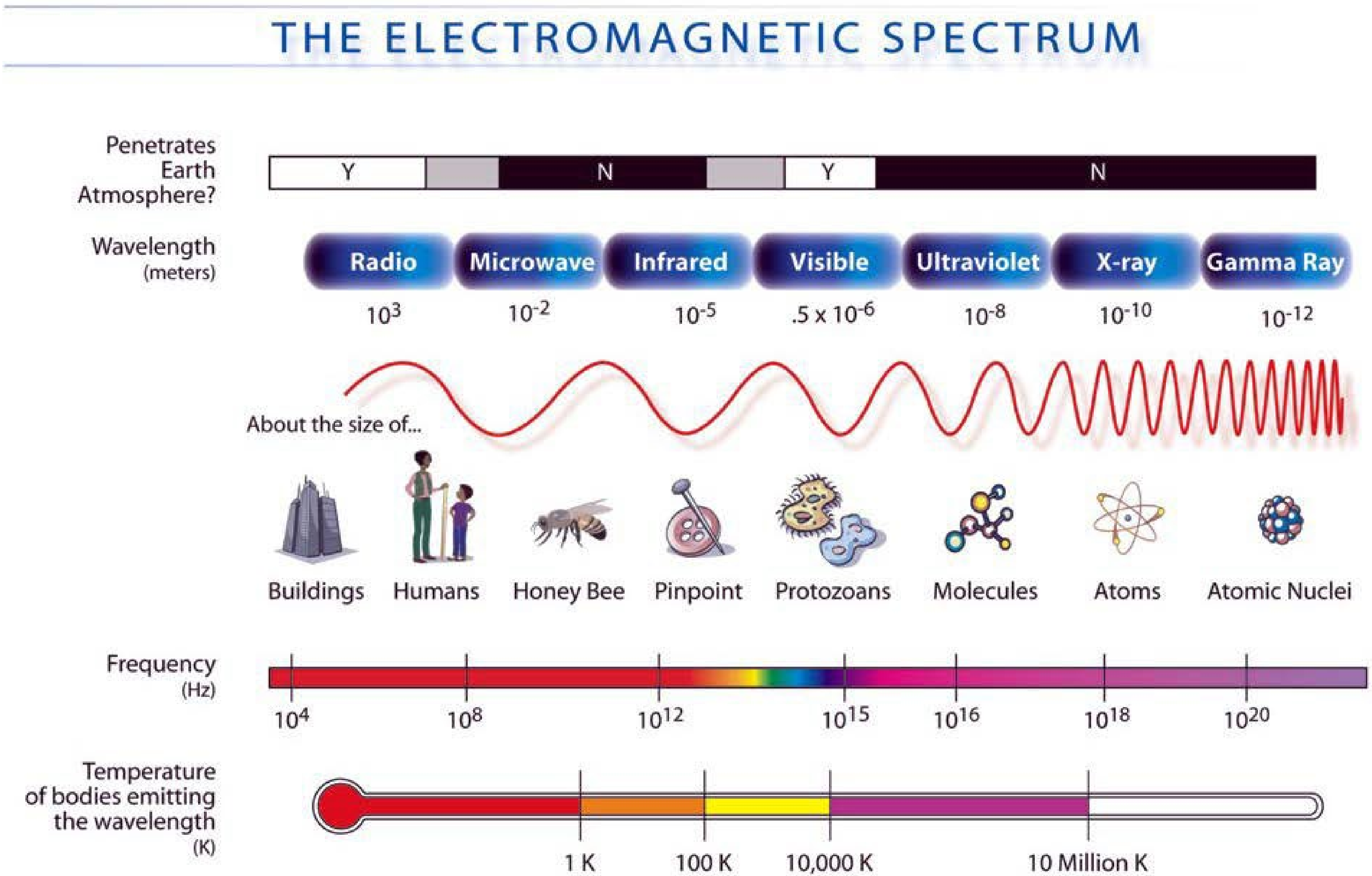
# Chapter 1

## Introduction to microwave device/circuit CAD

# SPECIFICITY OF RF/MICROWAVE FREQUENCIES



# SPECIFICITY OF RF/MICROWAVE FREQUENCIES

| Frequency | | Wavelength | |
|---|---|---|---|
| 1Hz | → | 300 000 000 m | |
| 300 Hz | → | 1 000 000 m | |
| 1 kHz | → | 300 000 m | |
| 300 kHz | → | 1000 m | |
| 1 MHz | → | 300 m | |
| 3 MHz | → | 100 m | |
| 10 MHz | → | 30 m | Decameter band |
| 30 MHz | → | 10 m | |
| 100 MHz | → | 3 m | Meter Band |
| 300 MHz | → | 1m | |
| 1 GHz | → | 30 cm | Decimeter Band |
| 3 GHz | → | 10 cm | |
| 10 GHz | → | 3 cm | Centimeter Band |
| 30 GHz | → | 1 cm | |
| 100 GHz | → | 3 mm | Millimeter Band |
| 300 GHz | → | 1 mm | |

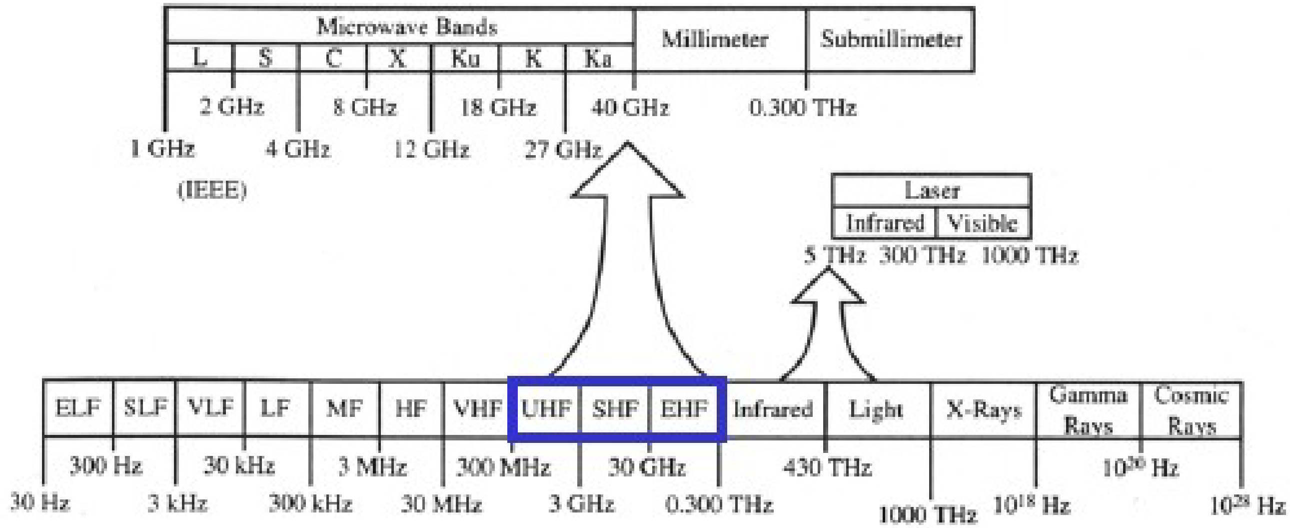
Relation between $f$ and $\lambda$,

highlighting the microwave bands

(formally from 300 MHz to 300 GHz)

