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TODO: check that 50 Ohms line impedance is used every were.

TODO: check that terminating resistance is 25 Ohms

TWO PORT MODELS FOR THE POWER LINE CHANNEL

THE ZIMMER MODEL

Zimmer proposed a multipath model for the power line channel in [1]. The model assumes the received signal is a sum o multiple signal that travel multiple distances and suffer different delay throw the channel. It assumes a dispersive channel with frequency dependent attenuation. The complex of a signal that travel a distance d is given by,

$$\mathrm{Zimmer}[d] = \mathrm{Exp}\left[-\left(\mathrm{a0} + \mathrm{a1} f^{k}\right) d - \frac{2\pi f d}{c} j\right]$$

The parameters values taken also from [1] table 3 for 100 meters line as in, corresponding to a0=9.40e-3 e a1=4.20e-7, k=0.7 will be used further in the paper. These correspond to 55.2 dB attenuation for a 100 m line at 20 MHz and -8.1 dB at DC.

For a k=1 the impulse response such a channel can be calculated. Ignoring the delay term,

$$\mathrm{h(t)} = \frac{2\mathrm{a1} d e^{-\mathrm{a0} d}}{\mathrm{a1}^2 d^2 + 4\pi^2 t^2}$$

Plotting this for d=100m and using the values above result in the chart in Figure 1. It can be seen that the pulse widens for greater distances, as in any dispersive channel, and resulting from the increase in attenuation. Note that that the delay for 100 m is 0.3 μs.

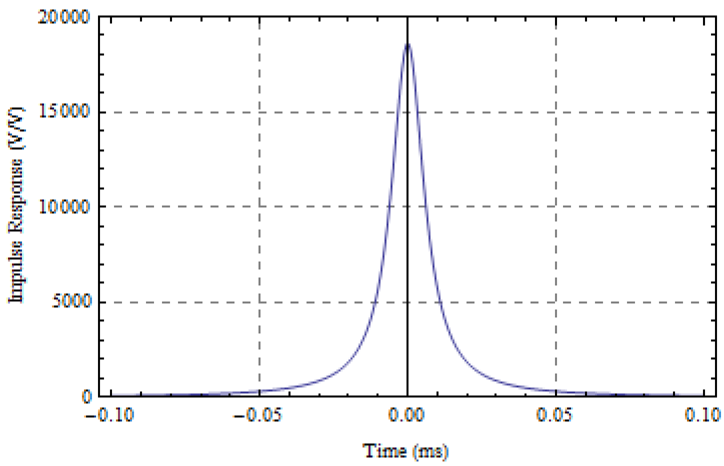


FIGURE 1 – IMPULSE RESPONDE OF A ZIMMER CHANNEL

This in fact corresponds to a non causal system since the impulse response is non zero for negative time. This will not be a problem in our implementation since one can delay the response to make it causal without reducing the usefulness of the model.

However, the minimum phase required to make the system causal can be calculated using Hilbert transform relations for the phase and amplitude. Given the Laplace transform of the system *H(s)*, if *Log(H(s))* is causal then *H(s)* is minimum phase. This is because if *Log(H(s))* has no poles (or singularities) in the right complex plane, then H(s) has no zeros on the right complex plane. The imaginary part of *Log(H(s))* is the phase of *H(s)*, and the real part is the magnitude. These mean that the phase of H(s) can be calculated as,

$$Phase\big(H(f)\big) = FT\big(IFT(Log|H(f)|)\mathrm{Sign(t)}\big).$$

Were FT is the Fourier Transform and IFT is the inverse Fourier Transform. This is equivalent to

$$Phase\big(H(f)\big) = \int_{-\infty}^{\infty} Log|H(f_0)| \frac{1}{(f-f_0)\pi\, i} \mathrm{df_0}$$

Were the integral must be calculated using the symmetry around singularity, these results that the phase of the Zimmer model will be,

$$Phase\big(Zimmer(d)\big) = \frac{2\mathrm{a1}df^k\mathrm{Gamma}[-k]\mathrm{Gamma}[1+k]\mathrm{Sin}[\frac{k\pi}{2}]^2}{\pi}$$

The resulting impulse response, for *k* equal to 0.7, is represented in Figure 2

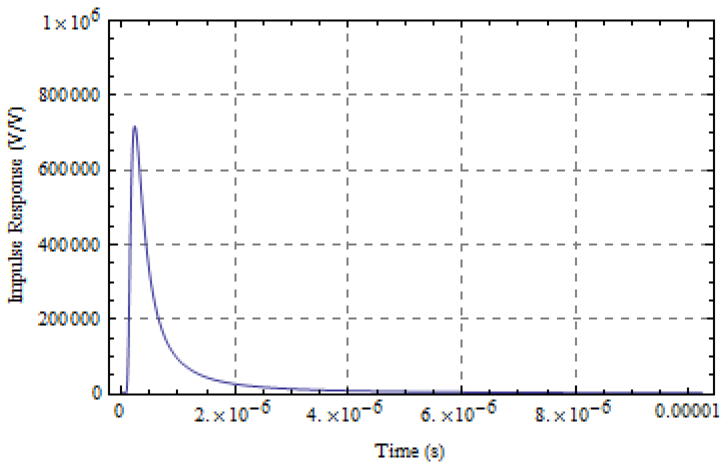
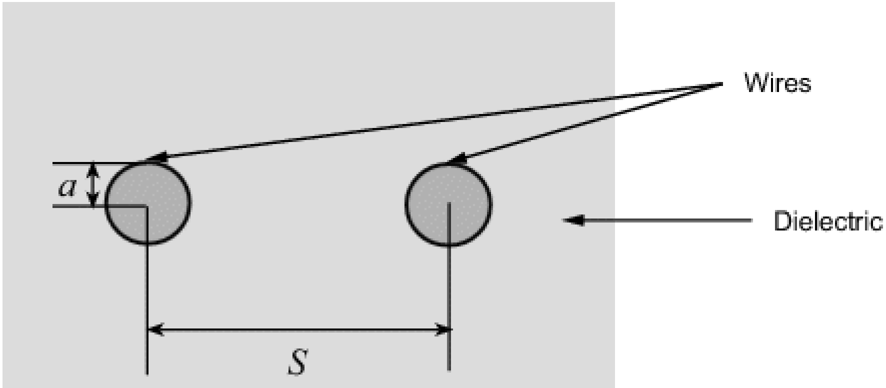


FIGURE 2 – CAUSAL IMPULSE RESPONSE OFF THE ZIMMER MODEL.

STUBLESS TRANSMISSION LINE

A two-wire transmission line is shown in cross-section in the following figure. Its physical characteristics include the radius of the wires *A*, the separation or physical distance between the wire centers *S*, and the relative permittivity and permeability of the wires.



The ABCD parameters of the line are as follows

$$A = \frac{e^{kd} + e^{-kd}}{2}$$
$$B = \frac{Z_0 * \left(e^{kd} - e^{-kd}\right)}{2}$$
$$C = \frac{e^{kd} - e^{-kd}}{2 * Z_0}$$
$$D = \frac{e^{kd} + e^{-kd}}{2}$$

The variables k and Z_0, are the propagation constant and the characteristic impedance of the line. They can be calculated from the resistance (R), inductance (L), conductance (G), and capacitance (C) per unit length (meters) as follows:

$$Z0 = \sqrt{\frac{R + j2\pi f L}{G + j2\pi f C}}$$

$$\gamma = \alpha + j\beta = \sqrt{(R + j\omega L)(G + j\omega C)}$$

Where

$$L = \frac{\mu}{\pi}\cosh^{-1}\left(\frac{D}{2a}\right)$$

$$C = \frac{\pi\,\varepsilon'}{\cosh^{-1}\left(\frac{D}{2a}\right)}$$

$$R = \frac{R_s}{\pi a}$$

$$G = \frac{\pi\omega\varepsilon''}{\cosh^{-1}\left(\frac{D}{2a}\right)}$$

$$R_s = \frac{1}{\sigma\, \delta_s}$$

$$\varepsilon = \varepsilon' - j\varepsilon'' = \varepsilon'(1 - j \tan \delta)$$

And ε' and ε'' are the real and imaginary parts of ε and $\tan \delta$ is the loss tangent of the dielectric. The conductivity of the dielectric is σ the permeability is μ and the permittivity is ε. The skin depth of the conductor is

$$\delta_s = \frac{1}{\sqrt{\pi f \mu \sigma_s}}$$

Note that assuming a constant value for $\tan(\delta)$ results that the attenuation constant that is proportional to frequency as obtained in practice.

