DVB-T

Receiver

DVB T
TERRESTRIAL

1.
Introduction

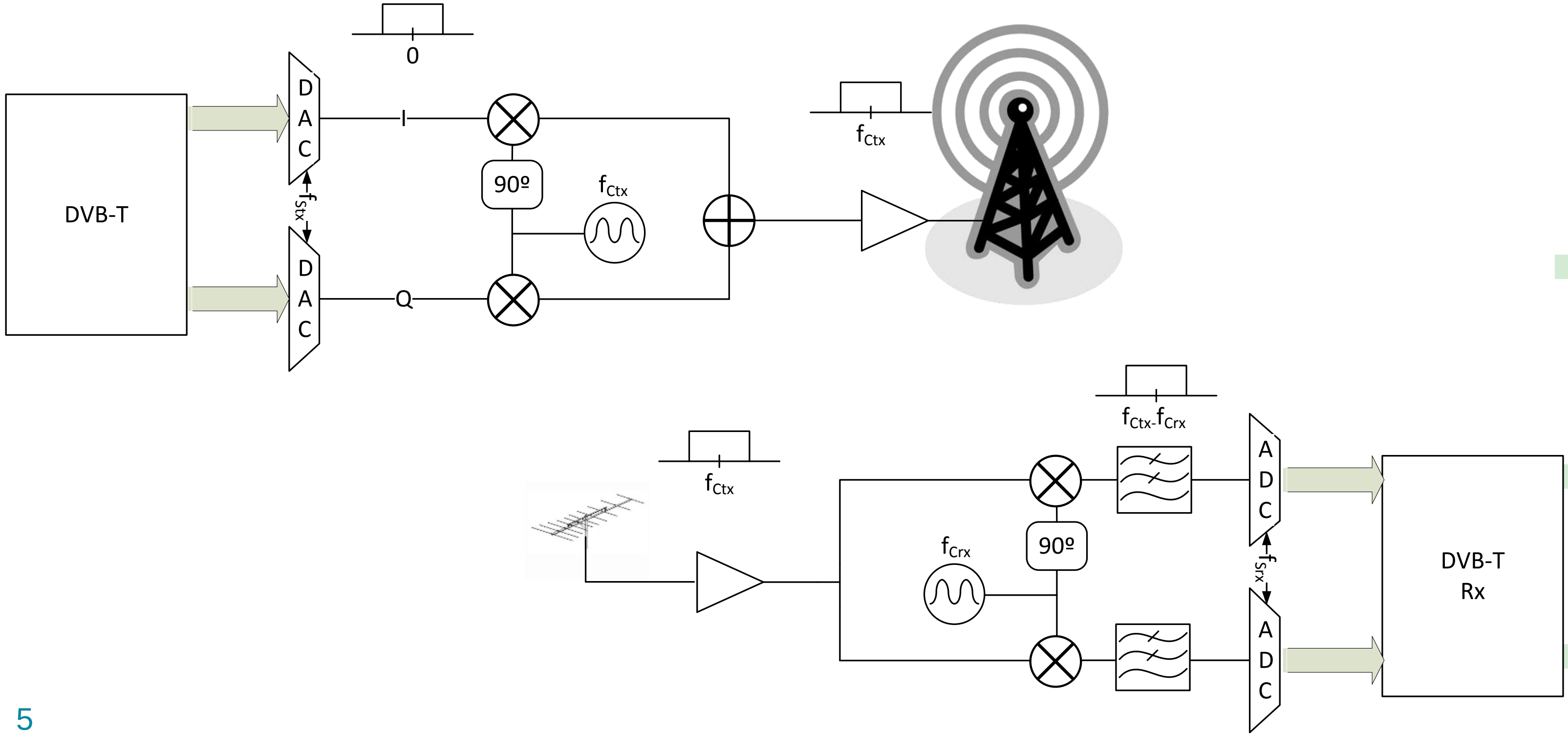
- The implementation of the receiver is not included in the DVB-T standard
  - Only the transmitter is fully detailed

- However the DVB Project provides as additional information the Implementation Guidelines
  - A guide to follow when implementing the receiver

http://www.etsi.org/deliver/etsi_tr/101100_101199/101190/01.03.02_60/tr_101190v010302p.pdf

2.
Frequency offset

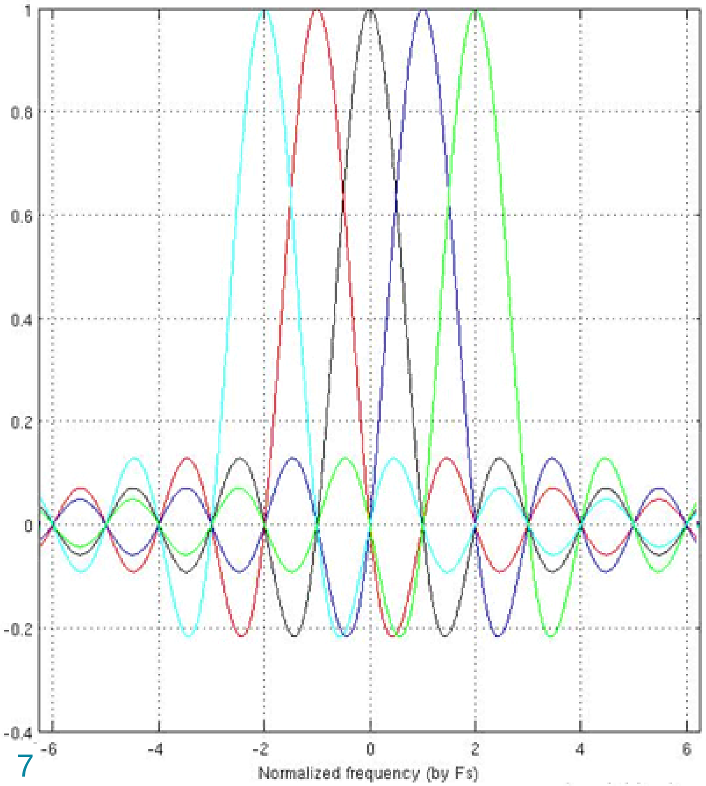
After the digital blocks of the transmitter