MICROWAVES

# Frequency spectrum and standard bands




## CCIR(Inter. Radio Consult. Comm.)

| Band | Frequency Range | Band Designation |
|---|---|---|
| 2 | 30–300 Hz | ELF (extremely low frequency) |
| 3 | 0.3–3 kHz | SLF/VF (voice frequency) |
| 4 | 3–30 kHz | VLF (very low frequency) |
| 5 | 30–300 kHz | LF (low frequency) |
| 6 | 0.3–3 MHz | MF (medium frequency) |
| 7 | 3–30 MHz | HF (high frequency) |
| 8 | 30–300 MHz | VHF (very high frequency) |
| 9 | 0.3–3 GHz | UHF (ultra high frequency) |
| 10 | 3–30 GHz | SHF (super high frequency) |
| 11 | 30–300 GHz | EHF (extremely high frequency) |
| 12 | 0.3–3 THz | Infrared light |
| 13 | 3–30 THz | Infrared light |
| 14 | 30–300 THz | Infrared light |
| 15 | 0.3–3 PHz | Visible light |
| 16 | 3–30 PHz | Ultraviolet light |
| 17 | 30–300 PHz | X-rays |
| 18 | 0.3–3 EHz | Gamma rays |
| 19 | 3–30 EHz | Cosmic rays |

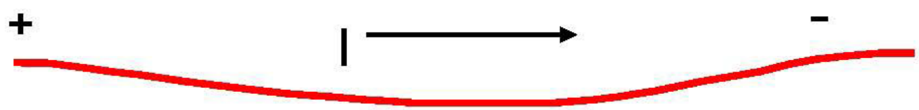
## IEEE Standard

| Band | Frequency Range, (GHz) |
|---|---|
| HF | 0.003–0.030 |
| VHF | 0.030–0.300 |
| UHF | 0.300–1.00 |
| L | 1.00–2.00 |
| S | 2.00–4.00 |
| C | 4.00–8.00 |
| X | 8.00–12.0 |
| Ku | 12.0–18.0 |
| K | 18.0–27.0 |
| Ka | 27.0–40.0 |
| Millimeter | 40.0–300.0 |
| Submillimeter | greater than 300 |

## New US Military Bands

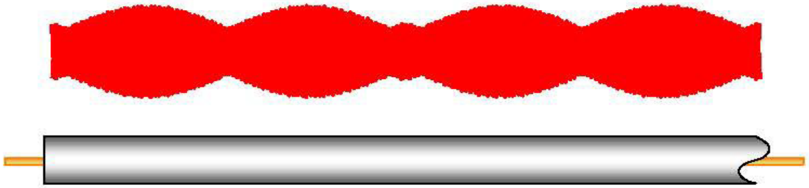
| Frequency Band | Frequency Range, (GHz) |
|---|---|
| A | 0.10–0.25 |
| B | 0.25–0.5 |
| C | 0.5–1.0 |
| D | 1.0–2.0 |
| E | 2.0–3.0 |
| F | 3.0–4.0 |
| G | 4.0–6.0 |
| H | 6.0–8.0 |
| I | 8.0–10.0 |
| J | 10.0–20.0 |
| K | 20.0–40.0 |
| L | 40.0–60.0 |
| M | 60.0–100.0 |
| N | 100.0–140.0 |

# RELATION BETWEEN WAVELENGTH AND PHYSICAL LENGTH




## *Low frequencies*

- wavelengths >> wire length
- current (I) travels down wires easily for efficient power transmission
- measured voltage and current not dependent on position along wire



## *High frequencies*

- wavelength ≈ or << length of transmission medium
- need transmission lines for efficient power transmission
- matching to characteristic impedance (Zo) is very important for low reflection and maximum power transfer
- measured envelope voltage dependent on position along line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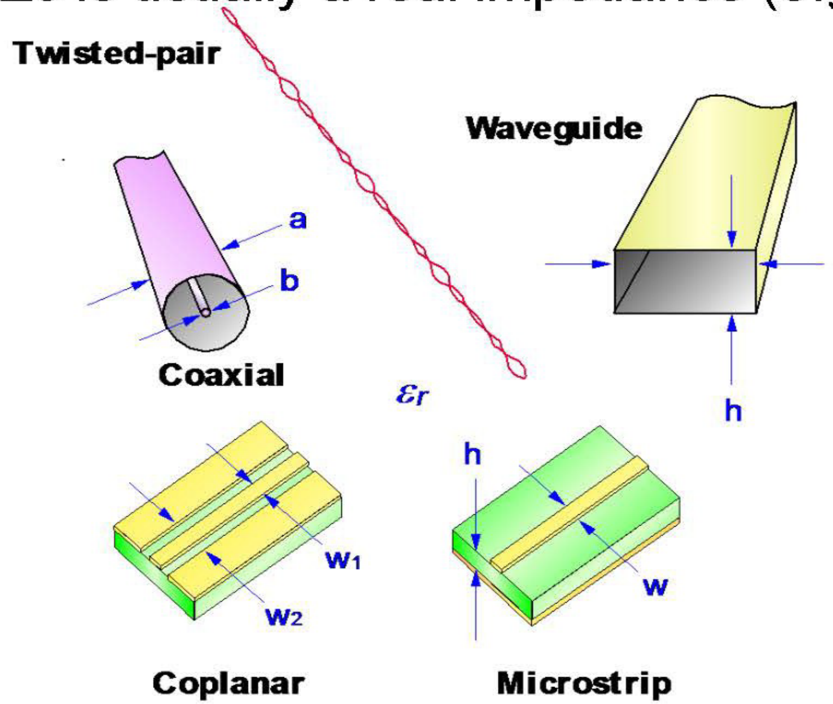
Question : **Why** referencing to a **characteristic impedance $Z_o$**?

# Transmission line Zo

- Zo determines relationship between voltage and current waves
- Zo is a function of physical dimensions and $\varepsilon_r$
- Zo is usually a real impedance (e.g. 50 or 75 ohms)



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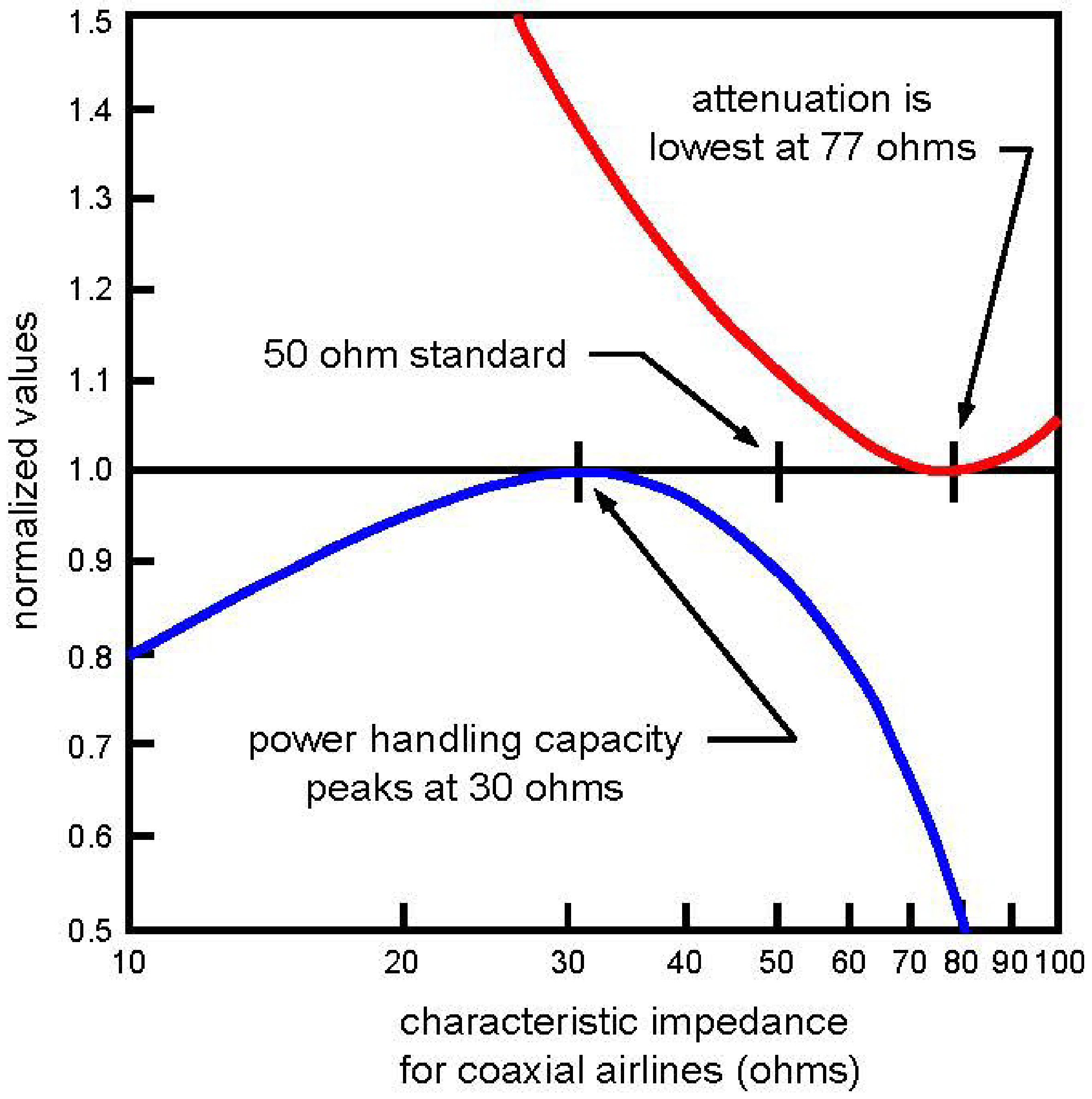
Question : **Why such particular values: 50 Ω or 75 Ω?**