Note that the vacuum permeability is,

$$\mu = \frac{\pi}{2500000} \frac{sV}{Am}$$

And the vacuum permittivity is,

$$\varepsilon = 8.85419 \times 10^{-12} \frac{As}{mV}$$

Resulting that the impedance in vacuum is,

$$Z_0 = \sqrt{\frac{\mu}{\varepsilon}} = 376.73\ \Omega$$

And the speed of light in vacuum is,

$$c = 1/\sqrt{\mu\varepsilon} = \ 2.99792 \times 10^8 \text{ m/s}$$

Calculating the values for the line impedance for different geometries of the line one gets [14] [5 page 413],

d/2a	**1**	**1,1**	**1.2**	**1.3**	**1.4**	**1.5**	**1.6**	**1.7**
Air : Z0 (Ω)	0	53	75	91	104	115	126	135
PVC : Z0 (Ω)	0	30	42	51	58	65	70	75

TABLE 1 – VARIATION OF THE IMPEDANCE OF A TWO WIRE LINE FOR DIFERENT GEOMETRIES

The value for PVC is obtained using a relative permittivity of 3.2.

The value usually assumed for the impedance of a power line is **50Ω**, this is the impedance of a LISN (Line Impedance Stabilization Network) of the CISPR standard. We will assume the at the receiver MODEM the line is terminated by an impedance of **25Ω.** Of course in a true MODEM the termination will probably be also 50Ω but this will allow accounting for the mismatch with actual values of the impedance of the line. Note that for telecommunications ports the ISN (Impedance Stabilization Network) is used that has a constant impedance of **150Ω**. For a no loss transmission line with 10m this will result in Figure 3 for the input impedance of the line.

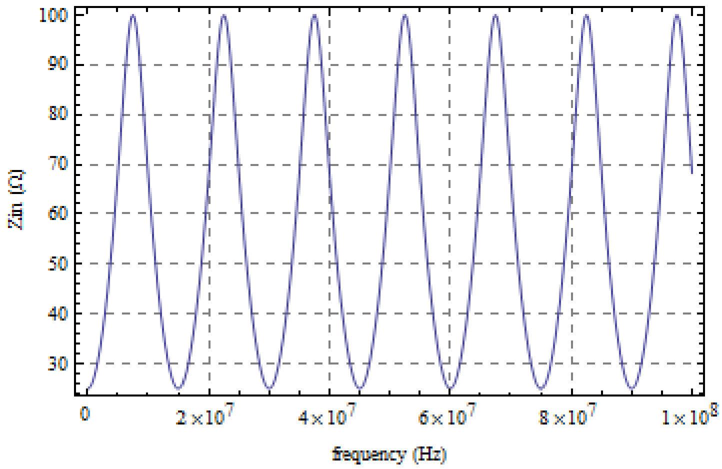


FIGURE 3 – INPUT INPEDANCE OF A NO LOSS TRANSMISSION LINE

Assuming a frequency dependent attenuation, increasing linearly with frequency, the input impedance of a open circuit terminated line with length, $l = 10m$, and attenuation -55.2391 dB or 40dB at 100m and 20MHz is represented in Figure 4.

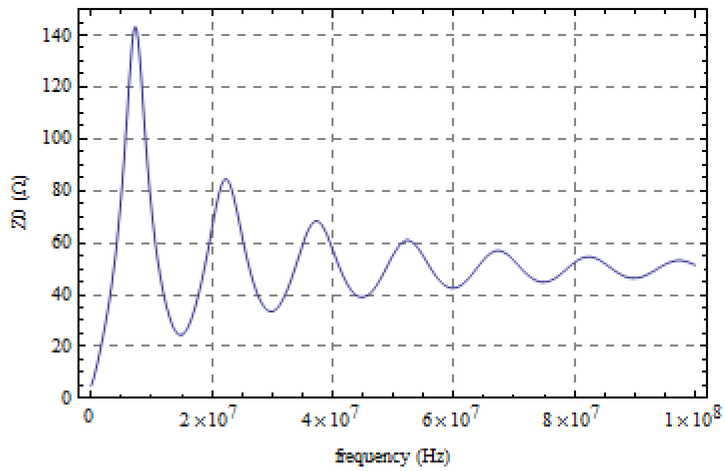


FIGURE 4 – INPUT IMPEDANCE OF AN OPEN-CIRCUIT TERMINATED LINE WITH FREQUENCY DEPENDENT ATENUATION

The result for a 2m line is plotted in Figure 5.

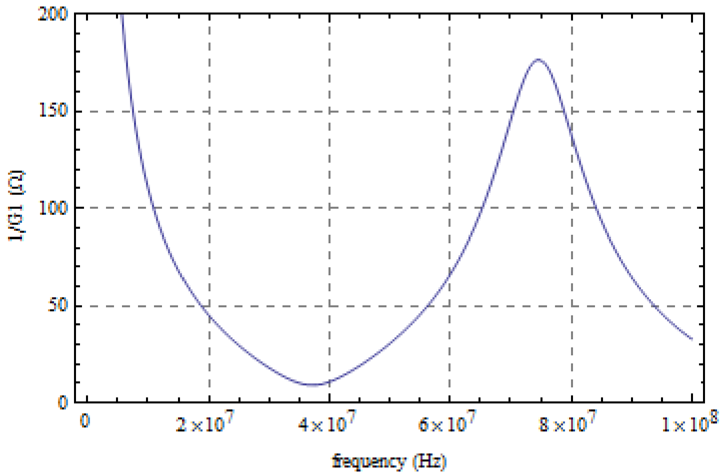


FIGURE 5 - INPUT IMPEDANCE OF A 2M OPEN-CIRCUIT TERMINATED LINE WITH FREQUENCY DEPENDENT ATENUATION

TWO PORT MODELS

As in PTDCEEA-TEL679792006 - PLC Noise - year 1.pdf the biport model for the power line is,

$$\begin{pmatrix} \text{Av} & \text{Z0} \\ \text{G1} & \text{Ag} \end{pmatrix} = \begin{pmatrix} \frac{2e^{l(ik+\alpha)}}{1+e^{2l(ik+\alpha)}} & \frac{(-1+e^{2l(ik+\alpha)})Z}{1+e^{2l(ik+\alpha)}} \\ \frac{-1+e^{2l(ik+\alpha)}}{(1+e^{2l(ik+\alpha)})Z} & -\frac{2e^{l(ik+\alpha)}}{1+e^{2l(ik+\alpha)}} \end{pmatrix}$$

with,

$$\begin{pmatrix} \text{V0} \\ \text{I1} \end{pmatrix} = \begin{pmatrix} \text{Av} & \text{Z0} \\ \text{G1} & \text{Ag} \end{pmatrix} \begin{pmatrix} \text{V1} \\ \text{I0} \end{pmatrix}$$

Or

$$\begin{pmatrix} \text{Sech}[l(ik+\alpha)] & Z\,\text{Tanh}[l(ik+\alpha)] \\ \frac{\text{Tanh}[l(ik+\alpha)]}{Z} & -\text{Sech}[l(ik+\alpha)] \end{pmatrix}$$

The inverse of this matrix is itself, that is,

$$\begin{pmatrix} \mathrm{V0} \\ \mathrm{I1} \end{pmatrix} = \begin{pmatrix} \mathrm{Av} & \mathrm{Z0} \\ \mathrm{G1} & \mathrm{Ag} \end{pmatrix} \begin{pmatrix} \mathrm{V1} \\ \mathrm{I0} \end{pmatrix}$$