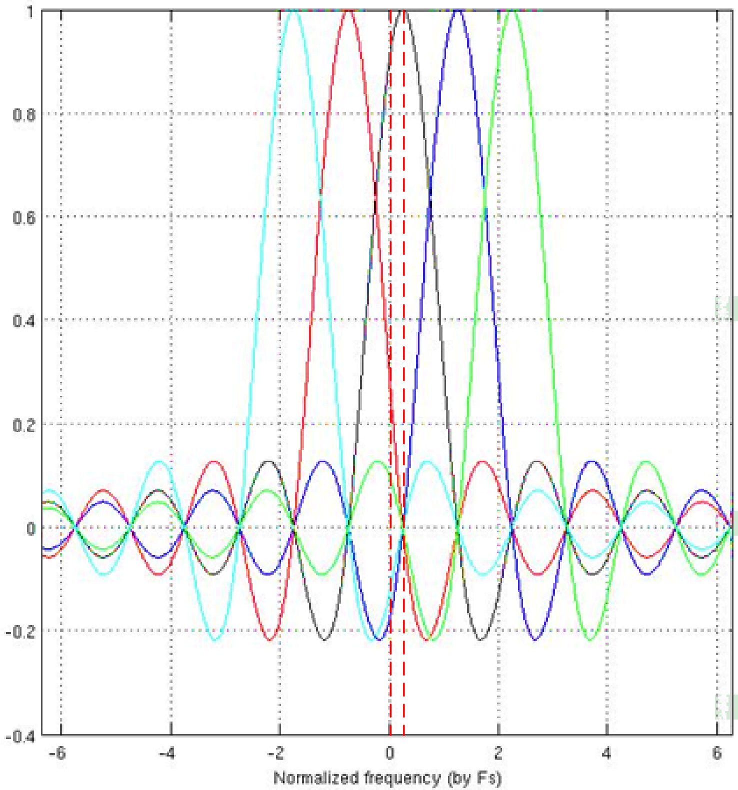
- Due to the implementation derive, frequencies at the Tx and Rx may differ even being nominally the same
  - Frequency offset $\Delta f = F_{Tx} - F_{Rx}$
  - This can vary with the temperature (drift)
  - Makes necessary a tracking

- The frequency offset error introduces ICI

Perfect synchronization (f_{Tx}-f_{Rx}=0)



Synchronization error (f_{Tx}-f_{Rx}=Δf != 0)



- The carrier separation is $1/T_s$

- In a general expression $\Delta f = n(1/T_s) + \delta f$
  - n is an integer
  - δf is a real number in $[-1/2T_s,\ 1/2T_s]$
- n doesn't generate ICI, but the carrier reference is lost
  - The 0 carrier becomes the nth
- δf causes the lost of orthogonality and the ICI
  - Needs to be corrected before working in the frequency domain

- Two approaches can be followed to correct it
  - Act over the analogue oscillator
  - Digital correction
- As a general truth it is better to work in the digital domain

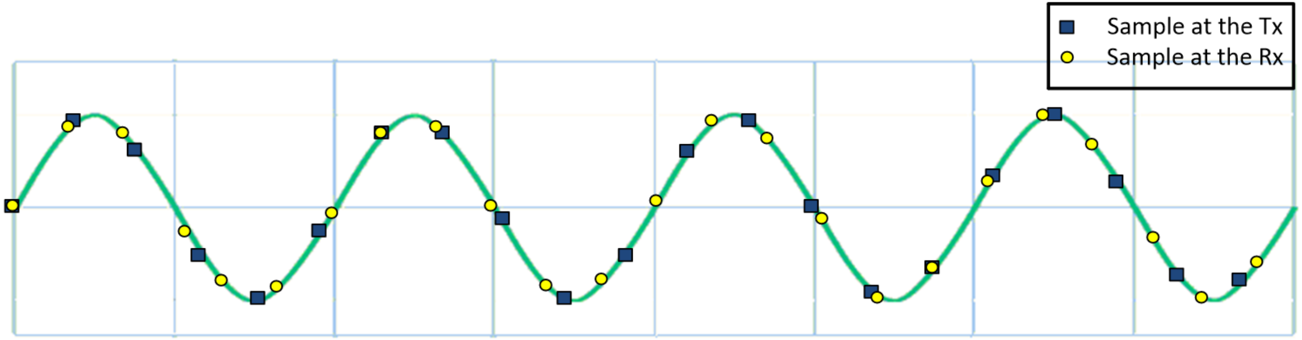
- If we consider
  - Desired signal $s(t)$
  - Received signal $s_{rx}(t) = s(t)e^{j2\pi\Delta f t}$

- If we knew the value of Δf, we would only need to multiply $s_{rx}(t)$ by $e^{-j2\pi\Delta f t}$, obtaining the desired signal

- Digitally I need to multiply $s_{rx}(kT)$ by $e^{-j2\pi\Delta f kT}$, being T the sampling time
  - By using a DSS (Digital Signal Synthesizer) we generate sine and cosine to generate the complex exponential

3.
Sampling frequency offset

- Again the difference between clocks
  - DAC at the Tx and ADC at the receiver have a frequency deviation
  - $f_{sampRx} = f_{sampTx}(1 + \delta)$
  - δ is measured in ppm (parts per million)
    - Depends on the precision of the crystal in the clock (typical value 50 ppm)
    - Varies in the time due to temperature fluctuations
      - A tracking of the offset is needed



- In the example the clock at the receiver is faster
  - The receiver will end up taking samples of a different OFDM symbol (ISI) due to the lost of the temporal synchronism
  - The first period has 6 samples at the transmitter and 7 at the receiver
  - For the receiver the signal is 6/7 slower
  - This produces a spectrum compression/expansion

- The effects of this are:
  - ISI, because of the lost of temporal synchronism
  - ICI, because of the expansion/compression
- However:
  - The deviation will be as higher as 100ppm
  - For a 10MHz clock this implies an error of ±1kHz
  - A sample of every 10000 is lost
  - The error will take a lot of time to be noticeable
  - A sample deviation in more than 2000 samples symbol is not a lot

- To ease this problem interpolation can be used to know what point was transmitted
    - Interpolation (Lagrange coefficients)

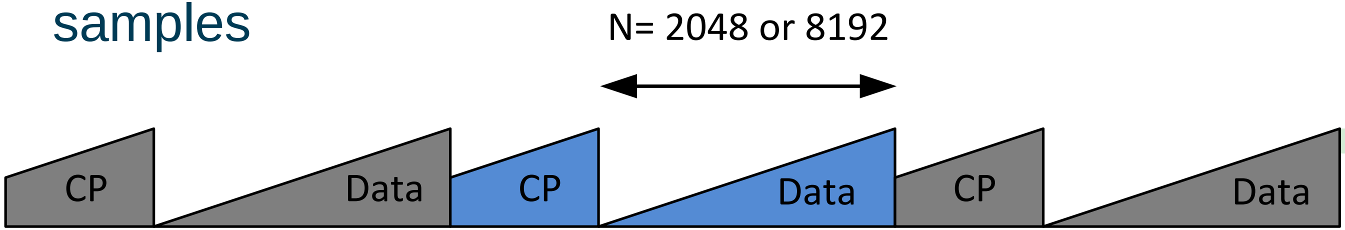
- This implies that the deviation (δ) of the sampling frequency offset needs to be estimated

2.
Synchronization

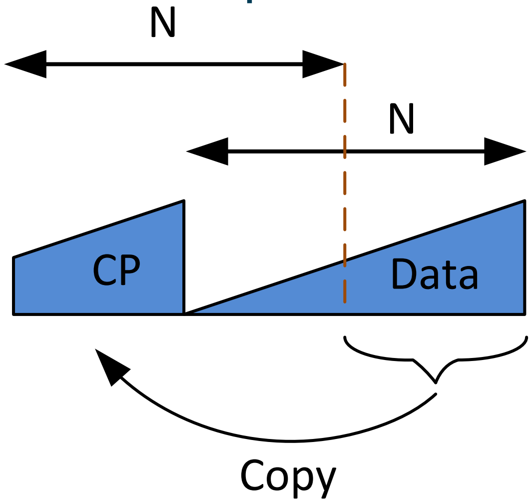
- When the reception starts:
  - The transmission mode is unknown (2k or 8k?)
  - The cyclic prefix is unknown (1/32, 1/16, 1/8, 1/4)
  - The temporal beginning of the symbol is unknown
  - Frequency offset error
  - Sampling frequency offset error
- This is all a mess!

