Part II: theory

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Contents

A note on mean values, standard deviations and dB's

FIXME Explain the differences between $\mathrm{E}[P_{dBm}]$ (or $\overline{P_{dBm}}$) and $\mathrm{E}_{dBm}[P_{mW}]$ (or $\overline{P}_{dBm}$); same for standard deviation: $10\log(\sigma(P_{mW}))$ and $\sigma(10\log(P_{mW}))$, and why the first one is not used (and therefore, why we also have to use $E[P_{dBm}]$).

Chapter 1

Electromagnetic waves

In this chapter, we will give a brief overview of electromagnetic waves. It is not intended as a complete, thorough and 100% correct analyses of the Maxwell's equations; it's main purpose is to make the understanding of Maxwell's equations a bit clearer and to derive the relation between the power of an electromagnetic wave and the amplitude of it's electrical field. This chapter is largely based on [Smo00] and [Kra53].

1.1 Maxwell's equations

The fundamental equations which descibe the propagation of electromagnetic waves, are the Maxwell's equations, which are given in time domain by (1.1).

$$\nabla \times \vec{\mathcal{E}}(\vec{r},t) = -\frac{\partial \vec{\mathcal{B}}(\vec{r},t)}{\partial t}, \nabla \times \vec{\mathcal{H}}(\vec{r},t) = \frac{\partial \vec{\mathcal{D}}(\vec{r},t)}{\partial t} + \vec{\mathcal{J}}_e(\vec{r},t), \nabla \cdot \vec{\mathcal{D}}(\vec{r},t) = \rho_e(\vec{r},t). \nabla \cdot \vec{\mathcal{B}}(\vec{r},t) = 0, \quad (1.1)$$

In these equations, $\vec{\mathcal{E}}$ is the electrical field in Volts per meter, $\vec{\mathcal{D}}$ is the electric flux density in Coulomb per m^2, $\vec{\mathcal{B}}$ is the magnetic induction in Weber per m^2, $\vec{\mathcal{H}}$ is the magnetic field in Ampere per meter, ρ_e is the electrical charge density in Coulomb per m^3 and $\vec{\mathcal{J}}_e$ is the current density of the conduction current in Ampere per m^2 (in a nonconducting medium, $\vec{\mathcal{J}}_e$ is zero).

Since the electromagnetic waves always propagate through air from the transmitter to the receiver, we can assume the following (for propagation in free space):

$$\vec{\mathcal{D}}(\vec{r},t) = \epsilon_0 \vec{\mathcal{E}}(\vec{r},t), \vec{\mathcal{B}}(\vec{r},t) = \mu_0 \vec{\mathcal{H}}(\vec{r},t). \quad (1.2)$$

(Please note that from here on, the $\vec{r}$ dependency is ommitted, unless in those places where there leaving it out can cause some confusion.)

The first equation of (1.1) can be read as: "a changing magnetic field ($\frac{\partial \mu_0 \vec{\mathcal{H}}(\vec{r},t)}{\partial t}$) will cause an electrical field ($\vec{\mathcal{E}}(\vec{r},t)$)". Similarly, the second equation can be read as: "both a displacement current ($\frac{\partial \epsilon_0 \vec{\mathcal{E}}(\vec{r},t)}{\partial t}$) and a conduction current ($\vec{\mathcal{J}}_e(\vec{r},t)$) cause a magnetic field ($\vec{\mathcal{H}}(\vec{r},t)$)". This is essentially what makes a wave propagate through a medium: by sending a harmonic current $\vec{\mathcal{J}}_e$ through an antenna, we induce a harmonic magnetic field $\vec{\mathcal{H}}$. Since this harmonic magnetic field is (by the definition of harmonic) a field which is changing in time, it will induce a harmonic electrical field $\vec{\mathcal{E}}$. This harmonic electrical field in turn will induce a harmonic magnetic field, which will induce a harmonic electrical field etc.

The thirth and fourth equation of (1.1) are easier to read in integral form (especially when one considers 1.2):

$$\oint_s \vec{\mathcal{D}} \cdot d\vec{s} = \int_v \rho_e \; dv, \oint_s \vec{\mathcal{B}} \cdot d\vec{s} = 0. \tag{1.3}$$

where ρ_e is the charge density enclosed in volume v which has a surface s. The second equation of 1.3 can be thought of as the Kirchoff current law and the first equation can be thought of as the Kirchoff voltage law. I will explain this a little further. The Kirchoff current law states that for every node in an electrical circuit, the sum of all currents going into the node equals the sum of all currents going out of that node. If the currents flowing into the node are taken to be negative, and the currents flowing out of the node positive, this results in the formula $\sum i = 0$. Now, for simplicy, think the magnetical field $\vec{\mathcal{H}}$ as the equivalent of current i and the node inside volume v with surface s. Kirchoff current law for magnetic fields would then say that the integral (sum) over the whole surface of the magnetic field going through that surface (currents going through the brances) should be 0. This is also the magnetic counterpart of Gauss's law for the electric field.

