Synchronization Scheme and Interference Cancellation for DVB System

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Outline

- ♦ Overview of DVB-T/H System
- ♦ Synchronization and Channel Estimation
- ♦ Typical OFDM system & ICI Effect
- ♦ Literature review
  - – ICI Reduction of OFDM Receiver Windowing method
  - – Linear time-varying channel approximation & an ICI self-cancellation algorithm
- ♦ Further research
  - – Equivalent linear channel time-variation mitigation
  - – Equivalent low-complex receiver windowing method
- ♦ Simulation results
- ♦ Conclusion

數位電視的優點

- 對雜訊有較高的容忍力
- 可加入錯誤更正碼，修正錯誤
- 可對信號作壓縮，降低資料量，減少所需傳送頻寬
- 較能對抗訊號衰落問題
- 所需發射功率較低
- 容易將信號鎖碼
- 可提供影像、聲音、資料的整合性數位服務內容
- 積體電路化

現有的數位電視廣播系統

- **Europe:**
  - European Telecommunications Standards Institute (ETSI) ➔ Digital Video Broadcasting (DVB)
- **America:**
  - Advanced Television Systems Committee (ATSC) ➔ ATSC DTV
- **Japan:**
  - Association of Radio Industries and Business (ARIB). ➔ integrated services digital broadcasting (ISDB).
- **Korea:**
  - Digital multimedia broadcasting (DMB)
- **China**

數位電視廣播系統比較

規格	歐規(DVB-T)	美規(ATSC)	日規(ISDB-T)
頻寬	6/7/8 MHz	6 MHz	6 MHz
調變方式	COFDM QPSK,16QAM,64QAM Guard interval : 1/32,1/16,1/8,1/4 of OFDM Symbol 2 modes;2k and 8k FFT	8VSB	BST-OFDM DQPSK,QPSK,16QAM and 64QAM 3 mode :2k,4k and 8k FFT
視訊壓縮	MPEG-2	MPEG-2	MPEG-2
音訊壓縮	MPEG-2(or AC3)	Dolby AC-3	MPEG-2(or AC3)
訊息碼率 (Mbps)	4.35~31.67	19.38	5.6 MHz : 3.68~21.46 432 KHz : 283~1.65K
移動接收	可	差	可
抗鬼影	可	尚可	可
單頻網路	可	否	可
外部編碼	RS(204,188,t=8)	RS(207,187,t=10)	RS(204,188,t=8)
外部交錯	12 RS block interleaver	52 RS block interleaver	12 RS block interleaver
內部編碼	迴旋碼	Rate 2/3 trellis code	迴旋碼
內部交錯	Bit-wise interleaving and frequency interleaving	12 to 1 trellis code interleaver	Bit-wise interleaving and frequency interleaving ,and selectable time interleaving
資料擾亂器	16-bit PRBS		

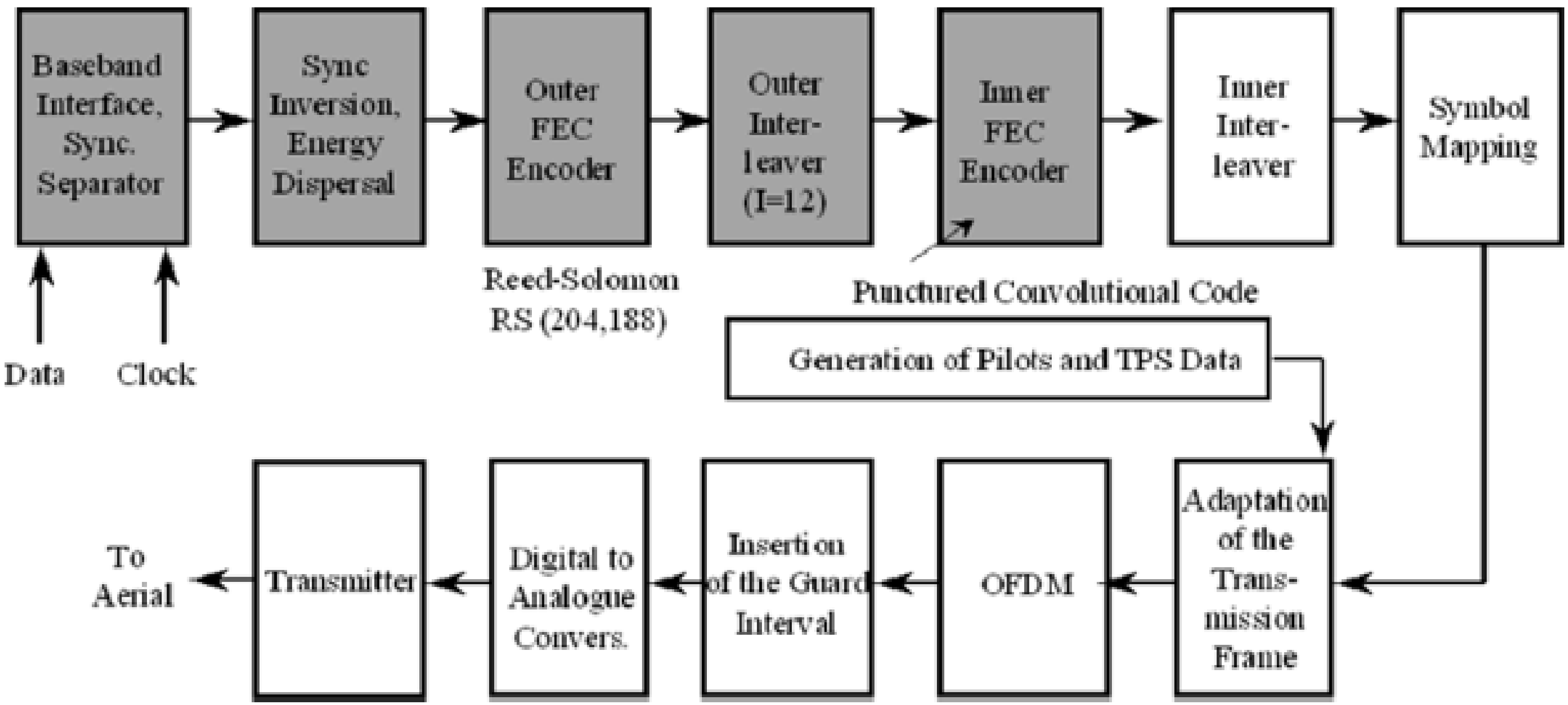
傳輸技術特點比較

- 8-VSB系統的特點：
  - 可在一個6MHz的通道傳送最高解析度的HDTV信號。
  - 在相同廣播大小範圍，所需的發射功率較小。
  - 處理短脈衝噪聲源的能力強。
- COFDM系統的特點：
  - 可行動接收，在車速達到130公里時仍可收看電視。
  - 抗多路徑干擾能力強，室內接收能力較佳。
  - 適用於單頻網路(Single Frequency Network, SFN)，所謂單頻網是指相鄰的發射站台使用相同的頻率來傳送資料，進而提升頻譜使用效率。
  - 更多、更靈活的選擇，能適應未來廣播不斷變化的需求。
  - 分層模式可解決臨界接收性能問題。

歐規DVB系列

- Satellite
  - DVB-S、 DVB-S2
- Terrestrial
  - DVB-T
- Cable
  - DVB-C
- Mobility
  - DVB-H

DVB系列傳輸技術



Block diagram of the encoder for DVB-T. Shaded blocks are used in DVB-C and DVB-S as well.

DVB-T、S、C 調變技術

- DVB-C:
  - the modulation chosen is QAM (with 16, 32, 64, 128, or 256 points in the constellation diagram)
- DVB-S:
  - QPSK or BPSK
- DVB-T:
  - OFDM was selected where each individual carrier can be modulated with QPSK, 16 QAM, or 64 QAM.

DVB規格特性

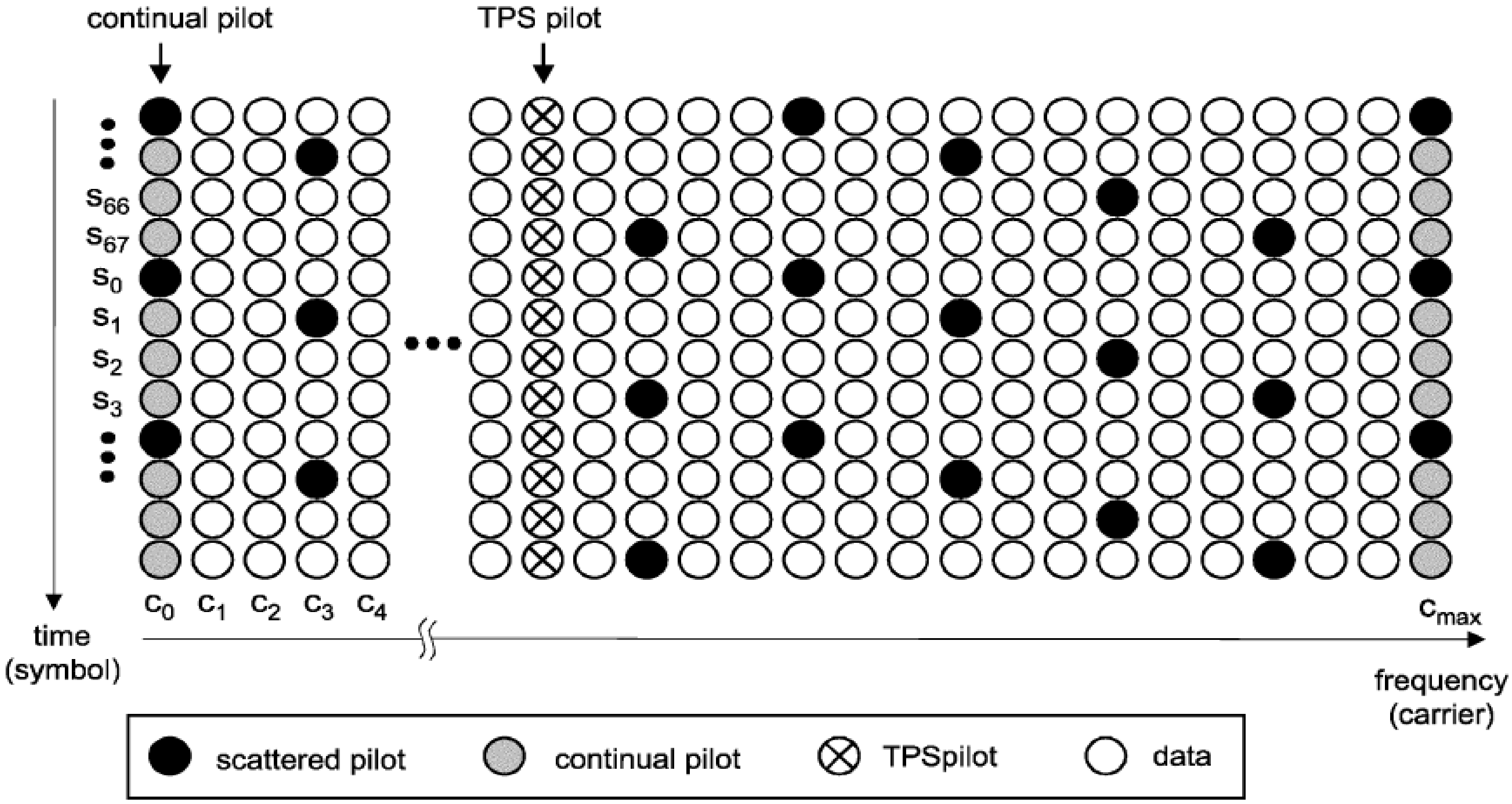
- Coded OFDM (COFDM)
- Single-Frequency Network (SFN)
- Multipath reception performance better than ATSC.
- Flexible (2K 、 8k mode, variable code rate)
- Digital Video Broadcasting-Handheld was published as (ETSI) Standard EN 302 304 in November 2004.
- Based on DVB-T for fixed and in-car reception of digital TV.
- DVB-H is designed for handheld device and mobile reception.
- DVB-H is backward compatible to DVB-T

Carriers Type

DVB-T contains the following types of carrier :

- Payload carriers with fixed position.
- Inactive carriers with fixed position.
- Continual pilots with fixed position.
- Scattered pilots with changing position in the spectrum.
- TPS carriers with fixed position.

Carriers Position



TPS Carriers

- The TPS carriers are located at fixed frequency positions.
- TPS stands for Transmission Parameter Signaling. These carriers represent virtually a fast information channel via which the transmitter informs the receiver about the current transmission parameters.
- All the TPS carriers in one symbol carry the same information, i.e. they are all either at 0 degrees or all at 180 degrees on the I axis.
- The complete TPS information is broadcast over 68 symbols and comprises 68 bits.

TPS Purpose & Content

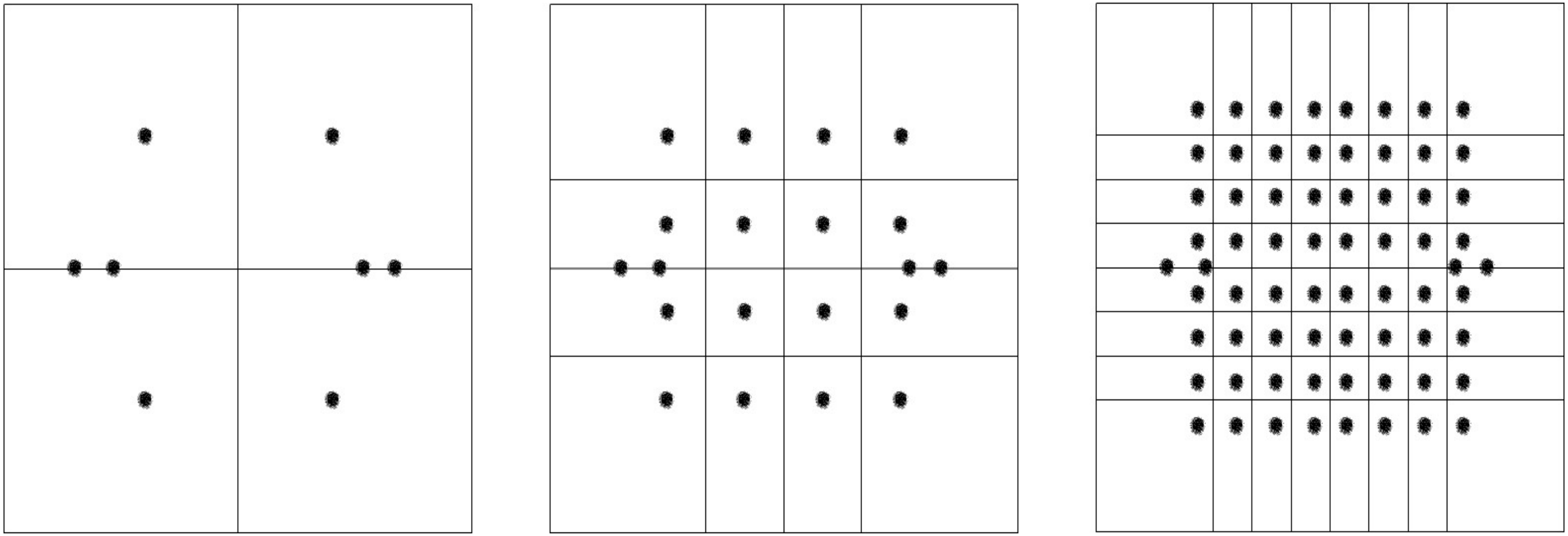
Bit number	Format	Purpose/Content
s_0		Initialization
$s_1 - s_{16}$	0011010111101110 or 1100101000010001	Synchronization word
$s_{17} - s_{22}$	010 111	Length indicator
s_{23}, s_{24}		Frame number
s_{25}, s_{26}		Constellation 00=QPSK/01=16QAM/10=64QAM
s_{27}, s_{28}, s_{29}		Hierarchy information 000=Non hierarchical, 001=α=1, 010=α=2, 011=α=4
s_{30}, s_{31}, s_{32}		Code rate, HP stream 000=1/2, 001=2/3, 010=3/4, 011=5/6, 100=7/8
s_{33}, s_{34}, s_{35}		Code rate, LP stream 000=1/2, 001=2/3, 010=3/4, 011=5/6, 100=7/8
s_{36}, s_{37}		Guard interval 00=1/32, 01=1/16, 10=1/8, 11=1/4
s_{38}, s_{39})		Transmission mode 00=2K, 01=8K
$s_{40} - s_{53}$	all set to "0"	Reserved for future use
$s_{54} - s_{67}$	BCH code	Error protection

TPS Carry Information

Thus, the TPS carriers keep the receiver informed about:

- The mode (2k, 8k).
- The length of the guard interval (1/4, 1/8, 1/16, 1/32).
- The type of modulation (QPSK, 16QAM, 64QAM).
- The code rate (1/2, 2/3, 3/4, 5/6, 7/8).
- The use of hierarchical coding.

DVB-T Constellation Diagram



DVB-T Constellation Diagrams for QPSK,16-QAM and 64-QAM

Carrier Type Value

2K Mode	8K Mode	
2048	8192	carrier
1705	6817	used carrier
142/131	568/524	scattered pilots
45	177	continual pilots
17	68	TPS carrier
1512	6048	payload carrier

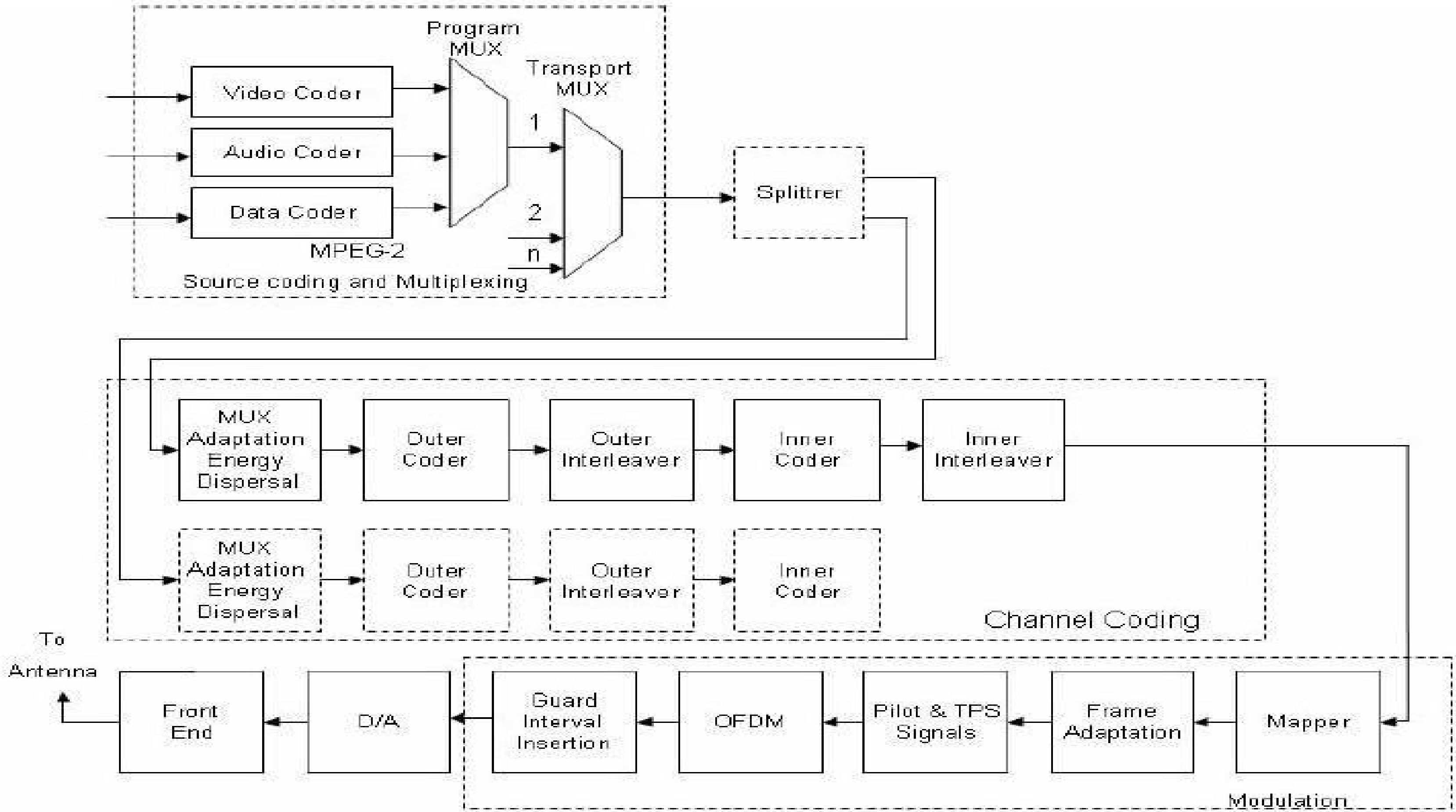
DVB-T可選擇之傳輸參數

OFDM 模式		2K			8K		
載波數 K		1705(0….1704)			6817(0….6816)		
RF通道頻寬		6 MHz	7 MHz	8 MHz	6 MHz	7 MHz	8 MHz
區間 K_0 and K_{max}		5.71 MHz	6.66 MHz	7.61 MHz	5.71 MHz	6.66 MHz	7.61 MHz
載波區間		3348 Hz	3906 Hz	4464 Hz	837 Hz	977 Hz	1116 Hz
期間 T_{want}		299μs	256μs	224μs	1195μs	1024μs	896μs
保護區間	1/4	75μs	64μs	56μs	299μs	256μs	224μs
	1/8	37μs	32μs	28μs	149μs	128μs	12μs
	1/16	19μs	16μs	14μs	75μs	64μs	56μs
	1/32	9μs	8μs	7μs	37μs	32μs	28μs
載波調變方式		QPSK,16-QAM,64-QAM					
內部編碼率		1/2,2/3,3/4,5/6,7/8					

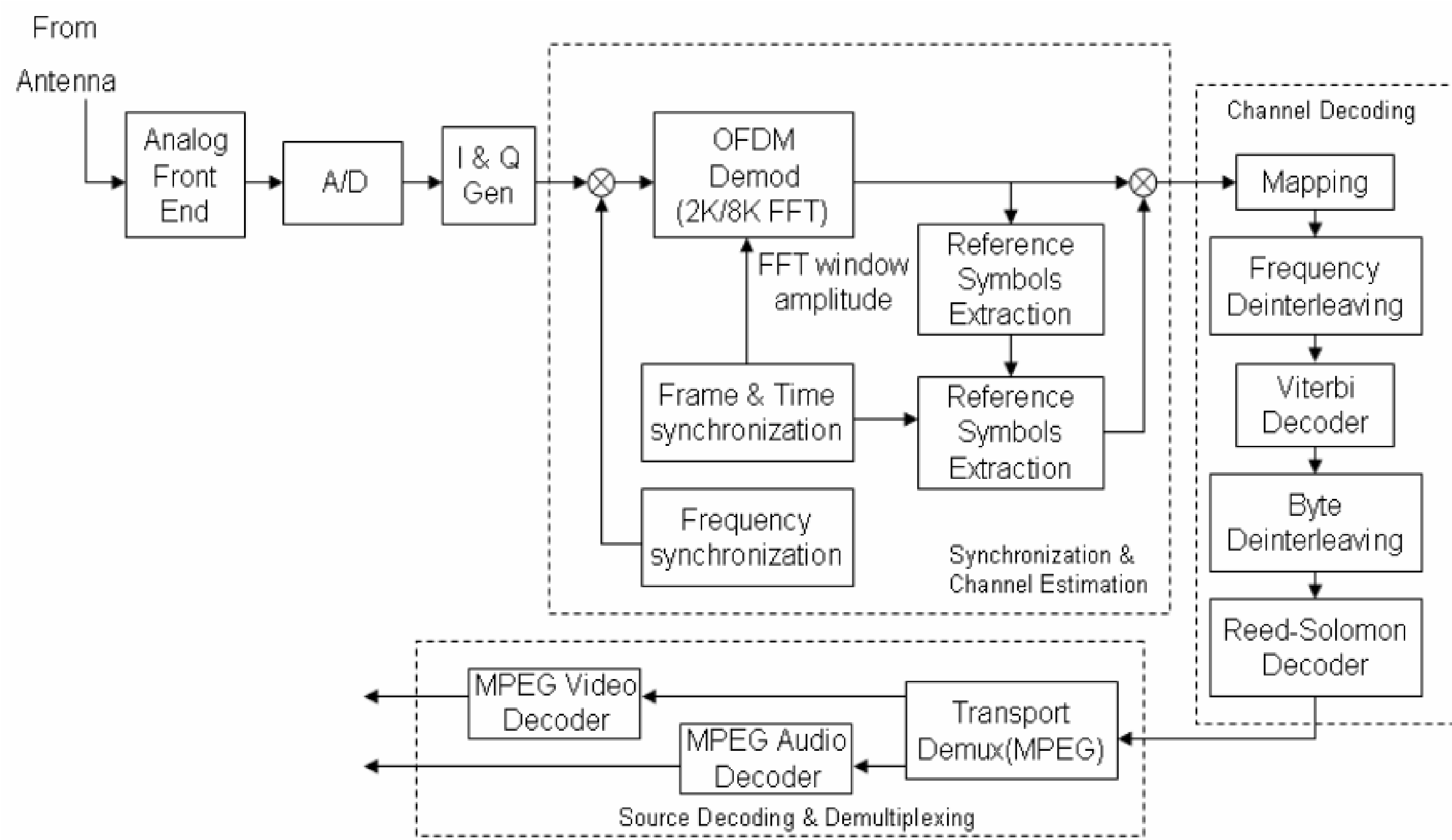
6MHz通道下的網路資料傳輸率(Mbps)

