Radio Communication Circuits
(Communication Electronics)

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Goal of the course

- To develop skills in **component-level circuit** construction, as well as **modular interconnection of subsystems**, needed to build **physical communications systems**.

- To use industry-relevant software communications systems simulation methods for the purpose of evaluating overall communication system performance.

- To understand the functionality of analog and digital communications modulation and demodulation by building, testing and analyzing circuits.

- To study and implement essential subsystems such as carrier acquisition and recovery, receiver front-end, and super-heterodyne receiver architectures.

Outline (1)

Chapter 1: Introduction to Communication Systems

Elements of Communication Systems.

Radio Frequency Metrics.

Parallel-Tuned Circuit, Series-Tuned Circuit.

Impedance Matching.

Chapter 2: Radio Frequency (RF) Power Amplifiers

Class C Amplifier.

Class D Amplifier

Chapter 3: Low Noise Amplifier (LNA)

Chapter 4: Frequency Conversion Circuits (Mixers)

Outline (2)

Chapter 5: RF Filters

Chapter 6: Oscillators and Frequency Synthesizers

RF Oscilators, Voltage-Controlled Oscillators (VCO)

Phase-Locked Loops (PLLs) and Applications

Chapter 7: Analog Modulation Circuits

Amplitude Modulation

Frequency Modulation

Phase Modulation

Chapter 8: Digital Modulation Circuits

ASK, FSK, PSK, QPSK, M-ary PSK

DPSK

M-ary QAM

References

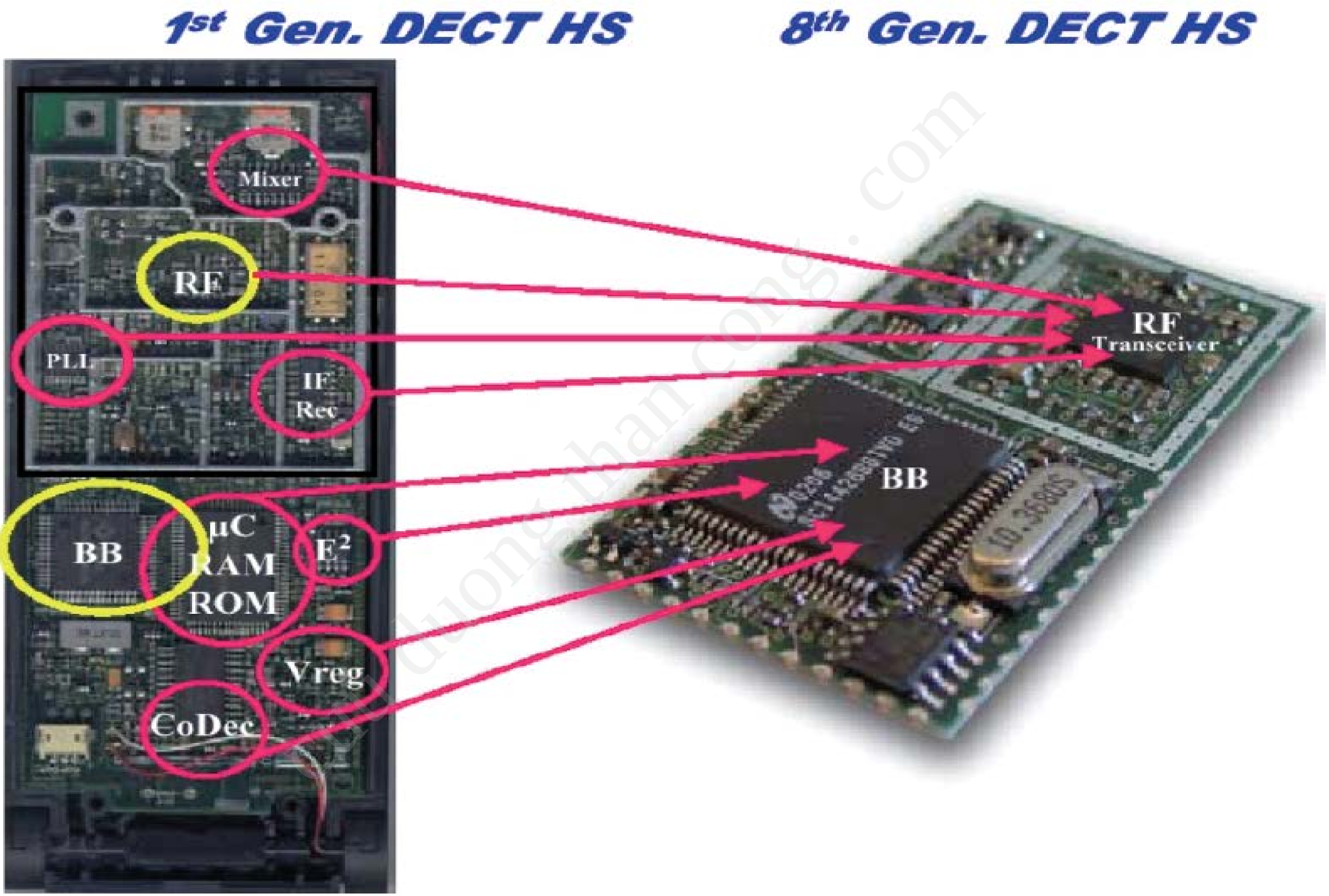
- P. H. Young, *Electronic Communication Techniques*, Fifth Edition, Prentice-Hall, 2004.
- C. W. Sayre, *Complete Wireless Design*, McGraw Hill, 2001.
- J. G. Proakis, M. Salehi and G. Bauch, *Contemporary Communication Systems Using MATLAB and Simulink*, Second Edition, Thomson Engineering, 2004.
- **J. Rogers, C. Plett, *Radio Frequency Integrated Circuit Design*, Artech House, 2003**
- **M. Albulet, *RF Power Amplifier*, Noble Publishing, 2001.**
- **F. Ellinger, *RF Integrated Circuits and Technologies*, Springer Verlag, 2008.**
- M. C. Jeruchim, P. Balaban and K. S. Shanmugan, *Simulation of Communication Systems*, Plenum Press, 1992.
- C. Bowick, *RF Circuit Design*, Newnes Publishing, 1982.
- S. R. Bullock, *Transceiver and System Design for Digital Communications*, Second Edition, Noble Publishing, 2000.
- K. McClaning and T. Vito, *Radio Receiver Design*, Noble Publishing, 2000.
- W. Tomasi, *Advanced Electronic Communications Systems*, Fifth Edition, Prentice-Hall, Inc., 2001.
- S. Haykin, *Communication Systems*, Fourth Edition, John Wiley and Sons, Inc., 2001.

Grading

- 30% for midterm examination.
- 20% for in-class quizzes
- 10% assignments
- 40% for final examination.

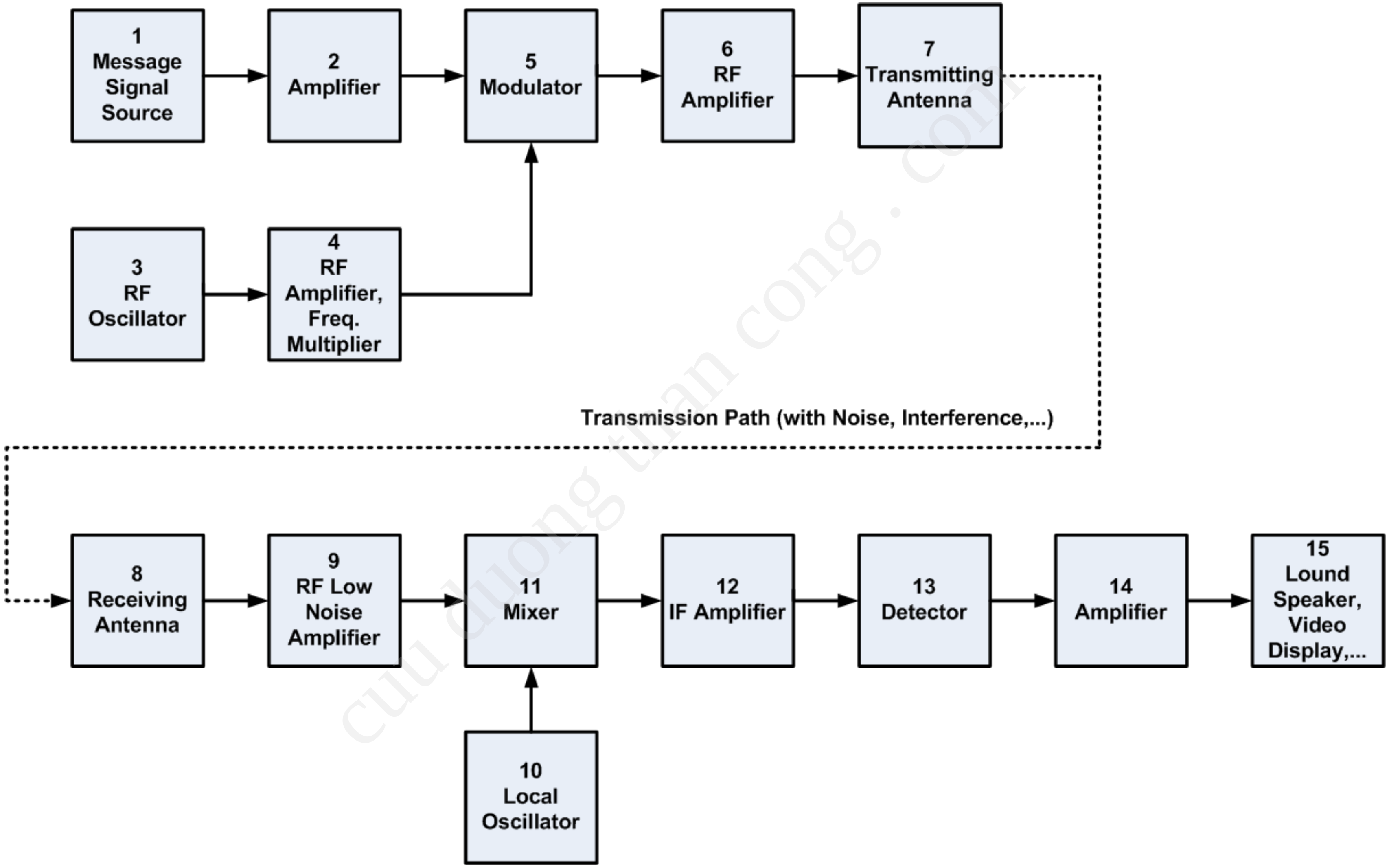
Chapter 1:

Introduction to Communication Systems



1st Gen. DECT HS
8th Gen. DECT HS
Mixer
RF
PLL
IF
Rec
BB
μC
RAM
ROM
E2
Vreg
CoDec
RF
Transceiver
BB

Elements of Communication Systems (1)



Elements of Communication Systems (2)

1. The source of the message signal may be analogue or digital information transformed into an electrical signal.

2. The signal is amplified and often passed through a low-pass filter to limit the bandwidth.

3. The RF oscillator establishes the carrier frequency. Since good frequency stability is required to keep the transmitter on its assigned frequency, the oscillator is often controlled by a quartz crystal (**Chapter 6**).

4. One or more amplifier stages increase the power level of the signal from the oscillator to that needed for input to the modulator.

5. The modulator combines the signal and carrier frequency components to produce one of the varieties of modulated waves (**Chapter 7 (8)**).

Elements of Communication Systems (3)

6. Additional amplification may be required after modulation to bring the power level of the signal to the desired value for input to the antenna (**Chapter 2**).

7. The transmitting antenna converts the RF energy into an electromagnetic wave of the desired polarization. If a single (fixed) receiver is to be reached, the antenna is designed to direct as much of the radiated energy as possible toward the receiving antenna.

8. The receiving antenna may be omni-directional for general service or highly directional for point-to-point communication. The wave propagated from the transmitter induces a small voltage in the receiving antenna. The range of amplitudes of the induced antenna voltage may be from tens of millivolts to less than 1 microvolt, depending upon a wide variety of conditions.

Elements of Communication Systems (4)

9. The RF amplifier stage (RF low noise amplifier) increases the signal power to a level suitable for input to the mixer and it helps to isolate the local oscillator from the antenna. This stage does not have a high degree of frequency selectivity but does serve to reject signals at frequencies far removed from the desired channel. The increase in signal power level prior to mixing is desirable because of the noise that is inevitably introduced in the mixer stage (**Chapter 3**).

10. The local oscillator in the receiver is tuned to produce a frequency f_{LO} that differs from the incoming signal frequency f_{RF} by the intermediate frequency f_{IF} that is, f_{LO} can be equal to $f_{RF} + f_{IF}$ or $f_{RF} - f_{IF}$ (**Chapter 6**).

11. The mixer is a nonlinear device that shifts the received signal at f_{RF} to the intermediate frequency f_{IF}. Modulation on the received carrier is also transformed to the intermediate frequency (**Chapter 4**).

Elements of Communication Systems (5)

12. The IF amplifier increases the signal to a level suitable for detection and provides most of the frequency selectivity necessary to "pass" the desired signal and filter out the undesired signals that are found in the mixer output. Because the tuned circuits in blocks 11 and 12 always operate at a fixed frequency (f_{IF}), they can be designed to provide good selectivity. Ceramic or crystal filters are often used (**Chapter 5**).

13. The detector recovers the original message signal from the modulated IF input (**Chapter 7** (**8**)).

14. The audio or video amplifier increases the power level of the detector output to a value suitable for driving a loudspeaker, a television tube, or other output device.

15. The output device converts the signal information back to its original form (analogue or digital sound waves, picture, etc.).

Classification of RF Applications

RF applications

- **Broadcast (un-idirectional)**
  - FM/AM radio
  - VHF/UHF TV
  - Ku satelite TV
- **Wireless communications (bi-directional)**
  - Cellular telephony
  - Wireless local area network (WLAN)
  - Wireless personal area network (WPAN)
  - Cordless telephony
  - Global positioning system (GPS)

Wireless Communication Standards (1)

Standard	Digital cellular telephony (voice/data)	
	GSM	cdmaOne
Frequency range (MHz)	GSM850: DL(869-894)-UL(824-849) GSM900: DL(935-960)-UL(890-915) GSM1800 DL(1805-1880)-UL(1710-1785) GSM1900: DL(1930-1990)-UL(1850-1910)	DL(869-894)-UL(824-849) DL(1930-1990)-UL(1850-1910)
Modulation	GMSK 8-PSK (EDGE only)	QPSK/OQPSK
Multiple access	TDMA/FDMA	CDMA/FDMA
Duplex (UL/DL)	FDD	FDD
Channel bandwidth	200 KHz	1.25 MHz
Peak data rate	14.4 kbit/s 53.6 kbit/s (GPRS) 384 kbit/s (EDGE)	14.4 kbit/s (IS-95-A) 115.2 kbit/s (IS-95-B)

Wireless Communication Standards (2)

Standard	Digital cellular telephony (voice/data)		
	cdma2000	WCDMA 3GPP/FDD	WCDMA 3GPP/TDD
Frequency range (MHz)	450; 700 800; 900 1700; 1800 1900: 2100	DL(2110-2170);UL(1920-1990) DL(1930-1990);UL(1850-1910) DL(1805-1880);UL(1710-1785)	2010-2025 1900-1920 1930-1990 1850-1910 1910-1930
Modulation	QPSK, OQPSK HPSK	UL:Dual BPSK DL:QPSK, 16QAM (HSDPA only)	UL+DL:QPSK DL:8PSK (HSDPA only)
Multiple access	CDMA	CDMA/FDMA	CDMA/TDMA
Duplex	FDD	FDD	TDD
Channel bandwidth	1.25 MHz	5 MHz	5 MHZ
Peak data rate	307.7 kbit/s (CDMA2000 1x) 2.4 Mbit/s (CDMA2000 3x)	2 Mbit/s 10 Mbit/s (HSDPA)	2 Mbit/s 10 Mbit/s (HSDPA)

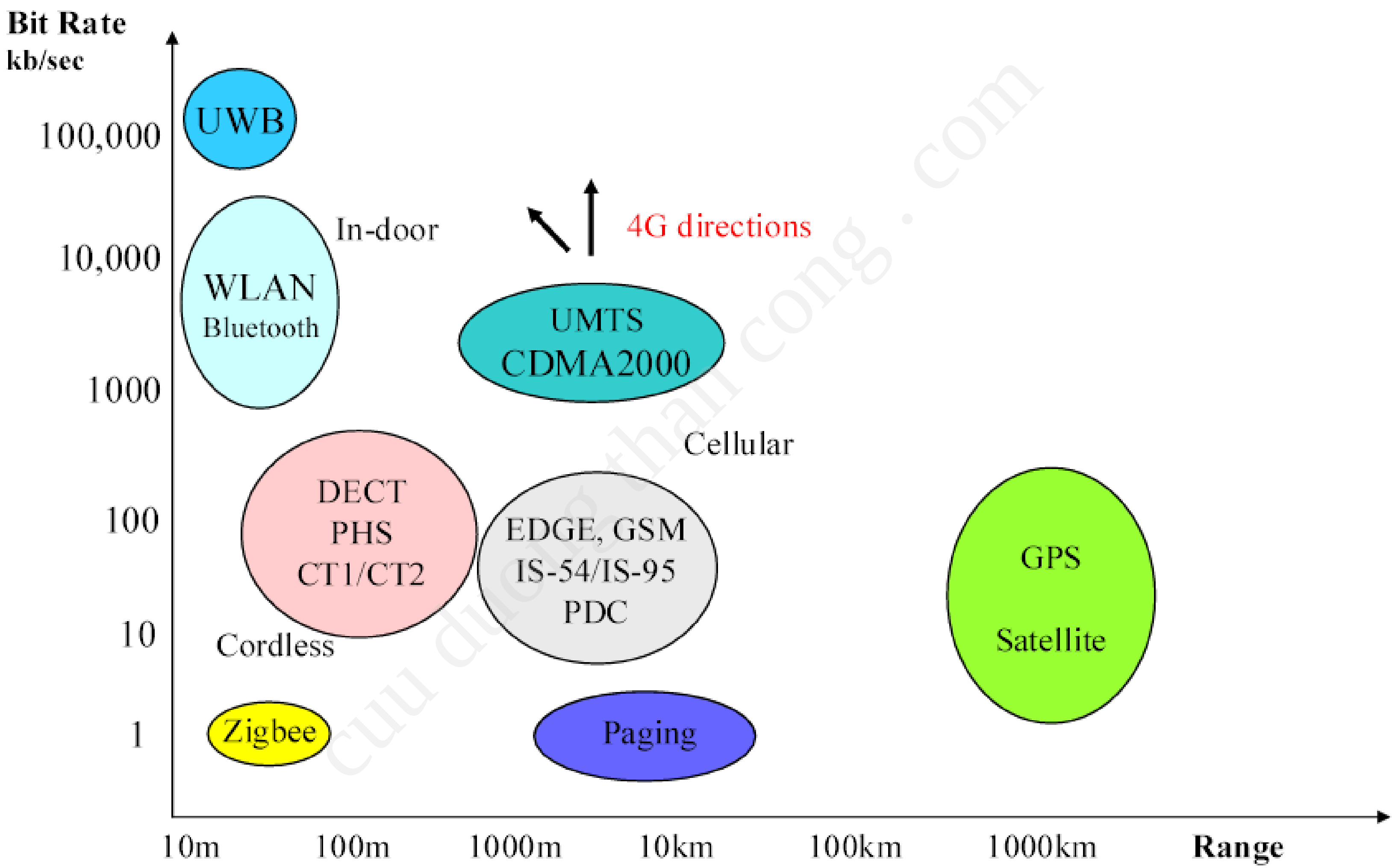
Wireless Communication Standards (3)