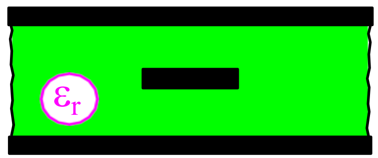


1.5
1.4
1.3
1.2
1.1
1.0
0.9
0.8
0.7
0.6
0.5
normalized values
attenuation is
lowest at 77 ohms
50 ohm standard
power handling capacity
peaks at 30 ohms
10
20
30
40
50
60
70
80
90
100
characteristic impedance
for coaxial airlines (ohms)

# RF/MICROWAVE TRANSMISSION LINES

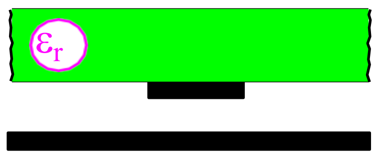
Waveguides and planar lines are the most widely used transmission supports in the cm/mm bands.

Strip-line



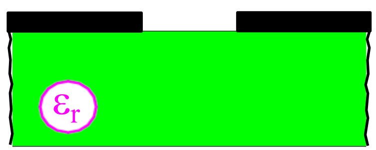

Microstrip


Inverted Microstrip

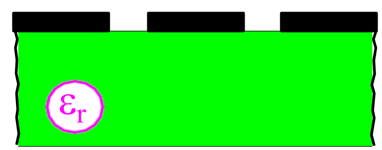



Suspended Microstrip


Slotline




Coplanar Waveguide




Fin Line

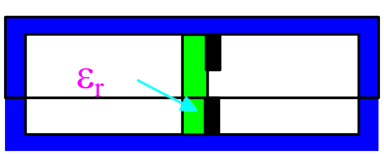
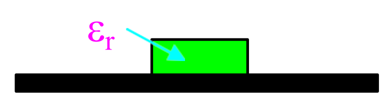



Image Line

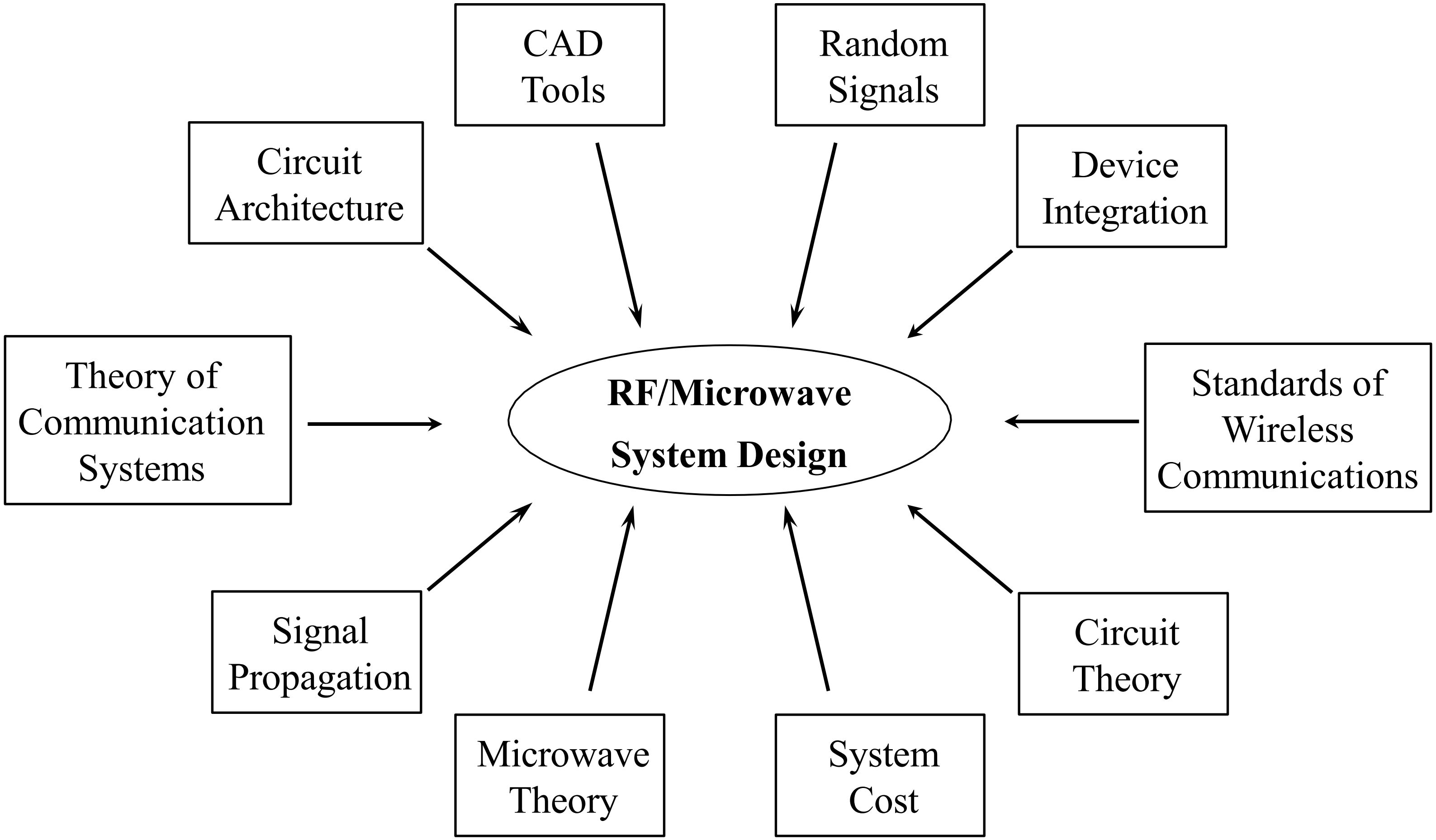


# MICROWAVE COMPUTER-AIDED DESIGN

Different aspects to consider in RF/microwave design:

- Theoretical/Technical issues

but also …



# ELEMENTS OF CIRCUIT THEORY

## Fundamental concepts

In electronics, a fundamental concept is that *all* electronic circuits/systems are nonlinear.

*Now, in <u>a</u> <u>specific</u> design, nonlinearities should be minimized or maximized?*

Mixers: mixing is possible *only if* the circuit *exploits* nonlinearities, and then, it is desirable to **maximize** them.

Small-signal amplifiers: nonlinearities degrade the system performance and then should be **minimized**.

***Small-signal*** ??

# ELEMENTS OF CIRCUIT THEORY

## Fundamental concepts

From a **general point of view**, we could say that (from Wikipedia)

- *Small-signal modeling* is a common analysis method used to describe nonlinear devices in terms of linear equations.

