RADIO SYSTEM PERFORMANCE

Since the early introduction of radio, the significance and number of applications of radio services have constantly been growing. From sophisticated satellite and microwave links which provide world-wide communication infrastructure, mobile radio and personal communication services that are providing seamless communication capabilities, all the way to consumer products like cordless phones, wireless local area networks, wireless toy communicators or garage door openers, radio services have changed everyday life in a manner that could have not been predicted decades ago. The growing sophistication of systems and usages is closely related to the technical understanding necessary for putting radio systems into operation. Challenges in radio system design have grown tremendously due to the overcrowding of radio spectrum. The need for co-ordinating radio services has been identified early and International Telecommunication Union (ITU) has established necessary regulations. The effective spectrum management is essential to maximizing the benefits that can be obtained from the radio spectrum (1). After restructuring of the ITU in 1993, these problems have been the focus of the ITU Radiocommunication Sector.

Radio systems operate in different radio frequency bands as shown in Table 1. (2).

This influence the characteristics of the radio channel. Signal propagation path may vary from clear line-of-sight to severely obstructed by buildings, mountains and other environmental effects. Since radio channels are random in nature, understanding of the radio propagation phenomena and consequently their modeling play a major role in the prediction of the radio reception quality which in turn enables efficient design of wireless systems. Large-scale propagation models characterize signal strength over long distances, while small-signal models describe the rapid fluctuation of the received signal over very short distances or short time duration (3).

Radio reception is influenced by different factors, summarized in Table 2. Radio propagation influences radio link design, with path loss models and statistics of large-scale fading being the major elements of power budget. In addition to power budget, radio link design includes geographical positioning of transmitter and receiver, frequency planning, radio link control scheme and handover schemes. Requirements for the type of radio service, quality of service, coverage and availability of the radio link provide an additional set of parameters which influence radio link design. On the other hand, fading affects the radio receiver design, selection of modulation and coding scheme, techniques for fading compensation, synchronization circuits, design of radio front-end, etc. Selection of the appropriate receiver techniques determines the radio sensitivity which is an integral part of the power budget. Results of radio link and radio receiver design are a part of wireless networks design. Networks architecture, network interfaces and network control are closely related to radio reception which defines the physical layer of the network (4).

RADIO LINK DESIGN

Power Budget

Power budget (link budget) is the starting point for a radio communication link design. Major factors influencing the power budget are shown in Fig. 1. These factors include transmitting and receiving antenna gains G_T and G_R, path loss L_p (which depends on operating frequency, environment and type of radio service), fading margin M_F, interference margin M_I, and radio receiver sensitivity RS.

Transmitting antenna gain (usually specified as the gain relative to isotropic radiator) and receiving antenna gain are dependant on the carrier frequency and physical size of the antenna. For a given radio system, selection of the equipment determines antenna gains. Additional sources of signal loss at the antenna may include noise due to protective cover, pointing loss, feeder loss and antenna efficiency described as the ratio of the effective aperture to the physical aperture of the antenna (5).

Link budget analysis of the radio system defines the maximum acceptable path loss

$$L_p[dB] = P_T[dBm] + G_T[dBi] + G_R[dBi] - M_F[dB] - M_I[dB] - RS[dBm]$$

which can be used to estimate the coverage area of the system by determining the maximum transmitter-receiver distance.

Relationship between path loss and coverage area is crucial for the design of radio system. Other sources of signal loss in radio propagation are caused by environmental effects such as rain, clouds, fog, etc.

Fading margin determines the outage probability of the system due to large-scale and/or small-scale fading. In a similar way, interference margin can be built into the link budget to maintain desired performance in the presence of the interference, regardless of the interference mechanism: adjacent channel interference (ACI), cochannel interference (CCI) in cellular systems, intermodulation distortion (IMD) products created by the large interfering signals, etc.. The amount of margin depends on the required quality of service. It is expressed in terms of signal to noise ratio (SNR) for analog systems or bit error rate (BER) for digital systems, and availability of the link which is the measure of long-term link utility stated on an average basis, usually on the annual basis.

Radio receiver sensitivity describes the ability of the radio receiver to detect weak signals, and is derived from the performance requirements of the given radio service. Sensitivity specification includes the impact of techniques that are integral part of radio reception at the detection level, such as diversity, coding, interleaving, as well as the impact of radio receiver such as noise figure and noise bandwidth.

Frequency Planning

Frequency planning is a part of link control and is essential part of radio link design. Since the radio spectrum is the limited resource, coexistence of different services will be practically impossible without the planning procedure. It is effective way to optimize spectrum usage, enhance channel

Table 1. Summary of frequency bands for operating radio systems

Frequency band	Frequency range	Mode of radio propagation	Radio services	Major characteristics
Very low frequency (VLF)	3–30 KHz	Ionospheric	navigation, long distance radio telegraphy	small bandwidth, large antennas,
Low frequency (LF) Medium frequency (MF)	30–300 KHz 300 KHz - 3 MHz	groundwave ionospheric	navigation, AM broadcasting, aeronautical, maritime comm.	quite large antennas, high power transmitters
High frequency (HF)	3–30 MHz	ionospheric	amateur radio, broadcasting, aeronautical maritime comm.	high variability of channel conditions
Very high frequency (VHF) Ultra high frequency (VHF)	30–300 MHz 300 MHz-3 GHz	direct waves, ground reflected waves	radio and television broadcasting, land mobile radio, paging, GPS	relatively small antennas, considerable bandwidth
Super high frequency (SHF)	3–30 GHz	direct waves	satellite, radars, short-range communications, microwave terrestrial links	high losses, directional antennas
Extra high frequency (EHF)	30–300 GHz	direct wave	very short range comm, satellite to satellite links	enormous bandwidths, rain absorption

Table 2. Radio system design elements relevant for radio reception

Radio propagation	Radio link design	Radio receiver design	Radio network design
Large-scale fading	Requirements	-modulation	-architecture
-path loss	-radio services	-multiple-access	-interference
-shadowing	-quality of service	-coding	-network control
	-coverage/range	-anti-fading techniques	
Small-scale fading	Design	-interfaces	
(multipath fading)	-power budget	suppression	
-delay spread	-TX/ RX	-synchronization	
-coherence time	allocation	-receiver hw design	
	-link control		

Table 3. Parameters for regression coefficients calculation

Frequency[GHz]	k_H	k_V	α_H	α_V
2	0.000154	0.000138	0.963	0.923
6	0.00175	0.00155	1.308	1.265
10	0.0101	0.00887	1.276	1.264
15	0.0367	0.0335	1.154	1.128

capacity and reduce different types of interference.

Some of the benefits of frequency planning will be illustrated on the example of the mobile radio services. Frequency planning for mobile radio includes channel numbering, channel grouping into subsets, cell planning and channel assignment. It controls ACI by channel separation which provides adequate isolation, and CCI by selection of frequency reuse pattern. Frequency planning can be efficiently combined with cell sectorization to improve the system capacity. Different trade-offs are involved in frequency planning based on targeted channel capacity and signal to interference ratio (SIR). The most popular OMNI frequency plan in mobile radio is N=7 plan, which allows for ACI. This plan is outperformed by N=9 plan with respect to SIR, but the drawback is the reduced channel capacity. Detailed analysis of various frequency plans can be found in (21). Besides the impact on technical parameters of the system, frequency planning greatly influences radio system growth and economics.

Large-Scale Variations: Path Loss

The prediction of the path loss is the most important step in the radio system planning process since it determines radio coverage. Prediction models could include only finite number of parameters that can influence radio propagation,

Table 4. Tapped-delay fading channel models for GSM mobile radio (19)

a) typical case for rural area RA x, x-speed of mobile in km/h

Tap number	Relative time [microsec]	Average relative power [dB]	Doppler spectrum
1	0.0	0	Rice
2	0.1	−4	Class
3	0.2	−8	Class
4	0.3	−12	Class
5	0.4	−16	Class
6	0.5	−20	Class

b) typical case for hilly terrain HT x

Tap number	Relative time [microsec]	Average relative power [dB]	Doppler spectrum
1	0.0	0	Class
2	0.1	−1.5	Class
3	0.3	−4.5	Class
4	0.5	−7.5	Class
5	15.0	−8.0	Class
6	17.2	−17.7	Class

c) typical case for urban area TU x

Tap number	Relative time [microsec]	Average relative power [dB]	Doppler spectrum
1	0.0	−3.0	Class
2	0.2	0	Class
3	0.5	−2.0	Class
4	1.6	−6.0	Class
5	2.3	−8.0	Class
6	5.0	−10.0	Class

Table 5. Tapped-delay fading channel models for WCDMA mobile radio: classical Doppler spectrum for all taps (20)

Case 1,speed 3km/h		Case 2,speed 3 km/h		Case 3,speed 120 km/h		Case 4,speed 3 km/h		Case 5,speed 50 km/h*		Case 6,speed 250 km/h	
Relative Delay [ns]	Relative mean Power [dB]	Relative Delay [ns]	Relative mean Power [dB]	Relative Delay [ns]	Relative mean Power [dB]	Relative Delay [ns]	Relative mean Power [dB]	Relative Delay [ns]	Relative mean Power [dB]	Relative Delay [ns]	Relative mean Power [dB]
0	0	0	0	0	0	0	0	0	0	0	0
976	−10	976	0	260	−3	976	0	976	−10	260	−3
		20000	0	521	−6					521	−6
				781	−9					781	−9

Table 6. Parameters for calculation of the impulse noise median

Environmental category	c	d
Business	76.8	27.7
Residential	72.5	27.7
Rural	67.2	27.7
Quiet rural	53.6	28.6

Table 7. SNR improvement in dB for BER of 1% relative to to single channel reception, comparison of various diversity combining schemes (20)

Diversity branches	Selection	Maximal ratio	Equal gain
2	10.0	11.5	10.8
4	16.0	19.0	18.0
6	18.0	22.0	21.5

