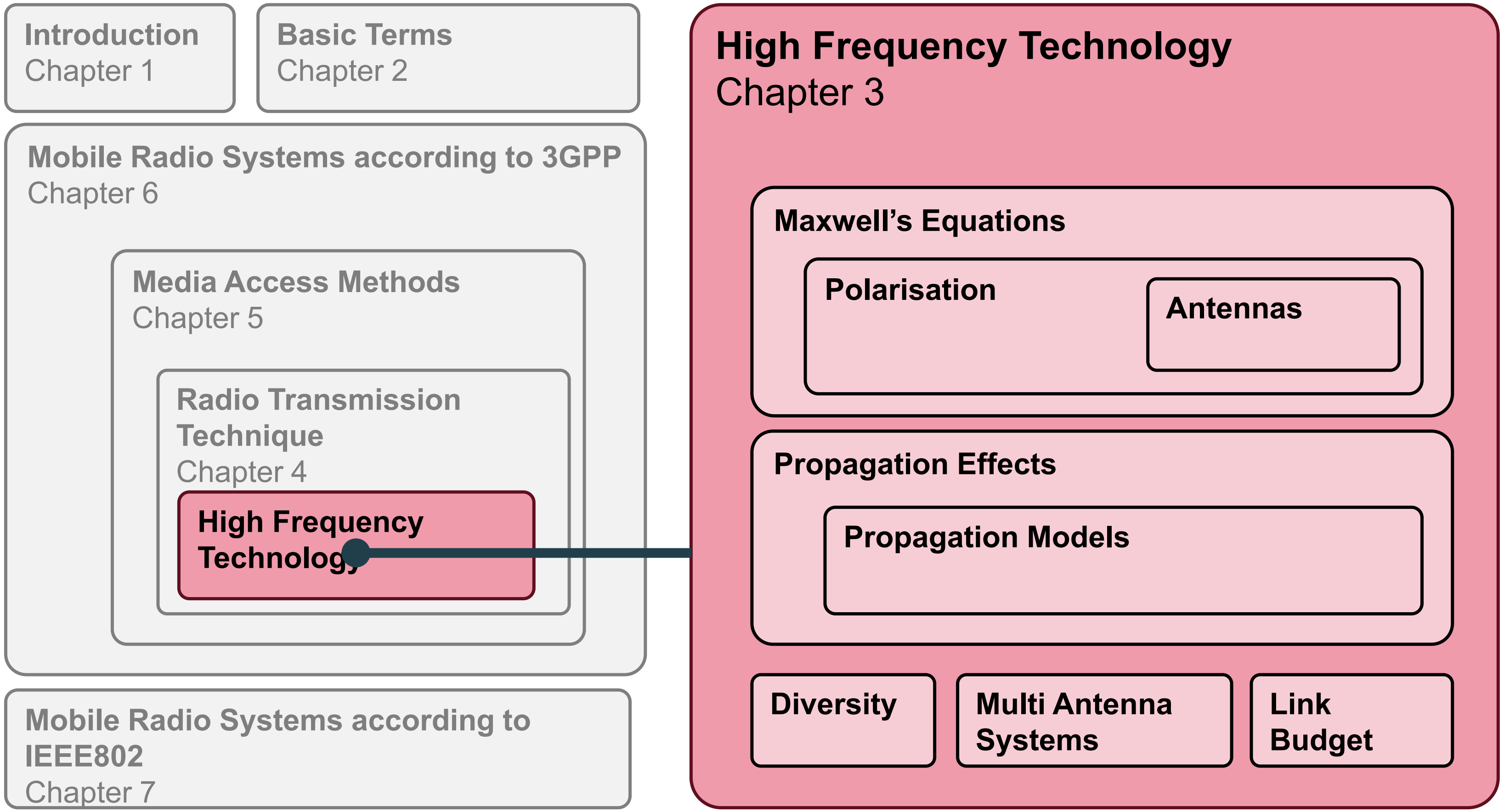
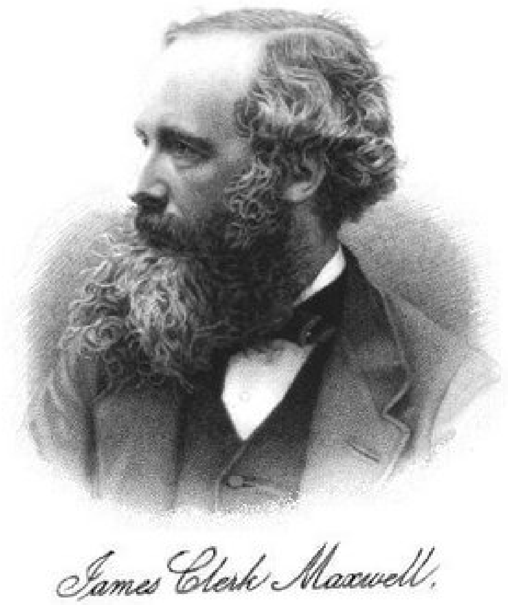
Chapter 3 – High Frequency Technology



3 High Frequency Technology
3.1 Maxwell's Equation



Heinrich Hertz

- Complete description of the electromagnetic fields and their time dependency by Maxwell's Equations

The Field Equations in Integral Form

$$\oint_C \vec{H}\mathrm{d}\vec{r} = \iint_A \vec{J}\mathrm{d}\vec{A}$$

(law of magnetic flux)
(3.1)

A magnetic field is induced by a time-variant electrical field or a current.

$$\oint_C \vec{E}\mathrm{d}\vec{r} = -\frac{\partial}{\partial t}\iint_A \vec{B}\mathrm{d}\vec{A}$$

(law of induction)
(3.2)

An electrical field is induced by a time-variant magnetic field.

$$\oiint_A \vec{B}\mathrm{d}\vec{A} = 0$$

(source-freedom)
(3.3)

Magnetic flux lines are closed.

$$\oiint_A \vec{D}\mathrm{d}\vec{A} = \iiint_V p\mathrm{d}V$$

(continuity equations)
(3.4)

Electrical flux lines either begin and end at charges or they are closed.

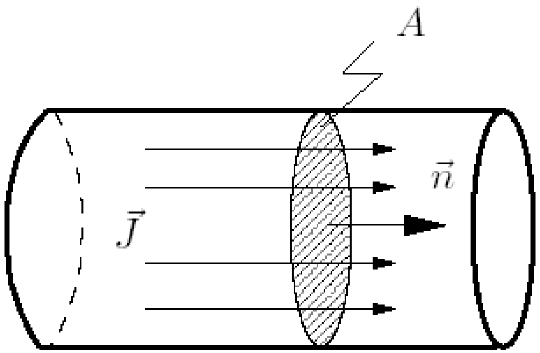
Material Equations

$\vec{D} = \epsilon\vec{E}$ $\vec{H} = \frac{1}{\mu}\vec{B}$ (Material equations) 3.5

Essential facts:

- For processes in the current-free and charge-free space, the following restrictions apply: $\vec{J} = 0$ und $\rho = 0$
- Electrical fields lines have a source and a sink.
- Magnetic fields lines are divergence free.
- Alternating electrical and magnetic fields induce each other (electro-magnetic wave).

Field Equation of the Stationary Electrical Field



$$= \vec{J}\vec{n} \cdot A \qquad (2.6)$$

$$= \kappa\vec{E} \qquad (2.7)$$

κ: electrical conductivity

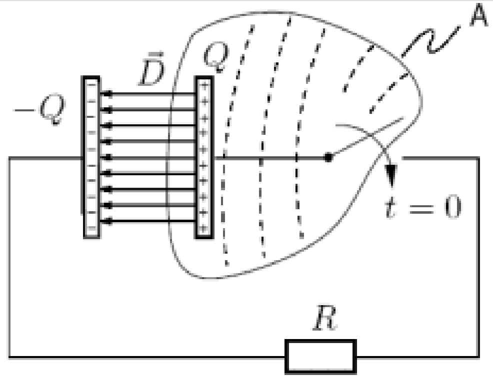
general form:

$$I = \iint_A \vec{J}\mathrm{d}\vec{A} \qquad (2.8)$$

and for a closed contour:

$$\oiint_A \vec{J}\mathrm{d}\vec{A} = 0 \qquad (2.9)$$

Displacement Current Density



- Plate capacitor with electrical charge Q

$$t < 0\colon\ Q = \oiint_A \vec{D}\mathrm{d}\vec{A} \qquad (2.10)$$

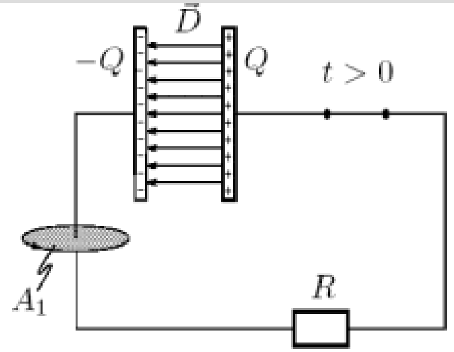
- At the time t = 0 the switch is closed.
- t > 0: The capacitor is discharged. Current flows through the hull

$$-\frac{\partial Q}{\partial t} = \oiint_A \vec{J}\mathrm{d}\vec{A} = -\frac{\partial}{\partial t}\oiint_A \vec{D}\mathrm{d}\vec{A} \to \dim\left[\frac{\partial}{\partial t}\vec{D}\right] = \dim[\vec{J}] \qquad (2.11)$$

$\frac{\partial}{\partial t}\vec{D}$ is called displacement current density.

- Overall current density $(\frac{\partial}{\partial t}\vec{D}+\vec{J})$ is source-free: $\oiint\left(\frac{\partial}{\partial t}\vec{D}+\vec{J}\right)d\vec{A} = 0$ (2.12)

Electromagnetic Fields (1)



- At the time t = 0 the switch is closed.
- The capacitor is discharged.

$$\iint_{A_1} \vec{J}\mathrm{d}\vec{A} \neq 0 \quad \text{, but} \quad \iint_{A_2} \vec{J}\mathrm{d}\vec{A} = 0 \qquad (2.13)$$

- According to the law of magnetic flux, the current density is connected with a magnetic field. This is clearly defined only if

$$\oint_C \vec{H}\mathrm{d}\vec{r} = \iint_A \left(\frac{\partial}{\partial t}\vec{D} + \vec{J}\right)\mathrm{d}\vec{A} \qquad \text{(1}^{\text{st}}\text{ Maxwell's equation)} \qquad (2.14)$$

- In terms of generation of magnetic fields, conduction current density and displacement current density are equivalent. A time-variant displacement field generates a magnetic field without current flow, i.e. also in a vacuum. In terms of its magnetic effect, a time-variant conduction current interrupted by an insulator can be theoretically continued through a displacement current of the same amperage.

Electromagnetic Fields (2)

- Assumption:
  - The law of induction also applies if the path of integration C is not along with a conductor.
  - Thus, for any closed circulation *C* the following applies:

$$\oint_C \vec{E}\,\mathrm{d}\vec{r} = -\frac{\partial}{\partial t}\iint_A \vec{B}\,\mathrm{d}\vec{A} \qquad \text{(2}^{\text{nd}}\text{ Maxwell‘s equation)} \qquad (2.15)$$

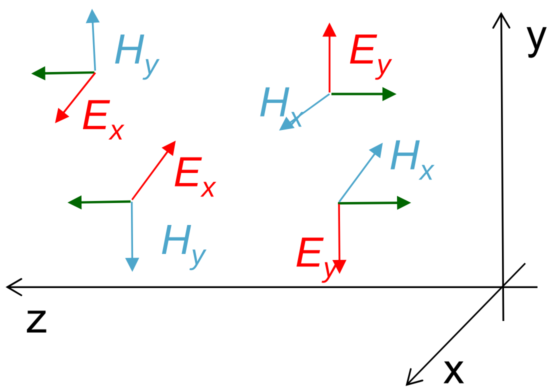
- From this follows:
  - The flux lines of a time-variant magnetic field are circularly surrounded by electrical flux lines.
  - The flux lines of a time-variant electrical field are surrounded by magnetic flux lines.

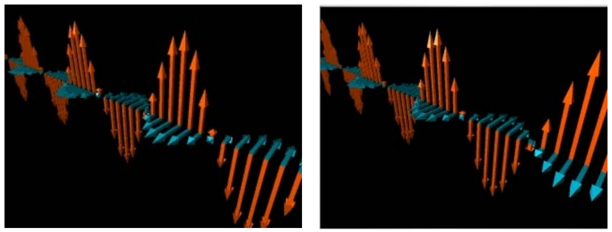
3.2 Several Electrodynamic Principles

- For Maxwell's equations, several solutions also exist following the a.m. restrictions. The simplest solution is the plane wave.
- We refer to a wave as a plane wave, if the instantaneous values of the **field values $\underline{\vec{E}}$ und $\underline{\vec{H}}$** do not depend on the coordinates vertically located to the direction of propagation.
- Since $\vec{E}$ and $\vec{H}$ are orthogonal in the free space and are in phase, they are proportional to each other.

$$\underline{E}_y = Z_{F0}\underline{H}_x \qquad (2.16) \qquad\qquad \underline{E}_x = -Z_{F0}\underline{H}_y \qquad (2.17)$$

Z_{F0}: wave impedance of the free space





$H_x(z,t_1), E_y(z,t_1)$ $\quad$ $H_x(z,t_2), E_y(z,t_2)$

Snap-shots of E_y and H_x

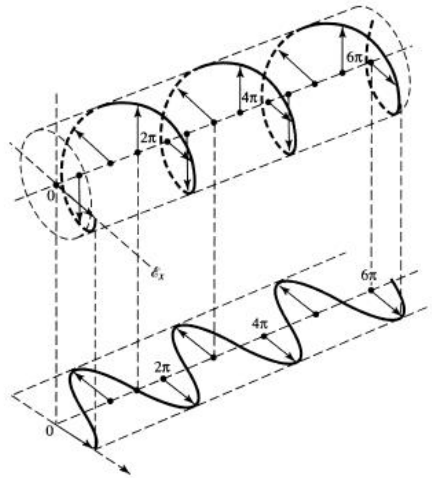
Polarisation

- Polarisation is the characteristic of a single-frequent electromagnetic wave that describes the amount and the direction of oscillation of the field vector as a function of place and time.
- In the far field of an antenna, for a plane wave the electric field vector is located in a plane vertical to the direction of propagation of the wave.
- Components of the electric field in x and y direction:

$$E_x = E_{x0}\cos(2\pi f t - \frac{2\pi}{\lambda} z + \phi_x) \qquad (2.18)$$

$$E_y = E_{y0}\cos(2\pi f t - \frac{2\pi}{\lambda} z + \phi_y) \qquad (2.19)$$

f: frequency λ: wavelength



General case of the elliptical polarisation

- The special case for linear polarisation is obtained for $\phi = \phi_x = \phi_y$.
- Surface of the earth as the reference for the direction of polarisation of a linearly polarised wave:
  vertical polarisation for E_x=0 and horizontal polarisation for E_y=0

Circular Polarisation

- For the unique characterisation of an elliptically and circularly polarised wave, the direction of rotation of the spiral has to be considered.
  - For this, a stationary plane vertical to the direction of propagation is considered and defined:
    - For an observer looking into the direction of propagation, the vector of the E-field in this plane for right-handed polarisation rotates clockwise

$$\Phi_x - \Phi_x = -\frac{\pi}{2} \qquad (2.20)$$

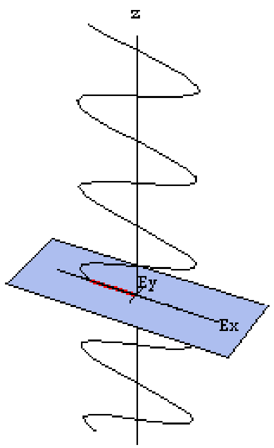
    - The common term in case of counter-clockwise rotation is left-hand polarisation

$$\Phi_x - \Phi_x = +\frac{\pi}{2} \qquad (2.21)$$

Examples of Polarisation

linear polarisation

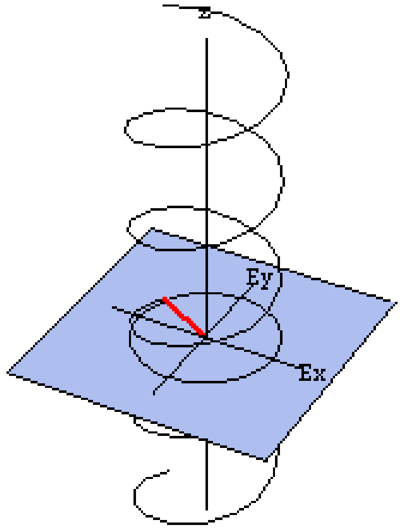
$E_{y0} = 0.$



left-handed circular polarisation

$E_{x0} = E_{y0}$

$\phi_x - \phi_y = \frac{\pi}{2}$

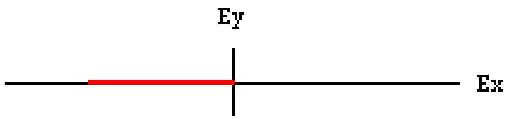


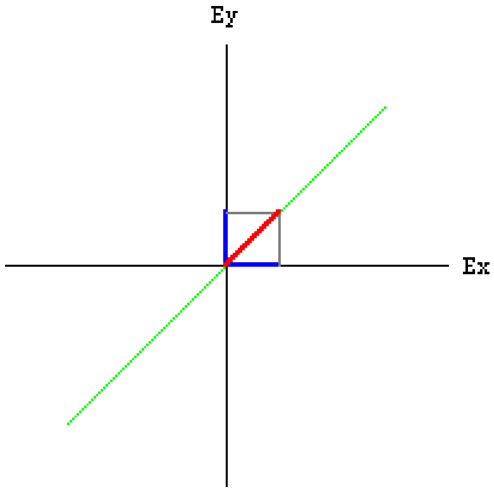
Created by Hsiu C. Han, 1996.

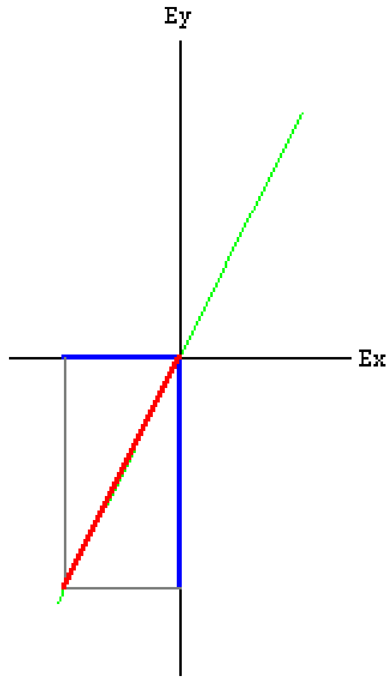
Source: http://vulcan.ece.iastate.edu/~hsiu/em_movies.html

Examples of Linear Polarisation

- Projection onto the x-y plane
  - Linear polarisation







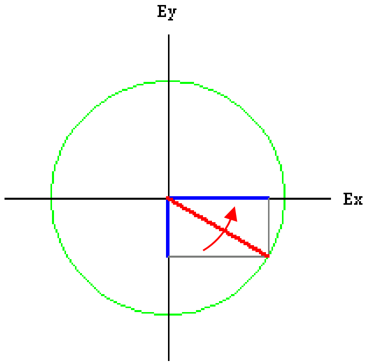
$E_{y0} = 0$

$E_{x0} = E_{y0}$
$\phi_x = \phi_y$

$E_{y0} = 2E_{x0}$
$\phi_x = \phi_y$

Examples of Circular Polarisation

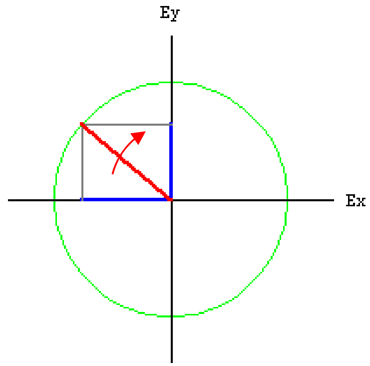
- Circular Polarisation



$$E_{x0} = E_{y0}$$

$$\phi_x - \phi_y = -\frac{\pi}{2}$$

right-handed circular polarisation



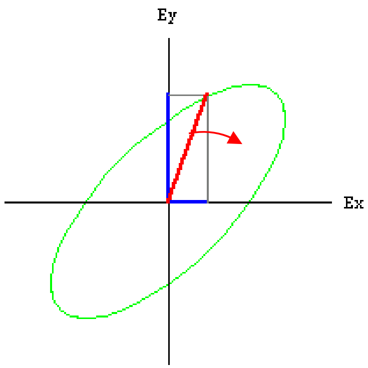
$$E_{x0} = E_{y0}$$

$$\phi_x - \phi_y = \frac{\pi}{2}$$

left-handed circular polarisation

Examples of Elliptic Polarisation

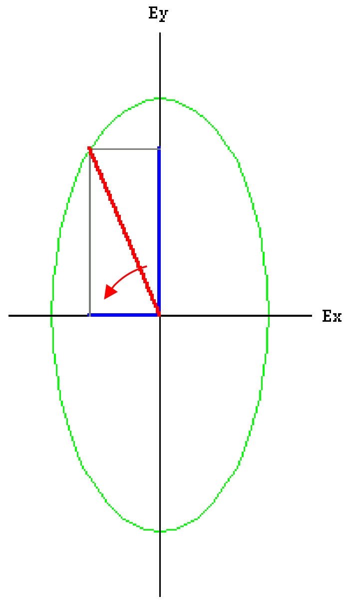
- General elliptic polarisation



$E_{x0} = E_{y0}$

$\phi_x - \phi_y = \frac{\pi}{4}$

left-handed elliptic polarisation



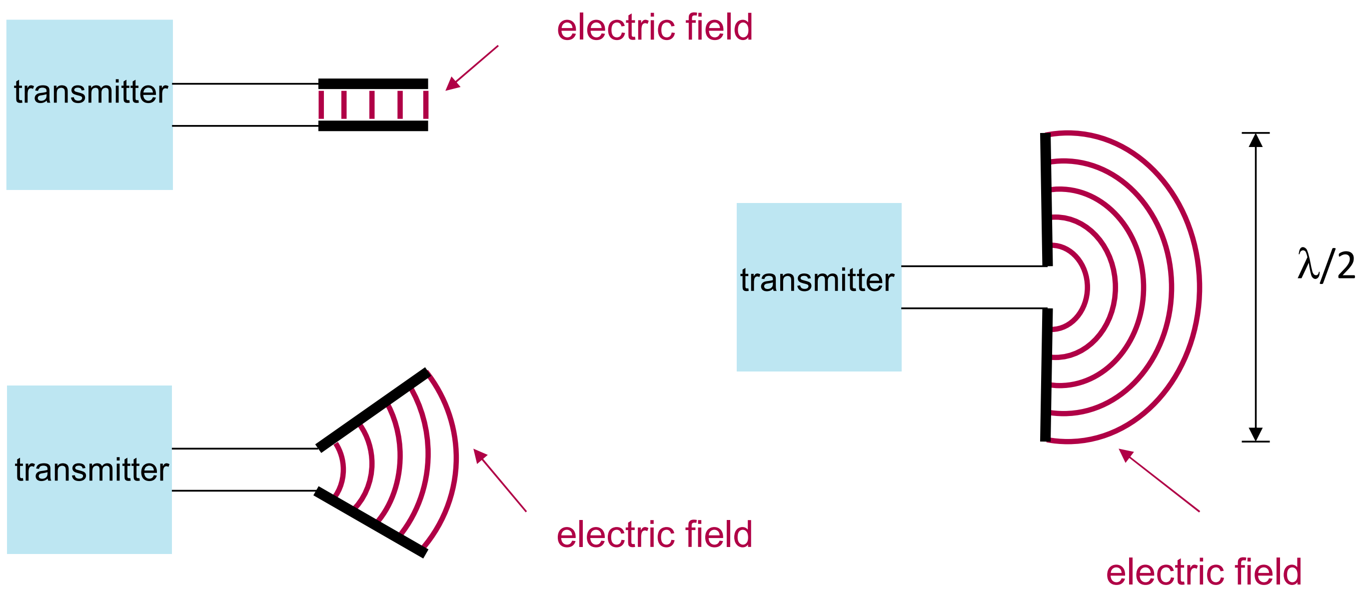
$E_{y0} = 2E_{x0}$

$\phi_x - \phi_y = -\frac{\pi}{2}$

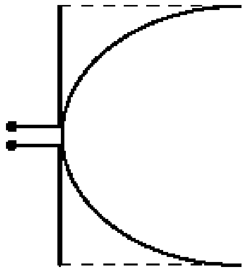
right-handed elliptic polarisation

3.3 Generation of Electromagnetic Waves

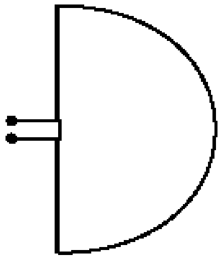
- Theoretical experiment:
  - Forming an antenna using a plate capacitor



