 11

Modulation

RADIO SIGNALS AND COMPLEX NOTATION

Simple model of a radio signal

- A transmitted radio signal can be written

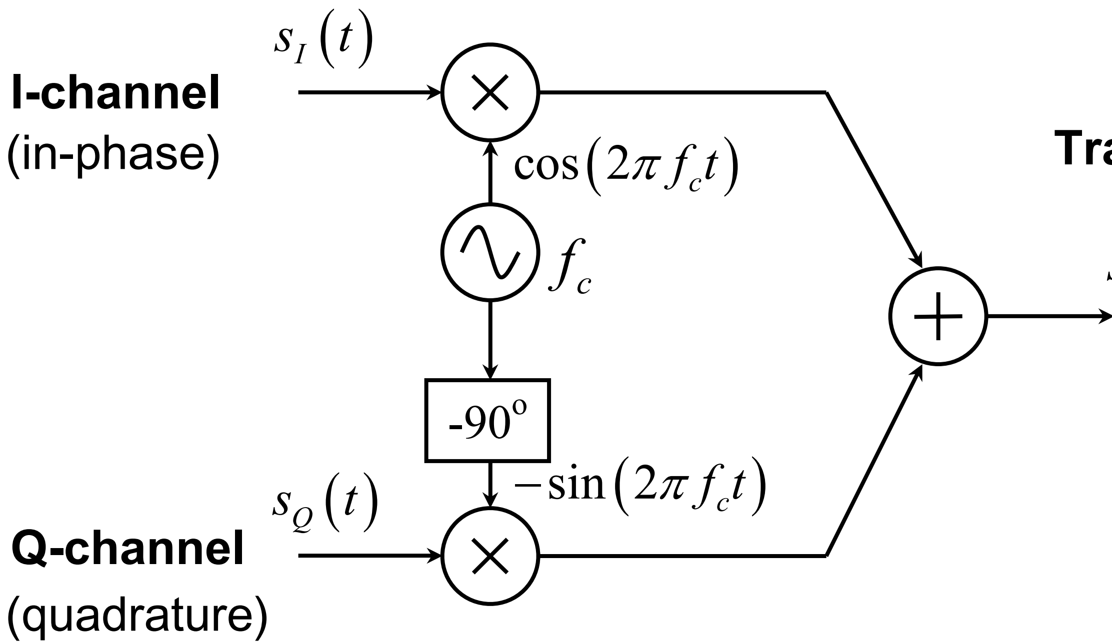
$$s(t) = \underbrace{A}_{\text{Amplitude}} \cos(2\pi \underbrace{f}_{\text{Frequency}} t + \underbrace{\phi}_{\text{Phase}})$$

- By letting the transmitted information change the amplitude, the frequency, or the phase, we get the tree basic types of digital modulation techniques

  – ASK (Amplitude Shift Keying)
  – FSK (Frequency Shift Keying)
  – PSK (Phase Shift Keying)

Constant amplitude

The IQ modulator



$$s(t) = s_I(t)\cos(2\pi f_c t) - s_Q(t)\sin(2\pi f_c t)$$

Take a step into the complex domain:

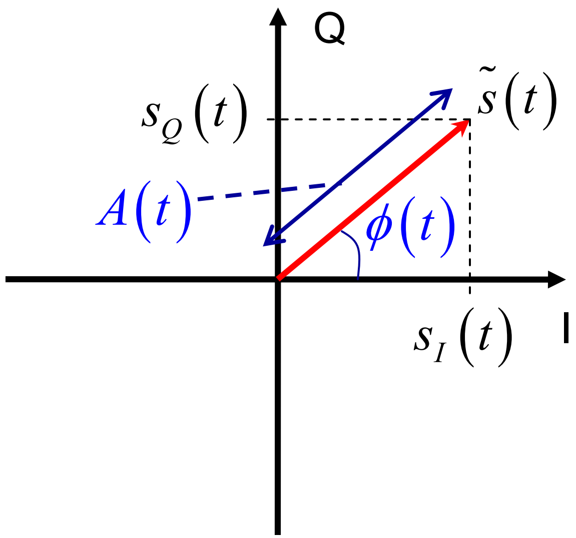
Complex envelope $\tilde{s}(t) = s_I(t) + j s_Q(t)$

Carrier factor $e^{j2\pi f_c t}$

$$\Rightarrow \quad s(t) = \mathrm{Re}\left\{\tilde{s}(t) e^{j2\pi f_c t}\right\}$$

Interpreting the complex notation

Complex envelope (phasor)



Polar coordinates:

$$\tilde{s}(t) = s_I(t) + j s_Q(t) = A(t) e^{j\phi(t)}$$

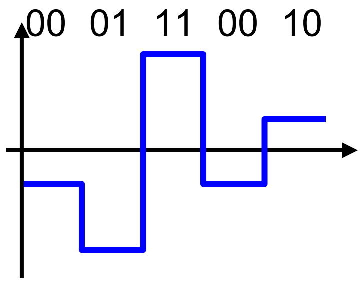
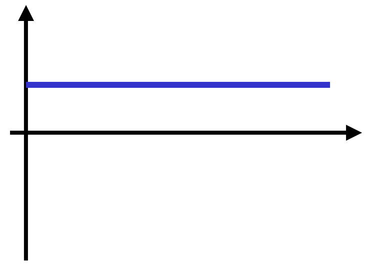
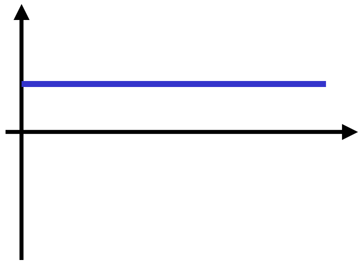
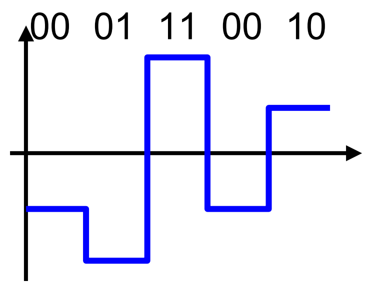
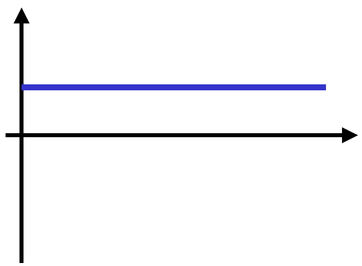
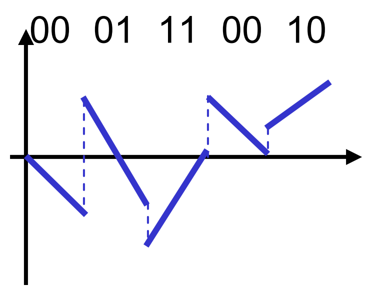
Transmitted radio signal

$$\begin{aligned} s(t) &= \mathrm{Re}\left\{ \tilde{s}(t) e^{j2\pi f_c t} \right\} \\ &= \mathrm{Re}\left\{ A(t) e^{j\phi(t)} e^{j2\pi f_c t} \right\} \\ &= \mathrm{Re}\left\{ A(t) e^{j\left(2\pi f_c t + \phi(t)\right)} \right\} \\ &= A(t) \cos\left(2\pi f_c t + \phi(t)\right) \end{aligned}$$

By manipulating the amplitude A(t) and the phase ϕ (t) of the complex envelope (phasor), we can create any type of modulation/radio signal.

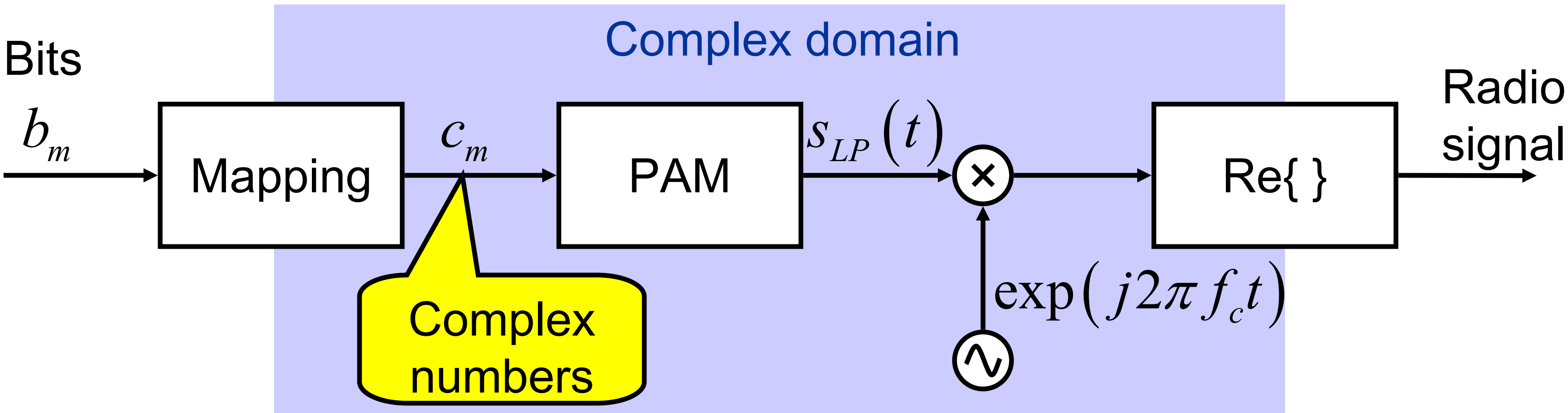
Example: Amplitude, phase and frequency modulation

$$s(t) = A(t)\cos\left(2\pi f_c t + \phi(t)\right)$$

	$A(t)$	$\phi(t)$	Comment:
4ASK			- Amplitude carries information - Phase constant (arbitrary)
4PSK			- Amplitude constant (arbitrary) - Phase carries information
4FSK			- Amplitude constant (arbitrary) - Phase slope (frequency) carries information

MODULATION BASICS

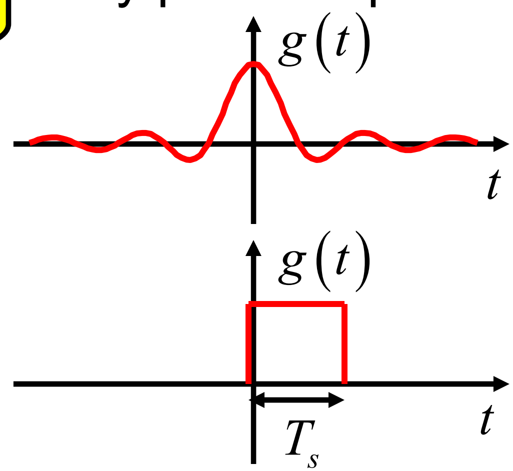
Pulse amplitude modulation (PAM)
The modulation process



PAM: $s_{LP}(t) = \sum_{m=-\infty}^{\infty} c_m g(t - mT_s)$

Symbol time

Many possible pulses



"Standard" basis pulse criteria

$\int_{-\infty}^{\infty} |g(t)|^2 \, dt = 1 \text{ or } = T_s$ (energy norm.)

$\int_{-\infty}^{\infty} g(t) g^*(t - mT_s) \, dt = 0, \; m \neq 0$ (orthogonality)

Pulse amplitude modulation (PAM)
Basis pulses and spectrum

Assuming that the complex numbers c_m representing the data are independent, then the power spectral density of the base band PAM signal becomes:

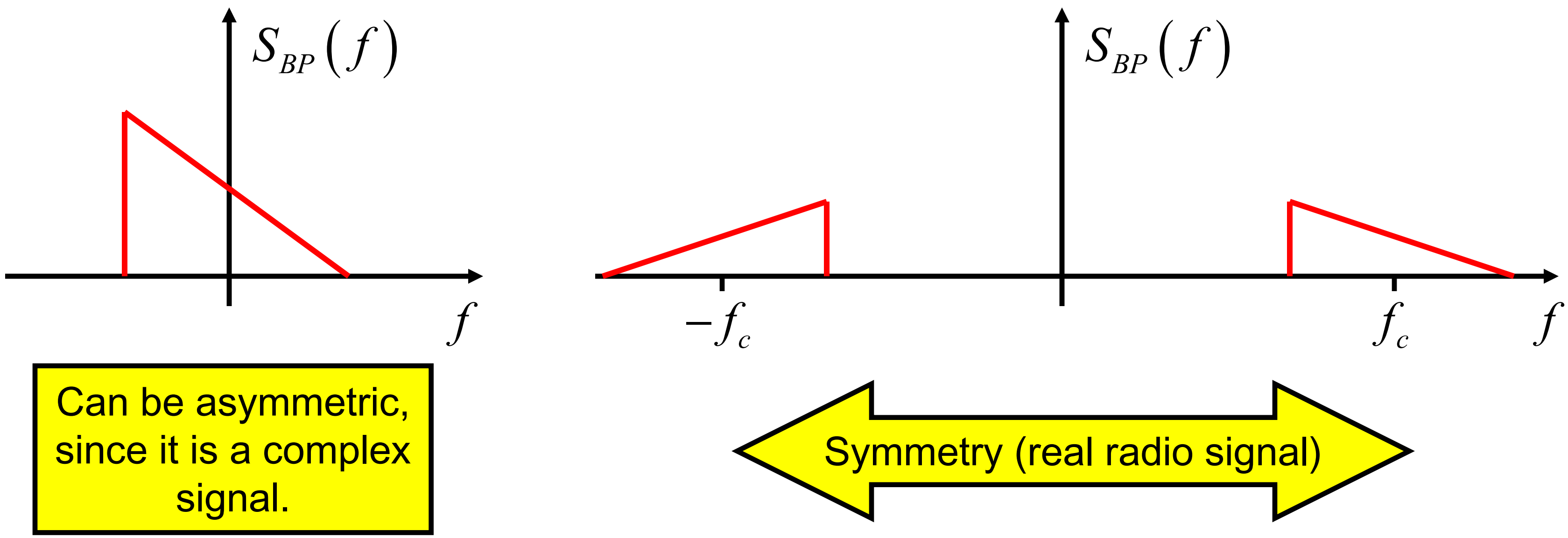
$$S_{LP}(f) \sim \left| \int_{-\infty}^{\infty} g(t) e^{-j2\pi ft} dt \right|^2$$

which translates into a radio signal (band pass) with

$$S_{BP}(f) = \frac{1}{2}\left(S_{LP}(f - f_c) + S_{LP}(-f - f_c) \right)$$

Pulse amplitude modulation (PAM) Basis pulses and spectrum

Illustration of power spectral density of the (complex) base-band signal, $S_{LP}(f)$, and the (real) radio signal, $S_{BP}(f)$.



What we need are basis pulses $g(t)$ with nice properties like:

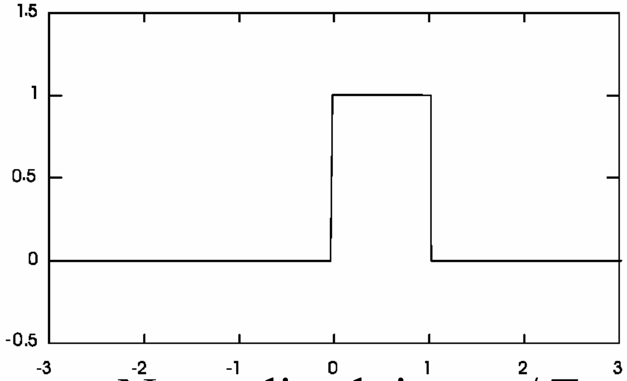
- Narrow spectrum (low side-lobes)
- Relatively short in time (low delay)

Basis pulses

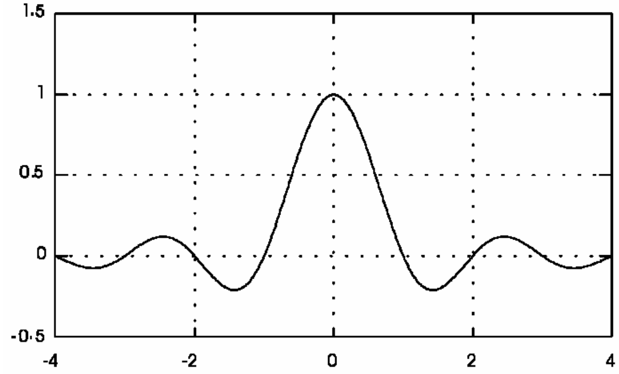
TIME DOMAIN

FREQ. DOMAIN

Rectangular [in time]

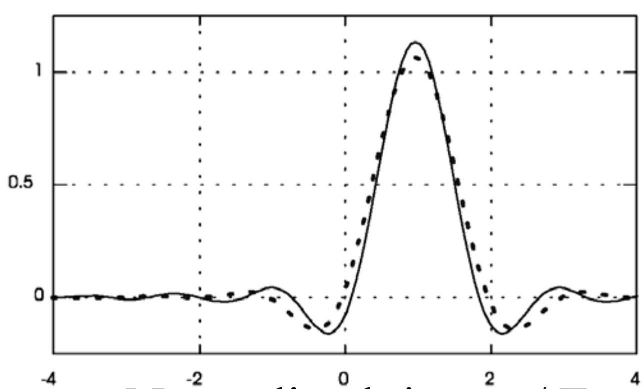


Normalized time t / T_s

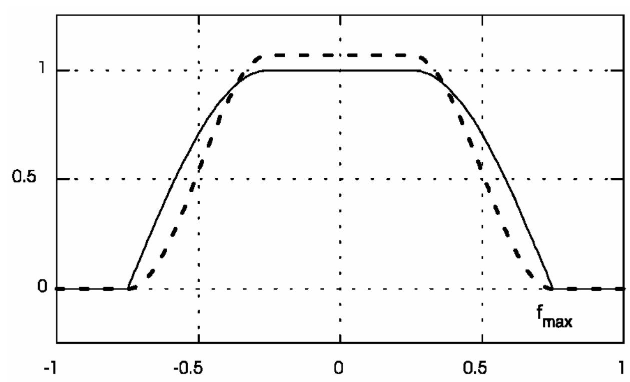


Normalized freq. $f \bullet T_s$

(Root-) Raised-cosine [in freq.]



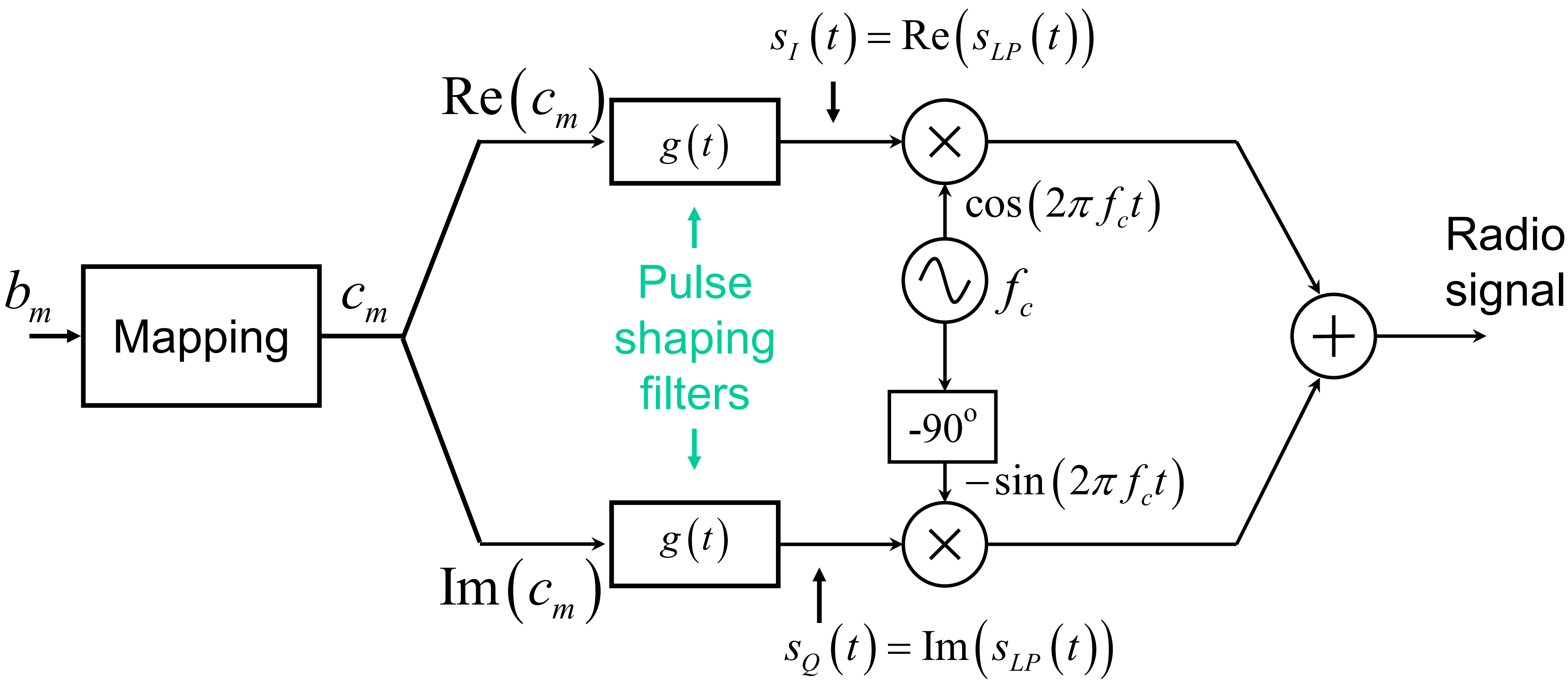
Normalized time t / T_s



Normalized freq. $f \bullet T_s$

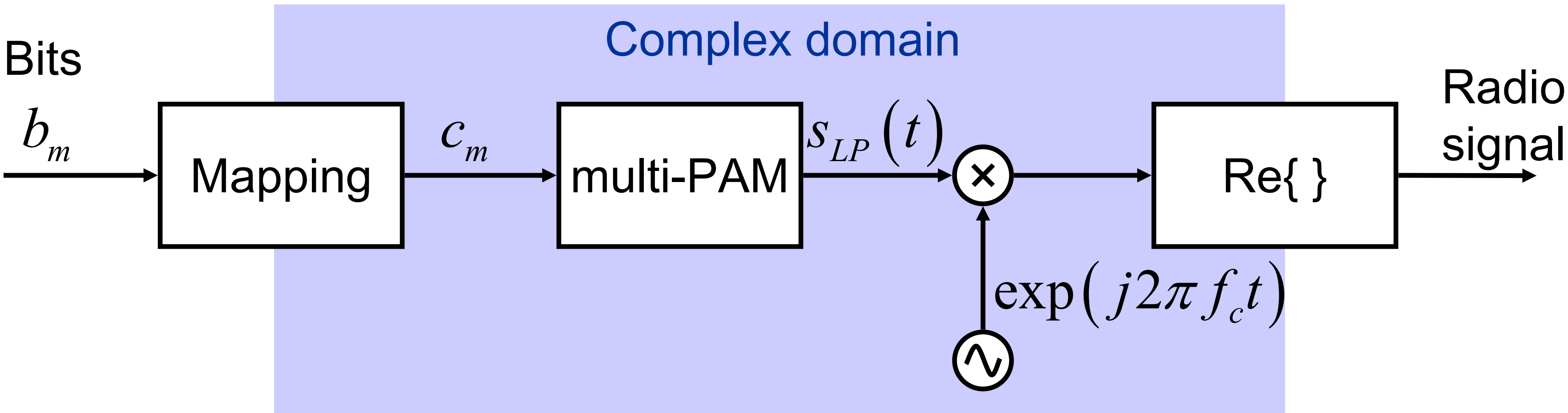
Pulse amplitude modulation (PAM)
Interpretation as IQ-modulator

For real valued basis functions $g(t)$ we can view PAM as:



(Both the rectangular and the (root-) raised-cosine pulses are real valued.)

Multi-PAM
Modulation with multiple pulses



$$\text{multi-PAM:} \quad s_{LP}(t) = \sum_{m=-\infty}^{\infty} g_{c_m}(t - mT_s)$$

Several different pulses

"Standard" basis pulse criteria

$$\int_{-\infty}^{\infty} \left| g_{c_m}(t) \right|^2 dt = 1 \text{ or } = T_s \qquad \text{(energy norm.)}$$

$$\int_{-\infty}^{\infty} g_{c_m}(t) g_{c_n}^*(t) dt = 0, \; c_m \neq c_n \qquad \text{(orthogonality)}$$

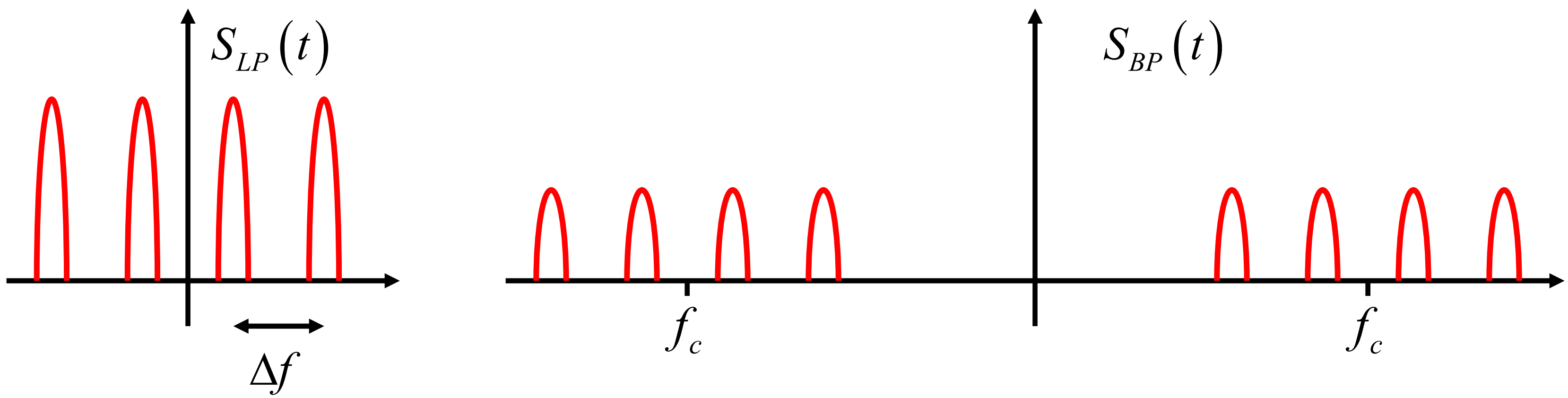
$$\int_{-\infty}^{\infty} g_{c_m}(t) g_{c_m}^*(t - nT_s) dt = 0, \; n \neq 0 \qquad \text{(orthogonality)}$$

Multi-PAM
Modulation with multiple pulses

Frequency-shift keying (FSK) with M (even) different transmission frequencies can be interpreted as multi-PAM if the basis functions are chosen as:

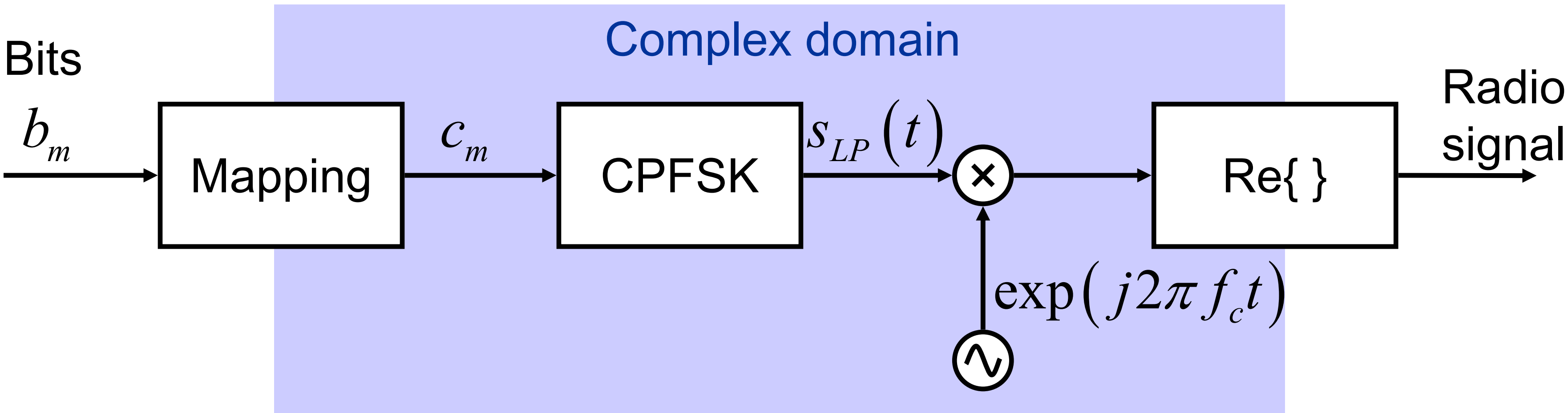
$$g_i(t) = \exp(-j\pi i \Delta f t) \text{ for } 0 \le t \le T_s$$

for i = +/- 1, +/-3, ... , +/-M/2



Bits: 00 01 10 11

Continuous-phase FSK (CPFSK)
The modulation process



CPFSK: $$s_{LP}(t) = A\exp\left(j\Phi_{CPFSK}(t)\right)$$

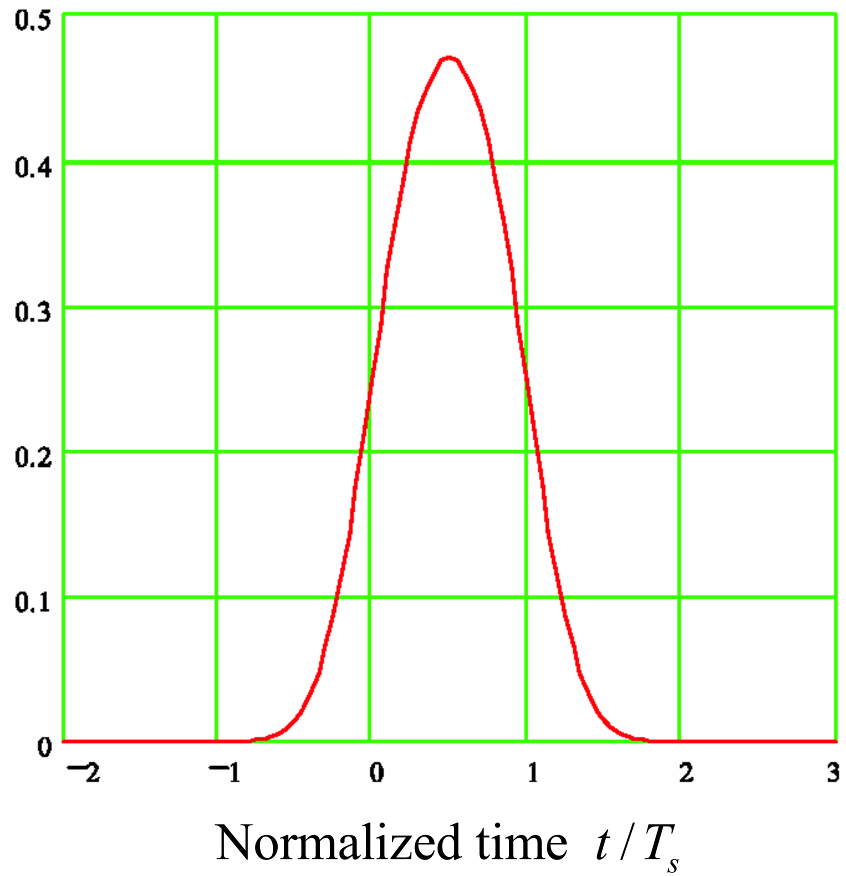
where the amplitude A is constant and the phase is

$$\Phi_{CPFSK}(t) = 2\pi h_{\text{mod}} \sum_{m=-\infty}^{\infty} c_m \int_{-\infty}^{t} \tilde{g}(u - mT)\,du$$

Phase basis pulse

where h_{mod} is the modulation index.

Continuous-phase FSK (CPFSK)
The Gaussian phase basis pulse



BT_s=0.5

SIGNAL SPACE DIAGRAM

Principle of signal-space diagram (1)

- Represent a continuous signal by a discrete vector
- Choice of expansion functions:
  - In passband, usually

$$\varphi_{\mathrm{BP},1}(t) = \sqrt{\frac{2}{T_{\mathrm{S}}}}\cos(2\pi f_{\mathrm{c}} t)$$

$$\varphi_{\mathrm{BP},2}(t) = \sqrt{\frac{2}{T_{\mathrm{S}}}}\sin(2\pi f_{\mathrm{c}} t) \quad .$$

  - In baseband, usually

$$\varphi_1(t) = \sqrt{\frac{1}{T_{\mathrm{S}}}} \cdot 1$$

$$\varphi_2(t) = \sqrt{\frac{1}{T_{\mathrm{S}}}} \cdot j.$$

Principle of signal-space diagram (2)

- Signal vector for m-th signal

$$s_{m,n} = \int_0^{T_S} s_m(t)\varphi_n^*(t)dt$$

- Energy contained in signal

$$E_{S,m} = \int_0^{T_S} s_{BP,m}{}^2(t)dt = \|\mathbf{s}_{BP,m}\|^2$$

$$E_{S,m} \approx \frac{1}{2}\int_0^{T_S}\|s_{LP,m}(t)\|^2 dt = \frac{1}{2}\|\mathbf{s}_{LP,m}\|^2$$

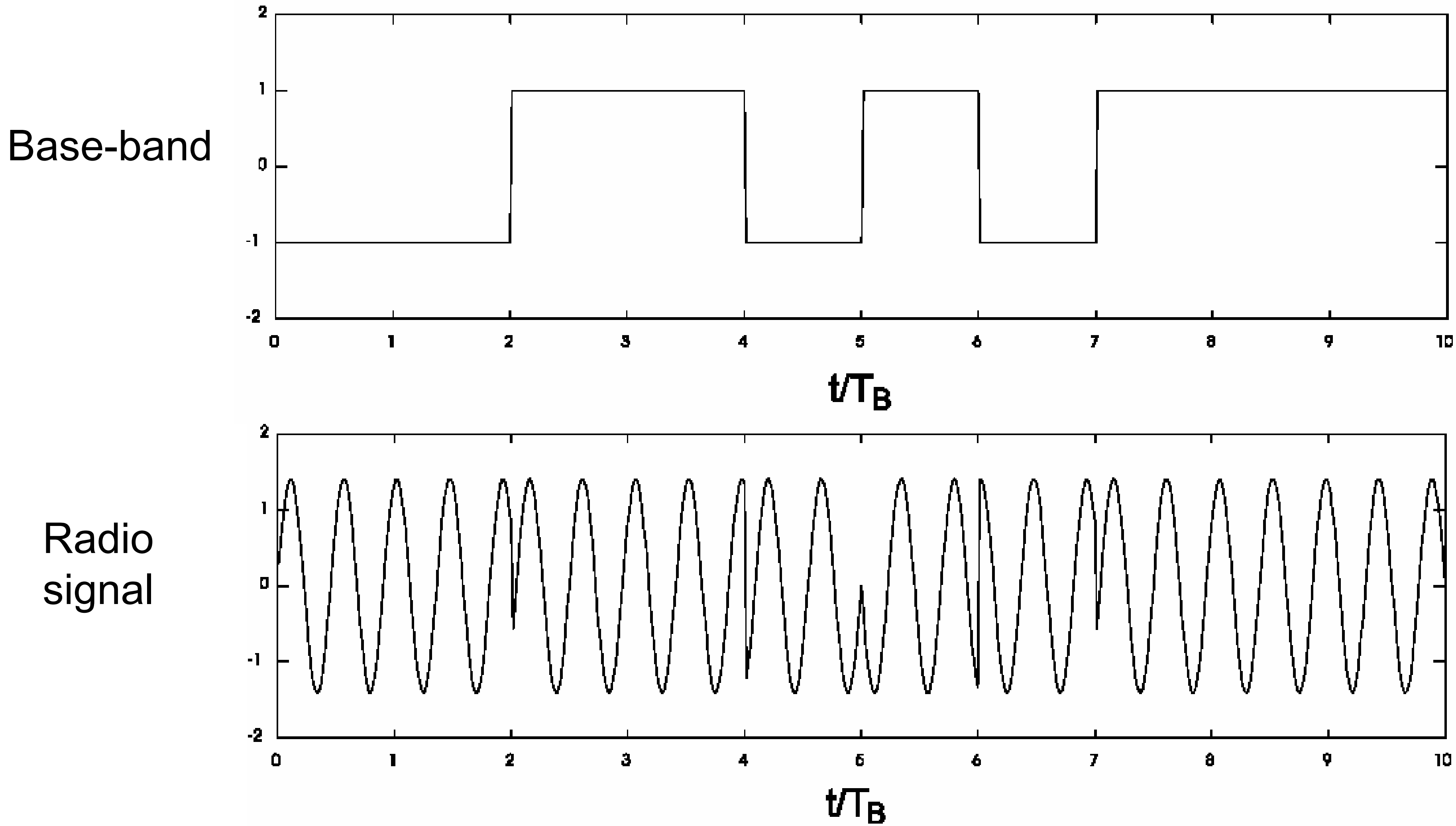
- Correlation coefficients between signals *k* and *m*

$$\mathrm{Re}\{\rho_{k,m}\} = \frac{\mathbf{s}_{BP,m}\mathbf{s}_{BP,k}}{\|\mathbf{s}_{BP,m}\|\|\mathbf{s}_{BP,k}\|}$$

- *Take care about normalization BP vs. LP*

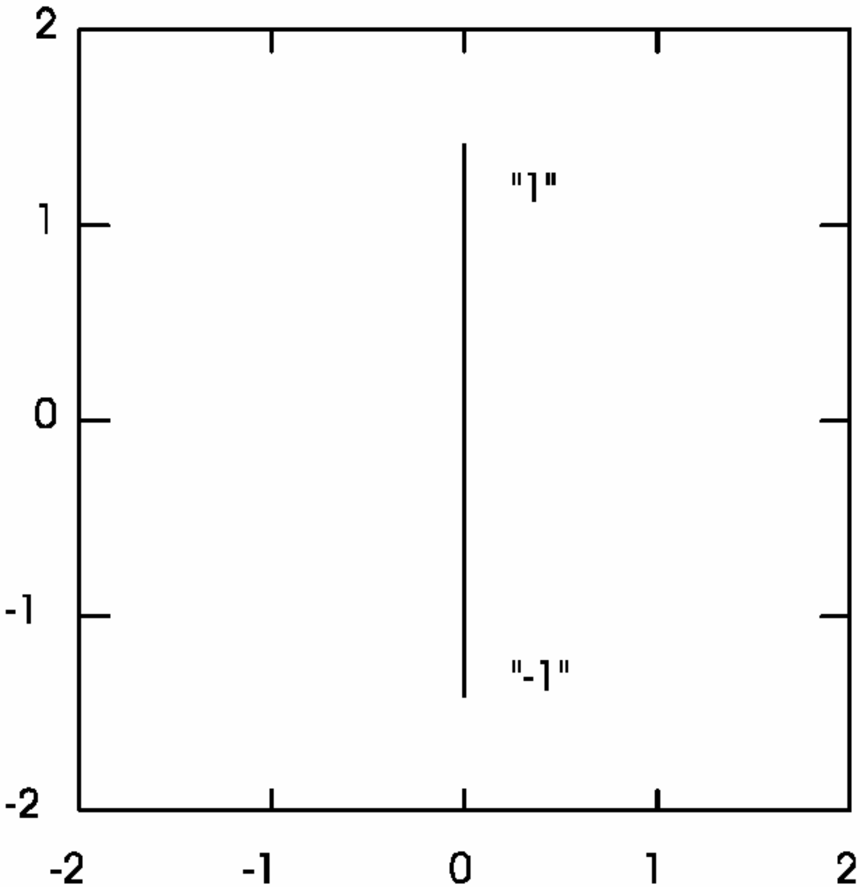
IMPORTANT MODULATION FORMATS

Binary phase-shift keying (BPSK) Rectangular pulses

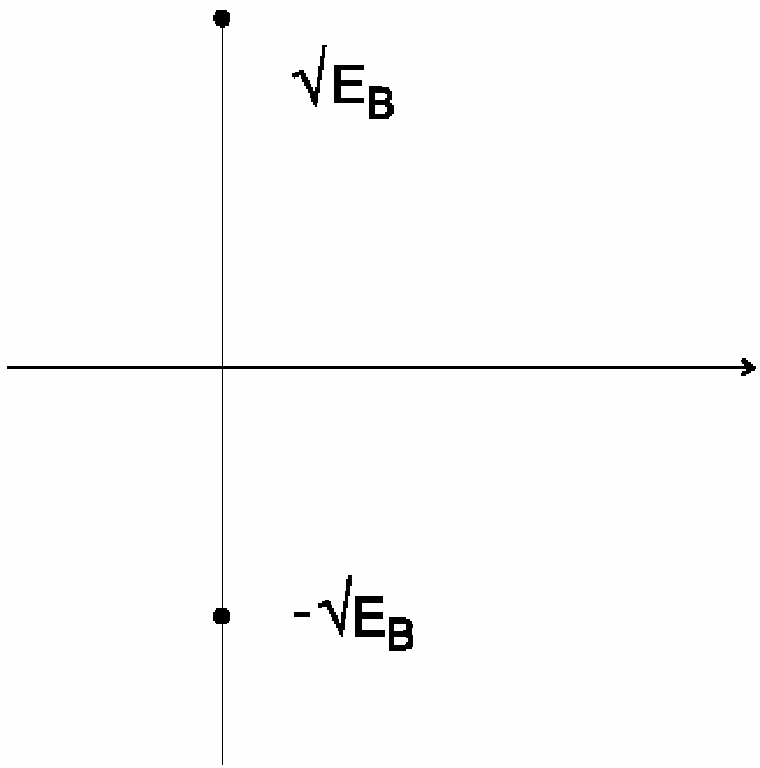


Binary phase-shift keying (BPSK)
Rectangular pulses

Complex representation

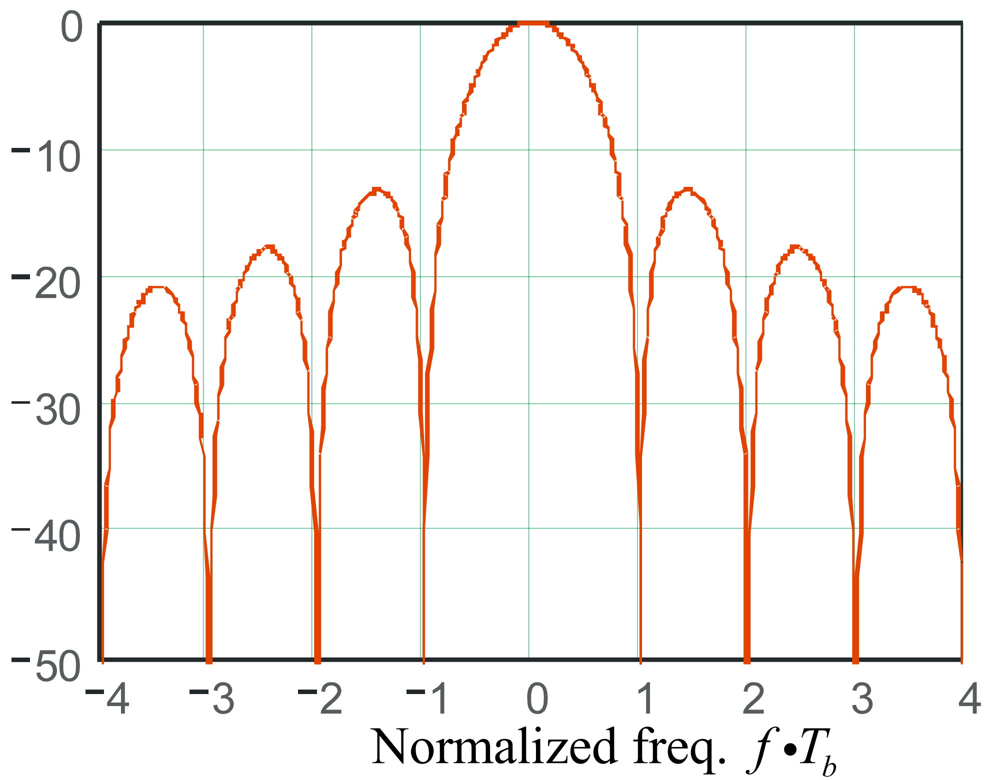


Signal space diagram



Binary phase-shift keying (BPSK) Rectangular pulses

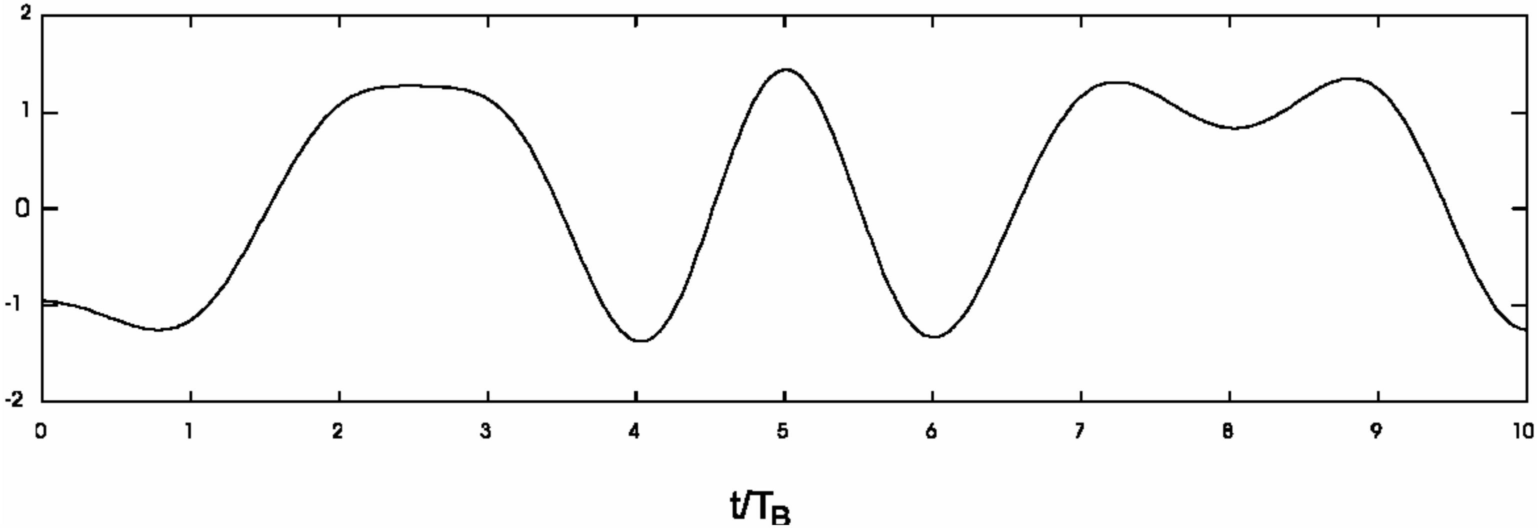
Power spectral density for BPSK



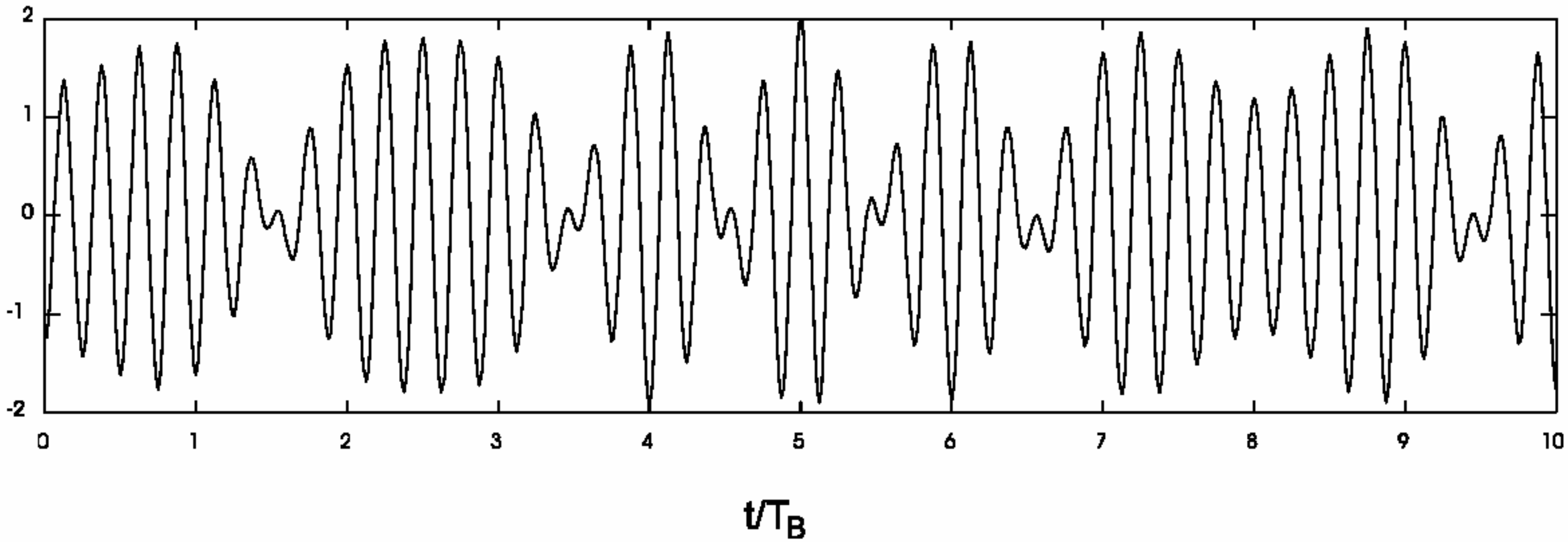
Contained percentage of total energy	spectral efficiency
90%	$0.59 Bit/s/Hz$
99%	$0.05 Bit/s/Hz$

Binary amplitude modulation (BAM)
Raised-cosine pulses (roll-off 0.5)

Base-band

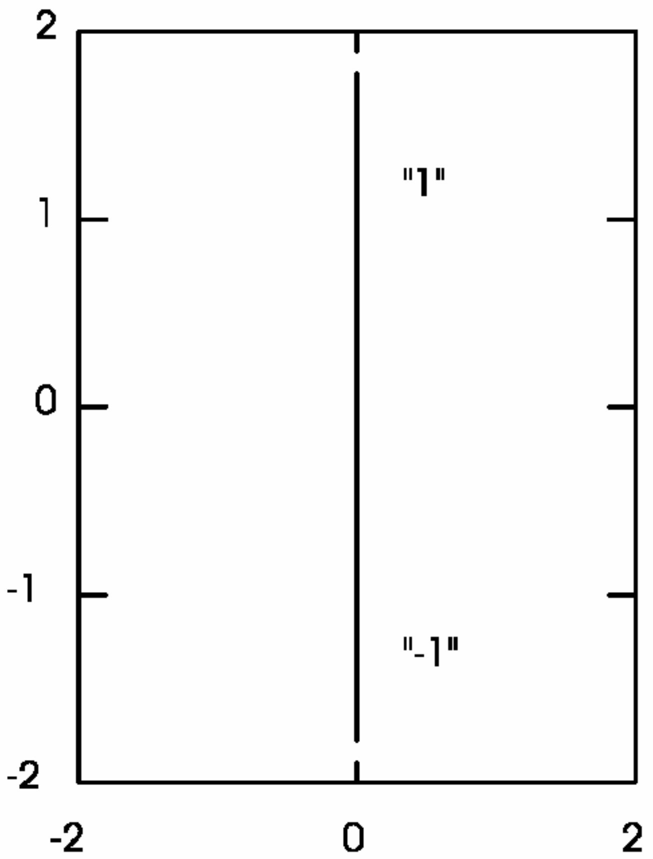


Radio signal

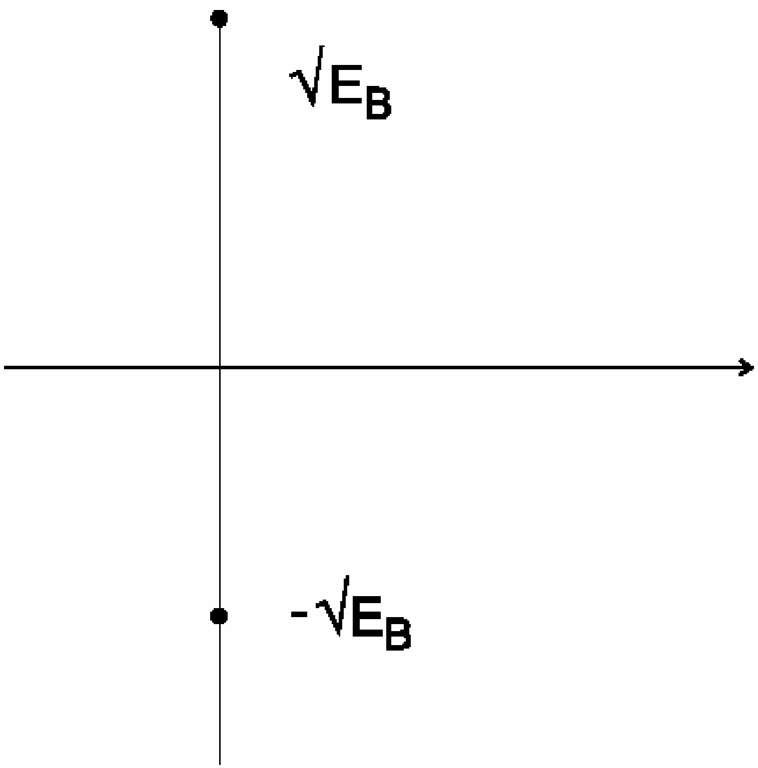


Binary amplitude modulation (BAM)
Raised-cosine pulses (roll-off 0.5)

Complex representation

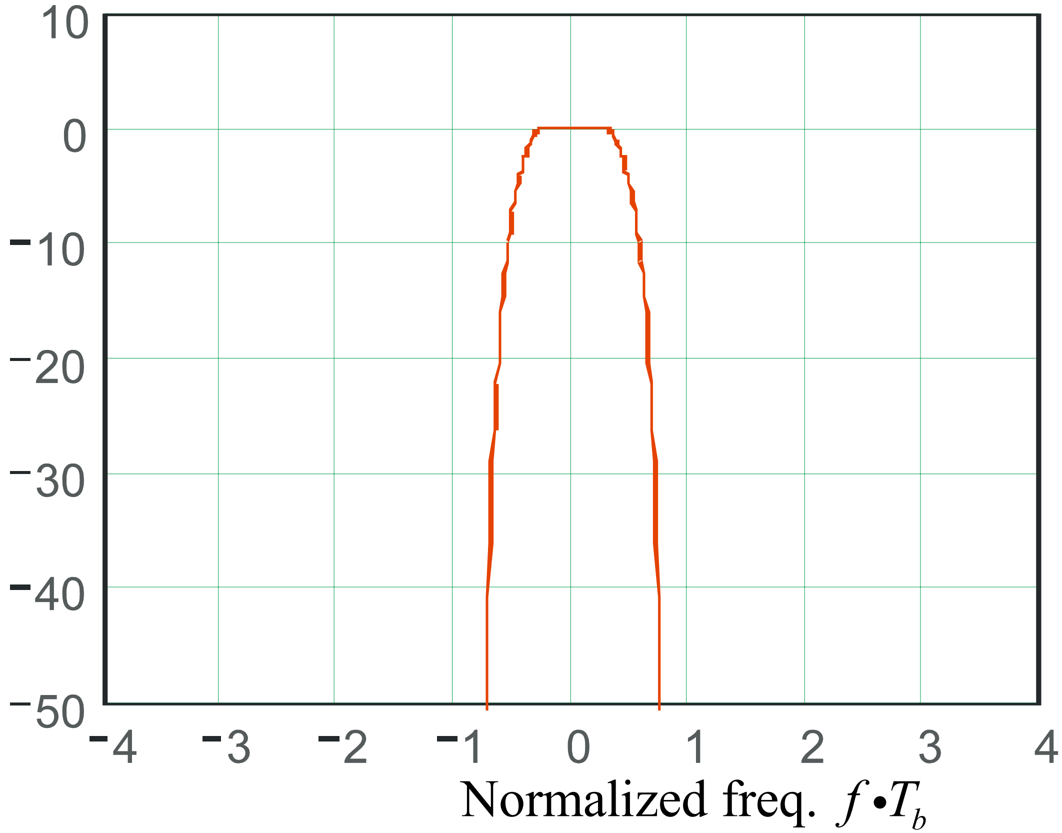


Signal space diagram



Binary amplitude modulation (BAM) Raised-cosine pulses (roll-off 0.5)

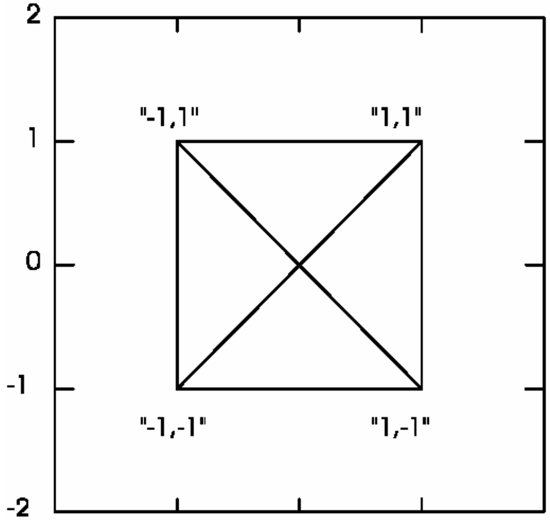
Power spectral density for BAM



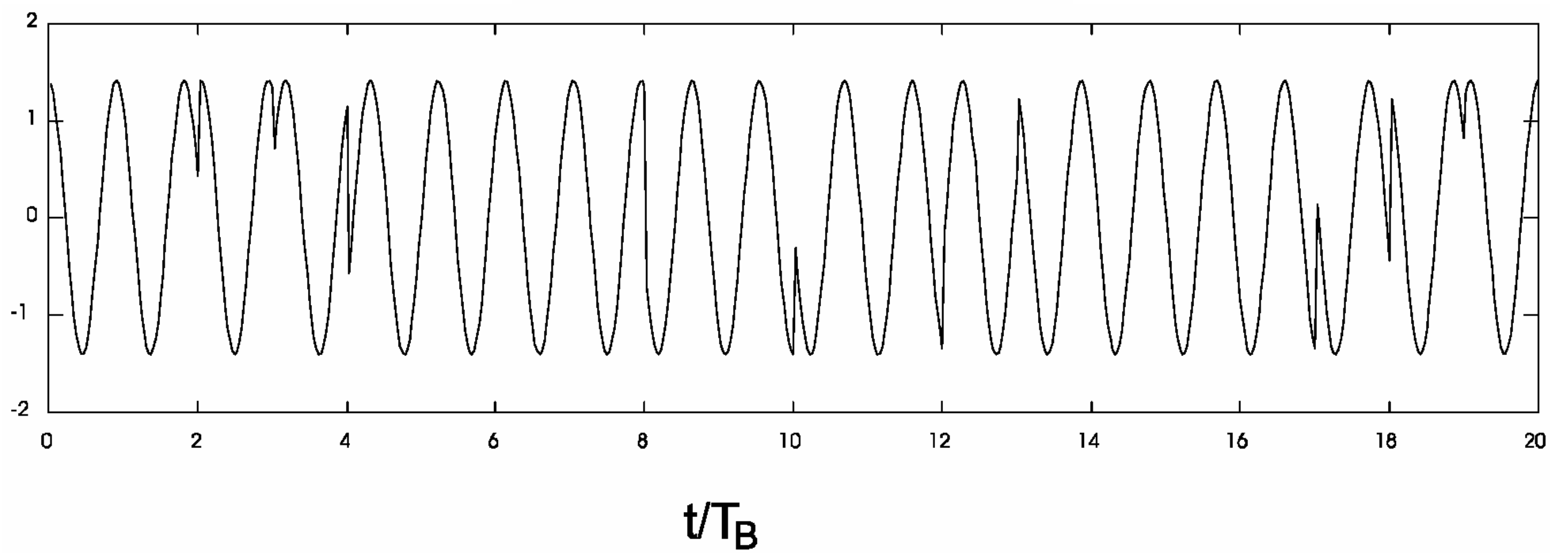
Contained percentage of total energy	spectral efficiency
90%	$1.02 Bit/s/Hz$
99%	$0.79 Bit/s/Hz$

Quaternary PSK (QPSK or 4-PSK) Rectangular pulses

Complex representation

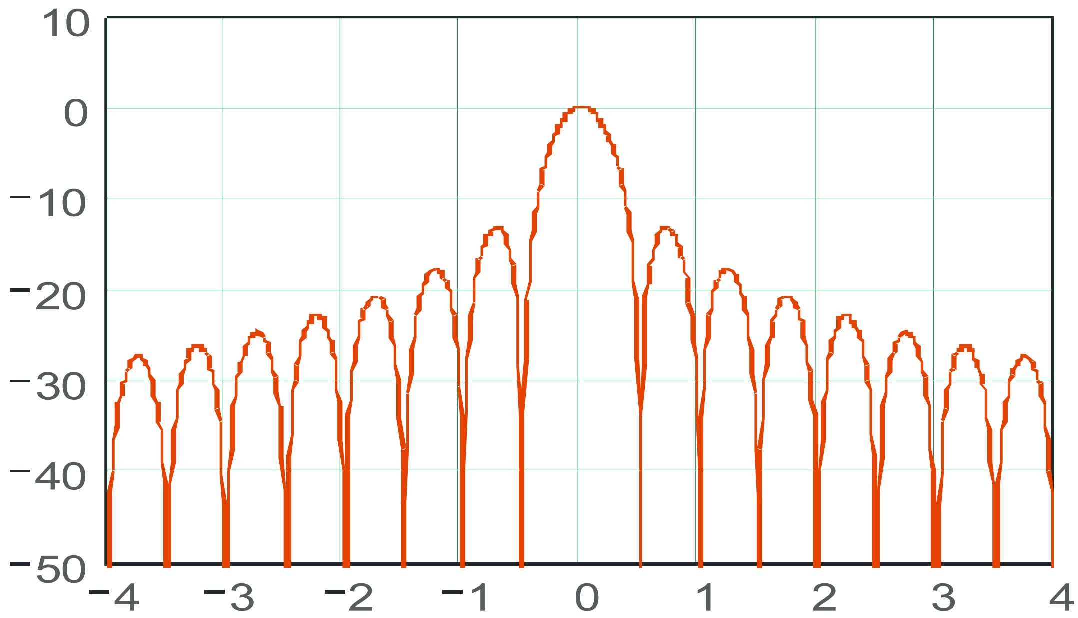


Radio signal



Quaternary PSK (QPSK or 4-PSK) Rectangular pulses

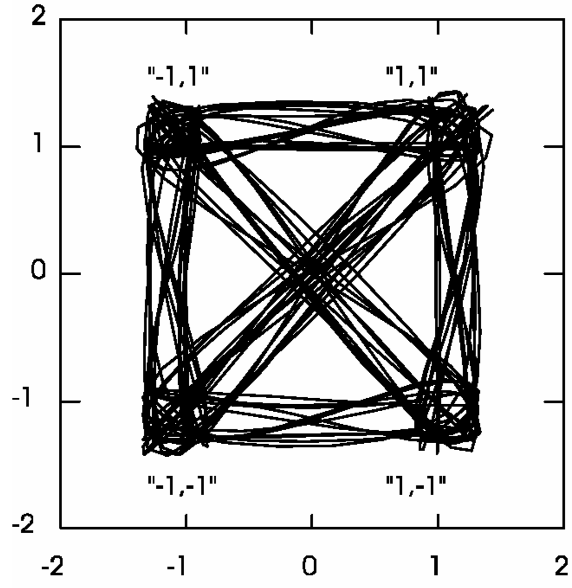
Power spectral density for QPSK



Contained percentage of total energy	spectral efficiency
90%	$1,18 Bit/s/Hz$
99%	$0.10 Bit/s/Hz$

Quadrature ampl.-modulation (QAM) Root raised-cos pulses (roll-off 0.5)

Complex representation

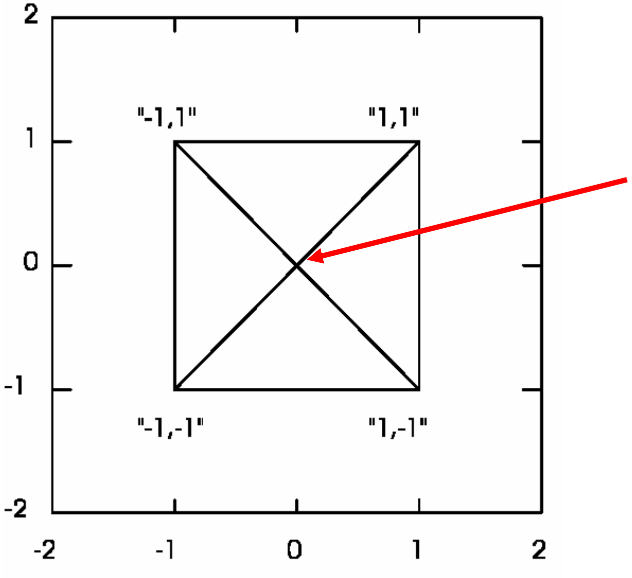


Contained percentage of total energy	spectral efficiency
90%	$2.04 Bit/s/Hz$
99%	$1.58 Bit/s/Hz$

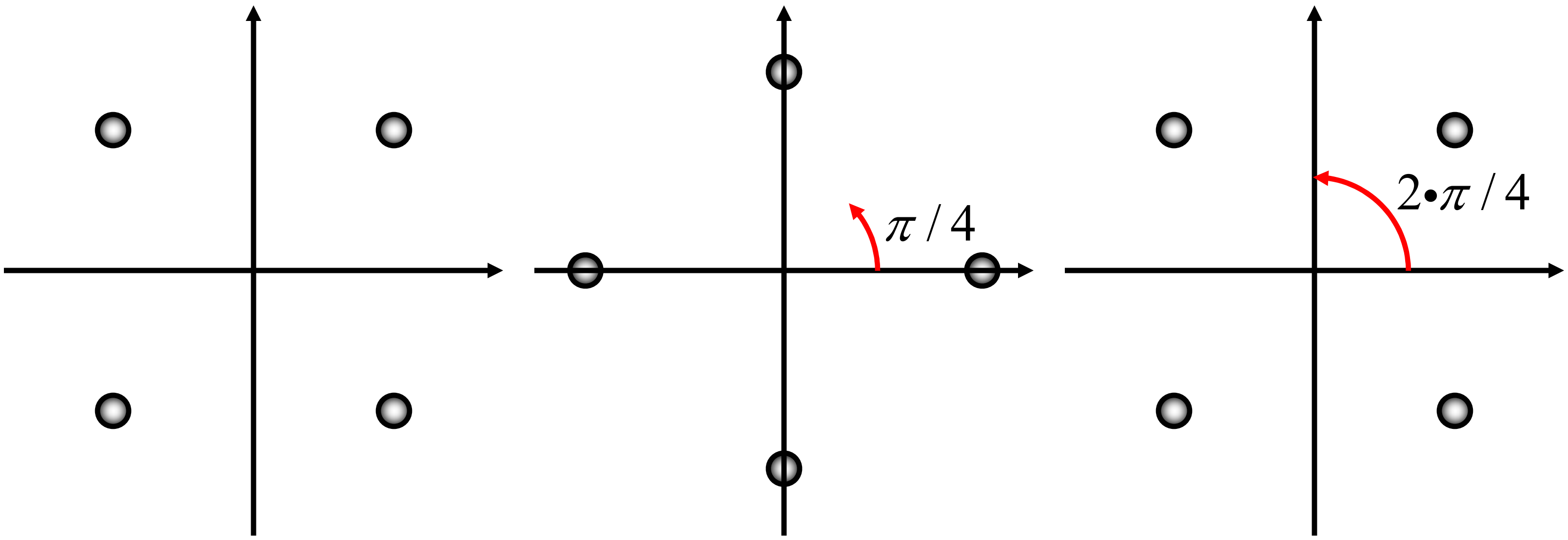
Amplitude variations
The problem

Signals with high amplitude variations leads to less efficient amplifiers.

Complex representation of QPSK



Amplitude variations
A solution

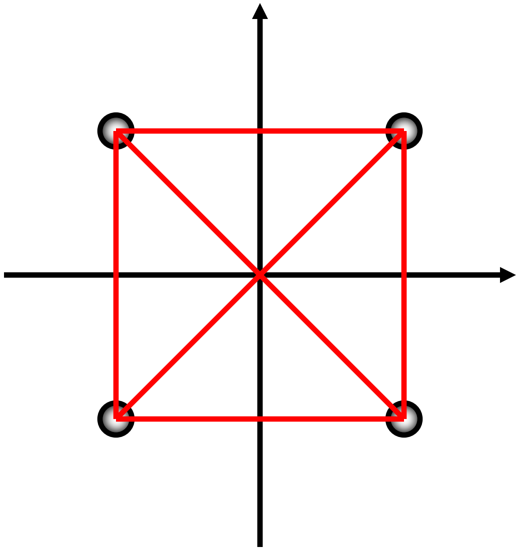


etc.

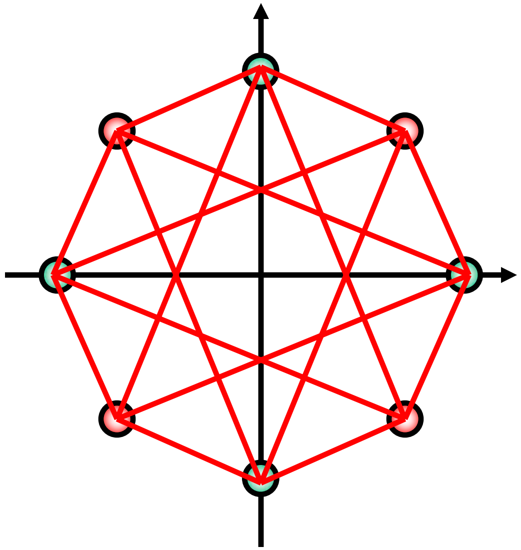
Amplitude variations
A solution

Looking at the complex representation ...

