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Cellular Concept-I

Introduction

i) Radio telephone system needs to
→Achieve **high user capacity** and **cover large areas**
→but with **limited radio spectrum**

ii) Using a **single high-powered** transmitter mounted on a tall tower
→Acheives **good** and **large coverage**
→However **impossible** to **reuse** the same **frequency**

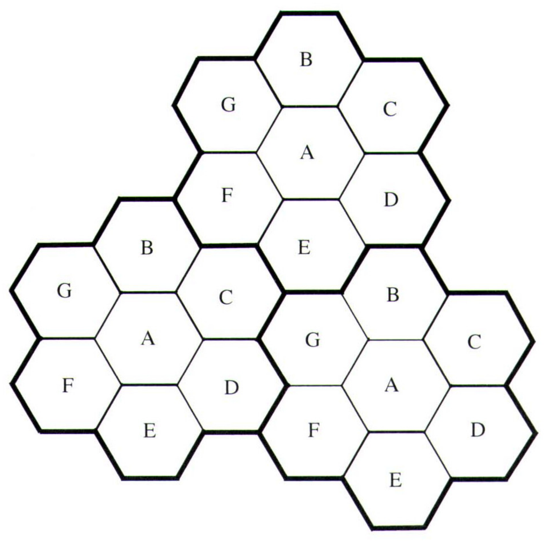
Frequency Reuse

i) Each BS is **allocated a group of radio channels** to cover a small area called ***cell***

ii) BS in adjacent cells are assigned channel groups which are **completely different**

iii) **Same channel group** is used to **cover different cells** that are separated by large distance to **keep interference** within **tolerable limits**

iv) The design process of allocating channel gruops to BSs is called **frequency reuse**



Channel Assignment Strategies

i) Objective of frequency reuse schemes are
  a) Increase the capacity of the system
  b) Minimize interference between co-channel

ii) Assignment Strategies can be classified as
  a) **Fixed** channel assignment
  b) **Dynamic** channel assignment

Handoff Strategies

i) MSC **automatically transfers** the call to a channel belonging to a new base station when a **mobile moves** into a different cell while in a conversation

ii) **Tasks** involved in a Handoff are
→ **Identifying** a new base station and
→ **allocate** voice and control signals channels of the new base station
→ **Handoff** requests are **prioritized** over call-initiation while allocating channels

iii) Handoffs must be **performed** as **infrequently** as possible and be **imperceptible** to the users

Handoff Scenario at cell boundary:

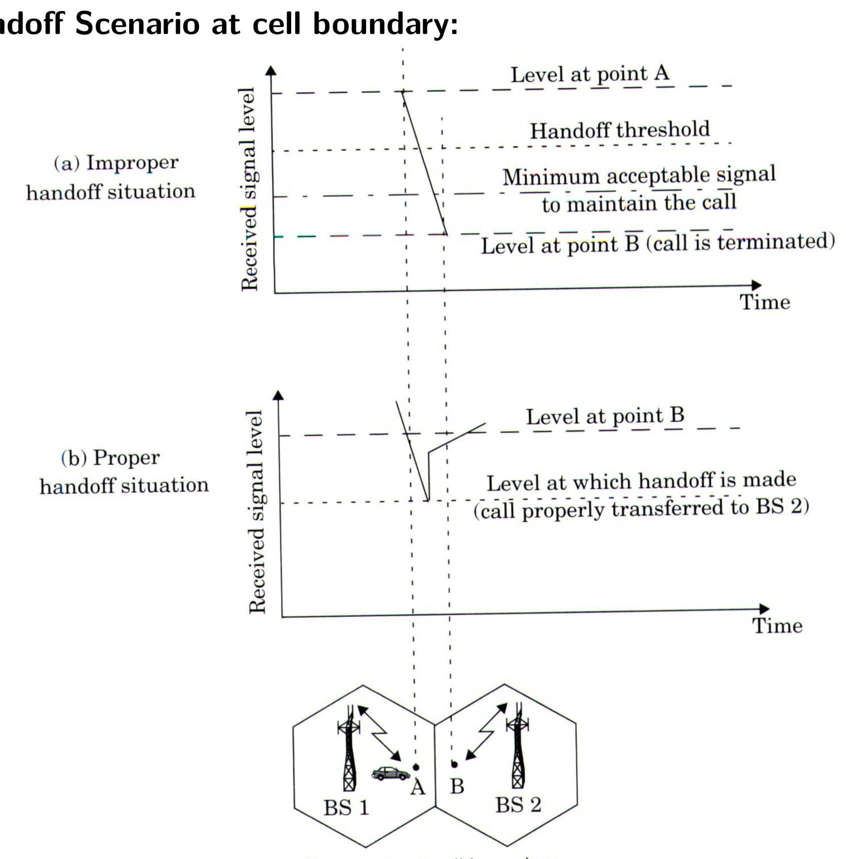


Figure 3.3 Illustration of a handoff scenario at cell boundary.

i) A **minimum usable signal** level ($P_r^{\text{minimum usable}}$) for **acceptable** voice quality is specified
→ a **slightly stronger signal** level (P_r^{handoff}) is used as a **threshold** at which a **handoff** is made.

ii) The **margin** Δ is given by $\Delta = P_r^{\text{handoff}} - P_r^{\text{minimum usable}}$, where
→ a **large** Δ with lead to **unnecessary handoffs** and burden the MSC
→ a **small** Δ will result in **insufficient time** to complete a handoff before a call is lost due to weak signal

Cellular Concept (Multiple low-power transmitters)

i) Is a system-level idea which **replaces** a **single high-power** transmitter (large cell) with **many low-power** transmitters (small cells)

ii) Each base station is **systematically spaced** and
→ provides **coverage** to **only a small portion** of the area
→ BS is **allocated a portion** of the total number of channels
→ **neighboring base stations** are allocated **different group** of channels
→ **channel groups** are **reused** while **co-channel interference** kept within **acceptable levels**

Cell Shape (Hexagon)

i) The radio coverage of a cell called footprint is **amorphous** in nature, is determined from field measurements or propagation models

ii) The **hexagonal cell shape** is a simple model of radio coverage and is universally adopted since it permits easy and manageable analysis
→ Hexagon shapes cover an **entire region** with **fewest number of cells** and without overlap
→ Base station transmitters are depicted in the center of the cell (**center-excited**) or on three of the six cell vertices (**edge-excited**)

Cell Cluster

i) Consider a celluar system with a total of S **duplex channels**,
→ these channels are divided among N cells into **unique and disjoint groups**,
→ each cell is allocated a **group of** k **channels**, such that $S = k \times N$

ii) The N cells which collectively use the **complete available** frequency is called **a cluster**

iii) If a cluster is **replicated** M times, the total number of duplex channels is
$$C = M \times k \times N = M \times S \quad \blacksquare$$
→ Note, **capacity** of the cellular system $C \propto M$

iv) The factor N is called ***cluster size*** and is typically $4, 7$ or 12
→ if N is **reduced** keeping the cell size constant, a **larger** M is required to cover an area, there by **increases** the capacity C
→ However, a **small cluster size** indicates that co-channel cells are located much closer resulting in **increase in interference**
$\implies$ N is a function of **tolerable interference** in order to **increase capacity**

iv) The ***frequency reuse factor*** is given by $\frac{1}{N}$,
since each cell in a cluster is only assigned $\frac{1}{N}$ of the total channels

Fixed channel assignment

i) Each cell is **allocated** a **predetermined** set of voice channels

ii) Any call attempt in the cell is served by the **unused** channels

iii) If all channels are occupied, the call is **blocked**

iv) In one variation of the Fixed strategy named **borrowing strategy**,
→ cell is allowed to **borrow** channels from **neighboring** cell
→ the **MSC supervises** the borrowing procedures and ensures that **no disruption** or **interference** occurs in the donor cell

Dynamic channel assignment

i) Voice channels are **not allocated** to a cell **permanently**

ii) For **each call attempt** in the cell, the MSC allocates a voice channel

iii) The MSC follows an **algorithm** that takes into account likelihood of future blocking, frequency of channel, the reuse distance and other cost functions

iv) This strategy **reduces** the likelihood of **blocking** and **increases** the **trunking capacity** as all the available channels are accessible to all the cells
→ the MSC is **required** to **collect real-time** data on channel occupancy, traffic distribution and RSSI (radio signal strength indications)
→ This **increases** the **storage** and computational **load** on the system

Dwell Time

i) **Excessive delay** by the MSC in assigning a new channel leads to **Call drop**

ii) It is important to **ensure** that the mobile is actually **moving away** from the serving base station
→ not that the drop in measured signal level is due to **momentary fading**

iii) **Before initiating** handoff, the base station **monitors** the signal level for a **certain period** of time
→ The **length of monitoring** depends largely on the **speed of mobile** units

iv) The **time** over which a call may be **maintained** within a cell **without handoff** is called **Dwell time**
→ it is governed by **factors** like propagation, interference, distance and time varying effects
→ **even** a **stationary** mobile may have a **random** and **finite** dwell time
→ The **statistics** of dwell time are important in **design** of handoff algorithms

Prioritizing Handoffs

i) If handoff requests in handled the **same way** as originating calls, the prob. that handoff is not served is equal to blocking prob. of incomming call

ii) To **improve** quality of service various methods are devised

With Increase in demand for service

i) When **more channels** are **needed**,
→ the **number** of base stations **is increased**
→ with a corresponding **decrease** in **transmitter power**
→ the **channels are reused** throughout the coverage region

ii) Every subscriber equipment is manufactured **with the same** set of channels

Co-channel Cells

i) There are only **certain** cluster sizes and cell layouts which are possible to connect the hexagon shaped cells **without gaps** between adjacent cells

ii) The geometry is such that N **can only have values** satisfying
$$N = i^2 + ij + j^2 \quad \blacksquare$$
where i, j are non-negative integers

iii) To find the **nearest** co-channel neighbors of a cell,
→ move i cells along any chain of hexagons and then
→ trun 60° counter-clockwise and move j cells

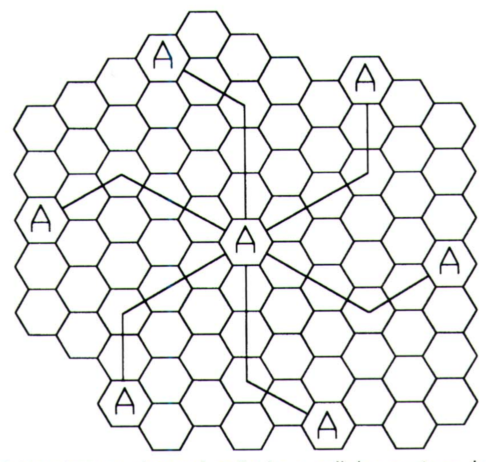


Figure 3.2 Method of locating co-channel cells in a cellular system. In this example, $N = 19$ (i.e., $i = 3$, $j = 2$). (Adapted from [Oet83] © IEEE.)

Example: Numerical Problem

Q.(a) Given total BW $= 30$ MHz and 2×25 kHz/duplex channel

Sol: $\implies$ Available channels $= \dfrac{30\text{ MHz}}{50\text{ kHz}} = 600$

For N	Channels per cell
4	$(600/4) = 150$
7	$(600/7) = 85$
12	$(600/12) = 50$

Q.(b) Given 1 MHz is dedicated for control channels

Sol: $\implies$ Dedicated control channels (CC) $= \dfrac{1\text{ MHz}}{50\text{ kHz}} = 20$ and
$\implies$ $(600 - 20) = 580$ voice channels (VC)

For N	Channels per cell
4	$(5$ CC $+ 145$ VC$) \times 4$ cells
7	$(3$ CC $+ 82$ VC$) \times 4$ cells $+$ $(2$ CC $+ 83$ VC$) \times 3$ cells
12	$(2$ CC $+ 48$ VC$) \times 8$ cells $+$ $(1$ CC $+ 49$ VC$) \times 4$ cells

1^{st} Generation Analog cellular systems (MSC initiated Hand-off)

i) Signal strength (RSSI) **measuremens** are made by the BS and supervised by MSC
→ **reverse voice channels** (RVCs) are monitored to determine **relative location** of each mobile user

ii) A **spare** location receiver is used to **determine** RSSI of users in **neighboring cells**

iii) Based on RSSI values received, the **MSC decides** if a handoff is necessary

2^{nd} Generation Digital cellular systems (Mobile-assisted Handoff, MAHO)

i) **Handoff decisions** are **mobile assisted**

ii) In **MAHO** (mobile-assited Handoff), every mobile **measures** the received power from **surrounding** BSs and report the results to the serving BS

iii) A handoff is **initiated** when the power received from **neighboring cell** begins to **exceed** the **serving BS** by a certain level or for a certain period of time

iv) The MAHO performs at a much **faster rate** and is particularly **suited** for **micro cellular** environment

Intersystem handoff (MSC-MSC Handoff)

i) An **intersystem handoff** becomes necessary when a **mobile moves** to a **different cellular system** controlled by a **different MSC**
→ a **local call** may become **long-distance call** and user becomes **roamer**
→ The two MSCs must be **compatible** for **implementing** intersystem handoff

Guard channel concept

i) A **fraction** of the total available channels is reserved for handoff requests

ii) It may reduce the total carried traffic, however offers efficient spectrum utilization when dynamic channel assignment strategies are used

Queuing of handoff requests

i) Handoff requests are **queued** instead of **forced** termination due to lack of available channels

ii) However, does not gaurantee a zero prob. of forced termination

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Cellular Concepts-II

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Cellular Concept-II

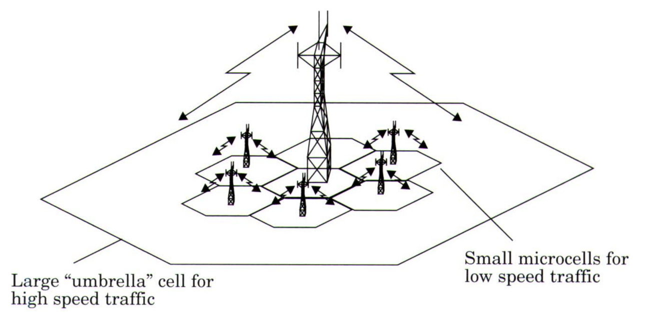
Practical Handoff Considerations

i) High **speed** vehicles pass through the cell within seconds (require large cells), where as **pedestrain** many never need a handoff (require micro cells)

→ MSC can quickly become **burdened** with addition of microcells

ii) Another practical limitation is the **ability** to obtain new cell sites

Umbrella Cell



i) By using **different** antenna heights (at the same tower) and different power levels, large and small cells are **co-located**

→ This approach is used to **provide large area** coverage for high speed users while providing **small area coverage** to the pedestrains

Cell Dragging

i) A **strong signal** is provided when there is line-of-sight radio path between a pedestrain and BS. A handoff may not be made even user travels away.

Hard Handoff (GSM)

i) Assigns **different** radio channels

ii) In 1^{st} generation analog systems, it takes about 10 seconds and the value of Δ is $6dB$ to $12dB$

iii) In 2^{nd} generation and digital systems, requires only $1-2$ seconds and the value of Δ is $0dB$ to $6dB$

Soft Handoff (CDMA)

i) Spread spectrum share the same channel in every cell.

→ **Handoff does not** mean a physical **change** in the assigned channel

→ Rather a **different base station** handles the radio communication task

ii) MSC **simultaneously** evalutes the receiver signals from a single subscriber at several neighboring base stations

MSC **decides** which version of the user's signal is best at any moment in time

Co-Channel Reuse Ratio Q

i) Co-channel interference is a function of the **radius of the cell** R, and the **distance** to the center of the nearest co-channel cell D

ii) **Co-channel reuse ratio** is defined as $Q \triangleq \frac{D}{R}$ which is equal to $\sqrt{3N}$ for hexagonal geometry

	Cluster Size (N)	CCRR (Q)	Remark
$i=1, j=1$	3	3	**High** capacity, but **high** CCI
$i=1, j=2$	7	4.58	
$i=2, j=2$	12	6	
$i=1, j=3$	13	6.24	**Low** CCI, but **low** capacity

Co-channel Interference and System Capacity

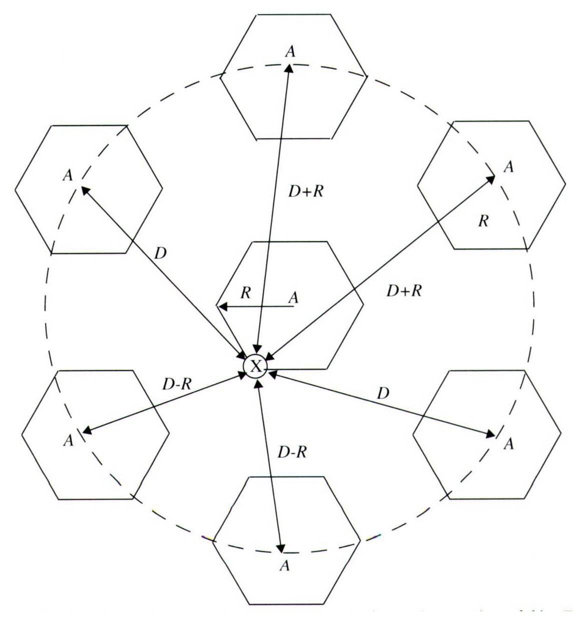
i) Due to frequency reuse, there are several cells that use the same set of freqs.

→ these cells are called **co-channel cells**, and interference between them is called **co-channel interference** (CCI)

ii) CCI can **not** be reduced by simply increasing the power at the transmitter

→ because increasing tx power increases the interference in neighboring co-channel cells

iii) Co-channel cells must be **physically separated** to reduce CCI



Signal-to-Interference ratio SIR

i) The SIR of a mobile receiver which monitors a forward channel is

$$SIR = \frac{S}{I} = \frac{S}{\sum_{i=1}^{i_0} I_i},$$

where S is the **signal power** from the **serving BS** and I_i is the **interference power** caused by the i^{th} **co-channel BS** out of the total i_0

ii) Due to the **propagation losses**, the average received signal power P_r at a distance d from the BS is

$$P_r = P_0\left(\frac{d}{d_0}\right)^{-n} \quad \text{or} \quad P_r(dBm) = P_0(dBm) - 10\,n\log\left(\frac{d}{d_0}\right)$$

P_0 is the power received at a close-in **reference point** at distance d_0 in the far field region and n is the path loss exponent

→ If R and D_i are the **distances** of the receiver from the serving BS and the i^{th} interferer respectively, we have

$$\frac{S}{I} \approx \frac{R^{-n}}{\sum_{i=1}^{i_0} D_i^{-n}}, \text{ further assuming equidistance } \frac{S}{I} = \frac{(D/R)^n}{i_0} = \frac{(\sqrt{3N})^n}{i_0} \blacksquare$$

→ **Example**: A cellular system requires $SIR = 18dB$ for sufficient voice quality,

$\implies$ the cluster size is atleast $N = 6.49$ assuming $n = 4$ and $i_0 = 6$

Example: Numerical Problem

Q. Given required $SIR > 15dB$ and equidistant co-channel cells $i_0 = 6$, compute cluster size N for a) $n = 4$ and b) $n = 3$

Sol: a) Using $n = 4$, for different values of N we have

For N	$\frac{D}{R} = \sqrt{3N}$	$SIR = \frac{(\sqrt{3N})^4}{i_0}$
4	$\sqrt{3\times4} = 3.46$	$\frac{3.46^4}{6} = 24.00 = 13.8\ dB$
7	$\sqrt{3\times7} = 4.58$	$\frac{4.58^4}{6} = 73.50 = 18.66\ dB$

$\implies$ cluster size of $N = 7$ is required.

b) Using $n = 3$, for different values of N we have

For N	$\frac{D}{R} = \sqrt{3N}$	$SIR = \frac{(\sqrt{3N})^3}{i_0}$
4	$\sqrt{3\times4} = 3.46$	$\frac{3.46^3}{6} = 6.93 = 8.41\ dB$
7	$\sqrt{3\times7} = 4.58$	$\frac{4.58^3}{6} = 16.03 = 12.10\ dB$
12	$\sqrt{3\times12} = 6$	$\frac{6^3}{6} = 36 = 15.56\ dB$

$\implies$ cluster size of $N = 12$ is required.

Interference and System Capacity

i) Interference is the major **limiting factor** in the performance of cellular system

→ major bottleneck in increasing capacity

ii) **Sources** of **interference** include

→ another mobile in the same cell,

→ a call in progess in a neighboring cell,

→ other base stations operating in the same freq. band

→ any noncellular system which leaks energy into the cellular freq. band

→ more severe in urban areas, due to the greater RF noise floor & large number of base stations and mobiles

iii) **Effects** of interference

→ on **voice channels** causes cross talk, subscriber hears interference in the background

→ Interference on **control channels** leads to missed and blocked calls due to errors in the digital signalling

v) **Types** of interferences

→ System-generated cellular interference include **co-channel** interference and **adjacent channel** interference

→ **Out-of-band** interference is caused mainly from the transmitters of competing cellular carriers (located in close proximity)

Channel Planning for Wireless Systems

i) Radio spectrum is **divided** into channels as per *air interface standard*

ii) **Two types** of channels a) **Control** channels (CC) and b) **Voice** channels

→ *Control channels*: used for initiating, requesting or paging a call

About 5% of the spectrum is devoted to control channels

Carry very brief and bursty messages

→ *Voice channels*: dedicated to carry revenue-generating traffic

About 95% of the spectrum is devoted to control channels

iii) Since control channels are **vital**, the freq. reuse strategy is more **conservative** than voice channels. Ex: CC are allocated 21-cell reuse, whereas VC are allocated 7-cell reuse

iv) In practice, propagation considerations require most CDMA systems to use some sort of limited frequency reuse e.g., $f1/f2$ cell planning.