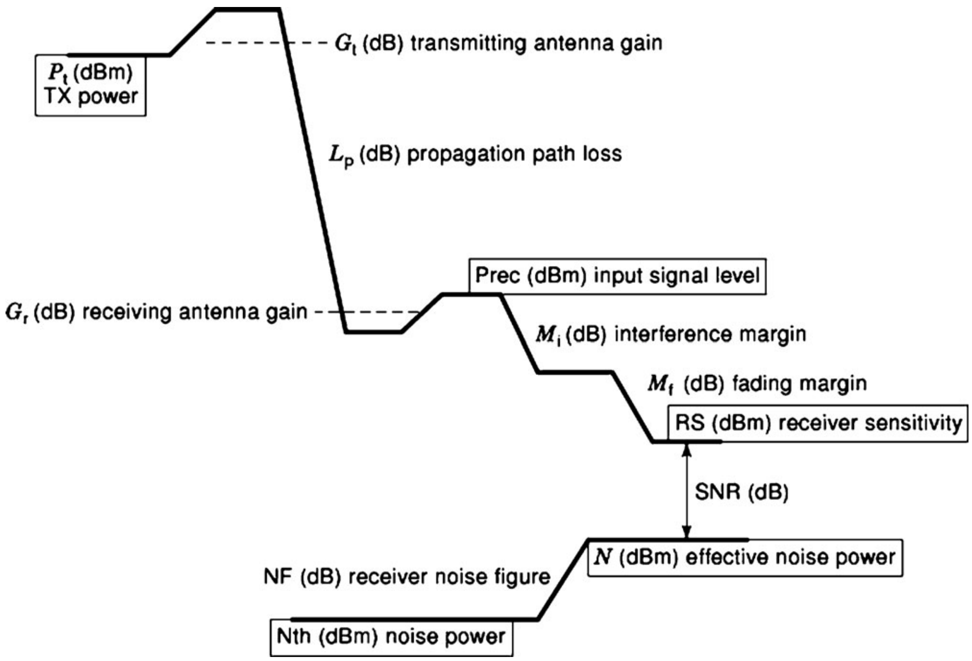

Figure 1. Power budget for radio link design.

and often factors such as environment (effects of buildings, man made obstacles, vegetation) are treated separately as well as multipath fading.

Three major mechanisms of radio propagation include reflection, diffraction and scattering. Rather than focusing on the details of radiowave propagation which are covered in (6), the models for large-scale path loss prediction will be summarized indicating critical parameters that are influencing radio reception.

Free space propagation model is applicable when there is unobstructed line-of-sight path between the transmitter and the receiver. This scenario is typical for satellite systems and microwave radio links. The path loss is given by

$$\begin{aligned} L_{free\,space}[dB] &= -10\log(G_T[dB]) - 10\log(G_R[dB]) \\ &\quad + 20\log(\,f[Hz]) + 20\log(d[m]) - 147.6 \end{aligned}$$

where G_T and G_R are transmitter and receiver antenna gains, f is the carrier frequency and d is the distance between the transmitter and receiver. Free-space path loss has inverse square law dependence with the distance so received power is reduced by 6 dB when the range is doubled (20 dB/decade decay with distance). In the same manner path loss increases with the increase of carrier frequency. While increasing the antenna gains may compensate for the loss due to high operating frequency in point-to-point links, in mobile radio links this is not possible due to required omnidirectional coverage. The equation for free-space path loss is only valid when distance d is in the far-field of the transmitting antenna.

In practical radio channels free-space model conditions do not apply and further corrections to the path loss have to be accounted. A simple but practical scenario includes the direct path and ground reflected path between transmitter and receiver. This model has been found to be accurate for path loss prediction over distances of several kilometres for mobile radio systems and for the line-of-sight microwave links. Assuming that transmitter and receiver antenna heights h_T, h_R are much smaller than the distance, path loss is given by

$$\begin{aligned} L_{ground\,re\,fl.}[dB] &= 40\log(d[m]) - 10\log(G_T[dB]) \\ &\quad - 10\log(G_R[dB]) - 20\log(h_T[m]) \\ &\quad - 20\log(h_R[m]) \end{aligned}$$

In this case path loss is independent from carrier frequency; however, the inverse fourth power law decay is observed when increasing distance. For distances that are larger than few tens of kilometres the Earth curvature should be taken into account and the values of the reflection coefficient should be modified (2). Also of importance is dependence of the reflection coefficient on polarization and materials of the reflected surface (2, 3).

Diffraction mechanism allows the radio link to be maintained even when receiver is in the obstructed area, shadowed by the object. Huygen's principle could be used to explain the diffraction (2). For practical applications Fresnel zones should be considered (4). Simple knife-edge diffraction model is often used to calculate signal attenuation. Objects within Fresnel zones may cause diffraction loss which is the function of the dimensionless Fresnel-Kirchoff diffraction parameter

$$\upsilon = h\sqrt{\frac{2(d_1+d_2)}{\lambda d_1 d_2}}$$

The geometry of knife-edge diffraction is depicted in Fig. 2. Diffraction loss can be translated into path loss relative to free-space loss, and it is given by nomograms or by approximations which are expressed as

$$L(\upsilon)(dB) = \begin{cases} 20\log(0.5 - 0.62\upsilon) & -0.8 < \upsilon < 0 \\ 20\log[0.5\exp(-0.95\upsilon)] & 0 < \upsilon < 1 \\ 20\log[0.4 - \sqrt{0.1184 - (0.38 - 0.1\upsilon)^2}] & 1 < \upsilon < 2.4 \\ 20\log(\frac{0.225}{\upsilon}) & \upsilon > 2.4 \end{cases}$$

In the cases when there are more than one objects affecting the radio propagation multiple knife-edge model should be introduced. Various approximate solutions have been presented, with different level of accuracy. Their comparison is given in (2).

Path Loss Prediction Models. Prediction models have evolved extensively in UHF/VHF bands addressing the growing importance of mobile radio services. Most models provide the prediction of the median path loss, the loss that is not exceeded at 50% of locations and for 50% of the time (2). Knowledge of the radio signal statistics further allows estimation of the signal variability and prediction of the area where specified signal strength is achieved for a given percentage of location. Path loss models differ in applicability to different terrain profiles and cover different level of details, from general to very specific scenarios. Most of the models are empirical, based on the interpretation of the measured data in the particular area by fitting curves or analytical expressions. The advantage of this approach is that it takes into account the variety of propagation factors through actual field measurements. However, the models are strictly applicable only to the environments characterized in measurements and additional measurements are necessary in different frequency bands and on different locations to provide correction factors. Some of the commonly used models for outdoor and indoor environments will be discussed in subsequent sections. For more detailed comparison reader is referred to (2, 3).

Outdoor Prediction Models. The significance of two models has been established in practice in VHF/UHF band: the Hata model and the Walfisch-Ikegami model (8). Both models are empirical, derived from experimental data, and are extensively used in commercial computer-aided prediction tools. In general usage and accuracy of models depends on the propagation environment: Hata's model provides good accuracy in urban and suburban environments, Walfisch-Ikegami model is widely used for dense urban environments and micro cells. Extensions to both models are provided by The European Co-operative for Scientific and Technical research (COST) to cover PCS band (9).

The Hata model, based on measurements of Okumura, has established empirical mathematical formulas for the path loss and considerably enhanced practical value of the Okumura method (2). It is restricted for the following range of parameters: frequency f 150-1000 MHz, height of the base station antenna h_{te} 30-200 m, height of the mobile unit antenna h_{re} 1-10 m and distance d 1-20 km. Standard formula for basic transmission loss in urban area is given by

$$L_{urban}[dB] = 69.55 + 26.16\log(f[MHz]) - 13.82\log(h_{te}[m]) \\ -a(h_{re}) + (44.9 - 6.55\log(h_{te}[m]))\log(d[km])$$

where f is the carrier frequency, h_{te} is the effective transmitter antenna height, h_{re} is the effective receiver antenna height, d is the distance and $a(h_{re})$ is the correction factor for effective receiver antenna height which is a function of the cell size. The correction factor for mobile antenna height for small to medium size city (urban area) is computed as

$$a(h_{re})[dB] = (1.1\log(f[MHz]) - 0.7)h_{re}[m] \\ - (1.56\log(f[MHz]) - 0.8)$$

and for a large city (dense urban area) as

$$a(h_{re})[dB] = \begin{cases} 8.29(\log(1.54h_{re}[m])^2 - 1.1, & f_c \leq 300MHz \\ 3.2(\log(11.75h_{re}[m]))^2 - 4.97, & f_c \geq 300MHz \end{cases}$$

Suburban area path loss can be calculated using the modified equation

$$L_{suburban}[dB] = L_{urban}[dB] - 2(\log(f[MHz]/28))^2 - 5.4$$

and path loss in open rural areas is given by

$$L_{rural}[dB] = L_{urban}[dB] - 4.78(\log(f[MHz]))^2 \\ - 18.33\log(f[MHz]) - 40.98$$

The Hata model is well suited for large cell mobile systems, but not for the PCS cells with the radius on the order of 1 km. The COST analysis of several measurements conducted in European cities for PCS band have resulted in extended range of parameters for Hata's model to include 1800 MHz frequency band (9). The COST-231 proposed model for path loss is given by

$$L_{urban}[dB] = 46.3 + 33.9\log(f[MHz]) - 13.82\log(h_{te}[m]) - a(h_{re}) \\ +(44.9 - 6.55\log(h_{te}[m]))\log(d[km]) + C_M$$

where all parameters are defined for original Hata's formula and

$$C_M = \begin{cases} 0dB, & for\ medium\ sized\ city\ and\ suburban\ areas \\ 3dB, & for\ dense\ urban\ area \end{cases}$$

COST-231-Hata's model is valid for scenarios with base station heights larger than roof tops in the vicinity of the base station. This is the case in large cells and small cells (maximum range 1–3 km), and path loss is determined largely by diffraction and scattering at roof tops in the vicinity of mobile unit. Main rays propagate in this case above the roof tops. In practical situation it is not recommended to extend the range of validity on base station heights below the roof tops, like in the micro cell scenario. Wave propagation in micro cells (maximum range 0.5–1 km) is determined by diffraction and scattering around buildings, and main rays propagate in street canyons. This phenomena are further addressed by the Walfish-Ikegami model.