- *Large-signal modeling* is a common analysis method used to describe nonlinear devices in terms of the underlying nonlinear equations from the **circuit design point of view**, we can say that

- *Small signal*: A small-signal amplifier is a *linear circuit* in which the input or output signal strength does not affect the circuit's electrical properties.

- *Large signal*: *Opposite of small signal*. For instance, a small-signal amplifier can be designed easily using S-parameter while large-signal amplifier not.

while from the **signal point of view**, we have

- *Small signal*: small time-varying signals carried over a constant bias.

  The signal *is small* relative to the nonlinearity of the device.

- *Large signal*: large time-varying signals carried over a constant bias.

  The signal *is large* enough that the nonlinearity of the device cannot be ignored.

As for the **concept of linear and nonlinear**, we can read that (from Wikipedia)

- Linear circuit:

  the output is a linear transform of the input.

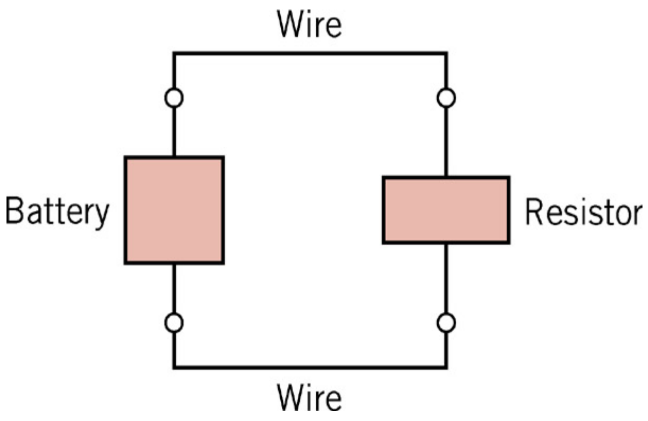
  <u>Superposition principle can be applied</u>.

- Nonlinear circuit:

  <u>does not satisfy the superposition principle</u>.

**Superposition?**

# SUPERPOSITION



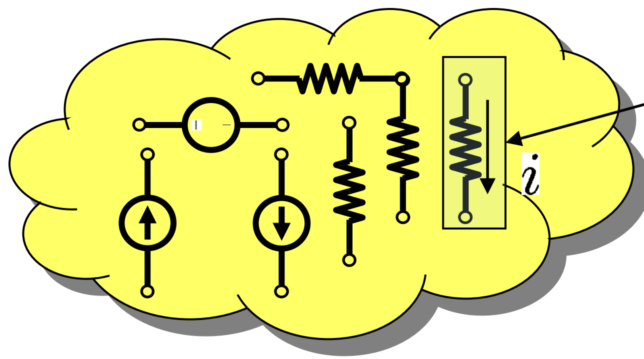

Let the input of a given circuit be $v_i$ and the output $v_o$

If the circuit is linear, the superposition theorem can be applied:

- for a given input $v_{i1}(t)$, the output will be $v_{o1}(t)$
- for a given input $v_{i2}(t)$, the output will be $v_{o2}(t)$
- for a given input $v_{i1}(t) + v_{i2}(t)$, the output will be $v_{o1}(t) + v_{o2}(t)$
- for a given input $a^*v_{i1}(t) + b^*v_{i2}(t)$, the output will be $a^*v_{o1}(t) + b^*v_{o2}(t)$

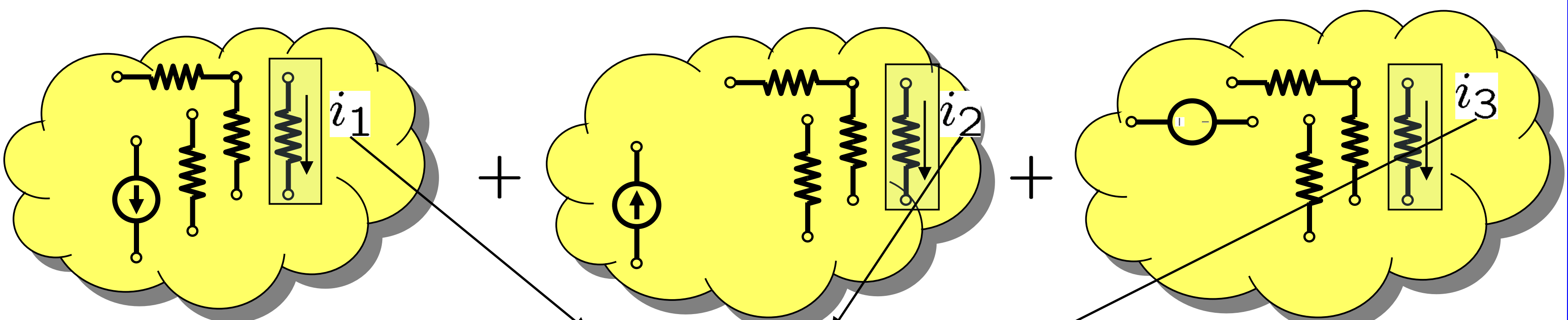
# SUPERPOSITION APPLIED TO CIRCUIT THEORY

Given a general circuit




Assume that we need the current $\boldsymbol{i}$ in one resistor, e.g., this one

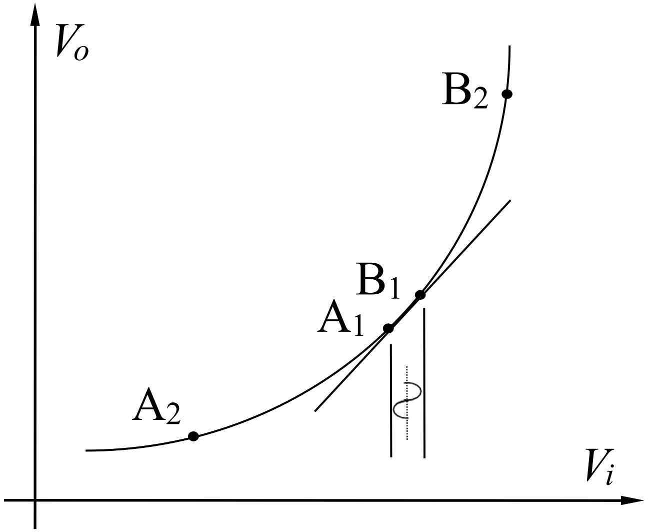
We can find any electrical quantity such as the current in any resistor, ***through*** finding the current due to each source individually while leaving out or ***deactivating*** the other sources.




$$i = \boxed{i_1} + \boxed{i_2} + \boxed{i_3}$$

## LINEARITY AND NONLINEARITY IN CIRCUIT THEORY

If an input signal $V_i$ imposes a small excursion around the DC point (Q point), the curve $A_1B_1$ is almost linear




$$V_o = G\,V_i$$

The input-output relationship for the nonlinear case will be

$$V_o = A V_i + B\,V_i^2 + C\,V_i^3 + \ldots$$

Each m$^{th}$ term " $(V_i)^m$ " generates the harmonic $mf_i$ of the fundamental input frequency $f_i$.

# LINEARITY AND NONLINEARITY IN CIRCUIT THEORY

*Commensurate and non-commensurate frequencies*

If the output frequency of a given circuit is the combination of the harmonics of two input frequencies, such as

$$\omega_{output} = \omega_{m,\,n} = m * \omega_1 + n * \omega_2$$

with $m, n = \ldots -2, -1, 0, 1, 2 \ldots$

This output frequency (also noted $\omega_{m,n}$) is called the *mixing frequency* and its output voltage component is a *mixing product*.

The sum of the absolute values of $m$ and $n$ ( i.e., $|m| + |n|$ ) is the order of the mixing product.

For the $\omega_{m,n}$ to be distinct, $\omega_1$ and $\omega_2$ must be ***non-commensurate***. A pair of frequencies is said to be commensurate if their ratio is a rational number.