In such cases, CDMA also makes hard handoffs

Adjacent Channel Interference

i) Interference resulting from signals which are **adjacent in freq.** to the desired signal is called **Adjacent Channel Interference** (ACI)

Table 3.2 AMPS Channel Allocation for A and B Side Carriers

1A	2A	3A	4A	5A	6A	7A	1B	2B	3B	4B	5B	6B	7B	1C	2C	3C	4C	5C	6C	7C	
1	2	3	4	5	6	7	8	9	10	11	12	13	14	15	16	17	18	19	20	21	
22	23	24	25	26	27	28	29	30	31	32	33	34	35	36	37	38	39	40	41	42	
43	44	45	46	47	48	49	50	51	52	53	54	55	56	57	58	59	60	61	62	63	
64	65	66	67	68	69	70	71	72	73	74	75	76	77	78	79	80	81	82	83	84	
85	86	87	88	89	90	91	92	93	94	95	96	97	98	99	100	101	102	103	104	105	
106	107	108	109	110	111	112	113	114	115	116	117	118	119	120	121	122	123	124	125	126	
127	128	129	130	131	132	133	134	135	136	137	138	139	140	141	142	143	144	145	146	147	
148	149	150	151	152	153	154	155	156	157	158	159	160	161	162	163	164	165	166	167	168	
169	170	171	172	173	174	175	176	177	178	179	180	181	182	183	184	185	186	187	188	189	
190	191	192	193	194	195	196	197	198	199	20	201	202	203	204	205	206	207	208	209	210	A
211	212	213	214	215	216	217	218	219	220	221	222	223	224	225	226	227	228	229	230	231	SIDE
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253	254	255	256	257	258	259	260	261	262	263	264	265	266	267	268	269	270	271	272	273	
274	275	276	277	278	279	280	281	282	283	284	285	286	287	288	289	290	291	292	293	294	
295	296	297	298	299	300	301	302	303	304	305	306	307	308	309	310	311	312	-	-	-	
313	314	315	316	317	318	319	320	321	322	323	324	325	326	327	328	329	330	331	332	333	
-	-	-	-	-	-	-	-	-	-	-	-	-	-	-	-	-	-	667	668	669	
670	671	672	673	674	675	676	677	678	679	680	681	682	683	684	685	686	687	688	689	690	
691	692	693	694	695	696	697	698	699	700	701	702	703	704	705	706	707	708	709	710	711	
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1003	1004	1005	1006	1007	1008	1009	1010	1011	1012	1013	1014	1015	1016	1017	1018	1019	1020	1021	1022	1023	
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565	566	567	568	569	570	571	572	573	574	575	576	577	578	579	580	581	582	583	584	585	
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628	629	630	631	632	633	634	635	636	637	638	639	640	641	642	643	644	645	646	647	648	
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-	-	-	-	-	717	718	719	720	721	722	723	724	725	726	727	728	729	730	731	732	
733	734	735	736	737	738	739	740	741	742	743	744	745	746	747	748	749	750	751	752	753	
754	755	756	757	758	759	760	761	762	763	764	765	766	767	768	769	770	771	772	773	774	
775	776	777	778	779	780	781	782	783	784	785	786	787	788	789	790	791	792	793	794	795	
796	797	798	799																		

Causes of adjacent channel interference

i) ACI results from the **imperfect receiver filters** which allow nearby freqs. to leak into the passband

ii) **Near-far** effect:

→ This occurs when an adjacent channel user is transmitting in very close range to a **subscriber**'s receiver, while the receiver attempts to receive a BS on the desired channel

→ Alternatively, the near-far effect occurs on the **base station** when a close-by mobile transmits on an adjacent channel, while the BS attempts to receive from a far mobile

Minimizing adjacent channel interference

i) ACI can be minimized through careful filtering and channel assignments

→ By keeping the **freq. separation** between each channel in a given cell as large as possible, the ACI may be reduced

→ If the **cluster size** N is **small**, the separation may not be sufficient to keep the ACI low. **Ex**: If a near mobile is 20 times closer to a BS than a far mobile, then the signal-to-interference ratio at the BS is $\frac{S}{I} = 20^{-n}$, which is $-52\ dB$ for $n = 4$

ii) Channels are divided into **subsets** and **allocated to different cells** so that ACI is minimized

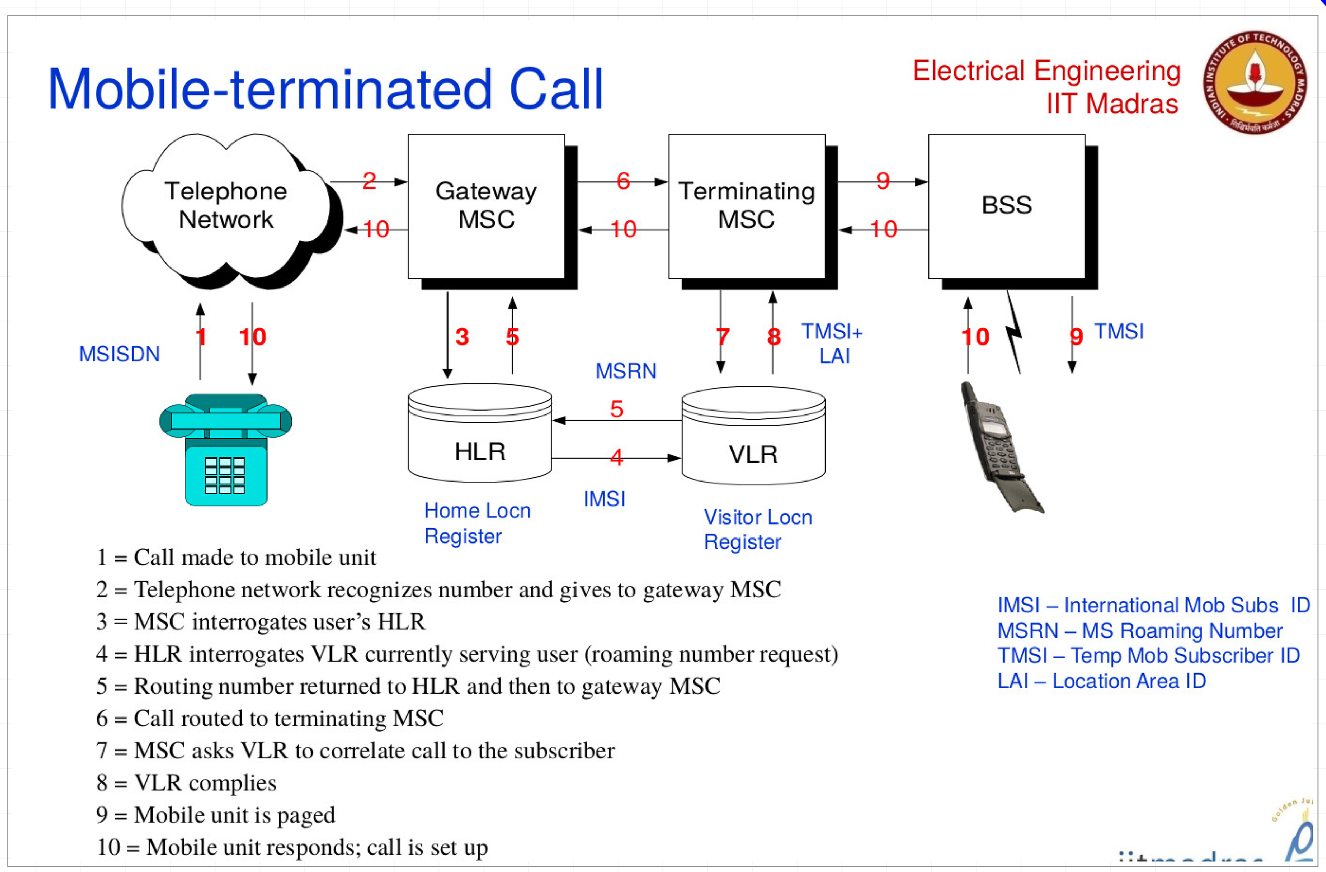
→ **Ex**: In a 7-cell reuse system, each cell uses channels in the subsets $iA + iB + iC$, where $1 \le i \le 7$

Power Control for Reducing Interference

i) The power levels transmitted by every mobile unit are under constant control by the serving base stations to ensure that each mobile transmits the smallest power on the reverse channel

ii) It helps prolong battery life and also reduces the interference

iii) It is especially important for CDMA systems, every user in the cell share the same radio channel



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Cellular Concept-III

Trunking

i) Cellular radio systems rely on **trunking** to accomodate a **large number** of users in a limited radio spectrum

ii) Each user is allocated a channel on a **per call basis**, and upon termination of the call, the occupied channel is **returned to the pool**

iii) Trunking exploits the **statistical behavior** of users so that a limited number of channels or circuits may accommodate a large, random user community

iv) **Trunking theory** principle is used in designing cellular radio system

Table 3.4 Capacity of an Erlang B System

Number of Channels C	Capacity (Erlangs) for GOS = 0.01	= 0.005	= 0.002	= 0.001
2	0.153	0.105	0.065	0.046
4	0.869	0.701	0.535	0.439
5	1.36	1.13	0.900	0.762
10	4.46	3.96	3.43	3.09
20	12.0	11.1	10.1	9.41
24	15.3	14.2	13.0	12.2
40	29.0	27.3	25.7	24.5
70	56.1	53.7	51.0	49.2
100	84.1	80.9	77.4	75.2

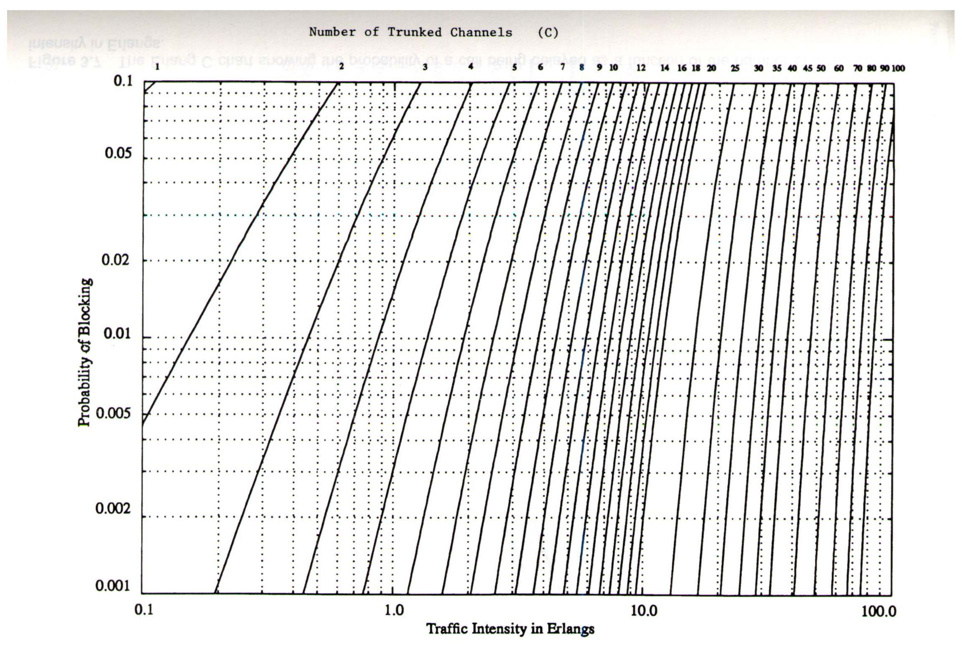


Figure 3.6 The Erlang B chart showing the probability of blocking as functions of the number of channels and traffic intensity in Erlangs.

Improving Coverage and Capacity

i) As the demand increases techniques such as a) **Cell splitting** b) **sectoring** and c) **coverage zone approaches**, are used in practice to expand the capacity of a system

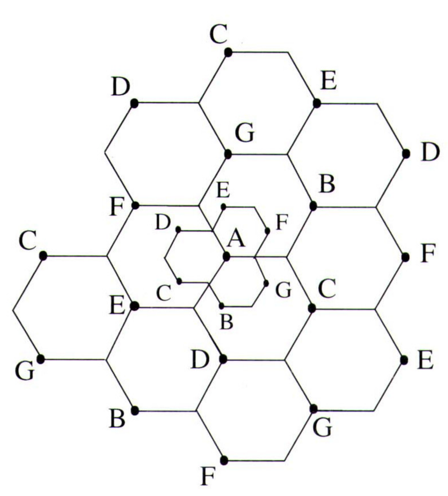
a) Cell Splitting:

i) Cell splitting is the process of **subdividing** a congested cell into smaller cells each with its base station and a corresponding reduction in transmitted power

→ Cell spiltting increase the capacity since it increases the channel reuse

ii) The radius of every cell is cut into **half** i.e., $\frac{R}{2}$, implying **four** times as many cells would be required

iii) Ex: The microcell BS labeled G is placed half way between A and C utilizing the same channel set G



iv) The transmit power of the new cell P_{t2} is found by examining the received power P_r at the new and old cell boundary and setting them equal to each other, i.e.,

P_r[at old cell boundary] $\propto P_{t1}R^{-n}$

P_r[at new cell boundary] $\propto P_{t2}(\frac{R}{2})^{-n}$, where P_{t1}, P_{t2} are transmit power of old and new cells. And using $n = 4$, we have

$$P_{t2} = \frac{P_{t1}}{16}$$

→ The transmit power is reduced by $12\ dB$ in order to fill the original coverage area with microcells while maintaining the SIR requirement.

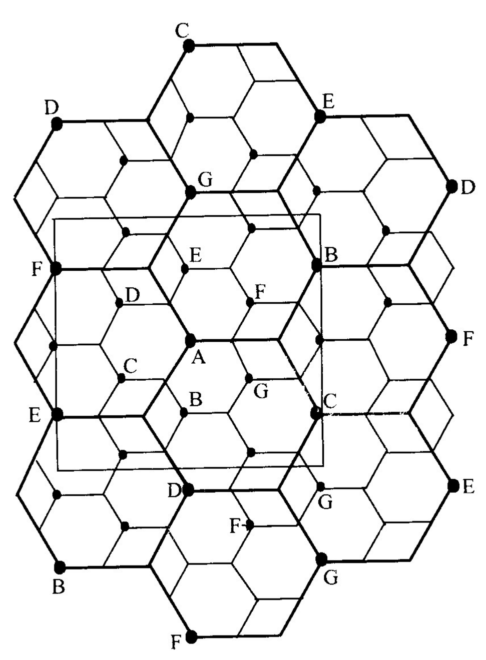


Figure 2.9
Illustration of cell splitting within a 3 km by 3 km square centered around base station A.

Erlang

i) **Erlang** is the **fundamental unit** used to measure the **traffic intensity**

ii) One Erlang represents the **amount of traffic intensity** carried by a channel that is **completely occupied** (i.e., one call-hour per hour or one call-minute per minute). **Ex**: A channel that is occupied for 30 minutes during an hour carries 0.5 Erlangs of traffic

Grade of Service (GOS)

i) A **measure** of the ability of a **user to access** a trunked system during the **busiest hour**

ii) It is typically given as the **likelihood** that a call is **blocked** (for Erlang B), or the **likelihood** of a call **experiencing a delay** (for Erlang C) greater than a certain queuing time

iii) GOS is a **benchmark** used to define the desired performance of a trunked system

→ Designer estimates the maximum required capacity in order to meet the GOS

→ The AMPS cellular system is designed for a GOS of 2% blocking, i.e., cell sites are designed so that 2 out of 100 calls will be blocked due to channel occupancy during the busiest hour

iv) **Trunking efficiency**: It is a measure of the number of users which can be offered a particular GOS for a given configuration of fixed channels.

Common Terms

i) **Set-up Time**: The time required to allocate a trunked radio channel to a requesting user

Blocked Call: Call which cannot be completed at time of request due to congestion

ii) **Holding Time**: Average duration of a typical call. Denoted by H (in seconds) (also called lost call)

Request Rate: The average number of call requests per unit time. Denoted by λ seconds^{-1}

→ Traffic **intensity** generated by **each user** $A_U = \lambda\, H$

iii) **Traffic Intensity**: It is a measure of channel time utilization, which is the average channel occupancy measured in Erlangs. Is denoted by A

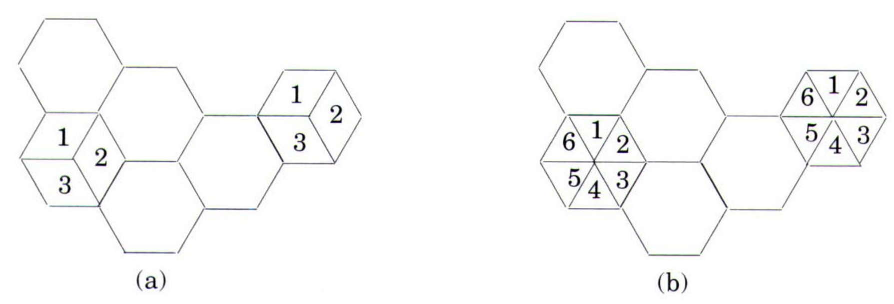
For a system containing U number of users and C channels,

→ total **offered intensity** is $A = UA_U$ and

→ traffic **intensity per channel** is $A_C = \dfrac{UA_U}{C}$

iv) **Load**: Traffic intensity across the entire truked radio system, measured in Erlangs. The maximum possible carried traffic is the total number of channels C in Erlangs

b) Sectoring



(a) 120° sectoring; (b) 60° sectoring.

i) Co-channel **interference** in a cellular system may be decreased by replacing a single **omnidirectional** antenna at the BS by several **directional** antennas, each radiating within a specified sector

ii) This technique for decreasing co-channel interference and thus increasing system performance by using directional antennas is called **sectoring**

iii) The factor by which the CCI is reduced depends on the amount of sectoring used. Ex: $120°$ sectors or $60°$ sectors

c) Coverage zone appr. (Repeaters for Range Extension)

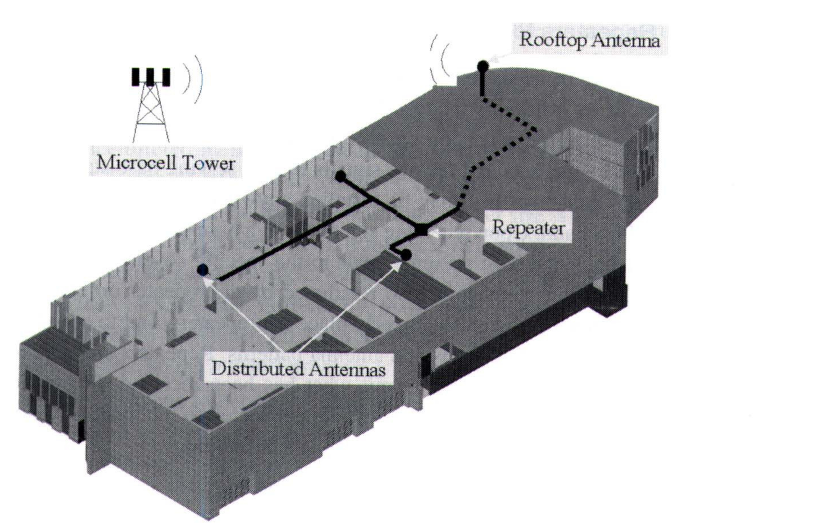


Figure 3.12 Illustration of how a distributed antenna system (DAS) may be used inside a building. Figure produced in SitePlanner®. (Courtesy of Wireless Valley Communications Inc.)

i) The repeater does not add capacity to the system, it simply serves to re-radiate the BS signal into specific locations that are hard-to-reach.

ii) They are bidirectional in nature, and simultaneously send signals to and receive signals from a serving base station

iii) In practice, directional antennas or distributed antenna systems (DAS) are connected to the inputs or outputs of repeaters for localized spot coverage particularly in tunnels or buildings

Blocked calls Cleared (Erlang B formula)

i) This type of trunking system offers **no queuing** for call requests

ii) The user is given immediate access to a channel if one is available. If **no channels** are **available**, the user is **blocked** . Further, it is assumed

→ that calls **arrive** as determined by a **Poisson** distribution

→ that there are an infinite number of users

iii) It assumes $M/M/m/m$ **queue**

a) there are **memoryless** arrivals of requests

b) the probability of a user **occupying** a channel is **exponentially** distributed

c) there are a **finite** number of channels available in the trunking pool

iv) The Erlang B formula determines the probability that a call is blocked and is given by $Pr\{\text{blocking}\} = \dfrac{\frac{A^C}{C!}}{\sum_{k=0}^{C}\frac{A^k}{k!}} = \text{GOS}$

Blocked calls Delayed (Erlang C formula)

i) This type of trunking system **provides a queue** to hold calls which are blocked

ii) If a channel is not available, the call request may be delayed until a channel becomes available

iii) Its **measure** of GOS is defined as the probability that a call is blocked after waiting a specific length of time in the queue

iv) The likelihood of a call not having immediate access to a channel is determined by the Erlang C formula, given as

$$Pr\{\text{delay} > 0\} = \frac{A^C}{A^C + C!(1-\frac{A}{C})\sum_{k=0}^{C-1}\frac{A^k}{k!}}$$

v) The GOS of a trunked system where blocked calls are delayed is given by

$$\begin{aligned} Pr\{\text{delay} > t\} &= Pr\{\text{delay} > 0\}\; Pr\{\text{delay} > t \mid \text{delay} > 0\} \\ &= Pr\{\text{delay} > 0\}\; \exp\left[\frac{-(C-A)\,t}{H}\right] \end{aligned}$$

vi) The average delay D for all calls in a queued system is given by

$$D = Pr\{\text{delay} > 0\}\frac{H}{C-A}$$

Example:

Q) Assuming each user generates 0.1 Erlang, compute the number of users supported in a blocked calls cleared system. Given GOS $= 0.5\%$ and trunked channels as a) 1, b) 5, c) 10, d) 20 and e) 100.

Sol)

C (Given)	A (from ErlangB plot)	Users $U = \frac{A}{A_U} = \frac{A}{0.1}$
1	0.005	$\frac{0.005}{0.1} = 0.05 \implies U = 1$
5	1.13	$\frac{1.13}{0.1} \approx 11$
10	3.96	$\frac{3.96}{0.1} \approx 39$
20	11.10	$\frac{11.1}{0.1} \approx 110$
100	80.9	$\frac{80.9}{0.1} \approx 809$

Illustration

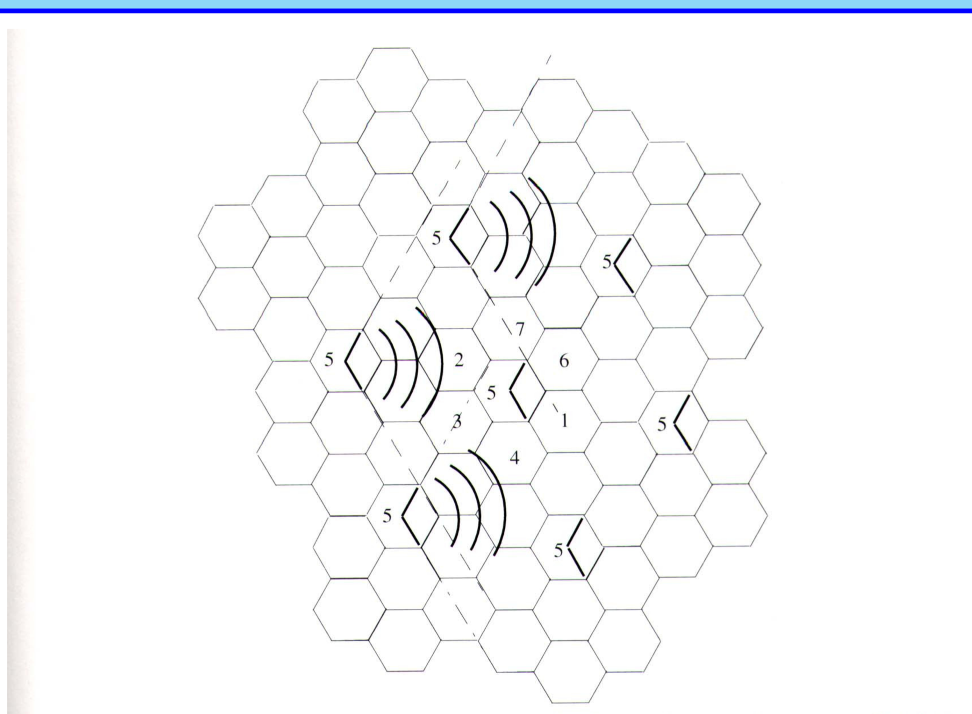


Figure 3.11 Illustration of how 120° sectoring reduces interference from co-channel cells. Out of the 6 co-channel cells in the first tier, only two of them interfere with the center cell. If omnidirectional antennas were used at each base station, all six co-channel cells would interfere with the center cell.

i) Note that a mobile in the sector 5 of the center cell will experience interference from only two sectors out of the six co-channel cells

ii) This SIR improvement allows the engineers to then decrease the cluster size N in order to improve the freq. reuse

iii) The **penalty** for improved SIR is an increased number of antennas at each BS and a decrease in trunking efficiency due to channel sectoring at the BS.

Microcell Zone Concept

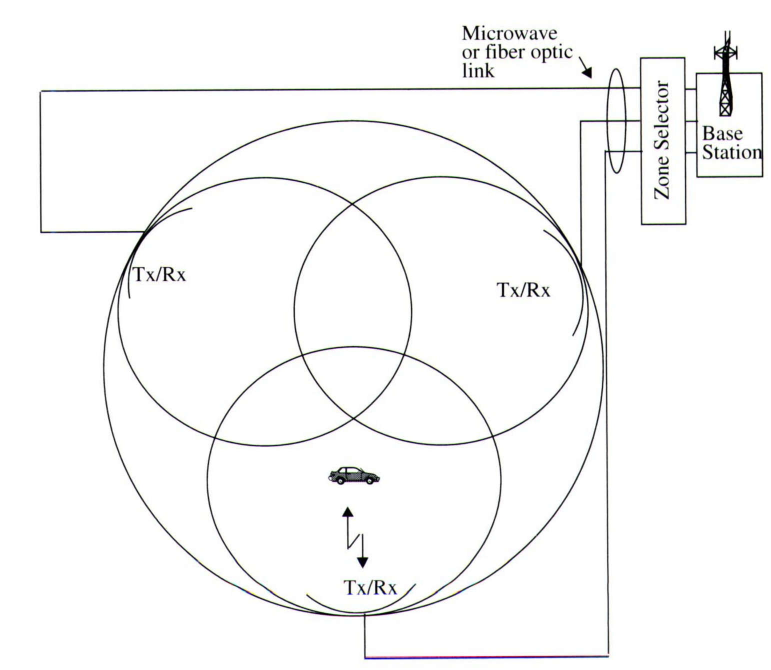


Figure 3.13 The microcell concept [adapted from [Lee91b] © IEEE].

i) In this scheme, each of the zone sites are connected to a single BS and share the same radio equipment

ii) The zones are connected by coaxial cable, fiberoptic cable or microwave link to the BS

iii) As a mobile travels within the cell, it is served by the zone with the strongest signal

iv) As a mobile travels from one zone to another within the cell, it retains the same channel. Thus, unlike in sectoring, a handoff is not required at the MSC when mobile travels between zones within a cell.

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Unit-II: Large-Scale Path Loss

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Large Scale PL-I

Introduction to Radio Wave Propagation

i) Unlike wireline channnels, the Radio channels are extremely random and not easy to analyze

→ Mobility and speed of motion impacts the received signal levels

ii) In urban regions the transmiter and receiver (T-R) are generally not in line-of-sight (NLOS)

→ EM waves propagation is generally attributed to reflection, diffraction and scattering

→ Due to different paths of varying length, the waves at a specific location combine constructively or distructively thereby resulting in fading

iii) The instantaneous path loss (PL_{dB}) at any specific location is a combination of losses due to propogation loss ($\overline{PL_{dB}}$), shadowing (X_σ) and fading.