The approach of Walfish-Ikegami model is restricted to radio paths that are obstructed by buildings and is not applicable if a line-of-sight path exist between the transmitter and receiver antennas within a street canyon. Few parameters are introduced to describe the character of the dense urban environment as depicted in Figs. 3 and 4: heights of buildings, widths of roads, building separation and road orientation with respect to the direct radio path.

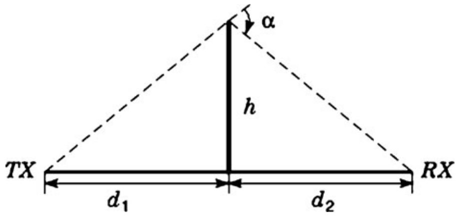

Figure 2. Knife-edge diffraction geometry.

However, the model still considers only characteristic values and there is no consideration of the topographical database of the buildings. The models can be applied even when no data on the urban parameters are available by using the default values, which increases the prediction error. For a line-of-sight case in a street canyon the path loss is equal to

$$L_b[dB] = 42.6 + 26\log(d[km]) + 20\log(f[MHz]),\ \ for \quad d \geq 20m$$

where the first constant in formula is determined so L_b is equal to free-space loss for 20 m. Otherwise the model is composed by three terms and restricted by the free space loss

$$L_b[dB] = \begin{cases} L_O + L_{rts} + L_{msd} \\ L_O \ \ for \quad L_{rts} + L_{msd} \leq 0 \end{cases}$$

The free space loss is given by

$$L_O[dB] = 32.4 + 20\log(d[km]) + 20\log(f[MHz])$$

the roof-top-to-street diffraction and scatter loss is

$$L_{rts}[dB] = -16.9 - 10\log(w[m]) + 10\log(f[MHz]) + 20\log(\Delta h_{mobile}[m]) + L_{ori}$$

where

$$L_{ori}[dB] = \begin{cases} -10 + 0.354\varphi[deg] \ \ for \quad 0 \leq \varphi < 35^o \\ 2.5 + 0.075(\varphi[deg] - 35) \ for \quad 35 \leq \varphi < 55^o \\ 4.0 - 0.114(\varphi[deg] - 55) \ for \quad 55 \leq \varphi \leq 90^o \end{cases}$$

and

$$\Delta h_{mobile} = h_{roof} - h_{mobile}$$
$$\Delta h_{base} = h_{base} - h_{roof}$$

The multi-screen diffraction loss

$$L_{msd}[dB] = L_{bsh} + k_a + k_d\log(d[km]) + k_f\log(f[MHz]) - 9\log(b[m])$$

consists of number of terms:

$$L_{bsh}[dB] = \begin{cases} -18\log(1 + \Delta h_{base}[m]) & h_{base} > h_{roof} \\ 0 & h_{base} \leq h_{roof} \end{cases}$$

with

$$k_a[dB] = \begin{cases} 54 & h_{base} > h_{roof} \\ 54 - 0.8\,\Delta h_{base}[m] & d \geq 0.5\,km \quad and \quad h_{base} \leq h_{roof} \\ 54 - 0.8\,\Delta h_{base}[m]\,d[km]\,/0.5 & d \leq 0.5\,km \quad and \quad h_{base} \leq h_{roof} \end{cases}$$

representing the increase of path loss for base station antennas below the roof tops of the adjacent buildings,

$$k_d[dB] = \begin{cases} 18 & h_{base} > h_{roof} \\ 18 - 15\,\dfrac{\Delta h_{base}}{h_{roof}} & h_{base} \leq h_{roof} \end{cases}$$

controling the dependence of the multi-screen diffraction loss versus distance and

$$k_f[dB] = -4 + \begin{cases} 0.7(f[MHz]/925 - 1) \ \ for \quad urban \quad and \quad suburban \quad ar \\ 1.5(f[MHz]/925 - 1) \ \ for \quad dense \quad urban \quad area \end{cases}$$

controling the dependence of the multi-screen diffraction loss versus radio frequency.

COST-231-Walfish-Ikegami model is restricted for the following range of parameters frequency f 800–2000 MHz, height of the base station antenna h_{base} 4–50 m, height of the mobile unit antenna h_{mobile} 1–3 m and distance d 0.02–5 km If the data on the structure of buildings and roads are unknown the default values may be used: for b 20–50 m, $w = b/2$, h_{roof} 3 m times the number of floors plus the roof height, roof height is 3 m for pitched, 0 m for flat and ϕ is 90 degrees.

The COST-Walfish-Ikegami model has been verified for frequencies in the 900 and 1800 MHz band and radio path lengths from about 100 m to 3 km (9). The path loss has very steep decay versus the height of the base station antenna when the later one is close to the height of adjacent buildings and this case generally results in large prediction errors. Prediction errors are larger when $h_{base} \approx h_{roof}$. The performance of the model is poor for $h_{base} << h_{roof}$ because it does not consider wave guiding in the street canyons and diffraction at corners. For good performance of small cell area coverage the base station antenna should be installed several meters (4 m or more) above the maximum roof tops of adjacent buildings within a radius of few hundred meters (e.g. 150 m). The prediction error may be quite large for micro cells which requires detailed knowledge on streets and buildings (9).

Indoor Prediction Models

Progress of personal communication services has established the need for accurate models that can support path loss prediction inside and into the buildings. Major characteristics of indoor radio channel are much smaller distances than in outdoor cases and higher variability of the environment. Building floor plan, construction materials and the building type greatly influence propagation characteristics. Partition losses at the same floor and partition losses between floors have been extensively measured and categorized (3).

For radio transmission being originated outdoors path loss due to signal penetration into building should be considered. Number of factors have been found to influence penetration loss including signal frequency, antenna pattern and antenna position. Typical values for penetration loss measured in 900 MHz band are on the order of 12 dB, with 6 dB lower loss when building front had the windows (3).

Losses within buildings and penetration losses can be combined in the model characterizing propagation into the building (8)

$$L[dB] = L_{mean}[dB] + 10\,n\log(d[m]) + kF[dB] + pW_I[dB] + W_E[dB]$$

where L_{mean} is the mean path loss from transmitter to the building, n is power exponent of the distance dependence, d is the distance into the building, k number of floors between transmitter and receiver, F floor loss factor, p number of interior walls between transmitter and receiver, W_I the internal wall loss factor and W_E the external wall loss. For inbuilding path loss L_{mean} becomes path loss at 1m distance from the transmitter antenna and W_E is not used. The wall loss factor depends on the construction materials and varies from 0.4 to 29 dB in the 900 MHz band.

Analysis of the experiments carried out in UK in the 900 MHz band in a typical office building with several floors resulted in path loss model

$$L[dB] = L_0[dB] + 10\,n\log(d[m]) + kF$$

where L_0 is path loss at 1m distance from the transmitter antenna, n is distance power law coefficient, d is vertical range in the building, k number of floors taken into consideration and F attenuation per floor in dB. At 900 MHz band n was approximately 4, $L_0 = 30dB$, and F is 5.4 dB per floor (8).

Recent advanced methods for coverage prediction within buildings rely on site specific propagation models and databases which support ray tracing method for deterministically modeling the propagation environment. As the three dimensional building models and databases become widespread, deterministic methods may prevail for determining the path loss in a wide range of operating conditions (3).

Coverage Prediction Using Path Loss Models

Propagation loss models can establish median path loss as a function of the distance between the transmitter and the receiver. Both theoretical and empirical models indicate that received signal power decreases logarithmically with distance with a slope of 10n dB per decade. The value of the n ranges from 2 for a free-space propagation to larger values for obstructed paths (e.g. between 3–5 for urban mobile radio, 4–6 for obstructed indoor propagation). However, this model does not consider different clutters that may results in different signals levels for the same distance. The value of the path loss at a particular location is random variable with log-normal distribution (Gaussian when measured in dB units) about the mean value which is distance dependant. Log-normal statistics describes the shadowing effect at different locations. In practice values of the exponent n and standard deviation of the log-normal fading are determined from measured data. Shadowing is often referred to as a long-term fading.

The random effects of shadowing cause the signal level at certain locations to be lower than the specified level. Therefore it is useful to compute the coverage area with a certain radius, in particular the percentage of the area with a received signal equal or greater than the specified level. Trade-off between the percentage of coverage (for a given exponent loss n and shadowing variance) and the amount of fading margin is important element of the power budget. Details on the coverage computation procedure could be found in (3, 4).

EFFECTS OF ENVIRONMENT

In addition to large-scale path loss different attenuation factors due to environment may influence the power budget. Additional signal attenuation may be result of natural phenomena or man created environmental conditions.

Weather Effects

For microwave radio with line of sight paths different weather related effects may contribute to fading and path loss. Examples include absorption and scattering by snow, hail, fog and rain. Dry snow does not have significant effect on the frequencies below 30 GHz, however wet snow may cause larger attenuation. In general this attenuation is not of the concern. However, degradation of antenna characteristics due to snow and ice build-up on the surface may influence antenna directivity and precautions are taken by using protective radomes. For different types of antennas attenuation loss due to accumulated ice is between 2–7 dB. Another effect of snow is that it may affect the reflection coefficient, and in most cases, depending on the type of snow reflection coefficients is close to 1 (7).

Molecular Absorption

The absorption from oxygen and water vapour in the atmosphere is the additional factor that has to be considered at certain frequencies. The atmosphere extends to an altitude of approximately 20 km, yet it represents a path loss source that cannot be neglected.

Specific attenuation (given in dB/km) is given in a form of figure in (8). Local maxima of attenuation occur at frequencies around 22 GHz for the water vapour, and 60 and 120 GHz for the oxygen. The respective values are around 0.2 dB/km for water vapour, and around 15 dB/km and 2 dB/km for the oxygen. However, system comparison indicate that water-vapour attenuation is far less important than the rain attenuation even in the frequency range of around 20 GHz.