調變方式	編碼率	Guard 1/4	Guard 1/8	Guard 1/16	Guard 1/32
QPSK	1/2	3.732353	4.147059	4.391003	4.524064
	2/3	4.976471	5.529412	5.854671	6.032086
	3/4	5.598529	6.220588	6.586505	6.786096
	5/6	6.220588	6.911765	7.318339	7.540107
	7/8	6.531618	7.257353	7.684256	7.917112
16QAM	1/2	7.464706	8.294118	8.782007	9.048128
	2/3	9.952941	11.058824	11.709343	12.064171
	3/4	11.197059	12.441177	13.173010	13.572193
	5/6	12.441176	13.823529	14.636678	15.080214
	7/8	13.063235	14.514706	15.368512	15.834225
64QAM	1/2	11.197059	12.441177	13.173010	13.572193
	2/3	14.929412	16.588235	17.564014	18.096257
	3/4	16.795588	18.661765	19.759516	20.358289
	5/6	18.661765	20.735294	21.955017	22.620321
	7/8	19.594853	21.772059	23.052768	23.751337

DVB-T Transmitter

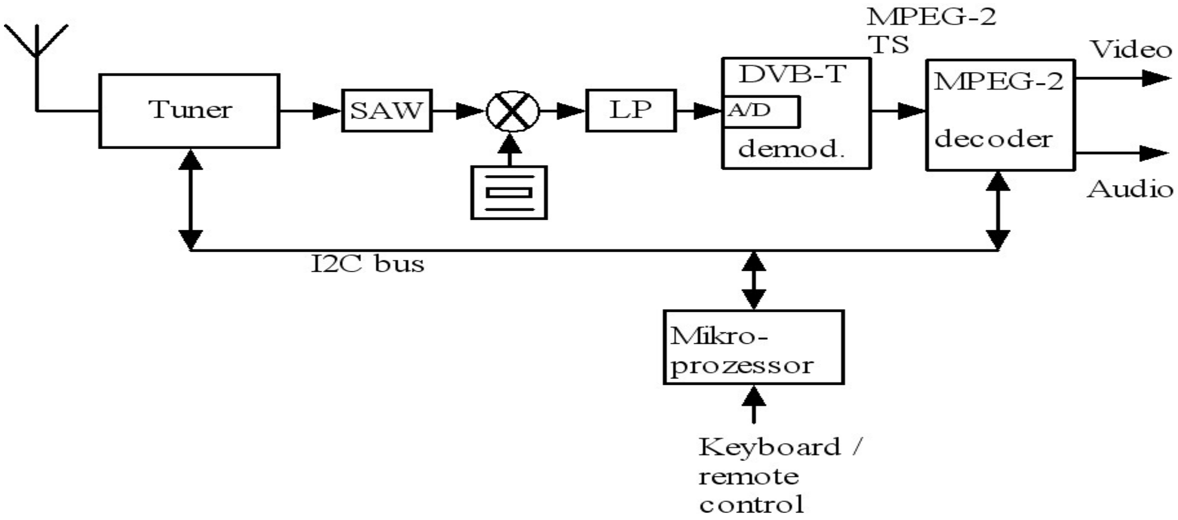


DVB-T Receiver



Set-Top Box

- The transport stream coming out of the DVB-T demodulator is fed into the MPEG-2 decoder where it is decoded back into video and audio.
- All these modules are controlled by a microprocessor via an I2C bus.



Conceptual Structure of DVB-H Receiver

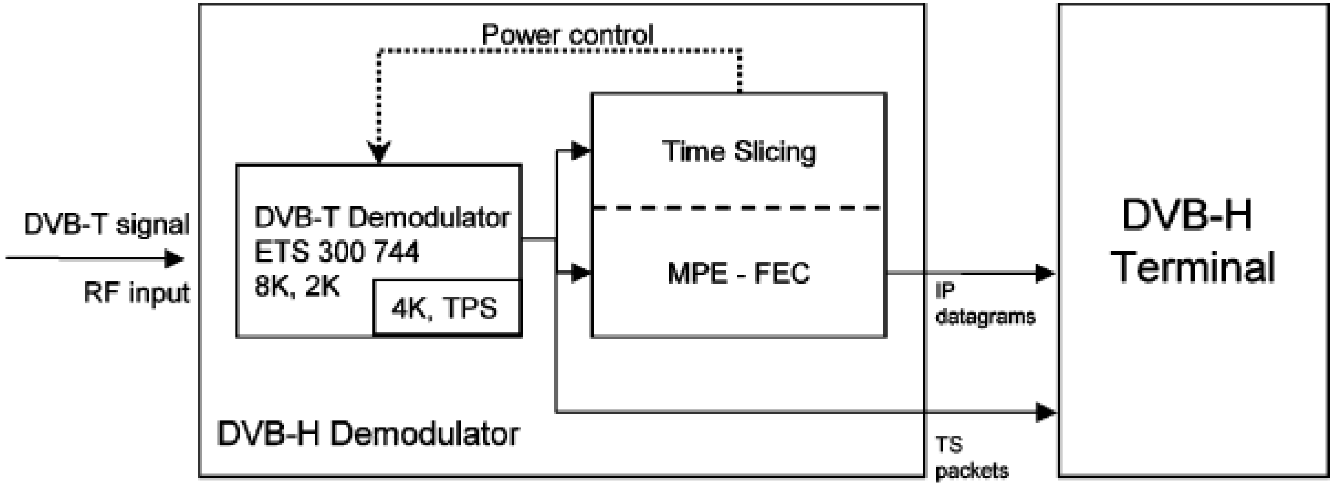


Fig. Conceptual structure of a DVB-H receiver.

- The time-slicing module controls the receiver to decode the wanted service and shut off during the other service bits. (reduce power consumption)

Conceptual Description of DVB-H System

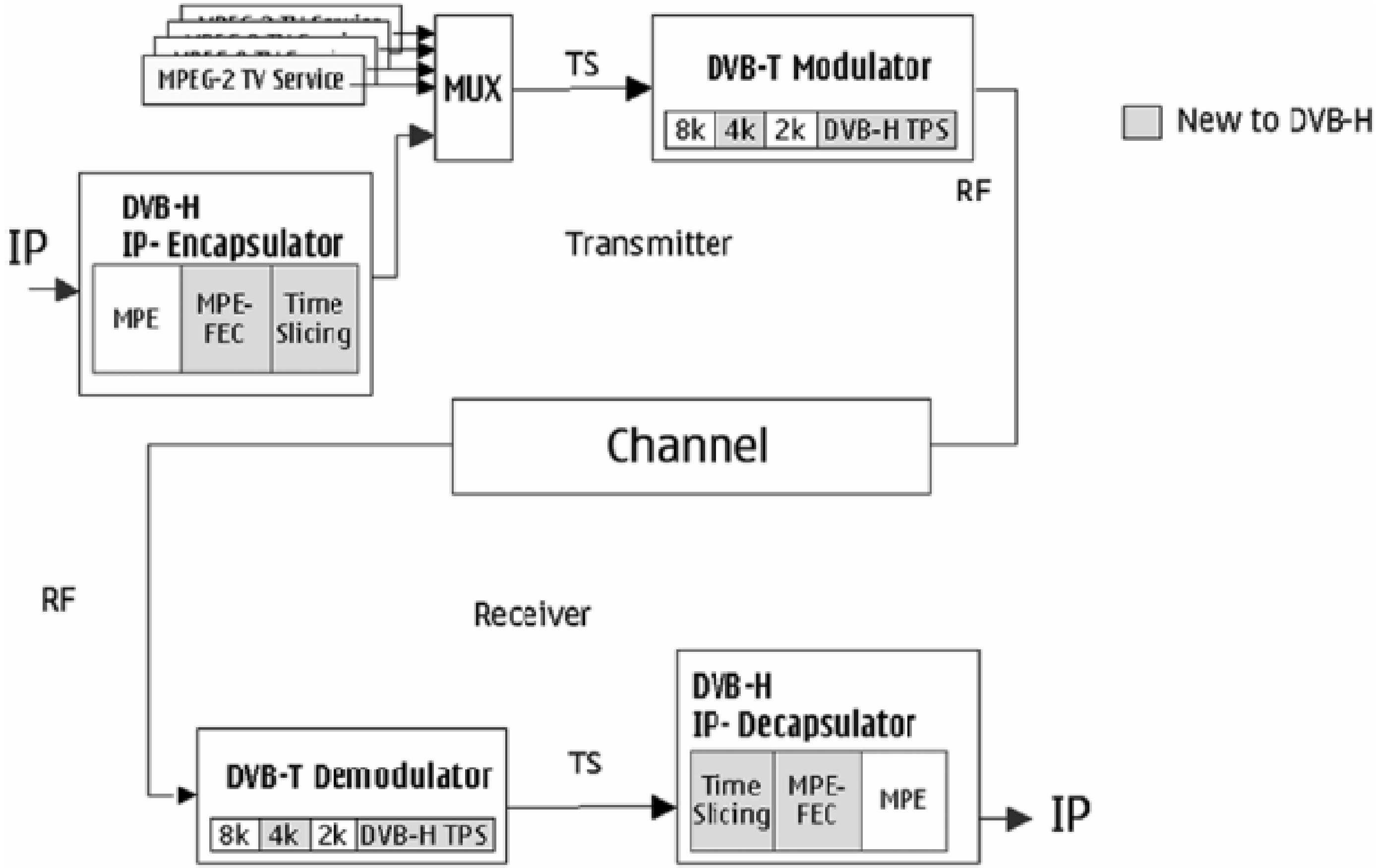


Fig. A conceptual description of using a DVB-H system (sharing a MUX with MPEG-2 services).

Time Slicing

- ♦ The receiver power off completely during off-time and will power on again to receive the next burst.

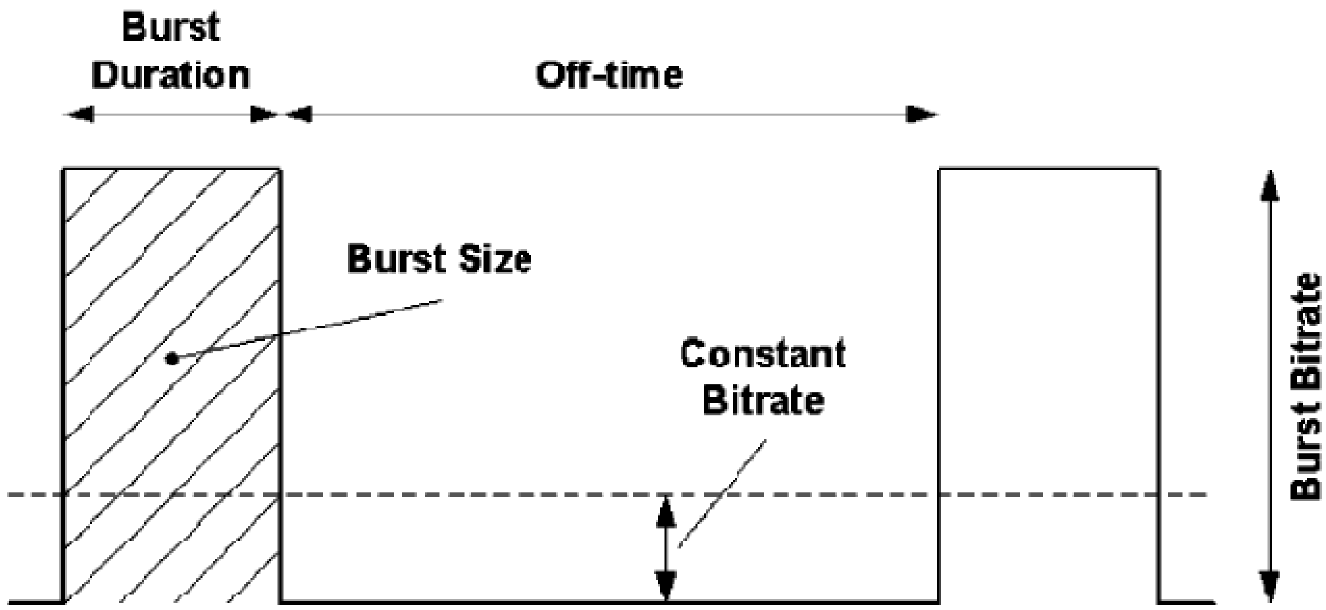
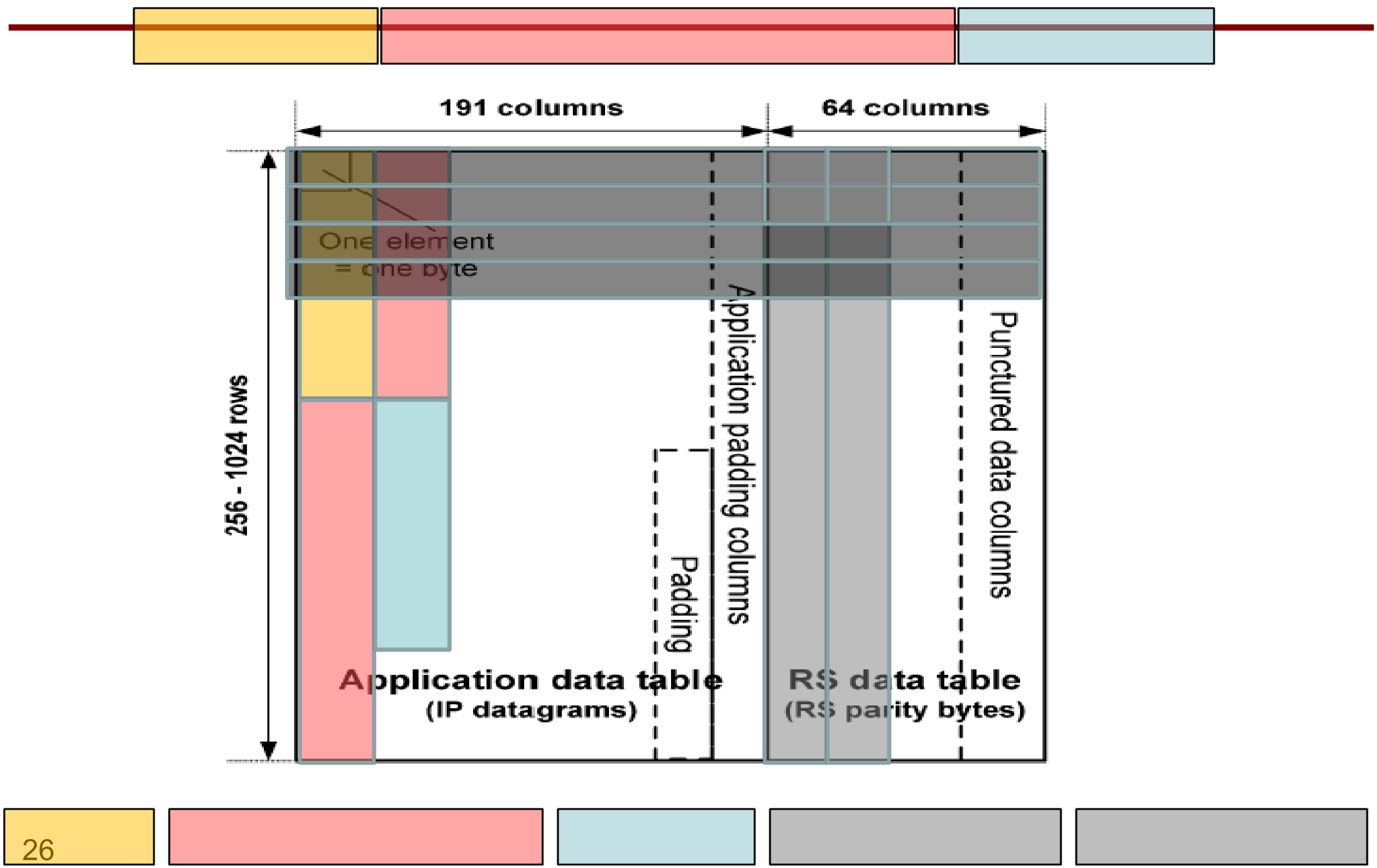


Fig. Principle of time slicing.

MPE-FEC Method



Sharing with DVB-T

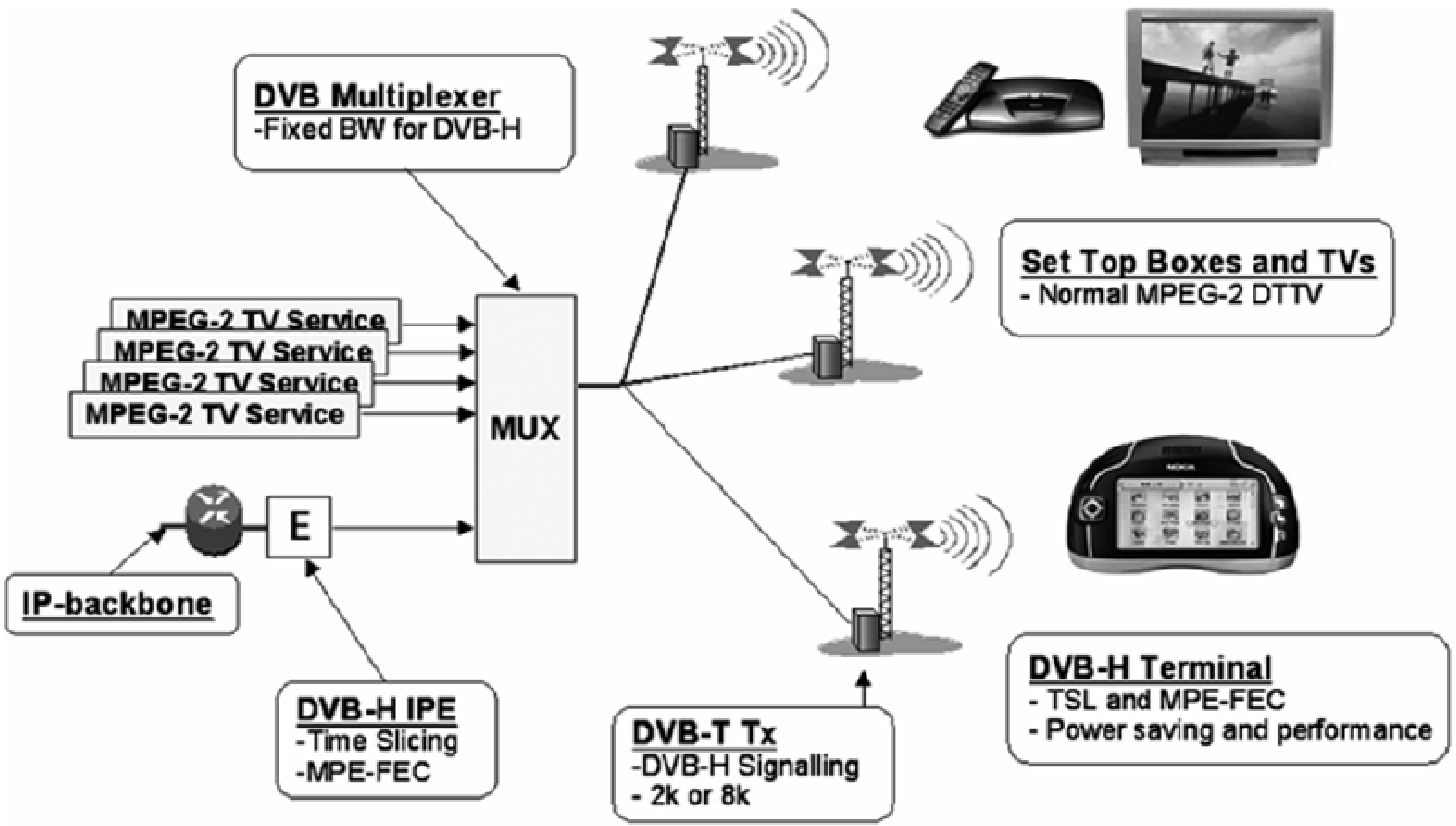


Fig. Sharing a network with DVB-T by multiplexing.

Dedicated DVB-H Networks

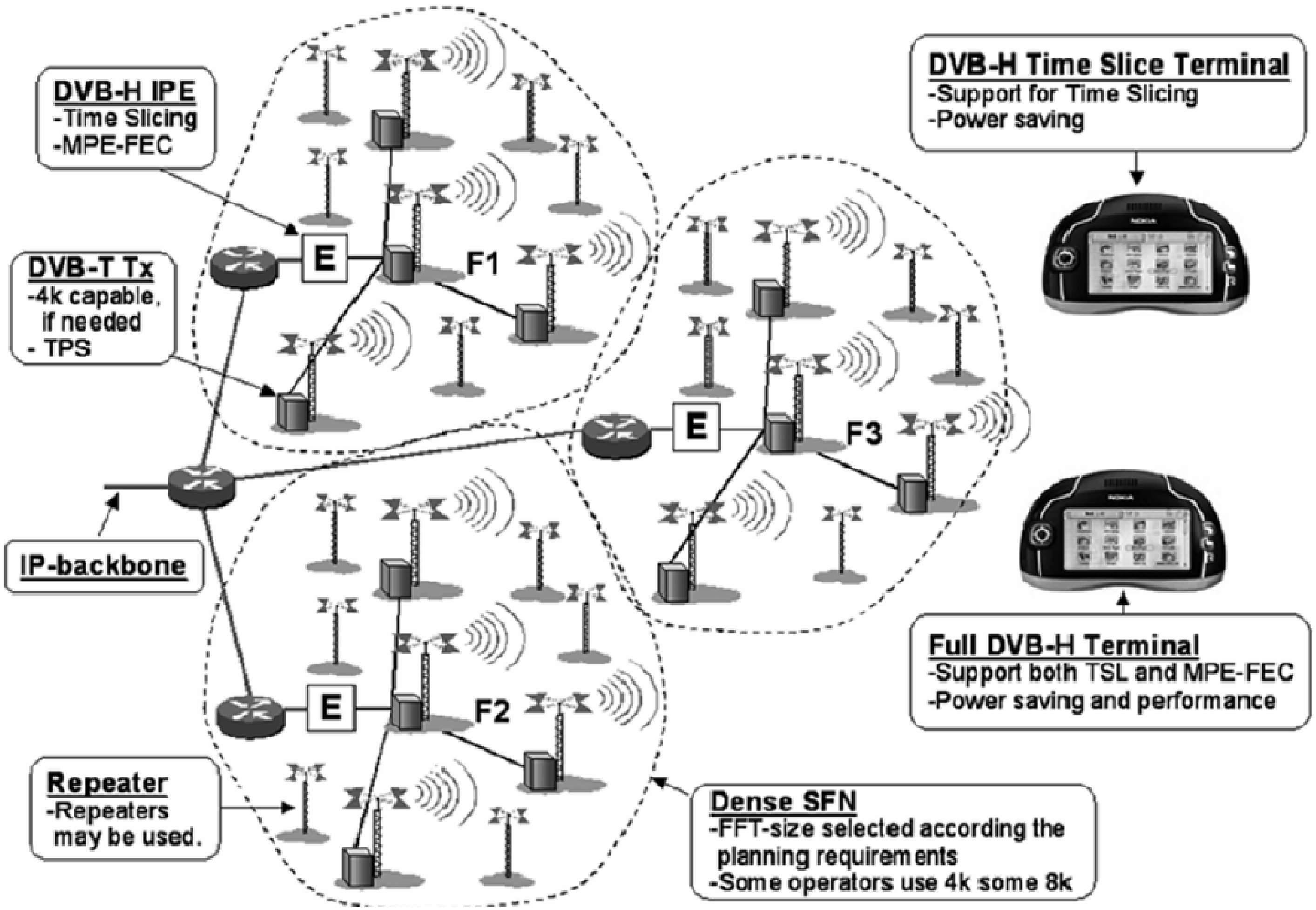


Fig. A dedicated DVB-H network.