Standard	Digital cordless telephony (voice/data)	WPAN	WLAN and broadband access	
	DECT	Bluetooth	IEEE 802.11a	IEEE 802.11b
Frequency range (GHz)	1.88-1.9 (Europe) 1.88-1.99 (World-) wide)	2.402-2.48	5.15-5.35 (USA) 5.725-5.825 (USA)	2.4-2.4835 (North America, Europe)
Modulation	GFSK	GFSK	BPSK, QPSK, 16QAM, OFDM, 64QAM	BPSK, DQPSK, (CCK, PBCC)
Multiple access	FDMA/TDMA	FHSS	CSMA/CA	TDD
Duplex	TDD	TDD	TDD	TDD
Channel bandwidth	1.728 MHz	1 MHz	20 MHz	1 MHz
Peak data rate	1152 kbit/s	723.2 kbit/s	54 Mbit/s	11 Mbit/s

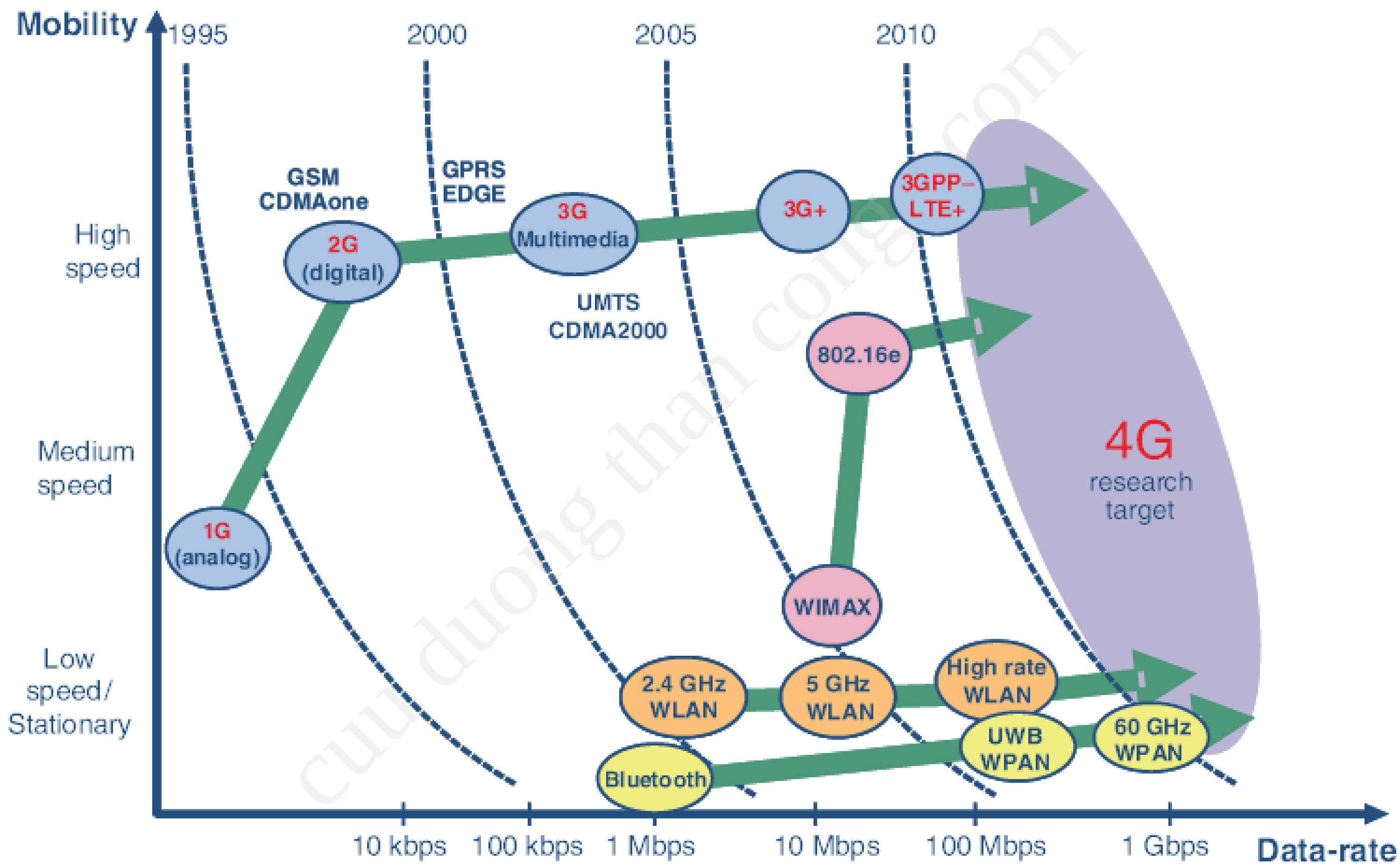
Wireless Communication Standards (4)

Standard	**WLAN and broadband access**		
	IEEE 802.11g	**IEEE 802.15.3a (UWB)**	**IEEE 802.15.4 (ZIGBEE)**
Frequency range (GHz)	2.4-2.4835	2.4-2.4835	2.4-2.4835 (World) 0.902-0.928 (America) 0.8683 (Europe)
Modulation	BPSK, QPSK, 16-64QAM, OFDM (CCK,PBCC)	QPSK, DQPSK, 16QAM, 32QAM 64QAM	16 QPSK
Multiple access	CSMA/CA	–	CSMA/CA
Duplex	TDD	–	–
Channel bandwidth	20 MHz	4.125 MHz	5 MHz
Peak data rate	54 Mbit/s	55 Mbit/s	20 kbit/s (868 MHz) 40 kbit/s (915 MHz) 250 kbit/s (2.4 GHz)

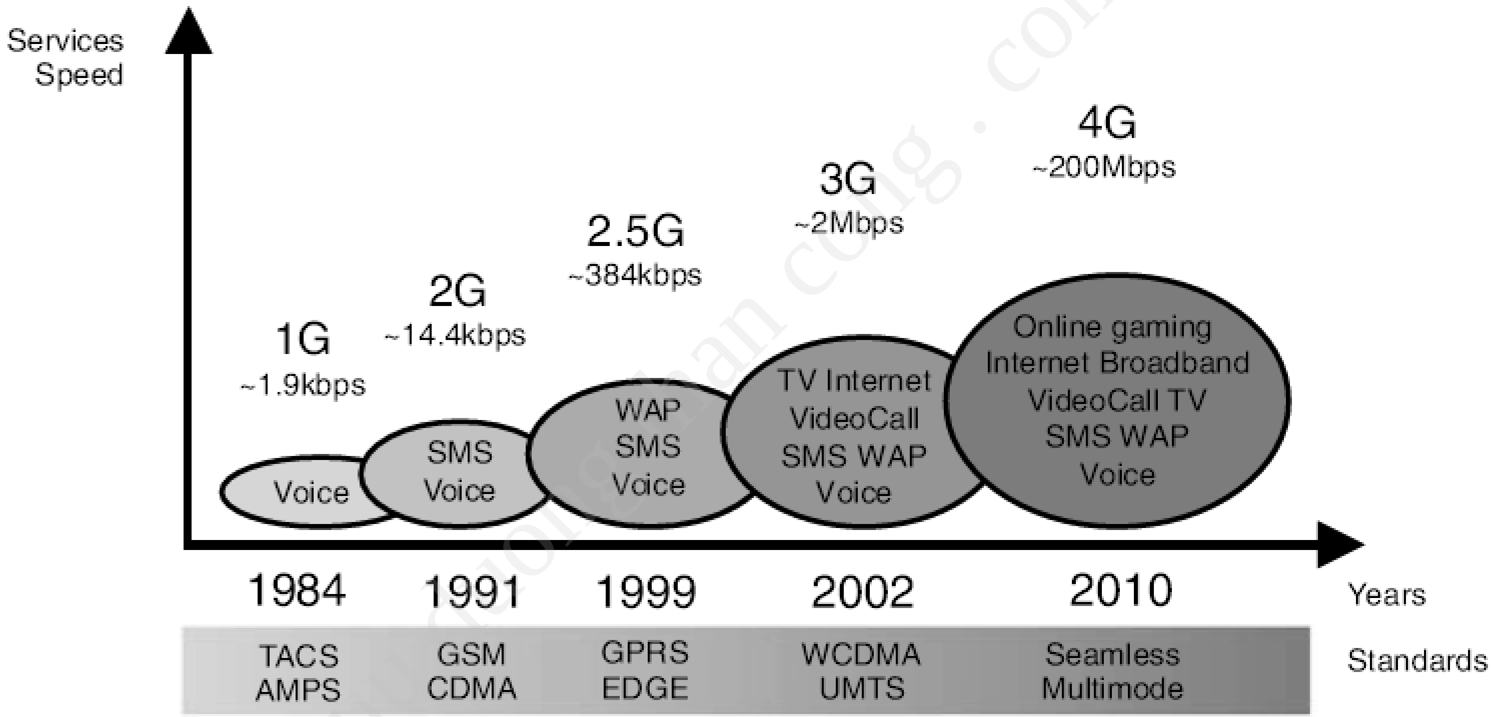
Wireless Communication Systems (1)



Wireless Communication Systems (2)



Wireless Communication Systems (3)



Frequency Band in Communication Systems (1)

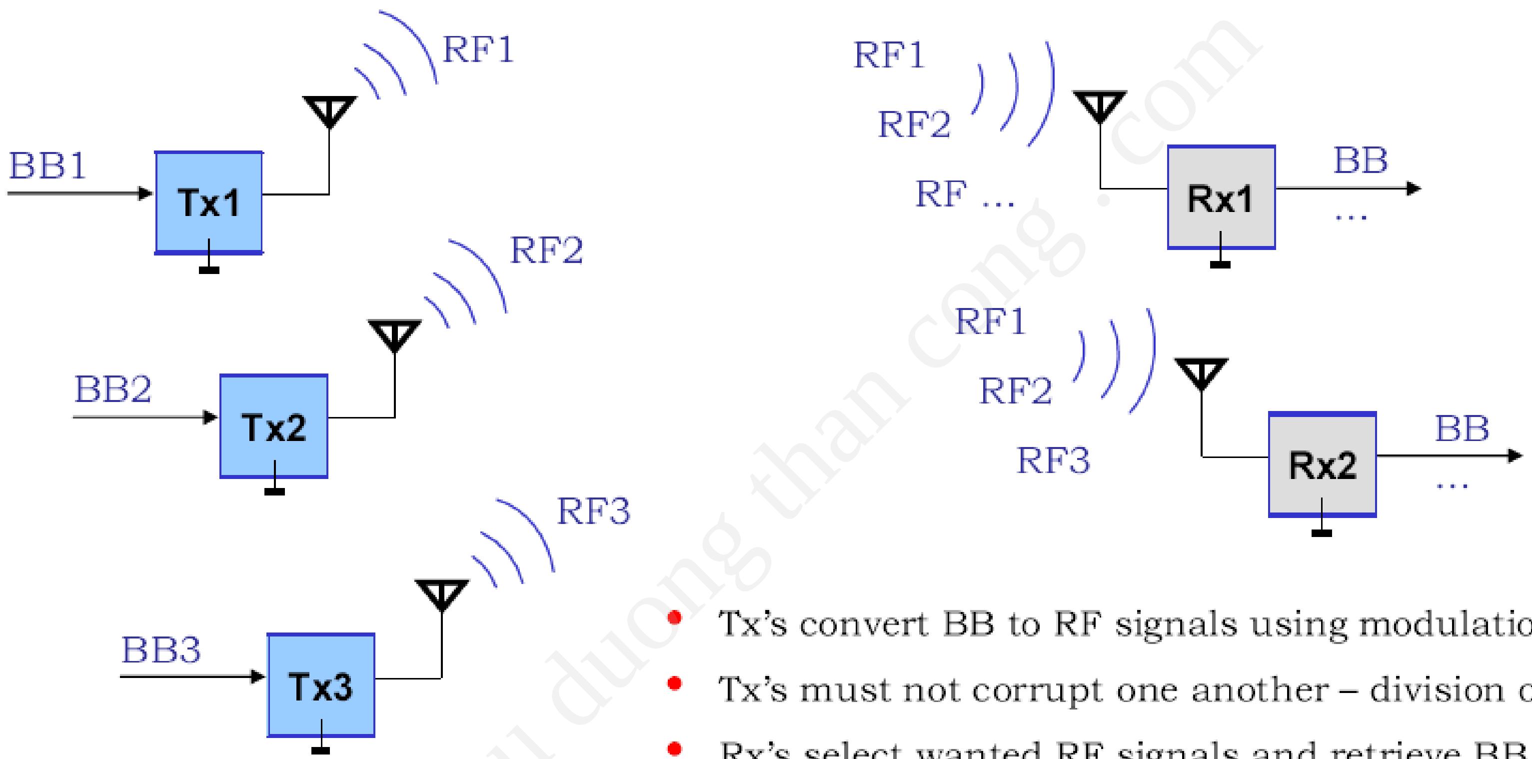
Band name	Abbr	ITU band	Frequency Wavelength	Example uses
			< 3 Hz > 100,000 km	
Extremely low frequency	ELF	1	3–30 Hz 100,000 km – 10,000 km	
Super low frequency	SLF	2	30–300 Hz 10,000 km – 1000 km	
Ultra low frequency	ULF	3	300–3000 Hz 1000 km – 100 km	
Very low frequency	VLF	4	3–30 kHz 100 km – 10 km	Military communication
Low frequency	LF	5	30–300 kHz 10 km – 1 km	Navigation, time signals, AM longwave broadcasting
Medium frequency	MF	6	300–3000 kHz 1 km – 100 m	AM broadcasts
High frequency	HF	7	3–30 MHz 100 m – 10 m	Shortwave broadcasts and amateur radio
Very high frequency	VHF	8	30–300 MHz 10 m – 1 m	FM and television broadcasts
Ultra high frequency	UHF	9	300–3000 MHz 1 m – 100 mm	television broadcasts, wireless LAN
Super high frequency	SHF	10	3–30 GHz 100 mm – 10 mm	microwave devices, mobile phones
Extremely high frequency	EHF	11	30–300 GHz 10 mm – 1 mm	
			Above 300 GHz < 1 mm	

Frequency Band in Communication Systems (2)

Microwave frequency allocations according to IEEE

Band	L	S	C	X	Ku	K	Ka	V	W
Frequency range	0.8–2 GHz	2–4 GHz	4–8 GHz	8–12 GHz	12–18 GHz	18–27 GHz	27–40 GHz	40–75 GHz	75–110 GHz

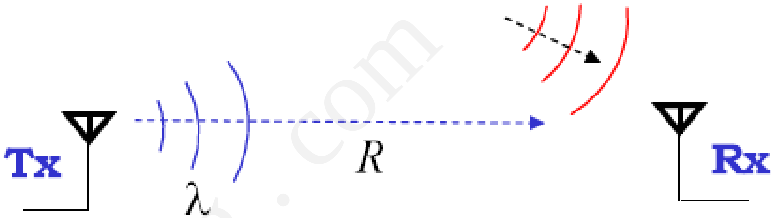
Wireless/ RF Communication Channels (1)



- Tx's convert BB to RF signals using modulation
- Tx's must not corrupt one another – division of RF band
- Rx's select wanted RF signals and retrieve BB by demodulation
- Rx's must suppress unwanted signals and noise

Wireless/ RF Communication Channels (2)

Propagation Effects



Path loss, interferers and external noise

Power loss in open area

$$L_p = 20\log(4\pi R/\lambda) \quad [\text{dB}]$$

Received power incl. gain of the antennas

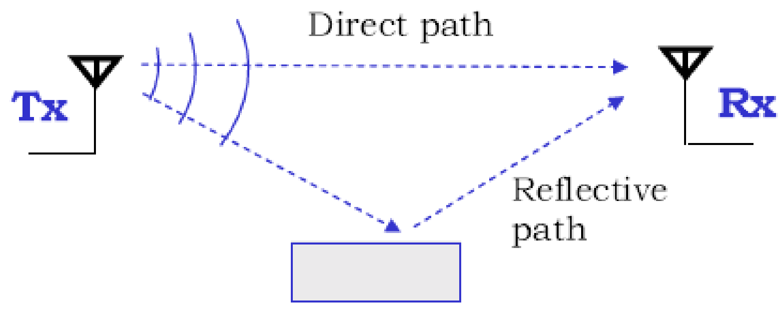
$$P_{Rx} = (P_{Tx} + G_{TxAnt}) - L_p + G_{RxAnt} \quad [\text{dBm}]$$

Wanted signal is corrupted by interferers and noise

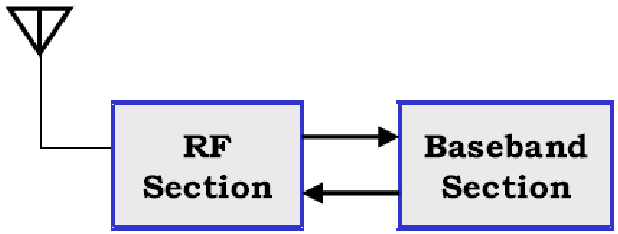
$$SNR = P_{sig} / P_{noise} \qquad SIR = P_{sig} / P_{int}$$

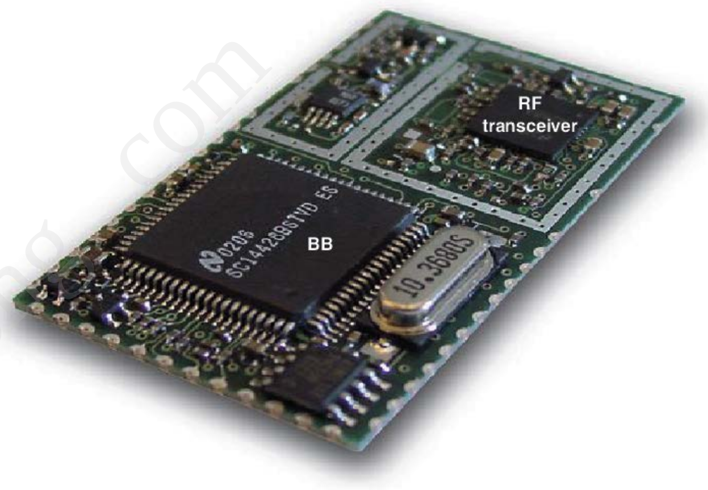
Multi-path and fading

Moving objects or Rx/Tx result in signal fluctuations, (different varying paths)



Digital Communication System (1)





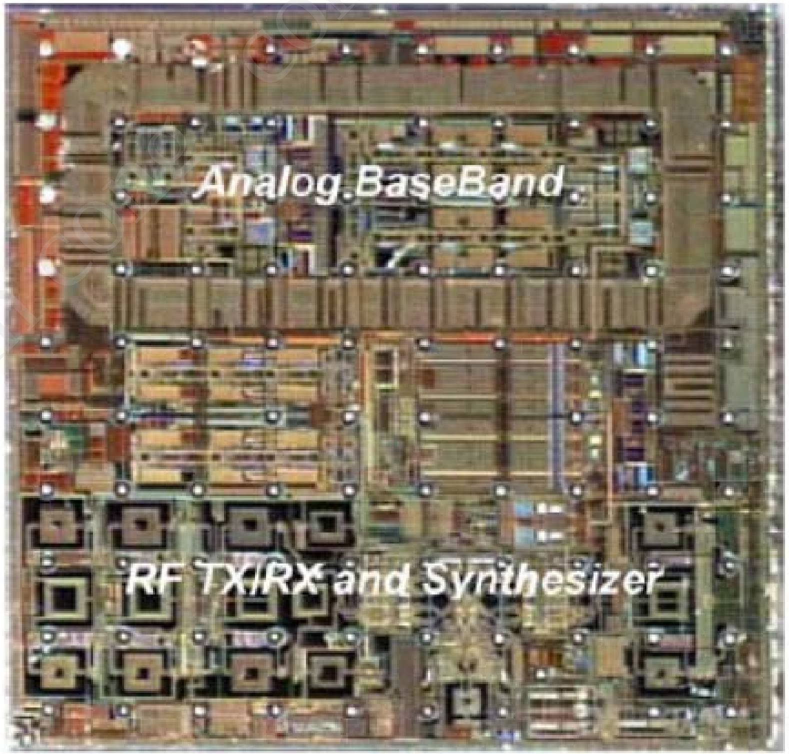
- RF Section – analog, high frequencies
- Baseband Section - mostly digital today (DSP), low frequencies

Terms DSP (digital signal processor) are used in a broad sense; therefore DSPs include digital signal processor (DSP), field programmable gate arrays (FPGA), reconfigurable computing (RC), etc.