The first equation of 1.3 can be tought of as the Kirchoff voltage law, which says that the sum of all voltages in a closed loop circuit must be 0. Since Kirchoff's voltage law works with voltage, which is the difference of 2 potentials, the absolute potential is not known, and thus the total net charge in the circuit is not known. Kirchoff's law for the electrical field would say that the integral (summation) of the voltages (electric field) over the whole surface (loop) should be equal to 0. This is true for voltages, but since electrical fields have an absolute value, we have to take the charge inside the volume into account. Thus, Kirchoff's law for the electrical field would be that the integral of the electric field over the whole surface should be equal to charge enclosed in that volume (conform Gauss's law for the electric field: *The surface of the normal component of the electric flux density $\vec{D}$ over any closed surface equals the charce enclosed.*)

Since the information that the transmitter wants to send to the receiver is usually modulated on a carrier, the bandwith of the signal compared to the carrier frequency is usually small. Therefore, we can approximate the time dependency of all components by the time dependency of the carrier, ie $e^{j\omega t}$. Then we can write:

$$\vec{\mathcal{E}}(\vec{r}, t) = Re\left[\vec{\mathcal{E}}(\vec{r})e^{j\omega t}\right]$$

A similar relation goes for the other components from 1.1. Therefore, the Maxwell's equations in frequency domain will become 1.4.

$$\nabla \times \vec{E}(\vec{r}) = -j\omega\mu_0 \vec{H}((\vec{r})), \nabla \times \vec{H}((\vec{r})) = j\omega\epsilon_0 \vec{E}(\vec{r}) + \vec{J_e}(\vec{r}), \nabla \cdot \vec{H}(\vec{r}) = 0, \nabla \cdot \vec{E}((\vec{r})) = \frac{\rho_e(\vec{r})}{\epsilon_0}. \tag{1.4}$$

1.2 Impedance of dielectric media

In general, in a 3-dimensional space, $\vec{E}$ and $\vec{H}$ have an x, y and z component. However, in a plane wave, $\vec{E}$ and $\vec{H}$ are perpendicular to each other and to the direction of propagation. For example, consider a vertically polarized plane wave[1] where

$$\vec{E} = \vec{e_x}E_x + \vec{e_y}E_y + \vec{e_z}E_z = \vec{e_x} \cdot 0 + \vec{e_y}E_y + \vec{e_z} \cdot 0 \vec{H} = \vec{e_x}H_x + \vec{e_y}H_y + \vec{e_z}H_z = \vec{e_x} \cdot 0 + \vec{e_y}0 + \vec{e_z} \cdot H_z \tag{1.5}$$

[1] In a vertically polarized plane wave, the electric field $\vec{E}$ only has a component E_y in the y direction, thus $E_x = 0$, $E_z = 0$ and $\frac{\partial E_y}{\partial z} = 0$. Similarly, $\vec{H}$ only has a component H_z in the z direction, therefore $H_x = 0$, $H_y = 0$ and $\frac{\partial H_z}{\partial y} = 0$

This is a plane wave which travels in the x direction (which is also the direction in which the energy flows); see also figure 1.1. We can usually asume that the electromagnetic field of a transmitter has a plane wave if we consider a distance large enough from the source (the antenna) and if the wave is horizontally of vertically polarized.

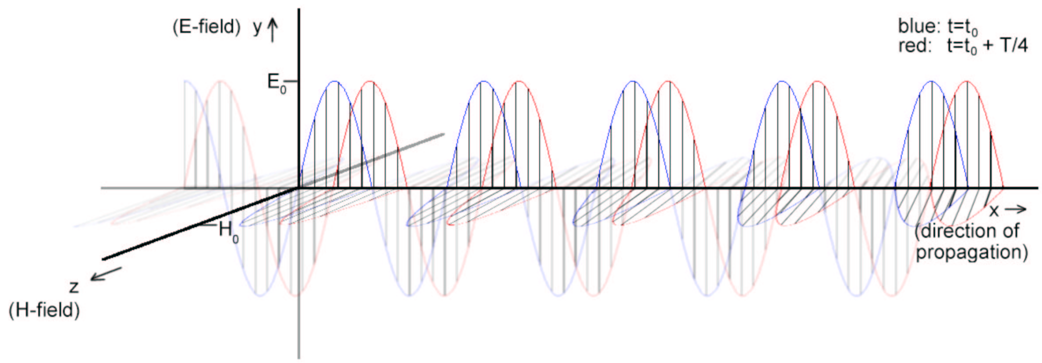


Figure 1.1: A plane electromagnetic wave at two points in time

Since air is a nonconducting medium, $\vec{J_e}$ is zero and we can write

$$\nabla \times \vec{H} = \frac{\partial \vec{D}}{\partial t}$$

as

$$\vec{e_x}\left(\frac{\partial H_z}{\partial y} - \frac{\partial H_y}{\partial z}\right) - \vec{e_y}\left(\frac{\partial H_z}{\partial x} - \frac{\partial H_x}{\partial z}\right) + \vec{e_z}\left(\frac{\partial H_y}{\partial x} - \frac{\partial H_x}{\partial y}\right) = \frac{\partial}{\partial t}(\vec{e_x}D_x + \vec{e_y}D_y + \vec{e_x}D_z)$$

With 1.5, we can reduce this to

$$-\vec{e_y}\frac{\partial H_z}{\partial x} = \vec{e_y}\frac{\partial D_y}{\partial t} = \epsilon\frac{\partial E_y}{\partial t} \tag{1.6}$$

Following the same steps for $\nabla \times \vec{E} = -\frac{\partial \vec{B}}{\partial t}$ and we get

$$\frac{\partial E_y}{\partial x} = -\mu\frac{\partial H_z}{\partial t} \tag{1.7}$$