- The synchronization is carried out by performing the following process
  - Time domain
    - Mode detection
    - Cyclic prefix detection
    - Coarse time synchronization
    - Fine frequency synchronization
  - Frequency domain (AFTER THE FFT!)
    - Coarse frequency synchronization
    - Frequency and sampling frequency offset error tracking
    - Fine time synchronization

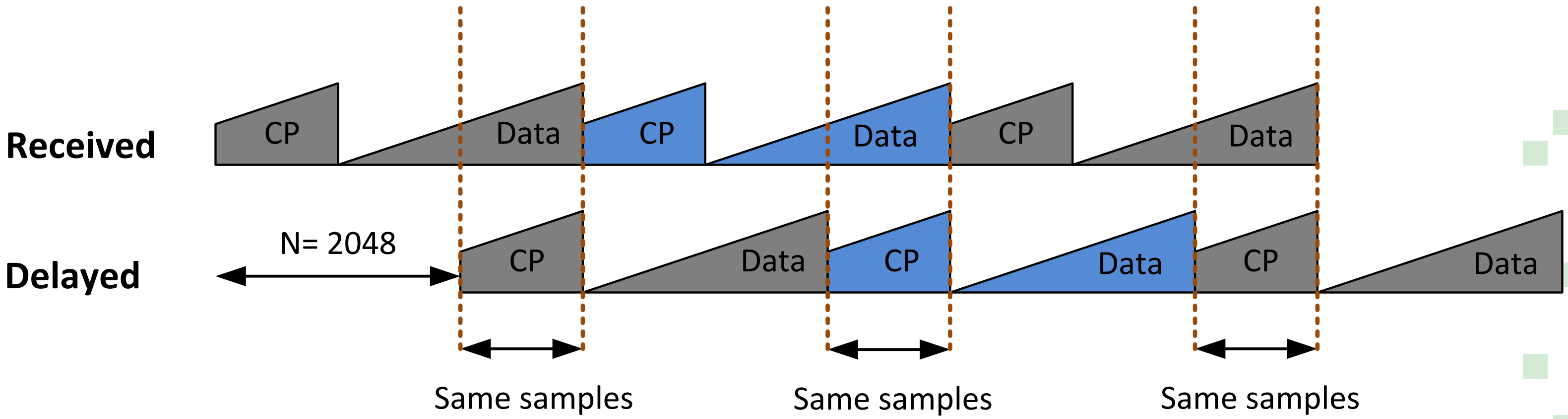
- Mode detection
  - Transmitted symbols may have 2048 or 8192 samples



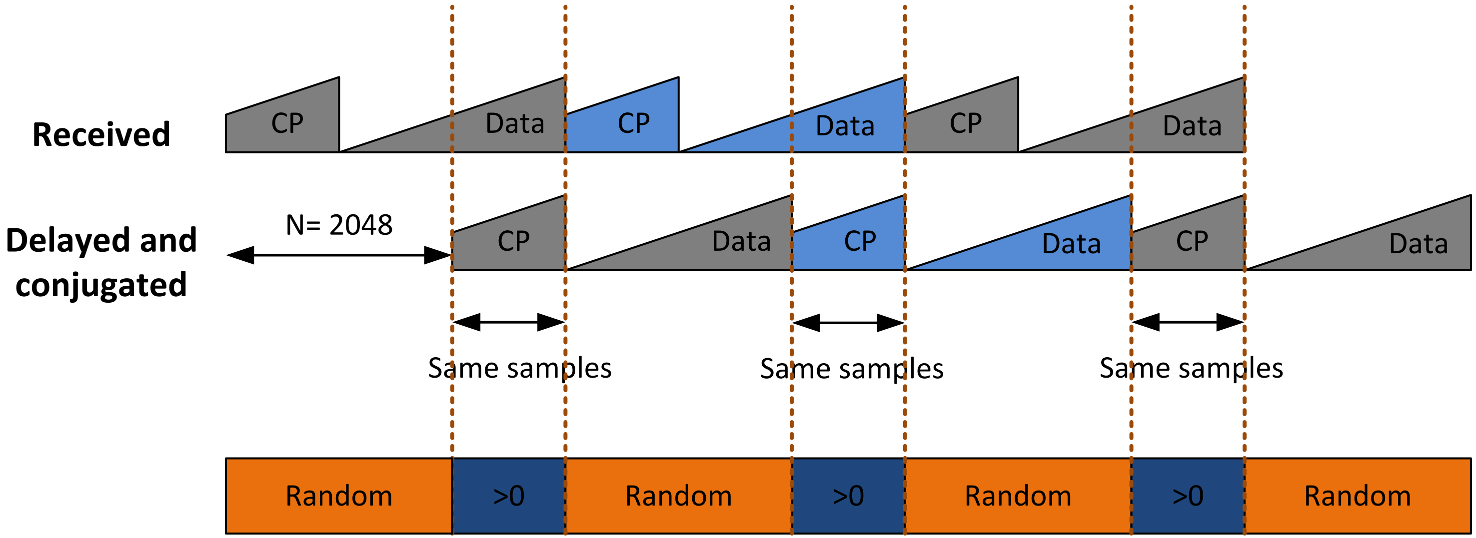
  - The CP is a copy of the last samples in the data part of the symbol



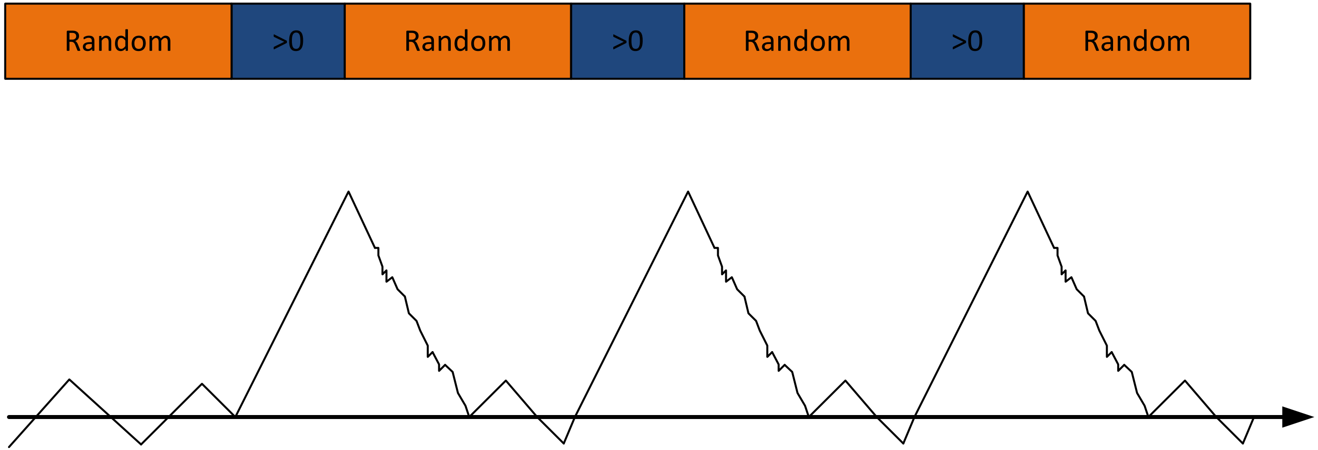
- Lets suppose that we have a 2k mode transmission (N=2048)
- By delaying the received signal 2048 samples (FIFO with length 2048)