Rain Attenuation

Two components are important when considering the rain attenuation. First one is polarization related loss due to non-spherical shape of rain drops. The horizontally polarized waves experience larger attenuation than the vertically polarized ones. This can also result in cross-polarization effect that may be harmful for microwave systems using channel planning based on different polarization. The frequency related attenuation depends on the drop-size distribution, and is not considered significant for the signals below 11 GHz for locations with rain climate similar to northern hemisphere. For tropical areas critical frequency is as low as 5 GHz.

Rain attenuation statistics prediction is based on rain-rate data, which depends on rainfall microstructure. The ITU-R model is based on rain rate $R_{0.01\%}$ exceeded for 0.01% of the time with an integration time of 1 min. This information can be obtained from local measurements, and in the case that measurements are not available from ITU-R reports. The specific attenuation is calculated as

$$\gamma_R = k\,R^{\alpha}$$

where R is the rain rate in mm/hour, and k, α are regression coefficients which are the function of frequency and polarization (11)

$$\begin{aligned} k &= \frac{k_H + k_V + (k_H - k_V)\cos^2(\theta)\cos(2\tau)}{2} \\ \alpha &= \frac{k_H\alpha_H + k_V\alpha_V + (k_H\alpha_H - k_V\alpha_V)\cos^2(\theta)\cos(2\tau)}{2k} \end{aligned}$$

where θ is the path elevation angle and τ is the polarization tilt angle relative to horizontal, $\tau = 45^\circ$ for circular polarization, and other parameters are given in a Table 3. The effective path length in this case is calculated by multiplying the actual path length L with a reduction factor

$$r = \frac{1}{(1 + L/L_0)}$$

where

$$L_0 = 35 \quad \exp(-0.015\,R_{0.01}\%)$$

The estimate of the path attenuation exceeded for the 0.01% of the time is given by

$$A_{0.01}[dB] = \gamma_R\, r\, L$$

Attenuation Due to Clouds and Fog

In Earth-space radio systems for frequencies higher than 10 GHz the attenuation due to clouds may be important factor affecting the system performance. The mechanism of attenuation is well understood since clouds and fog consist of small droplets and it is possible to express attenuation in terms of total water content per unit volume (12) as

$$\gamma[dB/km] = K_l[dB/km \quad x \quad g/m^3] \quad M[g/m^3]$$

where γ is specific attenuation within the cloud, K_l specific attenuation coefficient and M liquid water content of the cloud or the fog. The typical water content is about 0.05 g/m^3 for medium fog with visibility of the order of 300 m and 0.5 g/m^3 for thick fog with visibility of the order of 50 m. The specific attenuation coefficient for the frequencies up to 1000 GHz is given by

$$K_l = \frac{0.189\ f[GHz]}{\varepsilon''(1+\eta^2)} \qquad [dB/km/g/m^3]$$

where

$$\eta = \frac{2+\varepsilon'}{\varepsilon''}$$

and complex dielectric permitivity of water is

$$\begin{aligned} \varepsilon''(f) &= \frac{f(\varepsilon_0 - \varepsilon_1)}{f_p[1+(f/f_p)^2]} + \frac{f(\varepsilon_1 - \varepsilon_2)}{f_s[1+(f/f_s)^2]} \\ \varepsilon'(f) &= \frac{\varepsilon_0 - \varepsilon_1}{[1+(f/f_p)^2]} + \frac{\varepsilon_1 - \varepsilon_2}{[1+(f/f_s)^2]} + \varepsilon_2 \end{aligned}$$

where $\varepsilon_0 = 77.6 + 103.3(\theta - 1)$, $\varepsilon_1 = 5.48$, $\varepsilon_2 = 3.51$ and $\theta = 300/T$, T being temperature in Kelvin. The principal and secondary relaxation frequencies are given by

$$f_p[GHz] = 20.09 - 142(\theta - 1) + 294(\theta - 1)^2$$

$$f_s[GHz] = 590 - 1500(\theta - 1)$$

Attenuation due to fog becomes significant at frequencies around 100 GHz with specific attenuation of 0.4 dB/km for medium fog, and 4 dB/km for thick fog.

Foliage

The leaves of maple, oak, hickory and similar trees can cause additional signal loss in the frequency range above 400 MHz. In the winter leaves fall and received signal is generally stronger than in the summer time. Various factors contribute to this type of loss: operating frequency, type of leaves, height of the trees, thickness of the foliage. In general, design practice is to add about 10 dB of allowance in the forest area in addition to predicted path loss (7).

In tropical rain forest area leaves do not fall and their shape is different producing different attenuation. Major studies carried out in the 50 to 800 MHz band indicate that loss increases linearly in log scale from 35 dB/decade at 50 MHz up to 40 dB/decade at 800 MHz. The foliage loss is approximately proportional to the operating frequency raised to fourth power which is a good approximation for horizontal polarization. Vertical polarization experienced larger loss in general by 8 to 25 dB at 50 MHz and 1 to 2 dB at 800 MHz. For a transmitting antenna over the tree tops and receiving antenna located in the trees delay spread of $0.2\mu s$ has been reported due to foliage (7).

Constructions

Although path loss models tend to capture difference between build up and rural areas, few specific conditions should be emphasized. Street orientation has the channel effect on the received signal strength. For the mobile unit closer to the base station in the range of 1-2 miles the difference between signal strengths in the case when the street is in line with base station and when it is perpendicular to the base station is about 10 dB. This phenomenon diminishes for distances over 5 miles.

Mobile receiver moving into tunnel experiences a signal loss which is dependant on the frequency and the transmitter/receiver positioning. Experiments carried out for the transmitter at the entrance of the tunnel and the mobile receiver moving through the tunnel show that at 300 m inside the tunnel about 4 dB loss is observed at 1 GHz. Attenuation at lower frequencies was much higher approaching 20 dB at 400 MHz (7).

Results from study in New York have established inverse law dependence of the path loss versus the distance with a inverse law from 4 at 900 MHz reducing to 2 at 2400 MHz. Above this frequency loss is less than in free space indicating some sort of guiding mechanism.

The underpass effect results in signal drop between 5 to 10 dB when a mobile receiver drives through. The period of the attenuation depends on the vehicle speed. In cellular radio this type of effect usually does not affect the voice channel.

SMALL-SCALE VARIATIONS: MULTIPATH FADING

Small-scale **fading** (also referred as fast-fading, short-term fading or simply fading) describes the fluctuations of the received radio signal over a short period of time or a short travel distance. Substantial variations of the signal amplitude are mainly caused by local multipath propagation. Multiple replicas of the signal combine at the receiver antenna in different ways depending on the time of arrival and the phase of individual components, resulting in fluctuations that are fast compared to longer-term variation in mean signal level (shadowing). These fluctuations are seen in the received signal as large changes of amplitude over a small interval of time, random frequency modulation due to Doppler shifts on different multipath arrivals and possible time dispersion due to different propagation delays of multipath components. Although fading is basically spatial phenomenon, it is experienced as temporal phenomenon by a radio receiver moving through the multipath field. Motion of surrounding objects (reflectors, scaterrers) also impacts fading characteristics both in case of static transmitter/receiver pair, and or when it is comparable to the speed of the mobile. Fast fading is observed over distance of about half a wavelength, with fades of depth about 20 dB being frequent and larger fades of more than 30 dB being less frequent, but not uncommon (4).

In practice there is no clear border between slow and fast fading. It is commonly assumed that in the areas where multipath occurs radio signal consists of a local mean value which is constant over a small area, but varies slowly as the receiver moves, and a fast fading component superimposed on the slowly varying signal.

In mobile radio communications multipath fading occurs in urban areas where the height of the mobile antenna is much lower than that of the surrounding buildings and there is no line-of-sight path between the transmitter and the receiver. Propagation is mainly determined by scattering and diffraction around the buildings. In fixed radio links where the line of sight propagation exists, multipath propagation is caused by the reflection from the ground or surrounding objects.

Except for the multipath propagation, other factors may influence the fading characteristics. Relative motion between the transmitter and the receiver in mobile and low-orbit satellite communications results in a phase change due to the difference in path lengths, which can be observed as a Doppler frequency shift in each propagation path as illustrated in Fig. 5. The phase change is given by

$$\Delta\phi = \frac{2\pi\,\Delta l}{\lambda} = \frac{2\pi\,v\Delta t}{\lambda}\cos\theta$$

and the corresponding frequency shift is

$$f_d = \frac{1}{2\pi}\quad\frac{\Delta\phi}{\Delta t} = \frac{v}{\lambda}\cos\theta$$

where v is the speed of the mobile. If mobile is moving towards the transmitter, positive Doppler shift results in increase of the received signal frequency, and for the mobile moving away from the transmitter negative Doppler shift results in decrease of the received signal frequency. The angle between the incoming wave and the direction of motion determines the maximum rate of phase change which occurs when the waves are coming directly behind or ahead the mobile.

While characteristics of a fading channel are consequences of physical propagation phenomena, their impact on the received radio signal depends on the transmitted signal characteristics as well, primarily on the bandwidth of the transmitted signal. This relationship will be addressed in more details in the following section.

Statistical Characterization of Multipath Fading Channels

Two major characteristics of a multipath fading channel can be observed by transmitting a short pulse (ideally impulse). The first one is time spread (dispersion) of the channel resulting in more than one received pulse. They differ in time of arrival, amplitude and phase. Second characteristics is the time-variation; the structure of multipath varies with time, and the same sounding experiment will produce different number of echoes with different arrival times, amplitudes and phases if repeated at a different time.

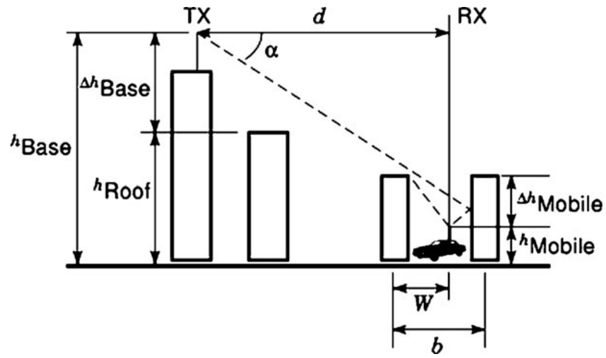

Figure 3. Parameters used in Walfish-Ikegami model.

