

Module 1: Basics of Digital Signal Processing

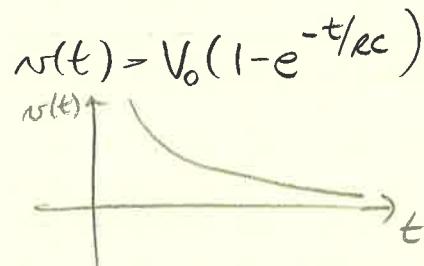
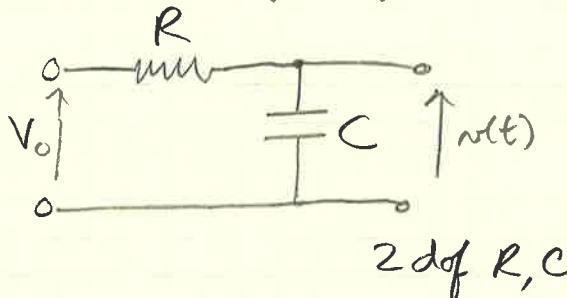
1.1 Introduction to digital signal processing

Signal: Description of evolution of a physical phenomenon

- Weather \rightarrow temperature
- Sound \rightarrow pressure
- Sound \rightarrow magnetic deviation
- Light intensity \rightarrow gray level on paper

Analysis: understanding the information carried by the signal

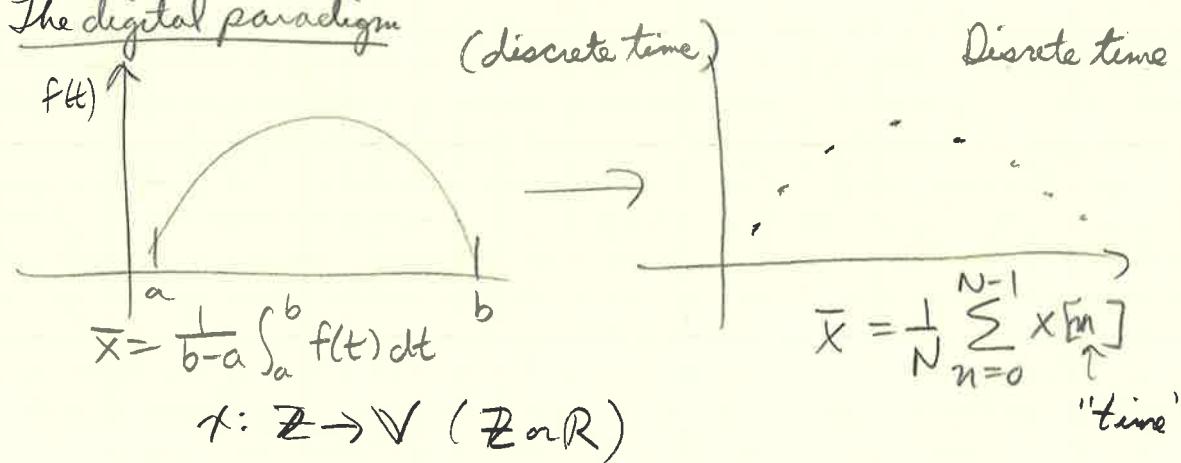
Synthesis: creating a signal to contain the given information



Analog signals $f: \mathbb{R} \rightarrow \mathbb{V}$

From analog to digital: $f(t) \rightarrow$ sample

The digital paradigm

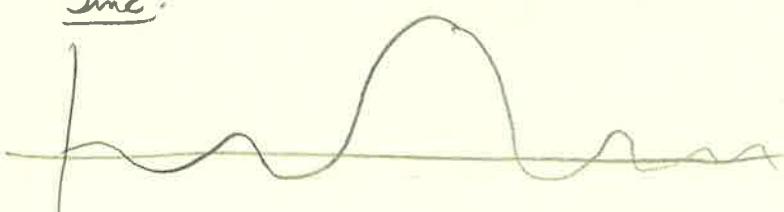


$x: \mathbb{Z} \rightarrow \mathbb{V} (\mathbb{Z} \text{ or } \mathbb{R})$

The Sampling Theorem (1920)

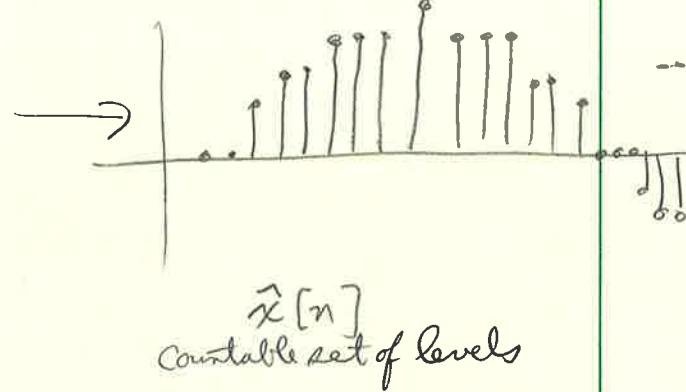
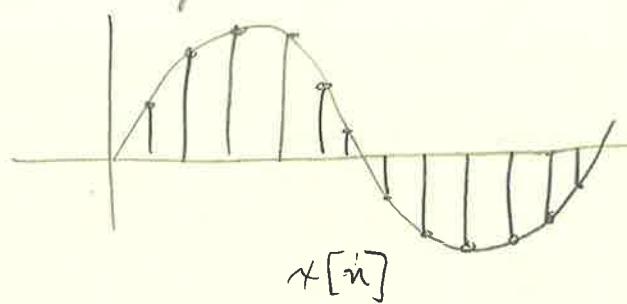
$$x(t) = \sum_{n=-\infty}^{\infty} x[n] \operatorname{sinc}\left(\frac{t-nT_s}{T_s}\right)$$

Sinc:



Infinite support

(discrete amplitude)



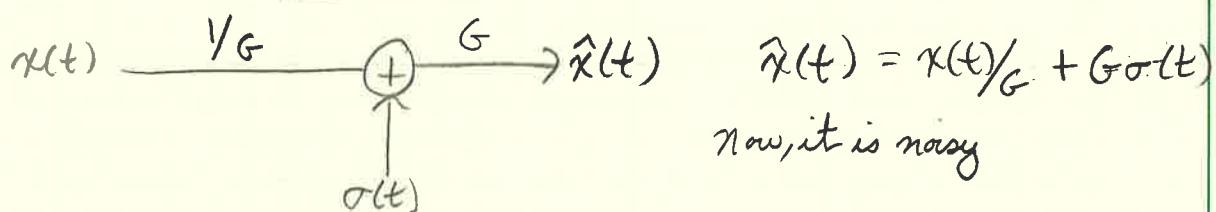
Why is it important?

- storage
- processing
- transmission

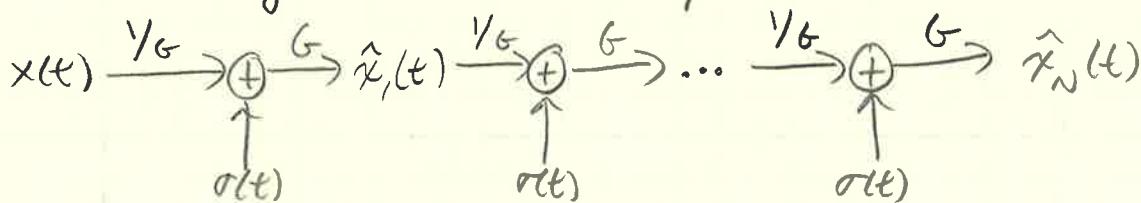
Digital Storage : $\{0, 1\}$

Data Transmission

TX \rightarrow channel \rightarrow RX

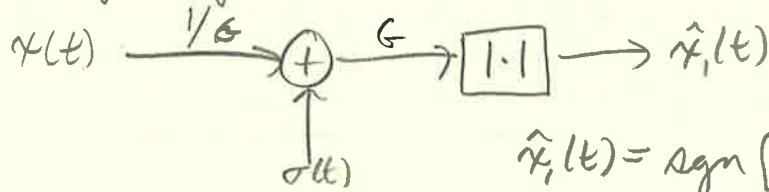


For a long channel, we need repeaters

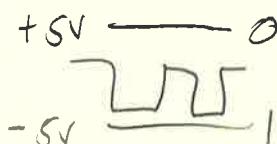


$$\hat{x}_N(t) = x(t) + NG\sigma(t)$$

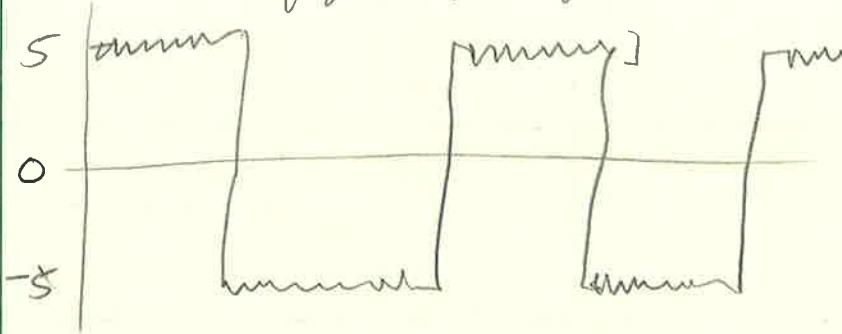
In digital signals, we can threshold



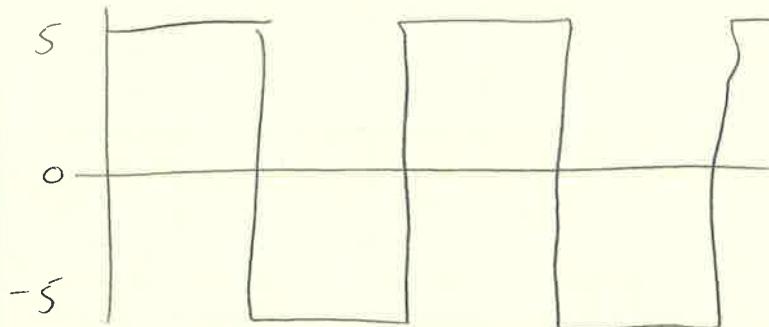
$$\hat{x}_1(t) = \text{sgn} [x(t) + G\sigma(t)]$$



Transmission of quantized signals



$$G(x(t)/G + \sigma(t)) = x(t) + G(\sigma(t))$$



$$\hat{x}(t) = G \operatorname{sgn}[x(t) + G\sigma(t)]$$

(after thresholding operator)

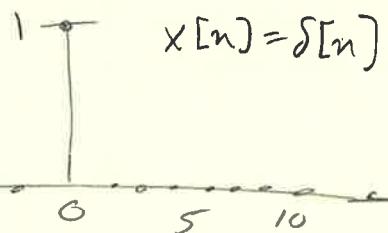
Digital Signal Processing: Key Ideas

- Discretization of time:
 - samples replace idealized models
 - simple math replaces calculus
- Discretization of values:
 - general-purpose of storage
 - general-purpose processing (CPU)
 - noise can be controlled

1.2 Discrete-time signals

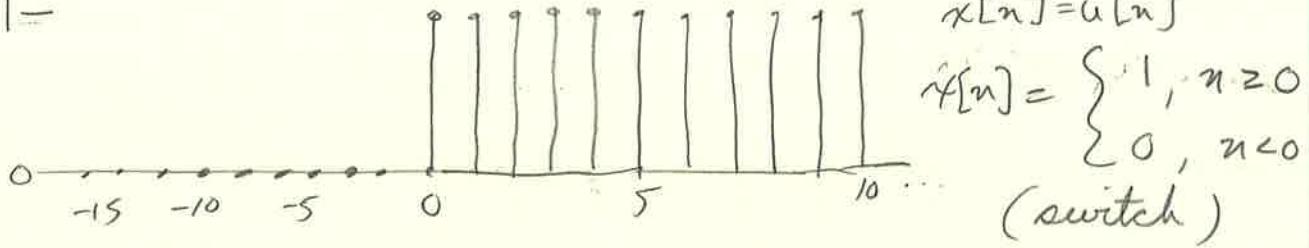
discrete-time signal : a sequence of complex numbers

- One dimension (for now)
- notation : $x[n]$
- two-sided sequences : $x: \mathbb{Z} \rightarrow \mathbb{C}$
- n is a-dimensional "time"
- analysis : periodic measurement
- synthesis : stream of generated samples

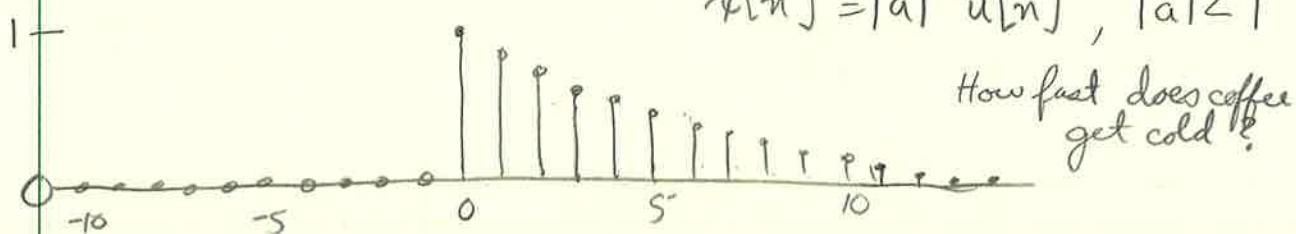


Ex: Used to synchronize audio and video in a movie

1-



Exponential decay



Newton's Law of cooling $\frac{dT}{dt} = -c(T - T_{\text{env}}) \Rightarrow T(t) = T_{\text{env}} + (T_0 - T_{\text{env}}) e^{-ct}$

Sine wave $x[n] = \sin(\omega_0 n + \theta)$, ω_0, θ in rad

Four signal classes

- finite-length
- infinite-length
- periodic
- finite-support

Finite-length signals

- sequence notation: $x[n], n=0, 1, \dots, N-1$
- vector notation: $\bar{x} = [x_0 \ x_1 \ \dots \ x_{N-1}]^T$
- practical entities, good for numerical packages (e.g. numpy)

Infinite-length signals

- sequence notation: $x[n], n \in \mathbb{Z}$
- abstraction, good for theorems

Periodic signals

- N -periodic sequence: $\tilde{x}[n] = \tilde{x}[n+kN], n, k, N \in \mathbb{Z}$
- same information as finite-length of length N
- "natural" bridge between finite and infinite lengths

Finite-support signals

- Finite-support sequence :

$$\bar{x}[n] = \begin{cases} x[n], & 0 \leq n \leq N \\ 0, & \text{otherwise} \end{cases} \quad n \in \mathbb{Z}$$

- same information as finite-length of length N
- another bridge between finite and infinite lengths

Elementary operators

- scaling : $y[n] = \alpha x[n], \alpha \in \mathbb{C}$
 - sum : $y[n] = x[n] + z[n]$
 - product : $y[n] = x[n] \cdot z[n]$
 - shift by k (delay) : $y[n] = x[n-k], k \in \mathbb{Z}$
- $\left. \begin{matrix} \\ \\ \end{matrix} \right\} 0 \leq n \leq N-1$

Shift of a finite-length : finite-support

$$\dots 000 \boxed{x_0 x_1 \dots x_7} 000 \dots$$

$\bar{x}[n]$

$$\dots 000 \boxed{0 x_0 x_1 x_2 x_3 x_4 x_5 x_6} x_7 0 0 \dots$$

$\bar{x}[n-1]$

$$\dots 000 \boxed{0 0 0 0 x_0 x_1 x_2 x_3} x_4 x_5 x_6 x_7 0 0 \dots$$

$\bar{x}[n-4]$

Shift of a finite length : periodic extension

$$\dots \boxed{x_0 x_1 x_2 x_3 x_4 x_5 x_6 x_7} \dots$$

$\bar{x}[n]$

$$\dots x_5 x_6 x_7 \boxed{x_0 x_1 x_2 x_3 x_4 x_5 x_6 x_7} x_8 x_9 \dots$$

$\tilde{x}[n]$

$$\dots x_4 x_5 x_6 \boxed{x_7 x_0 x_1 x_2 x_3 x_4 x_5 x_6} x_7 x_8 x_9 \dots$$

$\tilde{x}[n-1]$

$\cdots x_1 x_2 x_3 \boxed{x_4 x_5 x_6 x_7 x_8} x_1 x_2 x_3 x_4 x_5 x_6 \cdots$

Energy and power

$$E_x = \sum_{n=-\infty}^{\infty} |x[n]|^2$$

$$P_x = \lim_{N \rightarrow \infty} \frac{1}{2N+1} \sum_{n=-N}^{N} |x[n]|^2$$

Energy and power: periodic signals

$$\begin{aligned} E_{\tilde{x}} &= \infty \\ P_{\tilde{x}} &= \frac{1}{N} \sum_{n=0}^{N-1} |\tilde{x}[n]|^2 \end{aligned}$$

1.3 Basic signal processing

1.3.2 How your PC plays discrete-time sounds

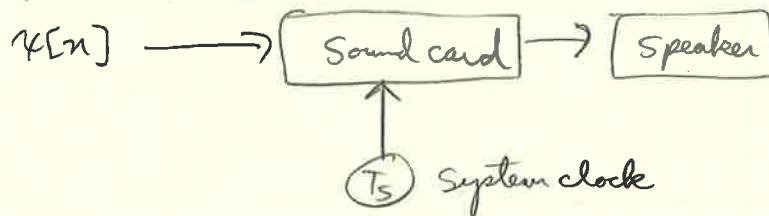
The discrete-time sinusoid

$$x[n] = \sin(\omega_0 n + \theta)$$

Digital vs. physical frequency

- Discrete time:
 - no: no physical dimension (just a counter)
 - periodicity: how many samples before pattern repeats
- Physical world:
 - periodicity: how many seconds before pattern repeats
 - frequency measured in Hz (s^{-1})

How your PC plays sounds



- set T_S , time in seconds between samples
- periodicity of M samples \rightarrow periodicity of MT_S seconds
- real world frequency: $f = \frac{1}{MT_S}$ Hz

- usually we choose F_s , the number of samples per second

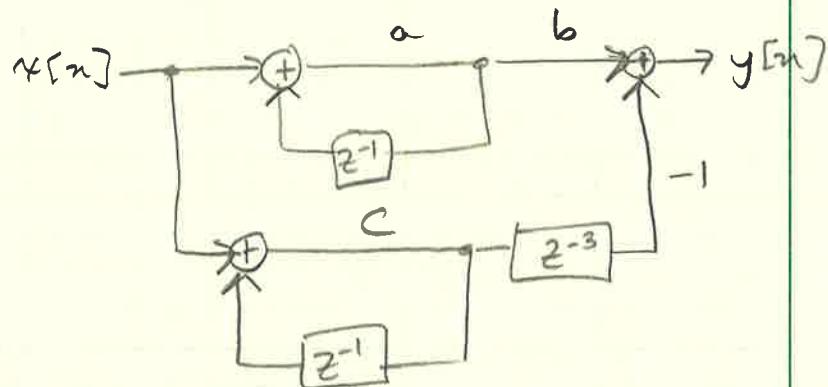
$$\cdot T_s = 1/F_s$$

Eg. for a typical value, $F_s = 48000 \text{ Hz}$, $T_s \approx 20.8 \mu\text{s}$.

If $M = 110$, $f \approx 440 \text{ Hz}$

1.3.6 The Kautus-Strong algorithm

DSP as Meccano



Building blocks:

- Adder: $x[n]$ $y[n]$ $\rightarrow x[n] + y[n]$

- Multiplier: $x[n] \xrightarrow{\alpha} \alpha x[n]$

- Unit Delay: $x[n] \rightarrow [z^{-1}] \rightarrow x[n-1]$

- Arbitrary Delay: $x[n] \rightarrow [z^{-N}] \rightarrow x[n-N]$

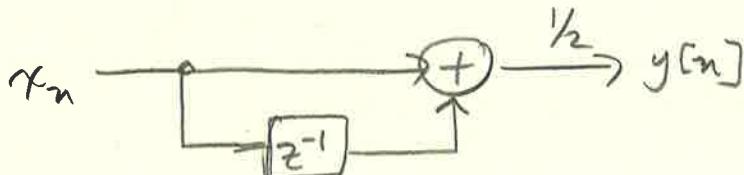
The 2-point Moving Average

- simple average: $M = \frac{a+b}{2}$

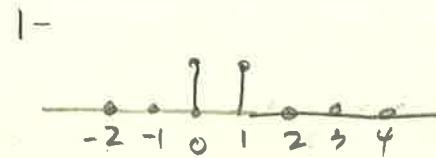
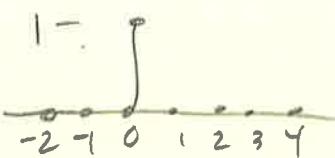
- moving average: take a "local" average

$$y[n] = \frac{x[n] + x[n-1]}{2}$$

- DSP Blocks:



Ex: $x[n] = \delta[n]$



$$y[0] = \frac{x[0] + x[-1]}{2} = \frac{1+0}{2} = \frac{1}{2}$$

$$y[1] = \frac{x[1] + x[0]}{2} = \frac{1+1}{2} = 1$$

- $x[n] = u[n]$

$$y[0] = \frac{x[0] + x[-1]}{2} = \frac{1+0}{2} = \frac{1}{2}$$

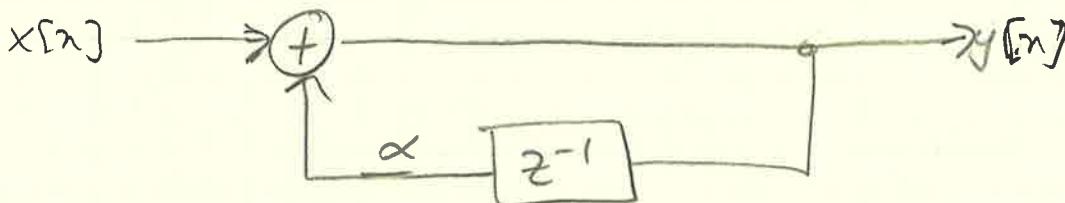
$$y[1] = \frac{x[1] + x[0]}{2} = \frac{1+1}{2} = 1$$

- $x[n] = \cos(\omega n)$, $\omega = \pi/10$

$$y[n] = \frac{\cos \omega n - \cos \omega(n-1)}{2} = \cos(\omega n + \theta)$$

- $x[n] = (-1)^n \Rightarrow y[n] = 0, \forall n$

What if we reverse the loop?



$$y[n] = x[n] + \alpha y[n-1], \alpha \in \mathbb{R}$$

(recursion)

How we solve the chicken-and-egg problem

Zero Initial conditions

• set a start time (usually $n_0 = 0$)

• assume input and output are zero for all time before n_0

Ex: A simple model for banking

A simple equation to describe compound interest:

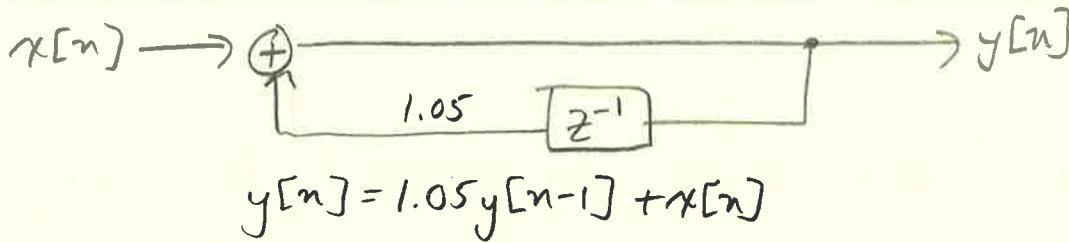
- constant interest/borrowing rate of 5% per year

- interest accrues on Dec 31

- deposits/withdrawals during year n : $x[n]$

- balance at year n :

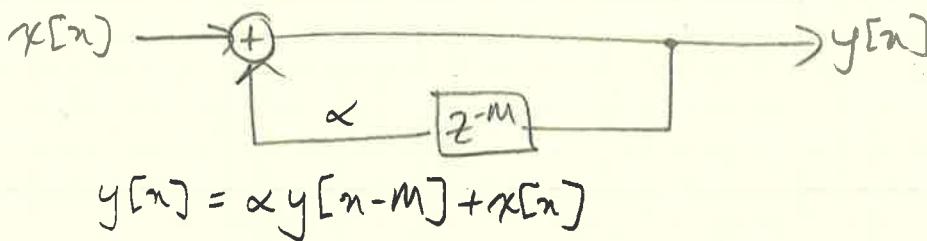
$$y[n] = 1.05y[n-1] + x[n]$$



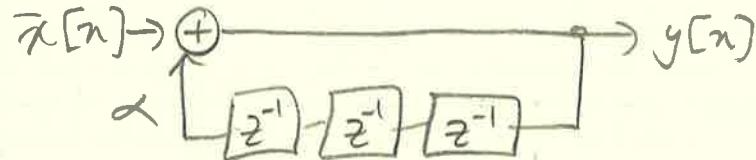
Ex: One-time investment $x[n] = 100\delta[n]$

- $y[0] = 100$
- $y[1] = 105$
- $y[2] = 110.25, y[3] = 115.7625$, etc.
- In general: $y[n] = (1.05)^n \cdot 100 u[n]$

An interesting generalization



• Creating loops



Ex: $M=3, \alpha=0.7, x[n] = \delta[n]$

- $y[0] = 1, y[1] = 0, y[2] = 0$
- $y[3] = 0.7, y[4] = 0, y[5] = 0$
- $y[6] = 0.7^2, y[7] = 0, y[8] = 0$, etc.

Ex: $M=3, \alpha=1, x[n] = \delta[n] + 2\delta[n-1] + 3\delta[n-2]$

- $y[0] = 1, y[1] = 2, y[2] = 3$
- $y[3] = 1, y[4] = 2, y[5] = 3$
- $y[6] = 1, y[7] = 2, y[8] = 3$, etc.

(We can make music with that!)