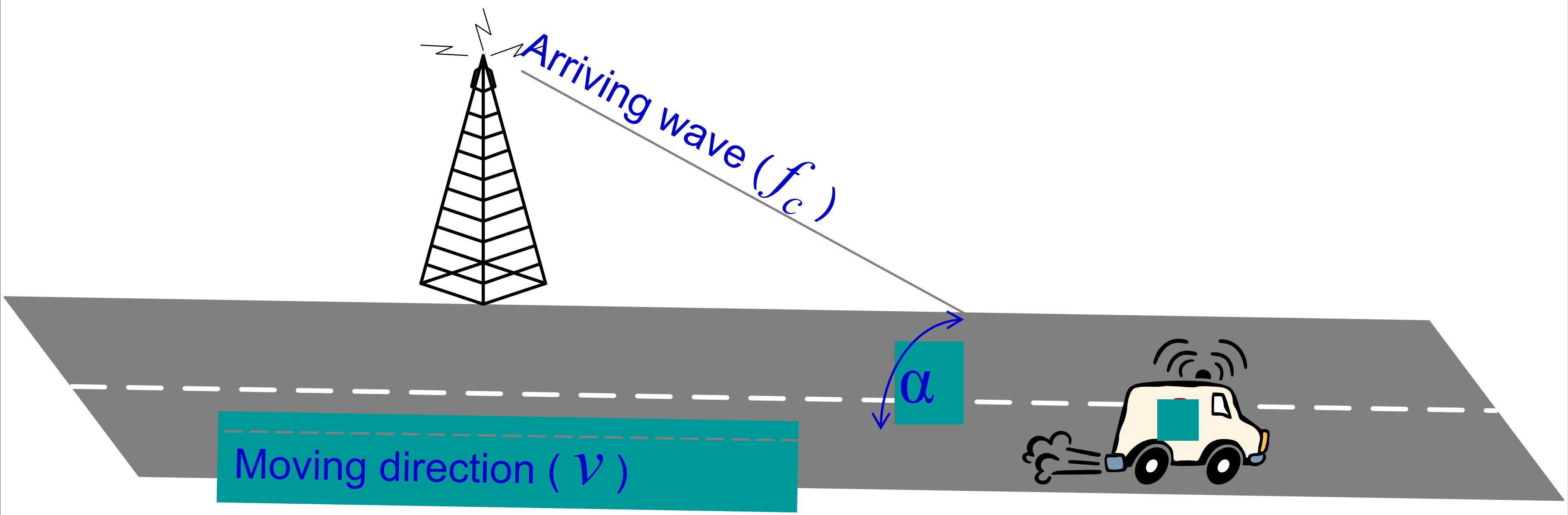
Mobile Reception (1)

- The channel is not characterized by just one direct and dominant signal path, but by multiple fast changing echo signals.
- These multipath signals can mathematically be described as a time-varying Rayleigh channel and require a higher carrier-to-noise ratio than needed for stationary reception.
- A specialty of mobile reception is the Doppler effect which is a frequency shift in the signals arriving at the receiver.

Mobile Reception (2)

- Doppler effect (frequency change)

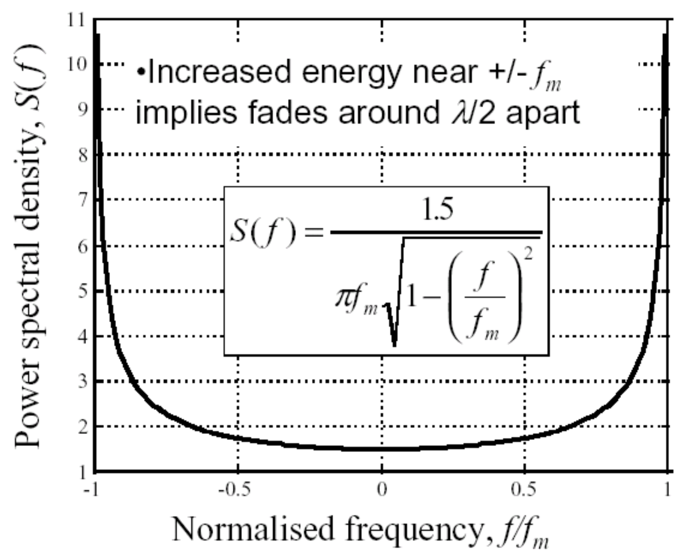
$$f_d = f_c \cdot \frac{v}{c} \cos\alpha \qquad\qquad f_{d,\max} = f_c \frac{v}{c}$$



Mobile Reception (3)

- Doppler effect (frequency shift)

- Classical Doppler Spectrum



Mobile Reception (4)

- To achieve successful mobile DVB-T reception, a number of factors need to be considered:
  - Need to track channel variations in time and frequency.
  - Correct channel estimation needs to be provided.
  - To be able to compensate for noise-like distortions called FFT leakage due to the time varying channel. (nonorthogonality of the DVB-T subcarriers)

Diversity Techniques for Mobile Receivers

- Utilize two or more reception antennas. (space diversity)
- Two approaches:
  1. Selection Combining:

     Select the highest carrier-to-noise-ratio.
  2. Maximum Ratio combining:

     Combine the signals of L diversity branches.
- More cost.

Selection Combining

♦ Practical used.

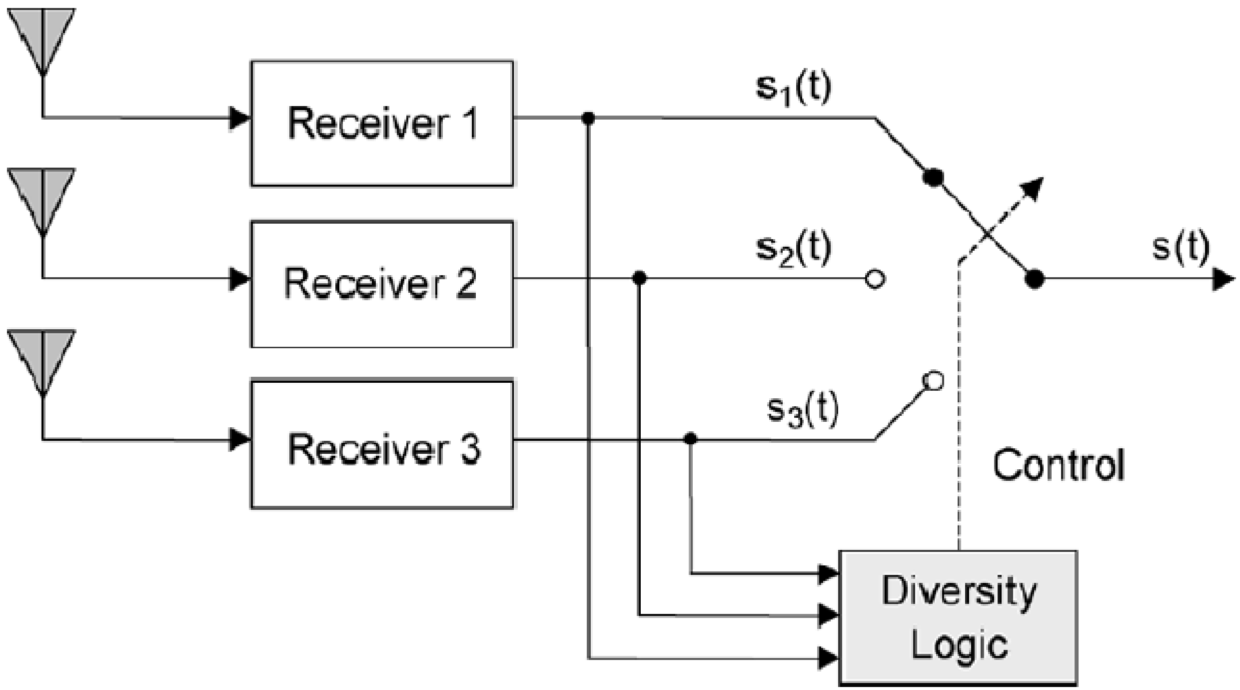


Fig. Selection Combining structure

Maximum Ratio Combining

♦ Synchronized in phase and multiply weight coefficients

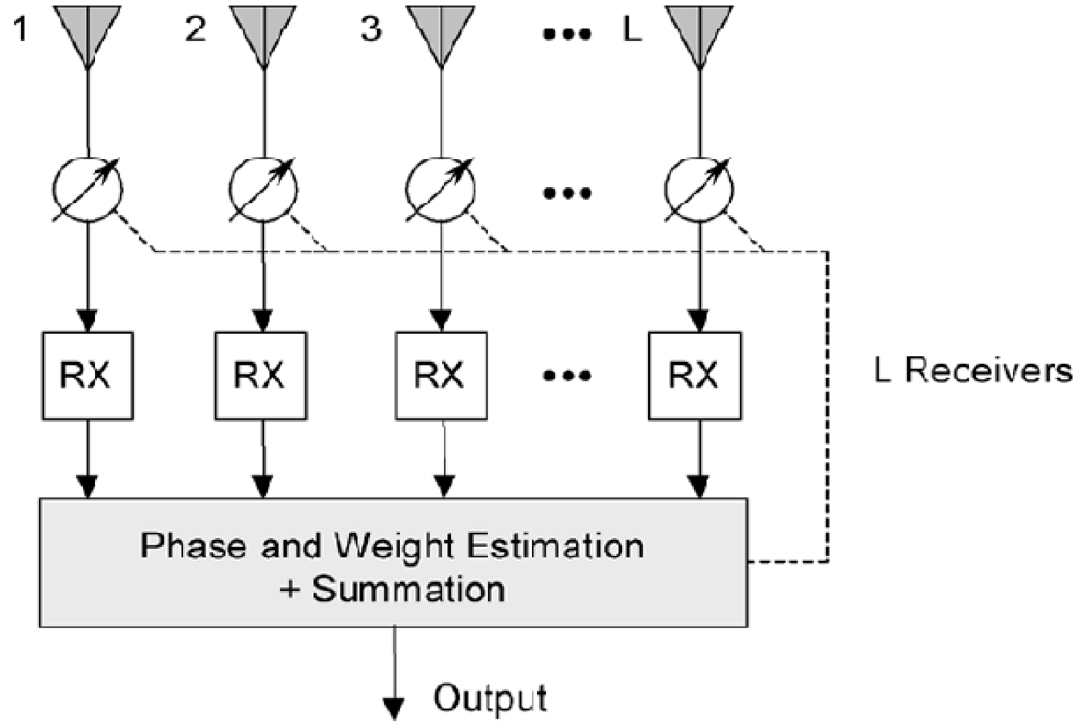
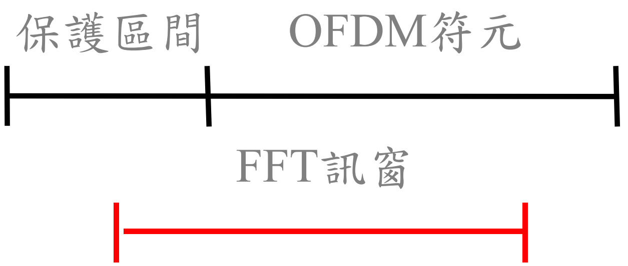


Fig. Maximum Ratio Combining structure

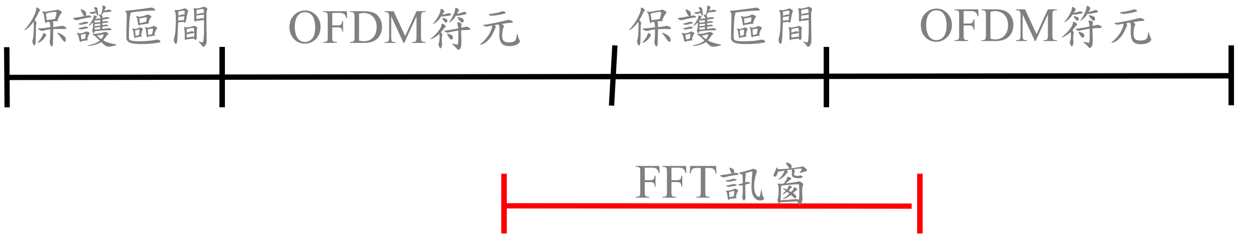
OFDM時間與頻率偏移

- 歐規數位電視系統所使用的傳輸架構為正交分頻多工，OFDM系統可以支援大量的資料傳輸，但在無線的傳輸上會造成時間偏移、相位偏移以及頻率偏移，這些問題對於接收訊號會有很大的影響，受到這些影響OFDM系統效能將明顯地降低，因此接收機對於符元同步和頻率同步的要求更加嚴格。

時間偏移

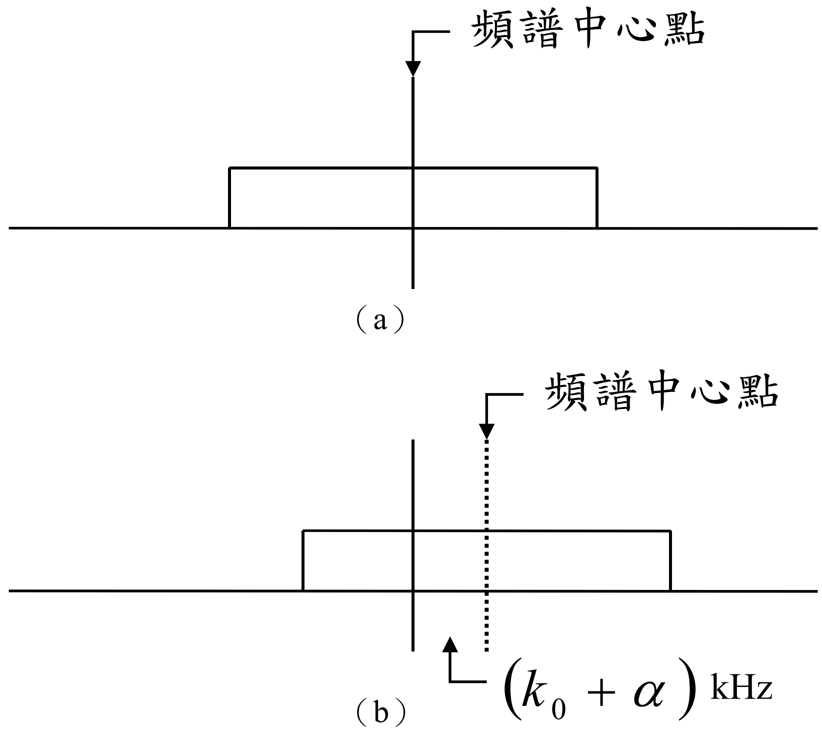


（a）



（b）

頻率偏移



Synchronization

- Synchronization
  - Time sync.
  - Frequency sync.

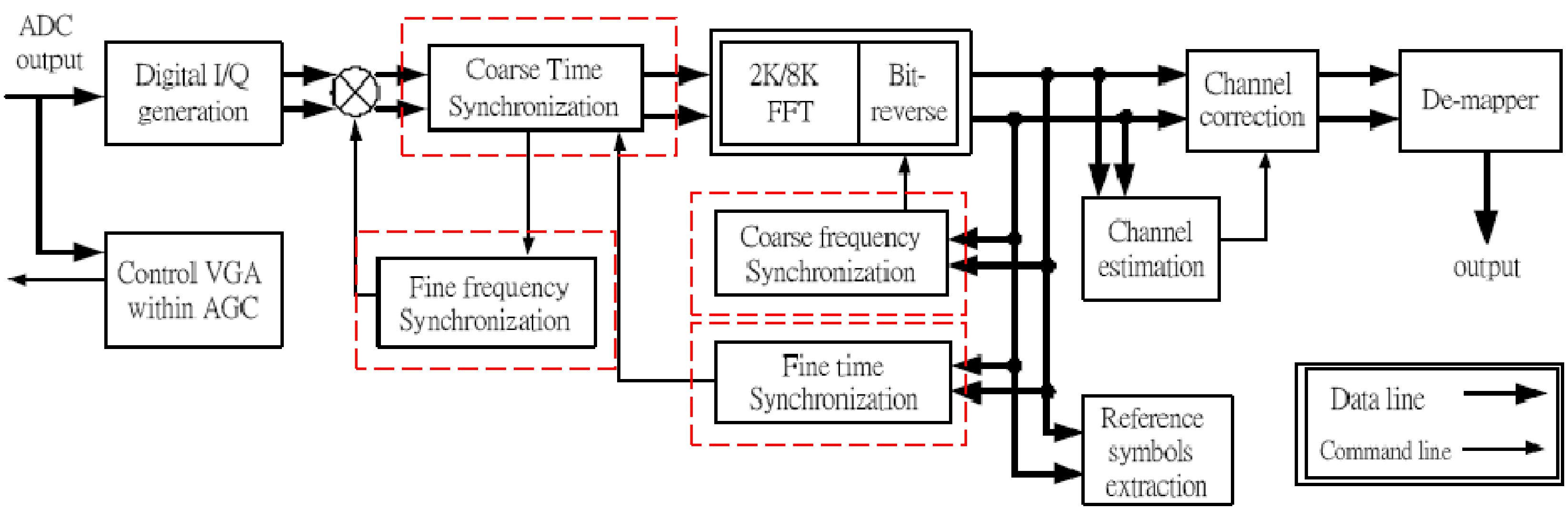
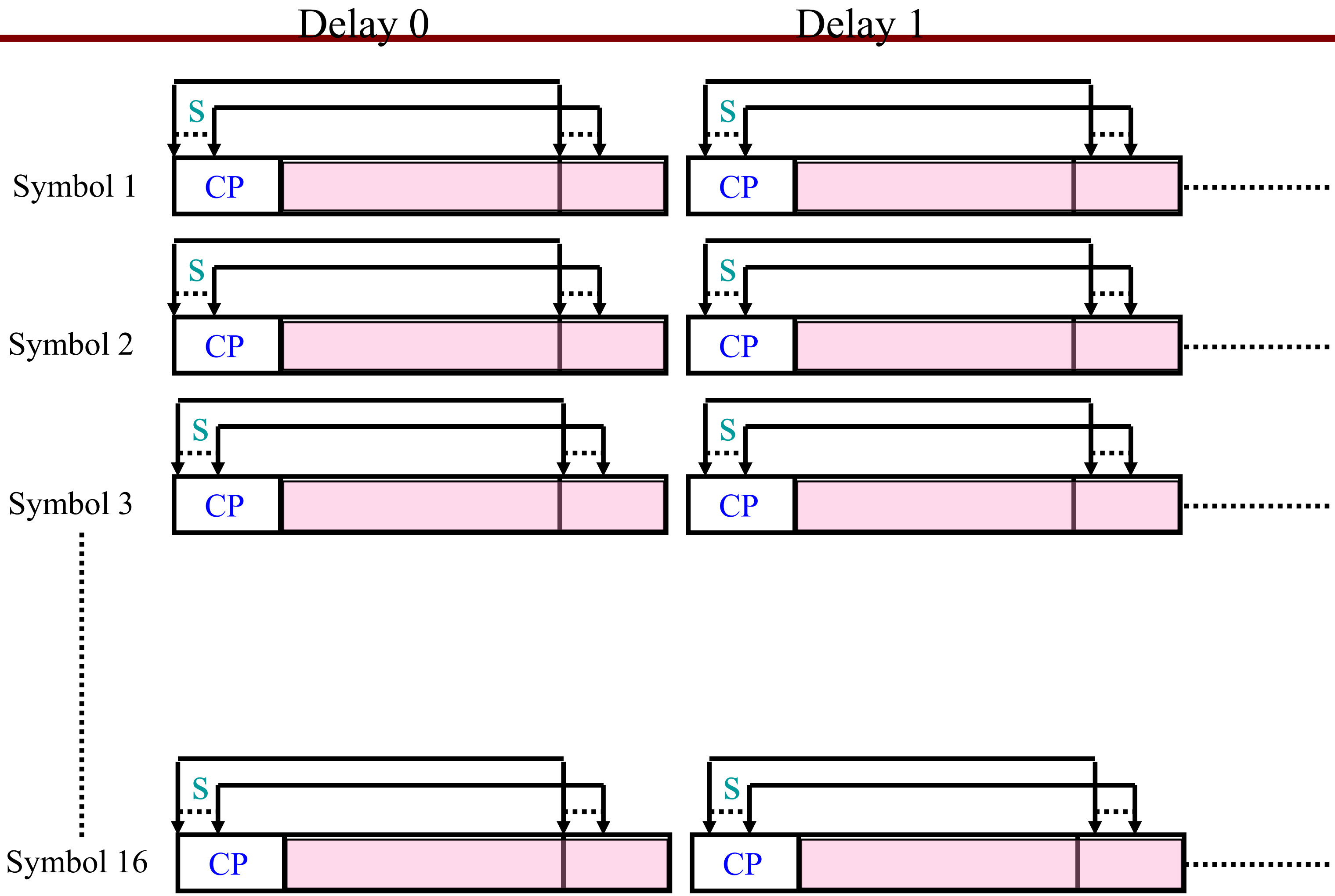


Fig. Receiver Baseband

部分相似估測法



部分相似估測法3

- 最大相似估測法表示式：

$$\Lambda_C(\theta)=\frac{\sum_{m=0}^{t-1}\sum_{k=\theta}^{\theta+s-1} y(k+m(N+L))y^*(k+N+m(N+L))}{\frac{1}{2}\sum_{m=0}^{t-1}\sum_{k=\theta}^{\theta+s-1}\left[|y(k+m(N+L))|^2+|y(k+N+m(N+L))|^2\right]}$$

$$\hat{\theta}=\arg\quad\max_{\theta}\{\Lambda_C(\theta)\}$$

- 時間取樣點估測值

$$\hat{\varepsilon}=-\frac{1}{2\pi}\angle\Lambda_C(\hat{\theta})$$

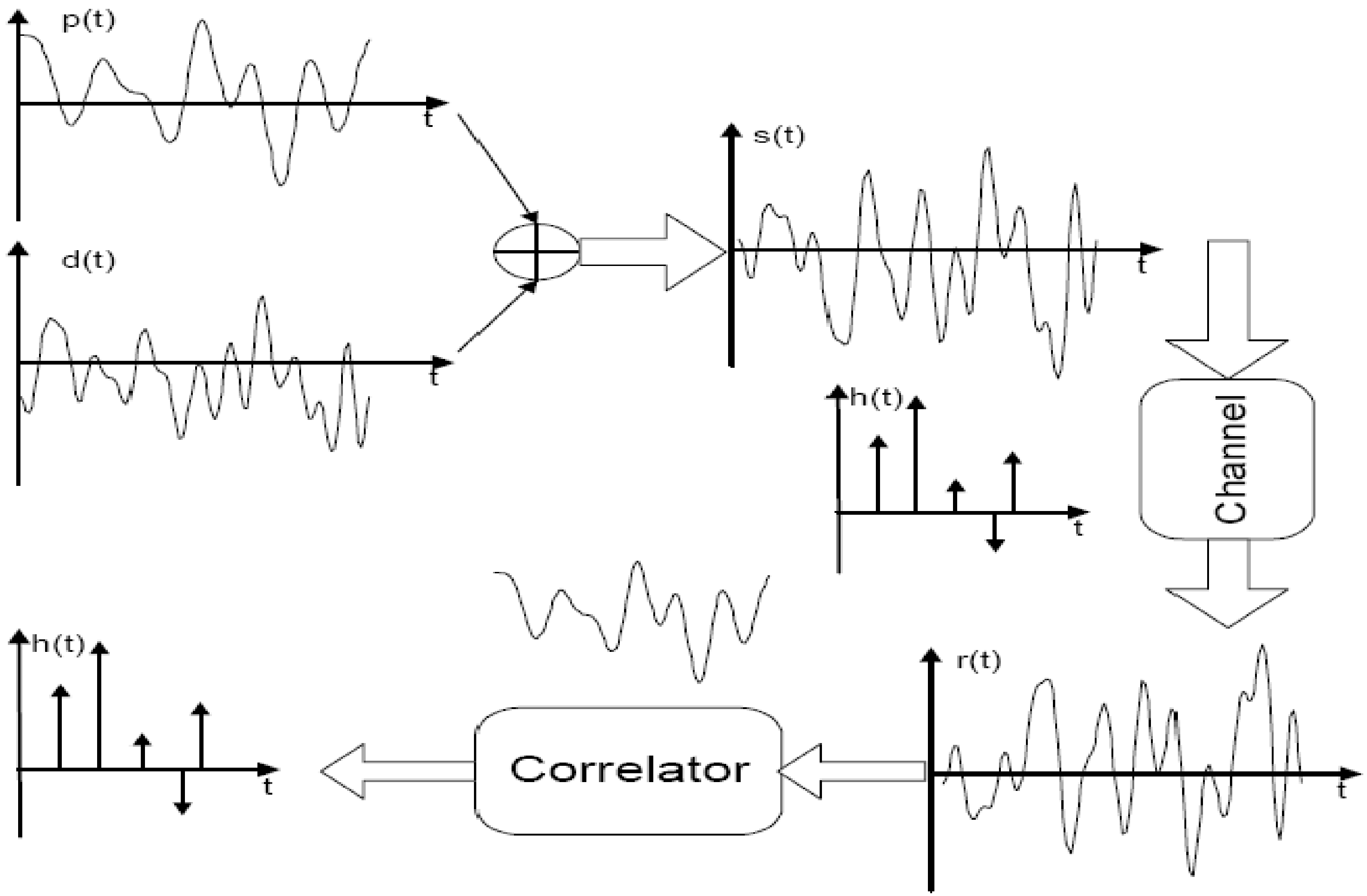
Why Need Channel Estimation

- 在無線通訊環境中，由於通道環境的不確定性，也就是通會受到地形地物和地貌影響，造成接收機在接收信號的同時會因為干擾，造成接收信號和發射信號不同而造成失真，失真就是信號經過通道後其信號的大小(amplitude)和相位(phase)不同。
- 通道估測(channel estimation)就是利用各種演算法(algorithm)解決通道的影響，求出通道響應(channel response)使接收端可以還原出發射端的訊號。
- 以台灣的現況而言，其數位電視信號是使用16-QAM的調變、6MHz的頻寬，因此如果信號振幅變得太小，則經過16-QAM的解調變後，可能會得到錯誤的結果。
- 此外，由於通道及其他因素（例如時間或取樣頻率有小幅偏差）的影響，進入16-QAM解調變器的符元，有一些會有相位偏差，而這些相位偏差，則又會造成解調變錯誤。

Channel Estimation

- To compensate channel effect (gain and phase)
- The known signal, Pilot signal, inserted in the transmitted signal.
- The receiver utilize the pilot signal to find the impairment of channel and compensate the whole received signal.