And

$$\begin{pmatrix} \mathrm{V1} \\ \mathrm{I0} \end{pmatrix} = \begin{pmatrix} \mathrm{Av} & \mathrm{Z0} \\ \mathrm{G1} & \mathrm{Ag} \end{pmatrix} \begin{pmatrix} \mathrm{V0} \\ \mathrm{I1} \end{pmatrix}$$

In order to make the system causable one has to write,

$$\begin{pmatrix} \mathrm{Av} & \mathrm{Z0} \\ \mathrm{G1} & \mathrm{Ag} \end{pmatrix} = \begin{pmatrix} \dfrac{2e^{-l(ik+\alpha)}}{e^{-2l(ik+\alpha)}+1} & \dfrac{(1-e^{-2l(ik+\alpha)})Z}{1+e^{-2l(ik+\alpha)}} \\ \dfrac{1-e^{-2l(ik+\alpha)}}{(1+e^{-2l(ik+\alpha)})Z} & -\dfrac{2e^{-l(ik+\alpha)}}{1+e^{-2l(ik+\alpha)}} \end{pmatrix}$$

Resulting in

$$\mathrm{Av} = \frac{2e^{-l(ik+\alpha)}}{1+e^{-2l(ik+\alpha)}}; \mathrm{Z0} = \frac{\left(1-e^{-2l(ik+\alpha)}\right)Z}{1+e^{-2l(ik+\alpha)}};$$

$$\mathrm{G1} = \frac{1-e^{-2l(ik+\alpha)}}{(1+e^{-2l(ik+\alpha)})Z}; \mathrm{Ag} = -\frac{2e^{-l(ik+\alpha)}}{1+e^{-2l(ik+\alpha)}};$$

For a transfer function of the form, A/(1 + B), the output can be written as,

$$A(1 - B + B^2 - B^3 + B^4 - B^5 + B^6 - B^7 + B^8 - B^9 + B^{10} - B^{11} + B^{12} - B^{13} + B^{14} - B^{15} + B^{16} + \cdots)\mathrm{Input}$$

These results in

$$\mathrm{Av}[t] = \sum_{i=0}^{\infty} 2(-1)^i \delta[t - l/c - \frac{2l}{c} i]$$

$$\mathrm{Z0}[t] = Z(\delta[t] - \sum_{i=0}^{\infty} 2(-1)^i \delta[t - 2l/c - \frac{2l}{c} i])$$

$$\mathrm{G1}[t] = 1/Z(\delta[t] - \sum_{i=0}^{\infty} 2(-1)^i \delta[t - 2l/c - \frac{2l}{c} i])$$

$$\mathrm{Ag}[t] = -\sum_{i=0}^{\infty} 2(-1)^i \delta[t - l/c - \frac{2l}{c} i]$$

Assuming a frequency dependent attenuation along the line given by,

$$\mathrm{Zimmer}[d] = \mathrm{Exp}\left[-\left(\mathrm{a0} + \mathrm{a1} f^k\right) d - \frac{2\pi f d}{c} j\right]$$

$$\mathrm{Av}[f] = \sum_{i=0}^{n} 2(-1)^i \mathrm{Zimmer}[l + 2l\, i]$$

$$\mathrm{Z0}[f] = Z(1 - \sum_{i=0}^{n} 2(-1)^i \mathrm{Zimmer}[2l + 2l\, i])$$

$$\mathrm{G1}[f] = 1/Z(1 - \sum_{i=0}^{n} 2(-1)^i \mathrm{Zimmer}[2l + 2l\, i])$$

$$\mathrm{Ag}[f] = -\sum_{i=0}^{n} 2(-1)^i \mathrm{Zimmer}[l + 2l\, i]$$

This model can be used to create blocks that simulate a portion of the power line, something like,

FIGURE 1 – A BIPORT MODEL OF THE POWERLINE.

This corresponds to the following circuit,

FIGURE 2 – THE CIRCUIT FOR THE BIPORT MODEL OF THE POWERLINE.

And this can be modeled as,

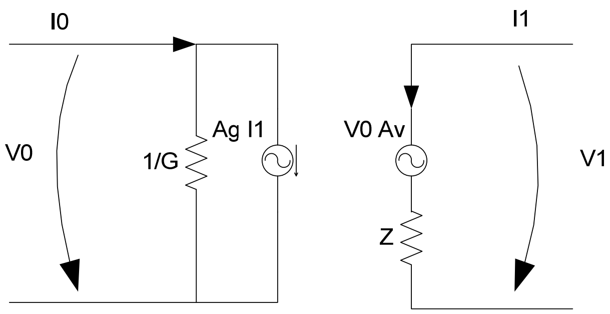


FIGURE 3 – DETAILED CIRCUIT FOR THE BIPORT MODEL OF THE POWERLINE.

Note that on the right side the current enters the box, resulting in a current that has the opposite direction of the voltage. The blocks can be connected to form more complex lines configurations as the one shown in Figure 4.

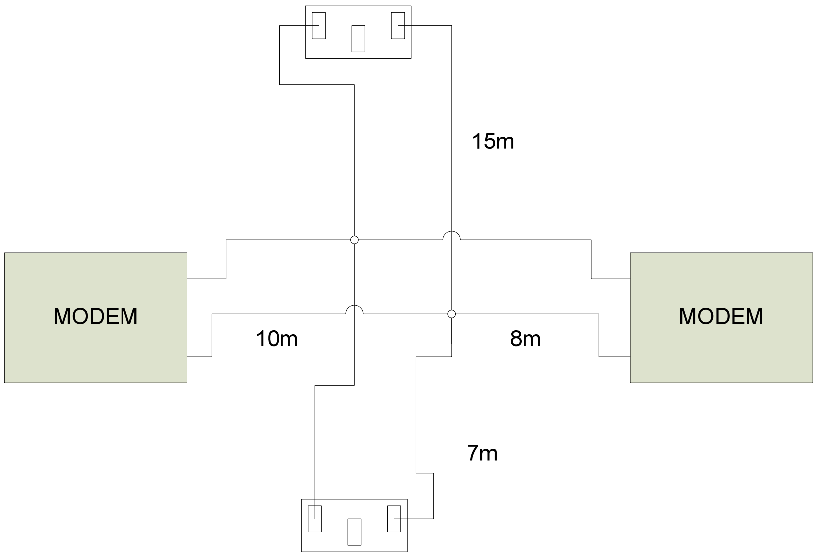


FIGURE 4 – AN EXAMPLE FOR A POWER LINE CONFIGURATION.

As an example the branches were chosen with lengths 10m, 7m, 8m and 15m. At the MODEM the line are terminated with a resistance while at the other end the circuits are assumed to be open.