$PL_{dB} = \overline{PL_{dB}} + X_\sigma +$ Small-scale fading

Free Space Propagation Model

i) This model is used to predict the received signal strength when T-R have clear line-of-sight (LOS) path. Ex: Microwave link, Satellite comm. etc

ii) The received signal power $P_r(d)$ at distance d from a transmitter with transmit power P_t is given by **Friis free space** equation as

$P_r(d) = \dfrac{P_t G_t G_r \lambda^2}{(4\pi)^2 d^2 L}$, where G_t, G_r are the antenna gains of the transmitter and the receiver, $L (\geq 1)$ is the miscellaneous system losses, $\lambda (= \frac{c}{f_c})$ is the carrier wavelength

→ Note that $P_r(d) \propto \frac{1}{d^2}$ for free space propagation (i.e., loss exponent $n = 2$)

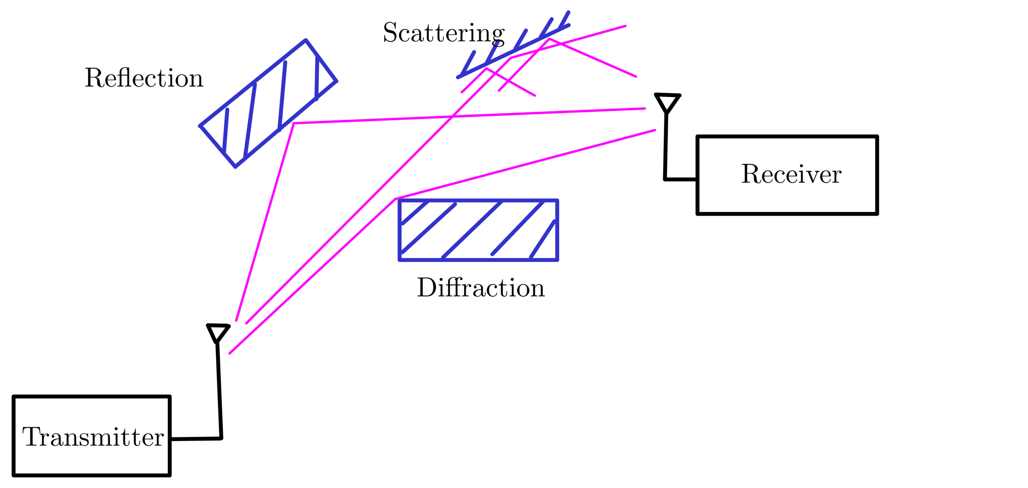
$\implies$ power decays with distance at a rate of $20dB/decade$

→ Gain of an antenna is related to its effective aperture A_e as $G = \dfrac{4\pi A_e}{\lambda^2}$

→ Effective isotropic radiated power EIRP (in dBi) is equal to $P_t G_t$

However, in practice effective radiate power (ERP in dBd) is used which is $2.15dB$ above EIRP

Three basic propagation mechanisms



I. Reflection (Propagation Mechanism)

i) Occurs when an EM wave impinges upon an object with very large surface compared to wavelength λ. Ex: Earth Surface, building and Walls

ii) When a radio wave propagating in one medium impinges upon another medium having different eletrical properties, the wave is partially reflected and partially transmitted

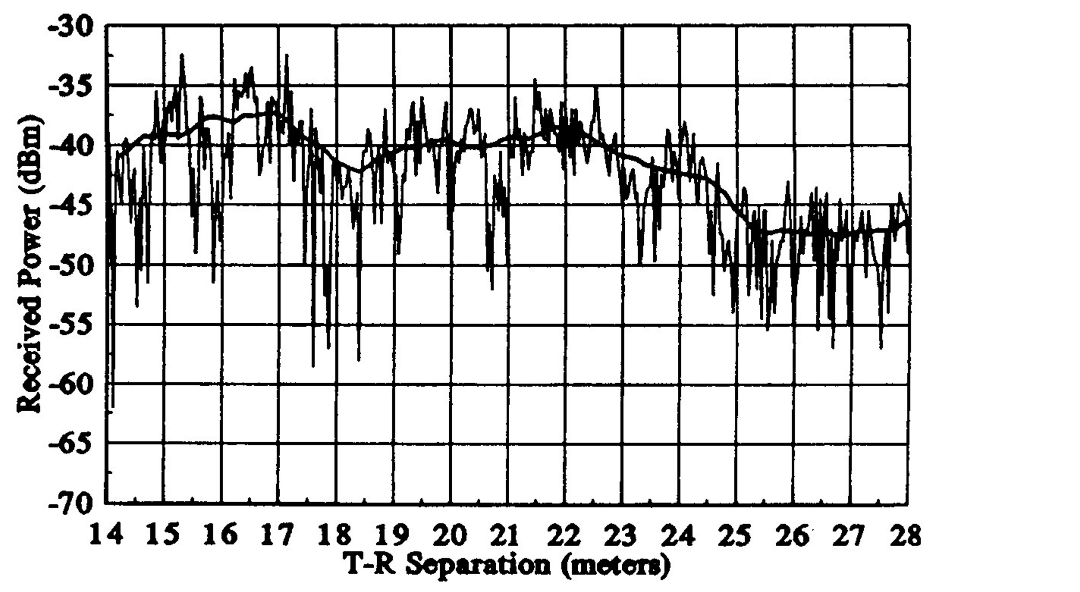
iii) The electric field intensity of the reflected and the transmitted waves are related to the incident wave through the Fresnel reflection coefficient Γ

→ The coefficient Γ is a function of the material properties, wave polarization, angle of incidence θ_i and the freq.

iv) Polarized EM waves have instantaneous electric field components in orthogonal directions in space

→ A polarized wave may be mathematically represented as the sum of two spatially orthogonal components, such as vertical and horizontal

Received power on Wireless channel



Path Loss equation

i) The difference (in dB) between the effective transmitted power and the received power is defined as the **path loss**

The path loss i.e., the signal attenuation in positive quantity is given by

$$PL(dB) = 10 \log \frac{P_t}{P_r} = -10 \log \frac{G_t G_r \lambda^2}{(4\pi)^2 d^2 L}$$
$$= -10 \log \frac{\lambda^2}{(4\pi)^2 d^2}, \text{ when } G_t = G_r = 1 \text{ and } L = 1$$

ii) Note that Friis equations are valid only when d is in far-field or Fraunhofer region, i.e., $d > d_f$, where $d_f = \dfrac{2D^2}{\lambda}$ and D is the largest dimension of the antenna. Additionally $d_f \gg D$ and $d_f \gg \lambda$.

Relating Power to Electric Field

i) In free space, the **power flux density** P_d (in W/m^2) is given by

$$P_d = \frac{EIRP}{4\pi d^2} = \frac{P_t G_t}{4\pi d^2} = \frac{E^2}{R_{fs}} = \frac{E^2}{\eta} \; W/m^2, \text{ where}$$

R_{fs} is the intrinsic impedance of the free space given by $\eta = 120\pi \, \Omega$

$\implies P_d = \dfrac{|E|^2}{120\pi} \; W/m^2$, where $|E|$ represents the magnitude of the radiating portion of the electric field in the far field.

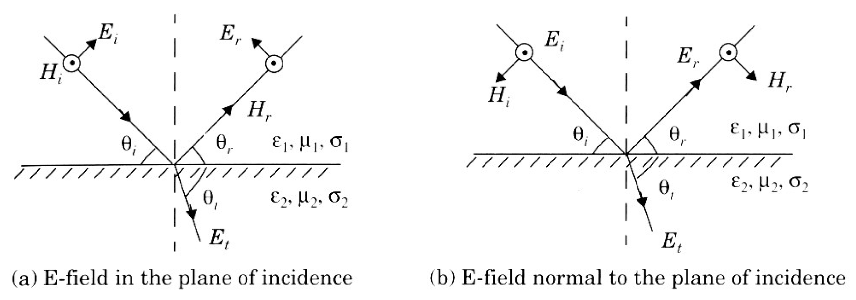
ii) The power received at distance d, $P_r(d)$, is given by

$P_r(d) = P_d A_e = \dfrac{|E|^2}{120\pi} A_e = \dfrac{P_t G_t G_r \lambda^2}{(4\pi)^2 d^2}$, where A_e is the effective aperture of the receiver antenna

Reflection from Dielectrics

i) If the plane wave is incident on a perfect dielectric, part of the energy is transmitted into the second medium and the remaining is reflected back.

→ There is no loss of energy in absorption in this case



(a) E-field in the plane of incidence (b) E-field normal to the plane of incidence

Figure 4.4 Geometry for calculating the reflection coefficients between two dielectrics.

ii) In the above figure,

→ $\theta_i, \theta_r, \theta_t$ are the incident, reflected and the transmitted angles

→ $\varepsilon_1, \mu_1, \sigma_1$ and $\varepsilon_2, \mu_2, \sigma_2$ represent the permittivity, permeability and conductance of the two media

The permittivity of lossless dielectric is $\varepsilon = \varepsilon_o \varepsilon_r$, where $\varepsilon_o = 8.85 \times 10^{-12} F/m$, ε_r is the relative value and for a lossy dielectric $\varepsilon = \varepsilon_o \varepsilon_r - j\varepsilon'$

→ the E-field is parallel polarized with the plane of incidence in fig-a and perpendicular polarized in fig-b

Reflection from Perfect Conductors

i) If the plane wave is incident on a perfect conductor, then all incident energy is reflected back into the first medium (without any loss of energy)

i.e., EM wave cannot pass through perfect conductors

ii) In order to obey Maxwell's equations, the reflected wave must be equal in magnitude to the incident wave and $\theta_i = \theta_r$. Regardless the value of θ_i,

→ for parallel polarization: $E_i = E_r$

→ for perpendicular polarization: $E_i = -E_r$

Ground Reflection (Two-Ray) Model

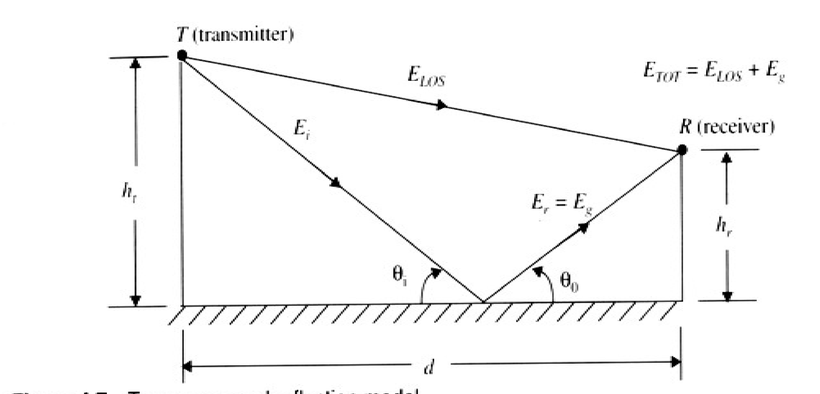


Figure 4.7 Two-ray ground reflection model.

i) This model is based on geometric optics, and considers both the direct path and a ground reflected propagation path between T-R

This is found usefull for several KMs distance

ii) The total received E-field E_{TOT} is a result of the direct line-of-sight component E_{LOS} and the ground reflected component E_g

iii) Considering E_0 as the free space E-field at a reference distance d_0 from the transmitter, the free space propagation E-field is given by

$E(d,t) = \dfrac{E_0 d_0}{d} \cos\left[\omega_c\left(t - \dfrac{d}{c}\right)\right]$ for $d > d_0$, where $\dfrac{E_0 d_0}{d}$ is the envelope

→ Similarly, the E-field of LOS and ground-reflected components are given by

$$E_{LOS}(d', t) = \frac{E_0 d_0}{d'} \cos\left[\omega_c\left(t - \frac{d'}{c}\right)\right] \text{ and}$$
$$E_g(d'', t) = \Gamma \frac{E_0 d_0}{d''} \cos\left[\omega_c\left(t - \frac{d''}{c}\right)\right] \text{ where } d', d'' \text{ are the distances}$$

travelled by each component and Γ is the reflection coefficient of ground

iii) For grazing incidence we have $\Gamma_\perp = -1$, $E_t = 0$ and $E_g = E_i$,

$$\implies E_{TOT}(d,t) = \frac{E_0 d_0}{d'} \cos\left[\omega_c\left(t - \frac{d'}{c}\right)\right] + (-1)\frac{E_0 d_0}{d''} \cos\left[\omega_c\left(t - \frac{d''}{c}\right)\right]$$

Propogation Models

i) Propagation models that predict the **mean** signal-strength for a T-R separation are called **large-scale** propogation models

→ Large-scale models characterize signal strength over large T-R separation, typically 100 - 1000 meters

→ The mean PL gradually decreases with T-R separation

ii) The **rapid** fluctuations over **short travel** (few λ) or for **short time** are called **small-scale fading**

→ Signal may vary 30 to 40 dB for a fraction of λ

Using a reference point at distance d_0

i) Consider the received power measured (or predicted) at a reference point at distance d_0 as $P_r(d_0)$, where $d_0 \geq d_f$,

→ Typically, d_0 is ≈ 1 m in indoor and $\approx 100m - 1km$ in outdoor for $1 - 2GHz$ freq.

ii) Then the received power at distance $d > d_0$ can be obtained by

$P_r(d) = P_r(d_0) \left(\dfrac{d_0}{d}\right)^2$, where $d > d_0 > d_f$

iii) In dB, we have the equation as

$P_r(d)(dB) = 10 \log P_r(d_0) + 20 \log \left(\dfrac{d_0}{d}\right)$ when $P_r(d_0)$ is in Watts

And in dBm, we have

$P_r(d)(dBm) = 10 \log P_r(d_0) + 20 \log \left(\dfrac{d_0}{d}\right)$ when $P_r(d_0)$ is in milli Watts

Example:

Q) Given $D = 1m$, $f_c = 900MH$, $G_r = G_t = 1$ and $L = 1$, compute Fraunhofer distance d_f and path loss PL at d_f.

Sol) We have

→ $\lambda = \frac{c}{f_c} = \frac{3\times10^8}{900\times10^6} = 0.33 \; m$,

→ $d_f = \frac{2D^2}{\lambda} = 6 \; m$, and

→ $PL(dB) = -10 \log \frac{\lambda^2}{(4\pi)^2 d^2} = 47 \; dB$

Relating Power to Electric Field (Continuation)

iii) Further, $P_r(d)$ is related to the open circuit rms voltage at receiver antenna as

$P_r(d) = \dfrac{V^2}{R_{ant}} = \dfrac{[V_{ant}/2]^2}{R_{ant}} = \dfrac{V_{ant}^2}{4R_{ant}}$, where R_{ant} is the resistance of the matched receiver.

Reflection Coefficient Γ

i) The reflection coefficient for the parallel polarized and perpendicular polarized E-fields is given by

$$\Gamma_\parallel = \frac{E_r}{E_i} = \frac{\eta_2 \sin\theta_t - \eta_1 \sin\theta_i}{\eta_2 \sin\theta_t + \eta_1 \sin\theta_i}, \qquad \Gamma_\perp = \frac{E_r}{E_i} = \frac{\eta_2 \sin\theta_i - \eta_1 \sin\theta_t}{\eta_2 \sin\theta_i + \eta_1 \sin\theta_t}$$

where η is the intrinsic impedance $= \sqrt{\mu/\varepsilon}$

ii) Using snell's law at the boundary, we get

$\theta_i = \theta_r$, $\quad E_r = \Gamma E_i \quad$ and $E_t = (1 + \Gamma) E_i$

For a case where first medium is free space and $\mu_1 = \mu_2$

i) We have

$$\Gamma_\parallel = \frac{-\varepsilon_r \sin\theta_i + \sqrt{\varepsilon_r - \cos^2\theta_i}}{\varepsilon_r \sin\theta_i + \sqrt{\varepsilon_r - \cos^2\theta_i}}, \qquad \Gamma_\perp = \frac{\sin\theta_i - \sqrt{\varepsilon_r - \cos^2\theta_i}}{\sin\theta_i + \sqrt{\varepsilon_r - \cos^2\theta_i}}$$

ii) Note that both $\Gamma_\parallel = 1$ and $\Gamma_\perp = -1$ when $\theta_i \to 0$, i.e., ground may be modelled as a perfect reflector with $|\Gamma| = 1$ when E_i grazes the earth

Brewster Angle

i) It is the angle at which no reflection occurs in the medium of origin i.e., it occurs when the incident angle θ_B is such that $\Gamma_\parallel = 0$

$$\implies \sin\theta_B = \sqrt{\frac{\varepsilon_1}{\varepsilon_1 + \varepsilon_2}}$$

ii) For a case when the first medium is free space and second medium has ε_r,

$$\sin\theta_B = \sqrt{\frac{\varepsilon_r - 1}{\varepsilon_r^2 - 1}}$$

iii) Note that the Brewster angle occurs only for parallel polarization.

iv) Given $f = 100MHz$ compute θ_B when a) $\varepsilon_r = 4$ b) $\varepsilon_r = 15$:

a) $\sin\theta_B = \sqrt{\frac{1}{5}} \approx 26.56°$, b) $\sin\theta_B = \sqrt{\frac{14}{224}} \approx 14.47°$

Two-Ray Model (continuation)

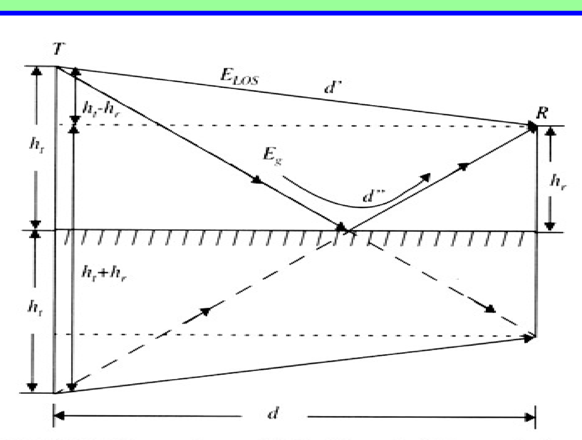


Figure 4.8 The method of images is used to find the path difference between the line-of-sight and the ground reflected paths.

iv) Using the method of images, when $d \gg (h_t + h_r)$, we get

$$\Delta \triangleq d'' - d' = \sqrt{(h_t + h_r)^2 + d^2} - \sqrt{(h_t - h_r)^2 + d^2} \approx \frac{2h_t h_r}{d}$$

$\implies$ The phase difference between E-fields and time delay of arrival are,

$$\theta_\Delta = \frac{2\pi\Delta}{\lambda} = \frac{\Delta\omega_c}{c} \text{ and } \tau_d = \frac{\Delta}{c} = \frac{\theta_\Delta}{2\pi f_c}$$

v) Note when d is large, Δ becomes small

$\implies E_{LOS}$ and E_g have identical amplitude and differ only in phase

i.e., $\left|\dfrac{E_0 d_0}{d}\right| = \left|\dfrac{E_0 d_0}{d'}\right| = \left|\dfrac{E_0 d_0}{d''}\right|$

vi) Further, the E-field evaluated at $t = \frac{d''}{c}$

$$E_{TOT}(d, t = \frac{d''}{c}) = \frac{E_0 d_0}{d'} \cos\left[\omega_c\left(\frac{d'' - d'}{c}\right)\right] - \frac{E_0 d_0}{d''} \cos 0$$
$$= \frac{E_0 d_0}{d'} \angle\theta_\Delta - \frac{E_0 d_0}{d''} = \frac{E_0 d_0}{d} [\angle\theta_\Delta - 1]$$

→ Using trignometric identities, we get $|E_{TOT}(d)| = 2\dfrac{E_0 d_0}{d} \sin\left(\dfrac{\theta_\Delta}{2}\right)$

Note that for increasing distance, $E_{TOT}(d)$ decays in an oscillatory fashion

vii) When $\frac{\theta_\Delta}{2}$ is small (i.e., $\frac{\theta_\Delta}{2} < 0.3 \; rad$), $\sin\left(\frac{\theta_\Delta}{2}\right) = \frac{\theta_\Delta}{2} = \frac{2\pi h_t h_r}{\lambda d}$, and received power $P_r = P_t G_t G_r \frac{h_t^2 h_r^2}{d^4}$. In terms of path loss, we have

$PL(dB) = 40 \log d - (10 \log G_t + 10 \log G_r + 20 \log h_t + 20 \log h_r)$

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Large Scale PL-II

II. Diffraction (Propagation Mechanism)

i) Diffraction occurs when EM wave is obstructed by a surface that has sharp edges.
present every where even behind the obstacle

ii) At high freq. diffraction depends on the geometry of the object as well as the amplitude, phase and polarization of the EM wave

iii) Diffraction allows radio signals to propagate around the curved surface beyond the origin and behind obstructions.

→ Received field decreases rapidly as receiver moves deeper into the shadow region

iv) This phenamenon is explained by Huygen's principle, which

→ states that all points on a wavefront can be considered as point sources for production of secondary wavelets

→ The wavelets combine to produce a new wavefront in the direction of propagation

→ Field strength in the shadowed region is the vector sum of the E-field components of all the wavelets

III. Scattering (Propagation Mechanism)

i) Scattering occurs due to objects with dimensions that are small compared to the wavelength and where the number of objects per unit volume is large. Ex: Trees, street signs and lamp posts etc.

ii) Also, when a radio wave impinges on a rough surface, the reflected energy is spread out (diffused) in all directions due to scattering. Thereby providing additional radio energy at a receiver.

iii) The roughness of large flat surfaces contribute to scattering

→ Roughness is often tested using the Rayleigh criterion defined by critical height (h_c) of surface protuberances for a given θ_i,

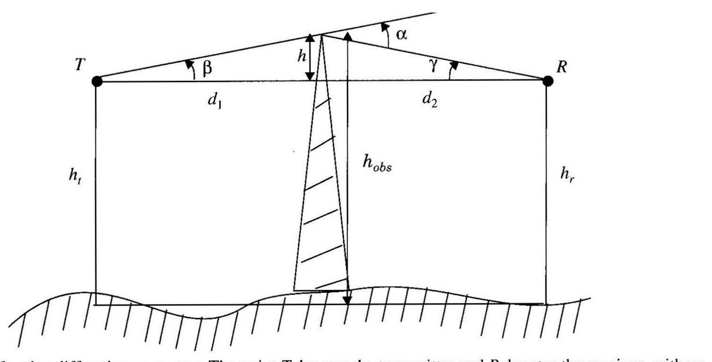
$$h_c \triangleq \frac{\lambda}{8\sin\theta_i}$$

iv) A surface is considered smooth if its minimum to maximum protuberance h is less than h_c and is considered rough if $h > h_c$

→ For rough surfaces, the reflection coefficient is given by,
$\Gamma_{rough} = \rho_s \Gamma$, where ρ_s is a scattering loss factor and
Γ is flat surface coefficient

→ When h is assumed to be Gaussian R.V with variance σ_h,
then we have $\rho_s = \exp\left[-8\left(\frac{\pi\sigma_h \sin\theta_i}{\lambda}\right)^2\right]$

Fresnel Zone Geometry



(a) Knife-edge diffraction geometry. The point T denotes the transmitter and R denotes the receiver, with an infinite knife-edge obstruction blocking the line-of-sight path.

i) Consider an obstructing screen of height h, be at a distance d_1 from the transmitter and d_2 from the receiver

→ The wave from the tx to rx via the top of the screen travels a longer distance than a LOS path.

→ Assuming $h \ll d_1, d_2$ and $h \gg \lambda$, the difference between the two paths Δ and the phase difference ϕ is given by

$$\Delta = \frac{h^2}{2}\frac{(d_1+d_2)}{d_1 d_2} \text{ and } \phi = \frac{2\pi\Delta}{\lambda} = \frac{2\pi}{\lambda}\frac{h^2}{2}\frac{(d_1+d_2)}{d_1 d_2}$$

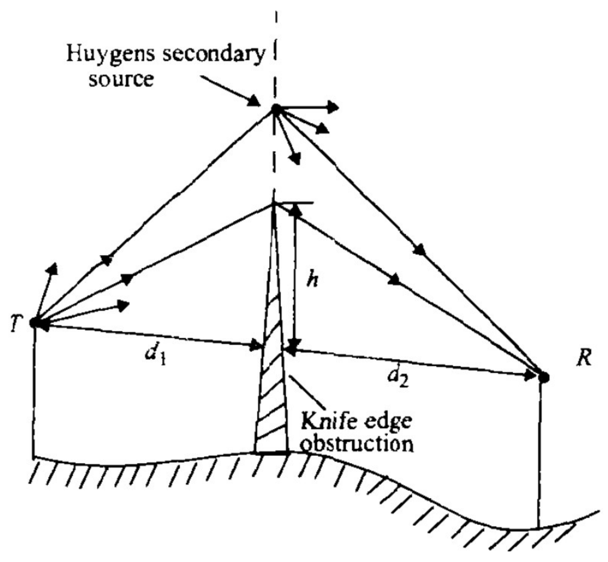
→ Further, when $\tan x = x$, then $\alpha = \beta + \gamma$ and $\alpha = h\frac{(d_1+d_2)}{d_1 d_2}$

ii) A dimensionless parameter ν called Fresnel-Kirchoff diffraction is defined as $\nu \triangleq h\sqrt{\frac{2(d_1+d_2)}{\lambda d_1 d_2}} = \alpha\sqrt{\frac{2 d_1 d_2}{\lambda(d_1+d_2)}}$

$\implies$ phase difference $\phi = \frac{\pi}{2}\nu^2$

(is a function of height h and distances d_1, d_2)

Knife-edge Diffraction Model



i) When shadowing is caused by a single object such as hill, the attenuation caused can be estimated by treating the obstruction as a diffracting knife-edge.
Diffraction loss in this case can be readily estimated using the Fresnel solution

ii) Consider a receiver at point R located in the shadowed region

→ The field strength at R is a vector sum of the fields due to all the secondary Huygen's sources in the plane above the knife edge.
The E-field strength E_d is given as

$$\frac{E_d}{E_0} = F(\nu) = \frac{1+j}{2}\int_{\nu}^{\infty} \exp\left[(-j\pi t^2)/2\right] dt$$

iii) The diffraction gain as compared to the free space E-field, is given by $G_d(dB) = 20\log|F(\nu)|$

Fresnel Zones

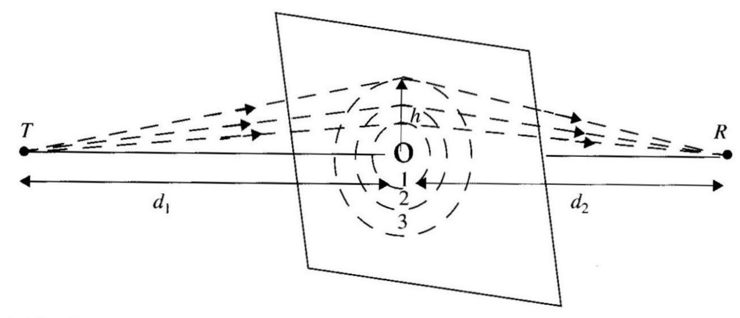


Figure 4.11 Concentric circles which define the boundaries of successive Fresnel zones.

i) The concept of diffraction loss as a function of the path difference is explained by Fresnel zones

→ These represent successive regions where secondary waves have a path length which are $\frac{n\lambda}{2}$ greater than the LOS path

→ The concentric circles on the plane represent the loci of the origins of wavelets such that the path length increases by $\frac{\lambda}{2}$ for successive circles (which are called Fresnel zones)

ii) Alternate Fresnel zone have the effect of providing constructive and destructive interference

iii) The radius of the n^{th} Fresnel zone is $r_n = \sqrt{\frac{n\lambda d_1 d_2}{d_1+d_2}}$

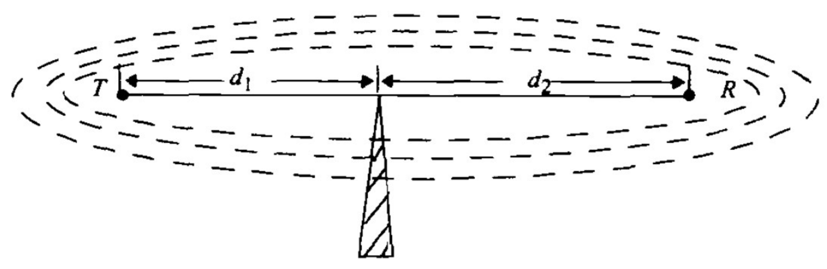
→ The path through the smallest circle $n = 1$ will have an excess path of $\frac{\lambda}{2}$. And for $n = 2, 3$, etc will have excess of $\lambda, \frac{2\lambda}{2}$ etc.