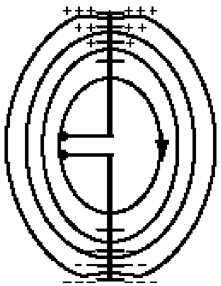
Field Distribution at a λ/2 Dipole



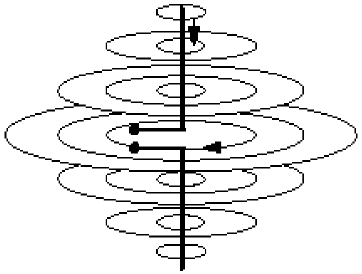
Spannungsverteilung (U)



Stromverteilung (I)



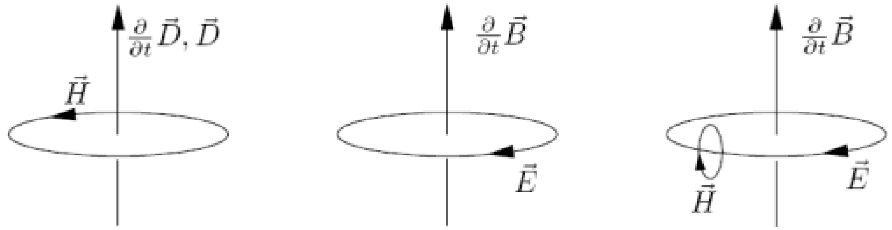
Elektrisches Feld (E)



Magnetisches Feld (H)

Radiation of Electromagnetic Waves from the Antenna

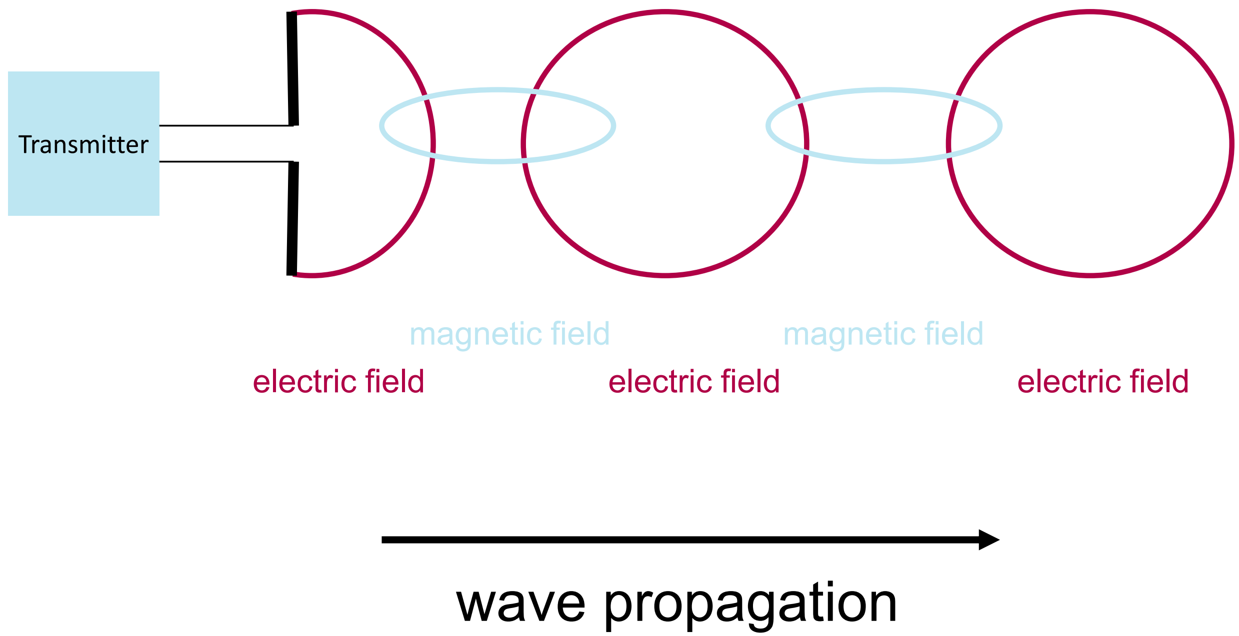
- Why do electromagnetic waves radiate from the antenna?
  - A time-variant electric field generates a magnetic rotational field.
  - A time-variant magnetic field generates an electric rotational field. In case that $\frac{\partial}{\partial t}\vec{B}$ is time-variant, the induced voltage generates another $\vec{H}$ field.
  - Electric and magnetic fields induce each other.



  - Electromagnetic waves are transverse waves propagating with the (phase) velocity

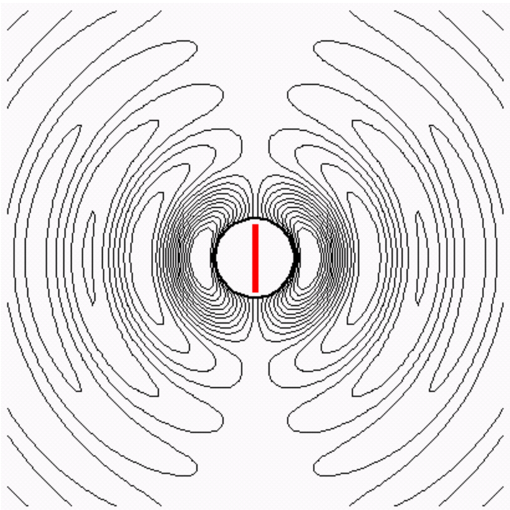
$$v = \frac{1}{\sqrt{\epsilon\mu}}\,.$$

Interaction between E- and H-Field

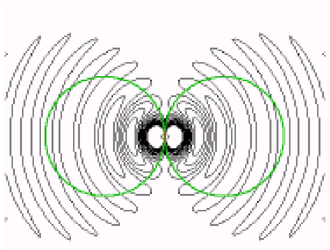


3.4 Antennas

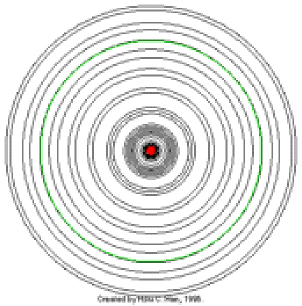
- Hertzian dipole
  - infinitesimal radiation source, e. g.
    - a short piece of wire of the length Δz, which is flown through by a current I_0
    - two spheres with the charge $+Q_0$ at intervals of Δz



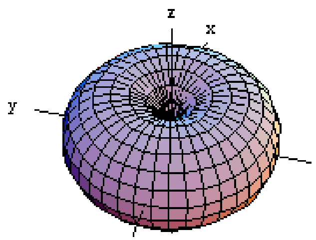
flux lines of the hertzian dipole



radiation pattern E plane



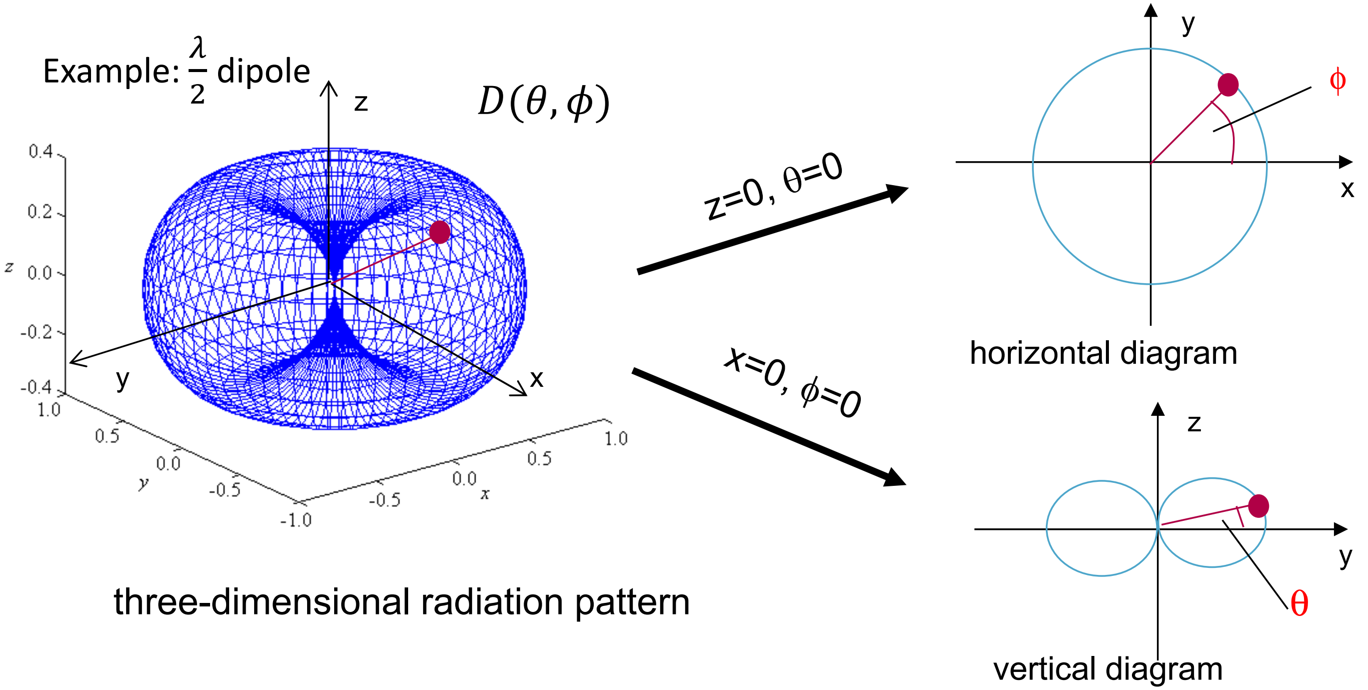
radiation pattern H plane



3D radiation pattern

Radiation Pattern of the λ/2 Dipole

- Directivity of antennas is described by a radiation pattern
  - radiation pattern $D(\theta,\phi)$: dimensionless quantity, absolute value between 0 and 1 (in linear presentation; in practice, often logarithmic values D_{dB})

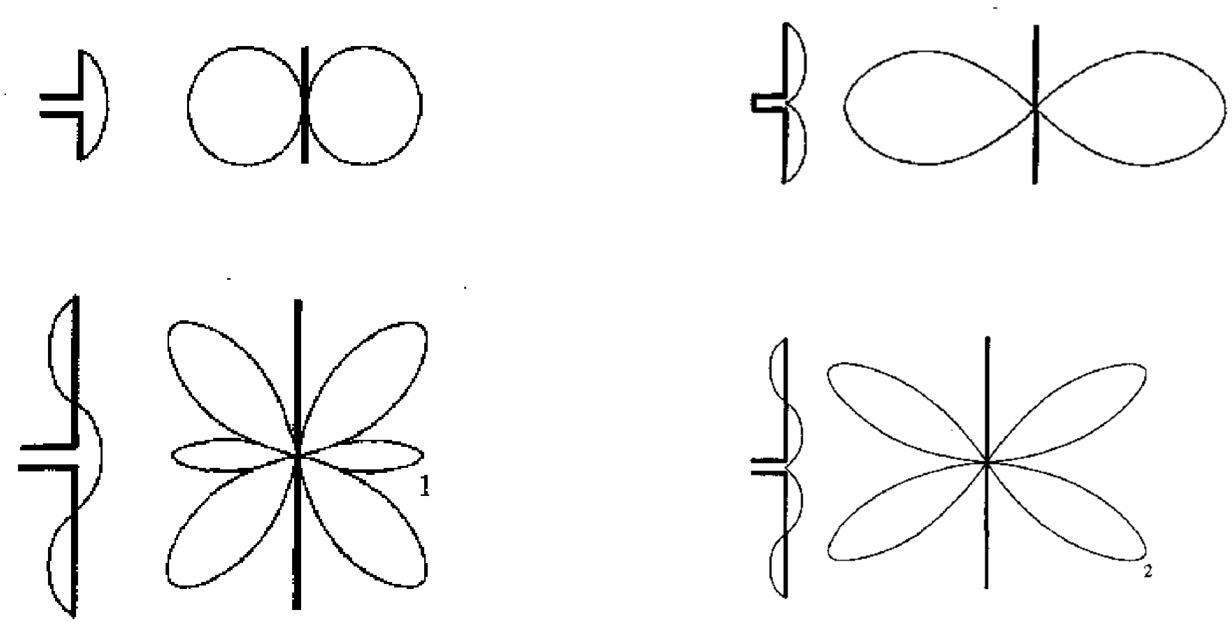


three-dimensional radiation pattern

Important Antenna Characteristics

- Half power beam width (3-dB width, HPBW):
  - characterises the angle around the radiation peak at the boundaries of which the radiation density is half as large as in the maximum (i.e. 3 dB less)
- Gain: $G = \dfrac{4\pi}{\int_0^{2\pi}\int_0^{\pi} D^2(\theta,\phi)\sin\phi \,\mathrm{d}\phi \,\mathrm{d}\theta}$ (2.22)
  - The gain is completely characterised by the radiation pattern. It is a measure for the characteristic of an antenna to preferably radiate energy to only one direction and to receive energy from only one direction, respectively.
  - The larger the dimension of an antenna (aperture) relating to the wavelength, the larger the gain and with it the directivity of this antenna.

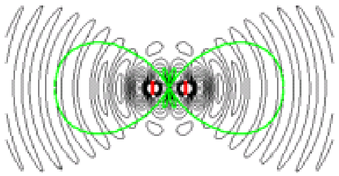
Current Distribution and Vertical Radiation Pattern of Simple Antennas

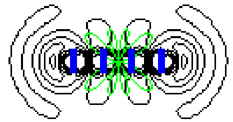


Groups of Antennas

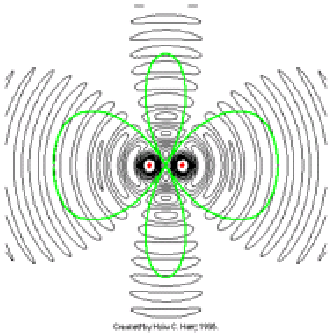
- Superposition of the fields of the single antennas

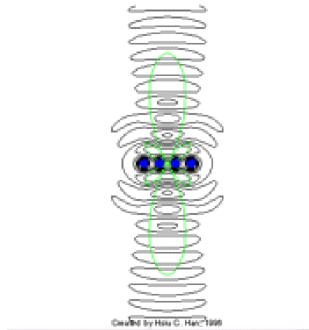
Radiation pattern E plane



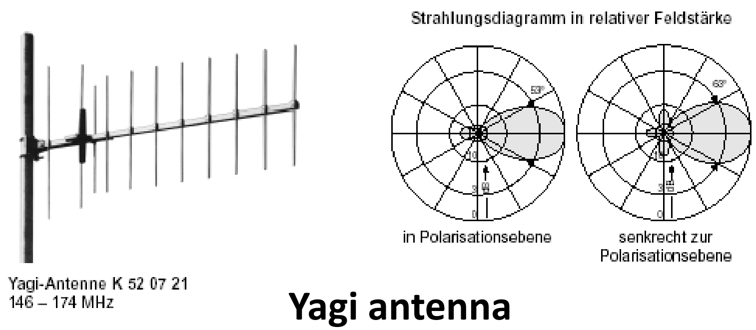


Radiation pattern H plane

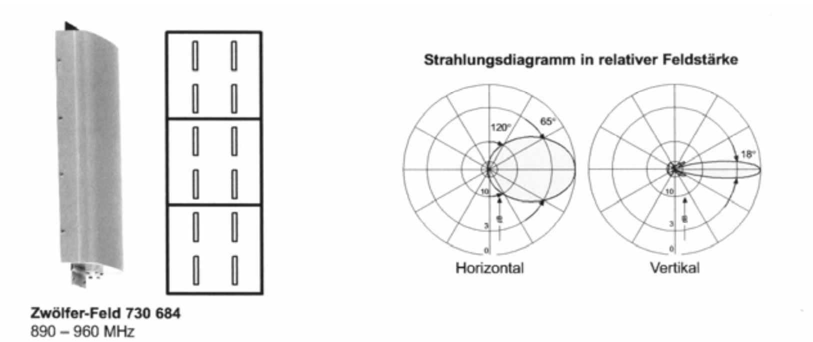




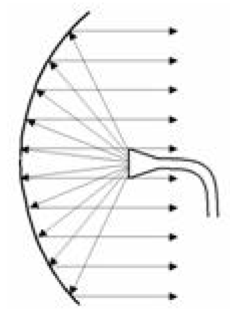
Antenna Designs (1)



Yagi antenna



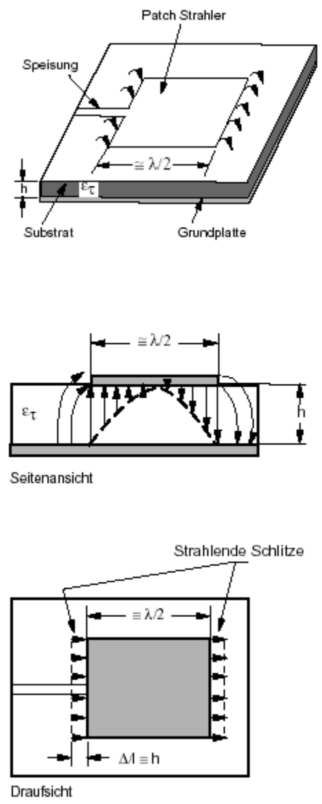
array of $\lambda/2$ dipoles



dish antenna

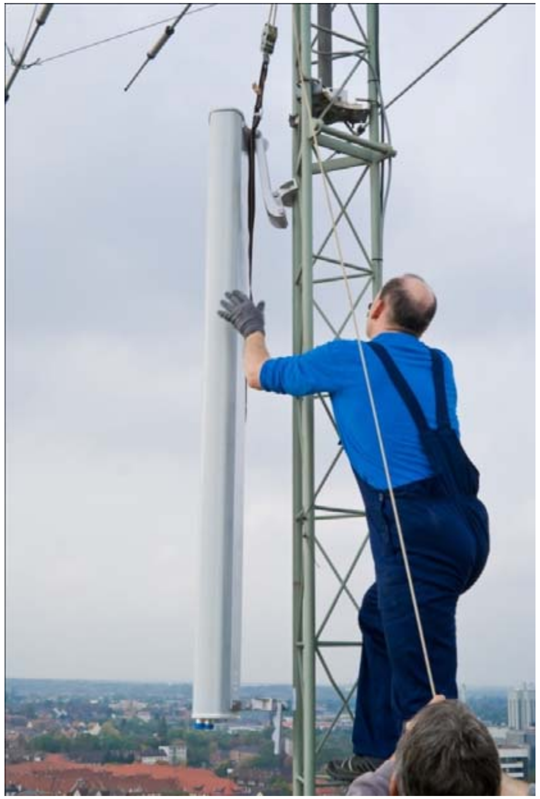
Antenna Designs (2)

patch antenna



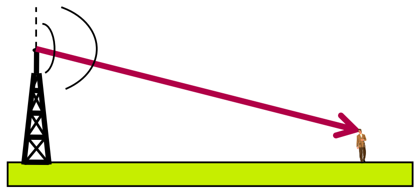
This is how an Antenna Installation for LTE800 looks like …

LTE800 test site at TU Braunschweig

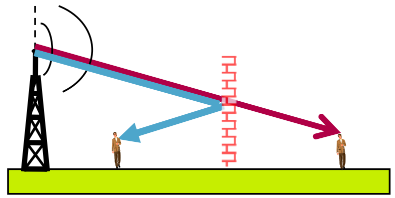


3.5 Propagation Mechanisms

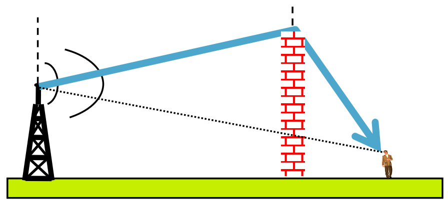
- Radio propagation is influenced by many factors and phenomena.
- In the frequency range interesting for mobile communications, the propagation characteristics can be considered as „quasi-optical“ in a first approximation.