Differentiating 1.6 with respect to time and 1.7 with respect to x and combining the result, we come to the wave equation for the electrical field:

$$\frac{\partial^2 E_y}{\partial t^2} = \frac{1}{\mu\epsilon}\frac{\partial^2 E_y}{\partial x^2} \tag{1.8}$$

Similarly, we can derive the wave equation for the magnetic field:

$$\frac{\partial^2 H_z}{\partial t^2} = \frac{1}{\mu\epsilon}\frac{\partial^2 H_z}{\partial x^2} \tag{1.9}$$

($v = \frac{1}{\sqrt{\mu\epsilon}}$ is the speed at which the wave propagates; for vacuum: $v = \frac{1}{\sqrt{\mu_o\epsilon_o}}$, which is the speed of light, c.)

It can be shown (see [Kra53]) that

$$E_y = E_0 \sin(\omega t - \beta x) \tag{1.10}$$

and

$$H_z = H_0 \sin(\omega t - \beta x) \tag{1.11}$$

are solutions to those wave equations (where E_0 and H_0 are the amplitudes of the electric resp. the magnetic field and $\beta = \omega/c = 2\pi/\lambda$). Equation 1.7 relates H_z to E_y. Differentiating 1.10 with respect to x, substituting the result in 1.7 and integrating that result with respect to time, we find H_z:

$$H_z = \frac{\beta}{\omega\mu} E_0 \sin(\omega t - \beta x) \tag{1.12}$$

Combining this equation with 1.10, we find:

$$E_y = \frac{\mu\omega}{\beta} H_z = \frac{\mu \cdot \frac{2\pi}{\lambda\sqrt{\mu\epsilon}}}{\frac{2\pi}{\lambda}} = \sqrt{\mu/\epsilon} H_z \tag{1.13}$$

Thus, $Z = \sqrt{\mu/\epsilon}$ is the intrinsic impedance of the medium. For free space, $Z_0 = \sqrt{\mu_0/\epsilon_0} \approx 120\pi$.

1.3 Poynting vector

Let us define the Poynting vector $\vec{S}$ as

$$\vec{S} = \vec{E} \times \vec{H}^* \quad [Watt/m^2] \tag{1.14}$$

This vector indicates the amount of enegy that flows at a given time through a unit area of the plane perpendicular to the direction of propagation. It can be thought of as the surface power density. The total power at that given time can be found by integrating S over the whole surface: $P = \oint_s S \cdot d\vec{s}$. With 1.13, the Poynting vector for the plane wave from 1.5 in free space becomes

$$\vec{S} = \vec{e_x} S_x = \vec{e_y} E_y \times \vec{e_z} H_z = \vec{e_x} \frac{E_y^2}{Z_0} \tag{1.15}$$

If we consider a harmonic plane wave, the instantaneous power per unit area is

$$S(t) = \frac{E^2(t)}{Z_0} = \frac{E_0^2 \sin(\omega t - \beta x)}{Z_0} \tag{1.16}$$

the peak power per unit area is the maximum of 1.16:

$$S_0 = \frac{E_0^2}{Z_0} \tag{1.17}$$

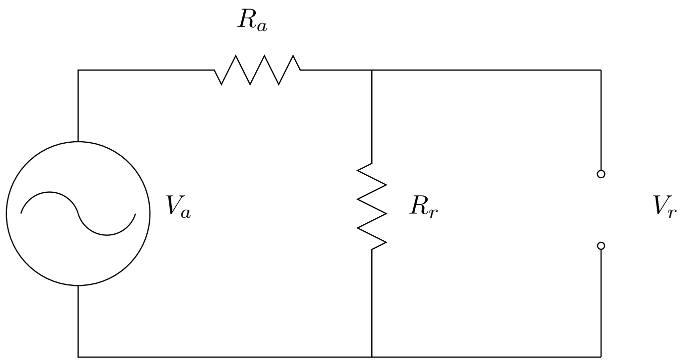
and the average power per unit area is

$$\bar{S} = \frac{1}{T} \int_0^T S(t) dt = \frac{E_0^2}{2Z_0} \tag{1.18}$$

1.4 Relating electrical field at the antenna to voltage at receiver

The received power from a plane wave at an antenna is the total power density received on the entire effective area A_e of the antenna:

$$P_a(d) = S A_e = \frac{A_e E^2}{Z_0} \tag{1.19}$$



Matched receiver: $R_r = R_a$

Figure 1.2: Model for an antenna with a matched receiver

We can model the antenna as a voltage source with an internal resistance R_a. If the receiver after the antenna is matched as a resistive load R_r to the antenna, as shown in figure 1.2, the voltage V_r at the receiver is half of the voltage V_a at the antenna (provided that the receiver measures V_r with a very high impedance voltmeter).

Of course, 1.19 should be equal to

$$P_a(d) = \frac{V_a^2}{R_a + R_r} = \frac{V_a^2}{2R_a} \tag{1.20}$$

The received power at the receiver is

$$P_r(d) = \frac{V_r^2}{R_r} = \frac{(V_a/2)^2}{R_a} = \frac{V_a^2}{4R_a} = P_a(d)/2 \tag{1.21}$$

which is half of the power at the antenna (since the other half of the power is dissipated by R_a inside the antenna). We can find the relation between the electrical field E and the input voltage V_r at the receiver by combining 1.21 with 1.19:

$$V_r = V_a/2 = E\sqrt{\frac{A_e R_a}{2Z_0}} \tag{1.22}$$

Thus, V_r is linear with E, and for a plane wave, we can evaluate the received power by evaluating V_r.