- By multiplying the conjugated delayed version of the received signal by the original received signal
  - In the zone with same samples the result is the modulus (always greater than 0)
  - In the other zones the product will be a random complex number



- Integrating the result in a window of length the number of samples in the CP (NCP)



- Mathematically this corresponds with the autocorrelation of a window of NCP samples

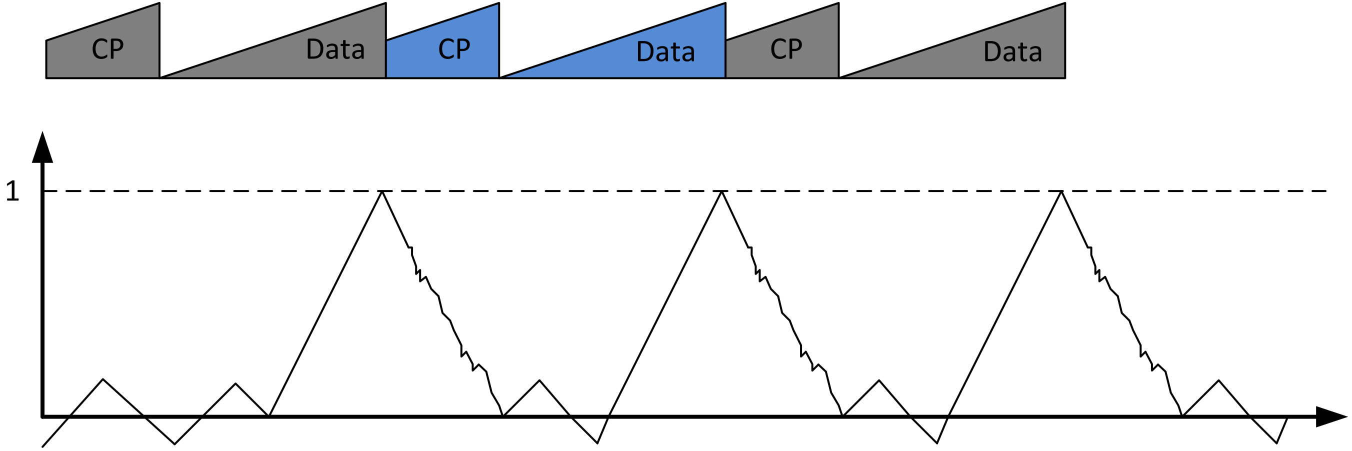
$$y(k) = \sum_{n=0}^{NCP-1} s(k-n)s^*(k-n-N)$$

- We can also define the energy of the signal as follows

$$e(k) = \sum_{n=0}^{NCP-1} s(k-n)s^*(k-n)$$

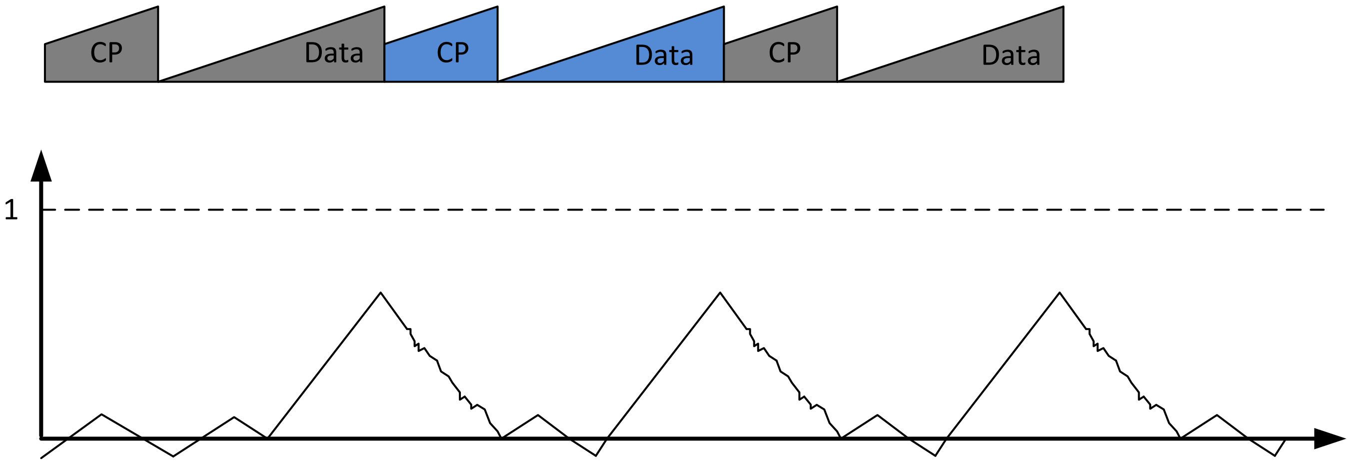
  - $e(k)$ is always positive and real
  - $e(k)$ and $y(k)$ only coincide in one sample, the last of the symbol

- We can normalize $y(k)$ by $e(k)$

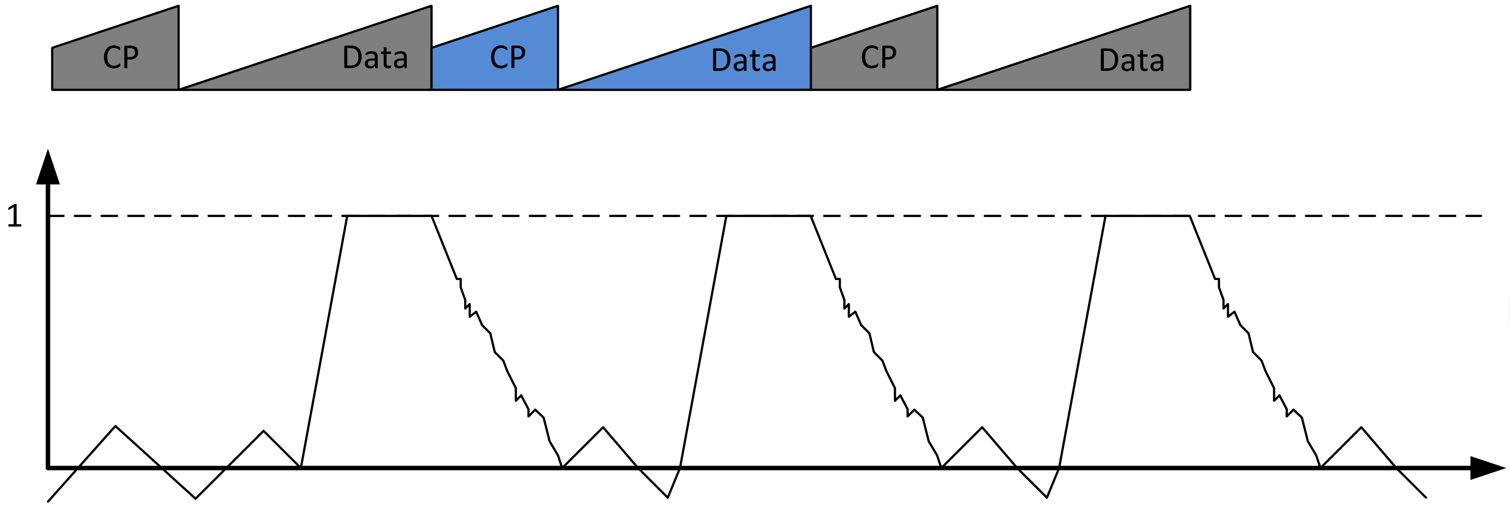


- Ideally the peaks will reach 1, but because of noise and interferences this won't happen, 0.5 is used as threshold