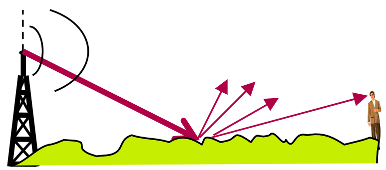
free-space propagation



reflection and transmission

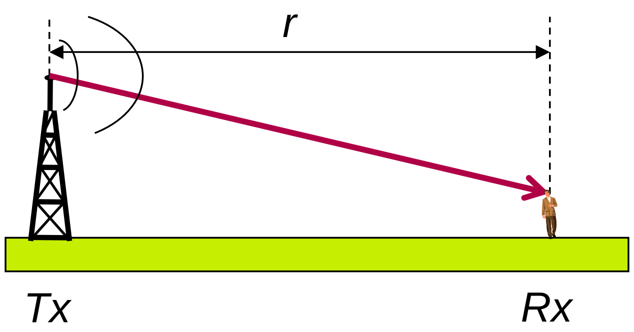


diffraction



scattering

Free-Space Propagation



$$P_\mathrm{R} = P_\mathrm{T}\underbrace{\frac{1}{r^2}\left(\frac{\lambda}{4\pi}\right)^2}_{}\underbrace{G_\mathrm{T}(\theta_\mathrm{T},\phi_\mathrm{T})G_\mathrm{R}(\theta_\mathrm{R},\phi_\mathrm{R})}_{} \quad (2.23)$$

— (green) influence of the free-space propagation

— (red) influence of the antenna

P_T: transmitting power of the transmitter (*Tx*)

P_R: received power at the receiver (*Rx*)

$G_\mathrm{T}(\theta_\mathrm{T},\phi_\mathrm{T})$: antenna diagram of the transmitter

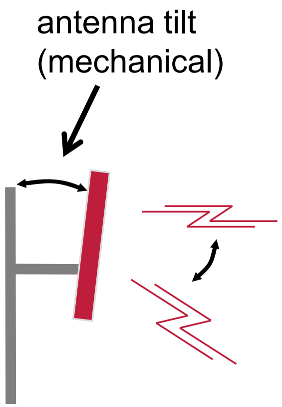
$G_\mathrm{R}(\theta_\mathrm{R},\phi_\mathrm{R})$: antenna diagram of the receiver

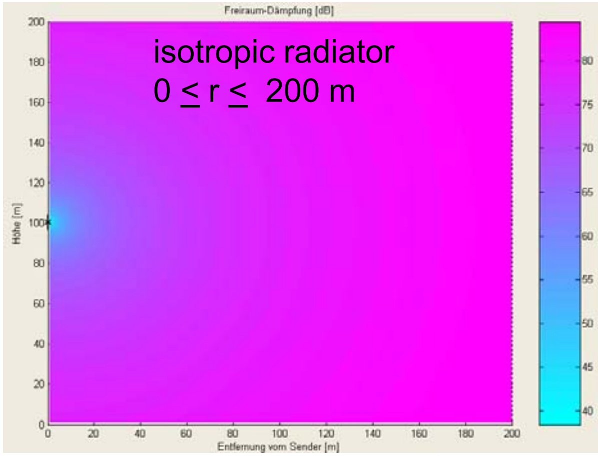
Note: Here and in the following, the downlink case is considered only. Since propagation is reciprocal, however, the same attenuations apply to the uplink.

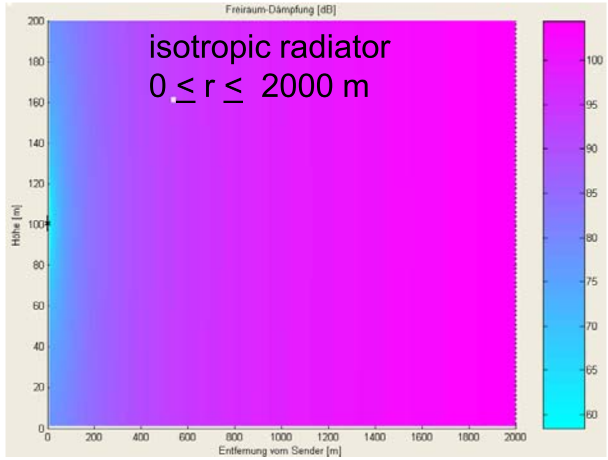
- free-space attenuation in dB: $L_\mathrm{dB,F} = 32{,}4 + 20\log\frac{r}{\mathrm{km}} + 20\log\frac{f}{\mathrm{MHz}}$ (2.24)

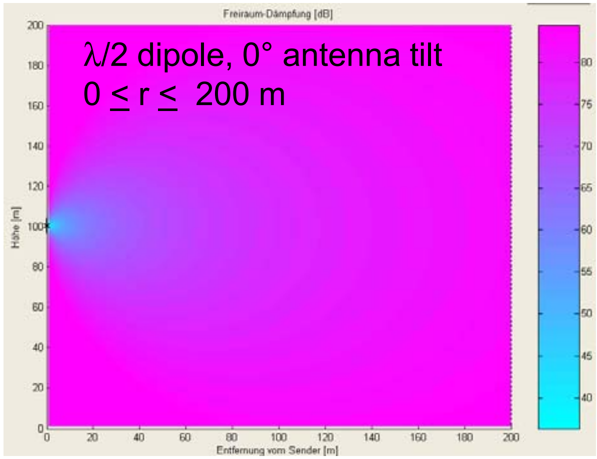
Examples of Free-Space Loss at 2 GHz

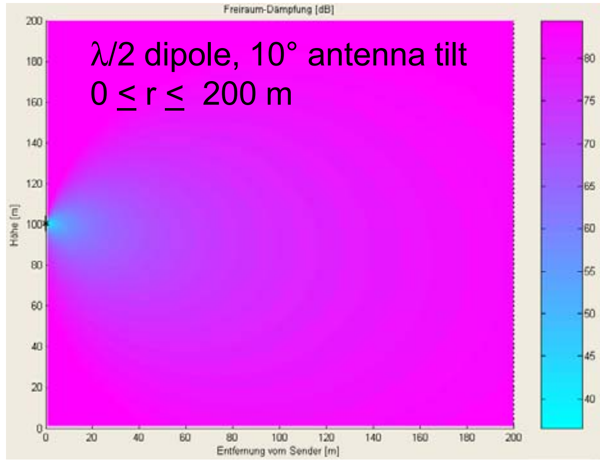
antenna height 100 m





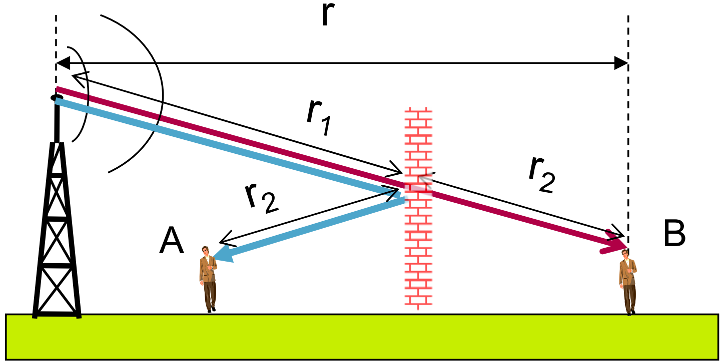






3.6 Reflection and Transmission

- Splitting of the field into incident, reflected and transmitted contributions
- Additional reflection loss (polarisation-dependent reflection factor $r_{\mathrm{TE/TM}}$) for the reflected contribution
- Additional transmission loss (polarisation-dependent transmission factor $t_{\mathrm{TE/TM}}$) for the transmitted contribution

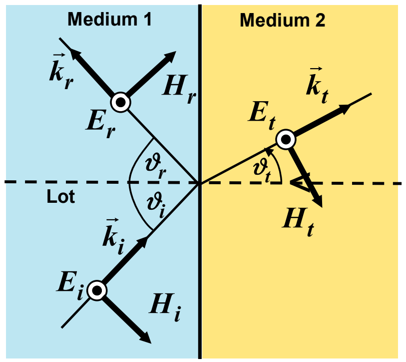


$$P_{\mathrm{R,A}} = P_{\mathrm{T}} \frac{1}{(r_1+r_2)^2} \left(\frac{\lambda}{4\pi}\right)^2 G_{\mathrm{T}}(\theta_{\mathrm{T}}, \phi_{\mathrm{T}}) G_{\mathrm{R}}(\theta_{\mathrm{R}}, \phi_{\mathrm{R}}) \underline{r}_{\mathrm{TE/TM}}\, \underline{r}^{*}_{\mathrm{TE/TM}} \quad (2.25)$$

$$R_{\mathrm{R,B}} = P_{\mathrm{T}} \frac{1}{(r_1+r_2)^2} \left(\frac{\lambda}{4\pi}\right)^2 G_{\mathrm{T}}(\theta_{\mathrm{T}}, \phi_{\mathrm{T}}) G_{\mathrm{R}}(\theta_{\mathrm{R}}, \phi_{\mathrm{R}}) \underline{t}_{\mathrm{TE/TM}} \underline{t}^{*}_{\mathrm{TE/TM}} \quad (2.26)$$

Fresnel‘s Reflection and Transmission Factors

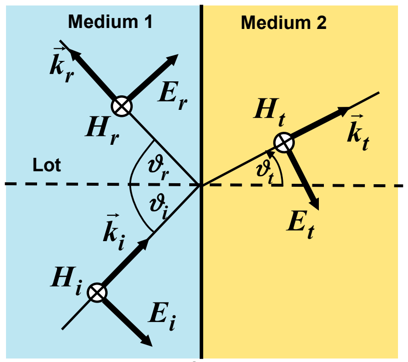
general case — special case for $\underline{\epsilon}_1 = 1; \underline{\epsilon}_2 = \underline{\epsilon}_r$



a) TE polarisation, E field⊥ incidence plane

$$\underline{r}_{\mathrm{TE}} = \frac{\cos(\vartheta_i) - \sqrt{\frac{\underline{\epsilon}_2}{\underline{\epsilon}_1}}\cos(\vartheta_t)}{\cos(\vartheta_i) + \sqrt{\frac{\underline{\epsilon}_2}{\underline{\epsilon}_1}}\cos(\vartheta_t)} \qquad \underline{r}_{\mathrm{TE}} = \frac{\cos(\vartheta_i) - \sqrt{\underline{\epsilon}_r - \sin^2(\vartheta_i)}}{\cos(\vartheta_i) + \sqrt{\underline{\epsilon}_r - \sin^2(\vartheta_i)}} \qquad (2.27)$$

$$\underline{t}_{\mathrm{TE}} = \frac{2\cos(\vartheta_i)}{\cos(\vartheta_i) + \sqrt{\frac{\underline{\epsilon}_2}{\underline{\epsilon}_1}}\cos(\vartheta_t)} \qquad \underline{t}_{\mathrm{TE}} = \frac{2\cos(\vartheta_i)}{\cos(\vartheta_i) + \sqrt{\underline{\epsilon}_r - \sin^2(\vartheta_i)}} \qquad (2.28)$$



b) TM polarisation, H field ⊥ incidence plane

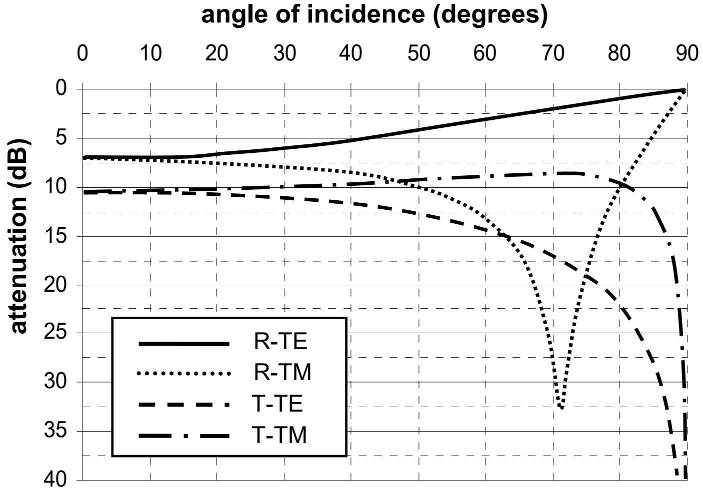
$$\underline{r}_{\mathrm{TM}} = \frac{\sqrt{\frac{\underline{\epsilon}_2}{\underline{\epsilon}_1}}\cos(\vartheta_i) - \cos(\vartheta_t)}{\sqrt{\frac{\underline{\epsilon}_2}{\underline{\epsilon}_1}}\cos(\vartheta_i) + \cos(\vartheta_t)} \qquad \underline{r}_{\mathrm{TM}} = \frac{\underline{\epsilon}_r\cos(\vartheta_i) - \sqrt{\underline{\epsilon}_r - \sin^2(\vartheta_i)}}{\underline{\epsilon}_r\cos(\vartheta_i) + \sqrt{\underline{\epsilon}_r - \sin^2(\vartheta_i)}} \qquad (2.29)$$

$$\underline{t}_{\mathrm{TM}} = \frac{2\cos(\vartheta_i)}{\sqrt{\frac{\underline{\epsilon}_2}{\underline{\epsilon}_1}}\cos(\vartheta_i) + \cos(\vartheta_t)} \qquad \underline{t}_{\mathrm{TM}} = \frac{2\sqrt{\underline{\epsilon}_r}\cos(\vartheta_i)}{\underline{\epsilon}_r\cos(\vartheta_i) + \sqrt{\underline{\epsilon}_r - \sin^2(\vartheta_i)}} \qquad (2.30)$$

For all cases applies: $\underline{\mu}_1 = \underline{\mu}_2 = 1$, incidence is stretched by the wave vector and the normal vector of the reflection plane.

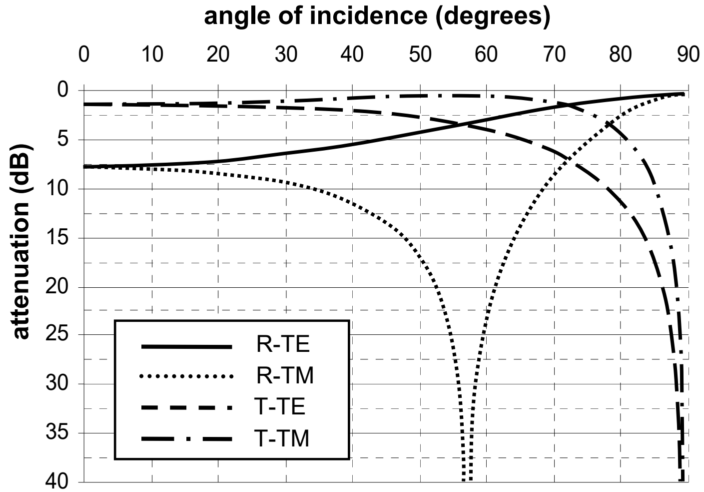
Examples of Reflection and Transmission Loss

- In the lossy medium, the transmission loss also depends on the thickness of the layer.
- Example: characteristics of concrete and wood at 1 GHz



Schichtdicke in cm	ε'	$\tan\delta$
30	9	0.1

concrete



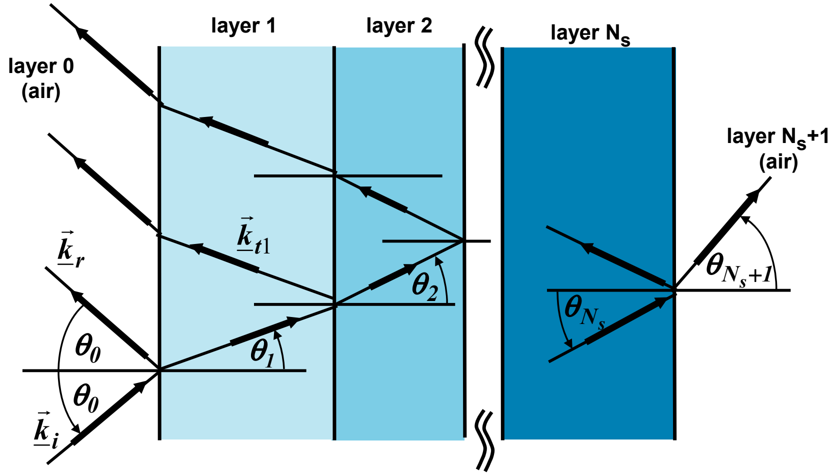
Schichtdicke in cm	ε'	$\tan\delta$
5	2.5	0.01

wood

Source: D. Cichon

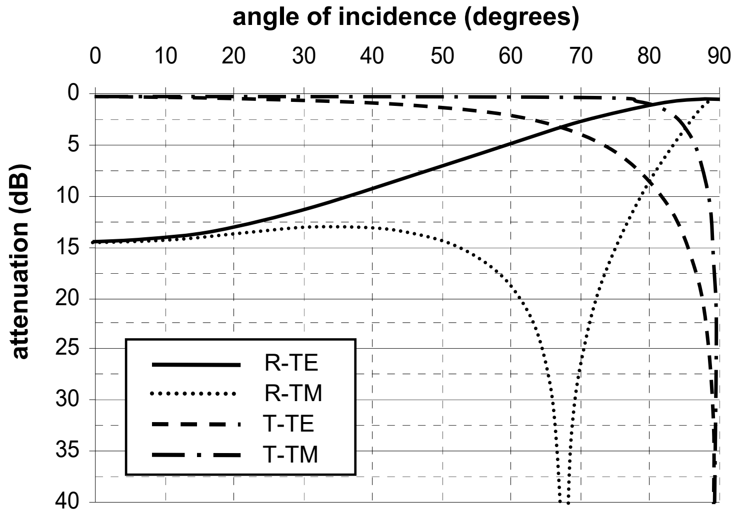
Modelling of Layers of Building Walls, Windows etc.

- A building wall or a window is described by a multi-layered arrangement of homogeneous layers of materials of a finite thickness
- Assumption that interfaces are plane and smooth
- Each layer is described by the real component of the relative permittivity, loss factor (tan δ), conductivity, thickness of layer



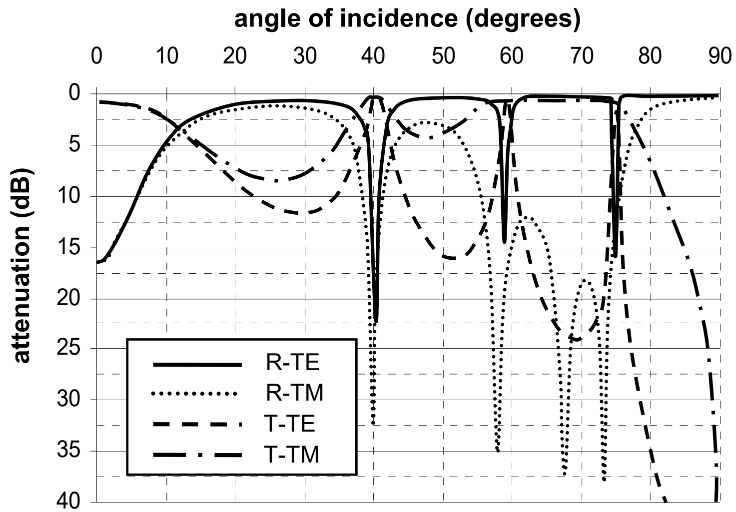
- Calculation by application of the so-called wave matrix method
- Special feature: explicit frequency dependence, even in case all layers consist of non-dispersive materials

Example: 3-Layer Window



Schichtdicke in cm	ε′	tanδ
0.3 / 2 / 0.3	6 / 1 / 6	0.008 / 0 / 0.008

2 GHz

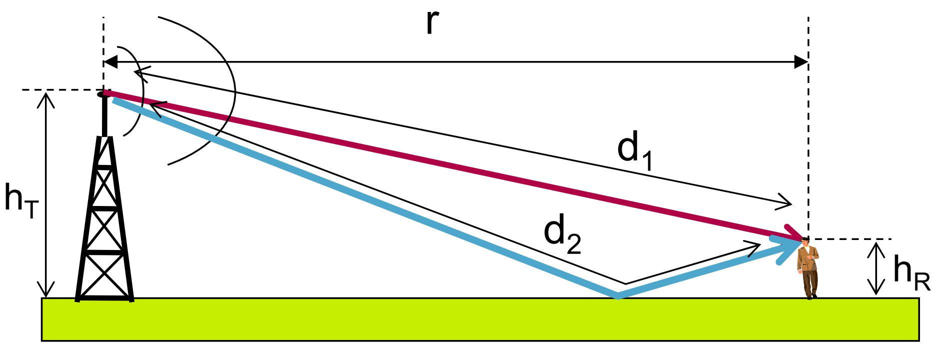


Schichtdicke in cm	ε′	tanδ
0.3 / 2 / 0.3	6 / 1 / 6	0.008 / 0 / 0.008

30 GHz

Source: D. Cichon

3.7 Ground Reflection and Two-Ray Theory



- At the receiver, superposition between the direct ray and a ray reflected on the surface of the earth can arise.
- For the received power, the following applies:

$$P_R = P_T \frac{1}{r^2} \left(\frac{\lambda}{4\pi}\right)^2 G_{ZS}(r, h_T, h_R) G_T(\theta_T, \phi_T) G_R(\theta_R, \phi_R) \qquad (2.31)$$

  - with the term $\frac{1}{r^2} G_{ZS}$ describing the attenuation resulting from the two rays

Standard Form of the Two-Ray Theory

- Neglecting the directional pattern of the antenna at first (valid for large distances r), G_{ZS} results from the complex superposition of both signal parts:

$$\tilde{G}_{\mathrm{ZS},v,h} = \left| \frac{\mathrm{e}^{-\mathrm{j}\frac{2\pi}{\lambda}d_1}}{\frac{d_1}{\mathrm{m}}} + \underline{r}_{\mathrm{TE,TM}}(\vartheta_i, \epsilon_r)\ \frac{\mathrm{e}^{-\mathrm{j}\frac{2\pi}{\lambda}d_2}}{\frac{d_2}{\mathrm{m}}} \right|^2 \qquad (2.32a)$$

with $$\tilde{G}_{\mathrm{ZS},v,h} = \frac{1}{r^2} G_{\mathrm{ZS},v,h} \qquad (2.32b)$$

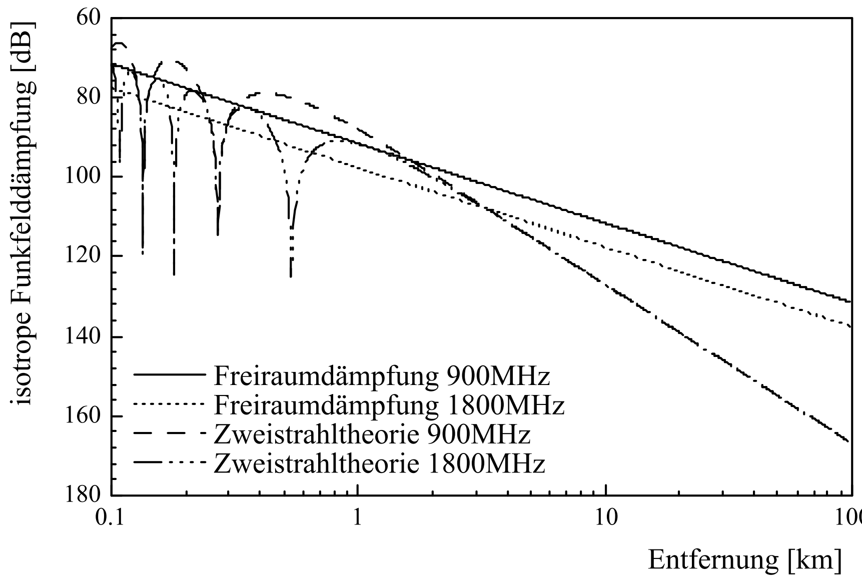
Approximation for large distances ($r >> h_T$ and h_R):