Digital Communication System (2)

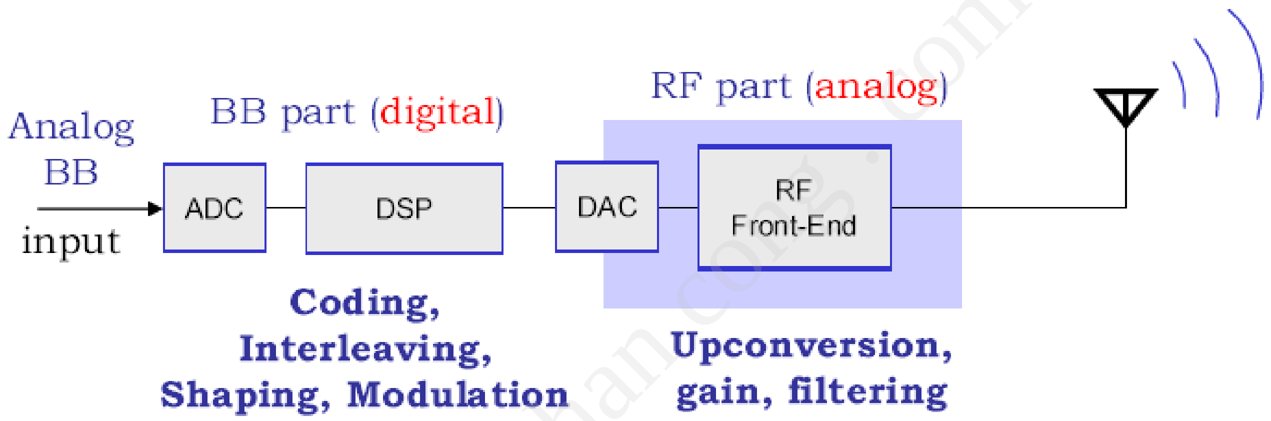
Intel RFIC transceiver on 0.18 μm TSMC CMOS technology (Taiwan Semiconductor Manufacturing Corporation).

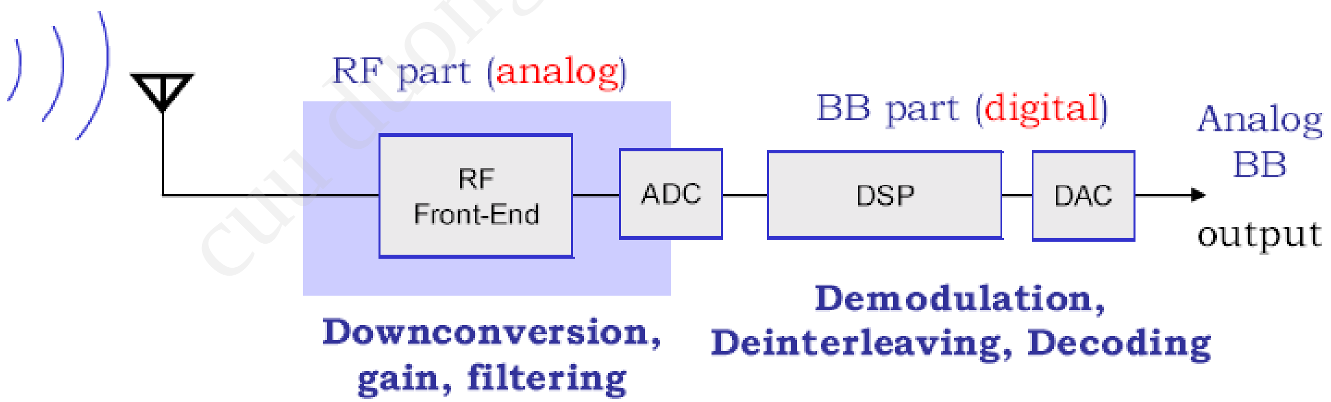
This IEEE 802.11a (in 5 GHz band) transceiver employs a direct-conversion architecture and includes an internal synthesizer. This is Intel's first RFIC used in a WLAN product.



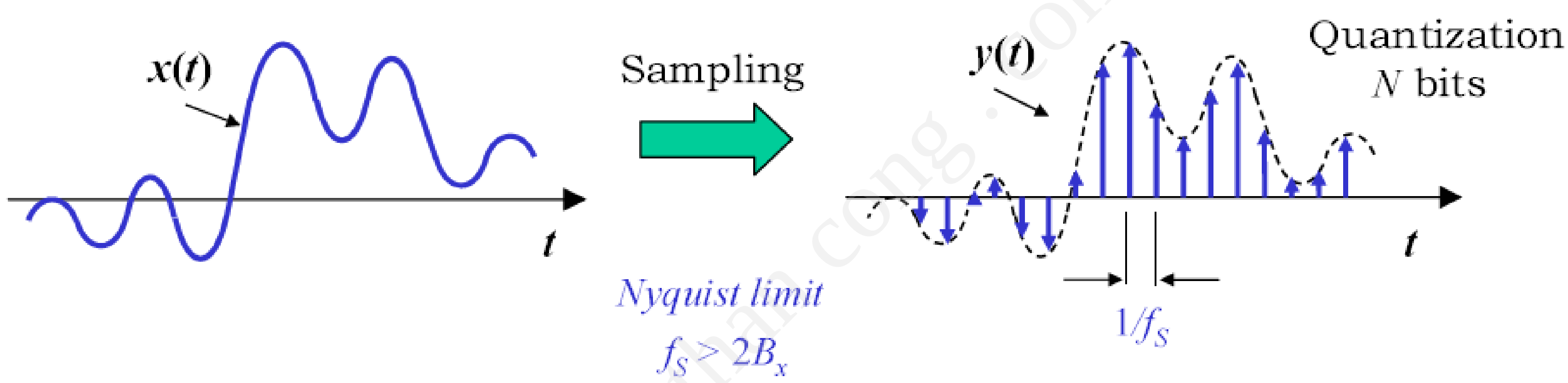
Digital Communication System (3)

Digital Tx & Rx





BB data rate



Number of bits per sample: N

Sampled BB data rate: $R = f_S N$ *bits/sec*

Example: For voice coding $B = 3.4$ kHz $f_S = 8$ kHz and $N = 8 \rightarrow R = 64$ *kb/sec.*
Next, compression with vocoders is used so $R = 2.4$ .. 9.6 *kb/sec*
but the transmitted data rate would be much higher for system arrangements and extra data needed, e.g. GSM – 270 kb/s, IS-95 (CDMA) – 1.23Mb/s

Shannon limits

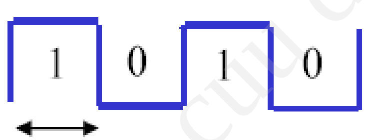
Information capacity: $C = 2B \log_2 M$ $[bits/sec]$

Channel bandwidth (B) — *Number of signal levels transmitted* (M)

Bandwidth efficiency: $C/B = 2 \log_2 M$ $[bits/sec/Hz]$

For 2-levels: $C/B = 2$, maximum possible to achieve,

$M = 2$

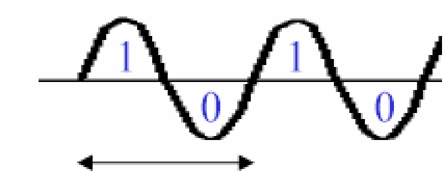


T_b

$Period = 2T_b$

Low pass channel

$B_{min} = 1/2T_b$



$2T_b$

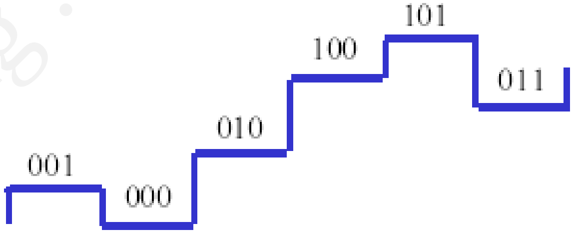
In Rx at least the first harmonic is needed

Digital Communication System (6)

Shannon limit due to noise

$$C = 2B \log_2 M \quad [bits/sec]$$

M-ary system



Information capacity $\uparrow$ if $B\uparrow$ or $M\uparrow$

The more levels the more noise harmful

$$C = B \log_2(1 + \text{SNR}) \quad [bits/sec]$$

Channel noise limits C, but M is not specified here.

In practice bit rate must be $\boldsymbol{R < C}$ to support transmission with an acceptable error rate

Digital Communication System (7)

RF systems vs channel capacity

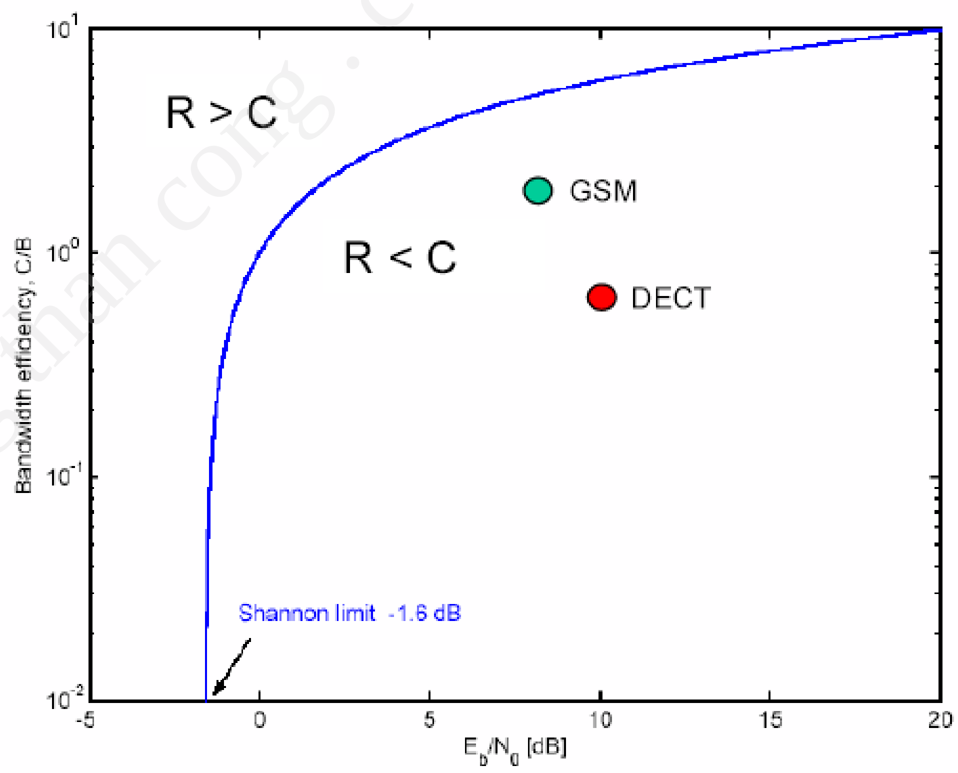
$$SNR = \frac{E_b R}{N_0 B}$$

$$\frac{C}{B} = \log_2\left(1 + \frac{E_b}{N_0} \times \frac{C}{B}\right)$$

Bit rate $R < C$ for any system

e.g. **for GSM**:
$R/B = 270\text{kbps}/200\text{kHz} = 1.35$
@ $SNR = 9\text{dB}$ for BER $< 10^{-3}$

for DECT:
$R/B = 1152\text{kbps}/1728\text{kHz} = 0.67$
@ $SNR = 10.3\text{dB}$ for BER $< 10^{-3}$



Tradeoff between signal BW and power

Digital Communication System (8)

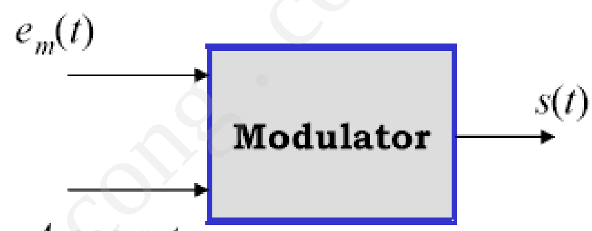
Digital modulation schemes

- Frequency shift keying FSK
- Phase shift keying PSK, QPSK, ...
- M-ary QAM
- Minimum shift keying MSK
- OFDM technique

Digital Communication System (9)

Basic view on modulation

Base-band signal
to be transmitted
low frequency



Sinusoidal carrier $x(t) = A_0 \cos \omega_0 t$
high frequency

Sinusoidal Carrier $x(t)$:

$$x(t) = A_0 \cdot \cos(\omega_0 t + \phi)$$

Amplitude

Frequency Phase

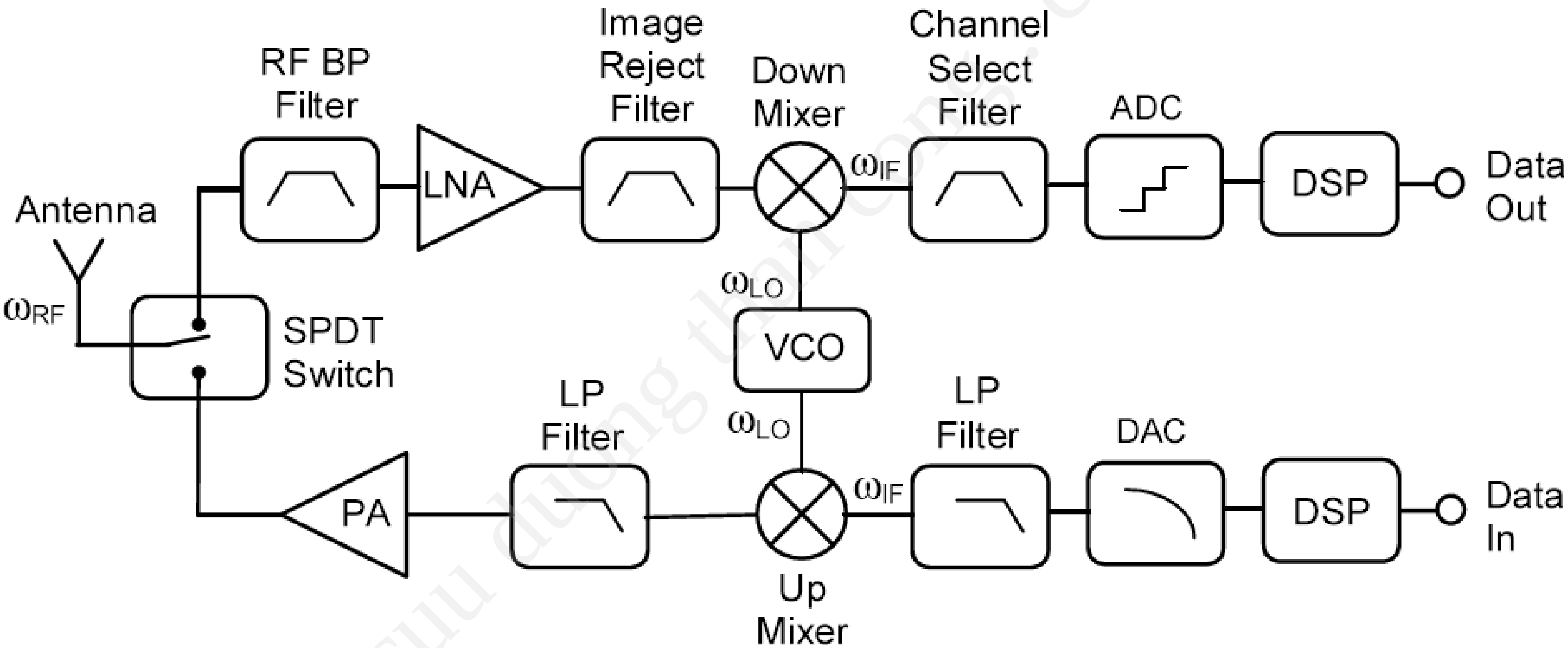
Angle

3 different parameters
available for modulation
by the base-band signal
(i.e. the modulating signal)

Angle modulation more useful in digital communication for its higher immunity to noise and interference

RF Transceiver Architecture (1)

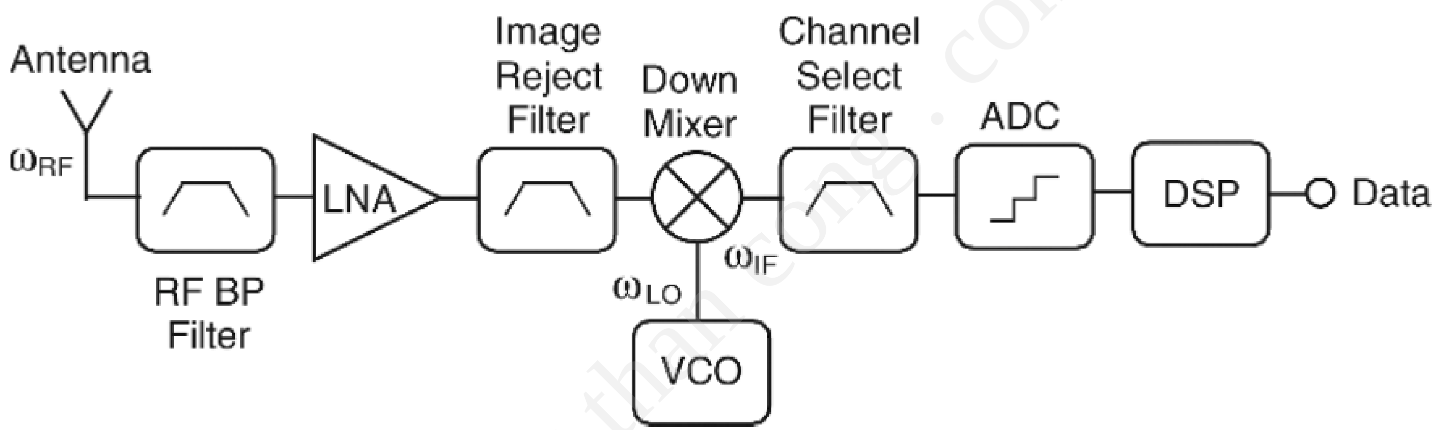
- A **transceiver** consists of a **trans**mitter and a re**ceiver**.
- Example of super-heterodyne transceiver:



(SPDT: Single Pole Double Throw)

RF Transceiver Architecture (2)

- A simplified architecture of the **super-heterodyne receiver** with **single down-conversion**:

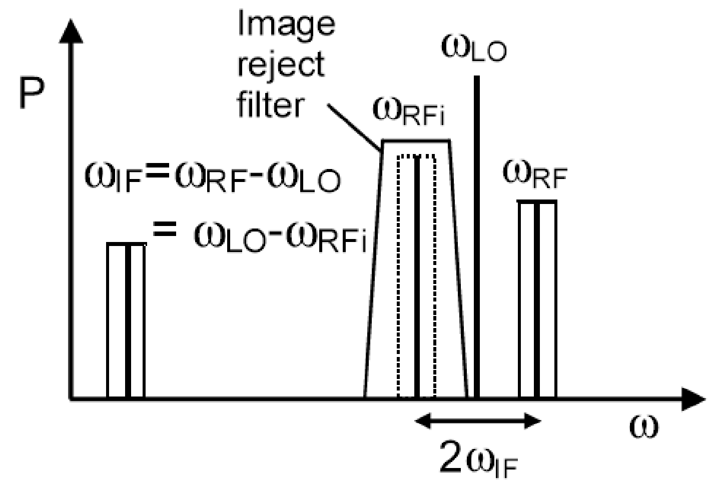


BP: Band pass, LNA: Low Noise Amplifier,

VCO: Voltage Controlled Oscillator,

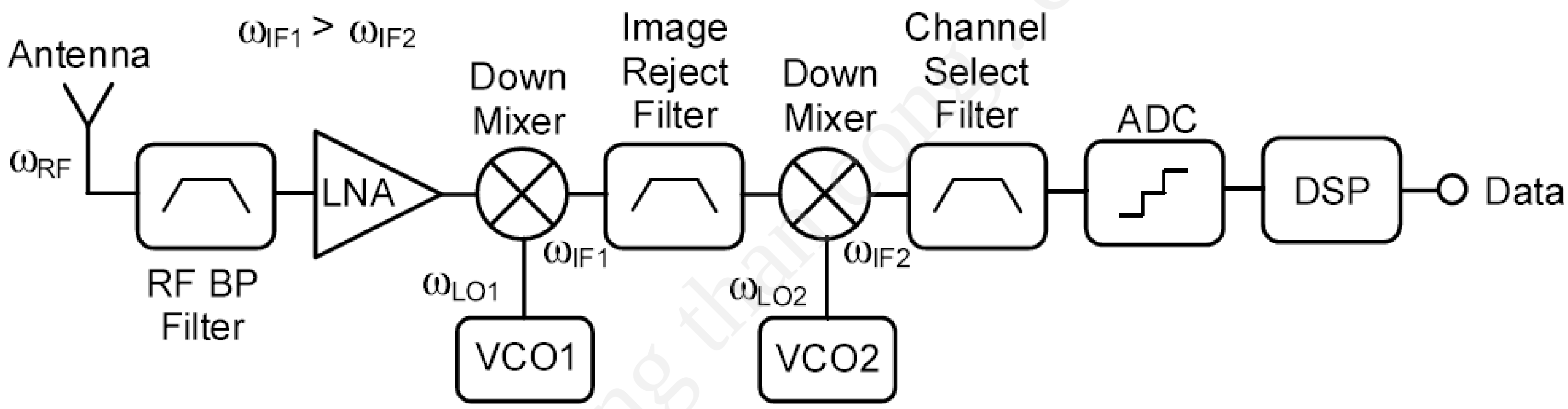
ADC: Analogue Digital Converter,

DSP: Digital Signal Processor



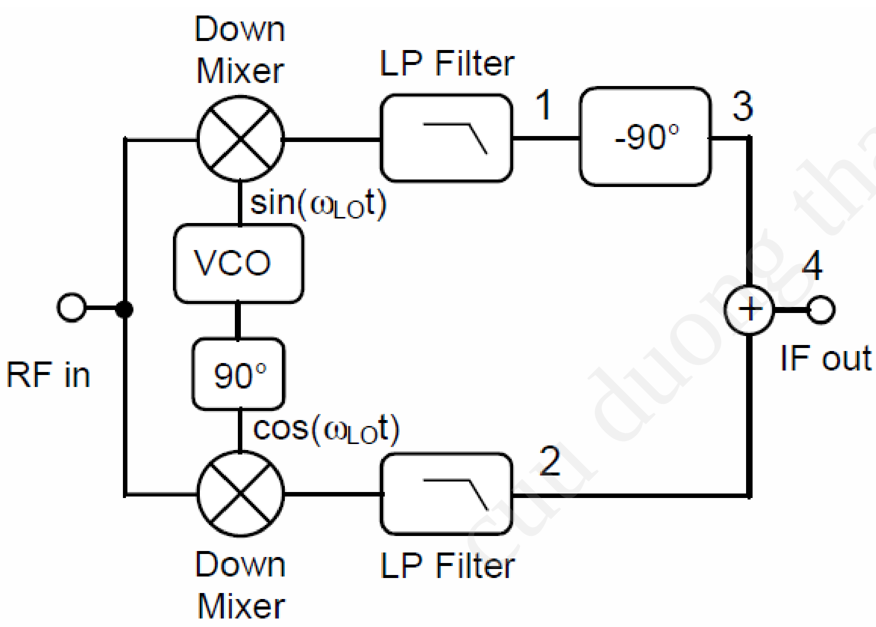
RF Transceiver Architecture (3)

- Simplified architecture of **super-heterodyne receiver** with **double down-conversion:**

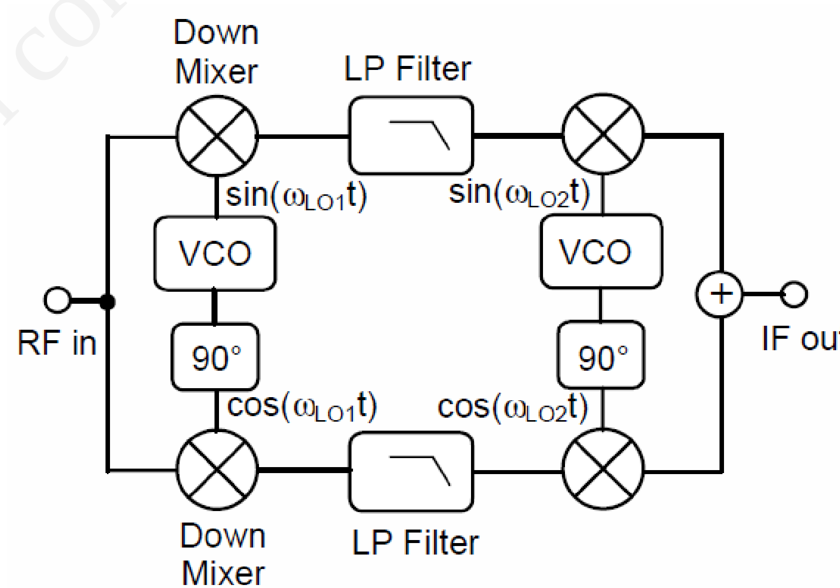


RF Transceiver Architecture (4)

- **Image rejection receiver:** smart techniques for the rejection of the image frequency without requiring sophisticated filters. Such techniques are especially useful for applications where the desired RF and the undesired image signal are so close in frequency that conventional filtering is not possible.



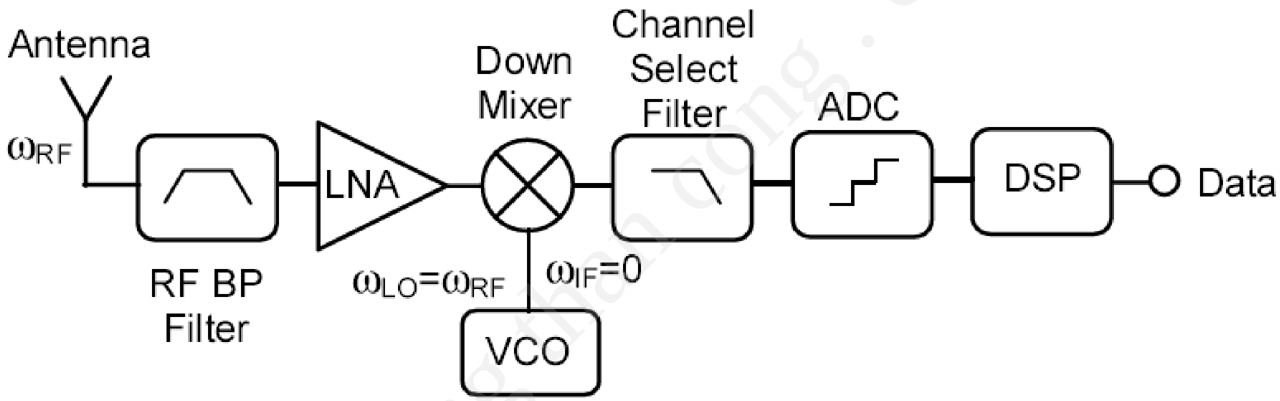
Hartley image rejection technique



Weaver image rejection technique

RF Transceiver Architecture (5)

- **Direct conversion receiver:** The motivation of increased integration has led to the direct conversion receiver, which is also referred to as **homodyne** or **zero-IF receiver**



The idea is to translate the RF signal directly to zero-IF frequency thereby exhibiting the following advantages: First, the channel filtering can be performed by a low pass filter (Recall that a more complex band pass filter is necessary for the super-heterodyne receiver). Second, the IF frequency of zero eliminates the image problem. Hence, no external high-Q image reject filter is required making fully integrated solutions feasible.

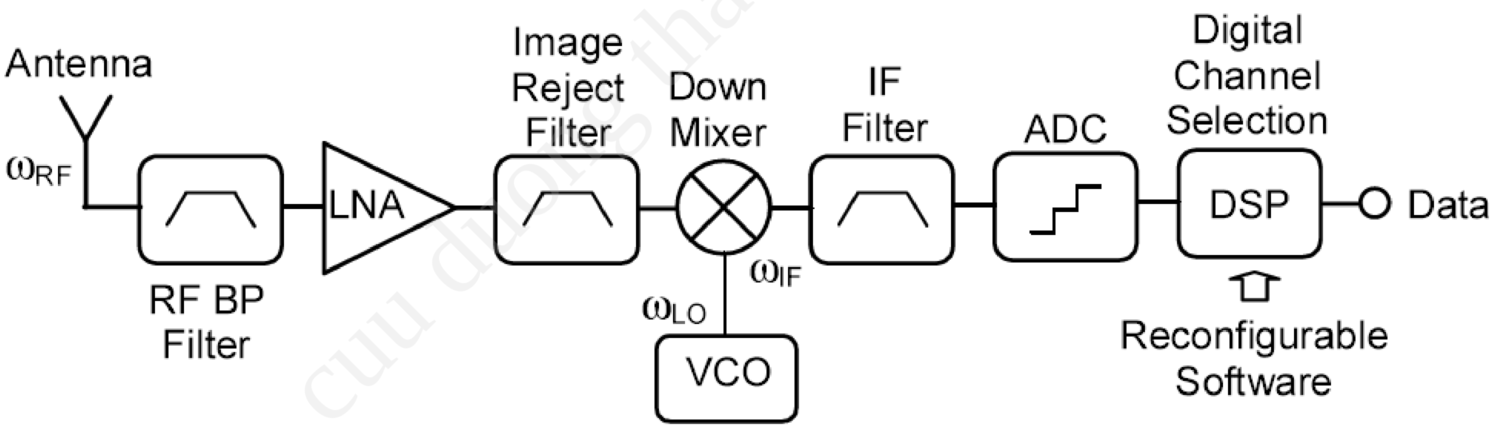
RF Transceiver Architecture (6)

- **Low-IF receiver:** Similar to the direct conversion receiver, a (quadrature) mixer is used to translate the desired channels to a low IF frequency. Typically, an IF frequency in the order of one up to two channel bandwidths corresponding to 50 kHz to 10 MHz are used as IF frequency. The image rejection can be performed by mixer topologies similar to the Hartley or Weaver architecture. Due to the low IF frequency, channel filtering is relatively simple.

  Unlike the zero-IF architecture, the low-IF receiver is not sensitive to the parasitic DC offset, LO leakage and flicker noise. The low IF topology is an excellent compromise between the zero-IF and the super-heterodyne architecture. Thus, the low-IF approach is quite popular in today's receivers.

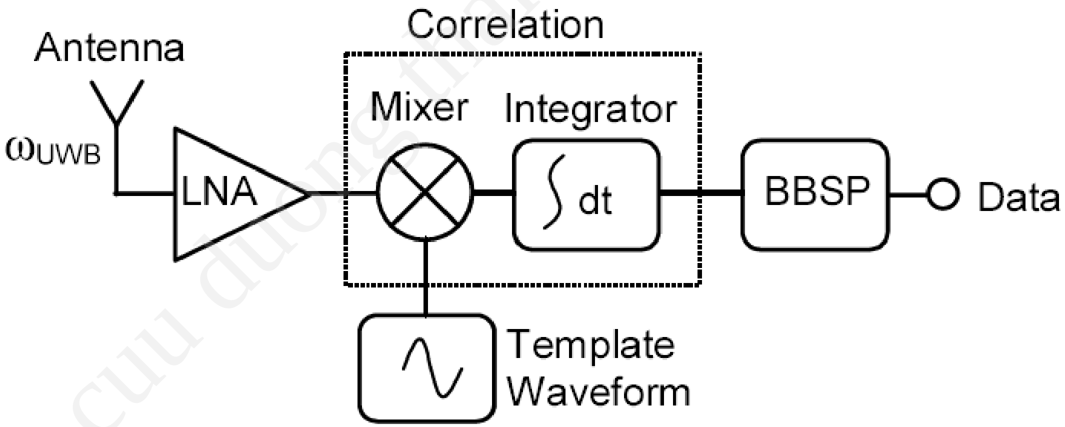
RF Transceiver Architecture (7)

- **Digital-IF receiver:** The idea is to perform the demanding channel filtering completely in the digital domain. Thus, simple RF filters may be employed for coarse band selection. The major advantage is the **flexibility of the architecture**. The receiver can be reconfigured for a variety of systems with different modulation types, channel frequencies and bandwidths meeting the demands of different standards.



RF Transceiver Architecture (9)

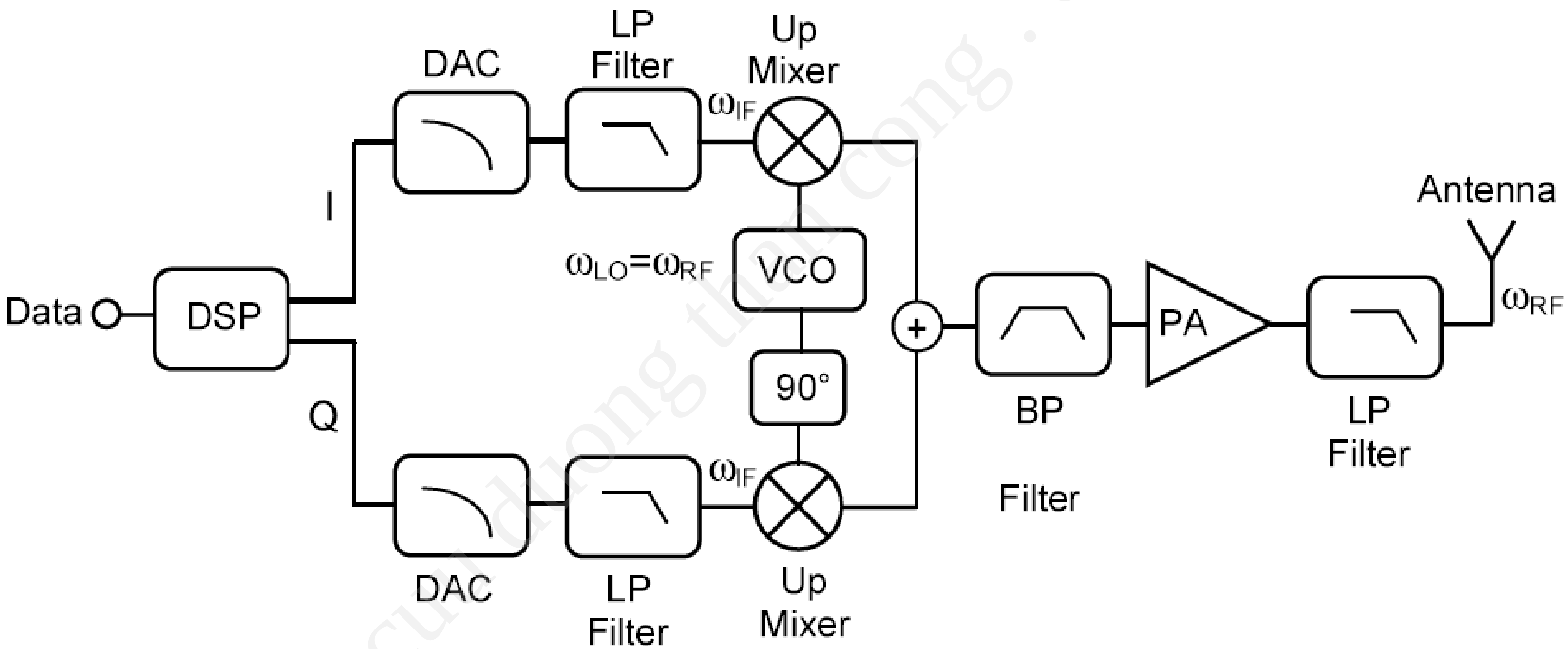
- **Impulse radio receiver:** In recent years, impulse based radios receive a revival due to its promising properties for short range, low power and high speed applications. In the USA, corresponding **UWB** (**Ultra-Wideband**) standards have already been published by the FCC (Federal Communications Commission). The UWB standard employs impulse transmission within a frequency band between 3.1 GHz and 10.6 GHz.



(BBSP: Baseband Signal Processing)

RF Transceiver Architecture (10)

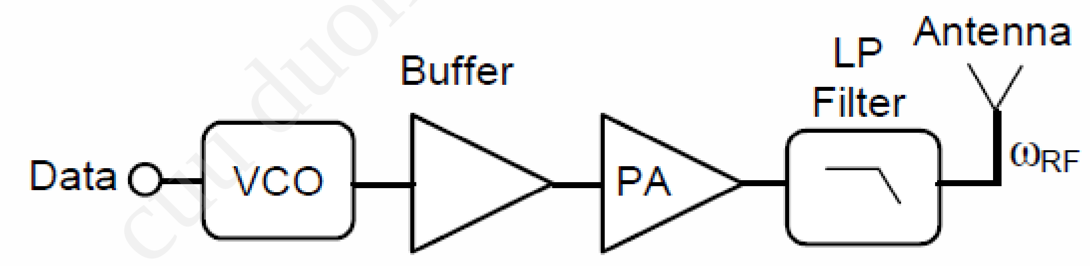
- **Direct conversion transmitter:** The baseband signal is up-converted to RF, band pass filtered, amplified and low pass filtered before the signal is emitted by the antenna.



(PA: Power Amplifier, DAC: Digital Analogue Converter)

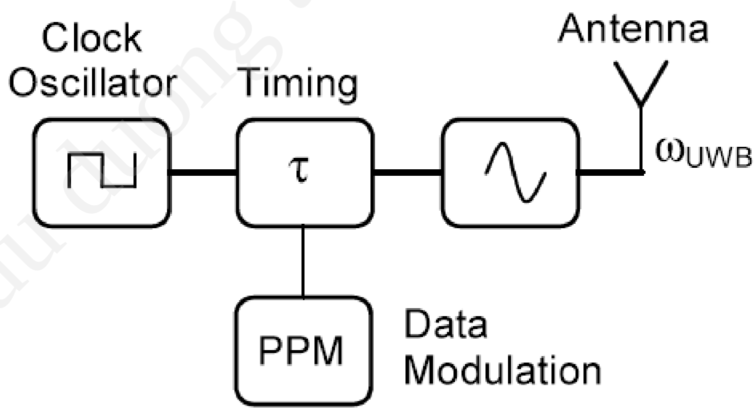
RF Transceiver Architecture (11)

- **Direct modulation transmitter:** The baseband signal is modulated and up-converted in one single step. By means of the frequency control voltage, the VCO is modulated by the applied data. Subsequently, the signal is amplified, low pass filtered and emitted via antenna. The architecture is well suited for frequency and phase modulations. Among the advantages of this approach are the low complexity, the increased ability for integration and the low power consumption. A PLL (Phase Locked Loop) is often added to improve the frequency stability and to reduce the content of harmonics and noise.



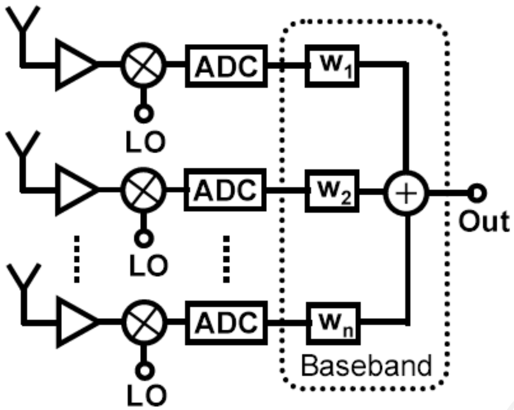
RF Transceiver Architecture (12)

- **Impulse Radio Transmitter:** an impulse radio transmitter is illustrated consisting of a pulse generator, a timing circuit and a clock oscillator. PPM (Pulse Position Modulation) is used for data modulation. A programmable delay circuit can be employed to determine the timing. The desired waveform is produced by the pulse generator, while the clock oscillator defines the pulse repetition frequency. Step, Gaussian or monocycle pulses are suited for UWB communication since they have a broadband frequency spectrum.