This will result in a configuration as,

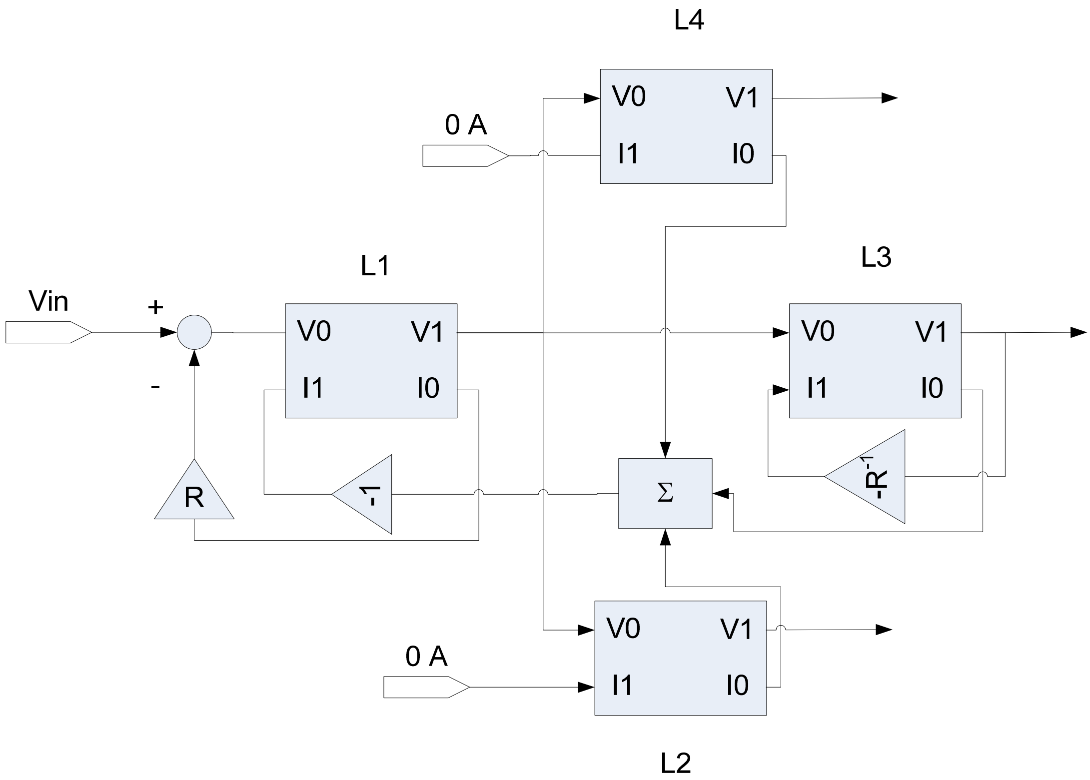


FIGURE 5 – MODEL OF THE POWERLINE CONFIGURATION IN FIGURE 4.

The direct feedback in the blocks is difficult to implement without solving the equations.

EQUATIONS FOR THE 4 LINE MODEL

In order to implement this model the equation for a four-stub configuration needs to be solved. We will solve the following set of equations,

$$I12=0;\ I14=0;\ I11 =-(I02+I03+I04);$$

$$V02=V11;\ V03=V11;\ V04=V11;$$

$$I13=-1/R\ V13;$$

$$V11==Av1\ V01 + Z01\ I11;\ I01==G11\ V01 + Ag1\ I11;$$

$$V12==Av2\ V02 + Z02\ I12;\ I02==G12\ V02 + Ag2\ I12;$$

$$V13==Av3\ V03 + Z03\ I13;\ I03==G13\ V03 + Ag3\ I13;$$

$$V14==Av4\ V04 + Z04\ I14;\ I04==G14\ V04 + Ag4\ I14;$$

This will result in,

I01 = -(V01 (G11 (-Ag3 Av3 Z01 +
R (1 + G12 Z01 + G13 Z01 + G14 Z01) + Z03 + G12 Z01 Z03 +
G13 Z01 Z03 + G14 Z01 Z03) +
Ag1 Av1 (Ag3 Av3 - (G12 + G13 + G14) (R +
Z03))))/(Ag3 Av3 Z01 -
R (1 + G12 Z01 + G13 Z01 + G14 Z01) - (1 + G12 Z01 + G13 Z01 +
G14 Z01) Z03),

V12 = (Av1 Av2 V01 (R + Z03))/(-Ag3 Av3 Z01 +
R (1 + G12 Z01 + G13 Z01 + G14 Z01) + (1 + G12 Z01 + G13 Z01 +
G14 Z01) Z03)

V14 = (Av1 Av4 V01 (R + Z03))/(-Ag3 Av3 Z01 +
R (1 + G12 Z01 + G13 Z01 + G14 Z01) + (1 + G12 Z01 + G13 Z01 +
G14 Z01) Z03)

I03 = (Av1 V01 (Ag3 Av3 -
G13 (R + Z03)))/(Ag3 Av3 Z01 -
R (1 + G12 Z01 + G13 Z01 + G14 Z01) - (1 + G12 Z01 + G13 Z01 +
G14 Z01) Z03)

I02 -> (Av1 G12 V01 (R + Z03))/(-Ag3 Av3 Z01 +
R (1 + G12 Z01 + G13 Z01 + G14 Z01) + (1 + G12 Z01 + G13 Z01 +
G14 Z01) Z03)

I04 = (Av1 G14 V01 (R + Z03))/(-Ag3 Av3 Z01 +
R (1 + G12 Z01 + G13 Z01 + G14 Z01) + (1 + G12 Z01 + G13 Z01 +
G14 Z01) Z03)

V11 = (Av1 V01 (R + Z03))/(-Ag3 Av3 Z01 +

R (1 + G12 Z01 + G13 Z01 + G14 Z01) + (1 + G12 Z01 + G13 Z01 +

G14 Z01) Z03)

V13 = (Av1 Av3 R V01)/(-Ag3 Av3 Z01 +

R (1 + G12 Z01 + G13 Z01 + G14 Z01) + (1 + G12 Z01 + G13 Z01 +

G14 Z01) Z03)

THE DIRECT FEEBACK PROBLEM

The direct feedback in the blocks of Figure 5 causes instabilities if the system is simulated as represented without solving the resulting system of equations, and this may be difficult if one wish to obtain a flexible system, that works with any configuration. However, if one only wishes to work with one or two models this is easy, and was the approach used in the remaining of the project. For completeness however one presents here the one technique that would allow a more flexible simulation system.

The direct feedback instabilities can be solved by reducing the gain from the inputs to the outputs at high frequencies. This corresponds to a low pass filter; however in general low pass filtering increases the delay in the block, increasing the stability problem. One solution is to use a non causal low pass filter.

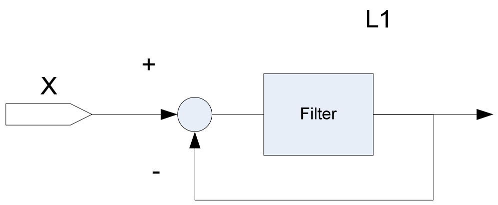


FIGURE 11 – DIRECT FEEDBACK IN A BLOCK

The system in Figure 10 corresponds to the equations,

$$y[n] = \sum_i h_i \left(x[n-i] - y[n-i]\right)$$

If the filter is non-causal i will take negative values. Non causal filters will use future values of y, but as long as the system is stable, values far into the future will have a small influence in the present value. This can be used to implement to filter, as long as a long processing delay is allowed. A vector of future values are processed at once, as in

$$y[n-M+j] = \sum_{i=1}^{M} h_i\,(x[n-i-M+j] - y[n-i-M+j])$$

with, j taking values from 0 to N. The resulting implementation would be as represented in Figure 11.

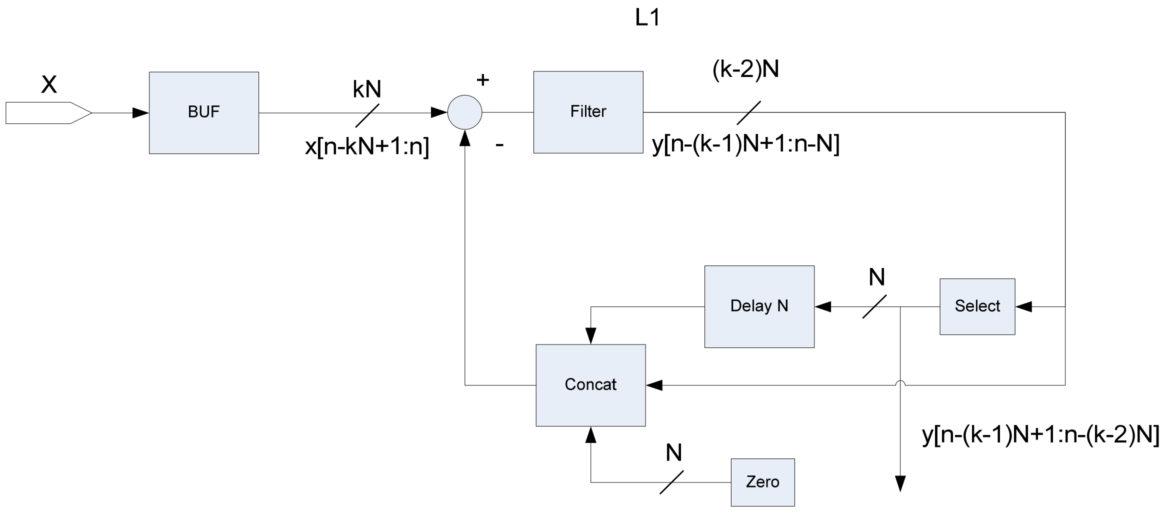


FIGURE 12 – IMPLEMENTINGS A NON CAUSAL FILTER WITH FEEBACK

RELATION TO ABCD PARAMETERS

These parameters are related to the so called ABCD parameter (as in the Mathworks RF blockset). These are defined by,

$$\begin{pmatrix} \mathrm{V1} \\ \mathrm{I1} \end{pmatrix} = \begin{pmatrix} \mathrm{A} & \mathrm{B} \\ \mathrm{C} & \mathrm{D} \end{pmatrix} \begin{pmatrix} \mathrm{V0} \\ -\mathrm{I0} \end{pmatrix}$$