- In all this we have supposed that we know NCP, but in a real situation WE DON'T KNOW!
- If any NCP is fixed and the used one is smaller
  - The integration of the energy, e(k), is bigger than the peak obtained in the autocorrelation, $y(k)$

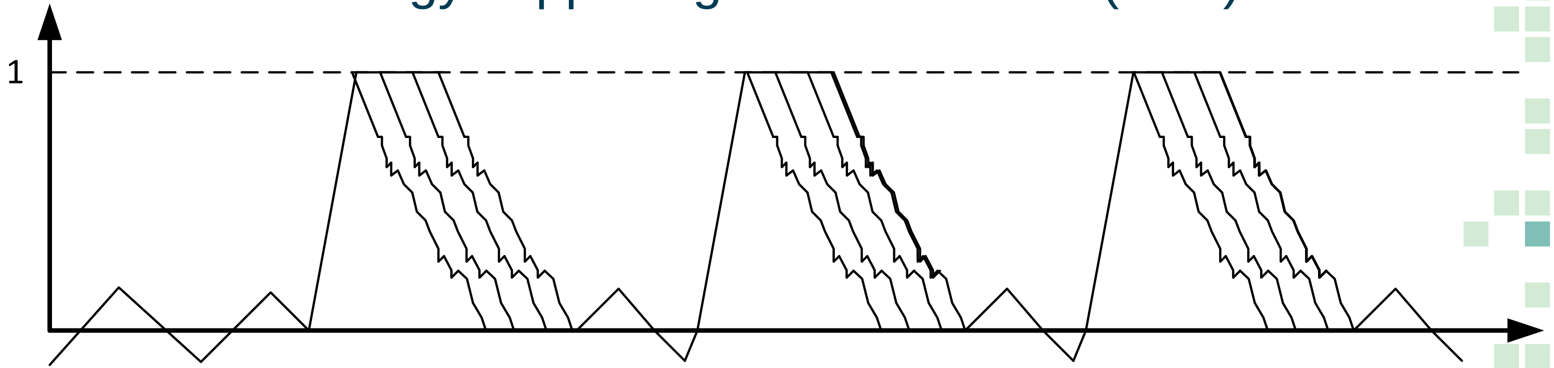


- If the opposite happens and the received CP is bigger than the one we set a mesa appears instead of a peak



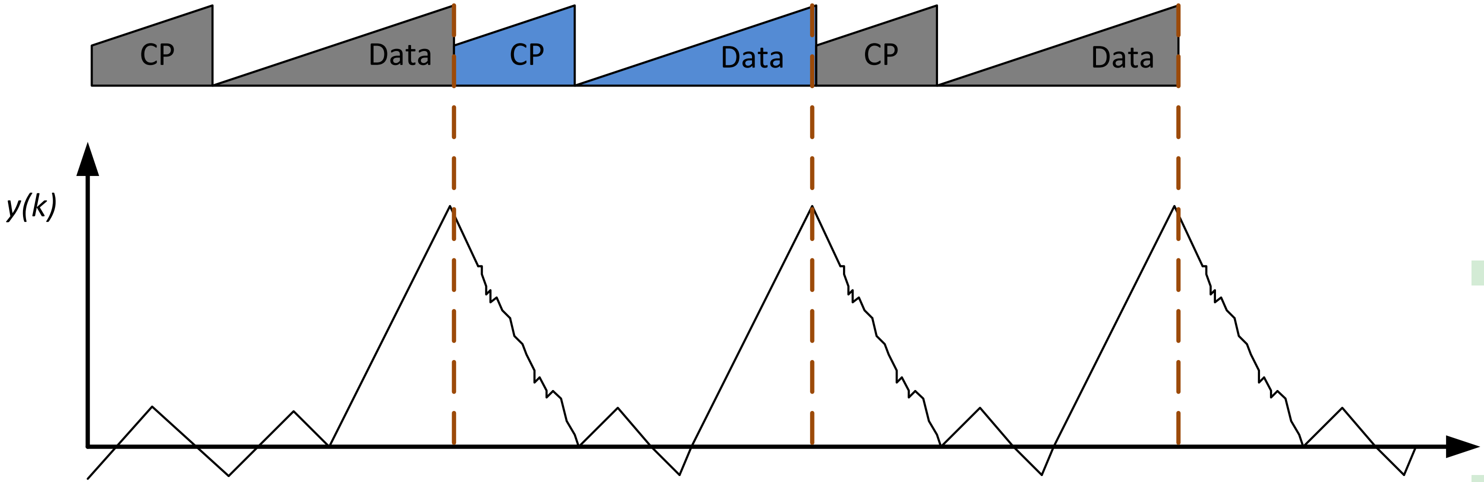
- As a method to detect the mode we can
  - Set as default the 2k mode with CP=1/32 (the smallest one, we don't want to have the situation where the peak is attenuated by averaging, we are using a threshold to detect)
  - Calculate the autocorrelation and the energy
  - If $y(k)/e(k) \geq 0.5$ the mode is 2k
  - After a timeout the 2k mode is not found change to 8k and CP=1/32 and repeat the process
  - If the time out is again surpassed … maybe there is no signal in the air!