- build a recursion loop with a delay of M
- choose a signal $\bar{x}[n]$ that is nonzero only for $0 \leq n < M$
- choose a decay factor
- input $\bar{x}[n]$ to the system
- play the output

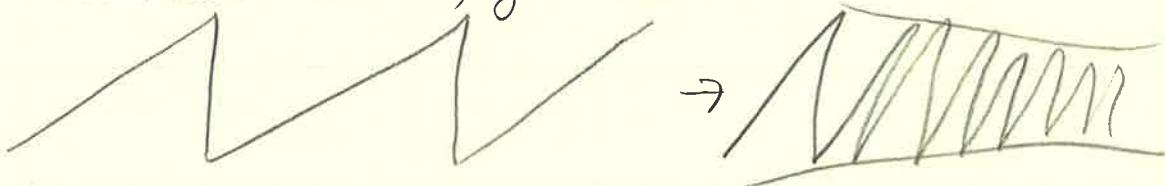
Ex: $M=100$, $\alpha=1$, $\bar{x}[n] = \sin(2\pi n/100)$ for $0 \leq n < 100$ and zero elsewhere

$$F_S = 48 \text{ kHz} \rightarrow 480 \text{ Hz}$$

Introducing some realism

- M controls frequency (pitch)
- α controls envelope (decay)
- $\bar{x}[n]$ controls color (timbre)

Proto-violin: $M=100$, $\alpha=0.95$, $\bar{x}[n]$: zero-mean sawtooth wave between 0 and 99, zero elsewhere



The Karplus - Strong Algorithm

$M=100$, $\alpha=0.9$, $\bar{x}[n]$: 100 random values between 0 and 99, zero elsewhere $\stackrel{\text{in } [-1, 1]}{\sim}$

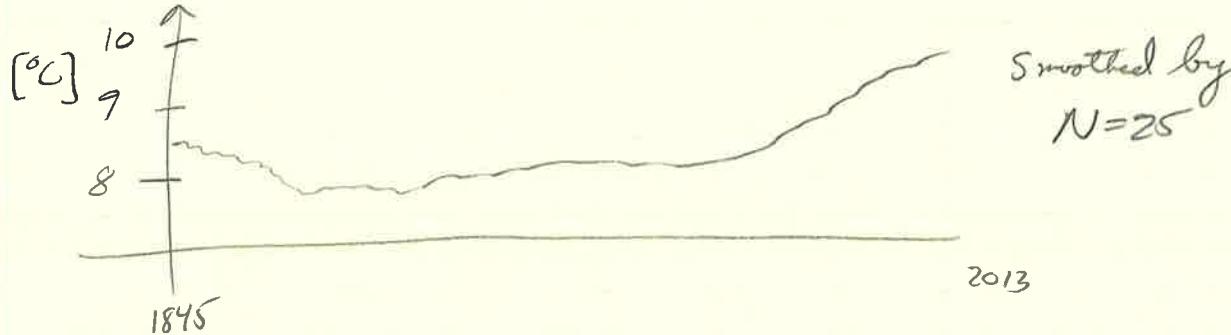
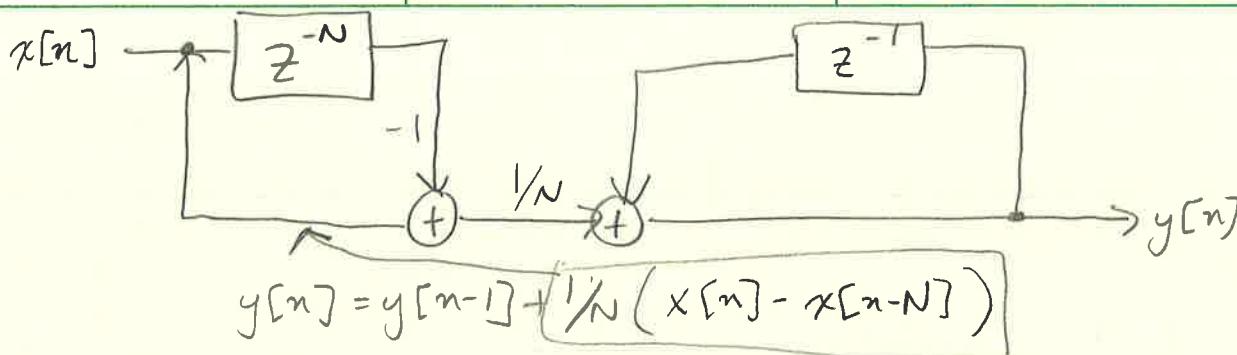
Similar to a harpsichord.

Signal of the Day: Goethe's Temperature Measurement

Smoothing { Moving average : $y[n] = \frac{1}{N} \sum_{m=0}^{N-1} x[n-m]$
 N : window of last observations over which the average is computed

A recursive method

$$\begin{aligned} y[n] &= \frac{1}{N} \sum_{m=0}^{N-1} x[n-m] \\ &= \frac{1}{N} x[n] + \frac{1}{N} \underbrace{\sum_{m=1}^{N-1} x[n-m]}_{y[n-1]} + \frac{1}{N} x[n-N] - \frac{1}{N} x[n-N] \\ &= y[n-1] + \frac{1}{N} (x[n] - x[n-N]) \end{aligned}$$

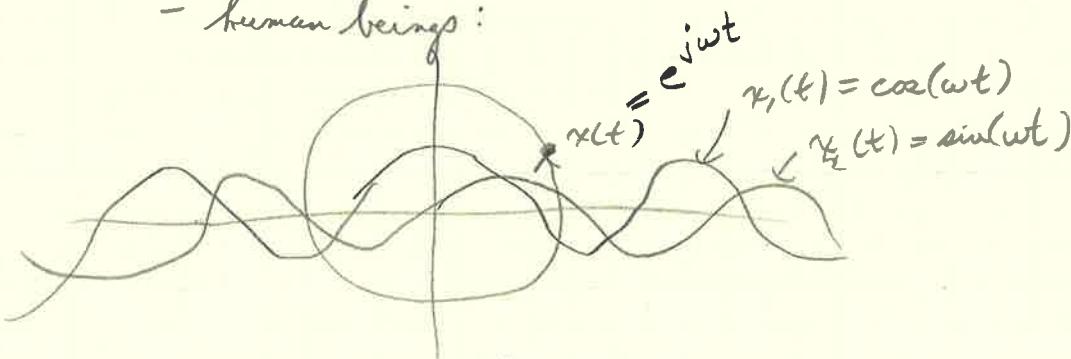


1.4 Complex exponentials

$$j = \sqrt{-1}$$

Oscillations are everywhere!

- Sustainable dynamic systems exhibit oscillatory behavior
- Intuitively: things that don't move in circles can't last:
 - bowls
 - rockets
 - human beings



- The discrete-time oscillatory heartbeat
Ingredients:

- a frequency ω (units: radians)
- an initial phase ϕ (units: radians)
- an amplitude A

$$x[n] = A e^{j(\omega n + \phi)}$$

$$= A [\cos(\omega n + \phi) + j \sin(\omega n + \phi)]$$

Why complex exponentials?

- we can use complex numbers in digital systems, so why not?
- it makes sense: every sinusoid can always be written as a sum of sine and cosine
- math is simpler: trigonometry becomes algebra

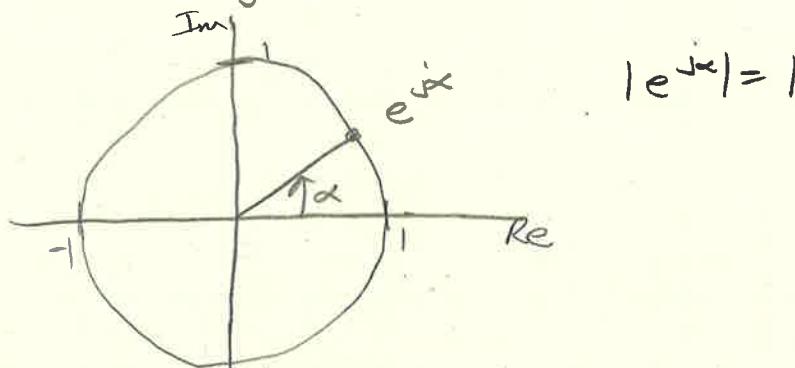
Ex: change the phase of a cosine the "old-school" way

$$\cos(\omega n + \phi) = a \cos(\omega n) - b \sin(\omega n), \quad a = \cos \phi, \quad b = \sin \phi$$

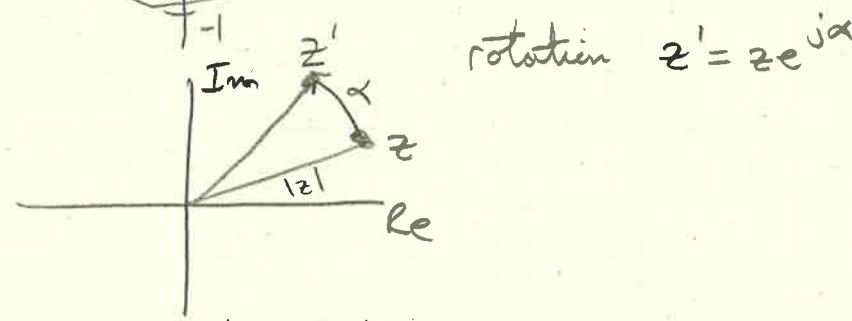
use complex exponentials

$$\cos(\omega n + \phi) = \operatorname{Re}[e^{j(\omega n + \phi)}] = \operatorname{Re}[e^{j\omega n} e^{j\phi}]$$

$$e^{j\alpha} = \cos \alpha + j \sin \alpha$$



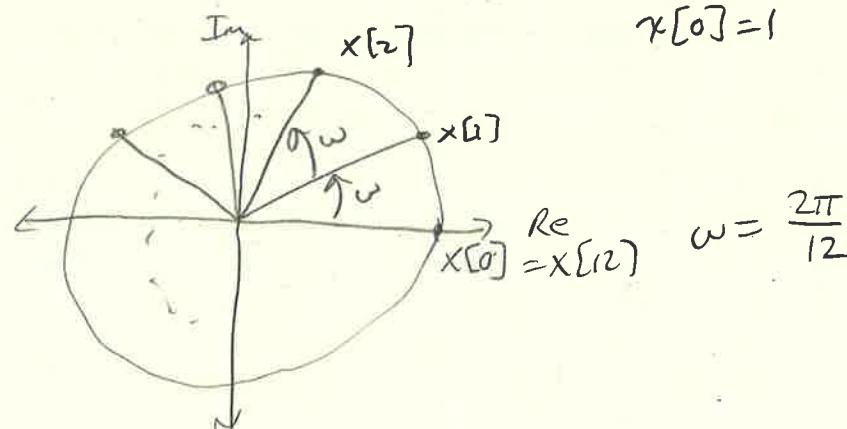
$$|e^{j\alpha}| = 1$$



$$\text{rotation } z' = z e^{j\alpha}$$

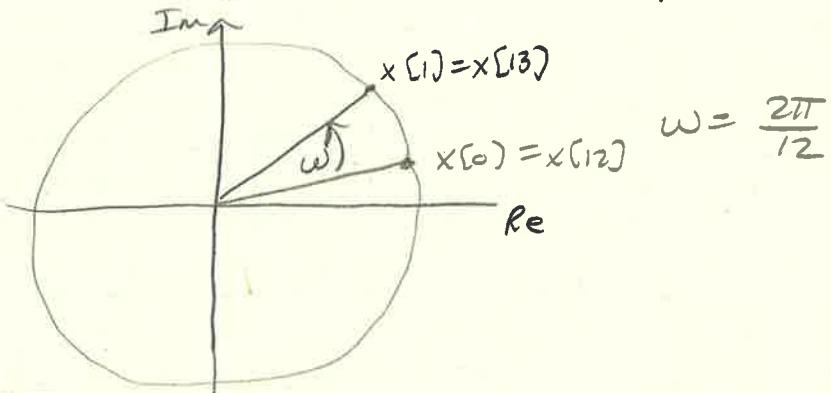
The complex exponential generating machine

$$x[n] = e^{j\omega n}; \quad x[n+1] = e^{j\omega} x[n]$$



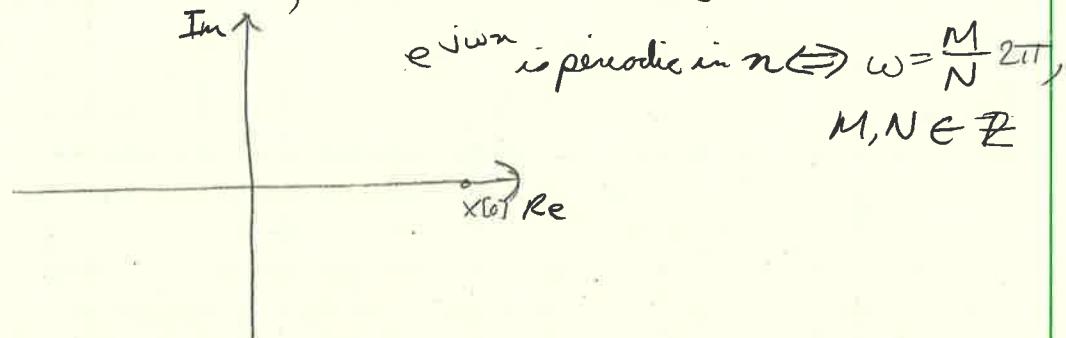
$$x[0] = 1$$

Initial phase
 $x[n] = e^{j(\omega n + \phi)}$; $x[n+1] = e^{j\omega} x[n]$, $x[0] = e^{j\phi}$



Careful: not every sinusoid is periodic in discrete time

$$x[n] = e^{j\omega n}; x[n+1] = e^{j\omega} x[n]$$



$$x[n] = x[n+N]$$

$$e^{j(\omega n + \phi)} = e^{j(\omega(n+N) + \phi)}$$

$$e^{j\omega n} e^{j\phi} = e^{j\omega n} e^{j\omega N} e^{j\phi}$$

$$e^{j\omega N} = 1 \Leftrightarrow \omega N = 2M\pi, M \in \mathbb{Z}$$

$$\omega = \frac{M}{N} 2\pi$$

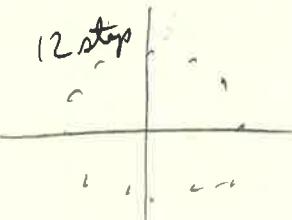
2π-periodicity: one point, many names

$$e^{j\alpha} = e^{j(\alpha + 2\pi k)}, \forall k \in \mathbb{Z}$$

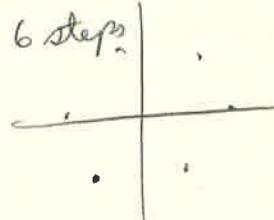
One point, many names: Aliasing

How "fast" can we go?

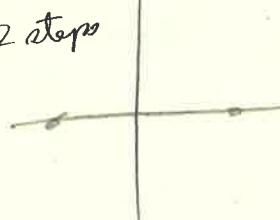
$$\omega = \frac{2\pi}{12}$$



$$\omega = 2\pi/6$$



$$\omega = 2\pi/2$$

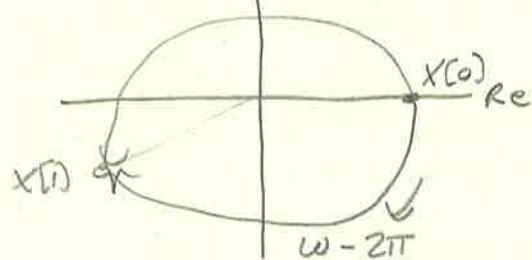


What if we go faster?

$$\pi < \omega < 2\pi$$

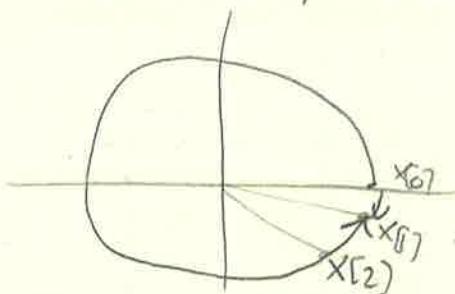
Im

corresponds to going slower
in opposite direction



$$\omega = 2\pi - \alpha, \alpha \text{ small}$$

very slow in opposite
direction



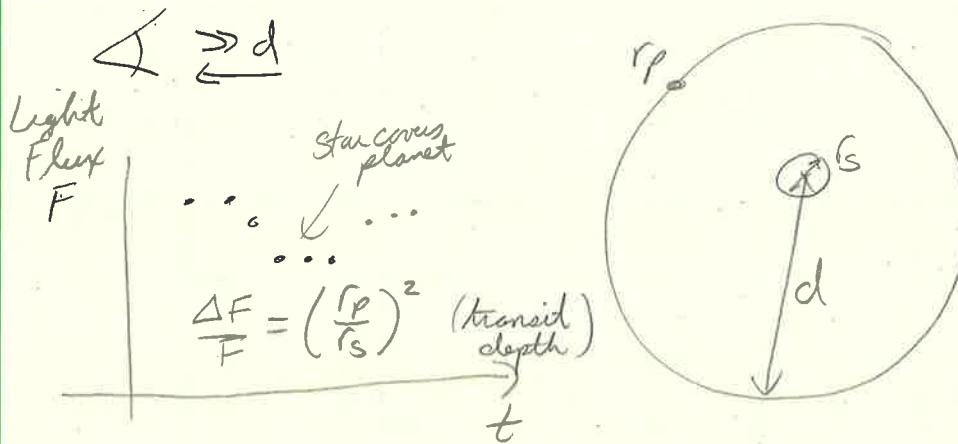
2.1 Signal processing and vector spaces

2.1

Common framework: vector space

- vector spaces are very general objects
- vector spaces are defined by their properties
- once you know the properties are satisfied, you can use all the tools for the space

Signal of the day: exoplanet hunting



• Earth: $\frac{\Delta F}{F} = \left(\frac{r_p}{r_s}\right)^2 = \left(\frac{6,371}{696,000}\right)^2 \approx 0.01\%$

• Jupiter: $\frac{\Delta F}{F} = \left(\frac{69,911}{696,000}\right)^2 \approx 1\%$

• Best telescope today can detect a transit depth of 0.1%.



2.2 Vector Spaces

2.2a Vector space

Some familiar examples

• $\mathbb{R}^2, \mathbb{R}^3$: Euclidean space

• $\mathbb{R}^N, \mathbb{C}^N$: linear algebra

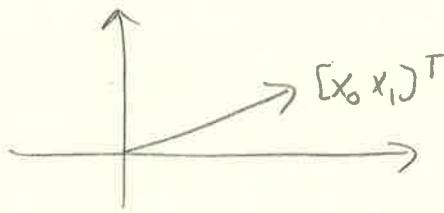
Other examples:

• $\ell_2(\mathbb{Z})$: space of square-summable infinite sequences

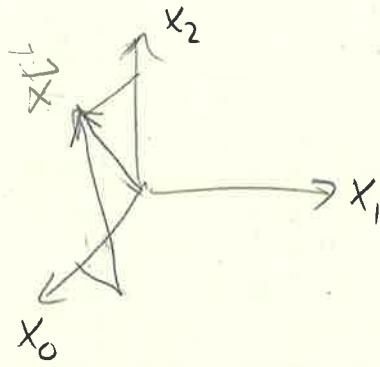
• $L_2([a,b])$: space of square-integrable functions over an interval

Some can be represented geometrically

$$\mathbb{R}^2: \vec{x} = [x_0 \ x_1]^T$$

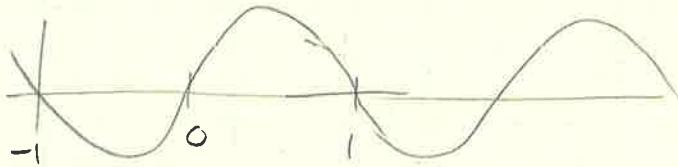


$$\mathbb{R}^3: \vec{x} = [x_0 \ x_1 \ x_2]^T$$



$$L_2([-1, 1]): \vec{x} = x(t), t \in [-1, 1]$$

$$\vec{x} = \sin(\pi t)$$



Can't plot $\mathbb{R}^N, N > 3$ or $\mathbb{C}^N, N > 1$

Ingredients

- the set of vectors V
- a set of scalars (say \mathbb{C})

We need at least to be able to:

- resize vectors, i.e., multiply a vector by a scalar
- combine vectors together, i.e., sum them

Formal properties: For $\vec{x}, \vec{y}, \vec{z} \in V$ and $\alpha, \beta \in \mathbb{C}$:

$$\vec{x} + \vec{y} = \vec{y} + \vec{x}$$

$$(\vec{x} + \vec{y}) + \vec{z} = \vec{x} + (\vec{y} + \vec{z})$$

$$\alpha(\vec{x} + \vec{y}) = \alpha\vec{y} + \alpha\vec{x}$$

$$(\alpha + \beta)\vec{x} = \alpha\vec{x} + \beta\vec{x}$$

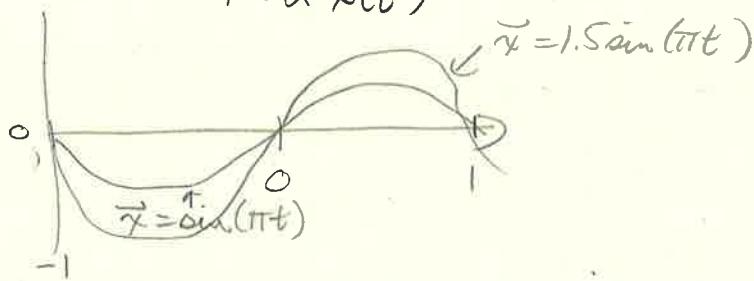
$$\alpha(\beta\vec{x}) = (\alpha\beta)\vec{x}$$

$$\exists 0 \in V : \vec{x} + 0 = 0 + \vec{x} = \vec{x}$$

$$\forall \vec{x} \in V, \exists (-\vec{x}) : \vec{x} + (-\vec{x}) = 0$$

Scalar multiplication in $L_2[-1, 1]$

$$\alpha \vec{x} = \alpha x(t)$$



We need something more: inner product (aka dot product)

$$\langle \cdot, \cdot \rangle: V \times V \rightarrow \mathbb{C}$$

- measure of similarity between vectors

- inner product is zero? vectors are orthogonal (maximally different)

Formal properties of the inner product

For $\vec{x}, \vec{y}, \vec{z} \in V, \alpha \in \mathbb{C}$:

- $\langle \vec{x} + \vec{y}, \vec{z} \rangle = \langle \vec{x}, \vec{z} \rangle + \langle \vec{y}, \vec{z} \rangle$

- $\langle \vec{x}, \vec{y} \rangle = \langle \vec{y}, \vec{x} \rangle^*$

- $\langle \alpha \vec{x}, \vec{y} \rangle = \alpha^* \langle \vec{x}, \vec{y} \rangle$

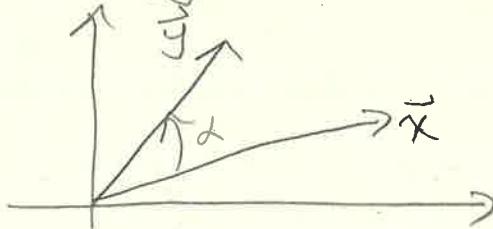
- $\langle \vec{x}, \alpha \vec{y} \rangle = \alpha \langle \vec{x}, \vec{y} \rangle$

- $\langle \vec{x}, \vec{x} \rangle \geq 0$

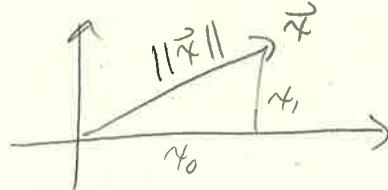
- $\langle \vec{x}, \vec{x} \rangle = 0 \Leftrightarrow \vec{x} = 0$

- If $\langle \vec{x}, \vec{y} \rangle = 0$ and $\vec{x}, \vec{y} \neq 0$, then \vec{x} and \vec{y} are called orthogonal

$$\langle \vec{x}, \vec{y} \rangle = x_0 y_0 + x_1 y_1 = \|\vec{x}\| \|\vec{y}\| \cos \varphi$$



$$\langle \vec{x}, \vec{x} \rangle = x_0^2 + x_1^2 = \|\vec{x}\|^2$$

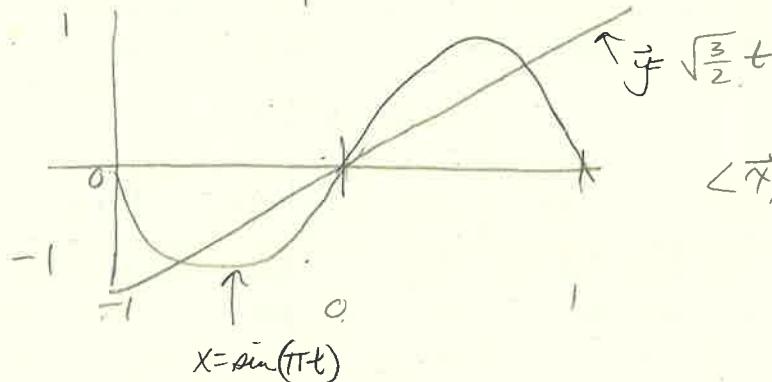


Inner product in $L_2[-1, 1]$

$$\langle \vec{x}, \vec{y} \rangle = \int_{-1}^1 x(t) y(t) dt$$

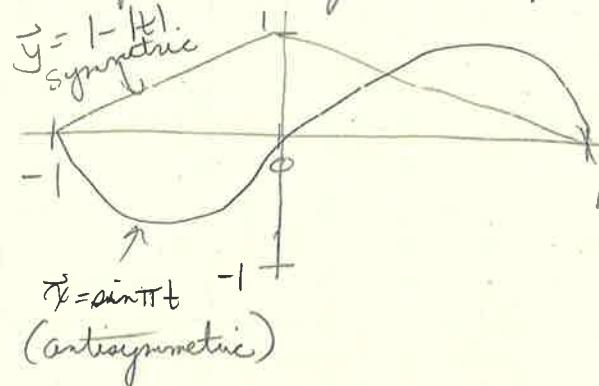
$$\| \sin(\pi t) \| = \sqrt{\int_{-1}^1 \sin^2 \pi t dt} = 1$$

$$\vec{y} = t: \| \vec{y} \| = \sqrt{\int_{-1}^1 t^2 dt} = \frac{2}{3}$$



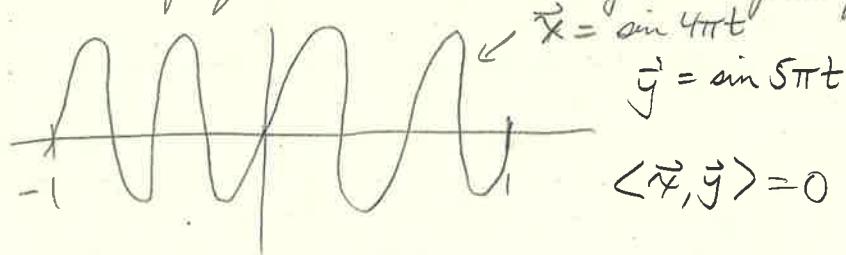
$$\begin{aligned}\langle \vec{x}, \vec{y} \rangle &= \int_{-1}^1 \sqrt{\frac{3}{2}} t \sin \pi t dt \\ &= \frac{2}{\pi} \sqrt{\frac{3}{2}} \approx 0.78\end{aligned}$$

\vec{x}, \vec{y} from orthogonal subspaces:



$$\langle \vec{x}, \vec{y} \rangle = 0$$

Sinusoids with frequencies that are integer multiples of a fundamental



$$\langle \vec{x}, \vec{y} \rangle = 0$$

Norm vs Distance

- inner product defines a norm: $\| \vec{x} \| = \sqrt{\langle \vec{x}, \vec{x} \rangle}$

- norm defines a distance: $d(\vec{x}, \vec{y}) = \| \vec{x} - \vec{y} \|$

Distance in $L_2[-1, 1]$: the Mean Square Error

$$\| \vec{x} - \vec{y} \|^2 = \int_{-1}^1 |x(t) - y(t)|^2 dt$$

$$\vec{x} = \sin 4\pi t, \quad \vec{y} = \sin 5\pi t, \quad \| \vec{x} - \vec{y} \|^2 = \int_{-1}^1 |\sin 4\pi t - \sin 5\pi t|^2 dt = 2$$

2.2.b Signal Spaces

Finite-Length Signals

finite-length and periodic signals live in \mathbb{C}^N

- vector notation: $\vec{x} = [x_0 \ x_1 \ \dots \ x_{N-1}]^T$

- all operations well-defined and intuitive

- space of N -periodic signals sometimes indicated by $\widetilde{\mathbb{C}}^N$

Inner product for signals

$$\langle \vec{x}, \vec{y} \rangle = \sum_{n=0}^{N-1} x^*[n] y[n]$$

well-defined for all finite-length vectors

Infinite Signals? $\langle \vec{x}, \vec{y} \rangle = \sum_{n=-\infty}^{\infty} x^*[n] y[n]$

We require sequences to be square-summable: $\sum |x[n]|^2 < \infty$
i.e. in $\ell_2(\mathbb{Z})$ (finite-energy)

Many interesting signals are not in $\ell_2(\mathbb{Z})$, such as,

$$x[n] = 1, \quad x[n] = \cos(\omega n), \text{ etc.}$$

Completeness

Limiting operations must yield vector space elements

An incomplete space: \mathbb{Q} $x_n = \sum_{k=0}^n \frac{1}{k!} \in \mathbb{Q},$

but $\lim_{n \rightarrow \infty} x_n = e \notin \mathbb{Q}$

Hilbert Space

1. A vector space: $H(V, \mathbb{C})$

2. An inner product: $\langle \cdot, \cdot \rangle: V \times V \rightarrow \mathbb{C}$

3. Complete

Bases

Linear combination is the basic operation in vector spaces:

$$\vec{g} = \alpha \vec{x} + \beta \vec{y}$$

Can we find a set of vectors $\{\vec{w}^{(k)}\}$ so that we can write any vector as a linear combination of the $\{\vec{w}^{(k)}\}$?