# LINEARITY AND NONLINEARITY IN CIRCUIT THEORY

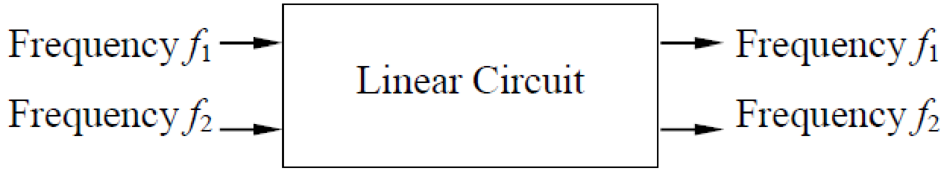
IF

the response of the circuit to $N$ linear excitations is equal to the sum of their responses,

*The circuit is linear*

In other words, the response of a linear circuit includes only the frequencies present in the excitation waveforms.

***A linear circuit does not generate new frequencies*.**



# LINEARITY AND NONLINEARITY IN CIRCUIT THEORY

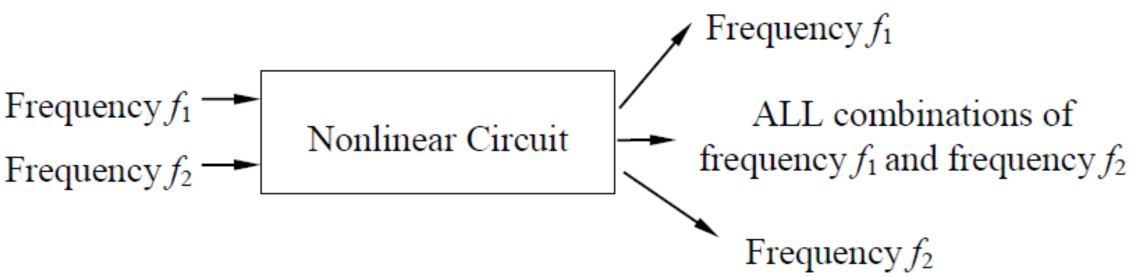
IF

the response of the circuit to *N* linear excitations includes the excitation frequencies as well as all possible combinations of them ( $m*\omega_1 + n*\omega_2$ )

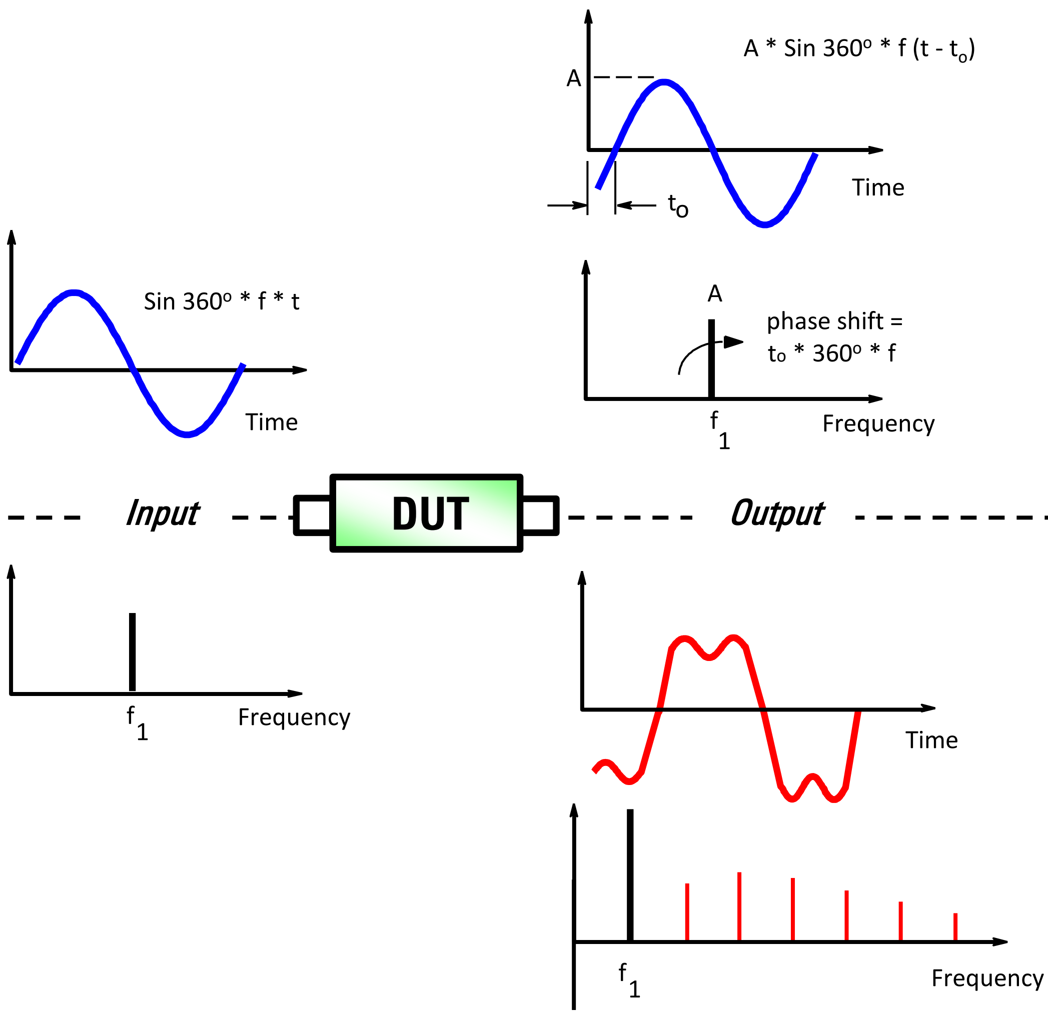
*The circuit is nonlinear*

In other words, the response of a nonlinear circuit includes the frequencies present in the excitation waveforms and their harmonics.

***A nonlinear circuit generates new frequencies***.



# Linear Versus Nonlinear Behaviour




## *Linear behaviour:*

- input and output frequencies are the same (no additional frequencies created)
- output frequency only undergoes magnitude and phase change

## *Nonlinear behaviour:*

- output frequency may undergo frequency shift (e.g. with mixers)
- additional frequencies created (harmonics, intermodulation)



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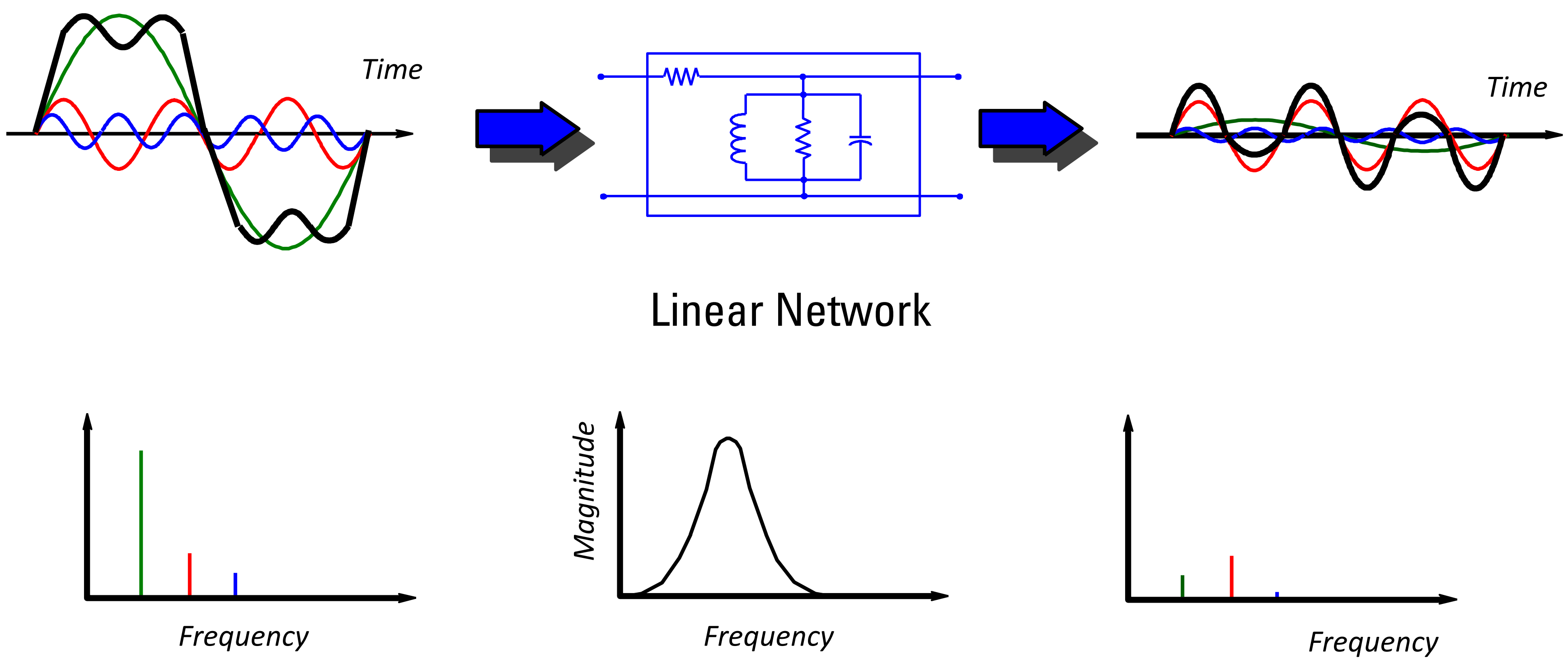
# Magnitude Variation with Frequency