- In the terms for the amplitude applies: $d_1 \approx d_2 \approx r$ (2.33)
- In the terms for the phase applies: $d_2 - d_1 \approx \frac{2h_\mathrm{R}h_\mathrm{T}}{d}$ (2.34)
- A metallic interface is assumed: $\underline{r}_{\mathrm{TE,TM}} = \pm 1$

Two-Ray Model

- Transmission loss of the two-ray model

$$G_{ZS,v,h} = \begin{cases} 4\cos^2\frac{2\pi h_T h_R}{\lambda r}, \text{vertical polarisation} \\ 4\sin^2\frac{2\pi h_T h_R}{\lambda r}, \text{horizontal polarisation} \end{cases} \quad (2.35)$$

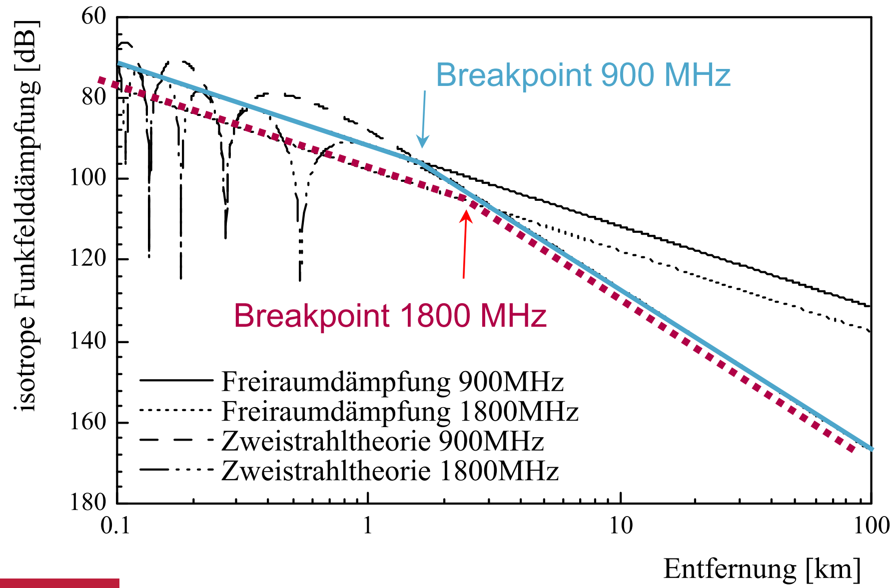


Source: N. Geng

- Example: isotropic transmission loss according to the two-ray theory for horizontal polarisation (h_T = 30 m, h_R = 1.5 m)

Break Point and Dual-Slope Approach

- Consideration of the envelope at horizontal polarisation
  - for small distances $G_{ZS}/r^2 \sim 1/r^2$ (complies with 20 dBd/decade)
  - for large distances $G_{ZS}/r^2 \sim 1/r^4$ (complies with 40 dB/decade)



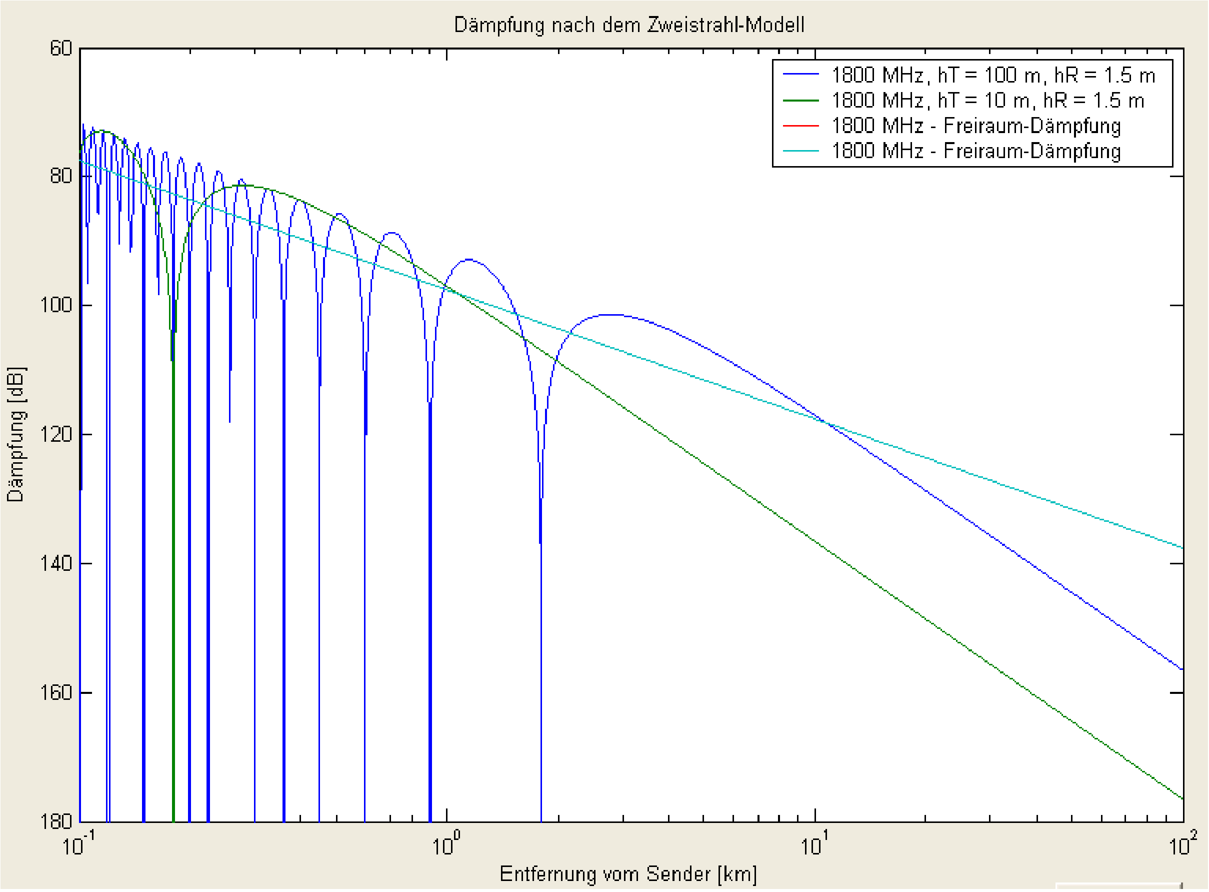
- break point distance:

$$r_{\text{break point}} = \frac{4\pi h_R h_T}{\lambda} \qquad (2.36)$$

- vitally practical importance; for real ground characteristics r, approximately valid for horizontal and vertical polarisation!

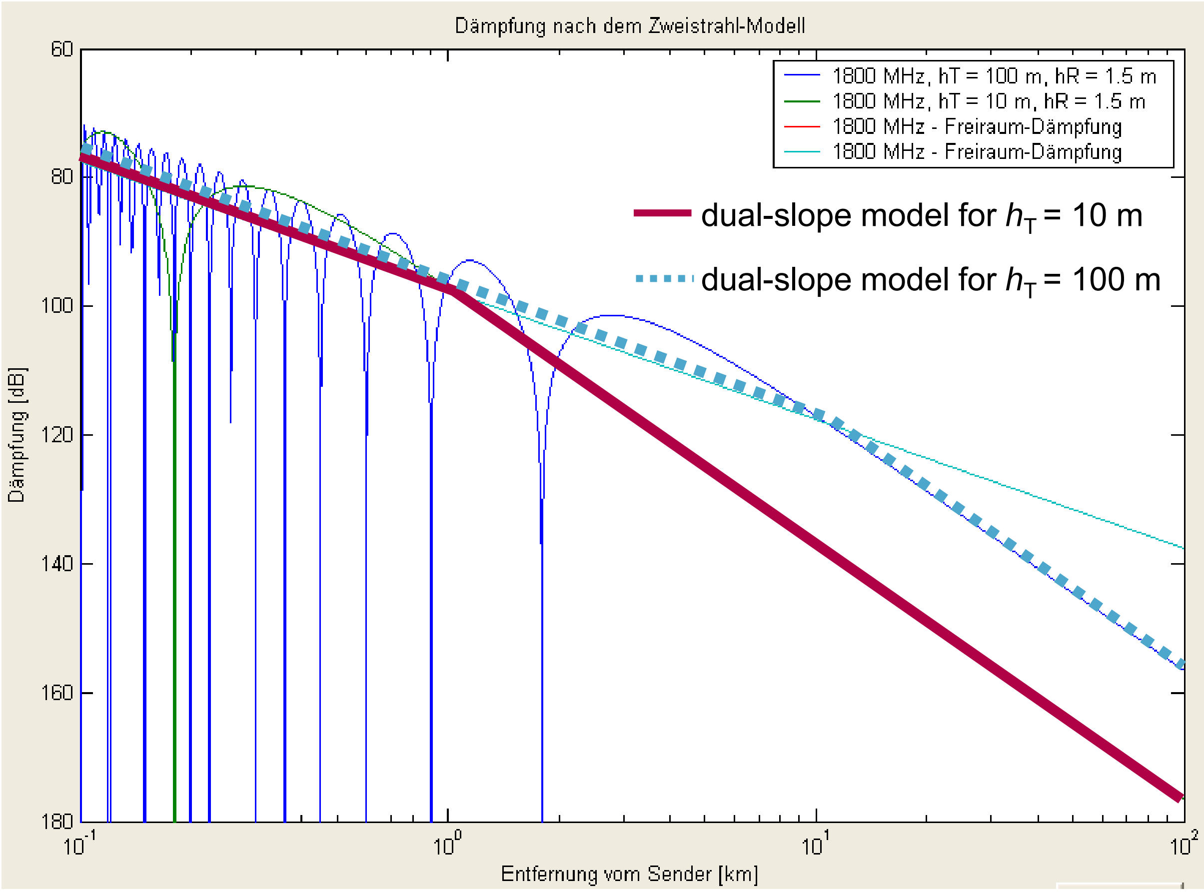
Example of Attenuation with the Two-Ray Model

dependency of the attenuation characteristic on h_T



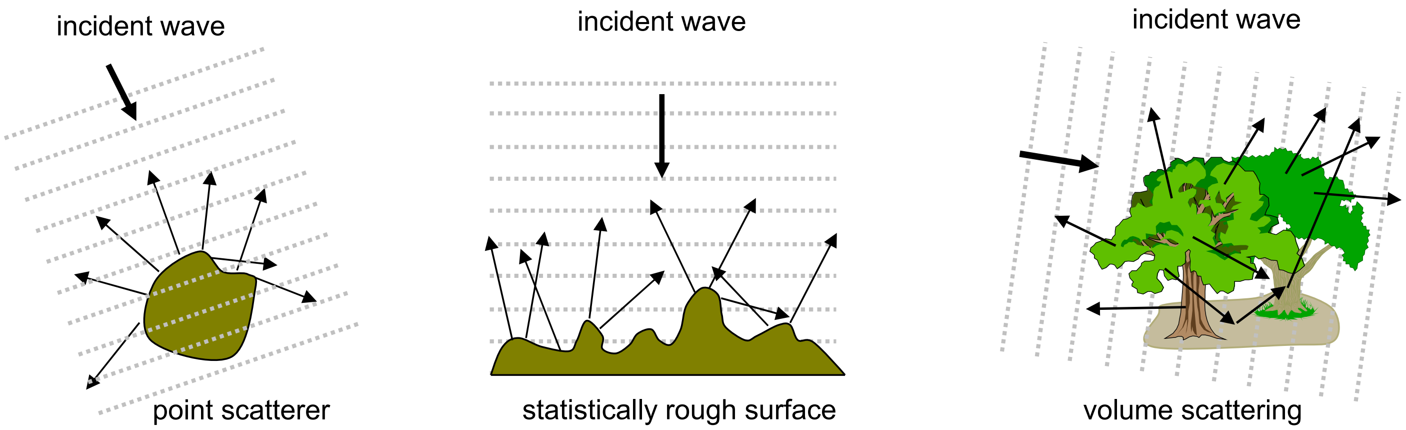
Example of Attenuation at the Two-Ray Model

dependence of the attenuation characteristic on h_T



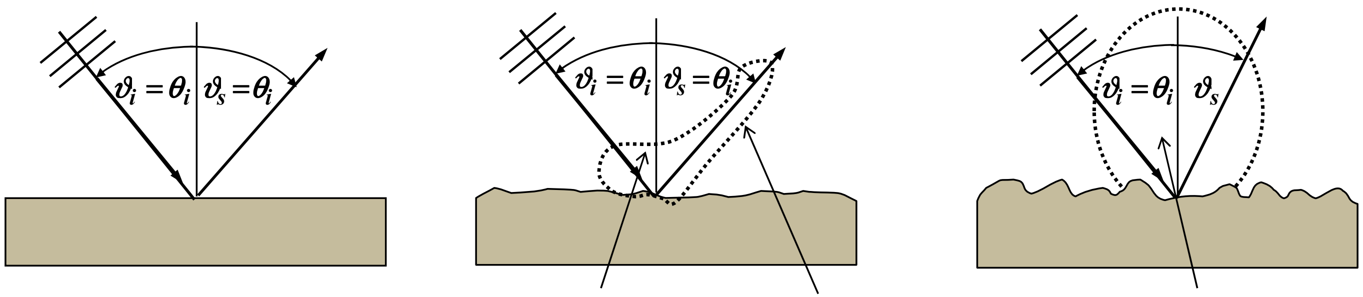
3.8 Scattering

- Scattering denotes the deviation of radiant energy from the original propagation direction occurring in an inhomogeneous medium, i.e. the complete power not reflected to mirror direction is scattered to other directions.
- The smaller the ratio object size to wavelength, the larger the deviation of the interaction of a wave with a single object from the behaviour considering
- Also for rough surfaces and groups of scatterers, treatment only be reflection and transmission is not sufficient.



Scattering at Rough Surfaces

- An ideally smooth, plane interface reflects an incident, locally plane wave in a single direction given by the reflection law.
- With increasing roughness, scattering in other directions (purely coherent scattering for slightly rough surfaces)
- For very high roughness or varying heights and slopes, the preferred direction of the mirrored reflection is no longer observable.



Criteria for Roughness

- Standard deviation σ of the surface roughness as a measure for the mean variation in height compared with the mean value as well as the correlation length I_{corr} as a measure for the statistical relation of two points on the rough surface at intervals of Δr
- The ratio σ to wavelength λ is decisive for the differentiation scattering/reflection.
- A surface can be considered smooth in case the Fraunhofer criterion is fulfilled:

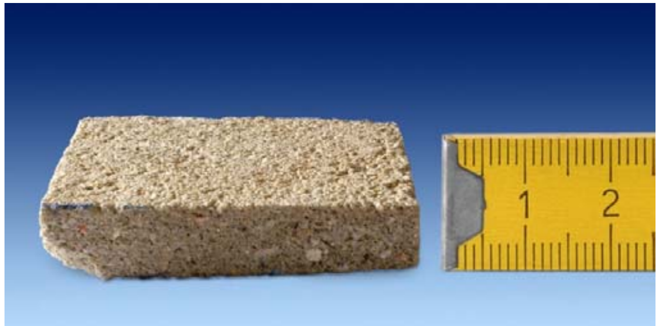
$$\sigma < \frac{\lambda}{32 \cos \theta_i} \tag{2.37}$$

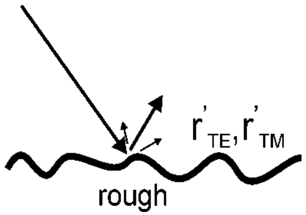
- For very large areas (measurements >> λ) and low roughness σ << λ, an approximation for the reduction of the field strength amplitude in the reflected direction (modified Fresnel reflection factors) exists:

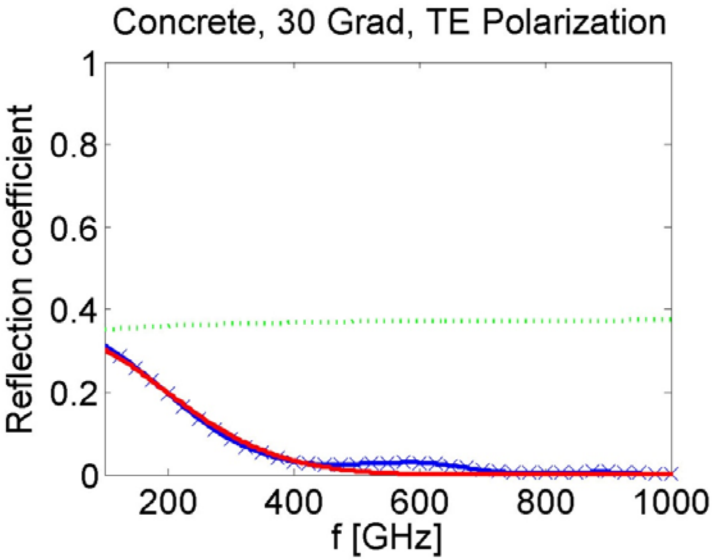
$$\underline{r}_{\mathrm{TE/TM}}^{\mathrm{mod}} = \underline{r}_{\mathrm{TE/TM}} \mathrm{e}^{-8\pi^2 \left(\frac{\sigma}{\lambda}\right)^2 \cos^2 \theta_i} \tag{2.38}$$

- For definite validity ranges of σ/λ and $I_{\mathrm{corr}}/\lambda$, analytical methods of calculation (Kirchhoff theory, small perturbation method) for calculation of the scattered parts exist.

Example: Reflection on Rough Plaster (σ=0.15mm) Beyond 100 GHz





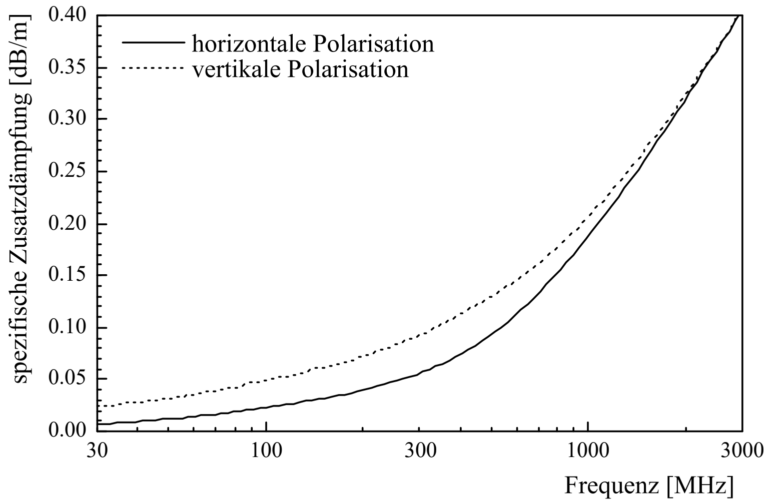


Volume Scattering and Attenuation due to Vegetation

- Estimation of the volume scattering
  - Modelling of the interaction of an electromagnetic wave with a volume, in which scattering objects of different shapes, sizes, orientation and material parameters are statistically distributed, is very complex (e. g. by Radiative Transfer Theory) and, mostly, only possible by approximation.
- Estimation of the attenuation without vegetation
  - simple empirical approximation procedure for the loss on passing through a vegetation layer (CCIR-Rep. 236-5):

$$L_{\mathrm{dB,veg}} = \eta \left(\frac{f}{\mathrm{MHz}}\right)^{\nu} \left(\frac{d}{\mathrm{m}}\right)^{\gamma} \qquad (2.39)$$

  - $L_{\mathrm{dB,veg}}$ denotes an excess loss that has to be added to the transmission loss arising without vegetation.
  - typical values: $\eta = 0.187; \nu = 0.284; \gamma = 0.588$



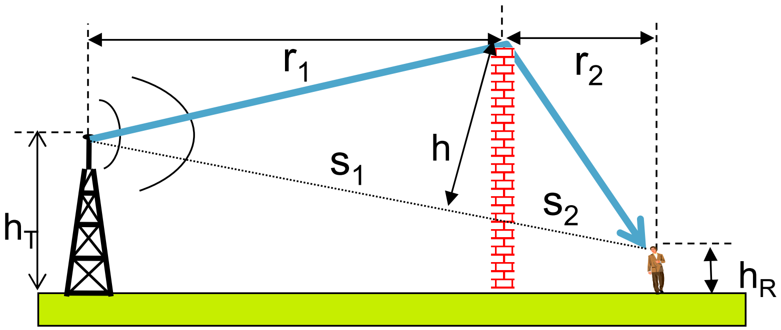
specific excess loss according to ITU-R Rec. 833

3.9 Diffraction

- Electromagnetic waves can move round obstacles and thus can reach the geometrical coverage gap. It depends on the wavelength as well as on the shape of the obstacle how much the field is diffracted, i.e. how much it penetrates the shadow region.

As a first approximation,

diffraction effects at sharp-edged obstacles can be described with the help of absorbing half-planes (knife edges) that are introduced vertically to the connecting line transmitter-receiver.