→ Zones will have maximum radii if $d_1 = d_2$ and smaller radii when the plane is moved towards the tx or rx

iv) An obstruction causes a blockage of energy from some of the Fresnel zones. The received energy will be a vector sum of the energy contributions from all unobstructed Fresnel zones.

v) A family of ellpsoids can be constructed between a tx and rx by joining all the points for which the excess path delay is an integer multiple of $\lambda/2$. The ellipsoids represent Fresnel zones.

→ In general, if an obstruction does not block the volume contained within the first Fresnel zone, then the diffraction loss will be minimal



(b) α and ν are equal to zero, since h is equal to zero

Multiple Knife-edge Diffraction

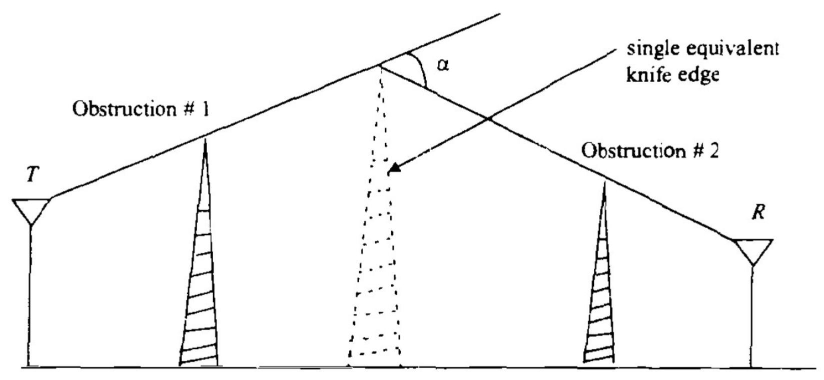


Figure 3.15
Bullington's construction of an equivalent knife edge [From [Bul47] © IEEE]

i) In many practical situations, the propagation path may consist of more than one obstruction

→ In such case, to compute the diffraction loss it is suggested to replace the series of obstructions by a single equivalent obstacle
This method simplifies the calculations

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Unit-II: Large-Scale Path Loss

Outdoor Propagation Model

i) The terrain profile of a particular area needs to be taken into account for estimating path loss.

→ Terrain may vary from curved earth to highly mountainous

ii) Different models aim to predict signal strength at a particular receiving point or a sector

Longley-Rice Model

i) This is applicable to point-to-point communication in freq. range from $40\ MHz$ to $100\ GHz$ over different kinds of terrain

ii) Median transmission loss is predicted using path geometry of the terrain profile and the refractivity of the troposhere

→ Primarly the two-ray ground reflection model is used to predict signal strengths within radio horizon.

→ The diffraction losses are estimated using the Fresnel-Kirchoff knife edge models

→ Forward scatter theory is used to make troposcatter predictions

iii) Longley-Rice method operates in two modes

→ when a detailed terrain profile is available, the prediction is called a point-to-point mode

→ If profile is not available, parameters are estimated and such prediction is called area mode

iv) One short-comming is that it does not provide corrections due to environment factors, buildings and foliage present in the immediate vicinity of the mobile receiver.

Okumura Model

i) This model is widely used for signal prediction in urban area for freq. in the range $150\ MHZ$ to $1920\ MHz$ (extrapolate to $3000\ MHz$) and distance of 1 Km to 100 Kms

→ Base station antenna heights from 30 m to 1000 m.

ii) Provides a set of curves giving the median attenuation (A_{mu}) relative to free spaces in an urban area over quasi-smooth terrain

→ with BS effective antenna height (h_{te}) of 200 m

→ and a mobile antenna height (h_{re}) of 3 m

→ curves were developed from extensive measurements using vertical omni directional antennas plotted as a function of freq (100 MHz to 1920 MHz) and distance (1 Km to 100 Kms)

Hata Model

i) This model is an empirical formulation of the graphical pathloss data provided by Okumura and is valid from 150 to 1500 MHz

ii) Standard formulae for median pathloss is

$$L_{50}(\text{urban})(dB) = 69.55 + 26.16 \log f_c - 13.82 \log h_{te} - a(h_{re}) + (44.9 - 6.55 \log h_{te}) \log d,$$

where f_c is freq, h_{te} is effective BS antenna (30 to 200 m), h_{re} from 1 to 10 m, d in Kms and $a(h_{re})$ is the correction factor which is function of the coverage area

→ For small city

$a(h_{re}) = (1.1 \log f_c - 0.7)h_{re} - (1.56 \log f_c - 0.8)$ dB

iii) For sub-urban area, the modified eq is

$L_{50}(dB) = L_{50}(\text{urban}) - 2\left[\log(f_c/28)\right]^2 - 5.4$

→ And for open rural area

$L_{50}(dB) = L_{50}(\text{urban}) - 4.78\left[\log(f_c)\right]^2 + 18.33 \log f_c - 40.94$

iv) The above eqs have significant practical values. Hata model compares very closely with the original Okumura model

→ Suited for large cell mobile systems, but not for personal communication system (PCS) with radius < 1 Km.

Walfisch and Bertoni Model

i) This model considers the impact of rooftops and building heights by using diffraction to predict avg. signal strength at street level. Path loss (S) is given by

→ $S = P_0 Q^2 P$, where

P_0 represents free space path loss given by $P_0 = \left(\frac{\lambda}{4\pi R}\right)^2$

Q^2 gives the reduction in rooftop signal due to the row of buildings which shadow the receiver

P_1 term is based upon diffraction to determine the loss from rooftop to the street

ii) In dB, $S(dB) = L_0 + L_{rts} + L_{ms}$, where L_0 is free space loss, L_{rts} is rooftop-to-street diffraction and L_{ms} is multi-screen diffraction due to the rows of buildings

Wideband PCS Microcell Model

i) This work is developed from extensive measurements in LOS and obstructed (OBS) environment using 20 MHz pulsed transmission at 1900 MHz

→ It revealed that a two-ray model is good estimate for path loss in LOS microcells

→ And simple log-distance path loss model holds well for OBS microcell

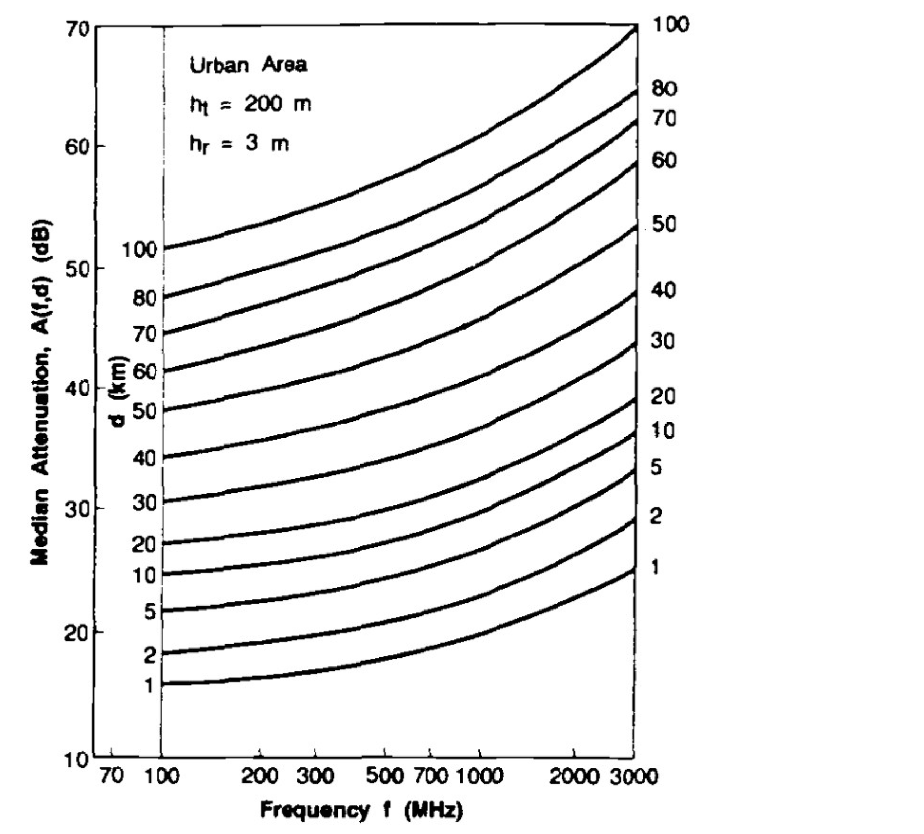
ii) For a flat earth ground reflection model, the distance d_f at which the first Fresnel zone just becomes obstructed by ground is given by $d_f = \frac{1}{\lambda}\sqrt{16 h_t^2 h_r^2 - \lambda^2(h_t^2 + h_r^2) + \frac{\lambda^4}{16}}$

iii) For LOS cases, a double regressing PL model fit well. It predicts avg PL as

$$\overline{PL}(d) = \begin{cases} 10 n_1 \log(d) + P_1 & \text{for } 1 < d < d_f, \\ 10 n_2 \log(d/d_f) + 10 n_1 \log d_f + P_1 & \text{for } d > d_f. \end{cases}$$

iv) For OBS case, standard log-distance PL fits well and is given by $\overline{PL}(d)[dB] = 10 n \log(d) + P_1$, where n is the OBS path loss exponent

Okumura Model (Continuation)



Okumura Model (Continuation)

iii) The pathloss is given by

$L_{50}(dB) = L_F + A_{mu}(f,d) - G(h_{te}) - G(h_{re}) - G_{Area}$, where

L_{50} is the median value of propagation path loss

L_F is the free space propagation loss

A_{mu} is median attenuation obtained from curves

$G(h_{te}), G(h_{re})$ are antenna gain factors and

G_{Area} is environment gain factor.

iv) Further, Okumura found that

→ $G(h_{te}) = 20 \log\left(\frac{h_{te}}{200}\right)$ for $1000m > h_{te} > 30m$

and varies by 20 dB/decade

→ $G(h_{re}) = 10 \log\left(\frac{h_{re}}{3}\right)$ for $h_{re} < 3m$

and varies by 10 dB/decade

v) Advantages: Simplest and best in terms of accuracy for mature cellular and land mobile radio systems in cluttered environments.

→ Disadvantages: It has a slow response to rapid changes in terrian and this model is not good in rural areas

PCS Extension to Hata Model

i) The European COST-231 committe developed an extended version of the Hata Model to handle 2 GHz. The proposed model for path hoss is

$$L_{50}(\text{urban}) = 46.3 + 33.9 \log f_c - 13.82 \log h_{te} - a(h_{re}) + (44.9 - 6.55 \log h_{te}) \log d + C_M$$

where C_M is 0 dB for medium city, suburban area and 3 dB for metropolitan centers

ii) This extension is restricted to the following range

freq: 1500 to 2000 MHz

h_{te}: 30 to 200 m

h_{re}: 1 to 10 m

d: 1 to 20 Km

Propagation geometry for Walfisch model

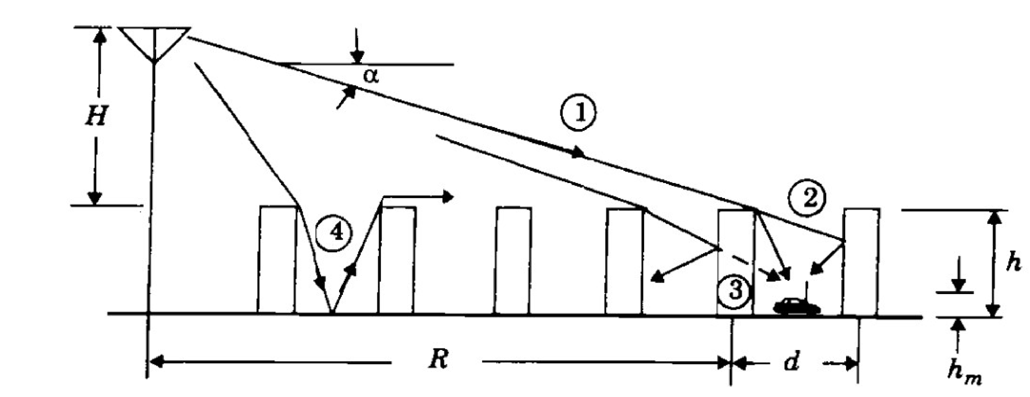


Figure 3.25
Propagation geometry for model proposed by Walfisch and Bertoni [From [Wal88] © IEEE]

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Large Scale PL

Indoor Propagation Model

i) These models deal with radio propagation characteristics inside the buildings
ii) The indoor channels differ from the traditional channels in
→ The distances covered are much smaller
→ The variability of the environment is much greater for a much smaller range of T-R separation
iii) The propagation within buildings is influenced by specific features such as building layout, construction materials and the building types
iv) Propagation is dominated by the same mechanisms of reflection, diffraction and scattering as outdoor. However, conditions are much more variable. Ex:
→ Signal levels vary greatly on whether interior door are open or closed
→ Whether antennas are mounted at desk level or ceiling level

Signal Penetration into Buildings

i) The signal strength received inside a building increases with height
→ At lower floors of a building, the urban clutter induces greater attenaution and reduces the level of penetration
→ At higher floors a LOS may exist thus causing a stronger incident signal
ii) Further, RF penetration has been found to be a function of
→ Frequency: Penetration loss decreases with increasing frequency
→ Antenna pattern in the elevation plane
→ Measurements made in front of windows indicated less penetration loss compared to building parts without window.
→ Metallic tints provide 3 to 30 dB attenuation for a single pane
→ Shadowing effects of adjacent buildings in high floors

Ray Tracing and Site Specific Modeling

i) This model of predicting RF signal coverage involves the use of site specific (SISP) propagation models and graphical information system (GIS) databases
→ SISP models ray tracing as a means of deterministically modeling any indoor or outdoor propagation environment
ii) For outdoor propagation, ray tracing is used with aerial photographs so that 3-D representation of buildings may be integrated with software to carry out reflection, diffraction and scattering models
iii) Using building databases, wireless systems are developed using computer aided design tools that provide deterministic, rather than statistical, prediction models.
→ For example, in the Siteplanner CAD environment, measurements and predictions of coverage and capacity can be simultaneously viewed, manipulated and optimized

Partition Losses (same floor)

i) Buildings have a wide variety of partitions and obstacles within their structure
→ Partitions that are formed as part of the building structure are called hard partitions
→ Partitions that may be moved and which do not span to the ceiling are called soft partitions
ii) Partitions vary widely in their physical and electrical characteristics, making it difficult to apply general models to specific indoor installation

Partition Losses between floors

i) The losses between floors of a building are determined
→ By the external dimensions and materials of the building
→ Type of construction used to create floors
→ Number of windows in a building and the presence of tinting
ii) The attenuation between floor is termed as floor attenuation factor (FAF)
→ After about 5 or 6 floor separation, very little additional path loss is experienced

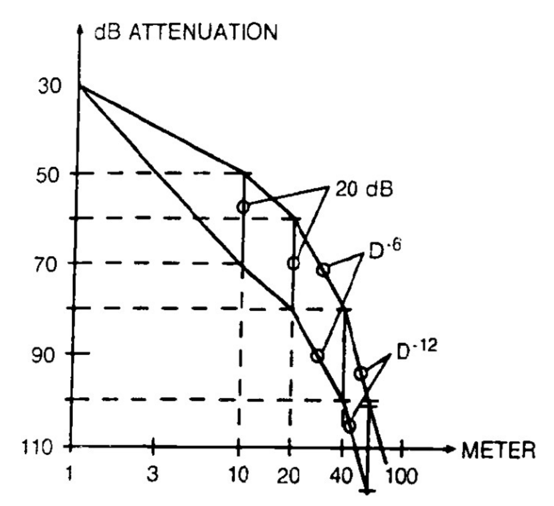
Log-distance Path Loss Model

i) Indoor path loss has been shown to obey the distance power law given by
$PL(dB) = PL(d_0) + 10n\log(\frac{d}{d_0}) + X_\sigma$, where
n depends on the surroundings and building type
X_σ represents a normal random variable in dB having a standard deviation of σ dB

Ericsson Multiple Breakpoint Model

i) This model was obtained by measurements in a multiple floor office building
→ The model has four breakpoints and considers both an upper and lower bound on the path loss
→ The model assumes that there is 30 dB attenuation at $d_0 = 1$m
ii) Rather than assuming a log-normal shadowing component, the Ericsson model provides a deterministic limit on the range of path loss at a particular distance
→ Further, uniform distribution was used to generate path loss values within the max and min range as a function of distance

Ericsson in-building pathloss model



Attenuation Factor Model

i) This model includes the effect of building type as well as the variation caused by obstacles and has been used to accurately deploy indoor and campus networks
→ It was shown to reduce the standard deviation between measured and predicted path loss to around 4 dB
ii) This model is given by
$$\overline{PL}(d)[dB] = \overline{PL}(d_0)[dB] + 10n_{SF}\log(\frac{d}{d_0}) + FAF[dB] + \sum PAF[dB]$$, where
→ n_{SF} represents the exponent value for the same floor measure
→ FAF represents a floor attenuation factor for a specified number of building floors,
→ PAF represents the partition attenuation factor for a specific obstruction encountered by a ray drawn between the transmitter and receiver in 3-D using primary ray tracing
iii) Further FAF may be replaced by an exponent n_{MF} which already considers the effects of multiple floor separation as
$$\overline{PL}(d)[dB] = \overline{PL}(d_0)[dB] + 10n_{MF}\log(\frac{d}{d_0}) + \sum PAF[dB]$$
iv) Additionally, it was found that in-building PL obeys free space PL plus additional loss which increases exponentially with distance
$$\overline{PL}(d)[dB] = \overline{PL}(d_0)[dB] + 20\log(\frac{d}{d_0}) + \alpha d + FAF[dB] + \sum PAF[dB]$$, where
α is the attenuation constant with units dB/m

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Small-Scale Fading and Multipath

Small-Scale Multipath Propagation

i) Fading is used to describe the rapid fluctuations of the amplitude, phase or multipath delays of a RF signal on a short period of time or short travel distance (typically 10λ - 20λ)

ii) Fading is caused by interference between two or more versions of the signal which arrive at the receiver at slightly different times.

→ These waves called multipath waves combine at the receiver antenna to give a resultant signal which may vary widely

Effects of Small-scale Fading

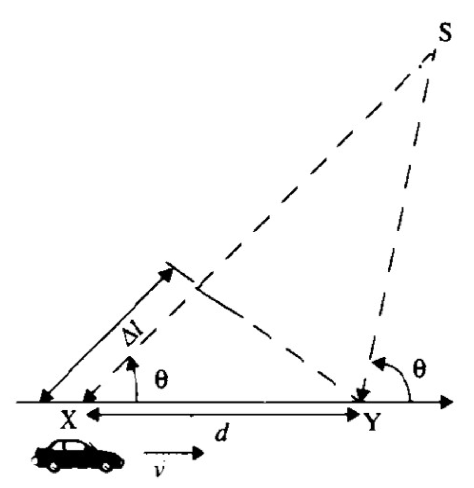
i) Multipath creates small-scale fading effects. The key effects are

→ Rapid changes in signal strength with short distance and short time interval

→ Random frequency modulation due to varying Doppler shifts on different multiplaths

→ Time dispersion (echoes) caused by multipath propagation delays

Doppler Shift Diagram



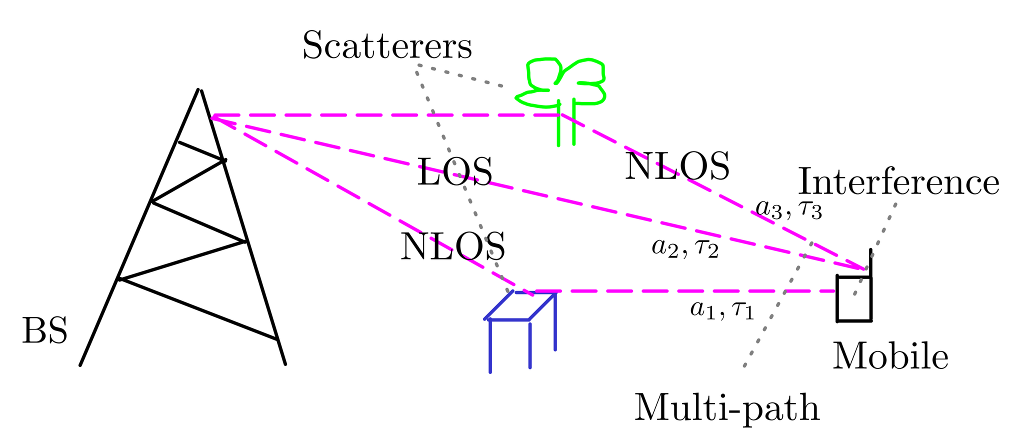
Impulse Response Model of a Multipath Channel

i) The small-scale variations of a radio signal can be directly related to the impulse response of the radio channel

→ The impulse response is the wideband characterization and contains all information to simulate or analyze radio transmissions

ii) The channel may be modeled as a linear filter with a time varying impulse response, where the time variation is due to receiver motion

Multipath Illustration



Factors Influencing small-scale Fading

I. Multipath propagation:

i) The presence of reflecting objects and scaterrers create a constantly changing environment

→ These result in multiple versions of the signal to arrive at the receiver that are displaced in time and spatial orientation

→ The random phase and amplitude of multipath components cause fluctuations in signal strength thereby inducing small-scale fading and/or distortion

II. Speed of the mobile:

i) The relative motion between the BS and mobile results in random freq. modulation due to different Doppler shifts on each of the path

→ Doppler shift will be positive or negative depending on the direction of mobile movement

Doppler Shift

i) Consider mobile moving with constant velocity v between X and Y separated by distance d, while receiving signal from source S

ii) The difference in path length travelled by the wave from S is
$\Delta I = d\cos\theta = v\,\Delta t\,\cos\theta$,
where Δt is time taken by mobile to travel from X to Y

→ The phase change in the received signal is given by
$\Delta\phi = \frac{2\pi\Delta I}{\lambda} = \frac{2\pi v\,\Delta t}{\lambda}\cos\theta$

→ Hence apparent change in the frequency (or Dopper Shift), f_d is
$f_d = \frac{1}{2\pi}\frac{\Delta\phi}{\Delta t} = \frac{v}{\lambda}\cos\theta$ ■

ii) If the mobile is moving towards the source, the f_d is positive (i.e., apparent freq is increase to $f_c + f_d$)
And if mobile is moving away, doppler shift is negative i.e., $f_c - f_d$

iii) Multipath components that arrive from different direction contribute to Doppler spreading, thus increasing the signal bandwidth

Impluse response at point d

i) For a fixed point d, the channel can be modelled as a linear time invariant system