F(t) = sin ωt + 1/3 sin 3 ωt + 1/5 sin 5 ωt



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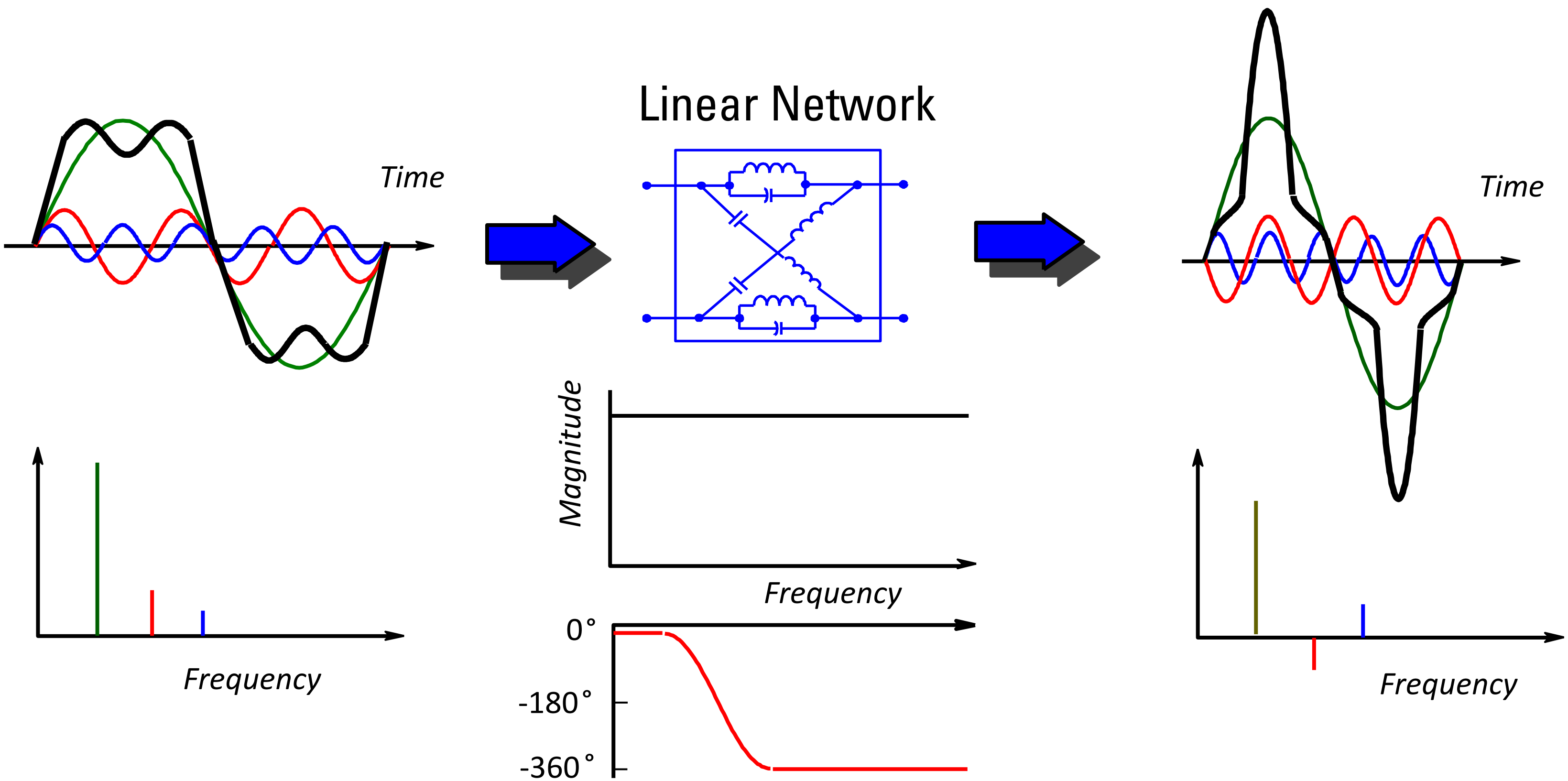
# Phase Variation with Frequency

$F(t) = \sin \omega t + 1/3 \sin 3\omega t + 1/5 \sin 5\omega t$





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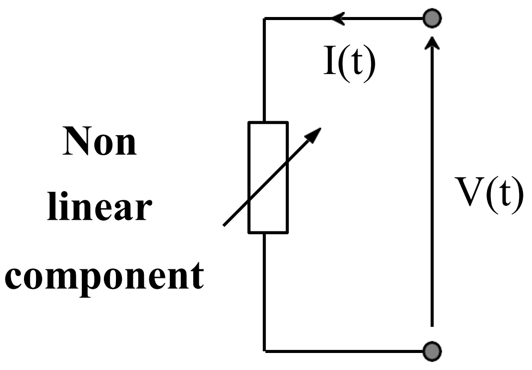
# REPRESENTATION OF A NONLINEAR CIRCUIT

The physical representation of a nonlinear system is in the time domain.



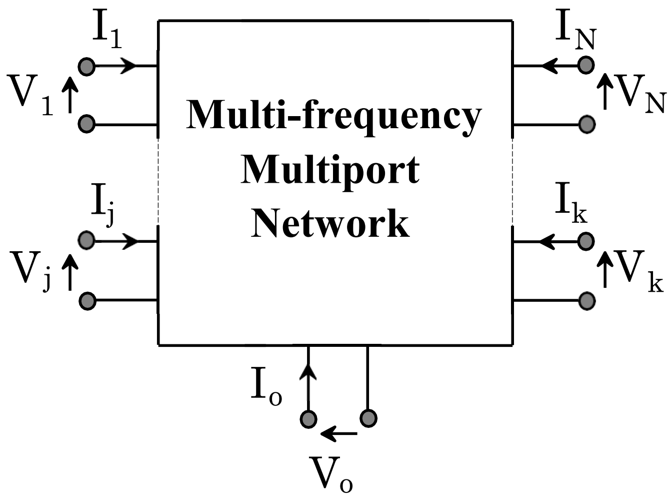

However, as a nonlinear behaviour generates harmonics,

it is common to represent a nonlinear network in the frequency domain:

a multifrequency multi-port network where each $k$th port is related to the $k$th harmonic of the excitation input frequency $f$.