See [Kra53] and others for more information about electromagnetic waves and the Poynting vector.

Chapter 2

Parameters of static mobile channels

A radio channel can be either static or dynamic. In a static channel, the transmitter, the receiver and all of the objects in the environment are standing still. In a dynamic channel, at least one of them is moving.

In this chapter, we will discuss what whill happen to the amplitude (A), the phase (ϕ) and the time of arrival (τ_A) when a very short pulse with frequency ω (as defined by equation 2.1) travels through a static channel. FIXME: discuss also the effect of polarization if we can find anything appropriate.

$$E(t) = \delta(t)\hat{E}e^{j\omega t} \tag{2.1}$$

2.1 Amplitude fading

When a radio signal travels from the transmitter to the receiver, some part of it's energy will be lost. This loss is caused by:

- free space loss (PL_f)
- large scale effects from objects in the environment (PL_l) (eg: buildings)
- normal scale effects from objects in the environment (PL_n) (eg: individual walls and windows in a building)
- small scale effects from objects in the environment (PL_s) (eg: individual walls and windows in a building and small objects like computer screens, books and glasses on a table)

The size of objects in the examples above is of course relative to the wavelength; the examples above are more or less correct for wavelengths in the order of 12 cm.

In formula, the received power will be:

$$P_{r,dBm}(d) = P_{t,dBm} - PL_f - PL_l - PL_n - PL_s \tag{2.2}$$

The difference between normal scale effects and small scale effects is basically that small scale effects are caused by interference from scattered waves and normal scale effects is the shadowing caused by objects when the small scale effects are averaged over a local area (eg 1 m^2). We will discuss those effects in the next sections.

2.1.1 Free space loss and large scale effects

We will treat free space loss and large scale effects together, since large scale effects require just a minor modification in the formula for the free space path loss, as we shall see.

According to the Friis free space model, the received power is

$$P_{r,dBm} = P_{t,dBm} + G_s + 10\log\frac{1}{d^2} = P_{t,dBm} + G_s - 20\log d \quad d \geq d_f \tag{2.3}$$

where

$$G_s = 10\log\left(\frac{G_t G_r \lambda^2}{16\pi^2 L}\right) \tag{2.4}$$

is the system gain in the Friis free space path loss formula and d_f is the far-field region of the transmit antenna. In 2.4, G_t and G_r are the gain of the transmit respectively the receive antenna and L is the system loss not related to propagation ($L \geq 1$).

The Friis free space model is correct if there is a real free space; that is, if there are no other objects in the environment besides the transmit and receive antennas. In reality, there is almost never a free space environment, and therefore, the exponent of d in (2.3) and (2.7) is not always 2 but it's value depends on the topography of the terrain. For example, in urban area, this exponent can vary between 2.7 and 3.5 for cellular radio, and in a building with a line of sight connection, the exponent can vary between 1.6 and 1.8 (according to [Rap96]). Therefore, in the log-distance path loss model, the received power is:

$$P_{r,dBm} = P_{t,dBm} + G_s - 10n\log d \quad d \geq d_f \tag{2.5}$$

where $n = 2$ for a free space environment, but can be any positive real number for other environments.

Thus, the pathloss caused by free space loss and large scale effects can be modelled as a slightly modified Friis free space loss:

$$PL_f(d) + PL_l(d) = 10n\log d - G_s \tag{2.6}$$

Note: in practical measurements, sometimes some of the parameters in G_s are unknown. In that case, one can choose a reference distance d_0 and take measurements at many different locations which have distance d_0 to the transmitter. The average received power $P_{r,dBm}(d_0)$ can then be used as a reference. If possible, d_0 is chosen close enough to the transmitter that one can assume a free space environment around and between d_0 and the transmitter, so that one can determine a correction factor for the unknown parameters of G_s. Of course the reference distance should be in the far field region of the transmit antenna: $d_0 \geq d_f$. For outdoor systems, d_0 is usually 100 m or 1 km; for indoor systems, $d_0 = 1$ m is often a practical value (assuming the system uses low gain antennas and the frequency is between 1 and 2 GHz). Equation 2.5 will then become:

$$P_{r,dBm}(d) = P_{r,dBm}(d_0) + 10\log\left(\frac{d_0}{d}\right)^n = P_{r,dBm}(d_0) - 10n\log\left(\frac{d}{d_0}\right) \quad d \geq d_f,\ d_0 \geq d_f \tag{2.7}$$

and equation 2.6 will be:

$$PL_f(d) + PL_l(d) = PL_f(d_0) + PL_l(d_0) + 10n\log\left(\frac{d}{d_0}\right) \tag{2.8}$$

2.1.2 Normal scale effects

The model presented in section 2.1.1 doesn't consider the fact that the average received power depends a lot on the objects close to the location where the received power is measured. We can apply a correction for that, by dividing the coverage area of the transmit antenna into N smaller sections. Then, section

i has it's own normal scale pathloss $PL_{n,i}$ (and i can be any natual value between 1 and N of course). We can assume that the objects around the location where the received power is measured are randomly positioned and thus, PL_n is a random process. However, this random process should not influence the mean received power of the total coverage area (since the mean received power on that large scale should be predicted by the large scale pathloss $PL_l(d)$ and the free space pathloss $PL_f(d)$). Thus, the mean value of PL_n should be 0 and we only need to determine it's standard deviation σ_n, by analyzing the path losses from all the sections. The formula for the normal scale path loss is

$$PL_n = X_{n,\sigma_n} \tag{2.9}$$

2.1.2.1 Common distributions for PL_n

FIXME talk about log-normal distribution.