→ Then channel impulse response can be expressed as $h(d,t)$, i.e., function of different locations

ii) Due to different multipath waves which have propagation delays, which vary over different locations, we have
$y(d,t) = x(t) \circledast h(d,t) = \int_{-\infty}^{t} x(\tau)h(d,t-\tau)d\tau$, where
$y(d,t)$ is received signal at position d and
$x(t)$ is the transmit signal

iii) Consider receiver moves at constant velocity v, i.e., $d = vt$,
$y(vt,t) = y(t) = x(t) \circledast h(vt,t) = \int_{-\infty}^{t} x(\tau)h(vt,t-\tau)d\tau$

⟹ Linear time-varying channel, where channel changes with time and distance

Discretizing the multipath delay axis τ

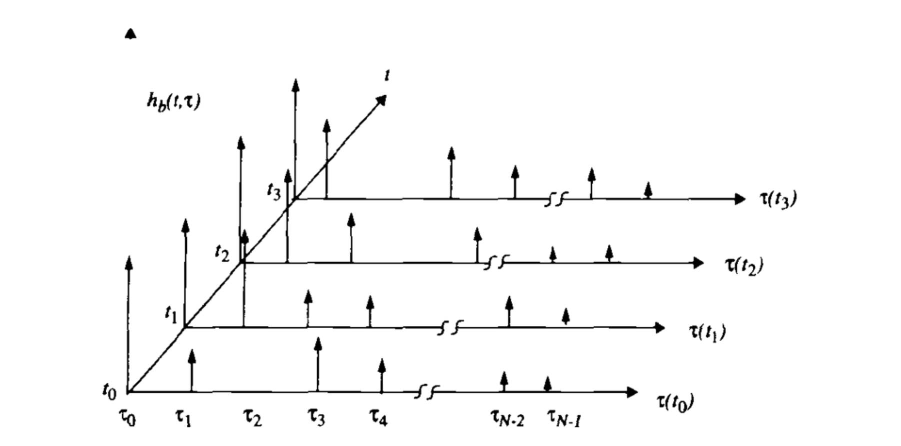


Figure 4.4
An example of the time varying discrete-time impulse response model for a multipath radio channel.

i) For simplifying the analysis, the multipath delay axis τ is discretized into equal time delay segments called excess delay bins, with delay width $\Delta\tau \triangleq \tau_{i+1} - \tau_i$

→ $\tau_0 = 0$ represents the first arriving signal at receiver and $\tau_i = i\Delta\tau$ for $i = 0, \cdots, N-1$

ii) All the multipath signals received within i^{th} bin are represented by a single component τ_i

→ This technique of quantization can analyze RF signals having BW $< \frac{2}{\Delta\tau}$

iii) Excess Delay: This is the relative delay of the i^{th} multipath component as compared to the first arriving component

→ The maximum excess delay of channel is $N\Delta\tau$

Factors Influencing small-scale Fading (continuation)

III. Speed of surrounding objects:

i) Moving objects induce a time varying Doppler shift on multipath components

→ If the objects move at a greater rate than the mobile, they dominate the small-scale fading, otherwise motion of objects can be ignored

ii) The staticness of the channel is defined by Coherence time, and is directly impacted by Doppler shift

IV. The transmission bandwidth of signal:

→ If the bandwidth of the signal is greater than the bandwidth of the channel, the received signal will be distorted, otherwise the signal experience small-scale fading (but not distortion)

Example

Q) Given $f_c = 900$ MHz and $v = 70$ Km/h, compute f_d for mobile moving a) towards source b) away from source and c) perpendicular to source

Sol) We have $\lambda = \frac{c}{f_c} = \frac{3\times10^8}{900\times10^6} = 0.33m$
$v = 70 \times \frac{1000}{3600} = 19.44$ m/sec and $f = f_c + f_d$

a) $f = 900\times10^6 + \frac{19.44}{0.33}\cos 0 = 900\times10^6 + \frac{19.44}{0.33}$

b) $f = 900\times10^6 + \frac{19.44}{0.33}\cos(\pi) = 900\times10^6 - \frac{19.44}{0.33}$

c) $f = 900\times10^6 + \frac{19.44}{0.33}\cos(\frac{\pi}{2}) = 900\times10^6 + 0$

Bandpass representation of Impluse response

i) With $x(t)$ are bandpass waveform, $y(t)$ received waveform and the time-varying multipath response $h(t,\tau)$ completely characterizes the channel and is a function of t and τ
i.e., $y(t) = x(t) \circledast h(t,\tau)$ where t represents time variation and τ represents channel multipath delay for fixed t

$$x(t) \rightarrow \boxed{h(t,\tau) = Re\{h_b(t,\tau)\,e^{j2\pi f_c t}\}} \rightarrow y(t)$$

→ which in terms of baseband signal are
$x(t) = Re\{c(t)e^{j2\pi f_c t}\}$, $y(t) = Re\{r(t)e^{j2\pi f_c t}\}$, where
$c(t), r(t)$ are complex envelope representation of $x(t), y(t)$ and
$h(t,\tau) = Re\{h_b(t,\tau)e^{j2\pi f_c t}\}$,
where $h_b(t,\tau)$ is complex baseband response

ii) The lowpass characterization removes the high freq. carrier variation and makes the signal analytically easier

$$c(t) \rightarrow \boxed{\tfrac{1}{2}h_b(t,\tau)} \rightarrow r(t)$$

Expression for baseband impulse response

i) Since the received signal is the sum of multiple time-delayed and phase shifted replicas of the tx signal, the baseband response can be expressed as
$h_b(t,\tau) = \sum_{n=0}^{N-1} a_n(t,\tau)\, e^{j[2\pi f_c\tau_n(t) + \phi_n(t,\tau)]}\,\delta(\tau - \tau_n(t))$

→ where $a_n(t,\tau)$ and $\tau_n(t)$ are the real amplitudes and excess delays of n^{th} component at time t,

→ $[2\pi f_c\tau_n(t) + \phi_n(t,\tau)]$ represents the phase shift

⟹ This is a Time-varying discrete-time channel response model

ii) If the channel is assumed to be time invariant over a local area, then
$h_b(\tau) = \sum_{n=0}^{N-1} a_n\, e^{j\theta_n}\,\delta(\tau - \tau_n)$

Power Delay Profile (PDP)

i) For measuring or predicting $h_b(\tau)$, a probing pulse $p(t)$ which approximates a unit impulse function is used at the transmitter
$p(t) \approx \delta(t - \tau)$

→ The power delay profile is found by spatial average of $|h_b(t,\tau)^2|$ over a local area

ii) By making several local area measurements, it is possible to build an ensemble of power delay profile $P(\tau)$
$P(\tau) = k\overline{|h_b(t,\tau)^2|}$, where

→ the gain k relates the power of input pulse to the received power

→ bar indicates the average over the local area of many snapshots

iii) Thus, PDP are the plots of relative received power as a function of excess delay

→ These are measured by channel sounding technique

→ They are found by averaging instantaneous power delay measurements over a local area

Local area: No greater than 6 m outdoor and no greater than 2m indoor

Samples taken at $\frac{\lambda}{4}$ meters approx

For frequency range of 450 MHz to 6 GHz

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Small-Scale Multipath Measurements

i) Multipath structure is the key factor for small-scale fading

→ Several channel sounding techniques are used to measure the power delay profile of a multipath wireless channel

Direct RF Pulse System

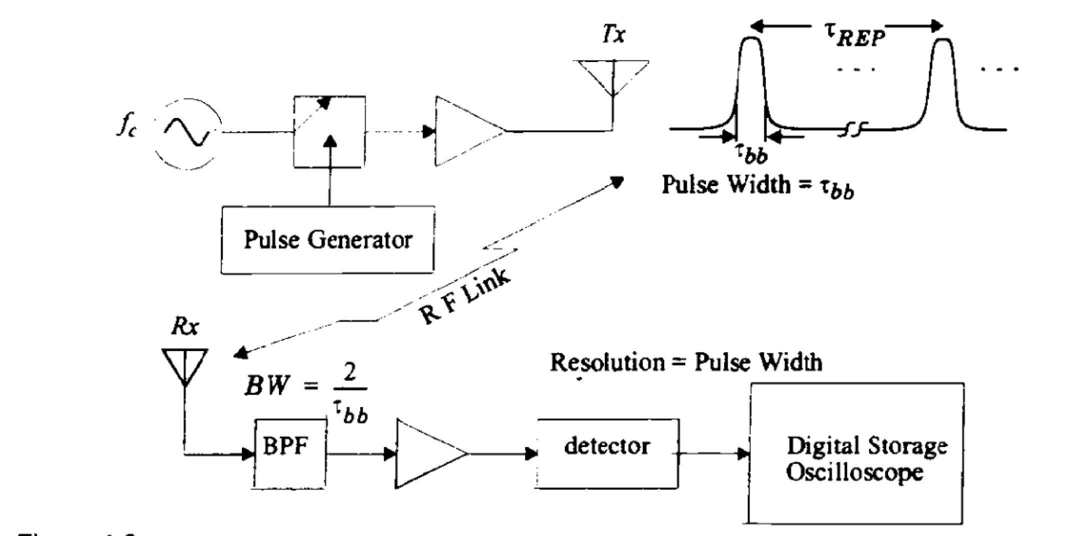

Figure 4.6
Direct RF channel impulse response measurement system.

Direct RF Pulse System (Continuation)

i) Is a simple technique to rapidly determine the PDP of any channel

ii) At Transmitter: A wideband pulsed bistatic radar transmits a repetitive pulse of width T_{bb} for sounding

iii) At Receiver: The signal is amplified by a wide bandpass filter $BW = \frac{2}{T_{bb}}$, then detected with an envelope detector and stored on a high speed oscilloscope

→ This gives instantaneous square of channel impulse response convolved with the probing pulse

→ The oscilloscope is set on average mode to provide local average power

iv) Disadvantages:

→ Due to wide passband filter, it is subjected to interference and noise

→ The system relies on the ability to trigger the oscilloscope on first arrival of received signal

→ The phase of multipath is not captured

Spread Spectrum Sliding Correlator Channel Sounding,

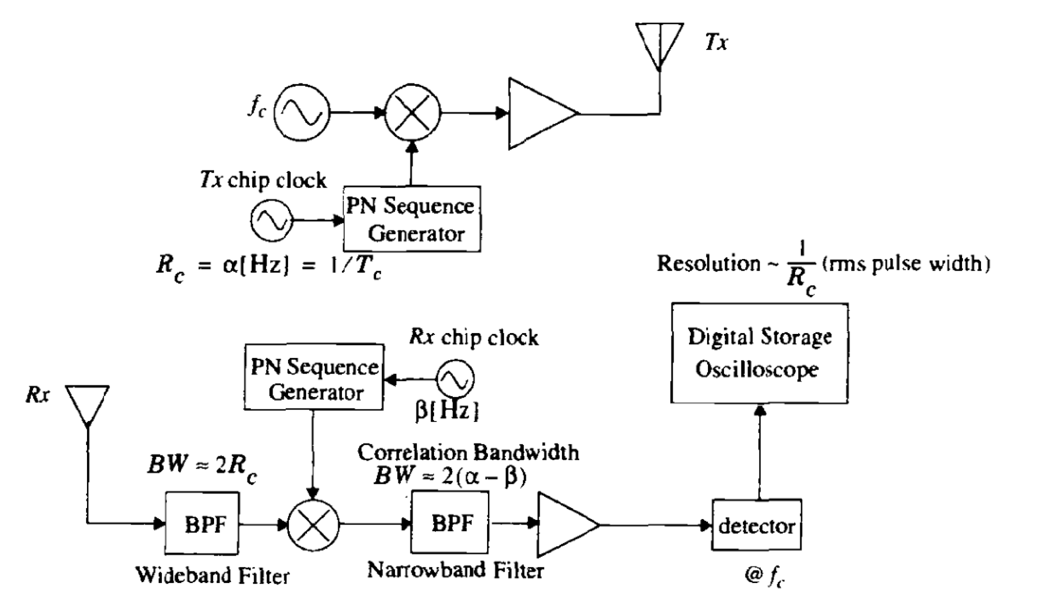

Figure 4.7
Spread spectrum channel impulse response measurement system.

Spread Spectrum Sliding Correlator Channel Sounding (Continuation)

i) While the probing signal may be wideband, the transmitted signal is detected using a narrowband receiver preceded by a wideband mixer, thus improving the range

ii) A carrier signal is spread over a large bandwidth by mixing it with a binary pseudorandom (PN) sequence having a chip duration of T_c, rate $R_c(=\frac{1}{T_c})$ Hz and $BW = 2R_c$

iii) At receiver the signal is despread using same PN

iv) The transmitter chip clock rate α is little faster than the receiver chip clock rate β, which results in sliding correlator

Futher, Processing Gian $= \frac{2R_c}{R_{bb}} = \frac{2T_{bb}}{T_c} = \frac{(S/N)_{out}}{(S/N)_{in}}$

Time resolution $\Delta\tau = 2T_c = \frac{2}{R_c}$

Sliding factor $\gamma = \frac{\alpha}{(\alpha-\beta)}$

v) Advantages:

a) Improves coverage range using same transmitter power,

b) Tx and Rx synchronization is eliminated using sliding correlator

→ Disadvantages: a) Measurements are not made in real time,

b) Associated time required is large,

c) Phase information is lost

Frequency Domain Channel Sounding,

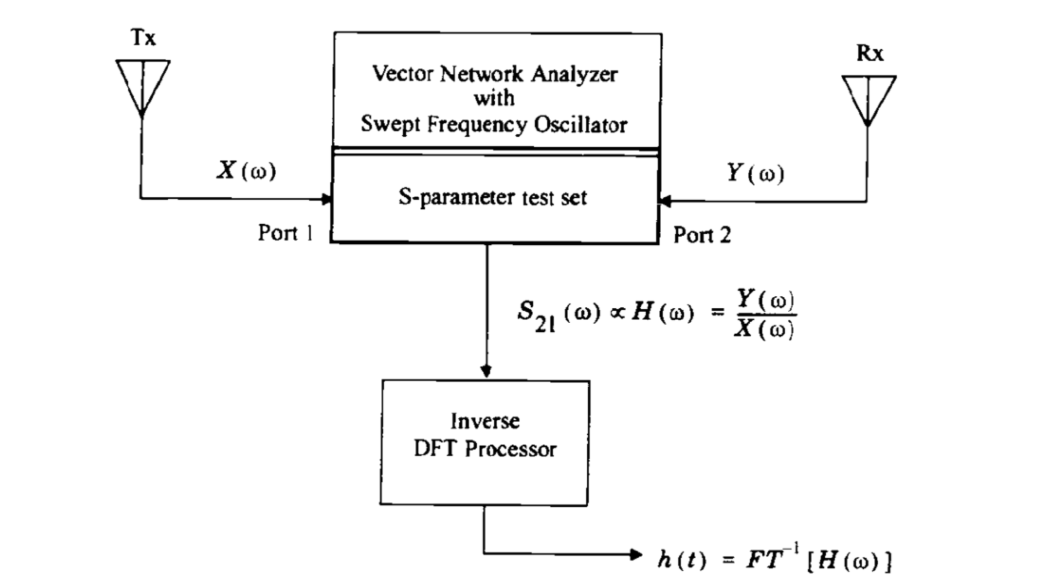

Figure 4.8
Frequency domain channel impulse response measurement system.

Frequency Domain Channel Sounding (Continuation)

i) The dual relationship between time and frequency is used to measure channel impulse response

ii) A vector network analyzer controls a synthesized frequency sweeper

→ A S-parameter test set is used to monitor the frequency response of the channel

→ The sweeper scans particular frequency band by stepping through discrete frequencies

→ The number and spacing of frequency steps impact the time resolution of impulse response measurements

→ The response is converted to time domain by using inverse Discrete Time Fourier Transform (IDFT)

iii) Advantages: Provides amplitude and phase information in time domain

→ Disadvantaes: a) Requires carefull calibration,

b) Needs hardwired synchroniation between Tx and Rx

c) Practical only for indoor measurements

d) Non-real time measurements

e) For time varying, channel response may change giving erroneous measurments

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Parameters of Mobile Multipath Channels

i) Many multipath channel parameters are derived from the power delay profile given by
$|r(t_0)|^2 = \sum_{k=0}^{N-1} a_k^2(t_0)$, where
$r(t_0)$ is the received signal at $t = t_0$ when the channel is sounded by a repetitive baseband pulse train $p(t)$

Time Dispersion Parameters (Due to Multipath delays)

i) These parameters grossly quantify the multipath channel. These are determined from a power delay profile

ii) **Mean excess delay**, $\overline{\tau}$ is the first moment of Power Delay Profile, and is given by $\overline{\tau} = \frac{\sum_k a_k^2 \tau_k}{\sum_k a_k^2} = \frac{\sum_k P(\tau_k)\tau_k}{\sum_k P(\tau_k)}$

iii) **RMS delay spread**, σ_τ is the square root of second central moment of the power delay profile, and is given by

$\sigma_\tau = \sqrt{\overline{\tau^2} - (\overline{\tau})^2}$, where $\overline{\tau^2} = \frac{\sum_k a_k^2 \tau_k^2}{\sum_k a_k^2} = \frac{\sum_k P(\tau_k)\tau_k^2}{\sum_k P(\tau_k)}$

→ Typically σ_τ is $\approx$ μ sec in outdoor and $\approx$ nano sec in indoor

iv) **Maximum excess delay** (X dB) of the PDP is the time delay during which multipath energy falls to X dB below the maximum i.e., $\tau_X - \tau$, where τ_X is the max delay at which multipath component is within X dB of the strongest component

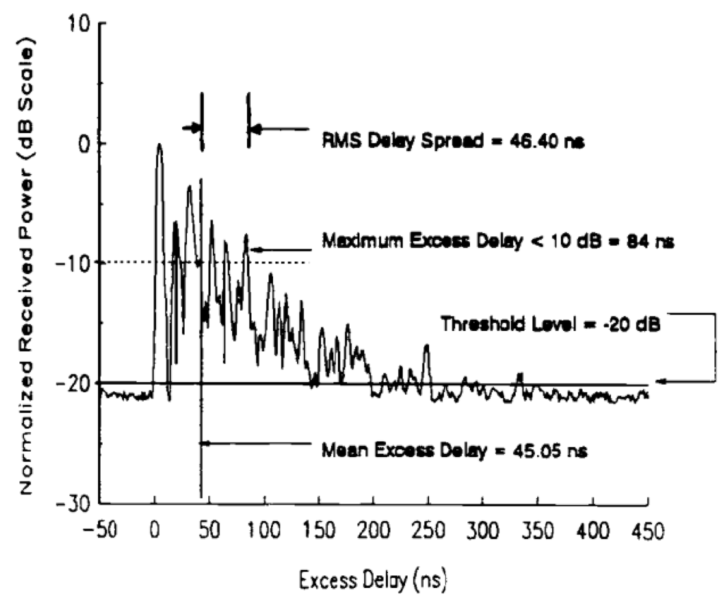


Figure 4.10
Example of an indoor power delay profile; rms delay spread, mean excess delay, maximum excess delay (10 dB), and threshold level are shown.

v) In practice, values for $\overline{\tau}, \overline{\tau^2}$ and σ_τ depend on the choice of noise threshold used to process $P(\tau)$

→ Noise threshold differentiates between received multipath components and thermal noise

→ If noise threshold is set too low, then noise will be processed as multipath component and thus causing the parameters to be higher

Doppler Spread (Due to mobility)

i) Doppler spread B_D is a measure of the sprectral broading caused by the time rate of change of the mobile channel (due to velocity)

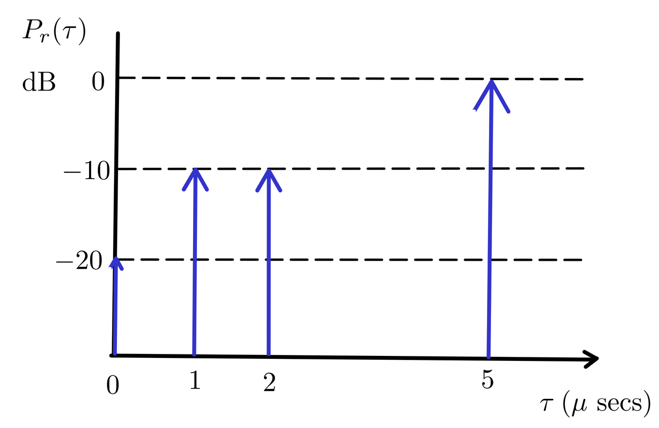
→ It is defined as the range of frequencies over which the received Doppler spectrum is non-zero

→ Ex: When a pure sinusoidal tone of f_c is transmitted, the received spectrum called Doppler spectrum will have components in the range $[f_c - f_d, f_c + f_d]$, where f_d is doppler shift

ii) The spectral broadening depends on f_d, which is a function of the relative velocity of the mobile and (θ) direction of motion

iii) If the baseband signal bandwidth is much greater than B_D, the effects of Doppler spread are negligible.
(This scenario is called Slow Fading)

Example: Time Dispersion



Q) Compute $\overline{\tau}, \overline{\tau^2}, \tau_{max}(10\ dB)$ and $50\,\%\ B_C$

Sol) $\overline{\tau} = \frac{10^{-2}\times 0 + 10^{-1}\times 1 + 10^{-1}\times 2 + 10^{0}\times 5}{10^{-2} + 10^{-1} + 10^{-1} + 10^{0}} = 4.38\mu\ sec$

→ $\overline{\tau^2} = \frac{10^{-2}\times 0 + 10^{-1}\times 1^2 + 10^{-1}\times 2^2 + 10^{0}\times 5^2}{10^{-2} + 10^{-1} + 10^{-1} + 10^{0}}$
$= 21.07\mu sec$

→ $\sigma_\tau = \sqrt{21.07 - (4.38)^2} = 1.37\mu sec$

→ $\tau_{max}(10\ dB) = 5\ \mu sec$

→ $B_C \approx \frac{1}{5\times 1.37\mu} = 146\ KHz$

Coherence Bandwidth, B_C

i) Analogous to the delay spread parameter in time domain, coherence bandwidth is used to characterize the channel in freq. domain

ii) It is a statistical measure of the range of freq. over which the channel can be considered *flat*

→ i.e., passes all spectral components with approx equal gain and linear phase

ii) Two sinusoids with freq. separation greater than B_C are affected quite differently by the channel

iii) If Coherence BW is defined as the range over which freq. correlation function

→ $C_{r1,r2} \geq 0.9$, then $B_C \approx \frac{1}{50\sigma_\tau}$

→ $C_{r1,r2} \geq 0.5$, then $B_C \approx \frac{1}{5\sigma_\tau}$ (also called 50% Coherence BW)

Delay Spread Vs Doppler Spread

i) Delay spread and coherence bandwidth describe the time dispersive nature of the channel in local area

→ However, they donot provide information about the time variation caused by relative motion

ii) Alternatively, Doppler spread and coherence time describe the time variation in a small-scale region

Coherence Time

i) Coherence time is the time domain dual of Doppler spread and characterizes the frequency dispersiveness in time domain

→ The doppler spread and coherence time are inversely proportional to one another i.e., $T_C \approx \frac{1}{f_m}$, where f_m is max Doppler shift $\frac{v}{\lambda}$

ii) Coherence time is actually a statistical measure of the time duration over which the channel impulse response is essentially invariant and quantifies the similarity of the channel response at different times

→ It is the time duration over which two received signals have a strong potential for amplitude correlation

→ If the reciprocal BW of the baseband signal is greater than the coherence time, then the channel will change during transmission of message, thus causing distortion (known as slow fading)

iii) Coherence Time is related to f_m as $T_C = \frac{0.423}{f_m}$

→ Two signals arriving with a time separation greater than T_C are affected differently by the channel

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Types of Small-Scale Fading

i) **Depending** on the relation between the **signal parameters** (i.e., bandwidth and symbol period) and the **channel parameters** (i.e., RMS delay spread and Doppler spread) different transmitted signals will undergo **different types** of **fading**.