QPSK without rotation



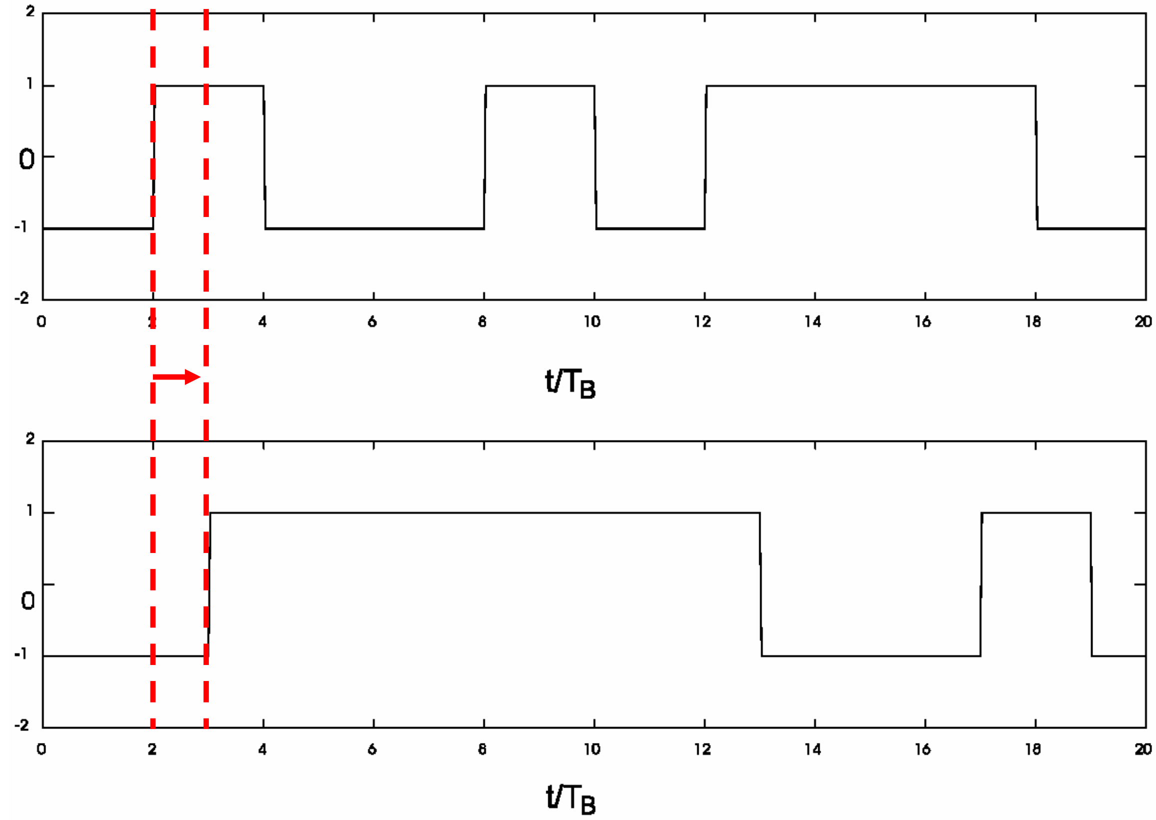
QPSK with rotation



Offset QPSK (OQPSK)
Rectangular pulses

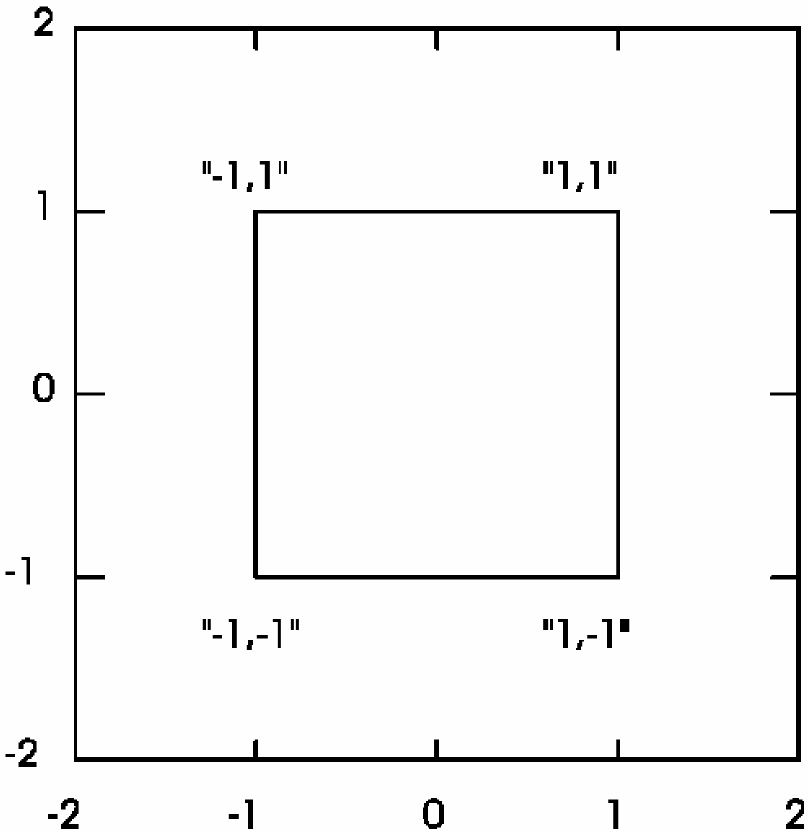
In-phase signal

Quadrature signal



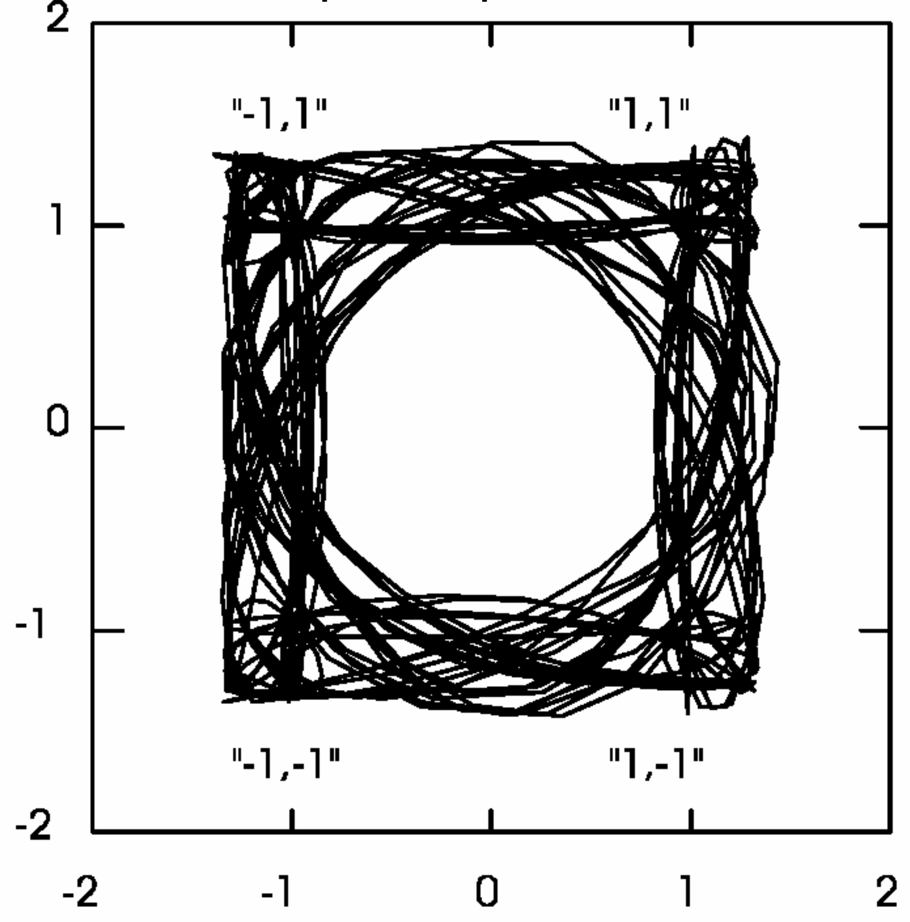
Offset QPSK
Rectangular pulses

Complex representation



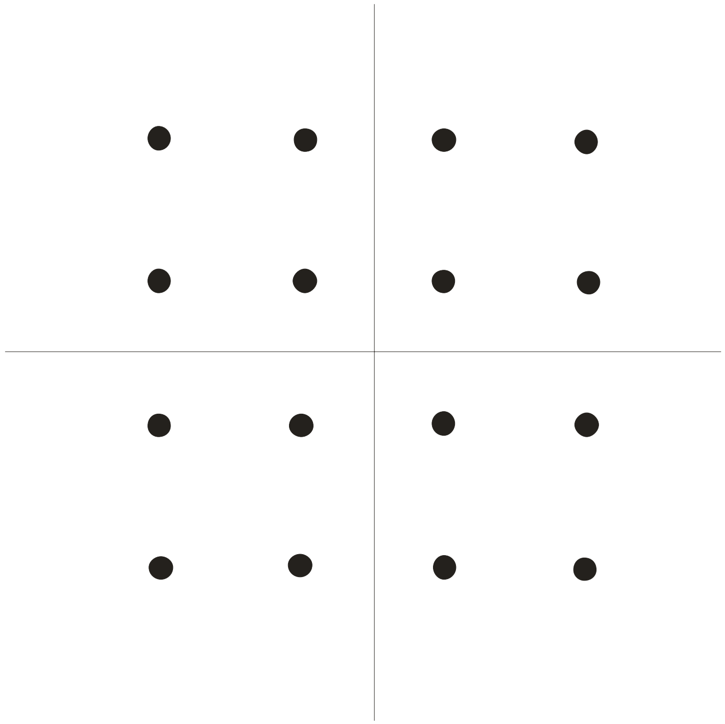
Offset QAM (OQAM)
Raised-cosine pulses

Complex representation



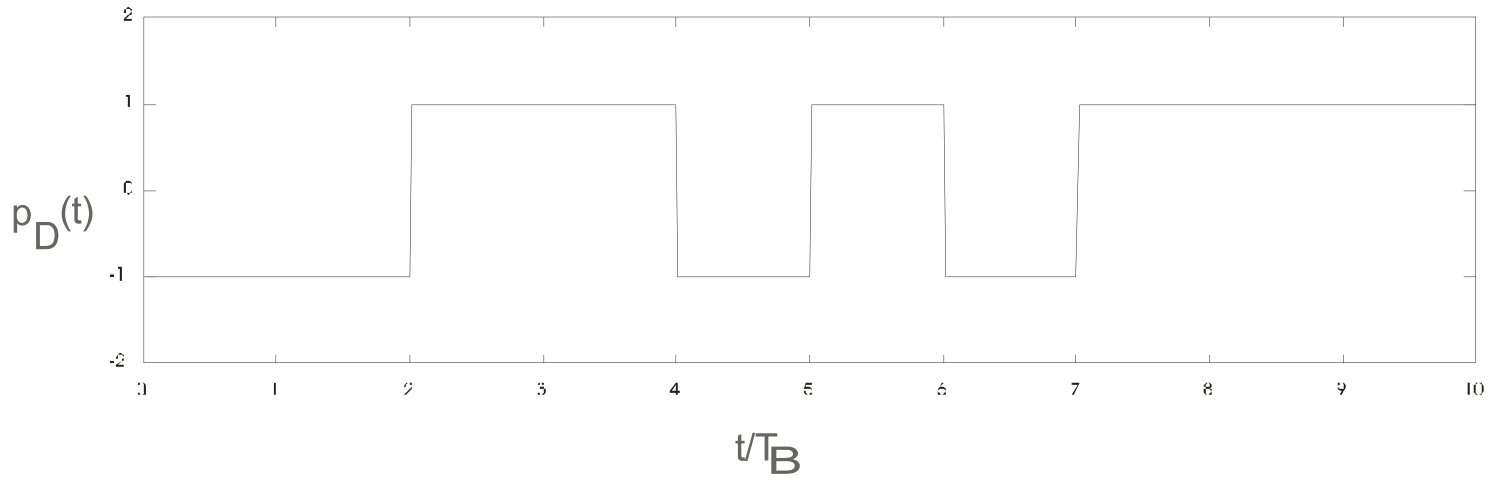
Higher-order modulation

16-QAM signal space diagram

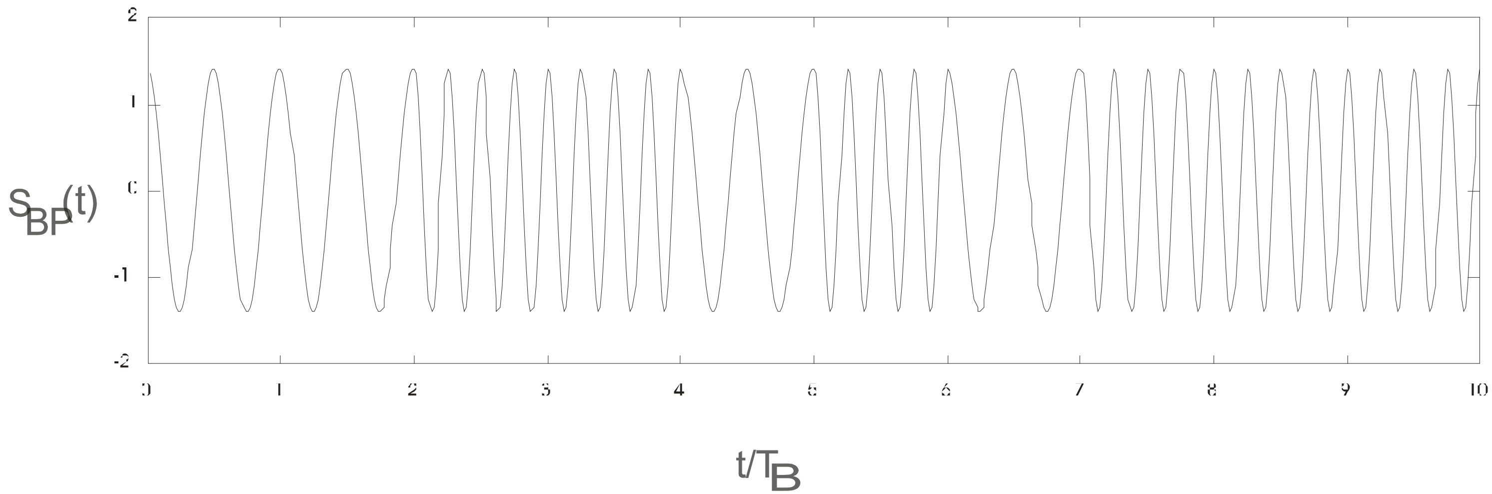


Binary frequency-shift keying (BFSK)
Rectangular pulses

Base-band

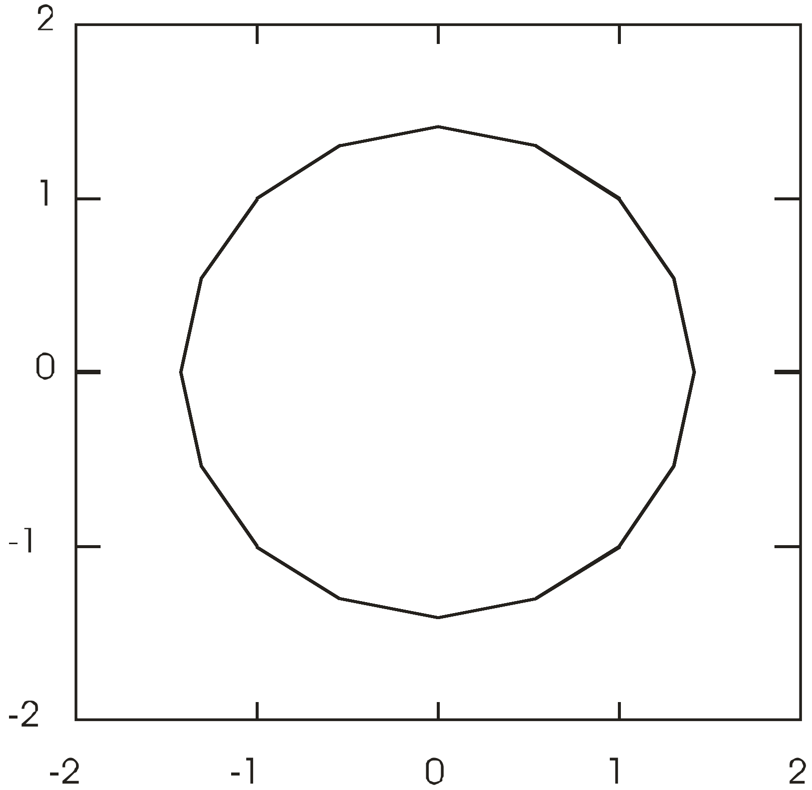


Radio signal

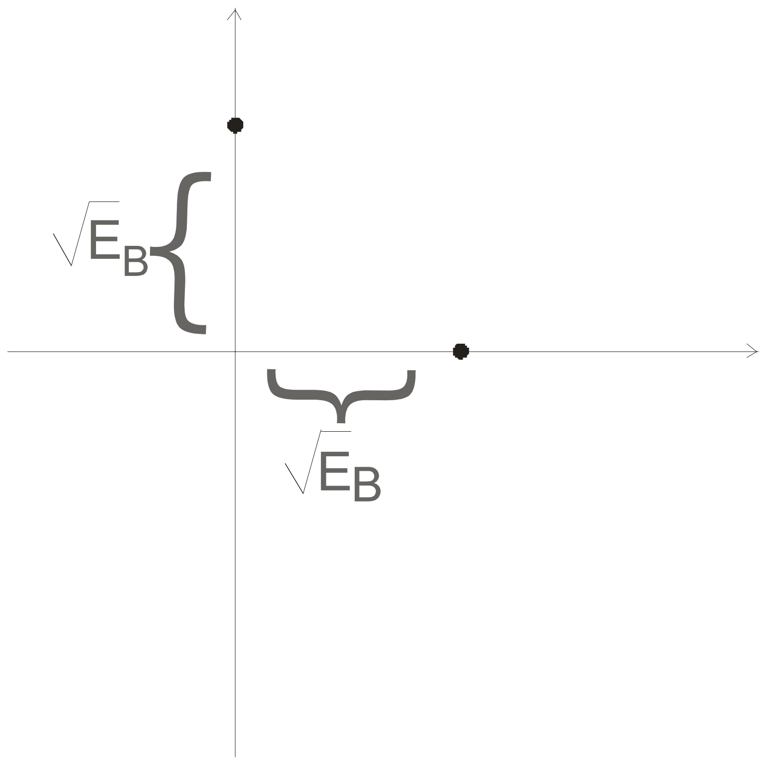


Binary frequency-shift keying (BFSK)
Rectangular pulses

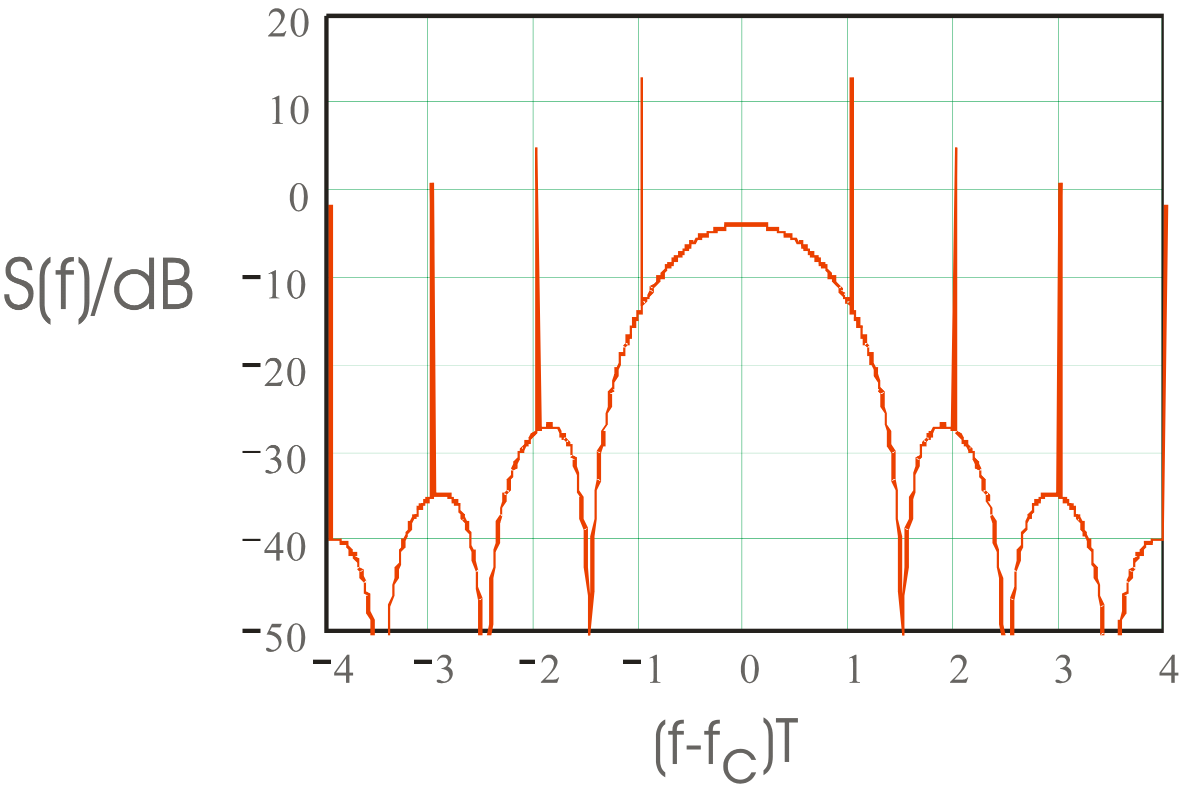
Complex representation



Signal space diagram



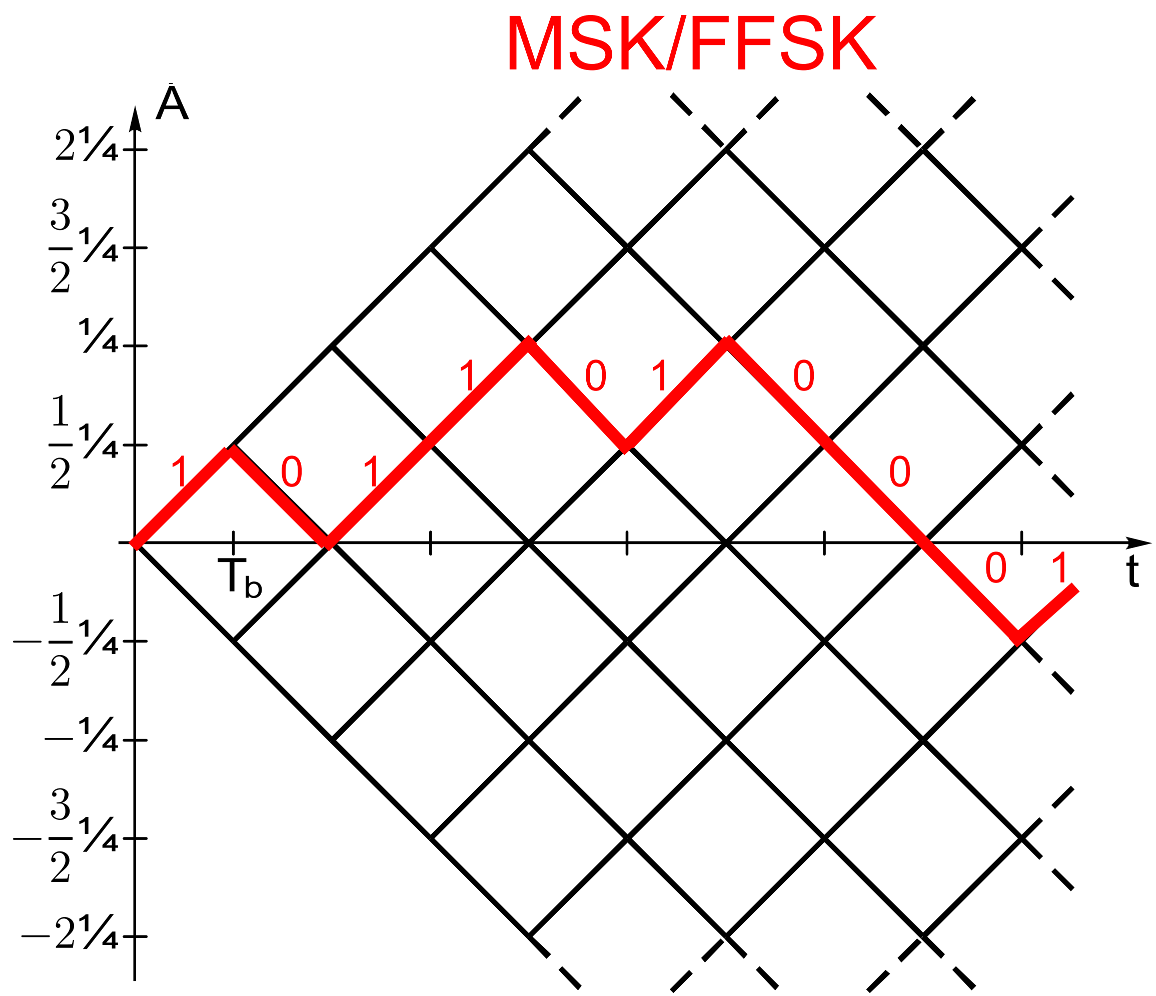
Binary frequency-shift keying (BFSK) Rectangular pulses



Contained percentage of total energy	spectral efficiency
90%	$0.59 Bit/s/Hz$
99%	$0.05 Bit/s/Hz$

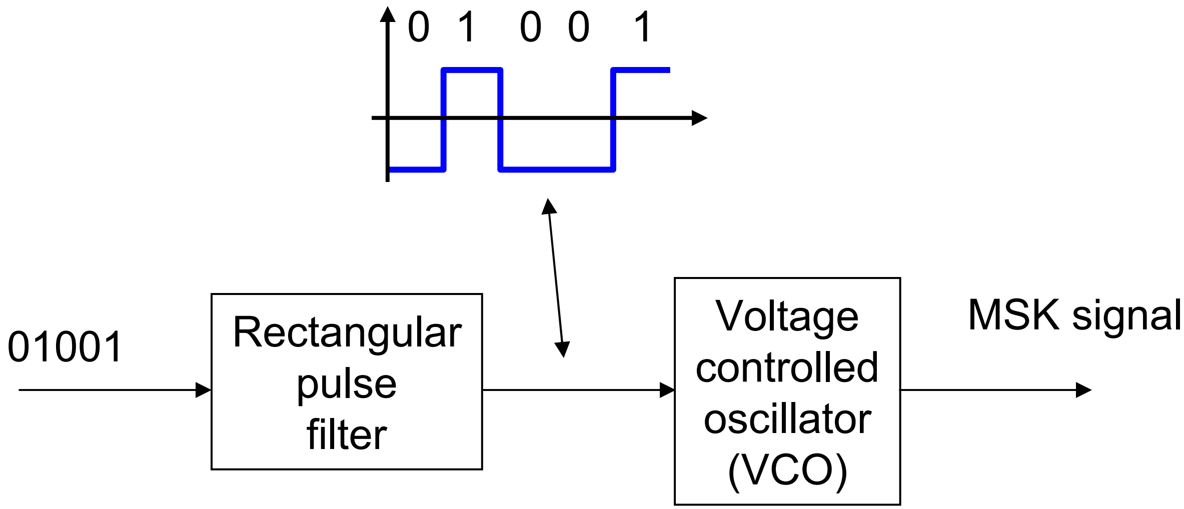
Continuous-phase modulation

Basic idea:
- Keep **amplitude constant**
- Change phase continuously



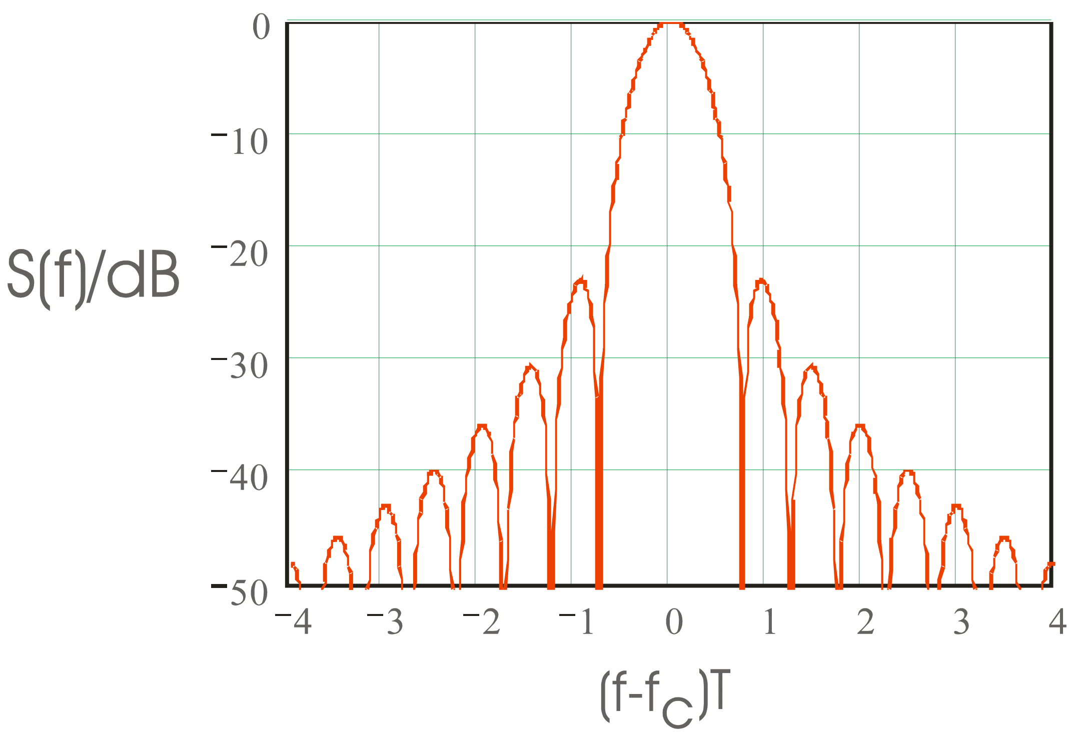
Minimum shift keying (MSK)

Simple MSK implementation



Minimum shift keying (MSK)

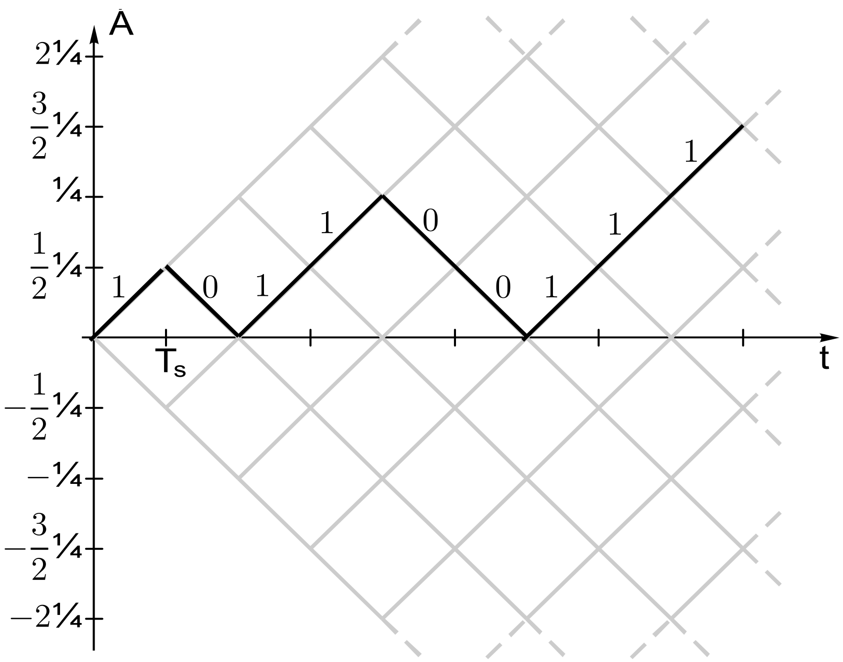
Power spectral density of MSK



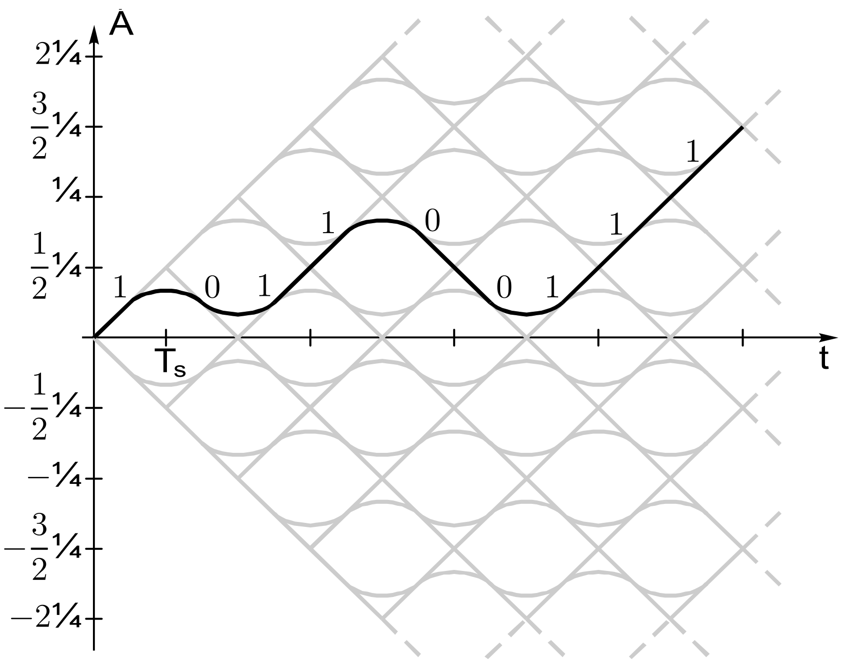
Contained percentage of total energy	spectral efficiency
90 %	1,29 Bit / s / Hz
99 %	0,85 Bit / s / Hz

Further improvement of the phase: Remove 'corners'

(Simplified figure)



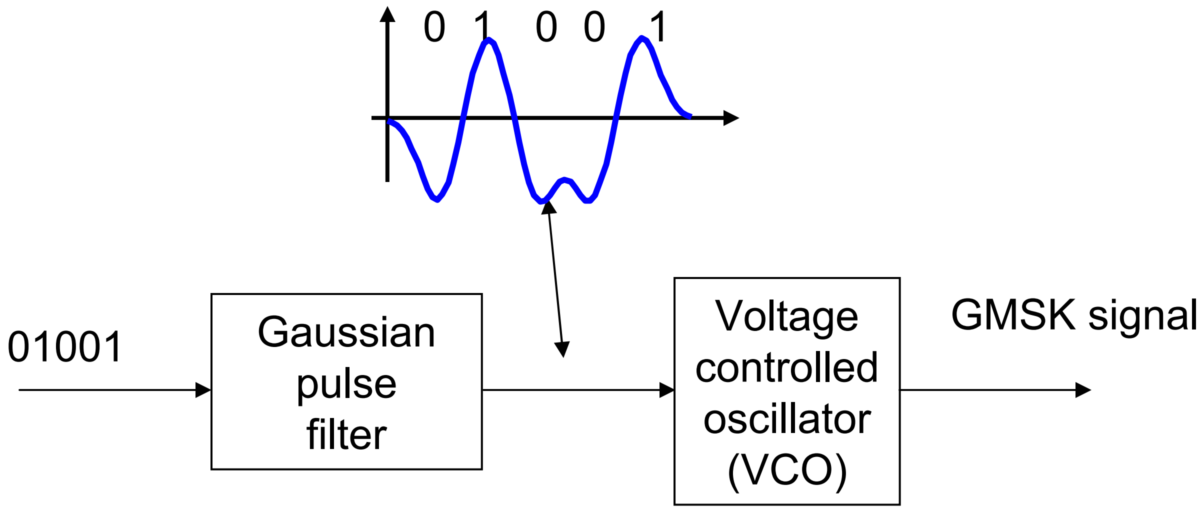
MSK
(Rectangular pulse filter)



Gaussian filtered MSK - GMSK
(Gaussian pulse filter)

Gaussian filtered MSK (GMSK)

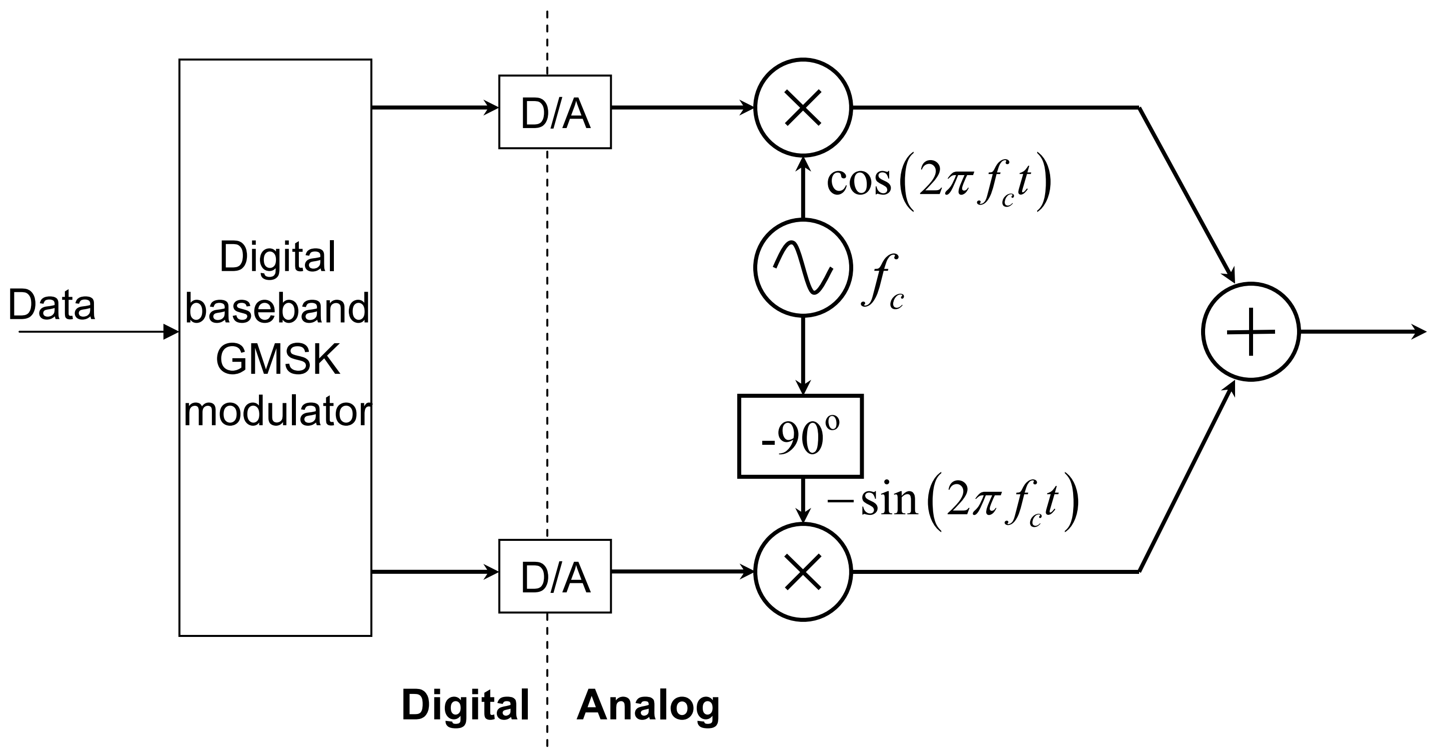
Simple GMSK implementation



GSFK is used in e.g. Bluetooth.

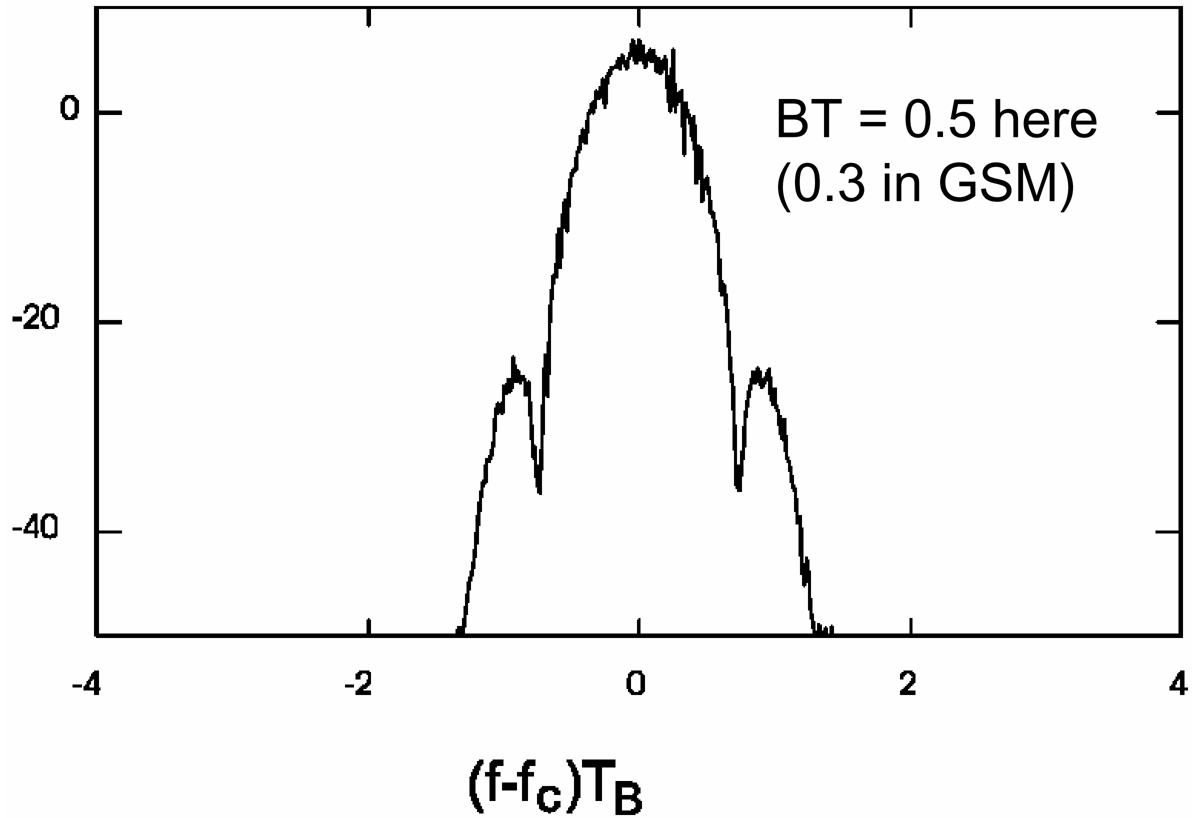
Gaussian filtered MSK (GMSK)

Digital GMSK implementation



Gaussian filtered MSK (GMSK)

Power spectral density of GMSK.



Contained percentage of total energy	spectral efficiency
90 %	1,45 Bit / s / Hz
99 %	0,97 Bit / s / Hz

How do we use all these spectral efficiencies?

Example: Assume that we want to use MSK to transmit 50 kbit/sec, and want to know the required transmission bandwidth.

Take a look at the spectral efficiency table:

Contained percentage of total energy	spectral efficiency
90 %	1,29 Bit / s / Hz
99 %	0,85 Bit / s / Hz

The 90% and 99% bandwidths become:

$$B_{90\%} = 50000/1.29 = 38.8 \text{ kHz}$$

$$B_{99\%} = 50000/0.85 = 58.8 \text{ kHz}$$

Summary

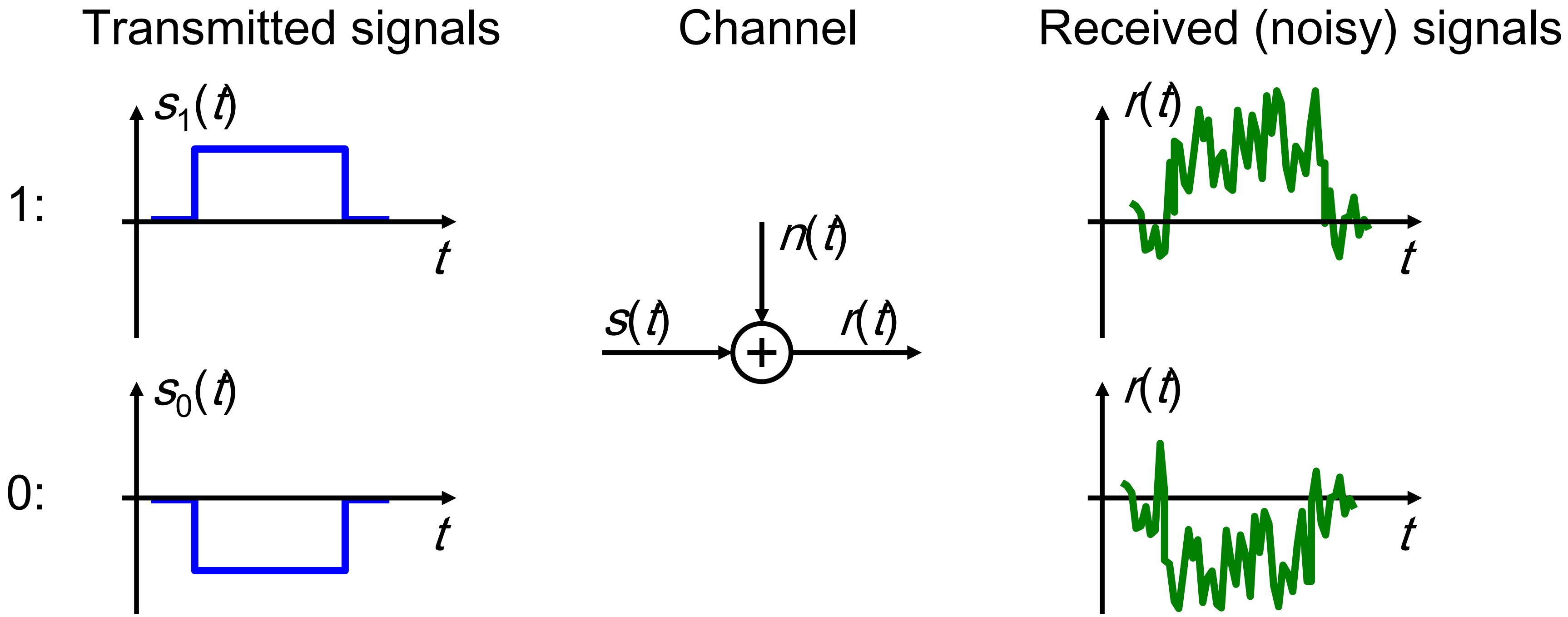
Modulation method	spectral efficiency for 90 % of total energy Bit / s / Hz	spectral efficiency for 99 % of total energy Bit / s / Hz	envelope variations w ratio of maximum and minimum amplitude
BPSK	0,59	0,05	1
BAM (α=0.5)	1,02	0,79	∞
QPSK, OQPSK, $\pi/4$-QPSK	1,18	0,10	1
MSK	1,29	0,85	1
GMSK (B_G T= 0.5)	1,45	0,97	1
QAM ($\alpha = 0.5$)	2,04	1,58	∞
OQAM ($\alpha = 0.5$)	2,04	1,58	2.6
FSK		$< 1/(2f_{\mathrm{D}}T_{\mathrm{B}})$	1

Chapter 12

Demodulation and BER computation

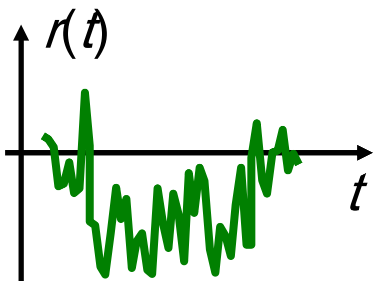
OPTIMAL RECEIVER
AND
BIT ERROR PROBABILITY
IN AWGN CHANNELS

Optimal receiver
Transmitted and received signal

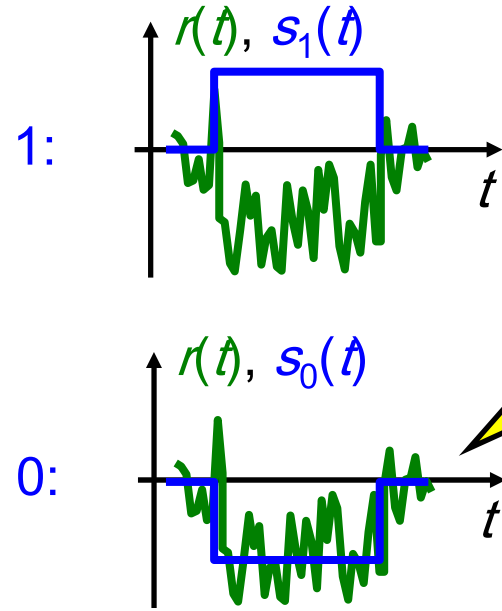


Optimal receiver
A first "intuitive" approach

Assume that the following signal is received:



Comparing it to the two possible noise free received signals:

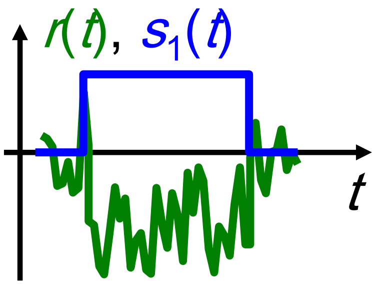


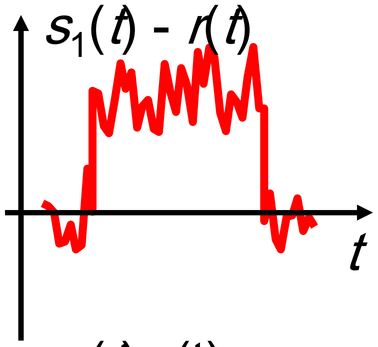
This seems to be the best "fit". We assume that "0" was the transmitted bit.

Optimal receiver
Let's make it more measurable

To be able to better measure the "fit" we look at the energy of the residual (difference) between received and the possible noise free signals:

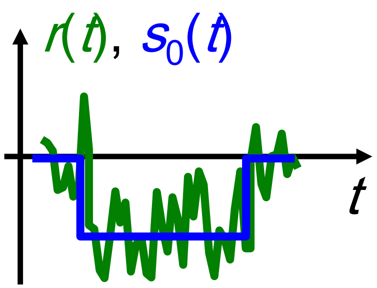
1:

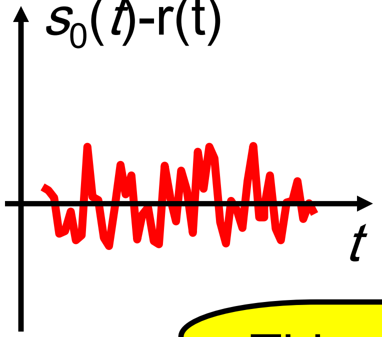




$$\int \left|s_1(t) - r(t)\right|^2 dt$$

0:



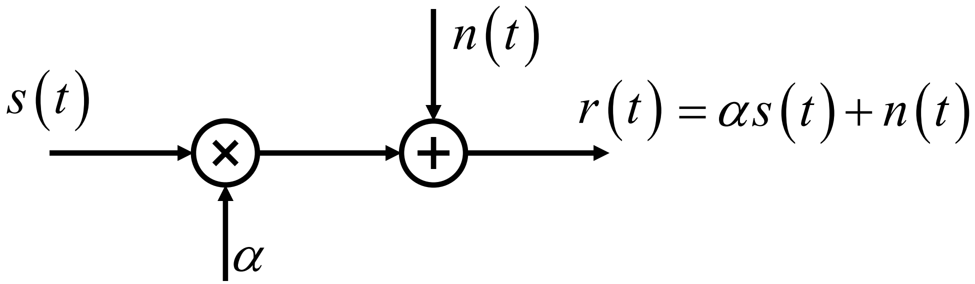


$$\int \left|s_0(t) - r(t)\right|^2 dt$$

This residual energy is much smaller. We assume that "0" was transmitted.

Optimal receiver
The AWGN channel

The additive white Gaussian noise (AWGN) channel



$s(t)$ - transmitted signal

α - channel attenuation

$n(t)$ - white Gaussian noise

$r(t)$ - received signal

In our digital transmission system, the transmitted signal $s(t)$ would be one of, let's say M, different alternatives $s_0(t)$, $s_1(t)$, ... , $s_{M-1}(t)$.

Optimal receiver
The AWGN channel, cont.

For a received $r(t)$, the residual energy e_i for each possible transmitted alternative $s_i(t)$ is calculated as

$$e_i = \int \left| r(t) - \alpha s_i(t) \right|^2 dt = \int \left(r(t) - \alpha s_i(t) \right) \left(r(t) - \alpha s_i(t) \right)^* dt$$

$$= \int \left| r(t) \right|^2 dt - \underbrace{2\,\mathrm{Re}\left\{ \alpha^* \int r(t) s_i^*(t)\, dt \right\}} + \left| \alpha \right|^2 \int \left| s_i(t) \right|^2 dt$$

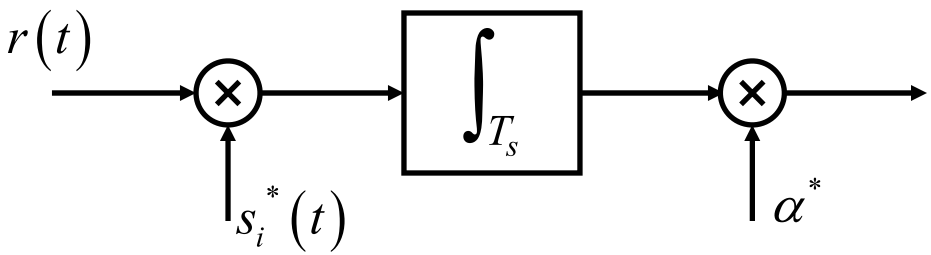
Same for all *i*

The residual energy is minimized by maximizing this part of the expression.

Same for all *i*, if the transmitted signals are of equal energy.

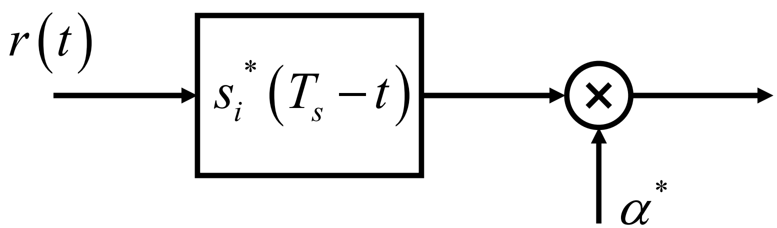
Optimal receiver
The AWGN channel, cont.

The central part of the comparison of different signal alternatives is a correlation, that can be implemented as a correlator:



The real part of the output from either of these is sampled at $t = T_s$

or a matched filter



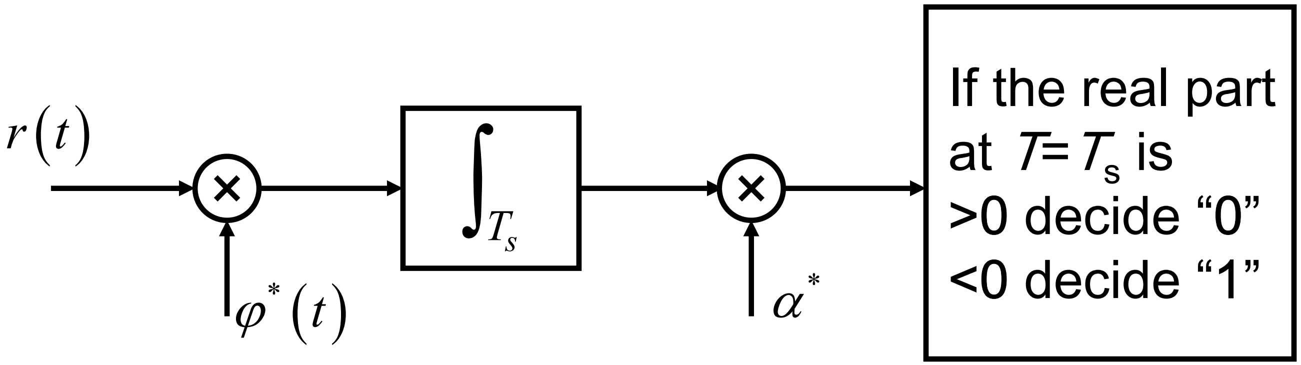
where T_s is the symbol time (duration).

Optimal receiver
Antipodal signals

In antipodal signaling, the alternatives (for “0” and “1”) are

$$s_0(t) = \varphi(t)$$
$$s_1(t) = -\varphi(t)$$

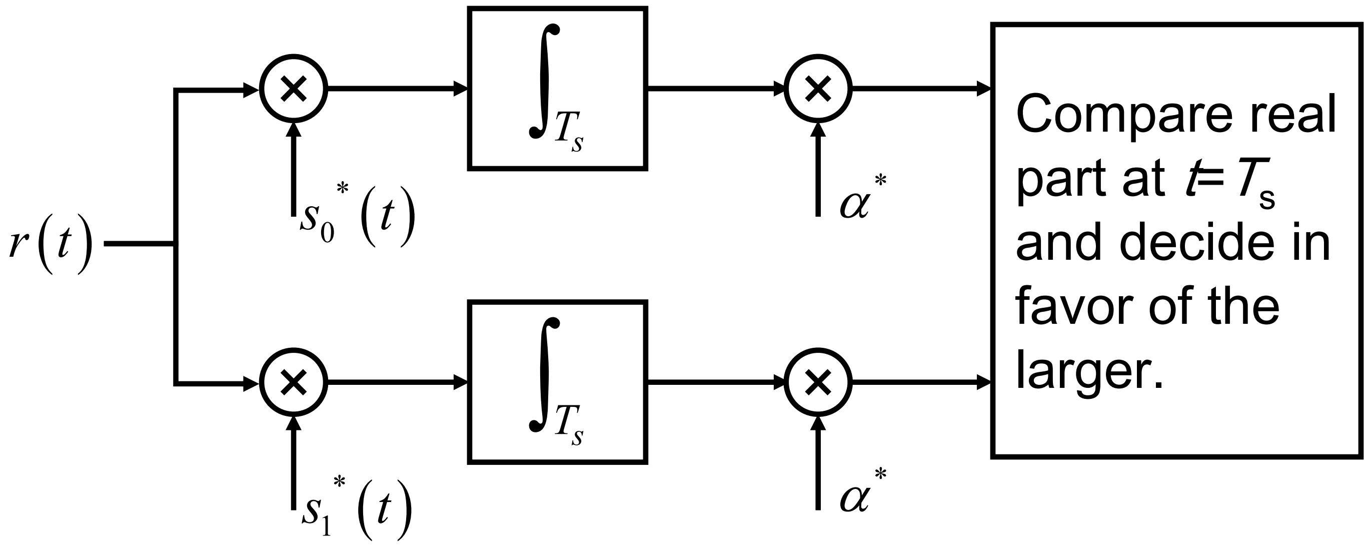
This means that we only need ONE correlation in the receiver for simplicity:



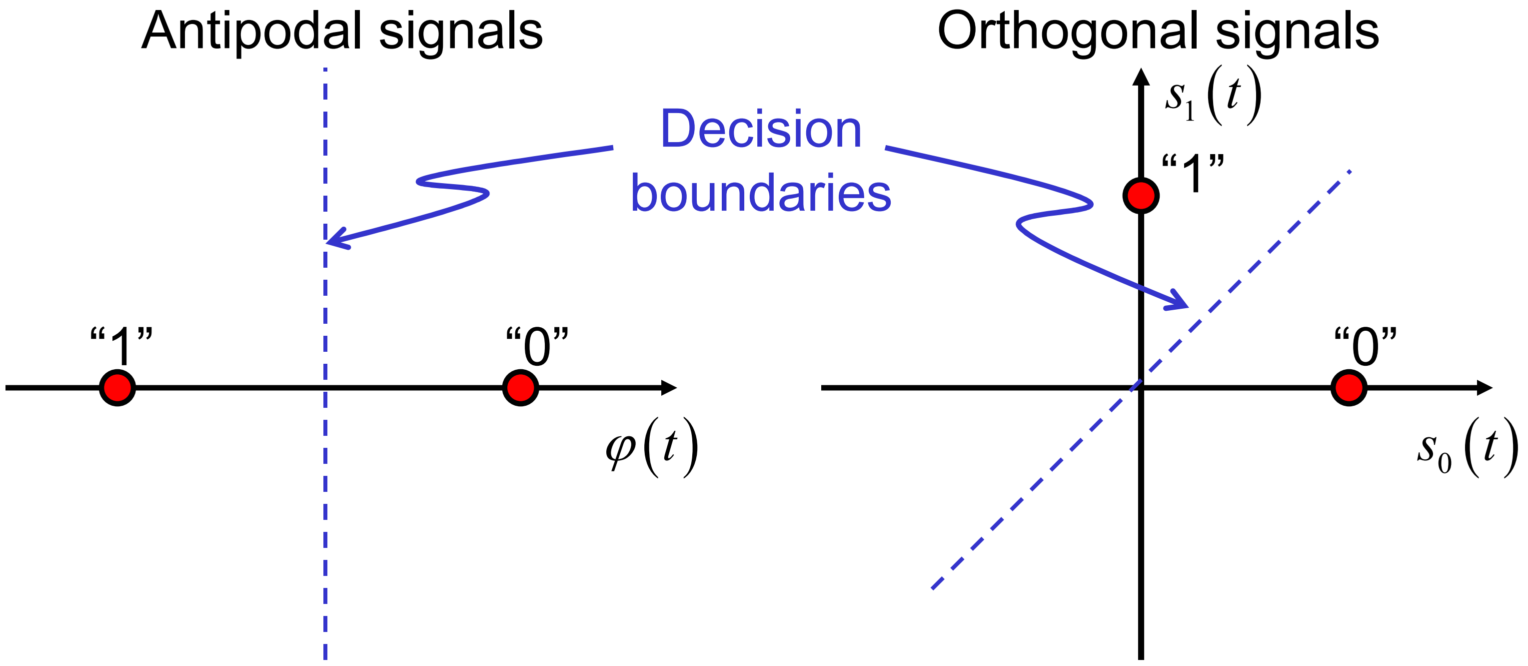
Optimal receiver
Orthogonal signals

In binary orthogonal signaling, with equal energy alternatives $s_0(t)$ and $s_1(t)$ (for "0" and "1") we require the property:

$$\langle s_0(t), s_1(t) \rangle = \int s_0(t) s_1^*(t)\, dt = 0$$

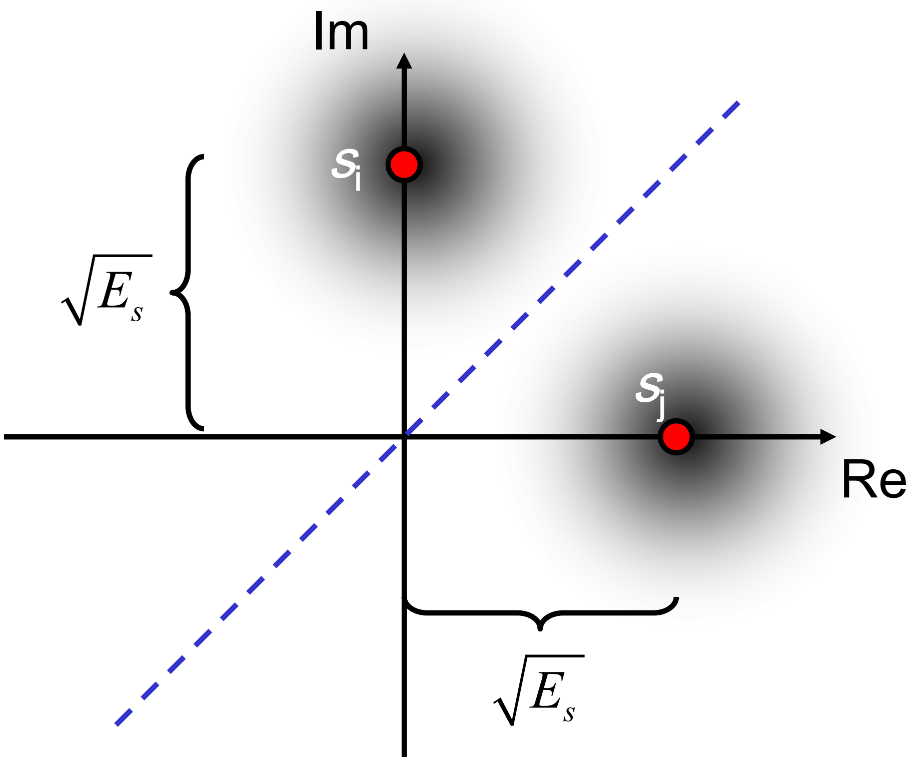


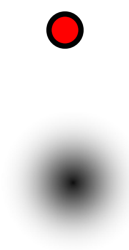
Optimal receiver
Interpretation in signal space



Optimal receiver
The noise contribution

Assume a 2-dimensional signal space, here viewed as the complex plane





This normalization of axes implies that the noise centered around each alternative is complex Gaussian

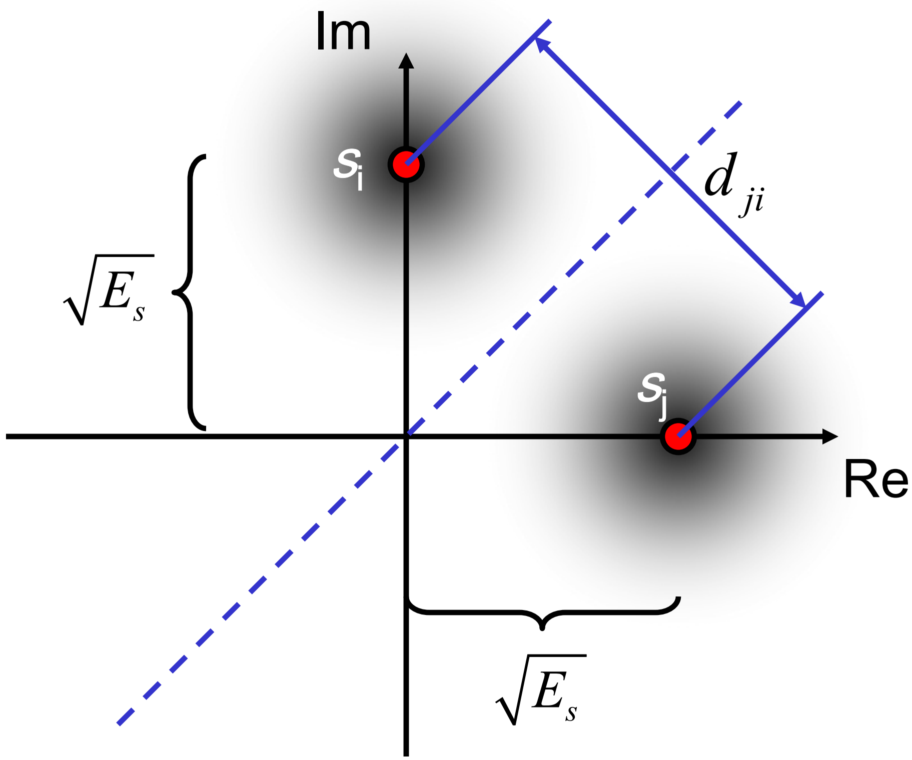
$$N\left(0,\sigma^2\right)+jN\left(0,\sigma^2\right)$$

with variance $\sigma^2 = N_0/2$ in each direction.

Fundamental question: What is the probability that we end up on the wrong side of the decision boundary?

Optimal receiver
Pair-wise symbol error probability

What is the probability of deciding s_i if s_j was transmitted?



We need the distance between the two symbols. In this orthogonal case:

$$d_{ji} = \sqrt{\sqrt{E_s}^2 + \sqrt{E_s}^2} = \sqrt{2E_s}$$

The probability of the noise pushing us across the boundary at distance $d_{ji}/2$ is

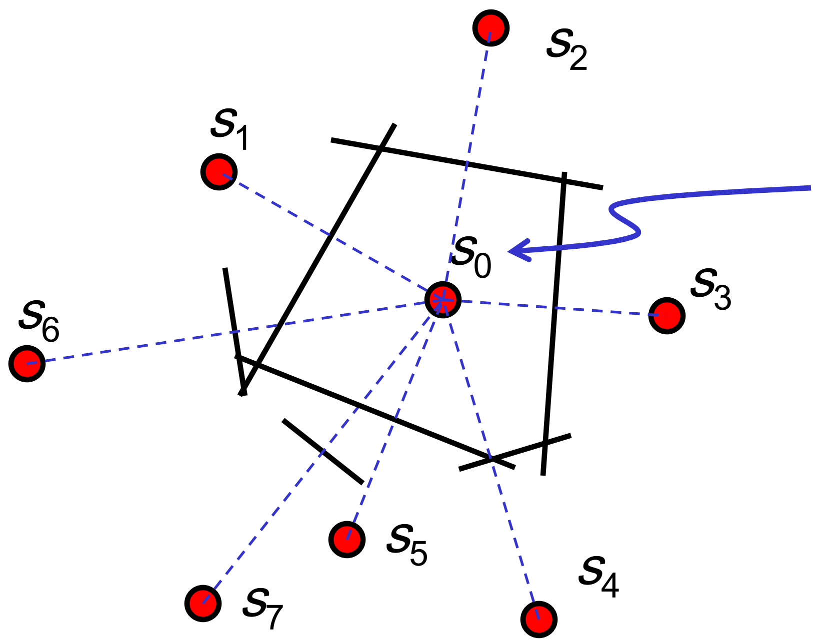
$$P\left(s_j \rightarrow s_i\right) = Q\left(\frac{d_{ji}/2}{\sqrt{N_0/2}}\right) = Q\left(\sqrt{\frac{E_s}{N_0}}\right)$$

$$= \frac{1}{2}\operatorname{erfc}\left(\sqrt{\frac{E_s}{2N_0}}\right)$$

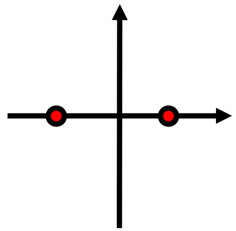
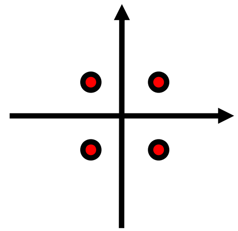
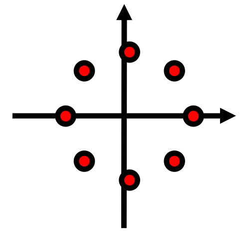
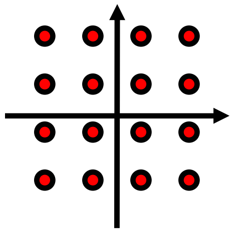
Optimal receiver
The union bound

Calculation of symbol error probability is simple for two signals!

When we have many signal alternatives, it may be impossible to calculate an exact symbol error rate.

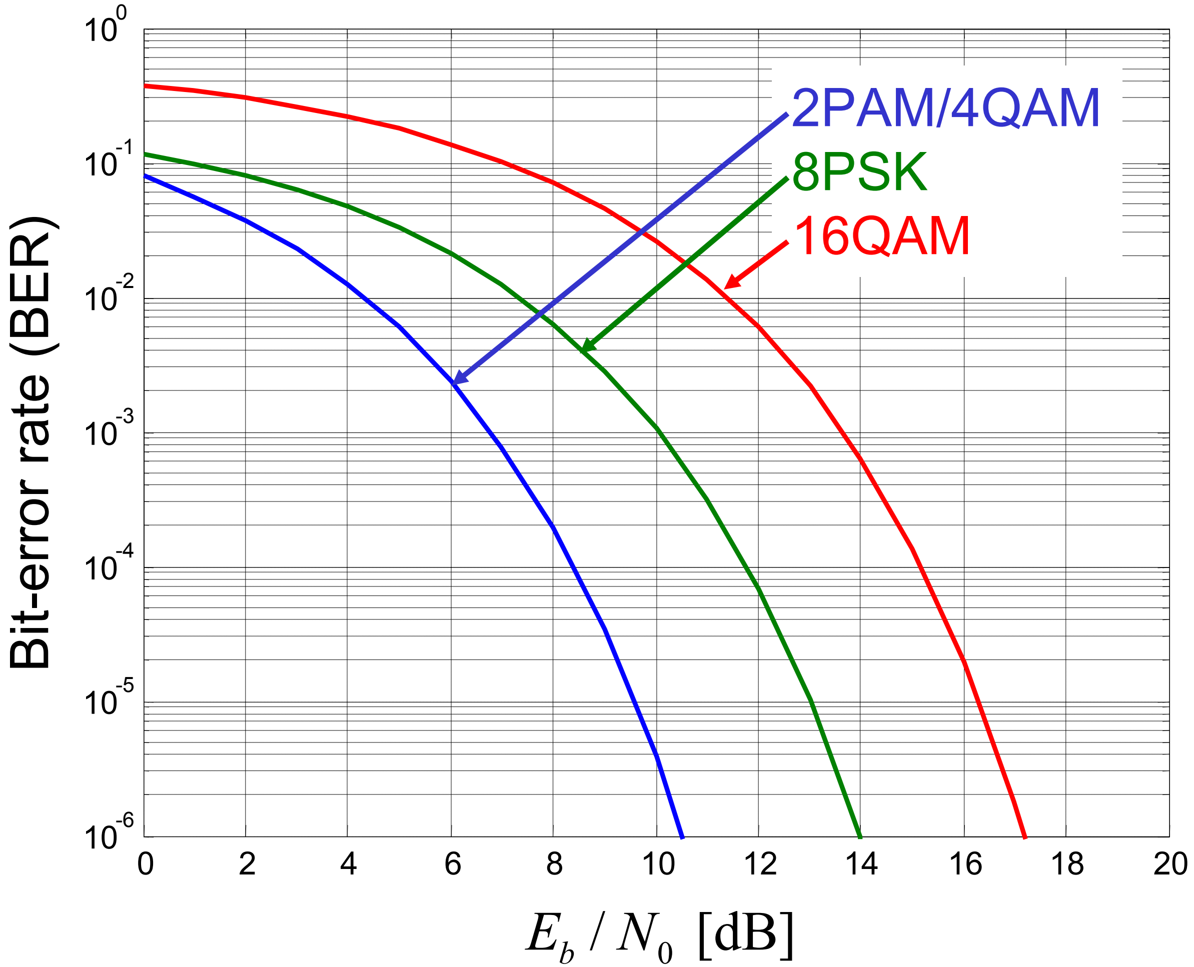


When s_0 is the transmitted signal, an error occurs when the received signal is outside this polygon.

EXAMPLES:	2PAM	4QAM	8PSK	16QAM
				
Bits/symbol	1	2	3	4
Symbol energy	E_b	$2E_b$	$3E_b$	$4E_b$
BER	$Q\left(\sqrt{\frac{2E_b}{N_0}}\right)$	$Q\left(\sqrt{\frac{2E_b}{N_0}}\right)$	$\approx \frac{2}{3} Q\left(\sqrt{0.87 \frac{E_b}{N_0}}\right)$	$\approx \frac{3}{2} Q\left(\sqrt{\frac{E_{b,\max}}{2.25 N_0}}\right)$

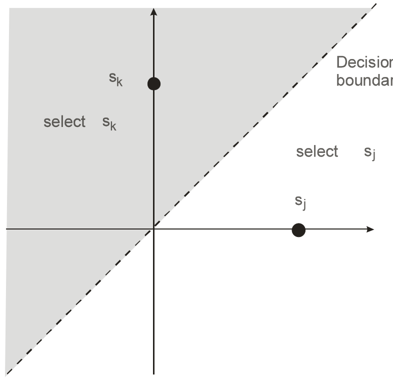
Gray coding is used when calculating these BER.

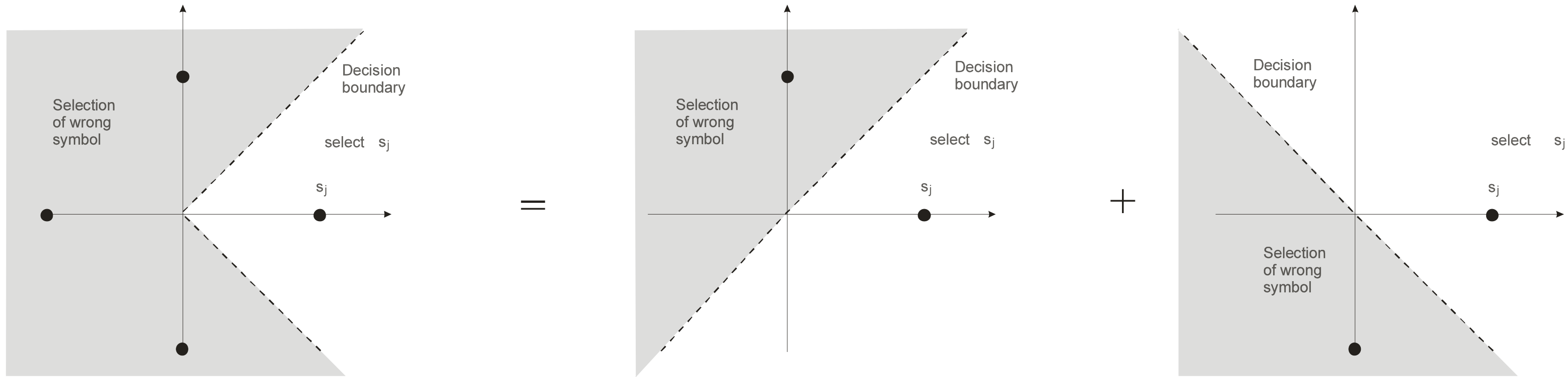
Optimal receiver
Bit-error rates (BER), cont.



Optimal receiver – BER of QPSK

- Compute via union bound
- Pairwise error probability $Q\left(\sqrt{2\gamma_B}\right)$
- Symbol error probability $SER \approx 2Q(\sqrt{2\gamma_B})$
- Bit error probability $BER = Q(\sqrt{2\gamma_B})$





Optimal receiver
Where do we get E_b and N_0?

Where do those magic numbers E_b and N_0 come from?

The noise power spectral density N_0 is calculated according to

$$N_0 = kT_0F_0 \Leftrightarrow N_{0|dB} = -204 + F_{0|dB}$$

where F_0 is the noise factor of the "equivalent" receiver noise source.

The bit energy E_b can be calculated from the received power C (at the same reference point as N_0). Given a certain data-rate d_b [bits per second], we have the relation

$$E_b = C / d_b \Leftrightarrow E_{b|dB} = C_{|dB} - d_{b|dB}$$

THESE ARE THE EQUATIONS THAT RELATE DETECTOR PERFORMANCE ANALYSIS TO LINK BUDGET CALCULATIONS!

BER for differential receiver

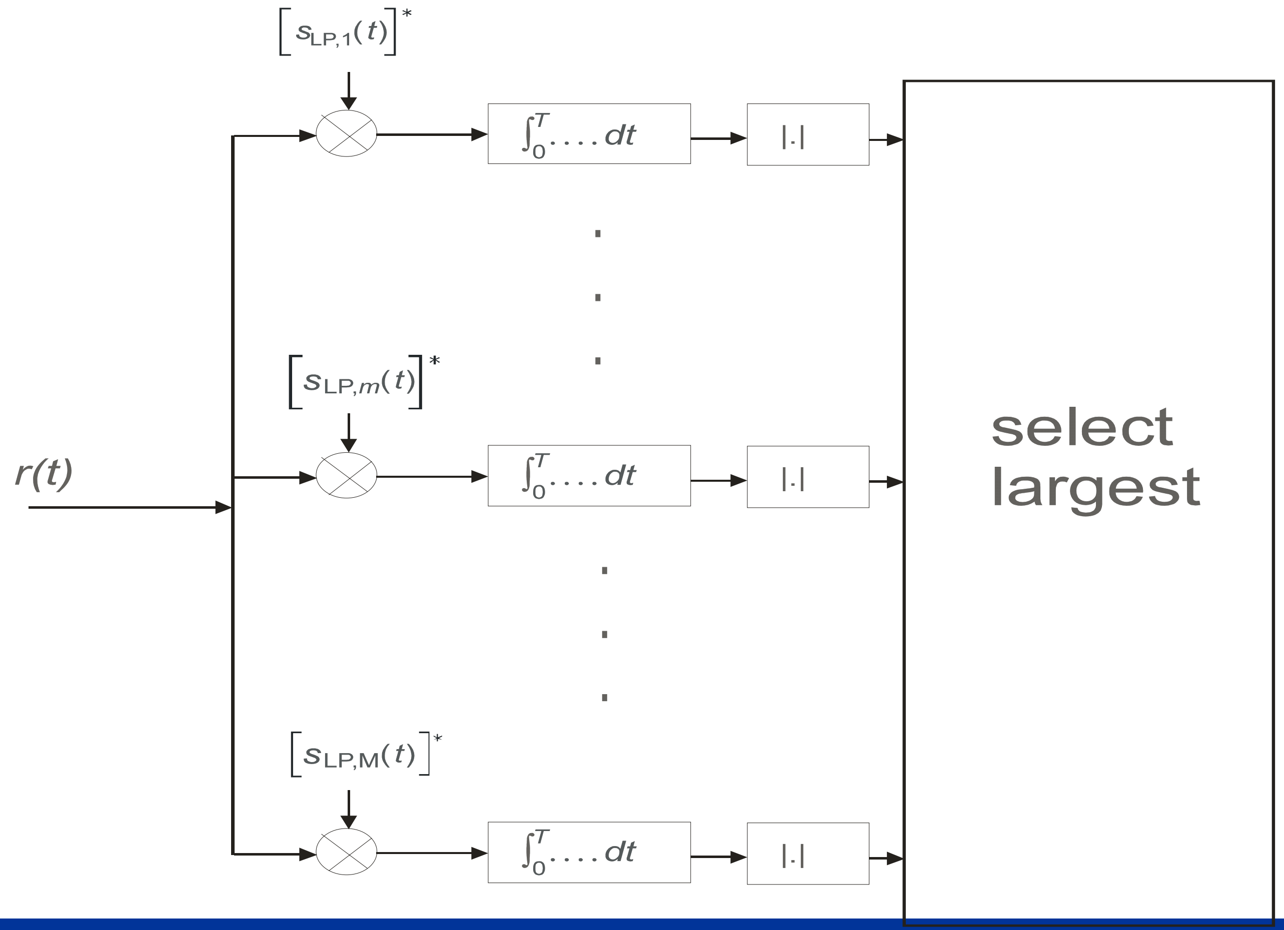
- Differential BPSK

$$\Phi_i = \Phi_{i-1} + \begin{cases} +\frac{\pi}{2} & b_i = +1 \\ -\frac{\pi}{2} & b_i = -1 \end{cases}$$

- BER for differentially detected BPSK:

$$BER = \frac{1}{2}\exp(-\gamma_b) \; .$$

Noncoherent detection (1)



Noncoherent detection (2)

- Error probability for noncoherent detection

$$BER = Q_{\mathrm{M}}(a,b) - \frac{1}{2} I_0(ab) \exp\left(-\frac{1}{2}(a^2 + b^2)\right)$$

$$a = \sqrt{\frac{\gamma_{\mathrm{B}}}{2}\left(1 - \sqrt{1 - |\rho|^2}\right)} \qquad b = \sqrt{\frac{\gamma_{\mathrm{B}}}{2}\left(1 + \sqrt{1 - |\rho|^2}\right)} \; .$$

- For phase modulation, |ρ|=1, therefore SNR=0

BER IN FADING CHANNELS AND DISPERSION-INDUCED ERRORS

BER in fading channels (1)

We have (or can calculate) BER expressions for non-fading AWGN channels.

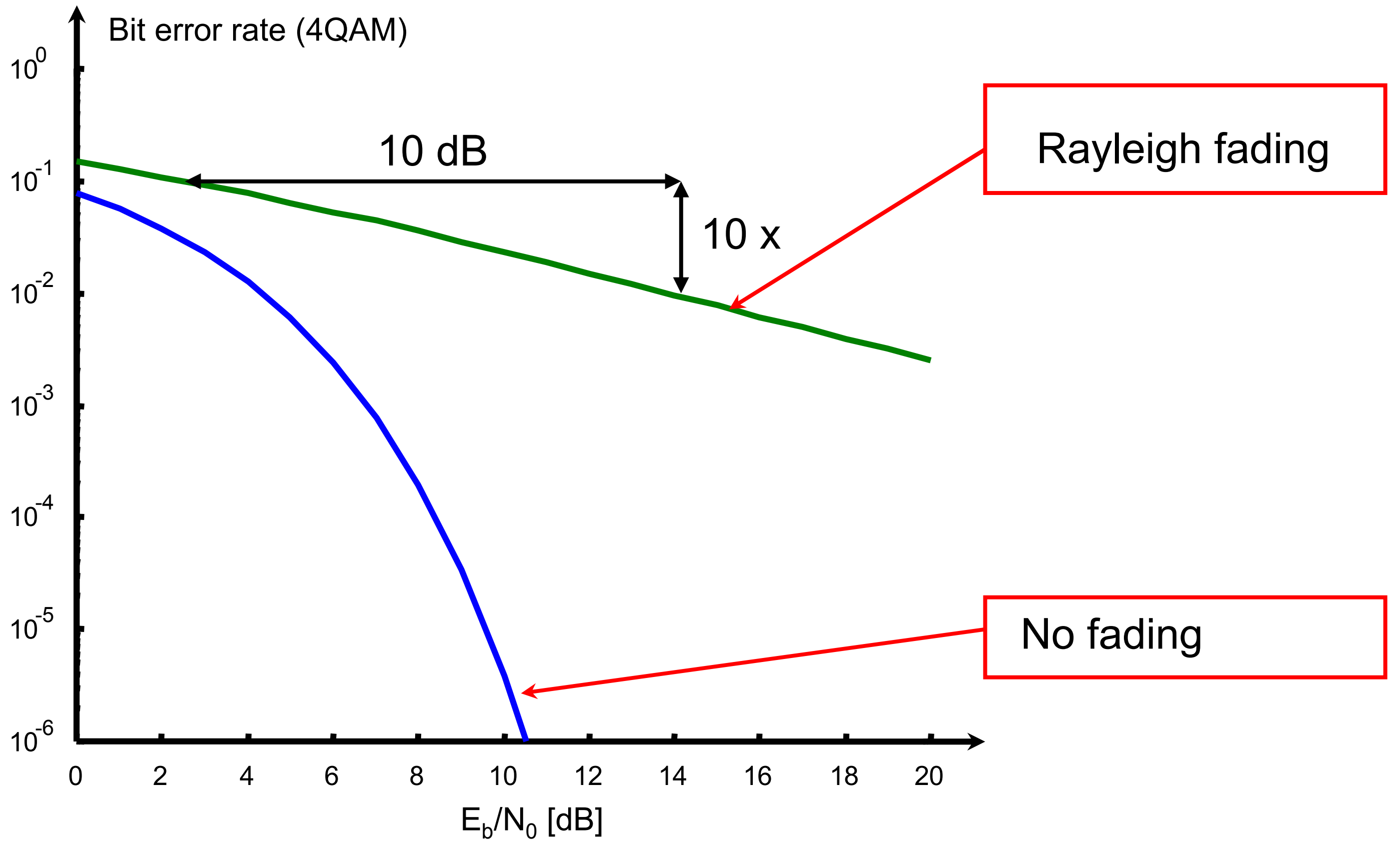
$$pdf\left(\gamma_b\right) = \frac{1}{\overline{\gamma_b}} e^{-\gamma_b/\overline{\gamma_b}}$$

γ_b	-- E_b/N_0
$\overline{\gamma_b}$	-- average E_b/N_0

$$BER_{Rayleigh}\left(\overline{\gamma_b}\right) = \int_0^{\infty} BER_{AWGN}\left(\gamma_b\right) \times pdf\left(\gamma_b\right) d\gamma_b$$

BER in fading channels (2)

THIS IS A SERIOUS PROBLEM!