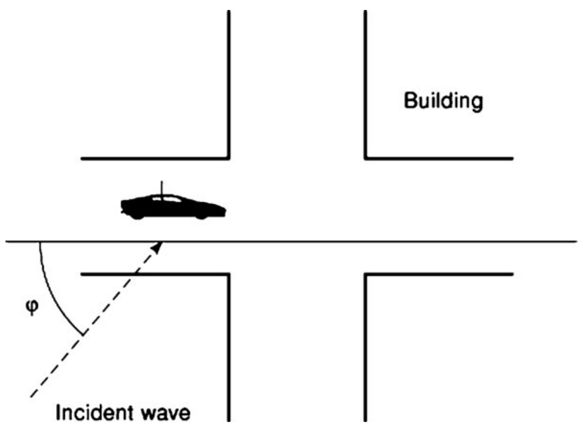

Figure 4. Definition of the street orientation in Walfish-Ikegami model.

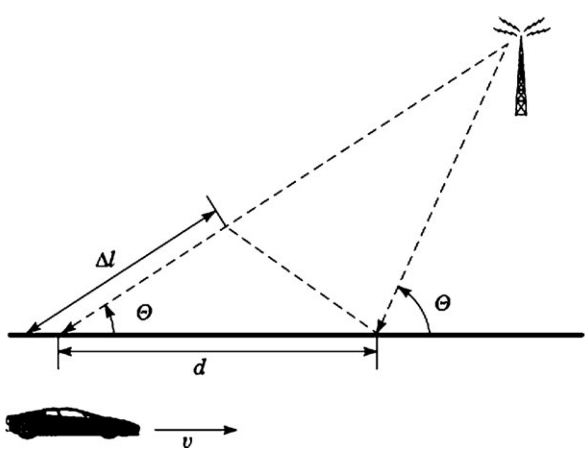

Figure 5. Geometry for illustration of Doppler effect.

Consequently a multipath fading channel is characterized in statistical terms. The impulse response of the time-varying multipath channel is represented by time-varying linear filter $h(\tau, t)$ where variable t represents time of transmission and τ represents the channel multipath delay, for a fixed value of t. Thus $h(\tau, t)$ represents the response of the channel at time t due to the impulse applied at time t-τ. When received signal contains discrete multipath components, the equivalent complex baseband impulse response model is given by

$$h(\tau, t) = \sum_n \alpha_n(t) \exp(-j2\pi f_c \tau_n(t)) \quad \delta[\tau - \tau_n(t)]$$

where $\alpha_n(t)$ is the attenuation factor for the n-th path, $\tau_n(t)$ is the propagation delay for the n-th path and f_c is the signal carrier frequency.

Statistical characterization of the fading channel is detailed in (15). In summary, channel impulse response $h(\tau, t)$ can be represented as a zero-mean complex-valued Gaussian process, so that its envelope at any time instant is Rayleigh distributed and the phase is uniform (16). In the case when there exist a strong signal component (fixed scatterer, strong reflection) $h(\tau, t)$ does not have a zero-mean and the statistics of the channel is Rician (17). Regardless of the probability distribution, a wide sense stationary channel is characterized by two functions.

Multipath intensity profile or *power delay profile* of the channel gives the average output power of the channel as a function of the time delay τ. The multipath spread T_m of the channel represents a range of values of τ over which the power delay profile is essentially nonzero.

In practice power delay profiles are found by averaging instantaneous power delay profile measurements over a local area to determine an average small-scale power delay profile. A typical spatial separation for sampling the channel is quarter of a wavelength, over receiver movements no greater than 6 m in outdoor channels and no greater than 2 m in indoor channels (1). Several statistical parameters quantify the delay characteristics of the channel. The *mean excess delay* is the first moment of the power delay profile

$$\tau_{mean} = \frac{\sum_n \alpha_n^2 \, \tau_n}{\sum_n \alpha_n^2}$$

and rms delay spread is the square root of the second central moment of the power delay profile

$$\sigma_\tau = \sqrt{\overline{\tau}^2 - \tau_{mean}^2}$$

where

$$\overline{\tau}^2 = \frac{\sum_n \alpha_n^2 \, \tau_n^2}{\sum_n \alpha_n^2}$$

All delays are measured relative to the first detectable signal arriving at the receiver at $\tau_0 = 0$. Also relative amplitudes of multipath components are used in computations. Typical values of rms delay spread are on the order of microseconds for outdoor mobile radio channels and nanoseconds for indoor radio channels (3).

The *maximum excess delay* is defined as $\tau_X - \tau_0$, where τ_X is the maximum delay at which the multipath component is X dB lower than the strongest multipath arrival, which is not necessarily at τ_0. It is important to note that all parameters highly depend on the noise threshold in the power delay profile which is used to differentiate multipath components and thermal noise.

The Fourier transform of the power delay profile determines frequency-coherence properties of the channel. The inverse of the multipath spread

$$(\Delta\ f)_c = \frac{1}{T_m}$$

is a measure of the *coherence bandwidth* of the channel. Effectively, two signal components separated in frequency by more than a coherence bandwidth will by affected differently by a channel i.e. they will fade independently. Coherence bandwidth is practically quantified by a level of frequency correlation function and the rms delay spread. For example, for a correlation function above 0.9 the coherence bandwidth is approximately (3)

$$(\Delta\ f)_c \approx \frac{1}{50\,\sigma_\tau}.$$

Time variations of the channel can be described by the *Doppler power spectrum* of the channel which gives the signal intensity as a function of a Doppler frequency. The range of Doppler frequency over which the Doppler power spectrum is essentially nonzero is called *Doppler spread* B_d of the channel. It measures the spectral broadening caused by the time variations. The inverse of the Doppler spread

$$(\Delta t)_c \approx \frac{1}{B_d}$$

is the *coherence time* of the channel, and defines the time interval over which the characteristics of the channel can be regarded as constant. If the signal components are separated in time more than a duration of a coherence time, they can be regarded as independent.

However, B_d is not uniquely defined. On the other hand, if a pure sinusoidal tone of frequency f_c is transmitted over a channel, the received signal is within $[\,f_0 - f_d,\ f_o + f_d]$, where f_d is the Doppler shift, showing that spectral broadening depends on the velocity of the mobile as well. A commonly used relationship describing the coherence time is given by (3)

$$(\Delta t)_c = \frac{0.423}{f_d}$$

Scattering function of the fading channel, representing power density as a two dimensional function of delay and Doppler frequency, and different correlation functions of the channel are related via different Fourier transform pairs (15).

In a case of Rayleigh fading channel, two important statistical parameters may clarify design choices of diversity techniques and error control codes. The *level crossing rate* is defined as the expected value of the rate at which the envelope of the Rayleigh fading process crosses a specified level in a positive going direction (3). The number of level crossings per second is given by

$$N_R = \sqrt{2\pi}\ f_d\, \rho\, e^{-\rho^2}$$

where f_d is the maximum Doppler frequency and $\rho = R/R_m$ is the normalized value of the specified level R by the local rms amplitude of the fading envelope. For example in 900 MHz band with vehicle speed of 48kmph, the maximum value is 39 crossings/sec.

The *average fade duration* defines the average period of time for which the received signal is below a given level R. For a Rayleigh fading, it is given by

$$\overline{\tau} = \frac{e^{\rho^2} - 1}{\rho\ f_d \sqrt{2\pi}}$$

showing the dependency on the mobile speed. For a given fading margin, determined in a power budget, it is important to evaluate the rate at which the received signal falls below the certain level and the average time during which it remains below that level. These parameters directly translate into a SNR reduction during the fade which can be used for the prediction of the radio performance.

Channel Classification

The impact of the fading channel on the received signal depends on the signal characteristics such as signal bandwidth and signal duration. The relationship between signal characteristics and channel parameters will serve as the criterion for channel classification. Let us consider digital communication system with signalling interval T.

For a signalling interval T much larger than the multipath spread of the channel, $T >> T_m$, channel introduces negligible amount of intersymbol interference. On the other hand for a bandwidth of a signalling pulse $W \approx 1/T$, previous relationship imply that the signal bandwidth is much smaller than the coherence bandwidth of the channel, resulting in *frequency-nonselective* or *frequency-flat* fading. This implies that amplitude attenuation and phase shift introduced by channel are the same for all signal components and consequently spectral characteristics of the transmitted signal are preserved at the receiver. Channel distortion is simply a multiplicative random process. In this case multipath components of the signal cannot be resolved. This case is typical for narrowband communication signals. Flat fading may result in deep fades.

In a case where signalling interval is smaller than the multipath spread of the channel, consecutive transmitted symbols interfere with each other at the receiver, producing intersymbol interference. In this case signaling bandwidth W is larger than the coherence bandwidth of the channel, and different signal frequencies are subject to different gains and phase shifts. This type of channel is referred to as the *frequency-selective* channel. Frequency-selective channels are typical for wideband communication systems such as time-division multiple-access systems or spread-spectrum systems.

For a signalling interval T smaller than the coherence time of the channel, $T \ll (\Delta t)_c$, channel characteristics can be regarded as fixed during the symbol interval. In the frequency domain, signaling bandwidth is much larger than the Doppler spread of the channel. This condition corresponds to a *slow fading channel*. However, the rate of change of the channel depends both on signalling bandwidth and the velocity of the transmitter or the receiver.

For a signaling interval larger than the coherence time, channel impulse response changes within the duration of the symbol. This results in spectral broadening which in turn leads to signal distortion since the signal bandwidth is smaller than the Doppler spread of the channel. This type of fading is referred to as *fast fading* or *time-selective fading*.