$\Downarrow$

A current $I_k$ and a voltage $V_k$ present in port $k$ are the Fourier coefficients of $v(t)$ and $i(t)$, respectively, for the harmonic " $kf$ ". A port "0" can be eventually added for the zero frequency (DC coefficients).




The multi-frequency multi-port network is **<u>LINEAR</u>** !!!

In the above time-domain circuit, the sign ↗ refers to a nonlinear element.

Question :

Why it is important to have **a linear** multi-frequency network in frequency domain?

Because:

- All circuit simulators are based on solving the voltage-current relationships in a circuit
- This implies solving relations like V = Z I

which …

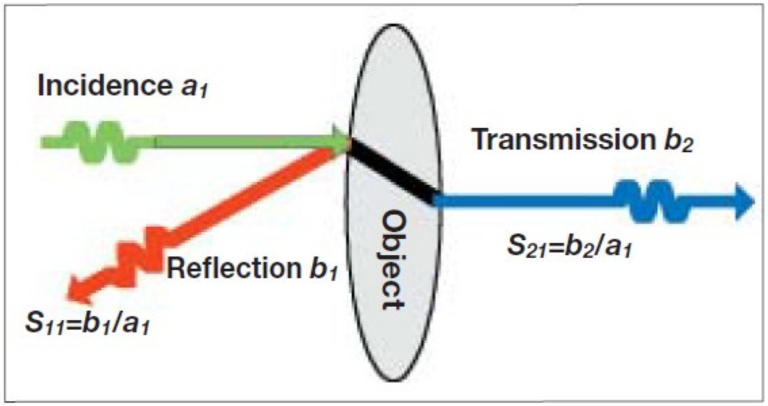
… are linear equations !!

# SCATTERING PARAMETERS OR S-PARAMETERS

## What are scattering parameters?

The scattering parameters are fixed properties of the (linear) circuit, which describe how the energy couples between each pair of ports or transmission lines connected to the circuit. Formally, S-parameters can be defined for any collection of **linear** electronic components, whether or not the wave view of the power flow in the circuit is necessary.

Analogy with optics :



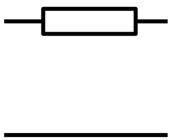
# SCATTERING PARAMETERS OR S-PARAMETERS

## Why scattering parameters?

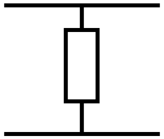
*I* - Impedance and admittance matrices are not defined for **all** circuits:

**Series impedance:** Using *Z*-parameters, serial impedance is defined as

$$Z_{11} = \left.\frac{V_1}{I_1}\right|_{I_2=0} \qquad \text{But} \qquad I_2 = 0 \rightarrow I_1 = 0 \rightarrow Z_{11} = ????$$



**Parallel admittance:** Using *Y*-parameters, parallel admittance is defined as



$$Y_{11} = \left.\frac{I_1}{V_1}\right|_{V_2=0} \qquad \text{But} \qquad V_2 = 0 \rightarrow V_1 = 0 \rightarrow Y_{11} = ????$$

$\Rightarrow$ The usual *Z*- or *Y*-parameters are not **universal**.

# SCATTERING PARAMETERS OR S-PARAMETERS

*II.- In HF, both voltages and currents cannot be measured in a direct manner.*

In high frequencies, the quantities that are directly measurable, by means of a small probe used to sample the relative field strength, are the <u>standing wave ratio, location of a field minimum and power</u>.

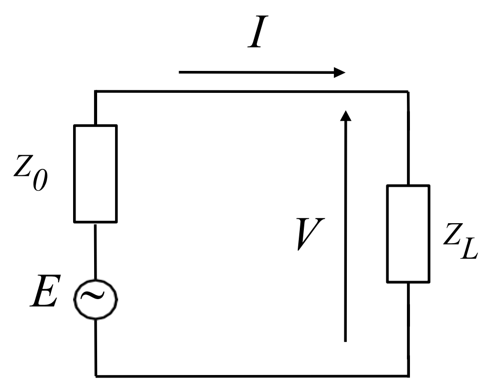
In fact, at high frequencies, we have a **non-uniform distribution** of the current along each branch and a **non-uniform distribution** of the voltage along each wire.

*III.- Short and open circuits cannot be assured accurate in high frequencies*

Therefore, designers have to use parameters that are linked to **power** and **reflection-transmission coefficients**

# SCATTERING PARAMETERS OR S-PARAMETERS

## How to get the S-parameters?




Let us consider a generator (a voltage source $E$ with an internal impedance $Z$ equal to the characteristic impedance $Z_o$).

If this one-port circuit is loaded by an impedance $Z_L$, the complex current and voltage can be defined as

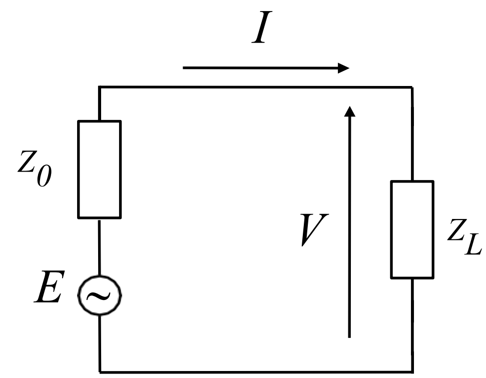
$$I = \frac{E}{Z_o + Z_L} \qquad V = \frac{E Z_L}{Z_o + Z_L}$$

For maximum transfer power (matching), the impedance load is equal to the complex conjugate of the generator internal impedance. Therefore, we get the **incident** quantities

$$I = I_i = \frac{E}{Z_o + Z_o^*} \qquad V = V_i = \frac{E Z_o^*}{Z_o + Z_o^*}$$

## SCATTERING PARAMETERS OR S-PARAMETERS

### How to get the S-parameters?




Similarly, we can define the **reflection** quantities as

$$V_r = V - V_i \qquad I_r = I_i - I$$

$$V_r = \frac{Z_o}{Z_o^*}\frac{Z_L - Z_o^*}{Z_L + Z_o} \cdot V_i \qquad I_r = \frac{Z_L - Z_o^*}{Z_L + Z_o^*} \cdot I_i$$

These expressions allow defining the voltage reflection coefficient $\Gamma_V$ and the current coefficient reflection $\Gamma_I$

$$\Gamma_V = \frac{V_r}{V_i} = \frac{Z_o}{Z_o^*}\frac{Z_L - Z_o^*}{Z_L + Z_o} \qquad \Gamma_I = \frac{I_r}{I_i} = \frac{Z_L - Z_o^*}{Z_L + Z_o}$$

# SCATTERING PARAMETERS OR S-PARAMETERS

## How to get the S-parameters?

If the characteristic impedance is $Z_o$, then

$$\Gamma = \Gamma_I = \Gamma_V = \frac{Z_L - Z_o}{Z_L + Z_o} = \frac{z_L - 1}{z_L + 1}$$

where $z_L$ is the normalized impedance of $Z_L$.