BER in fading channels (3)

- Coherent detection of antipodal signals

$$\overline{BER} = \frac{1}{2}\left[1 - \sqrt{\frac{\overline{\gamma}_B}{1+\overline{\gamma}_B}}\right] \approx \frac{1}{4\overline{\gamma}_B}$$

- Coherent detection of orthogonal signals

$$\overline{BER} = \frac{1}{2}\left[1 - \sqrt{\frac{\overline{\gamma}_B}{2+\overline{\gamma}_B}}\right] \approx \frac{1}{2\overline{\gamma}_B}$$

- Differential detection of antipodal signals

$$\overline{BER} = \frac{1}{2+\overline{\gamma}_B} \approx \frac{1}{\overline{\gamma}_B}$$

- Differential detection of orthogonal signals

$$\overline{BER} = \frac{1}{2(1+\overline{\gamma}_B)} \approx \frac{1}{2\overline{\gamma}_B}$$

Alternative computation of BER

- Alternative representation of Q-function

$$Q(x) = \frac{1}{\pi}\int_0^{\pi/2} \exp\left(-\frac{x^2}{2\sin^2\theta}\right) d\theta$$

- Example: SER of M-ary PSK in AWGN channel:

$$SER = \frac{1}{\pi}\int_0^{(M-1)\pi/M} \exp\left(-\frac{\gamma_S}{\sin^2\theta}\sin^2(\pi/M)\right) d\theta$$

- Averaged SER:

$$\overline{SER} = \frac{1}{\pi}\int_0^{(M-1)\pi/M}\int_0^{\infty} pdf_\gamma(\gamma)\exp\left(-\frac{\gamma_S}{\sin^2\theta}\sin^2(\pi/M)\right) d\theta$$

- This can be expressed in terms of the characteristic function of the fading distribution $M_\gamma(s)$

$$\overline{SER} = \frac{1}{\pi}\int_0^{(M-1)\pi/M} .M_\gamma\left(\frac{\gamma_S}{\sin^2\theta}\sin^2(\pi/M))\right) d\theta$$

Doppler-induced errors

- Distortion on the channel causes irreducible errors (cannot be eliminated by increasing transmit power)
- Frequency dispersion:
  - Due to Doppler effect
  - Instantaneous frequency can be computed as

$$f_{\mathrm{inst}}(t) = \frac{\mathrm{Im}\left(r^*(t)\frac{dr(t)}{dt}\right)}{|r(t)|^2}$$

  - Large frequency shift in fading dips
  - Resulting BER (for MSK)

$$\overline{BER}_{\mathrm{Doppler}} = \frac{1}{2}\pi^2(\nu_{\max}T_{\mathrm{B}})^2 \quad .$$

  - Mostly relevant for low datarates

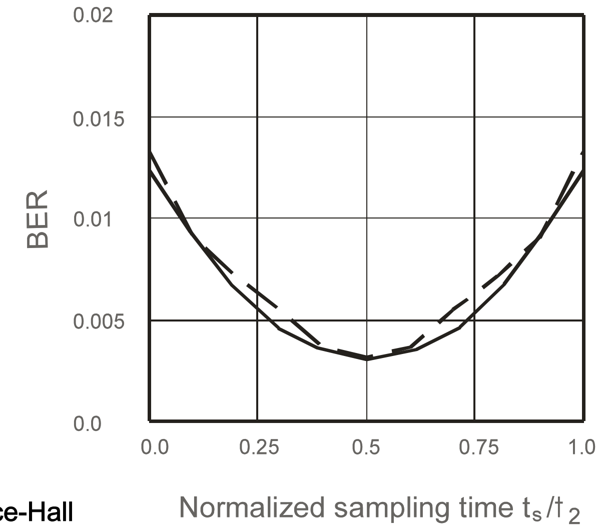
Errors induced by delay dispersion

- Delay dispersion causes intersymbol interference

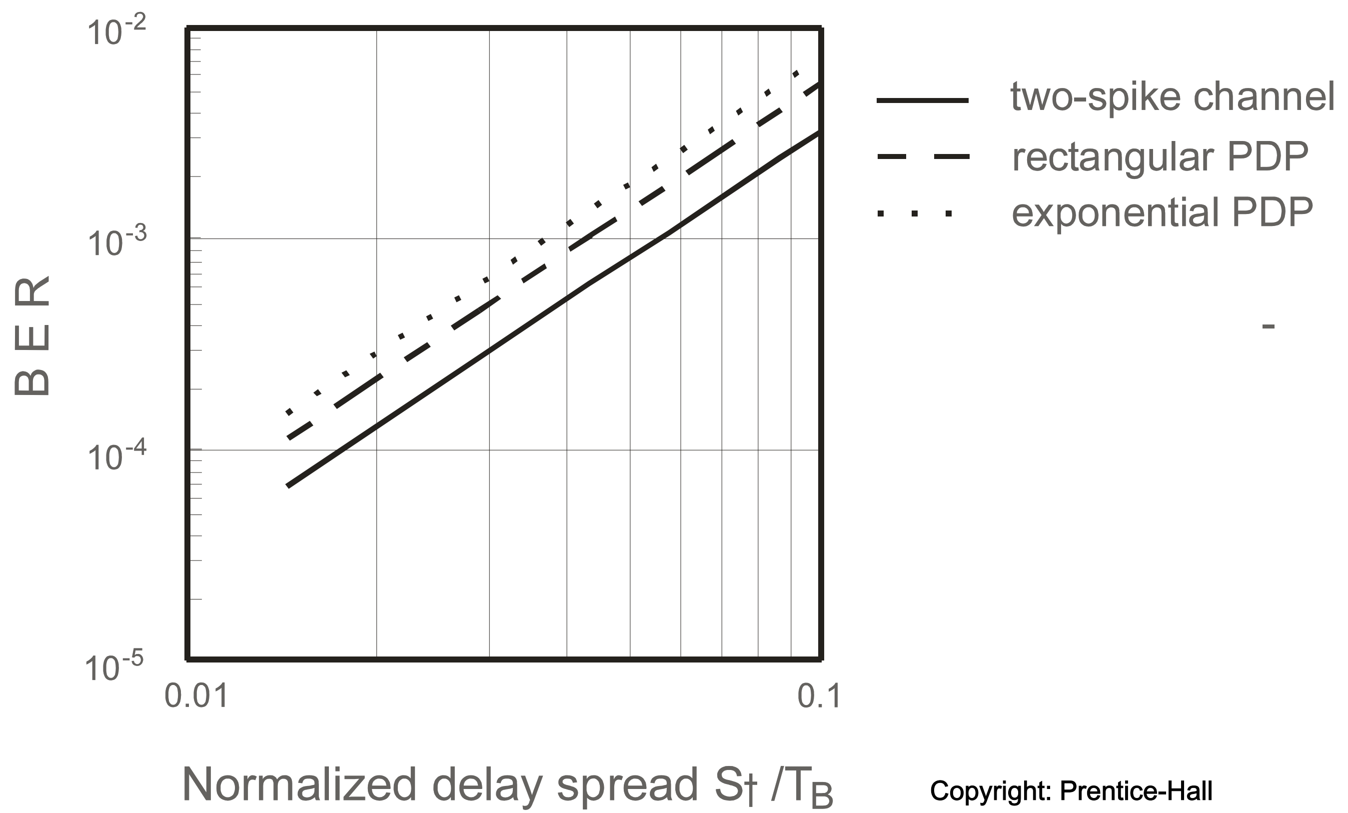
- Average BER

$$\overline{BER} = K\left(\frac{S_\tau}{T_\mathrm{B}}\right)^2$$

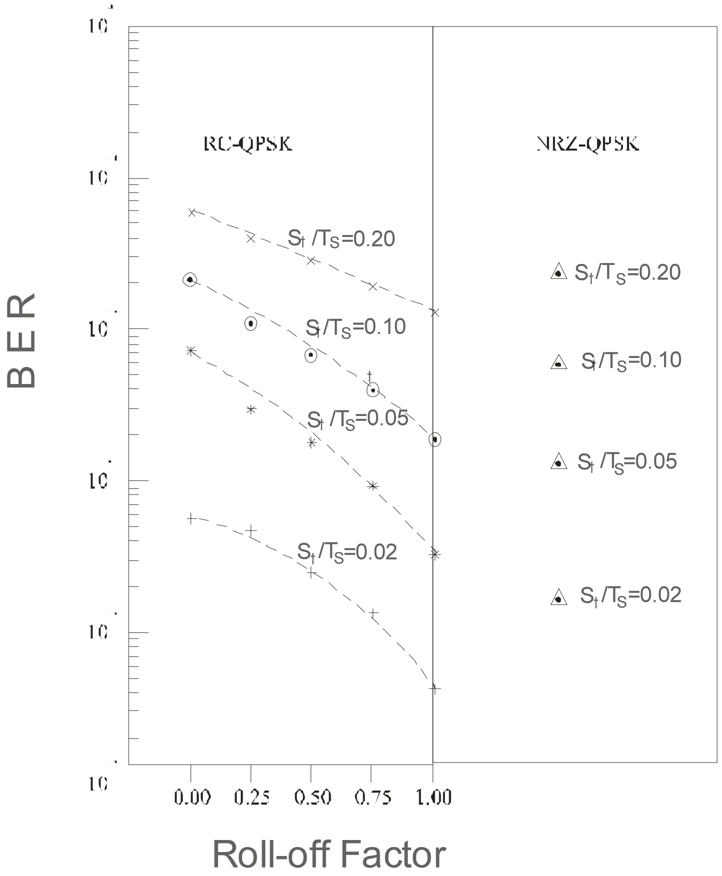
- Influenced by sampling instant



Errors induced by delay dispersion (2)



Impact of filtering



Computation methods (1)

- Group delay method: distortion of signal phase is related to group delay

$$\Phi_c(\omega) = \Phi_c(0) + \omega \frac{\partial \Phi_c}{\partial \omega}|_{\omega=0} + \frac{1}{2}\omega^2 \frac{\partial^2 \Phi_c}{\partial \omega^2}|_{\omega=0} + \ldots..$$

$$\approx \Phi_c(0) - \omega T_g$$

- Statistics of group delay

$$pdf_{T_g}(T_g) = \frac{1}{2S_\tau} \frac{1}{\left[1+\left(T_g/S_\tau\right)^2\right]^{3/2}}$$

- ->

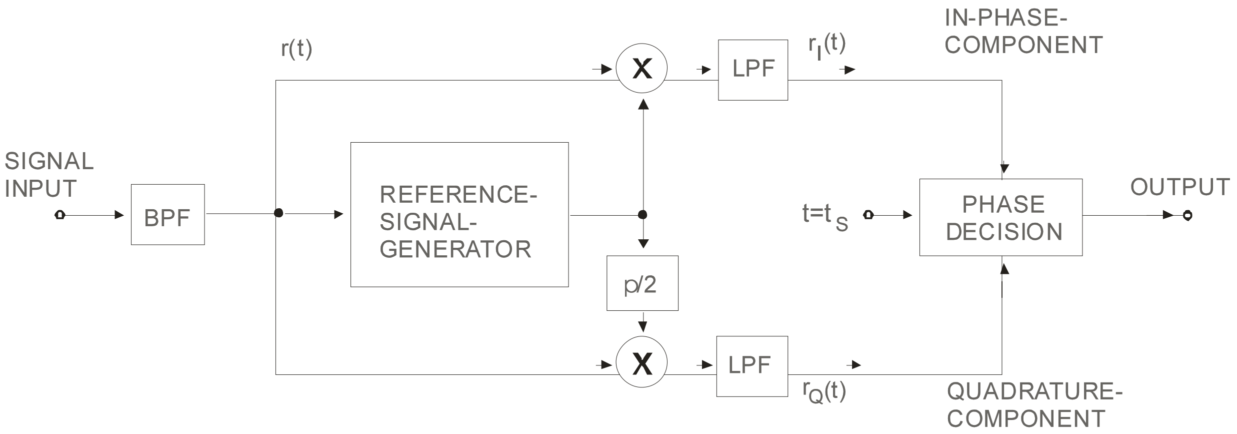
$$BER = \frac{4}{9}\left(\frac{S_\tau}{T_B}\right)^2 \approx \frac{1}{2}\left(\frac{S_\tau}{T_B}\right)^2.$$

Computation method (2)

- Quadratic form of Gaussian variables
- Formulate error event as

$$D = A|X|^2 + B|Y|^2 + CXY^* + C^*X^*Y < 0$$

- Canonical receiver



Computation method (3)

- Differentially-detected MSK

$$X = r(t_s) \; Y = r(t_s - T)$$

- Error condition is

$$\mathrm{Re}\{b_0 X Y^* \exp(-j\pi/2)\} \;\; < \; 0$$

- BER can be computed as

$$\overline{BER} = \frac{1}{2} - \frac{1}{2} \frac{b_0 \, \mathrm{Im}\{\rho_{XY}\}}{\sqrt{\mathrm{Im}\{\rho_{XY}\}^2 + (1 - |\rho_{XY}|^2)}}$$

Chapter 13

Diversity

Diversity arrangements
Let's have a look at fading again

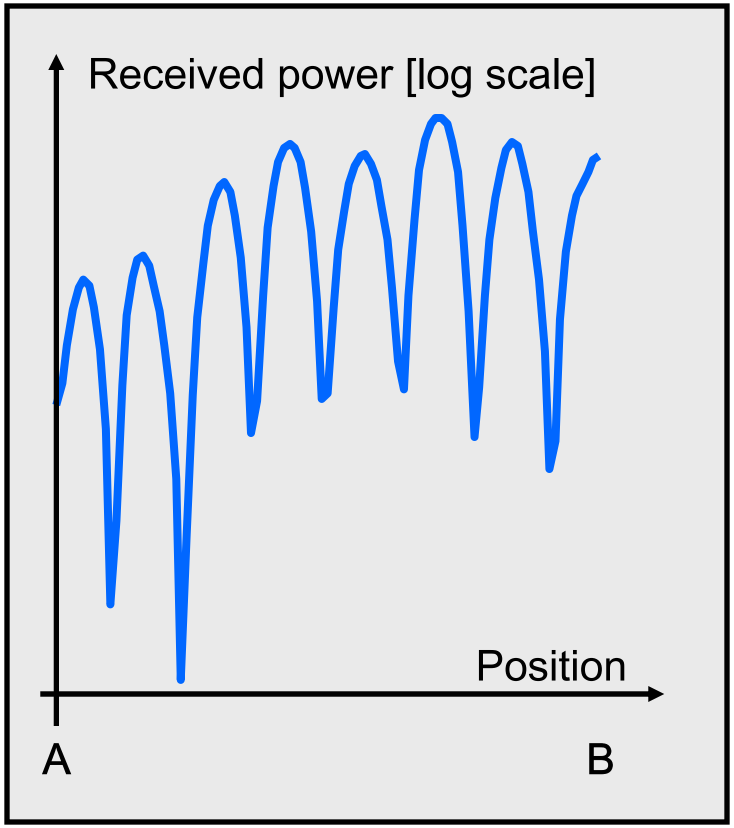
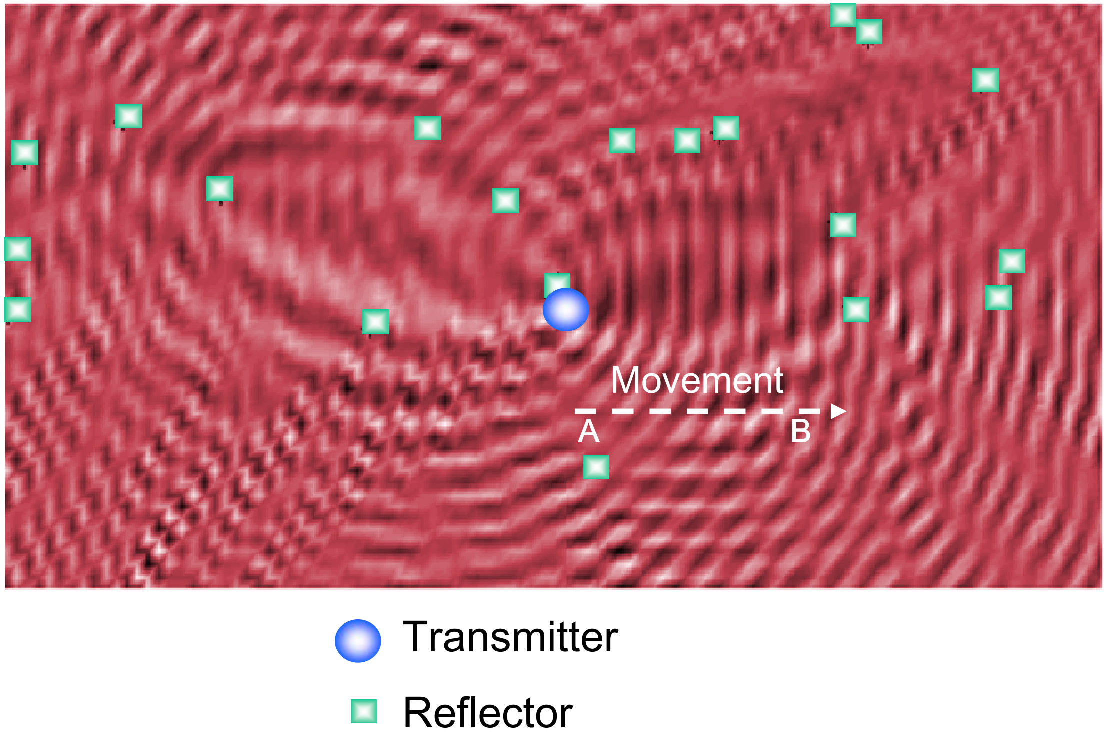


Illustration of interference pattern from above



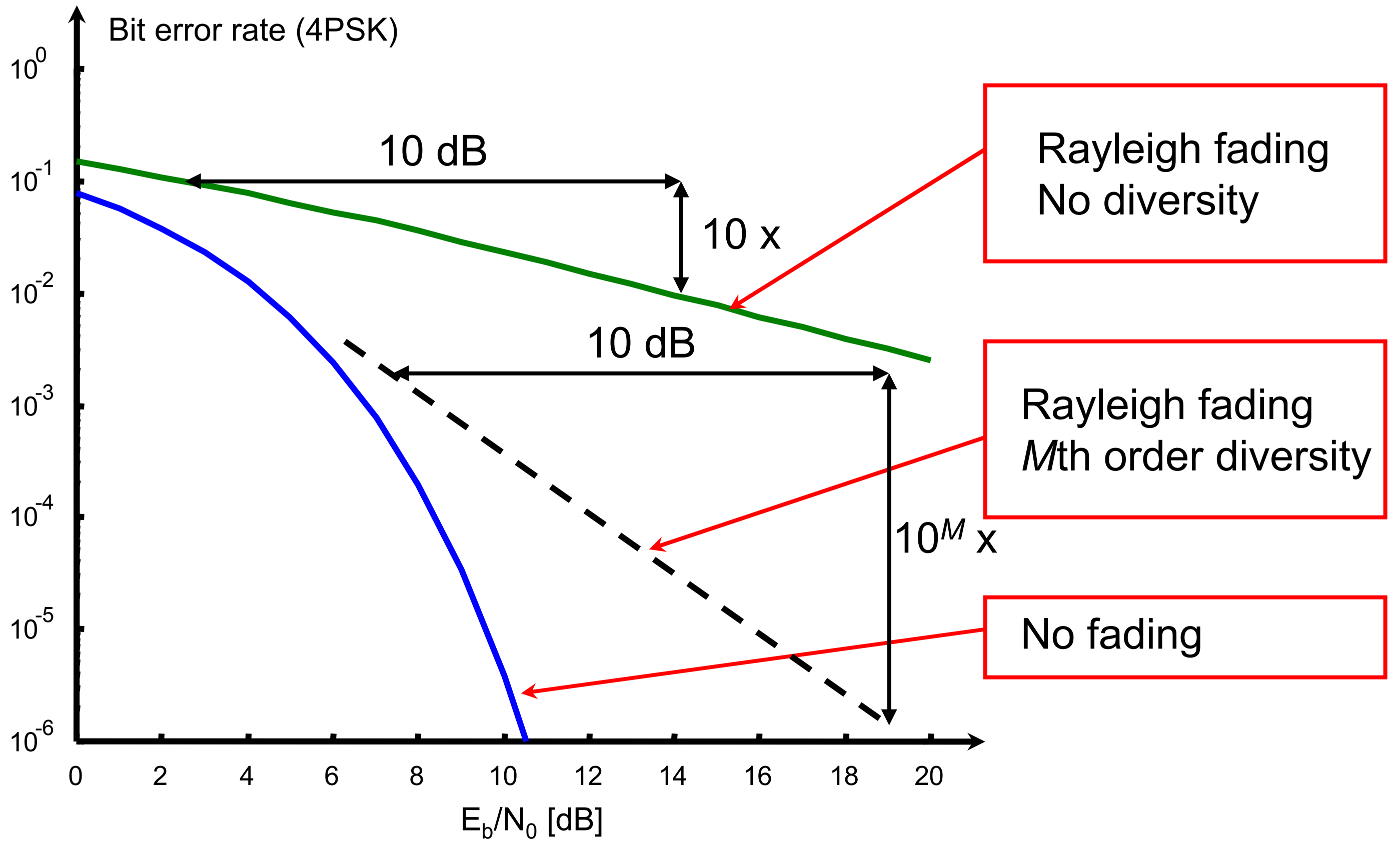
Diversity arrangements
The diversity principle

The principle of diversity is to transmit the same information on M statistically independent channels.

By doing this, we increase the chance that the information will be received properly.

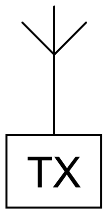
The example given on the previous slide is one such arrangement: antenna diversity.

Diversity arrangements
General improvement trend

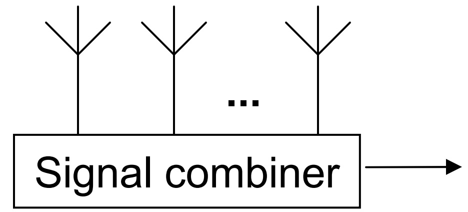


Diversity arrangements
Some techniques

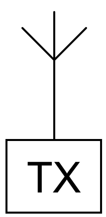
Spatial (antenna) diversity

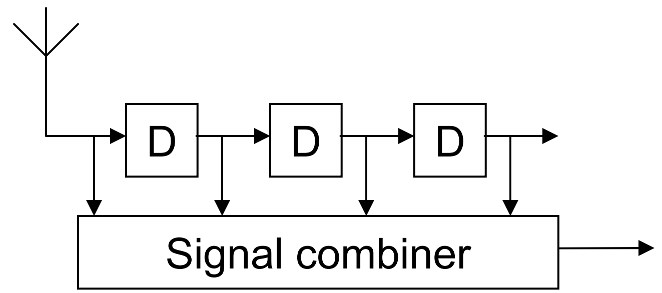


We will focus on this one today!

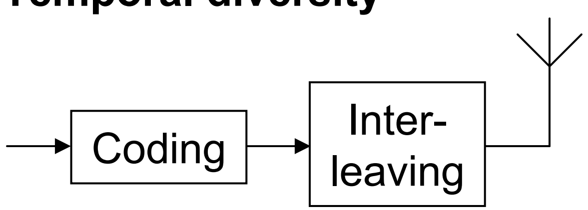


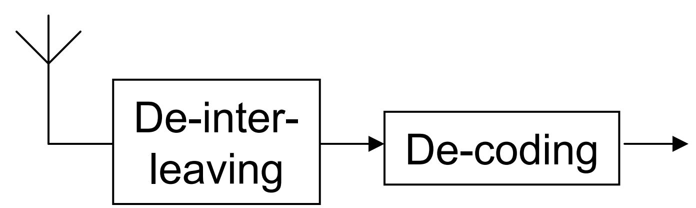
Frequency diversity





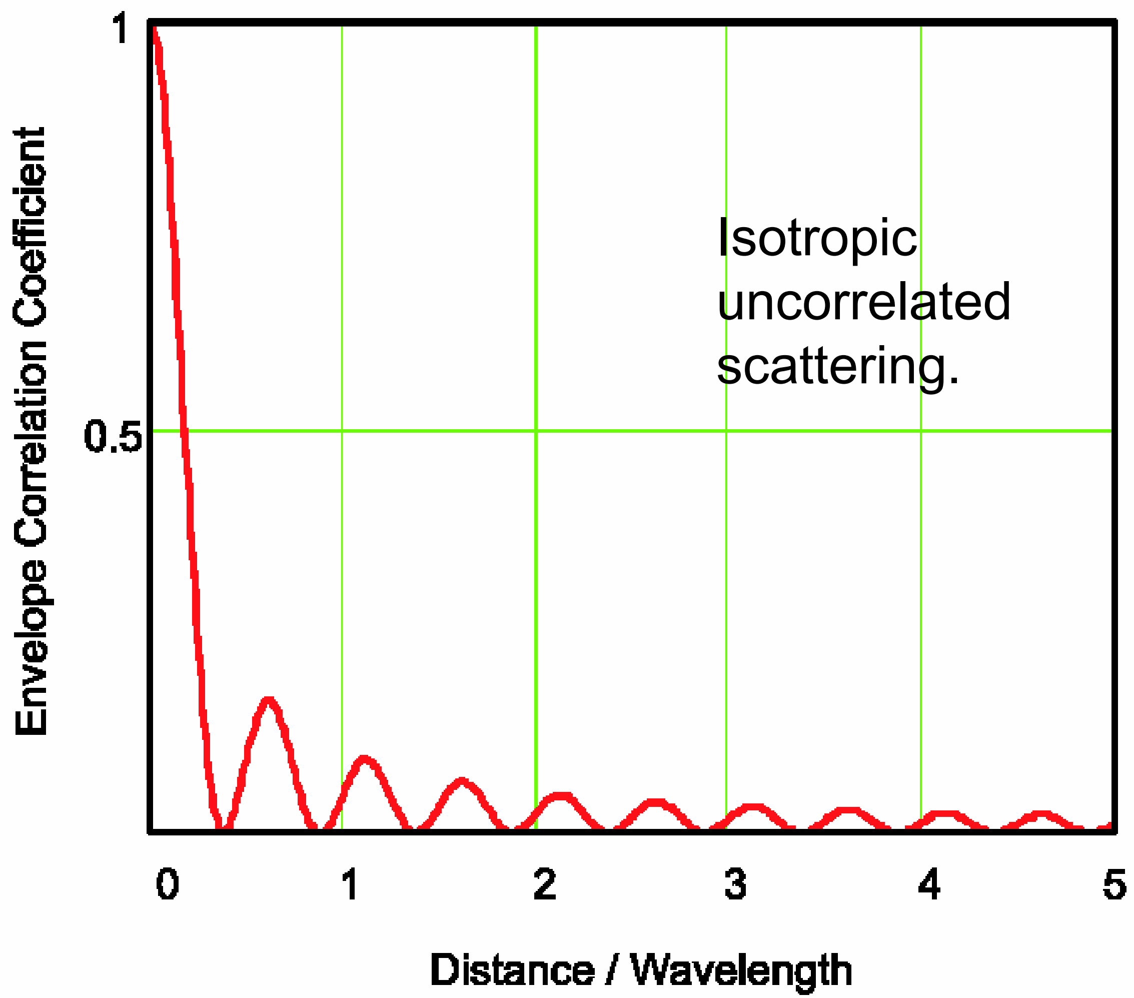
Temporal diversity



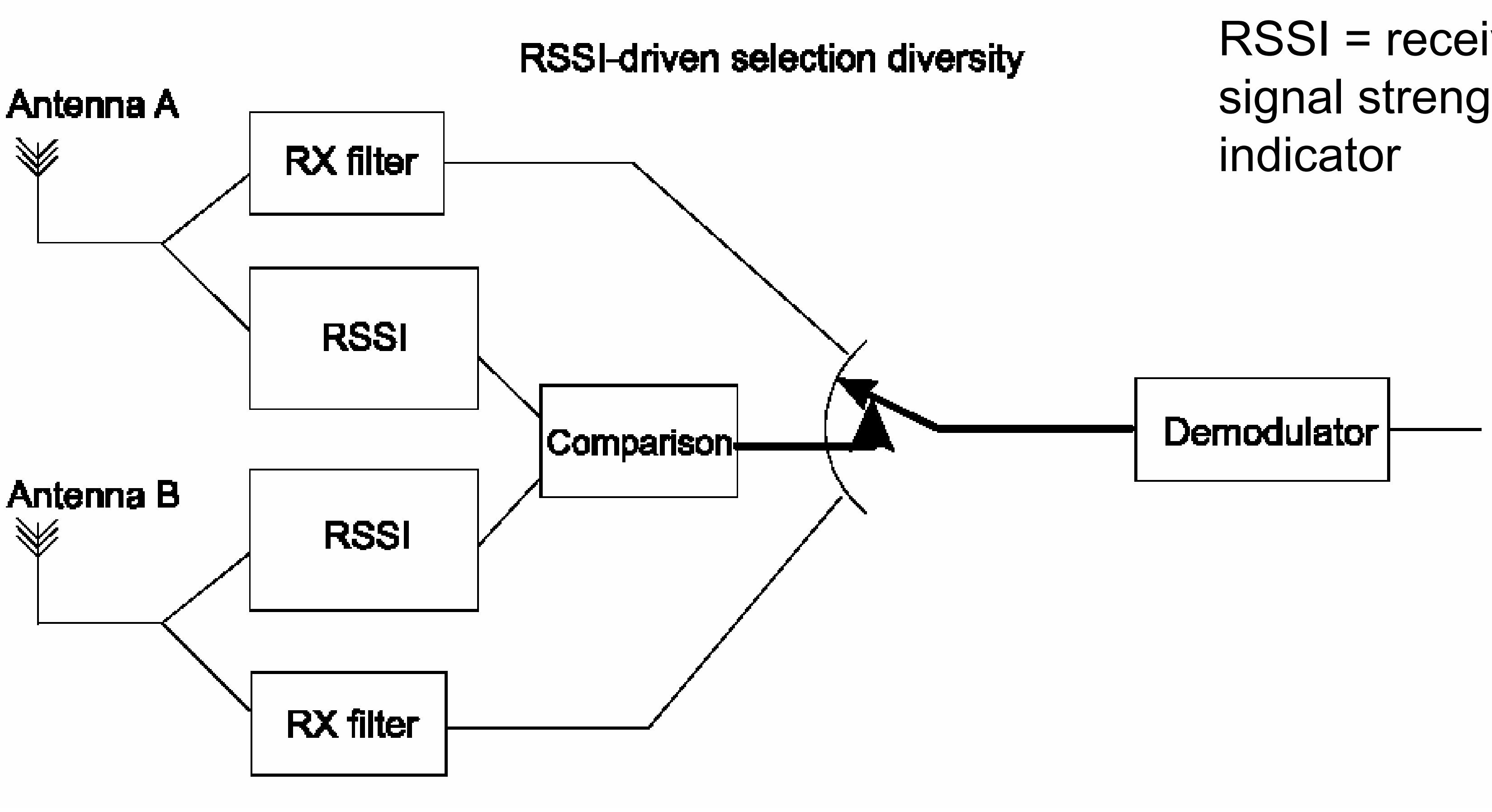


(We also have angular and polarization diversity)

Spatial (antenna) diversity
Fading correlation on antennas

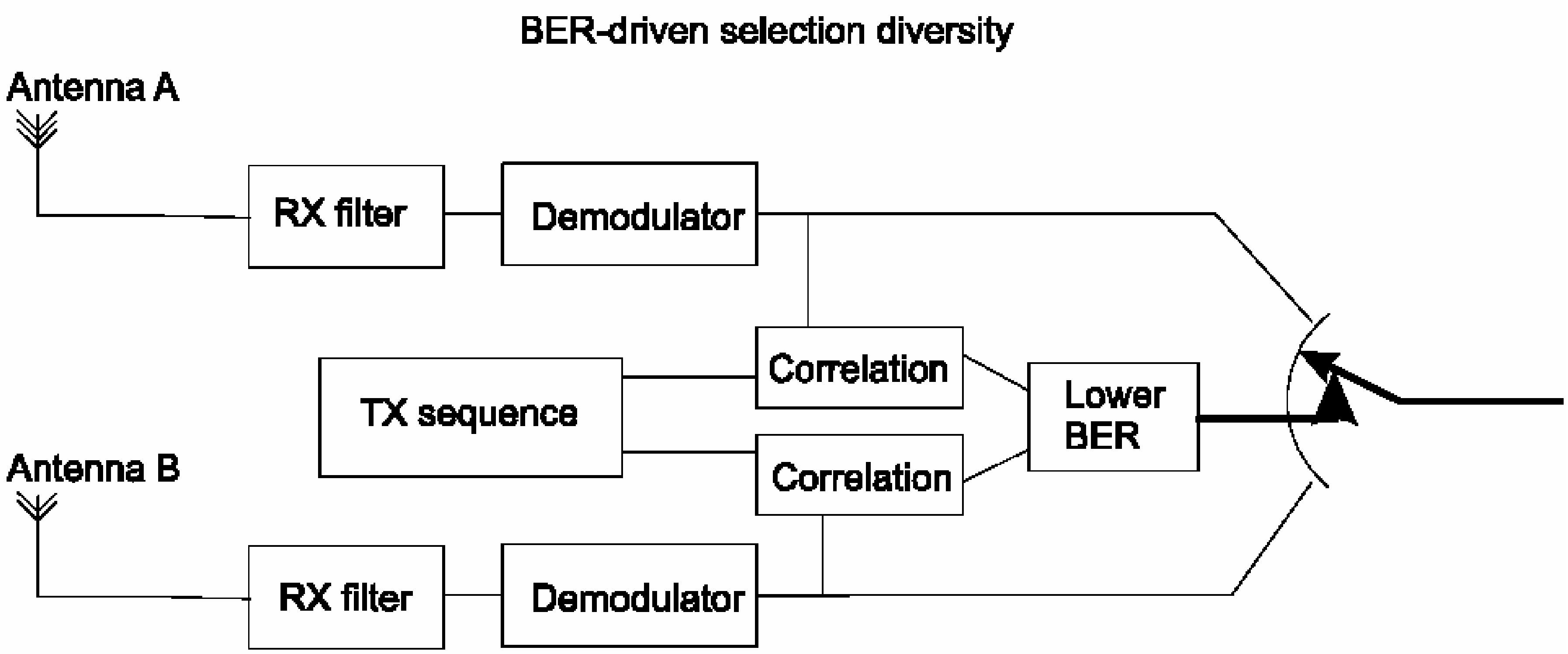


Spatial (antenna) diversity
Selection diversity

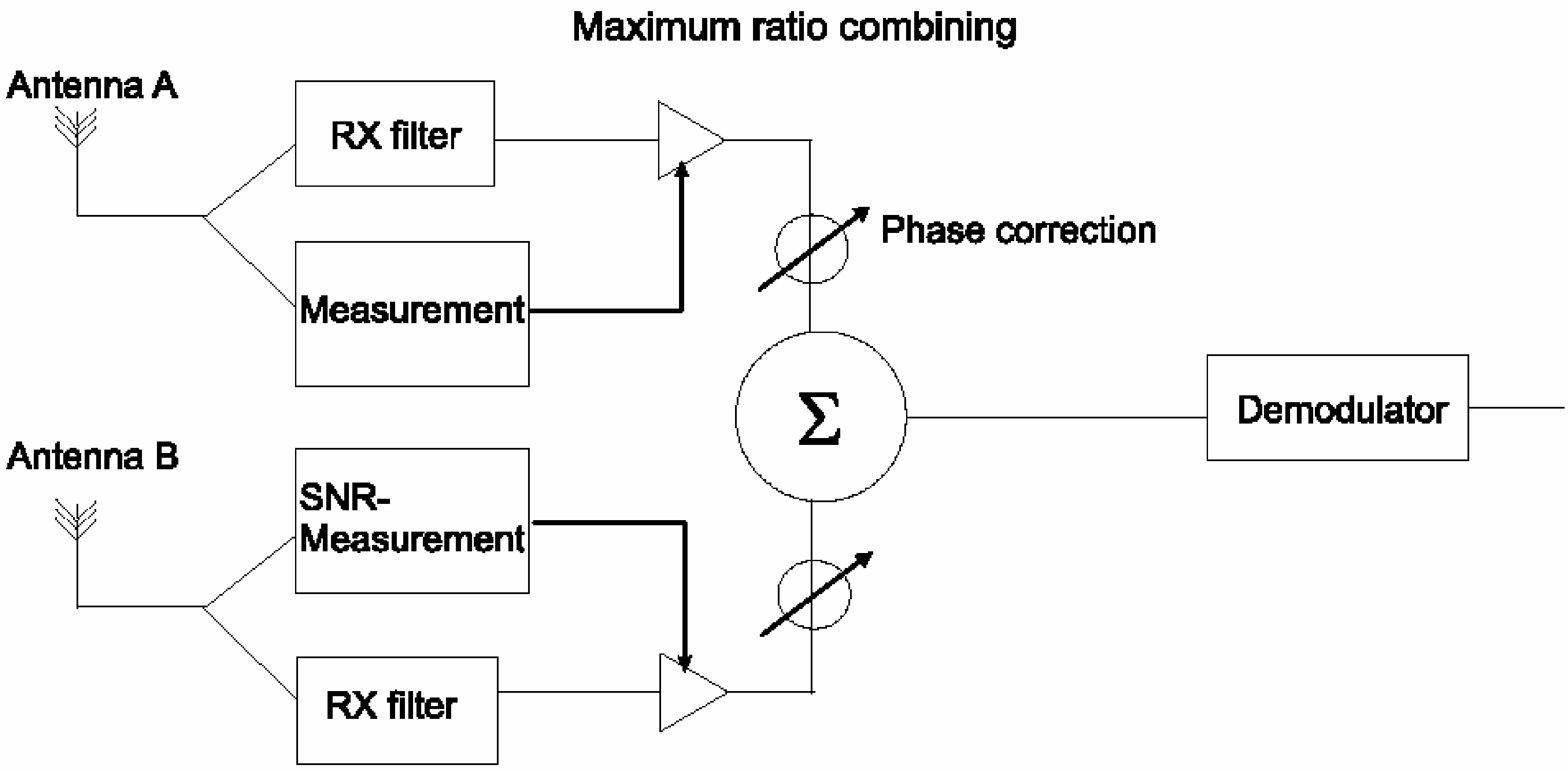


RSSI = received signal strength indicator

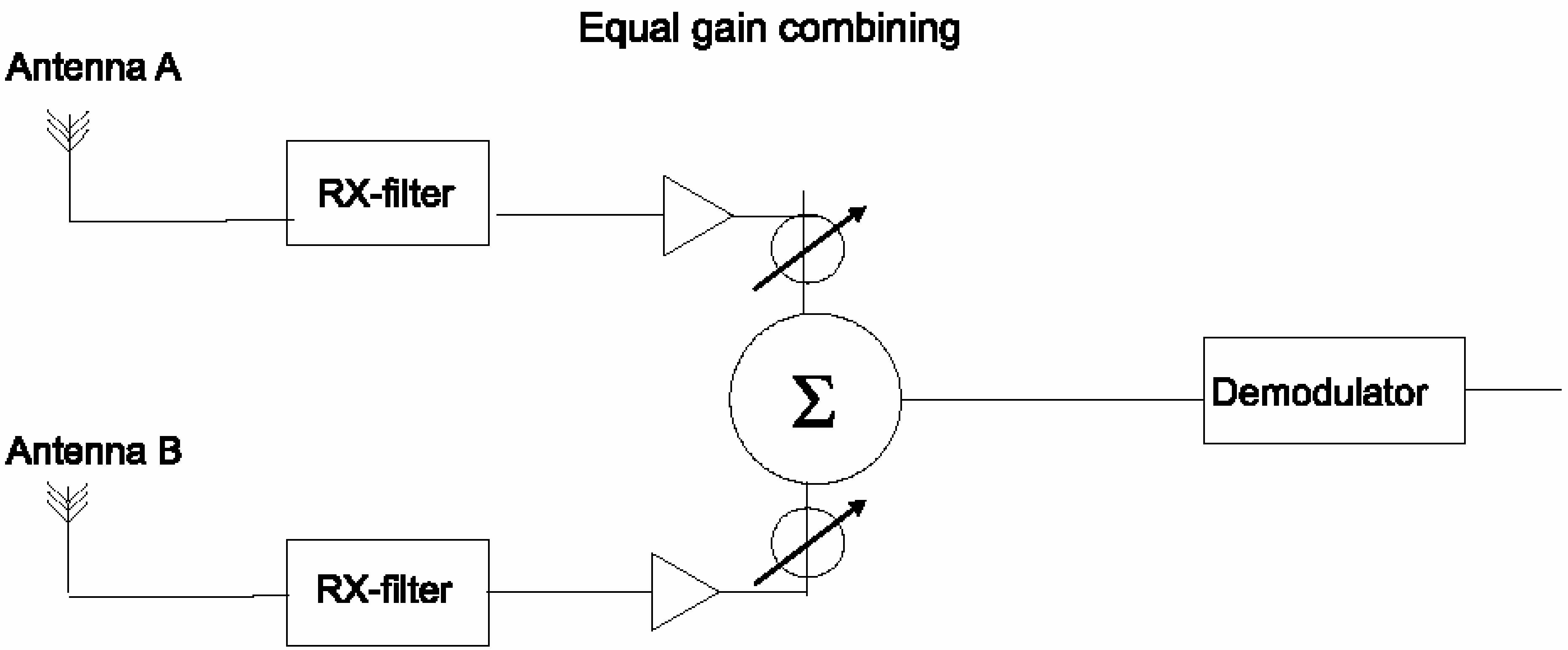
Spatial (antenna) diversity
Selection diversity, cont.



Spatial (antenna) diversity
Maximum ratio combining



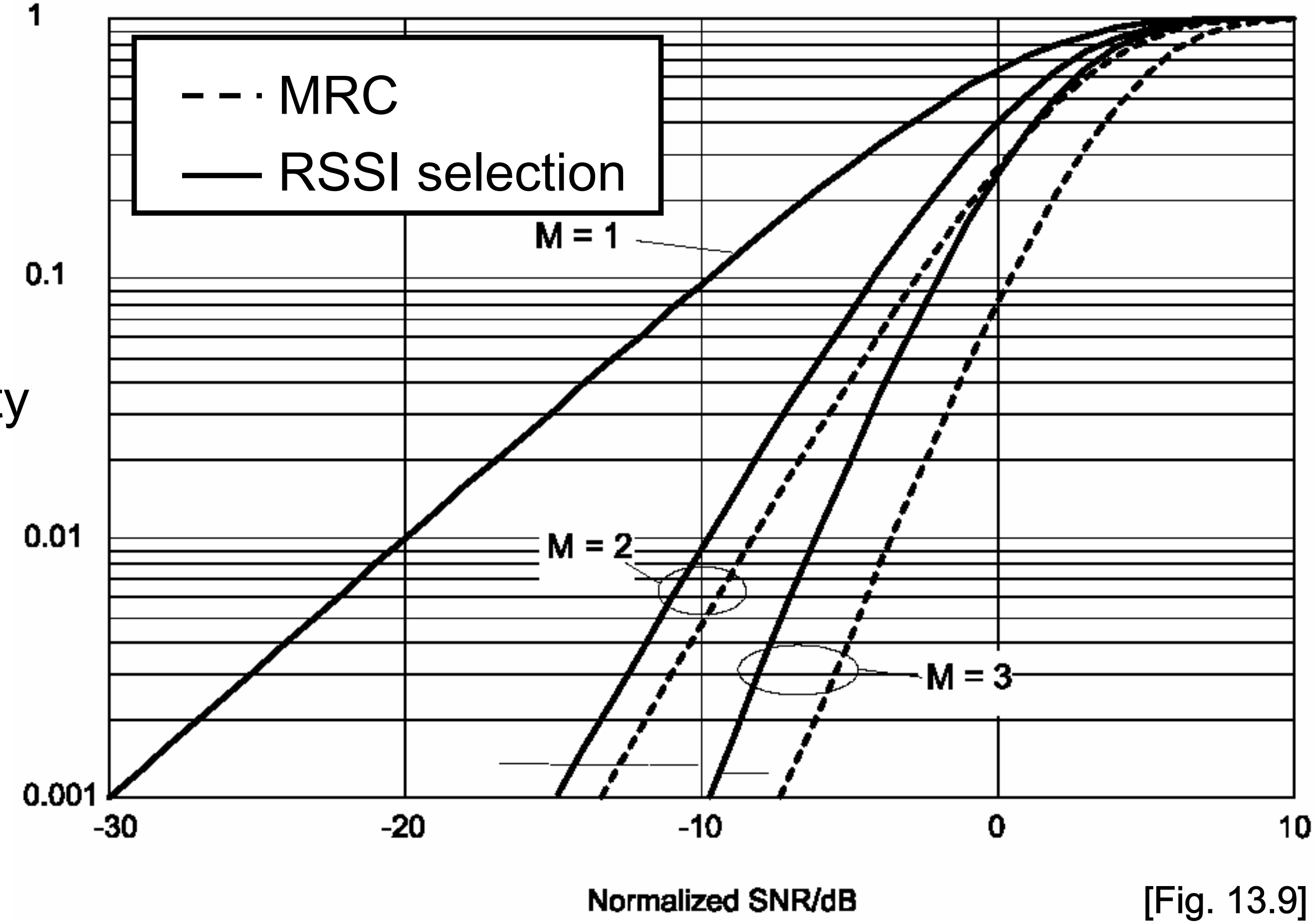
Spatial (antenna) diversity



Spatial (antenna) diversity Performance comparison

Comparison of SNR distribution for different number of antennas *M* and two different diversity techniques.

Cumulative distribution of SNR



[Fig. 13.9]

SNR statistics for diversity receivers

- Selection combining: easiest to compute cdf

$$cdf_\gamma(\gamma) = \left[1 - \exp(-\tfrac{\gamma}{\overline{\gamma}})\right]^{N_\mathrm{r}}.$$

- Maximum ratio combining:

$$pdf_\gamma(\gamma) = \frac{1}{(N_\mathrm{r}-1)!}\frac{\gamma^{N_\mathrm{r}-1}}{\overline{\gamma}^{N_\mathrm{r}}}\exp\left(-\frac{\gamma}{\overline{\gamma}}\right).$$

BER of diversity receivers

- Classical computation method: average BER over distribution of SNR output

$$\overline{SER} = \int_0^\infty pdf_\gamma(\gamma) SER(\gamma) d\gamma$$

- Use SNR distribution from previous slides
- For MRC and large SNR

$$\overline{BER} \approx \left(\frac{1}{4\overline{\gamma}} \right)^{N_\mathrm{r}} \binom{2N_\mathrm{r}-1}{N_\mathrm{r}}$$

Computation via moment-generating function

- BER in AWNG can be written as

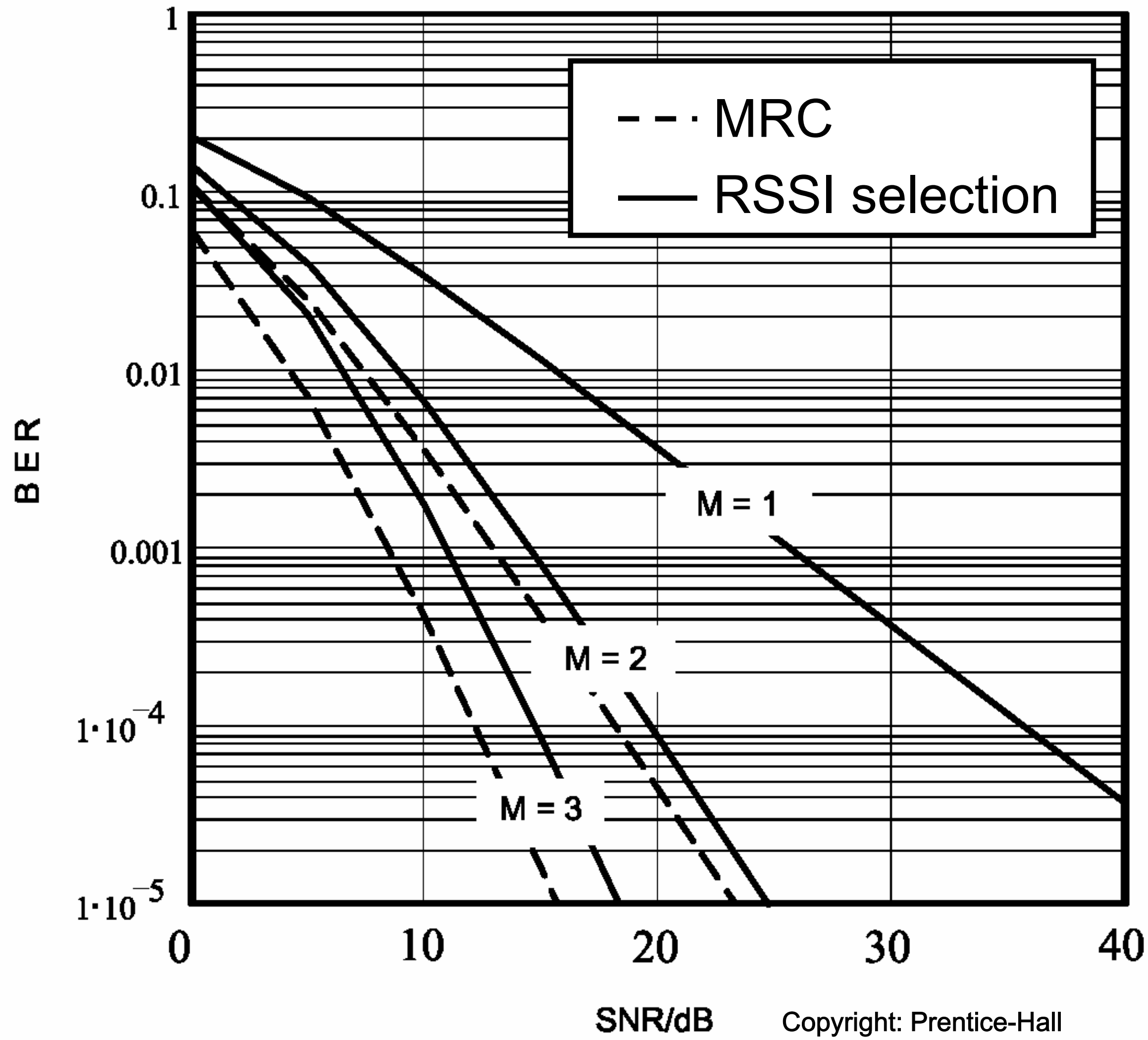
$$SER(\gamma) = \int_{\theta_1}^{\theta_2} f_1(\theta) \prod_{i=1}^{N_r} \exp(-\gamma_n f_2(\theta)) d\theta$$

- Averaging over SNR distribution

$$\begin{aligned}
\overline{SER} &= \int d\gamma_1 pdf_{\gamma_1}(\gamma_1) \int d\gamma_2 pdf_{\gamma_2}(\gamma_2) \ldots \int d\gamma_{N_r} pdf_{\gamma_{N_r}}(\gamma_{N_r}) \int_{\theta_1}^{\theta_2} d\theta\, f_1(\theta) \prod_{i=1}^{N_r} \exp(-\gamma_n f_2(\theta)) \\
&= \int_{\theta_1}^{\theta_2} d\theta\, f_1(\theta) \prod_{i=1}^{N_r} \int d\gamma_n pdf_{\gamma_n}(\gamma_n) \exp(-\gamma_n f_2(\theta)) \\
&= \int_{\theta_1}^{\theta_2} d\theta\, f_1(\theta) \prod_{i=1}^{N_r} M_\gamma(-f_2(\theta)) \\
&= \int_{\theta_1}^{\theta_2} d\theta\, f_1(\theta) [M_\gamma(-f_2(\theta))]^{N_r}
\end{aligned}$$

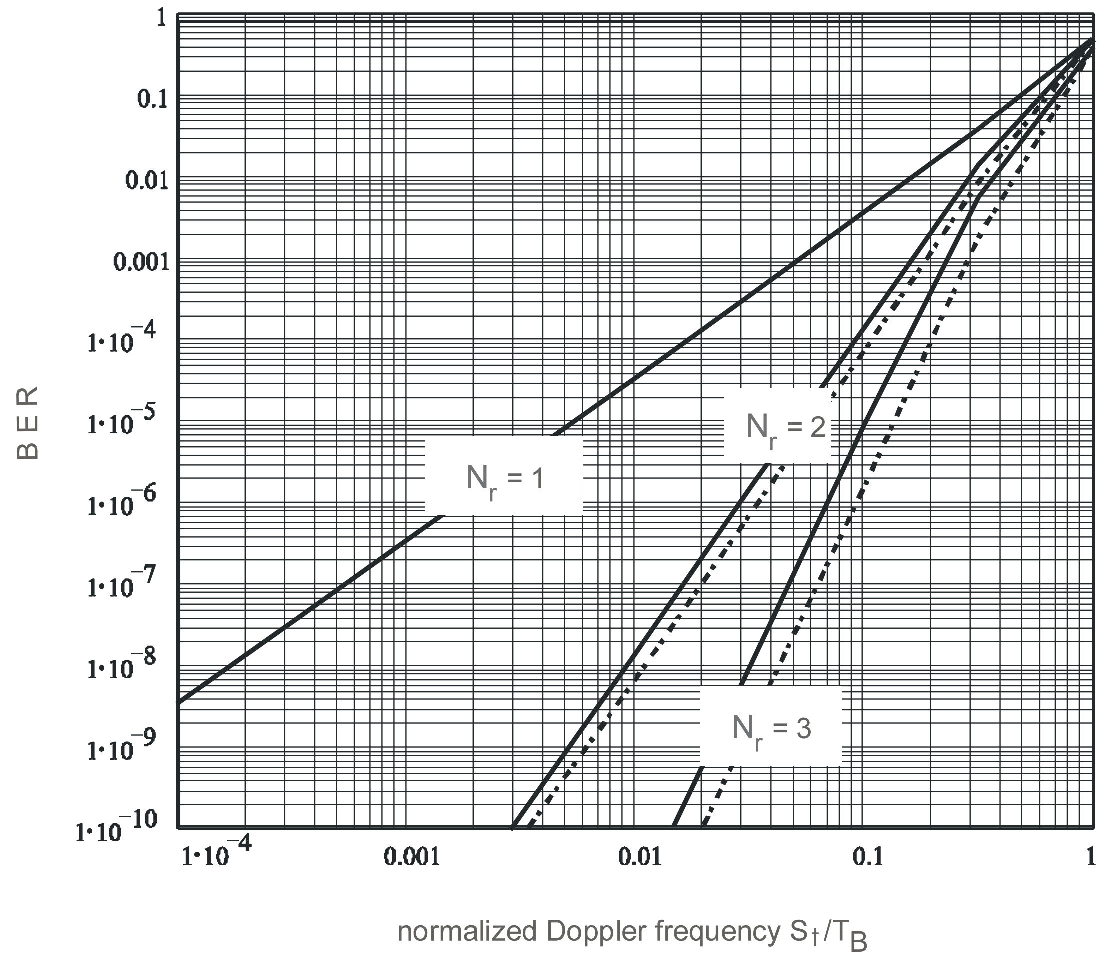
Spatial (antenna) diversity
Performance comparison, cont.

Comparison of 2ASK/2PSK BER for different number of antennas *M* and two different diversity techniques.



Spatial (antenna) diversity
Errors due to signal distortion

Comparison of 2ASK/2PSK BER for different number of antennas *M* and two different diversity techniques.



Optimum combining in flat-fading channel

- Most systems interference limited
- OC reduces not only fading but also interference
- Each antenna can eliminate one interferer or give one diversity degree for fading reduction:

  ("zero-forcing").
- MMSE or decision-feedback gives even better results
- Computation of weights for combining

$$\mathbf{w}_{\mathrm{opt}} = \mathbf{R}^{-1}\mathbf{h}_{\mathrm{d}}^{*} \qquad \mathbf{R} = \sigma_{\mathrm{n}}^{2}\mathbf{I} + \sum_{k=1}^{K} E\{\mathbf{r}_k^{*}\mathbf{r}_k^{T}\}$$

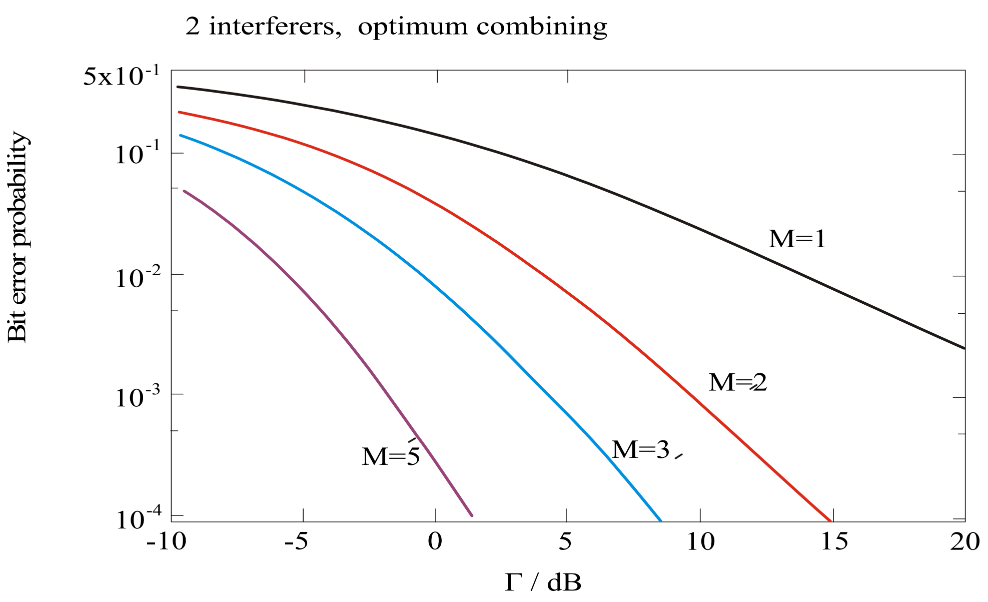
Performance of Optimum Combining

- Define channel matrix H:
  H_{km} is transfer function for k-th user to m-th diversity antenna
- Error of BPSK, QPSK for one channel constellation bounded as

$$BER_{static} \leq \exp\left[-\boldsymbol{h}_d^{\ H} \boldsymbol{R}_{ni}^{\ -1} \boldsymbol{h}_d\right]$$

- average behavior:

$$BER \leq \left[1 + SNR\right]^{-(M-K)}$$



From Winters 1984,

Chapter 14

Channel coding

Contents

- Overview
- Block codes
- Convolution codes
- Trellis-coded modulation
- Turbo codes and LDPC codes
- Fading channel and interleaving

OVERVIEW

Basic types of codes

Channel codes are used to add protection against errors in the channel.

It can be seen as a way of increasing the distance between transmitted alternatives, so that a receiver has a better chance of detecting the correct one in a noisy channel.

We can classify channel codes in two principal groups:

BLOCK CODES	CONVOLUTION CODES
Encodes data in blocks of *k*, using code words of length *n*.	Encodes data in a stream, without breaking it into blocks, creating code sequences.

Information and redundancy (1)

EXAMPLE

Is the English language protected by a code, allowing us to correct transmission errors?

When receiving the following sentence with errors marked by ´-´:

> “D- n-t w-rr- -b--t ---r d-ff-cult--s -n M-th-m-t-cs.
> \- c-n -ss-r- --- m-n- -r- st-ll gr--t-r.”

it can still be “decoded” properly.

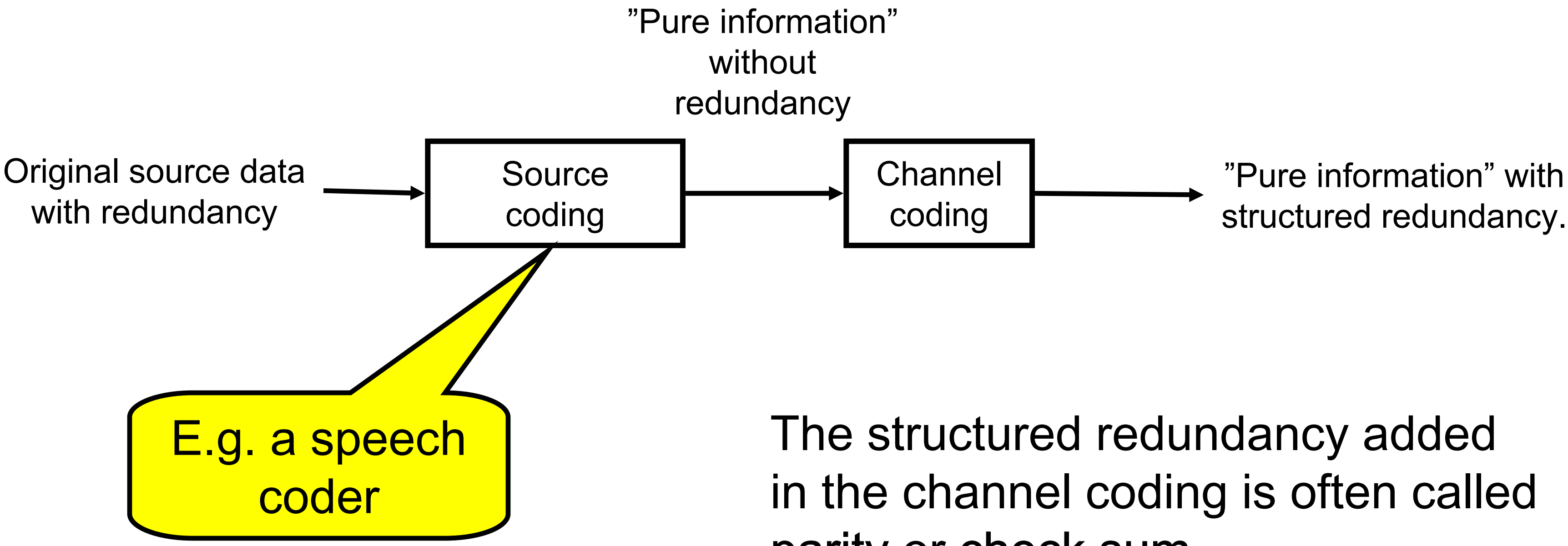
What does it say, and who is quoted?

There is something more than information in the original sentence that allows us to decode it properly, redundancy.

Redundancy is available in almost all “natural” data, such as text, music, images, etc.

Information and redundancy (2)

Electronic circuits do not have the power of the human brain and needs more structured redundancy to be able to decode “noisy” messages.

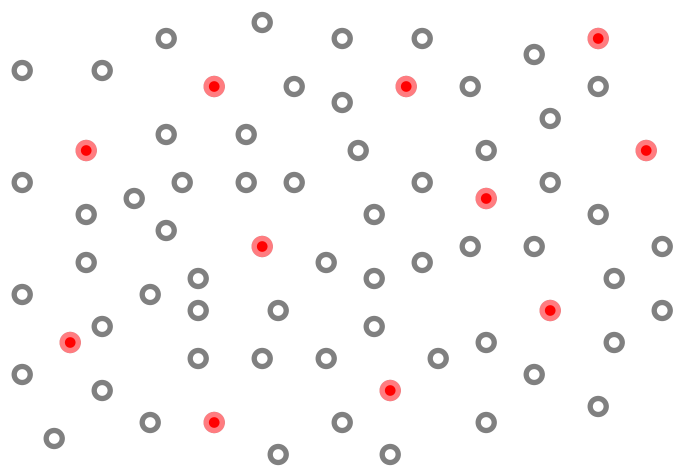


The structured redundancy added in the channel coding is often called parity or check sum.

Illustration of code words

Assume that we have a block code, which consists of k information bits per n bit code word ($n > k$).

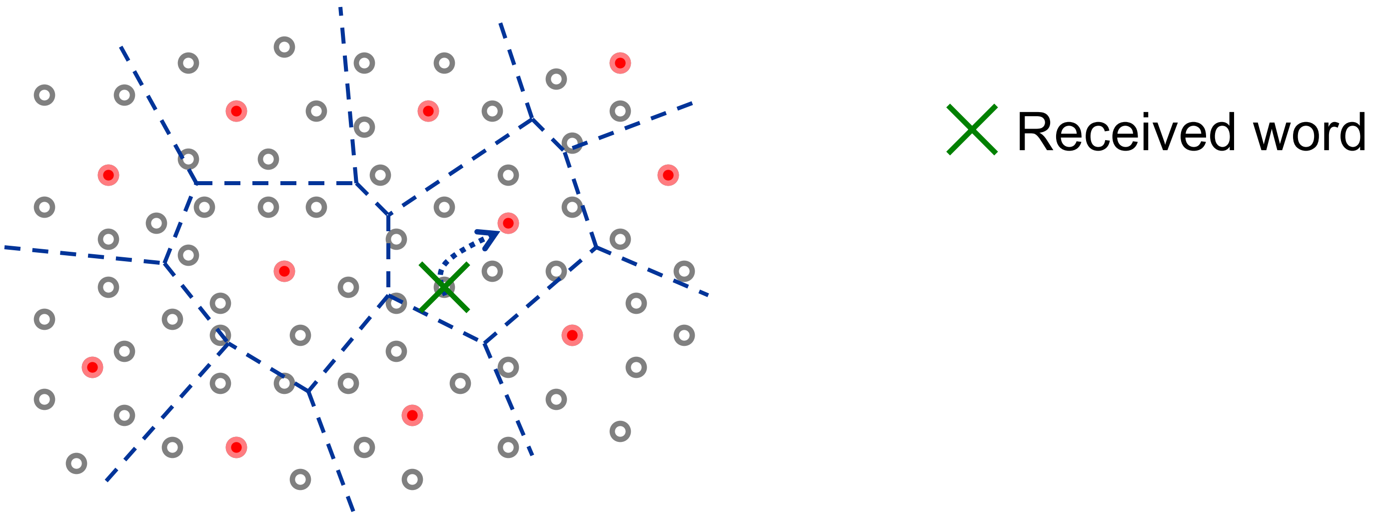
Since there are only 2^k different information sequences, there can be only 2^k different code words.



2^n different binary sequences of length n.

Only 2^k are valid code words in our code.

Illustration of decoding



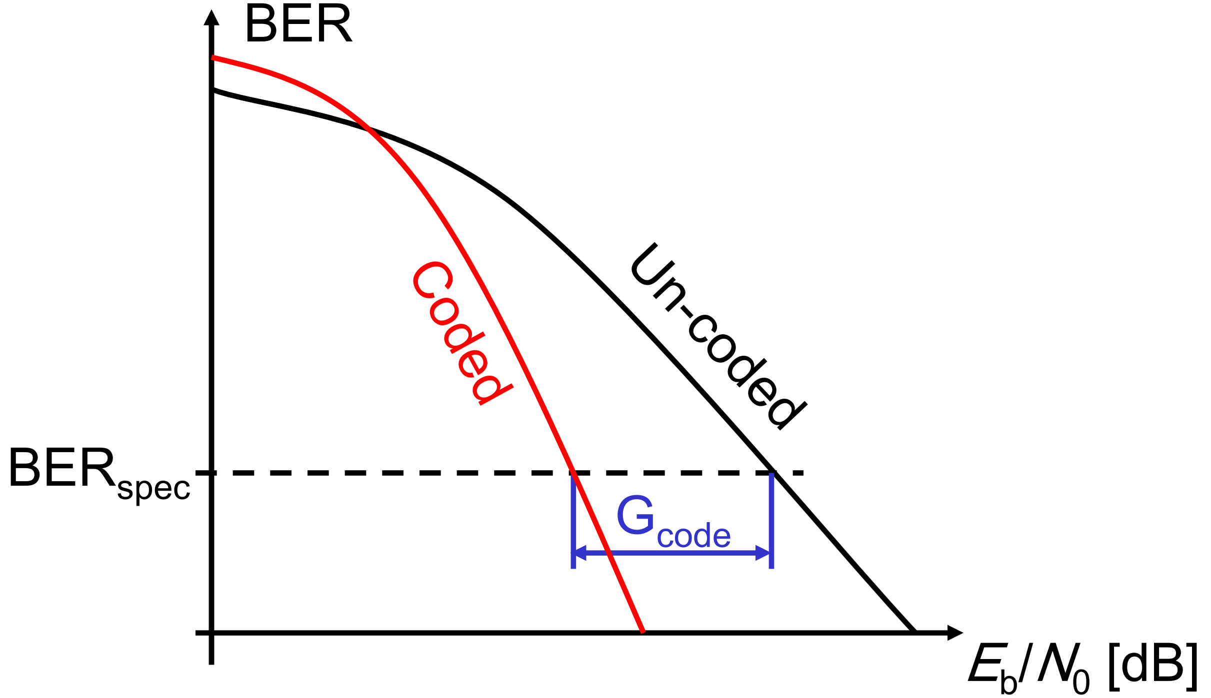
Distances

Two common ones:

Hamming distance	Measures the number of bits being different between two binary words.	Used for binary channels with random bit errors.
Euclidean distance	Same measure we have used for signal constellations.	Used for AWGN channels.

Coding gain

When applying channel codes we decrease the E_b/N_0 required to obtain some specified performance (BER).



BLOCK CODES

Channel coding
Linear block codes

The encoding process of a linear block code can be written as

$$\vec{x} = \underline{G}\vec{u}$$

where

$\vec{u}$	k - dimensional information vector
$\underline{G}$	n x k - dimensional generator matrix
$\vec{x}$	n - dimensional code word vector

Channel coding
Some definitions

Code rate:

$$R = \frac{\text{bits in}}{\text{bits out}} = \frac{k}{n}$$

Modulo-2 arithmetic (XOR):

$$\vec{x}_i + \vec{x}_j = \begin{bmatrix} 0 \\ 1 \\ 1 \end{bmatrix} + \begin{bmatrix} 0 \\ 0 \\ 1 \end{bmatrix} = \begin{bmatrix} 0 \\ 1 \\ 0 \end{bmatrix}$$

Hamming weight:

$$w(\vec{x}) = \text{number of ones in } \vec{x}$$

Hamming distance:

$$d(\vec{x}_i, \vec{x}_j) = w(\vec{x}_i + \vec{x}_j)$$

Minimum distance of code:

$$\begin{aligned} d_{\min} &= \min_{i \neq j} d(\vec{x}_i, \vec{x}_j) \\ &= \min_{i \neq j} w(\vec{x}_i + \vec{x}_j) \end{aligned}$$

Channel coding
Encoding example

For a specific (n,k) = (7,4) code we encode the information sequence 1 0 1 1 as

$$\underbrace{\begin{bmatrix} 1 & 0 & 0 & 0 \\ 0 & 1 & 0 & 0 \\ 0 & 0 & 1 & 0 \\ 0 & 0 & 0 & 1 \\ 1 & 1 & 0 & 1 \\ 1 & 0 & 1 & 1 \\ 0 & 1 & 1 & 1 \end{bmatrix}}_{\text{Generator matrix}} \begin{bmatrix} 1 \\ 0 \\ 1 \\ 1 \end{bmatrix} = \begin{bmatrix} 1 \\ 0 \\ 1 \\ 1 \\ 0 \\ 1 \\ 0 \end{bmatrix} \begin{matrix} \left.\begin{matrix} \\ \\ \\ \\ \end{matrix}\right\} \text{Systematic bits} \\ \left.\begin{matrix} \\ \\ \\ \end{matrix}\right\} \text{parity bits.} \end{matrix}$$

Channel coding
Encoding example, cont.

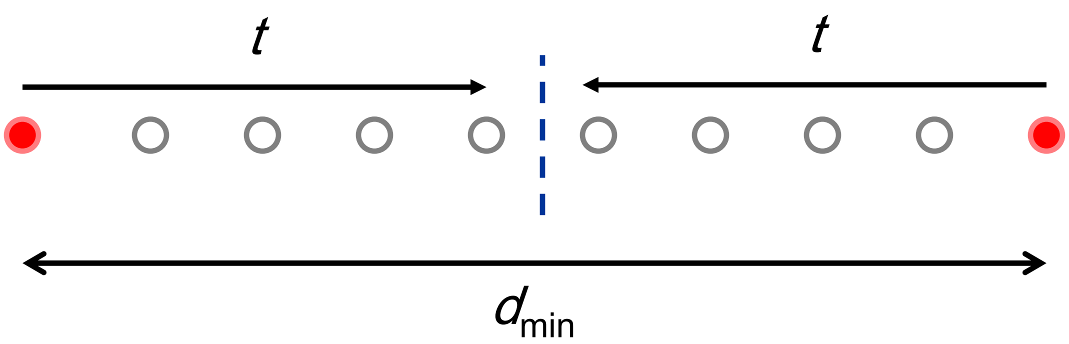
Encoding all possible 4 bit information sequences gives:

Information				Code word							Hamming weight
0	0	0	0	0	0	0	0	0	0	0	0
0	0	0	1	0	0	0	1	1	1	1	4
0	0	1	0	0	0	1	0	0	1	1	3
0	0	1	1	0	0	1	1	1	0	0	3
0	1	0	0	0	1	0	0	1	0	1	3
0	1	0	1	0	1	0	1	0	1	0	3
0	1	1	0	0	1	1	0	1	1	0	4
0	1	1	1	0	1	1	1	0	0	1	4
1	0	0	0	1	0	0	0	1	1	0	3
1	0	0	1	1	0	0	1	0	0	1	3
1	0	1	0	1	0	1	0	1	0	1	4
1	0	1	1	1	0	1	1	0	1	0	4
1	1	0	0	1	1	0	0	0	1	1	4
1	1	0	1	1	1	0	1	1	0	0	4
1	1	1	0	1	1	1	0	0	0	0	3
1	1	1	1	1	1	1	1	1	1	1	7

This is a (7,4) Hamming code, capable of correcting one bit error.

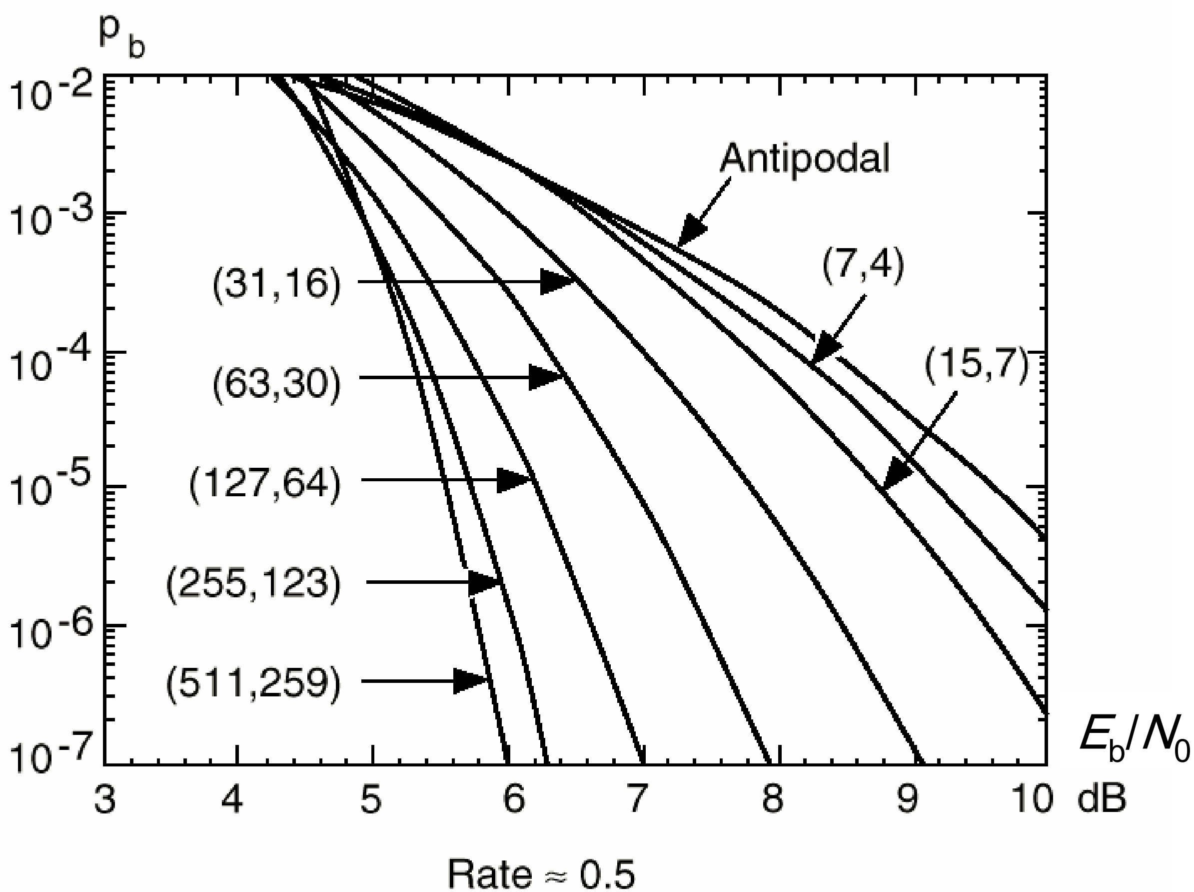
Channel coding
Error correction capability

$$t = \left\lfloor \frac{d_{\min} - 1}{2} \right\rfloor$$



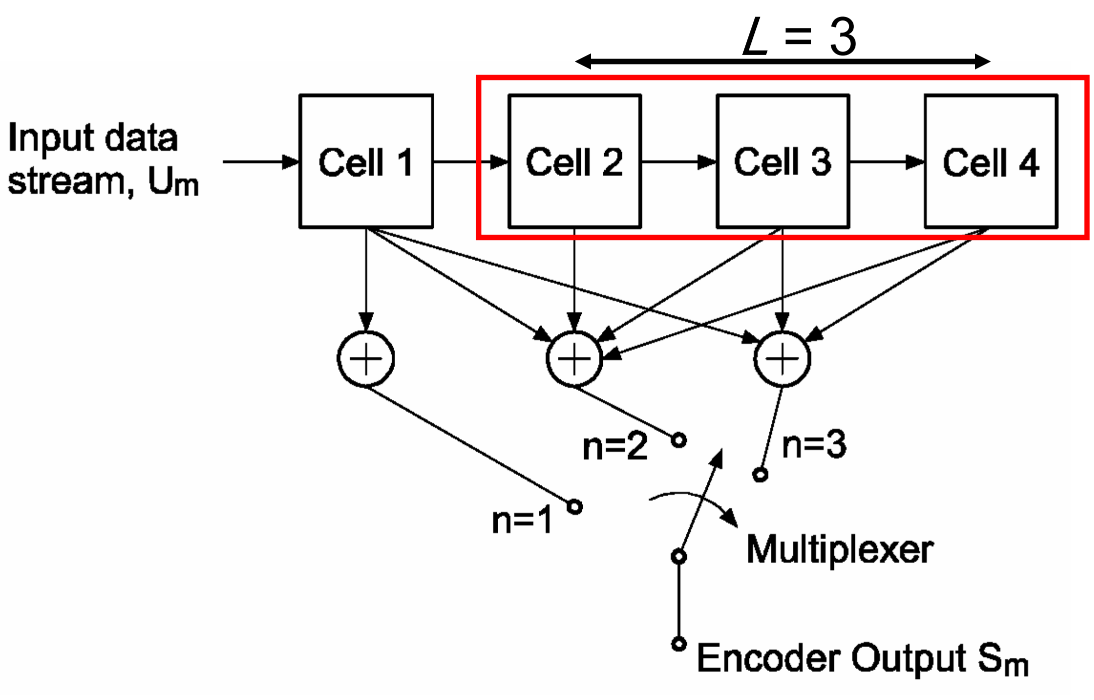
From Ericsson radio school

Channel coding
Performance and code length

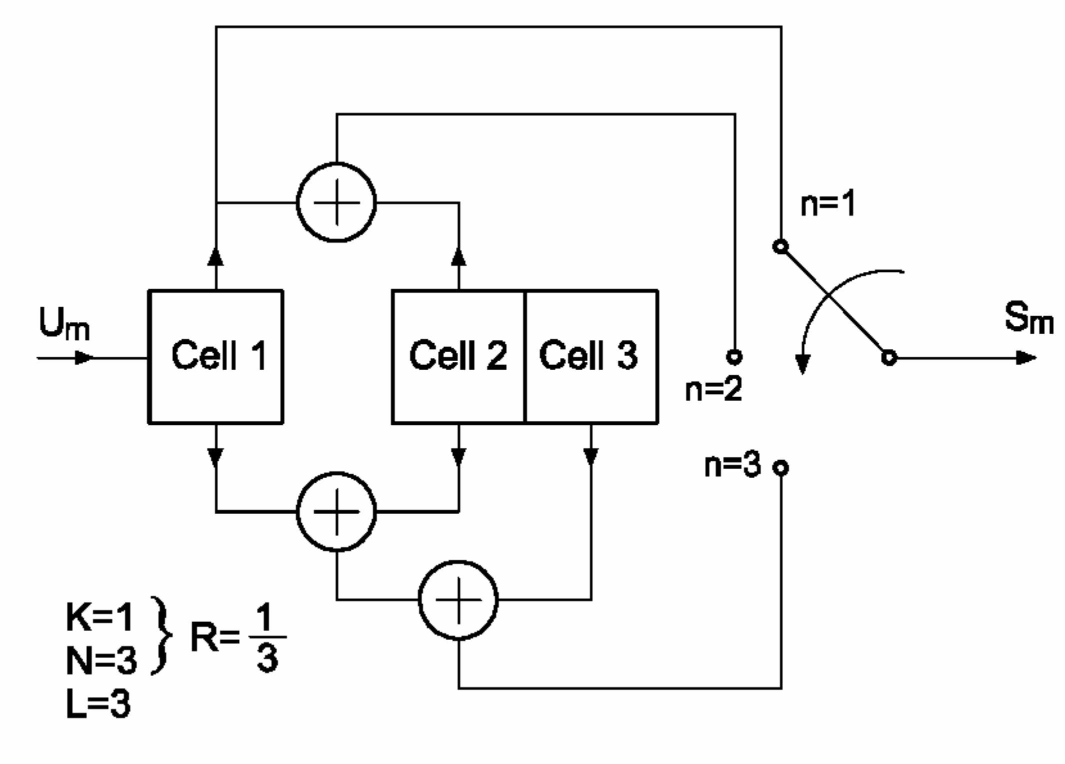


CONVOLUTION CODES

Channel coding
Encoder structure



Channel coding
Encoding example



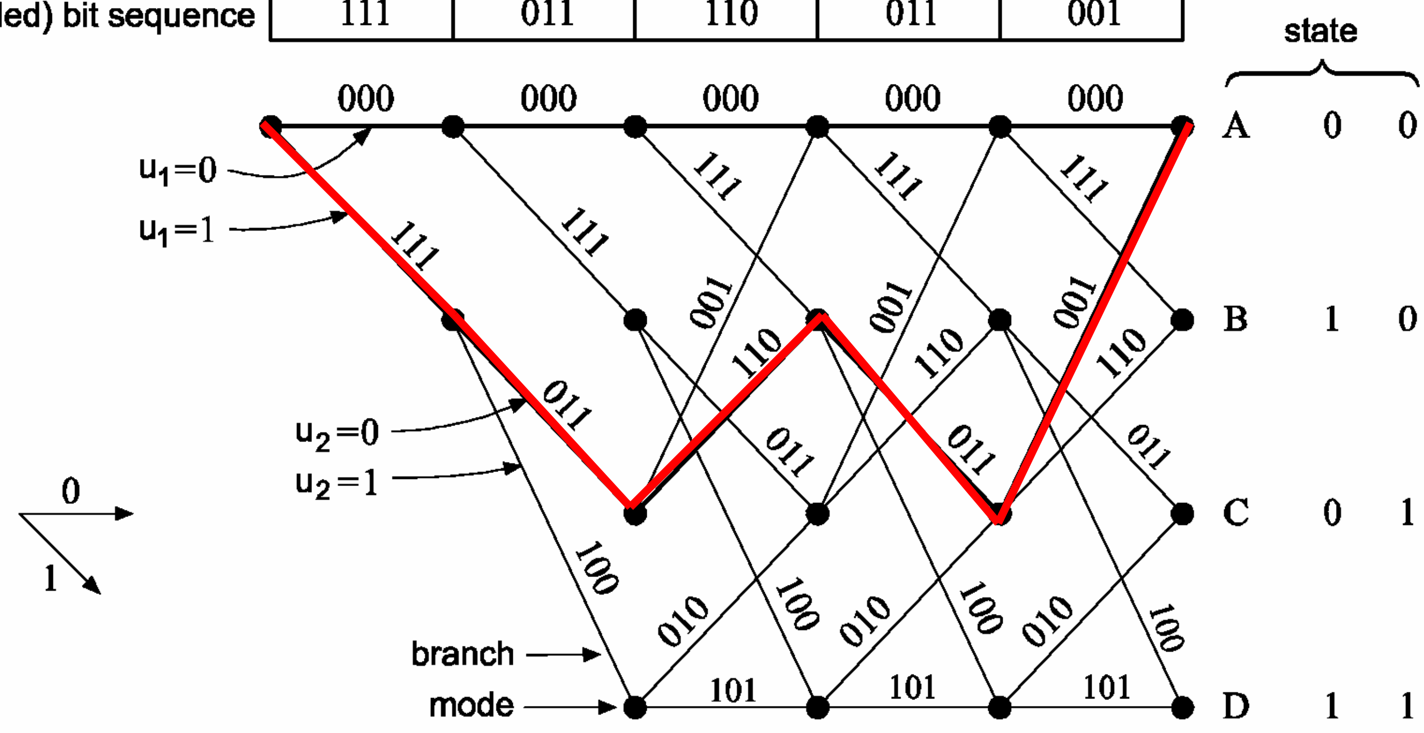
Input	State	Output	Next state
0	00	000	00
1	00	111	10
0	01	001	00
1	01	110	10
0	10	011	01
1	10	100	11
0	11	010	01
1	11	101	11

We usually start the encoder in the all-zero state!

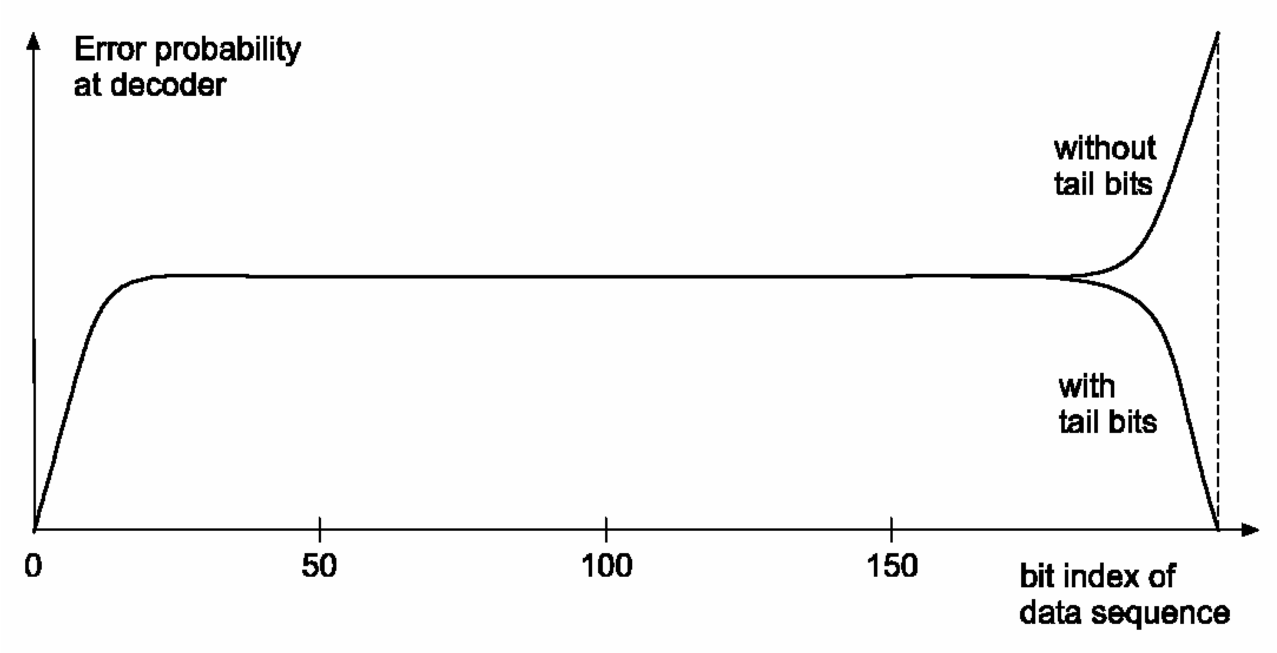
Channel coding
Encoding example, cont.