2.1.3 Small scale effects

Section 2.1.2 predicts the average received power for a small area of the complete coverage area. However, it does not predict what the actual received power inside such an area can be. The amplitude of a received electromagnetic wave is not only influenced by normal and large scale objects (and free space loss), but also by small scale objects which have a diameter of about one wavelength or less. For an office environment and a frequency of 2.4 GHz, objects like that could be mugs, tea cups and soda cans standing on a table, computer speakers or light bulbs etc. Those objects will cause scattering of the wave, but if there ar not many of those objects nearby, their influence can be neglected ([For95]).

However if we are transmitting a continous wave, small scale effects can also be caused by normal sized objects. For an office environment at 2.4 GHz, those objects could be walls, doors, a bookcase etc. Let us clearify the small scale effect by an example. Consider we are measuring inside an office. At a certain point, we can receive a wave reflected from a wall and another wave reflected from a door. If we were transmitting a very short pulse, we would either receive both reflections at a different time, or we would receive them at the same time. If we receive them at the same time (which would be very rare if the pulse is really short), the total power of the received pulse is the power of both individual pulses and the total power should be the same as predicted by the path loss model without small scale fading.

Yet, if we transmit a continuous wave, we would receive two continuous waves, which will interfere with each other. The phase of a wave is determined by the time it takes to travel to this point, and thus by the pathlength and the materialproperties of the obstacles which it penetrates. Since these two waves come from a different direction, they probably also have a different pathlength and thus a different phase (the phase of a wave can vary a lot if the total pathlength changes just a little bit). Therefore, those two waves can interfere constructively or destructively, and thus the power of the combined signal is not always the coherent sum of the powers of both signals. Let's assume now, that we are measuring at a location where there is a maximum, ie, the power of the resulting wave is the coherent sum of the powers of both incoming waves. If we then move just a little bit, about half a wavelength, the phase of both waves can change already so much that instead of constructive interference, we have destructive interference here, and the power of the combined signal is almost 0.

We can model this kind of fading in our pathloss model by adding an extra pathloss $X_{s,i}$ for every measurement i. Again X_s is a random variable with mean value 0 a standard deviation σ_s. The mean value is 0, since it should not influence the average received power measured at one section. The formula for the small scale path loss is

$$PL_s = X_{s,\sigma_s} \tag{2.10}$$

2.1.3.1 Common distributions for PL_s

FIXME Summarize Rayleigh and Rician distributions here. NOTE: if we combine log-normal and Rayleigh distributions, we get the Suzuki distribution.

2.1.4 An example

The different fading effects are all shown in figure 2.1. In the remainder of this section, we will see what happens when we transmit a continuous wave (unmodulated carrier) with a wavelength which is much larger than the distance between locations 1 and 2. Each section has a diameter of a few wavelengths.