The minus in I0 allows the calculation of the cascade of two lines by simply multiplying the two matrixes, but makes the line asymmetric. One has,

$$\begin{pmatrix} \mathrm{V1} \\ \mathrm{I1} \end{pmatrix} = \begin{pmatrix} \frac{1}{2}e^{-l(ik+\alpha)} + \frac{1}{2}e^{l(ik+\alpha)} & -\frac{1}{2}e^{-l(ik+\alpha)}Z + \frac{1}{2}e^{l(ik+\alpha)}Z \\ -\frac{e^{-l(ik+\alpha)}}{2Z} + \frac{e^{l(ik+\alpha)}}{2Z} & \frac{1}{2}e^{-l(ik+\alpha)} + \frac{1}{2}e^{l(ik+\alpha)} \end{pmatrix} \begin{pmatrix} \mathrm{V0} \\ -\mathrm{I0} \end{pmatrix}$$

Resulting in

$$A = \frac{1}{2}e^{-l(ik+\alpha)} + \frac{1}{2}e^{l(ik+\alpha)}$$

$$B = -\frac{1}{2}e^{-l(ik+\alpha)}Z + \frac{1}{2}e^{l(ik+\alpha)}Z$$

$$C = -\frac{e^{-l(ik+\alpha)}}{2Z} + \frac{e^{l(ik+\alpha)}}{2Z}$$

$$D = \frac{1}{2}e^{-l(ik+\alpha)} + \frac{1}{2}e^{l(ik+\alpha)}$$

We can solve the equations above to V0, I1 resulting in,

$$\begin{pmatrix} V0 \\ I1 \end{pmatrix} = \begin{pmatrix} \frac{1}{A} & \frac{B}{A} \\ \frac{C}{A} & \frac{BC-AD}{A} \end{pmatrix} \begin{pmatrix} V1 \\ I0 \end{pmatrix},$$

That relates the ABCD parameters with the hybrid (g) parameters used in the report.

LIMITING VALUES FOR SIGNALS

The values for the voltage and current signals in the power lines are limited by regulations related to electromagnetic compatibility, EMC. The following sections present these limit and limits for current values based on the emission limits.

In general values in Europe are regulated by CENELEC that references IEC, CISPR 22 norm. In the United States FCC Part 15 regulates electromagnetic compatibility.

European legislation, Directive 2004/108/CE of the European parliament and of the council, states that equipment that follows electromagnetic compatibility norm has the CE marking. The Official Journal of the European Union, 2008/C 280/05 specifies the norms that must be followed. These usually refer IEC/CISPR norms.

The International Electro-technical Commission (IEC) is an organization that promotes and coordinates international standardization and related matters, such as the assessment of conformity to standards, in the fields of electricity, electronics and related technologies.

CISPR 22 standard or its European equivalent EN 55022, entitled "Information technology equipment – Radio disturbance characteristics – Limits and methods of measurement". This standard defines limits for conducted emissions at mains ports and telecommunication ports for frequencies up to 30MHz as well as limits for radiated emissions between 30MHz and 1000MHz.

The voltage limit implies current limit dictated by the impedance of the ISN or LISN, but this usually assumes constant impedance for the frequency range of interest.

LIMITING VALUES DUE TO CONDUCTION INTERFERENCE

A Class B digital device is marketed for residential notwithstanding its use in industrial environments [9] while Class A is for industrial. In CISPR 22 [8] states that Class A limits can be used if the equipment states that it may cause radio interference.

According to FCC Part 15(2008), EN55022 and CISPR 22 the limits for the signal values in the mains port are:

Frequency range (MHz)	Limits in dB(μV)			
	Class A		**Class B**	
	Quasi-peak	Average	Quasi-peak	Average
0.15 – 0.50	79	66	66-56	56-46
0.50 – 5	73	60	56	46
5 – 30	73	60	60	50

TABLE 2 – LIMITS FOR THE SIGNAL VALUE AT THE MAINS PORT (FCC AND CISPR)

These values are from phase and neutral to ground [8] and measured using a Line Impedance Stabilization Network (LISN), while the values at the telecommunications port are common mode and measured using an Impedance Stabilization Network (ISN) but are planes to add a so called multi propose port with measurements made with an ISN.

The limits for the signal values in the telecommunications port for class B are:

Frequency Range (MHz)	Voltage limits (dBμV)		Current limits (dBμA)	
	Quasi-peak	Average	Quasi-peak	Average
0.15 - 0.50	84 – 74	74 – 64	40 – 30	30 – 20
0.5 – 30	74	64	30	20

TABLE 3 – LIMITS FOR THE SIGNAL VALUES AT THE TELECOMMUNICATIONS PORT (CISPR) – CLASS B

The limits for the signal values in the telecommunications port for class A are:

Frequency Range (MHz)	Voltage limits (dBμV)		Current limits (dBμA)	
	Quasi-peak	Average	Quasi-peak	Average
0.15 - 0.50	97 – 87	84 - 74	53 – 43	40 – 30
0.5 - 30	87	74	43	30

TABLE 4 – LIMITS FOR THE SIGNAL VALUES AT THE TELECOMMUNICATIONS PORT (CISPR) – CLASS A

LIMITING VALUES DUE TO EMISSION (FCC)

According to FCC Part 15(2008) the field strength of radiated emissions from unintentional radiators at a distance of 3 meters shall not exceed the following values:

Frequency of Emission (MHz)	Field Strength $(\mu V/m)$	
	Class A (at 10 m)	**Class B (at 3 m)**
30 – 88	90	100
88 – 216	150	150
216 – 960	210	200
Above 960	300	500

TABLE 5 – LIMITS FOR EMISSION DUE TO FCC

LIMITING VALUES DUE TO EMISSION (CISPR)

The limiting values for the radiated signal due to CISPR are as show in Table 7.

Frequency Range (MHz)	Voltage Limits (µV/m)	
	Class A (Quasi-peak at 30m)	**Class B (Quasi-peak at 10 m)**
30 – 230	31.6	31.6
230 – 1000	70.8	70.8

TABLE 6 – LIMITS TO EMISSION AS RULED BY CISPR

NOISE

According to [7] power line communications apparatus requires withstanding a noise level in the mains port represented in the following table.

	Frequency band	Level	
Class 1 (residential)	0.15 MHz 80 MHz	3V	80% AM (1 KHz)
Class2 (industrial)	0.15 MHz 80 MHz	10V	80% AM (1 KHz)

The 80% AM modulated signal corresponds to,

$$v(t) = A\big(1 + 0.8\, s(t)\big)\sin(2\pi f_O t + \varphi)$$

Were A is the signal amplitude, $|s(t)| < 1$ is a 1 KHz signal and f_o is the carrier frequency. The carrier frequency, f_o, should vary in the whole frequency band. These are narrow interferences that can be avoided using adaptive techniques. Historically AM modulations signal should have a modulation index below 100% so that they could be demodulated using an envelope detector.

In [2] channel measurements were made obtaining an average value of $\boldsymbol{-140dBV^2}$ for the background noise. Narrow band noise is superimposed on the background noise. In [13] noise levels are measured for several appliances. The results agree well with values used [2].

SIGNAL MEASURMENTS (CISPR 16)

Both FCC part 15 and CISPR use CISPR 16 for defining the measurements techniques. The values above should be measured using a super heterodyne receiver tuned at the frequency of interest and with the following values for the bandwidth (as in a spectrum analyzer) [11]. Each frequency should be measured by about one second for the peak detector. Since the peak detector will deal with less constant signal for the average detector this can reduced to about 1ms.