We can view the encoding process in a trellis created from the table on the previous slide.

Input (information) bit sequence	1	0	1	0	0
Output (coded) bit sequence	111	011	110	011	001

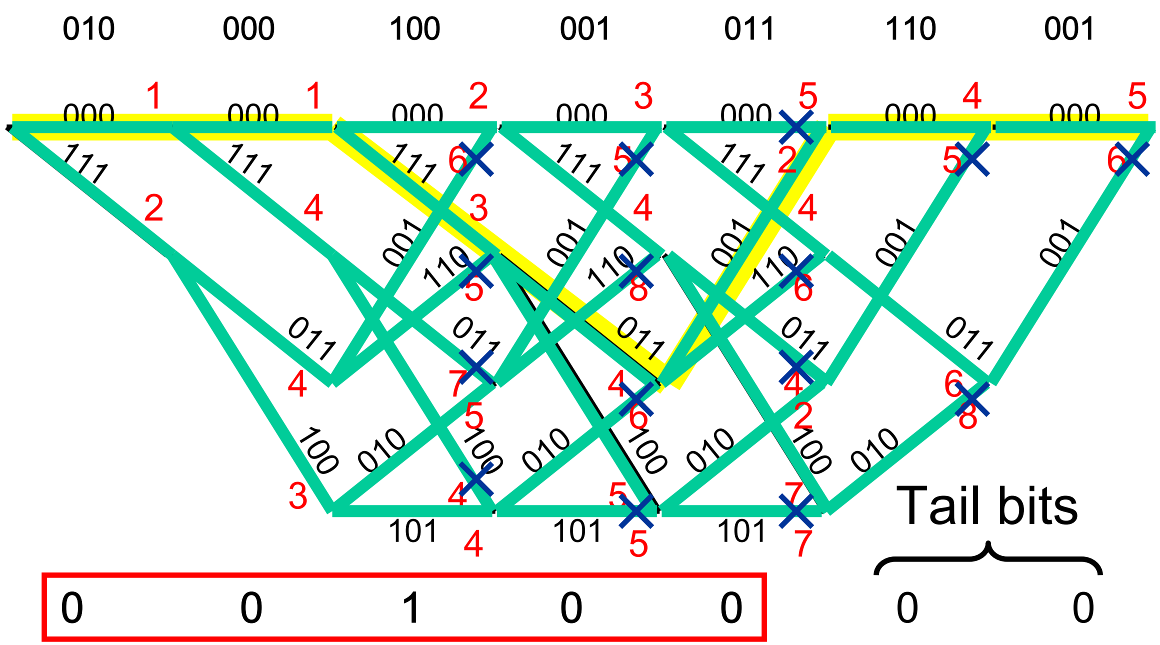


Channel coding
Termination



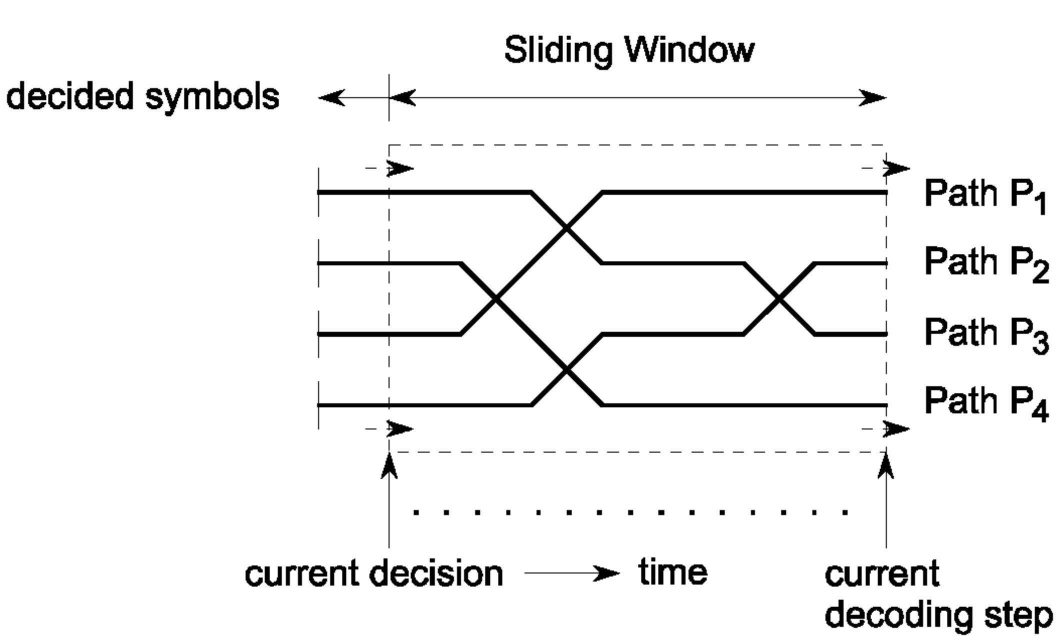
Channel coding
A Viterbi decoding example

Received sequence:



Decoded data:

Channel coding
Surviving paths



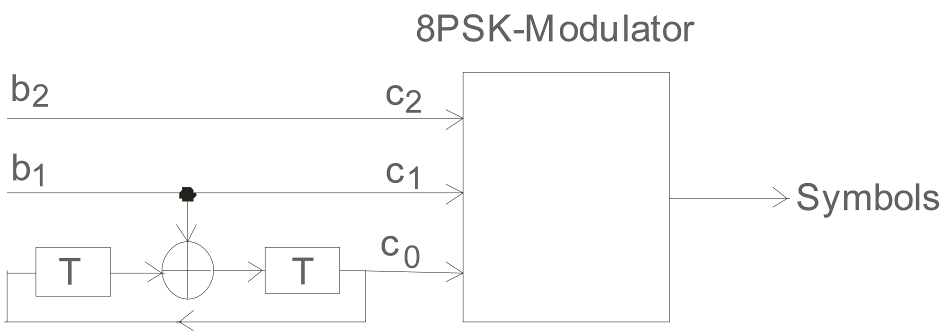
TRELLIS-CODED MODULATION

Principle of TCM

- Goal: improve BER performance while leaving the bandwidth requirement unchanged
- “Conventional” coding introduces redundancy, and therefore increases the requirement for bandwidth
- Therefore, TCM increases the constellation size of the modulation, while at the same time using a convolutional code

Trellis-coded modulation (1)

- Simple example: TCM with 8-PSK and rate 2/3 coding

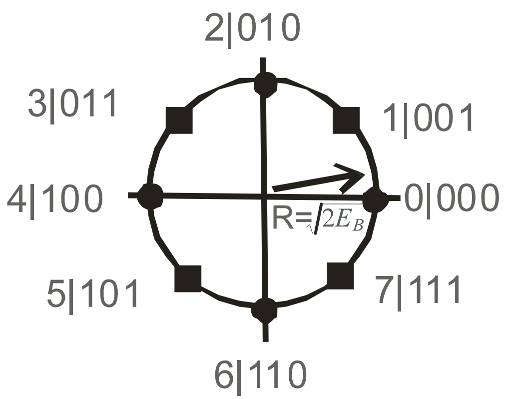


Trellis encoder for 8PSK-TCM

Trellis-coded modulation (2)

Signal-space diagram

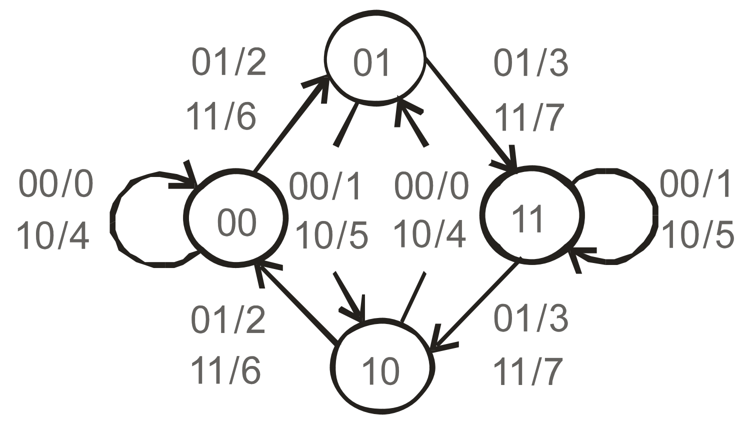
8PSK: A|C_2 C_1 C_0



8PSK signal constellation diagram

Admissible transitions

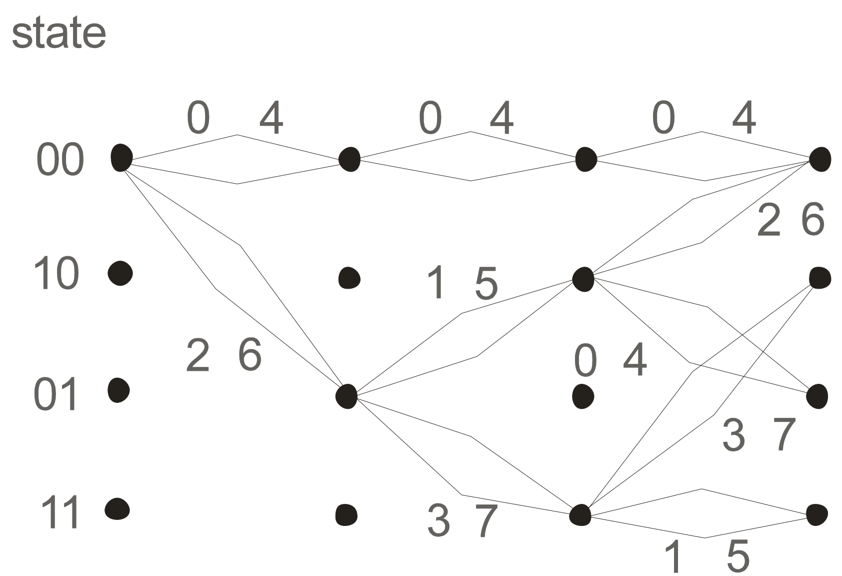
Infobits/8PSK-Symbol



state transition diagram

TCM: BER computation (1)

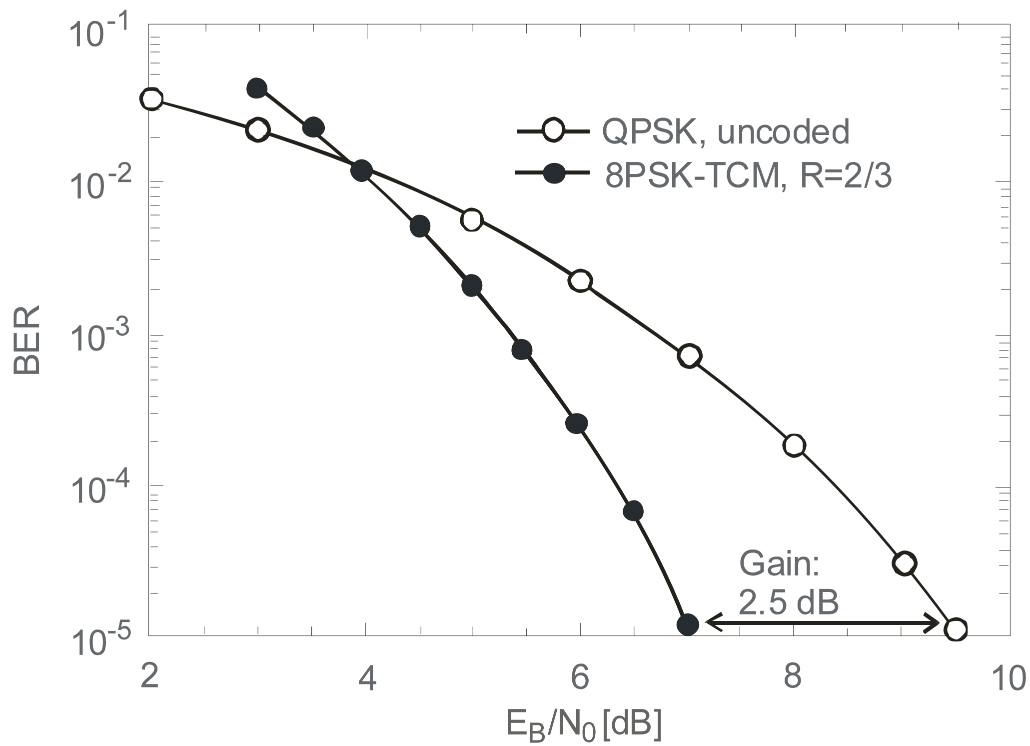
$$d^2 = 8E_{\mathrm{B}}$$



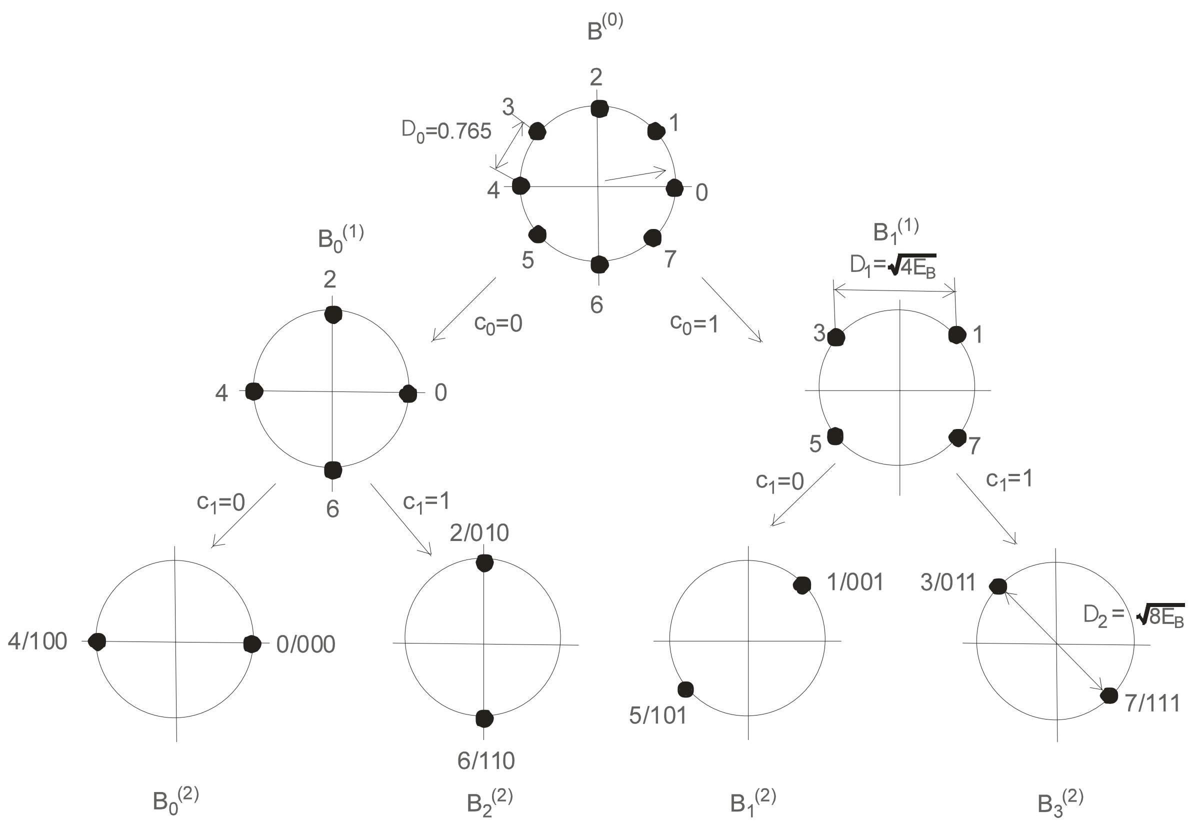
Trellis diagram for 8PSK-TCM

TCM: BER computation (2)

- Asymptotic coding gain of 3 dB
  - Euclidean distance is 8E, compared to 4E for QPSK



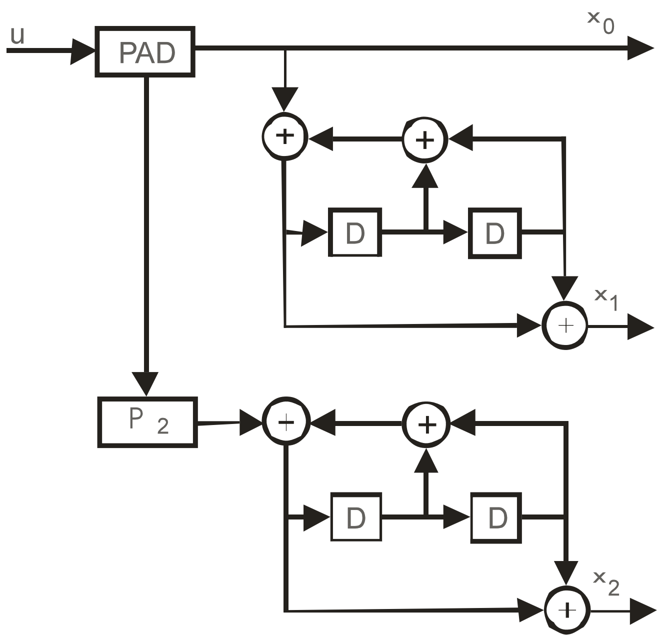
Set partitioning



TURBO CODES AND LDPC CODES

Turbocoders

- Generates long codewords by
  - encoding data with two different convolutional encoders
  - for each of the encoders, data are interleaved with different interleavers



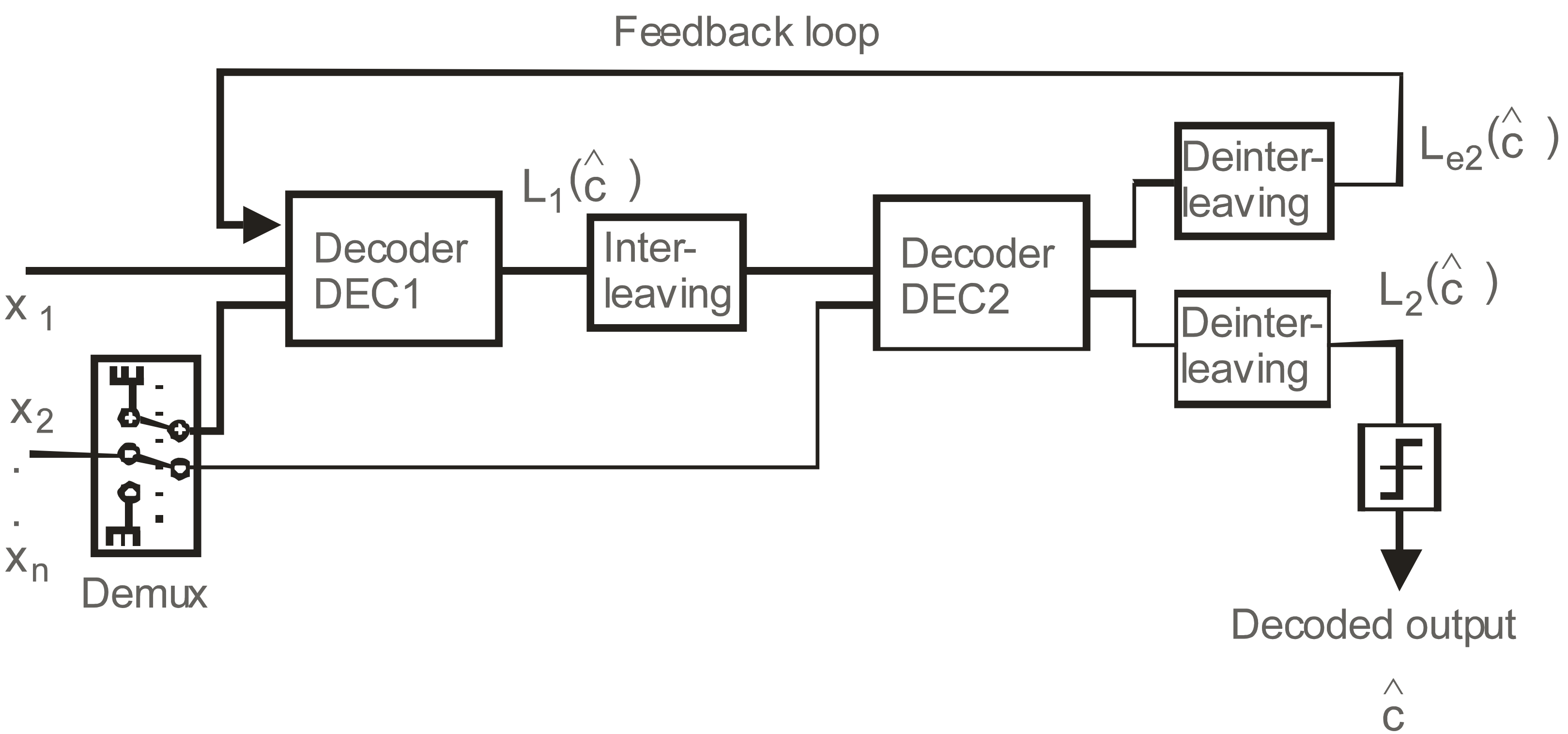
Decoding of turbocodes

- Iterative decoding
- Two separate decoders (corresponding to the two convolutional encoders) that exchange information

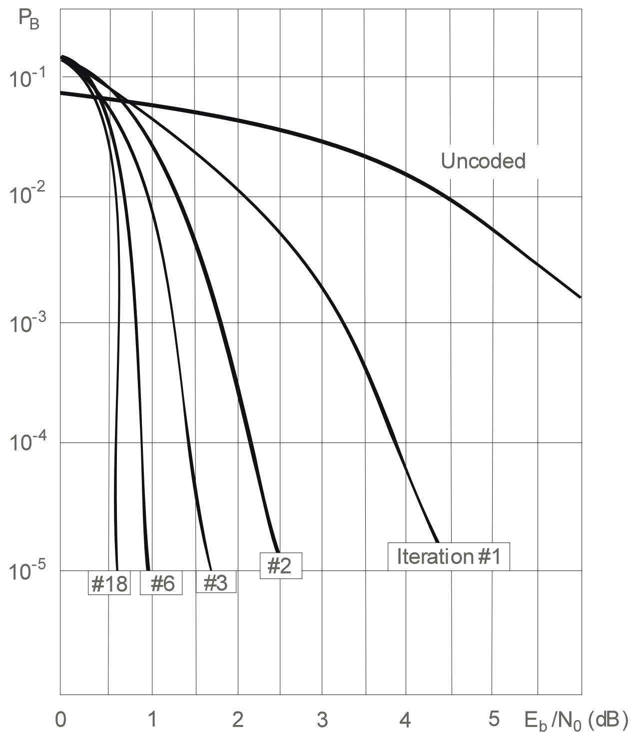
- Quantity of interest is the log-likelihood ratio

$$\log\left[\frac{\Pr(b_i=+1|x)}{\Pr(b_i=-1|x)}\right]$$

Block diagram of turbo decoder



Performance of turbo codes



Principle of LDPC codes

- LDPC: low density parity check codes
- Block codes with large block length
- Defined by the parity-check matrix H, not the generator matrix

Construction of parity-check matrix

1. Divide matrix horizontally into *p* submatrices
2. Put a “1” into each column of the submatrix. Make sure that there are *q* “1”s per row

$$\begin{bmatrix}
1 & 1 & 1 & 1 & 0 & 0 & 0 & 0 & 0 & 0 & 0 & 0 & 0 & 0 & 0 & 0 & 0 & 0 & 0 & 0 \\
0 & 0 & 0 & 0 & 1 & 1 & 1 & 1 & 0 & 0 & 0 & 0 & 0 & 0 & 0 & 0 & 0 & 0 & 0 & 0 \\
0 & 0 & 0 & 0 & 0 & 0 & 0 & 0 & 1 & 1 & 1 & 1 & 0 & 0 & 0 & 0 & 0 & 0 & 0 & 0 \\
0 & 0 & 0 & 0 & 0 & 0 & 0 & 0 & 0 & 0 & 0 & 0 & 1 & 1 & 1 & 1 & 0 & 0 & 0 & 0 \\
0 & 0 & 0 & 0 & 0 & 0 & 0 & 0 & 0 & 0 & 0 & 0 & 0 & 0 & 0 & 0 & 1 & 1 & 1 & 1
\end{bmatrix}$$

3. Let other submatrices be column permutations of first submatrix

$$\mathbf{H} = \begin{bmatrix}
1 & 1 & 1 & 1 & 0 & 0 & 0 & 0 & 0 & 0 & 0 & 0 & 0 & 0 & 0 & 0 & 0 & 0 & 0 & 0 \\
0 & 0 & 0 & 0 & 1 & 1 & 1 & 1 & 0 & 0 & 0 & 0 & 0 & 0 & 0 & 0 & 0 & 0 & 0 & 0 \\
0 & 0 & 0 & 0 & 0 & 0 & 0 & 0 & 1 & 1 & 1 & 1 & 0 & 0 & 0 & 0 & 0 & 0 & 0 & 0 \\
0 & 0 & 0 & 0 & 0 & 0 & 0 & 0 & 0 & 0 & 0 & 0 & 1 & 1 & 1 & 1 & 0 & 0 & 0 & 0 \\
0 & 0 & 0 & 0 & 0 & 0 & 0 & 0 & 0 & 0 & 0 & 0 & 0 & 0 & 0 & 0 & 1 & 1 & 1 & 1 \\
1 & 0 & 0 & 0 & 1 & 0 & 0 & 0 & 1 & 0 & 0 & 0 & 1 & 0 & 0 & 0 & 0 & 0 & 0 & 0 \\
0 & 1 & 0 & 0 & 0 & 1 & 0 & 0 & 0 & 1 & 0 & 0 & 0 & 0 & 0 & 0 & 1 & 0 & 0 & 0 \\
0 & 0 & 1 & 0 & 0 & 0 & 1 & 0 & 0 & 0 & 0 & 0 & 0 & 1 & 0 & 0 & 0 & 1 & 0 & 0 \\
0 & 0 & 0 & 1 & 0 & 0 & 0 & 0 & 0 & 0 & 1 & 0 & 0 & 0 & 1 & 0 & 0 & 0 & 1 & 0 \\
0 & 0 & 0 & 0 & 0 & 0 & 0 & 1 & 0 & 0 & 0 & 1 & 0 & 0 & 0 & 1 & 0 & 0 & 0 & 1 \\
1 & 0 & 0 & 0 & 0 & 1 & 0 & 0 & 0 & 0 & 0 & 1 & 0 & 0 & 0 & 0 & 0 & 1 & 0 & 0 \\
0 & 1 & 0 & 0 & 0 & 0 & 1 & 0 & 0 & 0 & 1 & 0 & 0 & 0 & 0 & 1 & 0 & 0 & 0 & 0 \\
0 & 0 & 1 & 0 & 0 & 0 & 0 & 1 & 0 & 0 & 0 & 0 & 1 & 0 & 0 & 0 & 0 & 0 & 1 & 0 \\
0 & 0 & 0 & 1 & 0 & 0 & 0 & 0 & 1 & 0 & 0 & 0 & 0 & 1 & 0 & 0 & 1 & 0 & 0 & 0 \\
0 & 0 & 0 & 0 & 1 & 0 & 0 & 0 & 0 & 1 & 0 & 0 & 0 & 0 & 1 & 0 & 0 & 0 & 0 & 1
\end{bmatrix}$$

Encoding of bits

- Generator matrix has to be computed
- First step:

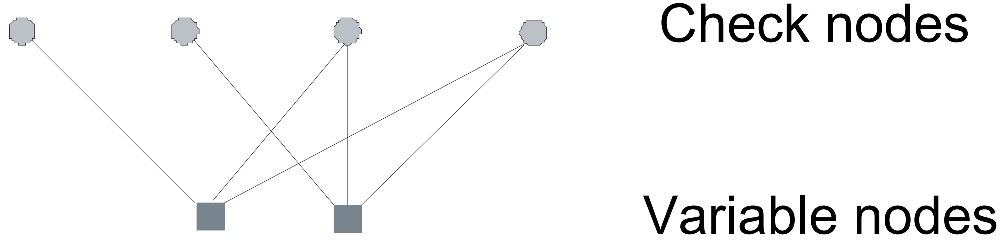
$$\widetilde{\mathbf{H}} = \begin{pmatrix} -\mathbf{P}^T & \mathbf{I} \end{pmatrix}$$

- Second step: generator matrix is

$$\mathbf{G} = \begin{pmatrix} \mathbf{I} & \mathbf{P} \end{pmatrix}$$

Decoding: Tanner graph

- Method for iterative decoding
- Represent code in a Tanner graph (bipartite graph)



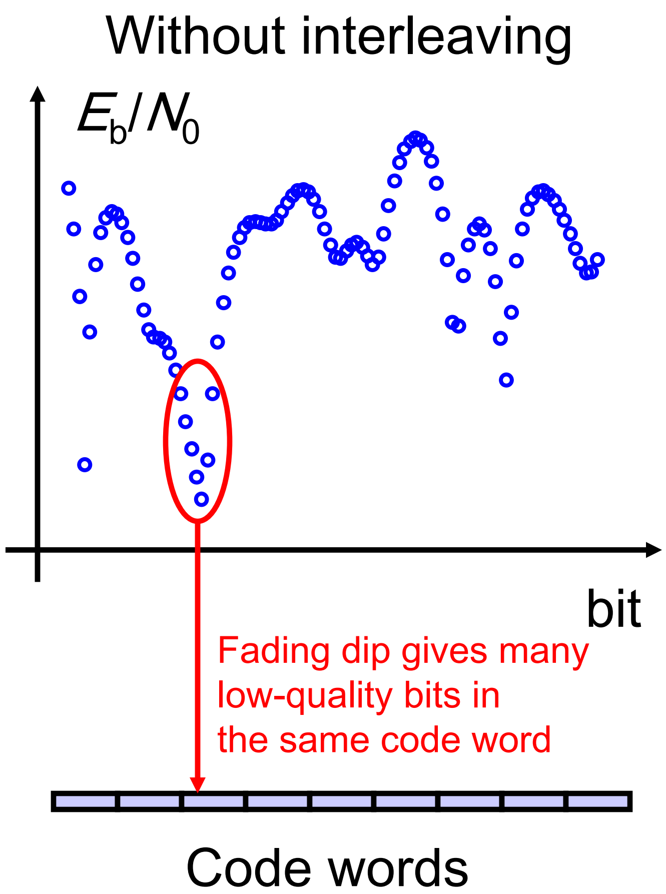
Tanner graph for parity check matrix $H = \begin{bmatrix} 1 & 0 & 1 & 1 \\ 0 & 1 & 1 & 1 \end{bmatrix}$

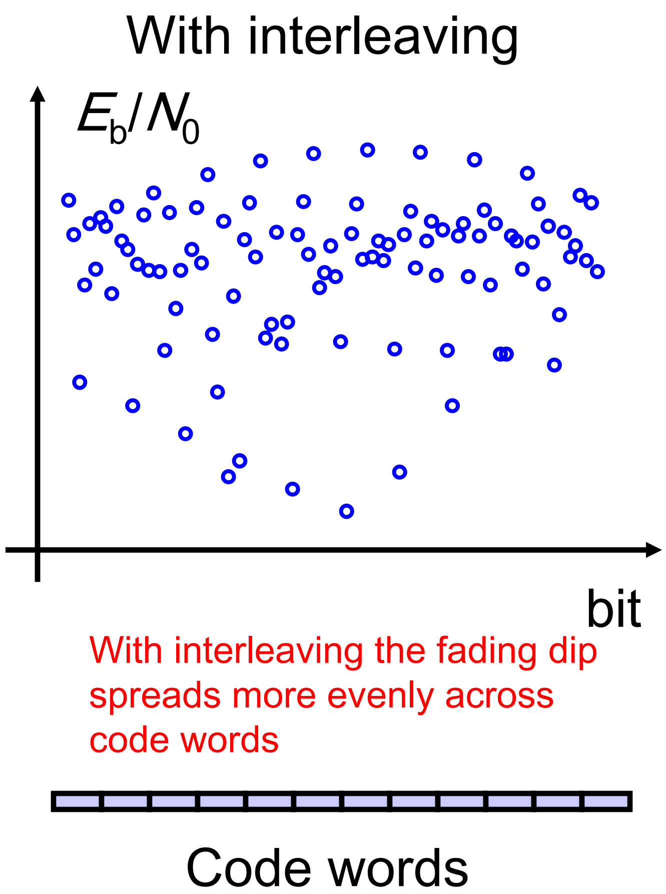
Decoding: step-by-step procedure

1. Variable nodes decide what they think they are, given external evidence only
$$\mu_{i,j}^{(0)} = 0, \text{ for all } i \qquad \lambda_{i,j}^{(0)} = (2/\sigma_n^2) r_j, \text{ for all } j$$
2. Constraint nodes compute what they think variable nodes have to be
$$\mu_{i,j}^{(l)} = 2 \tanh^{-1}\left(\prod_{\mathrm{k} \in \mathrm{A(i)-j}} \tanh\left(\frac{\lambda_{\mathrm{i,k}}^{(\mathrm{l}-1)}}{2} \right) \right)$$
$A(i) - j$ is "all the members of ensemble $A(i)$ with the exception of j"
3. Update opinion of what variable nodes have to be
$$\lambda_{i,j}^{(l)} = (2/\sigma_n^2) r_j + \sum\nolimits_{k \in B(j)-i} \mu_{\mathrm{k,j}}^{(\mathrm{l})}$$
$B(j) - i$ is "all variable nodes that connect to the j –th constraint node, with the exception of i."
4. compute the pseudoposterior probabilities that a bit is 1 or 0
$$L_j = (2/\sigma_n^2) r_j + \sum\nolimits_i \mu_{\mathrm{i,j}}^{(\mathrm{l})},$$
5. If codeword has syndrome 0, stop iteration; otherwise goto 2

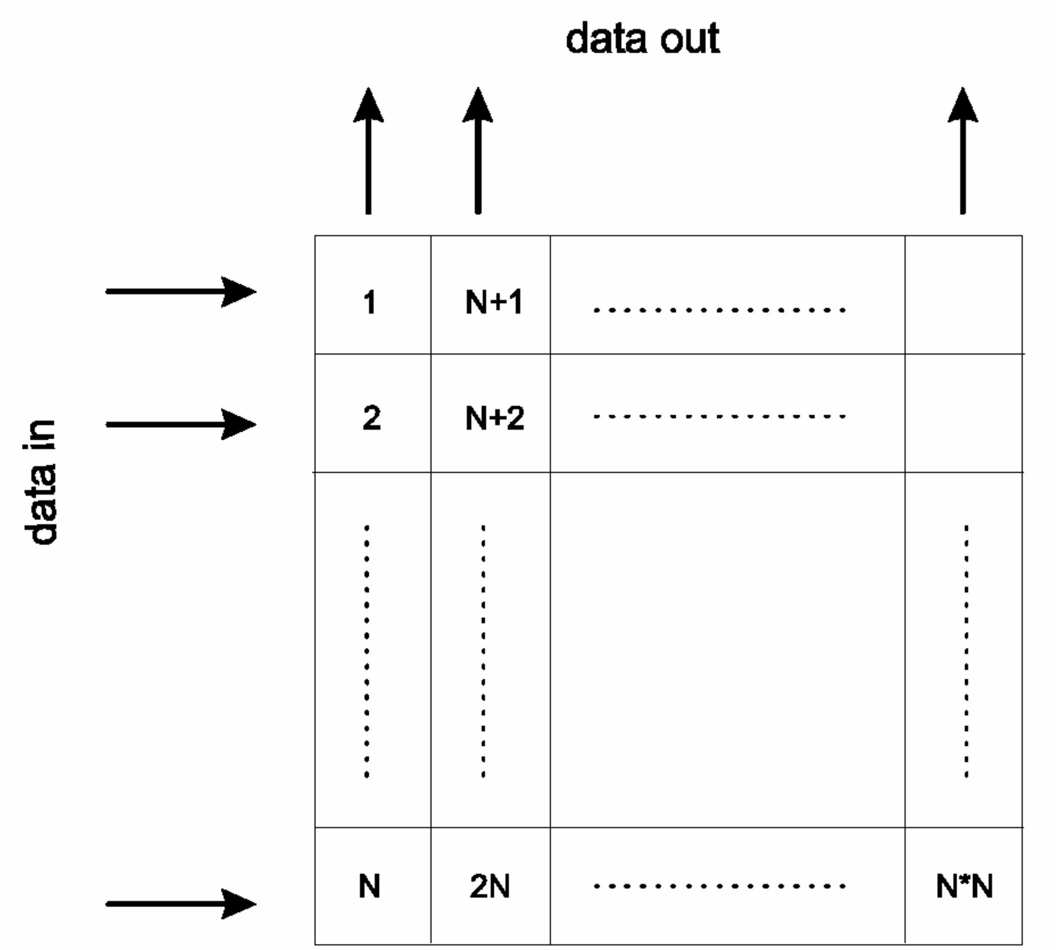
FADING CHANNELS AND INTERLEAVING

Channel coding
Distribution of low-quality bits



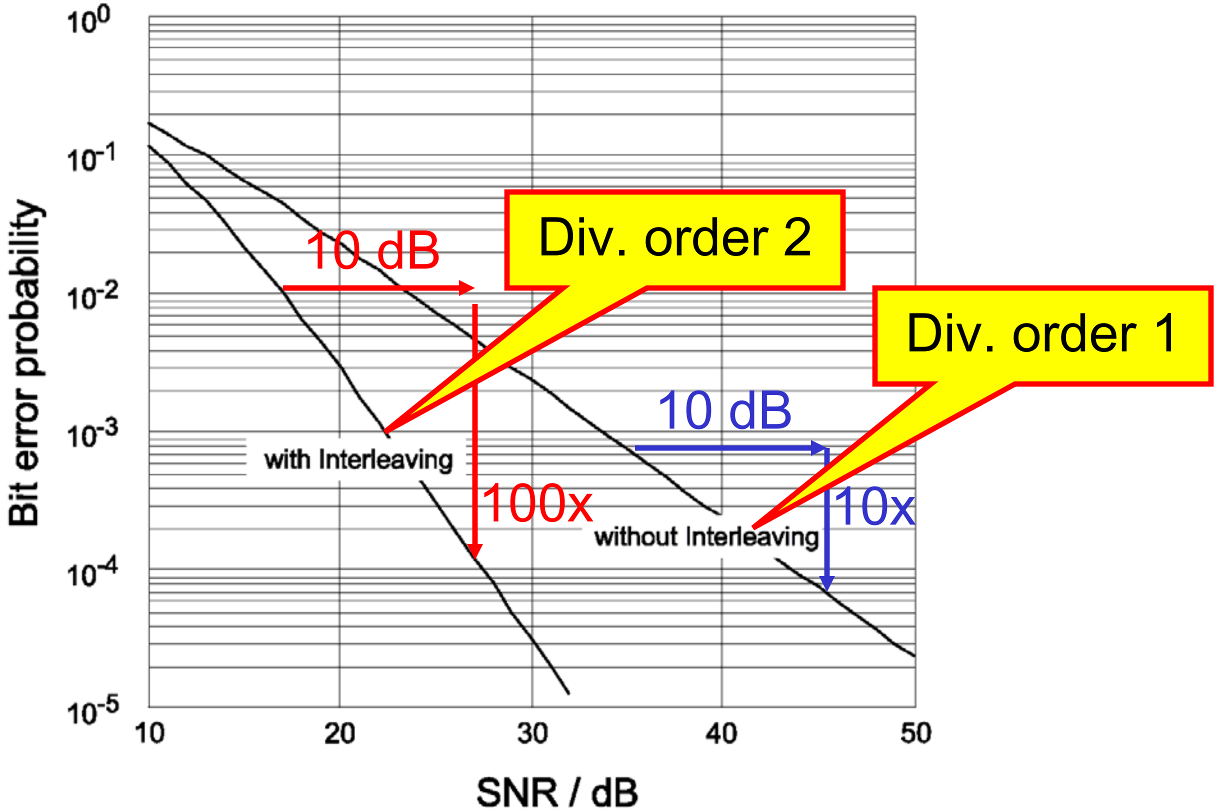


Channel coding
Block interleaver



Channel coding
Interleaving - BER example

BER of a R=1/3 repetition code over a Rayleigh-fading channel, with and without interleaving. Decoding strategy: majority selection.



Summary

- Channel coding is used to improve error performance
- For a fixed requirement, we get a coding gain that translates to a lower received power requirement.
- The two main types of codes are block codes and convolution codes
- Depending on the channel, we use different metrics to measure the distances
- Decoding of convolution codes is efficiently done with the Viterbi algorithm
- In fading channels we need interleaving in order to break up fading dips (but causes delay)

Chapter 16

Equalization

Contents

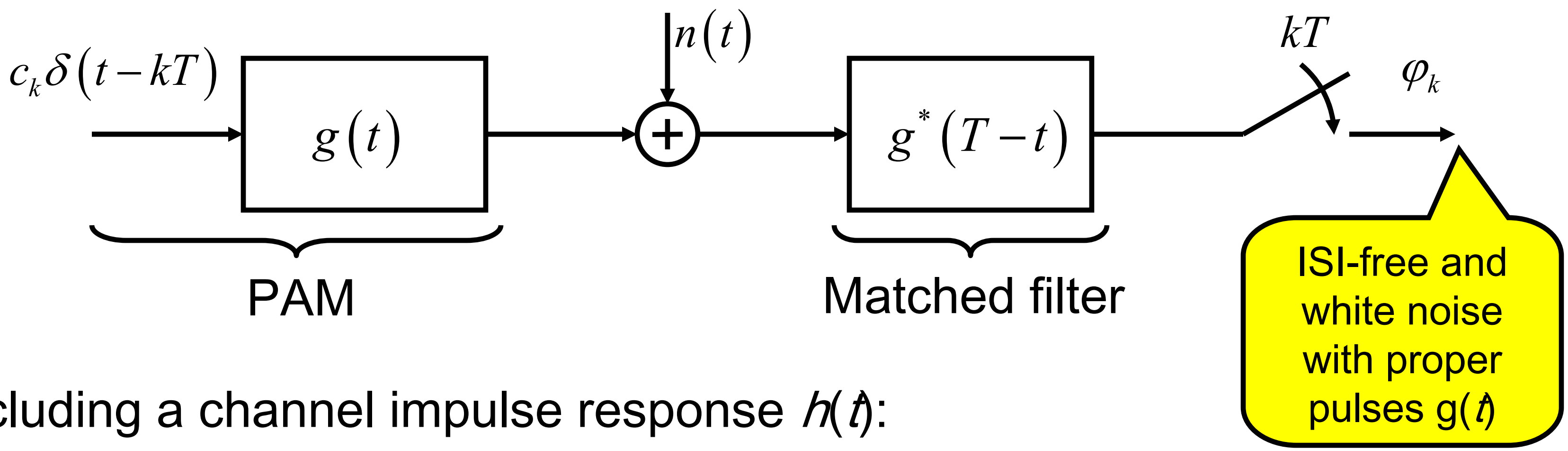
- Inter-symbol interference
- Linear equalizers
- Decision-feedback equalizers
- Maximum-likelihood sequence estimation

INTER-SYMBOL INTERFERENCE

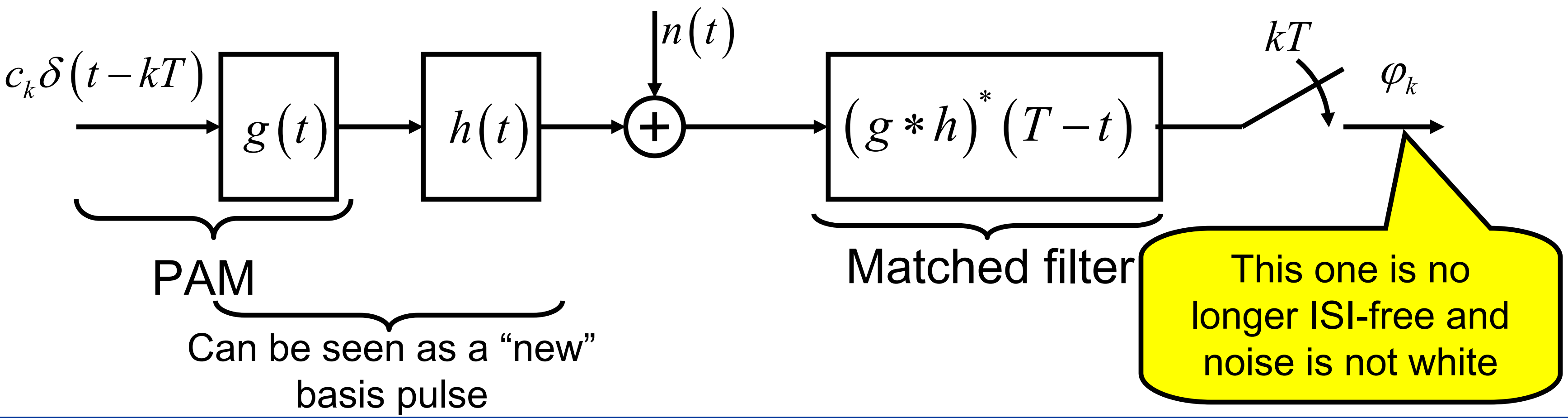
Inter-symbol interference - Background

Modeling of channel impulse response

What we have used so far (PAM and optimal receiver):

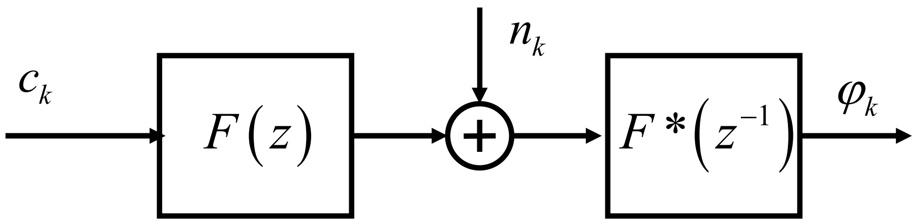


Including a channel impulse response *h*(*t*):



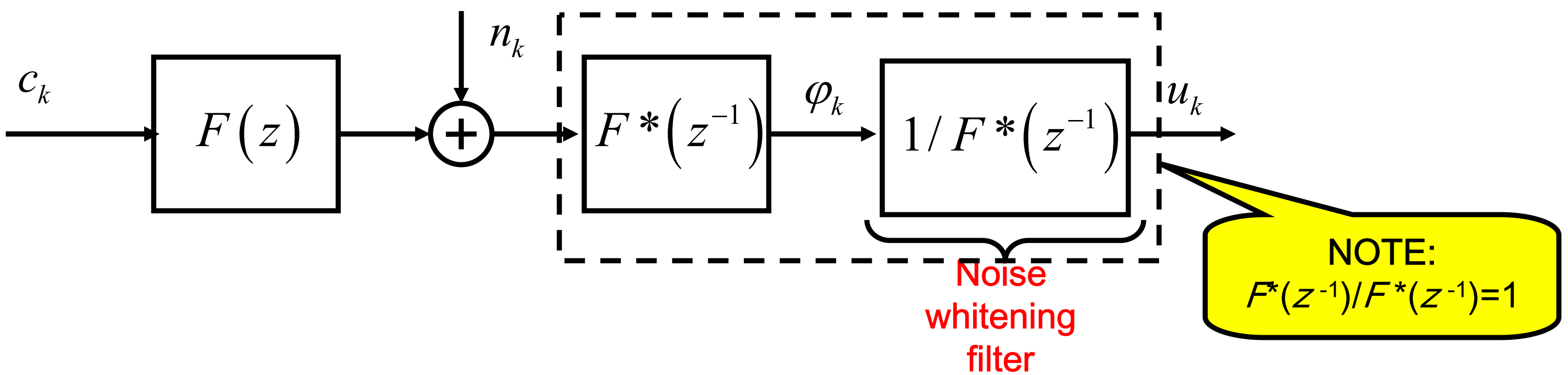
Modeling of channel impulse response

We can create a discrete time equivalent of the "new" system:



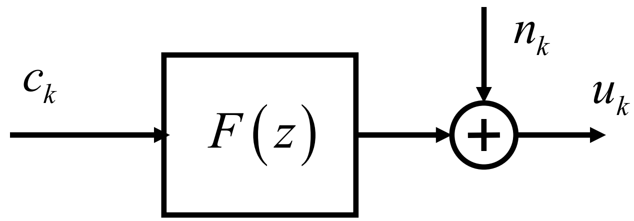
where we can say that $F(z)$ represent the basis pulse and channel, while $F^*(z^{-1})$ represent the matched filter. (This is an abuse of signal theory!)

We can now achieve white noise quite easily, if (the not unique) $F(z)$ is chosen wisely ($F^*(z^{-1})$ has a stable inverse) :



The discrete-time channel model

With the application of a noise-whitening filter, we arrive at a discrete-time model



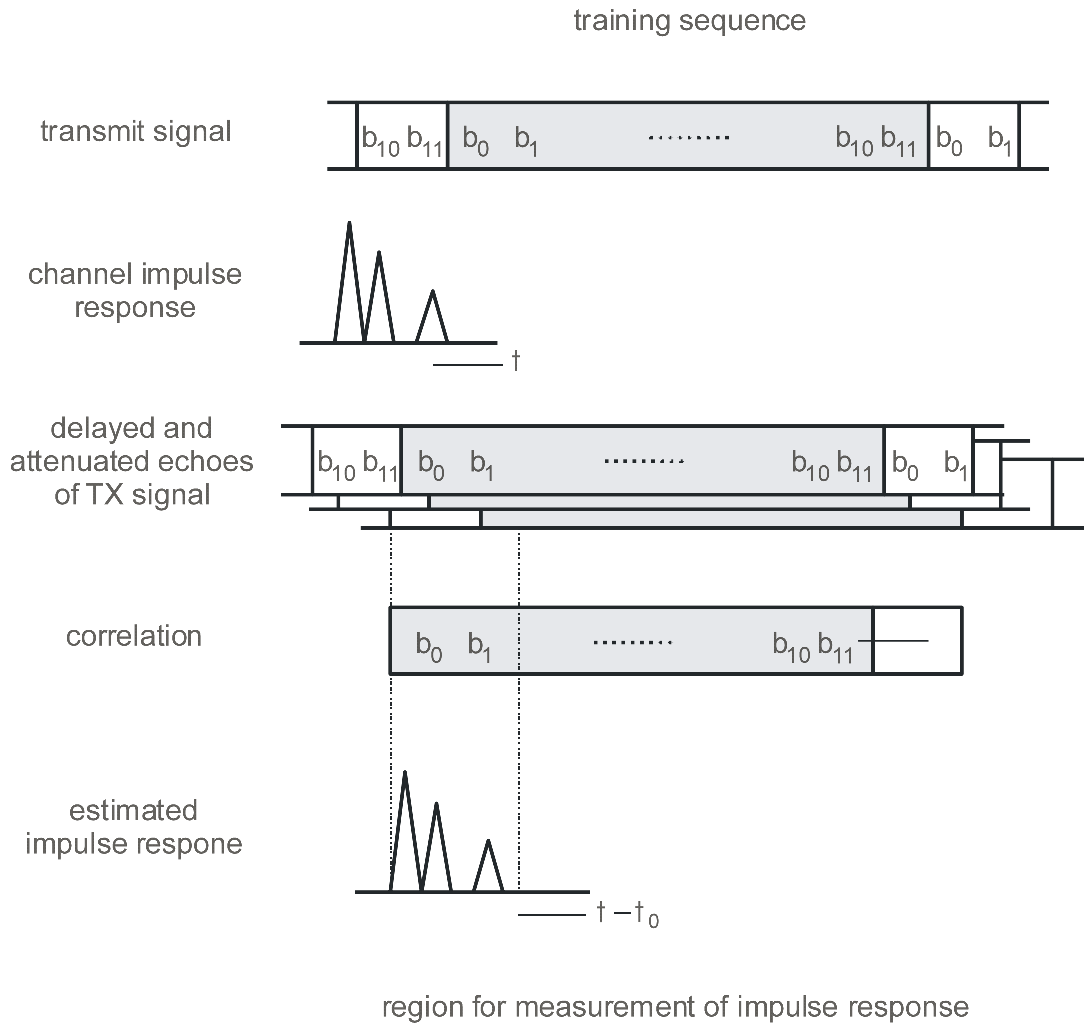
where we have ISI and white additive noise, in the form

$$u_k = \sum_{j=0}^{L} f_j c_{k-j} + n_k$$

This is the model we are going to use when designing equalizers.

The coefficients f_j represent the causal impulse response of the discrete-time equivalent of the channel $F(z)$, with an ISI that extends over L symbols.

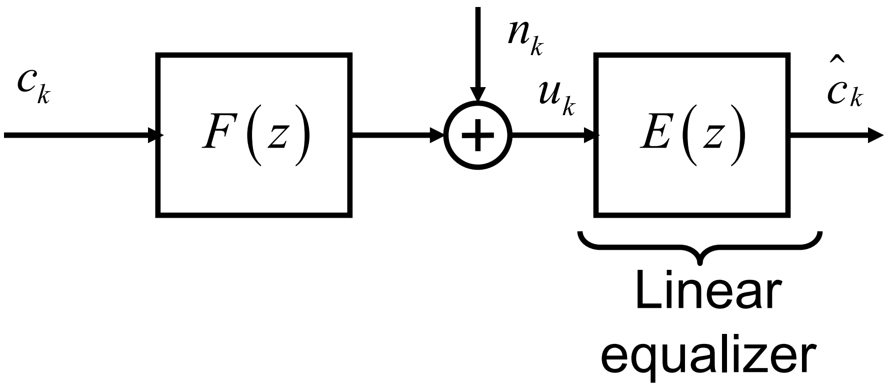
Channel estimation



LINEAR EQUALIZER

Principle

The principle of a linear equalizer is very simple: Apply a filter $E(z)$ at the receiver, mitigating the effect of ISI:

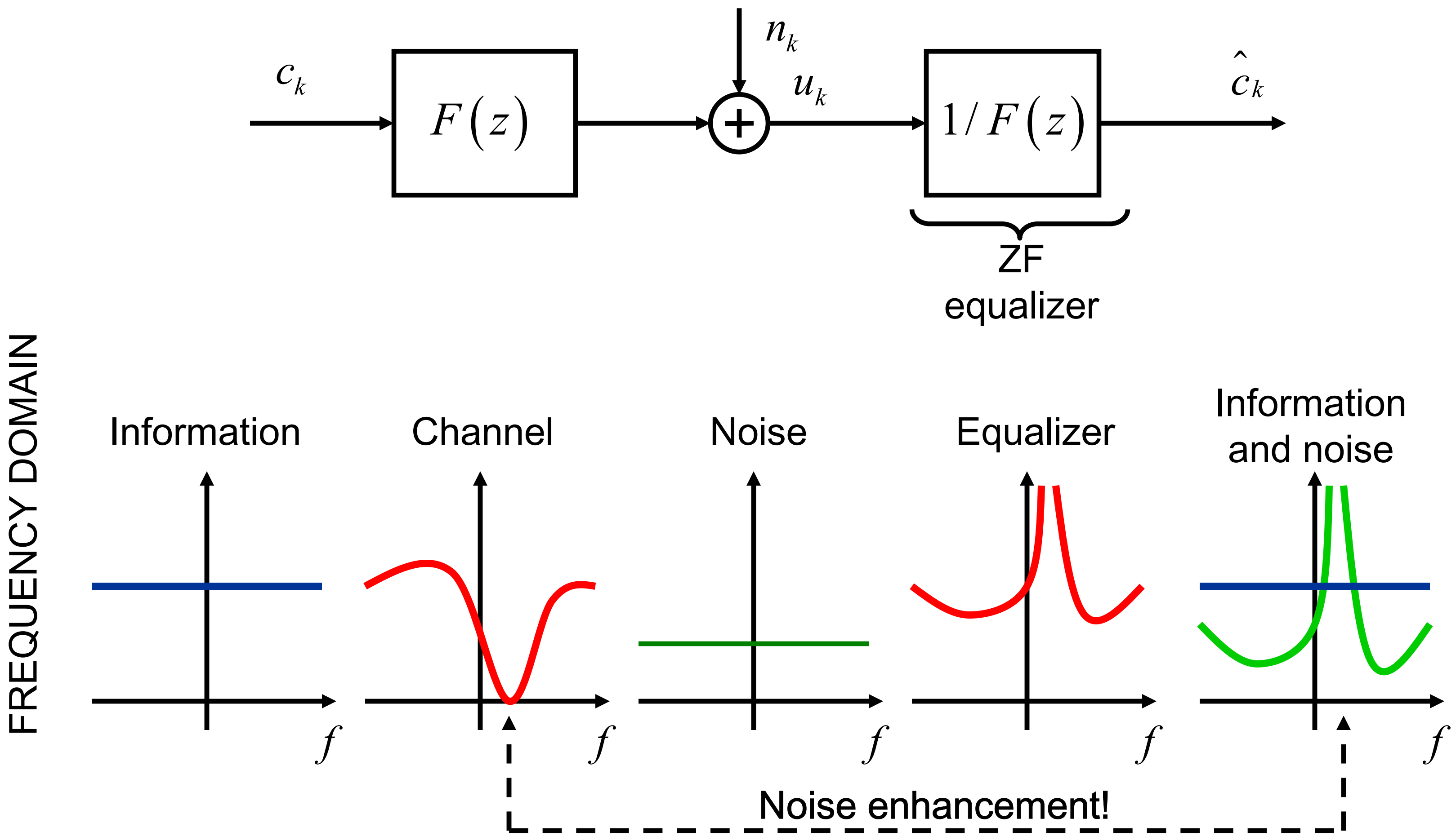


Now we have two different strategies:

1) Design $E(z)$ so that the ISI is totally removed **Zero-forcing**

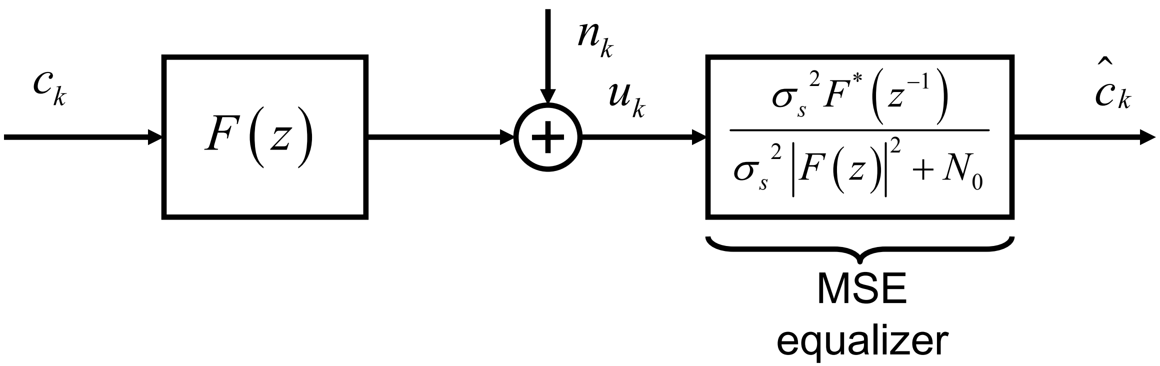
2) Design $E(z)$ so that we minimize the mean squared-error of $\varepsilon_k = c_k - \hat{c}_k$ **MSE**

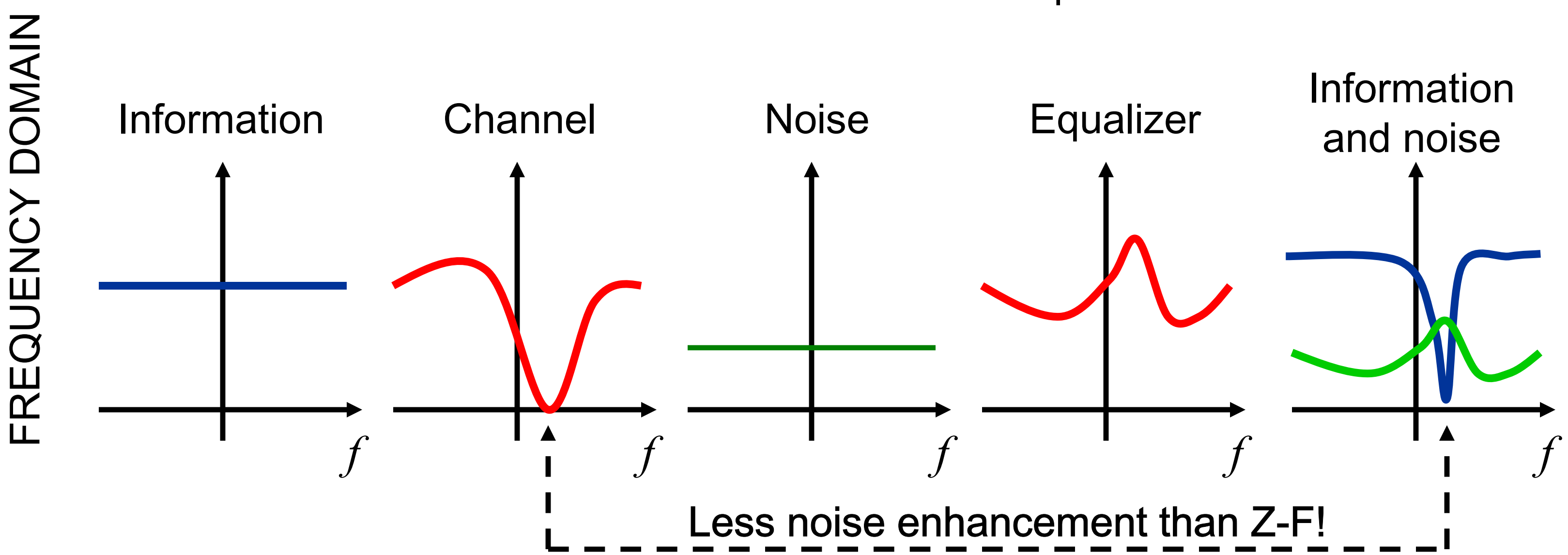
Zero-forcing equalizer



MSE equalizer

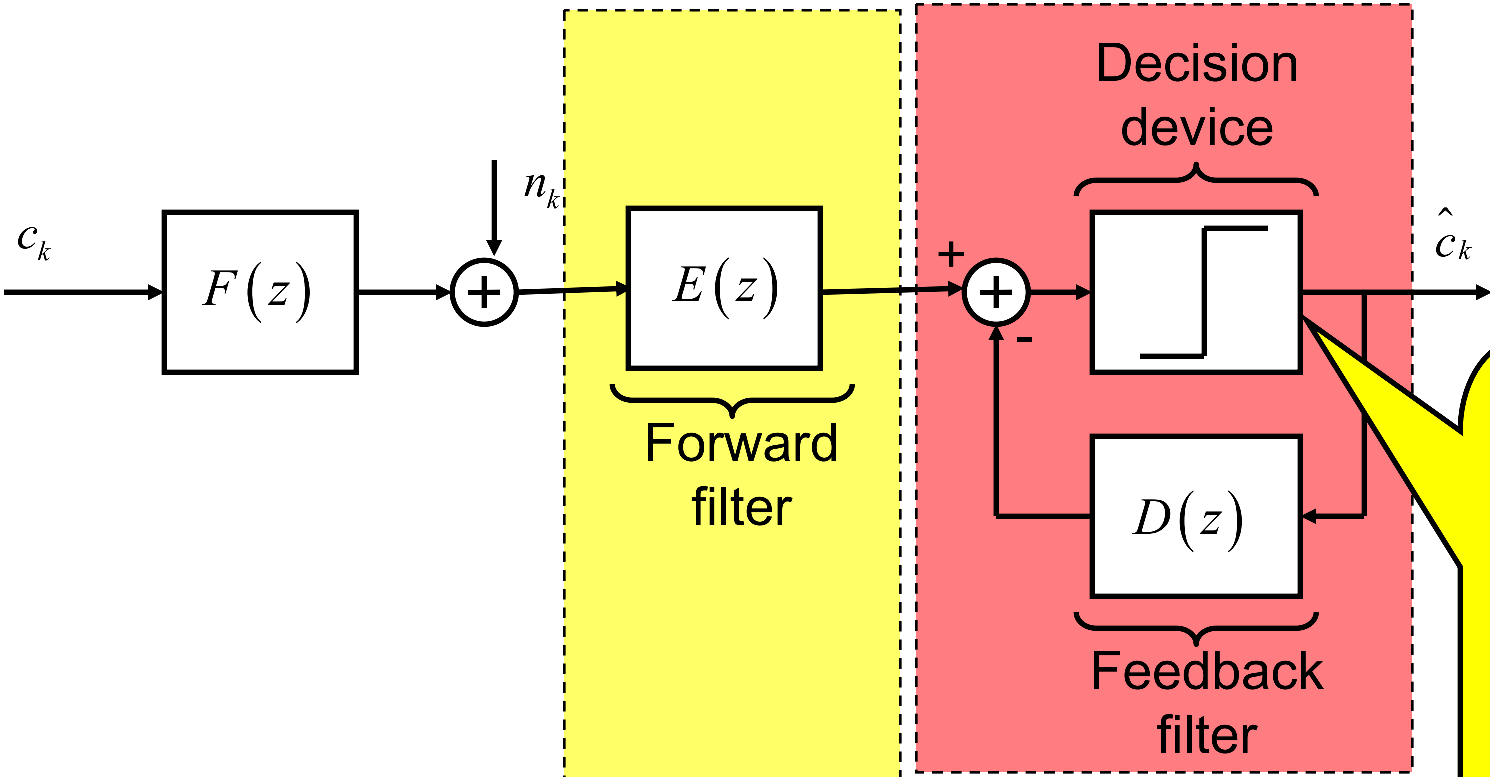
The MSE equalizer is designed to minimize the error variance





DECISION-FEEDBACK EQUALIZER

DFE – Principle



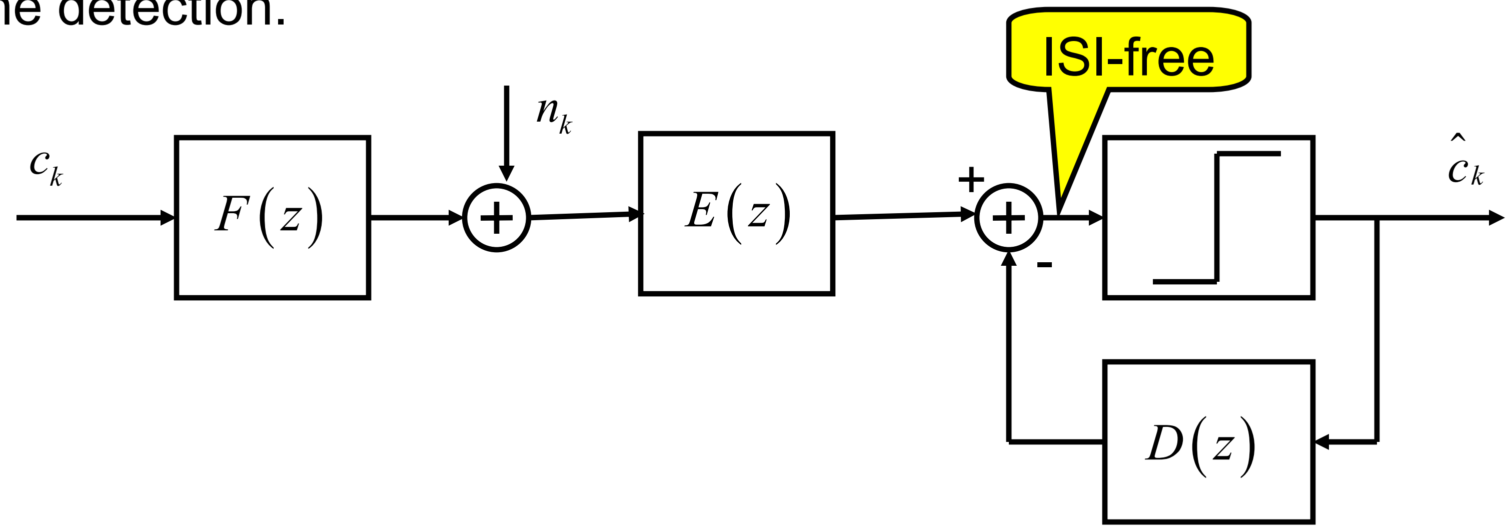
This part shapes the signal to work well with the decision feedback.

This part removes ISI on "**future**" symbols from the currently detected symbol.

If we make a wrong decision here, we may increase the ISI instead of remove it.

Zero-forcing DFE

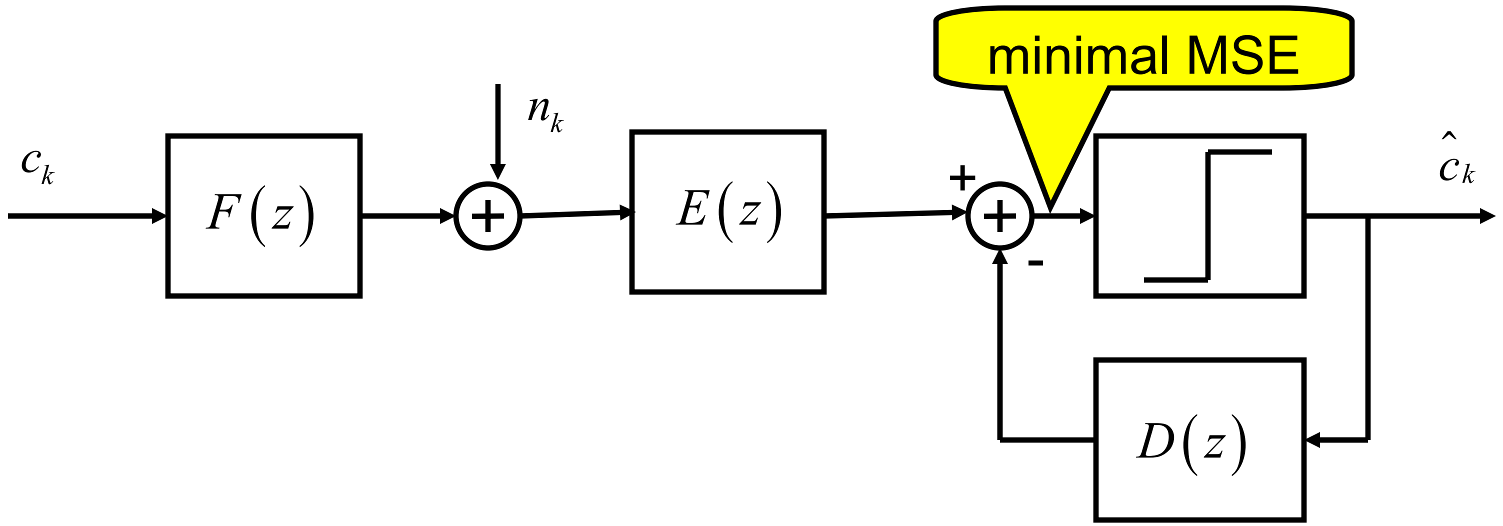
In the design of a ZF-DFE, we want to completely remove all ISI before the detection.



This enforces a relation between the $E(z)$ and $D(z)$, which is (we assume that we make *correct* decisions!)

$$F(z)E(z) - D(z) = 1$$

MSE-DFE



MAXIMUM-LIKELIHOOD SEQUENCE ESTIMATION

Principle

"noise free signal alternative"

$$u_m^{NF} = \sum_{j=0}^{L} f_j c_{m-j}$$

The squared Euclidean distance (optimal for white Gaussian noise) to the received sequence $\{u_m\}$ is

$$d^2\left(\{u_m\},\{u_m^{NF}\}\right) = \sum_m \left|u_m - u_m^{NF}\right|^2 = \sum_m \left|u_m - \sum_{j=0}^{L} f_j c_{m-j}\right|^2$$

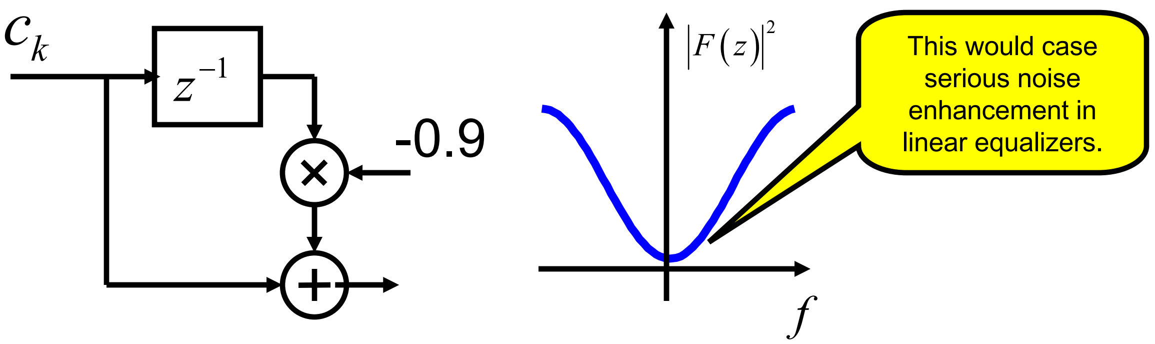
The MLSE decision is then the sequence of symbols $\{c_m\}$ minimizing this distance

$$\left\{\hat{c}_m\right\} = \arg\min_{\{c_m\}} \sum_m \left|u_m - \sum_{j=0}^{L} f_j c_{m-j}\right|^2$$

The Viterbi-equalizer

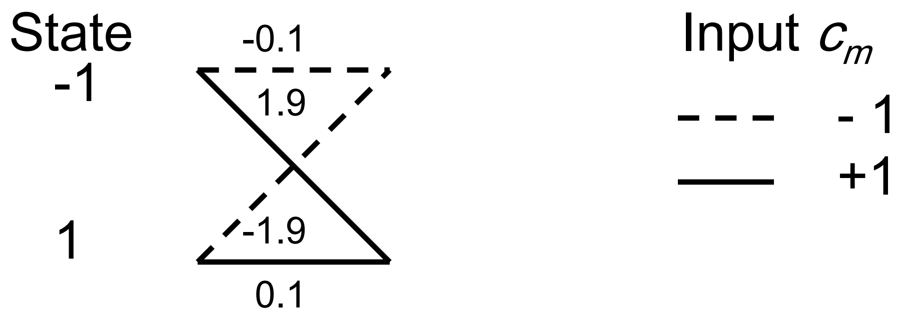
Let's use an example to describe the Viterbi-equalizer.

Discrete-time channel:



Further, assume that our symbol alphabet is –1 and +1 (representing the bits 0 and 1, respectively).

The fundamental trellis stage:

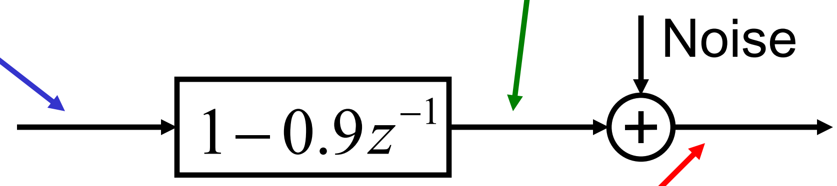


The Viterbi-equalizer (2)

Transmitted:
1 1 -1 1 -1

Noise free sequence:
1.9 0.1 -1.9 1.9 -1.9

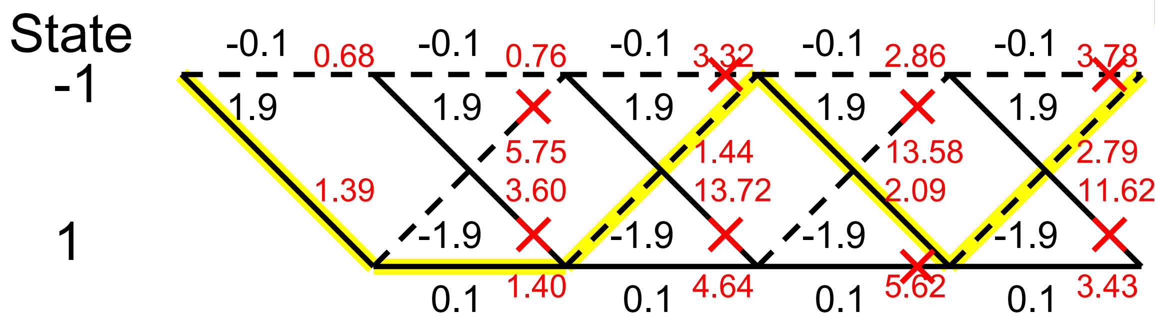
The filter starts in state –1.



Received noisy sequence:
0.72 0.19 -1.70 1.09 -1.06

At this stage, the path ending here has the best metric!

VITERBI DETECTOR



Detected sequence:
1 1 -1 1 -1

Correct!

Summary

- Linear equalizers suffer from noise enhancement.
- Decision-feedback equalizers (DFEs) use decisions on data to remove parts of the ISI, allowing the linear equalizer part to be less ”powerful” and thereby suffer less from noise enhancement.
- Incorrect decisions can cause error-propagation in DFEs, since an incorrect decision may add ISI instead of removing it.
- Maximum-likelihood sequence estimation (MLSE) is optimal in the sense of having the lowest probability of detecting the wrong sequence.
- Brut-force MLSE is prohibitively complex.
- The Viterbi-equalizer (detector) implements the MLSE with considerably lower complexity.

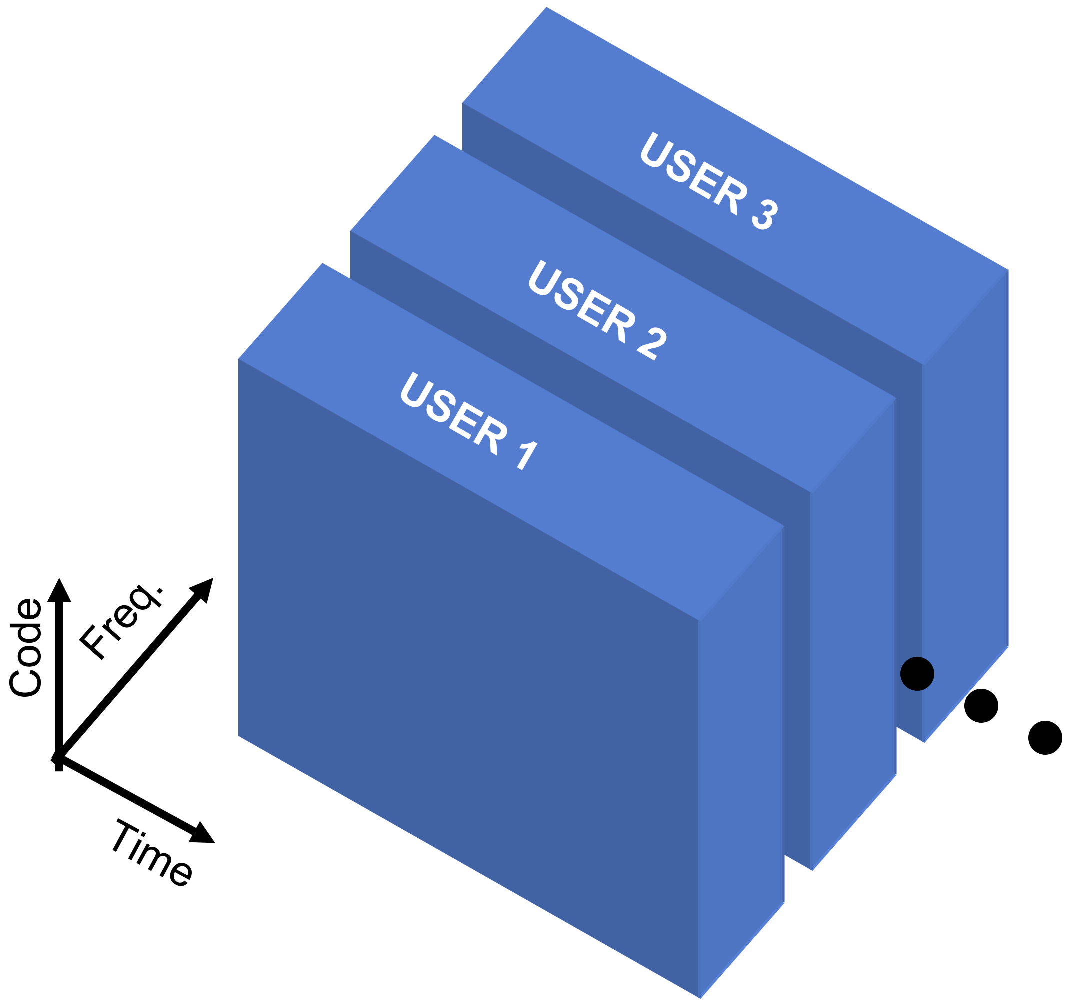
Chapter 17

Multiple access

Contents

- Interference and spectrum efficiency
- Frequency-division multiple access (FDMA)
- Time-division multiple access (TDMA)
- Packet radio

Freq.-division multiple access (FDMA)



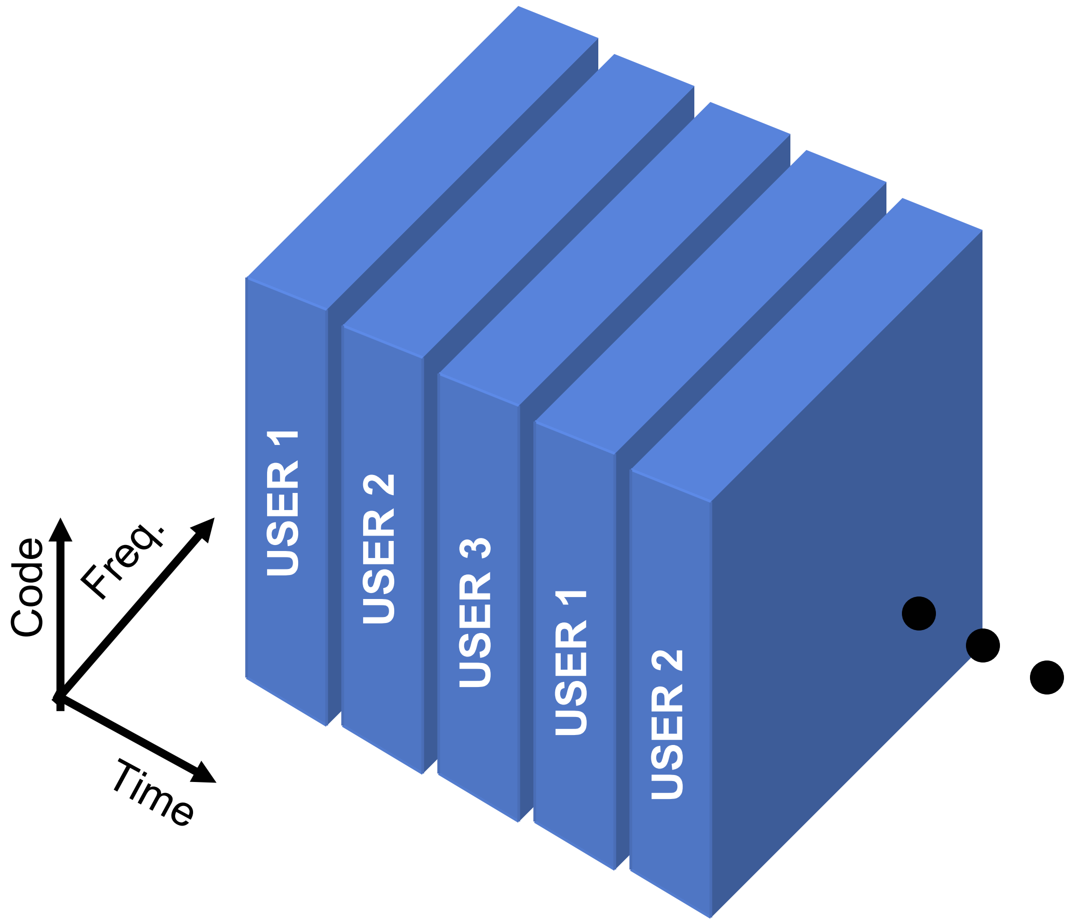
Assume that each channel has a bandwidth of B_{fch} Hz.