→ **Multipath delay** spread leads to **time dispersion** and **frequency selective** fading

→ **Doppler spread** leads to **frequency dispersion** and **time selective** fading

→ The two mechanisms are **independent** of each other

Fading Effects Due to Multipath Time Delay Spread

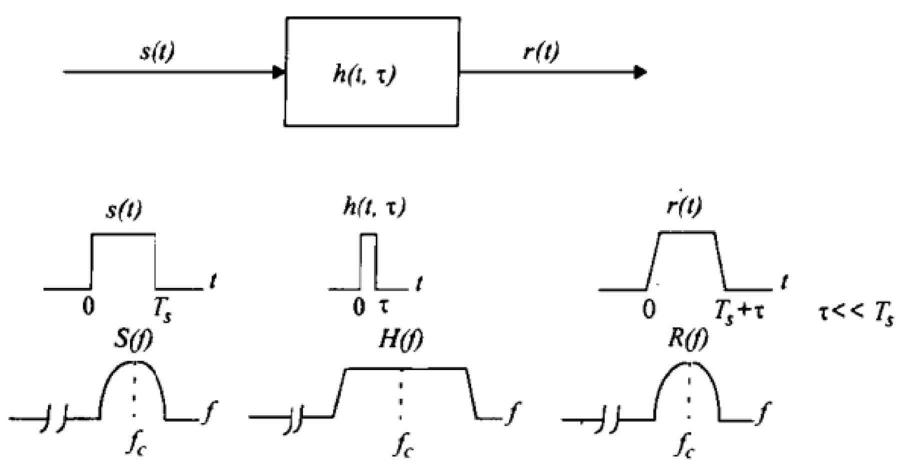
Small-Scale Fading
(Based on multipath time delay spread)

Flat Fading
1. BW of signal < BW of channel
2. Delay spread < Symbol period

Frequency Selective Fading
1. BW of signal > BW of channel
2. Delay spread > Symbol period

→ The **delay spread** of the channel and the **symbol duration** determine whether a signal undergoes **flat** fading or **freq. selective** fading

Flat Fading



i) If the channel has a **constant gain** and **linear phase** response on a **bandwidth** which **is greater** than the BW of the **transmitted signal**, the received signal will undergo **flat fading**. In this case,

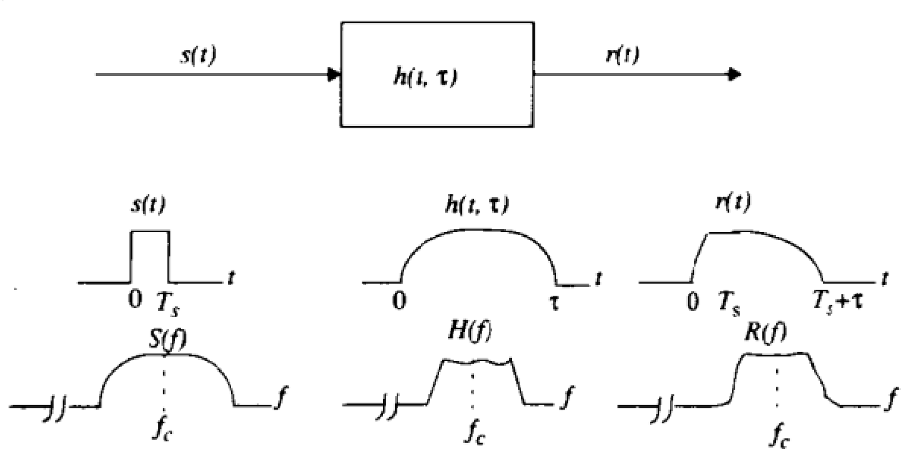
→ In the **time domain**, $T_s \gg \tau$, i.e., $T_s \gg \sigma_T$, where T_s is the symbol period (i.e., reciprocal bandwidth) and σ_T is the RMS delay spread

→ In the **freq. domain**, $B_s \ll B_c$, where B_s is the signal bandwidth and B_c is the coherence bandwidth

ii) The **spectral characteristics** of received signal are **preserved**

→ However the **strength** of the received signal **changes with time**

Frequency Selective Fading



i) If the channel possess a constant gain and linear phase response **over a BW** that is **smaller than** the **transmitted signal**, then the channel creates **freq. selective fading**.

→ In this case, in the **time domain**, $T_s < \tau$, i.e., $T_s < \sigma_T$

→ In the **freq. domain**, $B_s > B_c$

ii) The received signal includes **multiple version** of transmitted waveform which are **attenuated** (faded) and **delayed** in time, as a result received signal is **distored**. I.e., in the time-domain

→ the channel induces **intersymbol interference** (ISI)

→ in the frequency domain, the **gain is different** for **different** frequency components

Fading Efffects Due to Doppler Spread

Small-Scale Fading
(Based on Doppler spread)

Fast Fading
1. High Doppler spread
2. Coherence time < Symbol period
3. Channel variations faster than baseband signal variations

Slow Fading
1. Low Doppler spread
2. Coherence time > Symbol period
3. Channel variations slower than baseband signal variations

→ The **velocity** of the mobile (or velocity of surrounding objects) and the **baseband signalling** determines whether a signal undergoes **fast** fading or **slow** fading

Fast Fading

i) In a **fast fading channel**, the channel **impulse response changes** rapidly **within** the **symbol duration**. I.e., the **coherence time** T_c of the channel is **smaller** than the **symbol period** of the **transmitted signal** $T_s > T_c$

→ This is also called time selective fading and leads to signal distortion

ii) In freq. domain $B_D < B_d$ where B_D is the **Doppler spread**

→ signal distortion due to fast fading increases with increasing Doppler spread

iii) Fast fading only deals with rate of change of the channel **due to motion**

→ In practice, fast fading **only occurs** for very **low data rates**

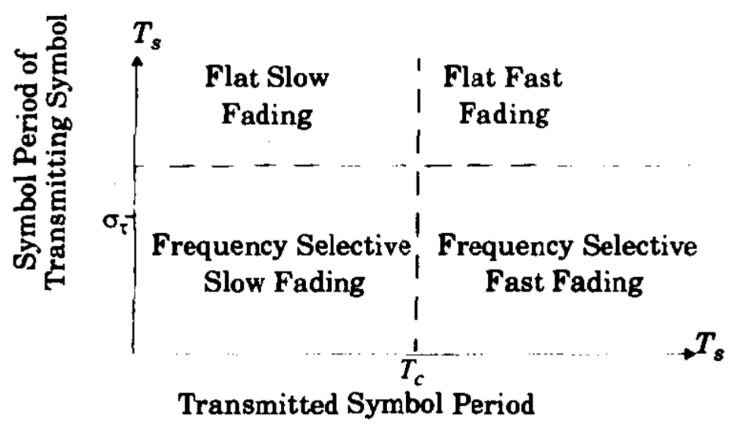
Slow Fading

i) In a slow fading channel, the **impulse response changes** at a rate **much slower** than the **transmitted** based band signal $s(t)$, i.e., $T_s \ll T_c$

→ In the freq. domain, this implies that the **Doppler spread** of the channel is much **less than** the **bandwidth** of the baseband **signal**, i.e., $B_s \gg B_D$

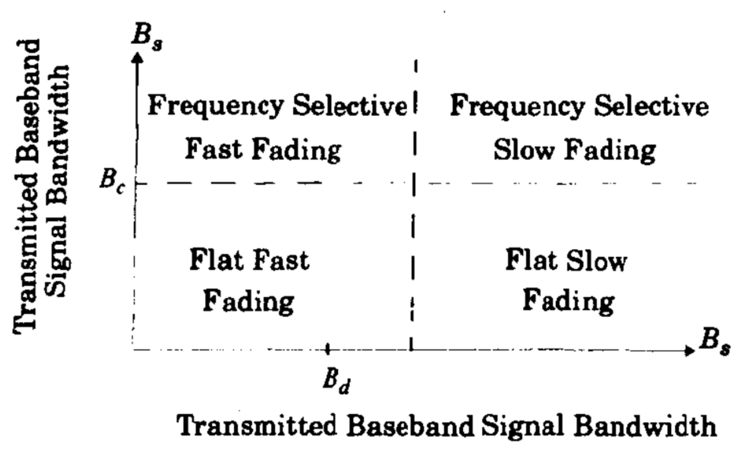
Summary Matrix

→ **Fading** experienced as a **function** of **symbol period**



Summary Matrix

→ **Fading** experienced as a **function** of **signal bandwidth**



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Small-Scale Fading and Multipath

Wireless Multipath Channel Gain

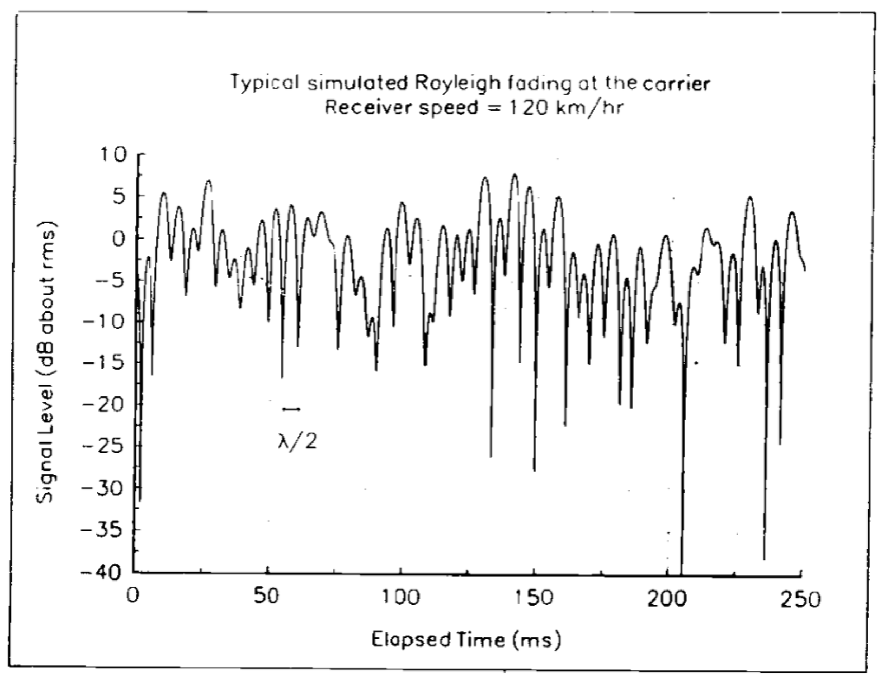
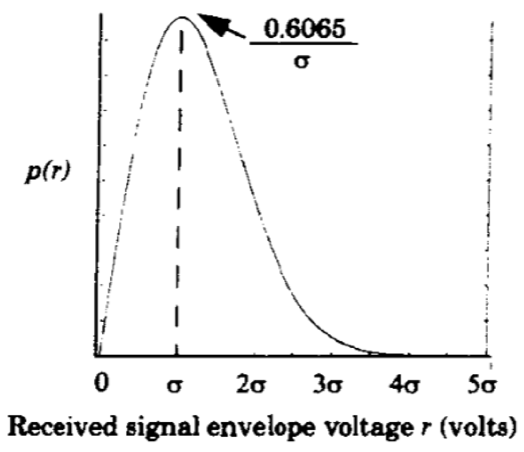
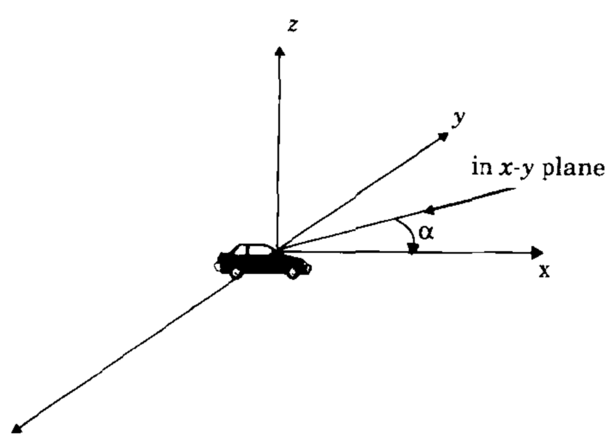


Figure 4.15
A typical Rayleigh fading envelope at 900 MHz [From [Fun93] © IEEE].

Rayleigh probability density function (pdf)



Clarke's Model - Diagram



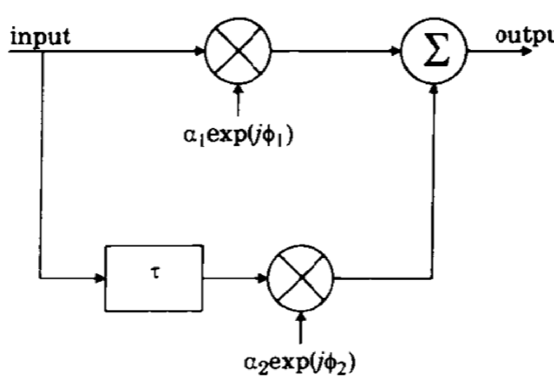
Statistical Models for Multipath Fading Channels

i) A statistical model explains the observed statistical nature of a mobile channel

ii) Ossana's model predicts flat fading power spectra for suburban areas

iii) Whereas Clarke's model is based on **scattering** and is widely used in **urban areas**

Block diagram of Two-Ray Rayleigh Fading Model



Rayleigh Fading Distribution

i) Rayleigh distribution is commonly used to describe the statistical **time varying nature** of the received **envelope** of a **flat fading** signal

→ The envelope of the **sum of two quadrature Gaussian** noise signals obeys a **Rayleigh** distribution

ii) Probability density function (**pdf**) is given by

$$p(r) = \begin{cases} \frac{r}{\sigma^2} e^{-\frac{r^2}{2\sigma^2}} & 0 \le r \le \infty \\ 0 & r \le 0 \end{cases}$$

where σ is the RMS value of the received voltage signal

σ^2 is the time-average power

→ The probability that the envelope **is within** a specified value R is given by

$Pr(r \le R) = \int_0^R p(r)dr = 1 - e^{-\frac{R^2}{2\sigma^2}}$

→ **Mean value** r_{mean} is given by

$r_{mean} = E\{r\} = \int_0^\infty r p(r) dr = \sigma\sqrt{\frac{\pi}{2}}$

→ **Variance** σ^2, which represents the AC power is given by

$\sigma^2 = E\{r^2\} - E\{r^2\}^2 = \int_0^\infty r^2 p(r) dr - \frac{\sigma^2\pi}{2} = \sigma^2\left(2 - \frac{\pi}{2}\right)$

→ **Median** value of r_{median} is computed as

$\frac{1}{2} = \int_0^{r_{median}} p(r)dr; \quad r_{median} = 1.77\sigma$

Clarke's Model for Flat Fading

i) This model uses the statistical characteristics of the EM fields of received signal from **scattering**

ii) Assumes that the mobile receives N **number of plane waves** with arbitrary carrier **phases**, arbitrary **angles** of arrival and having **equal average** amplitudes

→ Every wave that is incident on the mobile undergoes a **Doppler shift** $\left(f_n = \frac{v}{\lambda}\cos\alpha_n\right)$ due to receiver motion

→ Assumes that there is no excess delay due to multipaths

iii) Then, the vertically polarized plane waves arriving at the mobile have **E-field components** as

$E_z = E_0 \sum_{n=1}^{N} C_n \cos(2\pi f_c t + \theta_n)$, where

→ E_0 is the real **amplitude** of average E-field

→ C_n is a real **random variable** representing the amplitude of n^{th} wave, such that the ensemble **average** of is $\sum_{n=1}^{N} \overline{C_n^2} = 1$

→ f_c is the **carrier frequency**,

→ and the **random phase** θ_n of the n^{th} wave is $\theta_n = 2\pi f_n t + \phi_n$

Simulation of Clarke Fading Model

i) This simulation method uses the concept of **in-phase** and **quadrature** components to produce a **simulated** signal

$E_z = T_c(t) \cos 2\pi f_c t - T_s(t) \sin 2\pi f_c t$

ii) **Two independent Gaussian** low pass **noise** sources are used to **produce** in-phase and quadrature fading branches

→ The **complex Gaussian** source is formed by summing the two independent Gaussian random variables

→ Further **spectral filter** defined by

$$S_{E_z}(f) = \frac{1.5}{\pi f_m \sqrt{1 - \left(\frac{f - f_c}{f_m}\right)^2}}$$

is used to **shape** the random signal in the freq. domain

Fading Statistics

i) The level crossing rate (LCR) and average fade duration of a Rayleigh fading signal are two important statistics which are **useful** for designing **error control** codes and **diversity** schemes

Two-Ray Rayleigh Fading Model

i) This approach models the effects of multipath **delay spread** as well as **fading**

→ The block diagram represents 2-ray independent Rayleigh fading channel model

ii) The **impulse response** of the model is represented as

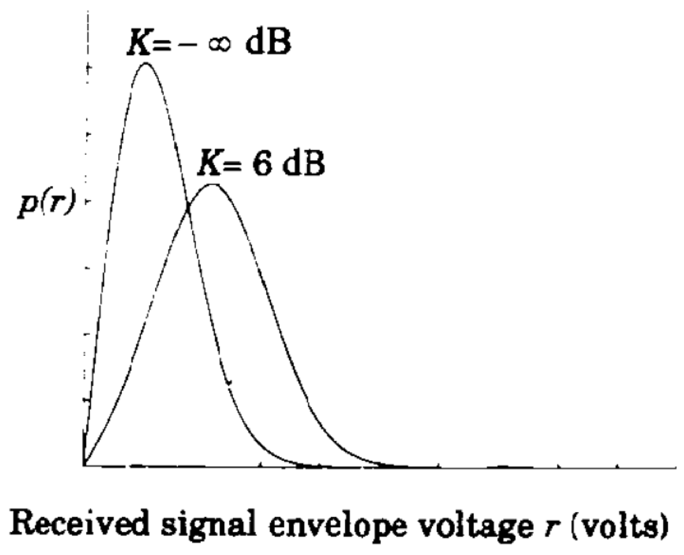
$h_b(t) = \alpha_1 e^{j\phi_1}\delta(t) + \alpha_2 e^{j\phi_2}\delta(t - \tau)$, where

α_1 and α_2 are independent **Rayleigh** distributed,

ϕ_1 and ϕ_2 are independent **uniformly** distributed over $[0, 2\pi]$

and τ is the time delay between the two rays

Ricean Fading Distribution



i) When there is a dominant stationary signal component present, such as a **LOS path**, the small-scale fading envelope distribution is **Ricean**

ii) As the **dominant signal** becomes **weaker**, the signal **resembles** an envelope that is **Rayleigh**

iii) The pdf of Ricean is given by

$$p(r) = \begin{cases} \frac{r}{\sigma^2} e^{-\frac{r^2 + A^2}{2\sigma^2}} I_0\left(\frac{Ar}{\sigma^2}\right) & \text{for } 0 \le A,\ 0 \le r \le \infty \\ 0 & \text{for } r \le 0 \end{cases}$$

→ A denotes the peak amplitude of the dominant signal and $I_0(\cdot)$ is the modified Bessel function of the first kind and zero-order

iv) **Ricean factor** K is ratio between the deterministic signal power and the variance of the multipath components, $K = \frac{A^2}{2\sigma^2}$

Clarke's Model for Flat Fading (cont ...)

iv) For sufficiently large N, E_z can be approximated as **Gaussian** random variable using CLT

→ Further, it can be modeled as **random process** and expressed in an in-phase and out-phase form as

$E_z = T_c(t) \cos 2\pi f_c t - T_s(t) \sin 2\pi f_c t$, where

$T_c(t) = E_0 \sum_{n=1}^{N} C_n \cos(2\pi f_n t + \phi)$ and

$T_s(t) = E_0 \sum_{n=1}^{N} C_n \sin(2\pi f_n t + \phi)$

and where $T_c(t), T_s(t)$ are **Gaussian** random processes

→ Implies, T_c and T_s are uncorrelated zero-mean **Gaussian random variables** with an equal variance given by

$\overline{T_c^2} = \overline{T_s^2} = \overline{|E_z|^2} = \frac{E_0^2}{2}$

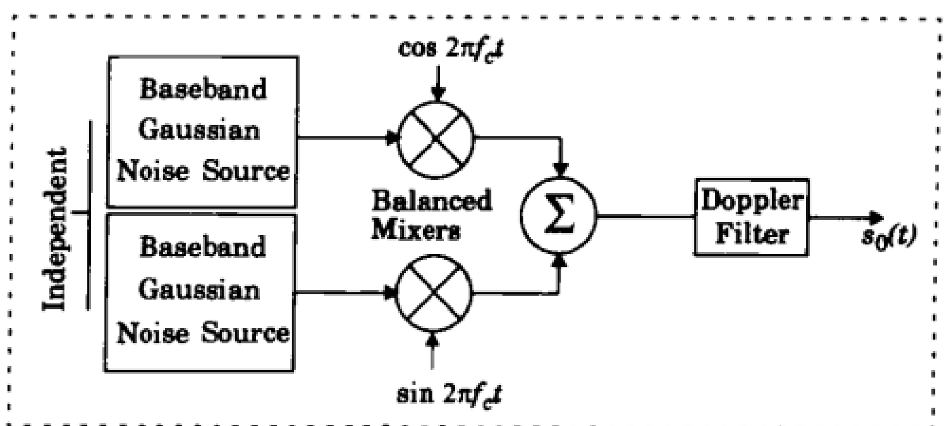
v) The **envelope** of the received E-field is given by

$|E_z(t)| = \sqrt{T_c^2(t) + T_s^2(t)} = r(t)$

→ The random signal envelope r has a **Rayleigh** distribution,

$$p(r) = \begin{cases} \frac{r}{\sigma^2} e^{-\frac{r^2}{2\sigma^2}} & 0 \le r \le \infty \\ 0 & r \le 0 \end{cases}, \text{ where } \sigma^2 = \frac{E_0^2}{2}$$

Simulator using Quadrature amplitude modulation



Level Crossing Rate

i) The **level crossing rate** (LCR) is defined as the expected rate at which the Rayleigh fading envelope (normalized to the local RMS signal level) **crosses** a **specified level** in a positive-going direction

ii) The **number** of **level crossings per second** is given by

$N_R = \sqrt{2\pi} f_m \rho e^{-\rho^2}$ at $r = R$, where

$\rho = \frac{R}{R_{rms}}$ is the value of the specified level

f_m is the maximum Doppler frequency

→ The LCR is a **function** of the **mobile speed**

→ There are few crossings at both high and low levels, with the **maximum rate** occuring at $\rho = \frac{1}{\sqrt{2}}$

→ The signal envelope experiences very deep fades only occasionally

Average Fade Duration

i) The average fade duration is defined as the **average period** of **time** for which the received **signal** is **below** a **specified level** R

→ For a Rayleigh fading signal, this is

$\overline{\tau} = \frac{1}{N_R} Pr[r \le R]$, where $Pr[r \le R] = \frac{1}{T}\sum_i \tau_i$

and where τ_i is the duration of the fade and T is the observation interval of the fading signal

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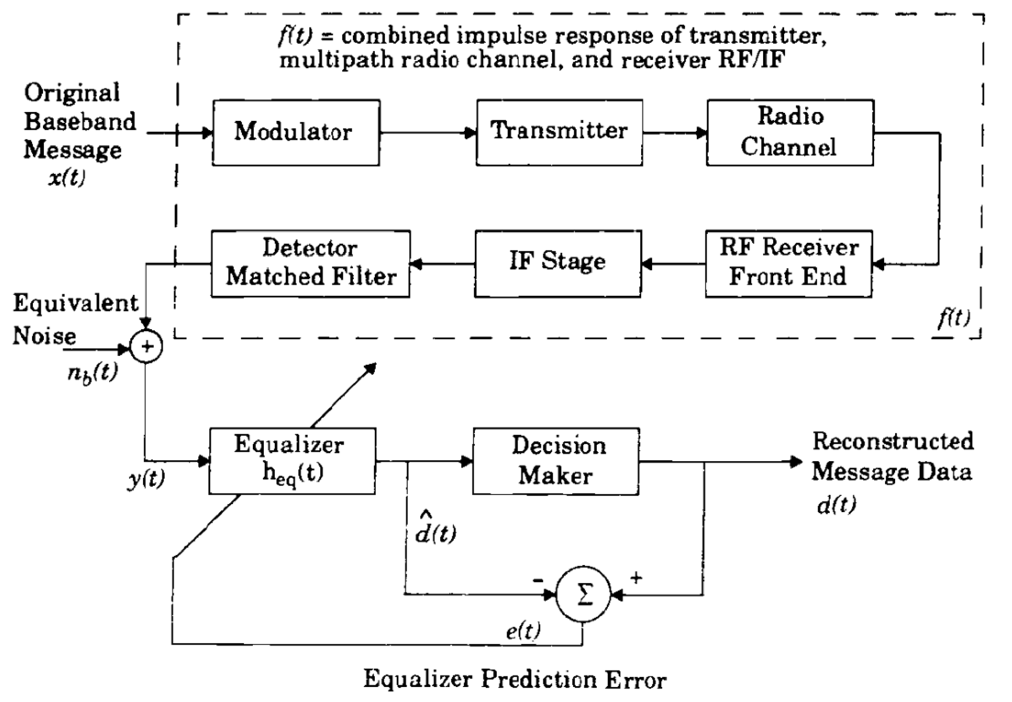
Introduction

i) The properties of mobile radio channels
→ **Multipath fading** leads to **time dispersion** and **ISI**
→ **Doppler Spread** leads to **time variation**
→ These have a strong **negative impact** on the **bit error rate** of any modulation
ii) The three techniques of **Equalization**, **Diversity** and **Channel Coding** are used to **improve** radio link performance i.e., to minimize the instantaneous bit error rate

Fundamentals of Equalization

i) **ISI** caused by multipath in time dispersive (i.e., freq. selective) channels **distorts** the transmitted signal causing **bit errors** at the receiver
→ This is a major **obstacle** to **high speed data** transmission
ii) **Equalization** is any **signal processing** operation that minimizes the ISI
→ Since the mobile fading **channel** is **random** and **time varying** equalizers must **track** the time varying **characteristics** of the channel, thus are called **adaptive** equalizers

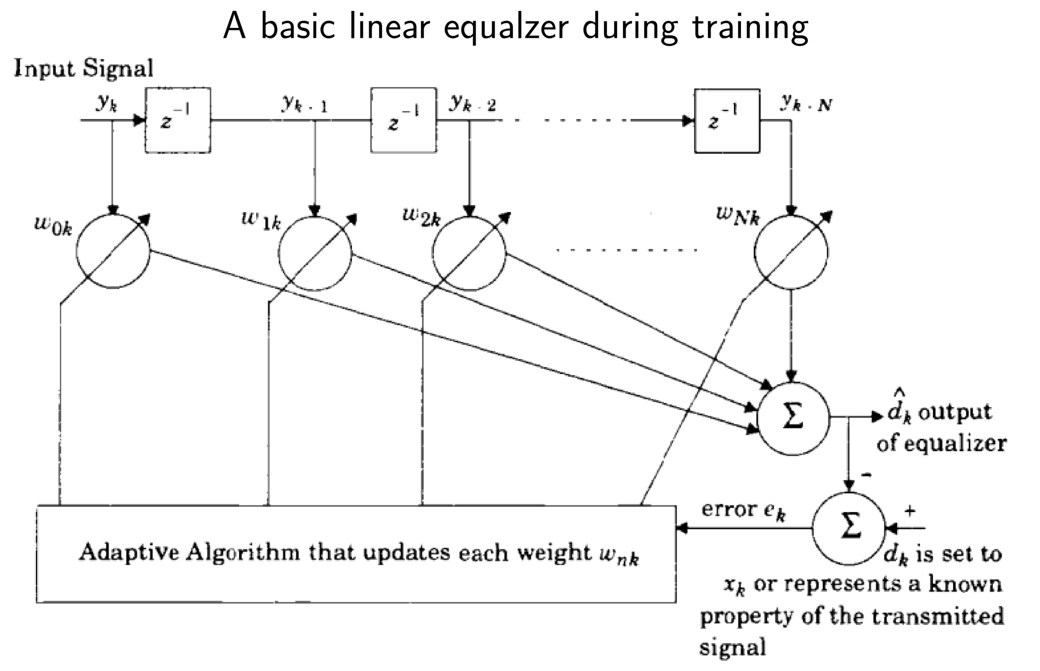


Equalizer Prediction Error

Equalization

Generic Adaptive Equalizer

A basic linear equalzer during training



i) An adaptive equalizer is a **time-varying** filter which must be constantly re-trained
ii) The **transversal filter** has a) N delay elements, b) $N+1$ taps c) $N+1$ tunable complex multipliers and d) $N+1$ weights
→ The **weights** are **updated** continously by the **adaptive algorithm**

Equalization Techniques

i) Equalization techniques can be subdivided into two general categories: a) **Linear** and b) **non-Linear**
→ Categories are determined from **how the output** of an equalizer is **used** for subsequent feedback of the equalizer
→ **Linear equalizer :** In this category $d(t)$ is **not used** in the feedback path to **adapt the equalizer**
→ **non-Linear equalizer:** In this category, $d(t)$ is fed back to change the subsequent outputs

Introduction (cont...)

i) **Equalization**: When the **signal BW** exceeds the **coherence** BW the received signal is **spread in time** and experiences **ISI**
→ Equalization technique compensates for ISI created by **multipath**
At the receiver, it **compensates** for the average range of expected **channel amplitude** and **delay** characteristics
→ Equalizers must be **adaptive** since the **channel** is generally unknown and **time varying**

Fundamentals of Equalization (cont...)