FPTC



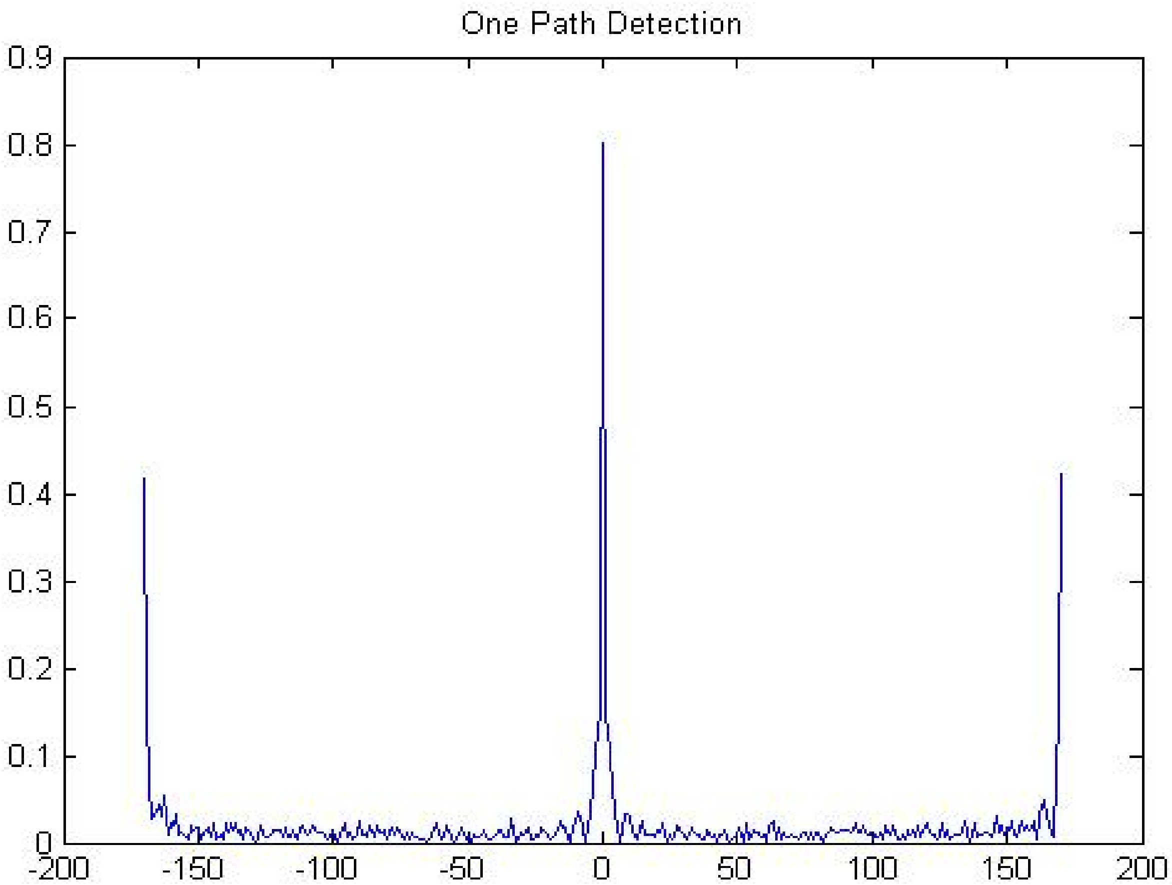
FPTC

一般而言一個在時域中有N點的OFDM符元可表示如下式：

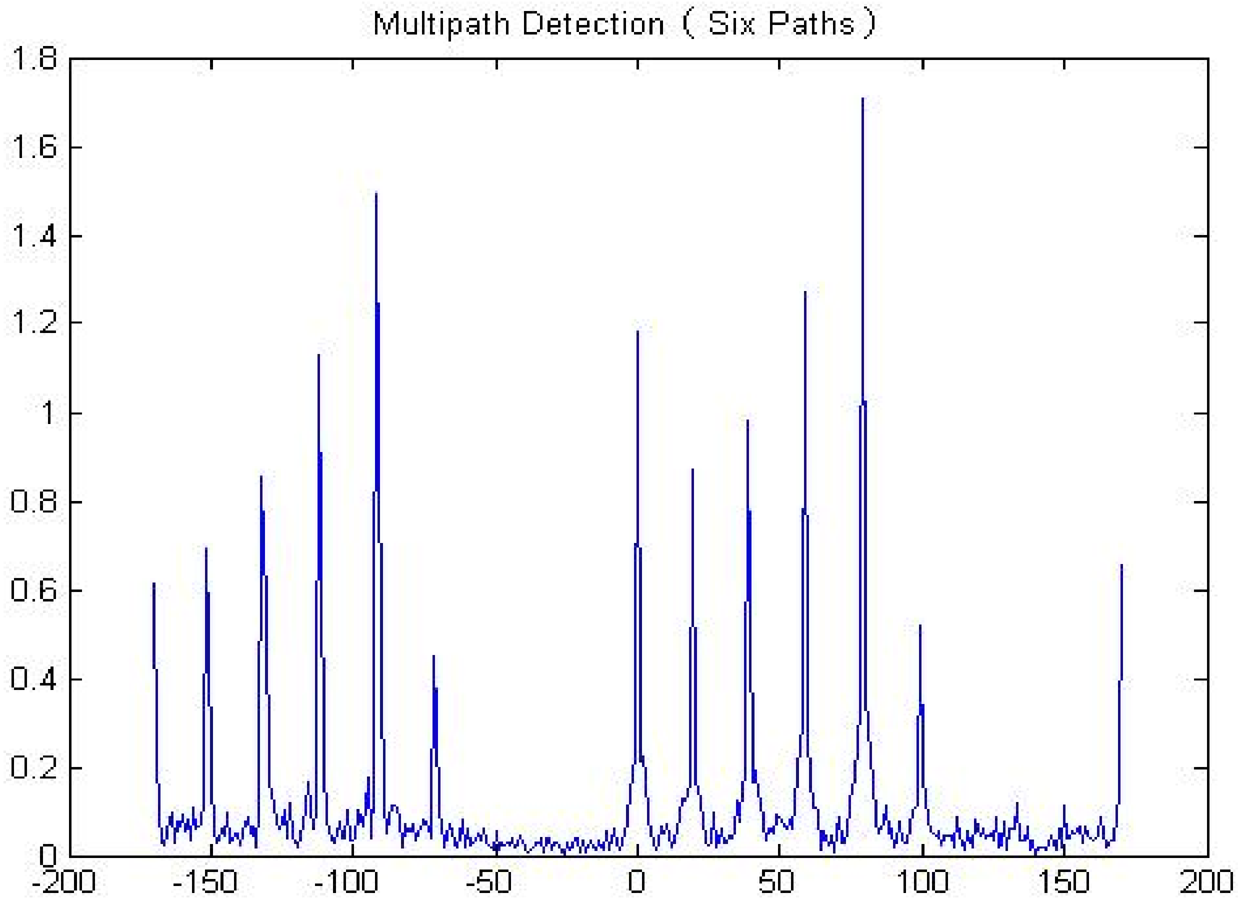
$$x(n) = \frac{1}{\sqrt{N}} \sum_{k=0}^{N-1} X_k e^{j\frac{2\pi nk}{N}}$$ ，n=0, 1, 2, …N-1

其中 $x(n)$ 代表通過IFFT後時域上的信號， X_k 爲原始輸入的頻域信號。

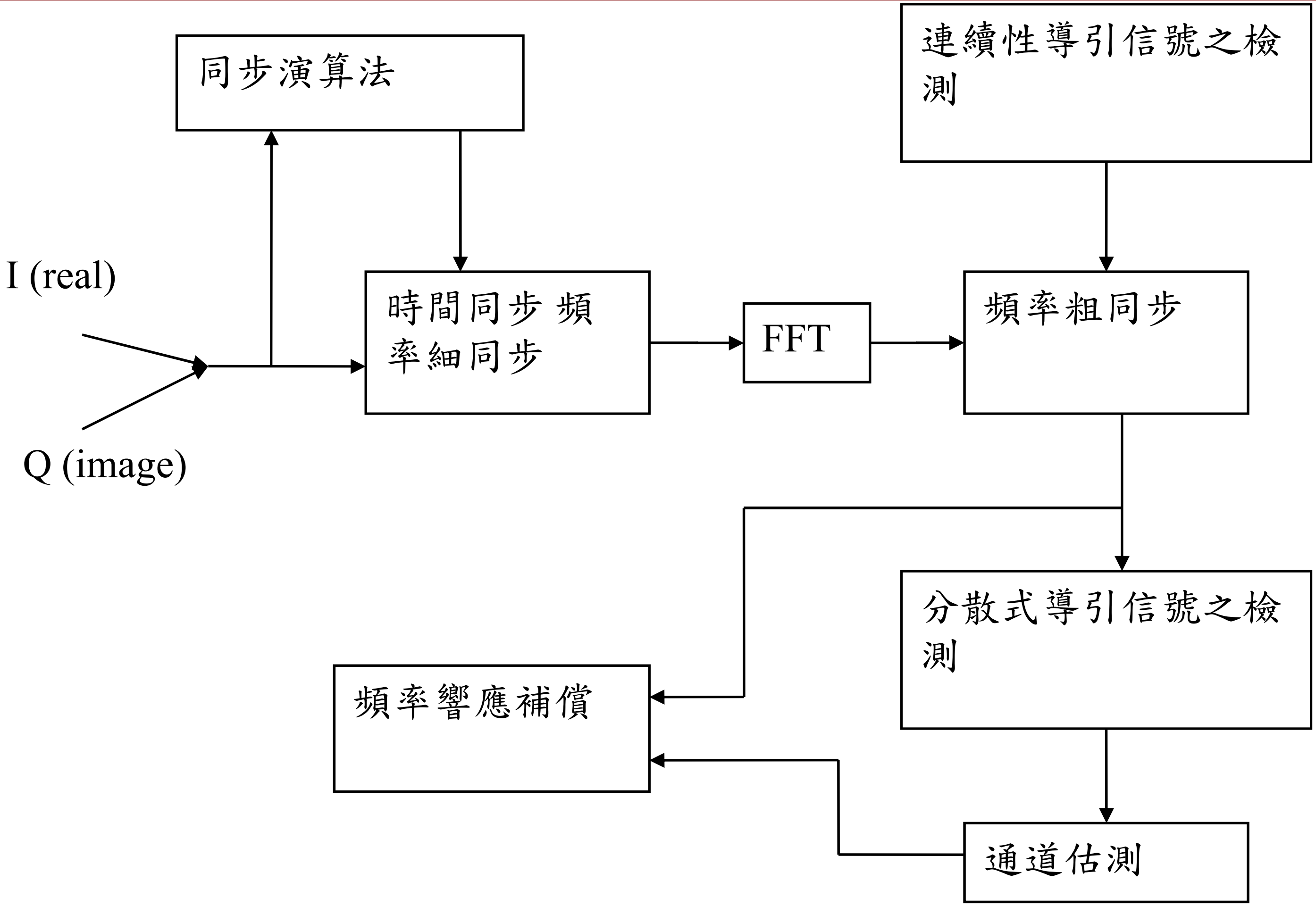
One Path Detection



Six Path Detection

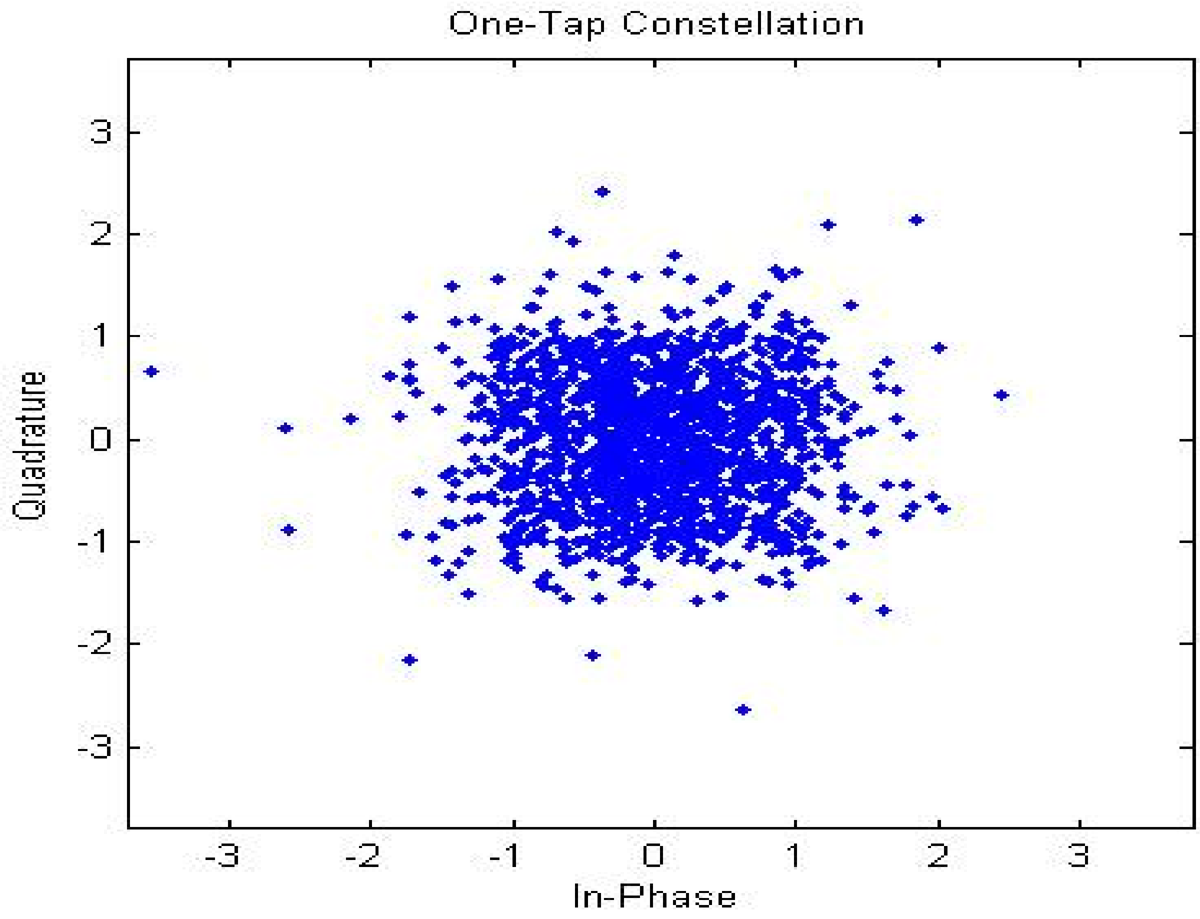


部分相似估測法結合FPTC

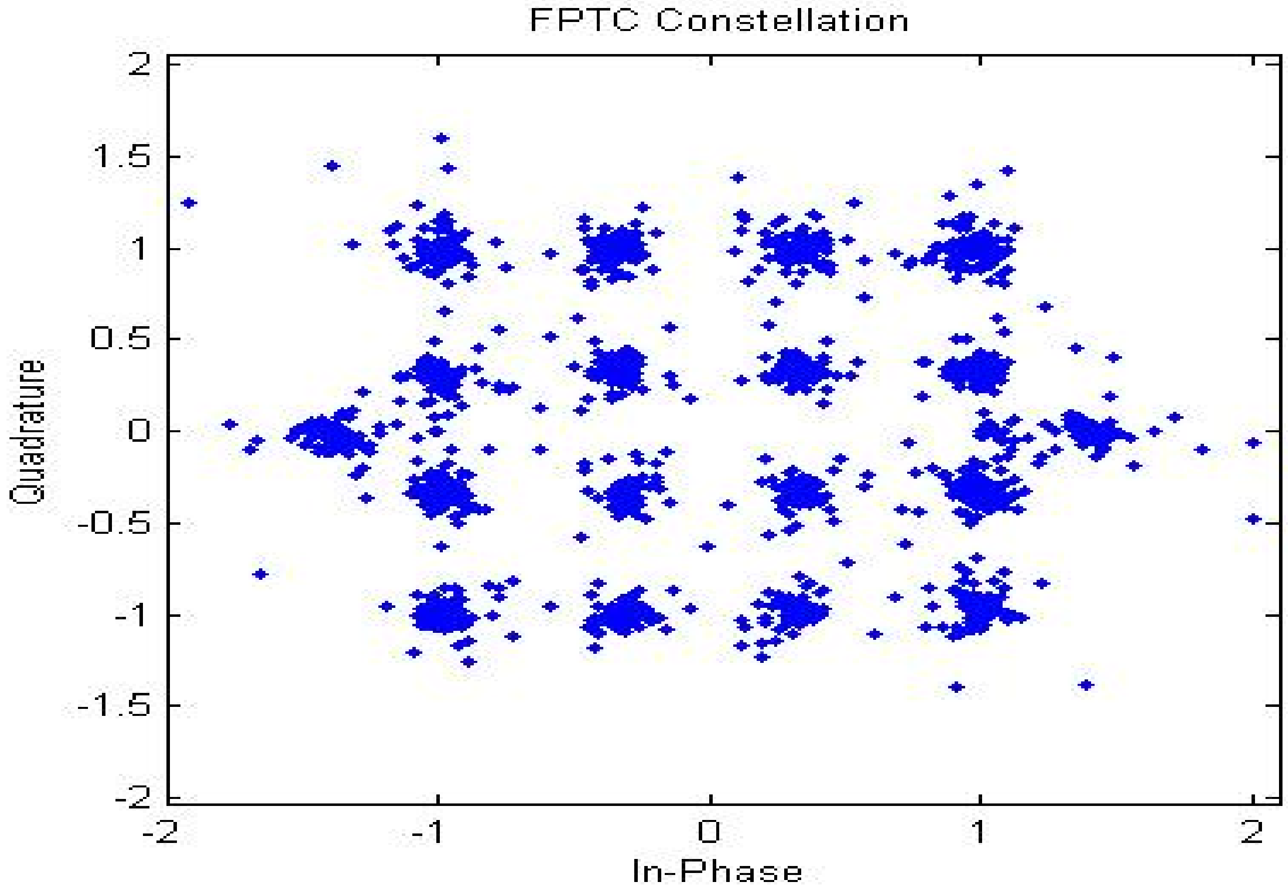


信號同步與通道估測流程圖

One-Tap Equalizer Constellation



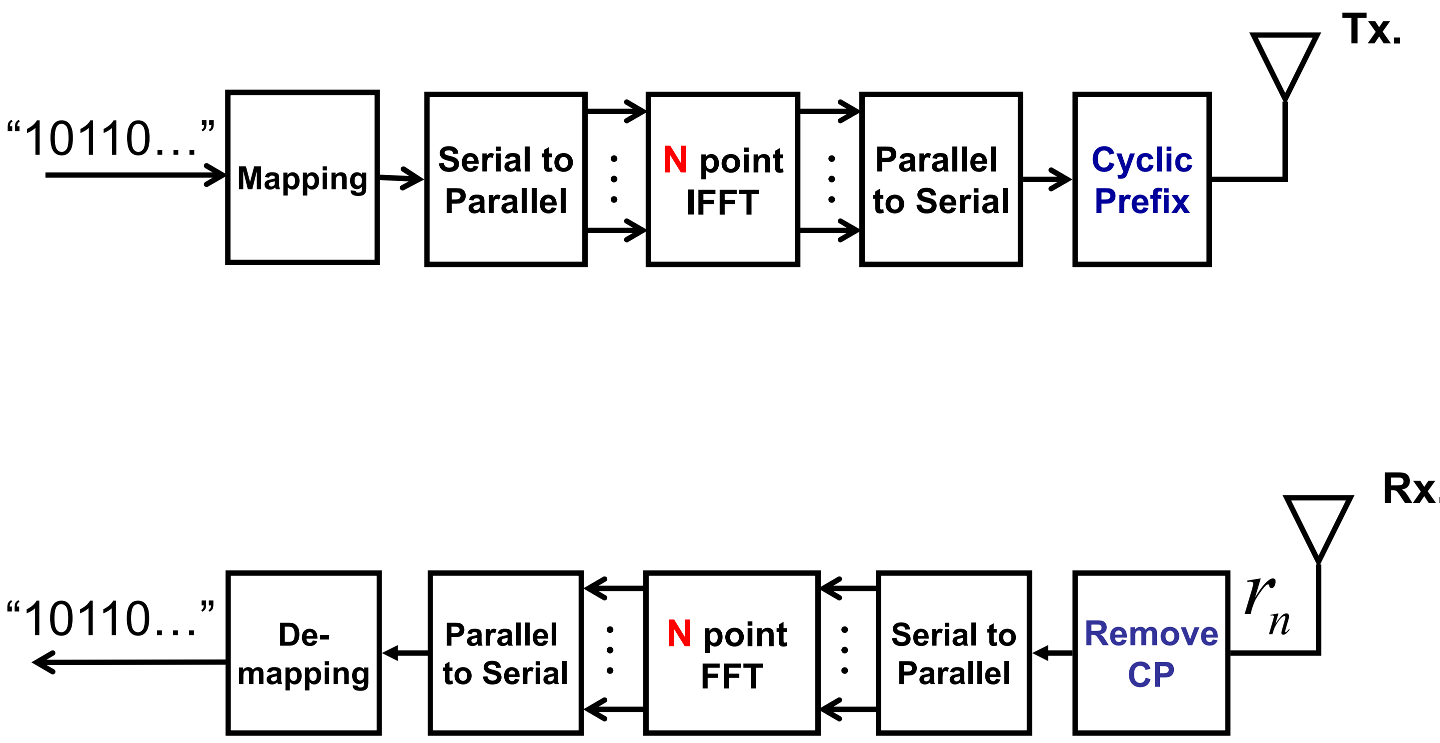
FPTC Constellation



ICI Effects

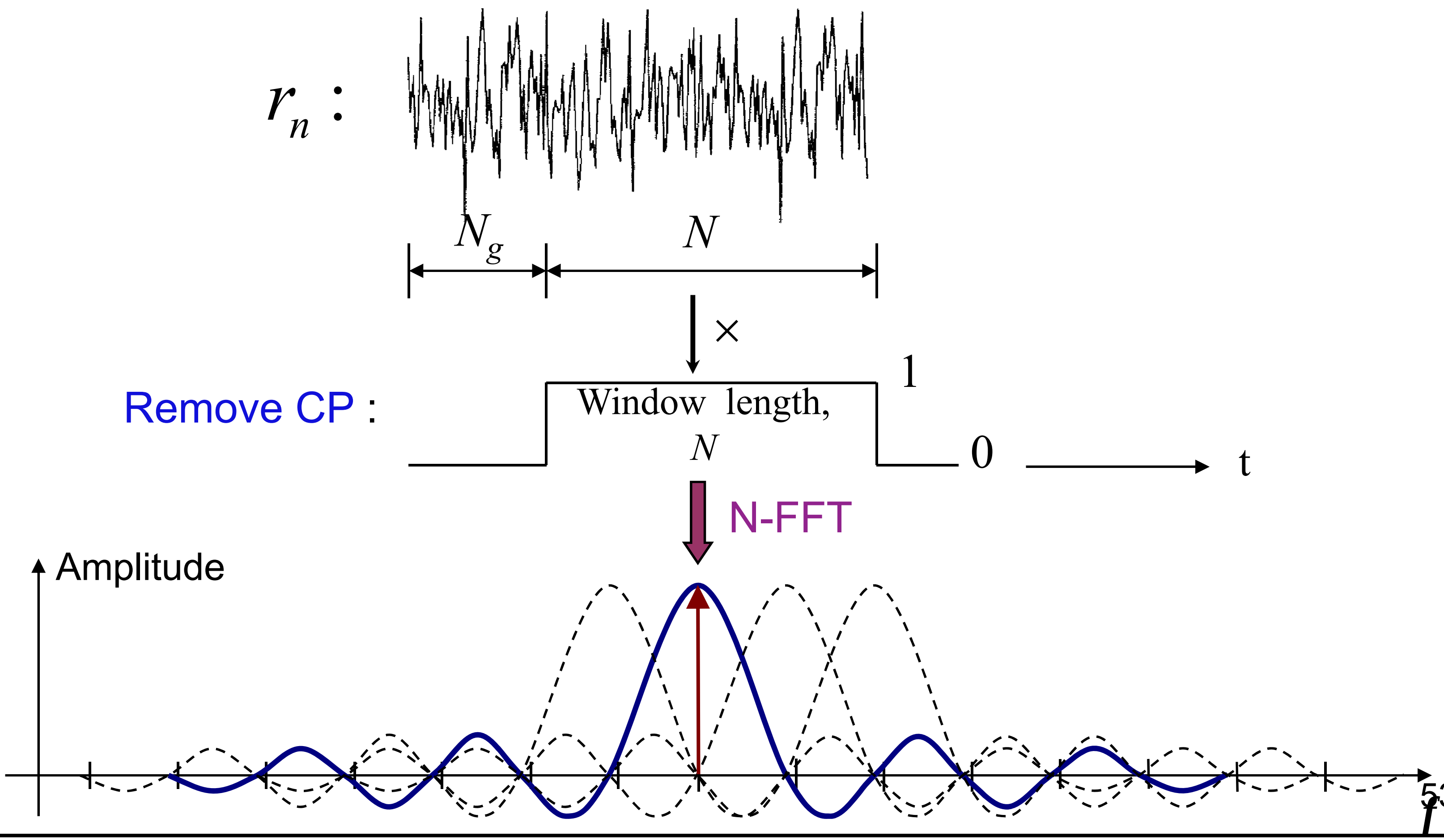
- The OFDM systems suffers ICI during time-varying channel or transceiver oscillator frequency mismatch.
- The typical OFDM systems using cyclic prefix (CP) as guard interval to prevent ISI.
- However,
- the ISI-free part of guard interval can be also utilized to suppress ICI.