If the system has a total bandwidth B_{tot}, then the number of available frequency channels is

$$N_{fch} = \frac{B_{tot}}{B_{fch}}$$

Applying a cellular structure, using frequency reuse, we can have more than N_{fch} simultaneous active users.

Time-division multiple access (TDMA)



TDMA is usually combined with FDMA, where each frequency channel is sub-divided in time to provide more channels.

Users within one cell use TDMA, while different cells share the radio resource in frequency.

One cell can have more than one frequency channel.

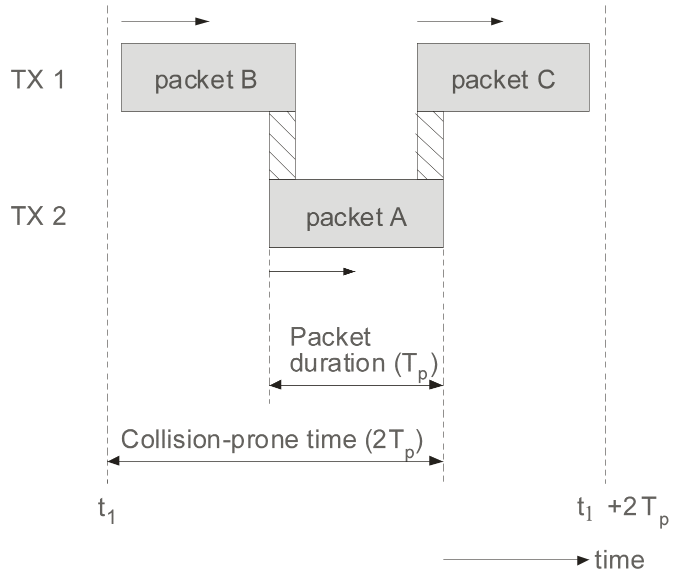
PACKET RADIO

Principle and application

- Data are broken into packets
- Each packet has to fight for its own resources
- Each packet can go from TX to RX via different relays
- Used for, e.g.,
  - Wireless computer networks: internet is packet radio by definition
  - Sensor networks: routing over different relay nodes gives better reliability
  - Voice over IP: allows to have one consistent MA principle for data and voice

ALOHA (1)

- Basic principle: send out data packets whenever TX has them, disregarding all other TXs
- When collision occurs, packet is lost



Copyright: IEEE

ALOHA (2)

- Probability that there are *n* packets within time duration *t*

$$\Pr(n,t) = \frac{(\lambda_{\mathrm{p}} t)^n \exp(-\lambda_{\mathrm{p}} t)}{n!}$$

  where λ_{p} is the packet rate of arrival
- Probability of collision

$$\Pr(0,t) = \exp(-\lambda_p t)$$

- Total throughput: $\lambda_{\mathrm{p}} T_{\mathrm{p}} \exp(-2\lambda_{\mathrm{p}} T_{\mathrm{p}})$
- Maximum throughput: 1/(2e)
- Slotted ALOHA: all packets start at certain discrete times

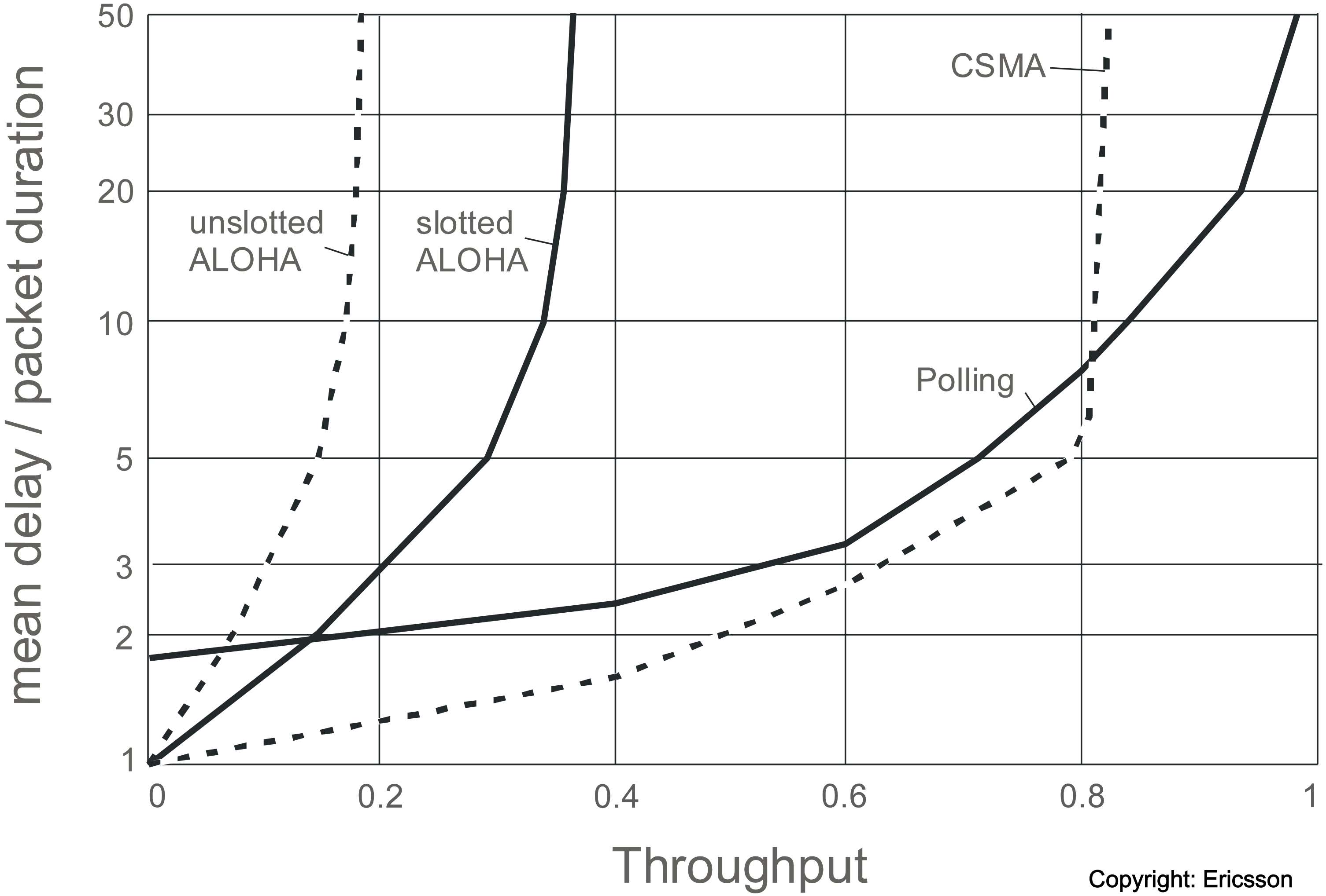
Carrier sense multiple access

- Principle: first determine whether somebody else transmits, send only when channel is free

- Why are there still collisions?
  - Delays are unavoidable: system delay and propagation delay
  - Collision, when there is a signal on the air, but device cannot sense it, because (due to delay) it has not reached it yet

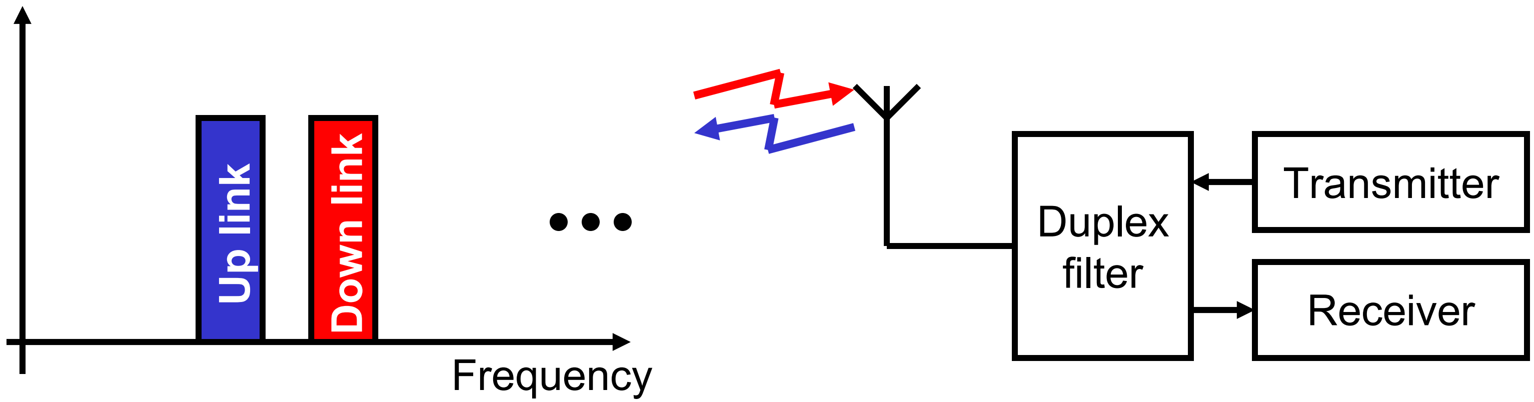
- What does system do when it senses that channel is busy?
  - WAIT
  - Different approaches to how long it should wait

Performance comparison



DUPLEXING

DUPLEX
Frequency-division Duplex (FDD)

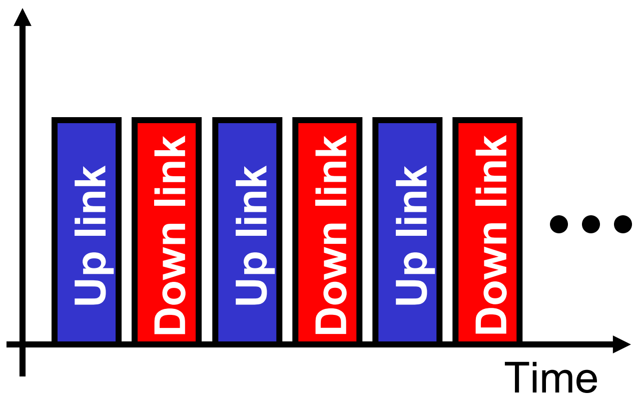


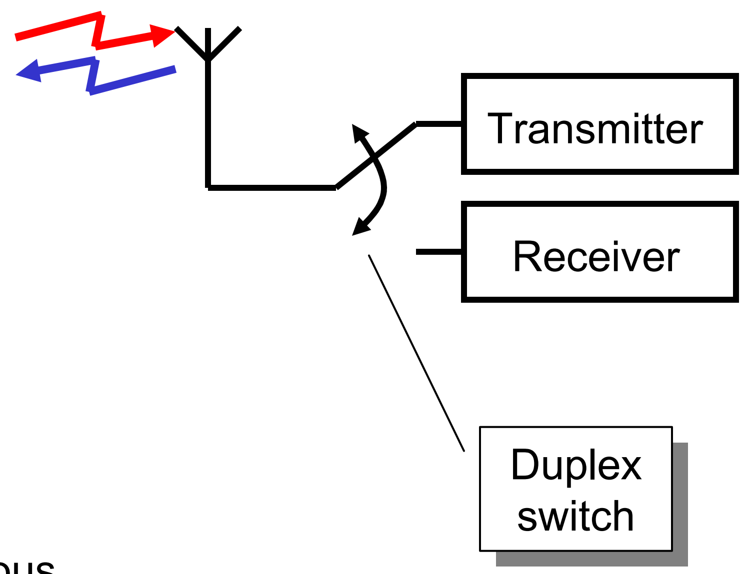
FDD gives a more complex solution (the duplex filter).

Can be used for continuous transmission.

Examples: Nodic Mobile Telephony (NMT), Global System for Mobile communications (GSM), Wideband CDMA (WCDMA)

DUPLEX
Time-division duplex (TDD)





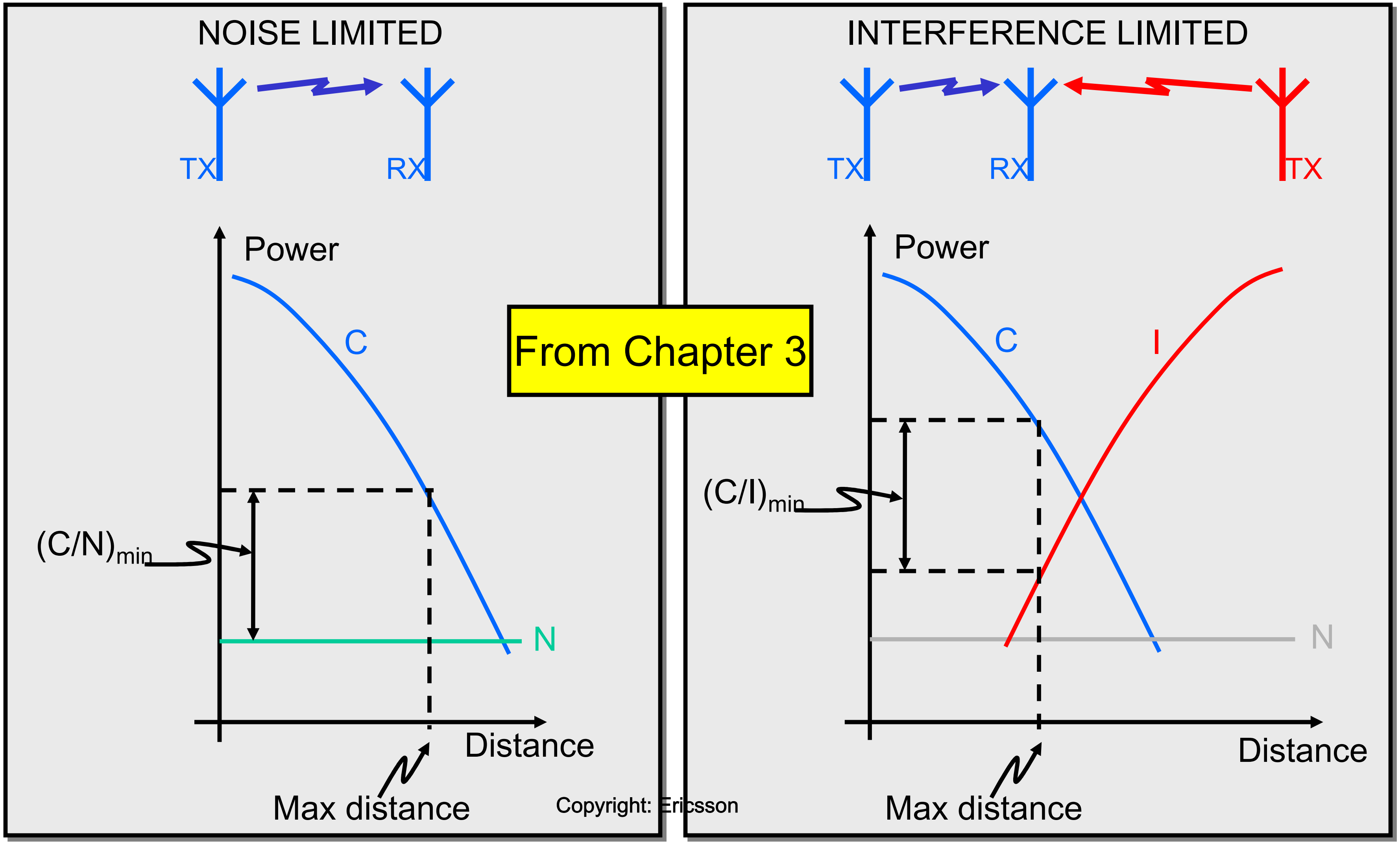
TDD gives a low complexity solution (the duplex switch).

Cannot be used for continuous transmission.

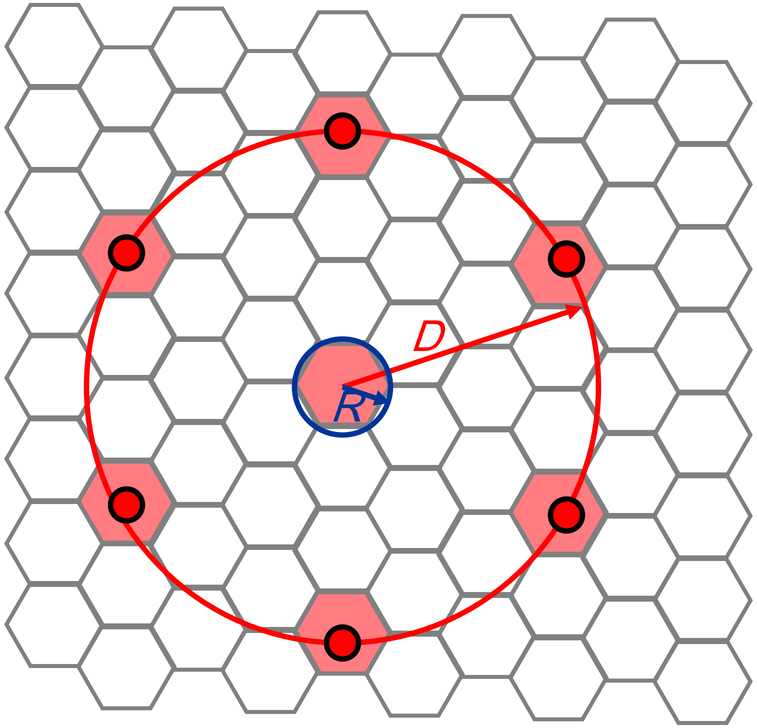
Examples: Global System for Mobile communications (GSM), Wideband CDMA (WCDMA)

INTERFERENCE AND SPECTRUM EFFICIENCY

Interference and spectrum efficiency
Noise and interference limited links



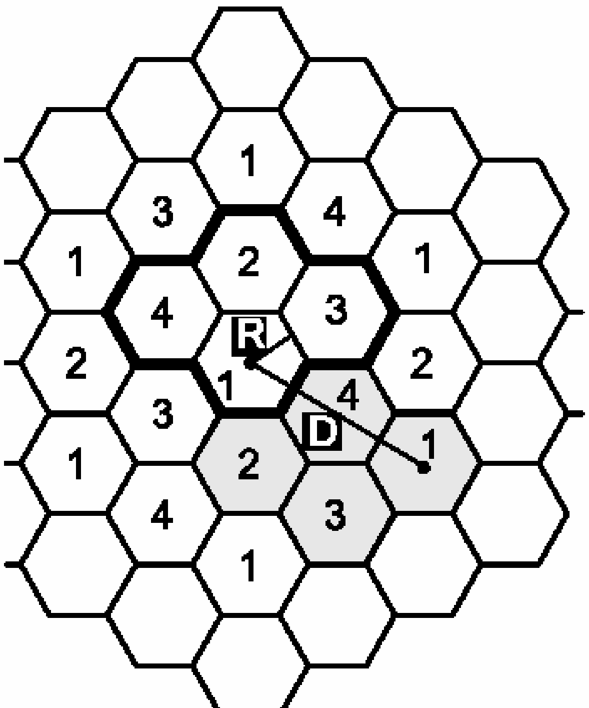
Interference and spectrum efficiency
Cellular systems



$$N_{cluster} = \frac{(D/R)^2}{3}$$

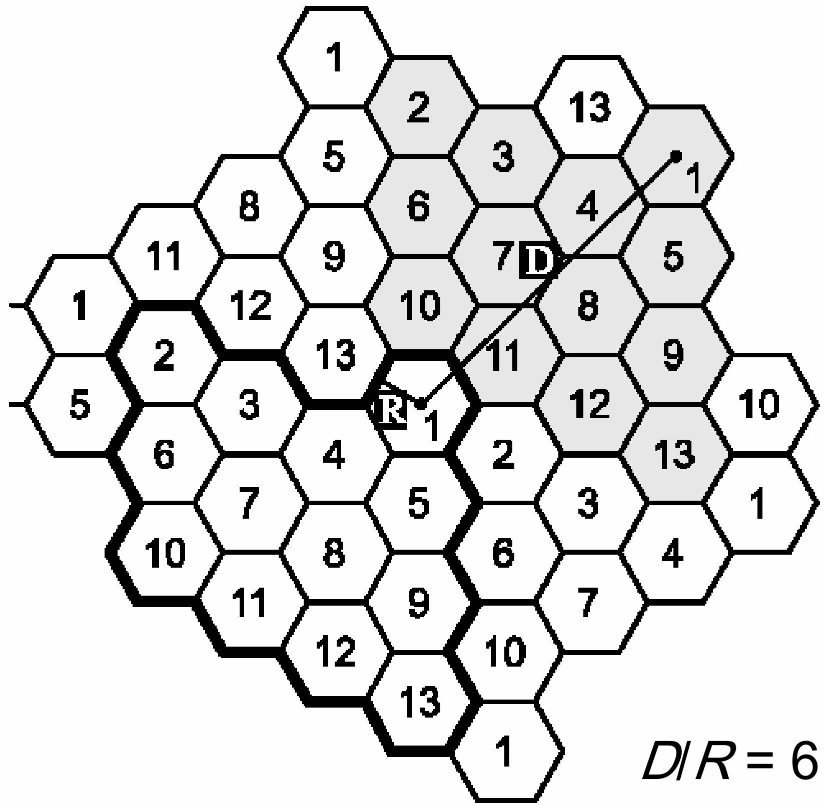
Interference and spectrum efficiency Cellular systems, cont.

Cluster size: $N_{cluster} = 4$



$D/R = 3.5$

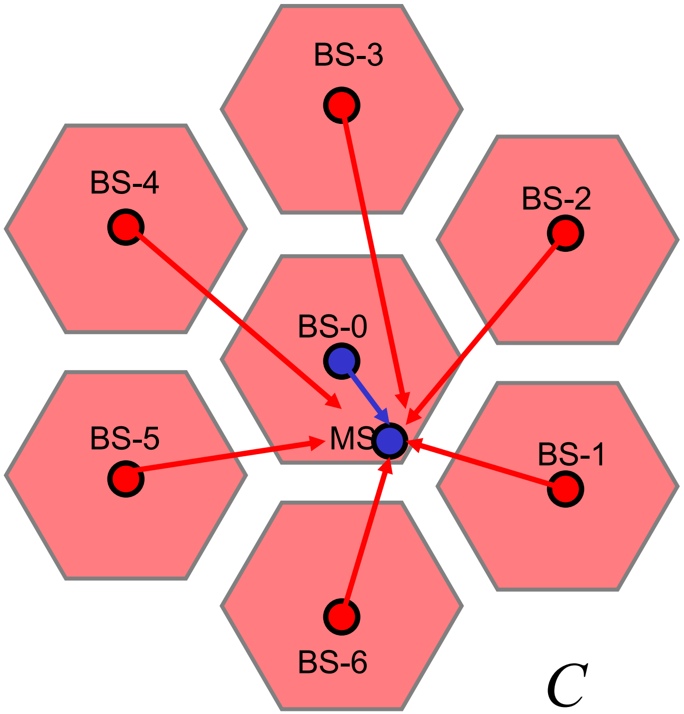
Cluster size: $N_{cluster} = 13$



$D/R = 6.2$

Interference and spectrum efficiency
Cellular systems, cont.

Where do we get the necessary D/R?



Received useful power is

$$C \propto P_{TX} d_0^{-\eta}$$

With 6 co-channel cells interfering, at distances d_1, d_2, ... d_6, from the MS, the received interference is

$$I \propto \sum_{i=1}^{6} P_{TX} d_i^{-\eta}$$

Knowing that $d_0 < R$ and $d_1,\ldots,d_6 > D - R$, we get

$$\frac{C}{I} = \frac{P_{TX} d_0^{-\eta}}{\sum_{i=1}^{6} P_{TX} d_i^{-\eta}} > \frac{P_{TX} R^{-\eta}}{\sum_{i=1}^{6} P_{TX} \left(D-R\right)^{-\eta}} = \frac{1}{6}\left(\frac{R}{D-R}\right)^{-\eta}$$

Interference and spectrum efficiency
Cellular systems, cont.

Assume now that we have a transmission system, which requires $(C/I)_{\min}$ to operate properly. Further, due to fading and requirements on outage we need a fading margin M.

Using our bound

$$\frac{C}{I} > \frac{1}{6}\left(\frac{R}{D-R}\right)^{-\eta}$$

we can solve for a "safe" D/R by requiring

$$\frac{1}{6}\left(\frac{R}{D-R}\right)^{-\eta} \geq M\left(\frac{C}{I}\right)_{\min}$$

We get

$$\frac{D}{R} \geq \left(6M\left(\frac{C}{I}\right)_{\min}\right)^{1/\eta} + 1$$

Interference and spectrum efficiency
Cellular systems, cont.

$N_{cluster}$	3	4	7	9	12	13	16	19	21	25	27
$D/R=\sqrt{3N_{cluster}}$	3	3.5	4.6	5.2	6	6.2	6.9	7.5	7.9	8.7	9

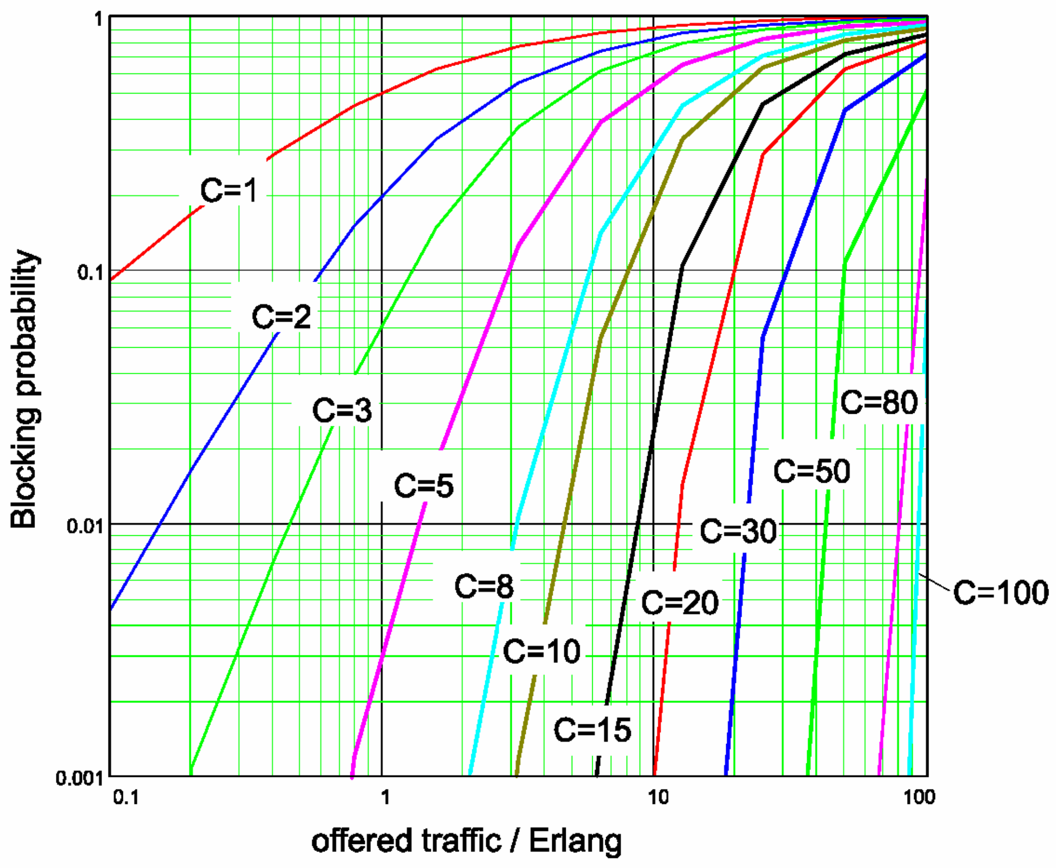
TDMA systems, like GSM

Analog systems, like NMT

Interference and spectrum efficiency Cellular systems, cont.

Erlang-B

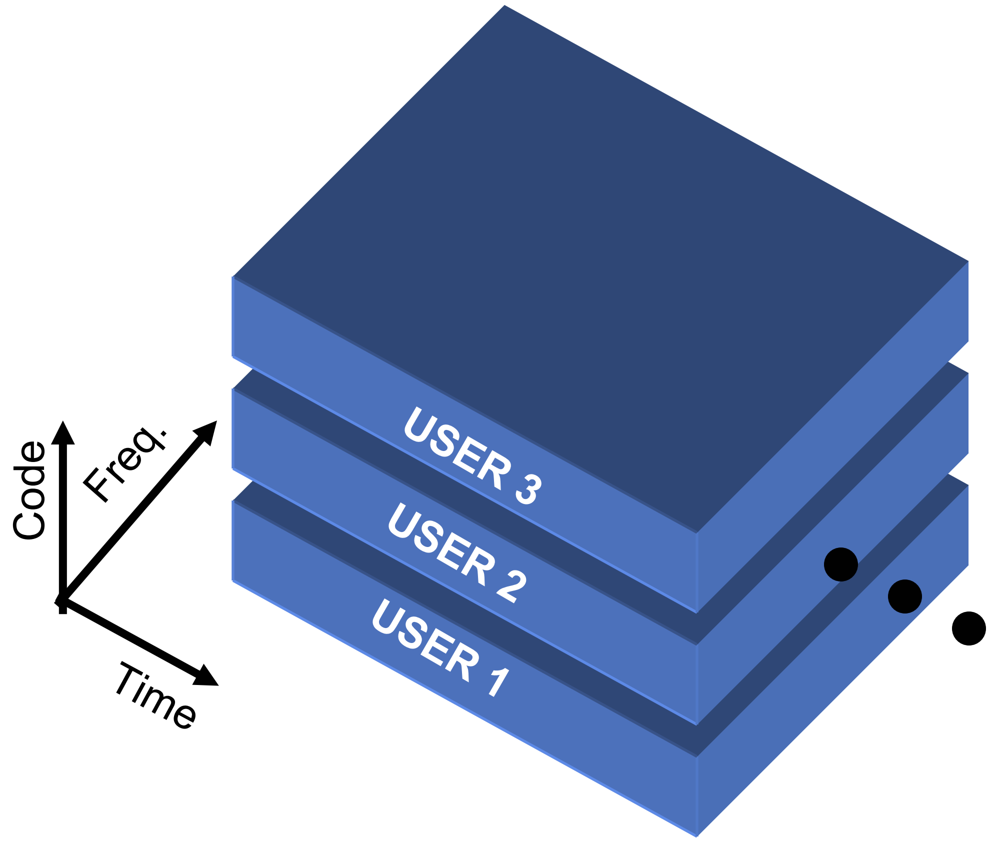
Relation between blocking probability and offered traffic for different number of available speech channels in a cell.



Spread Spectrum

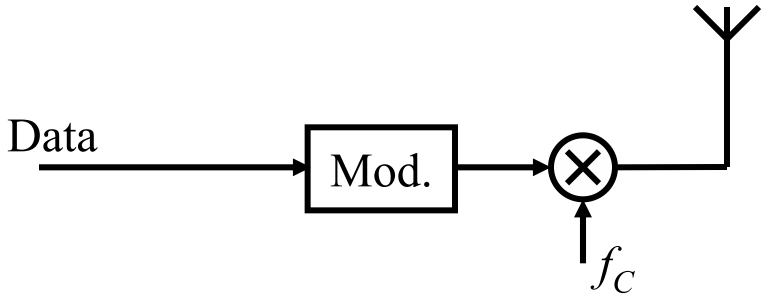
PRINCIPLES OF SPREAD SPECTRUM

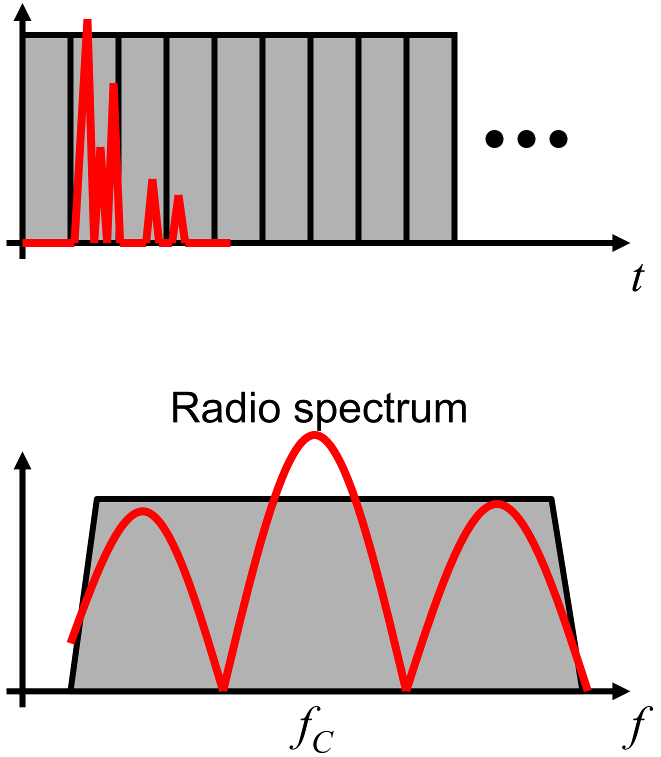
Spread spectrum for multiple access



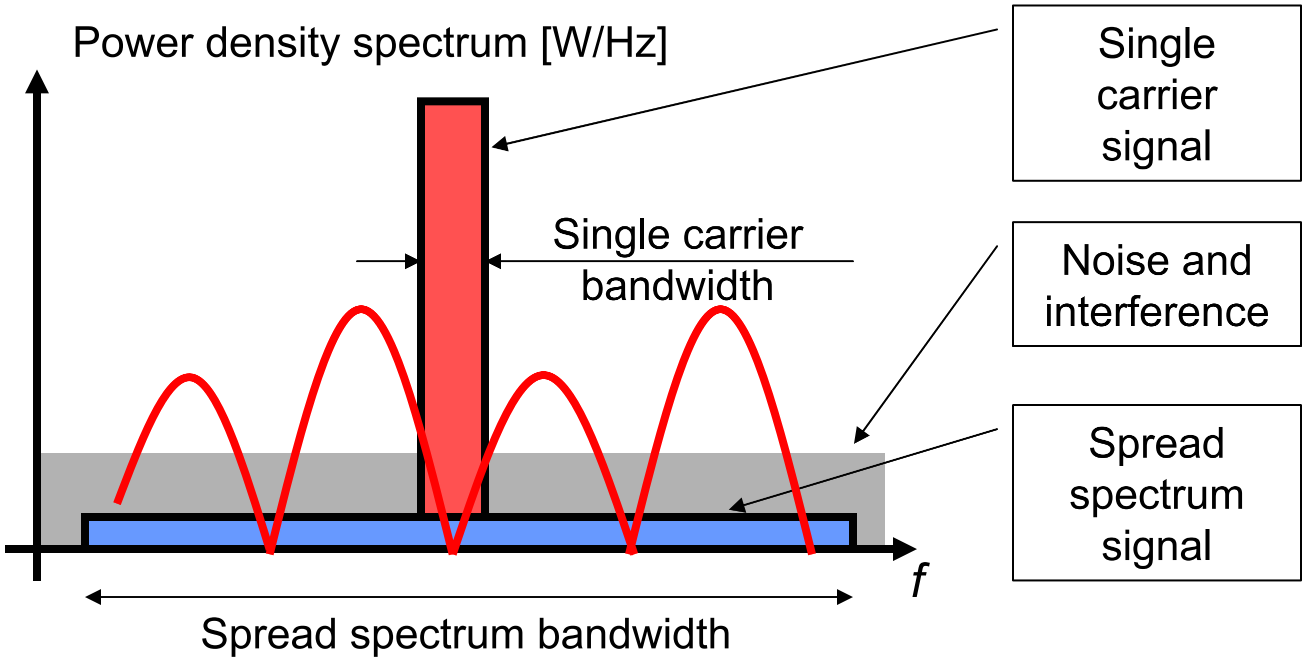
Single Carrier

The traditional way

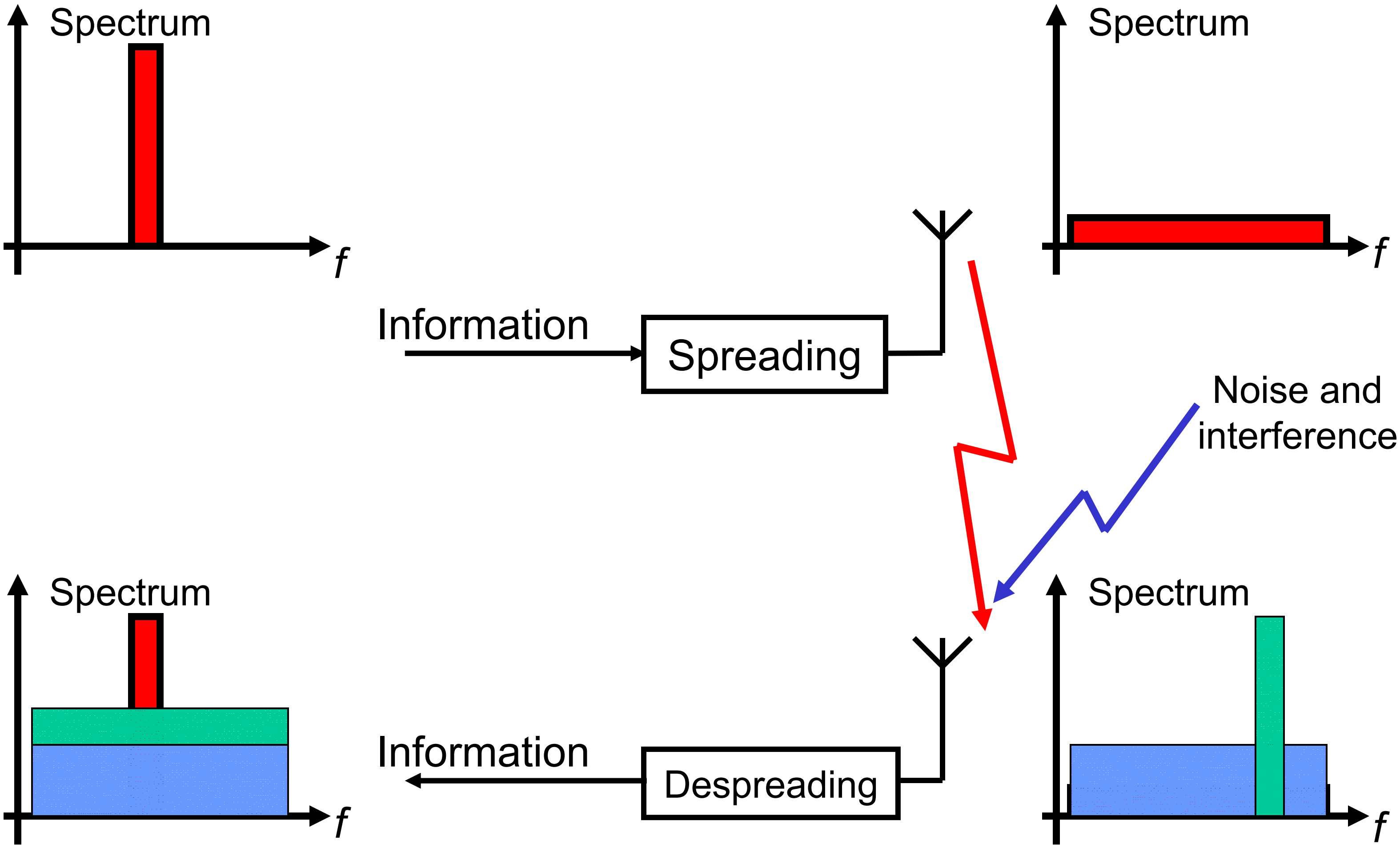




Spread Spectrum Techniques

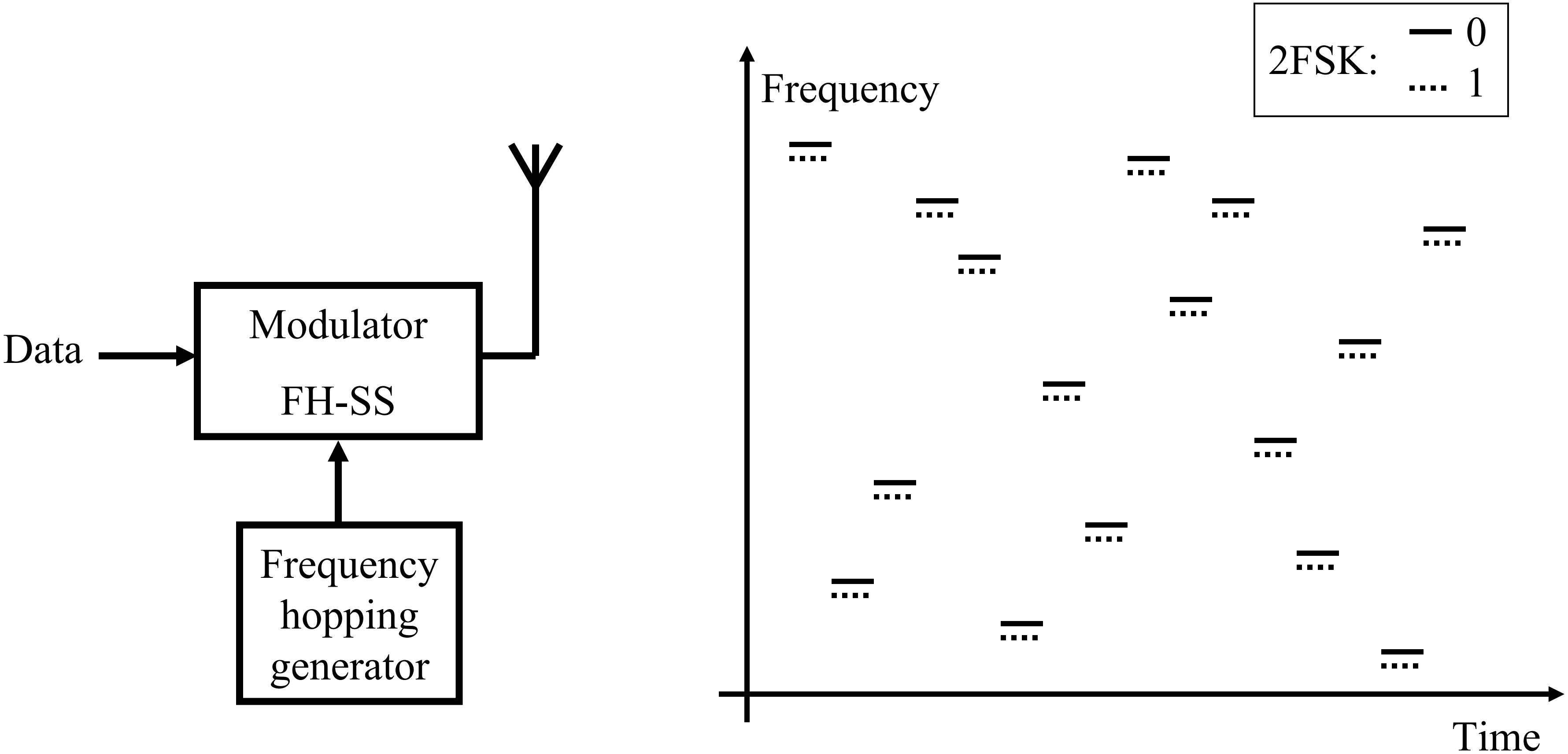


Spread Spectrum Techniques

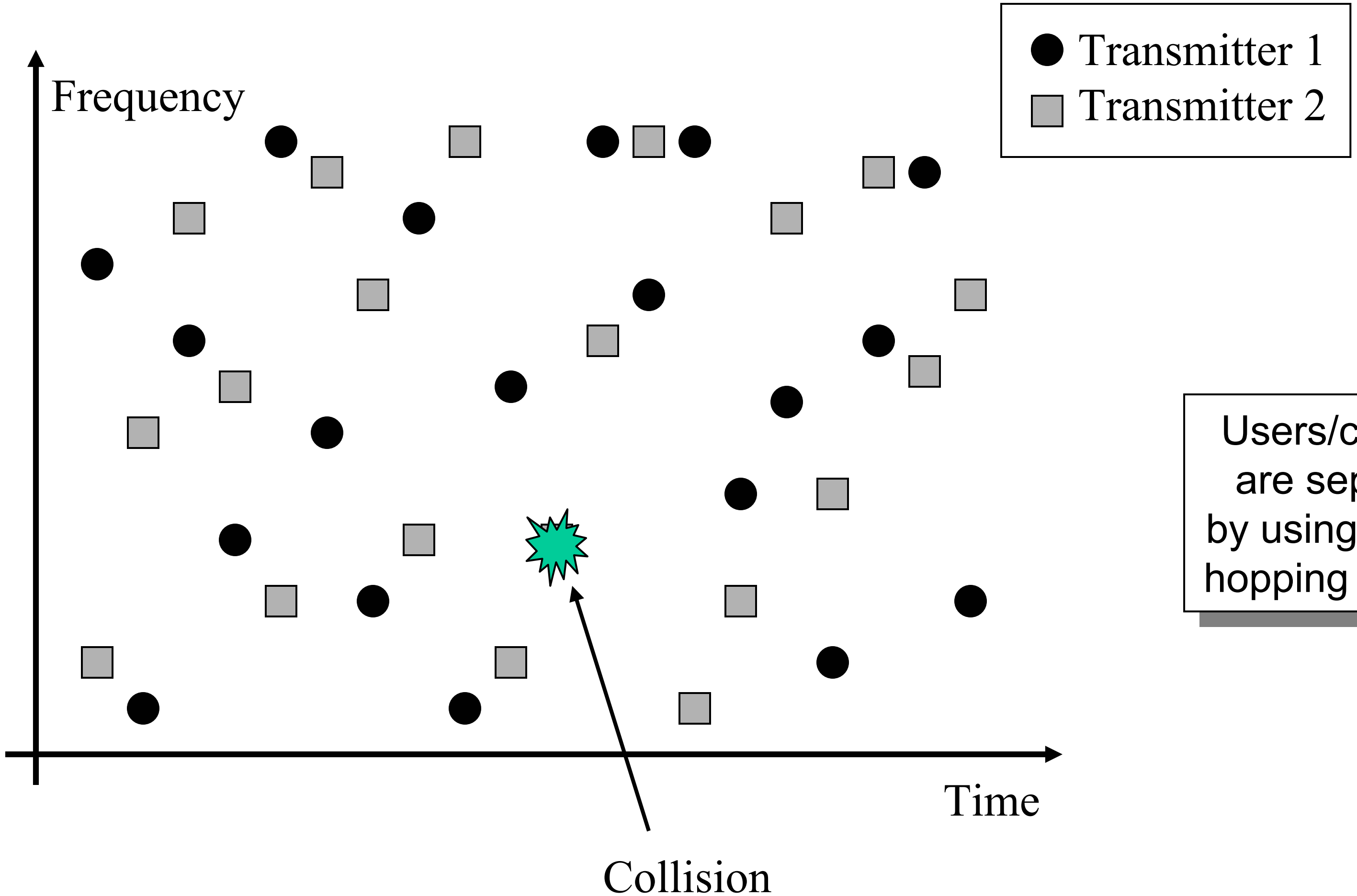


FREQUENCY HOPPING

Frequency-Hopping Spread Spectrum FHSS



Frequency-Hopping Spread Spectrum FHSS

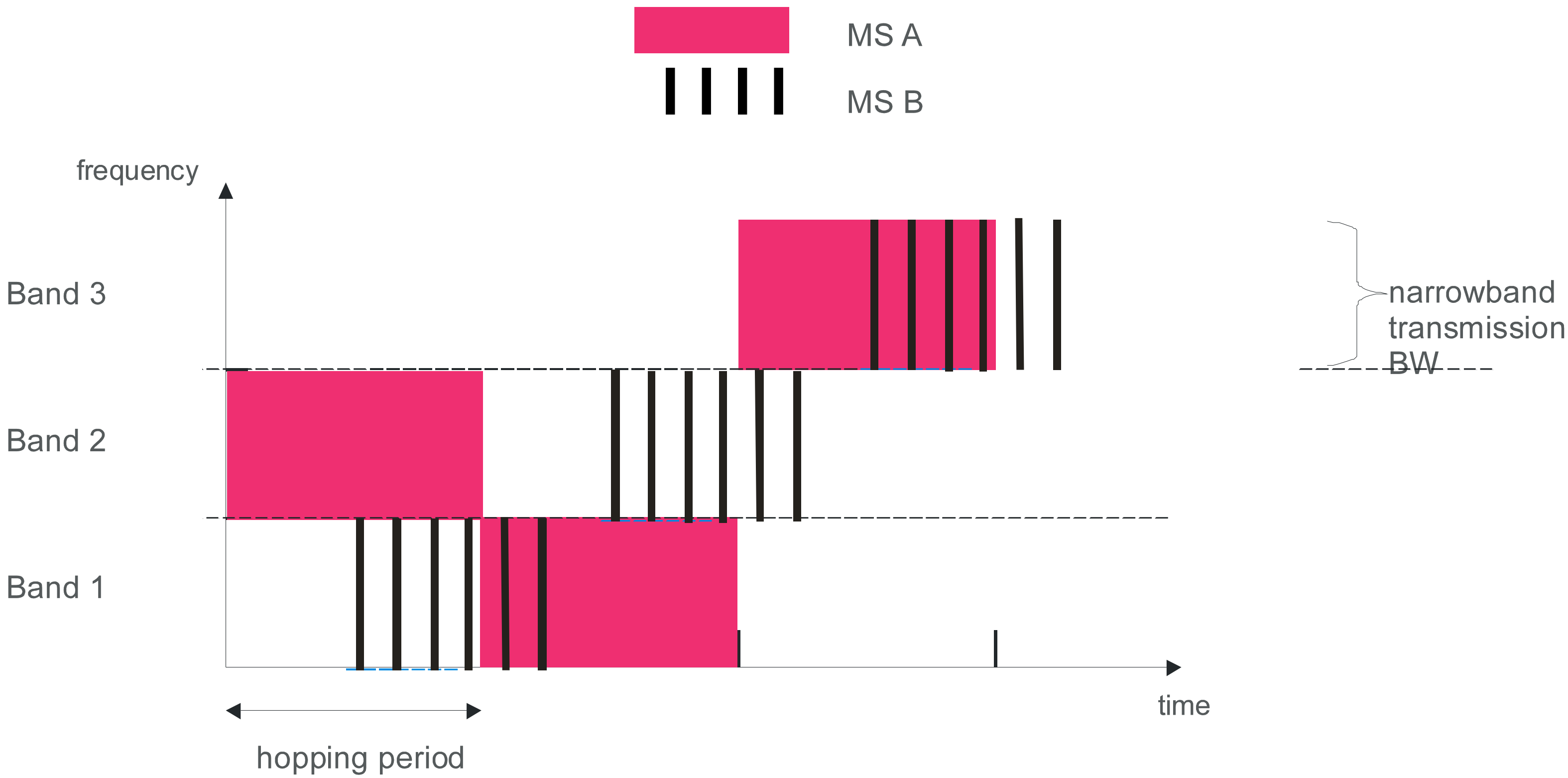


Users/channels are separated by using different hopping patterns.

FH codes (1)


MS A
MS B
frequency
Band 3
Band 2
Band 1
narrowband
transmission
BW
time
hopping period

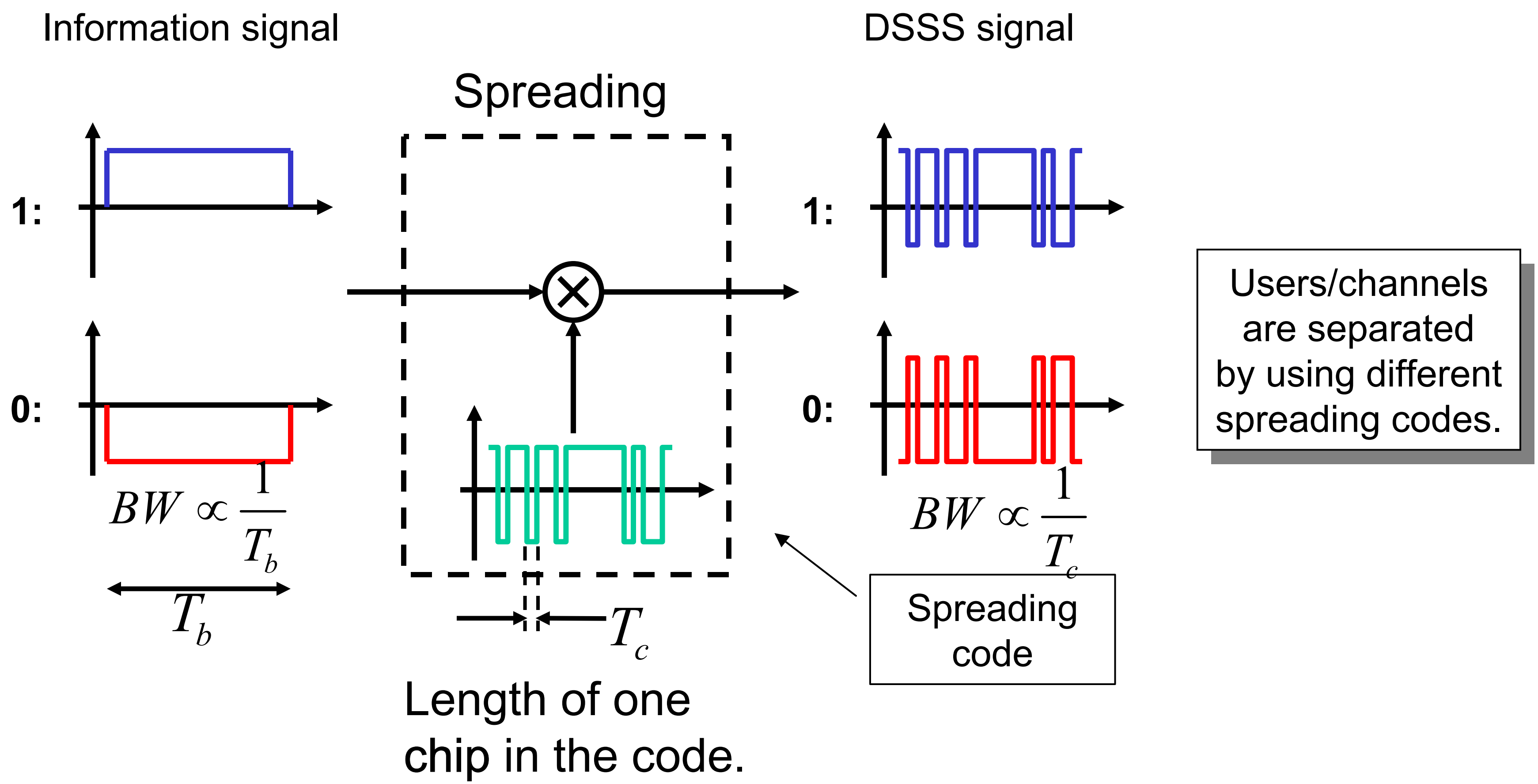
FH codes (2)



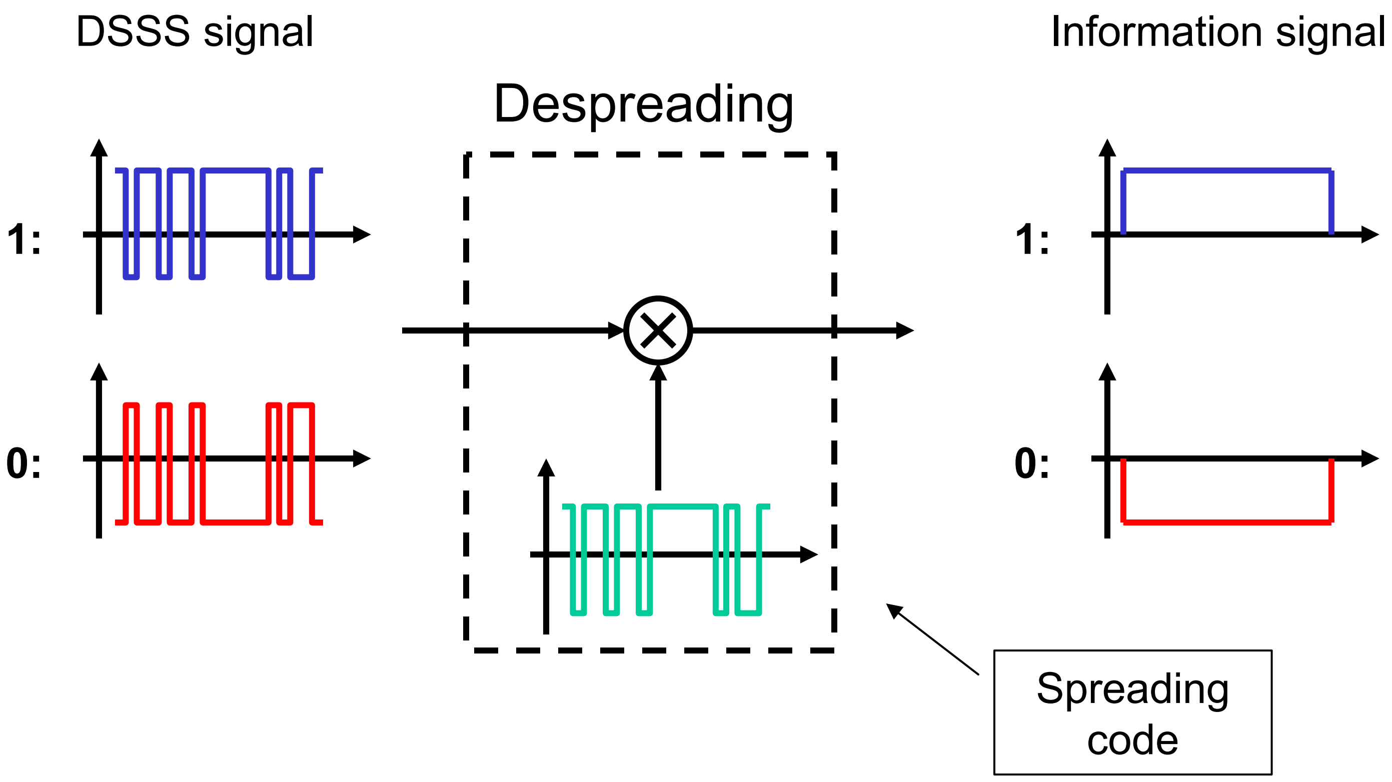
MS A
MS B
frequency
Band 3
Band 2
Band 1
narrowband
transmission
BW
time
hopping period

DIRECT SEQUENCE SPREAD SPECTRUM

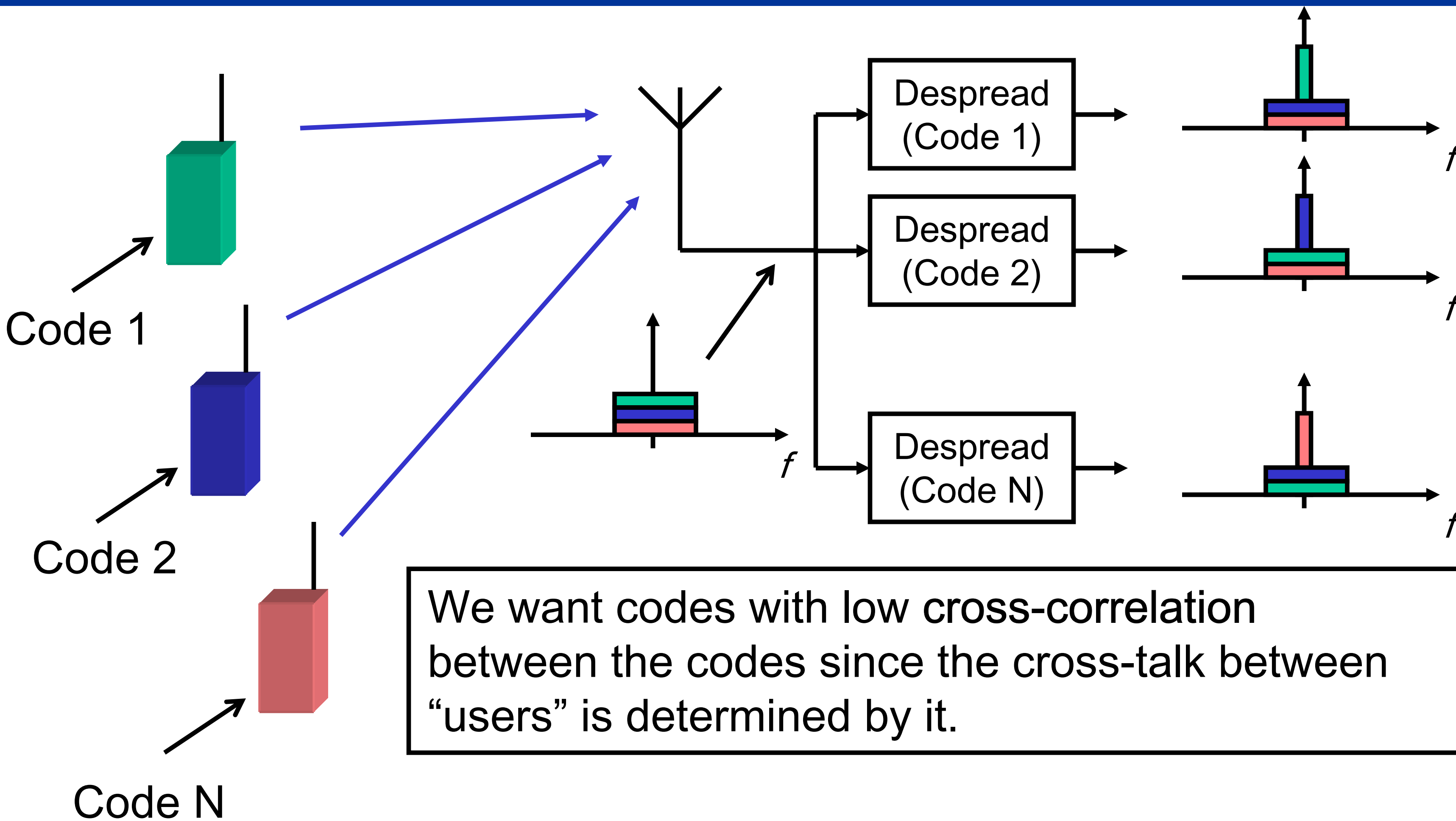
Direct-Sequence Spread Spectrum DSSS (1)



Direct-Sequence Spread Spectrum DSSS (2)



Code-division multiple access (CDMA)



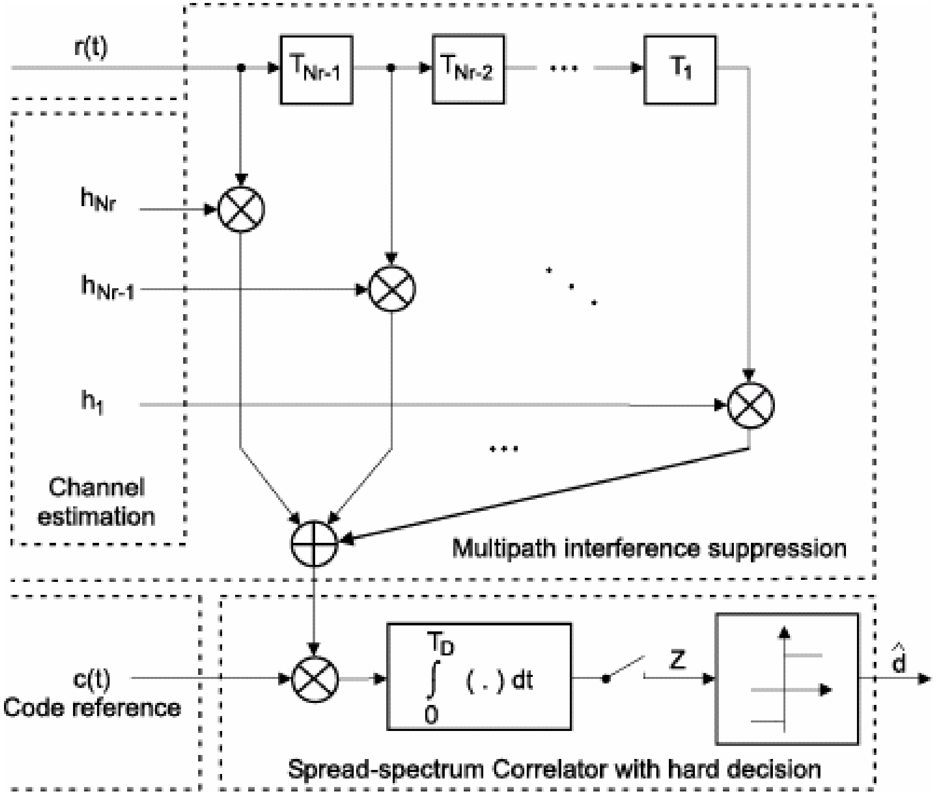
We want codes with low **cross-correlation** between the codes since the cross-talk between "users" is determined by it.

Impact of delay dispersion

- CDMA spreads signals over larger bandwidth -> delay dispersion has bigger impact.
- Two effects:
  - Intersymbol interference: independent of spreading; needs to be combatted by equalizer
  - Output of despreader is not impulse, but rather an approximation to the impulse response
- Needs Rake receiver to collect all energy

Rake receivers

Despreading becomes a bit more complicated ...



... but we gain frequency diversity.

Code families (1)

- Ideal goals:
  - Autocorrelation function is delta impulse

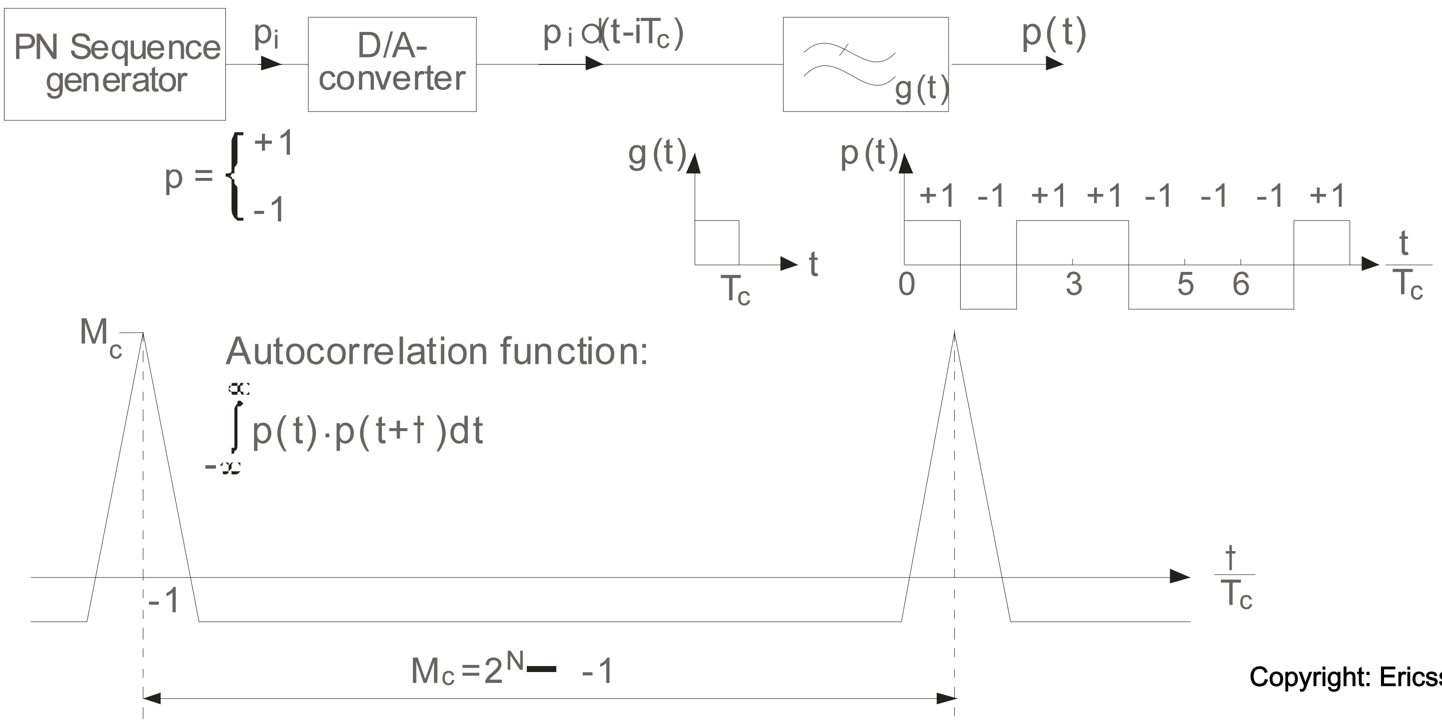
$$ACF(i) = \begin{cases} M_{\mathrm{C}} & \text{for } i = 0 \\ 0 & \text{otherwise} \end{cases}$$

  - Crosscorrelation function should be zero

$$CCF_{j,k}(t) = 0 \qquad \text{for } j \neq k$$

  - CCF properties should be (approx.) independent of relative shift between users

Code families (2)



Code families (3)

- Kasami-codes:
  - Larger family of codes that trades of number of codes vs. ACF and CCF properties
- Gold sequences
- Overview of results (for length 255):

Sequence	Number of codes		Maximum CCF	
PN-Sequence	$2^{N_{\mathrm{reg}}} - 1$	255		
Gold	$2^{N_{\mathrm{reg}}} + 1$	257	$\approx -3N_{\mathrm{reg}}/2 + 1.5$	$-10.5\mathrm{dB}$
S-Kasami	$2^{N_{\mathrm{reg}}/2}$	16	$\approx -3N_{\mathrm{reg}}/2$	$-12\mathrm{dB}$
L-Kasami	$2^{N_{\mathrm{reg}}/2}(2^{N_{\mathrm{reg}}} + 1)$	4112	$\approx -3N_{\mathrm{reg}}/2 + 3$	$-9\mathrm{dB}$
VL-Kasami	$2^{N_{\mathrm{reg}}/2}(2^{N_{\mathrm{reg}}} + 1)^2$	10^6	$\approx -3N_{\mathrm{reg}}/2 + 6$	$-6\mathrm{dB}$

Orthogonal codes

- Codes with perfect orthogonality are possible, but only for perfectly synchronized users

- Walsh-Hadamard codes:
  - Size 2x2

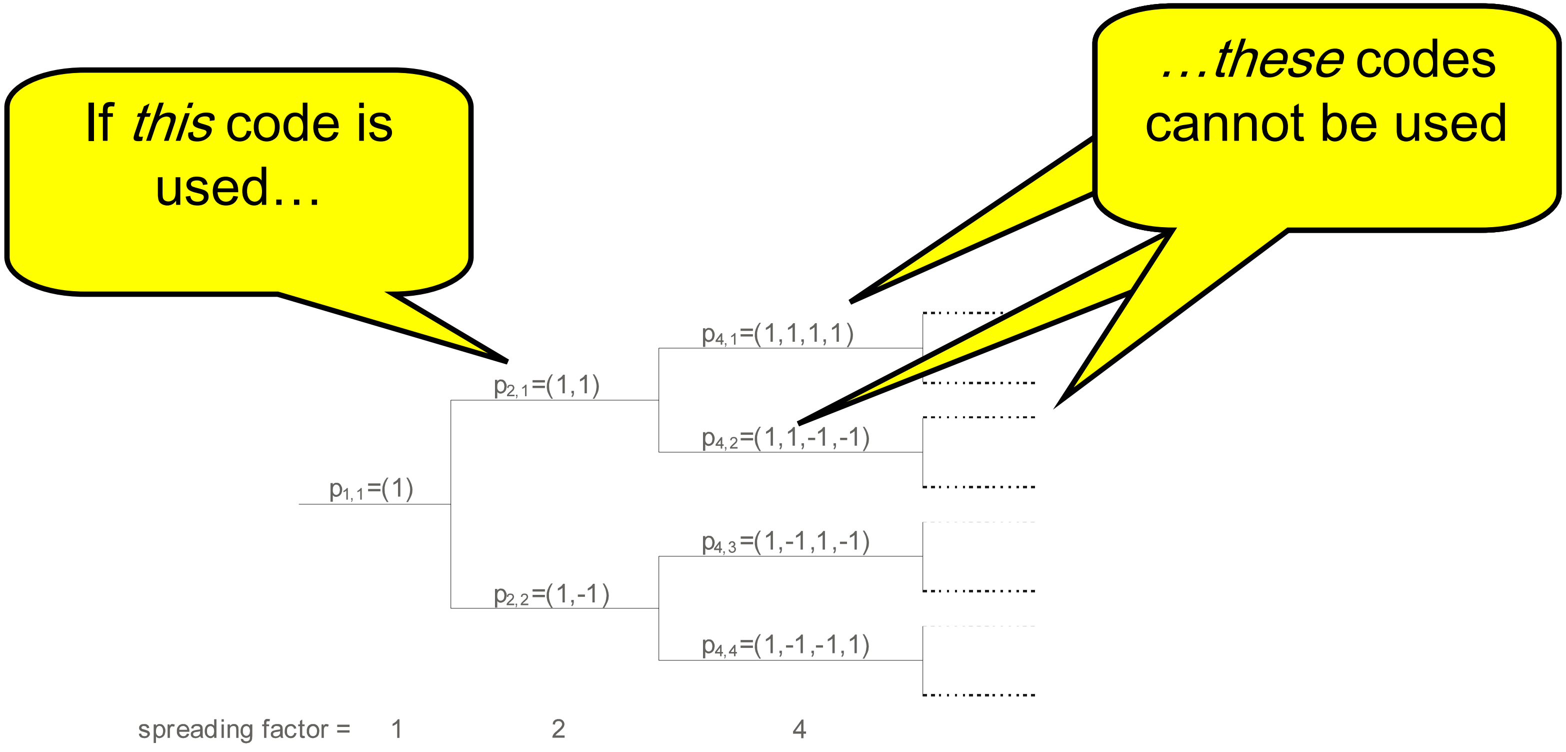
$$\mathbf{H}_{\mathrm{had}}^{(1)} = \begin{pmatrix} 1 & 1 \\ 1 & -1 \end{pmatrix}$$

  - Larger sizes: recursion

$$\mathbf{H}_{\mathrm{had}}^{(n+1)} = \begin{pmatrix} \mathbf{H}_{\mathrm{had}}^{(n)} & \mathbf{H}_{\mathrm{had}}^{(n)} \\ \mathbf{H}_{\mathrm{had}}^{(n)} & \overline{\mathbf{H}}_{\mathrm{had}}^{(n)} \end{pmatrix}$$

Orthogonal Variable Spreading Factor (OVSF) codes

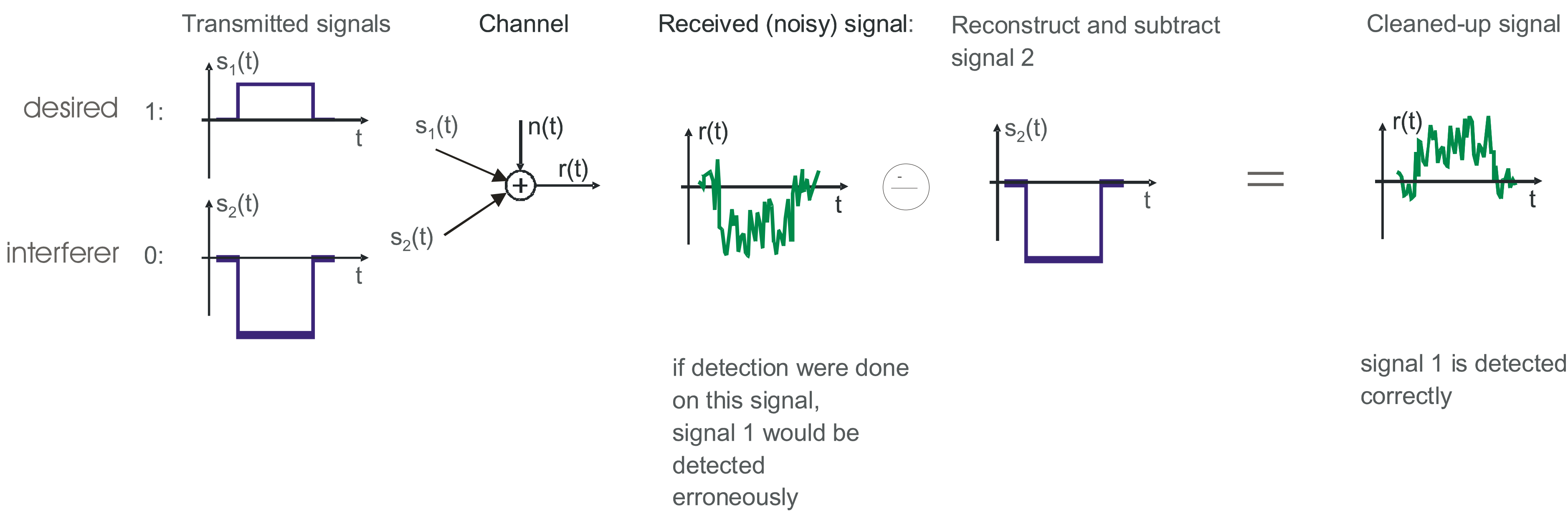
- When different spreading factors are needed



MULTIUSER DETECTION

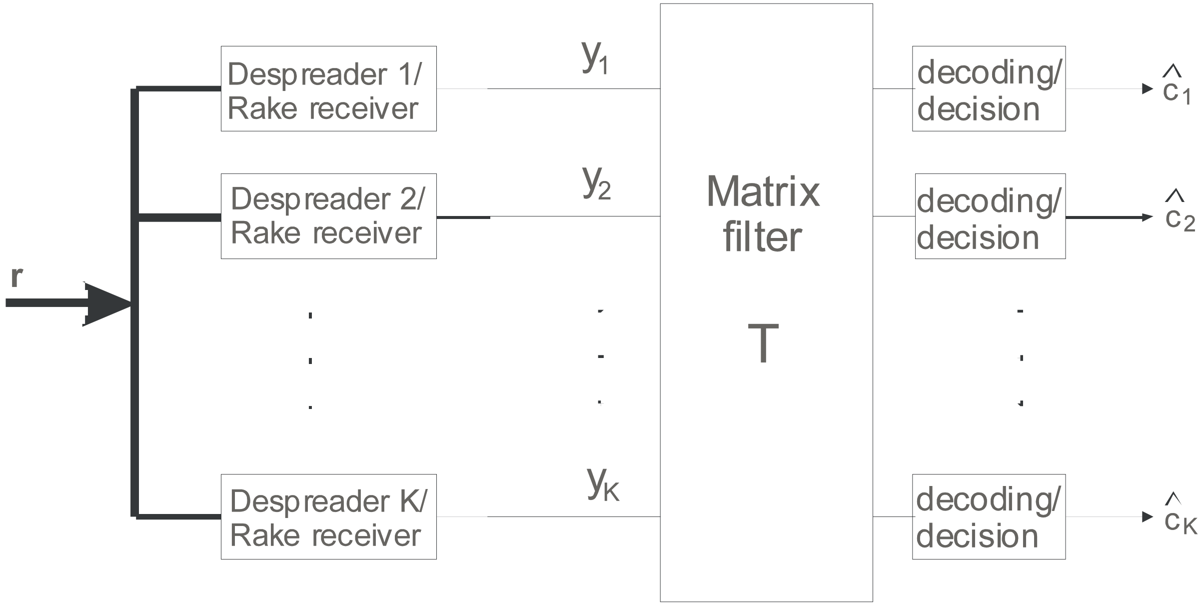
Principle of multiuser detection

- Conventional approach: treat interference like noise
- However: interference has structure that can be exploited



Linear MUD

- Receiver structure



- Zero-forcing: $\mathbf{T} = \mathbf{R}^{-1}$
  - Drawback: noise enhancement
- MMSE: $\mathbf{T} = [\mathbf{R}^{-1} + N_0\mathbf{I}]^{-1}$

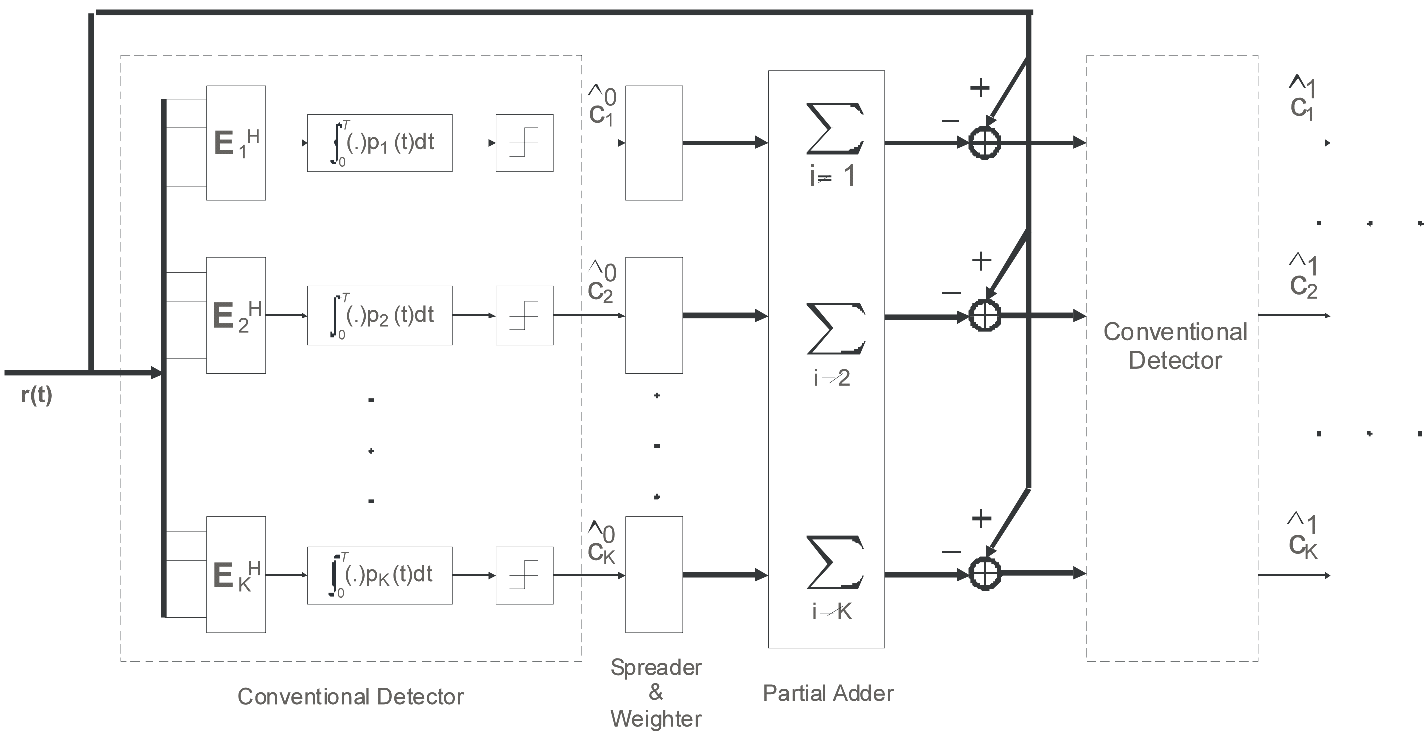
Nonlinear MUD (1)

- Multiuser MLSE:
  - optimally detect transmit sequences of all users
  - Number of states in trellis grow exponentially with number of users
  - Too complex for practical implementation

- Serial Interference cancellation:
  - Detect strongest user first; subtract its impact from signal, then detect second strongest,…
  - Drawback: error propagation

Nonlinear MUD (2)

- Parallel interference cancellation:
  - Detect all users at once; subtract *part* of their impact from signal, then repeat this

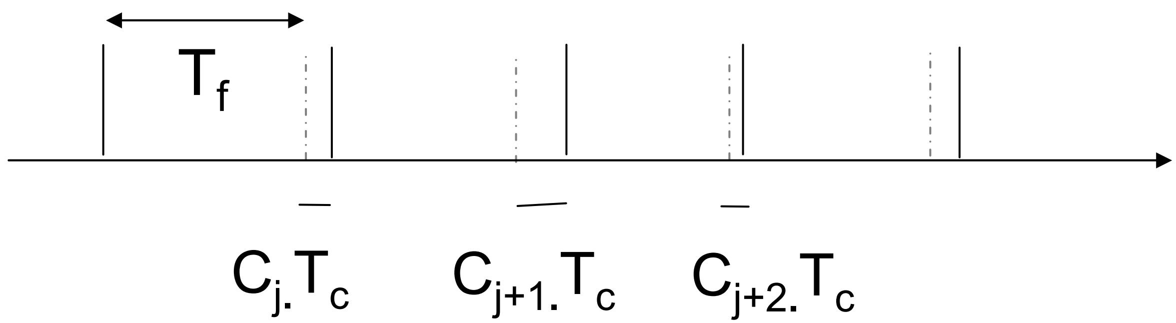


TIME HOPPING IMPULSE RADIO

Time Hopping

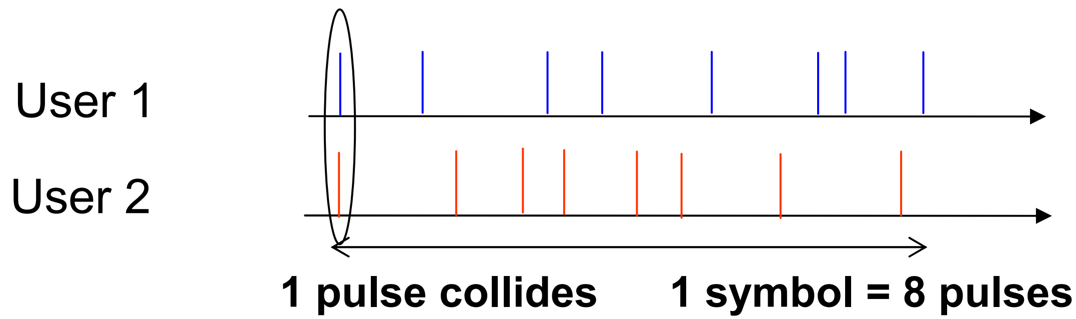
- Train of pulses

- $T_{Pul} \sim T_f/100$
- PN sequence $\{c_j\}$, NP code = N_p pulses, T_c: dither time



Interference Suppression

- Other impulse radio sources:
  - Relative delay of users cannot be influenced
  - Different users use different hopping codes
  - No “catastrophic collision” possible



- Narrowband interference
  - Receiver “sees” it only for duration of pulse
  - Suppression by factor Tf/Tp

Summary

- The available radio resource is shared among users in a multiple access scheme.
- When we apply a cellular structure, we can reuse the same channel again after a certain distance.
- In cellular systems the limiting factor is interference.
- For FDMA and TDMA the tolerance against interference determines the possible cluster size and thereby the amount of resources available in each cell.
- For CDMA systems, we use cluster size one, and the number of users depends on code properties and the capacity to perform interference cancellation (multi-user detection).