Quasi peak detector	Band A	Band B	Band C	Band D
	0-150 KHz	0.15-30 MHz	30-300 MHz	300-1000 MHz
6 dB Bandwidth	200 Hz	9 KHz	120 KHz	
Charge time	45 ms	1 ms	1 ms	

constant			
Discharge time constant	500 ms	160 ms	550 ms
Pre-detector overload factor	24 dB	30 dB	43.5 dB

The overload factor is the relation of the linear range of the detector at the input and at the output. These need to be high because impulse signal have a much higher value at the input than at the output. The charge and discharge time are for quasi-peak detectors.

AN IMPLEMENTATION OF THE LEGISLATION LIMITS

In the **5–30 MHz** band two cases were considered, current legislation and a possible future changes to accommodate multi-use ports for telecommunications and power supply. Since the average limits are lower this will be chosen, and we will use the values for class B. The average is taken as being the average of the amplitude.

Using current legislation the limit for the voltage signal will be,

$$V_{30MHz} = \frac{50\ dB\mu V}{\sqrt{9\ kHz}} = 3.3uV/\sqrt{Hz}$$

The total signal in the band is 18.3 mV. The impedance of a LISN network is about $50\ \Omega$ so limiting the current for values obtained using a 10 Ohms is reasonable. This will result in a current limit of

$$I_{30MHZ} = 0.33uA/\sqrt{Hz}$$

Possible **changes to legislation** may treat power line communications as a special multipurpose port. In telecommunications ports only the common mode signal is limited, since this has the highest contribution to radiation. The common mode signal, injected by MODEM, is not limited, but it is considered that this signal is converted to common mode by the line, so the differential signal is limited indirectly. The conversion from differential mode to common mode is given by the Longitudinal Conversion Loss (LCL) as in,

$$\mathrm{LCL} = \frac{V_{Diferential}}{V_{Common\ mode}}$$

This is not exactly true, but applies well to PLC. LCL is expressed in dB. Power lines have lower symmetry than telecommunications lines and so the LCL should be lower in telecommunications ports. Also the irregular nature of the line will create zones, bends for instance, were the radiation from phase and neutral will not cancel out. A value of **36 dB** is being considered [12].

The limits for the common mode of the telecommunications ports are,

$$V_{30MHz} = \frac{64\ dB\mu V}{\sqrt{9\ kHz}} = 16.7 \frac{uV}{\sqrt{Hz}}$$

Using the proposed value for the LCL the differential mode voltage limit will be given by

$$V_{30MHz} = \frac{100\ dB\mu V}{\sqrt{9\ kHz}} = 1.05 \frac{mV}{\sqrt{Hz}}.$$

The signal in the full band will be given by 5.8V. The limits for the current are obtained by dividing by the ISN (Impedance Stabilization Network) impedance, 150Ω. Resulting in

$$I_{30MHz} = 7.03 \frac{\mu A}{\sqrt{Hz}}.$$

The current in the full band will be 38.5mA.

In the **30–88 MHz** band only exists limiting values for the radiated signal. Also there isn't much work on the LCL factor. However, results in [12] indicate that it is constant in the range from 5 MHz to 30 MHz so we will assume also a value of 30 dB for higher frequencies. We will use the limits for the values of the radiated field for calculate limits for the current injected in the line.

For a half-wave antenna the amplitude of the radiated field in the in direction of greater radiation is given by,

$$|E| = \frac{I\, Z_0}{2\pi\, r}$$

In the near field the equations for the fields are:

$$E_r = \frac{Z}{2\pi} I_0\, \delta l \left(\frac{1}{r^2} - i\,\frac{\lambda}{2\pi\, r^3}\right) e^{i(\omega t - k\, r)}\ \cos(\theta)$$

$$E_\theta = i\frac{Z}{2\lambda} I_0\, \delta l \left(\frac{1}{r} - i\,\frac{\lambda}{2\pi\, r^2} - \frac{\lambda}{4\pi^2\, r^3}\right) e^{i(\omega t - k\, r)}\ \sin(\theta)$$

$$H_\phi = i\frac{1}{2\lambda} I_0\, \delta l \left(\frac{1}{r} - i\,\frac{\lambda}{2\pi\, r^2}\right) e^{i(\omega t - k\, r)}\ \sin(\theta)$$

At frequencies of 30MHz the half wavelength is 5m, and since the field is measured at a distance of 3m or 10m the far field assumption still holds approximately and the results should give a good estimate of the electromagnetic field.

FCC limits the signal to 100uV/m at 3m while CISPR limits to 50dBuV/m or 40dBuV/m class B after conversion to 3m distance. Using the lower value of 100uV/m for a band of 120 KHz results in,

$$E_{88\mathrm{MHz}} = 100 \frac{\mu\mathrm{V/m}}{\sqrt{120\mathrm{KHz}}} = 0.29 \frac{\mu\mathrm{V/m}}{\sqrt{\mathrm{Hz}}}$$

Using

$$\mathrm{E} = \frac{\mathrm{Z_0\ I}}{2\pi\,\mathrm{r}} \cong \frac{60\ \mathrm{I}}{\mathrm{r}}$$

As calculated above. For r=3m, results in,

$$\mathrm{I_{88MHz\ CM}} = 14.5\frac{\mathrm{nA}}{\sqrt{\mathrm{Hz}}}$$

This is the common mode value. Using an LCL value of 36dB results in the differential mode value of

$$\mathrm{I_{88MHz\ DM}} = 0.915\frac{\mathrm{\mu A}}{\sqrt{\mathrm{Hz}}}$$

For 50Ω line impedance and a bandwidth of 88MHz this will result in a voltage signal of 0.43V.

Implementing the current limits requires an estimate of the impedance. In order to do this we propose to do the channel estimation in two phases. Phase one is mostly to estimate the impedance and phase two is to better estimate the channel.

Phase two limits are the maximum allowed by the legislation. Phase one limits are the following. In the 5-30MHz band current legislation values are used, since there is no limit for the current. In the 30MHz to 88MHz band an impedance of 5Ω is assumed for calculating the voltage limits.

	Phase	Current 30MHz	Voltage 30MHz	Current 88MHz	Voltage 88MHz
Current legislation	One	--	$3.3uV/\sqrt{Hz}$	--	--
	Two	$0.33\mu A/\sqrt{Hz}$	$3.3uV/\sqrt{Hz}$	--	--
Legislation changes	One	--	$3.3uV/\sqrt{Hz}$	--	$4.57\mathrm{\mu V}/\sqrt{\mathrm{Hz}}$
	Two	$7.03uA/\sqrt{Hz}$	$1.05mV/\sqrt{Hz}$	$0.915\mathrm{\mu A}/\sqrt{\mathrm{Hz}}$	--

ALTERNATIVE LIMITS FOR THE CURRENT

The limits presented above are reasonable. They assume common mode has the major radiation contribution and this is usually the case. For completeness however we present values for limits to the current value when we consider radiation from the differential mode.

If we have two conductors, like faze and neutral, the field will be given by,

$$|E| = \frac{I\,Z_0}{2\pi\,r} - \frac{I\,Z_0}{2\pi\,(r+d)} \cong \frac{I\,Z_0\,d}{2\pi\,r^2}$$

were d is the distance between conductors. For a r=3m distance a impedance 376.73 Ω of and using once more a conservative value of 0.1m for d and this will result in the following limiting values for the current in the conductors. However, these were not the values used.

Instead the LCL (Longitudinal Conversion Loss) was used to calculate the conversion differential to common mode. These values are still shown only completeness.

Frequency of Emission (MHz)	Current	
	Class A (at 10 m)	**Class B (at 3 m)**
30 – 88	1.50mA	0.15mA
88 – 216	2.51mA	0.23mA
216 – 960	3.51mA	0.30mA
Above 960	5.01mA	0.75mA

TABLE 7 – CALCULATED CURRENTS LIMITS DUE TO EMISSIONS USING VALUES BY THE FCC. THESE VALUES ARE NOT USED. INSTEAD VALUES BASED ON THE LCL WERE USED.