It should be pointed out that there is a slight inconsistency in terminology. Term fast fading is used to describe multipath fading which is fast compared to shadowing effects (3). On the other hand, multipath fading is characterized as fast fading when signaling interval is larger than the coherence time of the channel (15). Although it might be somewhat confusing, precise meaning of the term can be deducted from the context in which it is used.

For a physical channel with given multipath delay spread and Doppler spread, different scenarios may arise depending on the signal design. Slow fading condition is desirable from the standpoint of receiver design since it allows coherent reception, i.e. synhcronization circuits are able to track changes of the signal. It is important to note that even in situations which are characterized as slowly fading i.e. for which $B_d T \ll 1$, receiver performance may be significantly affected by the channel fluctuations via non-perfect synchronization. The amount of degradation depends on the exact value of $B_d T$.

If signal bandwidth W cannot be selected independently of the signalling interval T, for a given fading rapidity, system may be subject to frequency selective of frequency nonselective fading. Slow frequency-nonselective fading introduces constant multiplicative distortion and no ISI. Both coherent and noncoherent receivers can be employed. In the case of slow frequency-selective fading, equalization techniques should be used to compensate for ISI and equalizer operation is possible due to slow variations of the fading. In the case when signal bandwidth can be selected independently of signalling interval (as in spread spectrum systems due to bandwidth expansion) one can simultaneously achieve no-ISI and slow fading condition.

In the case of fast fading and frequency-nonselective channels, noncoherent reception techniques may be employed. On the other hand fast frequency-selective fading presents serious obstacle for a communication system.

The product $B_d T_m$ is called the *spread factor* of the channel and has a fixed value for a given physical channel. For underspread channels, $B_d T_m < 1$, system can be designed to achieve the condition of negligible ISI and slow fading. For digital systems optimal data rate can be derived to satisfy conditions for slow frequency-nonselective fading (13)

$$R = \frac{1}{T} = \sqrt{\frac{B_d}{T_m}}$$

For overspread channels signal design choices are more restricted and noncoherent reception is usually used.

Statistical Models for Fading Channels

While path loss models play crucial role in power budget, determining the range and coverage zone of the radio system, statistical models for small-scale fading play important role in receiver design and evaluation. Performance in the presence of fading is one of the major factors influencing the receiver sensitivity. Few major fading models for different radio scenarios will be summarized in this section.

With the advances in multi-input multi-output wireless systems and space-time coding temporal characterization of the fading channels has been extended to cover space-time propagation effects as well. Details on space-time channel models can be found in reference (29).

Jake's Model. Jake's model of a Rayleigh fading path is based on scattering and is one of the most widely used models for a fading channel. It effectively models frequency-nonselective fading channel with a Doppler spectrum of the form

$$S(f) = \frac{K}{\pi \ f_d \sqrt{1 - (\frac{f - f_c}{f_d})}}$$

where K is the constant which depends on polarization (for detailed model derivation see (12)), f_c is the signal frequency and $f_d = v/\lambda$ is the maximum Doppler shift, v being the speed of vehicle, and λ signal wavelength. This type of Doppler spectrum can be simulated using a model with $N_0 = 8$ sinusoids where in-phase and quadrature components of the fading signal are given by

$$\begin{aligned} x_c(t) &= 2n = 1 \sum \cos(\beta_n)\cos(\omega_n t) + \sqrt{2}\cos\alpha\cos(\omega_d t) \\ x_s(t) &= 2n = 1 \sum \sin(\beta_n)\cos(\omega_n t) + \sqrt{2}\sin\alpha\cos(\omega_d t) \end{aligned}$$

and Doppler shift is given by $\omega_d = 2\pi\ f_d$, frequencies of the sinusoids are $\omega_n = \omega_d \cos(2\pi n/N)$, $N_0 = (N/2 - 1)/2$, and phases $\beta_n = \pi n/N_0$ and $\alpha = \pi/4$. This model shows excellent agreement with Rayleigh distribution, autocorrelation Bessel function and Doppler spectrum (12). This technique can be extended to generate up to N_0 independent fading signals, which is of importance for tapped-delay line fading channel models for frequency-selective fading.

Tapped-Delay Line Model for Outdoor Mobile Systems. Tapped-delay line models are commonly used to specify mobile radio channels. The structure of the model is straightforward, and measurements for a particular frequency band and geographical environment are used to develop model parameters. Model is specified in terms of number of multipath arrivals (usually fixed), relative delays, relative power of multipath components and Doppler spectrum for each component. For a given radio standard these parameters may widely vary not only due to different measurements but also due to different signal characteristics considered.

As an example of tapped-delay line model wideband propagation model for GSM digital cellular standard is summarized. Discrete number of taps (6 and 12) is determined by their time delay and their average power, amplitudes of each tap are Rayleigh distributed and varying according to a Doppler spectrum $S(f)$ (19). Two types of Doppler spectrum are defined: classical to be used in all but one case

$$S(f) = \frac{A}{\sqrt{(1 - \frac{f}{f_d})^2}} \qquad for \quad f \in \quad [-f_d,\ f_d]$$

where $f_d = v/\lambda$ is the maximum Doppler shift, and Rice which is the sum of a classical Doppler spectrum and one direct path

$$S(f) = \frac{0.41}{2\pi\ f_d \quad \sqrt{(1 - \frac{f}{f_d})^2}} + 0.91 \quad \delta(f - 0.7\ f_d)$$
$$for \quad f \in \quad [-f_d,\ f_d]$$

Depending on an environment few typical cases are specified: rural area, hilly terrain and typical urban environment. Example of six tap models are summarized in the Table 4.

With introduction of wider band systems such as Wideband Code Division Multiple Access the model of the propagation channel changes due to different structure of resolvable paths. Moreover, in addition to models where tap delay is fixed as summarized in Table 5, new models with variable tap delay has been introduced (32). The dynamic propagation condition consists of two paths of equal strengths and equal phases, first one is static while the time difference between the two paths is according to

$$\Delta\tau = B + \frac{A}{2}(1 + \sin(\Delta\omega \cdot t))$$

With parameters $B = 1\,\mu s$, $A = 5\,\mu s$ and $\Delta\omega = 40 * 10^{-3}\ s^{-1}$

Other system related test require further sophisticated models such as birth-death model (20) where two paths appear and disappear randomly according to the given statistical model.

Rummler Fading Model for Fixed Microwave Systems. Three path model developed for a line-of-sight microwave radio channels is based on a channel measurements on a typical link in 6 GHz band. Since the differential delay of two multipath components is small, channel transfer function is given by

$$H(f) = \alpha \quad [1 - \beta \quad \exp(-j2\pi(f - f_{\min})\tau_0)]$$

where α is the overall attenuation, β is a shape parameter resulting from multipath propagation, $f_{\min}$ is the frequency of the fade minimum and τ_0 is the relative time delay between the direct and multipath components. By fitting the model to the measurement data parameter α is modeled as log-normal distributed, while distribution for β is given as $(1-\beta)^{2.3}$. Two random variables are statistically independent. For $\beta > 0.5$ the mean of $-20\log\alpha$ is 25 dB, with standard deviation of 5 dB. For smaller values of β the mean decreases to 15 dB. Delay parameter was estimated to be $\tau_0 = 6.3 \quad ns$.

Saleh-Valenzuela Indoor Statistical Model. Saleh-Valenzuela model is based on statistical modeling of the measurements of indoor propagation with multipath resolution of 5 ns. Maximum reported multipath delay spread was on the order of 100–200 ns within the rooms in building and 300 ns in hallways. This model is based on multipath components arriving in clusters. Amplitudes of the components are independent Rayleigh distributed, with variance that decay exponentially with the cluster delay as well as with excess delay within a cluster. The difference from tapped delay line model is that clusters and multipath components within cluster are modeled as Poisson arrival processes with different rates and exponentially distributed interarrival time (3).

Spatial Channel Models. Spatial channel models have been developed over last decade to take into consideration increasingly important use of multiple antennas in radio systems, both at the transmitter and the receiver. The notion of space-time communications channels deserves a whole chapter both in terms of importance and complexity of the material. In addition to small-scale fading spatial channels take into consideration other effects such as angle spread due to space selective fading, scatterers both in local zone and remote, polarization, antenna array topologies etc (29). The most complex models are for Multiple

Input Multiple Output (MIMO) channels – example of one used in system evaluation has been presented in (33).

Effects of Fading on System Performance

Analog radio systems are relatively narrowband, even for high-capacity microwave links, and major factor that is contributing to the degradation of system performance is the fading depth (8). Therefore, flat-fading margin should be provided to insure robust performance. It defines the additional loss that can be tolerated before the system SNR reaches an unacceptable level.

For a digital system BER is the performance measure and it should be specified for particular fading condition. Important note is that performance degradation of the digital signal is not graceful as in the analog one. For a narrowband system performance may be specified in terms of flat-fading margin. For wideband systems, effects of frequency-selectivity are more emphasized and use of equalization techniques and/or other forms of diversity are necessary to insure desired performance. In that case, fading margin may not be specified explicitly, but the receiver sensitivity may be specified for fading conditions, and will include the effects of fading countermeasures.

MAN-MADE NOISE AND INTERFERENCE

Radio systems are either noise or interference limited. Characteristics of the noise greatly influence radio receiver design, however most of the radio systems nowadays are interference limited due to coexistence with other radio systems and/or multiple-access schemes employed. Interference level in the radio system is dependant on frequency, time, spatial location and signal separation. Some of the major interference mechanisms for radio systems will be described in this section.

Impulse Noise

Nature and characteristics of the noise are important for the radio reception in two ways. Understanding the noise phenomena leads to more efficient system design. Characterizing noise is crucial in performance prediction of any radio system since it directly affects the receiver sensitivity. Radio systems are subject to different type of noise sources. Effects of thermal noise and receiver noise are well understood and are described by noise density and noise figure. This type of noise is characterized as Gaussian. Atmospheric noise has non-flat spectral density and decreases rapidly with frequency. Galactic noise is due to energy radiation from stars and planets, however it is usually below the level of thermal noise.