Canonical \mathbb{R}^2 basis

$$\vec{e}^{(0)} = \begin{bmatrix} 1 \\ 0 \end{bmatrix}, \vec{e}^{(1)} = \begin{bmatrix} 0 \\ 1 \end{bmatrix} \Rightarrow \begin{bmatrix} x_0 \\ x_1 \end{bmatrix} = x_0 \begin{bmatrix} 1 \\ 0 \end{bmatrix} + x_1 \begin{bmatrix} 0 \\ 1 \end{bmatrix}$$

Another \mathbb{R}^2 basis

$$\vec{v}^{(0)} = \begin{bmatrix} 1 \\ 0 \end{bmatrix}, \vec{v}^{(1)} = \begin{bmatrix} 1 \\ 1 \end{bmatrix}$$

$$\begin{bmatrix} x_0 \\ x_1 \end{bmatrix} = \alpha_0 \begin{bmatrix} 1 \\ 0 \end{bmatrix} + \alpha_1 \begin{bmatrix} 1 \\ 1 \end{bmatrix} \Rightarrow \alpha_0 = x_0 - x_1, \alpha_1 = x_1$$

Not a basis for \mathbb{R}^2

$$\vec{g}^{(0)} = \begin{bmatrix} 1 \\ 0 \end{bmatrix}, \vec{g}^{(1)} = \begin{bmatrix} -1 \\ 0 \end{bmatrix} \quad \text{not linearly independent}$$

What about infinite-dimensional spaces?

$$\vec{x} = \sum_{k=0}^{\infty} \alpha_k \vec{w}^{(k)}$$

a basis for $l_2(\mathbb{R})$

$$\vec{e}^{(k)} = \begin{bmatrix} 0 \\ 0 \\ \vdots \\ 0 \\ 1 \\ 0 \\ \vdots \\ 0 \end{bmatrix}, \quad 1 \text{ in } k^{\text{th}} \text{ position, } k \in \mathbb{Z}$$

What about function vector spaces?

$$f(t) = \sum_k \alpha_k h^{(k)}(t)$$

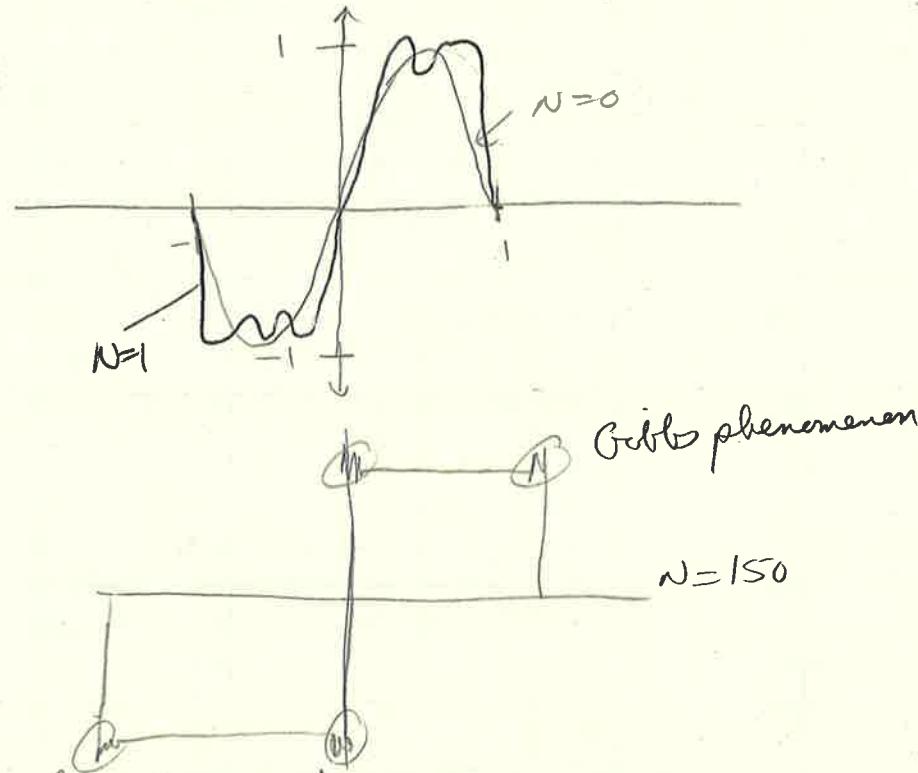
A basis for the functions over an interval?

the Fourier basis for $[-1, 1]$

$$\left\{ \frac{1}{\sqrt{2}}, \cos \pi t, \sin \pi t, \cos 2\pi t, \sin 2\pi t, \cos 3\pi t, \sin 3\pi t, \dots \right\}$$

Using the Fourier Basis (approximating a square wave)

$$\sum_{k=0}^N \frac{\sin((2k+1)\pi t)}{2k+1} = \sum_{k=0}^N \frac{w^{(4k+2)}}{2k+1}$$



Bases: formal definition

Given:

- a vector space H
- a set of K vectors from H : $W = \{\vec{w}^{(k)}\}_{k=0,1,\dots,K-1}$

W is a basis for H if:

1. We can write for all $x \in H$:

$$\vec{x} = \sum_{k=0}^{K-1} \alpha_k \vec{w}^{(k)}, \quad \alpha_k \in \mathbb{C}$$

2. the coefficients α_k are unique

Uniqueness implies linear independence

$$\sum_{k=0}^{K-1} \alpha_k \vec{w}^{(k)} = 0 \Rightarrow \alpha_k = 0, \quad k=0,1,\dots,K-1$$

Special bases

Orthogonal basis:
 $\langle \vec{w}^{(k)}, \vec{w}^{(n)} \rangle = 0, \quad k \neq n$

Orthonormal bases: $\langle \vec{w}^{(k)}, \vec{w}^{(n)} \rangle = \delta_{[n-k]}$

We can use Gram-Schmidt to normalize any orthogonal basis

Basis expansion

$$\vec{x} = \sum_{k=0}^{K-1} \alpha_k \vec{w}^{(k)}, \text{ how do we find the } \alpha's?$$

Orthonormal bases are the best: $\alpha_k = \langle \vec{w}^{(k)}, \vec{x} \rangle$

Change of basis

$$\vec{x} = \sum_{k=0}^{K-1} \alpha_k \vec{w}^{(k)} = \sum_{k=0}^{K-1} \beta_k \vec{v}^{(k)}$$

If $\{\vec{v}^{(k)}\}$ is orthonormal:

$$\begin{aligned} \beta_k &= \langle \vec{v}^{(k)}, \vec{x} \rangle \\ &= \left\langle \vec{v}^{(k)}, \sum_{h=0}^{K-1} \alpha_h \vec{w}^{(h)} \right\rangle = \sum_{h=0}^{K-1} \alpha_h \langle \vec{v}^{(k)}, \vec{w}^{(h)} \rangle \end{aligned}$$

$$= \sum_{h=0}^{K-1} \alpha_h C_{hk}$$

$$= \begin{bmatrix} c_{00} & c_{01} & \cdots & c_{0(K-1)} \\ & \vdots & & \\ c_{(K-1)0} & c_{(K-1)1} & \cdots & c_{(K-1)(K-1)} \end{bmatrix} \begin{bmatrix} \alpha_0 \\ \vdots \\ \alpha_{K-1} \end{bmatrix}$$

Change of basis: example

- canonical basis $E = \{\vec{e}^{(0)}, \vec{e}^{(1)}\}$

$$\vec{x} = \alpha_0 \vec{e}^{(0)} + \alpha_1 \vec{e}^{(1)}$$

- new basis $V = \{\vec{v}^{(0)}, \vec{v}^{(1)}\}$ with $\vec{v}^{(0)} = [\cos \theta \sin \theta]^T$
 $\vec{v}^{(1)} = [-\sin \theta \cos \theta]^T$

$$\vec{x} = \beta_0 \vec{v}^{(0)} + \beta_1 \vec{v}^{(1)}$$

- new basis is orthonormal: $C_{hk} = \langle \vec{v}^{(h)}, \vec{e}^{(k)} \rangle$

- in compact form:

$$\begin{bmatrix} \beta_0 \\ \beta_1 \end{bmatrix} = \begin{bmatrix} \cos \theta & \sin \theta \\ -\sin \theta & \cos \theta \end{bmatrix} \begin{bmatrix} \alpha_0 \\ \alpha_1 \end{bmatrix} = R \alpha$$

- R : rotation matrix

$$R^T R = I$$

Vector Subspace

- A subset of vectors closed under addition and scalar multiplication.
- Example: $\mathbb{R}^2 \subset \mathbb{R}^3$
- Subspace of symmetric functions over $L_2[-1, 1]$
 $\vec{x} = \cos \pi t$
 $\vec{y} = \cos 5\pi t$ to name a couple

- Subspaces have their own bases

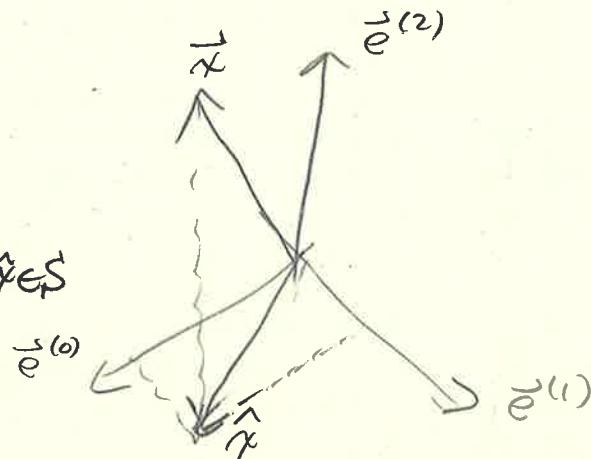
$$\left\{ \vec{e}^{(0)} = \begin{bmatrix} 1 \\ 0 \\ 0 \end{bmatrix}, \vec{e}^{(1)} = \begin{bmatrix} 0 \\ 1 \\ 0 \end{bmatrix} \right\} \text{ basis for a plane}$$

Approximation

Problem:

- vector $x \in V$
- subspace $S \subseteq V$

- approximate \vec{x} with $\vec{x} \in S$

Least-Squares Approximation

- $\{\vec{s}^{(k)}\}_{k=0, \dots, K-1}$ orthonormal basis for S

- orthogonal projection:

$$\hat{x} = \sum_{k=0}^{K-1} \langle \vec{s}^{(k)}, \vec{x} \rangle \vec{s}^{(k)}$$

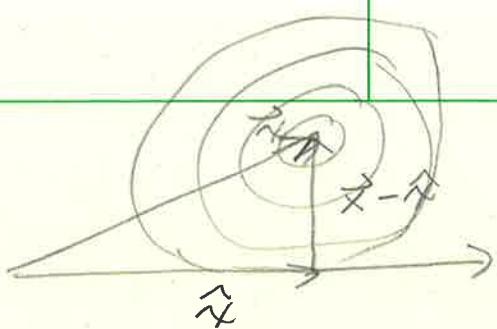
- orthogonal projection is the "best" approximation over S

- orthogonal projection has minimum-norm error:

$$\underset{y \in S}{\operatorname{argmin}} \| \vec{x} - \vec{y} \| = \hat{x}$$

- error is orthogonal to approximation:

$$\langle \vec{x} - \hat{x}, \hat{x} \rangle = 0$$



draw concentric circles
until hitting \vec{s} . This radius
vector is $\vec{x} - \vec{s}$.

Example: polynomial approximation

- vector space $P_N [-1, 1] \subset L_2 [-1, 1]$
- $\vec{p} = a_0 + a_1 t + \dots + a_{N-1} t^{N-1}$
- a self-evident, naive basis: $\vec{s}^{(k)} = t^k$, $k=0, 1, \dots, N-1$
- naive basis is not orthonormal

goal: approximate $\vec{x} = \sin t \in L_2 [-1, 1]$ over $P_3 [-1, 1]$

- build orthonormal basis from naive basis
- project \vec{x} over the orthonormal basis
- compute approximation error
- compare errors to Taylor approximation (well known but not optimal over the interval)

Building an orthonormal basis

Gram-Schmidt orthonormalization procedure:

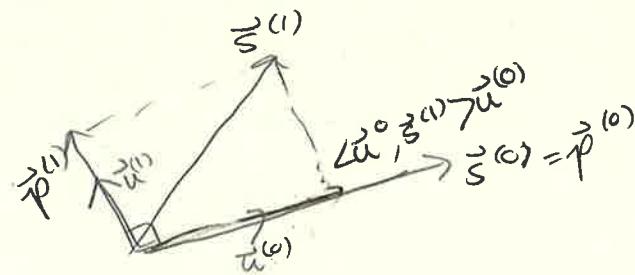
$$\left\{ \vec{s}^{(k)} \right\} \rightarrow \left\{ \vec{u}^{(k)} \right\}$$

original set orthonormal set

Algorithmic procedure: at each step k

$$1. \vec{p}^{(k)} = \vec{s}^{(k)} - \sum_{n=0}^{k-1} \langle \vec{u}^{(n)}, \vec{s}^{(k)} \rangle \vec{u}^{(n)}$$

$$2. \vec{u}^{(k)} = \vec{p}^{(k)} / \| \vec{p}^{(k)} \|$$



Apply Gram-Schmidt to $S = \{1, t, t^2, t^3, \dots\}$

$$\langle \vec{x}, \vec{y} \rangle = \int_{-1}^1 x(t) y(t) dt$$

$$\rightarrow \vec{s}^{(0)} = 1$$

$$\cdot \vec{p}^{(0)} = \vec{s}^{(0)} = 1$$

$$\cdot \| \vec{p}^{(0)} \|^2 = 2$$

$$\cdot \vec{u}^{(0)} = \vec{p}^{(0)} / \| \vec{p}^{(0)} \| = \frac{1}{\sqrt{2}}$$

$$\rightarrow \vec{s}^{(1)} = t$$

$$\cdot \langle \vec{u}^{(0)}, \vec{s}^{(1)} \rangle = \int_{-1}^1 \frac{t}{\sqrt{2}} dt = 0$$

$$\cdot \vec{p}^{(1)} = \vec{s}^{(1)} = t$$

$$\cdot \| \vec{p}^{(1)} \|^2 = \frac{2}{3}$$

$$\cdot \vec{u}^{(1)} = \sqrt{\frac{3}{2}} t$$

$$\rightarrow \vec{s}^{(2)} = t^2$$

$$\cdot \langle \vec{u}^{(0)}, \vec{s}^{(2)} \rangle = \int_{-1}^1 \frac{t^2}{\sqrt{2}} dt = \frac{2}{3\sqrt{2}}$$

$$\cdot \langle \vec{u}^{(1)}, \vec{s}^{(2)} \rangle = \int_{-1}^1 \frac{t^3}{\sqrt{2}} dt = 0$$

$$\cdot \vec{p}^{(2)} = \vec{s}^{(2)} - \frac{2}{3\sqrt{2}} \vec{u}^{(0)} = t^2 - \frac{1}{3}$$

$$\cdot \| \vec{p}^{(2)} \|^2 = \frac{8}{45}$$

$$\cdot \vec{u}^{(2)} = \sqrt{\frac{5}{8}} (3t^2 - 1)$$

Legendre Polynomials

The Gram-Schmidt algorithm leads to an orthonormal basis for $P_n([-1, 1])$

$$\vec{u}^{(0)} = \sqrt{\frac{1}{2}}, \vec{u}^{(1)} = \sqrt{\frac{3}{2}} t, \vec{u}^{(2)} = \sqrt{\frac{5}{8}} (3t^2 - 1), \vec{u}^{(3)} = \dots$$

Orthogonal projection over $P_3[-1, 1]$

$$\alpha_k = \langle \vec{u}^{(k)}, \vec{x} \rangle = \int_{-1}^1 u_k(t) \sin t dt$$

$$\cdot \alpha_0 = \langle \frac{1}{\sqrt{2}}, \sin t \rangle = 0$$

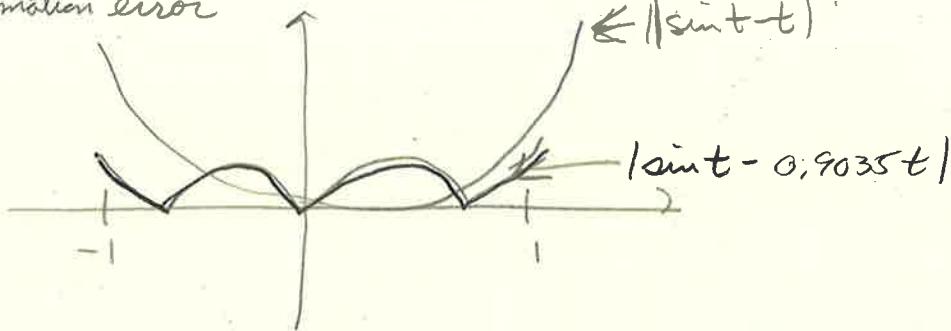
$$\cdot \alpha_1 = \langle \sqrt{\frac{3}{2}} t, \sin t \rangle \approx 0.7377$$

$$\cdot \alpha_2 = \langle \sqrt{\frac{5}{8}} (3t^2 - 1), \sin t \rangle = 0$$

$$\sin t \rightarrow \alpha_1 \vec{u}^{(1)} \approx 0.9035 t$$

Taylor Series: $\sin t \approx t$

Approximation error



Error norm:

Orthogonal projection over $P_3[-1, 1]$:

$$\| \sin t - \alpha_i \bar{u}^{(1)} \| \approx 0.0337$$

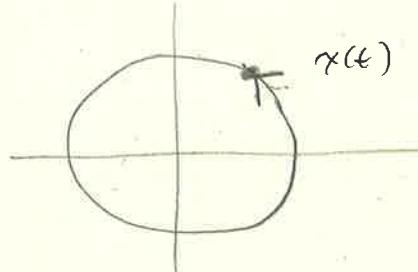
Taylor series: $\| \sin t - t \| \approx 0.0857$

3.1.a The frequency domain

- Oscillations are everywhere

- Sustainable dynamic systems exhibit oscillatory behavior
- Intuitively: things that don't move in circles don't last:
 - bombs
 - rockets
 - human beings...

Period P
Frequency $f = \frac{1}{P}$



- The intuition

- humans analyse complex signals (audio, images) in terms of their sinusoidal components
- We can build instruments that "resonate" at one or multiple frequencies (tuning fork vs. piano)
- the "frequency domain" seems to be as important as the time domain
- Fundamental question: can we decompose any signal into sinusoidal elements? Yes, using Fourier analysis

Analysis

- from time domain to frequency domain
- find the contribution of different frequencies
- discover "hidden" signal properties

Synthesis

- from frequency domain to time domain
- create signals with known frequency content
- fit signals to specific frequency regions

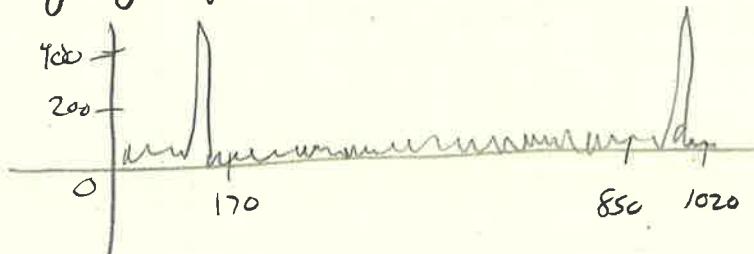
3.1.b The DFT as a change of basis

- The mathematical setup
 - let's start with finite-length signals (i.e. vectors in \mathbb{C}^N)
 - Fourier analysis is a simple change of basis
 - a change of basis is a change of perspective

- Mystery signal in time domain



- Mystery signal in the Fourier basis

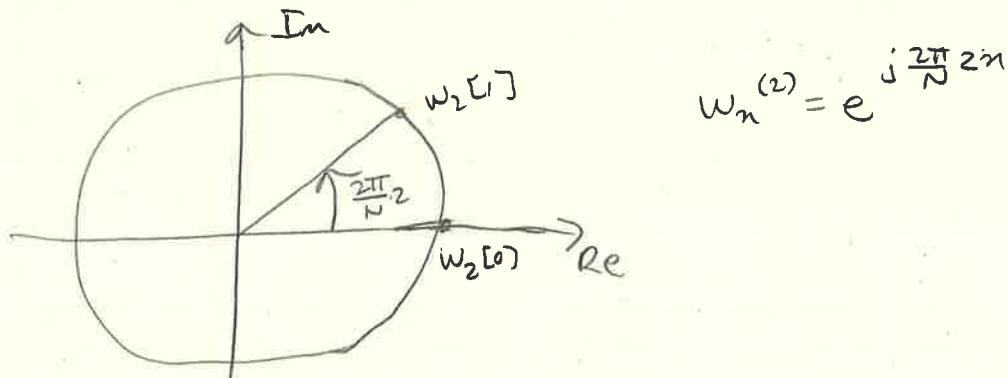
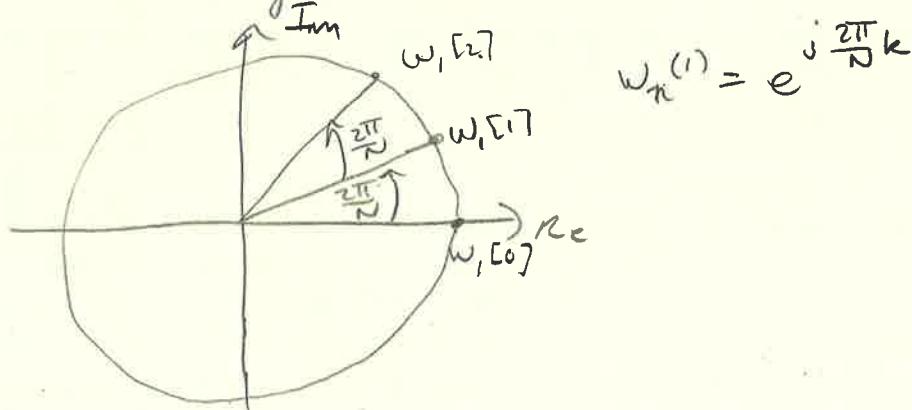


- The Fourier Basis for \mathbb{C}^N

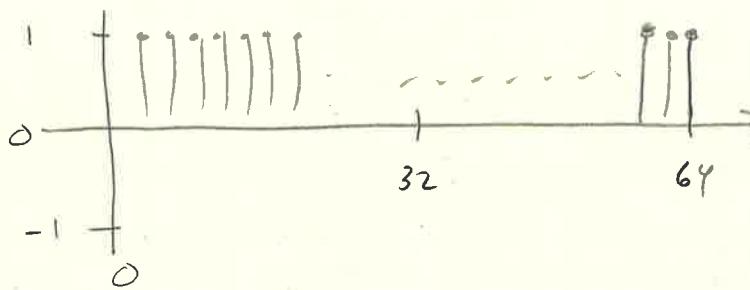
Claim: the set of N signals in \mathbb{C}^N

$w_n[n] = e^{j \frac{2\pi}{N} nk}$, $n, k = 0, 1, \dots, N-1$ is an orthogonal basis in \mathbb{C}^N . $\omega = \frac{2\pi}{N} k$

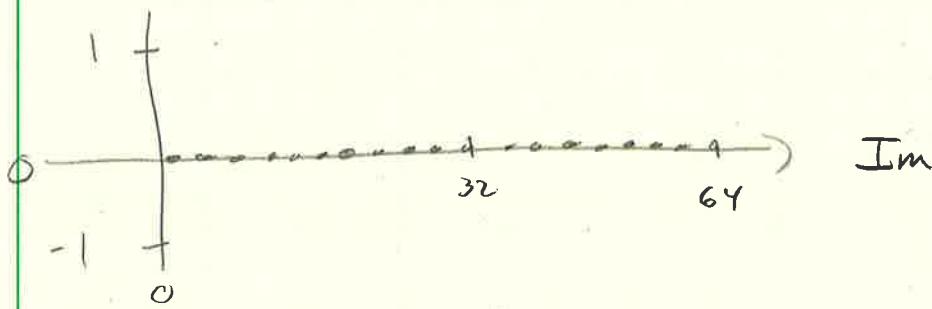
In vector notation: $\{\tilde{w}^{(k)}\}_{k=0,1,\dots,N-1}$ with $w_n^{(k)} = e^{j \frac{2\pi}{N} nk}$
is an orthogonal basis in \mathbb{C}^N .