Orthogonal Frequency Domain Multiplexing

Contents

- Principle and motivation
- Analogue and digital implementation
- Frequency-selective channels: cyclic prefix
- Channel estimation
- Peak-to-average ratio
- Inter-channel interference
- Adaptive modulation
- Multi-carrier CDMA

PRINCIPLE, MOTIVATION AND BASIC IMPLEMENTATION

Principle (1)

- For very high data rates, equalization and Rake reception becomes difficult
  - Important quantity: product of maximum excess delay and system bandwidth
  - Especially critical for wireless LANs and PANs

- Solution:
  - transmit multiple data streams with lower rates on several carriers
  - Have carriers multiplexed in the most efficient possible way:
  - Signals on the carriers can overlap and stay orthogonal

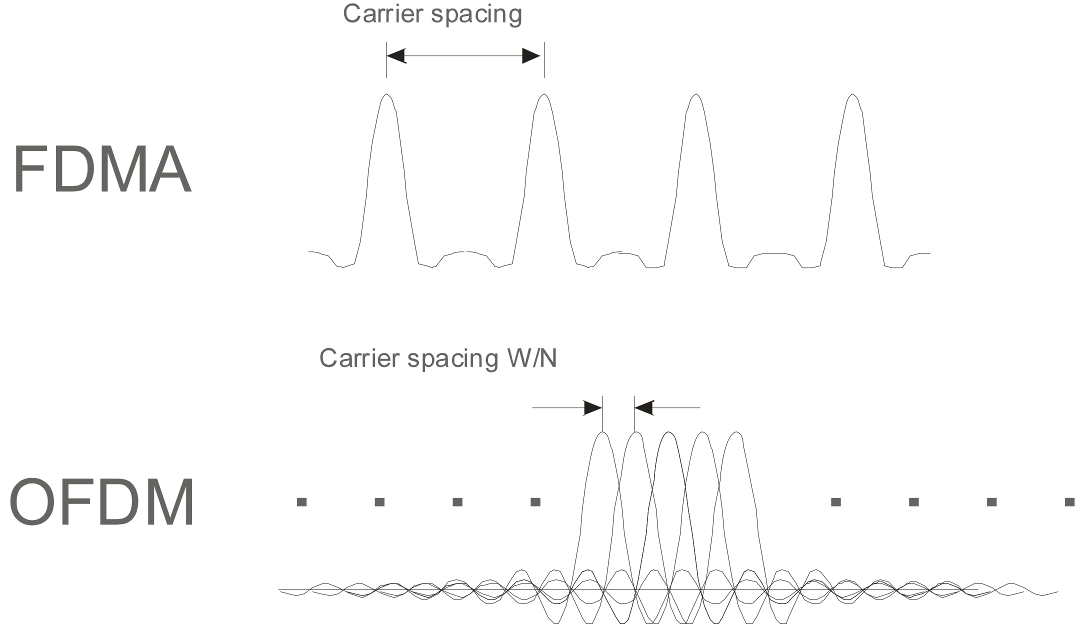
Principle (2)

- How close can we space the carriers?

$$f_n = nW/N \qquad W = N/T_S$$

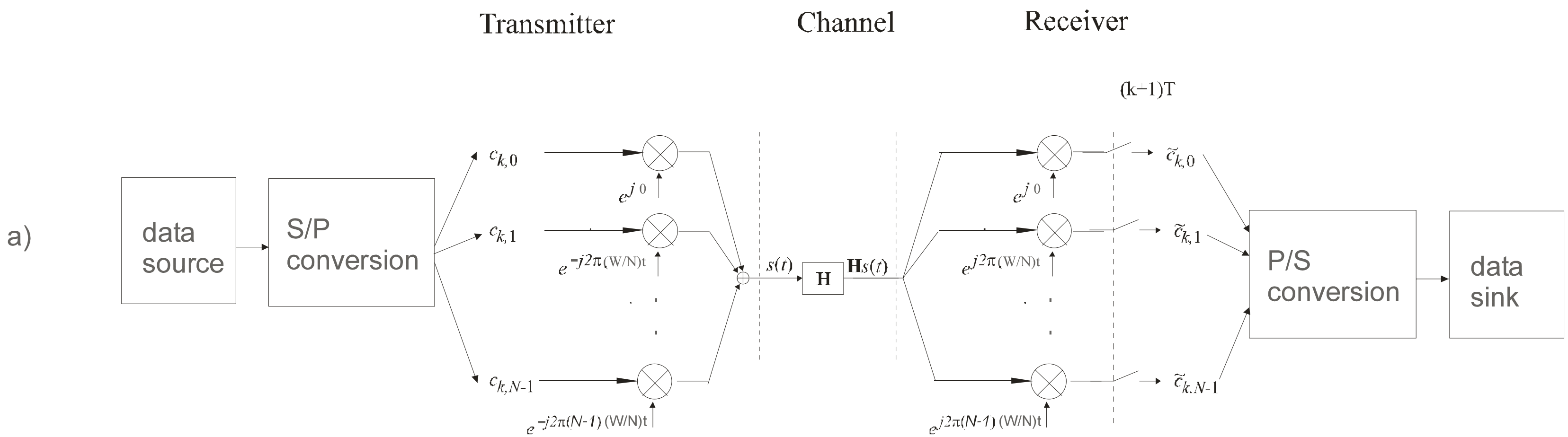
- Carriers are still orthogonal

$$c_n c_k \int_{iT_S}^{(i+1)T_S} \exp(j2\pi f_n t)\exp(-j2\pi f_k t)dt = c_n c_k \delta_{nk}$$

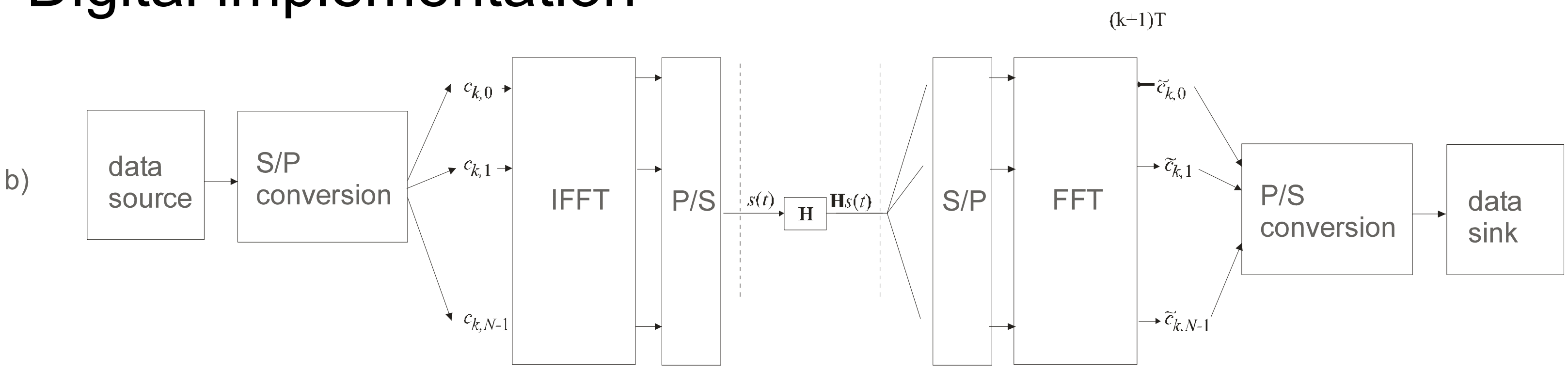


Analogue vs. digital implementation

- Analogue implementation



- Digital implementation



Why can we use an IFFT?

- Transmit signal is

$$s(t) = \sum_{i=-\infty}^{\infty} s_i(t) = \sum_{i=-\infty}^{\infty} \sum_{n=0}^{N-1} c_{n,i} g_n(t - iT_S)$$

- With basis pulse

$$g_n(t) = \begin{cases} \frac{1}{\sqrt{T_S}} \exp(j2\pi n \frac{t}{T_S}) & \text{for } 0 < t < T_S \\ 0 & \text{otherwise} \end{cases}$$

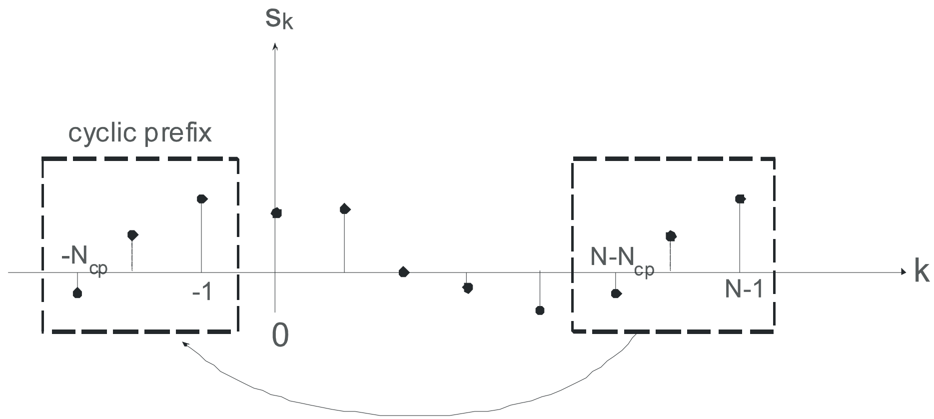
- Transmit signal sampled at $t_k = kT_S/N$

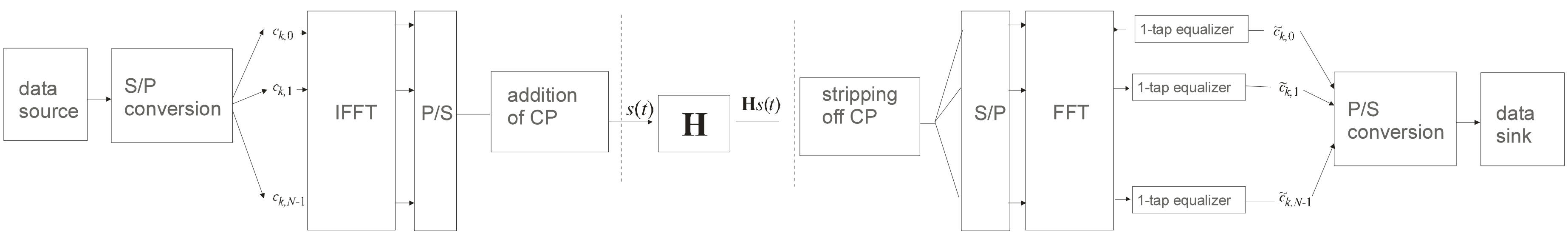
$$s_k = s(t_k) = \frac{1}{\sqrt{T_S}} \sum_{n=0}^{N-1} c_{n,0} \exp(j2\pi n \frac{k}{N}) \, .$$

- This is the definition of an IFFT

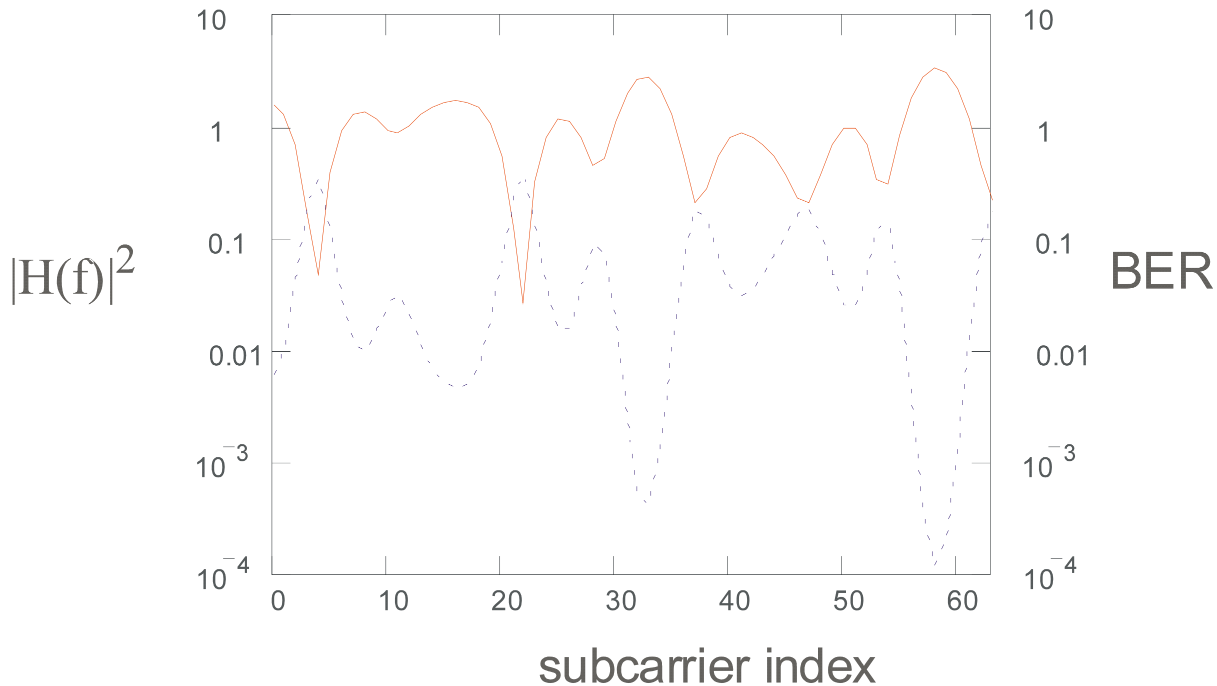
Frequency-selective channels

- Cyclic prefix, i.e., repeat last samples at beginning of symbol
- Converts linear to circular convolution

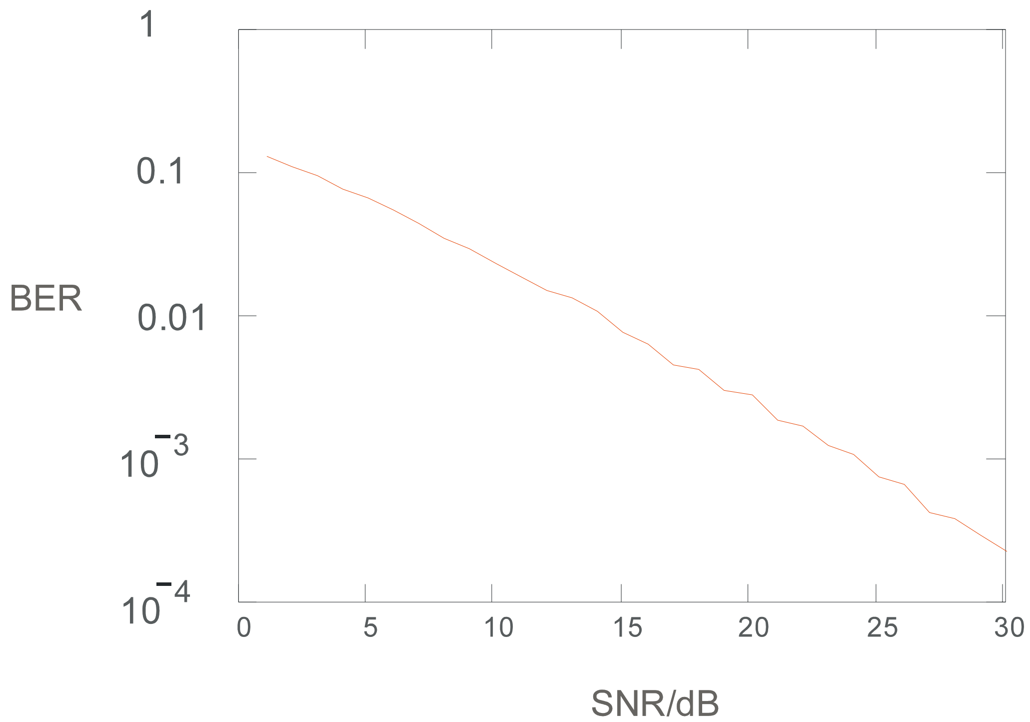




Performance in frequency-selective channels (1)



Performance in frequency-selective channels (2)



Performance in frequency-selective channels (3)

- How to improve performance?
  - adaptive modulation (different signal alphabets in different subcarriers)
  - spreading the signal over all tones (multicarrier CDMA)
  - Coding across different tones

ADVANCED IMPLEMENTATION ISSUES

Channel estimation (1)

- Easiest approach: dedicated pilot symbols
- Estimated channel gain on subchannel *n*

$$h_{n,i}^{\mathrm{LS}} = r_{n,i}/c_{n,i}$$

where *r* is the received signal and *c* the transmit signal

- Performance improvement:
  - Channels on subcarriers are correlated
  - Exploit that knowledge for noise averaging

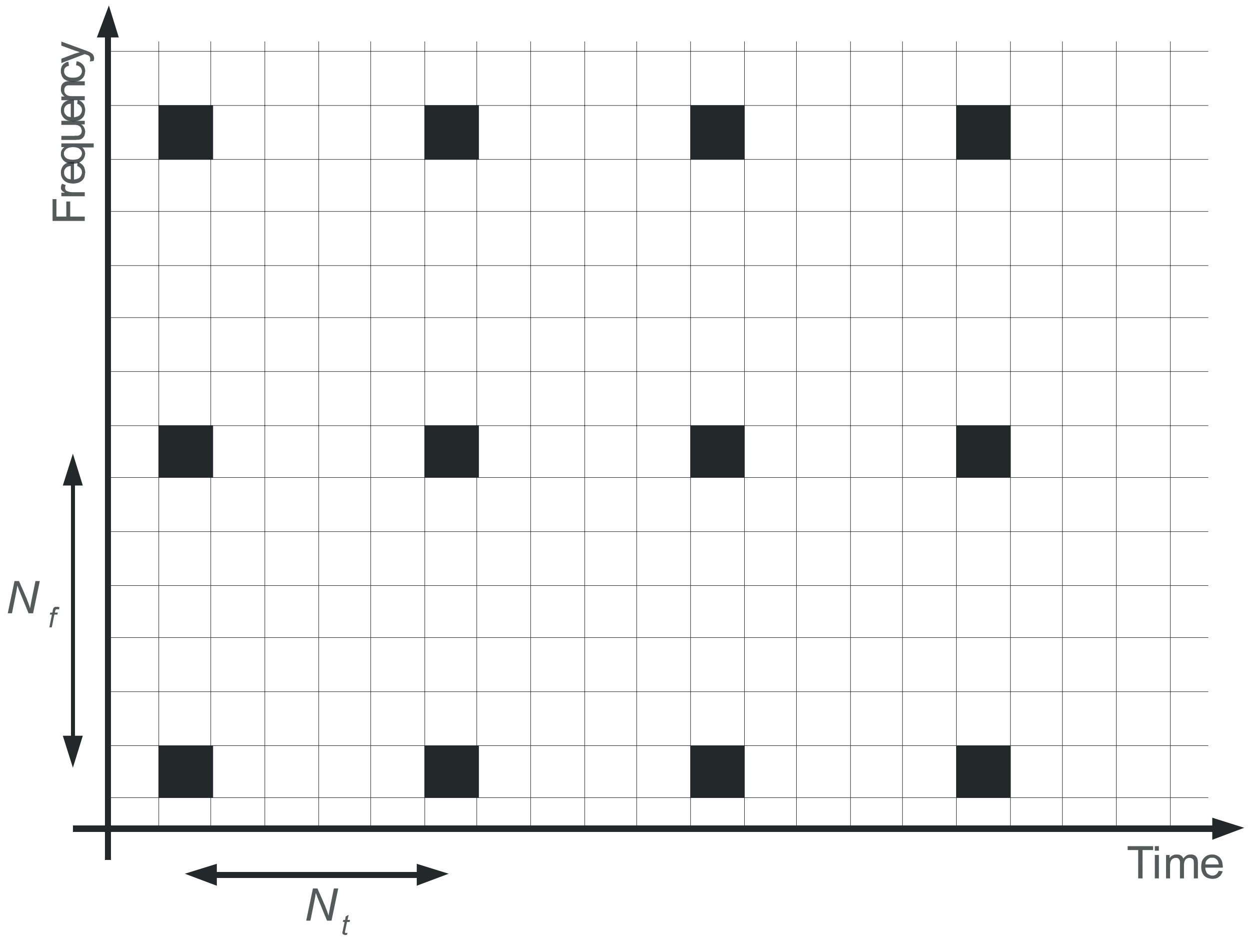
$$\mathbf{h}_i^{\mathrm{LMMSE}} = R_{hh^{\mathrm{LS}}} R_{h^{\mathrm{LS}}h^{\mathrm{LS}}}^{-1} \mathbf{h}_i^{\mathrm{LS}}$$

$R_{hh^{\mathrm{LS}}}$: covariance matrix between channel gains and least-squares estimate of channel gains,

$R_{h^{\mathrm{LS}}h^{\mathrm{LS}}}$: autocovarance matrix of least-squares estimates

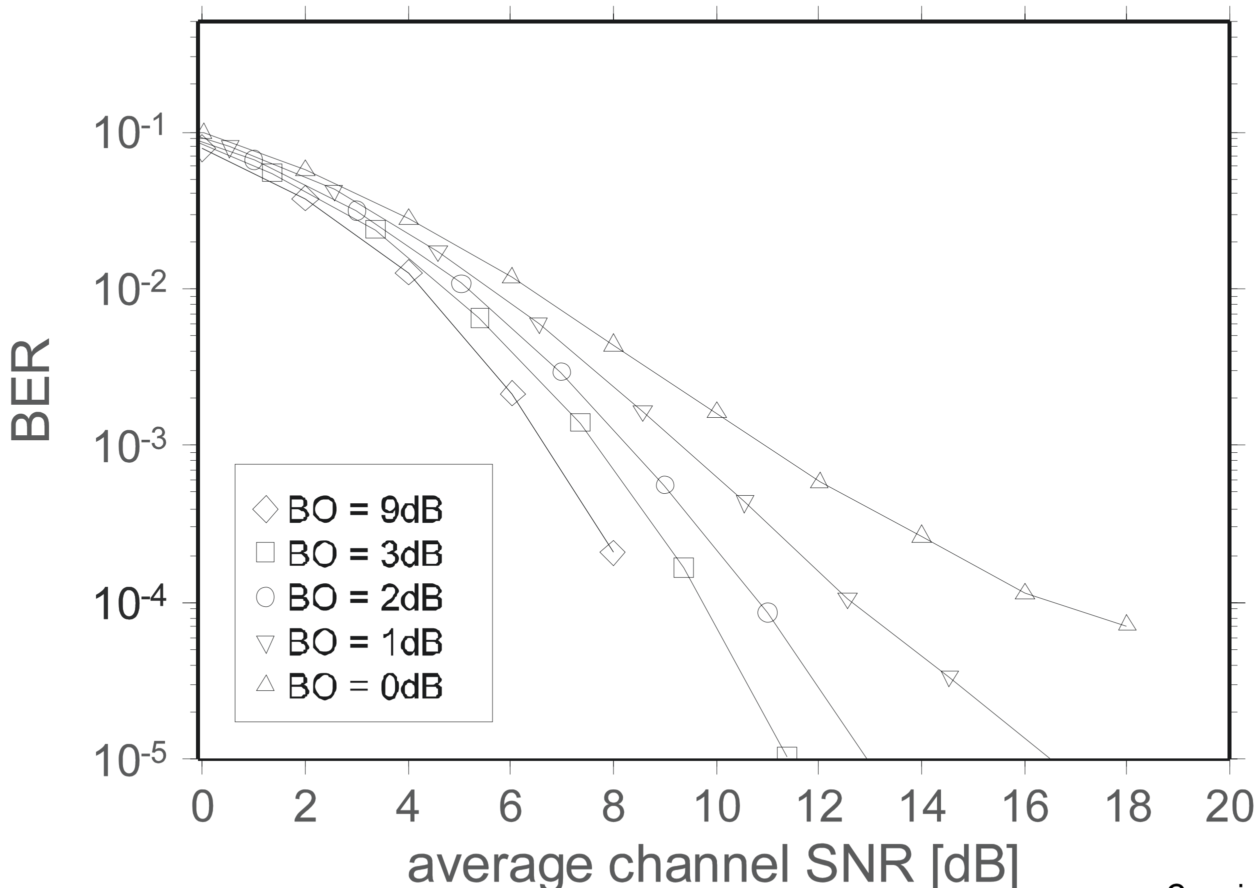
Channel estimation (2)

- Reduction of overhead by scatterered pilots



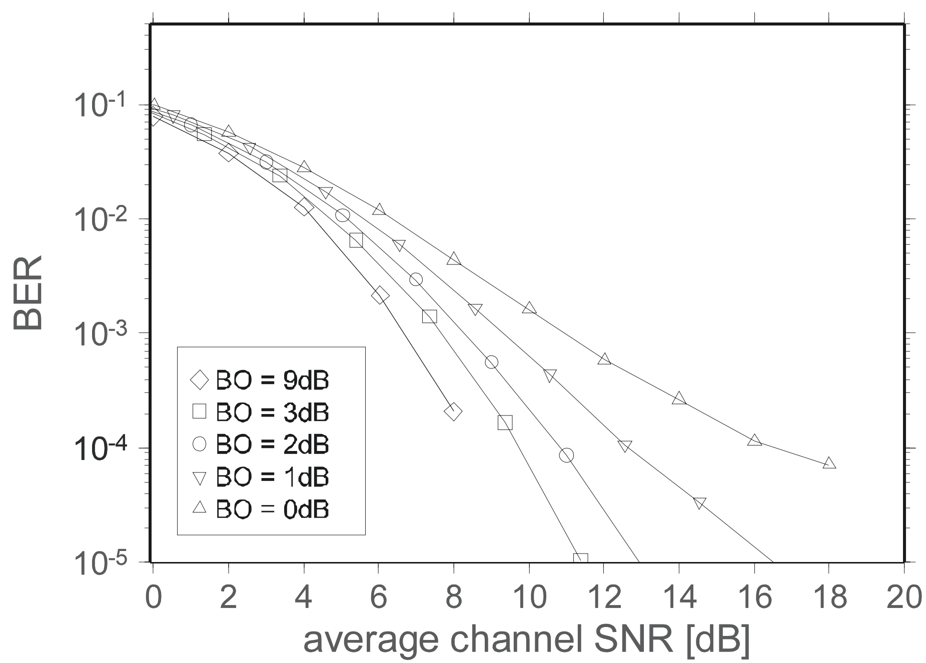
Effect of PAR problem

- Increases BER



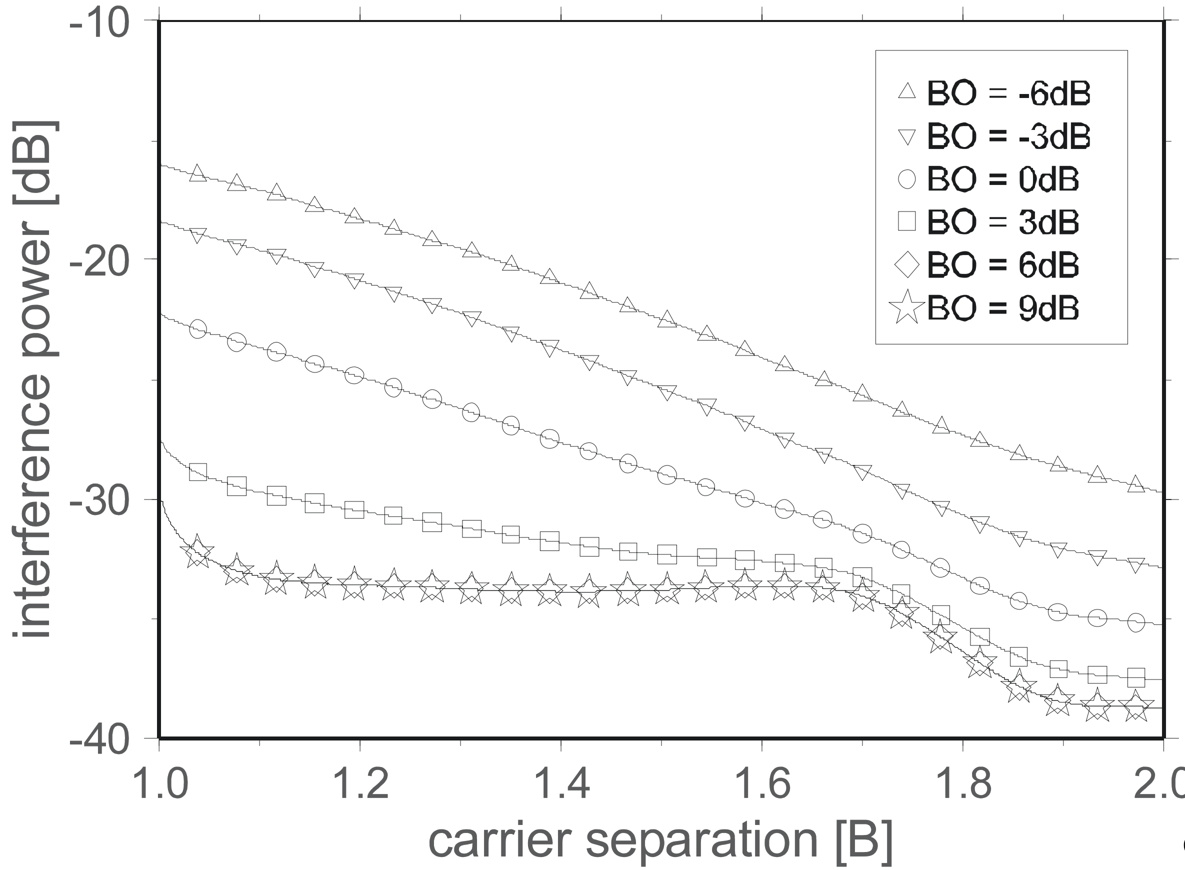
Remedies for the PAR problem (1)

- Backoff



Remedies for the PAR problem (2)

– Residual cutoff results in spectral regrowth



Remedies for the PAR problem (3)

- Coding for PAR reduction

- Phase adjustments
  - Cannot guarantee certain PAR

Remedies for PAR problem (4)

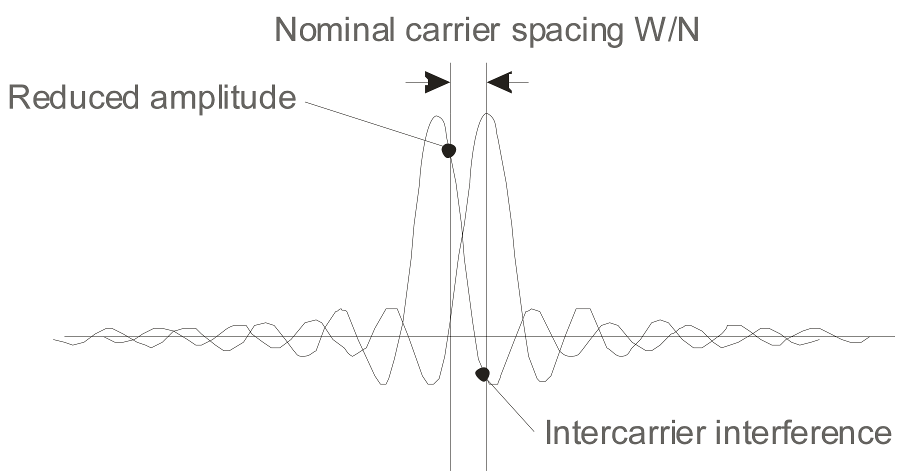
- Correction by multiplicative factor
  - Simplest case: clipping
  - More gentle: Gaussian functions

$$\hat{s}(t) = s(t)\left[1 - \sum_n \max\left(0, \frac{|s_k| - A_0}{|s_k|}\right) \exp\left(-\frac{t^2}{2\sigma_t^2}\right)\right]$$

- Correction by additive factor

Intercarrier interference (ICI)

- Intercarrier interference occurs when subcarriers are not orthogonal anymore



Remedies for ICI (1)

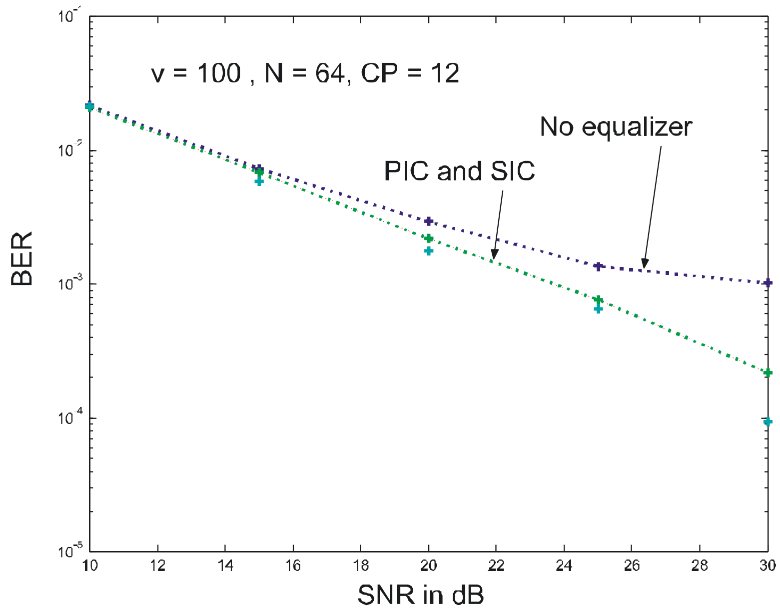
- Optimize the carrier spacing and symbol duration
  - Larger subcarrier spacing leads to smaller ICI
  - Larger spacing leads to shorter symbol duration: more sensitive to ICI; cyclic prefix makes it less spectral efficient
  - Maximize

$$SINR = \frac{\frac{E_S}{N_0} P_{sig} \frac{N}{N_{cp}+N}}{\frac{E_S}{N_0} P_{sig} \frac{N}{N_{cp}+N} \frac{P_{ISI}+P_{ICI}}{P_{sig}} + 1}$$

- Optimum choice of OFDM basis signals

Remedies for ICI (2)

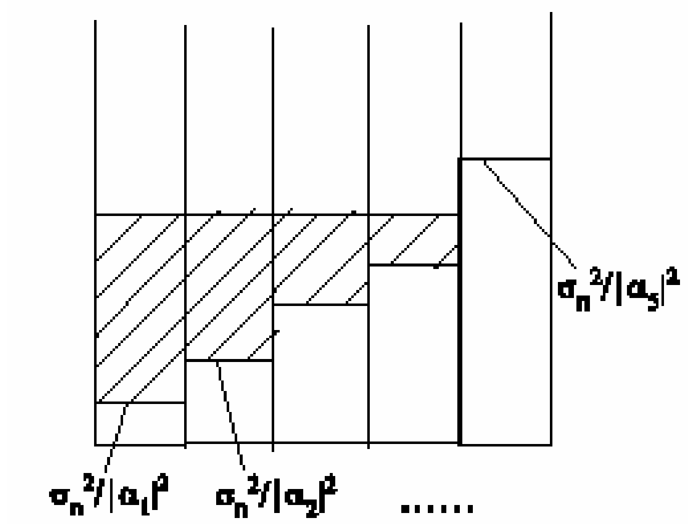
- Self-interference cancellation
- Frequency-domain equalizers



Waterfilling

- To optimize capacity, different powers should be allocated to the subcarriers
- Waterfilling:

$$P_n = \max\left(0, \varepsilon - \frac{\sigma_n^2}{|\alpha_n|^2}\right) \quad \text{with} \quad P = \sum_{n=1}^{N} P_n$$



MUTLICARRIER CDMA (MC-CDMA) AND SINGLE-CARRIER FREQUENCY-DOMAIN EQUALIZATION (SC-FDE)

And now for the mathematics…

- A code symbol *c* is mapped onto a transmit vector, by multiplication with spreading code p.
- For parallel transmission of symbols: a vector of transmit symbols c is mapped by multiplication with a *spreading matrix* P that consists of the spreading codes for the different symbols

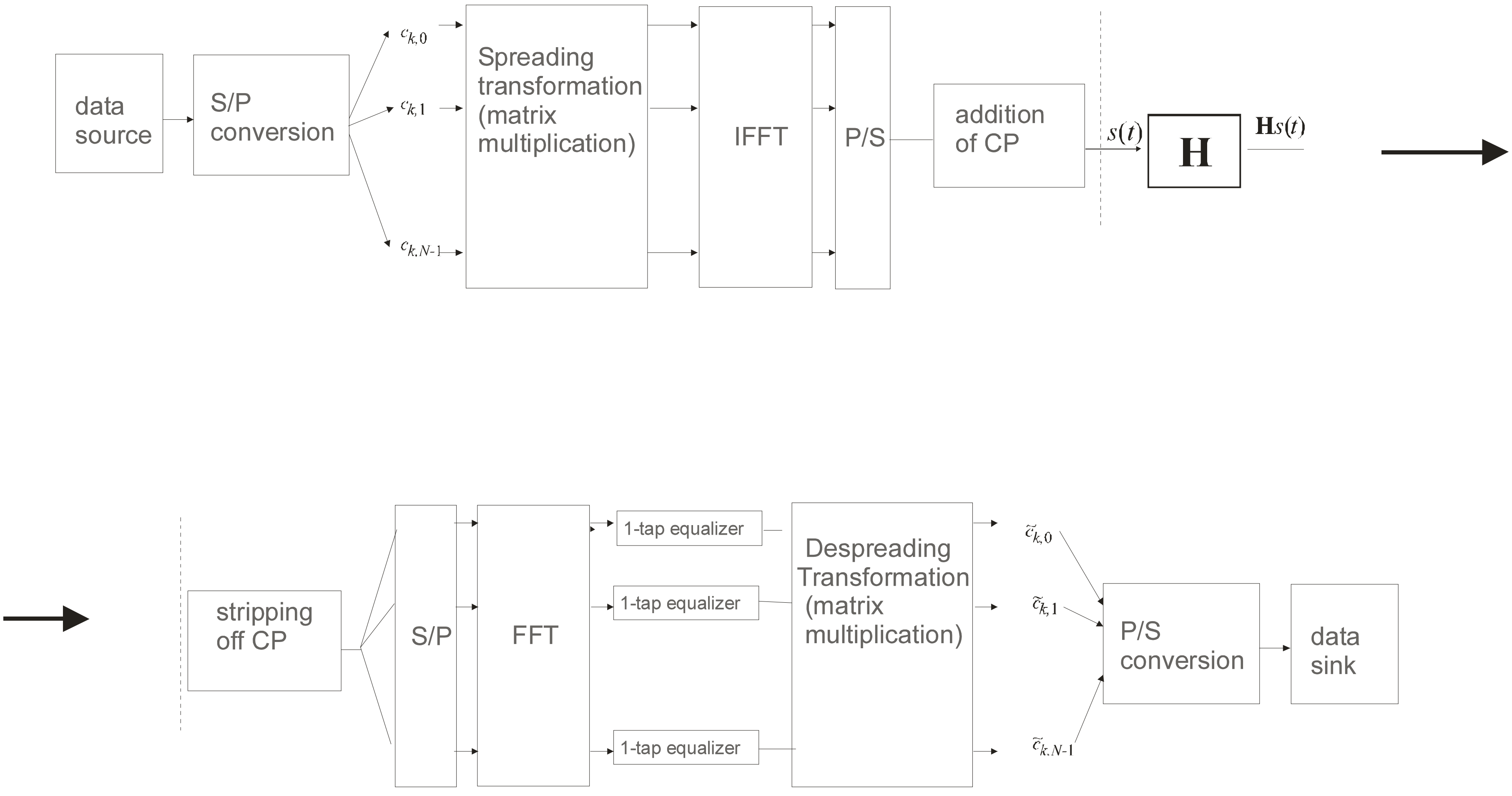
$$\tilde{\mathbf{c}} = \mathbf{P}\mathbf{c} \qquad \mathbf{P} = \left[\mathbf{p}_1 \ \mathbf{p}_2 \cdots\cdots \mathbf{p}_N\right]$$

- Symbol spreading is undone at the receiver

$$\tilde{\mathbf{r}} = \mathbf{H}\tilde{\mathbf{c}} + \mathbf{n}$$

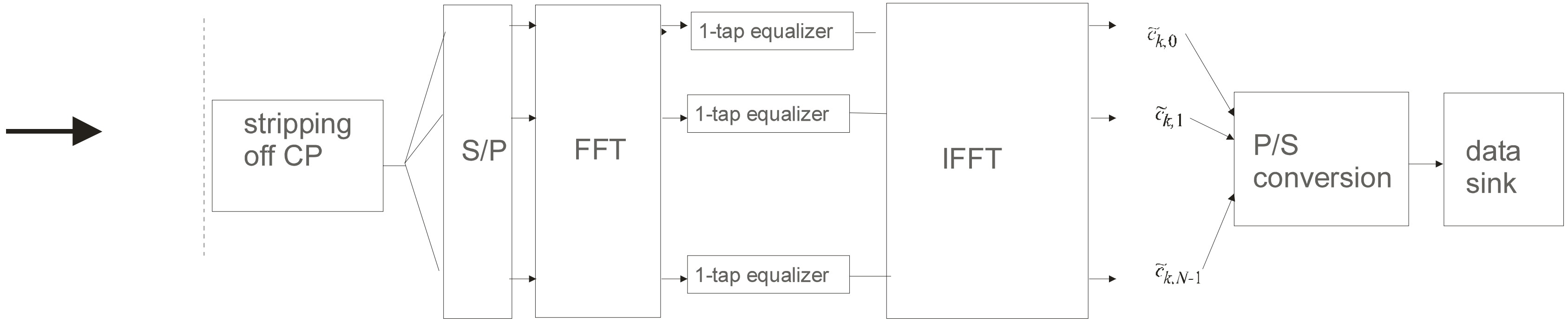
$$\begin{aligned} \mathbf{P}^\dagger\mathbf{H}^{-1}\tilde{\mathbf{r}} &= \mathbf{P}^\dagger\mathbf{H}^{-1}\mathbf{H}\mathbf{P}\mathbf{c} + \mathbf{P}^\dagger\mathbf{H}^{-1}\mathbf{n} \\ &= \mathbf{c} + \tilde{\mathbf{n}} \end{aligned}$$

Transceiver structure for MC-CDMA



SC-FDE Principle

- Move the IFFT from the TX to the RX

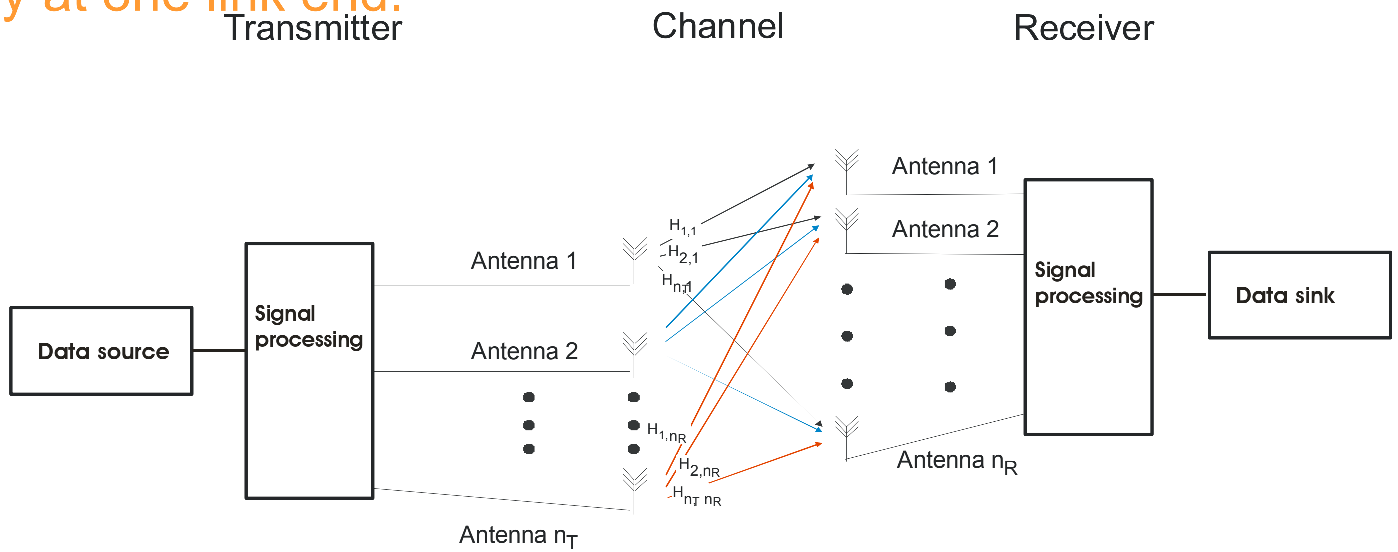


Multiple antenna systems

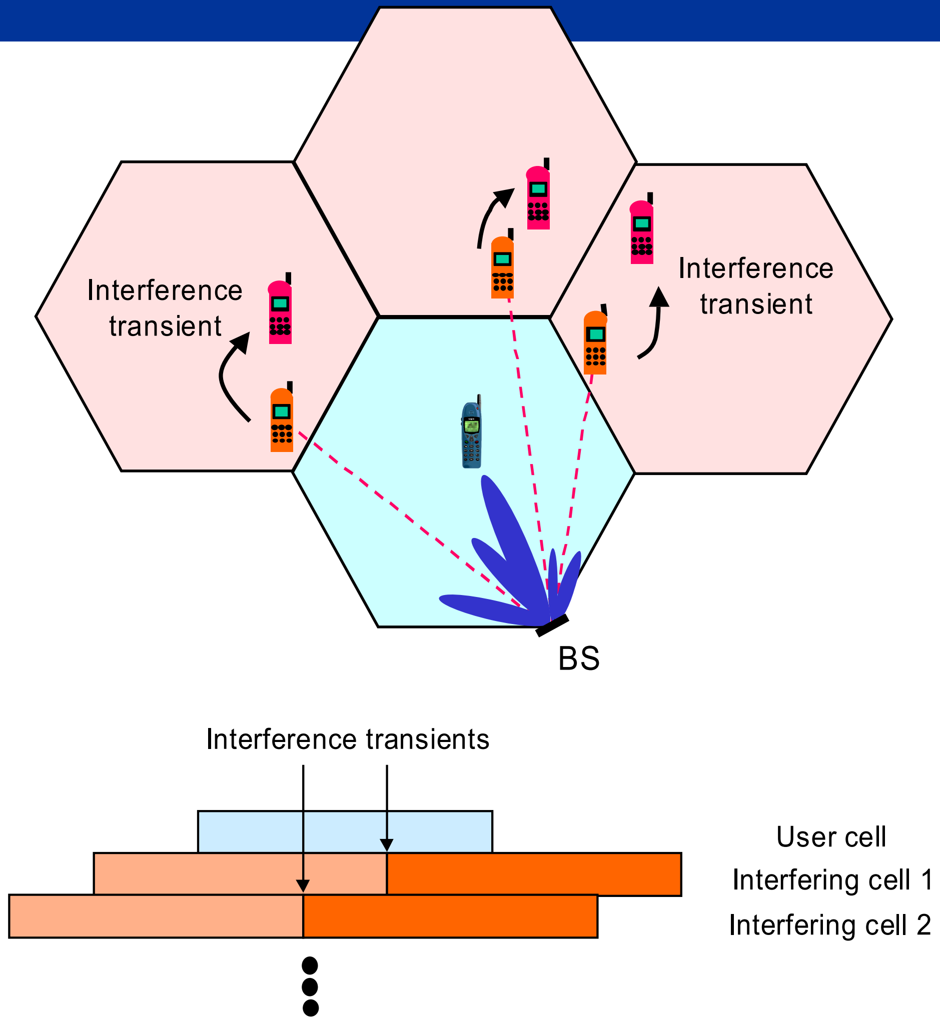
Definitions

- What are smart antennas and MIMO systems?

  A MIMO system consists of several antenna elements, plus adaptive signal processing, at both transmitter and receiver, the combination of which exploits the spatial dimension of the mobile radio channel. A smart antenna system is a system that has multiple antenna elements only at one link end.

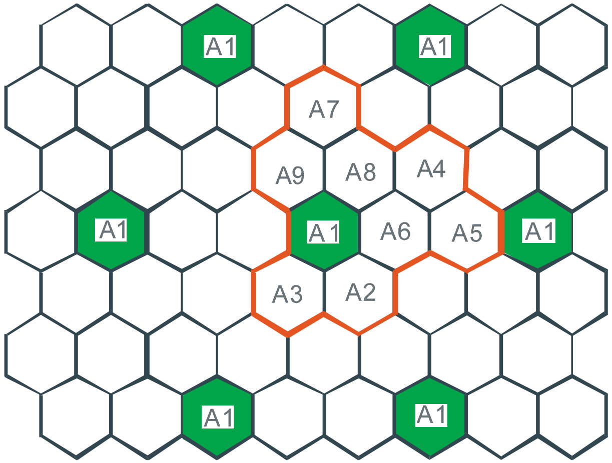


TDMA System with SFIR (1)

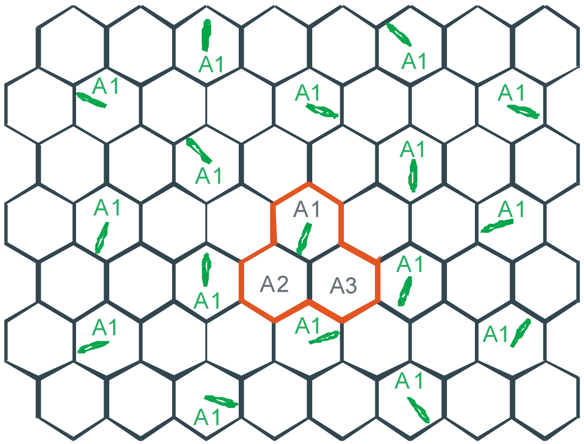


TDMA System with SFIR (2)

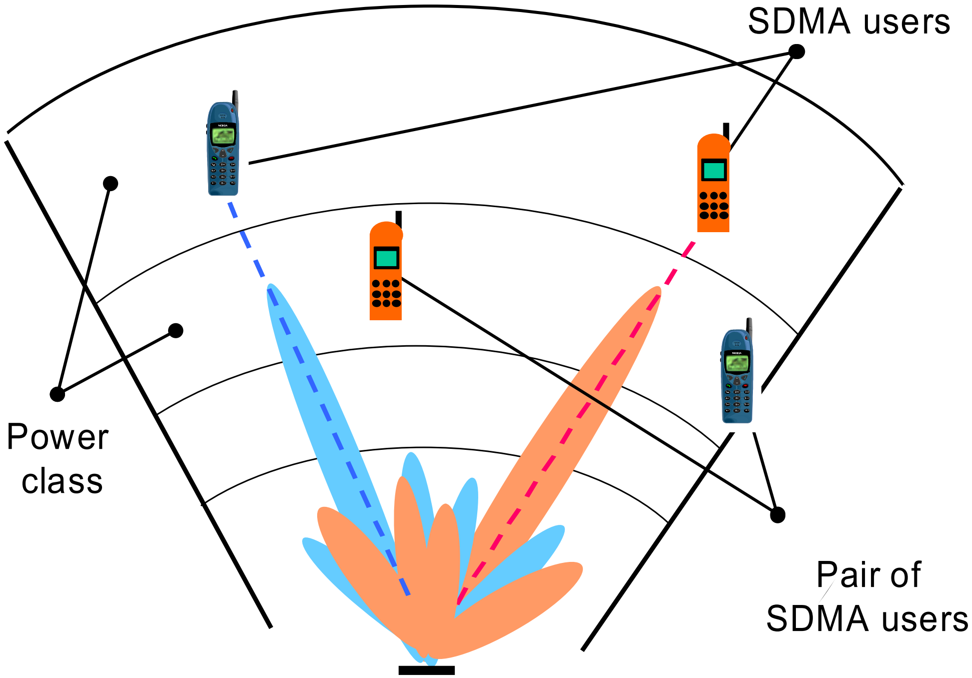
Conventional cell pattern



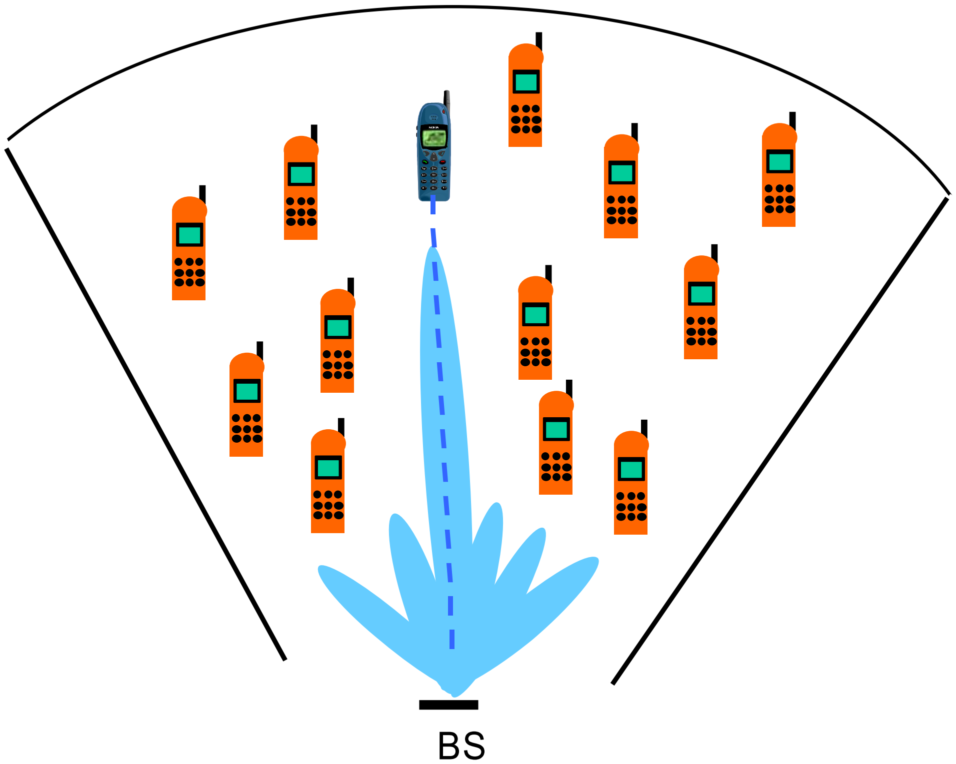
Spatial filtering



TDMA System with SDMA

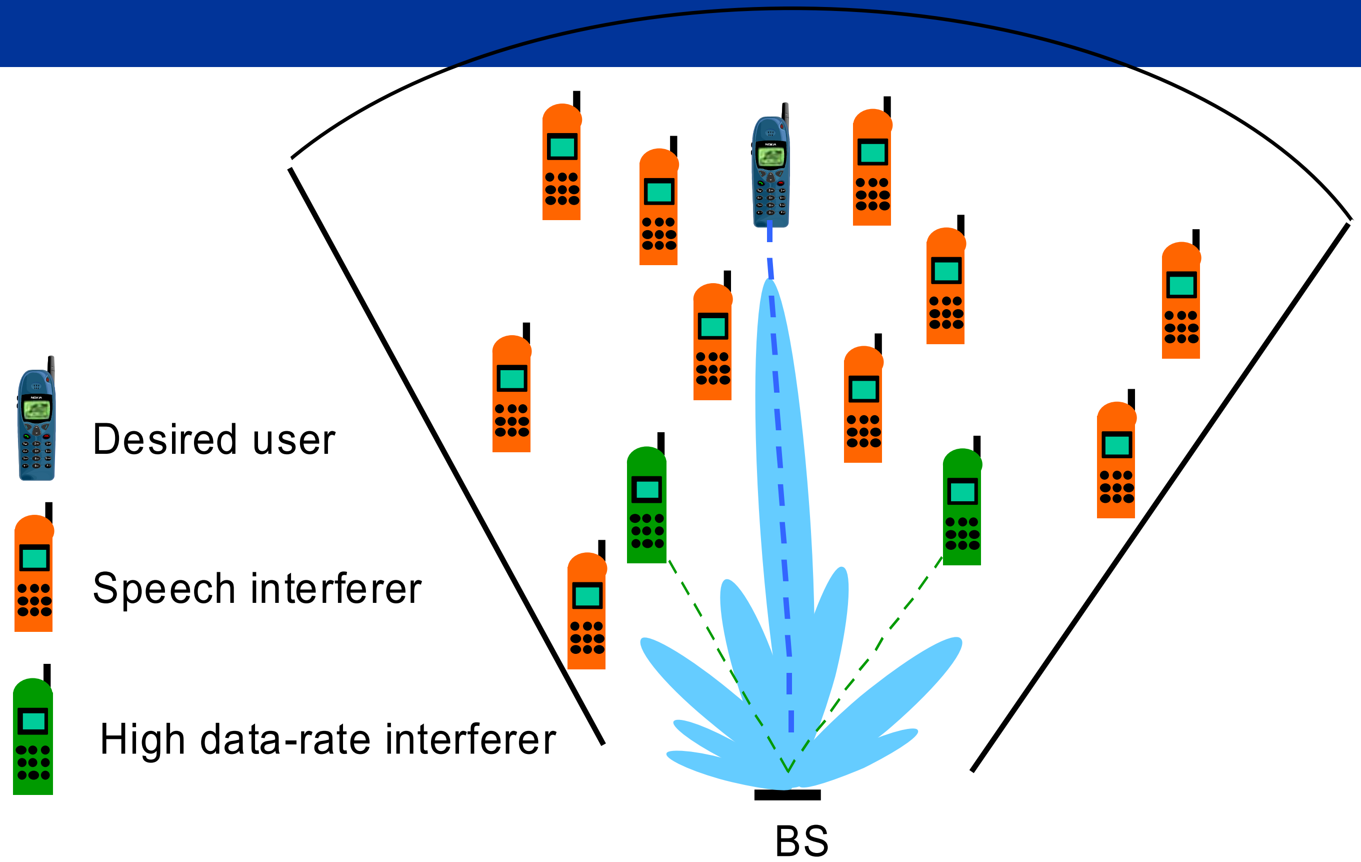


2G (single rate) CDMA System



$$K = \frac{M_c \cdot N_r}{SIR_{\text{threshold}}}$$

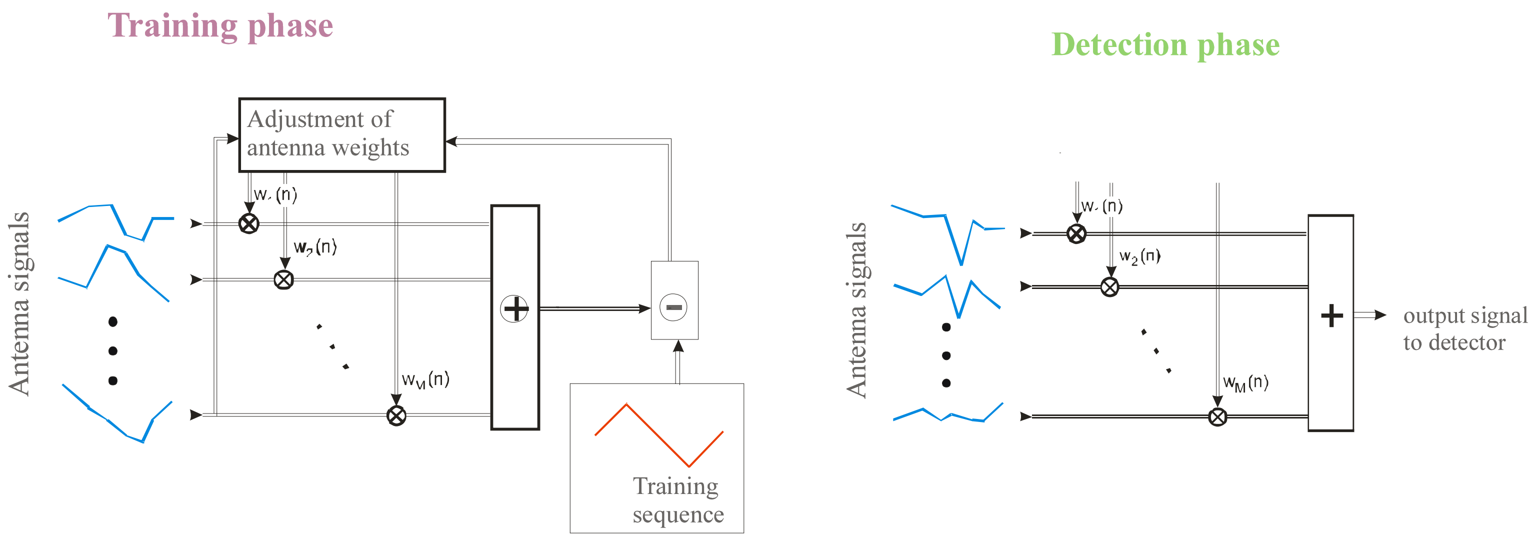
3G (Multirate) CDMA System



Temporal reference (TR) algorithms

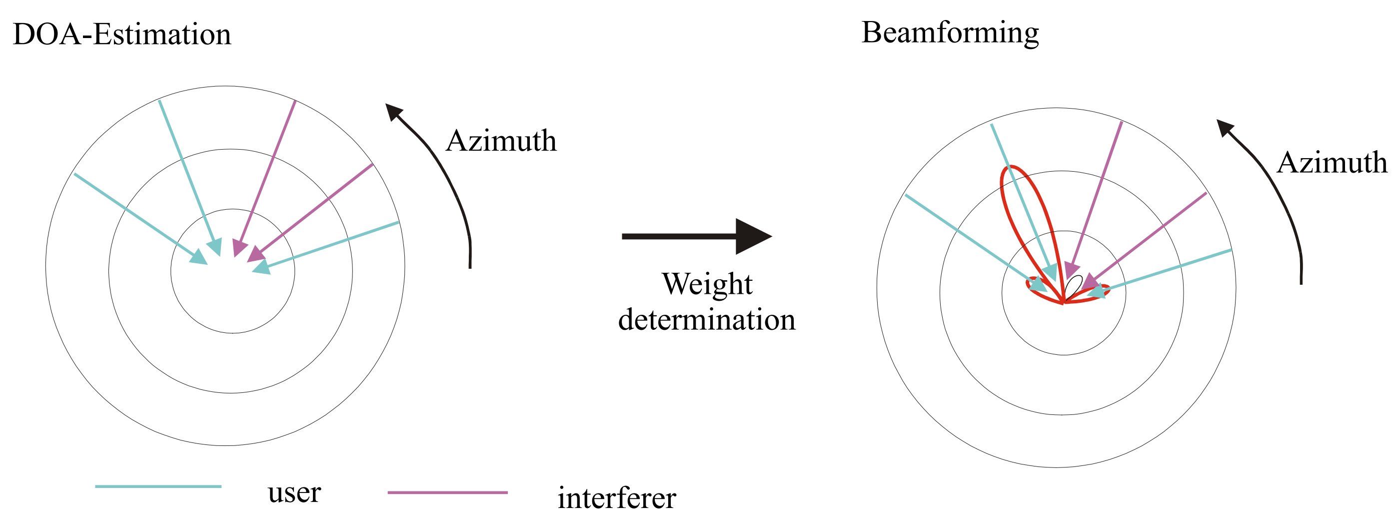
Basic idea:

- Choose antenna weights so that deviation of array output from transmit signal is minimized
- Needs training sequence



Spatial reference (SR) algorithms

- Determine DOAs, then do beamforming
- A priori information for DOA estimation: array structure
- Algorithms for DOA estimation:
  - Fourier analysis
  - Spectrum-based estimators
  - Parametric estimators



SR classification

DOA estimation

- SR algorithms
  - parametric
    - subspace-based methods
      - ESPRIT
    - maximum likelihood
      - SAGE
  - spectral-based
    - subspace-based methods
      - MUSIC
    - beamforming
      - MVM

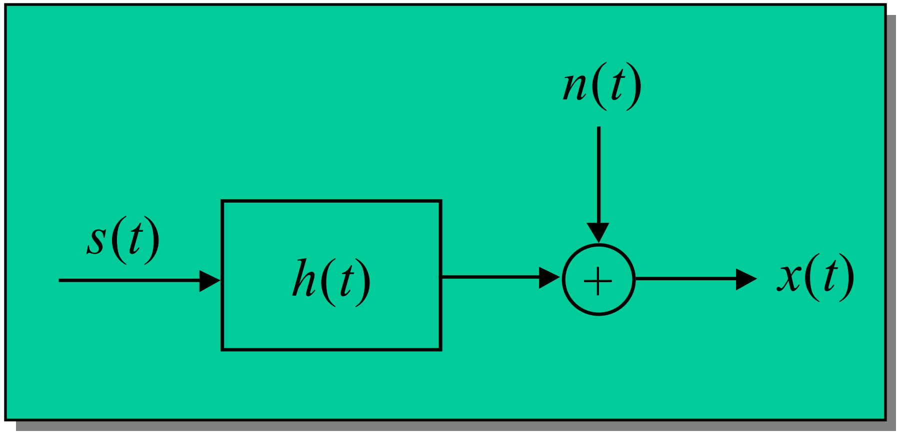
Blind Algorithms - Definition (1)

- Blind Estimation =

  Identification of the system parameters $h(t)$ or input $s(t)$ using only the output information

  (i.e. without access to the input sequence).

- Applications:
  - equalisation
  - speech processing
  - image processing
  - etc.



Blind Algorithms - Definition (2)

- Blind:

  no training sequences
  no known array properties (DOAs)

- but

  structural signal properties

- Semi-blind:

  both
  structural signal properties
  and
  known bit fields

Blind Algorithms - Identification problem

Array output data

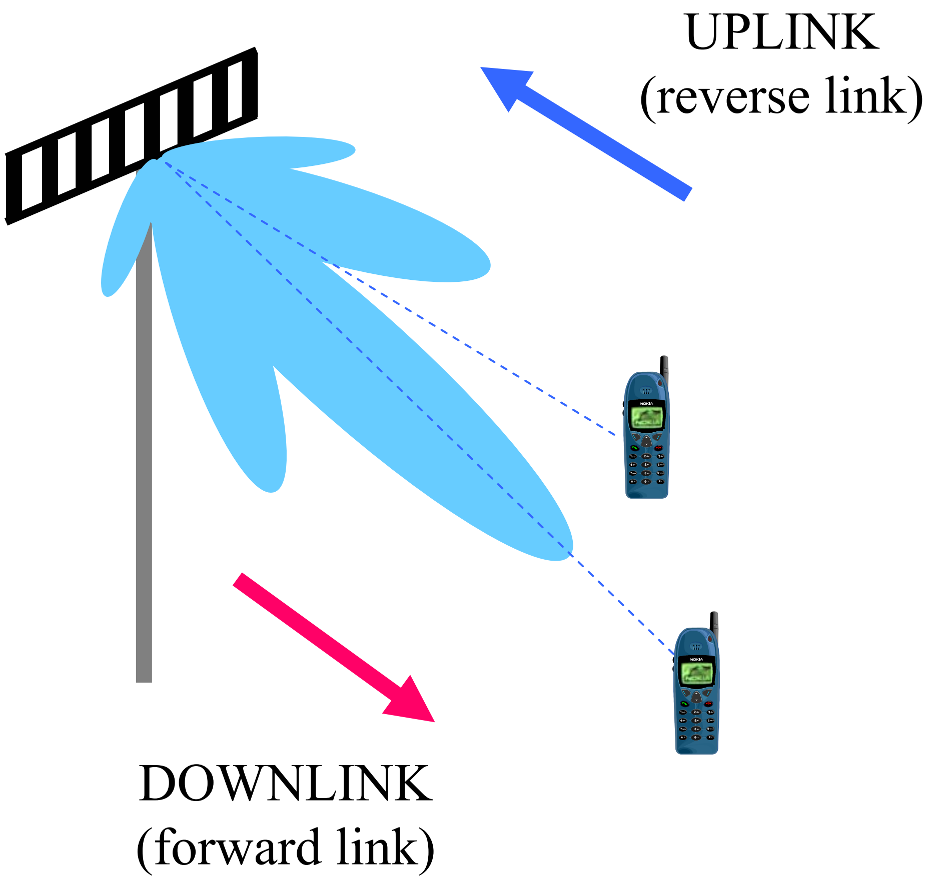
Unknown transmitted data

$$\boldsymbol{X} = \boldsymbol{H}\boldsymbol{S}$$

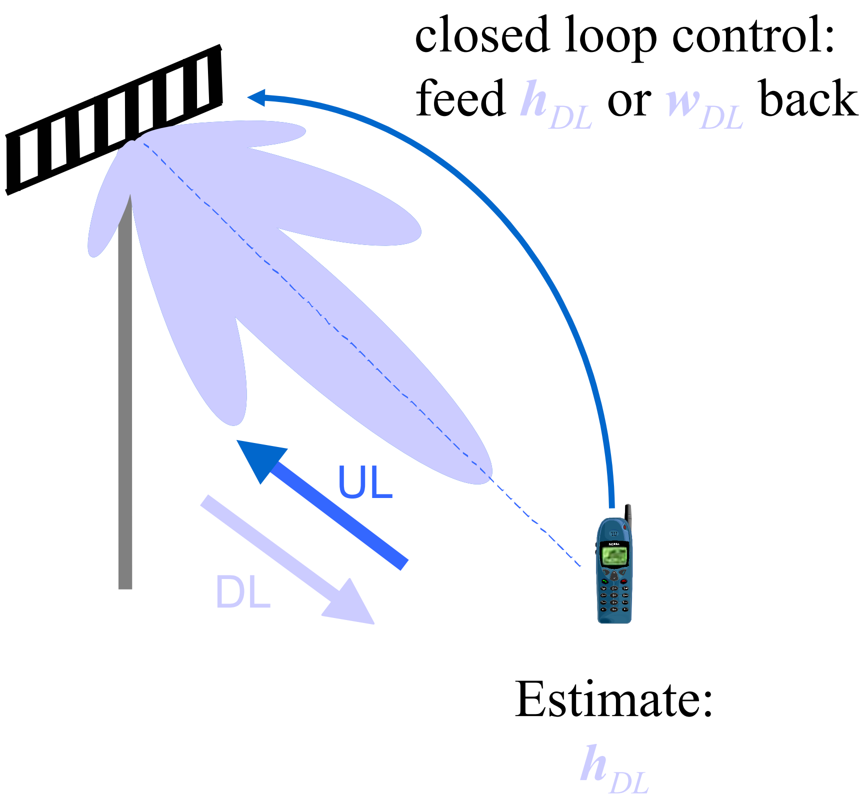
Unknown channel impulse response

- SIMO, MIMO problem (single or multiple inputs)
- separate or joint estimation of H and S :
  - space(-time) filter or space-time detector

Why downlink processing ?



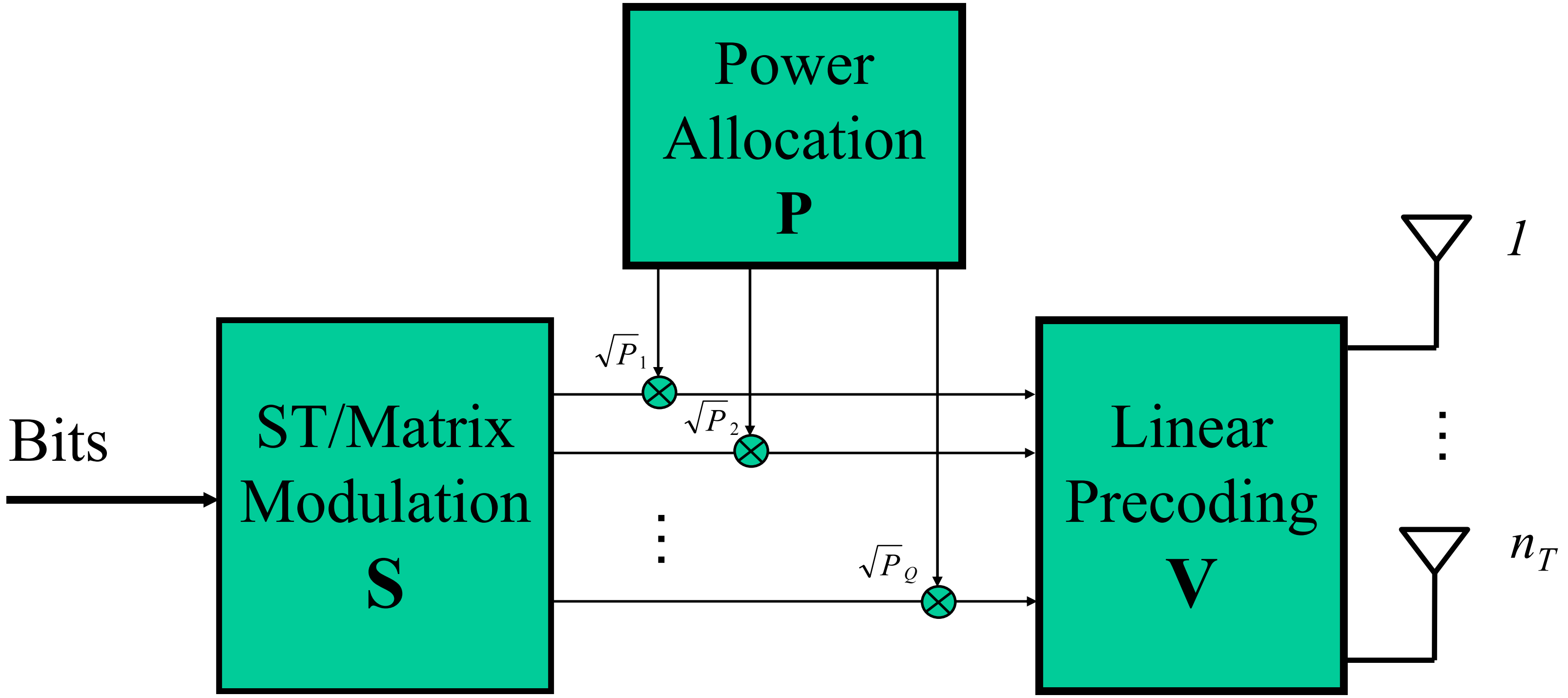
Mobile Feedback based Beamforming



- MS estimates h_{DL}
- Feedback of DL channel parameters (h_{DL} or w_{DL})

MIMO SYSTMES

MIMO Transmission – Generic Structure



MIMO Transmission – System Model

- Basic system model

$$\underbrace{\mathbf{Y}}_{n_R \times T} = \underbrace{\mathbf{H}}_{n_R \times n_T} \underbrace{\mathbf{X}}_{n_T \times T} + \underbrace{\mathbf{N}}_{n_R \times T} = \underbrace{\mathbf{H}}_{n_R \times n_T} \underbrace{\mathbf{V}}_{n_T \times Q} \underbrace{\mathbf{P}}_{Q \times Q} \underbrace{\mathbf{S}}_{Q \times T} + \underbrace{\mathbf{N}}_{n_R \times T}$$

  - S… ST modulation matrix containing the transmitted signals of Q transmission streams during T symbol periods
  - P… power allocation matrix P=diag($P_1^{1/2}$,...,$P_Q^{1/2}$) for Q streams
  - V… linear precoding matrix (e.g. for beamforming purpose)
  - H… MIMO channel matrix ($n_R \times n_T$) – assumed to be constant during T symbol periods
  - Y… received signal from n_R antennas during T symbols
  - X… TX signal from n_T antennas during T symbol periods
  - N… receiver noise at n_R antennas during T symbols

MIMO SYSTMES
WITH CSI AT TRANSMITTER

Decomposing the instantaneous channel

- Deterministic instantaneous channel can be decomposed via SVD:

$$\boldsymbol{H} = \boldsymbol{U}\boldsymbol{\Lambda}\boldsymbol{V}^{\mathrm{H}}$$

$$= \sum_{i=1}^{\min(n_{\mathrm{R}},n_{\mathrm{T}})} \lambda_i \boldsymbol{u}_i \boldsymbol{v}_i^{\mathrm{H}}$$

- Equivalent to $\min(n_{\mathrm{R}},n_{\mathrm{T}})$ independent parallel channels with powers λ_i.