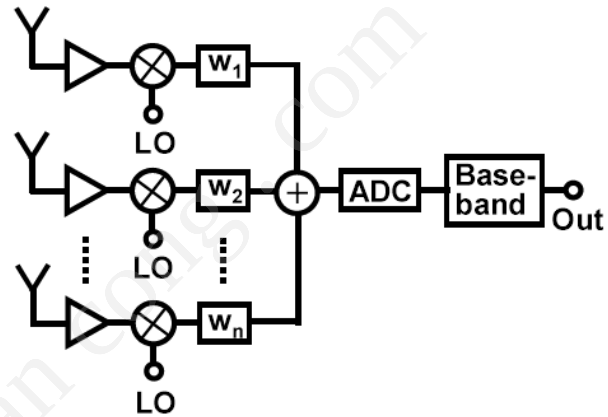


RF Transceiver Architecture (13)

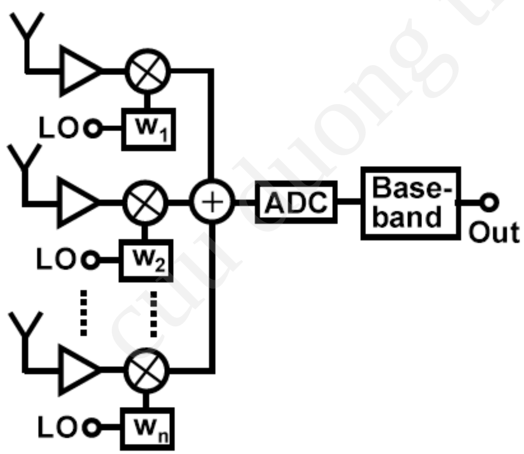
❑ **Smart antenna transceivers:**



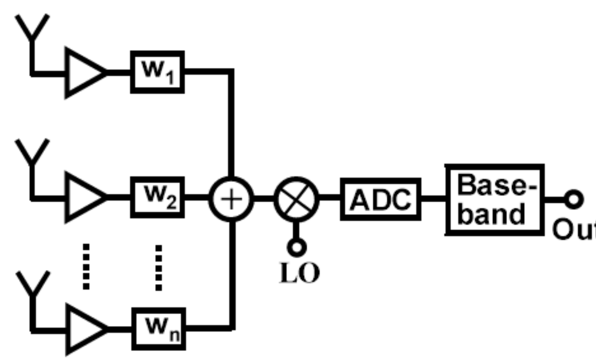
a.) Adaptive BB Combining



b.) Adaptive IF Combining



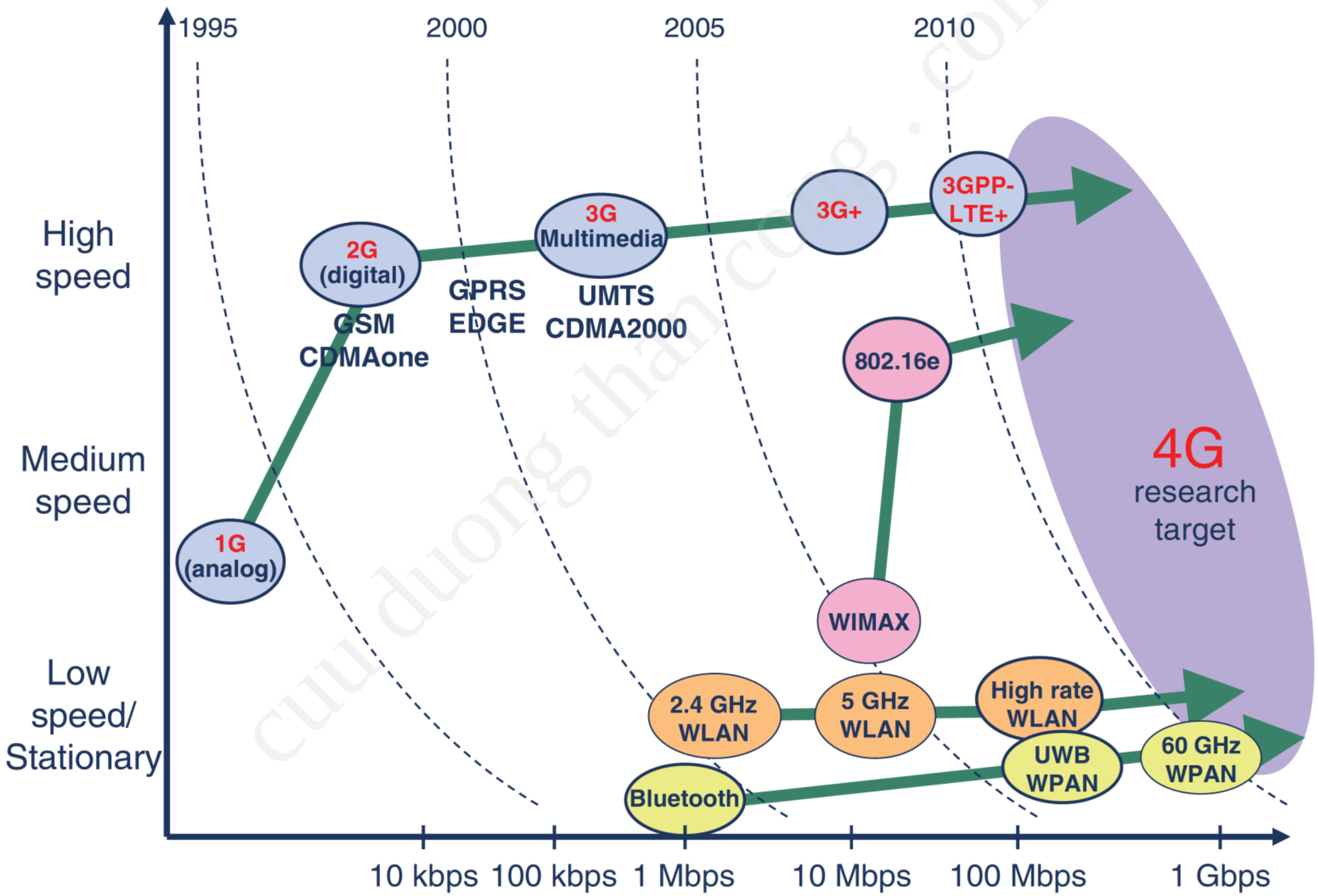
c.) Adaptive LO Combining



d.) Adaptive RF Combining

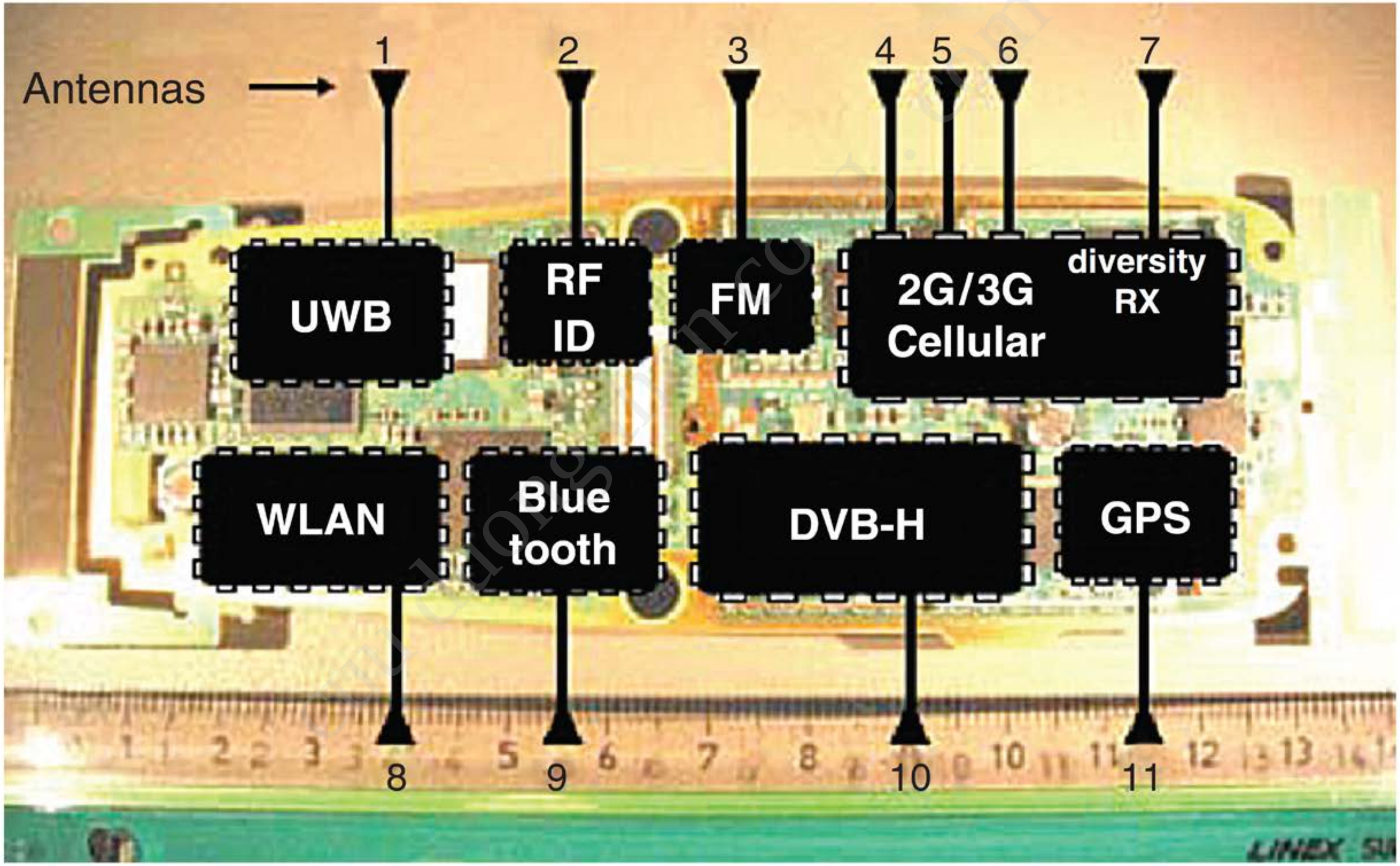
Software-defined radios (SDR) (1)

Variety of wireless access standards



Software-defined radios (2)

Multi-mode handset featuring separate radios



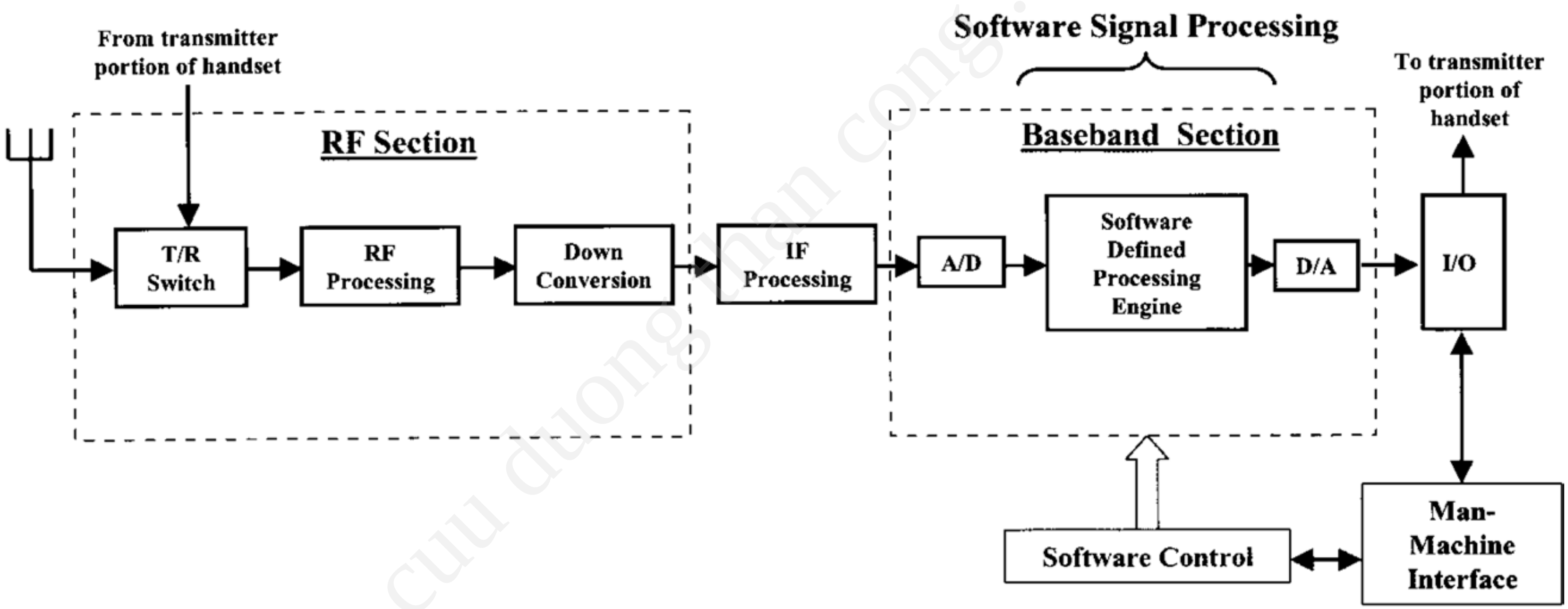
Software-defined radios (3)

- **<u>Software-defined radio (SDR)</u>** is a radio communication technology that is **based on software defined wireless communication protocols instead of hardwired implementations**. In other words, frequency band, air interface protocol and functionality can be upgraded with software download and update instead of a complete hardware replacement. SDR provides an efficient and secure solution to the problem of **building multi-mode, multi-band and multifunctional wireless communication devices**.
- An SDR is capable of being re-programmed or reconfigured to operate with different waveforms and protocols through dynamic loading of new waveforms and protocols. These waveforms and protocols can contain a number of different parts, including modulation techniques, security and performance characteristics defined in software as part of the waveform itself.

<u>Source</u>: http://focus.ti.com/docs/solution/folders/print/357.html

Software-defined radios (4)

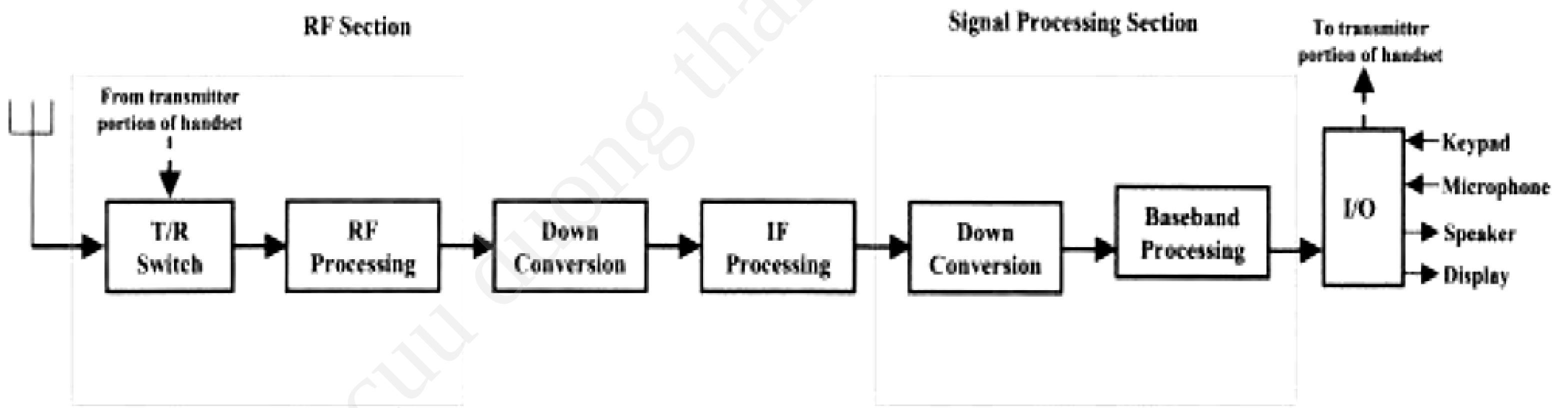
Conceptual definition of the software defined radio (applicable for wireless handset and base station architecture)



Software-defined radios (5)

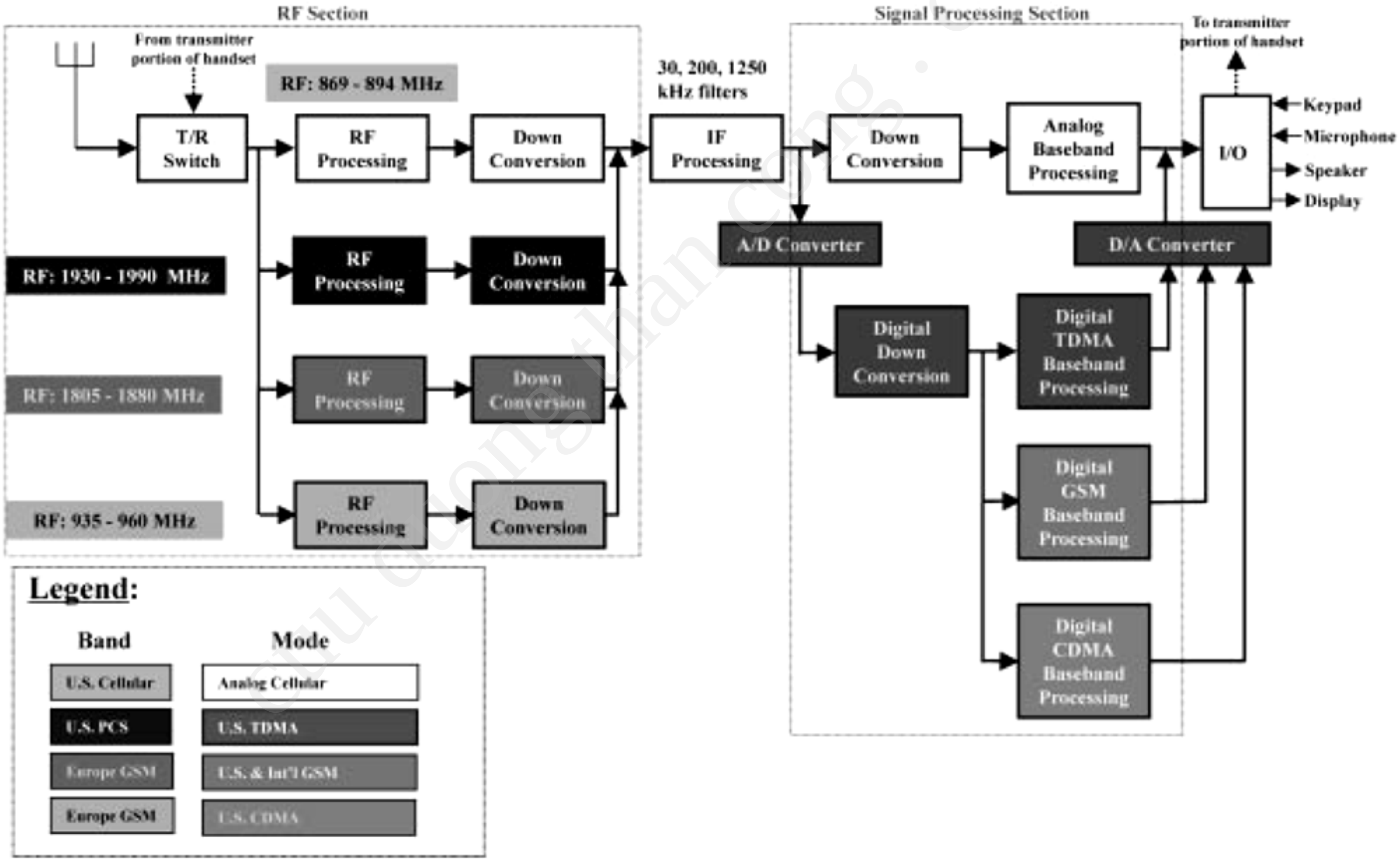
<u>Example</u>:

<u>SDR evolution – stage 1</u>: Cellular /PCS generic single mode, single band handset. This figure is representative of ANY single mode (i.e. AMPS, TDMA, CDMA, GSM, PHS, etc.) and single frequency band (i.e. 850, 900, 1800, 1900, etc.) handset. This is considered to be the traditional design product implementation.