- For the received power, the following applies:

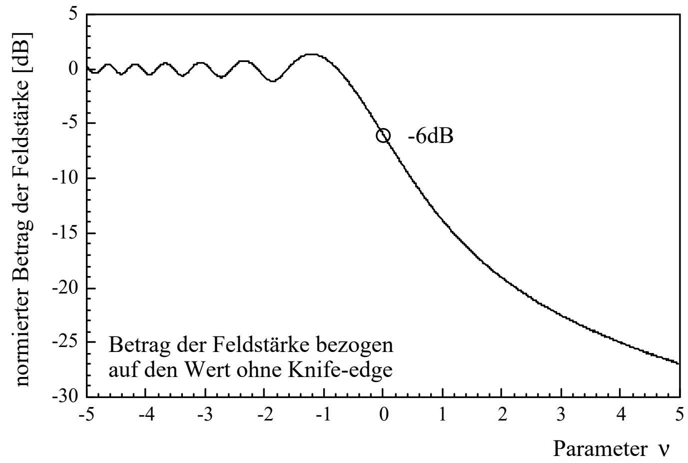
$$P_\mathrm{R} = P_\mathrm{T} \frac{1}{(r_1+r_2)^2} \left(\frac{\lambda}{4\pi}\right)^2 G_\mathrm{T}(\theta_\mathrm{T}, \phi_\mathrm{T}) G_\mathrm{R}(\theta_\mathrm{R}, \phi_\mathrm{R}) L_\mathrm{B} \qquad (2.40)$$

Attenuation for Knife-Edge Diffraction

$$L_{\mathrm{B}} = 10^{-\frac{\tilde{L}_{\mathrm{B}}}{10}} \qquad (2.41)$$

$$\tilde{L}_{\mathrm{dB,B}} = 6.9 + 20\log(\sqrt{(\nu-0.1)^2+1} + \nu - 0.1) \qquad (2.42)$$

$$\nu = h^* \sqrt{\frac{2}{\lambda}\left(\frac{s_1+s_2}{r_1 r_2}\right)} \qquad (2.43)$$

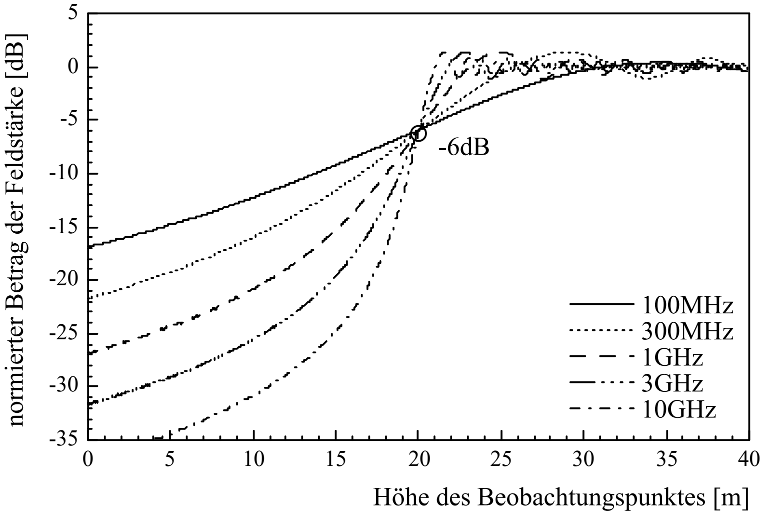


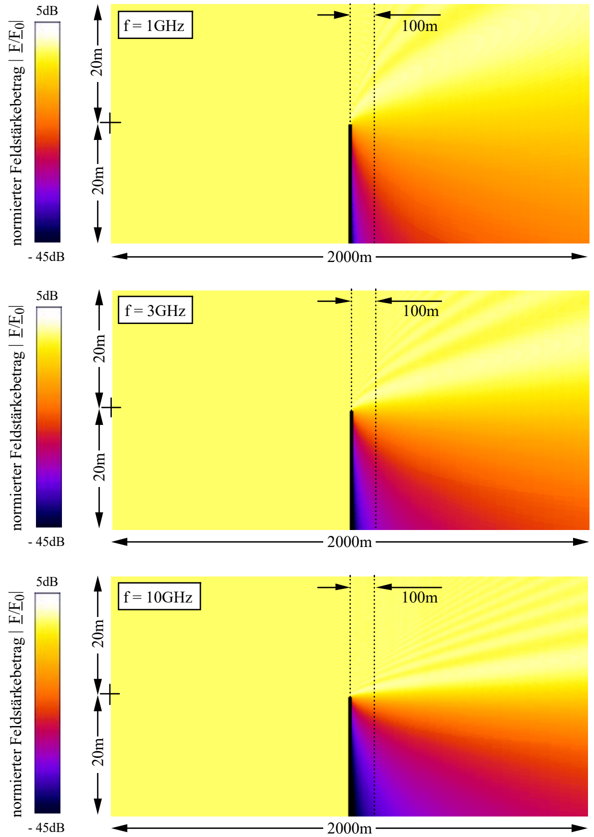
Source: N. Geng

- Although sharp-edged obstacles rarely occur in reality, the attenuation behaviour at realistic obstacles can be well approached by Knife-Edge diffraction.

Variation with Frequency

Normalised amplitude of the field strength as a function of the height of the observation point at a distance of 100m behind the half-plane (calculation with the Parabolic Equation Method):

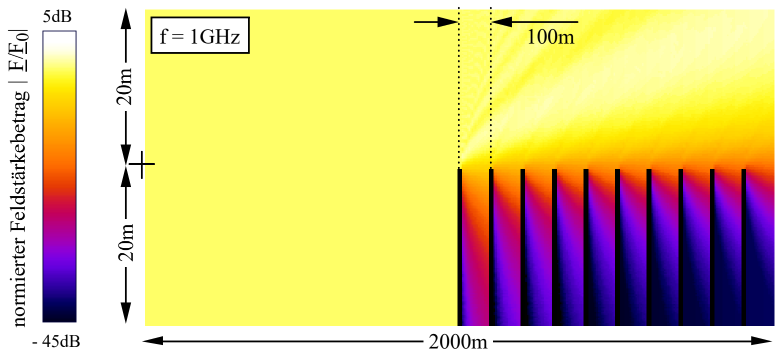




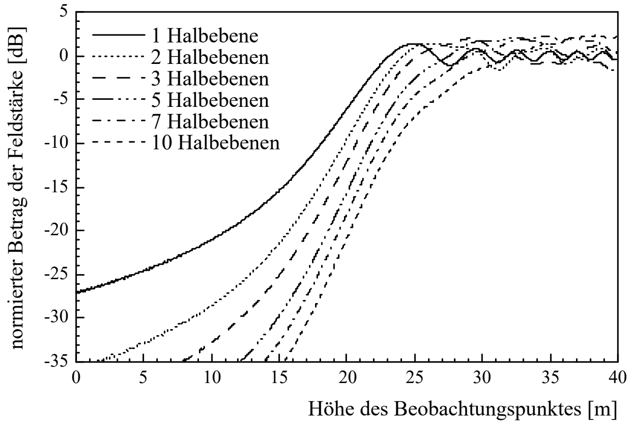
Source: N. Geng

Multiple Diffraction

- In practise, for lots of problems a half-plane is not sufficient.
- Substitution of obstacles by a set-up of several half-planes



Relative field strength in a vertical cutting plane at 1 GHz



Height dependency of the relative field strength at intervals of 100 m from one half-plane to the next

Source: N. Geng

Refraction and Effective Radius of the Earth

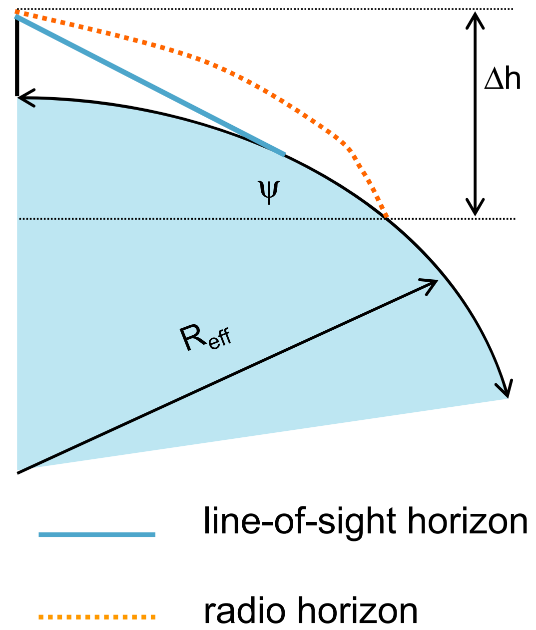
- Defining differences in height along a terrain profile, the curvature of the earth has to be considered.
  - projection of the height onto a sphere with the radius
    $R = R_{earth} = 6370\ \mathrm{km}$

- Tropospheric refraction effects enlarge the coverage of electromagnetic waves in the VHF/UHF range.

  - Extension of the radio horizon compared to the sight horizon can be considered by the effective radius of the earth

$$R_{\mathrm{Eff}} = k \cdot R_{\mathrm{earth}} \qquad (2.44)$$

  - For Europe $k = 4/3$ applies.



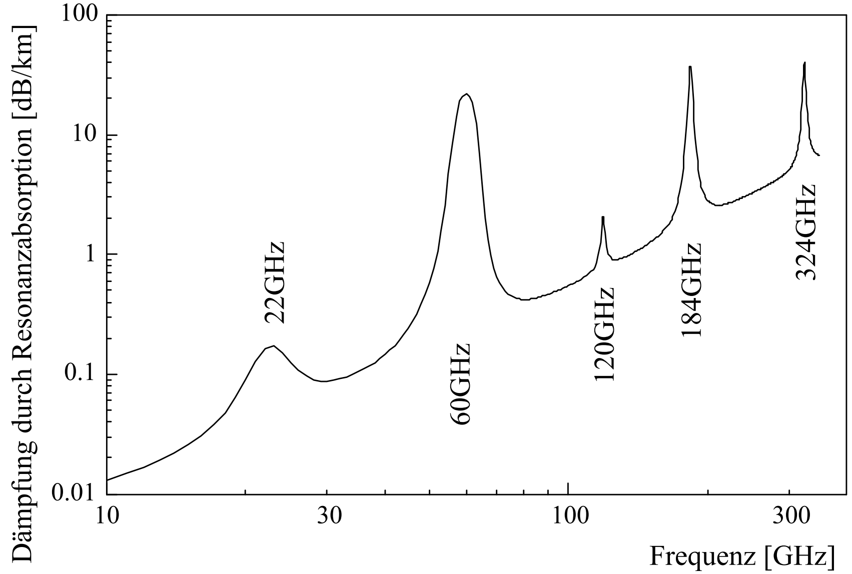
3.10 Resonant Absorption

- Resonant absorption occurs in case of discrete frequencies that correspond to the natural oscillation frequencies of the molecules included in the atmosphere.

water vapour:
22 GHz, 184 GHz, 324 GHz

oxygen:
60 GHz, 120 GHz

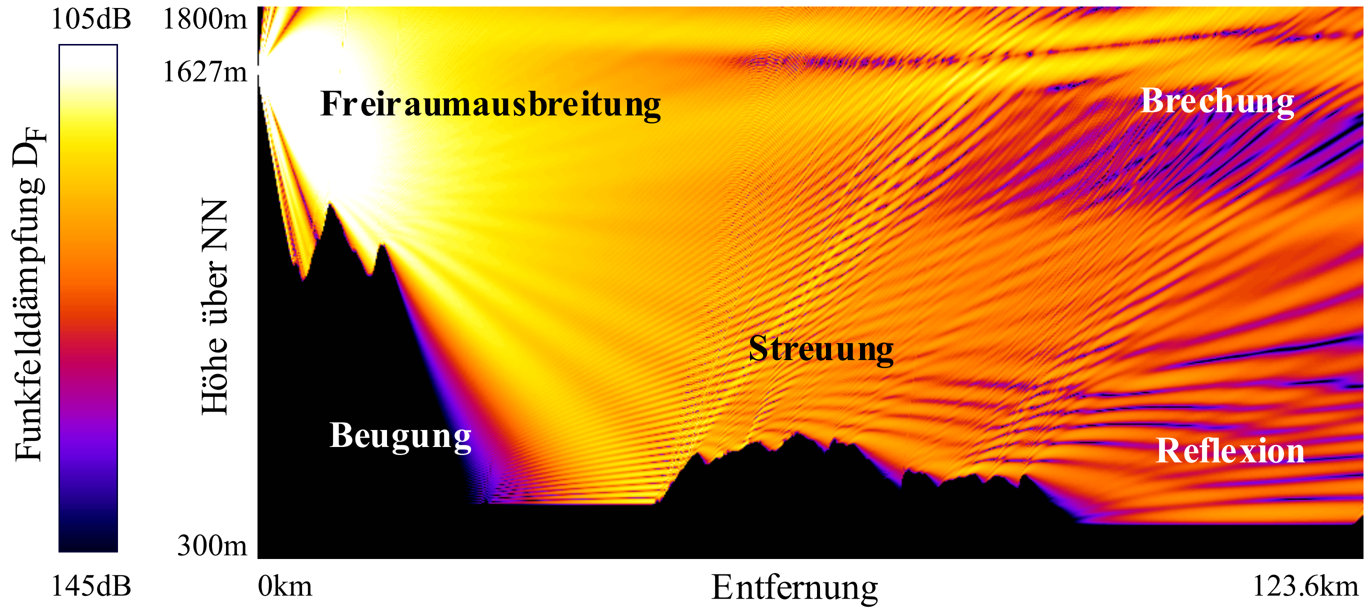
For future mobile radio systems partly intended for those frequencies, this effect has to be considered.



Source: N. Geng

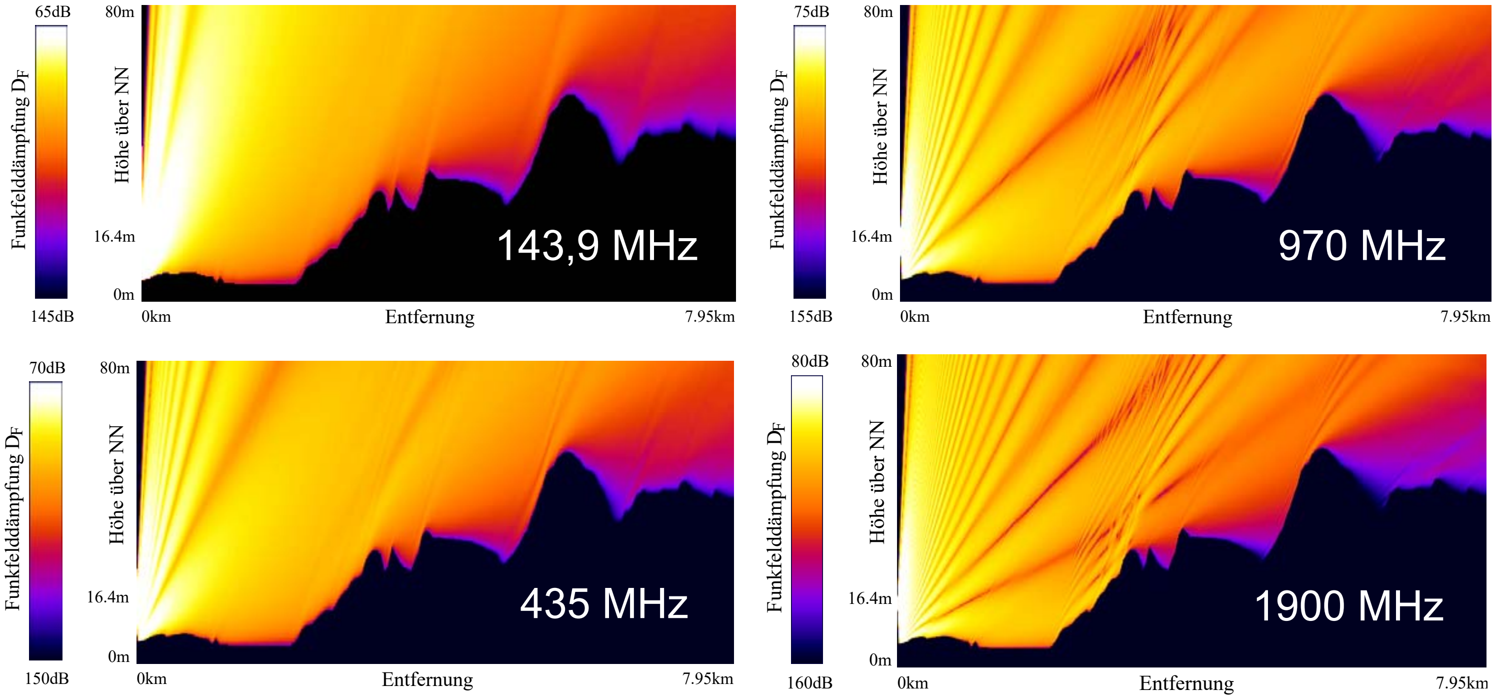
3.11 Superposition of the Propagation Effects

- Coherent superposition of the resulting multipath signals
  - example: propagation along a radio link at 300 MHz



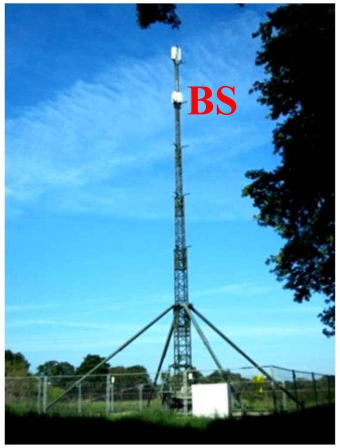
Source: N. Geng, W. Wiesbeck, Planungsmethoden für die Mobilkommunikation

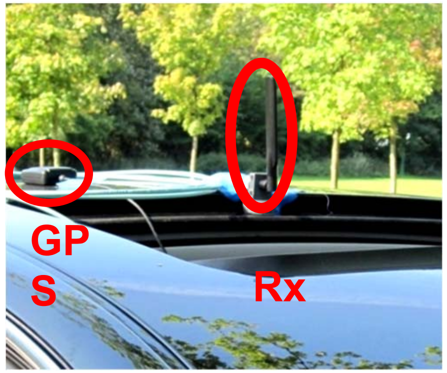
3.11 Superposition of the Propagation Effects
Frequency Dependence of the Propagation in a Hilly Terrain (v Pole)



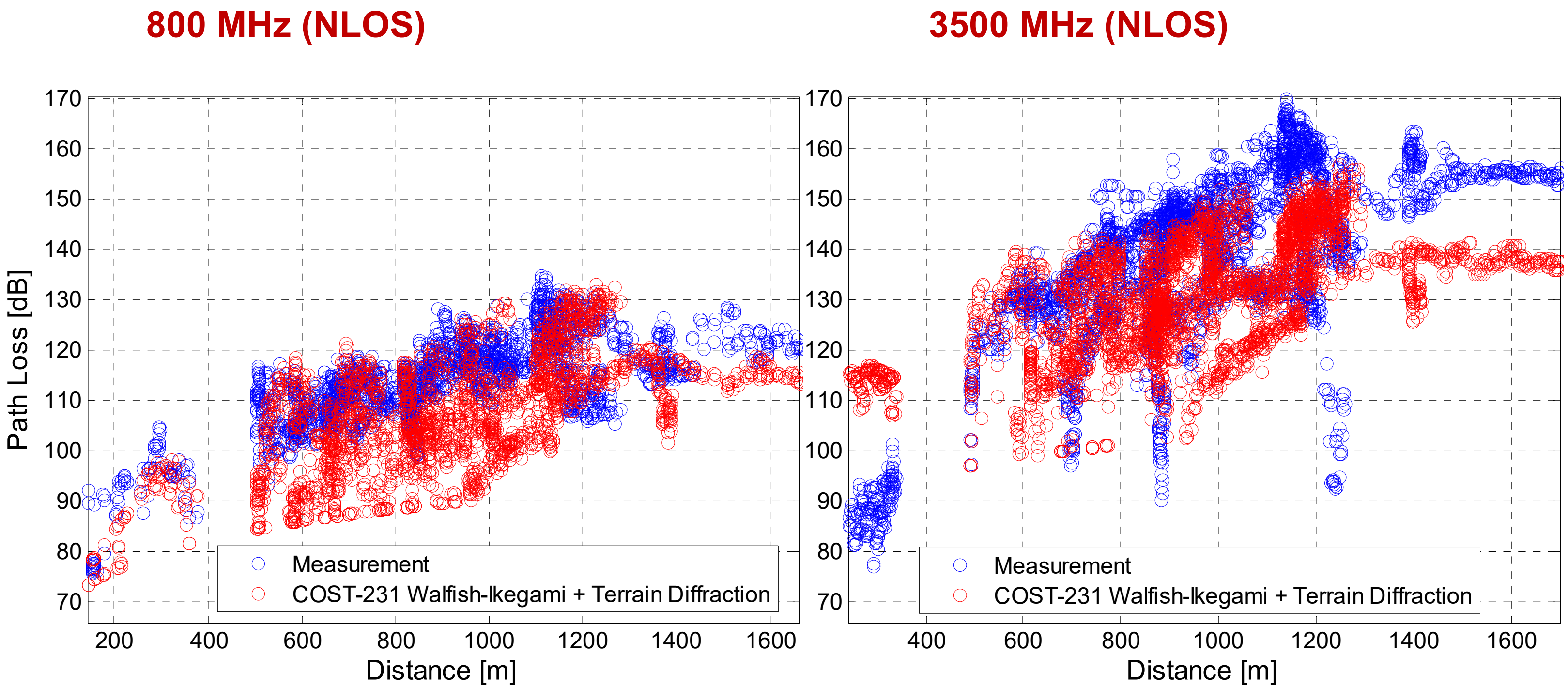
Source: N. Geng, W. Wiesbeck, Planungsmethoden für die Mobilkommunikation

Measurements at 800 MHz and 3500 GHz (WiMAX Pilot Test Lower Saxony)





Path Loss at 800 MHz and 3500 GHz (Measurement at Foliation)

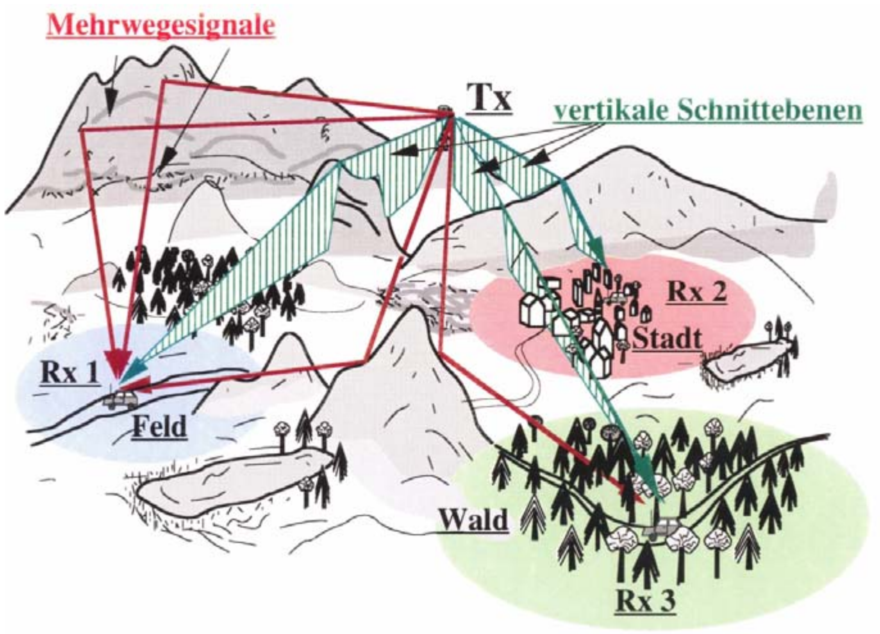


3.12 Propagation Models

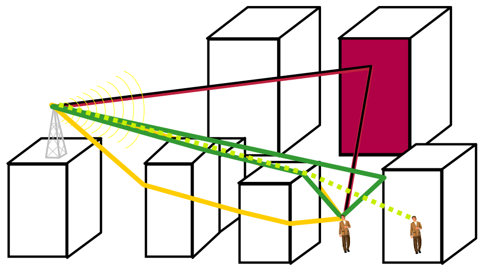
- For the planning of radio networks, models for the determination of the propagation behaviour of electromagnetic waves are required.
- In order to calculate the attenuation including the respective terrain structure and buildings as precise as possible, numerous deterministic, empirical and semi-empirical propagation models were developed in recent years.
- In the following, some simple empirical and semi-empirical propagation models are introduced for a quantitative representation of the propagation behaviour in principle in realistic terrain structures.