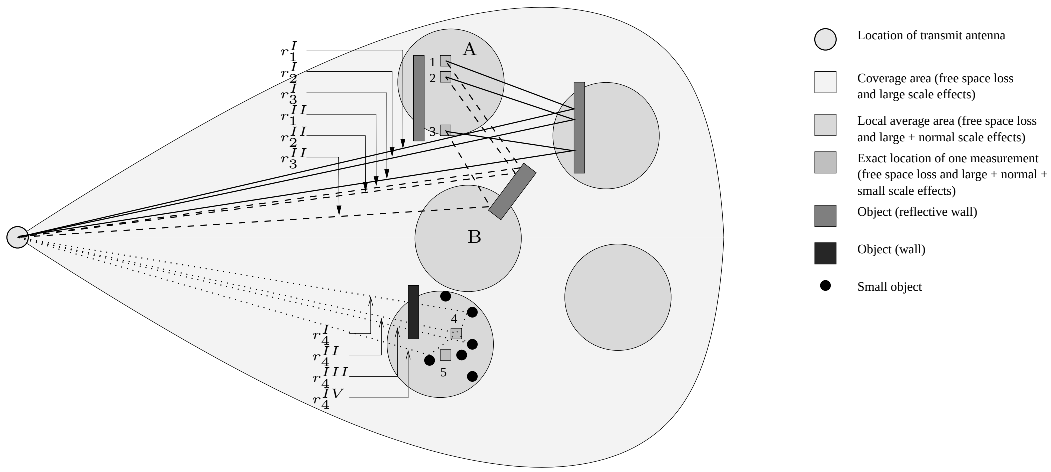


Figure 2.1: An illustration of the scale of the different fading effects

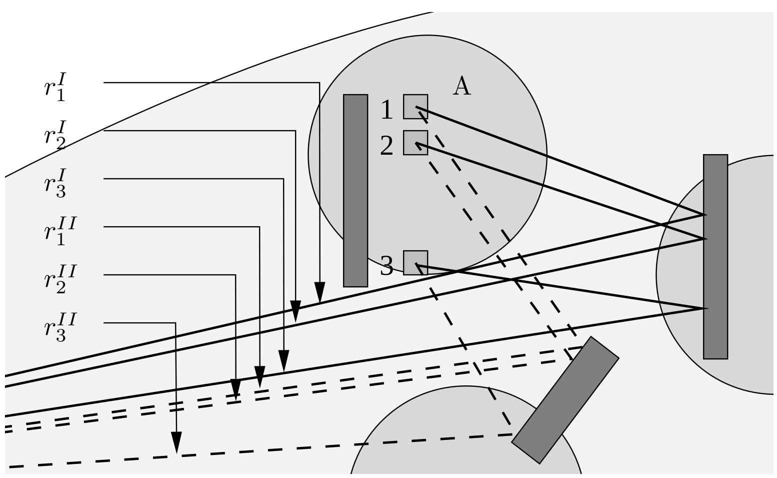


Figure 2.2: An illustration of the scale of the different fading effects, detail at section A

If we look at section A (figure 2.2), we see a number of locations. These locations are the locations for individual measurements. In location 1, there are two incoming waves, r_1^I and r_1^{II}. They are both reflected against a normal sized object and have more or less the same pathlength. Therefore, the amplitudes will probably have a strong correlation (assume here that this is the case). However, a small difference in pathlength can cause a large difference in phase and thus will determine whether we have constructive interference or destructive interference in the case of a continuous wave. If we look at location 2, there are also two incoming waves, r_2^I and r_2^{II}. Since r_1^I and r_2^I get reflected from the same wall and since they have almost the same pathlength, their amplitudes will be (almost) the same. Since the distance between location 1 and 2 is much shorter than the wavelength, the difference in pathlength will also be much shorther than one wavelength and therefore the phase difference between r_1^I and r_2^I is almost 0. The same reasoning goes for r_1^{II} and r_2^{II} and therefore, the resulting wave at location 2 will only differ a little in phase and amplitude with the resulting wave at location 1. For the two waves r_3^I and r_3^{II} at location 3, the amplitudes will again be very similar to the amplitudes of the waves arriving at location 1 and 2, since the pathlengts are comparable and since they bounce against the same walls. However, the difference in pathlength between r_3^I and r_1^I is not much less than one wavelength and thus, the phase of r_3^I is most probably not similar to the phases of r_1^I (the reasoning goes for r_1^{II}). Therefore, the resulting signal at location 3 will most probably have a different amplitude than the signal at locations 1 and 2. Thus, if we move slowly from 1 to 3, we will see a slow change in the amplitude of the resulting signal, caused by phase differences from reflections from normal sized objects (small scale effect).

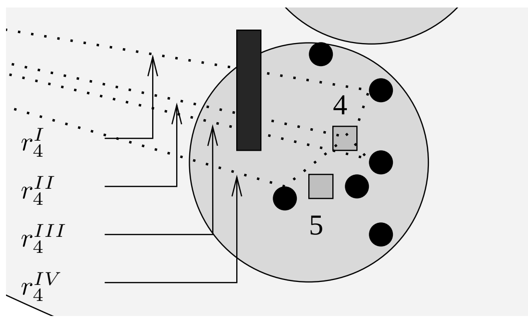


Figure 2.3: An illustration of the scale of the different fading effects, detail at locations 4 and 5

In a real world situation, there is also scattering from smaller objects in the surroundings of the measurement, as shown in location 4 (figure 2.3). Those objects have a size comparable to the wavelength or less and therefore, the radio wave will be more diffracted and "bend around" the object. The resulting waves arriving at the receiver can have a strong correlation in amplitude (like r_4^I and r_4^{III}), but that is not always the case (like r_4^I and r_4^{II} or r_4^I and r_4^{IV}). In any case, the phases of those waves will probably not have a strong correlation and therefore, if we are transmitting a continuous wave, there will be some interference, just like there is interference in location 1 from r_1^I and r_1^{II}. If we move slowly over a small distance from 4 to 5 however, it is very well possibly that the contribution of the received power due to this kind of scattering changes a lot, since at location 5, there are other small objects which will dominate this kind of scattering. However, this kind of interference does not severly affect the received power.

We can find the value for σ_s by analyzing each individual measurement within a small region. The size of a region should be a few wavelengths (eg, region A or B in figure 2.1) and the random process X_s should have a mean value of 0. Therefore, before one computes σ_s, he should substract the mean value of all the measurements in that region from all those measurements.

If we look at a little larger scale, we see regions A and B (and others which don't have a number). The average received power in section A can be different than the average received power in section B although A and B have about the same distance to the transmitter. Thus, we can find the value for σ_n by analyzing the mean values of all the regions that have about the same distance to the transmitter. Note however, that we should again substract the mean value of all those regions first, since the random process X_n should also have a mean value of 0.

Finally, if we look at an even larger scale, we should be able to find a value for n. (Note that if the environment changes too much, it could be necessary to determine multiple values for n, where each n_i

has it's own "operating range" from d_{i-1} to d_i. For an indoor environment however, this is most likely not the case.)