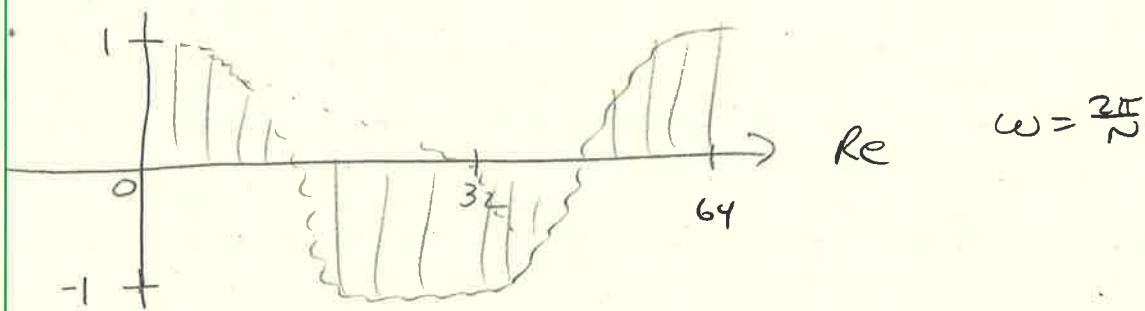
Basis vector $\vec{w}^{(0)} \in \mathbb{C}^{64}$



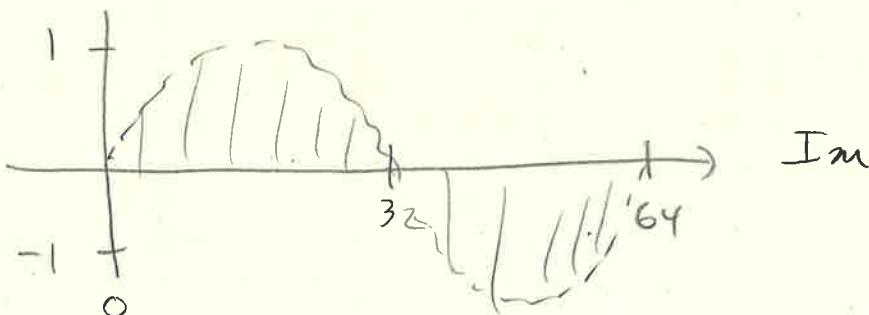
$$\vec{w}_n^{(0)} = e^{j \frac{2\pi}{N} 0n} = 1$$



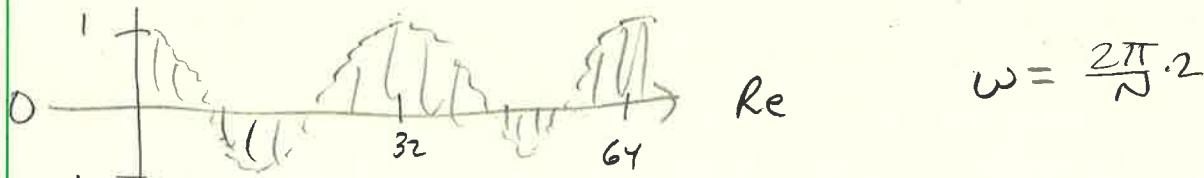
Basis vector $\vec{w}^{(1)} \in \mathbb{C}^{64}$, $\vec{w}_n^{(1)} = e^{j \frac{2\pi}{N} 1n}$



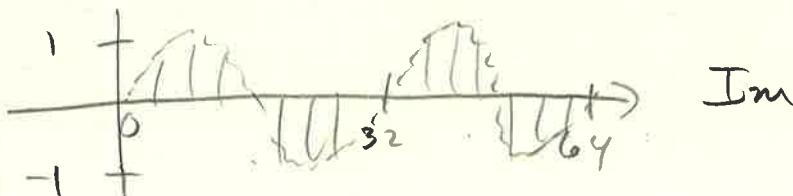
$$\omega = \frac{2\pi}{N}$$



Basis vector $\vec{w}^{(2)} \in \mathbb{C}^{64}$



$$\omega = \frac{2\pi}{N} \cdot 2$$



$$\vec{w}^{(3)} \in \mathbb{C}^{64}; \quad \omega = \frac{2\pi}{N} 3 = \frac{2\pi}{64} 3$$

⋮

$$\vec{w}^{(16)} \in \mathbb{C}^{64}; \quad \omega = \frac{2\pi}{64} 16 = \frac{\pi}{2}$$

⋮

$$\vec{w}^{(32)} \in \mathbb{C}^{64}; \quad \omega = \frac{2\pi}{64} 32 = \pi$$

⋮

$\vec{w}^{(64)} \in \mathbb{C}^{64}$ has same real part as $\vec{w}^{(2)}$ but the imaginary part is inverted

$\operatorname{Re}(\vec{w}^{(63)}) = \operatorname{Re}(\vec{w}^{(1)})$ but imaginary parts are inverted

- Proof of orthogonality

$$\langle \vec{w}^{(k)}, \vec{w}^{(n)} \rangle = \sum_{n=0}^{N-1} (e^{j \frac{2\pi}{N} nk})^* e^{j \frac{2\pi}{N} nh}$$

$$= \sum_{n=0}^{N-1} e^{j \frac{2\pi}{N} (n-k)n}$$

$$\left(\sum_{n=0}^{N-1} a^n = \frac{1-a^N}{1-a} \right) = \begin{cases} N & n=k \\ \frac{1-e^{j 2\pi(n-k)}}{1-e^{j \frac{2\pi}{N}(n-k)}} & \text{otherwise} \end{cases}$$

$$n-k \in \mathbb{N} \Rightarrow e^{j 2\pi(n-k)} = 1$$

- Remarks

- N orthogonal vectors \rightarrow basis for \mathbb{C}^N

- vectors are not orthonormal. Normalization factor would be \sqrt{N}

3.2 The Discrete Fourier Transform (DFT)

3.2a DFT definition

- Basis expansion

- Analysis formula: $X_k = \langle \vec{w}^{(k)}, \vec{x} \rangle$

- Synthesis formula: $\vec{x} = \frac{1}{N} \sum_{k=0}^{N-1} X_k \vec{w}^{(k)}$

- Change of basis in matrix form

Define $W_N = e^{-j\frac{2\pi}{N}}$ (or simply W when N is evident)

Change of basis matrix \underline{W} with $\underline{W}[n,m] = W_N^{nm}$

$$\underline{W} = \begin{bmatrix} 1 & 1 & 1 & 1 & \dots & 1 \\ 1 & W & W^2 & W^3 & \dots & W^{N-1} \\ 1 & W^2 & W^4 & W^6 & \dots & W^{2(N-1)} \\ \vdots & \vdots & \vdots & \vdots & \ddots & \vdots \\ 1 & W^{N-1} & W^{2(N-1)} & W^{3(N-1)} & \dots & W^{(N-1)^2} \end{bmatrix}$$

Analysis formula: $\underline{X} = \underline{W} \vec{x}$

Synthesis formula: $\vec{x} = \frac{1}{N} \underline{W}^H \underline{X}$

- Basis expansion (signal notation)

Analysis formula:

$$X[k] = \sum_{n=0}^{N-1} x[n] e^{-j\frac{2\pi}{N} nk}, \quad k=0, 1, \dots, N-1$$

N -point signal in the frequency domain

Synthesis formula:

$$x[n] = \frac{1}{N} \sum_{k=0}^{N-1} X[k] e^{j\frac{2\pi}{N} nk}, \quad n=0, 1, \dots, N-1$$

N -point signal in the "time" domain

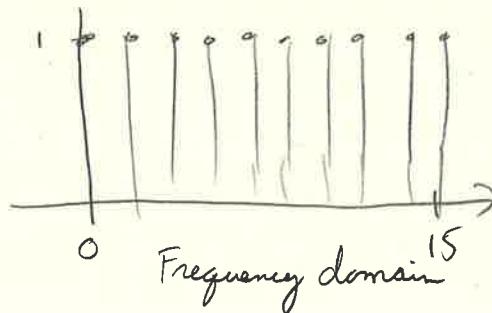
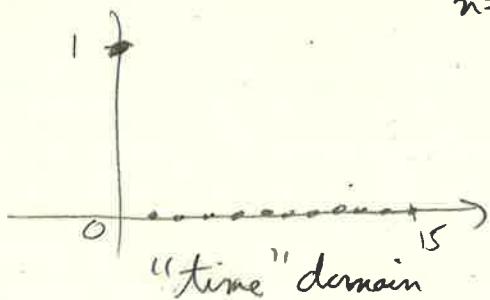
3.2 b Examples of DFT calculation

- DFT is obviously linear

$$\text{DFT}\{\alpha x[n] + \beta y[n]\} = \alpha \text{DFT}\{x[n]\} + \beta \text{DFT}\{y[n]\}$$

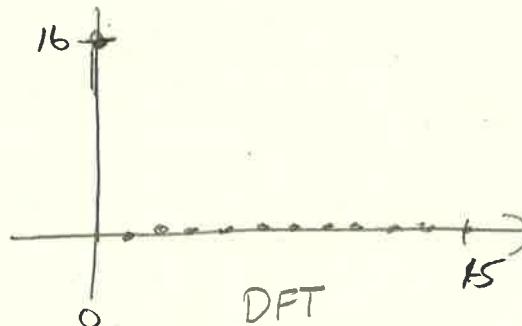
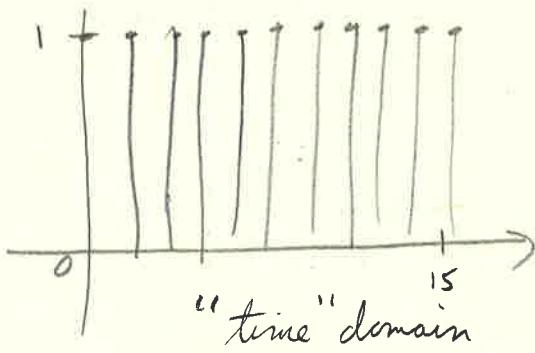
- DFT of $x[n] = \delta[n]$, $x[n] \in \mathbb{C}^N$

$$X[k] = \sum_{n=0}^{N-1} \delta[n] e^{-j\frac{2\pi}{N} nk} = 1$$



DFT of $x[n] = 1, x[n] \in \mathbb{C}^N$

$$X[k] = \sum_{n=0}^{N-1} e^{-j \frac{2\pi}{N} nk} = N \delta[k]$$



DFT of $x[n] = 3 \cos(\frac{2\pi}{16}n), x[n] \in \mathbb{C}^{64}$

$$x[n] = 3 \cos\left(\frac{2\pi}{16}n\right) = 3 \cos\left(\frac{2\pi}{64}4n\right) \quad \omega = \frac{2\pi}{64}$$

$$= \frac{3}{2} \left[e^{j \frac{2\pi}{64}4n} + e^{-j \frac{2\pi}{64}4n} \right]$$

$$= \frac{3}{2} \left[e^{j \frac{2\pi}{64}4n} + e^{j \frac{2\pi}{64}60n} \right] \quad -j \frac{2\pi}{64}4n = j \frac{2\pi}{64}60n$$

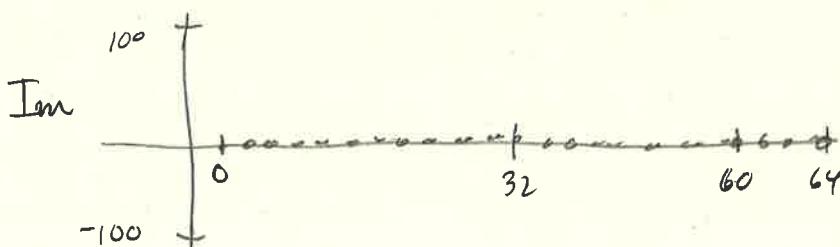
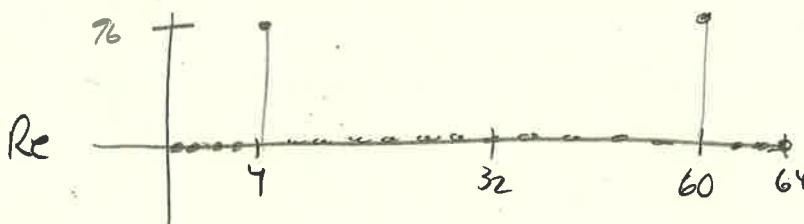
$$= \frac{3}{2} [w_4[n] + w_{60}[n]]$$

$$X[k] = \langle w_k[n], x[n] \rangle$$

$$= \langle w_k[n], \frac{3}{2}(w_4[n] + w_{60}[n]) \rangle$$

$$= \frac{3}{2} \langle w_k[n], w_4[n] \rangle + \frac{3}{2} \langle w_k[n], w_{60}[n] \rangle$$

$$= \begin{cases} \frac{3}{2} \cdot 64 = 96, & k = 4, 60 \\ 0, & \text{otherwise} \end{cases}$$



- DFT of $x[n] = 3 \cos\left(\frac{2\pi}{16}n + \pi/3\right)$, $x[n] \in \mathbb{C}^{64}$

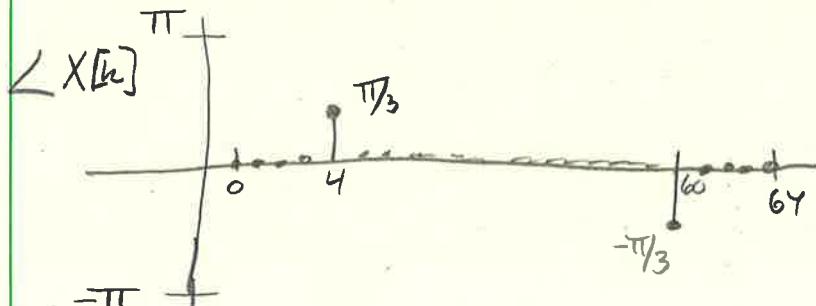
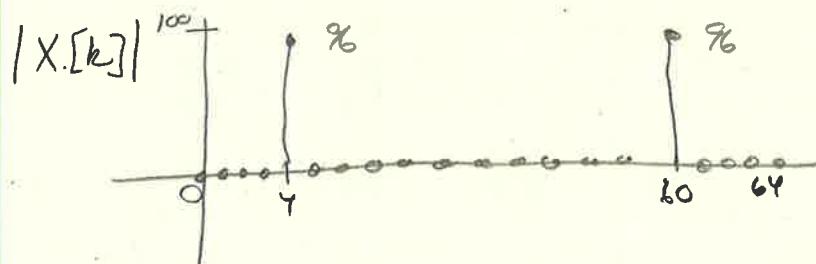
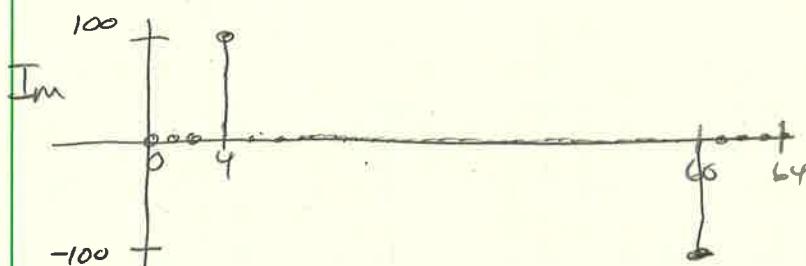
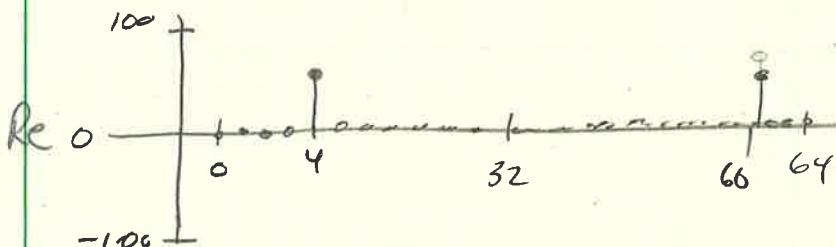
$$x[n] = 3 \cos\left(\frac{2\pi}{16}n + \frac{\pi}{3}\right)$$

$$= 3 \cos\left(\frac{2\pi}{64}4n + \frac{\pi}{3}\right)$$

$$= \frac{3}{2} \left[e^{j\frac{2\pi}{64}4n} e^{j\frac{\pi}{3}} + e^{-j\frac{2\pi}{64}4n} e^{-j\frac{\pi}{3}} \right]$$

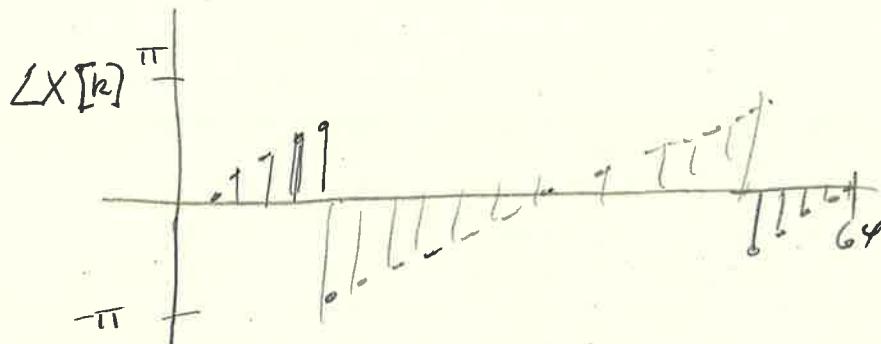
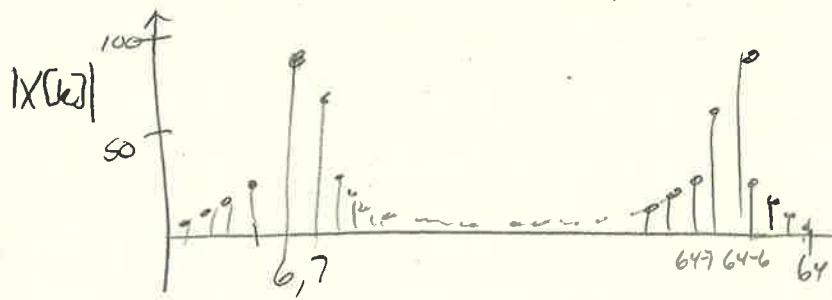
$$= \frac{3}{2} \left[e^{j\frac{\pi}{3}} w_4[n] + e^{-j\frac{\pi}{3}} w_{60}[n] \right]$$

$$X[k] = \langle w_n[n], x[n] \rangle = \begin{cases} 96e^{j\frac{\pi}{3}}, & k=4 \\ 96e^{-j\frac{\pi}{3}}, & k=6 \\ 0, & \text{otherwise} \end{cases}$$



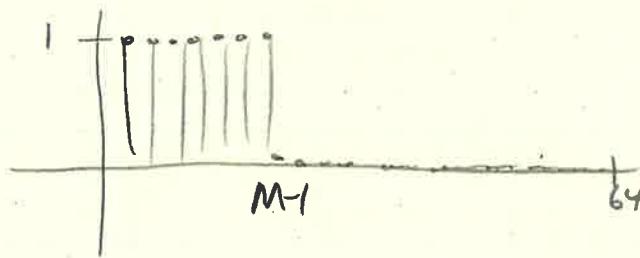
- DFT of $x[n] = 3 \cos\left(\frac{2\pi}{10}n\right)$, $x[n] \in \mathbb{C}^{64}$

$$\frac{2\pi}{64} 6 < \frac{2\pi}{10} < \frac{2\pi}{64} 7$$



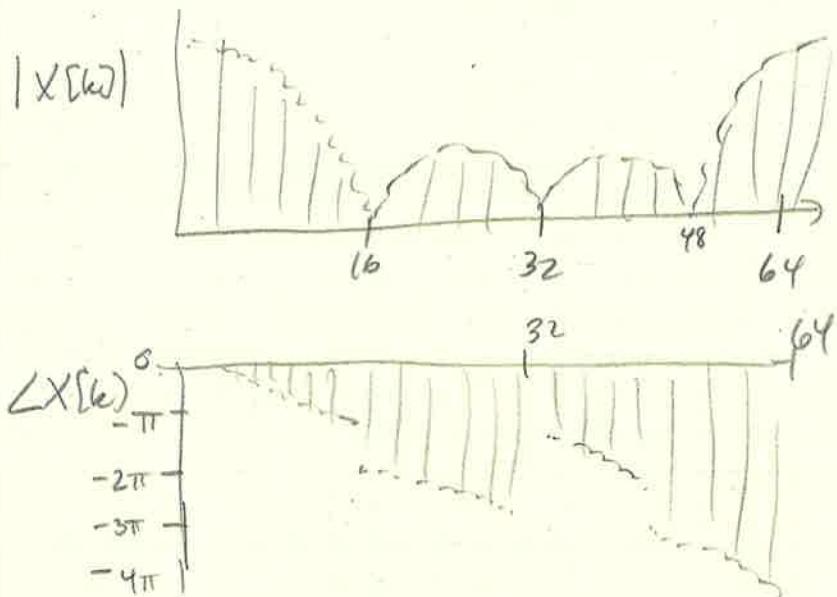
- DFT of length- M step in \mathbb{C}^N

$$x[n] = \sum_{h=0}^{M-1} \delta[n-h], \quad n=0, 1, \dots, N-1$$



$$\begin{aligned}
 X[k] &= \sum_{n=0}^{N-1} x[n] e^{-j \frac{2\pi}{N} nk} = \sum_{n=0}^{M-1} e^{-j \frac{2\pi}{N} nk} \\
 &= \frac{1 - e^{-j \frac{2\pi}{N} kM}}{1 - e^{-j \frac{2\pi}{N} k}} \quad \left(1 - e^{-j\alpha} = e^{-j\frac{\alpha}{2}} (e^{j\frac{\alpha}{2}} - e^{-j\frac{\alpha}{2}}) \right) \\
 &= \frac{e^{-j \frac{\pi}{N} kM} [e^{j \frac{\pi}{N} kM} - e^{-j \frac{\pi}{N} kM}]}{e^{-j \frac{\pi}{N} k} [e^{j \frac{\pi}{N} k} - e^{-j \frac{\pi}{N} k}]} \\
 &= \frac{\sin\left(\frac{\pi}{N} Mk\right)}{\sin\left(\frac{\pi}{N} k\right)} e^{-j \frac{\pi}{N} (M-1)k}
 \end{aligned}$$

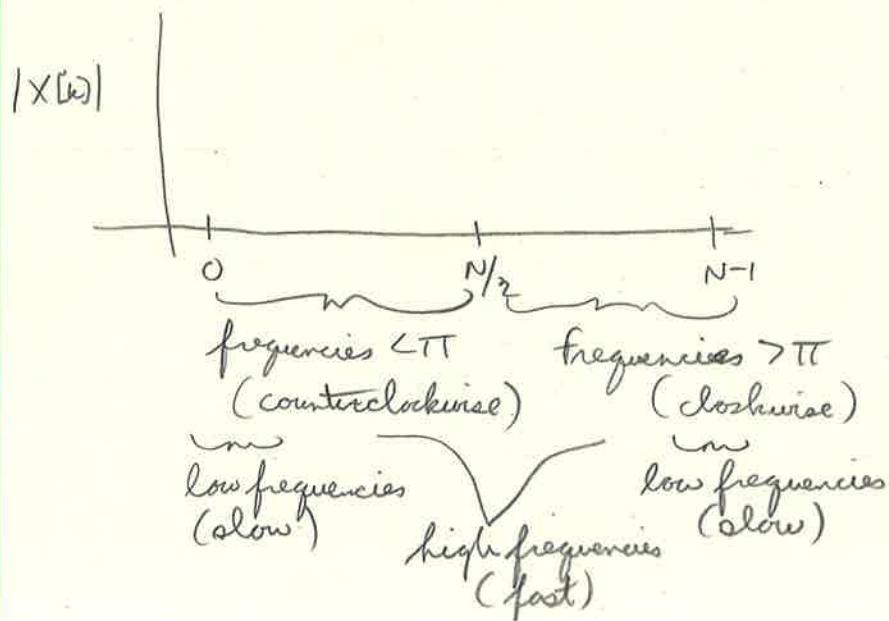
- $X[0] = M$, from the definition of the sum
- $X[k] = 0$, if M^k/N is an integer ($0 \leq k < N$)
- $\angle X[k]$ is linear in k (except at sign changes for the real part)
 - DFT of length-4 step in \mathbb{C}^{64}

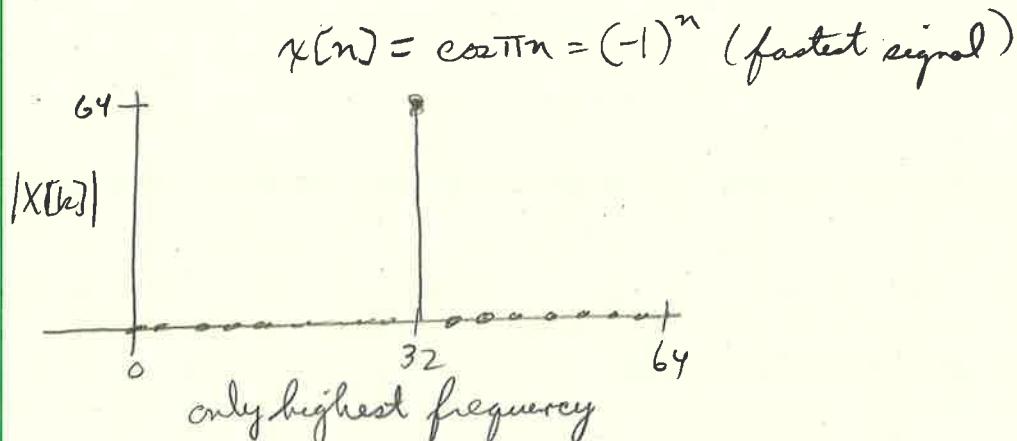
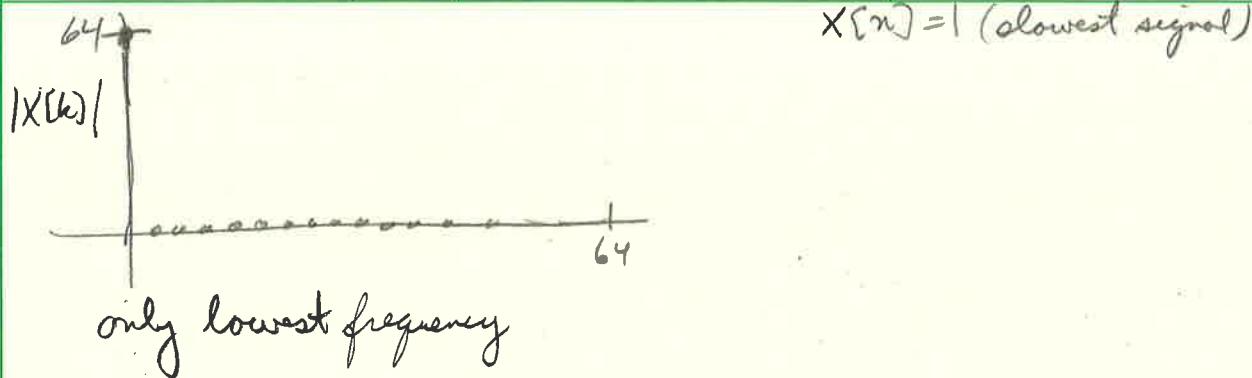