3.12 Propagation Models

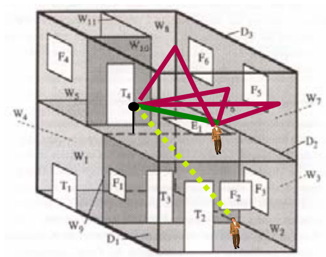
Typical Propagation Environments in Mobile Radio Communication



macro cell



small macro cell / micro cell



pico cell / femto cell

Simple Propagation Models for Macro Cells

- Attenuation ratios for macro cells (general equation)

$$L_{\mathrm{dB}}(r) = A + B \log\left(\frac{r}{\mathrm{km}}\right) + C + D \qquad (2.45)$$

L_{dB}: distance-dependent attenuation in dB

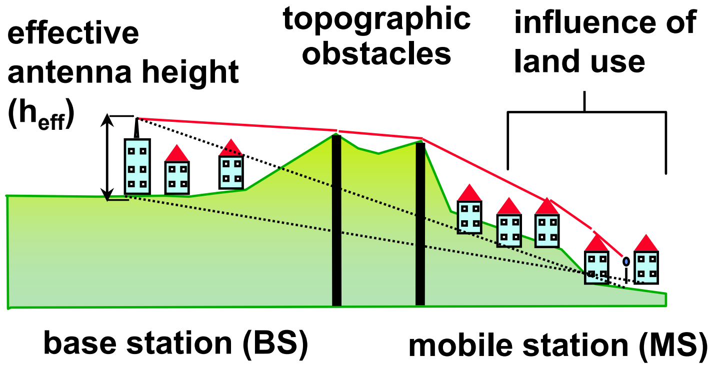
r: distance BS-MS in km

A: loss at a distance in 1 km

B: propagation coefficient (loss per decade)

C: diffraction loss through topographic obstacles

D: correction loss for land use (e. g. buildings)



Okumura-Hata Model and COST231-Hata Model, respectively

- For BS antenna heights (h_{BS}) between 30 and 200 m, a distance of 1 to 20 km and mobile station antenna heights (h_{MS}) of 1 to 10 m the following attenuation coefficients apply to urban areas:

  - from 150 MHz to 1000 MHz

$$A = 69.55 + 26.26 \log\left(\frac{f}{\mathrm{MHz}}\right) - 13.82 \log\left(\frac{h_{\mathrm{BS}}}{\mathrm{m}}\right) - a\left(\frac{h_{\mathrm{MS}}}{\mathrm{m}}\right) \qquad (2.46)$$

  - from 1500 to 2000 MHz

$$A = 46.3 + 33.9 \log\left(\frac{f}{\mathrm{MHz}}\right) - 13.82 \log\left(\frac{h_{\mathrm{BS}}}{\mathrm{m}}\right) - a\left(\frac{h_{\mathrm{MS}}}{\mathrm{m}}\right) \qquad (2.47)$$

  - for both a.m. frequency ranges

$$B = 44.90 - 6.55 \log\left(\frac{h_{\mathrm{BS}}}{\mathrm{m}}\right) \qquad (2.48)$$

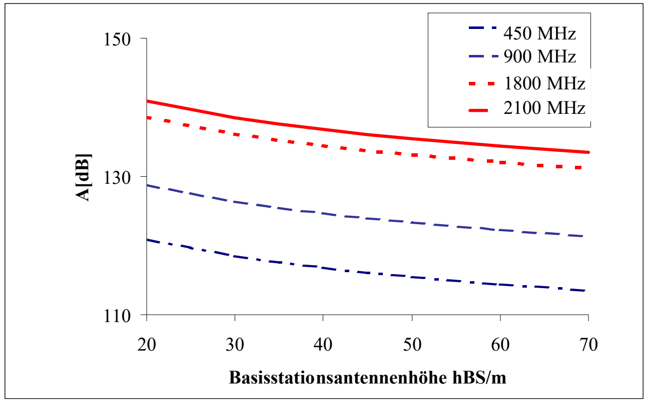
$$a(h_{\mathrm{MS}}) = \left(1.1 \log\left(\frac{f}{\mathrm{MHz}}\right) - 0.7\right)\frac{h_{\mathrm{MS}}}{\mathrm{m}} - \left(1.56 \log\left(\frac{f}{\mathrm{MHz}}\right) - 0.8\right) \qquad (2.49)$$

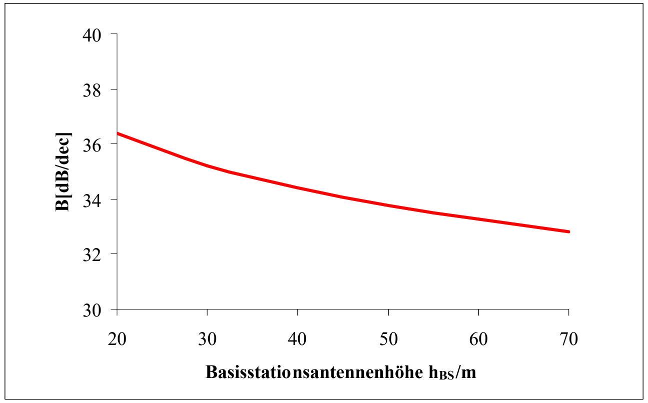
Attenuation Constants

- Examples of the values of *A* for 4 different frequencies and variations of the BS antenna height
- *B* is frequency independent
- *C* is usually described using the Knife-Edge model
- Determination of *D* for different land use classes

$$D = -9.42 \log\left(\frac{f}{\text{MHz}}\right) - 1.07 + E \qquad (2.50)$$

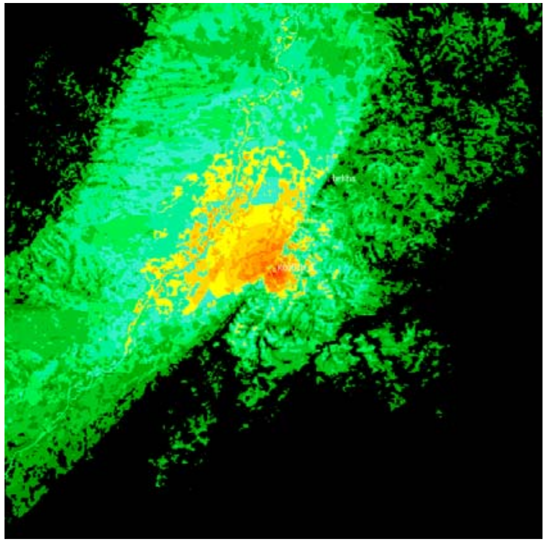
- The values of *E* depend on type and number of land use classes and have to be determined by calibration.
- Examples of *E* at 1800 MHz:
  - *E* = 19.6 dB in urban areas
  - *E* = 11.5 dB in open areas
  - *E* = 18.8 dB in forested areas



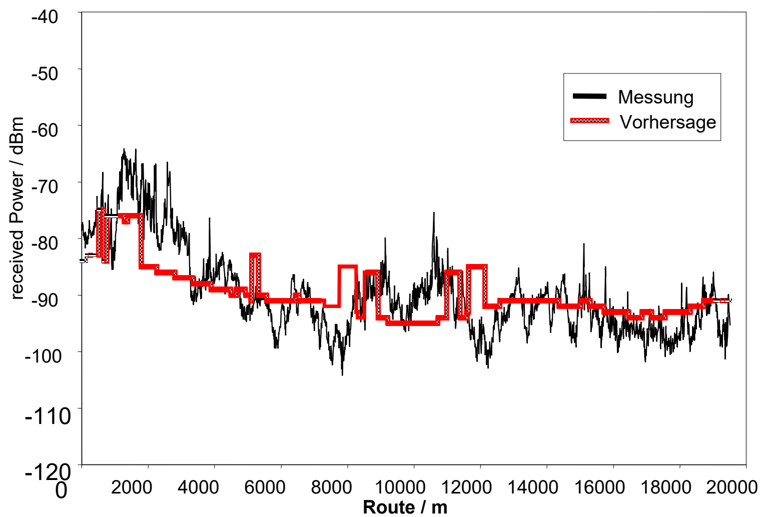


Example of Okumura-Hata Model

- Typical result based on a propagation model according to Okumura Hata + Knife Edge diffraction model + land use correction



Laminar prediction



Comparison prediction /measurement

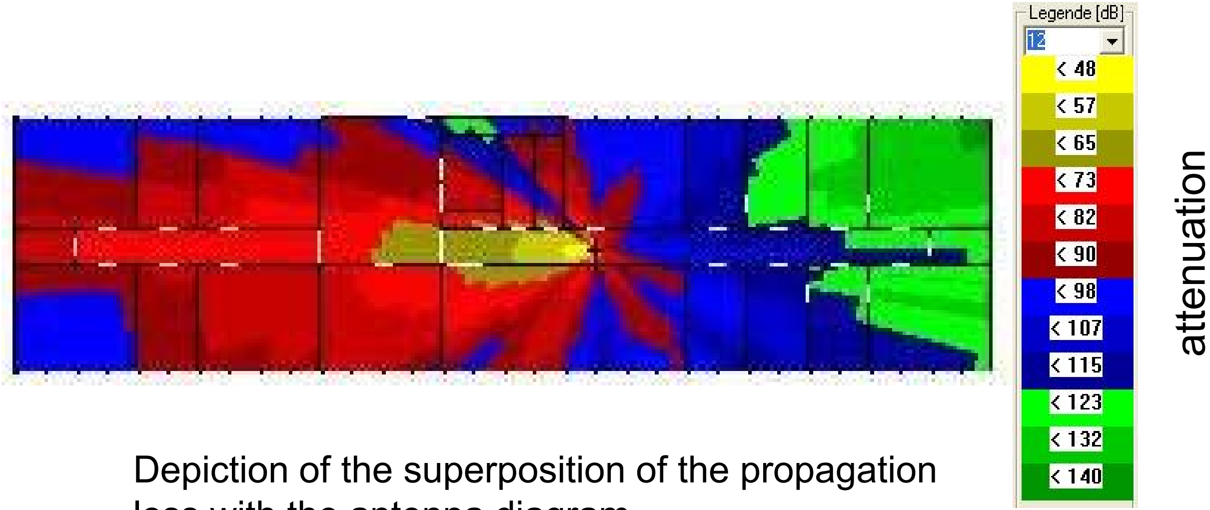
Simple Propagation Models for Indoor Coverage

- Multiwall model according to Motley-Keenan

$$L_{\mathrm{dB}} = L_{\mathrm{dB,FS}} + L_{\mathrm{dB,C}} + \sum_{i=1}^{I} k_{wi} L_{\mathrm{dB},wi} + \sum_{j=1}^{J} k_{fj} L_{\mathrm{dB},fj} \qquad (2.51)$$

$L_{\mathrm{dB,FS}}$	free-space attenuation
$L_{\mathrm{dB,C}}$	empirical determined constant
k_{wi}	number of penetrated walls of type *i*
k_{fj}	number of penetrated walls of type *j*
$L_{\mathrm{dB},wi}$	wall attenuation for wall of type *i*
$L_{\mathrm{dB},fj}$	floor attenuation for floor of type *j*
I	number of different types of walls
J	number of different types of ceilings

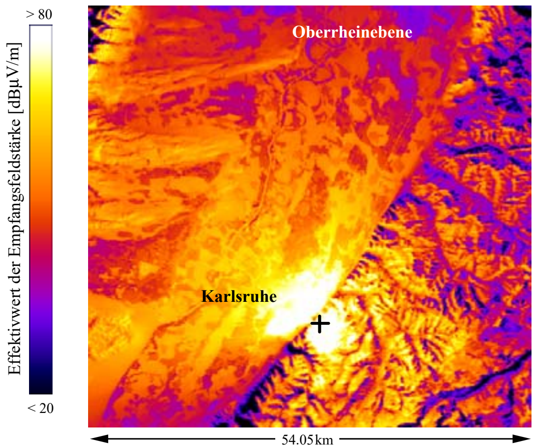
Example of the Prediction of the Indoor Coverage Using the Multiwall Model



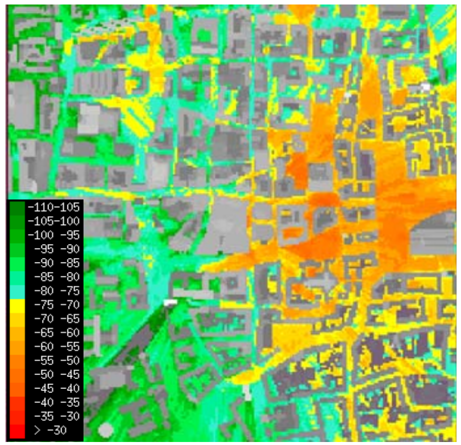
Depiction of the superposition of the propagation loss with the antenna diagram

Examples of Results Using more Komplex Models

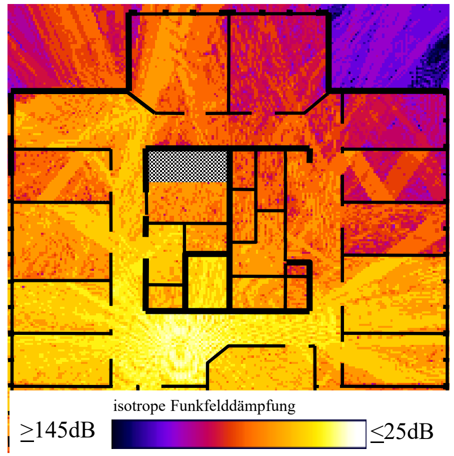
- For real radio network planning, deterministic propagation models are partly applied



macro cell

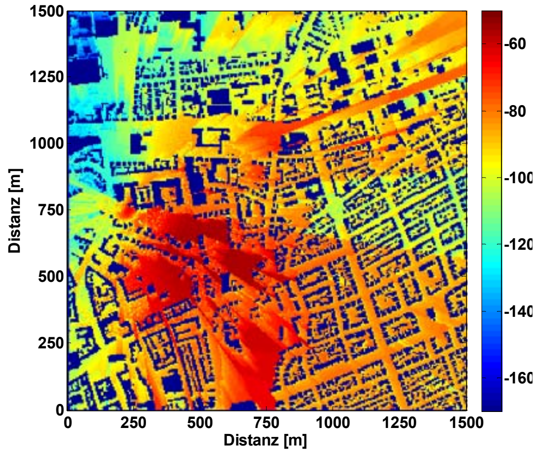


small macro cell

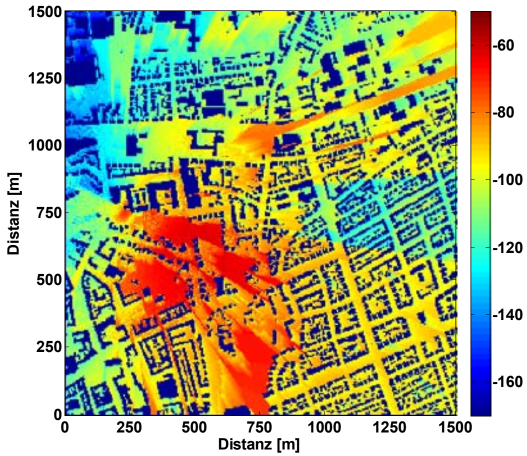


pico cell (indoor)

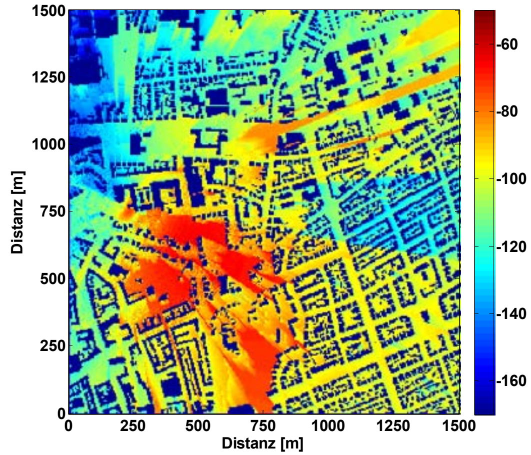
Prediction of the Received Power (Outdoor) on the Basis of Ray-Tracing



LTE800

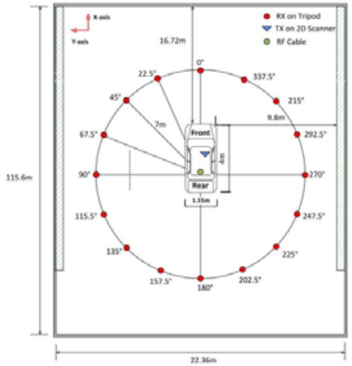


LTE1800

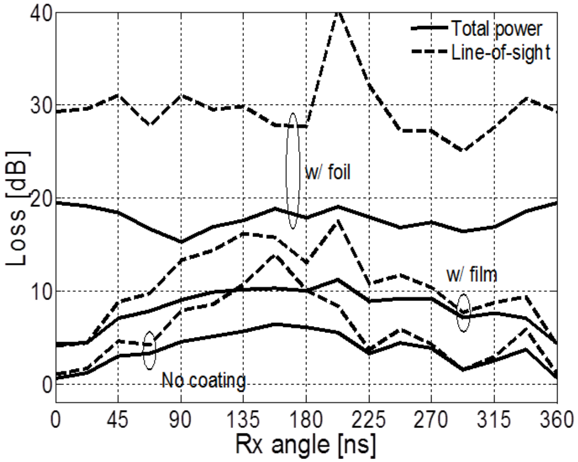


LTE2600

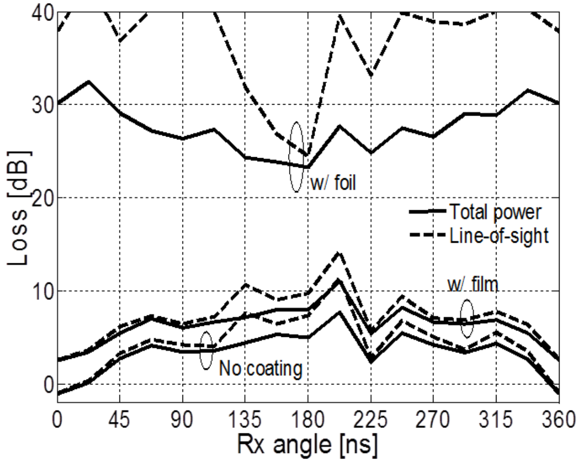
Penetration Loss at Radio Reception in a Vehicle without Outdoor Antenna



Source: Virk et. el, „Characterisation of Vehicle Penetration Loss at Wireless Communication Frequencies, EuCAP 2014



0.6 – 1.5 GHz

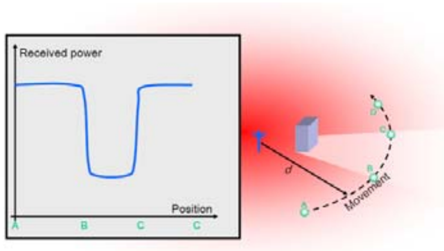


5.1 – 6 GHz

3.13 Fading Effects

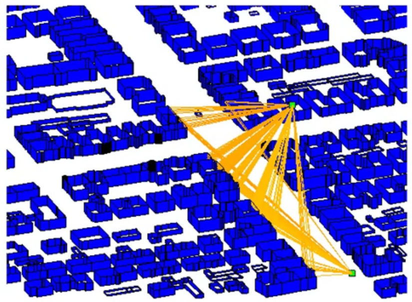
- The superposition of the different propagation paths results in additional statistical signal variations called fading.

  - Long-term fading:
    - deviations from the mean distance-dependent receive level due to varying shadowing effects (e.g. by buildings)
    - description of the variation of the attenuation in dB by a Gaussian normal distribution (lognormal fading)



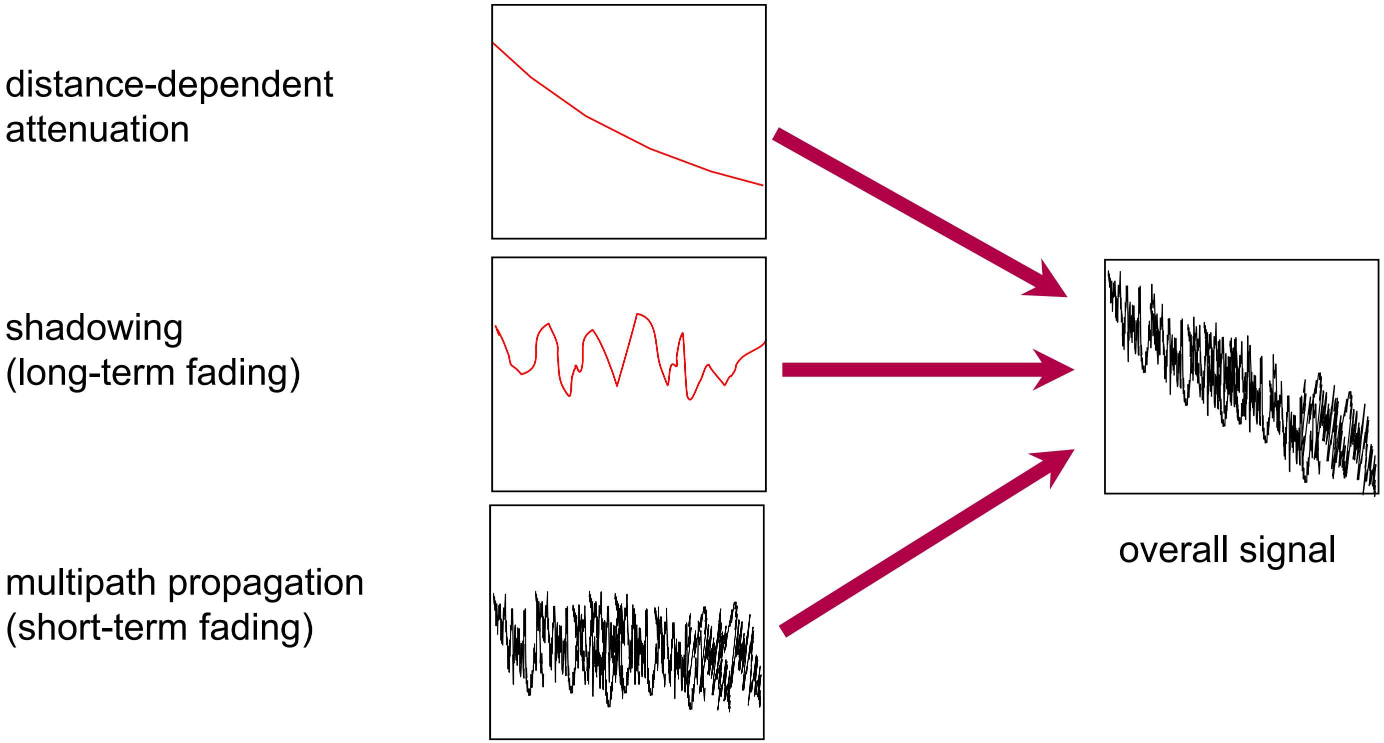
Source: Lecture slides for courses based on textbook A. F. Molisch, „Wireless Communications“

  - Short-term fading:
    - level fluctuations due to the coherent superposition of the multipath signals
    - description of the variation of the attenuation in dB by a Rayleigh distribution („Rayleigh-Fading“)



- These two fading effects superpose the distance-dependent attenuation.