Typical OFDM Block Diagram

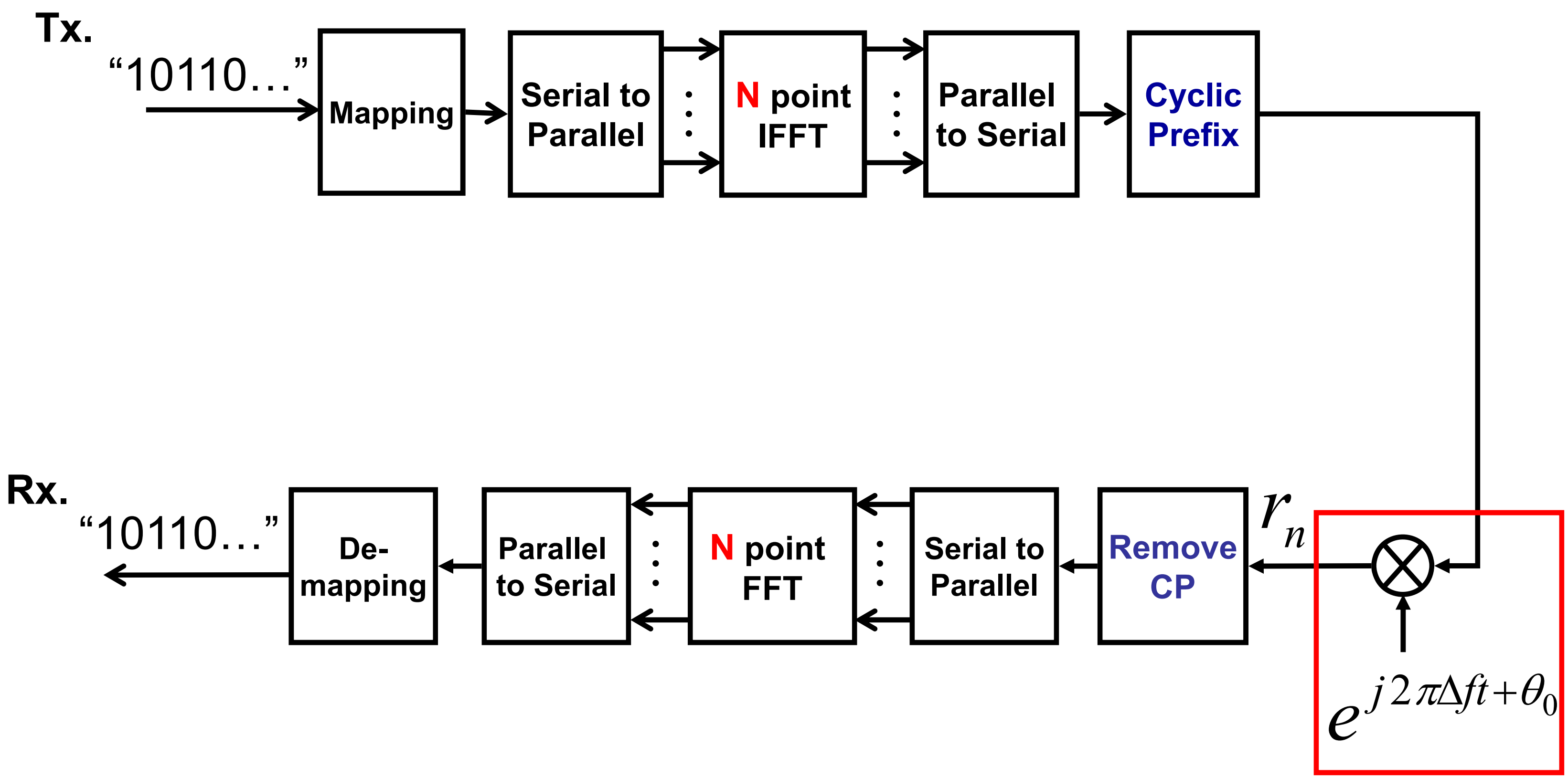


OFDM Demodulation (Non-windowing) (1/2)

- Without Frequency Offset

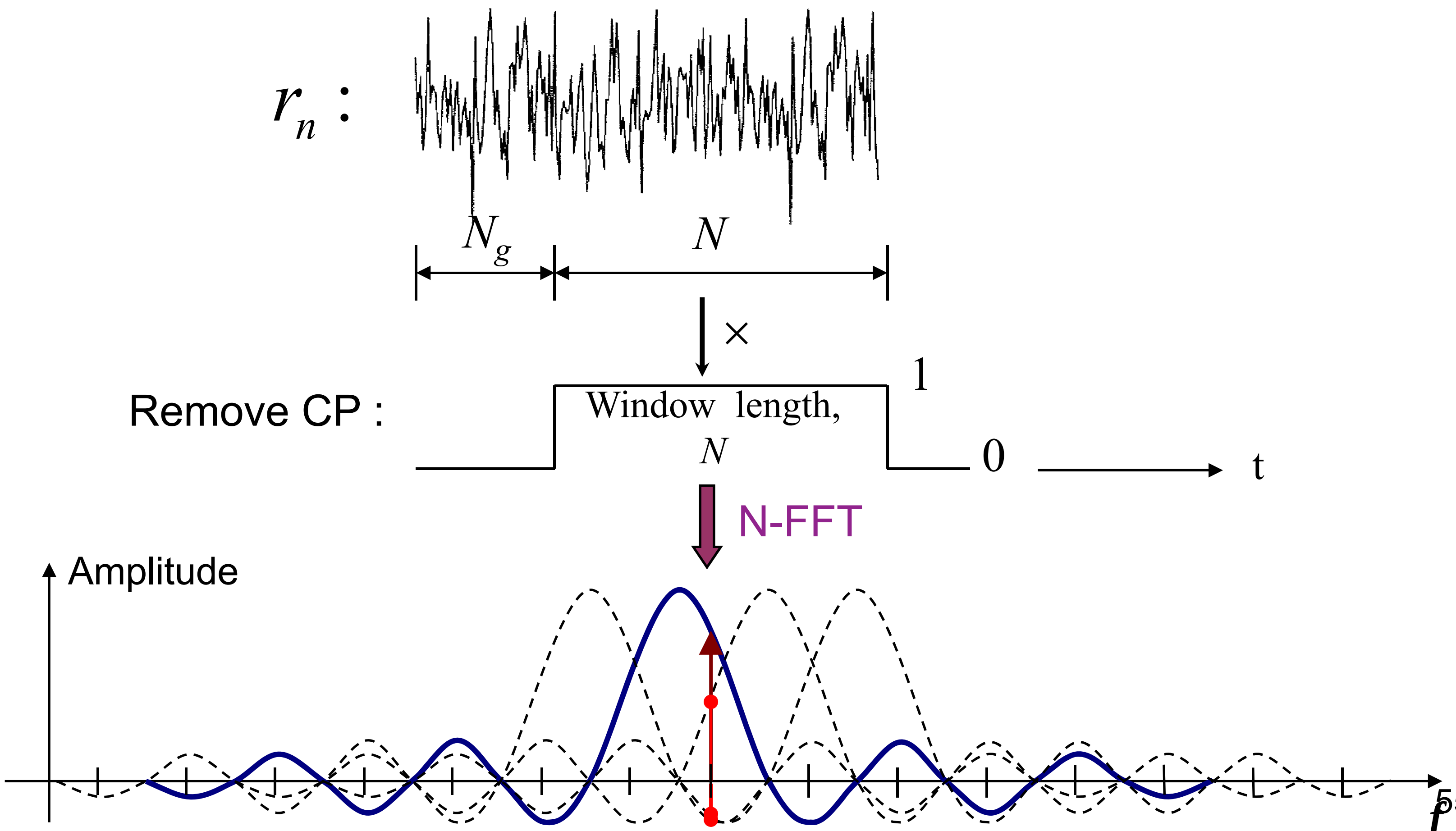


Oscillator Frequency Mismatch



OFDM Demodulation (Non-windowing) (2/2)

- With Frequency Offset (ICI)



ICI Induced by Constant Frequency Offset

$$Y_m = \frac{1}{N}\sum_{n=0}^{N-1} r_n e^{-j2\pi\frac{m}{N}n}$$

$$= \frac{1}{N}\sum_{n=0}^{N-1}\left(\sum_{k=0}^{N-1} X_k e^{j2\pi\frac{k+\Delta fT}{N}n+\theta_0}\right)e^{-j2\pi\frac{m}{N}n}$$

$$= \frac{1}{N}e^{j\theta_0}\sum_{k=0}^{N-1} X_k \sum_{n=0}^{N-1} e^{j2\pi\frac{(k-m)+\Delta fT}{N}n}$$

$$= e^{j\theta_0}\sum_{k=0}^{N-1} C_{k-m} X_k$$

where

$$C_{k-m} = \frac{1}{N}\sum_{n=0}^{N-1} e^{j2\pi\frac{(k-m)+\Delta fT}{N}n}$$

$$= \frac{1}{N}\cdot\frac{\sin\pi(k-m+\Delta fT)}{\sin\pi\left(\frac{k-m+\Delta fT}{N}\right)}\cdot e^{j\pi\frac{(N-1)(k-m+\Delta fT)}{N}}$$

Leakage Coefficients

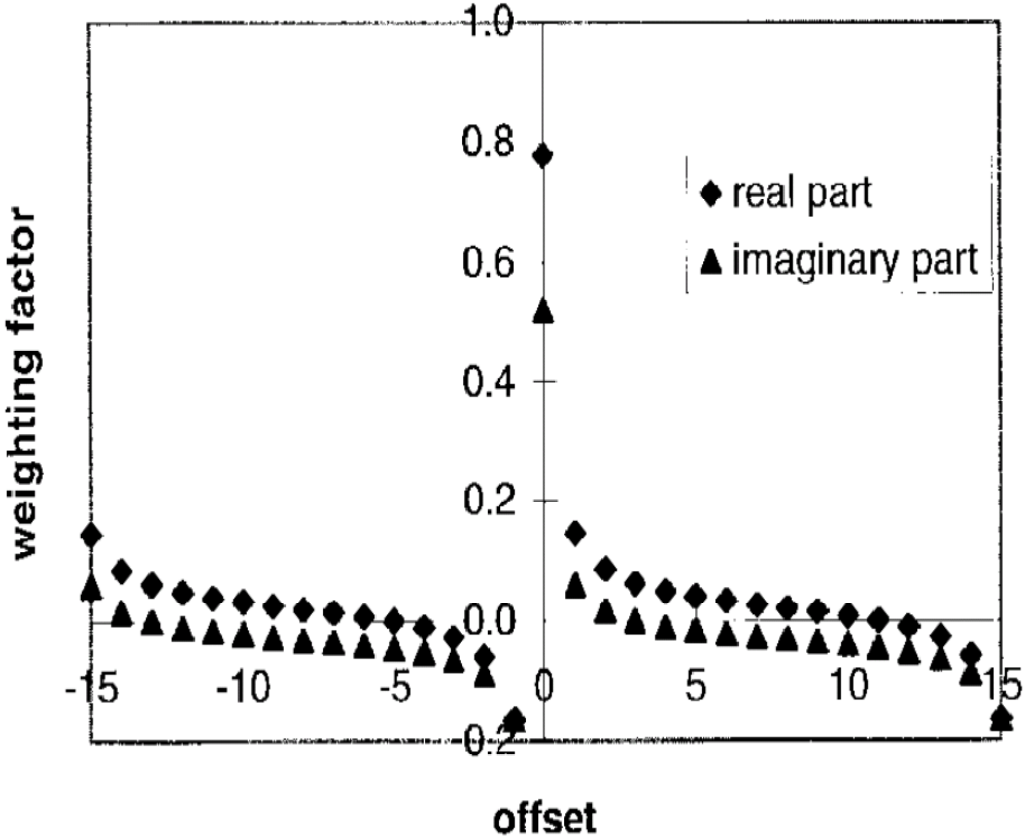


Fig. Leakage coefficients for N=16 and ΔfT=0.2

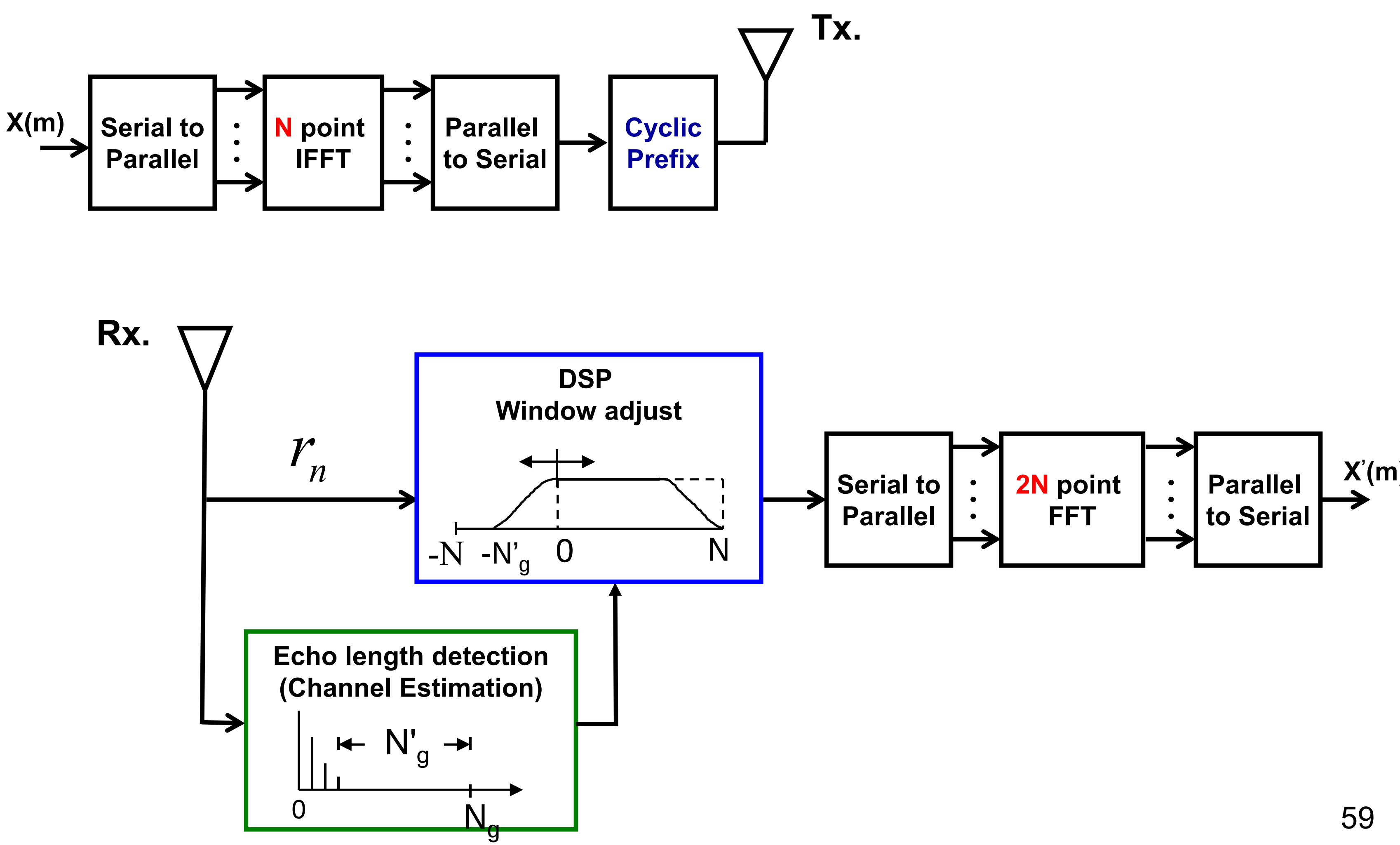
※ [] “Analysis of new and existing methods of reducing intercarrier interference due to carrier frequency offset in OFDM,” *IEEE Trans. Commun.* ,1999.

ICI Reduction of OFDM Receiver Windowing

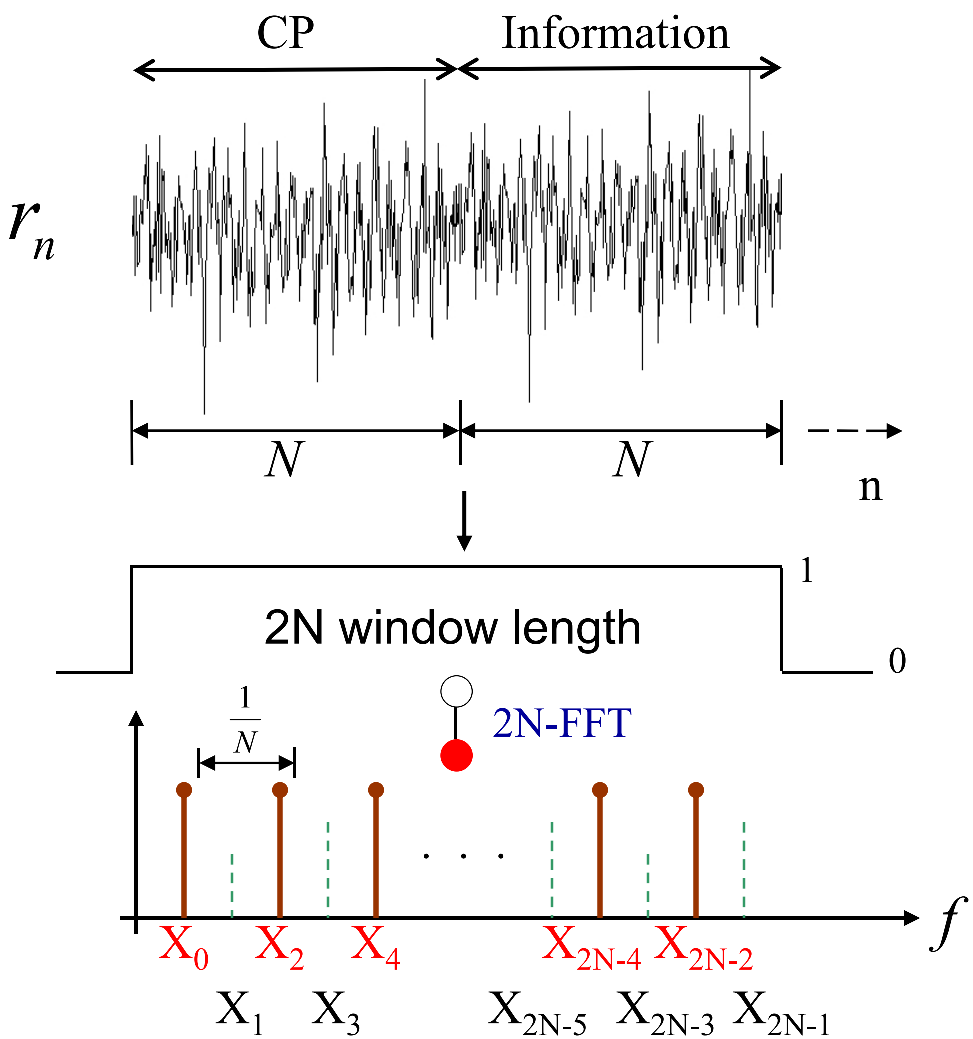
- [1] proposed the receiver windowing method by applying Raised-cosine window in time domain to suppress ICI effect.
- The windowing method utilizing the guard interval of ISI-free part to suppress the ICI effect.
- The Raised-cosine (Nyquist) window can preserve the orthogonality between subcarriers when there is no frequency offset.

※ [1] "Improving an OFDM reception using an adaptive Nyquist windowing," *IEEE Trans. Consumer Electron.*, vol. 42, pp. 259-269, Aug. 1996.