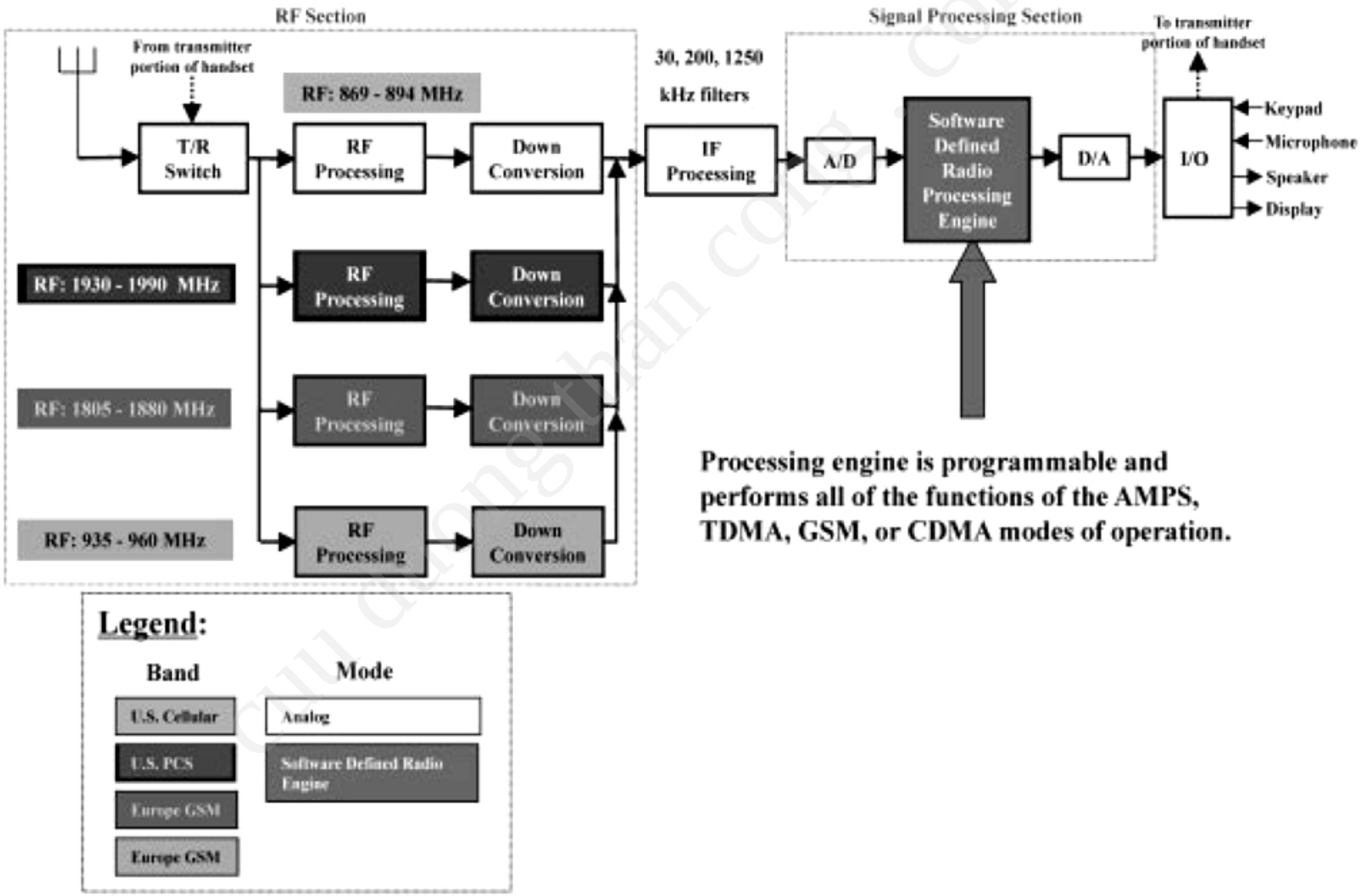
Software-defined radios (6)

<u>SDR Evolution – stage 2</u>: Quadruple-band (800, 900,1800, and 1900 MHz), quadruplemode (AMPS, TDMA, GSM, CDMA), traditional-design, multi-band, multi-mode handset.



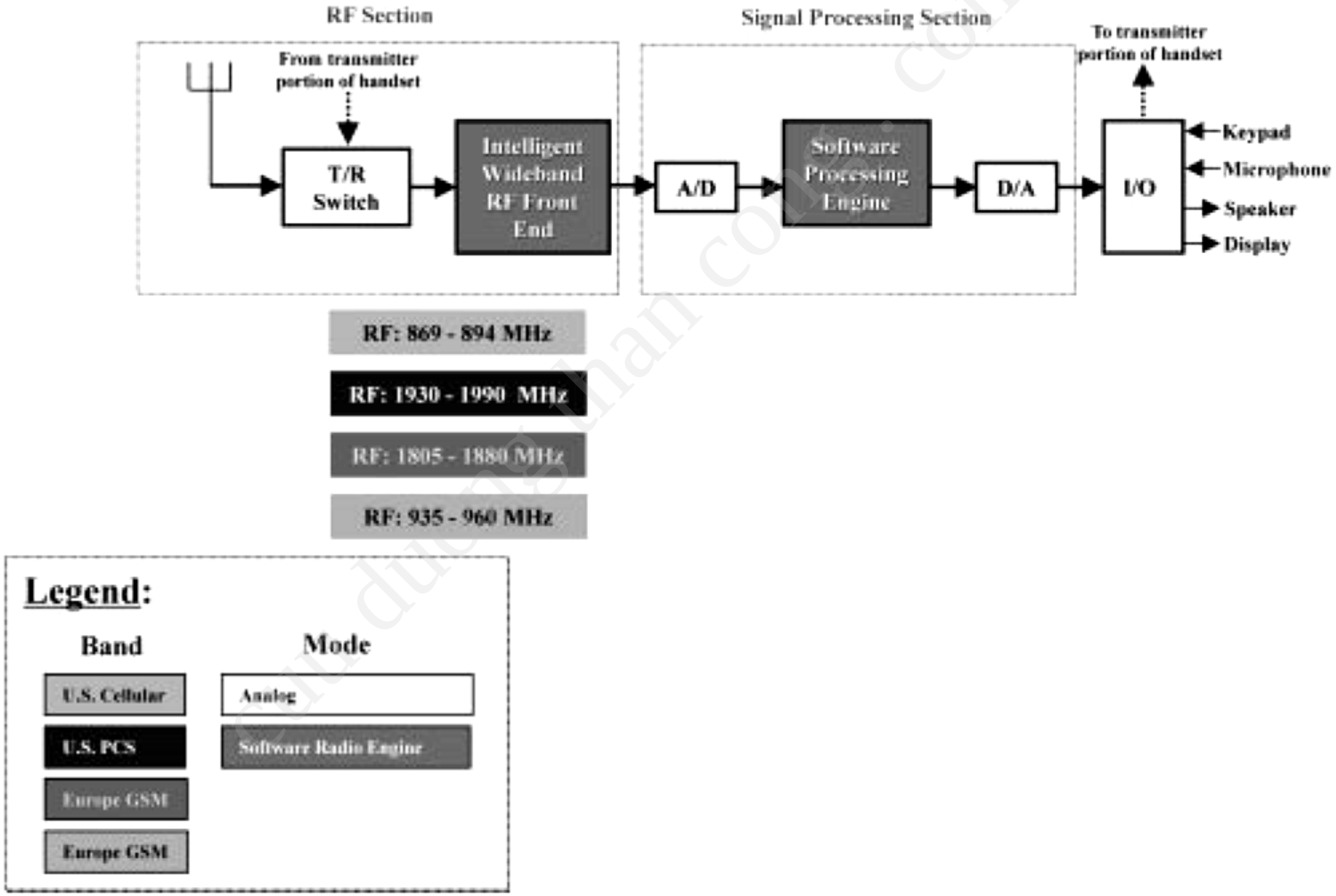
Software-defined radios (7)

SDR Evolution – stage 3: A/D, D/A, and signal processing chips currently have the capacity to perform this IF and baseband processing.



Software-defined radios (8)

<u>SDR Evolution – stage 4</u>: Future product as technology evolves in A/D capabilities, etc.



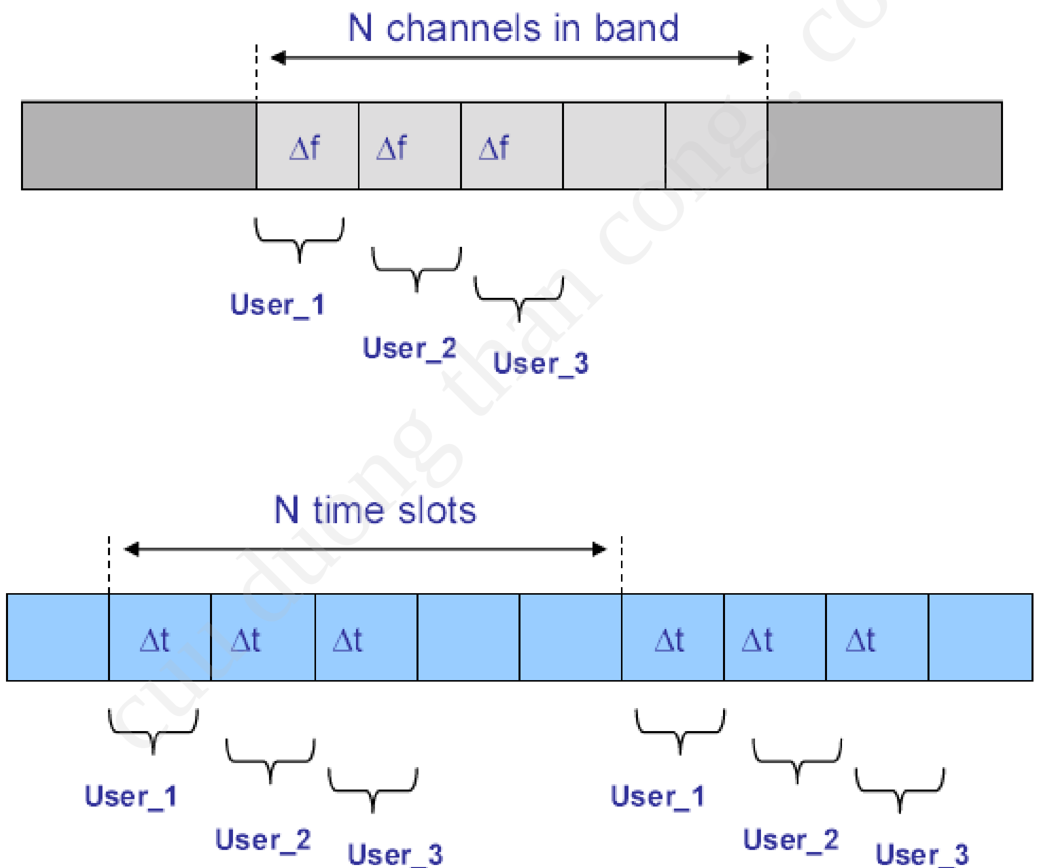
Multiple Access (1)

Multiple-Access techniques

- FDMA **(Frequency division)**
- TDMA **(Time division)**
- CDMA **(Code division)**
- Up-link and down-link TDD/FDD

Multiple Access (2)

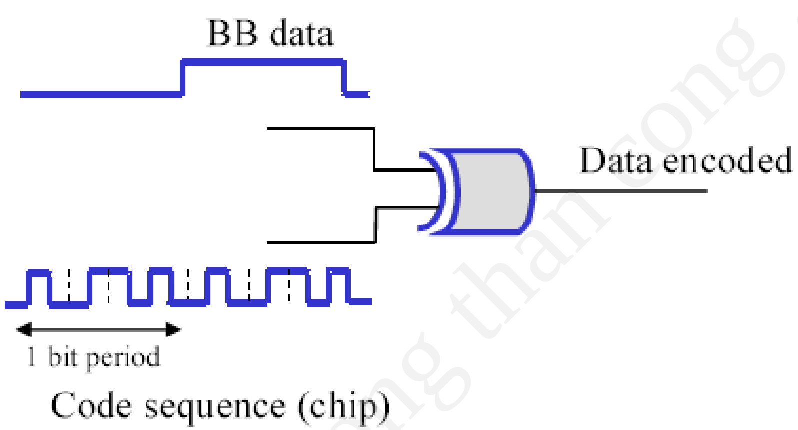
FDMA and TDMA systems



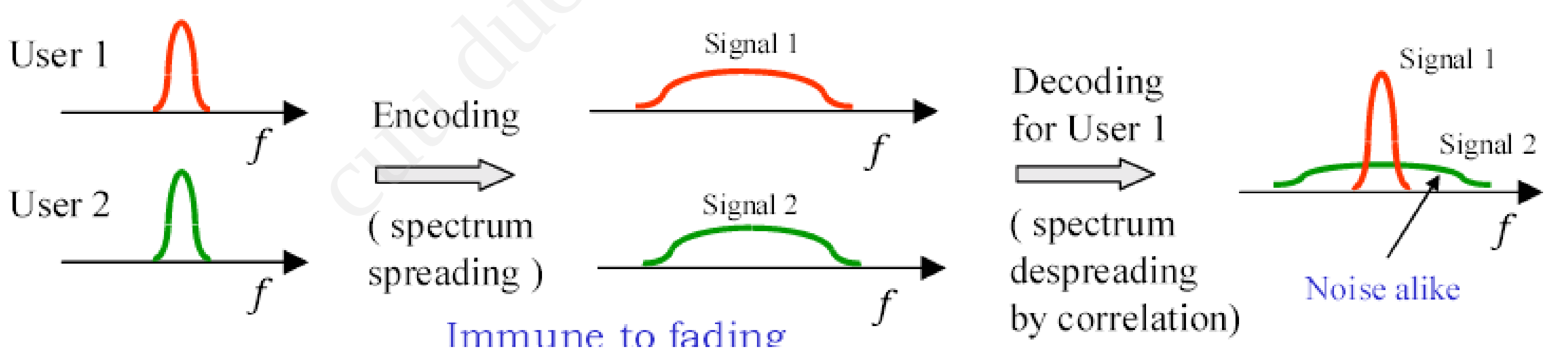
Multiple Access (3)

CDMA systems

Direct sequence CDMA



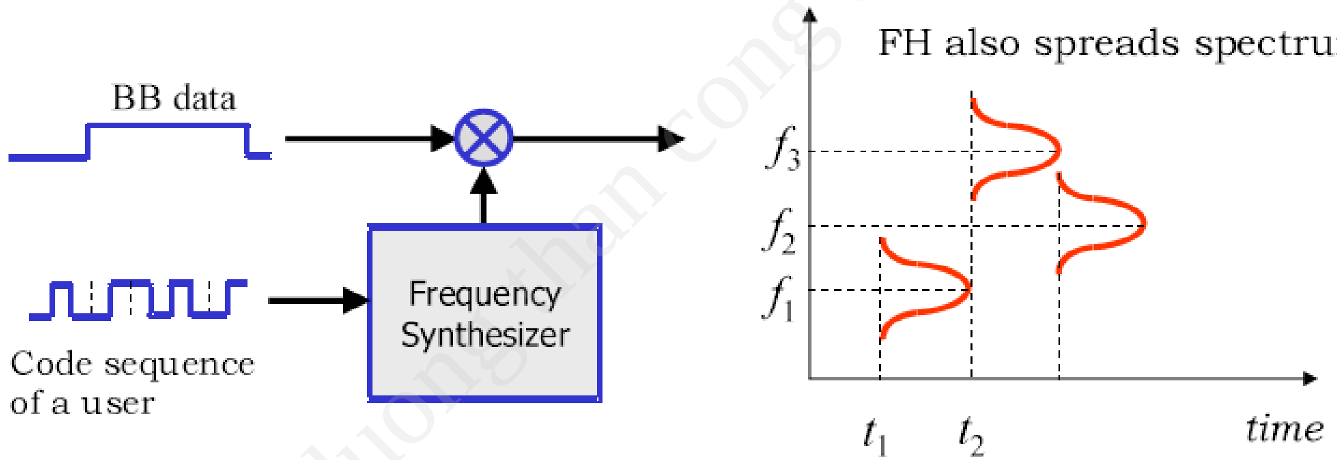
Coding sequences for different users are orthogonal (e.g. Walsh, Barker), signals overlap in frequency band and in time.