→ where $x(t)$ is the original information signal,
→ $f(t)$ is combined complex baseband **impulse response** of the transmitter, channel and the RF/IF section
i) Then the **signal received** by the equalizer is
$y(t) = x(t) \circledast f^*(t) + n_b(t)$, where $n_b(t)$ is the baseband noise
ii) Let $h_{eq}(t)$ be the **impulse response** of equalizer, then the output of the equalizer is

$$\begin{aligned}\hat{d}(t) = \ & x(t) \circledast f^*(t) \circledast h_{eq}(t) && + n_b(t) \circledast h_{eq}(t)\\ & x(t) \circledast g(t) && + n_b(t) \circledast h_{eq}(t)\end{aligned}$$

→ The complex baseband **impulse response** of a transversal filter equalizer is given by
$h_{eq}(t) = \sum_n C_n \delta(t - nT)$, where C_n are the complex filter coefficient of the equalizer
iii) The **desired output** of the equalizer is $x(t)$.
Assuming that $n_b(t) = 0$, for $\hat{d}(t)$ to be $= x(t)$, $g(t)$ must be
→ $g(t) = f^*(t) \circledast h_{eq}(t) = \delta(t)$, which implies in frequency domain
→ $H_{eq}(f)F^*(-f) = 1$. This indicates that an equlizer is an **inverse** filter of the channel
i.e., if the channel is **freq. selective**, the equalizer **enhances** the freq. components with **small amplitude**

Generic Adaptive Equalizer (cont...)

iii) The adaptive algorithm is **controlled** by the **error signal** e_k
→ where e_k is derived by **comparing** the **output** of the equalizer with same signal which is **exact** scaled **replica** of the **transmitted signal** x_k or which represents known property of x_k
→ The algo uses e_k to minimize a **cost function** and update the equalizer weights that iteratively reduces the cost function
→ Ex: The **Least Mean Square** (LMS) algo searches the optimum filter weights by performing
New weights = Prev. Weights +(Constant) × (Prev. error) ×(Current input vector)
where Prev. error = Prev. desired output − Prev. actual output
iv) This process is **repeated rapidly** in a programming loop while equalizer attempts to **converge**
→ Upon reaching the convergence, the adaptive algo **freezes** the filter weights until the error signal **exceeds** an acceptable level
v) Most common techniques used to minimize the error are a) **Gradient** decent algorithm and b) **Steepest** decent algorithm
→ Most common cost function used is **Mean Square Error** i.e., $E[e(k)\, e^*(k)]$

I. Linear Transversal Equalizer (LTE)

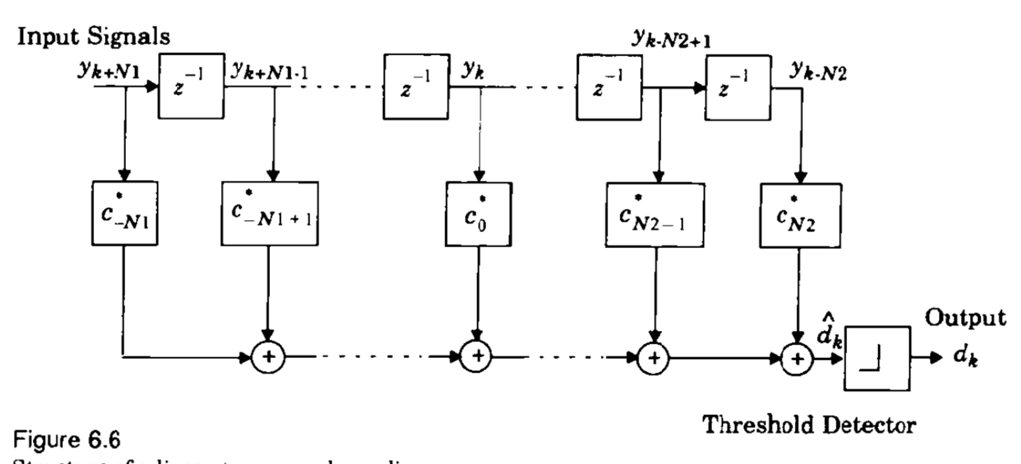


Figure 6.6
Structure of a linear transversal equalizer.

i) This structure can be implemented as a **FIR** filter which is also known as **transversal**
ii) **Current** and **past values** of the received signal are linearly **weighted** by the filter coefficient c_n and summed to produce the output
→ Implementation is usually carried out in **digital domain**
iii) The output before decision making (i.e., threshold detection)
$\hat{d}_k = \sum_{n=-N_1}^{N_2} c_n^* \, y_{k-n}$
iv) And the minimum **Mean Square Error** (MSE) it can acheive is

$$E\left[|e(n)|^2\right] = \frac{T}{2\pi}\int_{-\frac{\pi}{T}}^{\frac{\pi}{T}} \frac{N_0}{|F\left(e^{j\omega T}\right)|^2 + N_0}\, d\omega$$

where $F\left(e^{j\omega T}\right)$ is the **freq. response** of channel

non-Linear Equalizers

Introduction (cont...)

i) **Diversity**: This technique is usually employed to **reduce** the depth and duration of the **fades** experienced by a receiver in a **flat fading** channel without increasing the transmitted power or bandwidth.
→ This can be employed at both base station and mobile receiver

Operating Modes of an Adaptive Equalizer

i) **Training Mode** (First Stage):
→ A **known** fixed length **training sequence** is sent by the transmitter so that the equalizer may average a proper setting, i.e., **acquire** the proper **filter coefficients** in the worst possible channel conditions.
→ The training sequence is typically a **pseudorandom** binary signal or a fixed **bit pattern**
→ The time span over which an equalizer converges is a function of a) Equalizer **Algorithm**, b) Equalizer **Structure** and c) Time **rate of change** of the channel
→ Equalizer requires **periodic** re-training to maintain effective **ISI cancellation**
ii) **Tracking Mode** (Second Stage):
→ Immediately **after** the training sequence, the **user data is sent**
→ The **adaptive** algorithm **tracks** the changing channel and **adjusts** its filter characteristics even when the user data is received
→ **TDMA** wireless system are well **suited** for equalizers, where data is in fixed-length **time blocks**. The training sequence is usually sent at the **begining of a block**.
→ Equalizers can be **implemented** at baseband or at IF in a **receiver**

Blind Algorithms

i) More recent **Blind algorithms** are able to **exploit** characteristics of the transmitted signal and **do not require** training sequence
→ These provide equalizer **convergence** without **burdening** the transmitter with training **overhead**
→ They are able to acquire equalization through **property restoral** technique of the transmitted signal
ii) Examples:
→ **The Constant Modulus Algorithm (CMA)**: Used for constant envelope modulation
→ **Spectral Coherence Restoral Algorithm (SCORE)**: Exploits spectral redundancy or cyclostationarity in the transmit signal

Equalizers in a Communication Receiver

i) In a practical scenario, because of the presence of **noise** $n_b(t)$, an equalizer is **unable** to acheive **perfect performance**
→ The instantaneous combined freq. response **will not** always be **flat**, resulting in some **finite** prediction **error**. I.e.,

$$\begin{aligned}\hat{d}(n) &= x(n) \circledast g(n) + n_b(n) \circledast h_{eq}(n)\\ e(n) &= d(n) - \hat{d}(n)\end{aligned}$$

II. Linear Lattice Filter Implementation

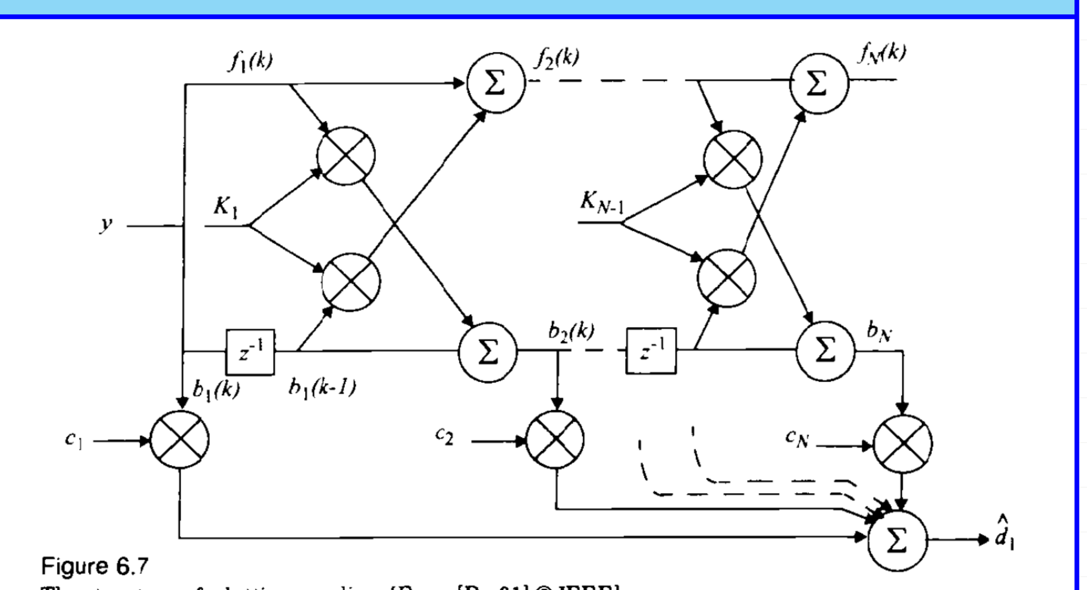


Figure 6.7
The structure of a lattice equalizer [From [Pro91] © IEEE].

→ Here $f_n(k)$ and $b_n(k)$ are **forward** and **backward** error signals
→ K_n is the reflection coefficient
→ And output $\hat{d}_k = \sum_{n=1}^{N} c_n(k) b_n(k)$
ii) Two main advantages of the lattice equalier are
→ Numerical stability and faster convergence
→ Allows **dynamic assignment** of stages as required.
I.e., when the channel is in more time dispersion, more stages are used
iii) Disadvantage: It is more complicated then LTE

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Equalization

non-Linear Equalizer

- i) Linear equalizers **do not perform well** on channels which have **deep spectral nulls** in the passband.
- → They place too **much gain** in the **vicinity** of the spectral null, thereby **enchancing** the **noise** present in those frequencies
- ii) **Non-linear** equalizers are used in applications where the channel **distortion** is **too severe** for a linear equalizer to handle
- iii) **Two** effective non-linear equalizers are:
- → Decision Feedback Equalizer (**DFE**)
- → Maximum Likelihood Sequence Estimation (**MLSE**)

I. Decision Feedback Equalization

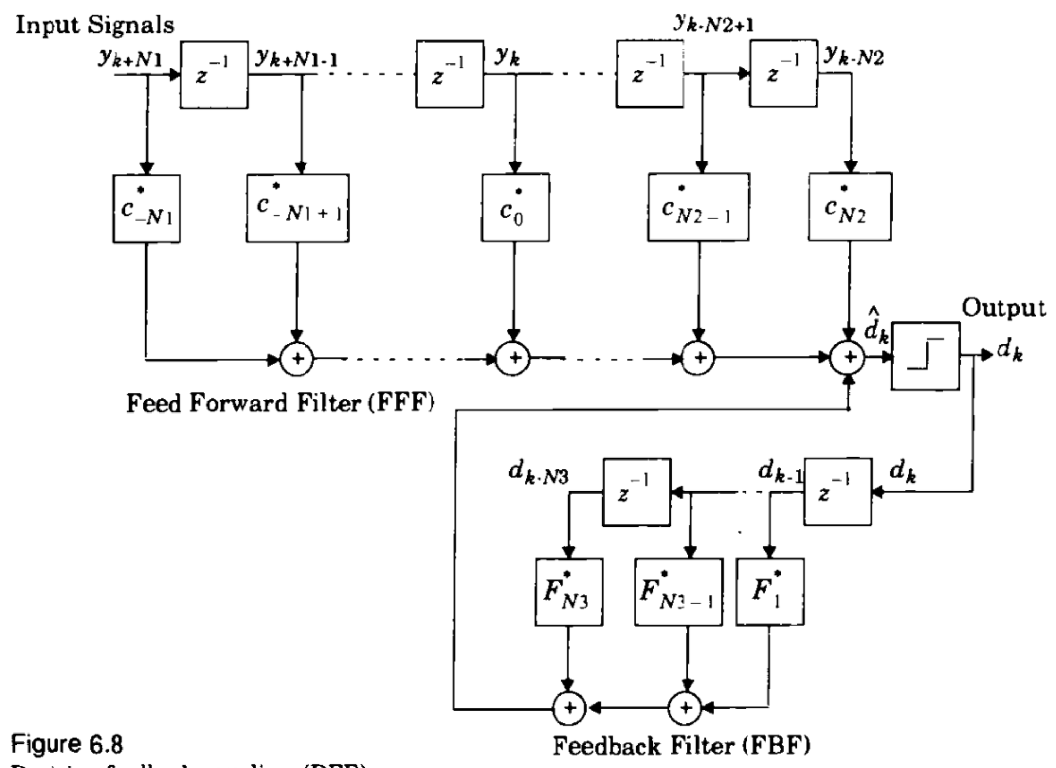

Figure 6.8
Decision feedback equalizer (DFE).

- i) The **basic idea** is that, once an information **symbol** has been **detected**, the **ISI** that is **induces** on **future** symbols can be **estimated** and **subtracted** out.
- ii) DFE can be realized in either the direct transversal (LTE) form or as a lattice filter

I. Decision Feedback Equalization (cont ...)

- iii) The LTE form consists of a **feedforward filter** (FFF) and a **feedback filter** (FBF)
- → The **FBF** is driven by **decisions** at the output of the detector, and its **coefficients** are **adjusted** to **cancel the ISI** on the **current** symbol from **past** detected symbols.
- → The equalizer has $N_1 + N_2 + 1$ taps in FFF and N_3 taps in the FBF
- → Output is expressed as $\hat{d}_k = \sum_{n=-N_1}^{N_2} c_n^* \, y_{k-n} + \sum_{i=1}^{N_3} F_i \, d_{k-i}$, where c_n are tap gains and y_n are inputs in FFF F_i are tap gains and $d_i \; (i < k)$ are inputs to FBF
- → The **minimum mean square error** of DFE is
$$E\left\{|e(n)^2|\right\}_{min} = \exp\left\{\frac{T}{2\pi}\int_{-\frac{\pi}{T}}^{\frac{\pi}{T}} \ln\left[\frac{N_0}{|F(e^{j\omega T})|^2 + N_0}\right] d\omega\right\}$$
- → DFE has **smaller min MSE** than **LTE** when there are **nulls** in $|F(e^{j\omega T})|$
- iv) **LTE** is **well behaved** when the channel **spectrum** is **flat**
- → **DFE** is **better** for **severely distorted** wireless channels
- → Further, the **LTE** has **difficulty** in **equalizing** a channel when the **strongest energy** arrives **after** the **first** arriving signal component

II. Maximum Likelihood Sequence Estimation (MLSE) Equalizer

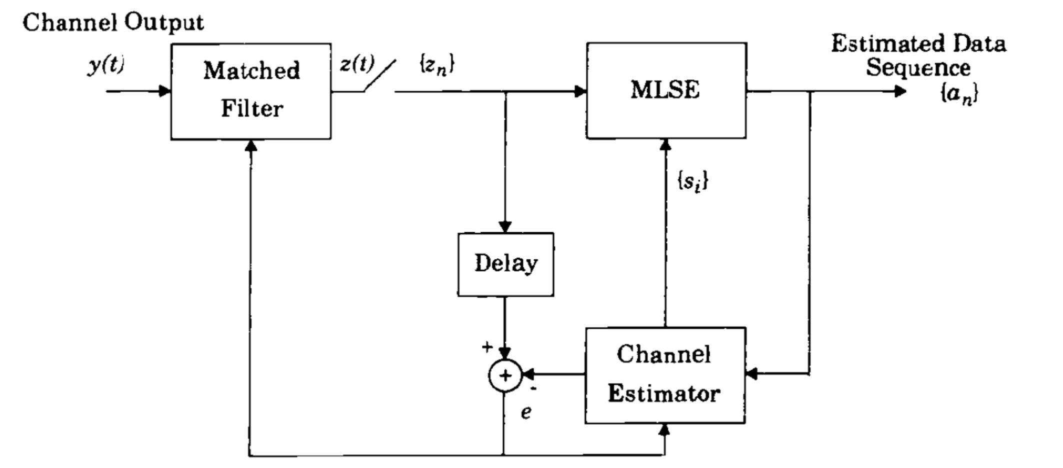

Figure 6.10
The structure of a maximum likelihood sequence estimator (MLSE) with an adaptive matched filter.

- i) The **MSE-based linear** equalizers are **optimum** with respect to minimum probability of **symbol error** only when the channel **does not** introduce any **amplitude distortion**
- → **MLSE** uses **maximum likelihood** receiver structure to over-come this **limitation** under amplitude distortion
- ii) Using a channel **impluse response simulator** within the algorithm, the **MLSE** tests all possible **data sequences** (**rather than** decoding each received **symbol individually**) and chooses the **data sequence** with the **maximum** probability as the output using **Viterbi** algorithm
- → As a result MLSE requires **huge computations**

II. Maximum Likelihood Sequence Estimation (MLSE) Equalizer (cont ...)

- iii) The MLSE can be viewed as a problem in **estimating** the state of a discrete time **finite state machine** with M^L states
- → where M is the size of the **symbol alphabet** of the modulation
- → and L is the number of most recent **input samples** based on which the channel **state is estimated**
- iv) An M^L **Trellis** is used by the receiver to model the channel over time
- → The **Viterbi** algorithm then **tracks** the state of the channel by the paths through the trellis
- → The MLSE is **optimal** in the sense that it **minimizes** the probability of a **sequence error**
- v) The MLSE **requires knowledge** of the channel **characteristics** in order to compute the **metrics** for making decisions
- → It also **requires knowledge** of the **statistical distribution** of the **noise** corrupting the signal
- → The matched filter operates on continuous time signal where as MLSE and channel estimator relay on discretized samples

Algorithms for Adaptive Equalization

- Zero Forcing Algorithm
- Least Mean Square Algorithm
- Recursive Least Squares Algorithm

Diversity Techniques

- Derivation of Selection Diversity Improvement
- Derivation of Maximal Ratio Combining Improvement
- Practical Space Diversity Considerations
  - Selection Diversity
  - Feedback or Scanning Diversity
  - Maximal Ratio Combining
  - Equal Gain Combining
- Polarization Diversity
- Frequency Diversity
- Time Diversity

RAKE Receiver

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Networks
Introduction
Advantages
Topologies

Wireless Networks I
Introduction

i) Wireless local area networks (**WLANs**) **transfer data** between **devices** like mobile PC workstations over **radio waves**
   → The key **advantages** of WLAN are that it **eliminates** the laying of **cables** and wiring **costs**
   → They use the industrial scientific and medical (**ISM**) frequency bands 900 MHz, 2.4 GHz and 5 GHz. WLAN products **do not require** a **license** to operate in this band
ii) Wireless networks are standardized by **IEEE**
   → Under IEEE 802 committe, the **wireless LAN** and metropolitan area network (**MAN**) standards are developed
iii) The first standard was created by IEEE in 1997 and was named **802.11**. It uses 2.4 GHz frequency and supported maximum bandwidth of **2 Mbps**

Wireless Networks II
Introduction

→ Wireless fidelity (**Wi-fi**) is a generic term used to refer IEEE 802.11 standard for WLANs.
→ Wi-fi is used to connect **computers** to **communicate** with each other, with the wired network and the Internet

iv) IEEE **802.11b** that uses the same frequency 2.4 GHz was created in July 1999 and supported **11 Mbps**
v) An updated version of the original 802.11 standard was created and called **802.11a**.
   → Due to its higher cost, 802.11a is usually found on **business** networks whereas 802.11b better serves the **home** market
vi) **802.11g** was created to **combine** the best of both 802.11a and 802.11b and supports network bandwidth upto **54 Mbps**
vii) **802.11n** is the newest IEEE standard in the 802.11 family

Wireless Networks III
Introduction

→ It **improves** on **802.11g** in the amount of **bandwidth** by using multiple antennas instead of one, i.e., **MIMO**
→ It also operates on the 2.4 GHz band

viii) The IEEE 802.11 standard is **related** to the medium acces control (**MAC**) layer and the physical (**PHY**) layer.
   → It contains **one** standard for MAC layer and **multiple** standards for the PHY layer
   → It defines several **modulation** methods involving
      infrared,
      direct-sequence spread spectrum (**DSSS**),
      frequency-hopping spread spectrum (**FHSS**),
      orthogonal frequency division multiplexing (**OFDM**)
   → and three different PHY layer technologies: IEEE 802.11a, 802.11b and 802.11g

Advantages and disadvantages of WLANs I

Advantages:

i) **Mobility**: Wireless LANs support mobility. This **improves** the real-time **access** to **information** even when the user is moving from one place to another within the range of an access point (AP)
ii) **Different Topologies:** Wireless networks are configured in two modes, the **ad-hoc** mode and the **infrastructure** mode
   → Ad-hoc mode provides **peer-to-peer** communication between wireless devices
   → Infrastructure mode provides communication **between** wireless **device** and a **central node** that can communicate with the wired nodes on that LAN. It is time consuming and expensive to run cables between computers
iii) **Flexible architecture:** It is **easier** to **add** or **remove** workstations

Advantages and disadvantages of WLANs II

iv) **Cost effective:** Although the initial investment for WLAN is similar to the **cost** of wired LAN, further **expansion** expenses can be significantly **lower**

Disadvantages:

i) **Less security:** Wireless LANs are **less secure** than wired LANs as the **signals** can easily be **intercepted** by others using the same frequency band
ii) **Low data rates:** The data transfer **rate decreases** with increase in the **number** of devices
iii) **Need for energy efficient devices:** In mobile devices, the **battery** power is a **scarce** resource. The devices must be designed to be **energy** efficient
iv) **Limited coverage:** Devices will **only** operate at a **limited distance** from an AP. The distance is determined by the standard used and the obstacles between the AP and the user