Conventional Receiver Windowing (1/2)



Double Length FFT and Square Windwoing



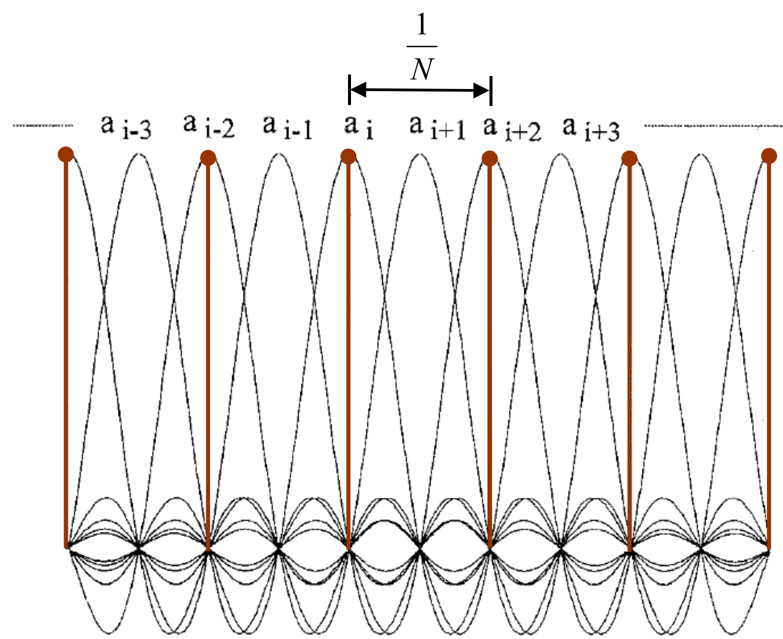
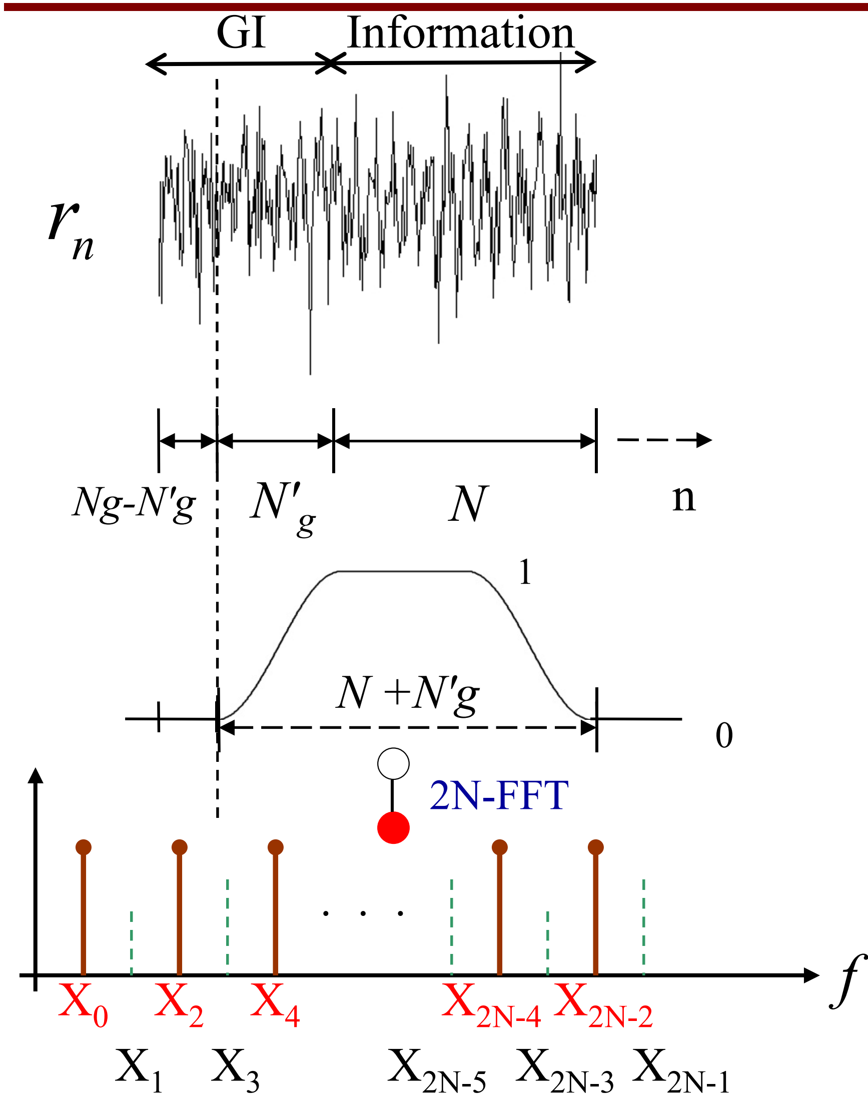


Fig. Filter bank of double-length DFT

Adaptive Nyquist Windowing



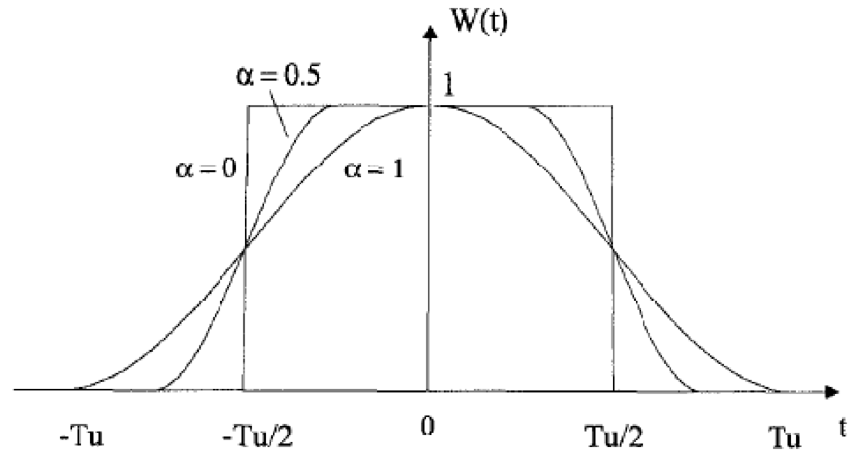


Fig. Raised cosine window function

$$\alpha \equiv \frac{N'_g}{N}$$

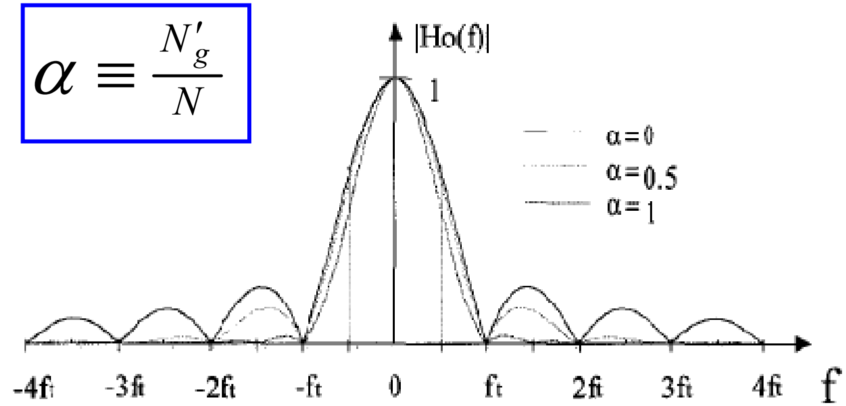
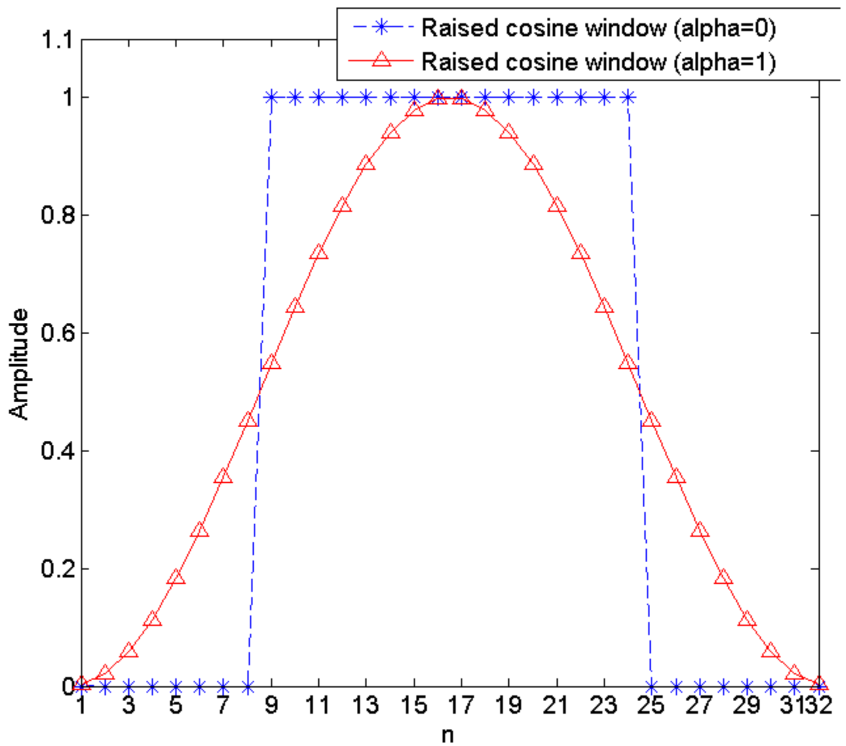


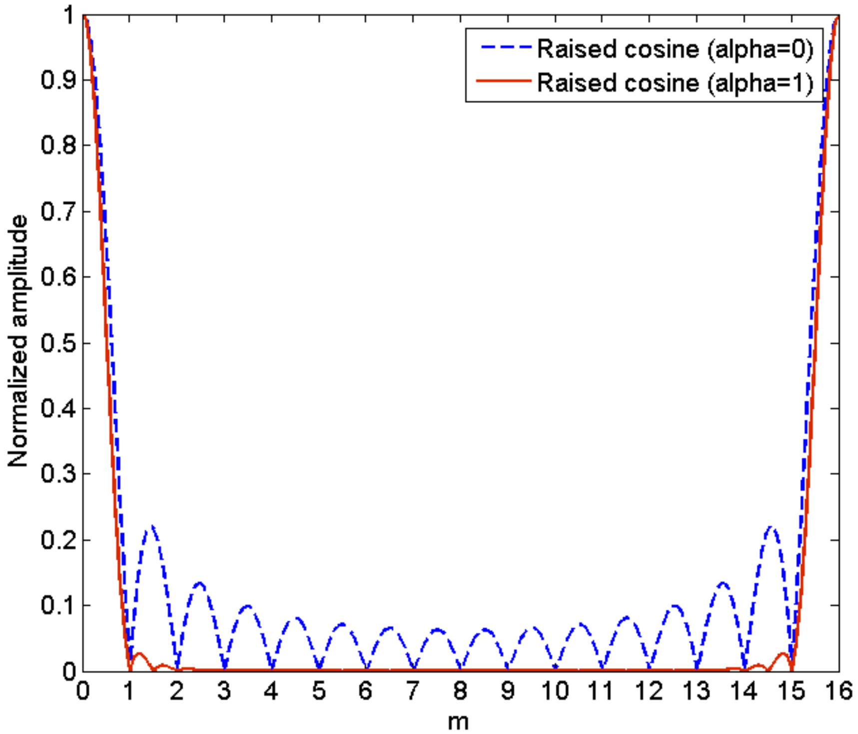
Fig. Frequency response of Raised cosine

Side Lobe Comparison

- Raised-cosine for different α
  - N=16 , $0 \leqq N'_g \leqq 16$



time domain



frequency domain

Signal to Interference Ratio

- The desired signal power

$$\sigma_m^2 = E\left\{\left|X_m\right|^2\right\}\left|\mathrm{W}(\Delta f)\right|^2$$

- The ICI power

$$\sigma_{ICI_m}^2 = \sum_{k=0,k\neq m}^{N-1}\sum_{p=0,p\neq m}^{N-1} E\left\{X_k X_p^*\right\}\mathrm{W}(\tfrac{m\text{-}k}{T_u}+\Delta f)\mathrm{W}^*(\tfrac{p\text{-}k}{T_u}+\Delta f)$$

- The average signal to Interference ratio

$$SIR = \frac{\left|\mathrm{W}(\Delta f)\right|^2}{\sum_{k=0,k\neq m}^{N-1}\left|\mathrm{W}(\tfrac{m\text{-}k}{T_u}+\Delta f)\right|^2}$$

SIR Comparison of Different Window

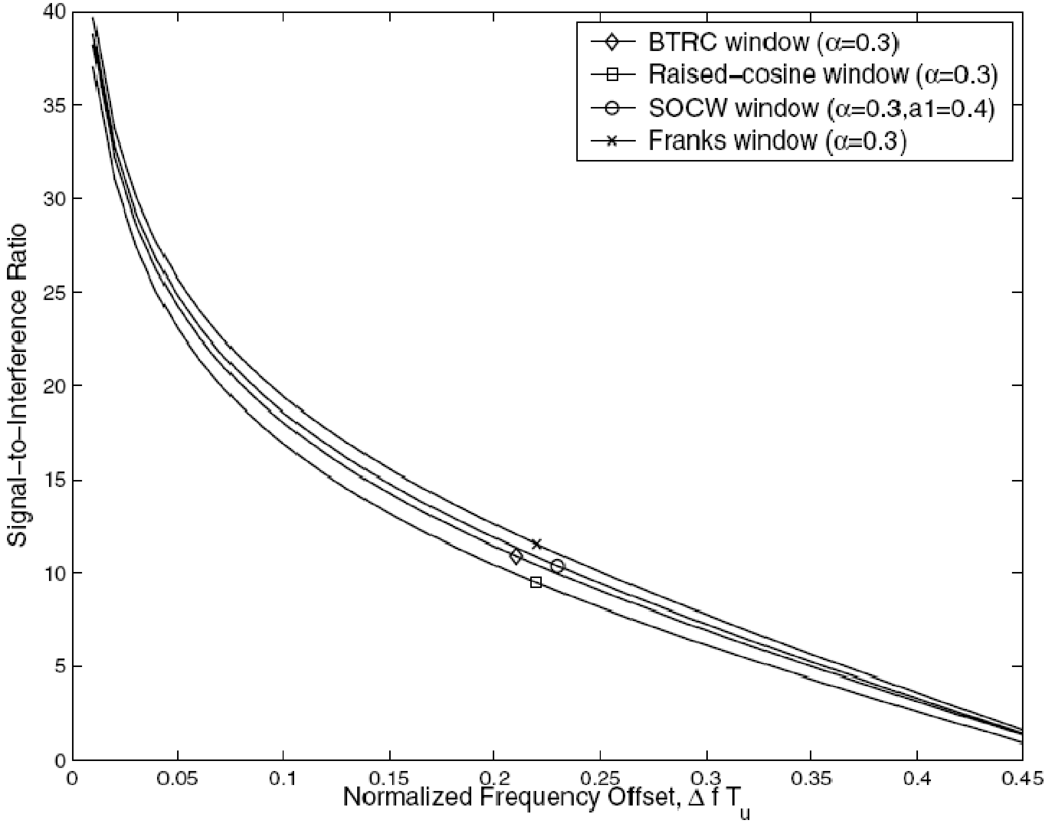
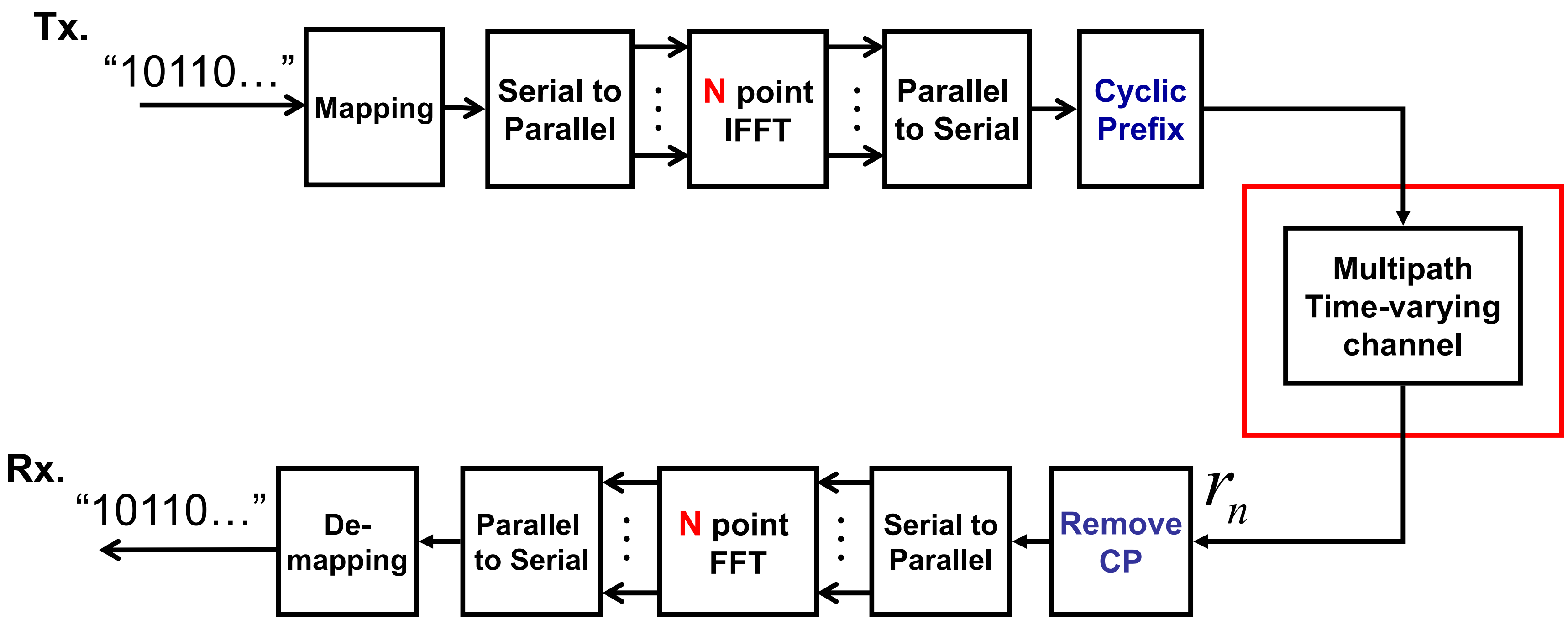


Fig. The SIR in 64-subcarrier OFDM receiver windowing with the employment of different window functions (α = 0.3).

※ "Receiver windowing for reduction of ICI in OFDM systems with carrier frequency offset," *in Proc. IEEE GLOBECOM Conf., vol. 5, pp. 2680-2684, Nov. 2005.*

In Time-varying Channel



ICI Induced by Time-varying Channel

$$Y_m = \frac{1}{N}\sum_{n=0}^{N-1} r_n e^{-j2\pi\frac{n}{N}m}$$

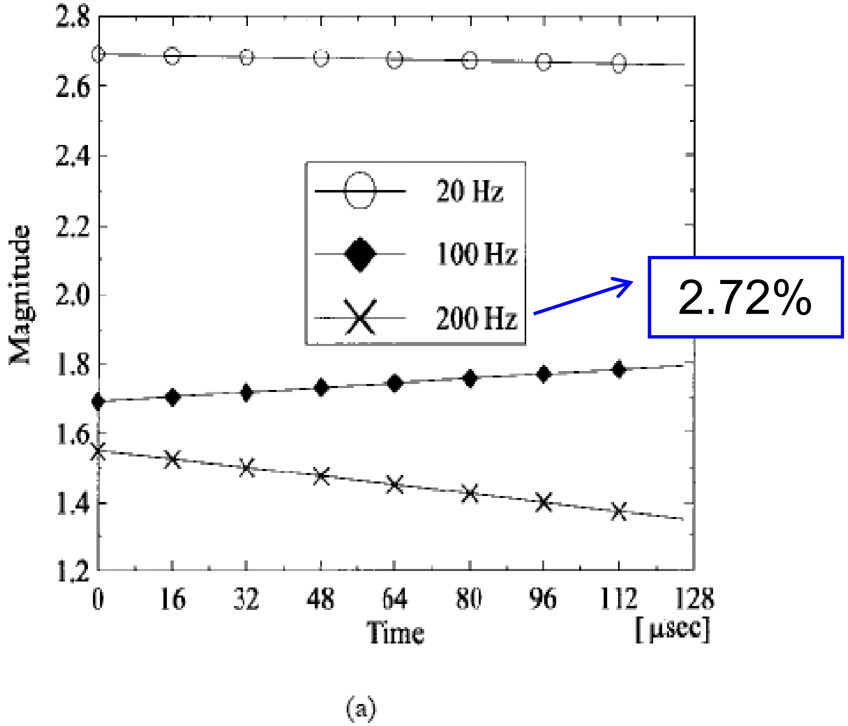
$$= \frac{1}{N}\sum_{n=0}^{N-1}\left(\sum_{l=0}^{L-1} h_n^{(l)} x_{n-\tau^{(l)}} + z_n\right) e^{-j2\pi\frac{n}{N}m}$$

$$= \frac{1}{N}\sum_{n=0}^{N-1}\left\{\sum_{l=0}^{L-1} h_n^{(l)}\left(\sum_{k=0}^{N-1} X_k e^{-j2\pi\frac{k}{N}\tau^{(l)}} e^{j2\pi\frac{n}{N}k}\right) + z_n\right\} e^{-j2\pi\frac{n}{N}m}$$

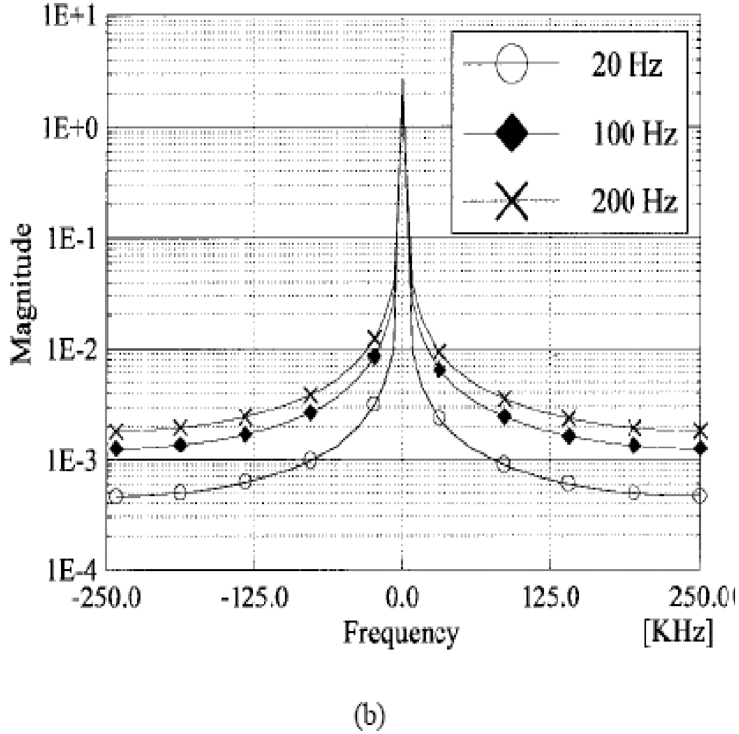
$$= \sum_{k=0}^{N-1} X_k \frac{1}{N}\sum_{l=0}^{L-1} e^{-j2\pi\frac{k}{N}\tau^{(l)}} \sum_{n=0}^{N-1} h_n^{(l)} e^{-j2\pi\frac{n}{N}(m-k)} + Z_m$$

$$= \underbrace{H_{m,m}X_m}_{desired\ \ signal} + \underbrace{\sum_{k=0,\ k\neq m}^{N-1} H_{m,k}X_k}_{ICI\ \ terms} + Z_m$$

Slow Time-varying Channel



(a) Time variation of the CIR for different Doppler frequencies within a block period



(b) Corresponding magnitude responses for different Doppler frequencies.