From the above equations we have

$$V_r = Z_o I_r \qquad V_i = Z_o^* I_i$$

$$\rightarrow \qquad [\mathbf{V_r}] = [\mathbf{Z_o}][\mathbf{I_r}] \qquad [\mathbf{V_i}] = \lfloor \mathbf{Z_o^*} \rfloor [\mathbf{I_i}]$$

# SCATTERING PARAMETERS OR S-PARAMETERS

## How to get the S-parameters?

Thus, it is possible to define the vector **[a]** of input waves or « *incident waves* » as follows

$$[a]=\left(\frac{[\mathbf{Z_o}]+[\mathbf{Z_o^*}]}{2}\right)^{\frac{1}{2}}[\mathbf{I_i}] \quad \rightarrow \quad \begin{bmatrix} a_1 \\ \vdots \\ a_n \end{bmatrix} = \begin{bmatrix} \sqrt{R_{o1}} & \cdots & 0 \\ \vdots & \ddots & \vdots \\ 0 & \cdots & \sqrt{R_{oN}} \end{bmatrix} \begin{bmatrix} I_{i1} \\ \vdots \\ I_{in} \end{bmatrix}$$

Similarly, we can define a vector **[b]** of output waves or « *reflected waves* » as

$$[\mathbf{b}]=\left(\frac{[\mathbf{Z_o}]+[\mathbf{Z_o^*}]}{2}\right)^{\frac{1}{2}}[\mathbf{I_r}]$$

Finally, the scattering matrix **[S]** of the network should satisfy to

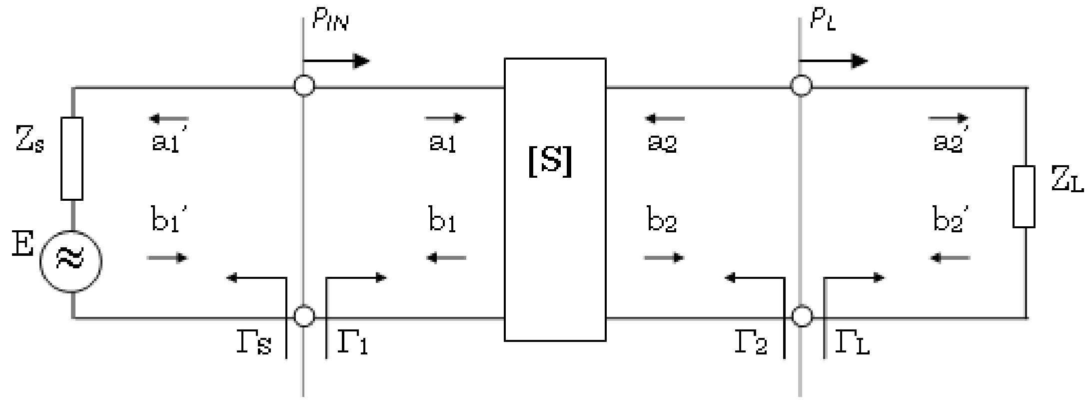
$$[\mathbf{b}] = [\mathbf{S}]\ [\mathbf{a}]$$

# SCATTERING PARAMETERS OR S-PARAMETERS

## Definitions

The $n$ incoming wave complex amplitudes are usually designated by the $n$ complex wave quantities $a_n$ (n-vector **[a]**) and the $n$ outgoing wave complex quantities $b_n$ (n-vector **[b]**)

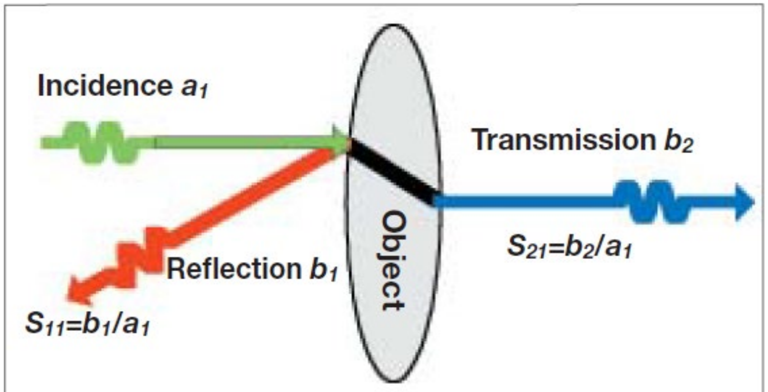
$$\begin{bmatrix} b_1 \\ \vdots \\ b_n \end{bmatrix} = \begin{bmatrix} S_{11} & \cdots & S_{1n} \\ \vdots & \ddots & \vdots \\ S_{n1} & \cdots & S_{nn} \end{bmatrix} \begin{bmatrix} a_1 \\ \vdots \\ a_n \end{bmatrix}$$




For a 2-port, we have

$$b_1 = S_{11}\, a_1 + S_{12}\, a_2$$

$$b_2 = S_{21}\, a_1 + S_{22}\, a_2$$



# SCATTERING PARAMETERS OR S-PARAMETERS

## Relation with power

$|a_1|^2$ Incident power on port 1 = Available power from the source

$|a_2|^2$ Incident power on port 2 = Available power from the load

$|b_2|^2$ Reflected power from port 1 = Available power from the source minus input power

$|b_1|^2$ Reflected power from port 2 = Incident power to the load

$$|S_{11}|^2 \quad \frac{\text{Power reflected from the network input}}{\text{Incident power on the network input}}$$

$$|S_{21}|^2 \quad \frac{\text{Power delivered to the load}}{\text{Power available from the source}} = \textit{Direct Power Gain}$$

$$|S_{22}|^2 \quad \frac{\text{Power reflected from the network output}}{\text{Incident power on the network output}}$$

$$|S_{12}|^2 \quad \frac{\text{Power delivered to the source}}{\text{Power available from the load}} = \textit{Inverse Power Gain}$$

# SCATTERING PARAMETERS OR S-PARAMETERS

## Physical meaning

If the network has no loss/no gain, the output power must equal the input power

$$|S_{11}|^2 + |S_{21}|^2 = 1$$

Let us consider a two-port network loaded by $Z_1$ and $Z_2$ respectively where the input and output port are matched by the characteristic impedances $Z_{o1}$ and $Z_{o2}$ respectively:

$$S_{11} = \left.\frac{b_1}{a_1}\right|_{a_2=0} \qquad S_{12} = \left.\frac{b_1}{a_2}\right|_{a_1=0} \qquad S_{21} = \left.\frac{b_2}{a_1}\right|_{a_2=0} \qquad S_{22} = \left.\frac{b_2}{a_2}\right|_{a_1=0}$$

$S_{11}$ is the input reflection coefficient when the output is matched : $Z_2 = Z_{o2}$

$S_{22}$ is the output reflection coefficient when the input is matched : $Z_1 = Z_{o1}$

$S_{21}$ is the direct transmission coefficient from input to output when the output is matched.

$S_{12}$ is the inverse transmission coefficient from output to input when the input is matched.

# SCATTERING PARAMETERS OR S-PARAMETERS

## Coming back to the definitions

Which kind of information can we get from this set of equations?

This system of equations is … **LINEAR** !!!!!!!!!!!

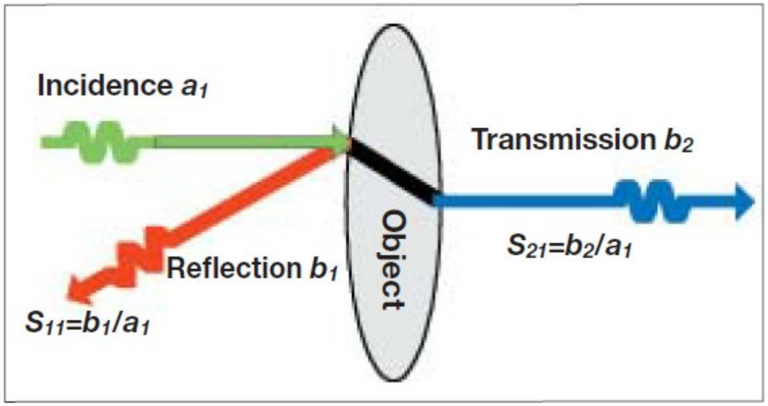
So S-parameters are **linear-based** quantities.

They cannot model nonlinear behaviours !!!!!!!!!!!

For a 2-port, we have

$$b_1 = S_{11}\, a_1 + S_{12}\, a_2$$

$$b_2 = S_{21}\, a_1 + S_{22}\, a_2$$



Question : **What to use instead of S-parameters?**

**X-parameters !**

**X-parameters** are a generalization of S-parameters !

They help characterizing the amplitudes and phase of harmonics generated by nonlinear components under large input power levels.

**X-parameters** are also referred to as the parameters of the Poly-Harmonic Distortion (PHD) nonlinear behavioral model.

**Thank you !**

**End of Chapter 1**