Transmitting on eigenmodes

- Transmit precoding is matched to Tx eigenmodes:

$$\boldsymbol{y} = \boldsymbol{H} \underbrace{\boldsymbol{V}}_{precoding} \underbrace{\boldsymbol{P}}_{power} \underbrace{\boldsymbol{s}}_{signal}$$

- The modulation matrix is just a serial to parallel conversion

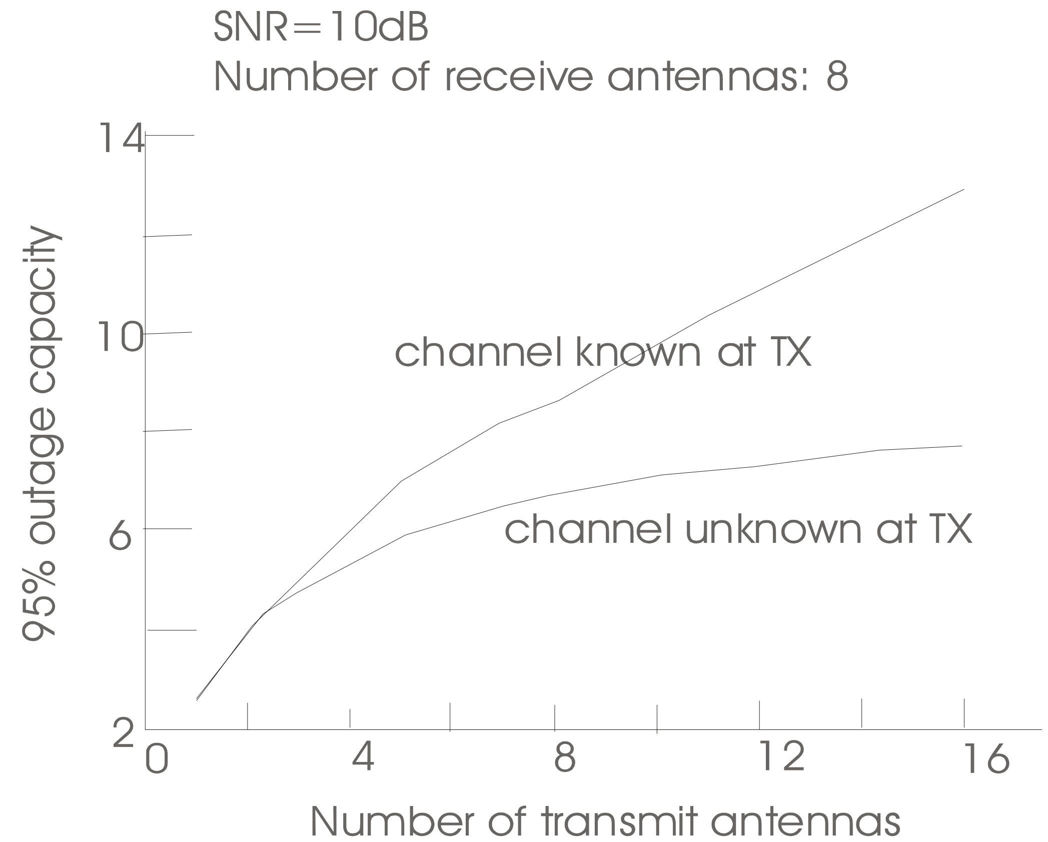
$$\boldsymbol{s} = \begin{bmatrix} s_1 \\ s_2 \\ \vdots \\ s_{\min(n_R, n_T)} \end{bmatrix}$$

Waterfilling

- Capacity formula for unequal power distribution

$$C = \log_2\left(\det\left[\boldsymbol{I}_{n_R} + \frac{\gamma}{n_T} \boldsymbol{HPH}^H \right]\right) bits/s/Hz$$

Performance



After [Burr 2001]

Many TX antennas: for unknown channel, TX power “wasted”

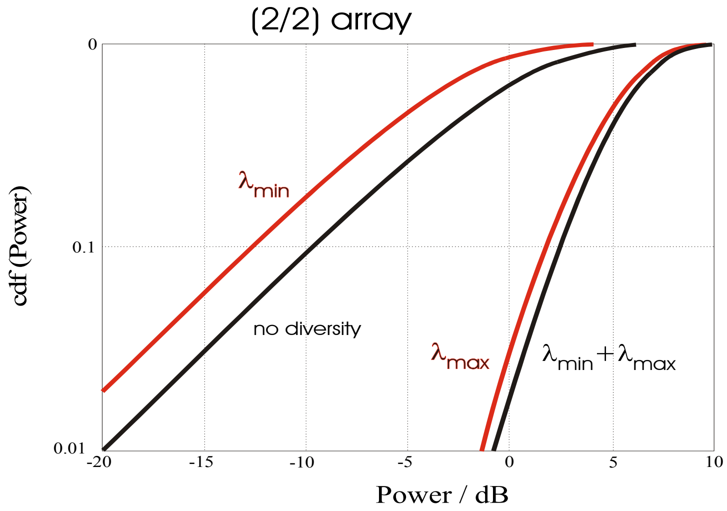
Diversity gain

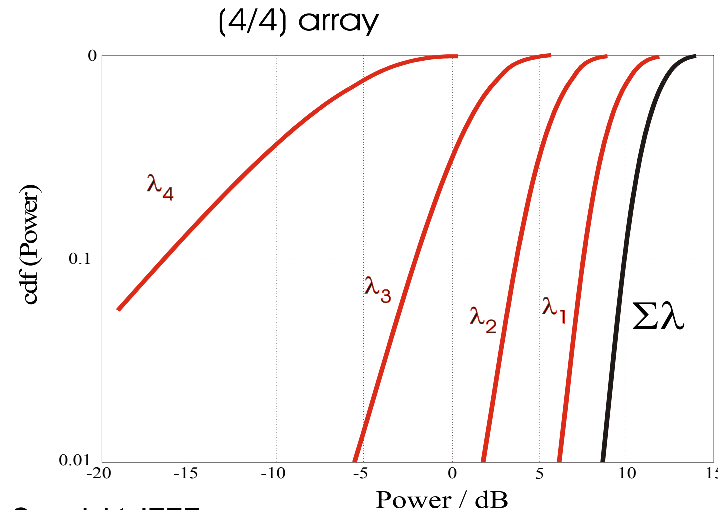
- Write channel matrix as

$$\boldsymbol{H} = \boldsymbol{U}\boldsymbol{\Lambda}\boldsymbol{V}^H$$

- Excite channel with V_i, receive with U_i^{H}
- Received power is λ_i^2
- Full benefit only for uncorrelated contributions
    - $n_T \cdot n_R$ diversity
- But: beamforming gain limited

  Upper bound: $(n_T^{\,1/2} + n_R^{\,1/2})^2$





MIMO SYSTMES
WITHOUT CSI AT THE TRANSMITTER

Capacity formula

- Instantaneous channel characterized by matrix *H*
- Shannon's formula (for two-dimensional symbols):

$$C = \log_2(1+\gamma\,|\,H\,|^2)\,bits\,/\,s\,/\,Hz$$

- Foschini's formula:

$$C = \log_2\left(\det\left[\boldsymbol{I}_{n_R} + \frac{\gamma}{n_T}\boldsymbol{H}\boldsymbol{H}^H\right]\right) bits\,/\,s\,/\,Hz$$

Capacity for fading channel (I)

- Rayleigh fading channel.
- Capacity becomes random variable.
- Channel not known at transmitter.
- χ^2_{2k}...random variable; chi-square with $2k$ degrees of freedom

- Transmit diversity

$$C = \log_2(1+(\gamma/n)\cdot\chi_{2n}^2)$$

- Receive diversity

$$C = \log_2(1+\gamma\cdot\chi_{2n}^2)$$

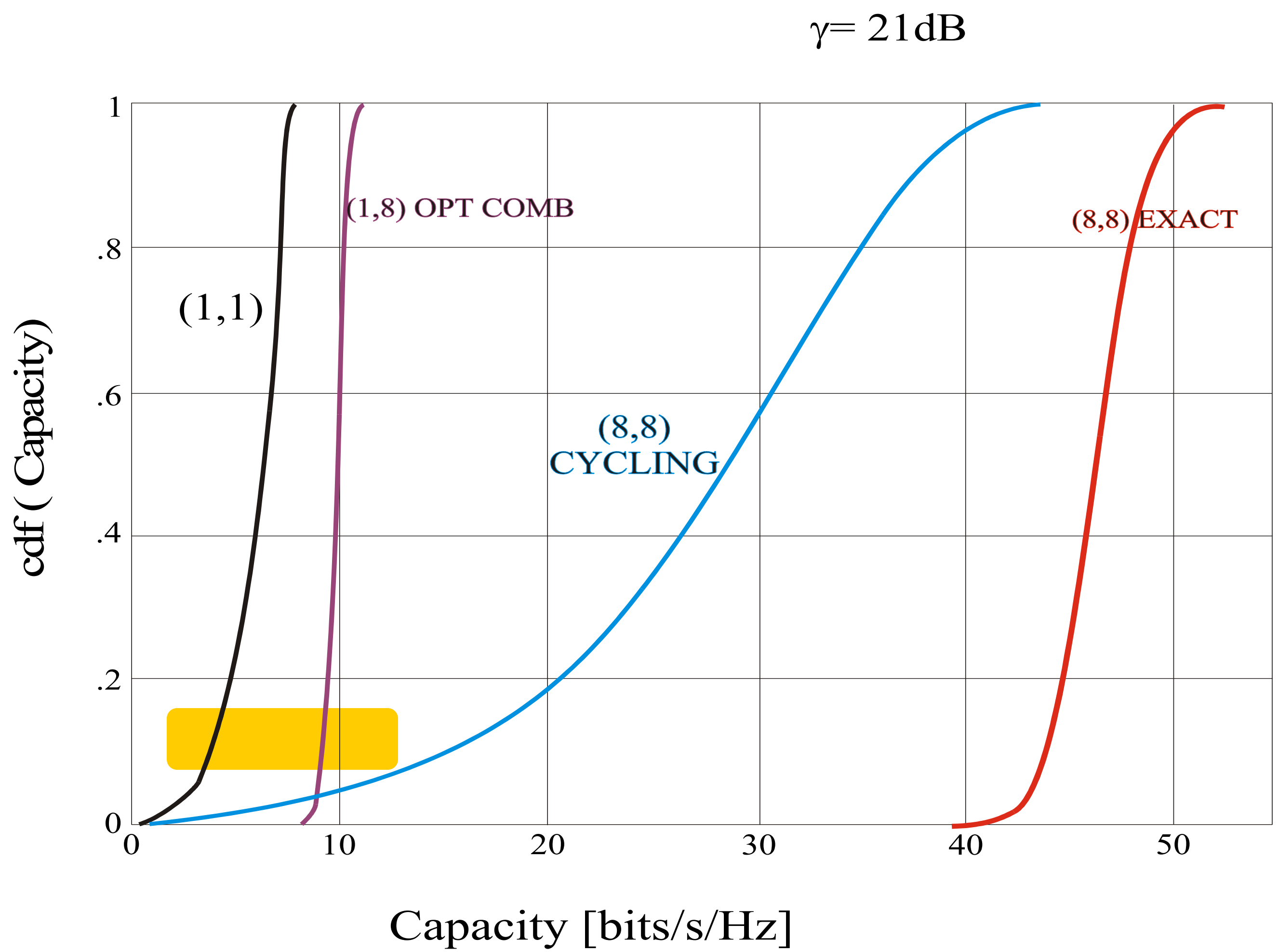
- Comb. transmit/receive diversity: linear with n for fixed outage

$$C > \sum_{k=1}^{n} \log_2\left[1+(\gamma/n)\chi_{2k}^{\ 2}\right]$$

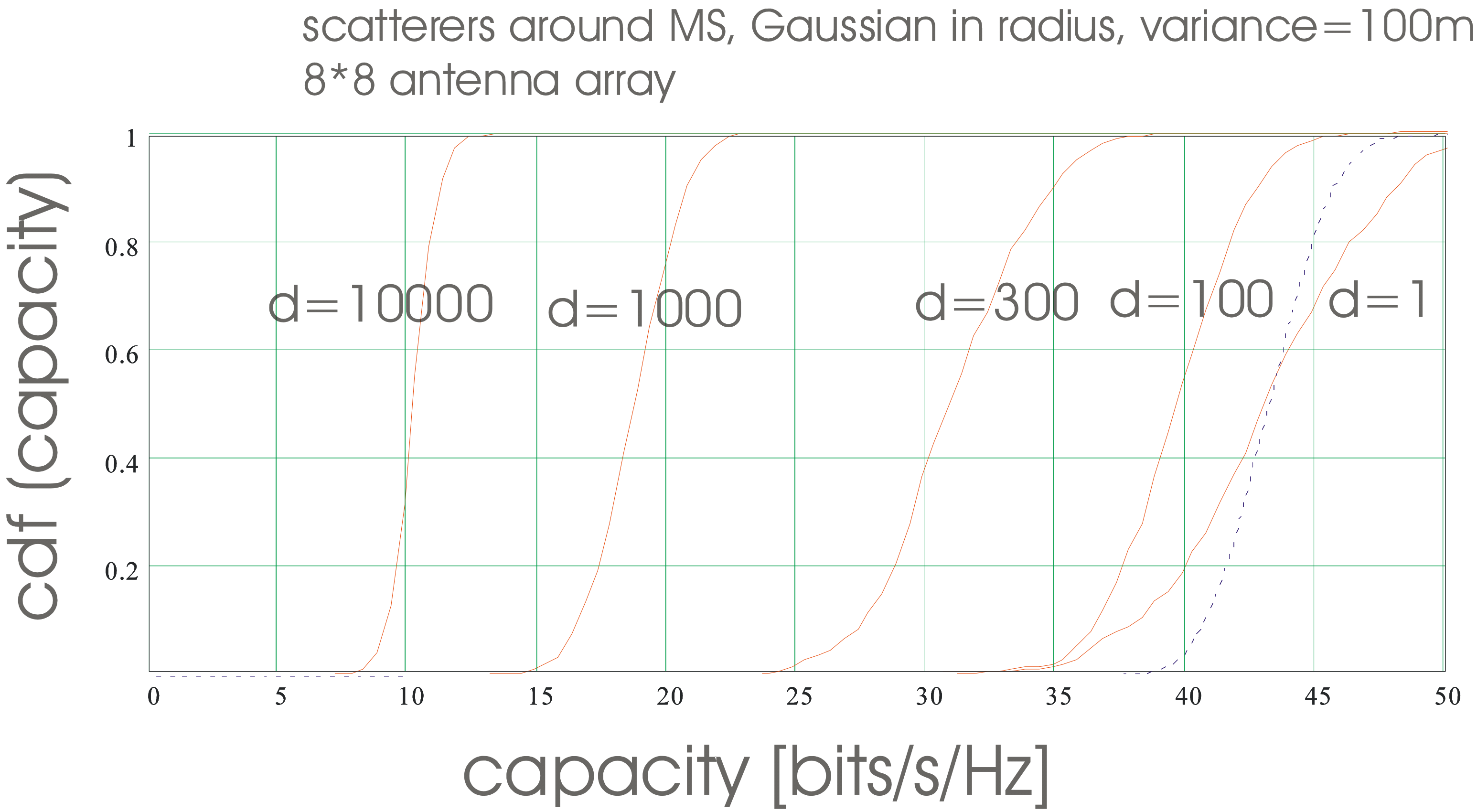
- Spatial cycling

$$C = \frac{1}{n}\sum_{k=1}^{n} \log_2\left[1+(\gamma)\chi_{2kn}^{\ 2}\right]$$

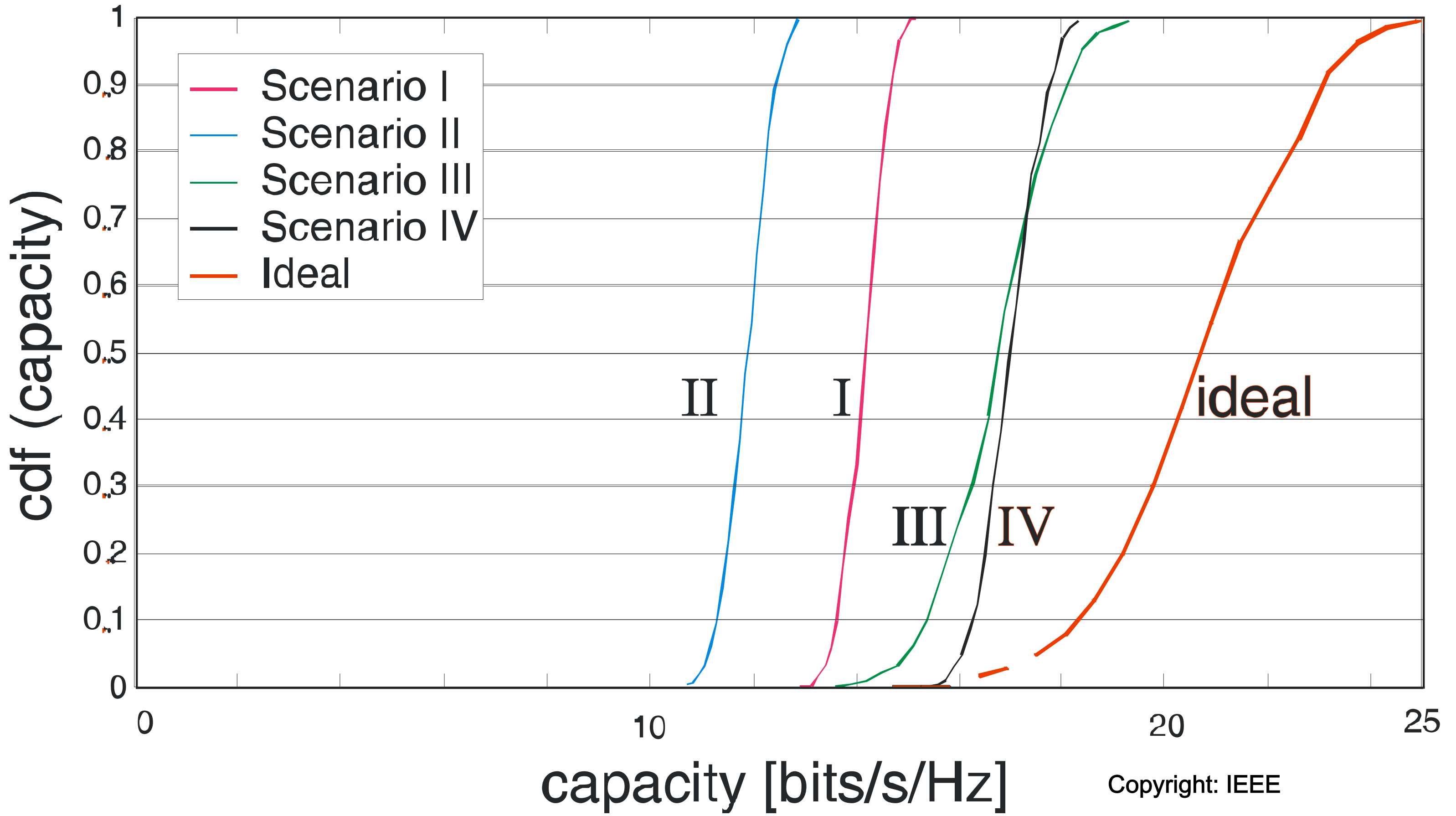
Capacity for fading channel (II)



Capacity with correlation



Measured capacities (LOS and NLOS)



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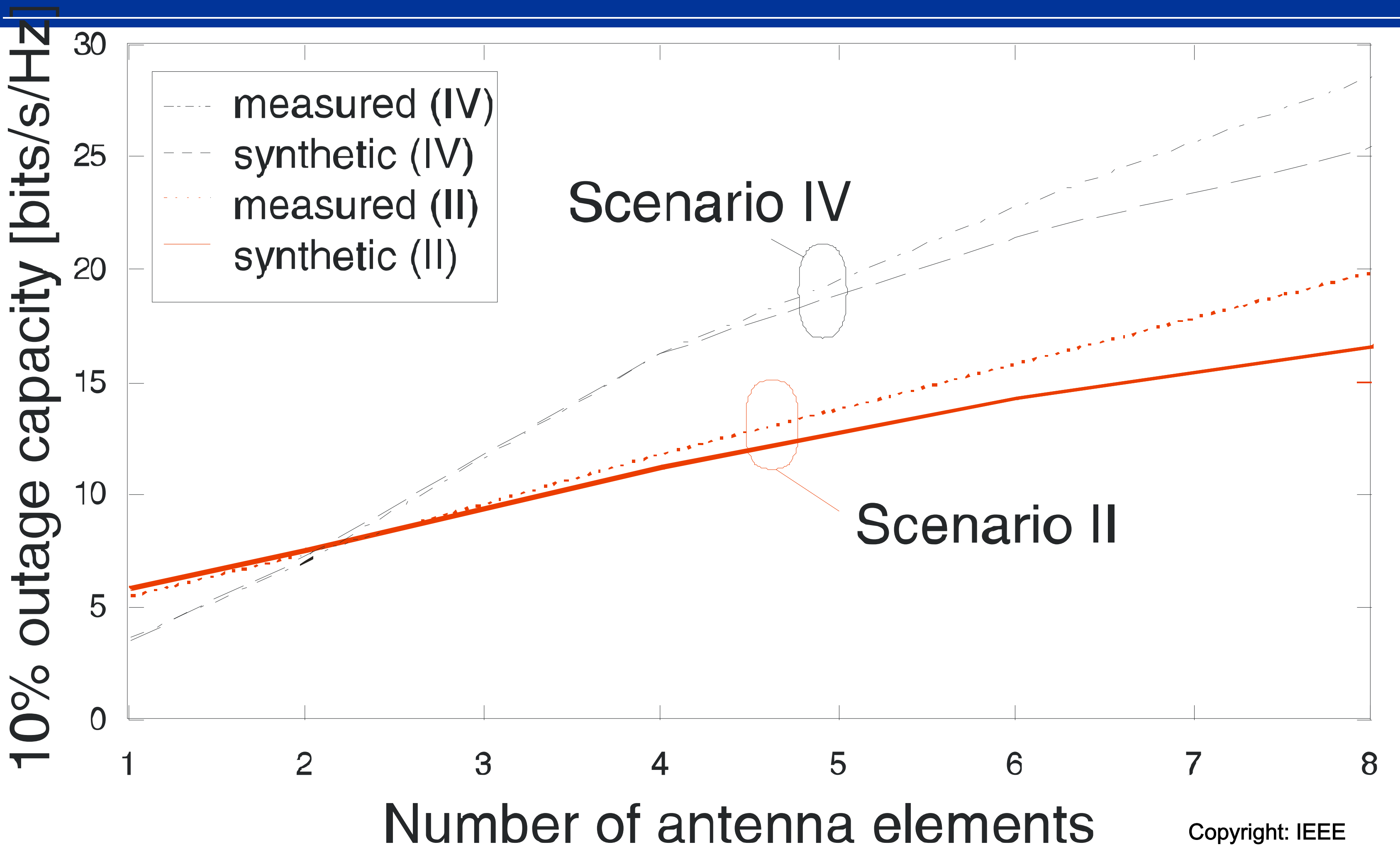
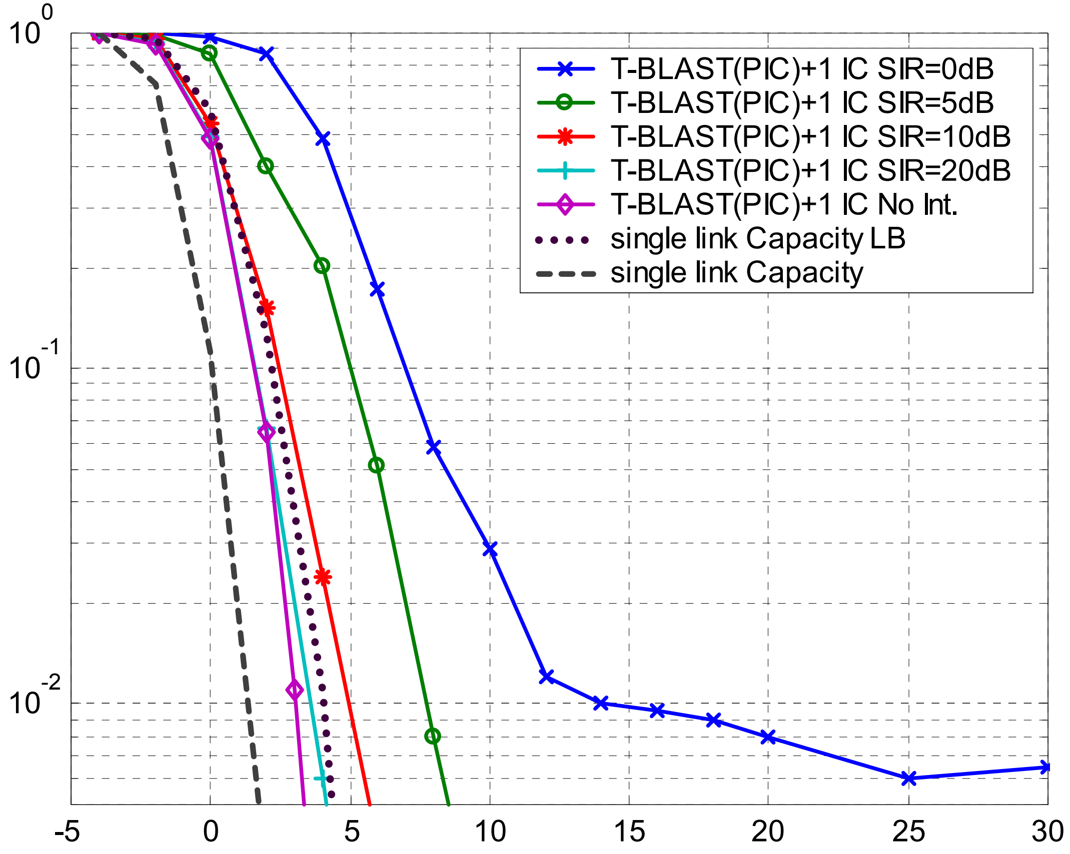
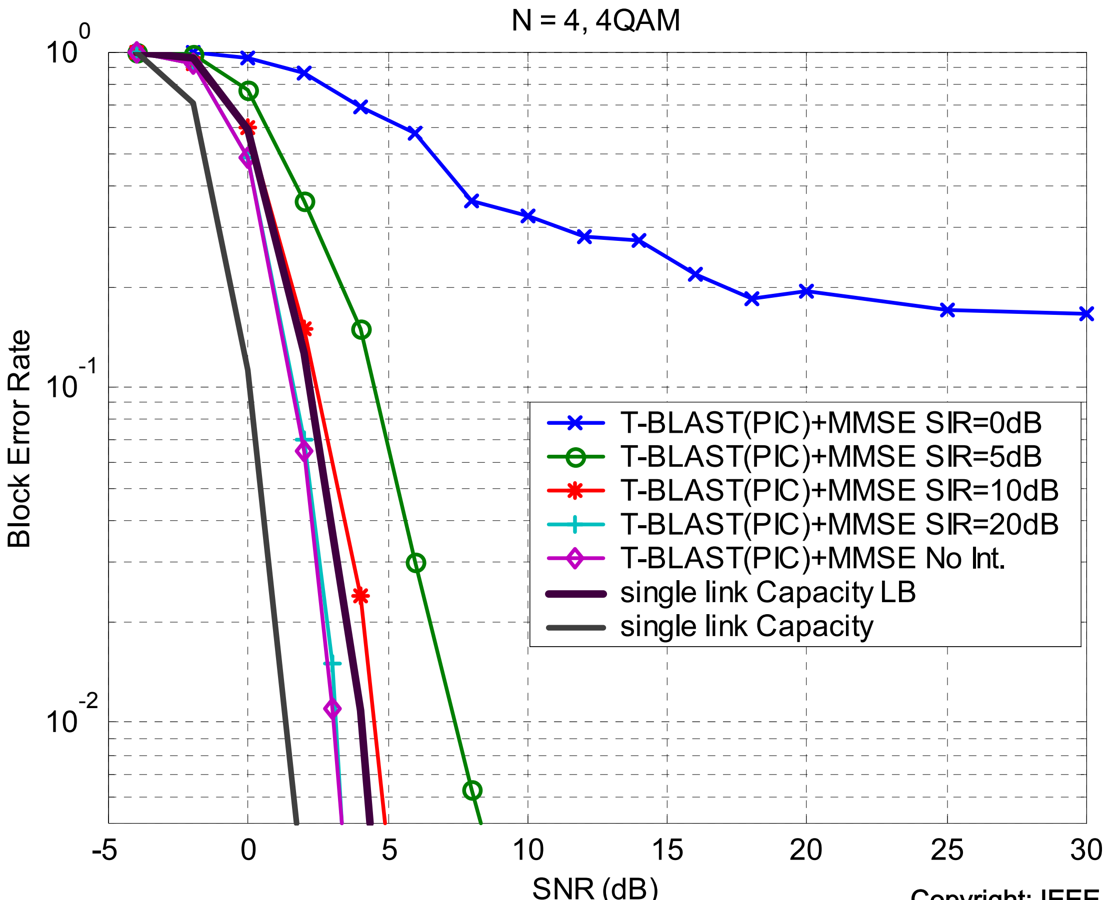
Limited number of scatterers



Performance when one interferer dominates

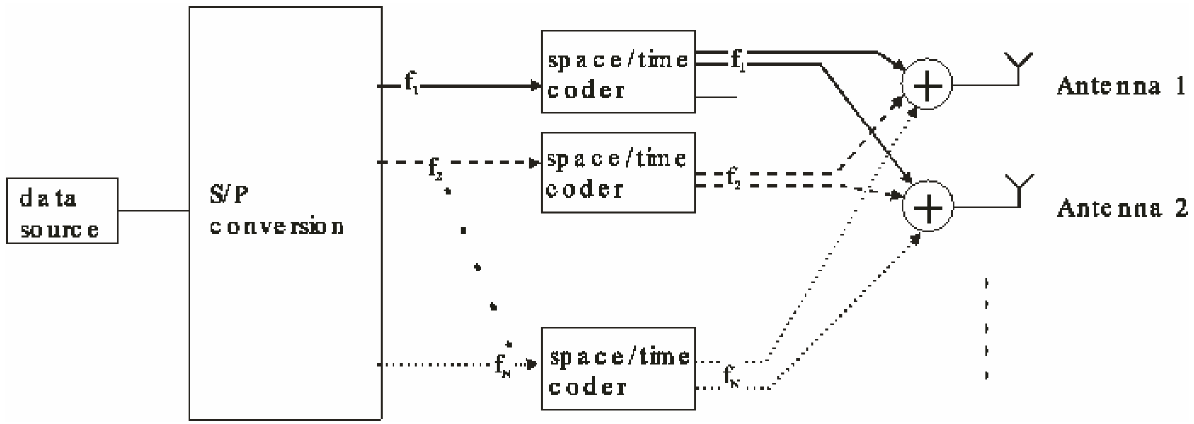


Performance when two interferers dominate

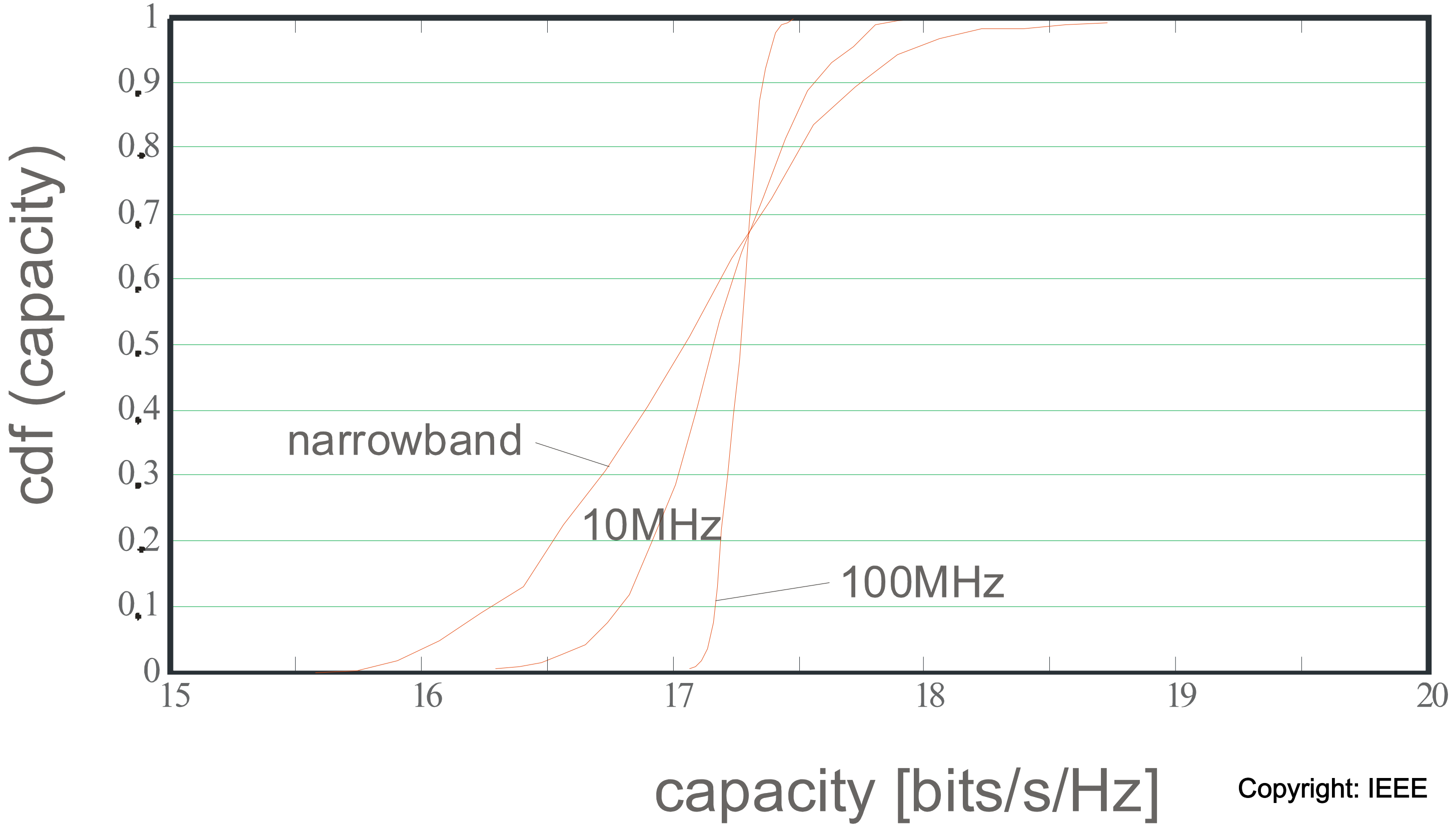


Frequency-selective environments

- Channel gives more diversity
- Equalizers: very complicated
- OFDM:
  - Subdivision into many frequency channels,
  - Flat-fading MIMO system on each tone
  - Efficient signal processing by using FFT
  - But: coding across tones required to exploit frequency diversity



Capacity in frequency-selective channels



Frequency diversity leads to smaller capacity fluctuations

BLAST TRANSCEIVERS

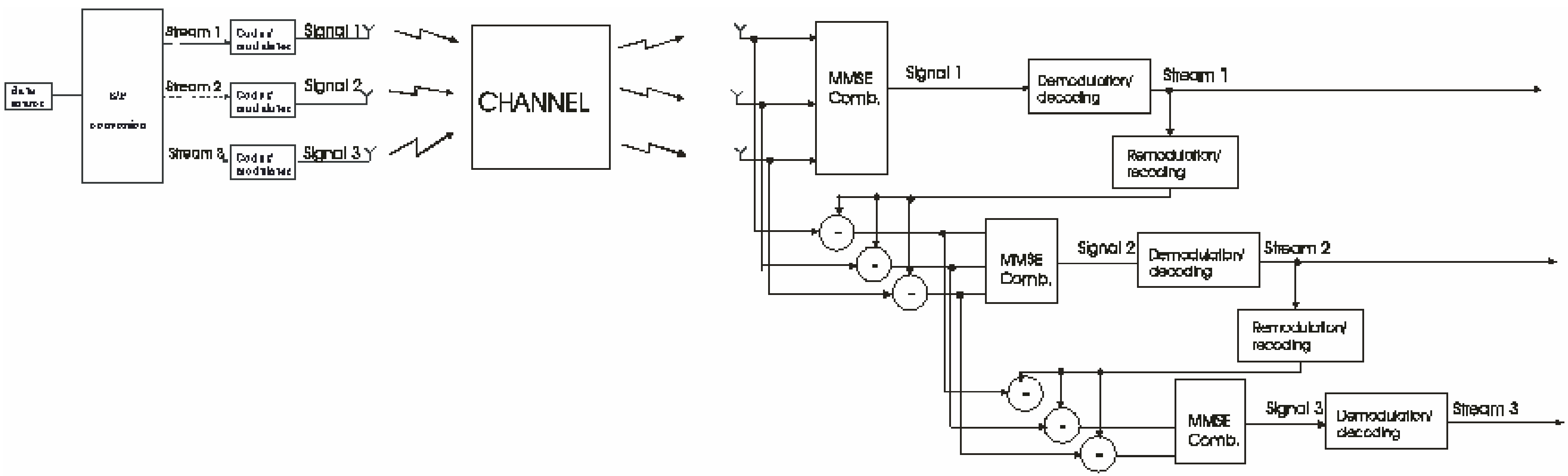
Spatial Multiplexing (H-BLAST)

$$\mathbf{S}_{VBLAST} = \begin{bmatrix} \boxed{s_1 \quad s_5 \quad s_9 \quad s_{13}} \\ \boxed{s_2 \quad s_6 \quad s_{10} \quad s_{14}} \\ \boxed{s_3 \quad s_7 \quad s_{11} \quad s_{15}} \\ \boxed{s_4 \quad s_8 \quad s_{12} \quad s_{16}} \end{bmatrix}$$

- Outer coding over T symbols (block length)
- Outer coding is independent for all streams
  → no spatial diversity
- No coding over the streams – is sometimes also called "vector modulation"

$$\mathbf{S} = \begin{bmatrix} s_1 & s_2 & s_3 & s_4 \end{bmatrix}^T$$

H-BLAST - principle



Spatial Multiplexing (D-BLAST)

- Diagonal BLAST
- Modulation matrix for the example $n_R = n_T = Q = T = 4$

$$\mathbf{S}_{VBLAST} = \begin{bmatrix} s_1 & s_8 & s_{11} & s_{14} \\ s_2 & s_5 & s_{12} & s_{15} \\ s_3 & s_6 & s_9 & s_{16} \\ s_4 & s_7 & s_{10} & s_{13} \end{bmatrix} \qquad \boldsymbol{Y} = \boldsymbol{H}\,\boldsymbol{V}\,\boldsymbol{P}\,\boldsymbol{S}$$

- Data streams are cycled through antennas
- Achieves spatial multiplexing gain (rate=4) and spatial diversity

SPACE-TIME CODING

Design rules for ST-coding

- Probability of picking wrong code symbol with ST-codes:

$$\left(\prod_{i=1}^{r}\lambda_i\right)^{-n_R}\left[\frac{4N_0}{E_s}\right]^{rn_R}$$

  r....rank of *A*

  λ...eigenvalues of *A*

$$A_{ik}=\sum_t\left(c_i(t)-c_i'(t)\right)\left(c_k(t)-c_k'(t)\right)$$

- Design rule:
  - for achieving full diversity effect, *A* must have full rank diversity order not decreased by frequency selectivity
  - for optimizing coding gain (with full diversity),

$$\underbrace{\min}_{c_i,c_k'}\left[\det(A)\right]$$

    must be maximized

Space Time Block Codes

- Example: Alamouti code (n_T=Q=T=2)

$$\mathbf{S}_{Alamouti} = \begin{bmatrix} s_1 & -s_2 \\ s_2^* & s_1^* \end{bmatrix} \qquad \boldsymbol{Y} = \boldsymbol{H}\,\boldsymbol{V}\,\boldsymbol{P}\,\boldsymbol{S}$$

- Linear reception:

$$\hat{s}_1 = h_1^* r_1 + h_2 r_2^* + n_1$$

$$\hat{s}_2 = h_2 r_2^* - h_1^* r_1 + n_2$$

- Two symbols are transmitted during two symbol periods (rate 1 – no spatial multiplexing)

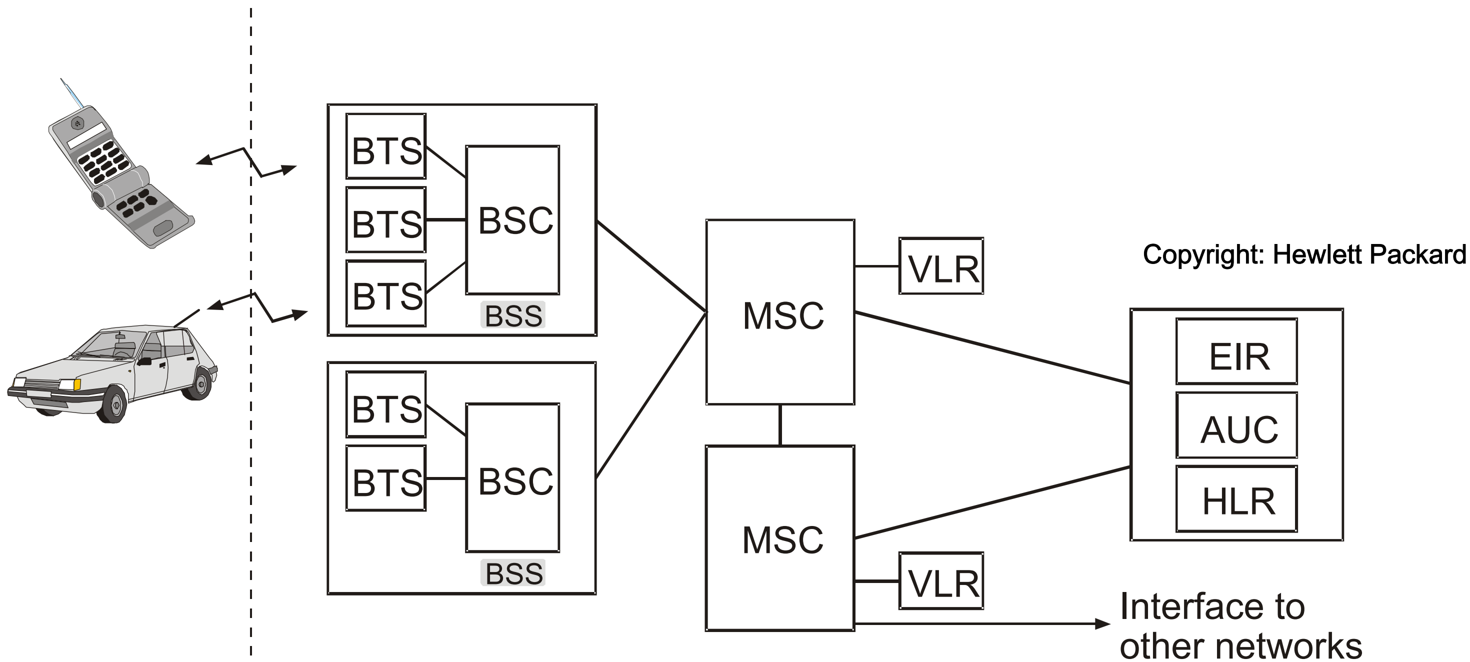
- Coding over the streams - achieves 2nd order TX diversity

- Reaches capacity only for n_R=1

Chapter 21

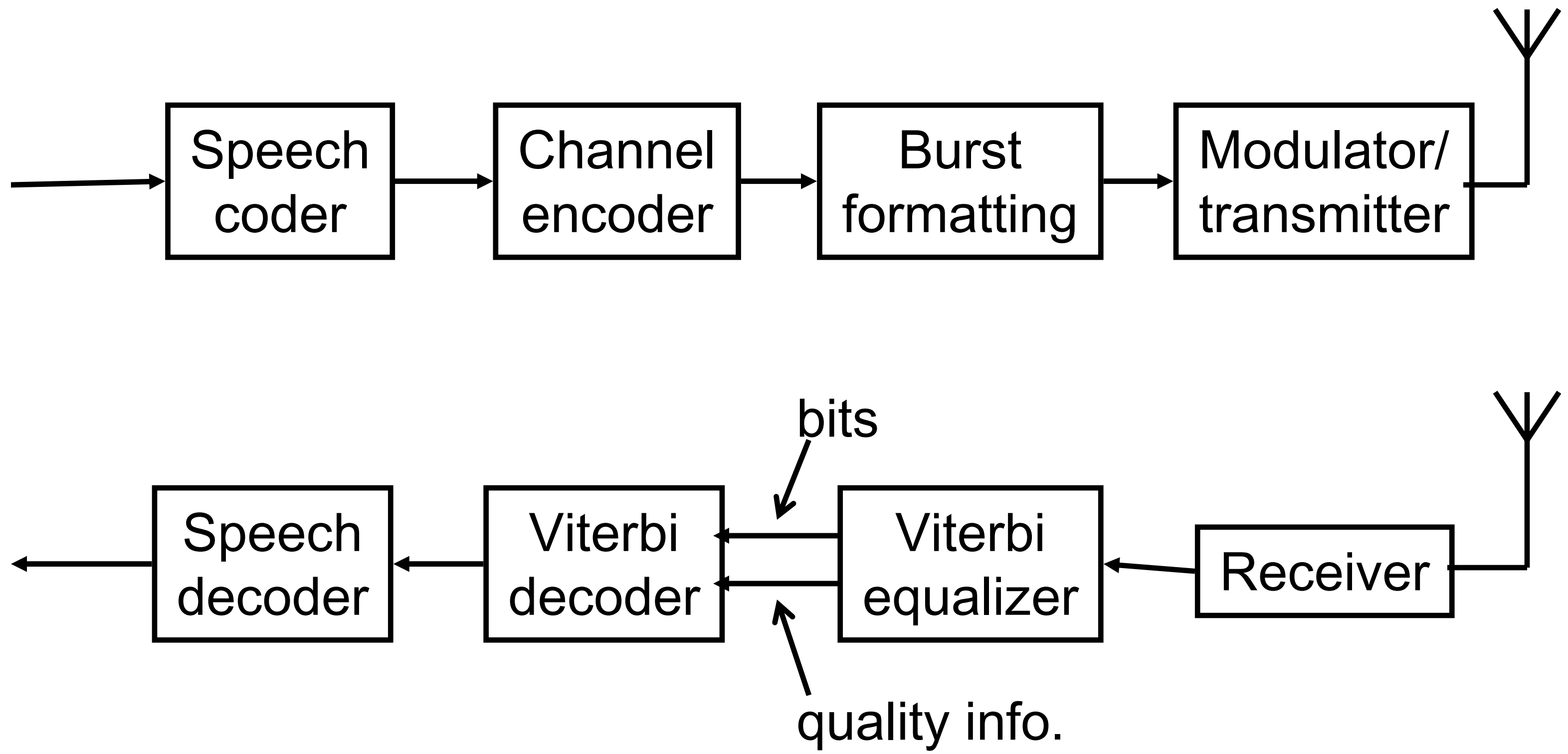
GSM

Simplified system overview



BTS	Base Transceiver Station	VLR	Visitor Location Register
BSC	Base Station Controller	EIR	Equipment Identity Register
BSS	Base Station Sub-system	AUC	AUthentication Center
MSC	Mobile Switching Center	HLR	Home Location Register

Simplified block diagram



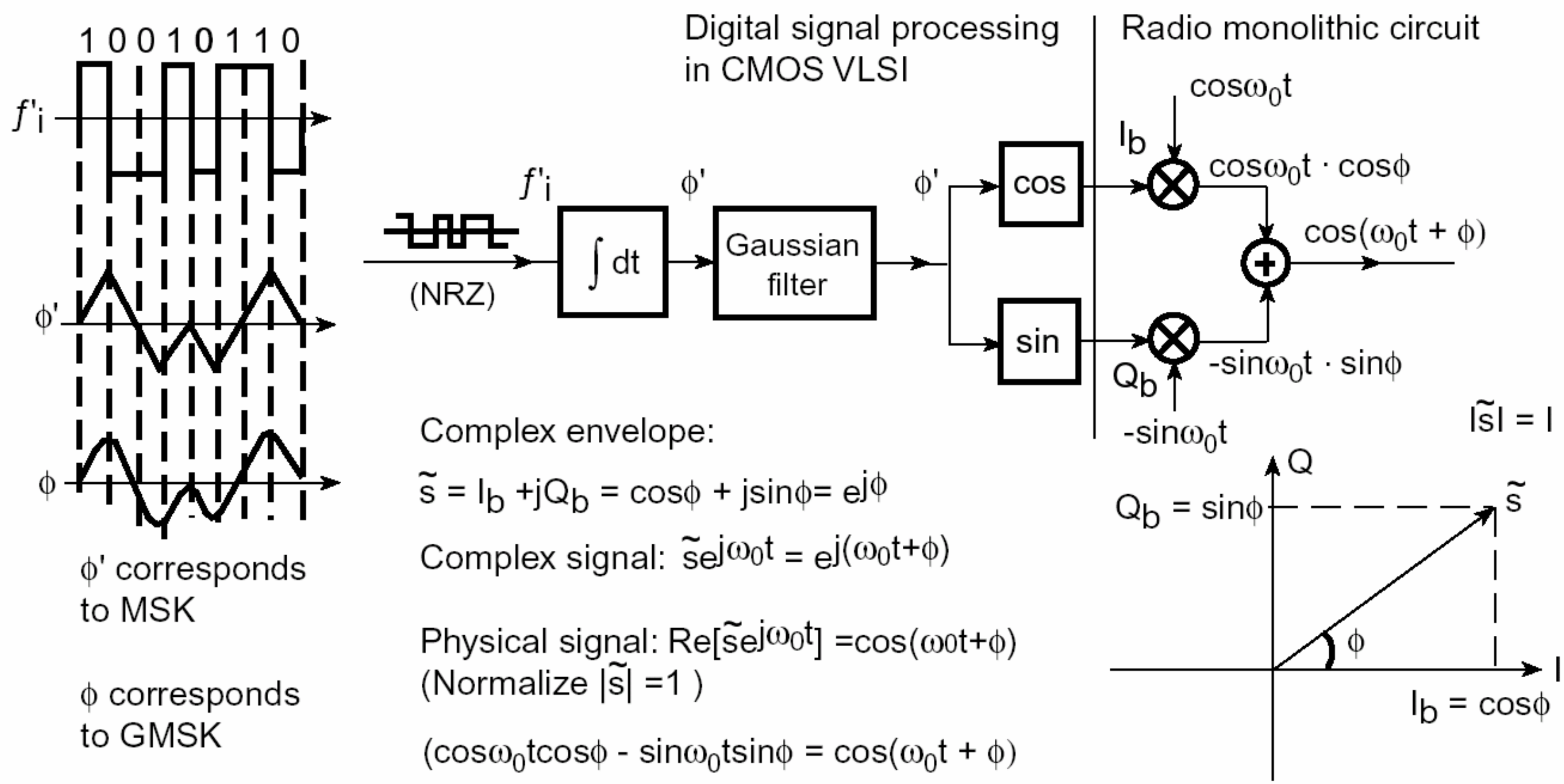
(Encryption not included in figure)

Some specification parameters

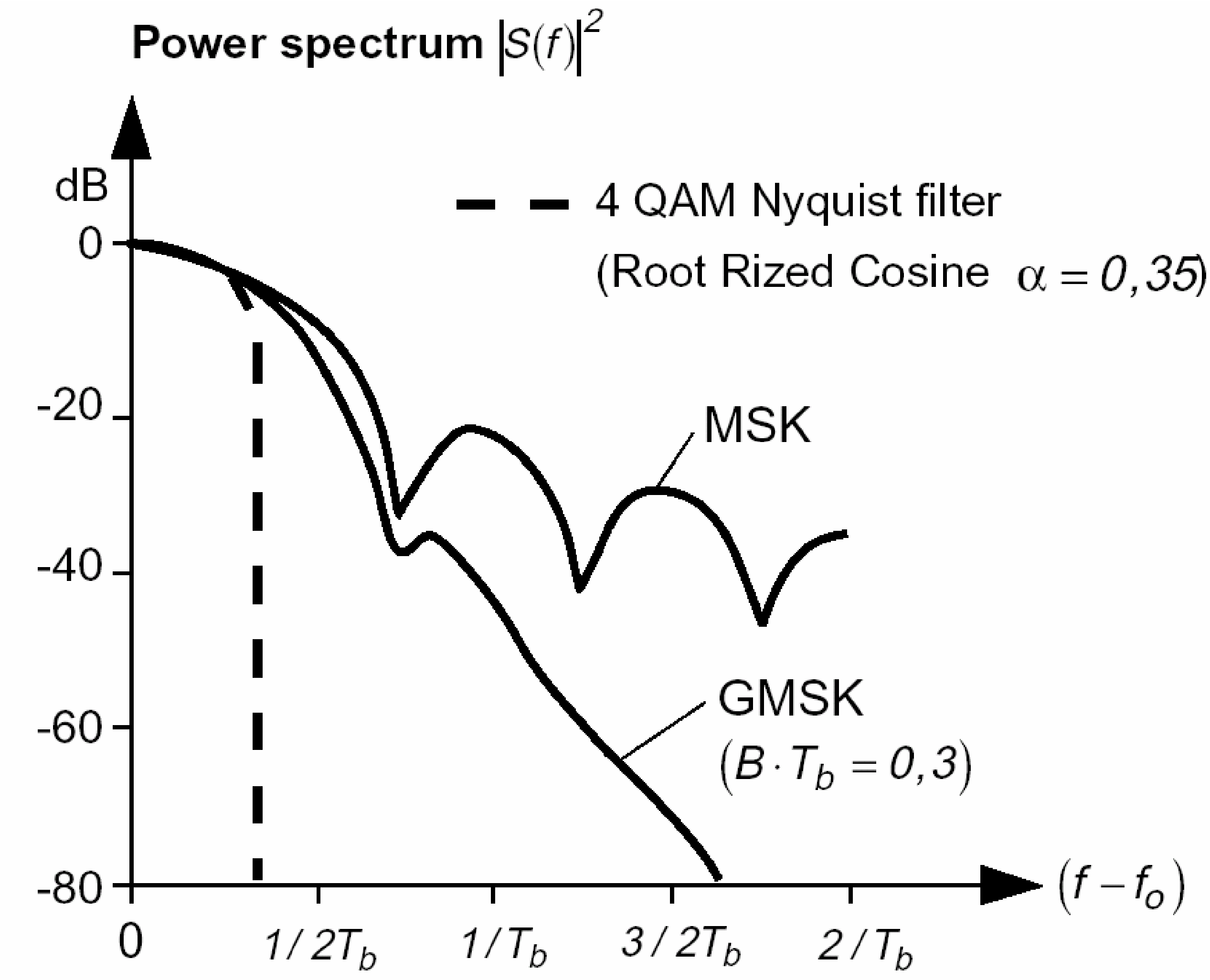
Frequency band:	890 - 915 MHz (uplink)
(frequency duplex)	935 - 960 MHz (downlink)
Channel spacing:	200 kHz
Modulation:	GMSK
System data rate:	271 kb/s
TDMA Frame:	4.6 ms
Time slots:	8 x 0.58 ms
Data rate (full-rate traffic channel):	22 kb/s
Speech coder:	Regular Pulse Exited LPC-LTP 13 kb/s
Diversity:	Channel coding
	Interleaving
	Frequency hopping
	Channel equalization

GMSK modulation

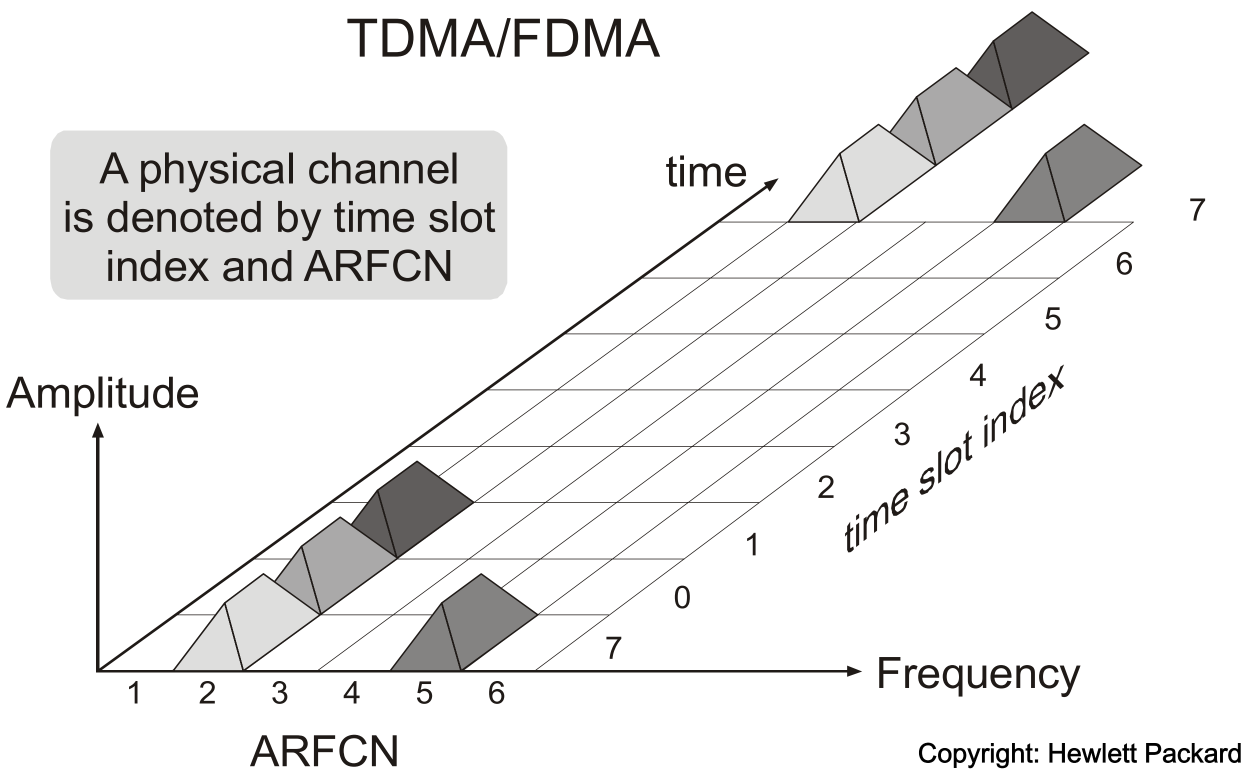
GMSK modulator (GMSK = Gaussian-filtered Minimum Shift Keying)
MSK interpreted as QAM (Complex signal representation)



Power spectrum

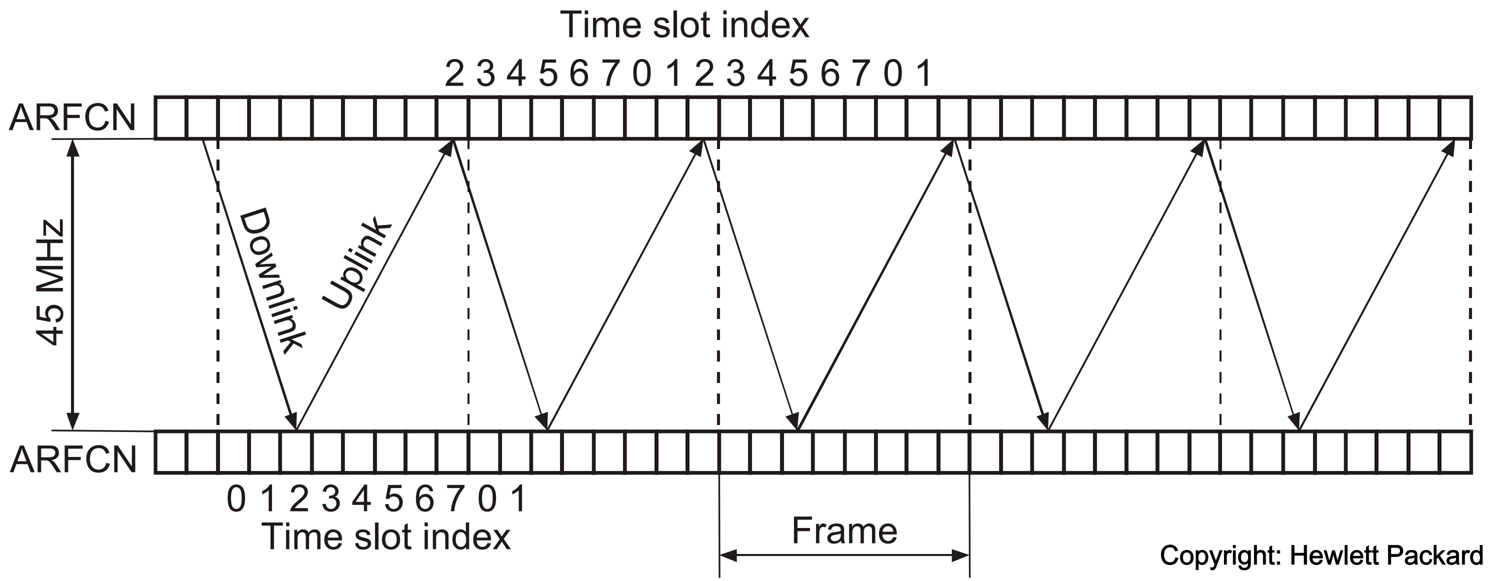


TDMA/FDMA structure



ARFCN Absolute Radio Frequency Channel Number
channels spaced 200 kHz apart

Up/down-link time slots



Some of the time slots

Normal

3 start bits	58 data bits (encrypted)	26 training bits	58 data bits (encrypted)	3 stop bits	8.25 bits guard period

FCCH burst

3 start bits	142 zeros	3 stop bits	8.25 bits guard period

SCH burst

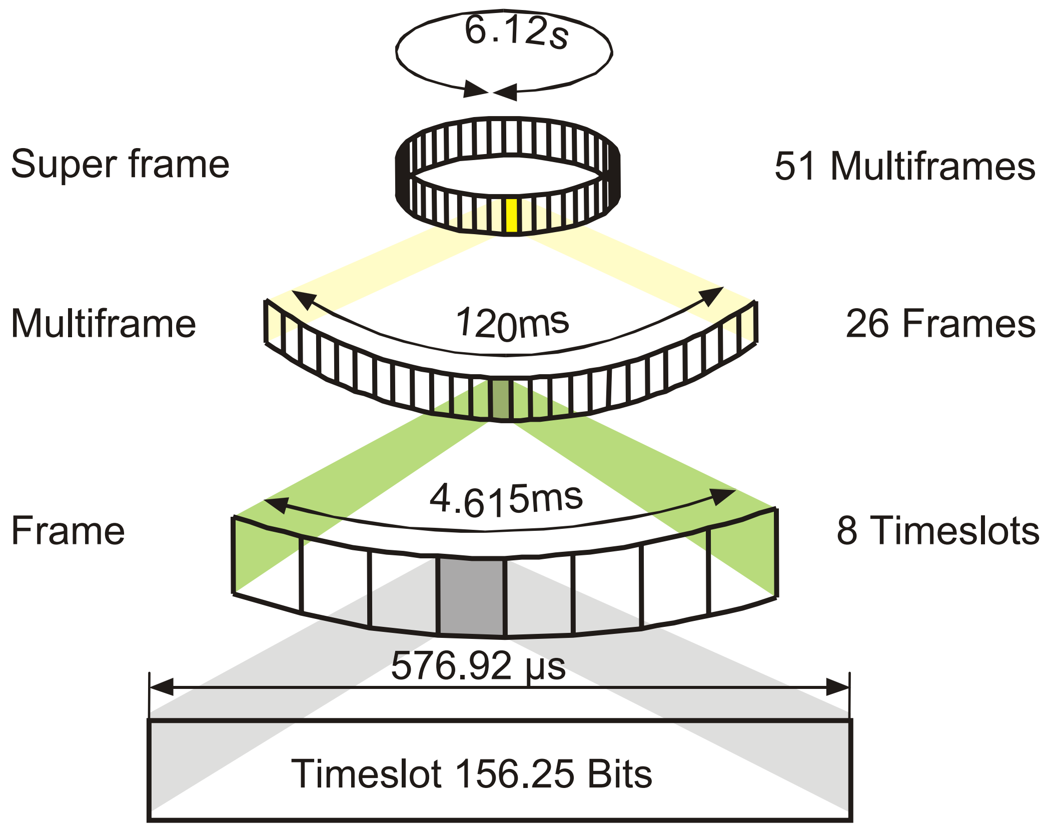
3 start bits	39 data bits (encrypted)	64 training bits	39 data bits (encrypted)	3 stop bits	8.25 bits guard period

RACH burst

8 start bits	41 synchronization bits	36 data bits (encrypted)	3 stop bits	68.25 bits extended guard period

FCCH	Frequency Correction CHannel
SCH	Synchronization CHannel
RACH	Random Access CHannel

Frames and multiframes



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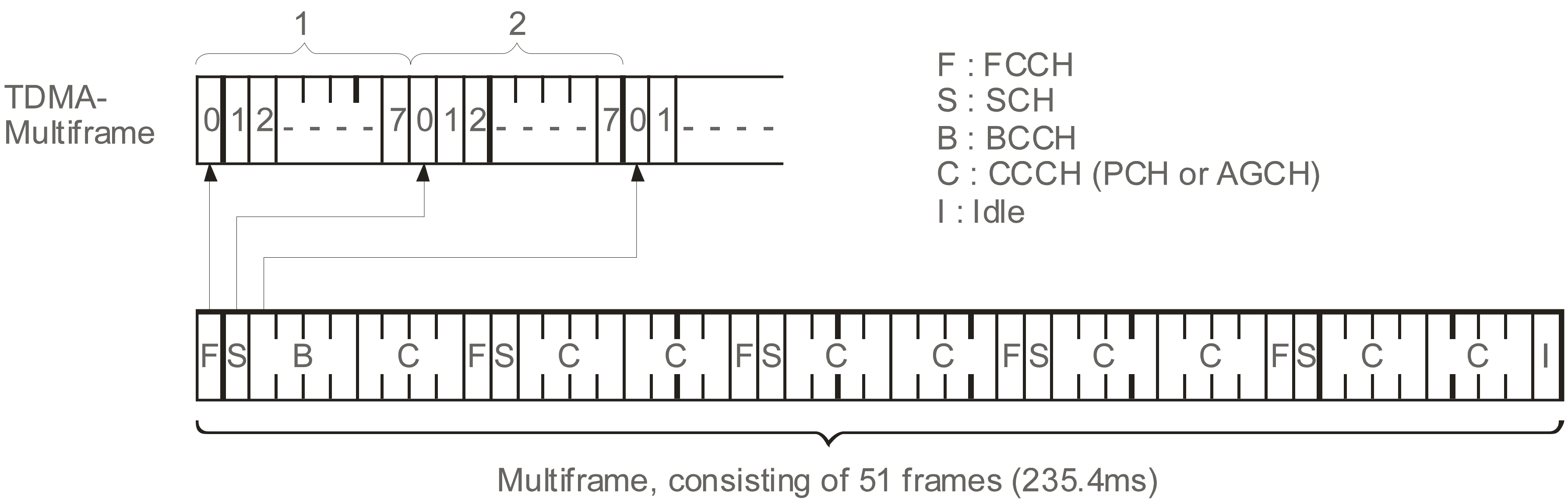
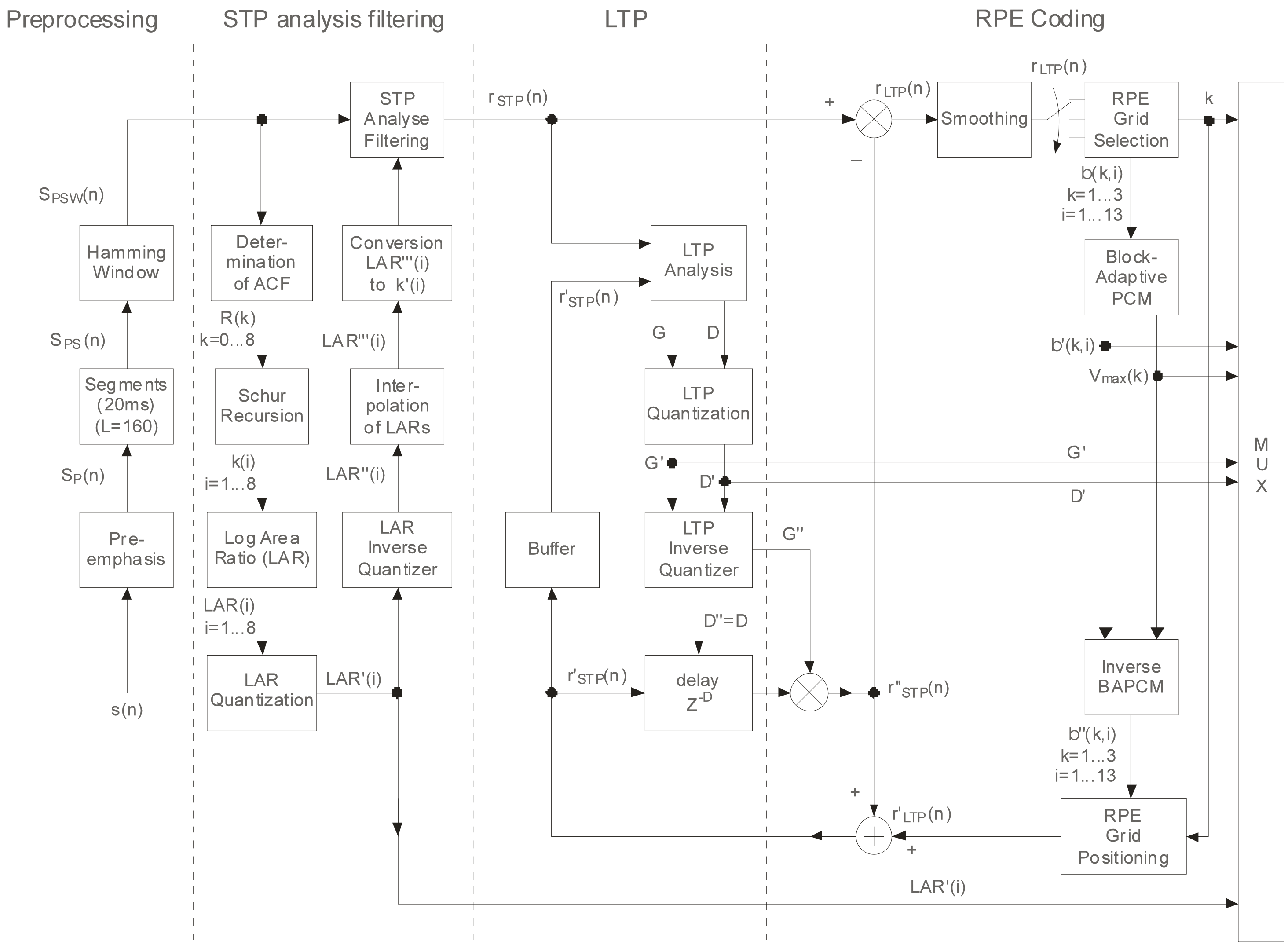
Mapping of logical channels to physical channels

- Logical channels transmitted in different frames/superframes/…



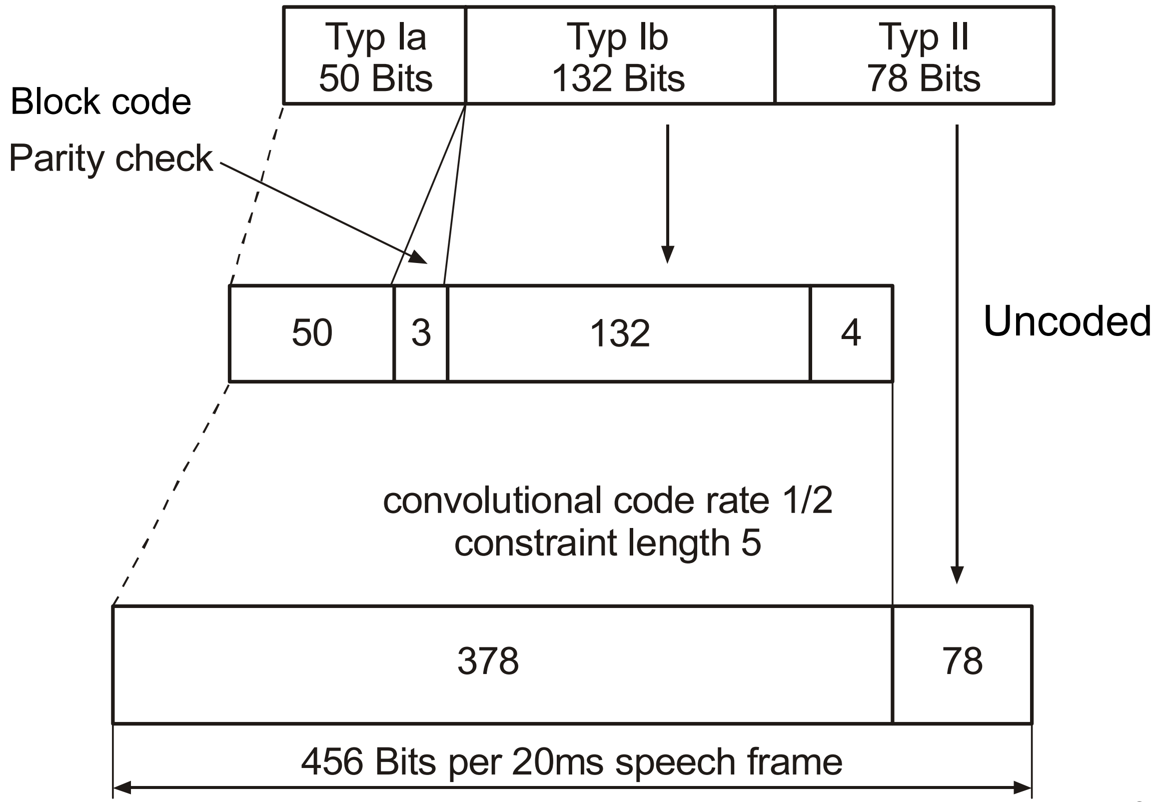
Vocoder



Preprocessing
STP analysis filtering
LTP
RPE Coding
STP Analyse Filtering
Hamming Window
Deter-mination of ACF
Conversion LAR'''(i) to k'(i)
Segments (20ms) (L=160)
Schur Recursion
Inter-polation of LARs
Pre-emphasis
Log Area Ratio (LAR)
LAR Inverse Quantizer
LAR Quantization
LTP Analysis
LTP Quantization
LTP Inverse Quantizer
Buffer
delay
Smoothing
RPE Grid Selection
Block-Adaptive PCM
Inverse BAPCM
RPE Grid Positioning
M U X
s(n)
LAR'(i)
G'
D'
G''
D''=D
k

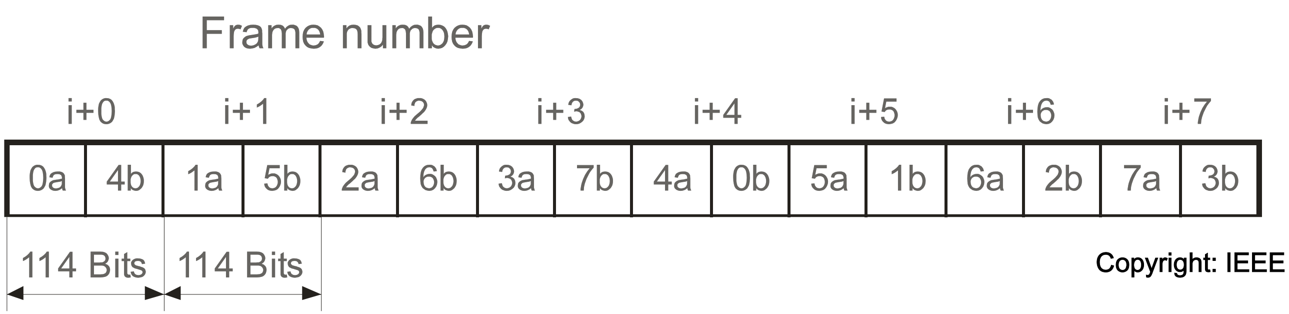
Channel coding of speech

The speech code bits are in three categories, with different levels of protection against channel errors.