Frequency Range (MHz)	Current (mA/m)	
	Class A (Quasi-peak at 30 m)	**Class B (Quasi-peak at 10 m)**
30 – 230	4.74	0.53
230 – 1000	10.6	1.18

TABLE 8 – CALCULATED CURRENT LIMITS DUE TO EMISSION USING VALUES RULED BY CISPR. THESE VALUES ARE NOT USED. INSTEAD VALUES BASED ON THE LCL WERE USED.

SIMULATION

POWER LINE ATTENUATION PARAMETERS

The values chosen are taken from [1] from 100 meters line as in table 3, corresponding to a0=9.40e-3 e a1=4.20e-7, k=0.7.

This corresponds to an attenuation of -8.16 dB at DC and -55.2 dB at 20MHz and more or less class 2 lines in [2] and in [3].

FREQUENCY BAND

The sampling frequency of the MODEM was chosen to be 83 MHz, while the analogue channel was simulated using a sampling frequency of 880MHz.

IMPLEMENTATION

The MODEM takes a signal X[k] and calculates the complex base band signal x[n] (that is still going to be modulated) using the IFFT as in,

$$x[n] = \frac{1}{\sqrt{N}} \sum_{i=0}^{N} X[k] e^{\frac{2\pi kn}{N} i}$$

Were N is the IFFT length (number of carriers). The power of the signal will be

$$P = \langle |x[n]|^2 \rangle = \langle \sum_{k=0}^{N} |X[k]|^2 / N \rangle.$$

The notation $\langle . \rangle$ represents the average. The $Power/\sqrt{Hz}$ ($2 \times PSD$ for negative and positive frequencies) at given frequency is given by

$$2 \times PSD\left(f = \frac{2\pi k\ FA}{N} + f_O\right) = \frac{\langle X[k]^2/N \rangle}{FA/N} = \frac{\langle X[k]^2 \rangle}{FA}.$$

The constant FA represents the sampling frequency. In this expression k will take negative values.

The average values used for the limits of the signals will be taken as the rms (root mean square) resulting in,

$$X[k] = V_{RMS} \sqrt{FA}.$$

Implementing voltage and current limits implies that the impedance must be knows. Using a voltage source, the impedance is required to implement the current limit. If a current source were used the impedance would be required to implement the voltage limit. The impedance must be estimated using a probing that should be within the limits. In order to solve this we propose to do the estimation in two phases. In the first phase o lower value signal is used, low enough so that the current is bellow the maximum value for every expected values for the impedance. Then using the first estimate for the impedance the limits for the probing signal are derived and the channel and impedance estimate are refined.

In the band of 5MHz to 30MHz one will use the current legislation limits in phase one, and the possible new values in phase two.

In the band 30MHz to 88MHz band we will start by assuming worst case line impedance of 5Ω .

PSEUDO-CODE (5MHZ TO 30MHZ)

```
Phase1_voltage_limit = 3.3e-6       // V/Sqrt(Hz)
Phase2_voltage_limit = 1.05e-3      // V/Sqrt(Hz)
Phase2_current_limit = 7.03e-6      // A/Sqrt(Hz)

Voltage_signal = Phase1_voltage_limit * FA

// Phase One

Simulate

// Update the values of:
```

```
//      current_signal1
//      received_voltage_signal
//      signal_noise_ratio

Channel_estimate = received_voltage_signal / Voltage_signal;
Input_impedance = Voltage_signal / current_signal1;

// Phase Two
Voltage_signal = Min(Phase2_voltage_limit,  Phase2_current_limit * Input_impedance)*FA

Simulate
// Updates the values of:
//      current_signal1
//      received_voltage_signal
//      signal_noise_ratio

Channel_estimate = received_voltage_signal / Voltage_signal;
Input_impedance = Voltage_signal / current_signal1;

n_bits = floor(log2(1+ signal_noise_ratio/Gamma)/2)*2;
```

PSEUDO-CODE (5MHZ TO 88MHZ)

```
FA=88-5=83MHz

// Phase One Constants
Phase1_voltage_limit _30 = 3.3e-6           // V/Sqrt(Hz)
Phase1_voltage_limit _88 = 2.18uV           // V/Sqrt(Hz) for 5 Ohms
// Phase1_current_limit _88 = 14.5e-9       // A/Sqrt(Hz),
// the current limit for 88MHz is only used to determine the voltage limit

Phase1_voltage_limit is a vector with all the carrier frequencies
Phase1_voltage_limit(0..30Hz)= Phase1_voltage_limit _30
Phase1_voltage_limit(30..88Hz)= Phase1_voltage_limit _88

// Phase Two Constants
Phase2_voltage_limit _30= 1.05e-3     // V/Sqrt(Hz)
Phase2_current_limit _30= 7.03e-6     // A/Sqrt(Hz)
Phase2_current_limit _88 = 53e-6      // A/Sqrt(Hz)

Phase2_current_limit is a vector with all the carrier frequencies
Phase2_current_limit(0..30MHz)= Phase2_current_limit _30
Phase2_current_limit(30..88MHz)= Phase2_current_limit _88
Phase2_voltage_limit is a vector with all the carrier frequencies
Phase2_voltage_limit(0..30MHz)= Phase2_voltage_limit _30
Phase2_voltage_limit(30..88MHz)=Infinity
// although there are no voltage limits practical consideration may limit the total value
```

```
Voltage_signal = Phase1_voltage_limit * sqrt(FA)

// Phase One
Simulate
// Updates the values of:
//      current_signal1
//      received_voltage_signal
//      signal_noise_ratio

Channel_estimate = received_voltage_signal /Voltage_signal;
Input_impedance = Voltage_signal / current_signal1;

// Phase Two
Voltage_signal = Min(Phase2_voltage_limit, Phase2_current_limit * Input_impedance)* sqrt(FA)

Simulate
// Updates the values of:
//      current_signal1
//      received_voltage_signal
//      signal_noise_ratio

Channel_estimate= received_voltage_signal /Voltage_signal;
Input_impedance = Voltage_signal / current_signal1;

// Data transmission

n_bits = floor(log2(1+ signal_noise_ratio/Gamma)/2)*2;

Voltage_signal = Min(Phase2_voltage_limit, Phase2_current_limit * Input_impedance)* sqrt(FA)
```

UP SAMPLING AND DOWN SAMPLING FILTERS

Using FIR filters with a Blackman window. These filters have an attenuation of 74dB. The transition band is:

$$\frac{12\pi}{filter\_length} \times \frac{FA\_A}{2\pi} = 6\, FA\_A/filter\_length$$

This implies that the cut frequency should be in the middle of the transition band, which is in

$$FA/2 - 3\, FA\_A/filter\_length$$

For *filter_length*=512, and FA_A=830MHz (FA=83MHz) results in a transition band of 9.7MHz. The transition bands may result in obtaining incorrectly very high values for the impedance of the channel. So we set these values for a minimum value in the transition band. This will be the impedance guard bands. The number of impedance guard bands in each side of the useful band will be:

$$6\frac{FA\_A\ N}{FA\ filter\_length}$$

We will assume that at the guard bands the impedance will be

$$Z/10$$

ESTIMATING THE CHANNEL TRANSFER FUNCTION SIGNAL TO NOISE RATIO AND INPUT IMPEDANCE

ESTIMATION WITH IMPULSE NOISE

temos uma medida da forma:

Y=X+V

em que Y é o valor medido, X é o parametro que pretendo minimizar e V é o ruido (ruído impulsivo, não gaussiano).

Pretende-se obter ^X, um estimativa de X tal que

E[|^X-X|^p] seja minimo,

ou seja minimizar o erros na estimativa. Seria interessante ter p elevado, para minimizar o pior caso para o erro mas p=2 já não é mau.

Sabes me dizer qual é o estimador que resolve este problema, ou me indicar um artigo ou livro sobre isto?