2.2 Time of arrival

One can predict the minimum amount of time it would take for a wave to travel distance d from transmitter to receiver with the formula $t = cd$ (where c is the speed of light and d is the distance between the transmitter and the receiver). However, if the wave has to penetrate objects, it will travel slower inside those objects. Also, if the wave is not going by line of sight, but instead bouncing against several objects before reacing the receiver, it will also have a longer delay. Thus, we can model the total time of arrival of a pulse or one of it's echos as follows:

$$t_a = cd + \tau_{\sigma_\tau} \tag{2.11}$$

where τ is a random distributed variable with standard deviation σ_τ.

An exact characterization for τ is not available [Yac93], however in general, the density function of τ is a negative exponential distribution:

$$p(\tau) = \frac{1}{\overline{\tau}} \exp\left(-\frac{\tau}{\overline{\tau}}\right) \tag{2.12}$$

An ideal delay profile is shown in figure 2.4, where a pulse is transmitted on $t = 0$ with height (amplitude) 3.5.

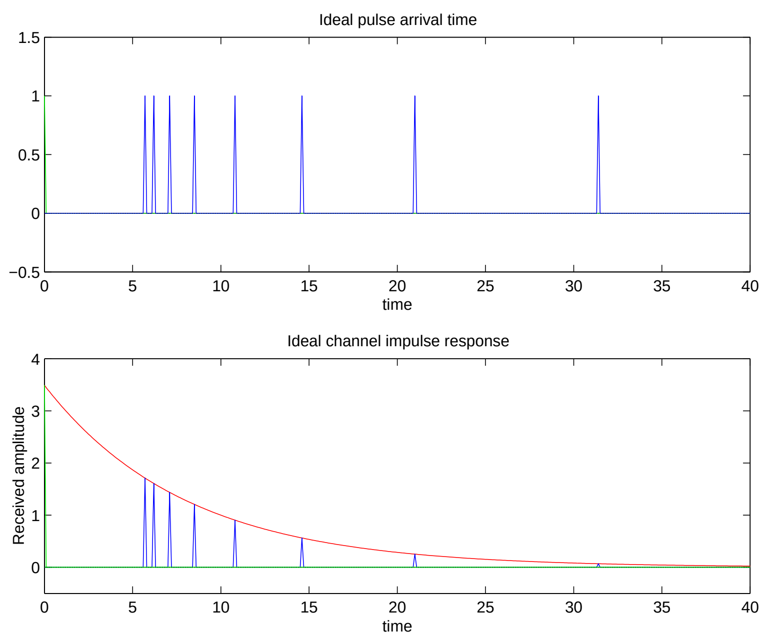


Figure 2.4: An ideal delay profile

FIXME explain this better: For such a distribution, the mean time delay as well as the delay spread are equal to $\overline{\tau}$.

FIXME Talk a little more about mean excess delay (first moment of power delay profile) and rms delay spread (root of second moment of power delay profile). Those values depend on the noise level that one chooses. Talk a bit about maximum excess delay, which depends on the noise level, but also on the threshold that one chooses.

2.2.1 Flat or frequency selective channel

The coherence bandwidth B_c of a channel is the maximum bandwidth between two frequencies where all of the frequencies within that band have a large chance of amplitude correlation. Thus, the channel deforms all of those frequencies in the same way with respect to the amplitude and phase and one can think of the channel as a filter with bandwidth B_c. The value for B_c depends of course on the amount of correlation that one requires and it depends on the rms delay spread σ_τ, but an exact relation ship is hard to give. If one chooses the output of the correltation function to be higher than 0.9 in order to have amplitude correlation, the coherence bandwith is ([Rap96])

$$B_c \approx \frac{1}{50\sigma_\tau} \tag{2.13}$$

and if the output should be higher than 0.5 to have amplitude correlation, it would be ([Rap96])

$$B_c \approx \frac{1}{5\sigma_\tau} \tag{2.14}$$

If the bandwidth of the signal B_s that one wants to transmit over the channel is smaller than B_c, the channel is a flat fading channel (the attenuation is almost the same for all the frequencies within B_s). This implies that the symbol time T_s is much larger than the rms delay spread σ_τ.

However, if the bandwidth of signal is larger than B_c, some of the frequencies in B_s will be subject to more attenuation than others and the channel is called a frequency selective channel. This implies that T_s is smaller than σ_τ.

2.3 Phase

FIXME phase is uniform distributed

2.4 Polarization

FIXME If I can find anything applicable here...

Chapter 3

Parameters of dynamic mobile channels

FIXME talk about coherence time and level crossing rate and random frequency modulation ([Yac93], figure 4.14 from [Rap96])

Chapter 4

Modulation techniques which are less sensitive to channel deformation

FIXME write introduction

4.1 FHSS

This is used in 802.11 for 1 and 2 Mbit modes. FIXME write this section

4.2 DSSS

This is used in 802.11 for 1 and 2 modes, and in 802.11b also for 5.5 and 11 Mbit modes. FIXME write this section

4.3 OFDM

FIXME write this section

Chapter 5

Introduction to ray tracing programs

In order to design a communication system, one has to have an idea about the parameters of the channel that can be expected for the operation of the system. Those parameters can be obtained by various methods:

- existing channel models
- channel sounding by measurements
- channel sounding by simulation

In this chapter, we will discuss those three methods.