The type of noise referred to as man-made noise is impulsive in nature and is generated by radiation of the electric equipment of various kinds such as vehicle ignition systems, alternator, power lines, neon lights, industrial noise form current switches and from the various domestic appliances. This type of noise can significantly affect the performance of radio system. Its spectral characteristics are very irregular, and also it varies in level with location and time. Impulsive noise may be viewed as a combination of successive impulses, with random amplitudes and random time spacing. Different parameters have been established to qualitatively describe this type of noise including mean or average voltage, peak voltage, amplitude probability distribution, pulse height distribution, level crossing rate, pulse duration distribution, etc. (2).

The median value of the man-made noise power is given as a function of frequency up to 200 MHz by (10)

$$F_{am} = c - d\log(f[MHz])$$

where c and d take values form Table 6. In the frequency range 200 MHz to 900 MHz the relationship is given by

$$F_{am} = 44.3 - 12.3\log(f[MHz])$$

As mentioned earlier, ignition impulses from motor vehicles produce the significant component of man-made noise, in particular in VHF band. Noise amplitude distribution in this case can be determined by

$$A[dB(\mu V/MHz)] = 106 + 10\log V[vehicles/km^2] - 28\log(f[MHz])$$

where V is the traffic density.

Predicting the performance of the receiver in the presence of impulse noise is extremely difficult compared to relatively straightforward method for Gaussian noise. Different techniques are presented in (2).

Cochannel Interference

Cochannel interference is the consequence of better spectrum utilization. It is generated in a scenario where two or more communication signals are assigned to the same frequency and operate at the same time. CCI may be caused by a number of scenarios depending on the radio service: bad route planning for microwave radio links, failure to carry out co-ordination procedures for the same type of service, cross-polarization degradation in cross-polar frequency plan, cochannel interference due to frequency reuse in mobile radio systems, and cochannel interference due to multiple-access method like in CDMA.

In any case result of the cochannel interference is the presence of the same-type modulated signal at the receiver input, which may be lower (in cellular TDMA and microwave systems) or on the same level (in satellite and cellular CDMA systems) as the desired signal. In most of the cases cochannel interference problem can be avoided by careful system planning. This will reduce the cochannel interference to the acceptable level such that system performance is not significantly degraded. Technical measures include proper cell design for cellular systems including cell sectorization, power control techniques, and use of orthogonal codes on the forward link in CDMA cellular systems (note that orthogonal codes in CDMA system can be used only to eliminate intracell interference, the interference from neighbouring cells will still be present). In the case when planning methods fail to reduce the amount of interference, solution may be found in more sophisticated receiver design using interference suppression techniques.

Adjacent Channel Interference

Adjacent channel interference is the results of closely spaced radio channel in frequency, so the spectral content of the adjacent channels is spilling into the bandwidth of the desired channel. Factors determining the level of the interference are signal power and spectral distribution, power and spectral distribution of the interfering signal intercepted by the receiver and distance dependence of the transmission losses between interfering transmitter and the desired signal receiver. ACI can be somewhat controlled by tight filtering, however limiting factors are spectral utilization of the system and, on the other hand, finite slope that can be achieved for filtering functions. Another way to improve adjacent channel performance is the design of spectrally efficient modulation techniques that have largely concentrated power in a limited bandwidth with good roll-off properties, which in turn requires more elaborated demodulation schemes. Also adjacent channel interference in cellular systems can be avoided by careful frequency planning, so adjacent channels are not used within the same cell or in the neighbouring cells.

Near-Far Interference

Near-far interference in a radio system occurs when two different signals arrive at the receiver input with largely different power levels and may be interpreted as a special form of cochannel interference. This may be result of difference in distance of different transmitters from the same receiver, or in some cases in may be the result of fading conditions on the link, so that closer transmitter may have lower power at the receiver when subjected to a deep fade. The consequence of near-far problem is the performance degradation for the weaker signal reception. Although near-far problem is general for the radio reception, it is emphasized in direct-sequence CDMA systems where all signals are present at the receiver input, and may significantly reduce the capacity of CDMA system. Countermeasures for near-far problem include tight power control which should insure that all signals arrive to the receiver with the power level required for desired operation. Another possibility for cellular systems is to develop a frequency management plan that can greatly reduce the possibility of near-far problem. Specific techniques for radio design include multi-user detection techniques described in (22).

RADIO RECEIVER DESIGN

Radio receiver consists of a channel interface followed by the demodulation part (23). Channel interface of the radio receiver is characterized by the receiver selectivity and its ability to process both desired signal and undesired (interfering) signal. In order to achieve specified performance radio receiver must be designed to operate in the presence of noise, fading and interference.

Basic Radio Receiver Parameters

Radio receiver sensitivity determines the input level of the weak radio signal that can be processed by a radio receiver to meet performance requirements. For analog receivers performance measure is given in terms of required SNR, usually specified at the input to the detection circuit. For digital systems performance measure is given in terms of BER or other parameters that can be derived from BER requirement, depending on the type of radio service. Detection algorithm includes coding, interleaving, equalization, diversity and other signal processing functions used to improve radio reception.

Receiver sensitivity is usually expressed by minimum detectable signal level for a required SNR, which includes the knowledge of the system bandwidth and is defined as

$$RS[dBm] = -174dBm + 10\log(NF[dB]) + 10\log(BW[Hz]) + SNR[dB]$$

where NF is the overall system noise figure referred to the input of the receiver and BW is the system noise bandwidth. For digital systems system requirement is commonly given in terms of energy per bit to noise density ratio Eb/No. Relation between Eb/No and SNR is given by

$$Eb/No = SNR \quad BW \quad T$$

where T is the bit duration (23). Depending on the particular radio system and targeted operating environment, performance measure of the system (SNR or BER) can be evaluated under fading conditions or under static conditions. In the first case fading margin is derived from the statistics of the large-scale signal variation such as log-normal fading. Small-scale signal variations are accounted for implicitly during BER evaluation. When BER of the system is specified under static conditions, fading margin is derived from combined statistics of both large-scale and small-scale variation. Fading margin effectively defines the increase in SNR required to preserve the same BER as in the static case.

Radio sensitivity specifies the lowest detectable signal level at the radio input; however, both desired and unwanted signals may be present at the radio receiver input at high levels, and it is important to specify the performance for undesired response rejection. The unwanted strong signal may reduce a nearby weak (desired) signal resulting in radio desensitization. Desensitization level is specified for a 1 dB weak signal reduction and is a consequence of radio front-end saturation.

Another factor that may severely limit the performance of a radio receiver is intermodulation distortion, where unwanted strong signals produce a number of intermodulation products due to receiver front-end nonlinearities. IMD products, particularly third-order products, may fall directly in the bandwidth of the desired signal. To limit the level of the IMD products receiver is required to have high intercept point of the particular order.

Dynamic range (DR) of the receiver defines the range of input signals that a radio receiver can process. Different considerations may be taken into account depending on the radio service; however, the most common spurious-free dynamic range (SFDR) is defined as

$$SFDR[dB] = \frac{2}{3}(IIP3[dBm] - RS[dBm])$$

where IIP3 is the third-order input intercept point.

Modulation and Coding

Choice of the modulation method for radio communication depends on many factors. In all cases the goal of the radio receiver design is to achieve the required quality of the output signal (voice, video, data) using minimum SNR or Eb/No. For power-limited systems, where bandwidth is not of concern (e.g. space communication links) noncoherent M-ary frequency shift keying (FSK) can be used to reduce the required Eb/No for a given BER, at the expense of bandwidth expansion. For a bandwidth-limited systems (like most of the modern radio systems) spectrally efficient, coherent modulation schemes like phase-shift keying (PSK) or quadrature amplitude modulation (QAM) can be used at the expense of increasing required Eb/No as the spectral efficiency increases (5). However, additional requirements for the modulation format in radio communications such as robustness to fading, interference, and nonlinear distortions, motivated development of constant-envelope or near constant-envelope modulation such as offset quadrature phase shift keying (OQPSK) and minimum shift keying (MSK). Different forms of such modulation formats can be found in wireless systems (3). Inherent resistance of a modulation method to the interference can be measured by determining required SIR for a specified BER. Performance functions for digital modulations in the presence of interference are presented for various types and statistical characteristics of interferers in (23). Further advances in signalling methods for radio channels include the application of coded modulation techniques (15).

Additional performance gain can be obtained by using coding techniques at the expense of introducing redundancy to the transmitted digital signal. Channel coding protects the signal from the effects of noise, fading and/or interference. Functions of channel codes are error detection and forward error correction. Both block and convolutional codes have been applied in radio communications (15). Lately increasing consideration has been given to space-time codes (29) and turbo codes (30).

The complexity of modern radio communications systems requires further means of efficiently utilizing channel based on the quality of the link. Different modulation-coding combinations have been developed as well as methods to switch between them as for example in EDGE cellular standard (19). This is one example of more general adaptive modulation and coding techniques that in principle requires a feedback path between the transmitter and receiver, with data transmission rates being varied relative to propagation channel conditions (34).

Diversity Reception for Fading Channels

Diversity reception techniques are used to reduce the effect of fading by exploiting the random nature of the radio channel. Independent or highly uncorrelated signals paths, known as diversity channels, are used to improve quality of the received signal. The concept of diversity is relatively simple: if one of diversity channels is subject to a deep fade another channel with higher signal level can be selected or appropriate combining techniques can be applied to increase the signal level. The probability of all diversity channels being simultaneously below a given level is much less than the probability of any of them being below that level. Therefore, a combination of different diversity channels will be subject to less severe fading conditions compared to any individual channel.

Diversity techniques may address different types of signal fluctuations. Large-scale fading is caused by shadowing due to variations in the propagation profile (hills, mountains) and the surrounding objects (buildings). To reduce the effect of the long-term fading, two geographically separated antennas may be used (for transmission or for reception) and the stronger signal may be selected. This type of diversity, known as macroscopic diversity, is often used in short-wave radio systems to reduce the effects of ionosphere and in cellular systems to improve the performance of the forward link (mobile unit is selecting to receive signal from base station that is not in the shadow) or of the reverse link (signals from two base stations may be combined). The selective combining techniques is the most usual approach for macroscopic diversity (7).