※ "An Equalization Technique for Orthogonal Frequency-Division Multiplexing Systems in Time-Variant Multipath Channels," *IEEE Trans. Commun.* ,1999

Pilot-Aided Channel Estimation

- Pilot extraction

$$\hat{H}_{k_q} = \frac{Y_{k_q}}{P_{k_q}} = H_{k_q} + \frac{I_{k_q} + W_{k_q}}{P_{k_q}} \quad , k_q = q \times \frac{N}{Q} \ , \ 0 \le q \le Q$$

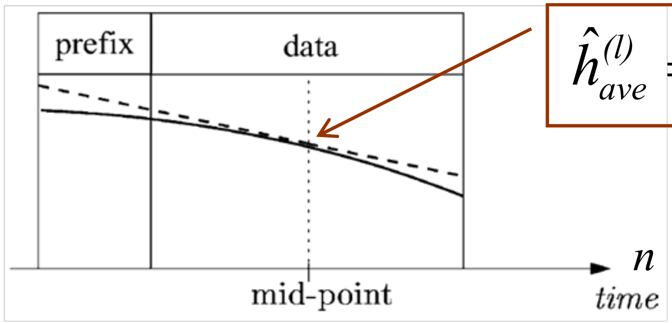
- The estimate of the l^{th} tap

$$\hat{h}_{ave}^{(l)} = \frac{1}{Q} \sum_{q=0}^{Q-1} \hat{H}_{k_q} e^{j\frac{2\pi q}{L} l} , \quad 0 \le l \le L\text{-}1$$

- Linear channel time-variation approximation

To minimize $E\left(\left| h_{ave}^{(l)} - h_n^{(l)} \right|^2 \right)$

➔ $n = \frac{N}{2} - 1$

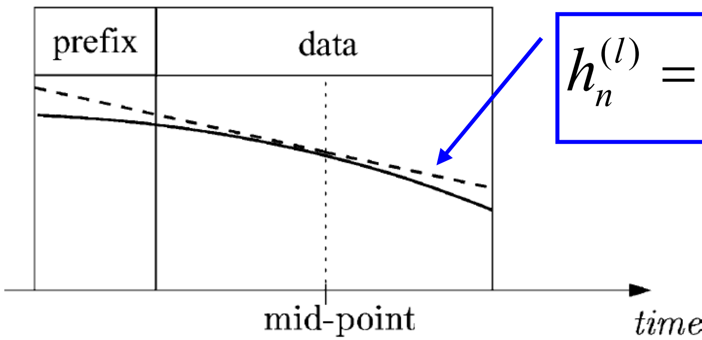


$$\hat{h}_{ave}^{(l)} = h_{\frac{N}{2}-1}^{(l)}$$

※ "ICI Mitigation for Pilot-Aided OFDM Mobile Systems," *IEEE Trans. Wireless Commun.* ,2005

Piece-wise Linear Time-variation Approximation (I)

- Linear time-varying channel approximation I :



$$h_n^{(l)} = h_{\frac{N}{2}-1}^{(l)} + \left(n+1-\tfrac{N}{2}\right)\times \alpha_l \times T_s \quad ,0 \le n \le N-1$$

- The signal model of piece-wise linear approximation in time domain

$$\vec{y} \approx (\mathbf{h}_{\mathrm{mid}} + \mathbf{M} \times \mathbf{A}) \times \vec{x} + \vec{w}$$

Piece-wise Linear Time-variation Approximation (I)

- The signal model of piece-wise linear approximation in frequency domain

$$\vec{Y} \approx \mathbf{H}_{\mathbf{mid}} \vec{X} + \mathbf{C} \times \mathbf{H}_{\mathbf{slope}} \vec{X}$$

where

$$\mathbf{H}_{\mathrm{mid}} = diag\left\{ FFT\left([h_0^{(\frac{N}{2}-1)} \; h_1^{(\frac{N}{2}-1)} \cdots h_{L-1}^{(\frac{N}{2}-1)} \; 0 \cdots 0] \right) \right\}$$

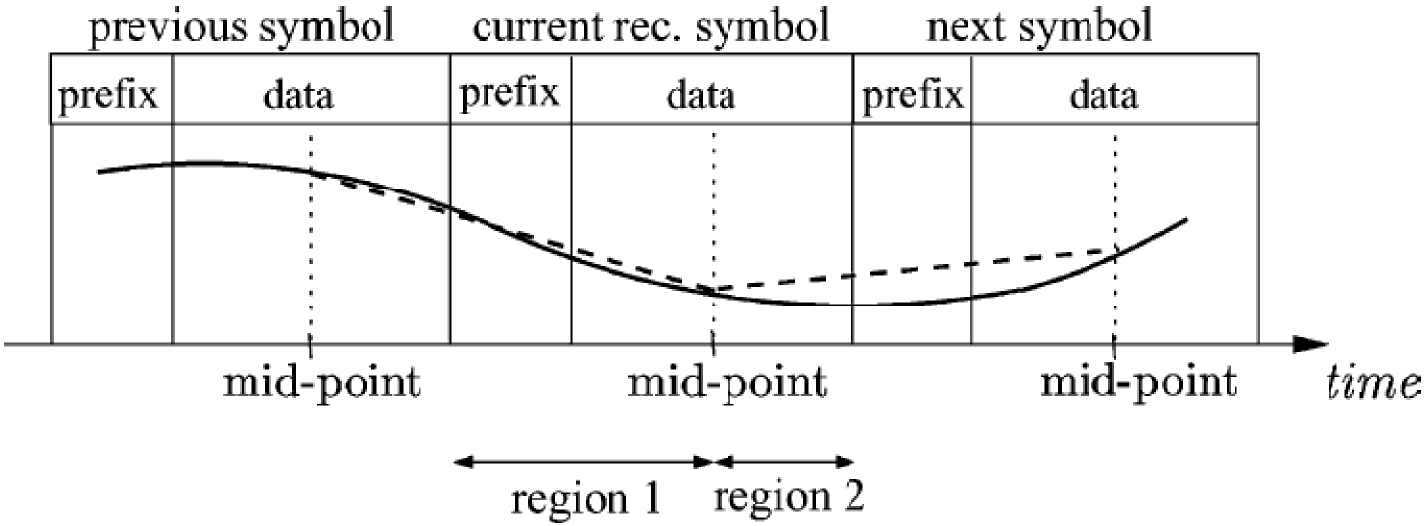
$$\mathbf{H}_{\mathrm{slope}} = diag\left\{ FFT\left([\alpha_0 \; \alpha_1 \cdots \alpha_{L-1} \; 0 \cdots 0] \right) \right\}$$

$$C_{m,m'} = \frac{B_{m-m'}}{N}$$

$$B_k = \mathrm{FFT}\left((n\text{-}\tfrac{N}{2}) \cdot T_s \right) = T_s \times N \times \begin{cases} -\dfrac{1}{1 - e^{-j\frac{2\pi k}{N}}} & k \neq 0 \\ 0.5 & k = 0 \end{cases}$$

Piece-wise Linear Time-variation Approximation (Ⅱ)

♦ Linear time-varying channel approximation Ⅱ：



$$\hat{\alpha}_l^{r_2} = \frac{h_l^{(\frac{N}{2}-1),next} - h_l^{(\frac{N}{2}-1)}}{T} \quad 0 \le l \le L-1$$

Piece-wise Linear Time-variation Approximation (Ⅱ)

- The signal model in time domain

$$\vec{y}_{\mathrm{method}_{II}} \approx (\mathbf{h}_{\mathrm{mid}} + \mathbf{M}^{r_1} \times \mathbf{A}^{r_1} + \mathbf{M}^{r_2} \times \mathbf{A}^{r_2}) \times \vec{x} + \vec{w}$$

- The signal model in frequency domain

$$\vec{Y}_{\mathrm{method}_{II}} \approx \mathbf{H}_{\mathrm{mid}} \vec{X} + \left(\mathbf{C}^{r_1} \times \mathbf{H}^{r_1}_{\mathrm{slope}} + \mathbf{C}^{r_2} \times \mathbf{H}^{r_2}_{\mathrm{slope}} \right) \times \vec{X} + \vec{W}$$

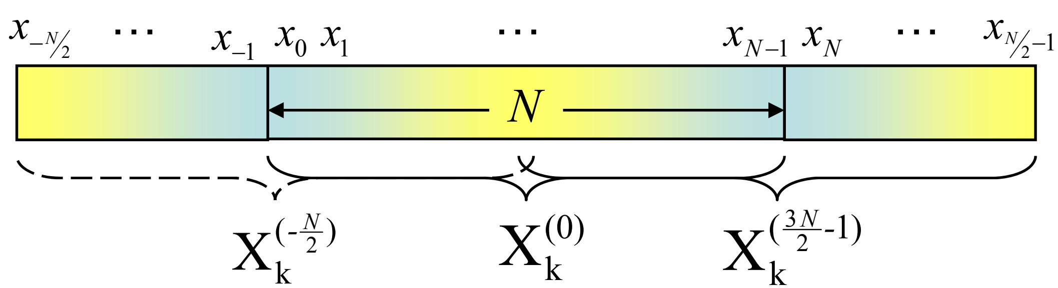
An ICI Self-cancellation Algorithm

- Periodic extend to 2N duration

$$x_n = \sum_{k=0}^{N-1} X_k e^{j2\pi\frac{kn}{N}} \quad , n = -\frac{N}{2}, \ldots, \frac{3N}{2} - 1$$

- Performs N-point FFT on shifting d samples

$$\mathrm{X}_\mathrm{k}^{(\mathrm{d})} = \mathrm{X}_\mathrm{k} \mathrm{e}^{\mathrm{j}2\pi\frac{\mathrm{kd}}{\mathrm{N}}} \quad , k = 0, \cdots, N\text{-}1$$



- $\mathrm{X}_\mathrm{k}^{(\mathrm{d})}$ can provide a diversity of X_k by multiplying $\mathrm{e}^{\mathrm{j}2\pi\frac{\mathrm{kd}}{\mathrm{N}}}$.

※ "A Novel Algorithm of Inter-Subchannel Interference Cancellation in OFDM Systems," in *Proc. IEEE Vehicular Technology Conf.*, vol. 1, pp. 460-464, Sept. 2004.

Combines $Y_m^{(d)}$ to Estimate Y_m

$$\hat{Y}_m = \frac{1}{N}\sum_{d=-N/2}^{N/2-1} Y_m^{(d)} \mathrm{e}^{-j2\pi\frac{m}{N}d}$$

$$= \frac{1}{N}\sum_{d=-N/2}^{N/2-1}\sum_{m'=0}^{N-1} H_d[m,m']\left(\mathrm{e}^{i2\pi\frac{(m'-m)d}{N}} X_{m'}\right)$$

$$+ \tilde{N}_m$$

$$= \left(\frac{1}{N}\sum_{d=-N/2}^{N/2-1} H_d[m,m]\right) X_m + \tilde{N}_m + \mathrm{ICI} \qquad (24)$$

and

$$\mathrm{ICI} = \frac{1}{N}\sum_{d=-N/2}^{N/2-1}\sum_{\substack{m'=0\\ m'\neq m}}^{N-1} H_d[m,m']\left(e^{j2\pi\frac{(m'-m)d}{N}} X_{m'}\right)$$

$$= \frac{1}{N^2}\sum_{l=1}^{1}\sum_{\substack{m'=0\\ m'\neq m}}^{N-1} e^{j2\pi\frac{m'}{T}\tau^{(l)}} X_{m'} \boxed{\sum_{d=-N/2}^{N/2-1}\sum_{k=0}^{N-1} h_{k+d}^{(l)} e^{j2\pi\frac{(m'-m)(k+d)}{N}}}$$

Self-Cancellation of the ICI

- ♦ Assume $h_k^{(l)}$ vary linearly within one symbol interval 2N.
- ♦ Let $h_{k+d}^{(l)} = (c_1^{(l)} k + c_1^{(l)} d + c_0^{(l)})$, where $c_0^{(l)}$, $c_1^{(l)}$ are constant coefficient.

Thus,

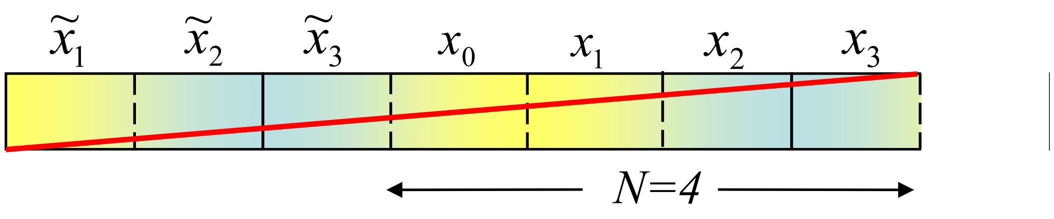
$$
\begin{aligned}
\mathrm{ICI} &= \sum_{d=-N/2}^{N/2-1} \sum_{k=0}^{N-1} h_{k+d}^{(l)} e^{j2\pi \frac{(m'-m)(k+d)}{N}} \\
&= \sum_{d=-N/2}^{N/2-1} \sum_{k=0}^{N-1} \left(c_1^{(l)} k + c_1^{(l)} d + c_0^{(l)} \right) e^{j2\pi \frac{(m'-m)k}{N}} e^{j2\pi \frac{(m'-m)d}{N}} \\
&= 0
\end{aligned}
$$

Further Extension from Chang's Method :

- Periodic extend to 2N-1 duration:

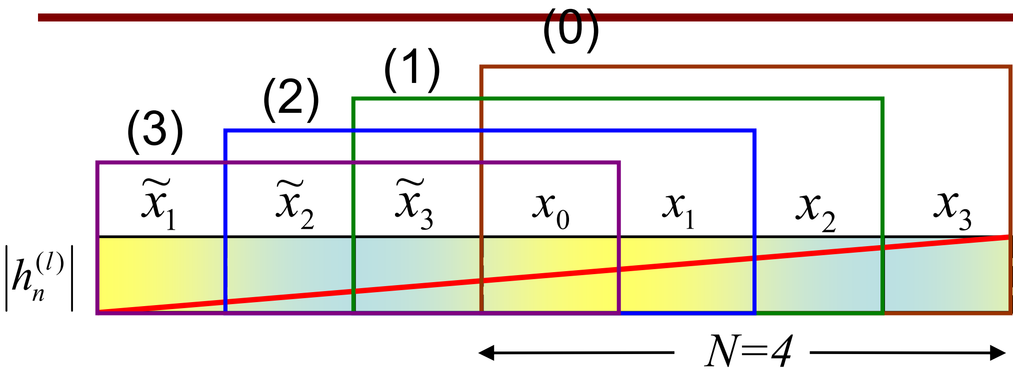
$$x_n = \sum_{k=0}^{N-1} X_k \cdot e^{j2\pi\frac{kn}{N}} \;,\; n = -(N-1),\ldots,N-1$$

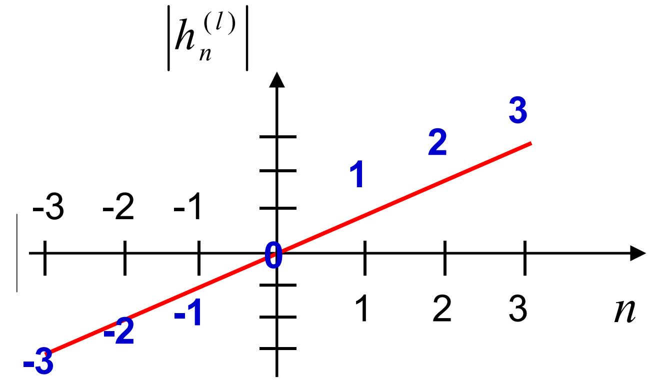
- If N=4 for extended OFDM symbol :



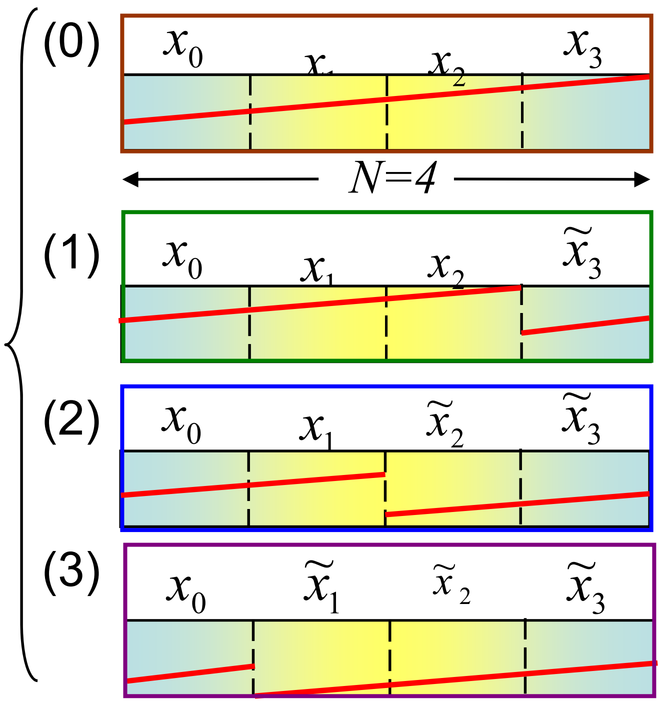
and assume linear time varying channel.

ICI self cancel – time domain combine





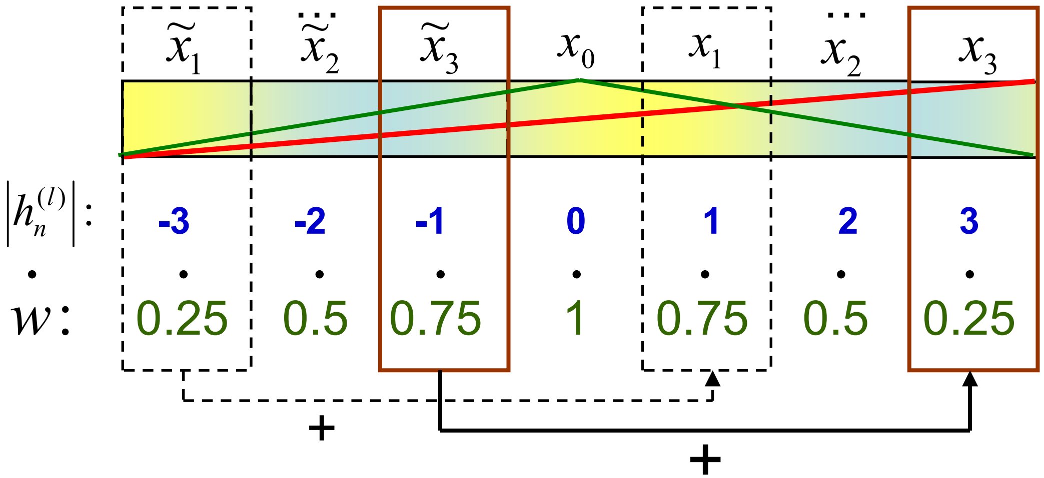
Circular shift & Combine



	x_0	x_1	x_2	x_3
(0) :	0	1	2	3
(1) :	0	1	2	-1
(2) :	0	1	-2	-1
+ (3) :	0	-3	-2	-1
	0	0	0	0
÷				4
$|h_n^{(l)}|$:	0	0	0	0

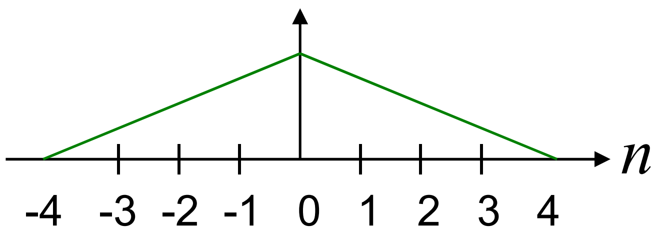
Windowing & Combining

- For N=4:



Triangular window shape

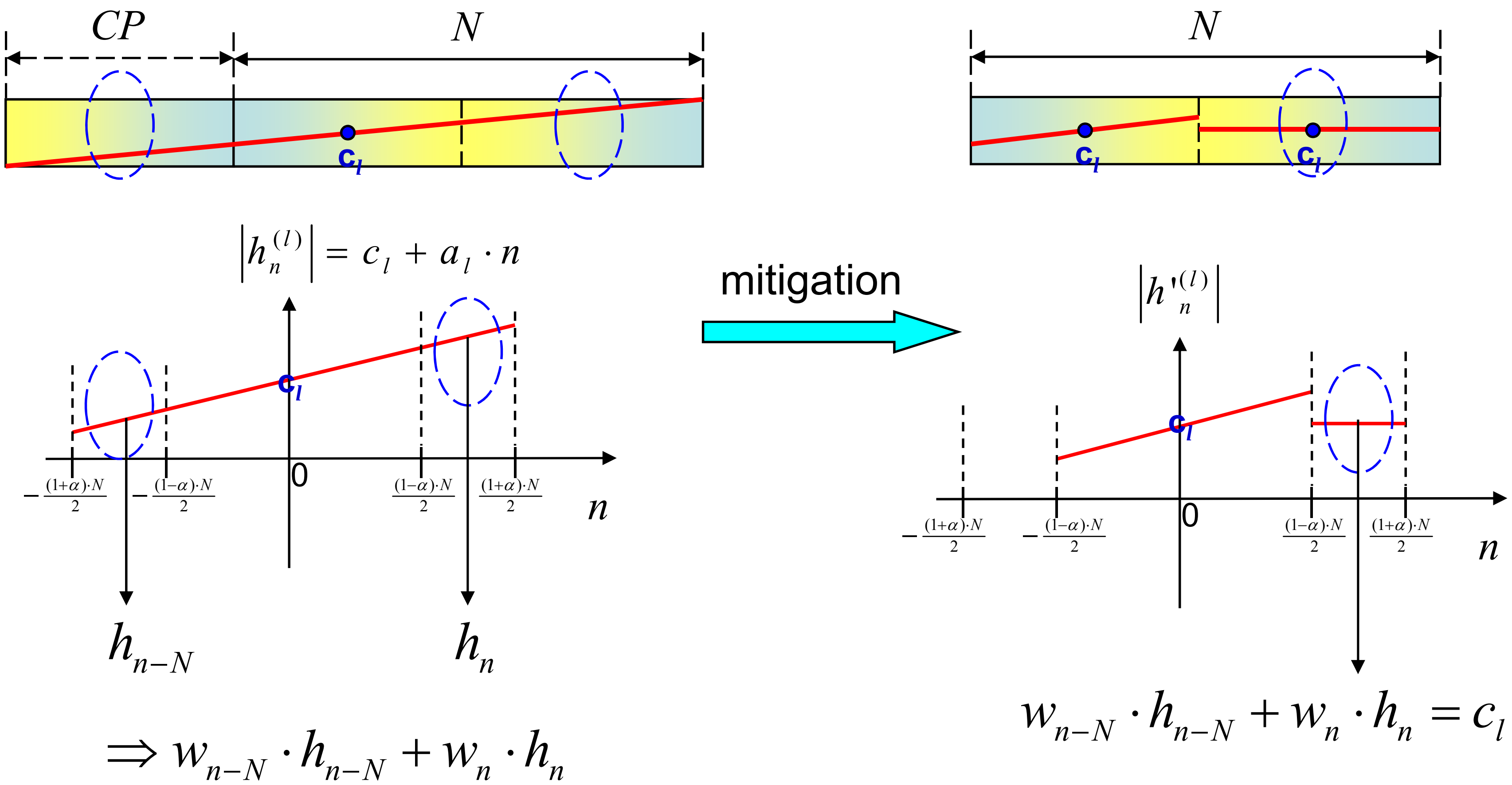
$$w_n = 1 - \frac{|n|}{N} = 1 - \frac{|n|}{4}$$



	x_0	x_1	x_2	x_3
	0	0.75	1	0.75
+		-0.75	-1	-0.75
$\|h_n^{(l)}\|$:	0	0	0	0

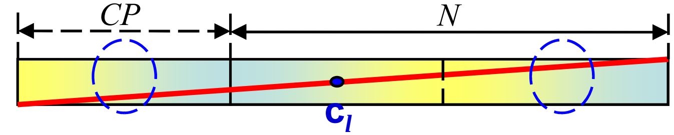
Window coefficients calculation (1/2)

- Assume linear time varying channel



Window coefficients calculation (2/2)

- For linear time varying channel : $\left|h_n^{(l)}\right| = c_l + a_l \cdot n$



- Combining,

$$w_n \cdot h_n + w_{n-N} \cdot h_{n-N} \quad , \frac{(1-\alpha)N}{2} \le n < \frac{(1+\alpha)N}{2}$$
$$= w_n \cdot (c_l + a_l \cdot n) + w_{n-N} \cdot [c_l + a_l \cdot (n-N)]$$
$$= \underline{(w_n + w_{n-N})} \cdot c_l + [\underline{w_n \cdot n + w_{n-N} \cdot (n-N)}] \cdot a_l$$

$$\text{Let} \Rightarrow \begin{cases} w_n + w_{n-N} = 1 & \cdots (1) \\ w_n \cdot n + w_{n-N} \cdot (n-N) = 0 & \cdots (2) \end{cases}$$

$$\therefore \begin{cases} w_{n-N} = w_{-(N-n)} = \frac{n}{N} \\ w_n = 1 - \frac{n}{N} \end{cases} , \frac{(1-\alpha)N}{2} \le n < \frac{(1+\alpha)N}{2}$$

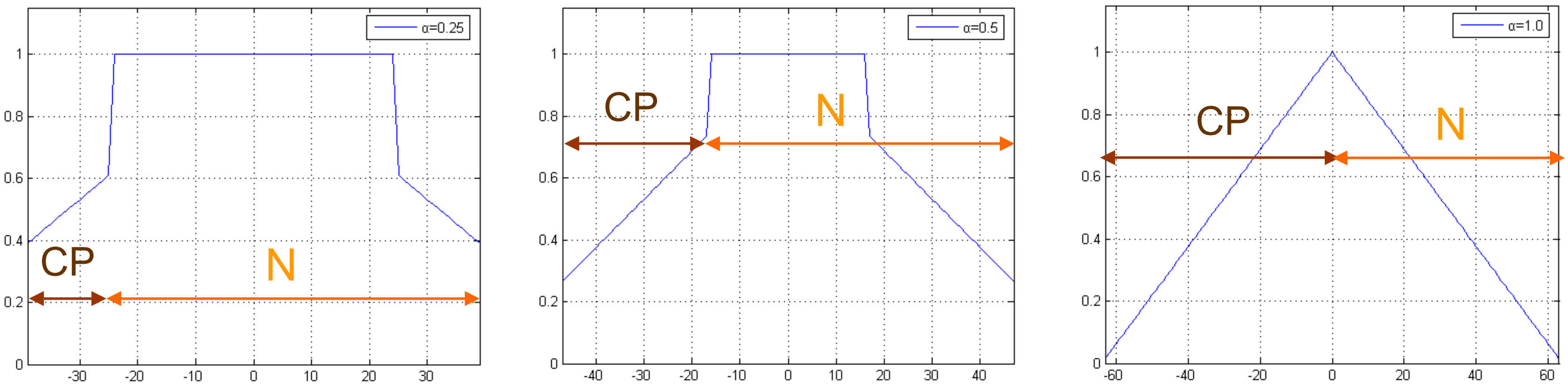
The coefficients are similar to Franks window