- Cyclic prefix detection
  - The mode is already known
  - Again the autocorrelation is calculated as well as the energy supposing the smaller CP (1/32)



  - Counting the samples in the mesa (until the normalized autocorrelation is lower than the threshold) we can guess what CP the transmitted signal is using

- Coarse time synchronization
  - Already knowing the mode and the CP
  - Again we calculate the autocorrelation



  - The peaks coincide with the last data sample of a symbol

- As a possible algorithm to have coarse time synchronization
  - Calculate the autocorrelation and energy knowing the mode and the CP
  - Define $y_{max} = 0$, $n = 0$, and $N_{SYM} = N + NCP$
  - As long as $y(k) < 0.5e(k)$ idle
  - Otherwise
    - If $y(k) > y_{max}$, $y_{max} = y(k)$ and $n = 0$
    - Else $n = n + 1$
    - If $n == N_{SYM}$ the next sample is the first sample of a CP of a DVB-T OFDM symbol

- The presented method has some problems
- The maximum is found in one sample
  - Ideally this would lead to a perfect synchronization
  - The transmitted signal is immersed in a noisy environment, the sample will not usually be the one we expected
- However this will be enough to allocate the FFT window

- Fine frequency synchronization
  - The carrier separation is $1/T_s$
  - In a general expression $\Delta f = n(1/T_s) + \delta f$
    - n is an integer
    - δf is a real number in $[-1/2T_s,\ 1/2T_s]$
  - n doesn't generate ICI, but the carrier reference is lost
  - δf causes the lost of orthogonality and ICI
    - Needs to be corrected before working in the frequency domain
  - The remaining par t of the frequency offset can be fixed afterwards

- Lets suppose that the received signal has a frequency offset of Δf :

$$s_{rx}(t) = s(t)e^{j2\pi\Delta f t}$$

- In discrete form

$$s_{rx}(k) = s(k)e^{j2\pi\Delta f kT}$$

  - Being T the sampling time
- We can rewrite the expression obtained for the autocorrelation

$$y(k) = \sum_{n=0}^{NCP-1} s_{rx}(k-n)s_{rx}^{*}(k-n-N) =$$

$$= \sum_{n=0}^{NCP-1} s(k-n)s^{*}(k-n-N)e^{j2\pi\Delta f(k-n)T}e^{-j2\pi\Delta f(k-n-N)T}$$

- Analysing the previous expression

$$e^{j2\pi\Delta f(k-n)T}e^{-j2\pi\Delta f(k-n-N)T} = e^{j2\pi\Delta fNT}$$

- Taking again that In a general expression $\Delta f = m(1/T_s) + \delta f$ and applying that $T_s = NT$

$$e^{j2\pi\Delta fNT} = e^{j2\pi\left(\frac{m}{NT}+\delta f\right)NT} = e^{j2\pi m}e^{j2\pi\delta fNT} = e^{j2\pi\delta fNT}$$

- And returning to the expression fore the autocorrelation

$$y(k) = \sum_{n=0}^{NCP-1} s(k-n)s^*(k-n-N)\,e^{j2\pi\delta fNT}$$

- If we consider the expression for the last sample

$$s(k-n) = s(k-n-N) \qquad 0 \le n \le NCP-1$$

$$s(k-n)s^*(k-n-N) = |s(k-n)|^2 \qquad 0 \le n \le NCP-1$$

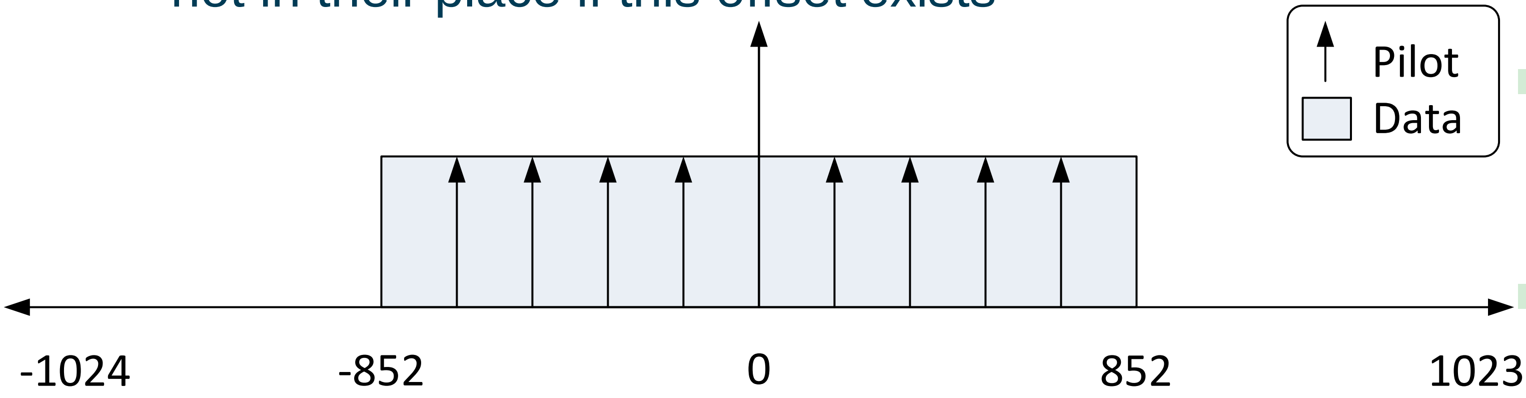
$$y(k) = \sum_{n=0}^{NCP-1} |s(k-n)|^2 \, e^{j2\pi\delta f NT} = Ke^{j2\pi\delta f NT} \qquad K \in \Re$$

- This position is the maximum of the autocorrelation already evaluated in the coarse time synchronization
- The estimation for δf is:

$$\delta f = \frac{\mathrm{angle}(y_{max})}{2\pi NT}$$

- Coarse frequency synchronization
  - We already know the mode an CP
  - We have an estimation for the beginning of the symbols
    - We know the samples to apply the FFT (bounding the ISI)
  - The fine frequency offset has been estimated and can be corrected by means of a DSS
    - The ICI is eliminated
  - We can already apply the FFT and perform the rest of the synchronization in the frequency domain

- In a general expression $\Delta f = n(1/T_s) + \delta f$
- δf has been already corrected
- The remaining to correct is a entire number of carriers shift
- Continual pilots are used for this purpose, if they are not in their place if this offset exists



OFDM symbol in the frequency domain (2k mode)

- As stated in the standard the location of the continual pilots is known and fixed always with the same information
- We denote as $S_n(k)$ the nth transmitted symbol (in the frequency domain)
- In reception $S_n^{rx}(k) = S_n(k)H(k)$, being $H(k)$ the frequency response of the channel, that can be expressed as $H(k) = h(k)e^{j\theta(k)}$
- Computing the following product

$$S_n^{rx}(k)S_{n-1}^{rx}(k)^* = S_n(k)S_{n-1}(k)^* h(k)^2$$

- In general this expression will be a random complex number, but in the continual pilot positions

- The transmitted pilots use a BPSK constellation [-4/3, 4/3]
- If *P* is the set of points corresponding to the continual pilots location