- Wrapping the phase

- Often the phase is displayed "wrapped" over the $[-\pi, \pi]$ interval
 - most numerical packages return wrapped phase
 - phase can be unwrapped by adding multiples of 2π

3.2c Interpreting a DFT plot





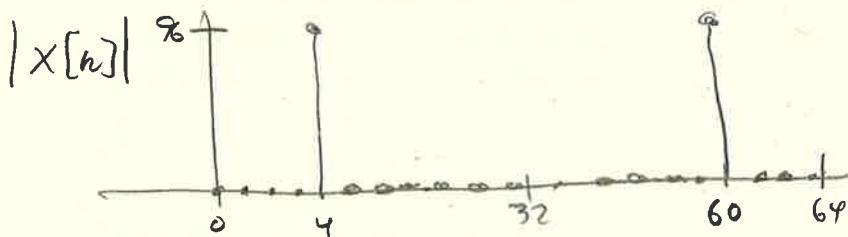
- Energy distribution

- Parseval : $\|\vec{x}\|^2 = \sum |x_k|^2$

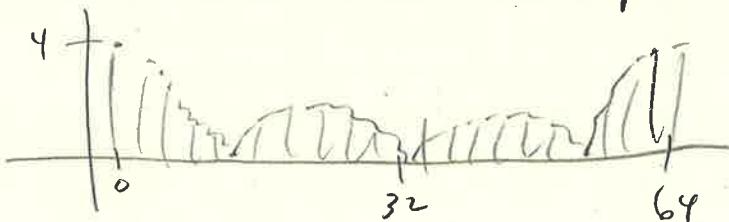
$$\sum_{n=0}^{N-1} |x(n)|^2 = \frac{1}{N} \sum_{k=0}^{N-1} |X(k)|^2$$

- square magnitude of k -th DFT coefficient proportional to signals energy at frequency $\omega = \frac{2\pi}{N} k$.

$$x(n) = 3 \cos\left(\frac{2\pi}{16} n\right) \quad (\text{sine wave})$$



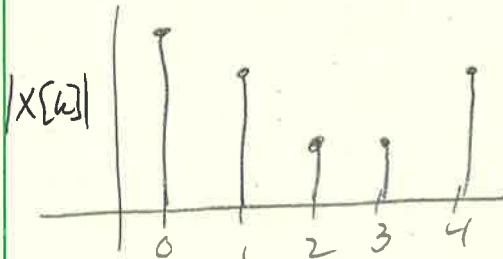
$$x(n) = u[n] - u[n-4] \quad (\text{step})$$



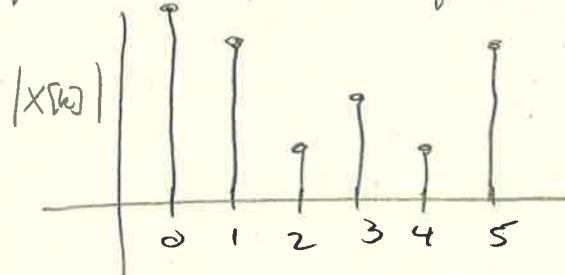
- DFT of real signals

For real signals the DFT is "symmetric" in magnitude:

$$|X[k]| = |X[N-k]| \text{ for } k=1, 2, \dots, \lfloor N/2 \rfloor \text{ (floor)}$$

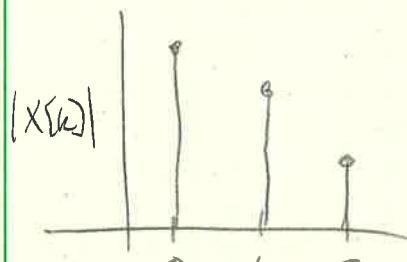


N=5, odd length

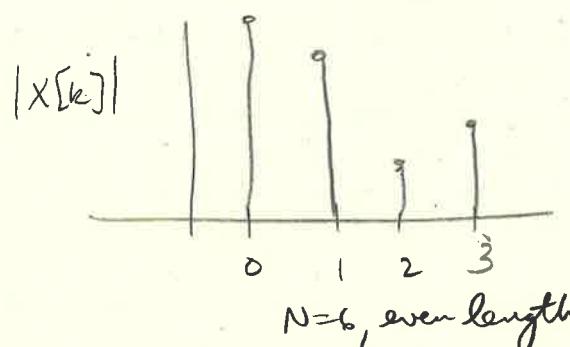


N=6, even length

For real signals, magnitude plots need only $\lfloor N/2 \rfloor + 1$ points



N=5, odd length



N=6, even length

3.3 : The DFT in practice

3.3a DFT analysis

- Mystery signal revisited

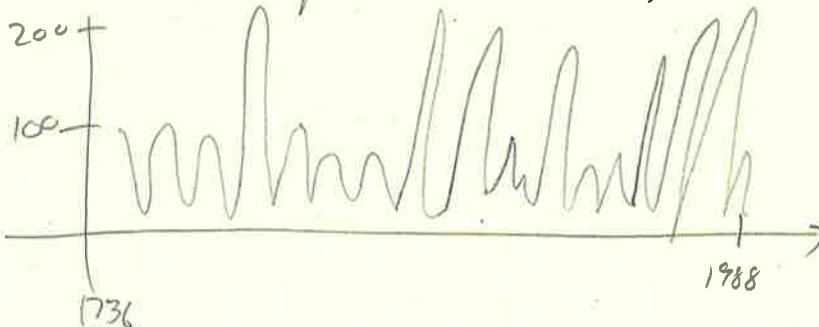
$$x[n] = \cos(\omega n + \phi) + \eta[n] \text{ with}$$

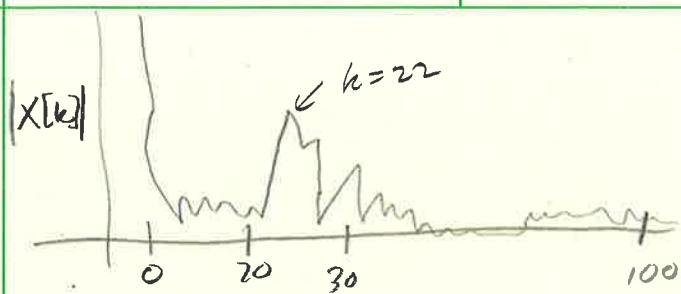
$$\phi=0, \omega = \frac{2\pi}{1024} 64 \quad \text{peak at } k=64$$

- Solar spots

- sunspot number : $S = 10 \times \# \text{ of clusters} + \# \text{ of spots}$

- data set from 1749 to 2003, 2904 months

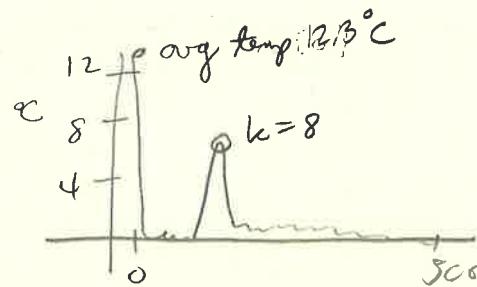
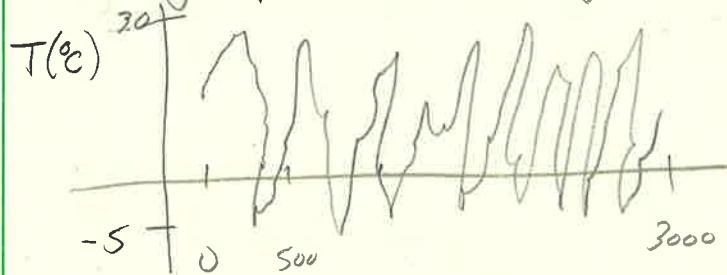




DFT of solar spots signal

- DFT main peak for $k=22$
- 22 cycles over 2904 months
- period: $\frac{2904}{22} \approx 11$ years

- Daily temperature (2920 days)

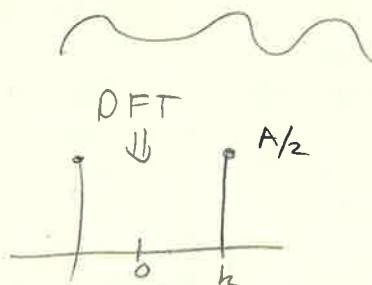


first few bunched DFT coefficients
(in magnitude and normalized by the length of the temperature vector)

- average value (0-th DFT coefficient) = 12.3°C normalized
- DFT main peak for $k=8$, value 6.4°C
- 8 cycles over 2920 days
- period: $\frac{2920}{8} = 365$ days
- temperature excursion: $12.3^\circ\text{C} \pm 12.8^\circ\text{C}$

$$X[0] = \sum_{n=0}^{N-1} X(n)$$

$$A = \cos(\omega n)$$



- Labeling the frequency axis

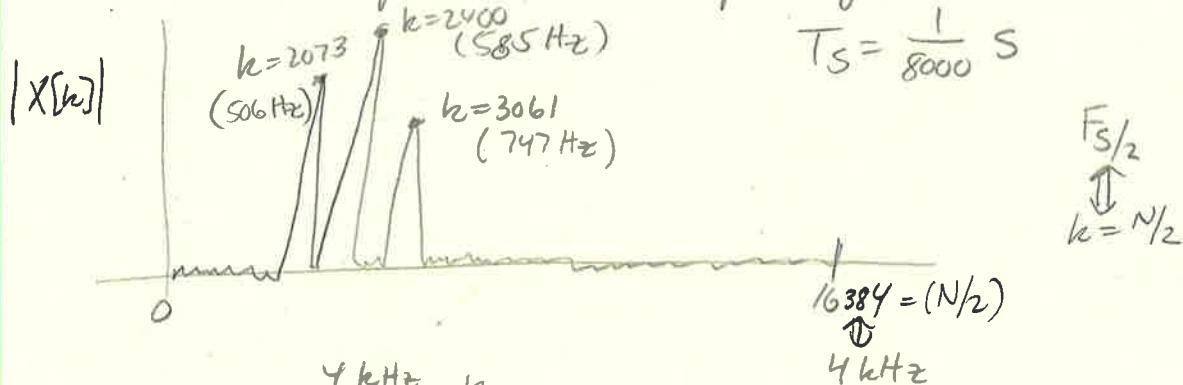
- If we know the "clock" of the system T_s
 - fastest (positive) frequency is $\omega = \pi$
 - sinusoid at $\omega = \pi$ needs two samples to do a full revolution
 - time between samples: $T_s = \frac{1}{F_s}$ seconds

- real-world period for fastest sinusoid: $2T_s$ seconds

- real-world frequency for fastest sinusoid: $F_s/2$ Hz

- Example: train whistle

32768 samples (the "clock" of the system $F_s = 8000 \text{ Hz}$)



$$\text{Frequency} = \frac{4 \text{ kHz}}{16384} \cdot k$$

If we look up the frequencies:



B minor chord

3.3b DFT example - analysis of musical instruments

- Analysis of musical instruments

"We all know about pitch..."

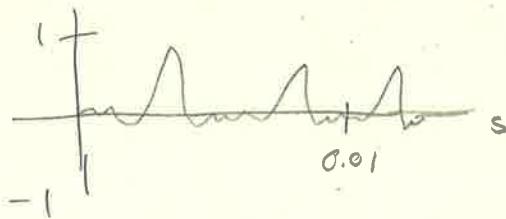
• It is really about frequency, or cycles per seconds

• How about harmonics?

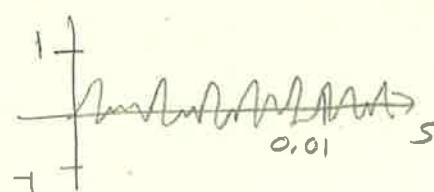
• That is what gives the timbre of an instrument!

- A difficult temporal analysis of musical instruments

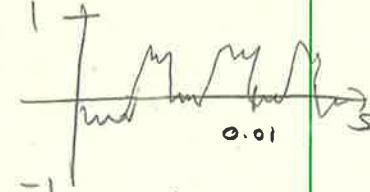
Sax



violin



cello

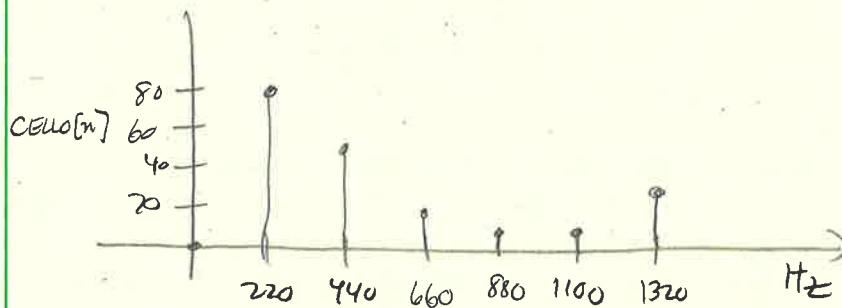
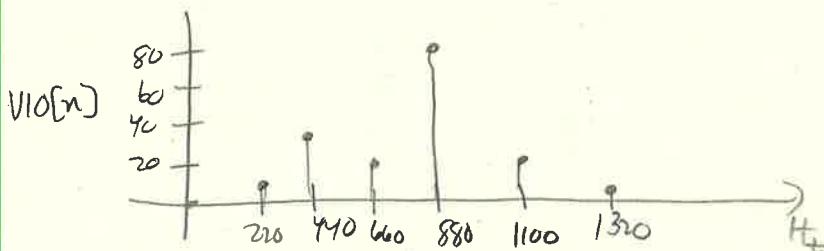
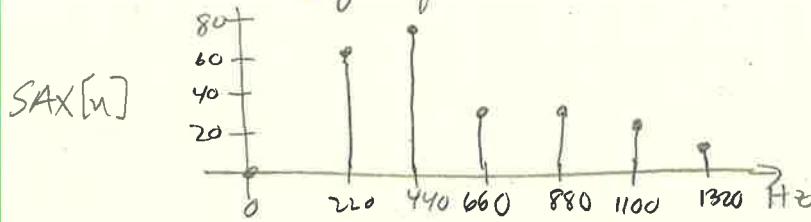


• What is the note?

• Could you guess the instrument from the temporal plot?

• In the time domain it is hard to process information of the sound

- A Fourier analysis of musical instruments

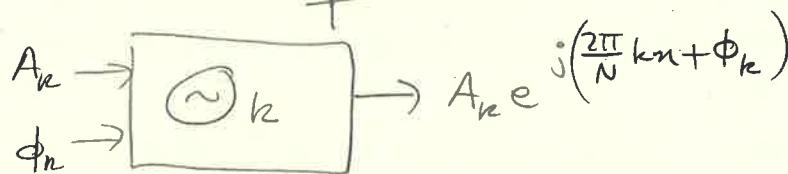
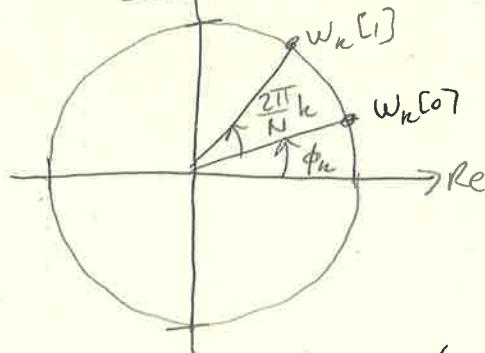


- The played note is the frequency of the first peak: 220 Hz in this case
- Other peaks are called harmonics: they define the typical sound of the instrument
- Without Fourier we would have been lost

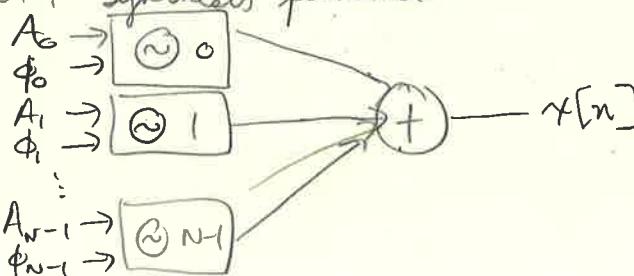
3.3c DFT synthesis

- Synthesis: the sinusoidal generator

$$w_k[n] = e^{j\left(\frac{2\pi}{N}kn + \phi_k\right)}$$



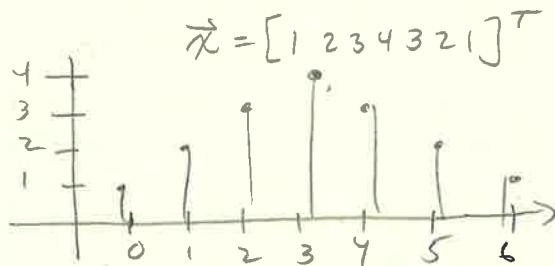
- DFT Synthesis formula



- Initializing the machine

$$A_k = |X[k]| / N$$

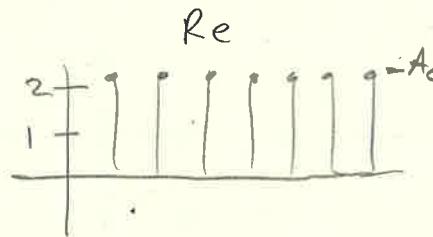
$$\phi_k = \angle X[k]$$



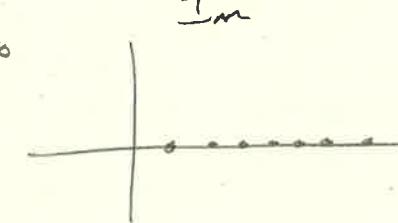
k	A_k	ϕ_k
0	2.2857	0
1	0.7213	-2.6928
2	0.0440	0.8976
3	0.0919	-1.7952
4	0.0919	-1.7952
5	0.0440	-0.8976
6	0.7213	2.6928

$$k=0$$

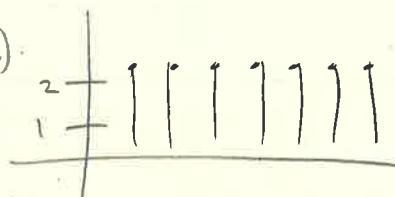
$$A_k e^{j(\frac{2\pi}{N} kn + \phi_k)}$$



Im

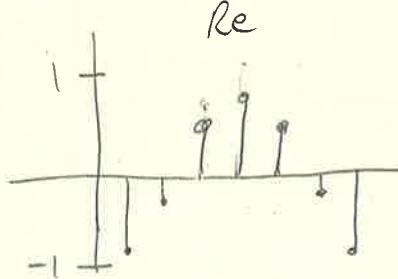


$$\sum A_k e^{j(\frac{2\pi}{N} kn + \phi_k)}$$

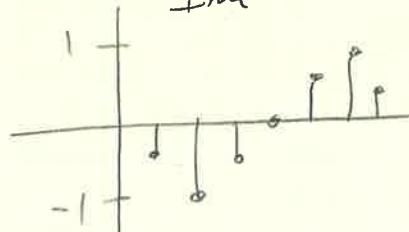


$$k=1$$

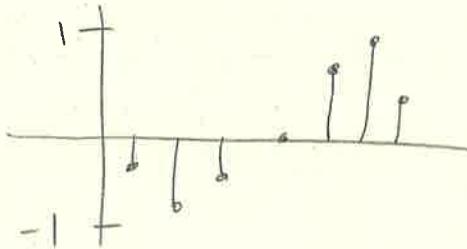
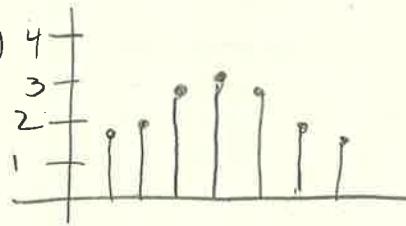
$$A_k e^{j(\frac{2\pi}{N} kn + \phi_k)}$$



Im



$$\sum A_k e^{j(\frac{2\pi}{N} kn + \phi_k)}$$



at $k=6$, we have reconstructed the signal

- Running the machine too long ...

$$x[n+N] = x[n] \quad \text{Output signal is } N\text{-periodic!}$$

- Inherent periodicities in the DFT

the synthesis formula:

$$x[n] = \frac{1}{N} \sum_{k=0}^{N-1} X[k] e^{j \frac{2\pi}{N} nk}, \quad n \in \mathbb{Z}, \text{ produces an}$$

N -periodic signal in the time domain

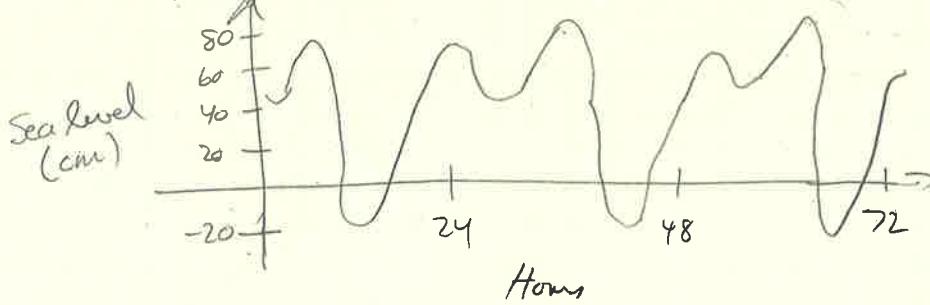
the analysis formula:

$$X[k] = \sum_{n=0}^{N-1} x[n] e^{-j \frac{2\pi}{N} nk}, \quad k \in \mathbb{Z}, \text{ produces } N\text{-periodic}$$

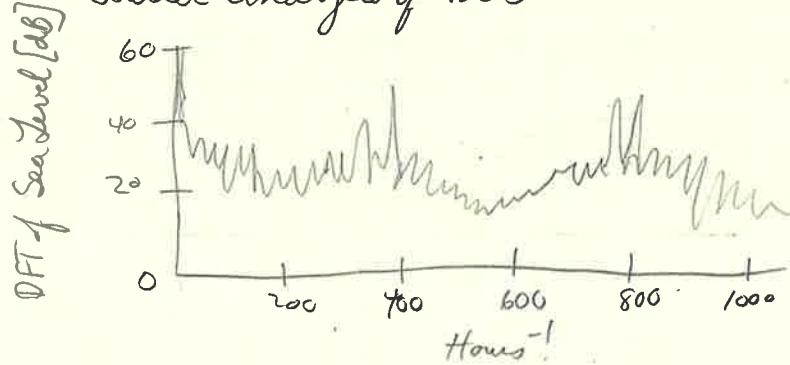
signal in the frequency domain.

3.3d DFT example - tide prediction in Venice

- Tides are due mostly to periodic astronomical phenomena
- Can we predict tides using Fourier? The first step is to approximate them
- We consider hourly measurements taken in Canal Grande during 2011
(3 days)



Fourier Analysis of Tides

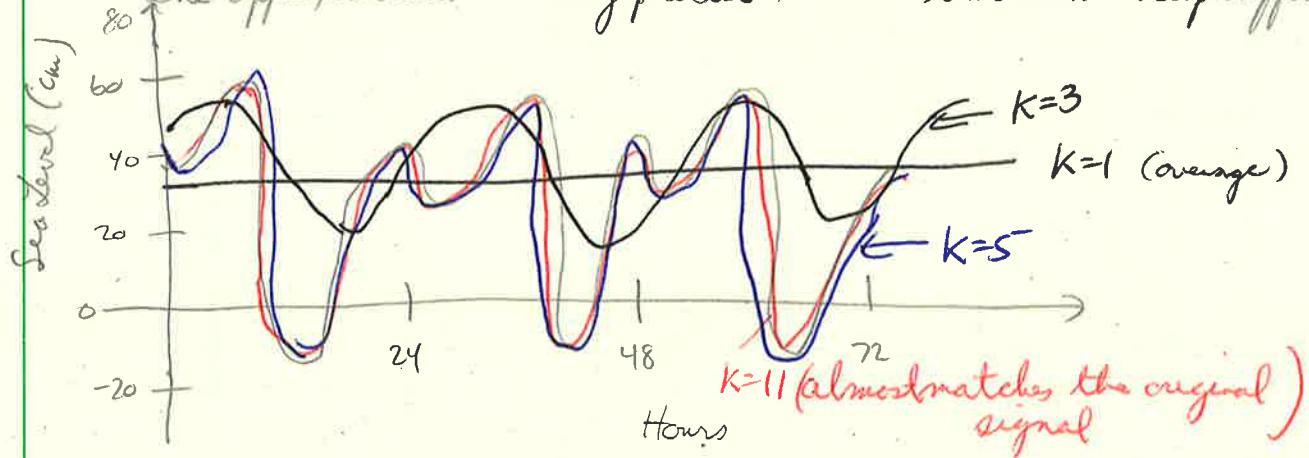


- Let us approximate tides using few Fourier coefficients

- We consider only $K=1, 3, 5, 11$ Fourier coefficients

- Darker gray represents approximation with more coefficients

- The approximations are very precise with a limited number of coefficients

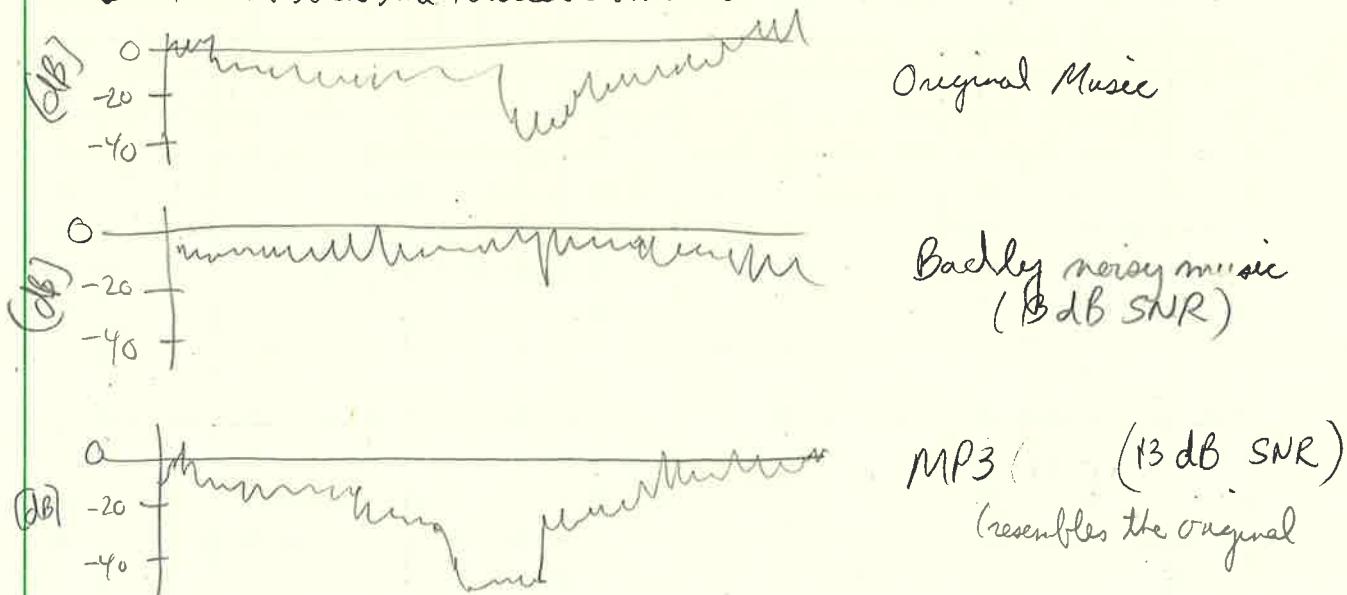


3.3e DFT example - MP3 compression

- MP3 Compression trick

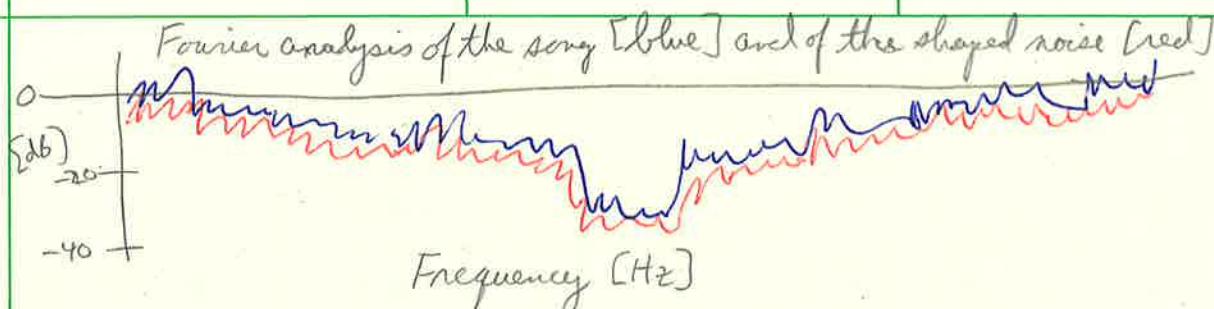
- How can an MP3 song sound so good while being so compressed?
- Compression introduces noise!
- The trick is to shape the noise!