The Franks window

- Definition of Franks window :

$$p_f(t) = \begin{cases} 1, & 0 \le |t| < \frac{T_u(1-\alpha)}{2} \\ 1 - \frac{|t|}{T_u}, & \frac{T_u(1-\alpha)}{2} \le |t| < \frac{T_u(1+\alpha)}{2} \\ 0, & \text{otherwise} \end{cases}$$

- For N = 64 :

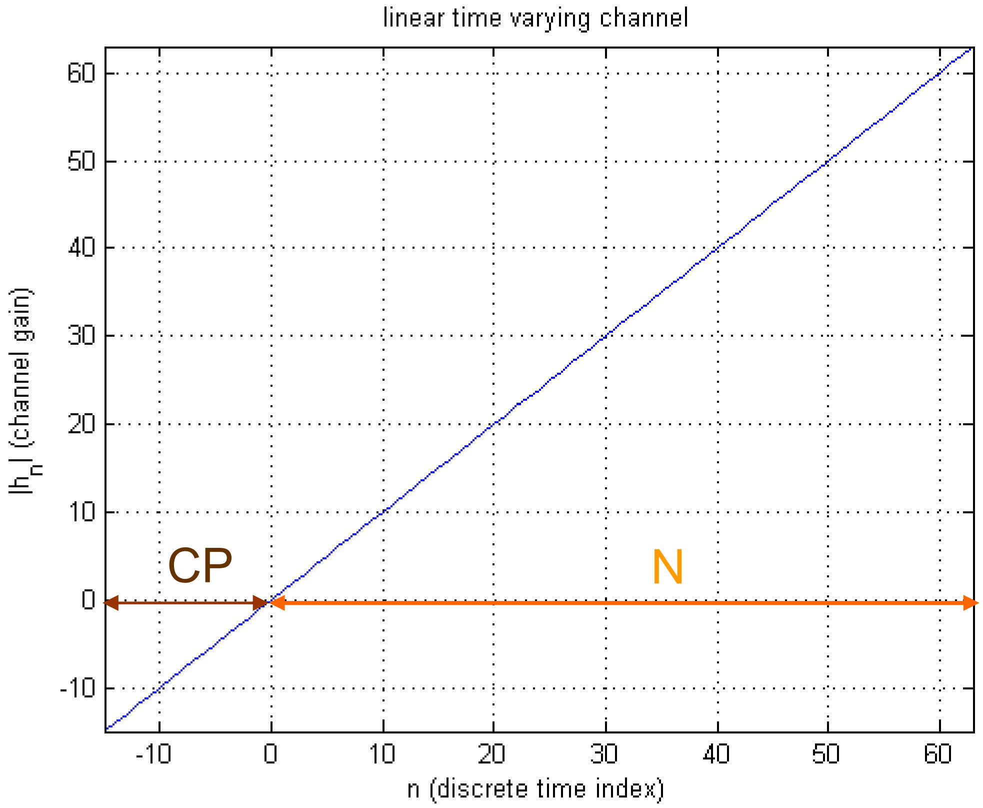


Useful CP = 15 points   Useful CP = 31 points   Useful CP = 63 points

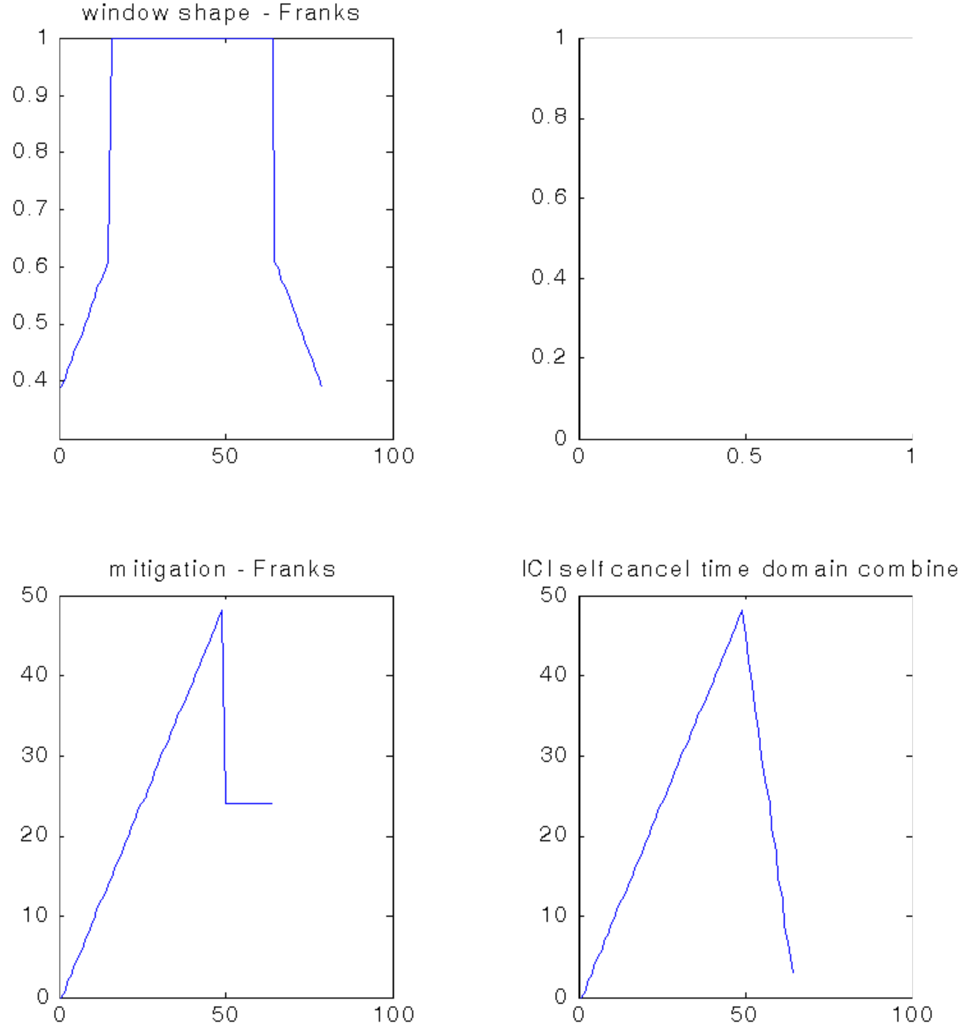
▲L.E. Franks, "Further results on Nyquist's problem in pulse transmission," *IEEE Trans. Commun. Technol.*, vol. 16, pp.337-340, Apr. 1968.

Mitigation for different window (1/2)

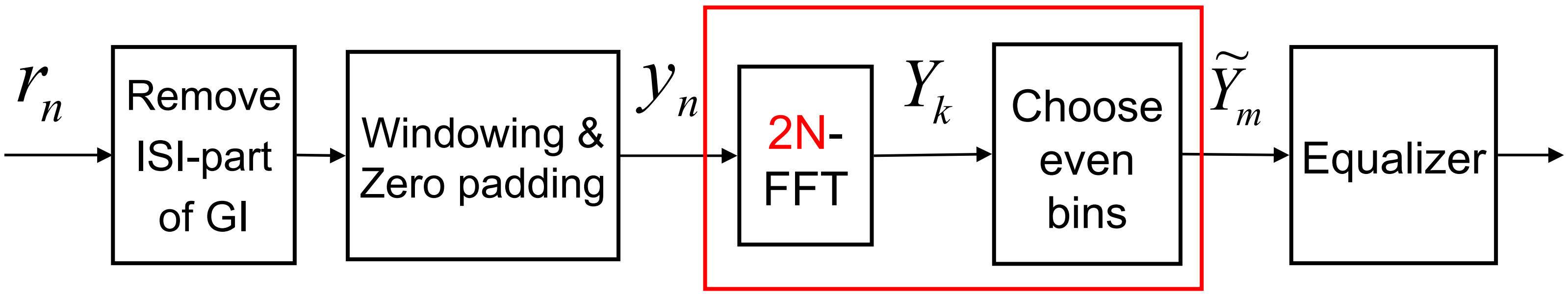
- Assume linear time varying channel : $|h_n| = n$ and $\alpha = 0.25$



Mitigation for different window (2/2)



Conventional Receiver Windowing



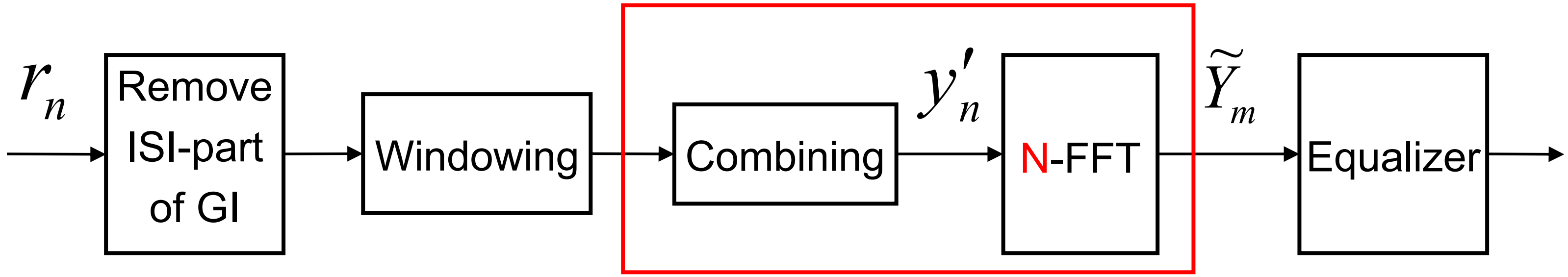
$$y_n = \begin{cases} 0 & , \ -N \le n \le -N'_g - 1 \\ w_n r_n & , \ -N'_g < n \le N - 1 \end{cases}$$

$$Y_k = \sum_{n=0}^{2N-1} y'_{n-N} e^{-j2\pi \frac{n}{2N} k} \ , \ 0 \le k \le 2N\text{-}1$$

$$\tilde{Y}_m = Y_{2m} \ , 0 \le m \le N\text{-}1$$

Low-complex Receiver Windowing Method

- The complexity of conventional receiver windowing method (2N-FFT) is twice as typical OFDM receiver.
- However, the 2N-FFT can be reduced to N-FFT by combing the useful GI to the corresponding information part.



Low-complex Receiver Windowing Method

$$\widetilde{Y}_m = Y_{2m}$$

$$= \sum_{n=0}^{2N\text{-}1} y_{n-N} e^{-j2\pi \frac{n}{2N} 2m}$$

$$= \sum_{n=0}^{N\text{-}1} \left[y_{n-N} e^{-j2\pi \frac{n}{N} m} + y_n e^{-j2\pi \frac{n+N}{N} m} \right]$$

$$= \sum_{n=0}^{N\text{-}1} \left[w_{n-N} r_{n\text{-}N} + w_n r_n \right] \cdot e^{-j2\pi \frac{n}{N} m}$$

$$= \sum_{n=0}^{N-N'_g-1} w_n r_n e^{-j2\pi \frac{n}{N} m} + \sum_{n=N-N'_g}^{N\text{-}1} (w_{n\text{-}N} r_{n\text{-}N} + w_n r_n) e^{-j2\pi \frac{n}{N} m}$$

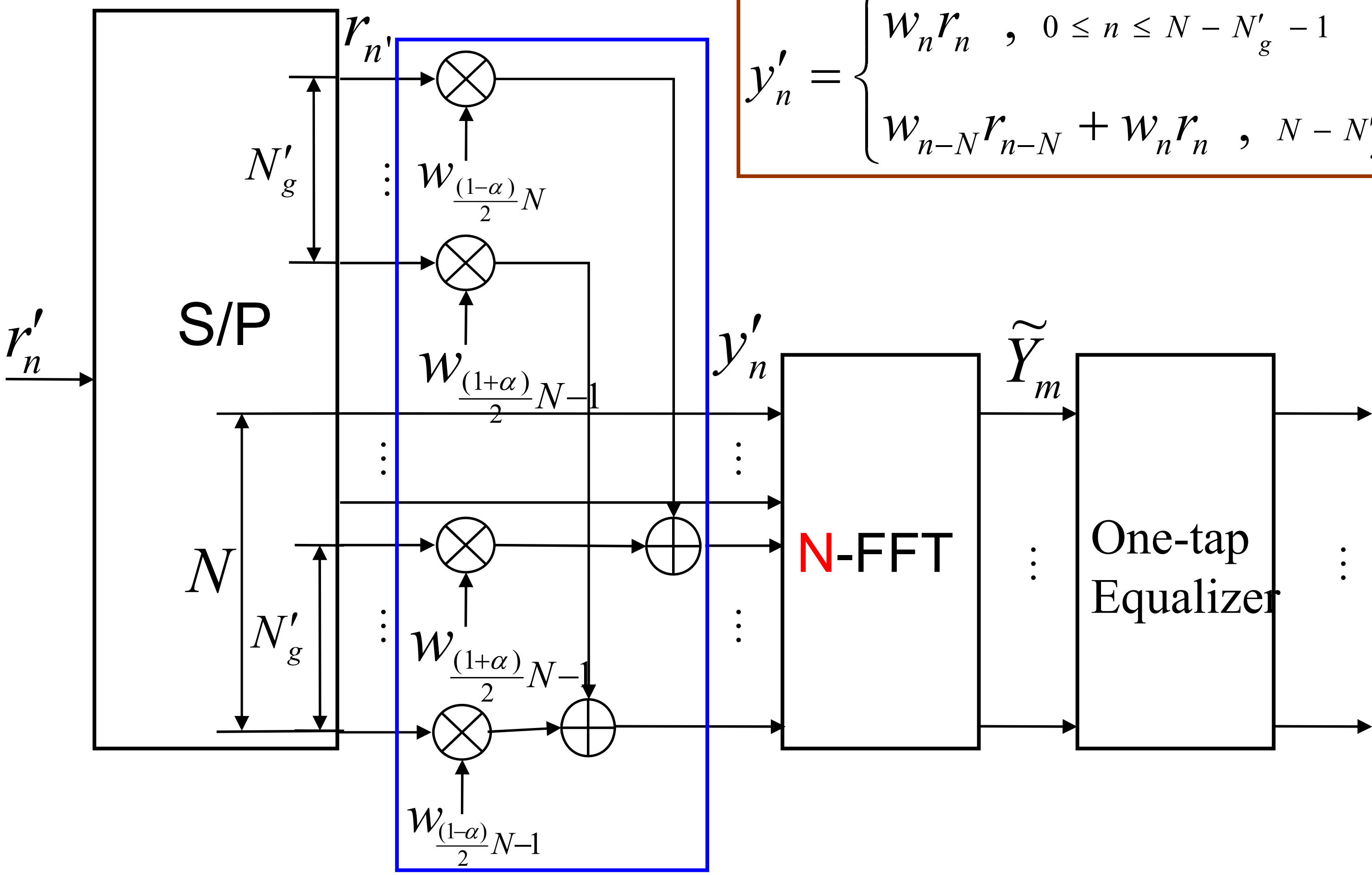
$$= \sum_{\mathrm{n}=0}^{\mathrm{N}\text{-}1} \mathrm{y}'_{\mathrm{n}} e^{-j2\pi \frac{n}{N} m} \quad , \; 0 \le m \le N-1$$

$$y_n = \begin{cases} 0 & , \; -N \le n \le -N'_g - 1 \\ w_n r_n & , \; -N'_g < n \le N-1 \end{cases}$$

$$y'_n = \begin{cases} w_n r_n & , \; 0 \le n \le N - N'_g - 1 \\ w_{n-N} r_{n-N} + w_n r_n & , \; N - N'_g < n \le N-1 \end{cases}$$

Low-complex Receiver Windowing Method

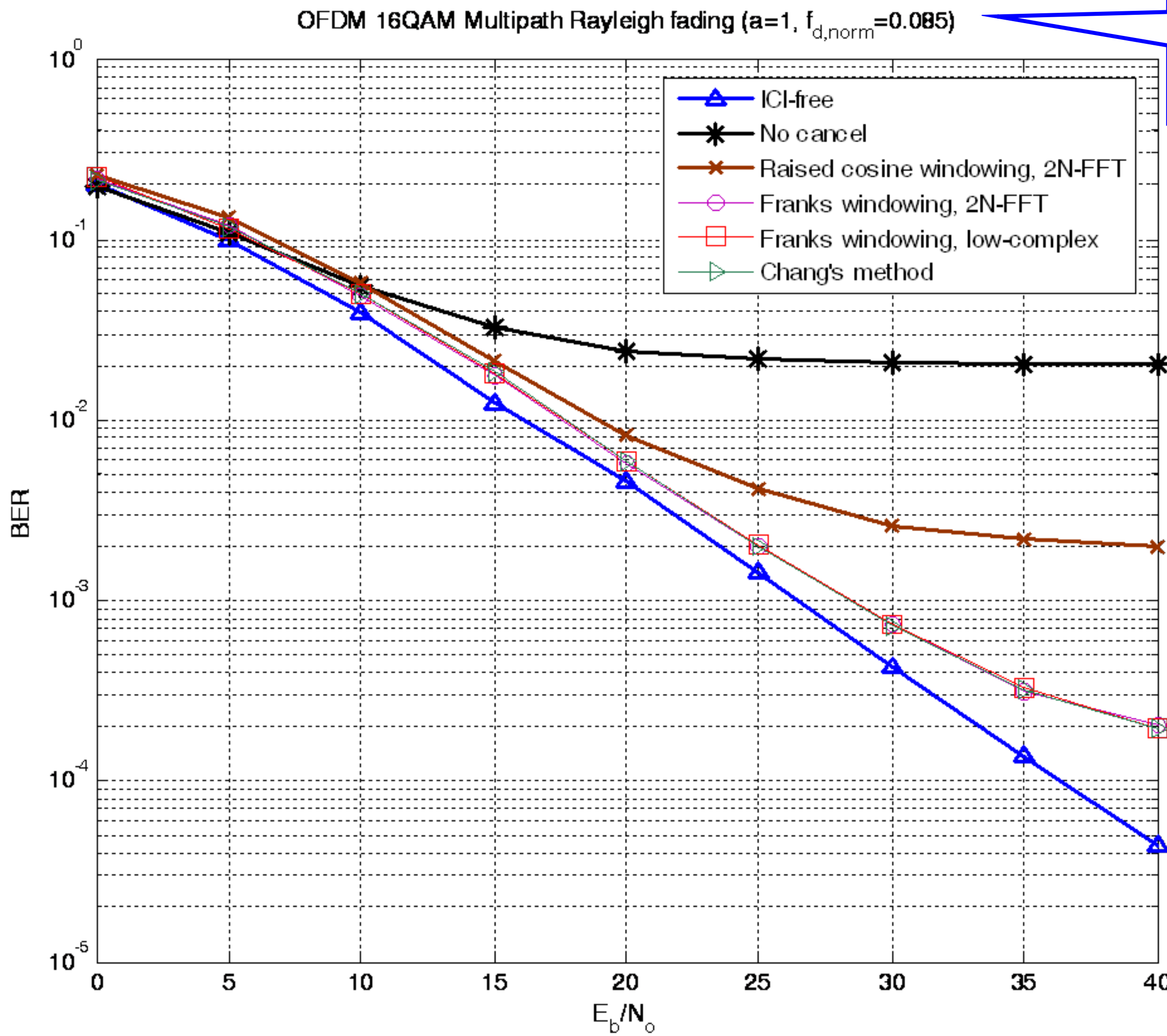
$$y'_n = \begin{cases} w_n r_n \quad , \; 0 \le n \le N - N'_g - 1 \\ w_{n-N} r_{n-N} + w_n r_n \quad , \; N - N'_g < n \le N - 1 \end{cases}$$



Simulation Parameter

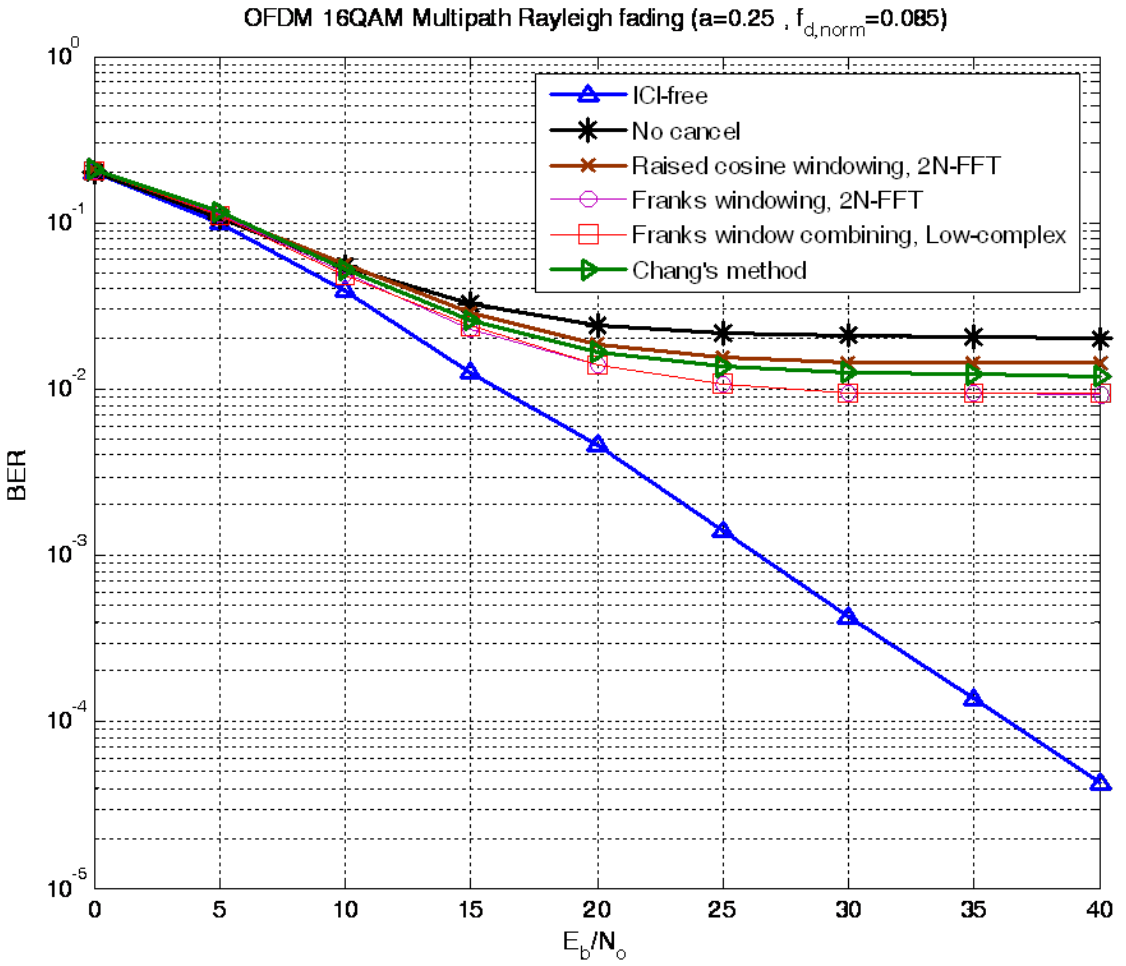
- DVB-T system parameters
- Time-varying multipath channel
  - TU6 channel model
  - Jake's model with Doppler
  - Rayleigh fading
- One-tap equalizer

Simulation Results (1/2)



127.5 km/Hr
(DVB-T 8Kmode 6Mhz)

Simulation Results (2/2)



Conclusion

- The complexity of conventional receiver windowing was reduced from 2N-FFT to N-FFT and receiving performance can be maintained.
- The Franks window can mitigate the channel time variation of corresponding part into constant in the assumption of linear time-varying channel.
- The performance of Franks window outperforms than raised-cosine window and Chang’s method.
- When **α** increased, the receiving performance can be more improved.

Thank you for your listening.