Superposition of Fading Effects with Distance-Dependent Attenuation

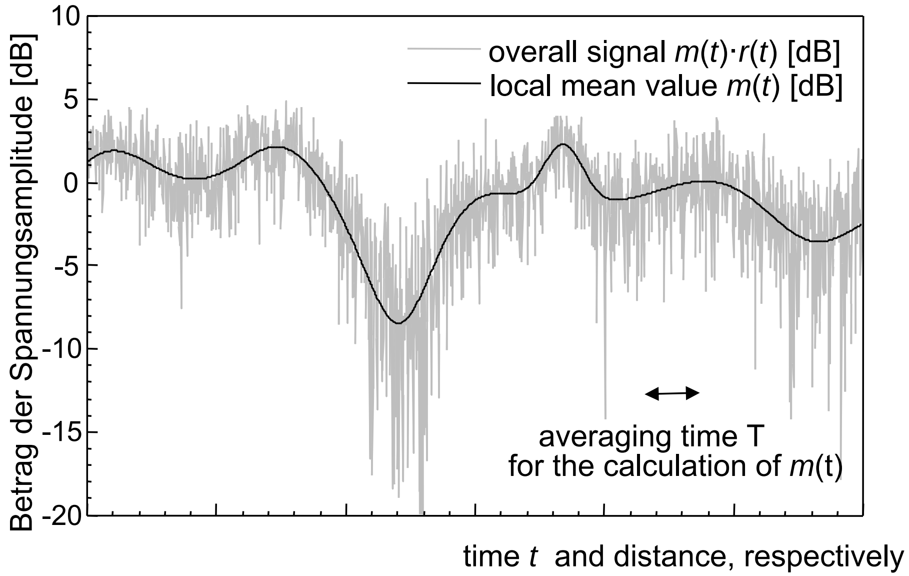


3.13 Fading Effects
Superposition of Long-Term and Short-Term Fading

- Superposition of long-term and short-term fading
- A mobile station moving through the radio field experiences spacial fading effects.

At coverage measurements, but also for measurements carried out permanently by a mobile station for the handover decision, the long-term fading is determined.

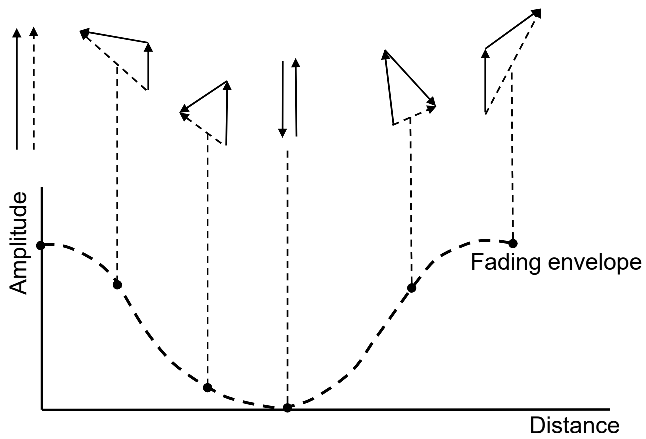
Determination of the level characteristic of the long-term fading through averaging of the overall signal

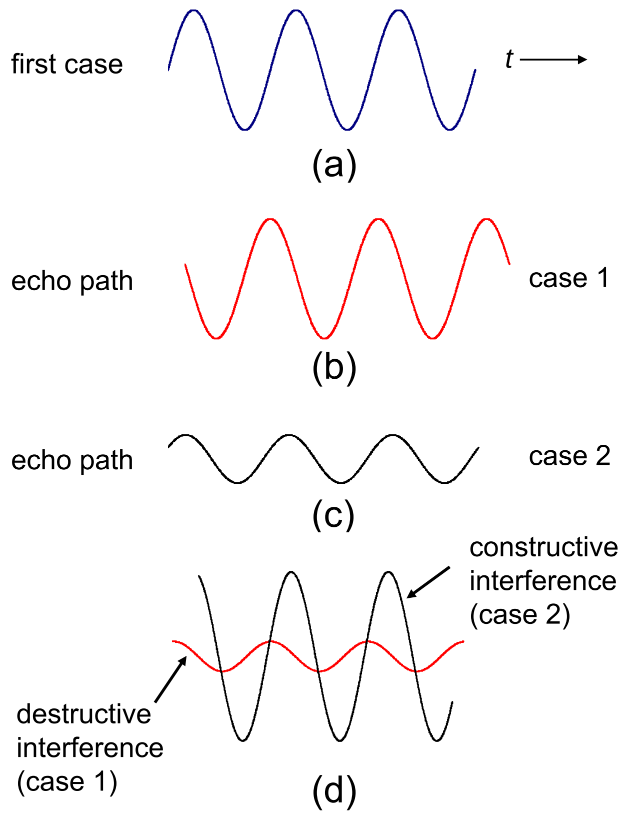


Source: N. Geng

3.14 Superposition of Multipath Signals

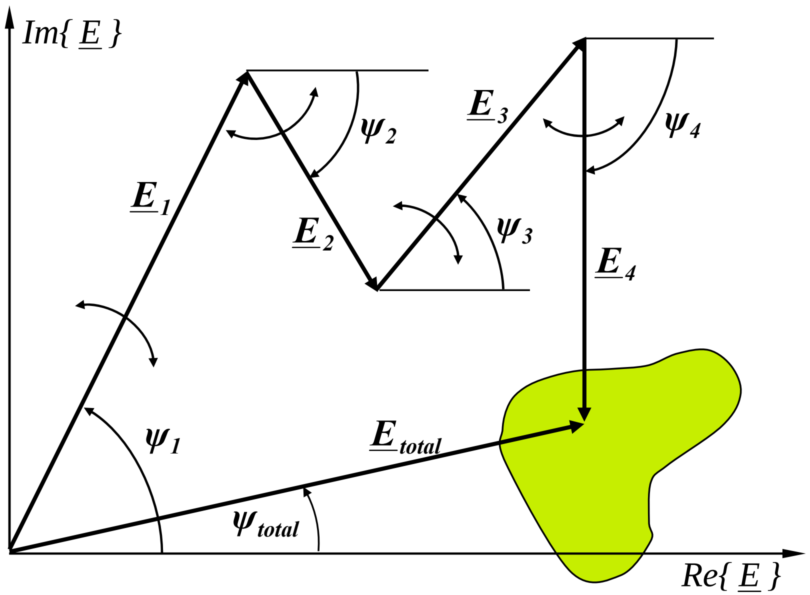
- Consideration of two-path propagation:
  - Depending on the amplitude and phase of both superposing signals, constructive and destructive interference can occur.

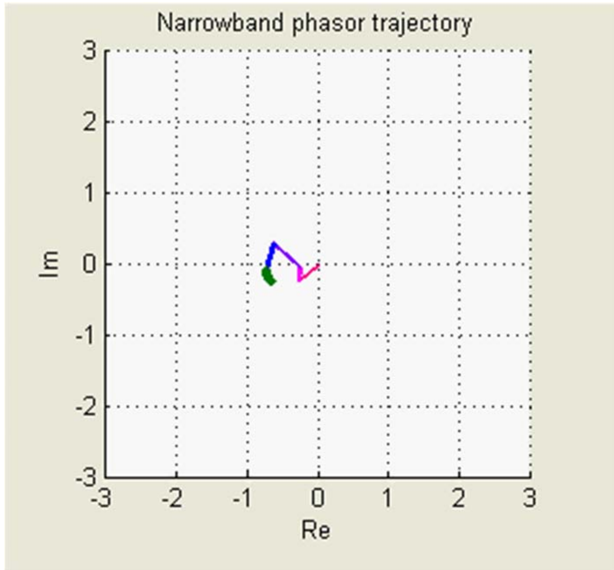




Superposition of Multipath Signals with Short-Term Fading

- Depiction of the signals as complex field strengths $\underline{E}_{\text{total}} = \sum_{i=1}^{n} \underline{E}_i = \sum_{i=1}^{n} E_i e^{j\psi_i}$ (2.52)





- Probability density of the resulting field strength depends on the statistical properties of the single signals

Probability Density Functions

- In case that the field strength indicators of the multipath signals consist of a determined indicator E_1 and many statistically independent indicators E_i (from multipath signals with comparable absolute values and statistically varying phases y_i), for the real part and the imaginary part of E_{total} a two-dimensional Gaussian distribution due to the central limit theorem results.

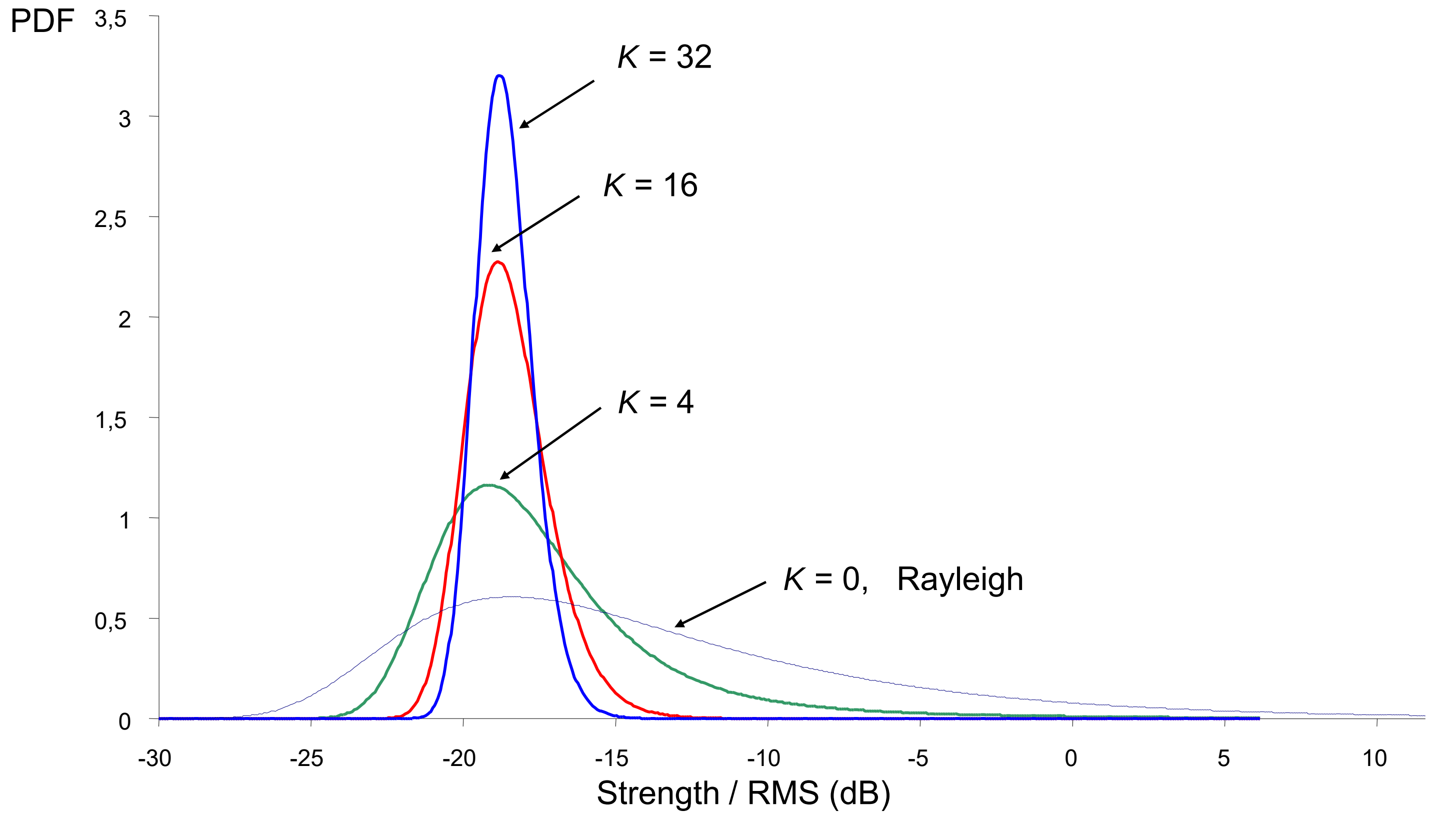
- For the absolute value of the field strength in dB, a Rice distribution is generally obtained.

$$P(E) = \frac{E}{\sigma^2} \mathrm{e}^{-\frac{E^2}{2\sigma^2}} \mathrm{e}^{-K} I_0\left(\frac{E}{\sigma}\sqrt{2K}\right) \qquad (2.53)$$

with

$$K = \frac{\text{power of the determined part}}{\text{power of the multipath signals}} \qquad (2.54)$$

Rice Distribution for Different Values of *K*



Special Cases of Rice Distribution

- 1. Special case: no determined signal, i.e. $K = 0$
  - Rice distribution passes into the Rayleigh distribution

$$P(E) = \frac{E}{\sigma^2} \mathrm{e}^{-\frac{E^2}{2\sigma^2}} \tag{2.55}$$

- 2. Special case: determined signal is much stronger than the sum of the multipath signals, i.e. $K \to \infty$
  - Rice distribution passes into the Gaussian distribution

$$P(E) = \frac{1}{\sigma\sqrt{2\pi}} \mathrm{e}^{-\frac{\left(E-\overline{E}\right)^2}{2\sigma^2}} \tag{2.56}$$

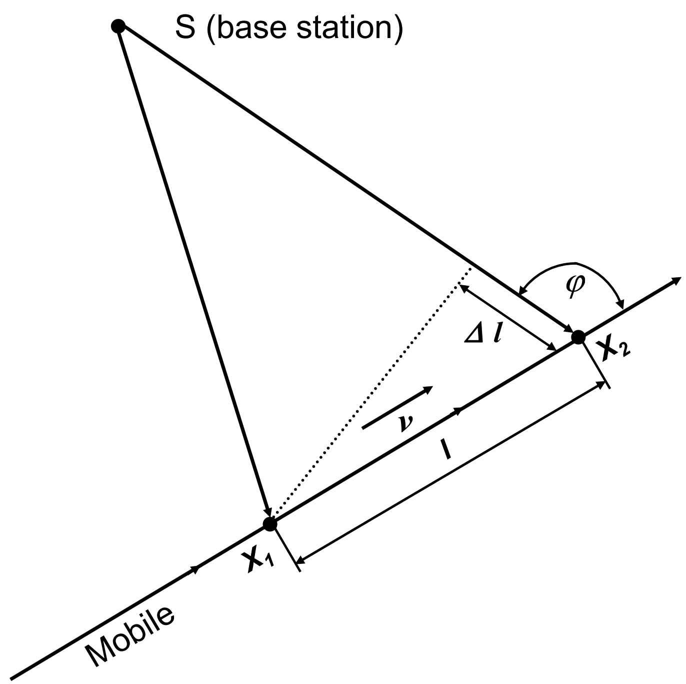
3.15 Doppler Shift

- In case of a moving receiver, for each multipath signal a Doppler shift occurs additionally.

$$f_{\mathrm{D}} = \frac{v}{\lambda}\cos\varphi \qquad (2.57)$$

v: speed of the receiver

λ: wave length



- Doppler shift results in a random frequency modulation

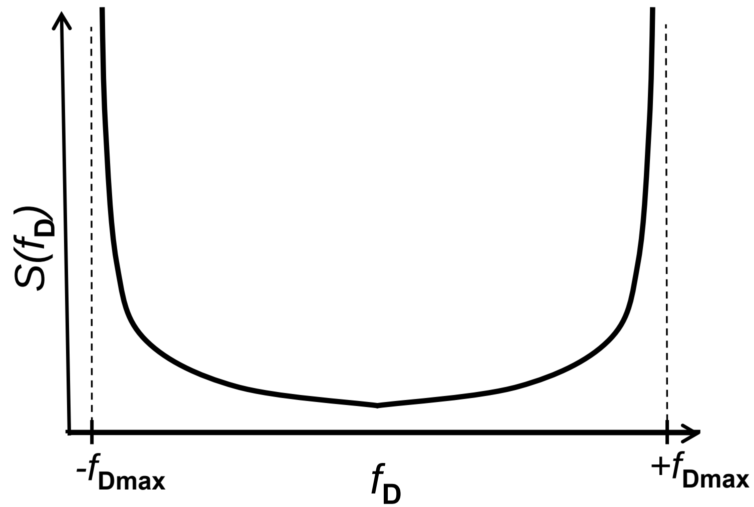
- Since the multipath signals arrive at different angles, a complete Doppler spectrum arises.

Doppler Spectrum (Example)

- The assumption of incoming multipath signals of the same amplitude uniformly distributed in the entire range of angles leads to the so-called Jakes spectrum

$$S(f_{\mathrm{D}}) = \frac{1}{\sqrt{1-\left(\frac{f_{\mathrm{D}}}{f_{\mathrm{Dmax}}}\right)^2}} \qquad (2.58)$$

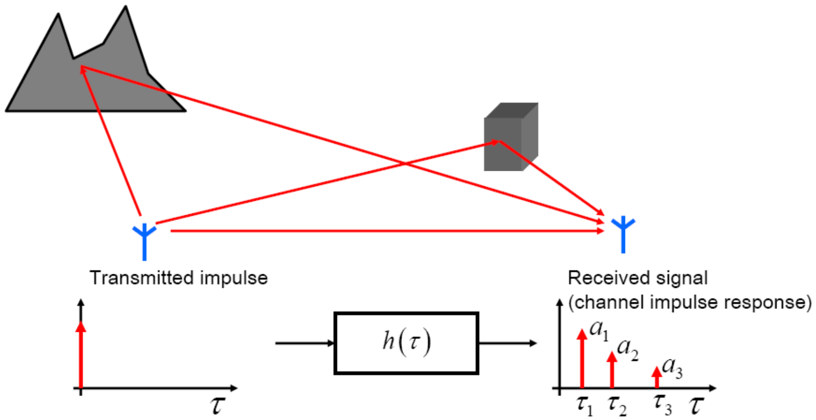
- f_{Dmax} is the maximum occurring Doppler frequency shift



- f_{Dmax} depends on the carrier frequency and determines the maximum velocity where a perfect reception of the signal is still possible.
- The higher the carrier frequency, the lower this maximum velocity

3.16 Broadband Radio Channel Characterisation

- Each multipath signal can be characterised by its amplitude, delay difference to the direct signal and its Doppler frequency shift.



Source: Lecture slides for courses based on textbook A. F. Molisch, „Wireless Communications“

- Due to the movement of the mobile station, the propagation conditions change location-dependently and thus time-dependently => the radio channel becomes time-dependent.

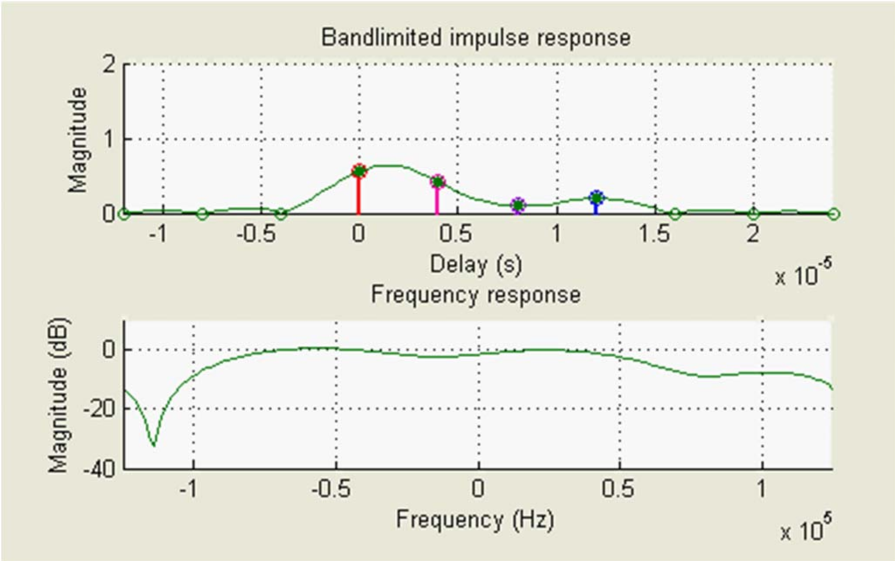
Effects on the Time and Frequency Domain

- The time-dependent impulse response $h(\tau,t)$ forms a Fourier pair with the time-dependent transfer function $H(f,t)$:

$$h(\tau,t) \;\circ\!\!-\!\!\!-\!\!\!-\!\!\bullet\; H(f,t) \qquad (2.59)$$

$h(\tau,t)$

$H(f,t)$



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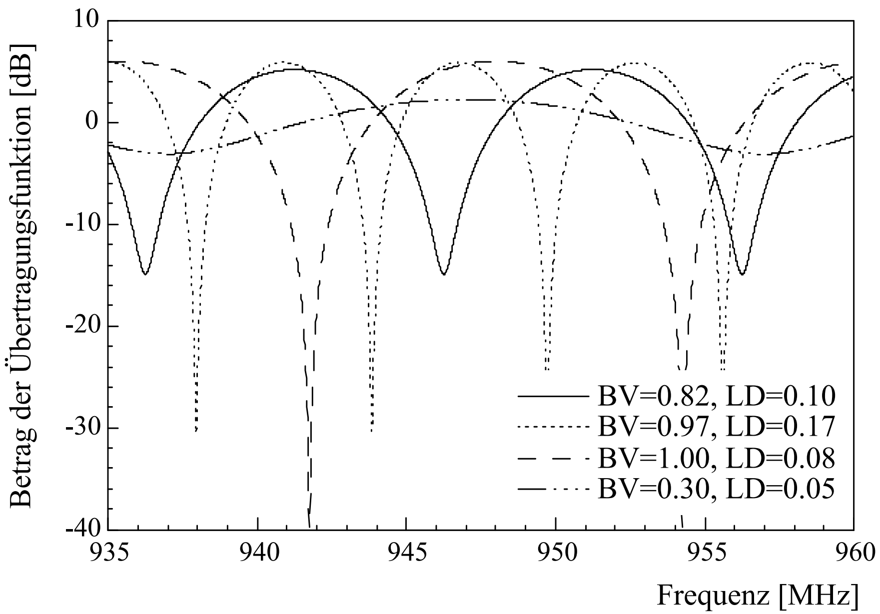
Institut für Nachrichtentechnik

Characteristics of the Transfer Function

- Constructive and destructive interferences arise with different frequencies.



Source: N. Geng

Special case of the two-path propagation
BV = a_2/a_1
LD = $\Delta t/\mu s$

3.17 Diversity

- Method for the averaging of fading notches (and interferences) are called diversity methods
  - Basic principle of all diversity methods: Redundant information arrives at the receiver via statistically independent radio channels.

- Time diversity
  - The data to be transferred are split into blocks of a definite length that are transferred temporally one after another.