The secret is in the Fourier domain!

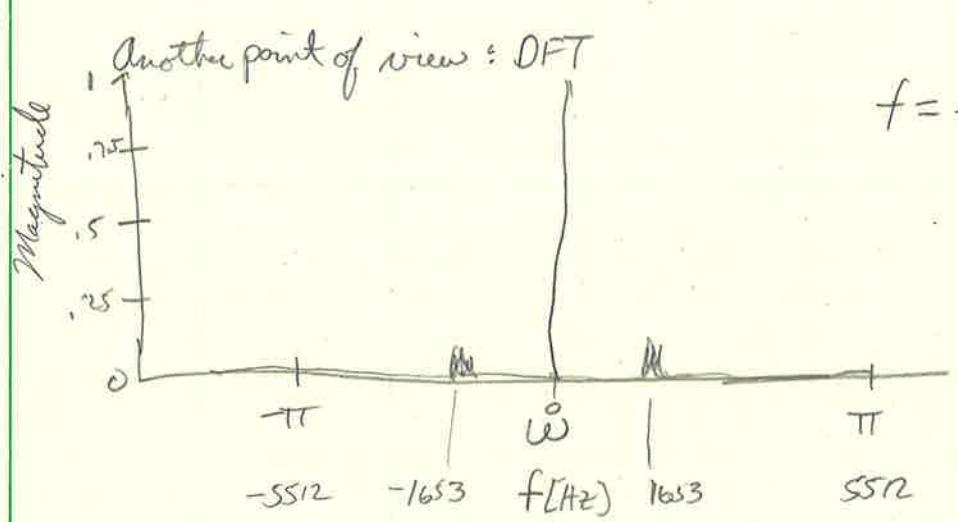
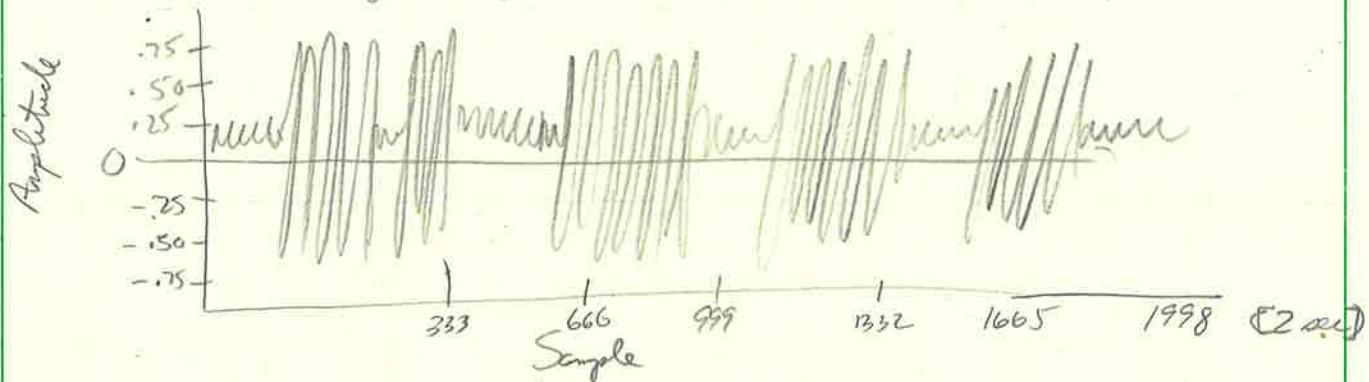


- Conclusions:

- MP3: complex compression algorithm that introduces errors
- Errors shaped as the song in the Fourier domain \rightarrow higher perceived quality
- MP3 minimizes the perceived quality decay by shaping the compression errors

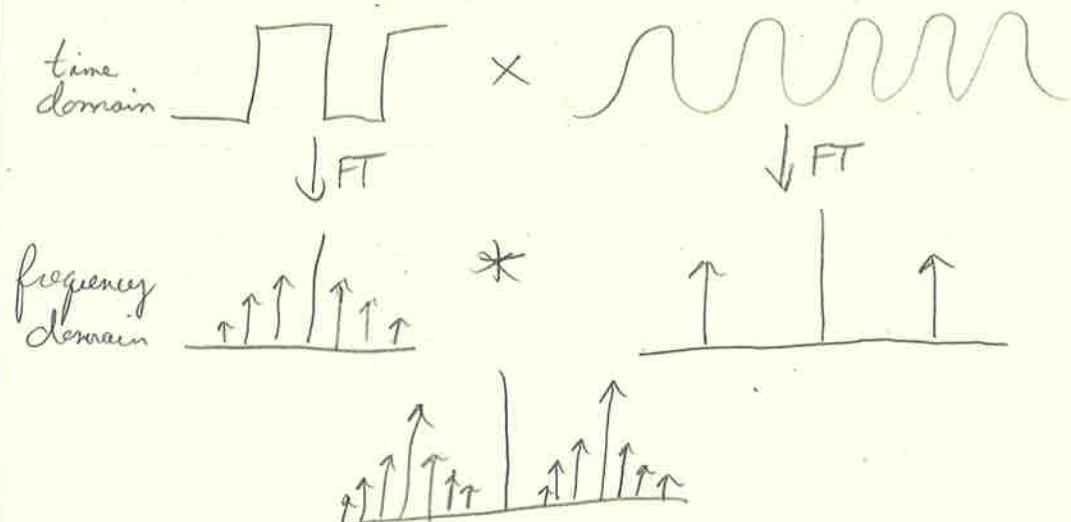


Signal of the Day: The first man-made signal from outer space (Sputnik)



$$f = \frac{\omega f_s}{2\pi} \leftarrow \begin{array}{l} \text{frequency at} \\ \text{which we} \\ \text{measure the signal} \end{array}$$

- Let's understand this...



Summary of Lesson 3.3

The DFT can be used as an analysis tool to understand the frequency components that a signal contains. If a signal has an associated system clock T_s (or a frequency $F_s = 1/T_s$), we can map the index k of the DFT coefficients to real frequencies. The largest digital frequency $N/2$ is associated with the largest continuous-time frequency $F_s/2$. Thus, the continuous frequency corresponding to index k is given by $\frac{kF_s}{N}$ and is measured in Hz.

The DFT synthesis can be seen as a series of up to N coupled sinusoidal generators:

- sinusoidal generator k has frequency $\frac{2\pi k}{N}$
- the amplitude of sinusoidal generator k is given by the magnitude of the DFT coefficient $|X[k]|$
- the phase of sinusoidal generator k is given by the phase of the DFT coefficient $\angle X[k]$

If we let the DFT synthesis run beyond $N-1$, we obtain an N -periodic signal, $x[n+N] = x[n]$. Likewise, the analysis formula produces also an N -periodic series of Fourier coefficients. This side comment will be very important when we study another form of Fourier transform for periodic sequences, namely discrete Fourier Series (DFS).

3.4 The Short-Time Fourier Transform (STFT)

3.4a The STFT

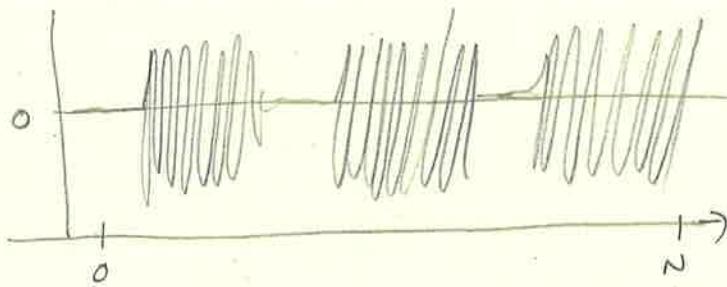
- Dual-Tone Multi Frequency dialing (DTMF)

	1209 Hz	1336 Hz	1477 Hz
697 Hz	1	2	3
770 Hz	4	5	6
852 Hz	7	8	9
941 Hz	*	0	#

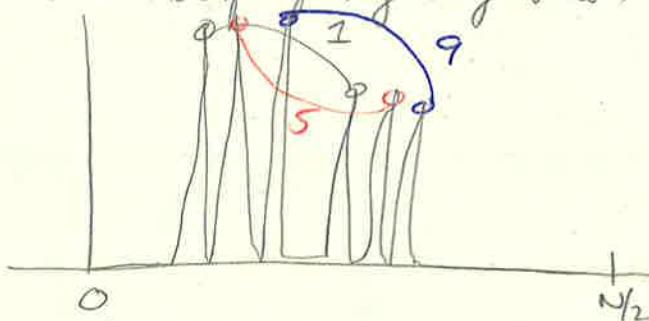
Analog
telephone

- 1) Frequencies are co-prime
- 2) No sum or difference of frequencies is in the set

1-5-9 in time

Can't tell the digit
in time domain

1-5-9 in frequency (magnitude)



- The fundamental tradeoff

- time representation obscures frequency
- frequency representation obscures time

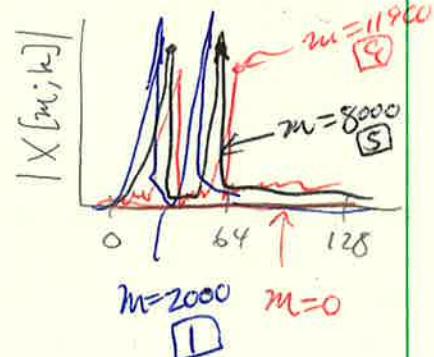
- Short-term Fourier Transform

Idea:

- take small signal pieces of length L
- look at the DFT of each piece

$$X[m; k] = \sum_{n=0}^{L-1} x[m+n] e^{-j \frac{2\pi}{L} nk}$$

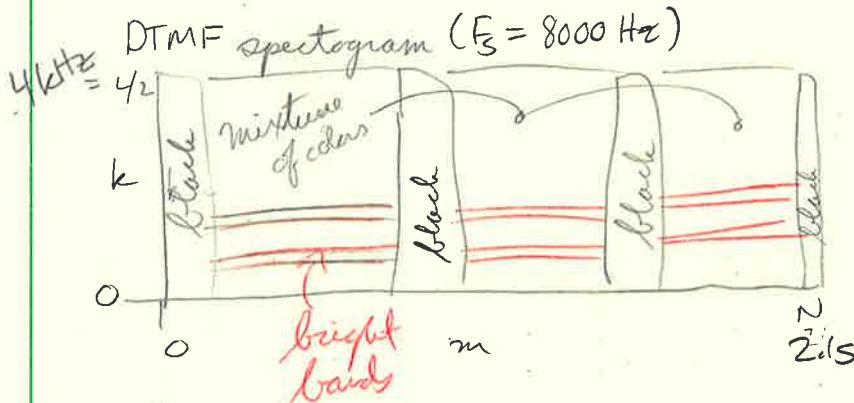
Starting point

STFT ($L=256$)

3.4b The spectrogram

Idea:

- Color-code the magnitude: dark is small, white is large
- use $10 \log_{10}(|X[m; k]|)$ to see better (power in dB)
- plot spectral slices one after another



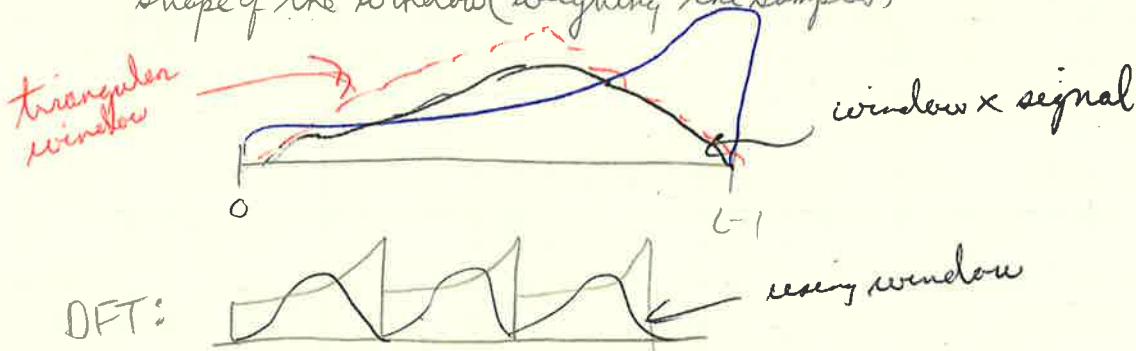
- Labeling the Spectrogram

- If we know the "system clock" $F_s = 1/T_s$, we can label the axis
 - highest positive frequency: $F_s/2 \text{ Hz}$
 - frequency resolution: $F_s/L \text{ Hz}$
 - width of time slices: $L T_s \text{ seconds}$

- The Spectrogram

Questions:

- width of the analysis window?
- position of the windows (overlapping?)
- shape of the window (weighing the samples)



- Wideband vs. Narrowband

Long window? narrowband spectrogram

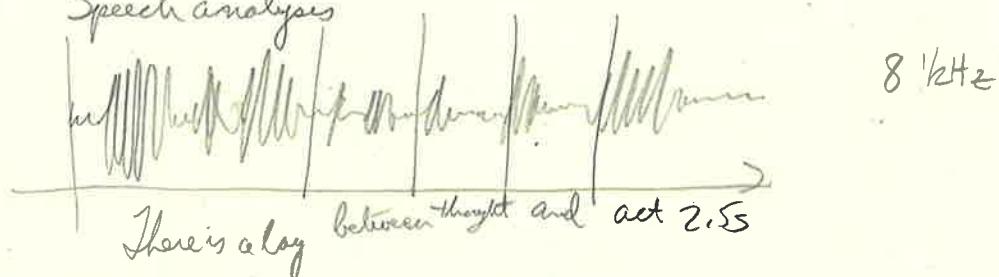
- long window \Rightarrow more DFT points \Rightarrow more frequency resolution $\frac{F_s}{L}$
- long window \Rightarrow more "things can happen" \Rightarrow less precision in time

Short window? wideband spectrogram

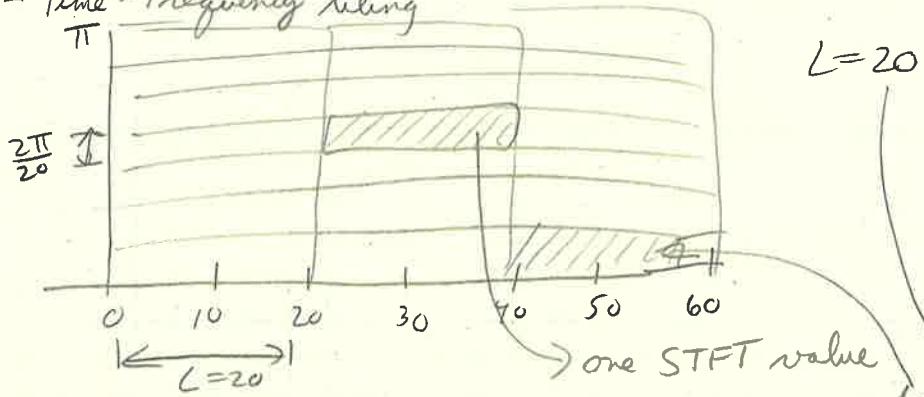
- short window \Rightarrow many time slices \Rightarrow precise location of transitions
- short window \Rightarrow fewer DFT points \Rightarrow poor frequency resolution

3.4c Time - frequency tiling

Speech analysis



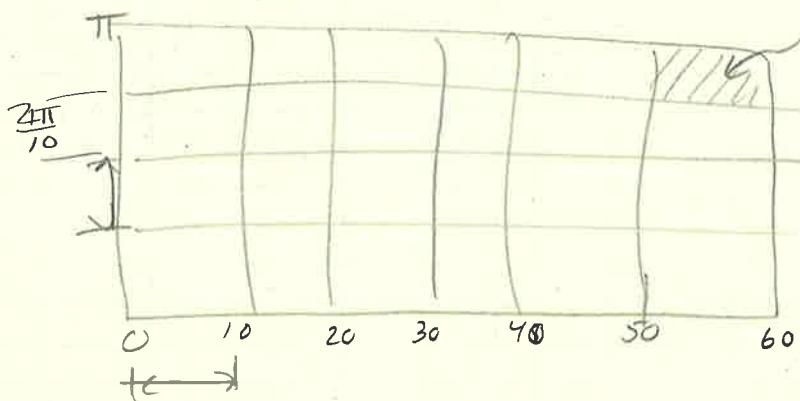
- Time - Frequency tiling



one STFT value

area of each tile = constant

$L = 10$



- Food for thought

- time "resolution" $\Delta t = L$
- frequency "resolution" $\Delta f = 2\pi/L$
- $\Delta t \Delta f = 2\pi$

uncertainty principle !

- Even more food for thought

- more sophisticated tilings of time - frequency planes can be obtained with the wavelet transform.

3.5 Discrete Fourier Series

3.5a Discrete Fourier Series

DFS = DFT with periodicity explicit

- The DFS maps an N -periodic signal onto an N -periodic sequence of Fourier coefficients
- The inverse DFS maps an N -periodic sequence of Fourier coefficients onto an N -periodic signal
- The DFS of an N -periodic signal is mathematically equivalent to the DFT of one period

- Finite-length time shifts revisited

The DFS helps us understand how to define time shifts for finite-length signals.

- For an N -periodic sequence $\tilde{x}[n]$

- $\tilde{x}[n-M]$ is well-defined for all $M \in \mathbb{N}$

$$\text{DFS} \{ \tilde{x}[n-M] \} = e^{-j \frac{2\pi}{N} M k} \tilde{X}[k], \quad \tilde{X}[k] = \text{DFS} \{ \tilde{x}[n] \}$$

$$-\text{IDFS} \left\{ e^{-j \frac{2\pi}{N} M k} \tilde{X}[k] \right\} = \tilde{x}[n-M]$$

delay factor

- For an N -point signal $x[n]$:

- $x[n-M]$ is not well-defined

- build $\tilde{x}[n] = x[n \bmod N] \Rightarrow \tilde{X}[k] = X[k]$

$$\begin{aligned} -\text{IDFT} \left\{ e^{-j \frac{2\pi}{N} M k} X[k] \right\} &= \text{IDFS} \left\{ e^{-j \frac{2\pi}{N} M k} \tilde{X}[k] \right\} \\ &= \tilde{x}[n-M] = x[(n-M) \bmod N] \end{aligned}$$

- Shifts for finite-length signals are "naturally" circular

3.5b Karpus-Strong revisited and DFS

- Periodic sequences: a bridge to infinite-length signals

- N -periodic sequence: N degrees of freedom

- DFS: only N Fourier coefficients capture all of the information

- Karpbus-Strong revisited



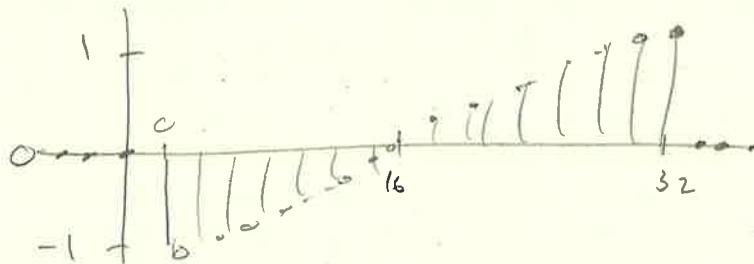
$$y[n] = \alpha y[n-M] + x[n]$$

- choose a signal $\bar{x}[n]$ that is nonzero only for $0 \leq n < M$

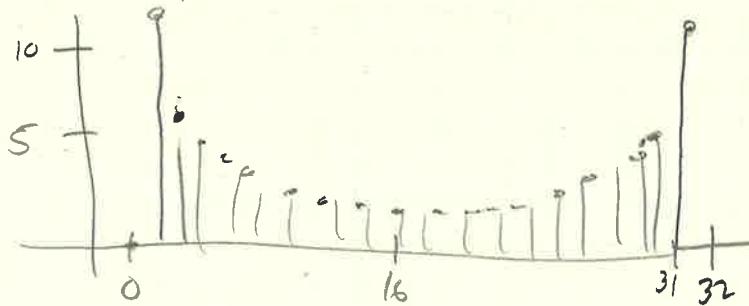
- $\alpha = 1$ (for now)

$$y[n] = \underbrace{\bar{x}[0], \bar{x}[1], \dots, \bar{x}[M-1]}_{\text{1st period}}, \underbrace{\bar{x}[0], \bar{x}[1], \dots, \bar{x}[M-1]}_{\text{2nd period}}, \bar{x}[0], \bar{x}[1], \dots, \dots$$

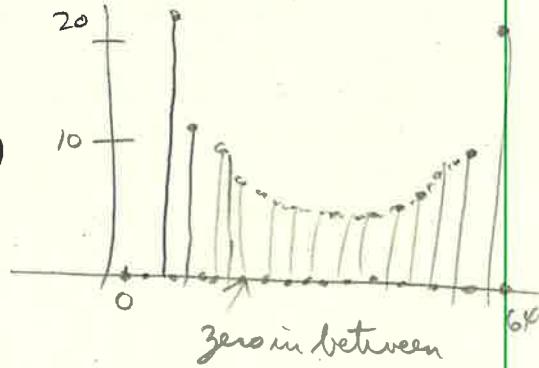
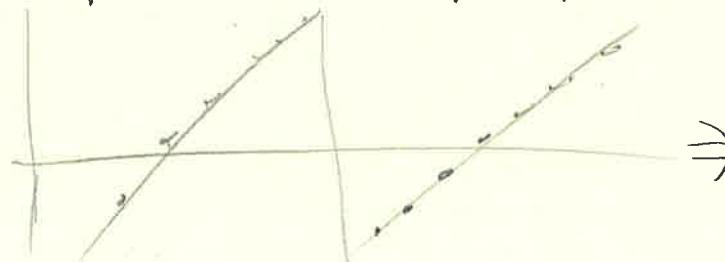
- Example : 32-tap Sawtooth wave

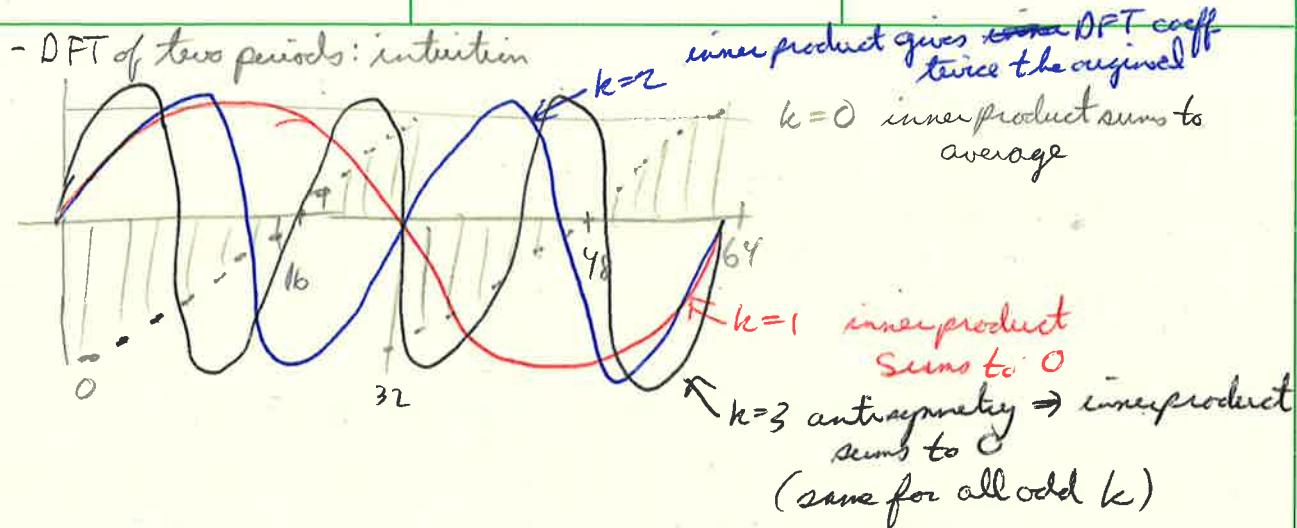


DFT of 32-tap sawtooth wave



- What if we take the DFT of two periods?