WLAN topologies

i) Wireless network **topology** is the **configuration** in which a mobile terminal (MT) **communicates** with another
ii) WLANs can be built with
   → the Peer-to-peer (**ad-hoc**) topology or
   → the **Infrastructure** topology

Peer-to-peer (ad-hoc) topology I
WLAN topologies

i) This topology **applies** to **reconfigurable** networks that can operate **without** need for a **fixed** infrastructure

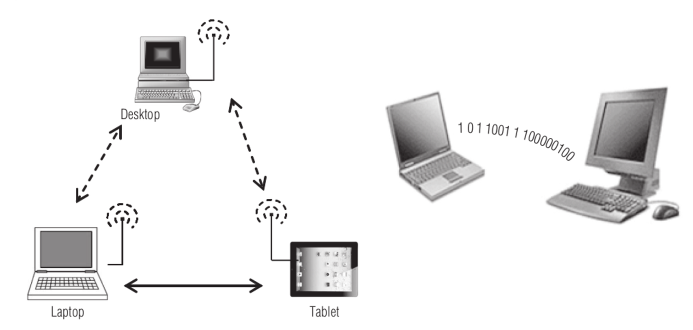

Figure 24.1 Ad-hoc network topology

→ This is the **easiest** WLAN mode to **configure** and requires the **least** hardware

Peer-to-peer (ad-hoc) topology II
WLAN topologies

→ An ad-hoc mode WLAN is comprised of **two or more** computers communicating directly with each other using **wireless** network **cards**
→ **Multi-hop** ad-hoc networks are the networks that are **distributed** over a **wide** area in which the user **terminals cooperate** in **relaying** the messages across the network

Infrastructure topology I
WLAN topologies

i) In this topology, there is a **fixed** (wired) **infrastructure** that **supports** communication between **MTs** and the **fixed terminals**

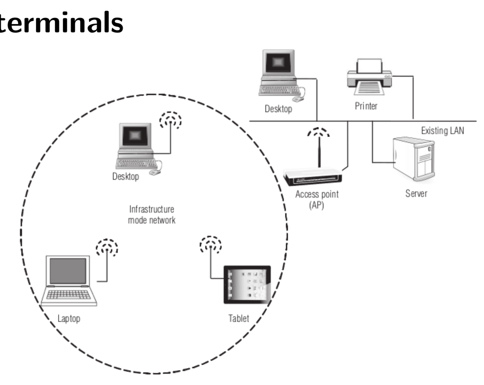

Figure 24.2 Infrastructure network topology

→ This topology is **designed** for **large coverage** areas and **multiple** base station or **AP** operations

Infrastructure topology II
WLAN topologies

ii) In this WLAN mode, a hardware or software **AP** is **configured** as part of the design
   → This AP **provides connectivity** for the systems on the WLAN
   The **wireless** network **card** on each computer is configured to use a specific **AP** and all traffic to other computers on the WLAN is directed **through** the AP
   → APs also **bridge** traffic onto a **wired** (Ethernet or token ring) or a wireless **backbone**
iii) **Access point:** The AP is a **wireless LAN transceiver** or base station that can **connect** one or many **wireless devices** simultaneously to the **internet**

Infrastructure topology III
WLAN topologies

→ The AP **coordinates** transmission and reception from **multiple** wireless devices within a **specific range**. The range and number of devices **depend** on the **standard** being used
→ **Multiple** APs can be used in a network to cover a large area

iv) All **standardized** cellular mobile telephone and wireless data systems use an **infrastructure** network **topology** to serve MTs withing the coverage area

Comparison of ad-hoc and infrastructure topologies I
WLAN topologies

i) **Flexibility**:
   → Operation of **infrastructure** networks **requires** deployement of a network infrastructure which is **time-consuming** and **expensive**
   → **Ad-hoc** networks are inherently **flexible** and can be set-up **instantly**. As a result, **ad-hoc** networks are always used for **temporary** applications where flexibility is important
ii) **Controllability**: To **coordinate** proper operation of a radio network, features like **time synchronization**, transmitted **power** need to be centrally **controlled**
   → In an **infra. network**, all these features are naturally **implemented** in the **AP**
   → An **ad-hoc** network **requires complicated** structures and changes in terminals to implement these features

Comparison of ad-hoc and infrastructure topologies II
WLAN topologies

iii) **Routing complexity**:
   → In a **multi-hop ad-hoc** network, each terminal should be able to route messages to other terminals. This requires **each** terminal to **monitor** the existence of **other** terminals. It requires a **routing algorithm**, that adds to the complexity of the terminals
   → This problem **does not** exist in **infra.** topology
iv) **Reliability**:
   → **Infr. networks** are **single failure point** networks. If the **AP fails**, the entire communication **network fails**
   → This problem **does not** exist in **ad-hoc** configuration

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Network

WLAN IEEE 802.11

WLAN Standard IEEE 802.11 I

Introduction

i) **IEEE 802.11** is an international **standard** that describes **WLAN characteristics**. The **WiFi** corresponds to the certification name issued by the Wi-Fi Allinace group
   - → This standard **ensures** the **compatibility** between hardware devices
   - → It provides **high-speed** connections to laptops, computers and PDAs located within a radius of 20-50 m **indoor** and 100 m **outdoor**

ii) The 802.11 is a specific standard that **defines** the **MAC** and **PHY** layers of a WLAN
   - → The **original 802.11** standard is a MAC plus a low data rate PHY standard which supports **1-2 Mbps** data rates
   - → It operates at the **2.4 GHz ISM** band and allows the vendors to choose between a DSSS and FHSS implementation

WLAN Standard IEEE 802.11 II

Introduction

iii) The **802.11b** is a **PHY extension** to the original 802.11. It also operate at the 2.4 GHz band and allows **higher** data rates of **5.5 and 11 Mbps**
   - → It uses a **technique** known as complementary code keying (**CCK**)

iv) The **802.11a** is another **PHY extension** to the 802.11 standard
   - → It operates at the **5 GHz** band and allows for data **rates** of **6-54 Mbps**
   - → It **uses** a **technique** known as orthogonal frequency-division multiplexing (**OFDM**)

v) The **802.11g** is the next extension of 802.11.

WLAN Standard IEEE 802.11 III

Introduction

- → It operates at the **2.4 GHz** band and supports data rates from **1-54 Mbps**. The 1 and 2 Mbps rates are operated in the DSSS mode whereas the 5.5 and 11 Mbps rates are operated in CCK mode. In addition, rates 6-54 Mbps are operated in OFDM
- → The 802.11g standard **borrows** the **OFDM** technique and data rates from the **802.11a** standard but operates at the **2.4 GHz ISM** band
- → It can therefore operate at **very high data rates** while being **backward compatible** with the **802.11b** standard

vi) All three versions share the same MAC layer that uses
   - → **carrier sense multiple access with collision avoidance** (**CSMA/CA**) for **contention data**,

WLAN Standard IEEE 802.11 IV

Introduction

- → a **request-to-send/clear-to-send** (**RTS/CTS**) mechanism to accommodate the **hidden terminal** problem and
- → an **optional** mechanism called **point coordination function** (PCF) to support **time-bounded** applications

vii) The 802.11 standard **supports both infrastructure** WLANs connection through an AP and **ad-hoc** operation allowing peer-to-peer communication between terminals

Challenges faced by WLAN

Introduction

i) Compared to wired LANs, WLANs operate in a **difficult medium** of communication, and they need to support mobility and security

ii) The wireless **medium** has serious **bandwidth limitations** and frequency **regulations**

iii) It suffers from time and location dependent **multi-path fading**

iv) It is subject to **interference** from **other WLANs**, radio and non-radio devices operating in the vicinity

v) The 802.11 body had to examine **connection** management, link **reliability** management and **power** management

vi) In addition, WLANs have no **PHY boundaries** and they **overlap** with each other, as a result need to provide provisions for **security** of the links

Comparison of IEEE 802.11a,b,g and n standards

The table gives **comparison** in terms of frequency, bandwidth, data rate, modulation and range

Table 24.2 Comparison of 802.11 protocols

802.11 protocol	Freq. (GHz)	Bandwidth (MHz)	Data rate per stream (Mbps)	Allowable MIMO streams	Modulation	Approx. indoor range (m)	Approx. outdoor range (m)
802.11	2.4	20	1,2	1	DSSS, FHSS	20	100
a	5	20	6, 9, 12, 18, 24, 36, 48, 54	1	OFDM	35	120
b	2.4	20	5.5, 11	1	DSSS	38	140
g	2.4	20	6, 9, 12, 18, 24, 36, 48, 54	1	OFDM, DSSS	38	140
n	2.4	20	7.2, 14.4, 21.7, 28.9, 43.3, 57.8, 65, 72.2	4	OFDM	70	250

IEEE 802.11 Physical layer

i) The **PHY** layer is the **interface** between the **MAC** and **wireless media** that provides transmission and reception of data frames over a **shared** wireless medium

ii) The IEEE 802.11 standard defines **different** transmission **techniques** with different PHY implementations
   - → Ifrared (**IR**)
   - → Direct-sequence spread spectrum (**DSSS**)
   - → Frequency-hopping spread spectrum (**FHSS**)
   - → Orthogonal frequency-division multiplexing (**OFDM**)
   - → Narrow band wicrowave LANs

IEEE 802.11 Medium Access Control I

i) The MAC is a **sublayer** of the **data link layer** in the OSI model

ii) It provides **addressing** and **channel access control** mechanisms that make it possible for **several** terminals to communicate
   - → **Each** node in an 802.11 network is **identified** by its **MAC** address

iii) The **functions** of MAC layer are as follows
   - → It provides a **reliable data delivery** service to the users of the MAC over wireless media through a **frame exchange protocol**

IEEE 802.11 Medium Access Control II

- → It **controls** the **access** to the **shared wireless medium** through two different access mechanisms
  a) The **basic** access mechanism called the **distributed coordination function** (DCF)
  b) The **centrally controlled** access mechanism called the **point coordination function** (PCF)
- → It **protects** the data delivery by providing a **privacy service** called **wireless equivalent privacy** (WEP), which **encrypts** the data

iv) *Study the reference material for further details related to MAC*

Wireless metropolitan area networks (WMAN) I

Introduction

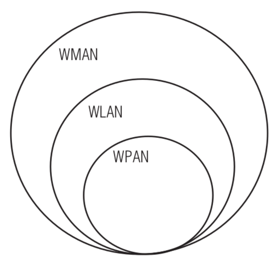

Figure 25.7 Wireless networks

i) **Need** for Wireless MAN (**WiMax**)
   - → **WLANs** and **WPANs restrict** the **mobility** of users to a few hundreds of meters from the source of the RF signal
   - → Also users are needed to stay within in LOS. As a result it is **confined** to **offices**, **homes** and **hotspots**

Wireless metropolitan area networks (WMAN) II

Introduction

- → **IEEE 802.16** and WiMAX (**worldwide interoperability for microwave access**) are designed as a **complimentary** technology to **Wi-fi** and **Bluetooth**
- → WMANs are a group of **technologies** that provide wireless connectivity across a **large geographical** area such as a large **metropolitan city**
- → These networks provide access of wired network **beyond** a **single location**

Table 25.2 Features of wireless networks

Parameter	WPAN	WLAN	WMAN
Protocol	802.15	802.11	802.16
Standards	Bluetooth, IrDA, UWD	Wi-fi	Wi-Max
Frequency range	2.4 to 2.483 Ghz	5.15 to 5.35Ghz	10–66 Ghz
Speed	1–4 Mbps	1–54 Mbps	2–70 Mbps
Cell radius	1–10 ms	1–500 ms	1–50km
Modulation	FHSS	OFDM, DSSS	QPSK

WMAN IEEE 802.16

IEEE 802.16 and its enhancements I

i) **IEEE 802.16** protocol of WiMax standard **focus** on transmitting **high data rates** for **large distances** and also maintain effective **QOS** and **security**

ii) IEEE 802.16 mainly **focuses** on the **PHY** and **Data Link layer** in OSI model

iii) Here, Physical Layer (PHY) can be **single-carrier** or **multi-carrier**

iv) The data link layer is divided into a) **Logical link control** (LLC) and b) **Medium access control** (MAC)

v) **MAC** is further divided into sub-layers
   - → **Convergence sub-layer** (CS)
   - → **Common part sub-layer** (CPS)
   - → **Security sub-layer** (SS)

IEEE 802.16 Physical Layer I

i) **PHY** sets the **connection** between the communicating **devices** and is **responsible** for transmitting the **bit sequence**

ii) It also **defines** the type of **modulation** and **demodulation** as well as the transmission **power**

iii) It considers two types of transmission techniques: **OFDM** and **OFDMA**

iv) **OFDM PHY:**
   - → The **OFDM** is developed to support **high data rate** and can handle **multi-carrier** signals
   - → It can **minimize** the **inter symbol interference** (ISI) much more compare to other multiplexing schemes
   - → The OFDM can also **handle multi-path effect** by converting serial data to several parallel data using **fast Fourier transform** (FFT) and **IFFT**
   - → For **fixed** WiMAX, the **FFT size** is fixed at **256**

IEEE 802.16 Physical Layer II

- → Whereas for **mobile** WiMAX, the **FFT size** can be 128, 512, 1024 and 2048 bits. This helps to combat **ISI** and **Doppler** spread
- → OFDM **splits** a single **high bit rate data** into several **low bit rate** of data sub-stream in parallel

v) **OFDMA PHY:**
   - → **OFDMA** which is also called **multi-user-OFDM** is **designed** for **fourth generation** wireless networks
   - → It is a **combination** of **FDMA**, **TDMA** and code division multiple access (**CDMA**), as it performs the **same function** like these **access methods**
   - → **OFDMA** can be seen as an **altertive** to **CDMA** where **each** user gets **different number** of **spreading code** with **different data rates**

IEEE 802.16 Physical Layer III

- → It **resembles** an alternative **TDMA** as **low data rate** users can **send** data with **low transmission power**, with constant and shorter delay
- → It can also be seen as a **combination** of **TDMA and FDMA** where the **resources** are **divided** according to **time-frequency spaces** and slots along with OFDM sub-carrier index
- → **Different number** of **sub-carrier** can be **allotted** to different number of **users** to maintain the **data rate** and **error** probability for **each** user
- → As result, **OFDMA** is an **efficient** access method for **multi-user** environment

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Network

WPAN IEEE 802.15

HiperLAN

WLL

WPAN Standard IEEE 802.15 I

i) A **wireless personal area network** (WPAN) is a **network** for **interconnecting devices** centred around an **individual person's workspace**, in which the connections are wireless
   → These might include a mobile **phone**, a laptop computer, pagers, PDAs and a **personal stereo**

ii) Typically, a **WPAN** uses **Bluetooth technology**, which was used as the basis for a new standard, **IEEE 802.15**
   → **Bluetooth** technology has been adopted as the **IEEE 802.15.1** WPAN standard
   → It permits communication within about **10 m**, which enables the use of **low power**, **low cost** and **extremely small-sized** devices

iii) The **concept** of PANs is that if **each** of the **devices** had a **short-range communications** tool built into them, they could **exchange information without wires** and **without** any **intervention** from the **user**

WPAN Standard IEEE 802.15 II

iv) The **key concept** in WPAN technology is known as **plugging in**. When any **two WPAN-equiped** devices come into **close proximity**, they can **communicate** as if connected by a cable

v) Another **important feature** is the ability of each device to **lock out** other devices **selectively**, **preventing** needless interference or **unauthoried access** to information

vi) The **IEEE 802.15** standards is a **family of protocols** to address the needs of **WPAN** at different data rates in **2.4 GHz ISM** band
   → Three IEEE 802.15 protocols (**IEEE 802.15.1, IEEE 802.15.3 and IEEE 802.15.4**) are developed **based** on **data rates**, **technology**, frequency band, channel **access** scheme and modulation used
   → They are referred to as **Bluetooth**, **high-rate WPAN** and **low-rate WPAN**

vii) **Applications**:

WPAN Standard IEEE 802.15 III

→ WPANs are **used to replace cables** between a **computer** and its **peripheral devices**

→ WPANs can be used for **transmitting images**, digitized **music** and other data

→ WPAN, **popularly** known as **Bluetooth technology** is a **short-range** wireless network formed around the **personal operating space** of a wireless terminal with built-in Bluetooth device

→ With the help of Bluetooth technology, **ad-hoc wireless pico-nets** can be formed, which are LANs with **very limited coverage** (about 10 m) and **without** the need for an **infrastructure**, offering asynchronous data and synchronous voice services at **1 Mbps**

→ **Bluetooth** also provides a **universal bridge** to **existing data networks** and a mechanism to form **small private mobile ad-hoc** networks (**MANETs**)

Hiper LAN

i) The **European** Telecommunications Standards Institute (ETSI) has come up with **HIgh Performance Radio LAN** (HIPER LAN), which is an **alternative** for the **IEEE 802.11 WLAN** standards

ii) The HiperLAN standards provide **features** and **capabilities similar** to those of the **IEEE 802.11**

iii) In HyperLAN, there are a **number** of base stations (or **APs**), and the devices **can communicate either** with the **base station** or **directly** with each other

iv) The **APs** can **automatically configure** their **frequency** so that there is **no need** for **manual** frequency **assignment**

v) The HiperLAN standard **family** has four different versions: **HiperLAN/1**, **HiperLAN/2**, HiperACCESS and HiperLINK

Comparison of WLAN technologies

Table 24.4 A Comparison between different WLAN technologies

Characteristic	802.11	802.11b	802.11a	HiperLAN2
Spectrum	2.4 GHz	2.4 GHz	5 GHz	5 GHz
Max PHY data rate	2 Mbps	11 Mbps	54 Mbps	54 Mbps
Max user data rate	1.2 Mbps	5 Mbps	32 Mbps	32 Mbps
MAC	CSMA/CA	CSMA/CA	To be determined	TDD/TDMA
Connection	Connectionless	Connectionless	Connectionless	Connection oriented
Frequency selection	Frequency hopped (FH) or direct sequence spread spectrum (DSSS)	DSSS	OFDM	OFDM with dynamic frequency selection

HiperLAN/1 I

i) **HiperLAN/1** is mainly **designed** to work **without** the need of any **infrastructure**
   → Two nodes may **exchange** data **directly**, **without** any interaction from a wired (or radio-based) infrastructure
   → The **simplest** HiperLAN/1 consists of **two nodes**
   → If two HiperLAN/1 nodes are not in radio contact with each other, they may use a **third node** which must **forward messages** between the two communicating nodes

ii) A **multi-hub topology** is considered to allow **overlay** of **two HiperLANs** to **extend** the communication beyond the radio **range** of a single node

HiperLAN/1 II

→ There are two **overlapping** HiperLANs A and B, and the **node 4** acts as a **bridge** between the two

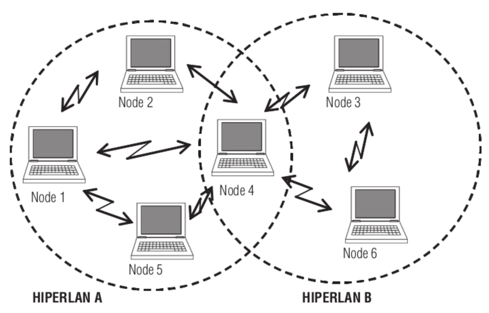

Figure 24.19 HIPERLAN/1 ad-hoc system architecture

→ **Each** node is **designated** either as a **forwarder** node or **non-forwarder** node

→ Each **non-forwarder** node should **select** at least one of its **neighbour** nodes as a **forwarder** node

HiperLAN/2 I

i) HiperLAN/2 has a **very high-transmission** rate upto **54 Mbps**
   → This is **acheived** by making use of **Orthogonal Frequency Division Multiplexing** (OFDM)

ii) HiperLAN/2 connections are **time-division multiplexed** and connection-oriented, either **bidirectional** point-to-point or **unidirectional** point-to-multipoint
   → There is also a dedicated **broadcast channel** through which the traffic from an AP reaches all terminals

iii) The HiperLAN/2 APs have a **built-in** support for **automatic** transmission **frequency allocation** within the APs coverage area

iv) Additionally, this network supports **authentication** and **encryption**

HiperLAN/2 II

→ Both the AP and the MT can **authenticate** each other to ensure authorized access to the network or to a valid network operator

→ Each node is given a **HiperLAN ID** (HID) and a **node ID** (NID).

→ The combination of these two IDs **uniquely** identify a station and **restricts** the connections to other HiperLAN nodes

→ All nodes with the **same HID** can **communicate** with each other using **dynamic routing** mechanism denoted **intra-HiperLAN** forwarding

v) The support for **handover** enables **mobility** of MTs
   → The **handover** scheme is **MT initiated**. The MT uses the AP with the **best signal** as the user moves around. All established **connections move** to the AP with the **best** radio transmission performance

HiperLAN/2 III

vi) The HiperLAN/2 architecture is easily adapted and integrated with a variety of fixed networks

vii) The **power save** mechanism in HiperLAN/2 is based on **MT-initiated** negotiation of **sleep** periods
   → The MT **requests** the AP for a **low power state** and a specific **sleep** period
   → The MT receives any **pending** data as the **sleep period expires**

Wireless Local Loop (WLL) I

i) Traditionally, the voice and data communication services to the end subscribers over the **local loop** has been provided by **wired systems**
   → For **residential** subscribers, **twisted pair** cable has been the standard means of connection
   → For **business** and government subscribers, **twisted** pair cable, **coaxial** cable and **optical** fiber cable are in use

ii) **WLL** is a system that **privdes** a **wireless** connection between **subscribers** and the **local telephone station**. The other names of WLL are **radio in the loop** (RITL) or **fixed-radio access** (FRA)
   → WLL is the **use of radio** to **provide** a **telephone connection** to the home

Wireless Local Loop (WLL) II

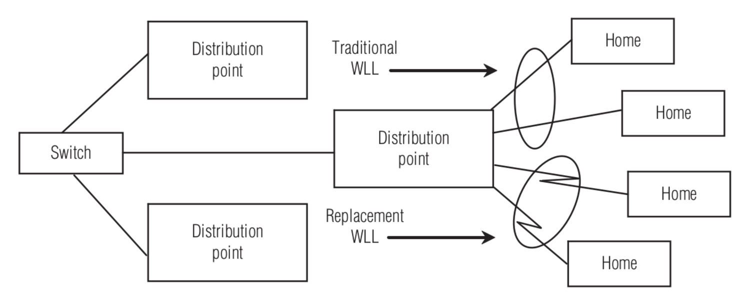

Figure 27.7 Role of WLL

iii) Historically, the local loop was **copper** cable **buried** in the ground
   → WLL **replaces** the **local loop** section with a **radio path** rather than a **copper** cable
   → It is **concerned** only with the **connection** from the **distribution** point to the **house**, all the other parts of the network are left **unaffected**

Wireless Local Loop (WLL) III

→ The **distribution point** is connected to a **radio transmitter** and a **radio receiver** is mounted on the side of the **house**. A cable is run down inside the house to a **socket** to which the home telephone is connected

iv) Modern WLL systems use **CDMA technology**
   → A WLL service provider serves **one or more cells**
   → Each cell includes a **base-station antenna** mounted on top of a tall building or tower
   → Individual **subscribers** have a **fixed antenna mounted** on a building or a high pole that has a **LOS** to the base-station
   → From the **base station**, there is a wired or wireless link to a **switching center**

v) **Advantages** of WLL over wired approach
   → WLL systems are **less expensive** than wired systems, as a result new requirements are met with WLL approach

Wireless Local Loop (WLL) IV

→ WLL systems can be **installed rapidly**

→ With a **wired** system, a **cable** is **laid** out in **anticipation** of serving every **potential** subscriber in a local area, whereas the subscriber **radio** units are **installed** only when the **subscriber** has **registered** for a service

→ A large geographical **area** is **still** not covered with **landline** telephone service

→ A **major advantage** of WLL over the cellular mobile system is that the **fixed subscriber** can use a **directional antenna pointed** at the base-station antenna, providing **improved** signal **quality**

vi) WLL has been allocated a frequency band of **2-40 GHz**, especially the unused frequency bands available above 25 GHz

vii) WLL **applications** include **local multipoint distribution service** (LMDS) and **multichannel multipoint distribution service** (MMDS)

Wireless Local Loop (WLL) V

→ **LMDS** is a relatively new WLL **service** capable of **providing video**, **telephony** and **high-data rate** of the order of several Mbps within the short range from the base station

→ LMDS is **useful** for **organization** requiring larger bandwidth

→ **MMDS** operate at **lower band of millimeter** range and can operate in considerably **larger cells** within a radius of **50 km**. MMDS is likely to be used by residential subscribers and small business

Comparison of WLL, mobile and wireline technologies

Table 27.5 Comparisons of WLL, mobile, wireless and wireline technologies

WLL	Mobile wireless	Wireline
Good LOS component	Mainly diffuse components	No diffuse components
Rician fading	Rayleigh fading	No fading
Narrowbeam directed antennas	Omnidirectional antennas	Expensive wires
High channel reuse	Less channel reuse	Reuse limited by wiring
Simple design, constant channel	Expensive DSPs, power control	Expensive to build and maintain
Low in-premises mobility only, easy access	High mobility allowed, easy access	Low in-premises mobility, wiring of distant areas is cumbersome
Weather conditions effects	Not very reliable	Very reliable

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