Small-scale fading caused by multipath propagation in the vicinity of the receiver results in deep and rapid signal fluctuations. They occur over a very short distances of few wavelengths, and microscopic diversity techniques are employed to reduce the deep fades of the received signal. Different methods may be used to derive the diversity branches (channels).

Space diversity methods use physically separated antennas that can provide signals with low correlations. Typically, a separation of few wavelengths is sufficient to obtain independently fading channels. Antenna separation is also dependant on antenna height, and can be reduced as the signal frequency increases. This type of diversity is widely used in microwave radio links and in cellular radio, particularly for base station reception. For base stations in mobile radio, recommended ratio between antenna height and antenna separation is about 11, suggesting about 9 feet of antenna separation for a 100 feet antenna in 900 MHz band (7). Antenna diversity at the mobile unit can be achieved with the antenna separation as low as 17.5% the signal wavelength. Practical antenna separation for microwave links is on the order 150–200 signal wavelengths. Receiver configuration for space diversity is relatively simple. Number of branches is selectable and no extra bandwidth or power is needed to achieve diversity, making the space diversity one of the most popular schemes (4).

Frequency diversity relies on the diversity channels being separated in frequency by more than the coherence bandwidth of the channel, so the signal replicas will have very little correlation. Typical values used for mobile radio in 900 MHz band are larger than 50 KHz in urban areas, and larger than 300 KHz for suburban areas (7). This technique is often applied in microwave links, where diversity frequencies are available as the backup frequencies in the case of deep fading. Beside the requirements for additional bandwidth, disadvantage of frequency diversity is the need for multiple receivers at different frequencies.

Polarization diversity is based on the fact that horizontally and vertically polarization signal components fade independently and may be combined (selected) at the receiver. This technique is primarily used for microwave links, where channel conditions vary slowly in time, although recently it has been increasingly considered for base station reception in mobile radio. Two differently polarized antennas may be at the same place, hence there is no separation requirement as in space diversity. The drawback of the technique is the 3 dB power reduction at the transmitter since the power must be split between differently polarized antennas. Also only two diversity branches are available in polarization diversity (7).

Angle diversity may be used at high operating frequencies (over 10 GHz) by pointing directional antennas in different directions at the receiver site. It is predominantly used by mobile units in cellular radio, since most of the multipath is created by local scatterers. Directive antennas also help in reducing the Doppler spread for each diversity branch.

Time diversity relies on transmitting the same signal at different times, so these signals fade independently. instants so the fading characteristics will be changed and signals will be uncorrelated. The two transmissions should be separated by more than the coherence time of the channel. This diversity method is effective in applications where fading mechanism is not related to the movement of the receiver. However, this method fails in the scenario where mobile unit may stay still at the given location which is the subject of the deep fade, and the repeated signal components are highly correlated. While this method requires more spectrum and large buffer memory to store different signal replicas, the advantage is in simple implementation.

Interleaving is the techniques used to obtain time diversity effect in modern digital communication systems. The function of the interleaver is to spread transmitted bits in time so that a block of bits is not subject to a deep fade at the same time. This helps to randomize burst errors and aid channel coding techniques in reducing the overall BER (3).

Path diversity technique relies on receiver capability to resolve different multipath components in time domain, and subsequently combine them. This type of diversity is also referred to as the implicit diversity as opposed to other techniques which explicitly define diversity channels, since multipath diversity channels are obtained after signal reception. Adaptive equalizers and RAKE receiver (15) are receiver techniques that provide path diversity, and are very effective in frequency-selective channels. Diversity gain largely depends on the delay profile of the channel.

Diversity techniques are usually applied at the receiver side, although some of the implementation complexity can be moved to a transmitter. For example, base station in the mobile radio scenario may use the antenna that can provide better path propagation to a mobile at the specific location. Transmitter diversity allows the receiver to obtain diversity gain while operating as a standard nondiversity receiver (4, 31).

Once the diversity channels are created it remains to be decided how are they going to be combined in order to utilize diversity effect. While macrodiversity schemes use almost exclusively selection combining, microdiversity schemes may employ one of the combining techniques outlined below. The methods of linear combining are derived based on different optimization criteria, different level of information available at the receiver and different complexity constraints.

Selective combining simply selects the diversity branch with the highest signal level. This type of combining is easy to implement, additional requirements are antenna switch and monitoring circuitry. While in selective combining only one branch is used for subsequent signal processing and others are discarded, maximal ratio combining performs coherent combining of all diversity channels. This is optimal method in a sense that provides the maximal signal to noise ratio for signal detection. The weights applied to different diversity branches are proportional to the signal to noise ratios in each of the branches, and can be obtained either by utilizing pilot signals or sophisticated channel estimation techniques. When variable weighting is not available at the receiver, equal-gain combining may be utilized by coherently summing the signals from different diversity branches without weighting. This method still allows the receiver to exploit all diversity channels. Performance of the equal-gain combining is better than the selection combining and worse than maximal ratio combining, thus representing viable technique as a trade-off between performance and complexity. Performance improvement obtained by different combining schemes is summarized in Table 7.

Diversity reception may be interpreted as the simple case of coding i.e. repetitive coding since the information is simply repeated on each diversity channel. Better spectrum utilization can be achieved by employing more sophisticated coding techniques for fading channels, in particular concatenated coding (15).

All of the diversity reception techniques discussed so far rely on the processing on the reception. Alternative approach might be to move complexity of signal processing to the transmitter side, typically base station in the cellular radio system, and preserve simple design of the handset unit while still obtaining diversity gain. These techniques have been commonly termed as transmit diversity and have been utilized to improve performance in WCDMA systems (31).

Interference Suppression Techniques

Every radio system operates in the presence of some form of interference. Regardless of the origin of interference (intentional or unintentional) some form of interference suppression must be employed in order to preserve system performance. Due to the increased number of services, most of the radio systems in operation are interference-limited rather than noise-limited. For power budget calculation, in addition to radio sensitivity which is a function of the noise bandwidth of the receiver and the required SNR, interference margin should be specified. Alternatively signal to interference ratio should be specified to satisfy required BER. Depending on the type of interference (CCI, ACI) different requirements may be imposed on the receiver.

In order to improve the efficiency of a radio system different approaches may be pursued: reducing the channel spacing (increasing ACI), narrowband and wideband system overlay (creating narrowband cochannel interference to the wideband system), smaller cell design in mobile radio (increasing CCI), overlapping footprints of the satellites, etc.

Based on the signal observation methods, interference suppression techniques could be broadly divided into multichannel techniques where multiple sensors (antenna array) are employed for signal reception and single-channel techniques where only one antenna is employed (25). In a multichannel scenario, spatial filtering techniques may realized via adaptive antenna arrays, relying on the different spatial distribution of the interfering signals. Single-channel techniques rely purely on temporal processing, usually using adaptive filtering. The most promising path for the modern receiver design is to incorporate spatio-temporal processing, efficiently combining capabilities of antenna arrays and temporal signal processing (26).

Spatial processing for interference signal suppression can be carried out using different techniques, including directional antennas, tilted beams and height adjustment. However, adaptive antenna arrays with capability of adaptive beamforming have shown great potential for congested radio environments. Different algorithms and optimization criteria for adaptive beamforming are presented in (27).

Single-channel interference suppression techniques can be applied both in cases of spread spectrum signalling and conventional techniques. Although all forms of spread spectrum signals (direct sequence, frequency hopping, time hopping) provide certain interference margin due to their inherent processing gain, additional techniques can be applied to further improve performance. Depending on the nature of interfering signal, different interference suppression algorithms have been developed. For narrowband interference rejection in direct sequence systems solutions include adaptive notch filters, decision feedback techniques, adaptive analog-to-digital conversion and other nonlinear techniques. In the case of CDMA signals, where interference has the same characteristics as the desired signal, multi-user detection techniques are developed either in decentralized form (detection of single user of interest) or in centralized form (joint detection of all active users) (22). Special techniques have been analyzed for frequency hoping systems as well (25).

In nonspread spectrum systems different techniques have been developed to cope with CCI, ACI and ISI. Adaptive equalization techniques were shown to be efficient in combating all three types of interference. Other methods include self-optimizing or blind receivers (based on different techniques like constant modulus, higher-order statistics, probabilistic approaches etc.), neural network receivers and other nonlinear techniques (25).

Most of the abovementioned interference suppression algorithms are performed in baseband. In addition to this different practical interference suppression schemes are developed at RF/IF stages of the receivers (28).

Signal Losses in Radio Receiver

Due to the inherent complexity of radio receiver number of signal loss sources are associated with the receiver design. Bandlimiting loss in transmitter and receiver is the result of finite bandwidth usage, usually restricted by the radio regulations. Intersymbol interference in which received digital symbols overlap with each other may result from different effects: bandlimited operation, physical multipath radio propagation and usage of partial response modulation formats. Local oscillator phase noise results in the phase jitter in the receiver and consequently degrades the performance of the detection algorithm. Other impairments of radio front-end also result in loss of receiver performance – for example DC offset, gain and phase imbalance of in-phase and quadrature branches. Nonlinear distortions in radio equipment include AM/AM and AM/PM conversion, limiter loss and intermodulation distortion. AM/PM conversion produces phase variations in the signal, mostly after nonlinear amplification. Signal sidelobes growth is another effects that can produce excessive interference despite tight filtering at the transmitter. Hard limiting stages in receiver may cause suppression of weak signal components in the presence of stronger ones. Intermodulation products are results on multiple signals interaction in a nonlinear device and creates additional noise source which contributes to total noise level. Also receiver operation can be largely affected by several imperfections associated with the demodulation part such as imperfections of the synchronization circuits that are producing noisy estimates of received signal parameters necessary for the detection process and finite numeric precision effects associated with programmable of fixed logic implementation of the demodulation functions.

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