Eu estou habituado a minimizar algo como ||Y-^X||^2 que é um problema diferente, a que corresponderia simplesmente a achar a média de Y, mas sei que por exemplo a mediana minimiza SUM_I|YI-^XI| e tem algumas vantagens para certos tipos de ruido.

$$E[(\hat{x} - X)^2|y] = \hat{x}^2 - 2E[X|y]\hat{x} + E[X^2|y]$$

Is the best estimate of the signal to noise ratio obtained by averaging? Improvements may be obtained by given less weight to samples with more noise.

PROBING SIGNALS

The Zadoff-Chu sequences have optimum correlation properties and are a special case of generalized chirp-like polyphase sequences [6][10]. They are also known as a CAZAC sequence (Constant Amplitude Zero Autocorrelation). Note that the sequences have complex values in the time domain. They are defined as:

$$S[k] = \begin{cases} e^{-\frac{2\pi r}{N}\left(\frac{k^2}{2}+qk\right)i} & N \text{ even} \\ e^{-\frac{2\pi r}{N}\left(\frac{k(k+1)}{2}+qk\right)i} & N \text{ odd} \end{cases}$$

where q is any integer (this is only a delay), $k = 0, \dots, N-1$ and r is any integer relatively prime to N. Two numbers are relative primes if the greatest common divisor is one, so $r = 1$ is always a solution. Their autocorrelation is zero for $n \neq 0$ and the time domain amplitude is one. This is better explained as follows.

If N is odd then the cross-correlation between two sequences formed using two different values for r is $\sqrt{N}$ [10]. This is a minimum for the cross-correlation of two zero autocorrelation sequences. If N is a prime number then the sequences for different values for r are all CAZAC and form the Zadoff-Chu set.

The frequency domains signal has constant amplitude,

$$\text{Abs}[S[k]] = 1$$

Defining the time domains signal as

$$s[n] = \frac{1}{\sqrt{N}} \sum_{i=0}^{N} S[k] e^{\frac{2\pi k n}{N} i}$$

As used for OFDM modulator and that makes the time domain power equal to the frequency domain power (Parseval relation). This implies that the periodic autocorrelation function,

$$r[n] = \sum_{i=0}^{N} \tilde{s}[i] \text{Conjugate}[\tilde{s}[i+n]]$$

were $\tilde{s}[i]$ is the periodic repetition of $s[i]$, is

$$r[n] = \begin{cases} N & \Leftarrow n = 0 \\ 0 & \Leftarrow n \neq 0 \end{cases}$$

The amplitude of the time domain signal is also unit,

$$\mathrm{Abs}[s[n]] = 1$$

The parameters values used in the project were,

$$r = 1$$

$$N = 1024$$

$$q = 0$$

TABLE 9 – PARAMETERS FOR THE CAZAC SEQUENCES

The sequence generated according to Table 9, are represented in Figure 1.

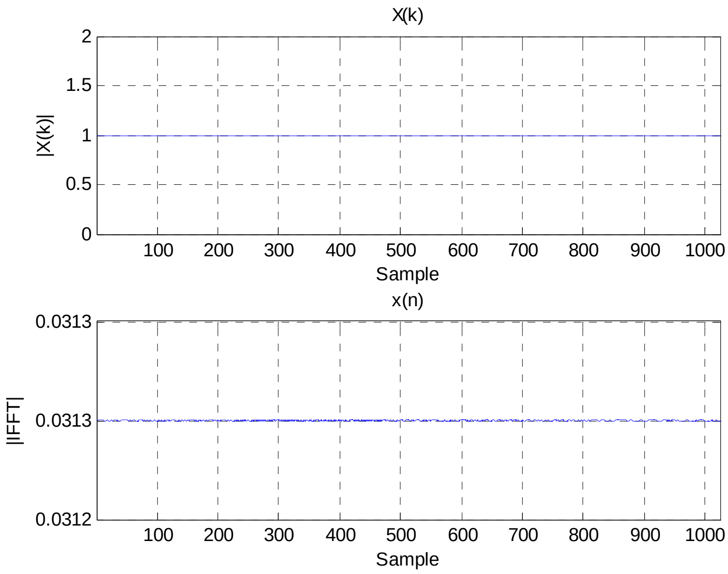


Figure 1 – Generated Sequence

The constants frequency domains and time domain amplitude of the CAZAC sequences make then good training sequences for OFDM systems, allowing a simple implementation of the receiver and maintaining a unity peak to average power ratio.

SOURCE CODING

n – code word length (in symbols)

q=2^b symbols (for binary codes q=2)

a code is forming by selecting

2^k code words

from the possible values.

(n,k) code

Code rate Rc = k / n, corresponds to the decrease of symbol rate due to coding

Redundancy symbols are n - k

At the modulator **k** symbols are mapped to **n** symbols

At the demodulator **n** symbols are mapped to **k** symbols

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Bose-Chaudhuri-Hocquenghem (BCH) codes

n=2^m-1

n-k<= mt, m>=3, (in the optimum case we have n-k=2t, so they are not optimum)

d_mim=2t+1

In the table t are the bits the code corrects.

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Reed-Solomon Codes are a subset of BCH codes. There is not much about the codes. They are non binary and

N=q-1=2^k-1

K=1, 2, 3, ..., N-1

Dmin = N-K+1 (they are optimum)

Rc = K/N

k is the number of bits per symbol

Were k (not caps) is the number of bits for a symbol.

Corrects up to

t=1/2 (Dmin-1) = 1/2(N-K)

Symbols (not bits!!!)

There are efficient hard decisions decoding algorithms.

PROBABILITY OF BIT ERROR WITH ERROR CORRECTING

In a word of dimension n, the probability of having t errors or more is, (P(A+B)=P(A)+P(B)-P(A.B)).

(Pe.n)^t (1-Pe)^(n-t) (n*...*(n-t+1)) ≈ (Pe.n)^t

As long as Pe.n < 0.1

Reed Solomon codes can correct more than two t errors if the errors are in the same symbol, but only t in the worst case.

The BCH code with n=63, k=45, t=1, g=1701317. How complex is this???

REFERENCES

1. Manfred Zimmerman and Klaus Dostert, "A Multipath Model for the Powerline Channel," IEEE Transaction on Communications, vol. 50, n. 4, 2002
2. Langfeld, Patrick J., "The Capacity of Typical Power Line Channels and Strategies for System Design," Proc. 5th Int'l. Symp. Power-Line Commun., Malmö, Sweden, 4.4.-6.4. 2001, pp. 271–78.
3. Matthias Götz, Manuel Rapp, and Klaus Dostert, "Power Line Channel Characteristics and Their Effect on Communication System Design," IEEE Communications Magazine, vol. 42, n. 4, pp. 78-86, April 2004
4. ETSI Systems Reference Document; Broadband Powerline Communication; Workitem V1.5; http://www.etsi.fi; 2000-06
5. M. de Abreu Faro, "Propagação guiada," _______
6. David C. Chu, "Polyphase Codes With Good Periodic Correlation Properties," IEEE Transactions on Information Theory, July 1972
7. "Power line communications apparatus and systems used in low-voltage installations in the frequency range 1,6 MHz to 30 MHz," CENELEC EN 50412-2-1:2005
8. Christian Hensen, Sascha Schwarze, "CISPR 22 Compliance Test of Power-Line Transmission Systems," International Symposium on Power Line Communications and Its Applications (ISPLC), Athens, Greece, March 2002
9. FCC part 15
10. Branislav M. Popovic, "Generalized chirp-like polyphase sequence with optimum correlation properties", IEEE Transaction on Information and Theory Vol 38 Nº4, July 1992 pp: 1406-1409
11. Tim Williams, "EMC for product designers," 4th edition, Newnes, 2007
12. Michel Ianoz, "Standardisation Problems Related to Conducted Limits for Power Line Communication Equipment," IEEE International Symposium on Electromagnetic Compatibility, EMC 2003, Istanbul, Turkey, 2003
13. H. Meng, Y. L. Guan, "Modeling and Analysis of Noise Effects on Broadband Power-Line Communications," IEEE Transaction and Power Delivery, Vol. 20, No. 2, April 2005
14. Pozar, David M. Microwave Engineering, John Wiley & Sons, Inc., 2005.