Multiple Access (4)

CDMA systems (cont'd)

Frequency-hopping CDMA



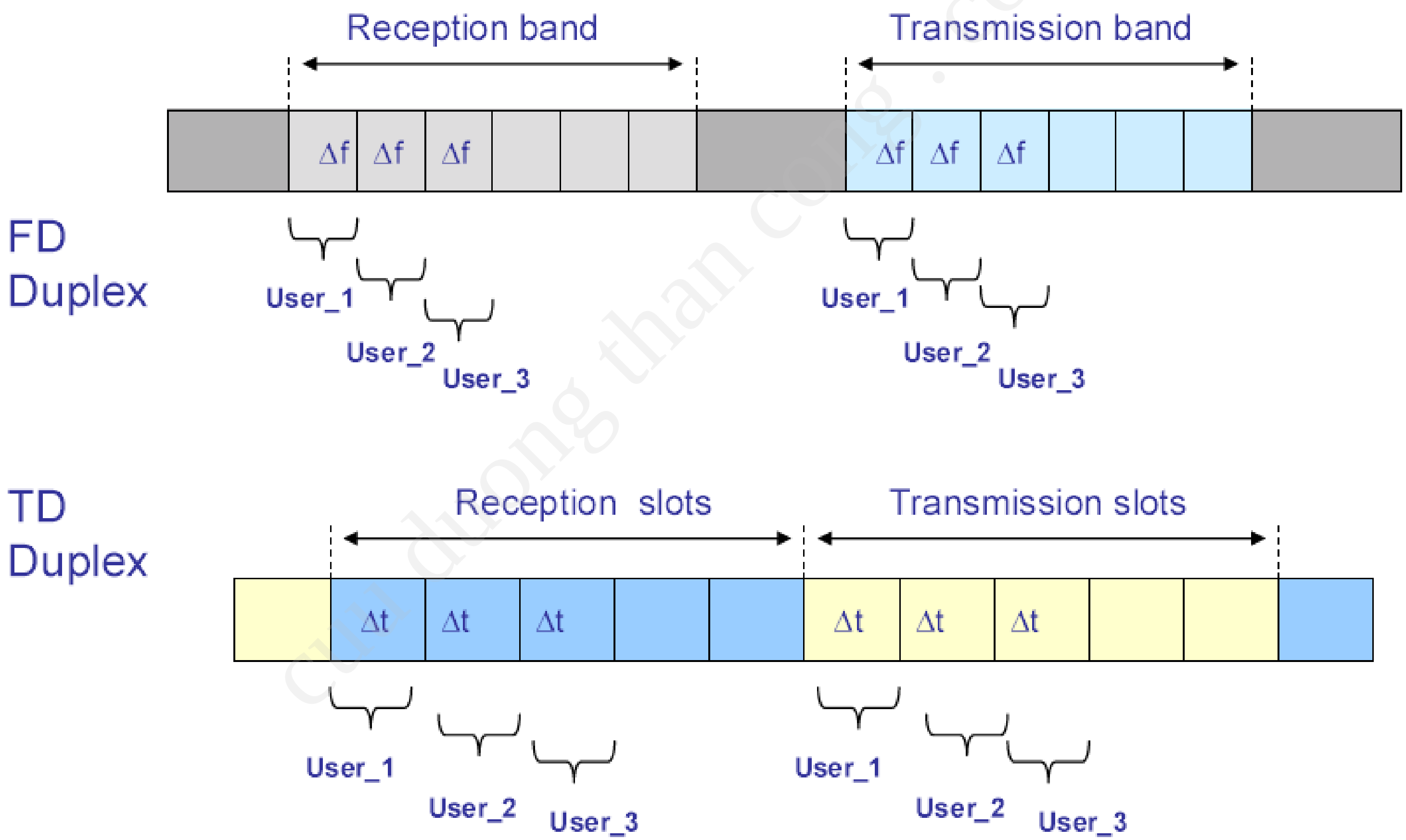
More resistant to strong interferers than DS CDMA, since it is similar to FDMA

In CDMA systems power level control of transmitters is critical, feedback is provided by the base station

Multiple Access (5)

Up-link and down-link by FDD/TDD

/Duplex – ability to transmit and receive simultaneously/



Radio Frequency Metrics (1)

- **<u>Units for RF design</u>:** In RF (microwave) circuits, **power** is usually used to describe signals, noise, or distortion with the typical unit of measure being **decibels above 1 milliwatt** (**dBm**). Voltage and current are expressed as **peak**, **peak-to-peak**, or **root-mean-square** (**rms**). Power in dBm, P_{dBm}, can be related to the power in watts, P_{watt}, as

$$P_{\mathrm{dBm}} = 10\ \log_{10}\left(\frac{P_{\mathrm{watt}}}{1\ \mathrm{mW}}\right)$$

Assuming a sinusoidal voltage waveform, P_{watt} is given by

$$P_{\mathrm{watt}} = \frac{v_{\mathrm{rms}}^2}{R}$$

where R is the resistance the voltage is across. Note also that v_{rms} can be related to the peak voltage v_{pp} by

$$v_{\mathrm{rms}} = \frac{v_{\mathrm{pp}}}{2\sqrt{2}}$$

Radio Frequency Metrics (2)

Example (R=50 Ohm):

v_{pp}	v_{rms}	P_{watt} (50Ω)	P_{dBm} (50Ω)
1 nV	0.3536 nV	2.5×10^{-21}	–176
1 μV	0.3536 μV	2.5×10^{-15}	–116
1 mV	353.6 μV	2.5 nW	–56
10 mV	3.536 mV	250 nW	–36
100 mV	35.36 mV	25 μW	–16
632.4 mV	223.6 mV	1 mW	0
1V	353.6 mV	2.5 mW	+4
10V	3.536V	250 mW	+24

Radio Frequency Metrics (3)

❑ Distortion

Consider a nonlinear system, e.g. LNAs, mixers, described by the following equation:

$$y(t) = \alpha_0 + \alpha_1 x(t) + \alpha_2 x^2(t) + \alpha_3 x^3(t)$$

where $y(t)$ and $x(t)$ is the output and input of the system respectively.

Assume $x(t) = A\cos(\omega t)$, then from equation we get:

$$y(t) = \left(\alpha_0 + \frac{\alpha_2 A^2}{2}\right) + \left(\alpha_1 A + \frac{3\alpha_3 A^3}{4}\right)\cos(\omega t) + \left(\frac{\alpha_2 A^2}{2}\right)\cos(2\omega t) + \left(\frac{\alpha_3 A^3}{4}\right)\cos(3\omega t)$$

Note that the DC (fundamental) magnitude is affected by the even (odd) harmonic components.

The term with the input frequency is called the **fundamental** and the higher order terms the **harmonics**.

Radio Frequency Metrics (4)

Harmonic distortion factors (HD_i) provide a measure for the distortion introduced by each harmonic for a given input signal level (using a single tone at a given frequency).

HD_i is defined as the ratio of the output signal level of the *i*th harmonic to that of the fundamental. The THD is the geometric mean of the distortion factors.

Assuming $\alpha_1 A \gg \dfrac{3\alpha_3 A^3}{4}$, the second harmonic distortion HD_2,

the third harmonic distortion HD_3 and the total harmonic distortion THD are defined as:

$$HD_2 = \frac{\alpha_2 A}{2\alpha_1} \qquad HD_3 = \frac{\alpha_3 A^2}{4\alpha_1}$$

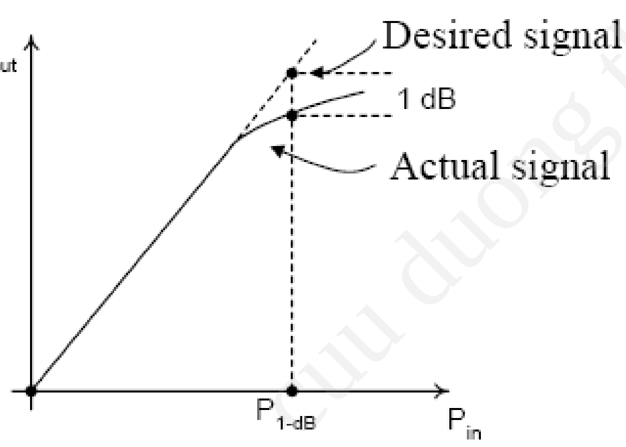
$$THD = \left(HD_2^2 + HD_3^2 + HD_4^2 + \ldots \right)^{1/2}$$

Radio Frequency Metrics (5)

The **1-dB compression point** is defined as the point where the fundamental gain deviates from the ideal small signal gain by 1 dB:

$$20\log\left(\alpha_1 A_{1-dB} + \frac{3\alpha_3 A_{1-dB}^3}{4}\right) = 20\log\left(\alpha_1 A_{1-dB}\right) - 1 = 20\log\left(0.89125\alpha_1 A_{1-dB}\right)$$

(Note that 20 log 0.89125= -1dB, |1-0.89125| = 0.10875)



Definition of the 1-dB compression point

$$\Rightarrow \quad A_{1-dB}^2 = 0.10875\frac{4}{3}\frac{|\alpha_1|}{|\alpha_3|} = k\frac{|\alpha_1|}{|\alpha_3|}$$

Radio Frequency Metrics (6)

Intermodulation distortion (IM):

Consider input signal of a nonlinear system as $x(t) = A\cos\omega_1 t + A\cos\omega_2 t$, then output signal is given by

$$y(t) = \left(\alpha_0 + \alpha_2 A^2\right) + \left(\alpha_1 A + \frac{9\alpha_3 A^3}{4}\right)\cos(\omega_1 t) + \left(\alpha_1 A + \frac{9\alpha_3 A^3}{4}\right)\cos(\omega_2 t) +$$

$$+\left(\frac{\alpha_2 A^2}{2}\right)\cos(2\omega_1 t) + \left(\frac{\alpha_2 A^2}{2}\right)\cos(2\omega_2 t) + \left(\alpha_2 A^2\right)\cos\left[(\omega_1 + \omega_2)t\right] +$$

$$+\left(\alpha_2 A^2\right)\cos\left[(\omega_1 - \omega_2)t\right] + \left(\frac{3\alpha_3 A^3}{4}\right)\cos\left[(2\omega_1 - \omega_2)t\right] + \left(\frac{3\alpha_3 A^3}{4}\right)\cos\left[(2\omega_2 - \omega_1)t\right] +$$

$$+\left(\frac{3\alpha_3 A^3}{4}\right)\cos\left[(2\omega_1 + \omega_2)t\right] + \left(\frac{3\alpha_3 A^3}{4}\right)\cos\left[(2\omega_2 + \omega_1)t\right] +$$

$$+\left(\frac{\alpha_3 A^3}{4}\right)\cos(3\omega_1 t) + \left(\frac{\alpha_3 A^3}{4}\right)\cos(3\omega_2 t)$$

Third order input intercept point IIP_3 is defined as the intercept point of the fundamental component with the third order intermodulation component as

Radio Frequency Metrics (7)

$$\alpha_1 A_{IIP3} = \frac{3\alpha_3 A_{IIP3}^3}{4} \quad \Rightarrow \quad A_{IIP3}^2 = \frac{4}{3}\frac{|\alpha_1|}{|\alpha_3|}$$

Therefore, the input IIP_3 is: $A_{IIP3} = \left[\frac{4}{3}\frac{|\alpha_1|}{|\alpha_3|}\right]^{1/2}$

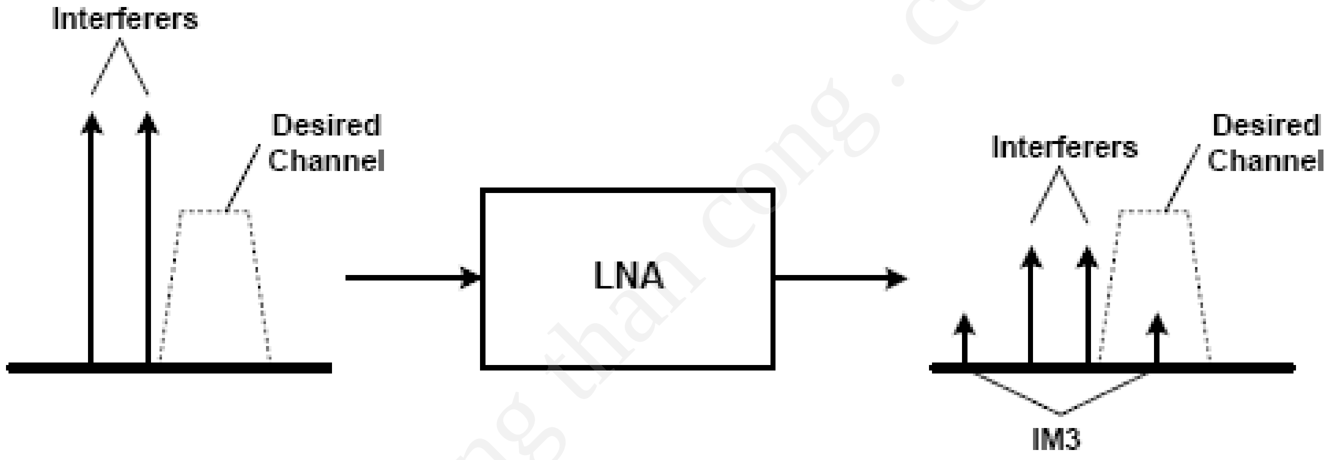
and the output IIP_3: $\alpha_1 A_{IIP3}$

The **third order intermodulation distortion** IM_3 is defined as:

$$IM_3 = \frac{3}{4}\frac{|\alpha_3|}{|\alpha_1|}A^2 = 3HD_3$$

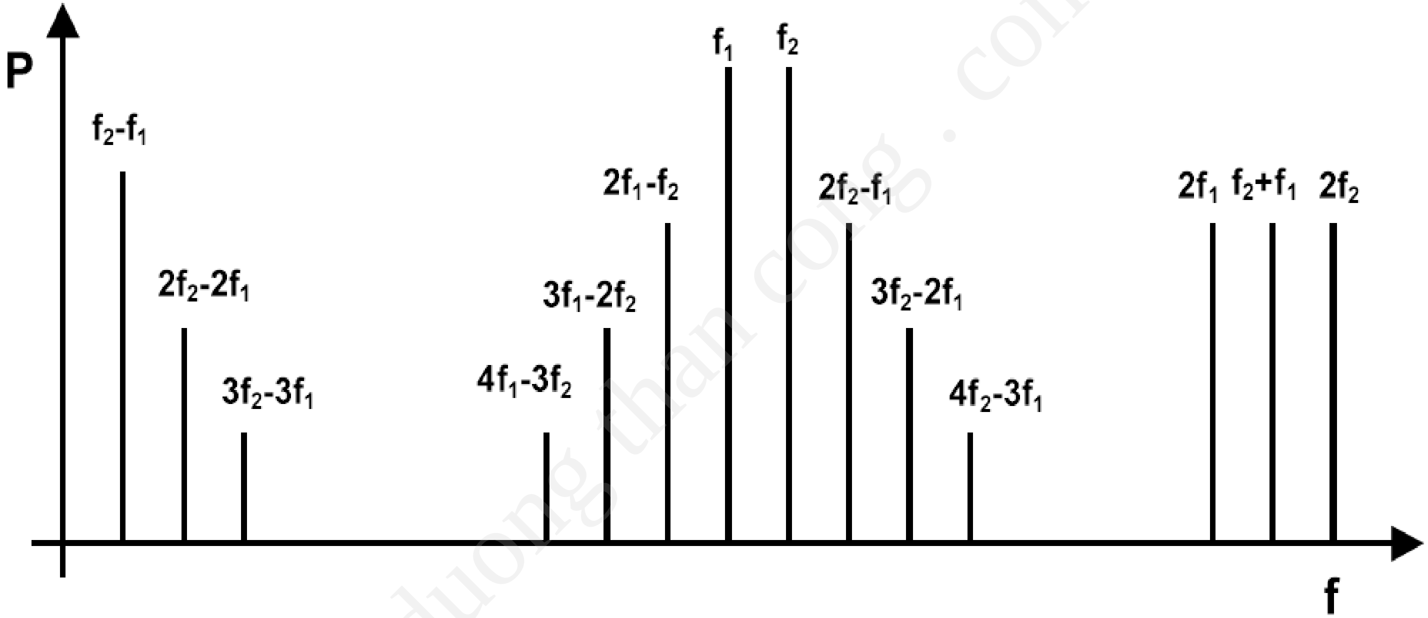
Note that $\frac{A_{IIP3}^2}{A_{1-dB}^2} = 9.195 \quad \Rightarrow \quad A_{IIP3}(dB) \cong A_{1-dB}(dB) + 10$

Radio Frequency Metrics (8)



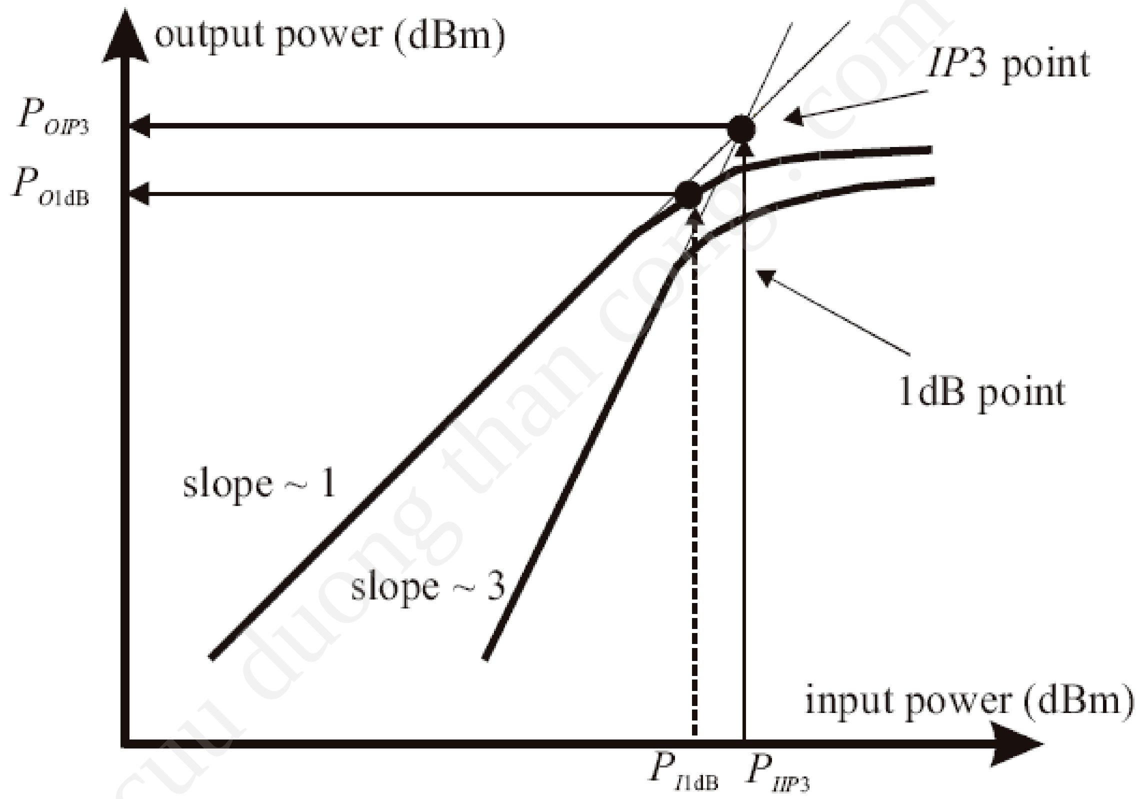
Third order intermodulation (IM3) components corrupt the signal resulting in distortion

Radio Frequency Metrics (9)



Schematic spectrum showing two signals with frequencies f_1 and f_2 and their intermodulation products.

Radio Frequency Metrics (10)

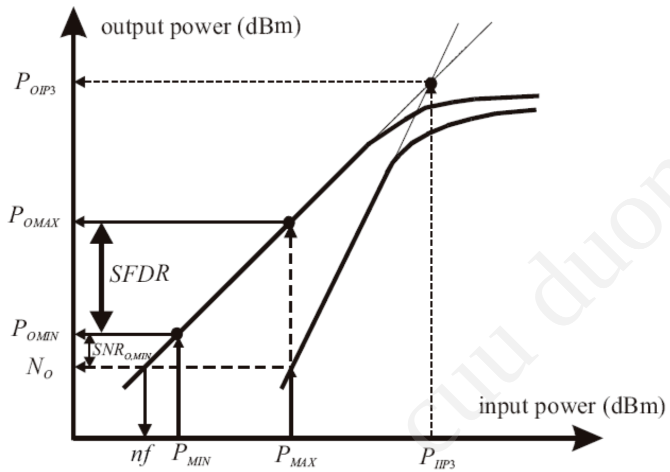


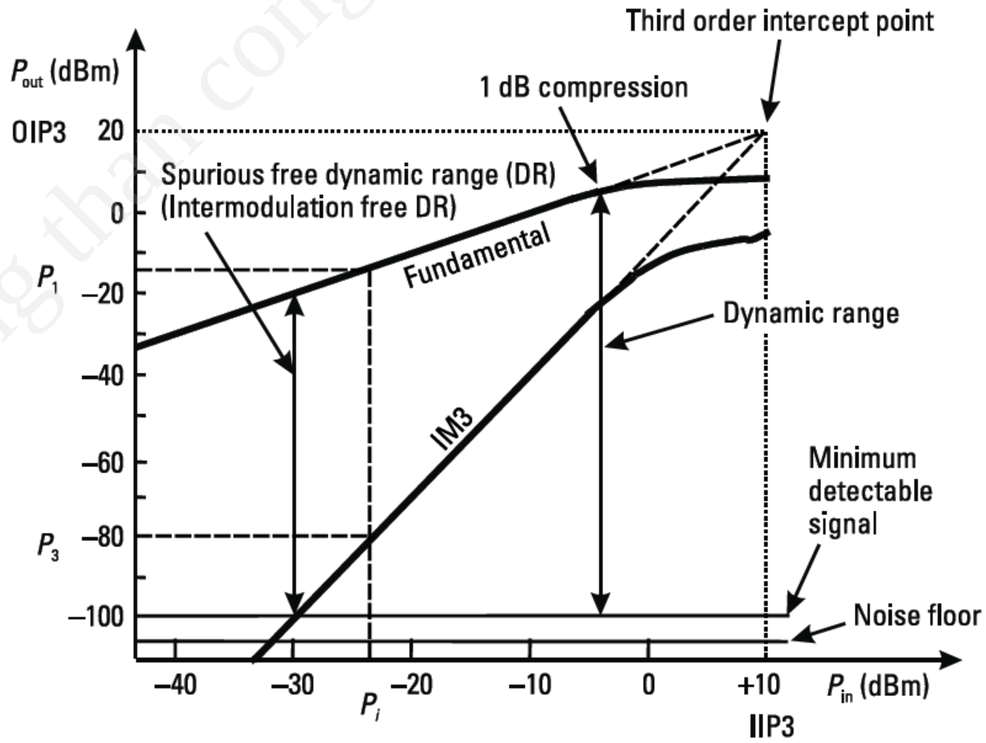
Definition of the third order intercept point

Radio Frequency Metrics (11)

<u>Dynamic range</u>:

There are many definitions for the dynamic range. We define here **spurious free dynamic range** (**SFDR**). The SFDR is the difference, in dB, between the fundamental frequency and the highest spur, which could be an intermodulation harmonic, in the bandwidth of interest.