5.1 Existing channel models

By using channel models, such as the "plane earth path loss model", the "knife-edge diffraction loss model" or the "Okumura/Hata model", one can usually get a general idea of the channel parameters for large scale fading. However, if no suitable model is available, or if one needs a better estimate, other methods have to be considered.

5.2 Channel sounding by measurements

One can get the real channel parameters of the system by doing measurements at the location where the system is to be deployed. Though, if one wants to do those measurements, he needs to have proper equipment to do the channel sounding. There are four ways to do channel measurements:

- direct RF pulse system channel sounding
- spread spectrum sliding correlator channel sounding
- frequency domain channel sounding
- poor-mans power measurement

5.2.1 Direct RF pulse channel sounding

The basic idea of direct RF pulse channel sounding is to transmit a very short probing pulse, ideally a delta function. The receiver should be able to distinguish each incoming echo of the pulse and record the time of arrival and the received power of each echo. Since each echo can be completely and individually distinguished, there is no overlap between the pulses (this can only be achieved if the probing pulse is short enough of course). Thus, the received power of a pulse cannot be subject to fading caused by interference from another echo. This means, that the total received power will not vary much over a small distance and the "normal scale" and small scale received power is the sum of the powers received in each multipath component. (However, since one knows the time of arrival and the amplitude of each echo, it is possible to compute the small scale received power for a continuous wave (carrier) at a certain location, and by doing that, determining the small scale fading for a continous wave.)

The direct result from direct RF pulse channel sounding is a channel impulse response graph. By doing a Fourier transform on that data, one gets the frequency response for that channel, ie, how each frequency will be attenuated by the channel.

5.2.2 Spread spectrum sliding correlator channel sounding

5.2.3 Frequency domain channel sounding

FIXME summary of rappaport 4.3.3

The direct result of frequency domain channel sounding is a graph which shows how much each frequency is attenuated. One can get the channel impulse response of the channel by doing an inverse Fourier transorm on the data from this measurement.

5.2.4 Poor-mans power measurement

In some cases, existing (communication-) systems can be used to do simple measurements of the received power and noise level. For example, with most 802.11b cards, it is possible to do measurements like this. Then, by taking measurements of the pathloss at a certain location, it is possible to determine some of the channel parameters. However, it is usually not possible to do real channel impulse response measurements with those systems, so one can only get information about the amplitude fading. Also, with the card that I tested, it was not possible to transmit short pulses; the system could only measure the received power from the carrier. Since the carrier is (ideally) a continuous wave, it can be subject to very large fading over a very short distance due to destructive interference. Therefore, one needs to take a lot of measurements on a very small grid to be able to determine the small scale, normal scale and large scale fading parameters.

The advantage of this system, however, is that one doesn't need expensive systems. In combination with a simulator, he can use the measurements to get information about the amplitude fading and use that to verify the computer model. After that, it is possible to use the computer model to get information about the phase, direction of arrival an time of arrival. See A for a practical description of this type of measurement.

5.3 Channel sounding by simulation

In theory, it could be possible to compute exactly how an electromagnetic wave propagates in buildings or cities using Maxwell's equations and the boundary conditions defined by the buildings. Unfortunately, the computing power of todays computers is not enough to do those calculations. However, if one is designing a wireless communication system which needs to be deployed in even just a few slightly different environments (eg, one needs to design a system to operate in a few offices), one is generally more interested in large and “normal” scale parameters. The small scale variations can be characterized statistically and can be taken into account in the link fade margins. By using the ray-tracing approach, it is possible to predict the large and normal scale parameters.

Advantages when compared to measurements:

- much faster than taking measurements
- no need for expensive measurement system
- one can get all parameters (including direction of arrival at receiver and direction of transmission at transmitter, which are hard to get using measurements)

Disadvantages:

- the accuracy of the model is limited and it is unknown how correct the model is
- for as far as I know, there are no free or inexpensive simulators on the market today

Since this project uses a ray-trace program very often, we will look in to that a little closer.

5.3.1 Ray-tracing

FIXME write this

5.3.1.1 Algorithm

FIXME write summary about Valenzuela’s paper

5.3.1.2 Accuracy of the algorithm

FIXME show some extreme cases here where the algorithm is not (very) correct

5.3.1.3 Accuracy of the model

Appendix A

A practical guide for poor-mans power measurement

FIXME: write this

Bibliography

[Smo00] Smolders, A. B.: Elektromagnetische Antennes, course notes from the course ”Elektromagnetische golven en antennes” at Eindhoven University of Technology.

[Kra53] Kraus, J. D. : Electromagnetics

[Rap96] Rappaport, T. S.: Wireless communications principles and practice

[Yac93] Yacoub, M. D.: Foundations of mobile radio engineering

[Val94] Valenzuela, R.A.: Ray Tracing Prediction of Indoor Radio Propagation. Personal, Indoor and Mobile Radio Communications, 1994. Wireless Networks - Catching the Mobile Future. 5th IEEE International Symposium on, Volume: 1 , 18-23 Sep 1994 Page(s): 140 -144 vol. 1)

[For95] Fortune, S.J.; Gay, D.M.; Kernighan, B.W.; Landron, O.; Valenzuela, R.A.; Wright, M.H.: WISE design of indoor wireless systems: practical computation and optimization. 1995 IEEE Computational Science and Engineering, Volume: 2, Issue: 1, Spring 1995, pp. 58-68