- Frequency diversity
  - Averaging of the frequency-dependent short-term fading in the frequency range, e. g. through large transmission bandwidth or by using frequency-hopping methods.

- Antenna diversity
  - Reception of the signal with several antennas (direction, polarisation and space diversity)
  - In the following, different antenna diversity methods are explained.

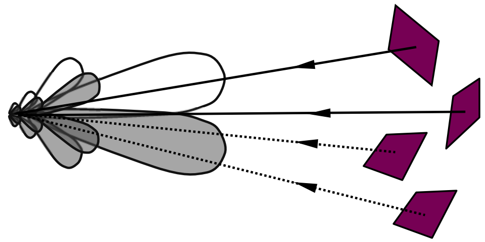
Direction Diversity

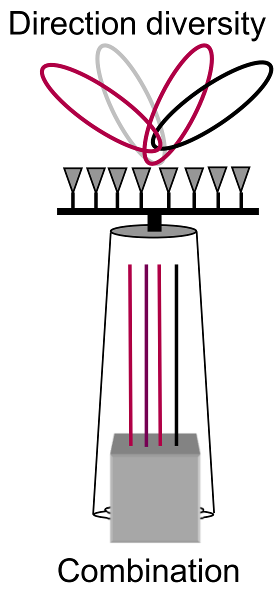
- The multipath signal starting from the MS arrives at the BS from different directions with different fading behaviour.

- By application of special antenna techniques, firstly separate reception and later combination

- Application for smart antennas providing several advantages:

  - higher antenna gain by bunching
  - reduction of fading
  - capacity gain by reducing interferences

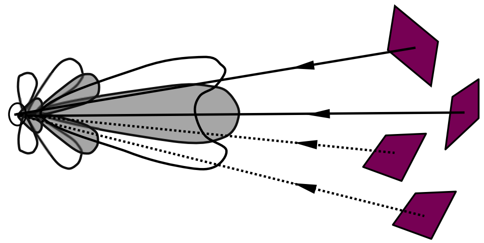




3.17 Diversity

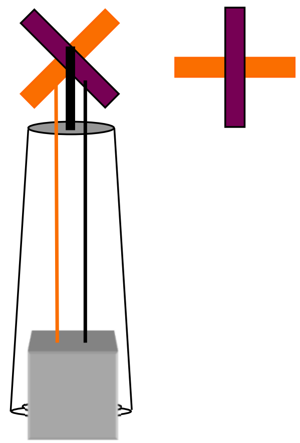
Polarisation Diversity

- In general, the polarisation changes with scattering processes.
- In environments with very distinctive scattering (e. g. urban areas), signals with different polarisation arrive at the receiver.
- Space-saving possibility for the realisation of diversity



Polarisation diversity

as **X** or as cross



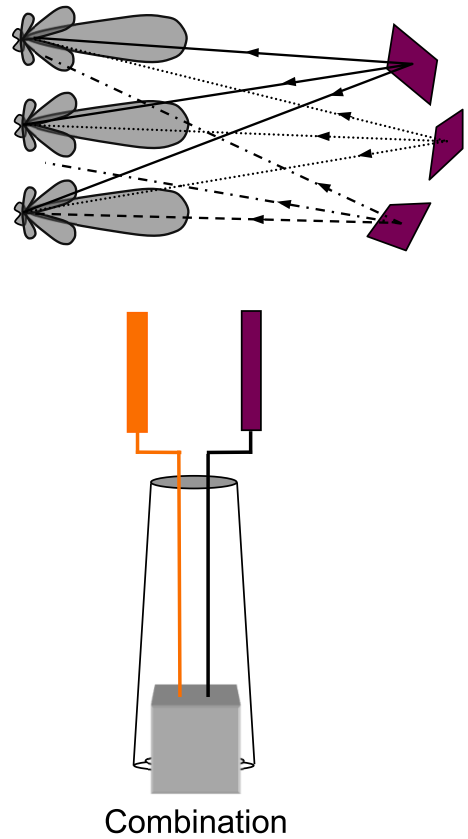
Combination

Space Diversity

- Reception via – normally two – spacially separated antennas

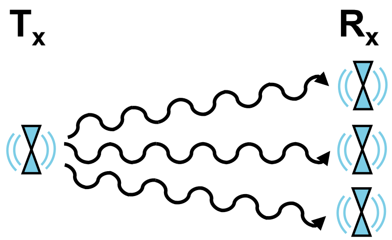
- Efficiency of space diversity depends on
  - number of antennas
  - method for the combination of the single signals
  - distance between the antennas

- Distinction of horizontal and vertical space diversity



Combination Methods

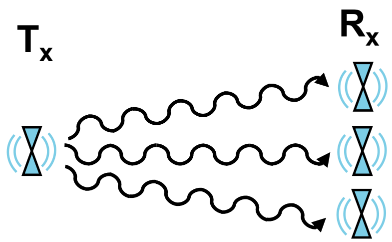
- There are different methods for combination of the single signals received via different radio channels.

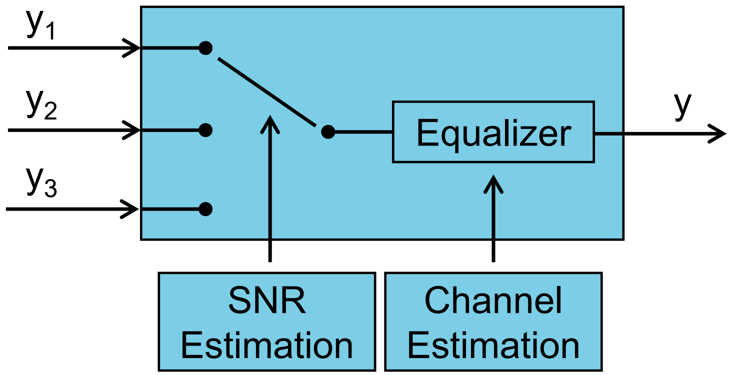


- These methods are most commonly used:
  - Selection Combining
  - Maximum-Ratio Combining
  - Equal Gain Combining

Selection Combining

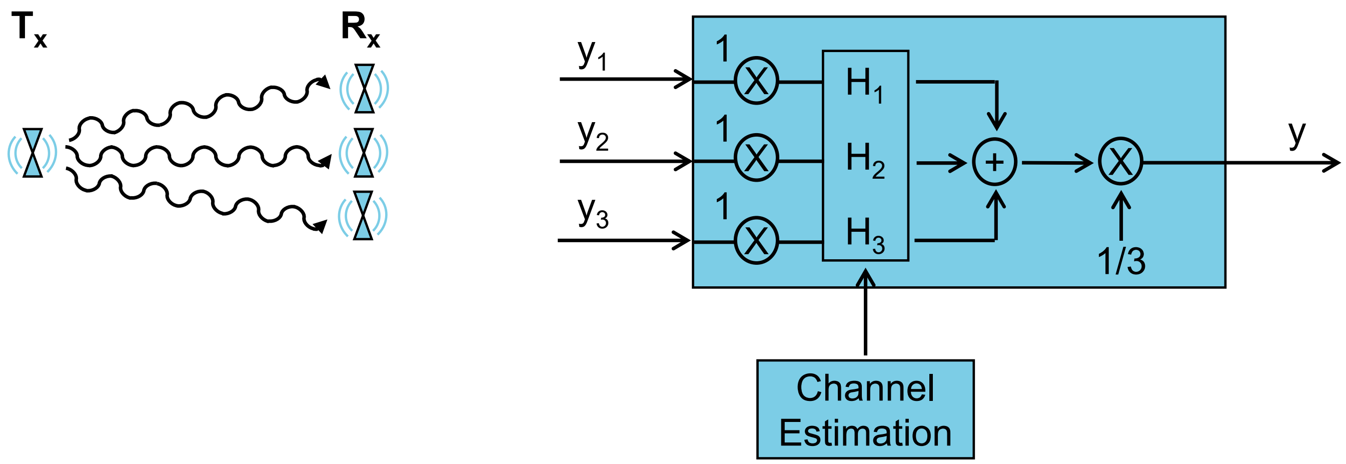
- Selection of a transmission channel
- Selection criterion: SNR conditions of the channels





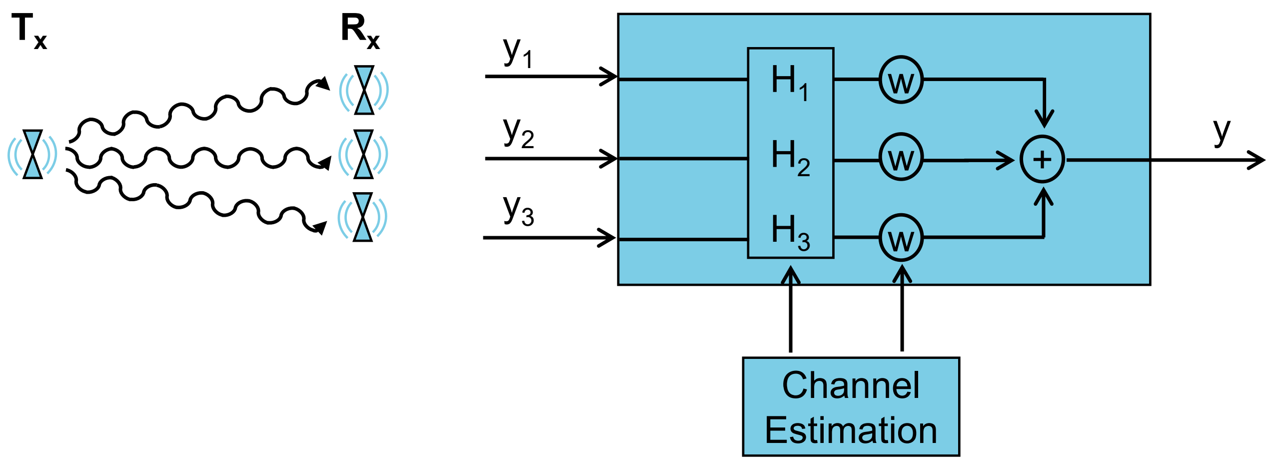
Equal Gain Combining

- All receive signals are considered.
- All channels have the same weighting factor.

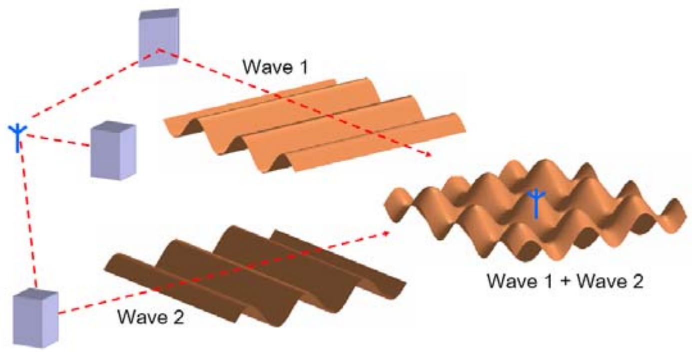


Maximum-Ratio-Combining Method

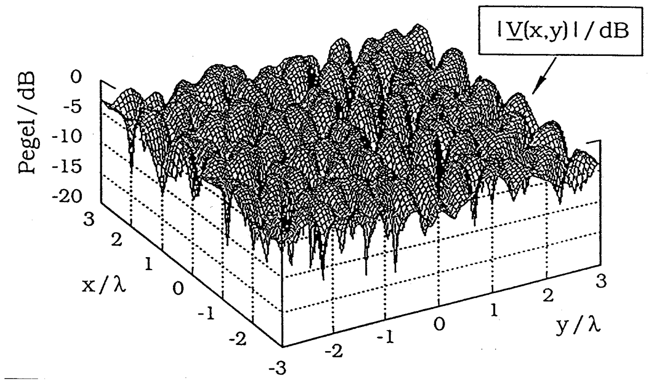
- All receive signals are considered.
- Weighting depends on the channel estimation.
- Sum of the weighting factors = 1



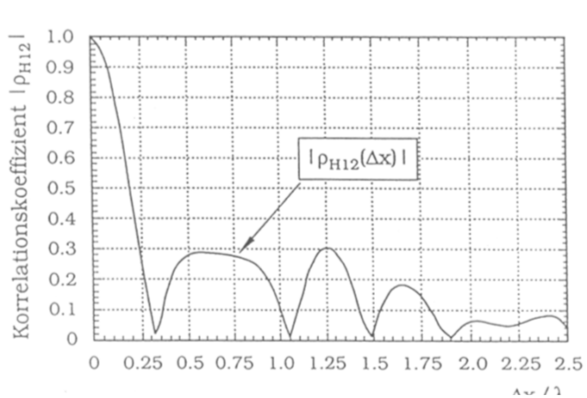
Example of Space Diversity



Source: Lecture slides for courses based on textbook A. F. Molisch, „Wireless Communications“



spatial field distribution



Source: Norbert Geng

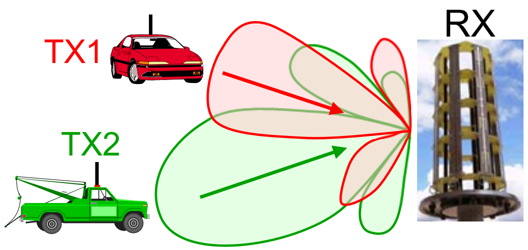
envelope cross correlation coefficient

Source: N. Geng

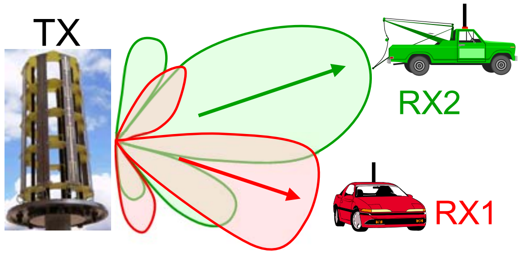
- Diversity gain
  - A diversity gain is only possible in case of uncorrelated signals.
  - Diversity gain can only be stated as estimated value for the gain to be expected.

3.18 Multi Antenna Systems

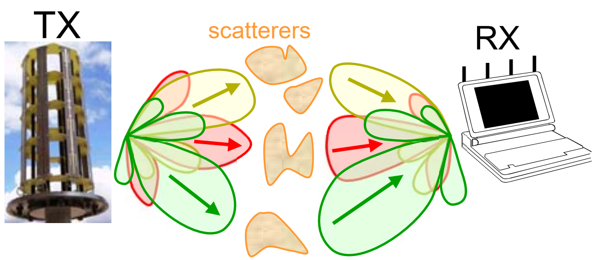
- Rough classification of smart antenna systems
- Input-output classification takes place from the viewpoint of the radio channel



SIMO: Single Input Multiple Output



MISO: Multiple Input Single Output

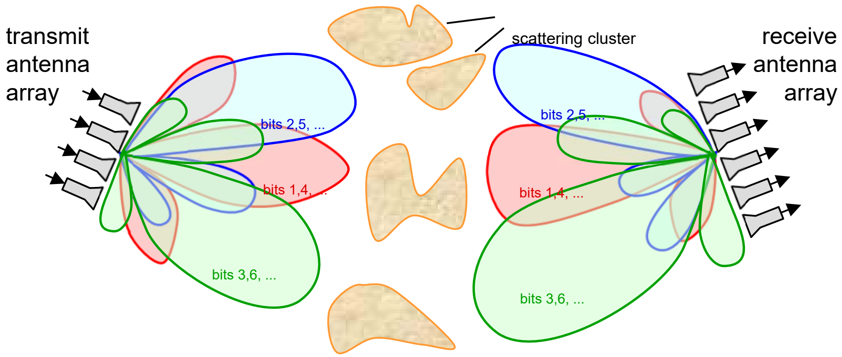


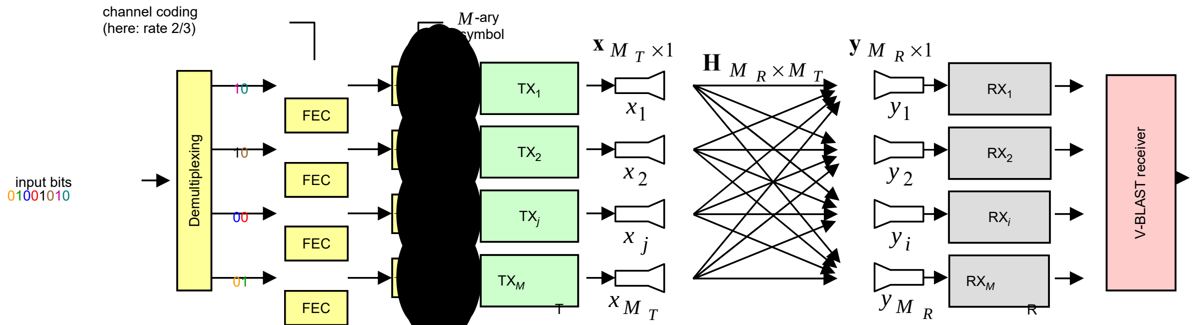
MIMO: Multiple Input Multiple Output

Increase in Capacity by Space-Division Multiplex

- Simplified model for
  optimal space-division multiplex

- Channel capacity increases with
  increasing multipath propagation





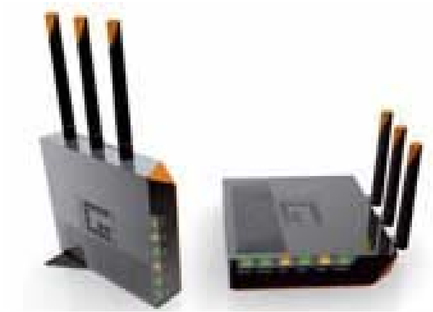
$$\mathbf{y} = \mathbf{H}\,\mathbf{x} + \mathbf{n} \longrightarrow \hat{\mathbf{x}} = „\mathbf{H}^{-1}“\,\mathbf{y}$$

Source: Prof. Dr. Norbert Geng

Examples of MIMO Antennas

Smart antenna
TD-SCDMA base station
(Roke Manor)

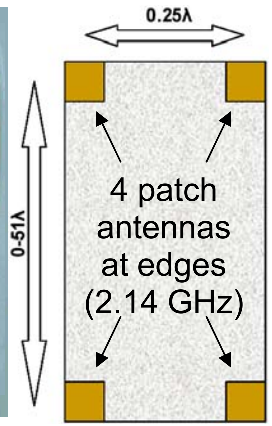
2-element Yagi
for LTE800



Source: http://www.chip.de

WLAN router

4-Element
MIMO terminal
(Univ. of Aalborg
and Nokia)



3.20 Link Budget

- Available power at the output of the receiving antenna under ideal conditions:

$$P_{\mathrm{dBm,R}} = P_{\mathrm{dBm,T}} + G_{\mathrm{dBm,R}} + G_{\mathrm{dBi,T}} - L_{\mathrm{dB,F}} \qquad (2.60)$$

- Required minimum received power level at the input of the receiverf P_{rmin} (corrsponds noise level + required signal-to-noise-ratio)
- Require´ment on isotropic path loss $L_{dB,F}$ under real conditions:

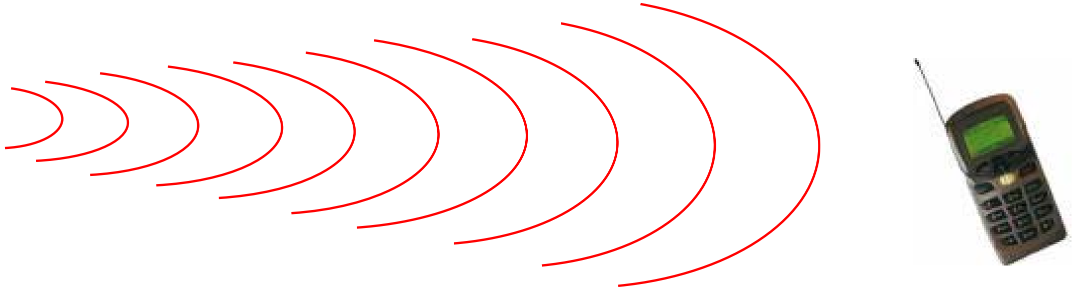
$$L_{\mathrm{dB,F}} \leq L_{\mathrm{dB,F,max}} = P_{\mathrm{dBm,T}} - P_{\mathrm{dBm,R,min}} + G_{\mathrm{dBi,R}} + G_{\mathrm{dBi,T}} + \text{other gains} - \text{other losses} - \text{safety margin}$$
(2.61)

- Other losses: for example lien losses between outpower of the power amplifier and the input of the antenna, imkpedance mismatch

Other Gains

- Other gains: e. g. diversity, antenna preamplifier
- Taking special measures, as mentioned in „other gains“ above, the feasible transmission loss can be increased.
- The link budget has to be set up for UL and DL separately.
- With GSM, a balanced link budget for UL/DL is mandatory.

Link Budget in the Downlink

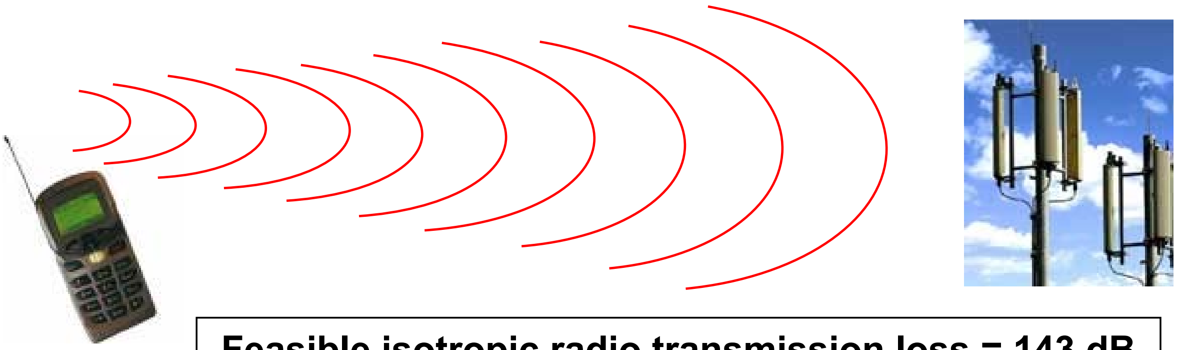


Feasible isotropic radio transmission loss = 143 dB

transmit power	42 dBm	receive antenna gain	0 dBi
cable losses	2 dB	diversity gain	0 dB
transmitter losses	3 dB	losses antenna-human	3 dB
transmit antenna gain	18 dBi	fading margin	6 dB
		interference margin	3 dB
		minimum receive power	-100 dBm

GSM Linkbudget UL/DL; ETSI Technical Report ETR103, GSM03.30

Link Budget in the Uplink



Feasible isotropic radio transmission loss = 143 dB

transmit power	30 dBm	receive antenna gain	18 dBi
transmit antenna gain	0 dBi	diversity gain	5 dB
losses antenna-human	3 dB	cable losses	2 dB
further losses	0 dB	fading margin	6 dB
		interference margin	3 dB
		minimum receive power	-104 dBm

GSM Linkbudget UL/DL; ETSI Technical Report ETR103, GSM03.30