$$S_n(k)S_{n-1}(k)^* = 16/9 \quad k \in P$$

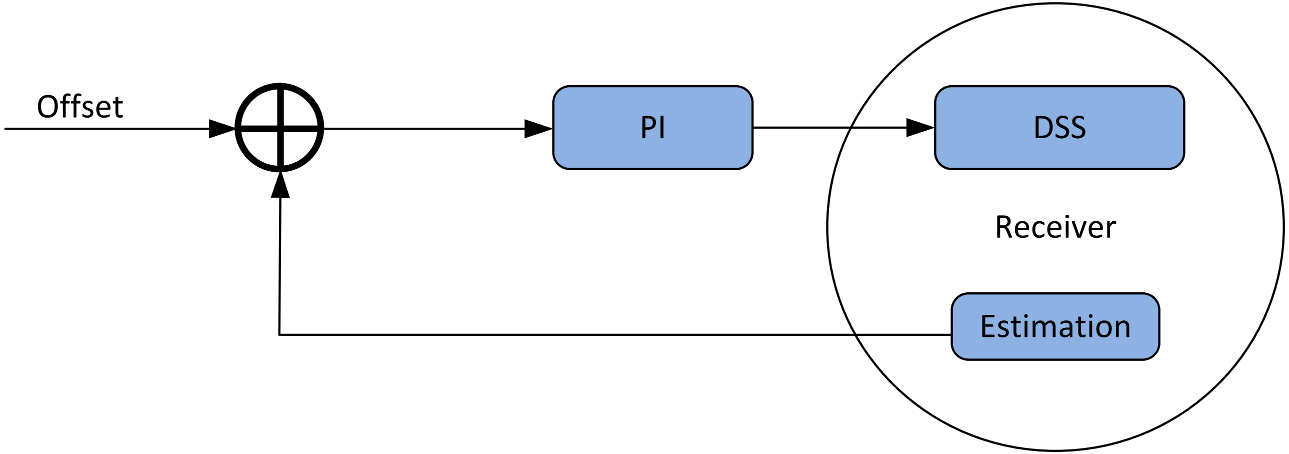
$$S_n^{rx}(k)S_{n-1}^{rx}(k)^* = S_n(k)S_{n-1}(k)^* h(k)^2 = (16/9)h(k)^2 \quad k \in P$$

  - A real and positive number
- Otherwise the result will be a random complex number
- If there is a frequency offset the pilots will be in *P+m* instead of in *P*

- A possible algorithm to correct the coarse frequency offset
  - We define the search interval $m \in [-M_{max}, M_{max}]$
  - The sum $S_n(k)S_{n-1}(k)^*$ for $k \in P + m$ is calculated
  - If the sum is higher than for the previous maximum the value and the index m are stored
  - The final maximum's index will be the frequency offset
  - The frequency offset will be corrected with a DSS

3.
Frequency tracking

- The frequency offset varies in the time (drift)
- With the previous method we have stated an instantaneous offset value
- This needs to be tracked to correct its possible deviations
- A PI (proportional integrator) control is used



- Frequency offset estimation
  - We use a similar process to the one for coarse frequency
  - $S_n^{rx}(k)S_{n-1}^{rx}(k)^* = S_n(k)S_{n-1}(k)^* h(k)^2$
  - The result should be a positive real number for $k \in P$
  - If a frequency offset exists the phase of the previous product is different from zero and proportional to the frequency offset

- The algorithm for the frequency offset tracking
  - For every couple of received symbols
    - The sum Y of the products $S_n(k)S_{n-1}(k)^*$ for $k \in P$ is calculated
    - The frequency offset is $\Delta f = \frac{angle(Y)}{2\pi\left(1+\frac{NCP}{N}\right)}$
    - This estimation actuates over the DDS through the PI control
  - The control loop constants must be calculated taking into account
    - Must be fast to follow the drift of the clock
    - Must be slow in comparison with the channel temporal fadings

- The sampling frequency offset also needs to be tracked and corrected
- The correction is very similar to the frequency offset one but instead of a DDS a Farrow filter is applied
- However it is much more complex

4.
Channel estimation

- After the channel propagation

$$S_{rx}(f) = S(f)H(f) + n(f)$$

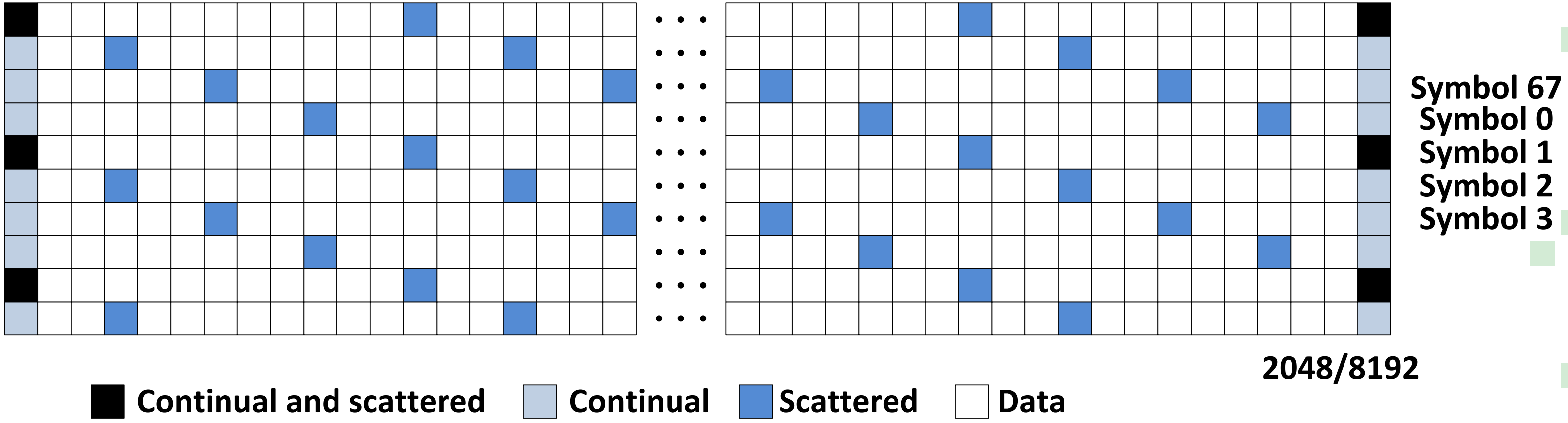
  - Being $n(f)$ the noise in the channel
- In the discrete domain (after the FFT)

$$S_{rx}(k) = S(k)H(k) + n(k)$$

- In order to demodulate the received signal $H(k)$ must be calculated
  - $\widetilde{H}(k)$ represents the estimation of $H(k)$
  - The estimation of $S(k)$ is obtained by equalizing

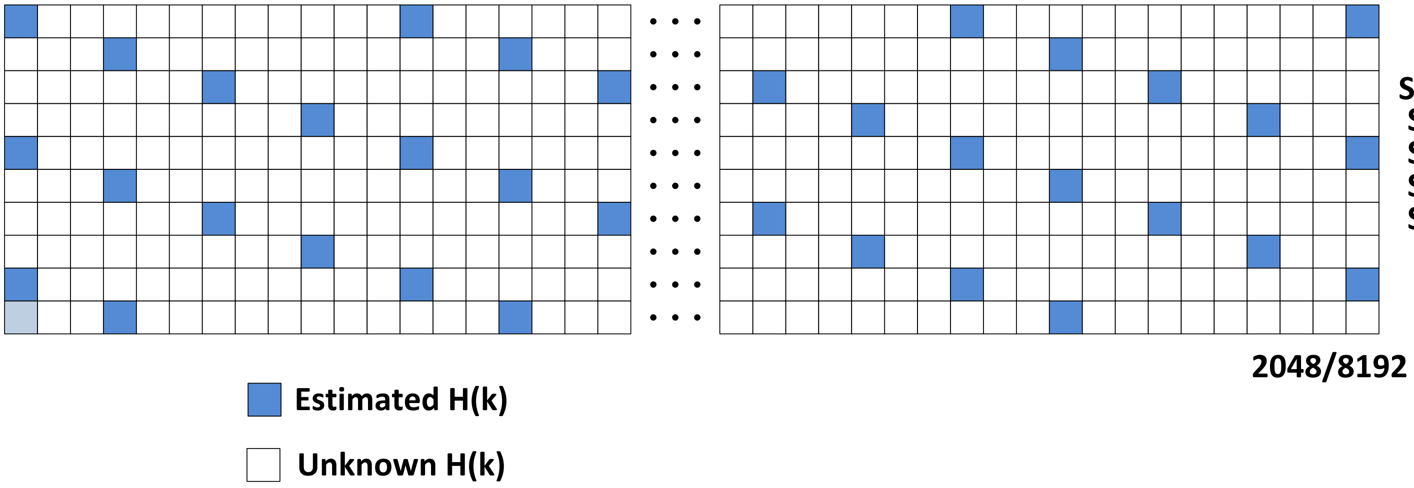
$$\tilde{S}(k) = S_{rx}(k)/\widetilde{H}(k) = S(k)H(k)/\widetilde{H}(k) + n(k)/\widetilde{H}(k)$$