- DFT of L periods

$$X_L[k] = \sum_{n=0}^{L-1} y[n] e^{-j \frac{2\pi}{L} nk}, \quad k=0, 1, 2, \dots, L-1$$

$$(k=n+pM) = \sum_{p=0}^{L-1} \sum_{n=0}^{M-1} y[n+pM] e^{-j \frac{2\pi}{L}(n+pM)k}$$

$$\bar{x}[n] \quad \begin{array}{|c|c|c|c|}\hline & & & \\ \hline \end{array} \quad M$$

$$= \sum_{p=0}^{L-1} \sum_{n=0}^{M-1} y[n] e^{-j \frac{2\pi}{L} nk} e^{-j \frac{2\pi}{L} pk}$$

$$y[n] \quad \begin{array}{|c|c|c|c|}\hline & & & \\ \hline \end{array} \quad L \text{ periods}$$

$$= \left(\sum_{p=0}^{L-1} e^{-j \frac{2\pi}{L} pk} \right) \sum_{n=0}^{M-1} \bar{x}[n] e^{-j \frac{2\pi}{L} nk}$$

- We've seen this before

$$\sum_{p=0}^{L-1} e^{-j \frac{2\pi}{L} pk} = \begin{cases} L, & \text{if } k \text{ is a multiple of } L \\ 0, & \text{otherwise} \end{cases}$$

(remember the orthogonality proof for the DFT basis)

- DFT of L periods

$$X_L[k] = \begin{cases} L \bar{x}[k/L], & \text{if } k=0, L, 2L, \dots \\ 0, & \text{otherwise} \end{cases}$$

- DFT and DFS

- again, all the spectral information for a periodic signal is contained in the DFT coefficients of a single period.
- To stress the periodicity of the underlying signal, we use the term DFS

Summary of Lesson 3.5

The discrete Fourier series is just a different flavor of the DFT applied to periodic series. N -periodic sequences in the time domain are mapped onto N -periodic sequences in the frequency domain. Furthermore, the definition of the DFS retrospectively better justifies the use of circular shifts as the natural extension of shifts for finite-length sequences.

Later in the lesson, we have revisited the Karples-Strong algorithm to illustrate a key point about the DFS. If we take the DFT of L repetitions of a finite length sequence of length N , we obtain a series which is non-zero only at multiple integers of L . Moreover, these non-zero coefficients are just scaled versions of the DFT coefficients of the original finite-length sequence. Therefore, all the spectral information of a N -periodic sequence is entirely captured by the DFT coefficients of one period.

3.6 The Discrete-Time Fourier Transform

- The situation so far

- Fourier representation for signal classes:

- N -point finite-length: DFT

- N -point periodic: DFS

- infinite length: ?

- Karples-Strong revisited, part 2

- consider now $\alpha < 1$

- generated signal is infinite-length but not periodic:

$$y[n] = \underbrace{\bar{x}[0], \bar{x}[1], \dots, \bar{x}[M-1]}_{\text{1st period}}, \underbrace{\alpha \bar{x}[0], \alpha \bar{x}[1], \dots, \alpha \bar{x}[M-1]}_{\text{2nd period}}, \underbrace{\alpha^2 \bar{x}[0], \alpha^2 \bar{x}[1], \dots}_{n}$$

- What is a good spectral representation?

- DFT of increasingly long signals

- Start with the DFT. What happens when $N \rightarrow \infty$?

- $(2\pi/N)k$ becomes denser in $[0, 2\pi]$... C^N , $\omega = \frac{2\pi}{N}$

- In the limit $(\frac{2\pi}{N})k \rightarrow \omega$: $\sum_n x[n] e^{-j\omega n}$, $\omega \in \mathbb{R}$

- Discrete-Time Fourier Transform (DTFT)

- Formal definition:

- $x[n] \in l_2(\mathbb{Z})$

- define a function of $\omega \in \mathbb{R}$

$$F(\omega) = \sum_{n=-\infty}^{\infty} x[n] e^{-j\omega n}$$

- inversion (when $F(\omega)$ exists):

$$x[n] = \frac{1}{2\pi} \int_{-\pi}^{\pi} F(\omega) e^{j\omega n} d\omega, n \in \mathbb{Z}$$

- DTFT periodicity and notation

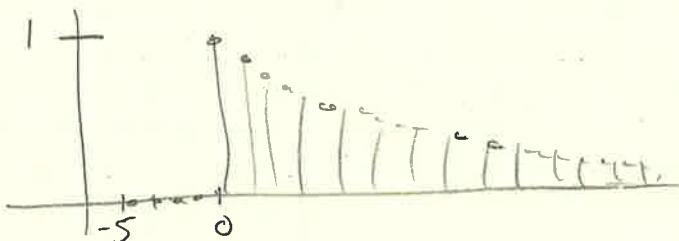
- $F(\omega)$ is 2π -periodic

- to stress periodicity (and for other reasons) we will write

$$X(e^{j\omega}) = \sum_{n=-\infty}^{\infty} x[n] e^{-j\omega n}$$

- by convention, $X(e^{j\omega})$ is represented over $[-\pi, \pi]$

$$- x[n] = \alpha^n u[n], |\alpha| < 1$$



- DTFT of $x[n] = \alpha^n u[n], |\alpha| < 1$

$$X(e^{j\omega}) = \sum_{n=-\infty}^{\infty} x[n] e^{-j\omega n}$$

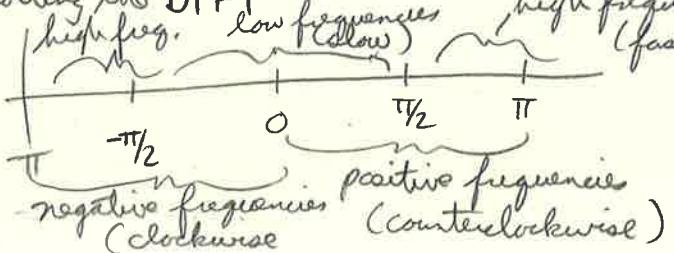
$$= \sum_{n=0}^{\infty} \alpha^n e^{-j\omega n}$$

$$= \sum_{n=0}^{\infty} (\alpha e^{-j\omega})^n$$

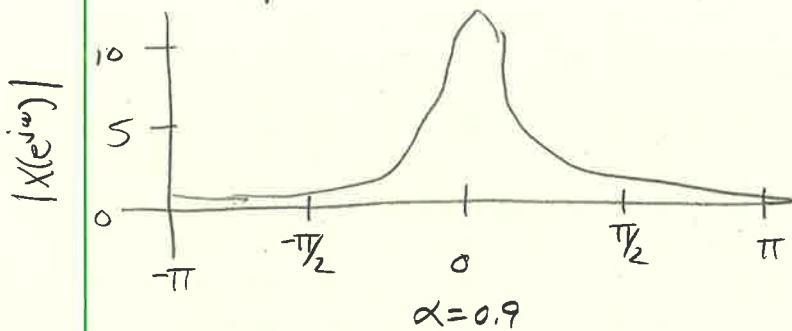
$$= \frac{1}{1 - \alpha e^{-j\omega}}$$

$$|X(e^{j\omega})|^2 = \frac{1}{1 + \alpha^2 - 2\alpha \cos \omega}$$

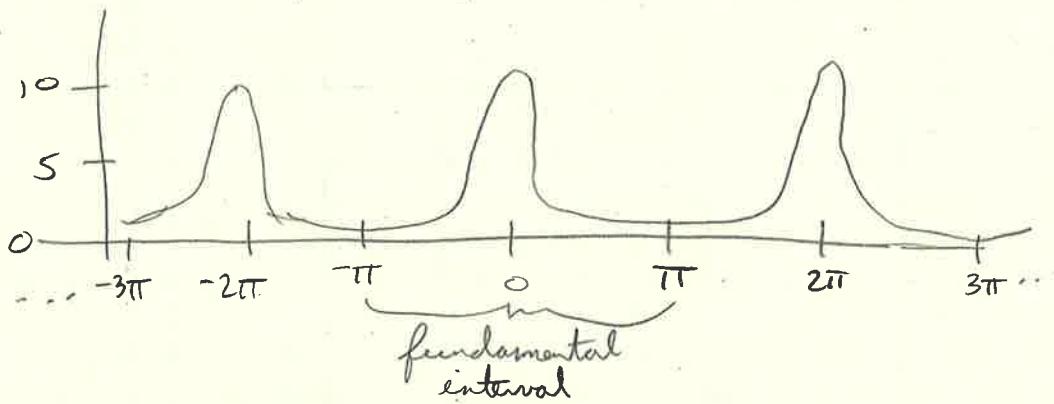
- Plotting the DTFT



- DTFT of $x[n] = \alpha^n u[n]$, $|\alpha| < 1$



- Remember the periodicity!

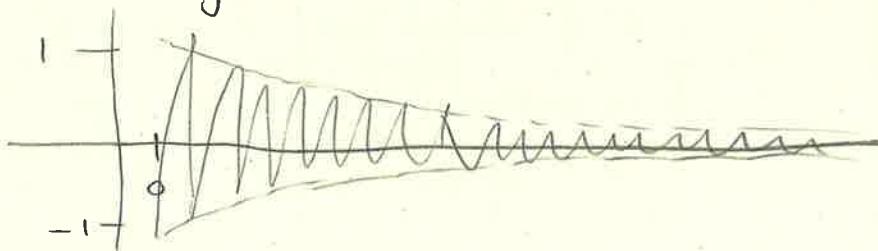


- KS revisited, part 2: 32-tap sawtooth wave

$$x[n] = \frac{2n}{M-1} - 1, \quad n=0, 1, \dots, M-1 \quad (M=32)$$

- KS revisited, part 2: decay $\alpha = 0.9$

$$y[n] = \alpha^{\lfloor \frac{n}{M} \rfloor} \bar{x}[n \bmod M] u[n]$$

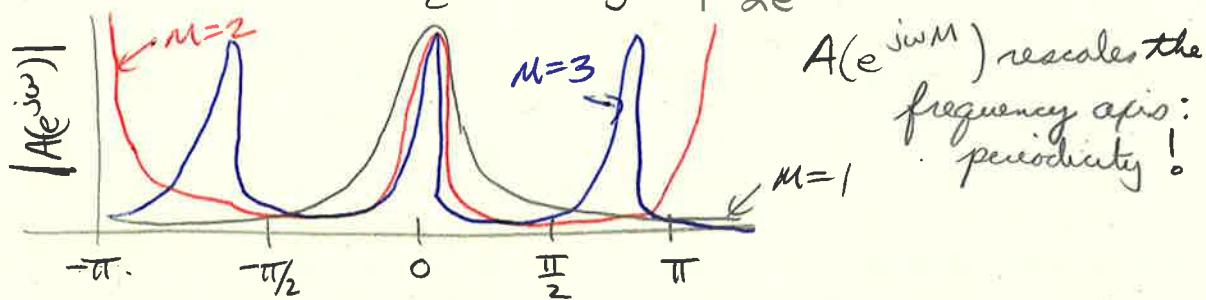


- DTFT of KS signal

$$\begin{aligned} Y(e^{j\omega}) &= \sum_{n=-\infty}^{\infty} y[n] e^{-j\omega n} \\ &= \sum_{p=0}^{\infty} \sum_{n=0}^{M-1} \alpha^p \bar{x}[n] e^{-j\omega(pM+n)} \\ &= \sum_{p=0}^{\infty} \alpha^p e^{-j\omega Mp} \sum_{n=0}^{M-1} \bar{x}[n] e^{-j\omega n} \\ &= A(e^{j\omega M}) \bar{X}(e^{j\omega}) \end{aligned}$$

- We know the first term

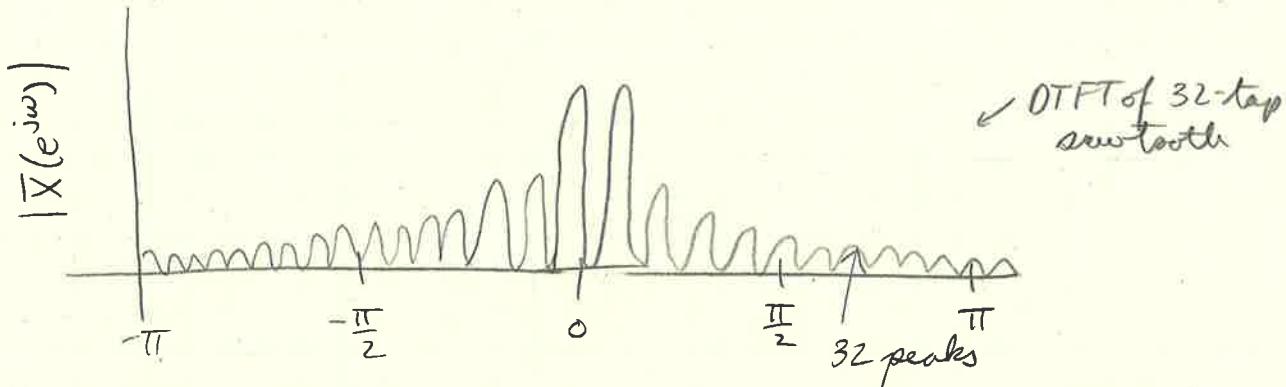
$$A(e^{j\omega}) = \text{DTFT}\{\alpha^n u[n]\} = \frac{1}{1-\alpha e^{-j\omega}} \quad (n=1)$$



$A(e^{j\omega M})$ rescales the frequency axis: periodically!

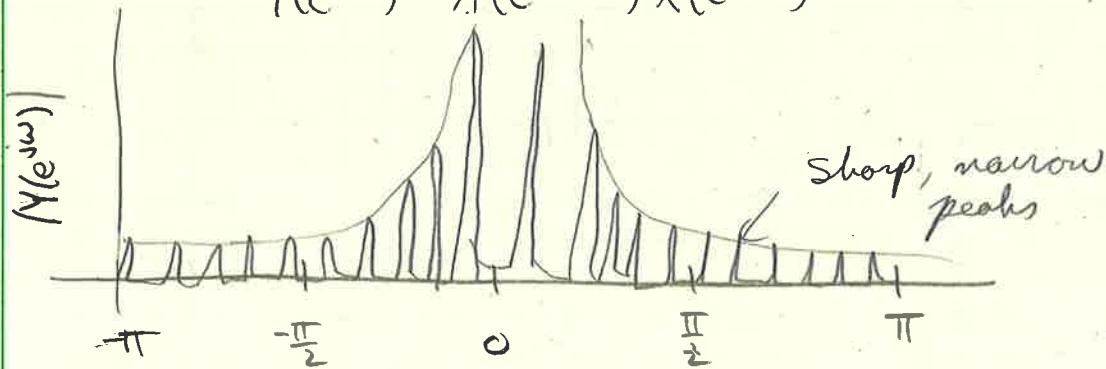
- Second term

$$\bar{X}(e^{j\omega}) = e^{-j\omega} \left(\frac{M+1}{M-1} \right) \frac{1-e^{-j(M-1)\omega}}{(1-e^{-j\omega})^2} - \frac{1-e^{-j(M+1)\omega}}{(1-e^{-j\omega})^2}$$



- DTFT of KS with decay

$$Y(e^{j\omega}) = A(e^{j\omega M}) \bar{X}(e^{j\omega})$$



3.6b Existence and properties of the OTFT

- Existence easy for absolutely summable sequences

$$\begin{aligned} |\bar{X}(e^{j\omega})| &= \left| \sum_{n=-\infty}^{\infty} x[n] e^{-j\omega n} \right| \\ &\leq \sum_{n=-\infty}^{\infty} |x[n]| e^{-j\omega n} = \sum_{n=-\infty}^{\infty} |x[n]| < \infty \end{aligned}$$

- Inversion easy for absolutely summable sequences

$$\begin{aligned} \frac{1}{2\pi} \int_{-\pi}^{\pi} X(e^{jw}) e^{jwn} dw &= \frac{1}{2\pi} \int_{-\pi}^{\pi} \left(\sum_{k=-\infty}^{\infty} x[k] e^{-jkw} \right) e^{jwn} dw \\ &= \sum_{k=-\infty}^{\infty} x[k] \underbrace{\int_{-\pi}^{\pi} \frac{e^{jw(n-k)}}{2\pi} dw}_{n} \\ &= x[n] \quad = 0 \text{ unless } n=k \end{aligned}$$

- A formal change of bases

• Formally DTFT is an inner product in \mathbb{C}^∞ :

$$\sum_{n=-\infty}^{\infty} x[n] e^{-jwn} = \langle e^{jwn}, x[n] \rangle$$

- "basis" is an infinite, uncountable basis: $\{e^{jwn}\}_{w \in \mathbb{R}}$
- something "breaks down": we start with sequences but the transform is a function
- we used absolutely summable sequences but DTFT exists for all square-summable sequences (proof is rather technical)

- Review: DFT

$$X[k] = \langle e^{j \frac{2\pi}{N} nk}, x[n] \rangle$$

basis: $\{e^{j \frac{2\pi}{N} nk}\}_k$

$$x[n] = \frac{1}{N} \sum X[k] e^{j \frac{2\pi}{N} nk}$$

- Review: DPS

$$\tilde{X}[k] = \langle e^{j \frac{2\pi}{N} nk}, \tilde{x}[n] \rangle$$

basis: $\{e^{j \frac{2\pi}{N} nk}\}_k$

$$\tilde{x}[n] = \frac{1}{N} \sum \tilde{X}[k] e^{j \frac{2\pi}{N} nk}$$

- DTFT

$$X(e^{j\omega}) = \langle e^{j\omega n}, x[n] \rangle$$

$\ell_2(\mathbb{R})$

"basis": $\{e^{j\omega n}\}_\omega$

$\ell_2([-T, T])$

$$x[n] = \frac{1}{2\pi} \int_{-\pi}^{\pi} X(e^{j\omega}) e^{j\omega n} d\omega$$

- DTFT properties

- linearity: $\text{DTFT}\{\alpha x[n] + \beta y[n]\} = \alpha X(e^{j\omega}) + \beta Y(e^{j\omega})$
- time shift: $\text{DTFT}\{x[n-M]\} = e^{-j\omega M} X(e^{j\omega})$
- modulation(dual): $\text{DTFT}\{e^{j\omega_0 n} x[n]\} = X(e^{j(\omega-\omega_0)})$
- time reversal: $\text{DTFT}\{x[-n]\} = X(e^{-j\omega})$
- conjugation: $\text{DTFT}\{x^*[n]\} = X^*(e^{-j\omega})$

- Some particular cases:

- if $x[n]$ is symmetric, the DTFT is symmetric:

$$x[n] = x[-n] \Leftrightarrow X(e^{j\omega}) = X(e^{-j\omega})$$

- if $x[n]$ is real, the DTFT is Hermitian-symmetric:

$$x[n] = x^*[n] \Leftrightarrow X(e^{j\omega}) = X^*(e^{-j\omega})$$

- special case: if $x[n]$ is real, the magnitude of the DTFT is symmetric:

$$x[n] \in \mathbb{R} \Rightarrow |X(e^{j\omega})| = |X(e^{-j\omega})|$$

- more special case: if $x[n]$ is real and symmetric, $X(e^{j\omega})$ is also real and symmetric.

3.6c The DTFT as a change of basis

- DTFT as basis expansion

• Some things are OK:

$$\text{DTFT}\{\delta[n]\} = 1$$

$$\text{DTFT}\{\delta[n]\} = \langle e^{j\omega n}, \delta[n] \rangle = 1$$

- Some things aren't:

$$\text{- DFT } \{1\} = N S[k]$$

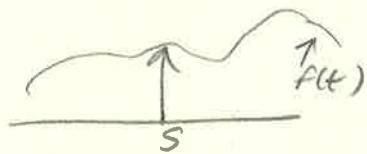
$$\text{- DTFT } \{1\} = \sum_{n=-\infty}^{\infty} e^{-j\omega n} = ?$$

- problem: too many interesting sequences are not square-summable!

- The Dirac delta functional

- Defined by the "sifting" property:

$$\int_{-\infty}^{\infty} \delta(t-s) f(t) dt = f(s), \quad \forall \text{ functions } t \in \mathbb{R}, s \in \mathbb{R}$$



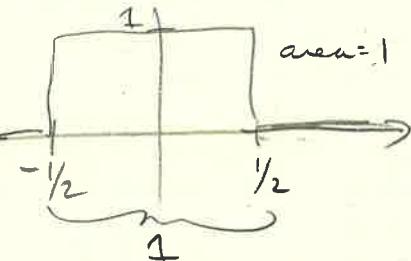
- Intuition

- Family of localizing functions $r_k(t)$ with $k \in \mathbb{N}$ and $t \in \mathbb{R}$

- Support inversely proportional to k

- Constant area

$$\text{rect}(t) = \begin{cases} 1, & |t| < \frac{1}{2} \\ 0, & \text{otherwise} \end{cases}$$

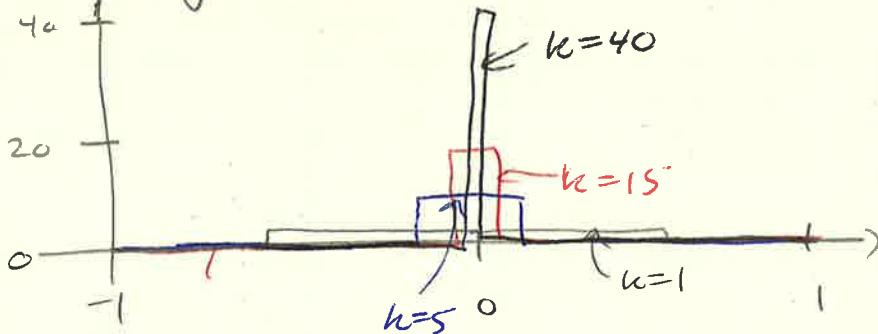


- Consider the localizing family $r_k(t) = k \text{rect}(kt)$:

- nonzero over $[-\frac{1}{2k}, \frac{1}{2k}]$, i.e., support is $\frac{1}{k}$

- area is 1

- The family $r_k(t) = k \text{rect}(kt)$



- Extracting a point value

- By the mean value theorem:

$$\int_{-\infty}^{\infty} r_k(t) f(t) dt = k \int_{-\frac{1}{2k}}^{\frac{1}{2k}} f(t) dt = f(\gamma) \Big|_{\gamma \in [-\frac{1}{2k}, \frac{1}{2k}]}$$

and so:

$$\lim_{k \rightarrow \infty} \int_{-\infty}^{\infty} r_k(t) f(t) dt = f(0)$$

- The Dirac delta functional

The delta functional shorthand. Instead of writing

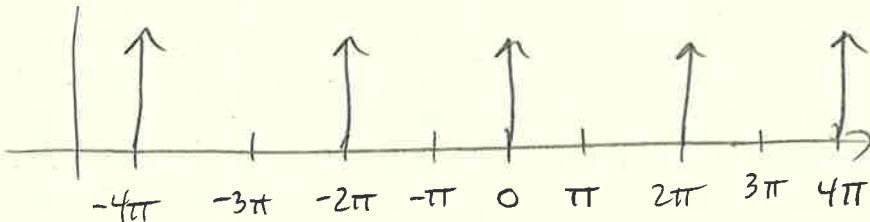
$$\lim_{k \rightarrow \infty} \int_{-\infty}^{\infty} r_k(t-s) f(t) dt$$

we write $\int_{-\infty}^{\infty} \delta(t-s) f(t) dt$, as if $\lim_{k \rightarrow \infty} r_k(t) = \delta(t)$.

- The "pulse train"

$$\tilde{S}(\omega) = 2\pi \sum_{k=-\infty}^{\infty} \delta(\omega - 2\pi k)$$

just a technique to use the Dirac delta in the space of
\$2\pi\$-periodic functions



$$\text{IDFT}\{\tilde{S}(\omega)\} = \frac{1}{2\pi} \int_{-\pi}^{\pi} \tilde{S}(\omega) e^{j\omega n} d\omega$$

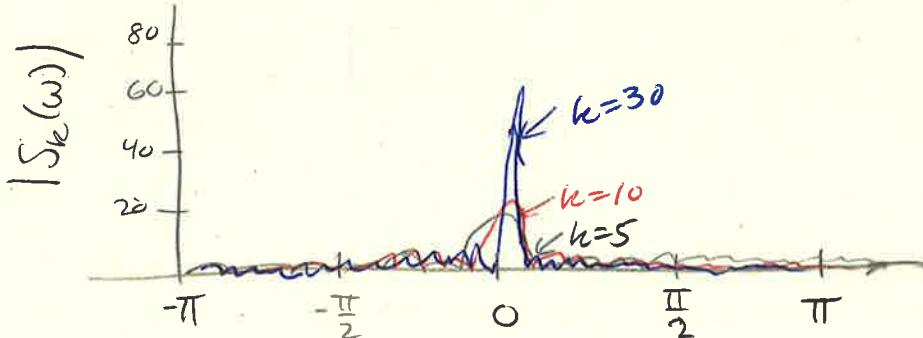
$$= \int_{-\pi}^{\pi} \delta(\omega) e^{j\omega n} d\omega = e^{j\omega n} \Big|_{\omega=0} = 1$$

$$(\text{IDFT}\{N \delta[k]\} = 1)$$

$$\Rightarrow \text{DTFT}\{1\} = \tilde{S}(\omega)$$

- Does this make sense?