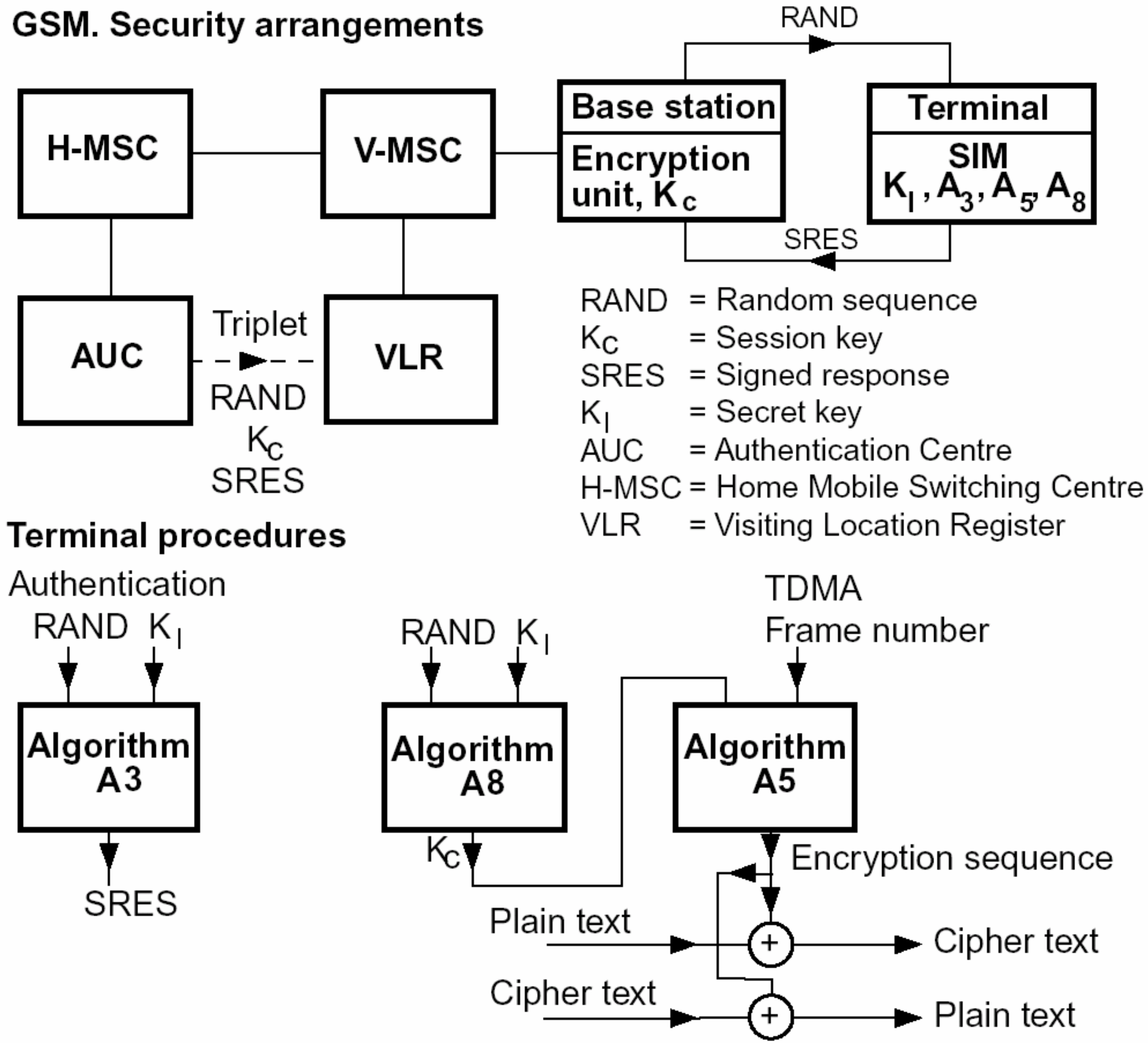
Interleaving and frequency hopping

- Bits interleaved over different frames

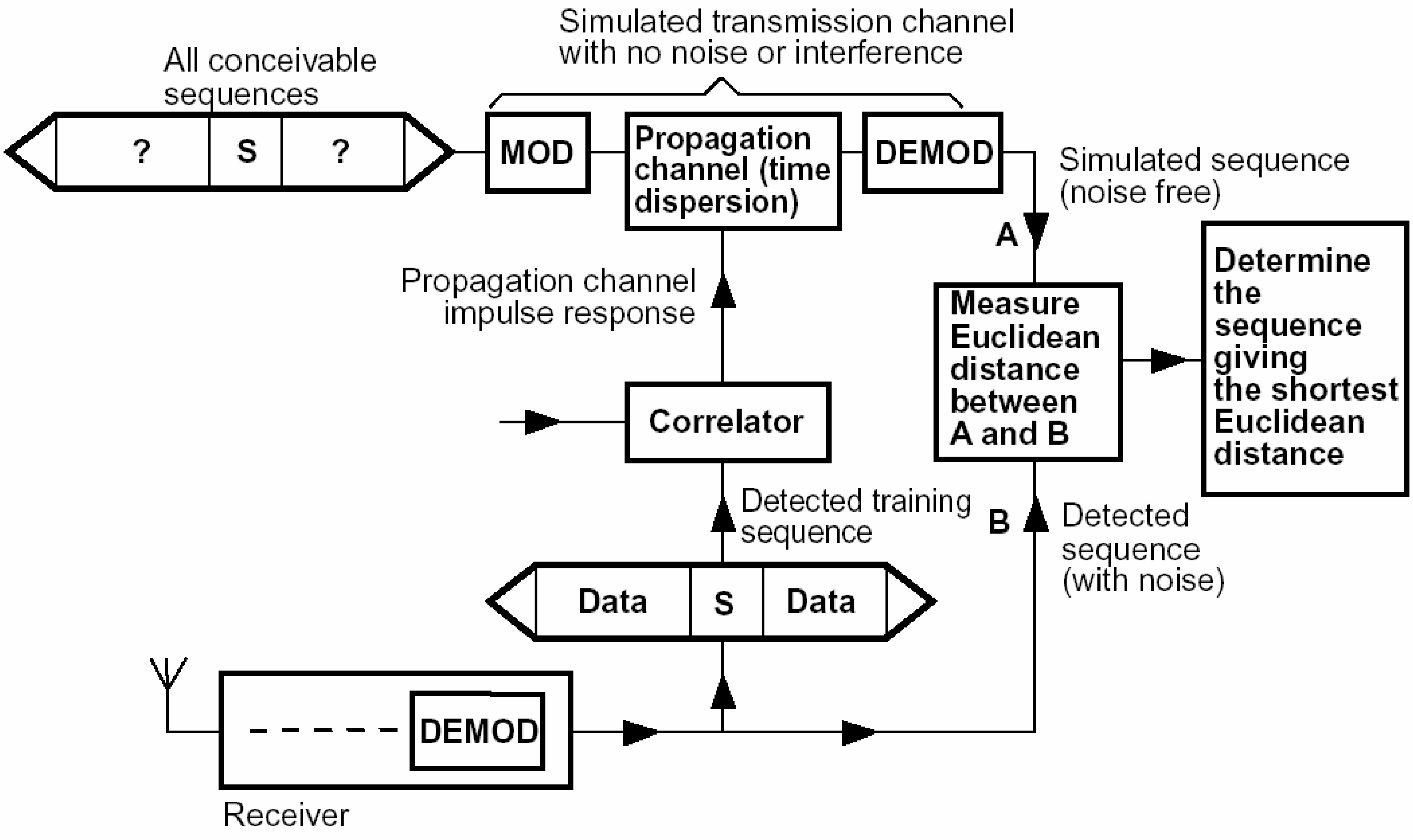


- Optional: frequency hopping, so that each frames sees different channel and interference

Encryption

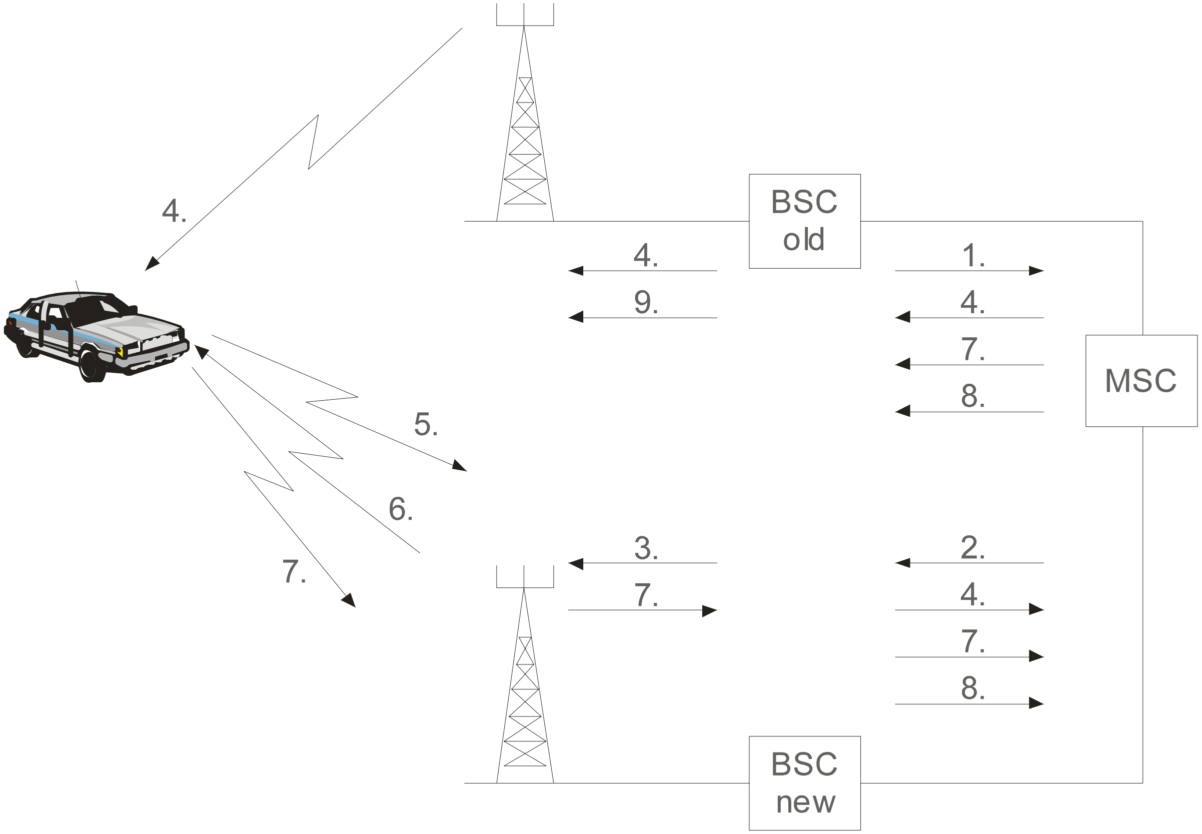

GSM. Security arrangements
RAND
H-MSC
V-MSC
Base station
Encryption unit, K_c
Terminal
SIM
K_I, A_3, A_5, A_8
SRES
AUC
Triplet
RAND
K_C
SRES
VLR
RAND = Random sequence
K_C = Session key
SRES = Signed response
K_I = Secret key
AUC = Authentication Centre
H-MSC = Home Mobile Switching Centre
VLR = Visiting Location Register
Terminal procedures
Authentication
RAND K_I
Algorithm A3
SRES
RAND K_I
Algorithm A8
K_C
TDMA
Frame number
Algorithm A5
Encryption sequence
Plain text
Cipher text
Cipher text
Plain text

Viterbi equalizer



Example for handover

- Handover between BTSs controlled by same MSC but different BSCs



GPRS and EDGE

GSM has evolved into a high-speed packet radio system in two steps

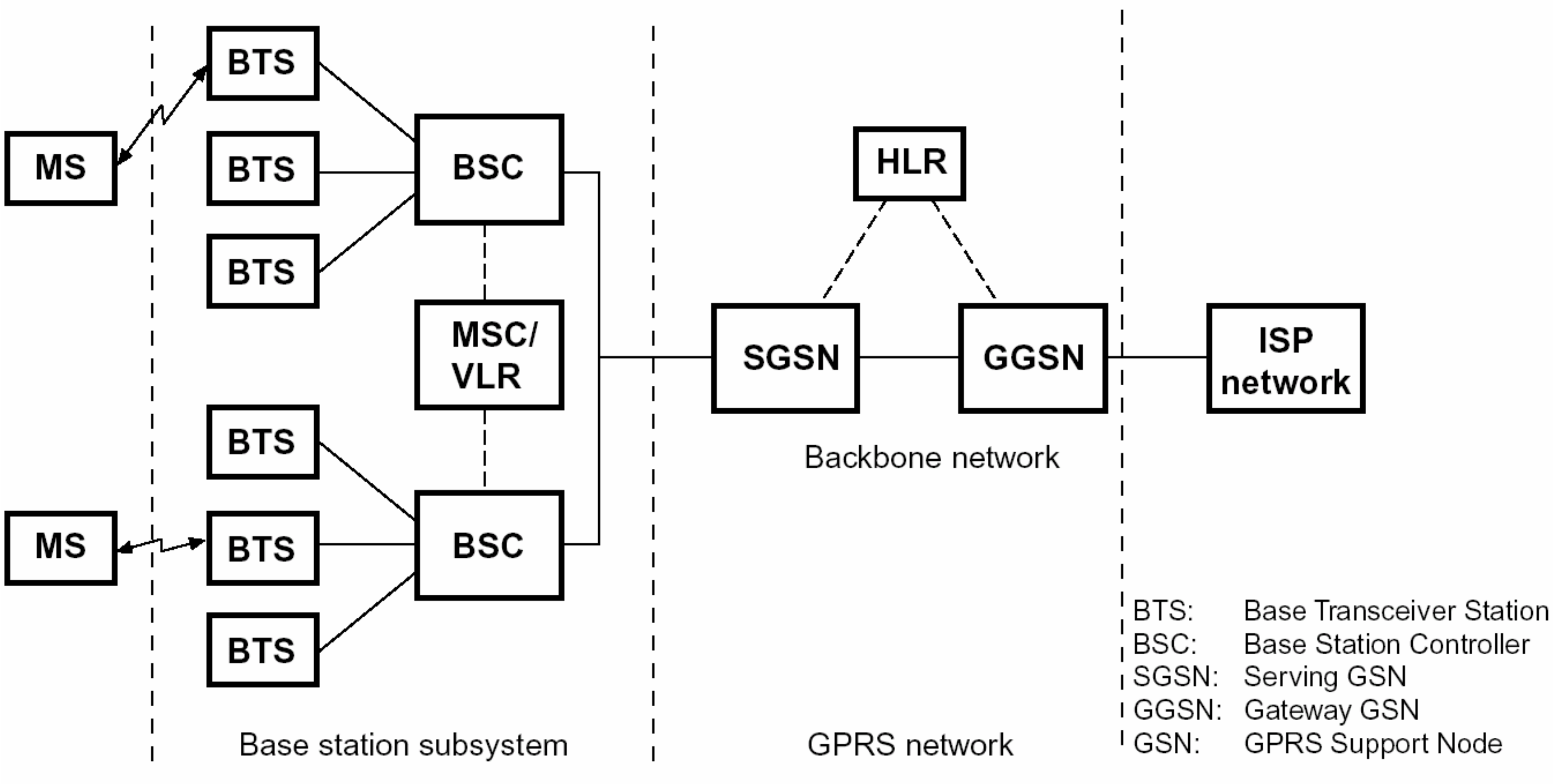
GPRS — **General Packet Radio Services**
where empty time slots can be used to transmit data packets.
Four new coding schemes are used (CS-1, ..., CS-4) with different levels of protection.

Up to 115 kbit/sec

EDGE — **Enhanced Data-rate for GSM Evolution**
where, in addition to GPRS, a new 8PSK modulation is introduced.
Eight new modulation and coding schemes are used (MCS-1, ..., MCS-8) with different levels of protection.

Up to 384 kbit/sec

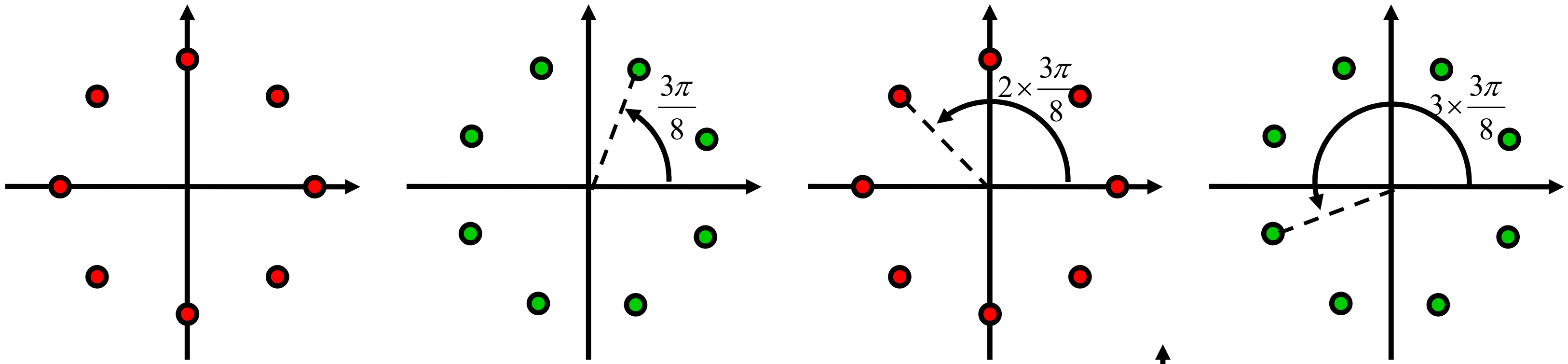
GPRS network



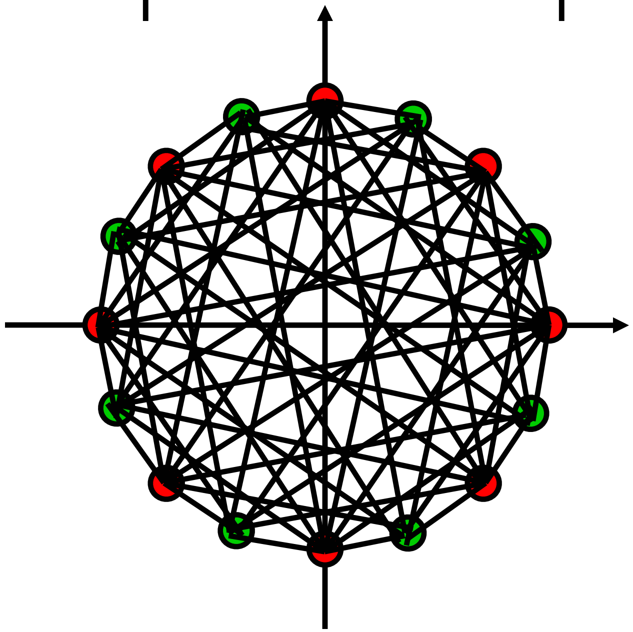
SGSN Serving GPRS Support Node
GGSN Gateway GPRS Support Node
ISP Internet Service Provider

EDGE 8PSK modulation

Linear 8-PSK ... but with rotation of signal constellation for each symbol



We avoid transitions close to origin, thus getting a lower amplitude variation!



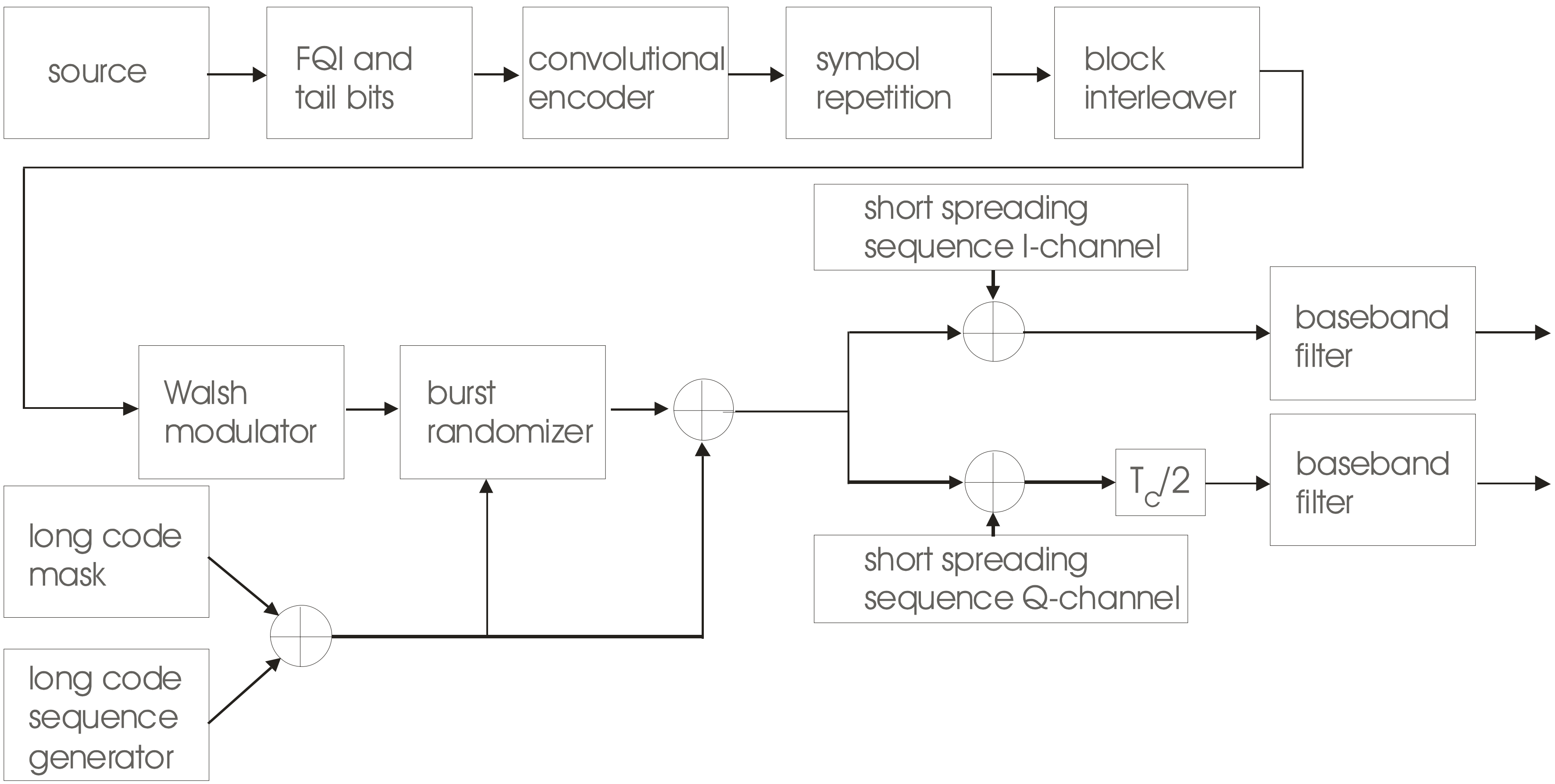
Chapter 22

IS-95 and CDMA 2000

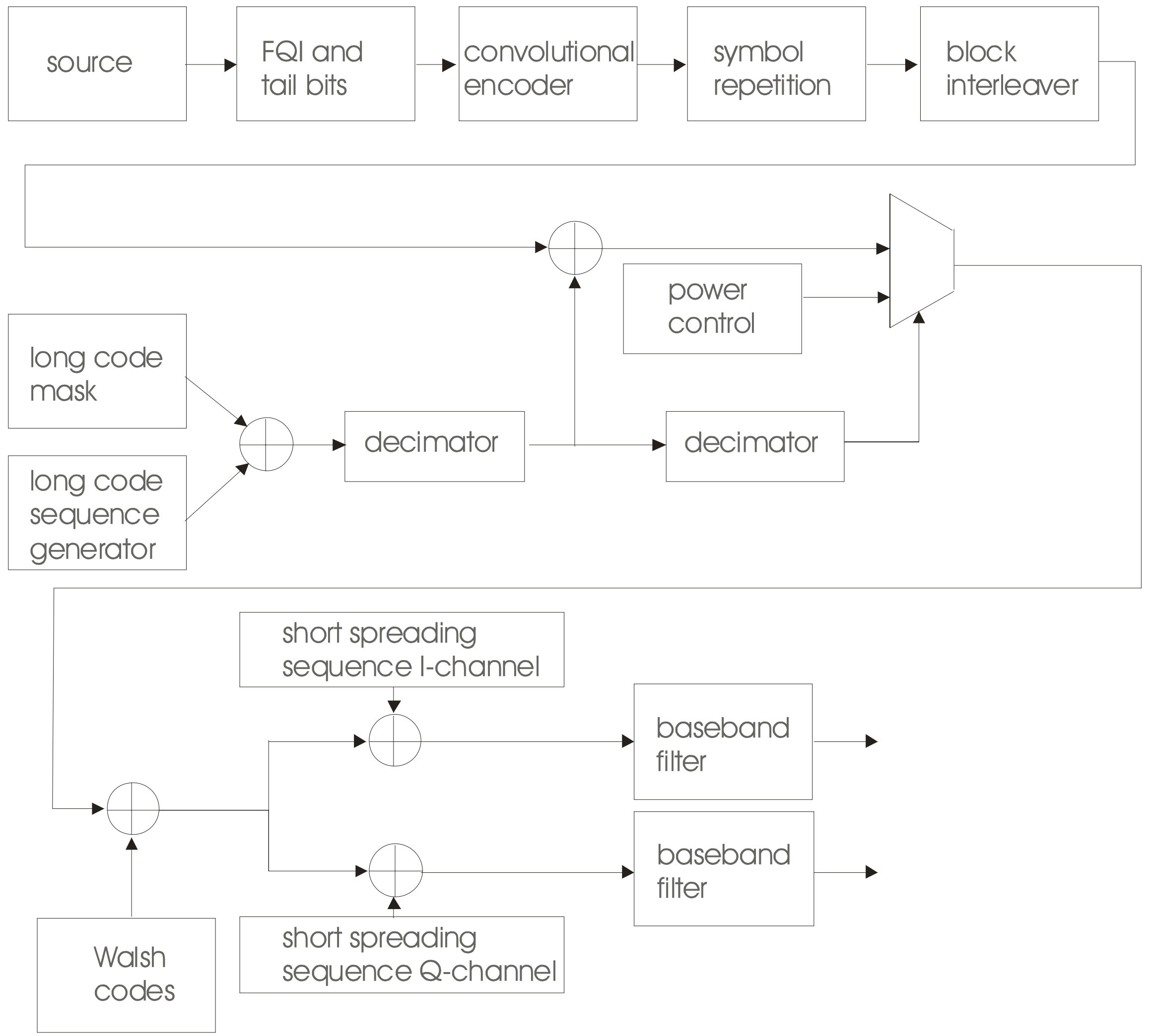
Speech coding

- Original speech codec: IS-96A
  - 8.6 kbit/s
  - Bad speech quality
- Enhanced speech codec: CDG-13
  - 13 kbit/s
  - Code excited linear prediction (CELP) principle
  - Much better speech quality
- Further enhancement: Enhanced Variable Rate Coder EVRC
  - Uses fewer number of bits both during speech pauses and during active period
  - 8.6 kbit/s

Spreading and modulation for uplink



Spreading and modulation for downlink



Logical channels (1)

- Traffic channels:
  - for transmission of user data
  - Depending on speech codec, use of rate set 1 or rate set 2
- Access channel:
  - Only in uplink
  - Allows MS that does not have current connection to transmit control messages: security messages, page response, origination, and registration

Logical channels (2)

- Pilot channel
- Synchronization channel
  - Transmits system details that allow MS to synchronize itself to the networks
- Paging channel
- Power control subchannel
- Mapping of logical channels to physical channels:
  - Assignment of different Walsh codes for different channels

Improvements in CDMA 2000

- Enhanced supplemental channels that can transmit data with higher rates
- Dedicated and common channels for packet data
- Walsh codes with variable length (OVSF codes)
- Faster power control for downlink
- Pilot for each uplink channel
- Enabling of smart antennas and transmit diversity

Chapter 23

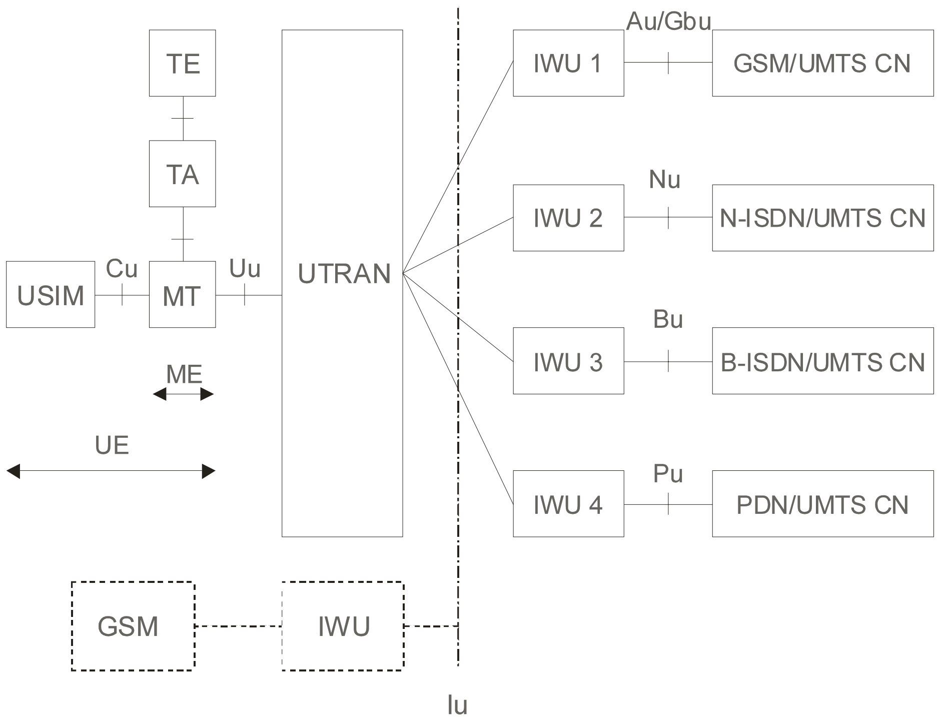
Wideband Code-Division Multiple Access (WCDMA)

Third-generation systems

- IMT-2000 established by International Telecommunications Union
- 3GPP and 3GPP2 are two organizations developing standards for IMT-2000
- 3GPP allows several "modes"
  - Wideband CDMA
  - C-TDMA

  } UMTS
  - DECT
  - EDGE
  - S-CDMA (China)
- Goals
  - Higher spectral efficiency
  - More flexibility, better suited for data transmission

UMTS simplified system overview



USIM:	User Service Identity Module
MT	Mobile Terminal
TA	Terminal Adapter
MT	Mobile Termination
IWU	Interworking Unit
UTRAN	UMTS Radio Access Network
UE	User equipment
CN	Core Network

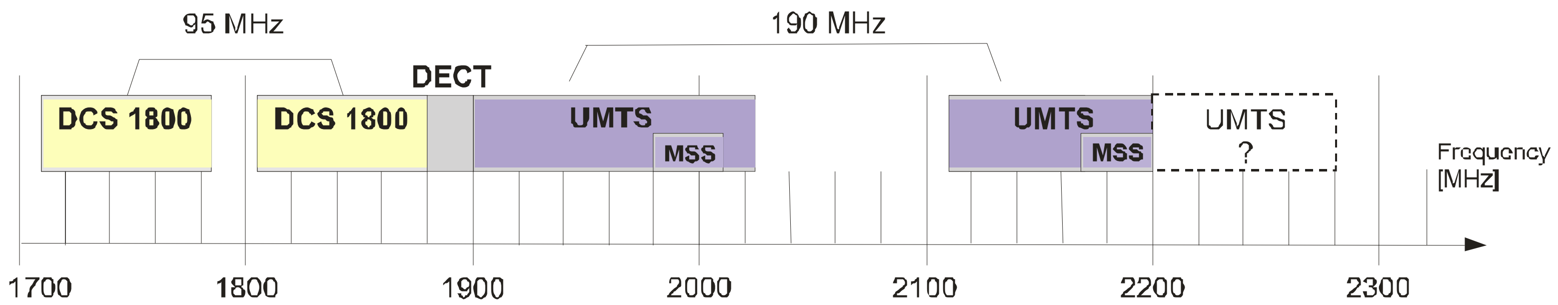
WCDMA - some parameters

Carrier spacing	5 MHz
Chip rate	3.84 Mchips/sec
Uplink spreading factor	4 to 256
Downlink spreading factor	4 to 512

All cells use the same frequency band!

RF aspects

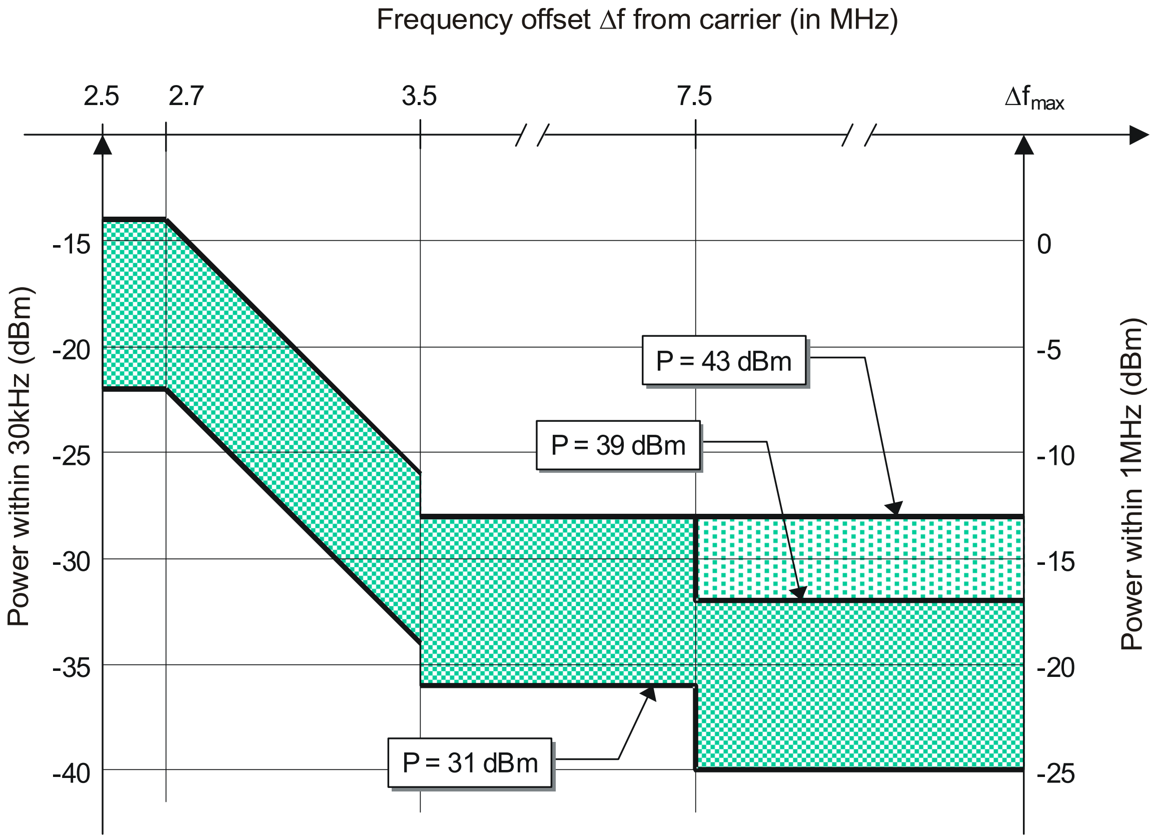
- Frequency bands



  - USA: uplink 1850-1910 MHz; downlink 1930-1990

- Transmit power
  - MS: 33, 27, 24, 21 dBm
  - BS: not specified in standard; typically 40-46 dBm

Spectrum mask

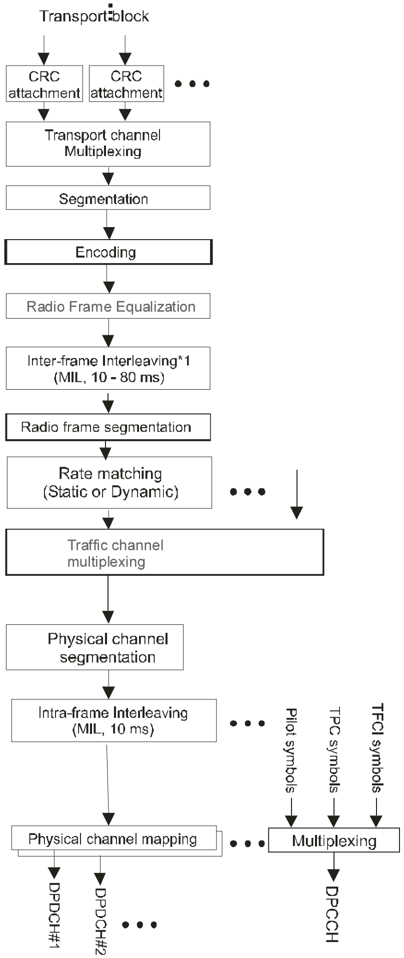


Mapping of logical to physical channels

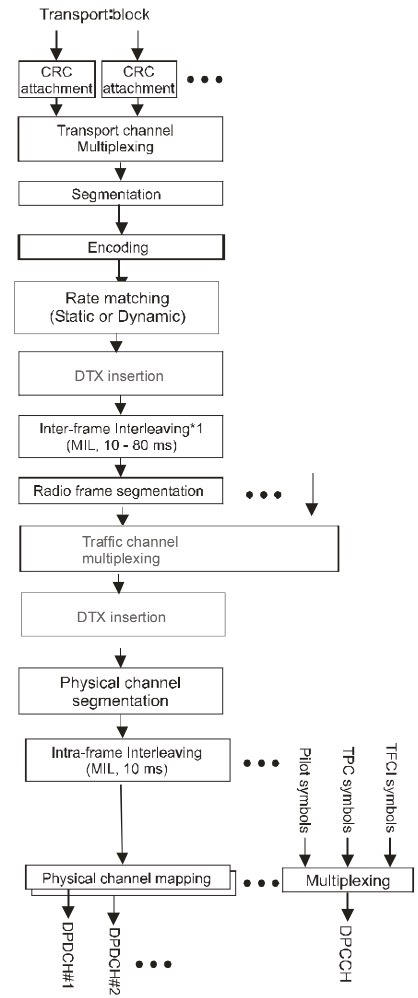
- Some physical channels have no equivalent logical channel

Transport Channels	Physical Channels
DCH	Dedicated Physical Data Channel (DPDCH)
	Dedicated Physical Control Channel (DPDCH)
RACH	Physical Random Access Channel (PRACH)
CPCH	Physical Common Packet Channel (PCPCH)
	Common Pilot Channel (CPICH)
BCH	Primary Common Control Physical Channel (P-CCPCH)
FACH	Secondary Common Control Physical Channel (S-CCPCH)
PCH	
	Synchronisation Channel (SCH)
DSCH	Physical Downlink Shared Channel (PDSCH)
	Acquisition Indication Channel (AICH)
	Page Indication Channel (PICH)

Multiplexing



a) uplink



b) downlink

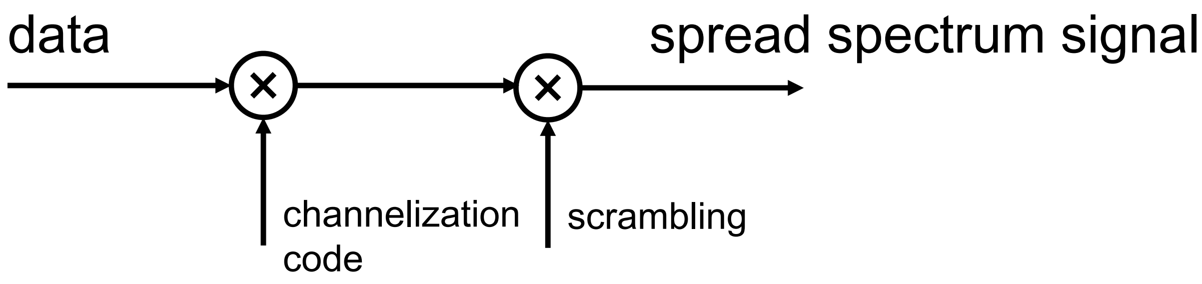
Coding

- CRC added for error detection

- Convolutional codes:
  - Rate ½ for common channels
  - Rate 1/3 for dedicated channels

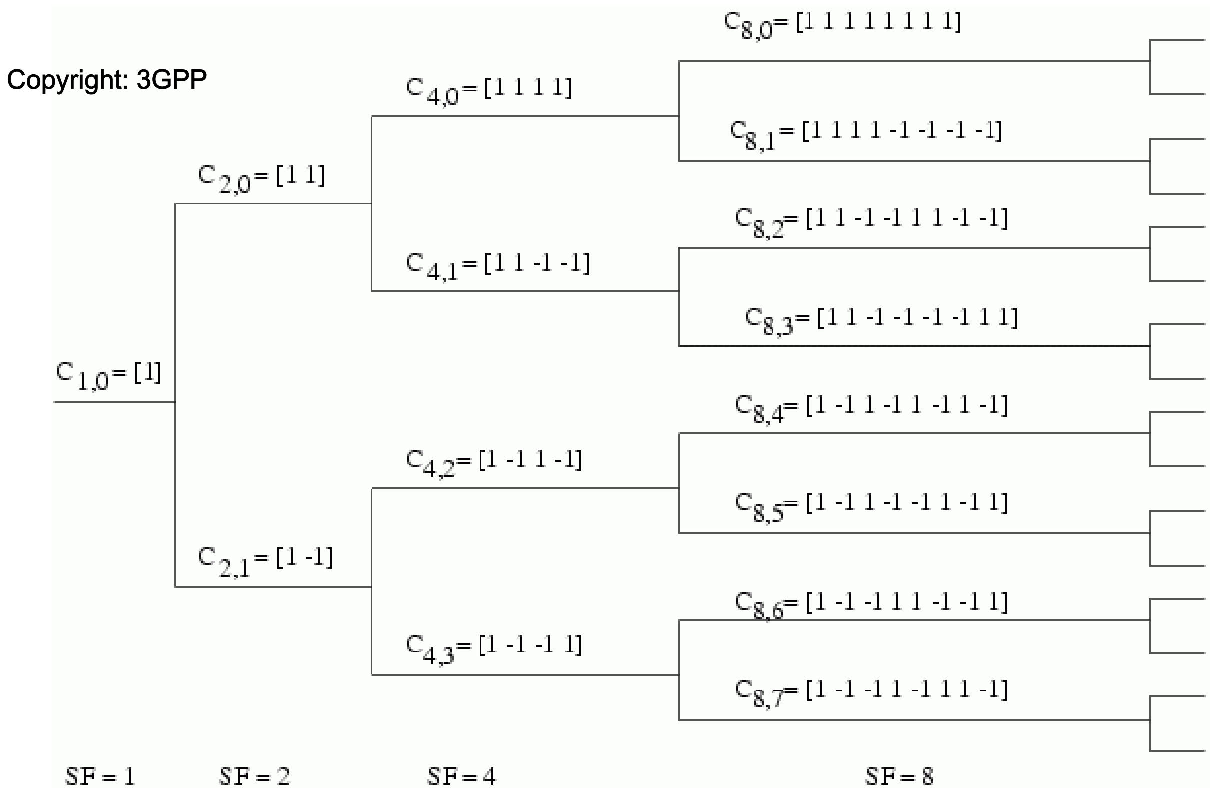
- Turbo codes
  - Code rate 1/3
  - Mainly for high-data-rate applications

Channelization and scrambling



Orthogonal Variable Spreading Factor

The OVSF codes used for variable rate spreading can be viewed as a code tree.

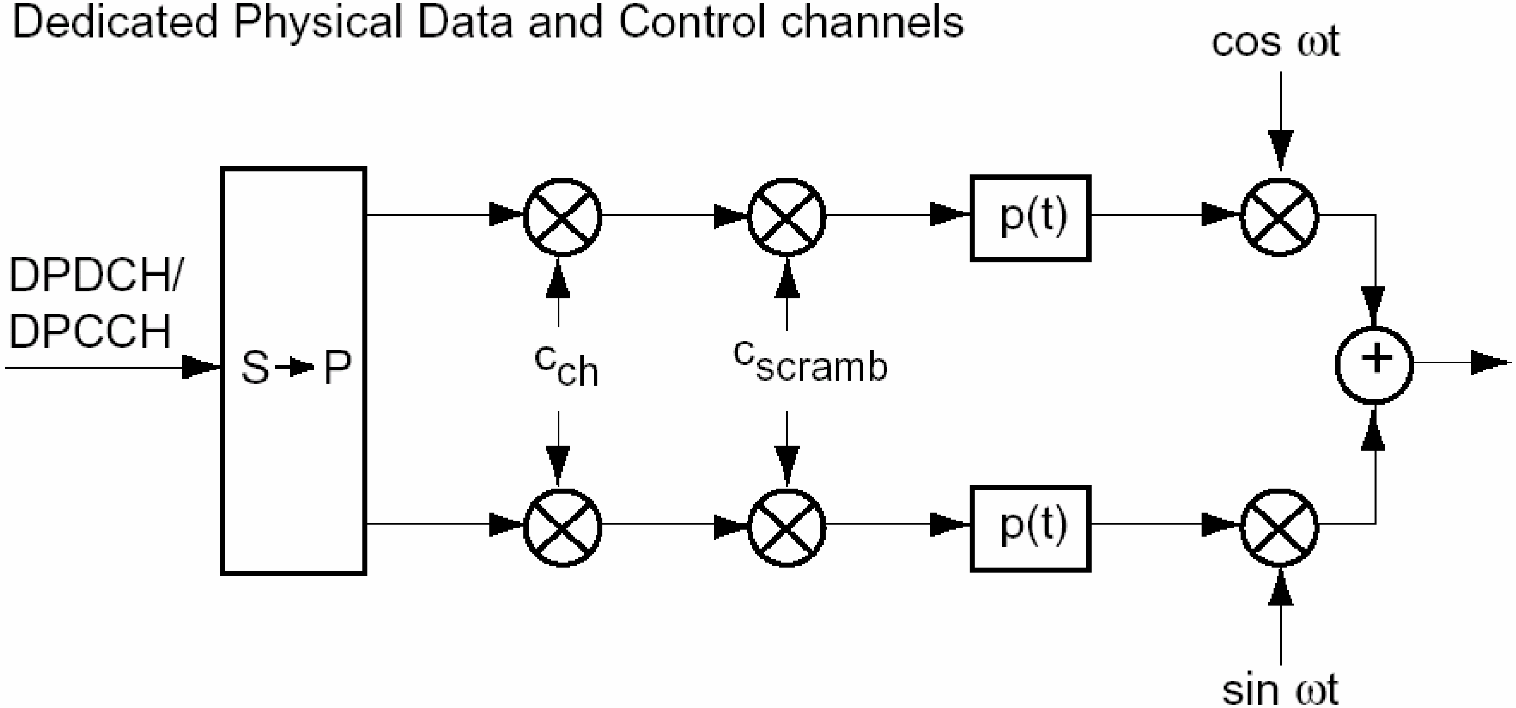


We can create several orthogonal channels by picking spreading codes from different branches of the tree.

Downlink

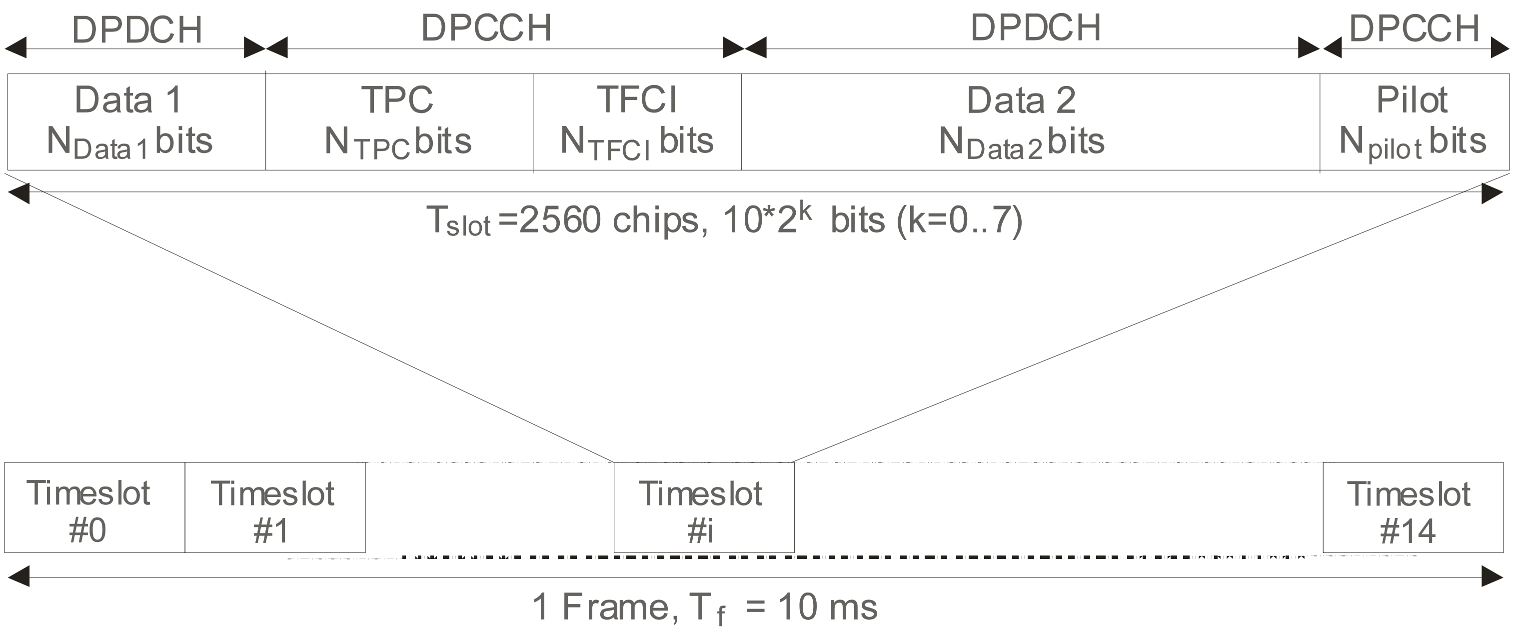
Downlink Spreading and Modulation

Dedicated Physical Data and Control channels



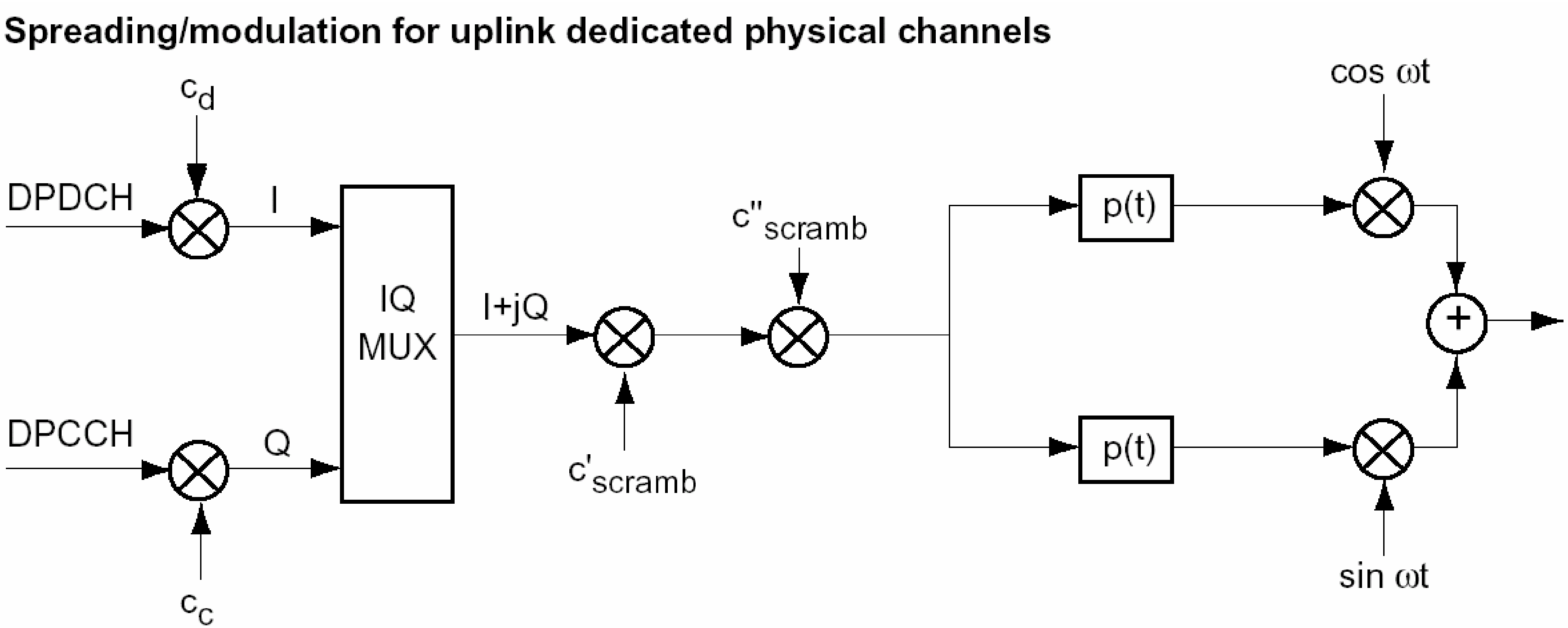
c_{ch}	Channelization code (OVSF)
c_{scramb}	Scrambling code (10 ms) 2^{18}-1 Gold code (40 960 chips)
p(t)	Root-raised cosine pulse shaping roll off 0.22
OVSF:	Orthogonal Variable Spreading Factor

Structure of downlink packet



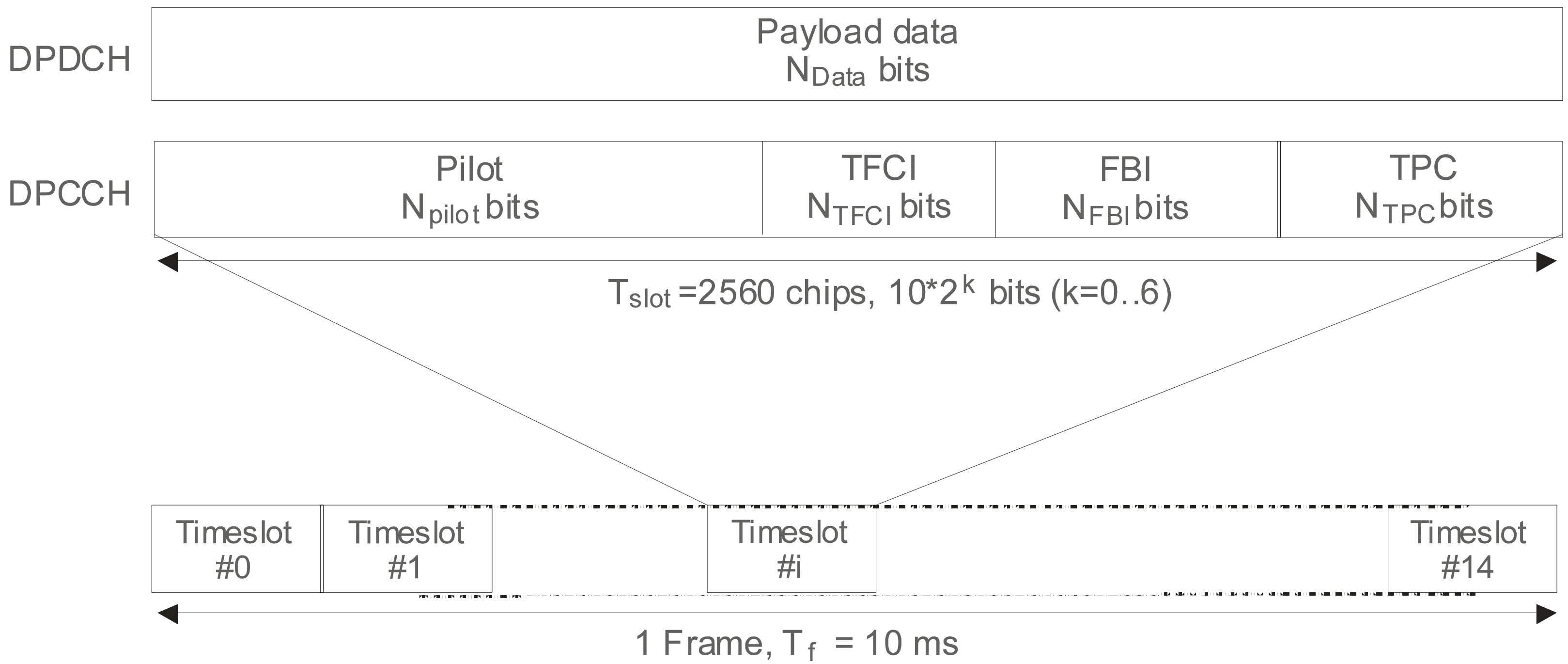
Copyright: 3GPP

Uplink



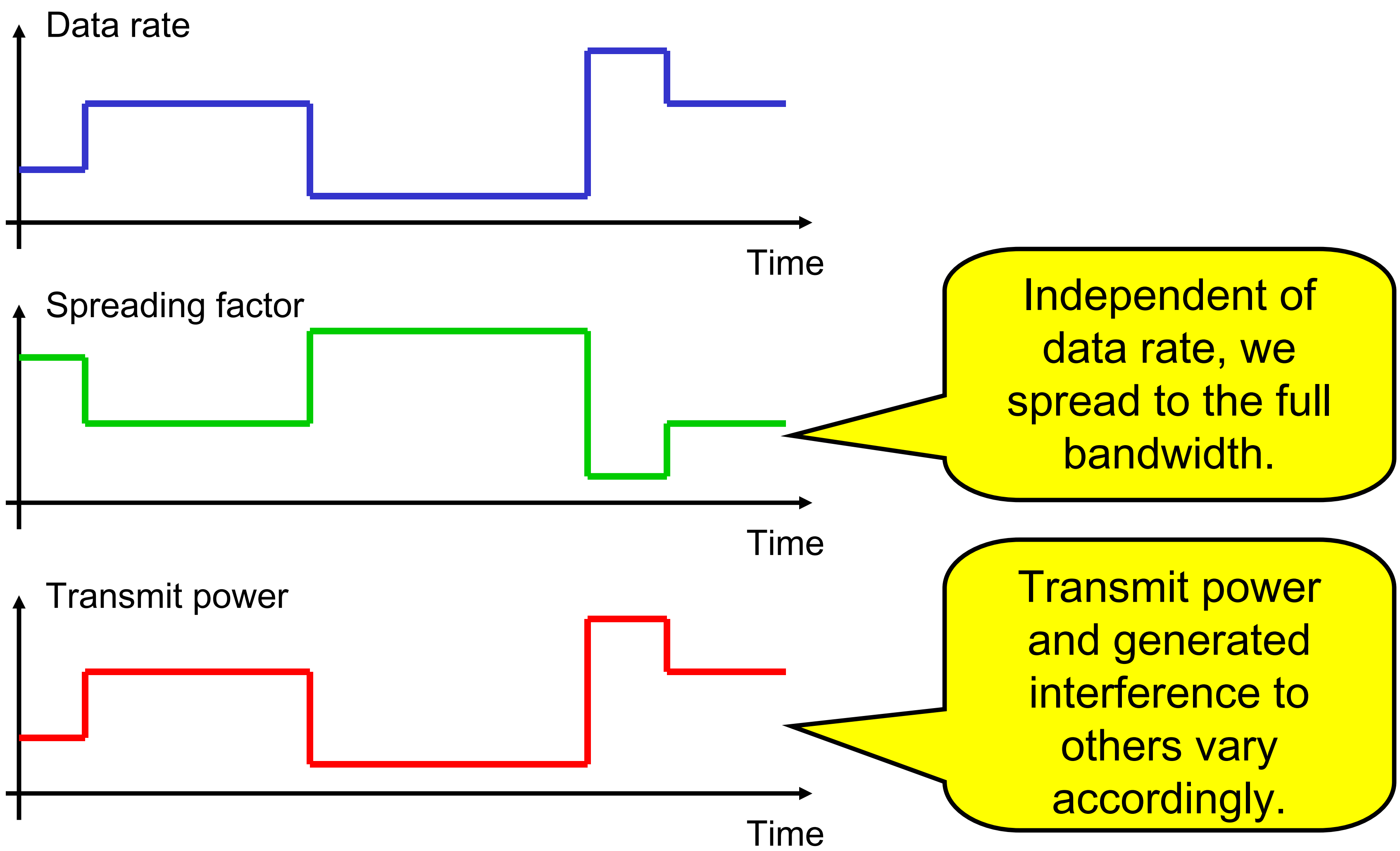
c_c, c_d	Channelization codes (OVSF)
c'_{scramb}	Primary scrambling code (256 chips) VL-KASAMI code (2 codes)
c''_{scramb}	Secondary scrambling code (10 ms optional) 2^{41}-1 Gold code (40 960 chips)
p(t)	Root-raised cosine pulse shaping, roll-off 0.22

Structure of uplink packet



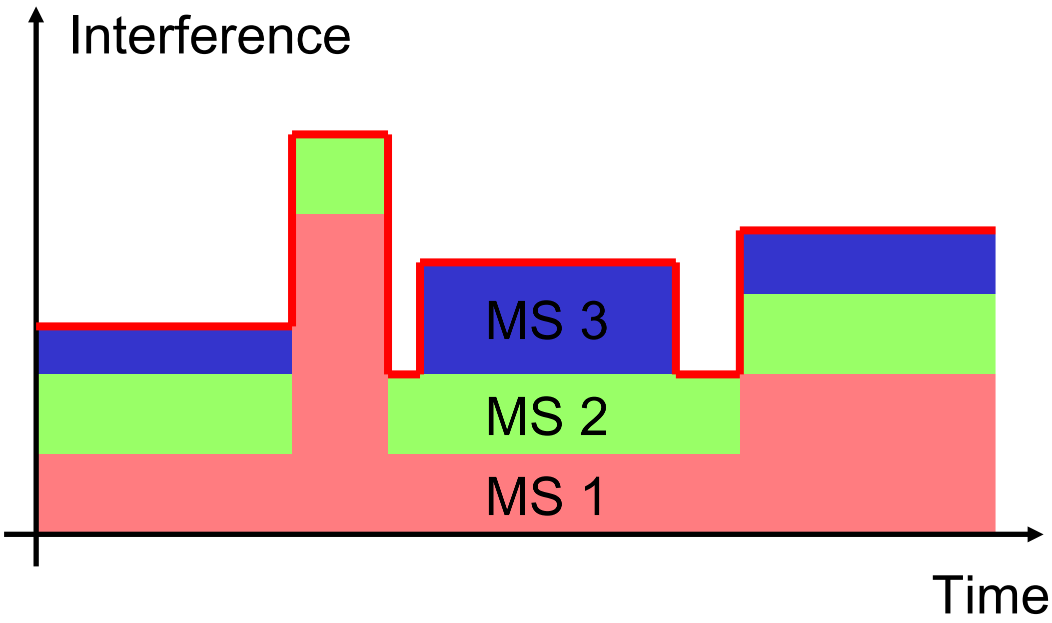
Copyright: 3GPP

Data rate and spreading factor



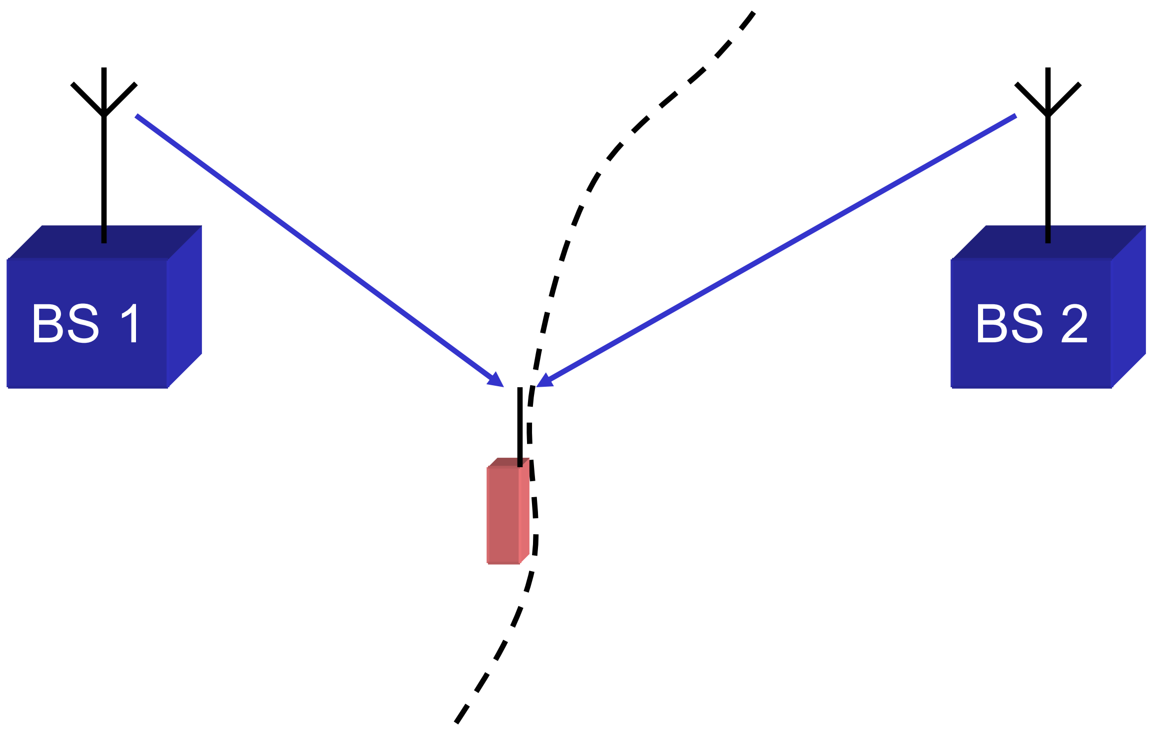
Data rate and interference

In simple words, with a limited interference allowed, we can have many low data-rate channels or a few high data-rate channels.



Soft handover

Since all base stations used the same frequency band, a terminal close to the cell boundary can receive “the same” signal from more than one base station and increase the quality of the received signal.



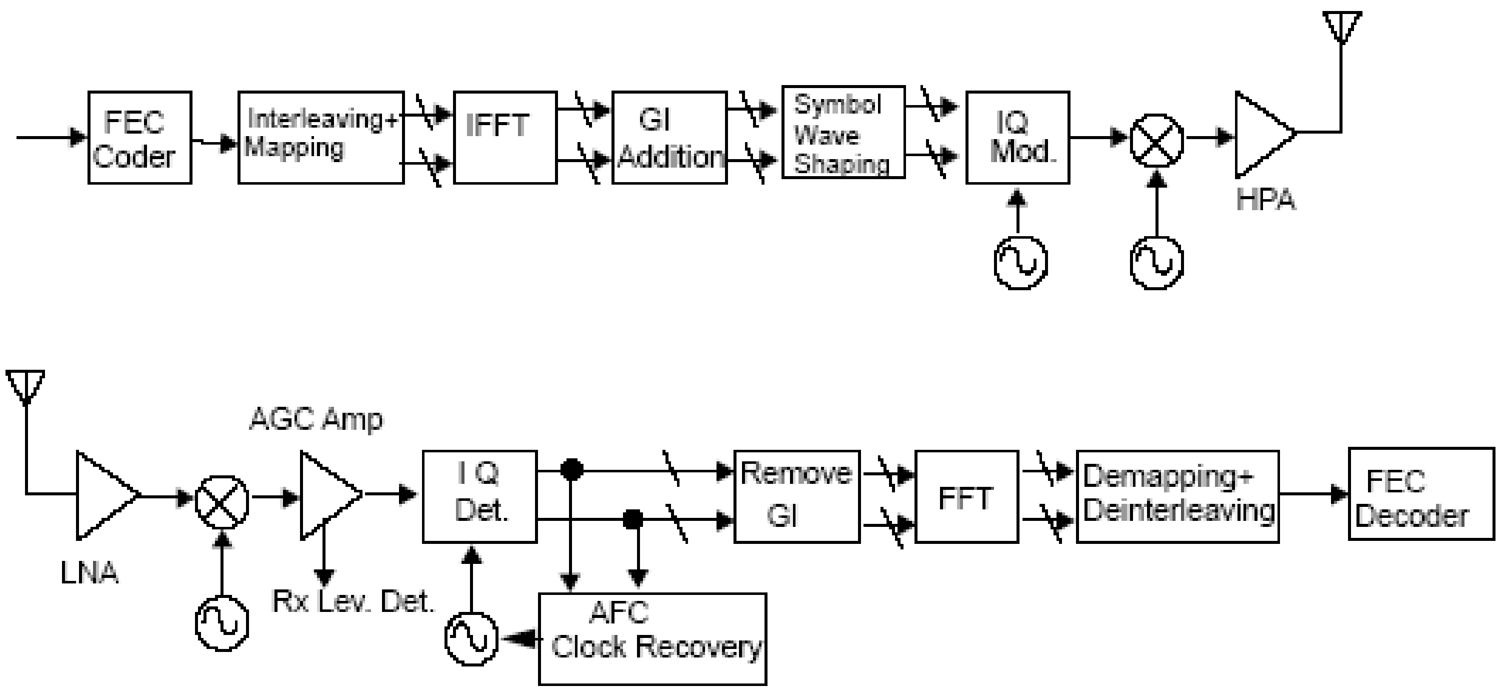
Wireless LANs
IEEE 802.11

History

- Wireless LANs became of interest in late 1990s
  - For laptops
  - For desktops when costs for laying cables should be saved
- Two competing standards
  - IEEE 802.11 and HIPERLAN
  - IEEE standard now dominates the marketplace
- The IEEE 802.11 family of standards
  - Original standard: 1 Mbit/s
  - 802.11b (WiFi, widespread after 2001): 11 Mbit/s
  - 802.11a (widespread after 2004): 54 Mbit/s
  - 802.11e: new MAC with quality of service
  - 802.11n: > 100 Mbit/s

802.11a PHY layer

- Transceiver block diagram



Copyright: IEEE

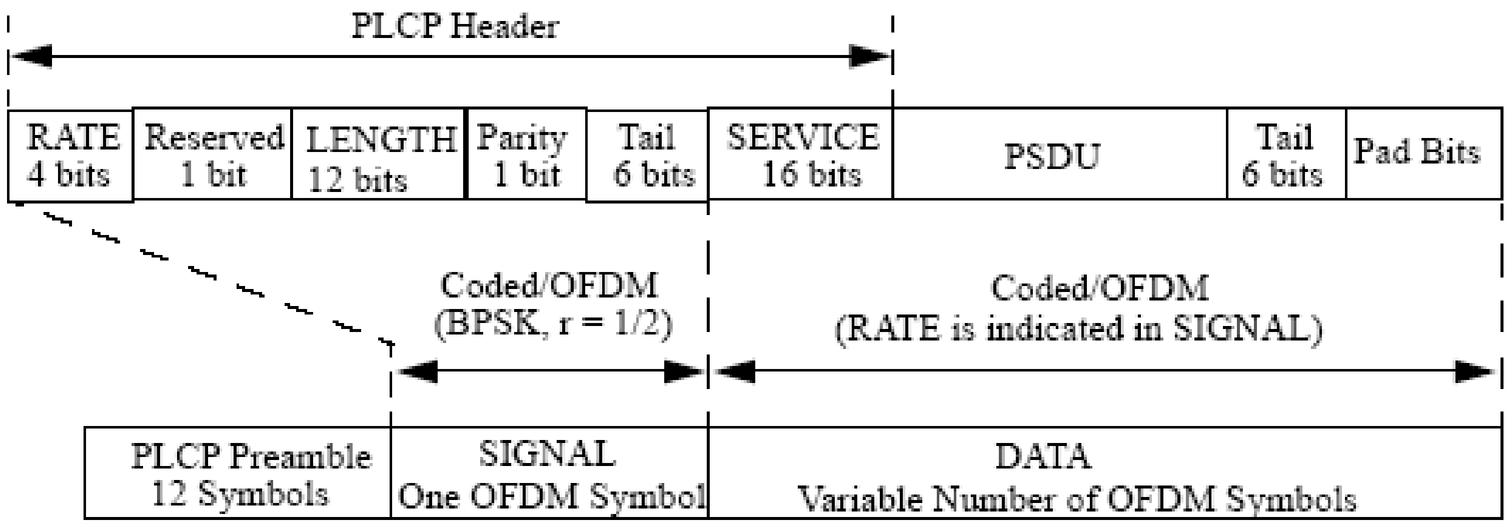
802.11a PHY layer

- The following data rates are supported:

Data rate (Mbit/s)	Modulation	coding rate	coded bits per subcarrier	coded bits per OFDM symbol	data bits per OFDM symbol
6	BPSK	1/2	1	48	24
9	BPSK	3/4	1	48	36
12	QPSK	1/2	2	96	48
18	QPSK	3/4	2	96	72
24	16-QAM	1/2	4	192	96
36	16-QAM	3/4	4	192	144
48	64-QAM	2/3	6	288	192
54	64-QAM	3/4	6	288	216

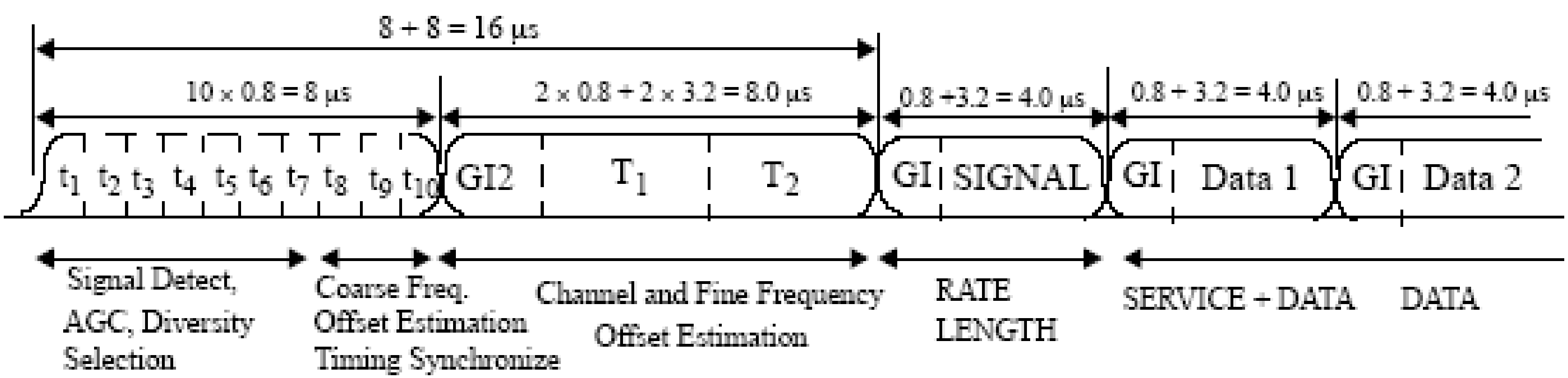
11a header and preamble

- Header conveys information about data rate, length of the data packet, and initialization of the scrambler



11a header and preamble

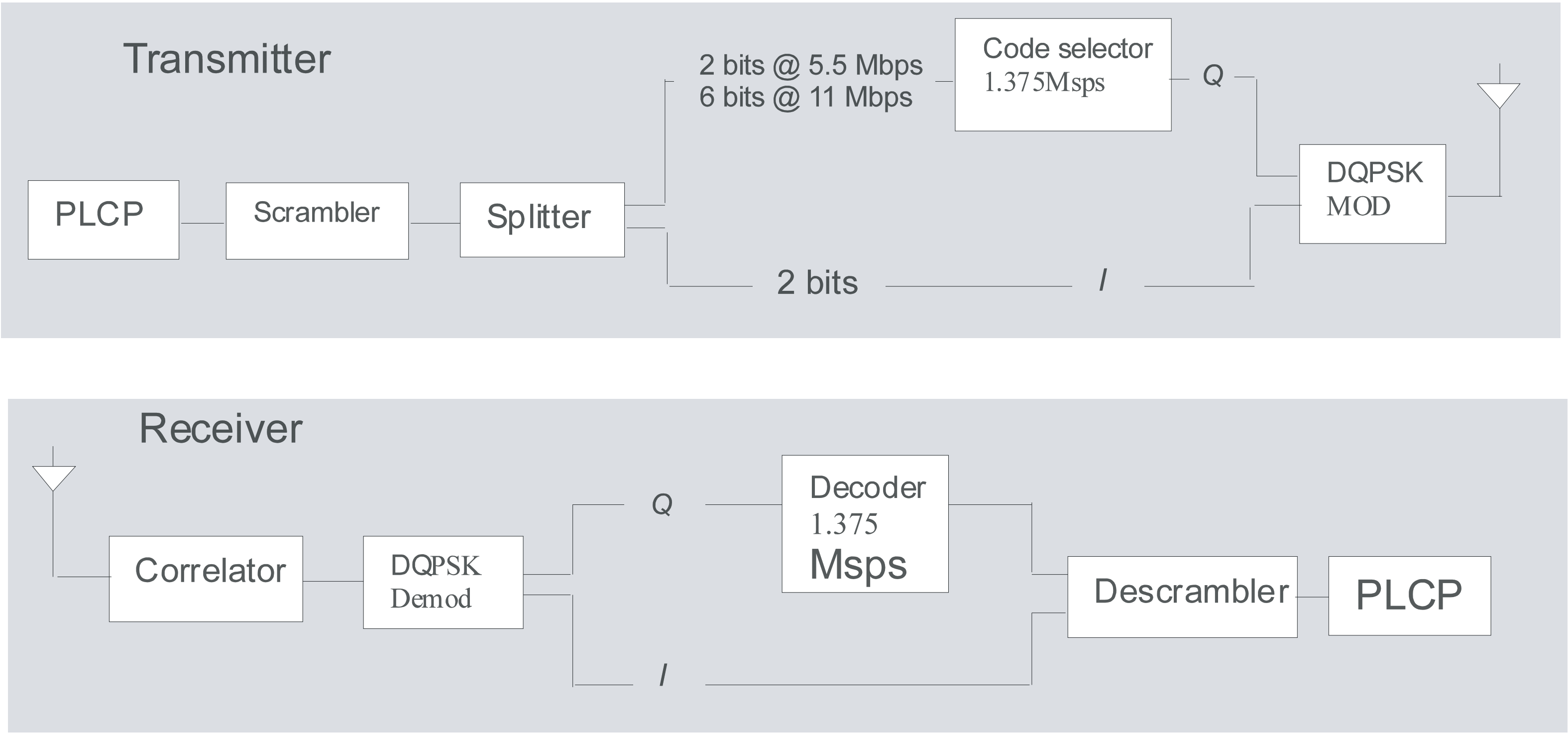
- PLCP preamble: for synchronization and channel estimation



802.11b air interface

- Key air interface parameters
  - Frequency range: 5.40-2.48 GHz
  - Carrier spacing: 20-25 MHz
  - Data rates: 1, 2, 5.5, 11 Mbps
- Modulation and multiple access:
  - for low data rates, as well as for header and preamble (1 Mbit/s):
    - Direct-sequence spreading with Barker sequence
    - Differential phase shift keying modulation
  - For high data rates: complementary code keying (CCK)
  - Multiple access by FDMA and packet radio access
- Channel coding:
  - Convolutional coding with rate ½ is option

Transceiver structure for 802.11b



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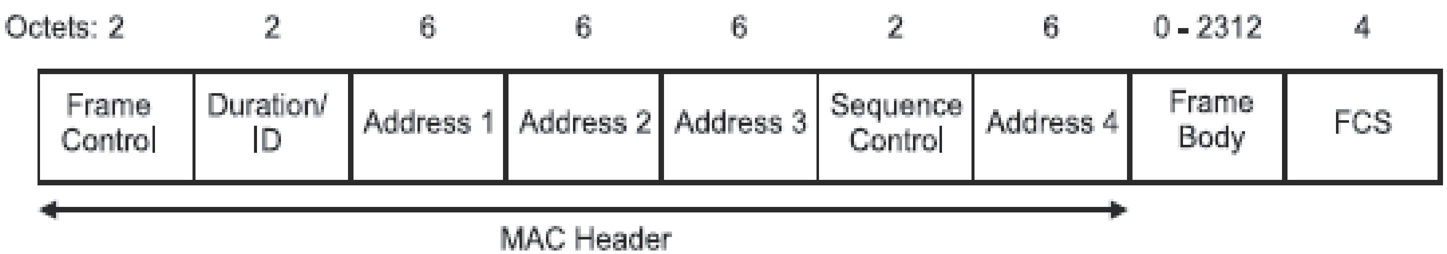
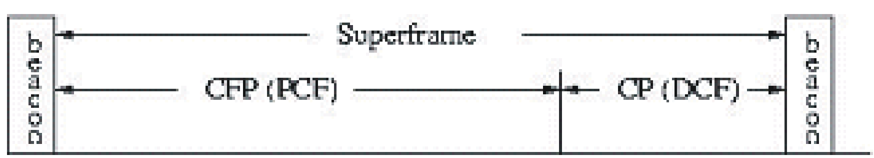
MAC and multiple access

- Frame structure:
  - Contains payload data, address, and frame control into



- Multiple access: both contention-free and contention-based access



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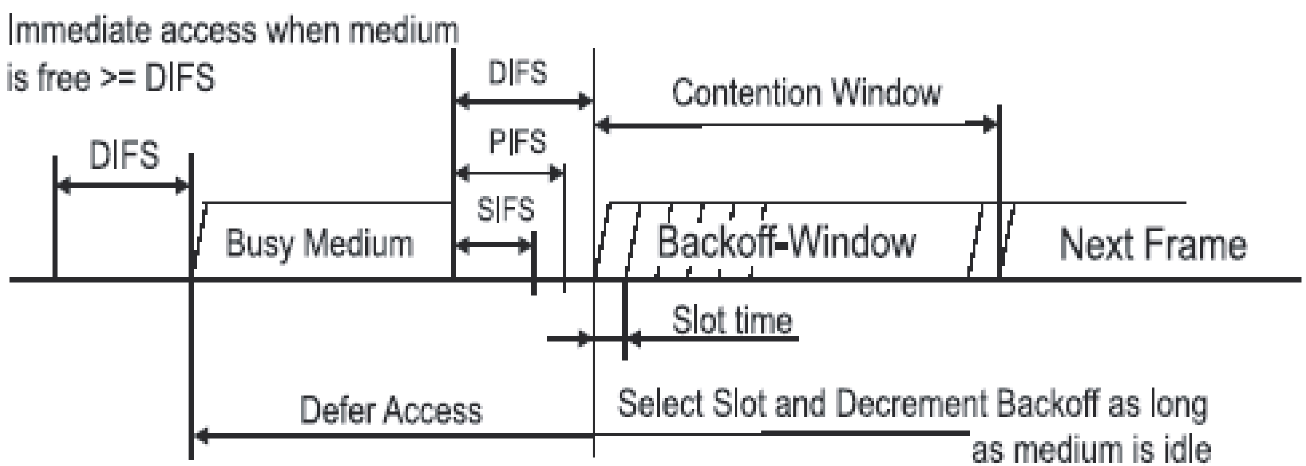
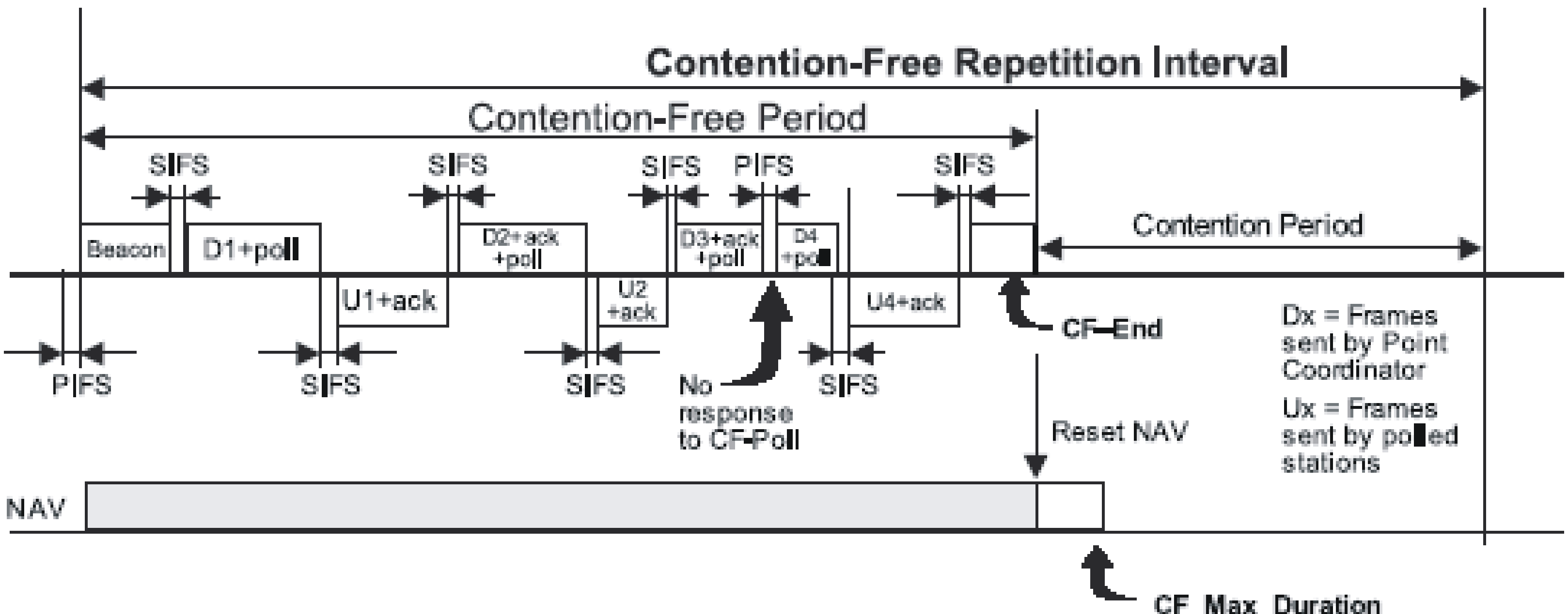
Contention-based access

- CSMA (carrier-sense multiple access):



Contention-free access

- Polling:



Further improvements

- 802.11e: improvements in the MAC; provides quality of service
  - CSMA/CA-based Enhanced Distributed Channel Access (EDCA) manages medium access during CP.
  - Polling-based HCF (Hybrid Coordination Function) Controlled Channel Access (HCCA) handles medium access during CFP.
  - BlockACK and delayed blockACK reduce overhead
  - Contention Free Burst (CFB) and Direct Link Protocol (DLP) improve channel efficiency.
- 802.11n: higher throughput by using multiple antenna elements