- The channel is estimated by using the scattered pilots inserted in the OFDM symbol

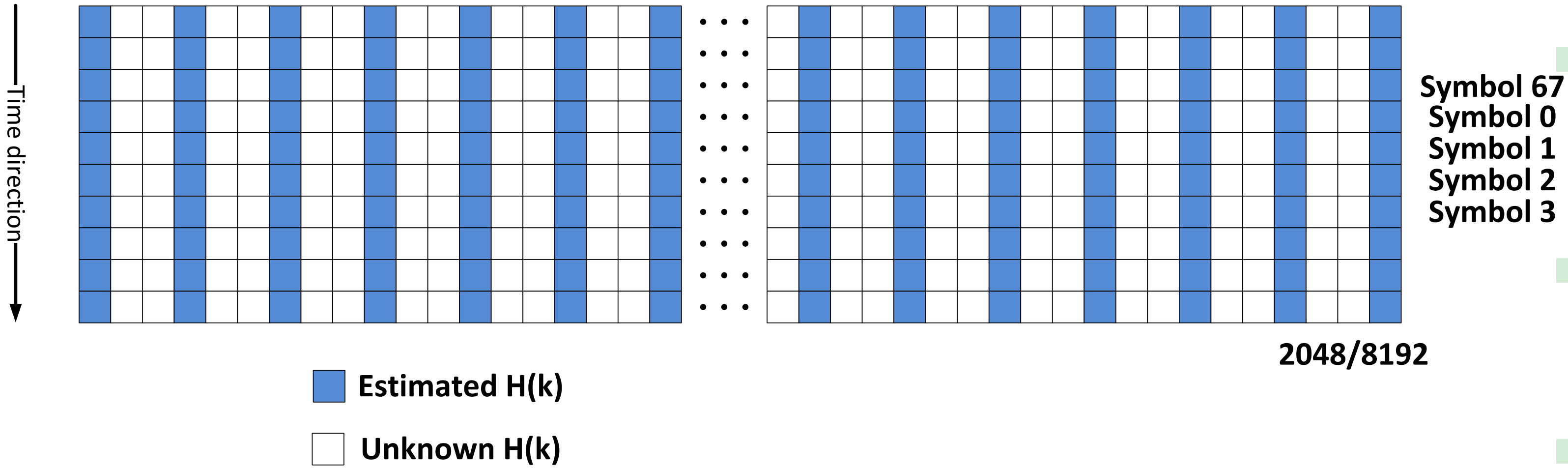


- At the receiver $S_{rx}(k) = S(k)H(k) + n(k)$
  - If k is a scattered pilot carrier, the value of $S(k)$ is known
  - So the estimation in those positions can be obtained by dividing by the value of $S(k)$

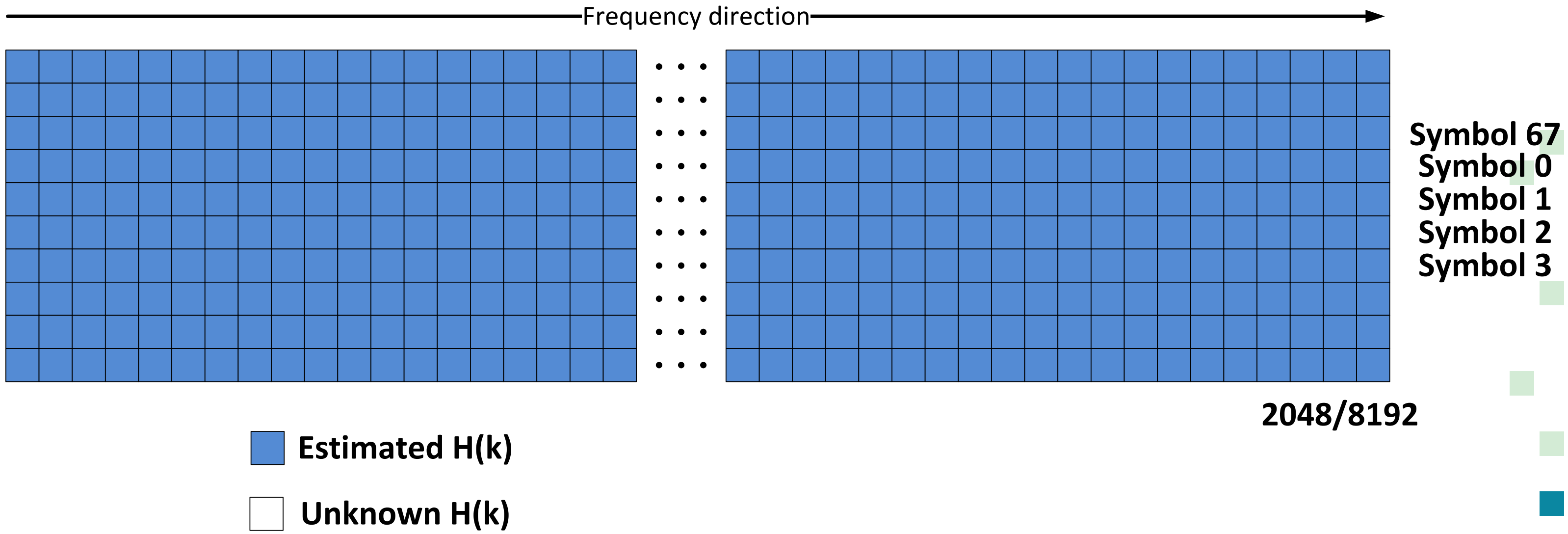
$$\widetilde{H}(k) = S_{rx}(k)/S(k)$$



- The remaining positions are estimated in two steps
  - Channel estimation in time direction
  - Channel estimation in frequency direction
- For the temporal estimation a carrier k is fixed and by means an interpolator filter the unknown positions are calculated



- After the temporal interpolation, a frequency interpolation is carried out
  - A symbol is selected and the unknown positions are interpolated



- The interpolator filters are designed taking into account the characteristics of the radio channel
  - A WSS-US (Wide Sense Stationary-Uncorrelated Scattering)

- As the signals propagates through a noisy channel wiener filters are used to minimize the square root mean error

THANKS!

Any questions?

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