Partial DTFT sum: $S_k(\omega) = \sum_{n=-k}^k e^{-j\omega n}$



- Using the same technique

$$\text{IDTFT} \left\{ \tilde{\delta}(\omega - \omega_0) \right\} = e^{j\omega_0 n}$$

- ⇒
 - DTFT $\left\{ 1 \right\} = \tilde{\delta}(\omega)$
 - DTFT $\left\{ e^{j\omega_0 n} \right\} = \tilde{\delta}(\omega - \omega_0)$
 - DTFT $\left\{ \cos \omega_0 n \right\} = [\tilde{\delta}(\omega - \omega_0) + \tilde{\delta}(\omega + \omega_0)]/2$
 - DTFT $\left\{ \sin \omega_0 n \right\} = -j[\tilde{\delta}(\omega - \omega_0) - \tilde{\delta}(\omega + \omega_0)]/2$

3.7: Sinusoidal Modulation

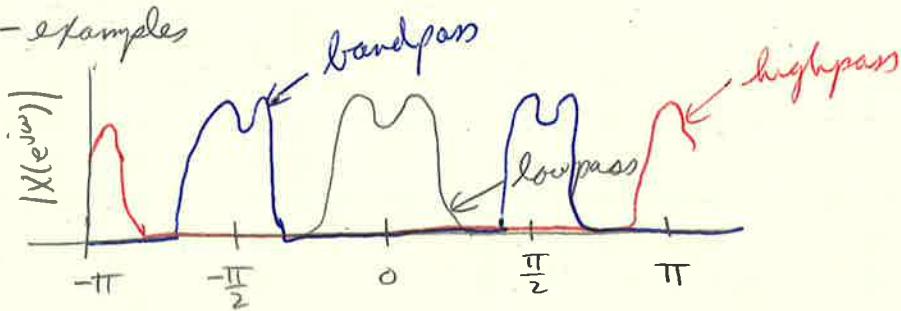
3.7a Sinusoidal modulation

- Classifying signals in frequency

* Three broad categories according to where most of the spectral energy resides:

- lowpass signals (also known as "baseband" signals)
- highpass signals
- bandpass signals

- examples



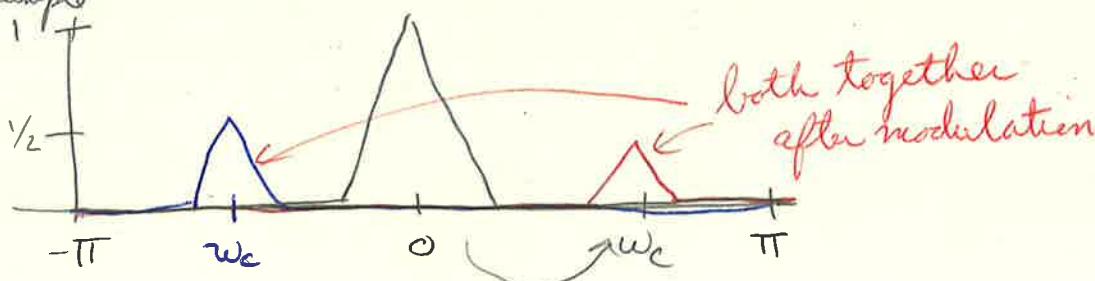
- Sinusoidal modulation

$$\begin{aligned} \text{DTFT} \left\{ x[n] \cos(\omega_c n) \right\} &= \text{DTFT} \left\{ \frac{1}{2} e^{j\omega_0 n} x[n] + \frac{1}{2} e^{-j\omega_0 n} x[n] \right\} \\ &= \frac{1}{2} [X(e^{j(\omega - \omega_c)}) + X(e^{j(\omega + \omega_c)})] \end{aligned}$$

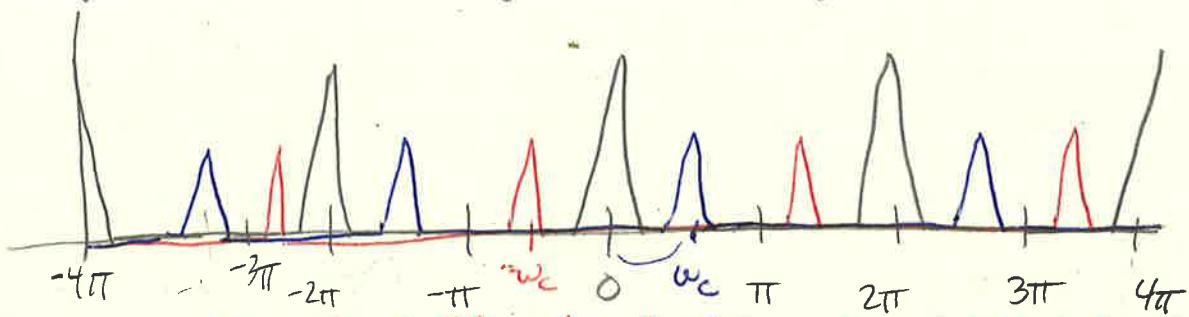
* usually $x[n]$ baseband

* ω_c is the carrier frequency

- example

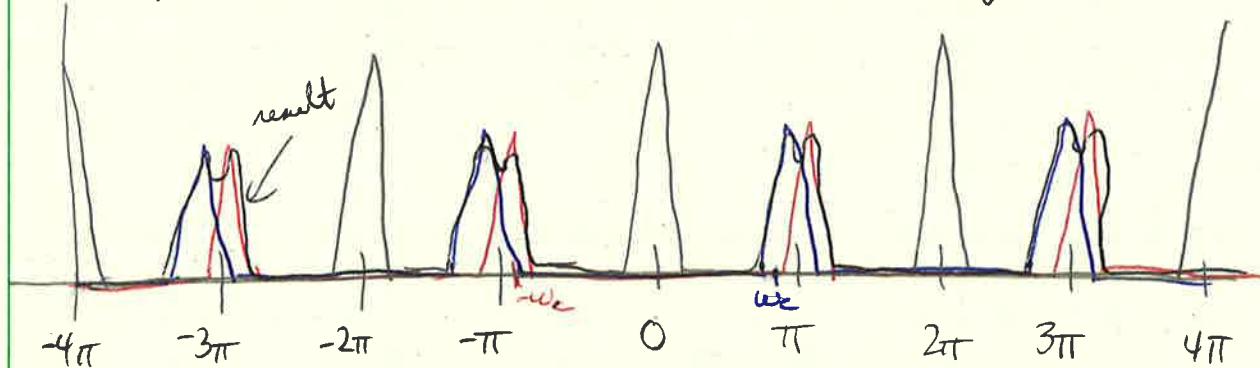


- Again, explicitly showing the periodicity of the spectrum



combine red and blue to get the resulting modulated spectrum

- Careful when the modulation frequency is too large!



- Sinusoidal modulation: applications

- voice and music are lowpass signals
- radio channels are bandpass, in much higher frequencies
- modulation brings the baseband signal in the transmission band
- demodulation at the receiver brings it back

- Sinusoidal demodulation

just multiply the received signal by the carrier again

$$y[n] = x[n] \cos(\omega_c n), \quad Y(e^{j\omega}) = \frac{1}{2} [X(e^{j(\omega-\omega_c)}) + X(e^{j(\omega+\omega_c)})]$$

$$\begin{aligned} \text{DTFT}\{y[n] \cdot 2 \cos(\omega_c n)\} &= Y(e^{j(\omega-\omega_c)}) + Y(e^{j(\omega+\omega_c)}) \\ &= \frac{1}{2} [X(e^{j(\omega-2\omega_c)}) + X(e^{j\omega}) + X(e^{j\omega}) + X(e^{j(\omega+2\omega_c)})] \end{aligned}$$

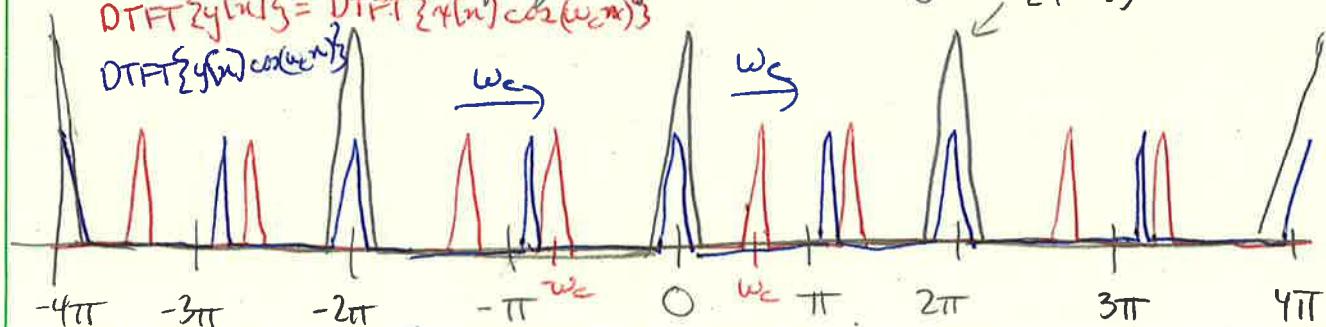
$$= X(e^{j\omega}) + \frac{1}{2} [X(e^{j(\omega-2\omega_c)}) + X(e^{j(\omega+2\omega_c)})]$$

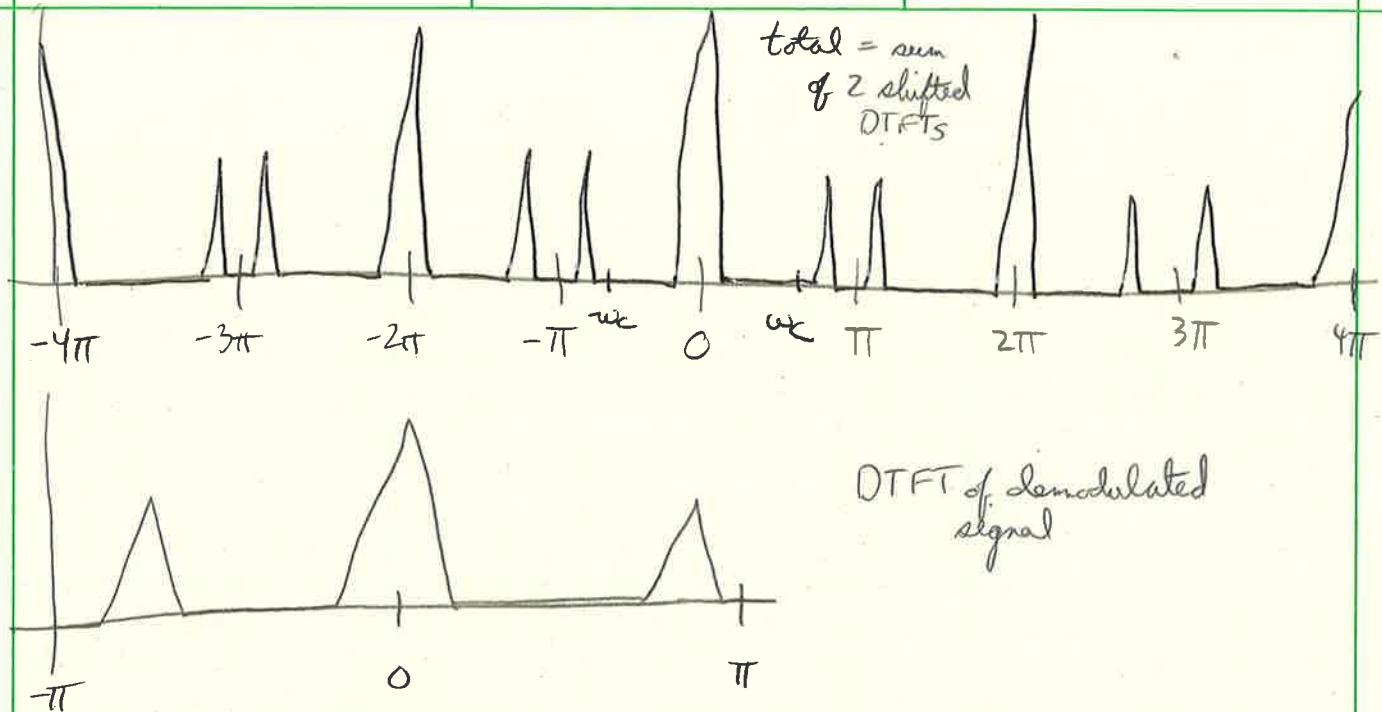
- Demodulation in the frequency domain

$$\text{DTFT}\{y[n]\} = \text{DTFT}\{x[n] \cos(\omega_c n)\}$$

$$\text{DTFT}\{x[n] \cos(\omega_c n)\}$$

$$\text{DTFT}\{x[n]\}$$





- we recovered the baseband signal exactly ..
- but we have some spurious high-frequency components
- in the next Module we will learn how to get rid of them!

3.7b Tuning a guitar

- Problem (abstraction):

- reference sinusoid at frequency ω_0
- tunable sinusoid at frequency ω
- make $\omega = \omega_0$ "by ear"

- The procedure

1. bring ω close to ω_0 (easy)
2. when $\omega \approx \omega_0$ play both sinusoids together
3. trigonometry comes to the rescue

$$\begin{aligned} x[n] &= \cos(\omega_0 n) + \cos(\omega n) \\ &= 2 \cos\left(\frac{\omega_0 + \omega}{2} n\right) \cos\left(\frac{\omega_0 - \omega}{2} n\right) \\ &\approx 2 \cos(\Delta\omega n) \cos(\omega_0 n) \end{aligned}$$

- Let's see what's happening

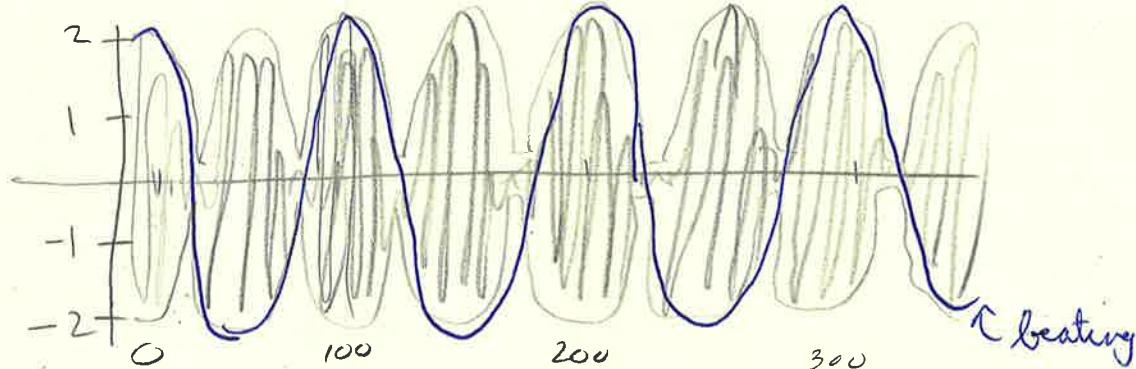
$$x[n] \propto 2 \cos(\Delta\omega n) \cos(\omega_0 n)$$

"error" signal

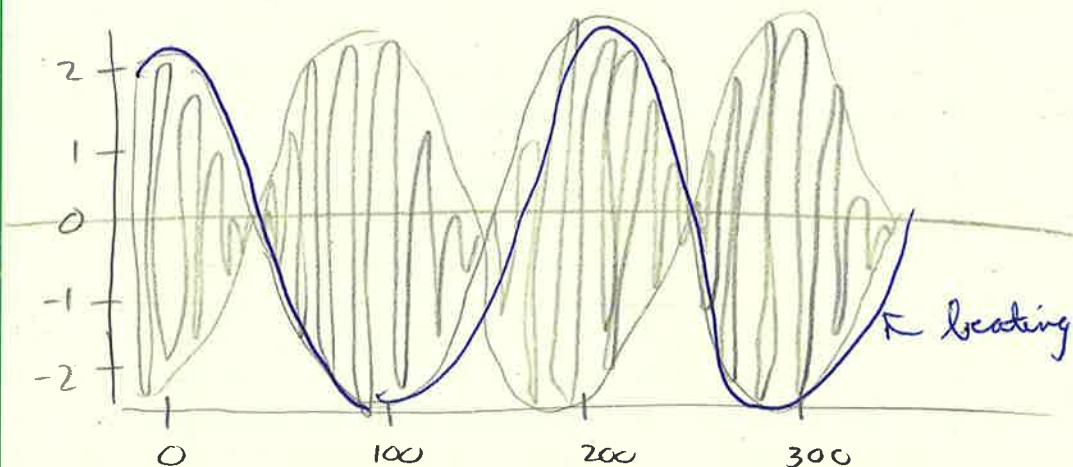
modulation at ω_0

when $\omega \approx \omega_0$, the error signal is too low to be heard; modulation brings it up to hearing range and we perceive it as amplitude oscillations of the carrier frequency

- In the time domain...



$$\omega_0 = 2\pi(0.2), \omega = 2\pi(0.22), \Delta\omega = 2\pi(0.0100)$$



$$\omega_0 = 2\pi(0.2), \omega = 2\pi(0.21), \Delta\omega = 2\pi(0.0050)$$

- A Detour on Western Musical Conventions

- Each note has a unique frequency, $A_3 = 220 \text{ Hz}$, $A_4 = 440 \text{ Hz}$, $A_5 = 880 \text{ Hz}$
- An octave corresponds to doubling/halving frequency
- Octaves are separated into 12 evenly spaced half-tones

$$f_h = f_0 \cdot 2^{h/12}$$

reference note

3.8 Relationship between transforms

- Overview

- DFT, DFS, DTFT
- DTFT of periodic sequences
- DTFT of finite-support sequences
- Zero padding

- Transforms

- DFT, DFS: change of basis in \mathbb{C}^N
- DTFT: "formal" change of basis in $l_2(\mathbb{R})$
- basis vectors are "building blocks" for any signal
- DFT: numerical algorithm (computable)
- DTFT: mathematical tool (proofs)

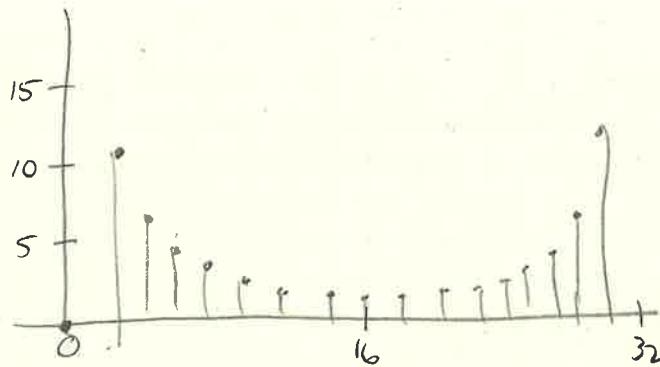
- Embedding finite-length signals

- N -tap signal $x[n]$
- natural spectral representation: DFT $X[k]$
- two ways to embed $x[n]$ into an infinite sequence:
 - periodic extension: $\tilde{x}[n] = x[n \bmod N]$
 - finite support extension: $\tilde{x}[n] = \begin{cases} x[n], & 0 \leq n < N \\ 0, & \text{otherwise} \end{cases}$
- how does $X[k]$ relate to the DTFT of the embedded signals?

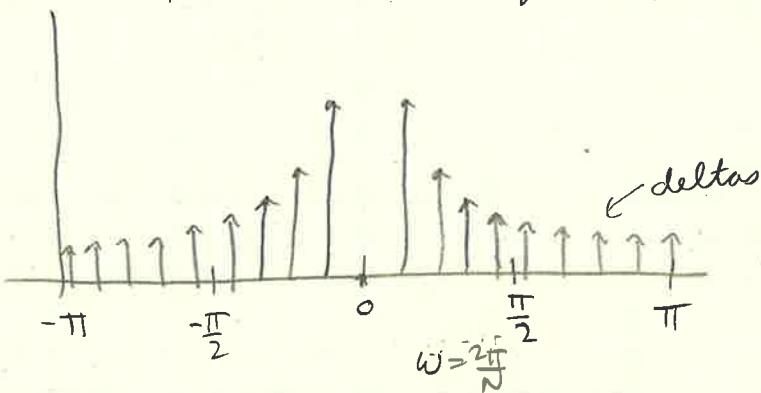
- DTFT of periodic signals

$$\begin{aligned}
 \tilde{x}[n] &= x[n \bmod N] \\
 \tilde{X}(e^{j\omega}) &= \sum_{n=-\infty}^{\infty} \tilde{x}[n] e^{-j\omega n} \\
 &= \sum_{n=-\infty}^{\infty} \left(\frac{1}{N} \sum_{k=0}^{N-1} X[k] e^{j\frac{2\pi}{N} nk} \right) e^{-j\omega n} \\
 &= \frac{1}{N} \sum_{k=0}^{N-1} X[k] \left(\sum_{n=-\infty}^{\infty} e^{j\frac{2\pi}{N} nk} e^{-j\omega n} \right) \\
 &\quad \left(\sum_{n=-\infty}^{\infty} e^{j\frac{2\pi}{N} nk} e^{-j\omega n} = \text{DTFT} \left\{ e^{j\frac{2\pi}{N} nk} \right\} \right. \\
 &\quad \left. = \tilde{g}\left(\omega - \frac{2\pi}{N} k\right) \right) \\
 \tilde{X}(e^{j\omega}) &= \frac{1}{N} \sum_{k=0}^{N-1} X[k] \tilde{g}\left(\omega - \frac{2\pi}{N} k\right)
 \end{aligned}$$

- DFT of 32-tap sawtooth



- DTFT of periodic extension of 32-tap sawtooth



- DTFT of finite-support signals

$$\bar{x}[n] = \begin{cases} x[n], & 0 \leq n < N \\ 0, & \text{otherwise} \end{cases}$$

$$\bar{X}(e^{j\omega}) = \sum_{n=-\infty}^{\infty} \bar{x}[n] e^{-j\omega n} = \sum_{n=0}^{N-1} x[n] e^{-j\omega n}$$

$$= \sum_{n=0}^{N-1} \left(\frac{1}{N} \sum_{k=0}^{N-1} x[k] e^{j\frac{2\pi}{N} nk} \right) e^{-j\omega n}$$

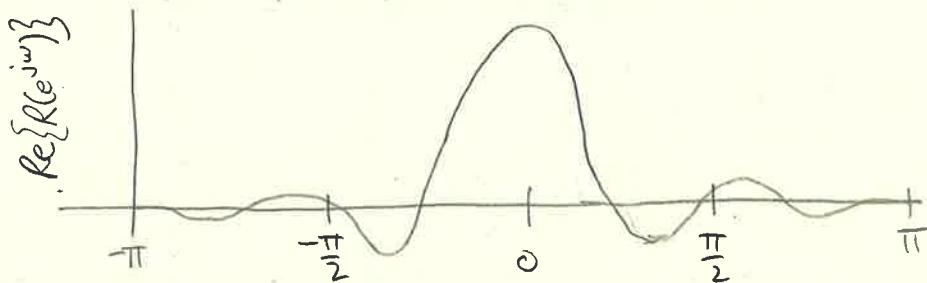
$$= \frac{1}{N} \sum_{k=0}^{N-1} X[k] \left(\sum_{n=0}^{N-1} e^{-j(\omega - \frac{2\pi}{N} k)n} \right)$$

$$\left[\sum_{n=0}^{N-1} e^{-j(\omega - \frac{2\pi}{N} k)n} = \bar{R}(e^{j(\omega - \frac{2\pi}{N} k)}) \right], \text{where } \bar{R}(e^{j\omega}) \text{ is the DTFT of } \bar{x}[n], \text{ the interval indicator signal: } \bar{x}[n] = \begin{cases} 1, & 0 \leq n < N \\ 0, & \text{otherwise} \end{cases}$$

- DTFT of interval signal

$$\begin{aligned}\bar{R}(e^{j\omega}) &= \sum_{n=0}^{N-1} e^{-j\omega n} = \frac{1 - e^{-j\omega N}}{1 - e^{-j\omega}} \\ &= e^{-j\frac{\omega N}{2}} \left[e^{j\frac{\omega N}{2}} - e^{-j\frac{\omega N}{2}} \right] \\ &\quad \overline{e^{-j\frac{\omega}{2}} \left[e^{j\frac{\omega}{2}} - e^{-j\frac{\omega}{2}} \right]} \\ &= \frac{\sin(\frac{\omega}{2}N)}{\sin(\frac{\omega}{2})} e^{-j\frac{\omega}{2}(N-1)}\end{aligned}$$

- DTFT of interval signal ($N=9$)

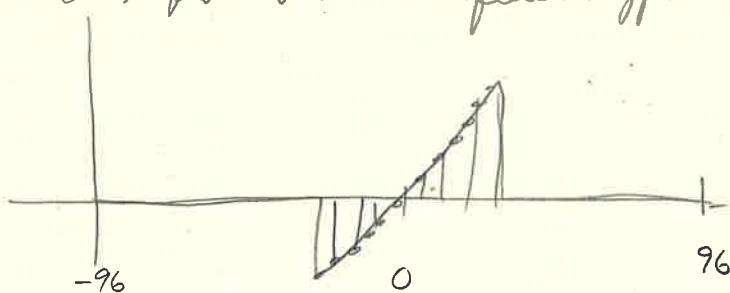


- DTFT of finite-support signals cont'd

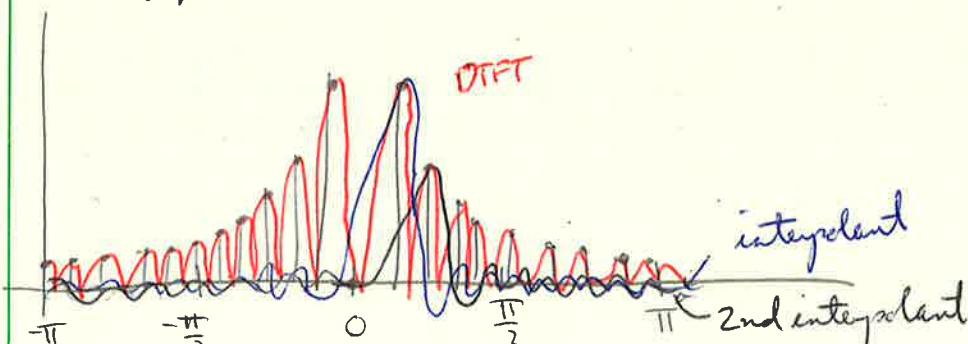
$$\bar{X}(e^{j\omega}) = \sum_{k=0}^{N-1} X[k] \Lambda\left(\omega - \frac{2\pi}{N}k\right), \text{ where } \Lambda(\omega) = \frac{1}{N} \bar{R}(e^{j\omega}).$$

smooth interpolation of DFT values.

- 32 tap sawtooth with finite support extension



- DTFT of finite support extension (sketch)



- About zero-padding

When computing the DFT numerically, one may "pad" the data vector with zeros to obtain "nicer" plots.

$$\begin{aligned}
 X_M[h] &= \sum_{n=0}^{M-1} x[n] e^{-j \frac{2\pi}{M} nh} = \sum_{n=0}^{N-1} x[n] e^{-j \frac{2\pi}{M} nh} \\
 &\stackrel{\text{extension of } x[n]}{=} \sum_{n=0}^{N-1} \left(\frac{1}{N} \sum_{k=0}^{N-1} X_N[k] e^{j \frac{2\pi}{N} nk} \right) e^{-j \frac{2\pi}{M} nh} \\
 &= \frac{1}{N} \sum_{k=0}^{N-1} X_N[k] \left(\sum_{n=0}^{N-1} e^{-j \left(\frac{2\pi}{M} h - \frac{2\pi}{N} k \right) n} \right) \\
 &= \overline{X(e^{j\omega})} \Big|_{\omega = \frac{2\pi}{M} h}
 \end{aligned}$$

- zero padding does not add information
- a zero-padded DFT is simply a sampled DTFT of the finite-support extension
- DFT of 32-top sawtooth, zero-padded