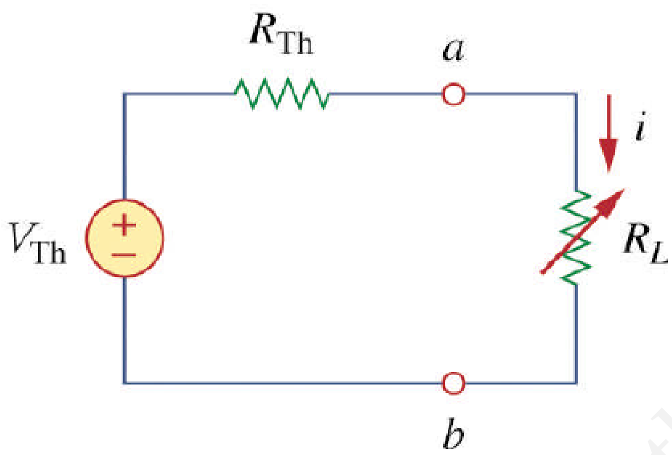


Radio Frequency Metrics (12)

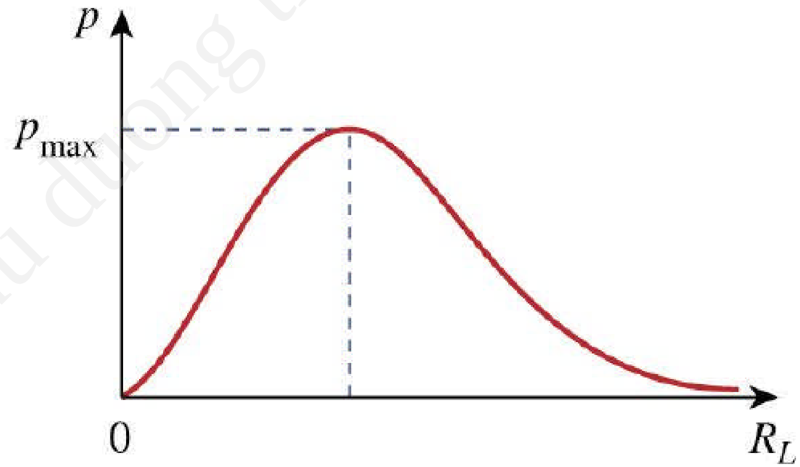
- **Blockers and blocker filtering**: Large unwanted signals can block the desired signal. This can happen when the **desired signal** is **small** and the **undesired signal** is **large**, for example, when the desired signal is far away and the undesired signal is close. If the result is that the receiver is overloaded, the desired signal cannot be received. This situation is known as **blocking**. If the blockers are in the desired frequency band, then filters do not help.

Maximum Power Transfer (1)

❑ **Maximum power transfer:**



$$p_L = i^2 R_L = \left(\frac{V_{Th}}{R_{Th} + R_L} \right)^2 R_L$$



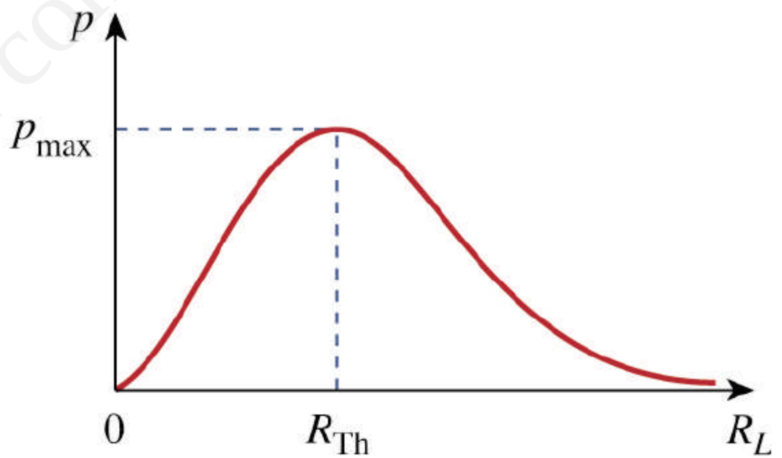
Maximum Power Transfer (2)

Taking the derivative of p_L and setting it equal to zero, we find that

$$\frac{dp_L}{dR_L} = \frac{d}{dR_L}\left[\left(\frac{V_{\text{Th}}}{R_{\text{Th}} + R_L}\right)^2 R_L\right] = V_{\text{Th}}^2\left[\frac{R_{\text{Th}} - R_L}{(R_{\text{Th}} + R_L)^3}\right] = 0$$

This implies $R_{\text{Th}} - R_L = 0$

which yields $R_L = R_{\text{Th}}$



The power delivered when $R_L = R_{\text{Th}}$ is

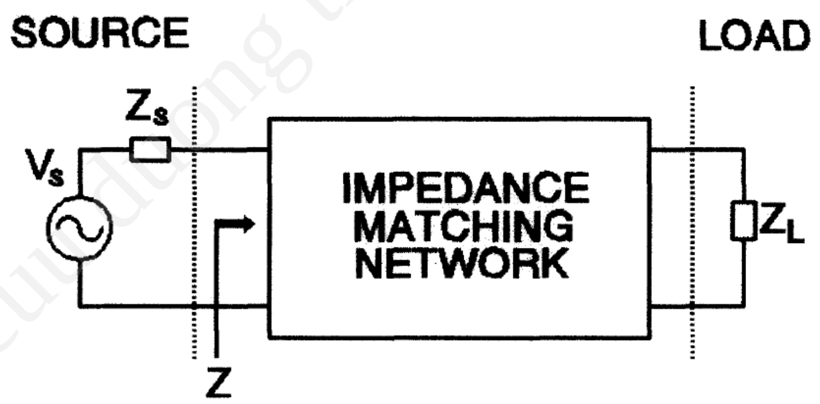
$$p_{\max} = \frac{V_{\text{Th}}^2}{4R_{\text{Th}}}$$

In general, if R_L and R_{Th} are the impedances, then the load impedance R_L will be the **complex conjugate** of the source impedance R_{Th}.

Impedance Matching (1)

- **Impedance matching** is a major problem in RF/microwave circuit design. Impedance matching consists of transforming a load impedance, Z_L , in the optimal working impedance of the signal source Z.

  - Depending on the specific purpose of the circuit, the optimal working impedance (Z) may assure maximum power delivered to the load, maximum efficiency or power gain, minimum distortion of the signal across the load and more.



Impedance Matching (2)

- In a specific case, this optimal impedance may be the **complex conjugate** of the source impedance (Z_s), assuring a **maximum power transfer**, as is usual in **small-signal amplifiers**.
- As an almost general rule, the reactive component of the source impedance must be compensated by a convenient reactance seen at the input of the matching network, so the signal source operates into a **purely resistive load**.
- Mismatching in RF power amplifiers may cause reduced efficiency and/or output power, increased stresses of the active devices, distortion of the output signal and so on.

Impedance Matching (3)

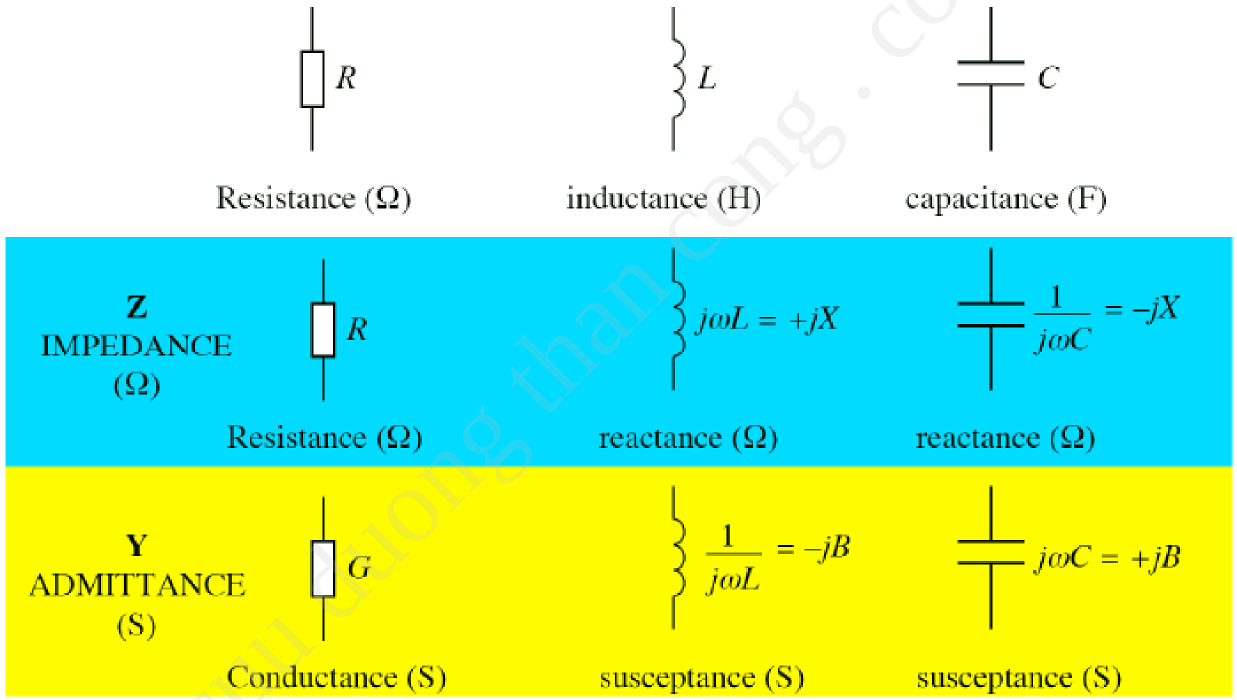
- If the RF circuit operates at a fixed frequency or over a narrow frequency band in comparison with the carrier frequency, the above requirements must be met at only one frequency, and **narrowband matching networks** should be used. Obviously, the matching circuit must contain L and C in order to specify the matching frequency ω_0.

- If the circuit operates over a wide frequency band, the matching requirements (or at least some of them) must be met over the entire frequency range. This requires the use of **broadband matching network.**

- At low frequencies (HF, VHF and UHF), the narrowband impedance matching is usually achieved with **lumped element** circuits (will be studied in this course). At higher frequencies, **distributed element** networks are most often required.

Impedance Matching (4)

❑ Essential revision

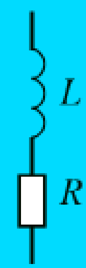


Impedance Matching (5)

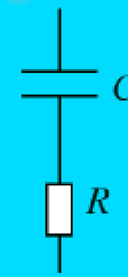
❑ Practical components are lossy

Series:

$$Q = \frac{X}{R} = \frac{1}{RB} = \frac{G}{B}$$



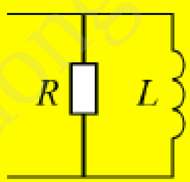
$$Q = \frac{\omega L}{R}$$



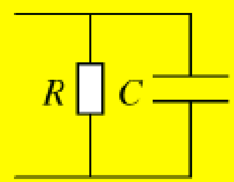
$$Q = \frac{1}{\omega CR}$$

Parallel:

$$Q = \frac{R}{X} = RB = \frac{B}{G}$$



$$Q = \frac{R}{\omega L}$$



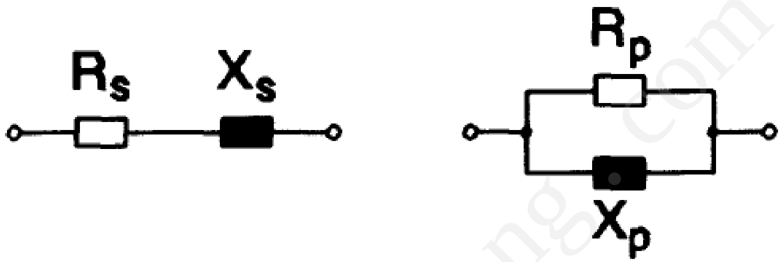
$$Q = \omega CR$$

Higher Q means that it is closer to the ideal L or C.

(Q: Quality factor)

Lumped Element Narrowband Matching Networks (1)

- **Series to parallel conversion and vice versa:**



Assuming that X_s and X_p in the figure are similar elements (i.e., both are either capacitances or inductances), the relations between the elements of the two circuits are given by:

$$R_p = R_s\left[1+\left(\frac{X_s}{R_s}\right)^2\right] \qquad X_p = X_s\left[1+\left(\frac{R_s}{X_s}\right)^2\right]$$

$$R_s = \frac{R_p}{1+\left(\frac{R_p}{X_p}\right)^2} \qquad X_s = \frac{X_p}{1+\left(\frac{X_p}{R_p}\right)^2}$$

Lumped Element Narrowband Matching Networks (2)

Note that taking into account that the quality factor:

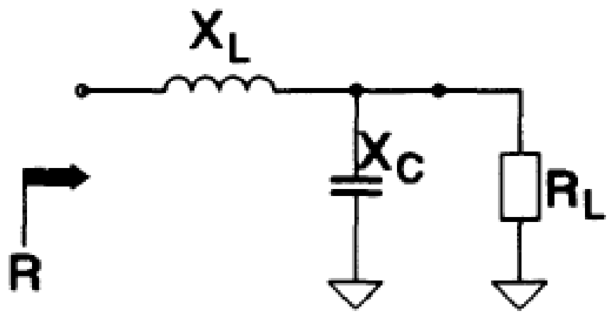
$$Q = \frac{X_s}{R_s} = \frac{R_p}{X_p}$$

Then

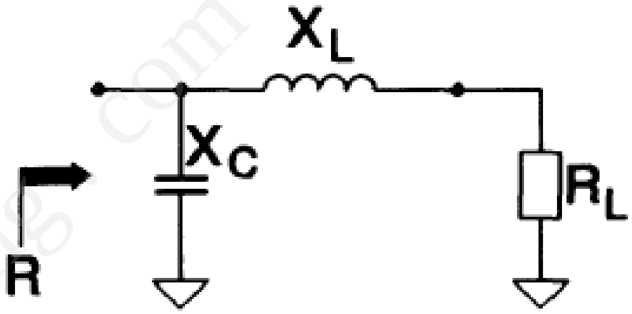
$$R_p = R_s\left(1 + Q^2\right) \qquad X_p = X_s\left(1 + \frac{1}{Q^2}\right)$$

Lumped Element Narrowband Matching Networks (3)

- **<u>Two-reactance matching networks (L matching network):</u>**



(for $\boldsymbol{R < R_L}$)



(for $\boldsymbol{R > R_L}$)

$$X_L = R\sqrt{\frac{R_L}{R} - 1} \qquad X_C = \frac{R_L}{\sqrt{\frac{R_L}{R} - 1}}$$

$$X_L = R_L\sqrt{\frac{R}{R_L} - 1} \qquad X_C = \frac{R}{\sqrt{\frac{R}{R_L} - 1}}$$

Lumped Element Narrowband Matching Networks (4)

The L matching networks in the previous slide have several drawbacks:

a. The design problem has no solution for some combinations of matched impedances.

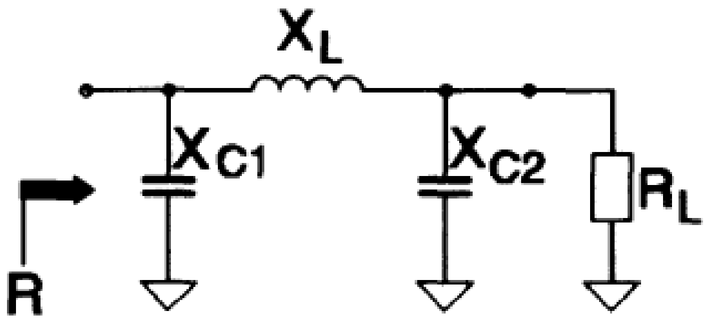
b. The values obtained may be impractical; the values of the capacitors and inductors may be too large or too small.

c. There is no design flexibility. Designers may wish to optimize their designs for other parameters of practical interest, such as harmonic attenuation, power losses, or bandwidth.

The **three-reactance matching networks** are most widely used because they are simple and provide flexibility. Although each network has limitations, one of the circuits usually meets the design requirements with practical component values.

Lumped Element Narrowband Matching Networks (5)

Three-reactance matching networks: Pi matching network



$$X_{C1} = \frac{R}{Q} \qquad X_{C2} = \frac{R_L}{\sqrt{\frac{R_L}{R}\left(1+Q^2\right)-1}}$$

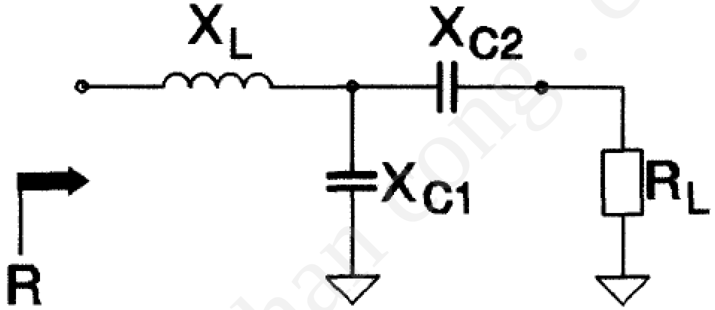
$$X_L = \frac{QR}{1+Q^2}\left[1+\frac{1}{Q}\sqrt{\frac{R_L}{R}\left(1+Q^2\right)-1}\right]$$

- ✓ This circuit can be used only if: $R_L\left(1+Q^2\right) > R$
- ✓ Recommended values of Q usually range from 1 to 10.
- ✓ The pi matching network is widely used in vacuum-tube transmitters to match large resistance values. For small resistance values, the inductance of L becomes unpractically small, while the capacitance of both C_1 and C_2 becomes very large. This circuit is generally not useful in solid-state RF Power Amplifiers where the matched resistances are often small.

Lumped Element Narrowband Matching Networks (6)

- **<u>Three-reactance matching networks: T matching network</u>**

  The T matching network in the below figure is applicable to most solid-state RF Power Amplifiers.



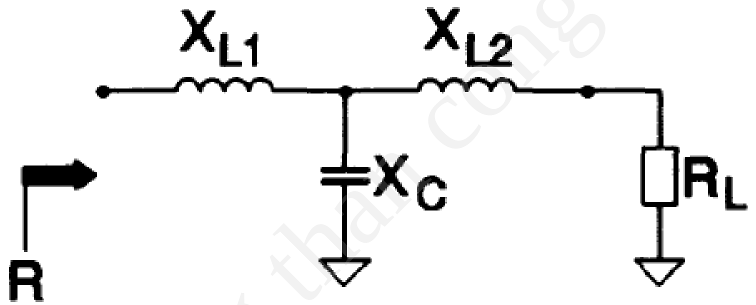
Its design equations are:

$$X_L = QR$$

$$X_{C1} = \frac{R\left(1+Q^2\right)}{Q - \sqrt{\frac{R}{R_L}\left(1+Q^2\right)-1}} \qquad X_{C2} = R_L\sqrt{\frac{R}{R_L}\left(1+Q^2\right)-1}$$

- **<u>Three-reactance matching networks: Two-inductance T matching network</u>**

  Another T matching network with two inductances and is also applicable to many solid-state RF Power Amplifiers.



The design equations are:

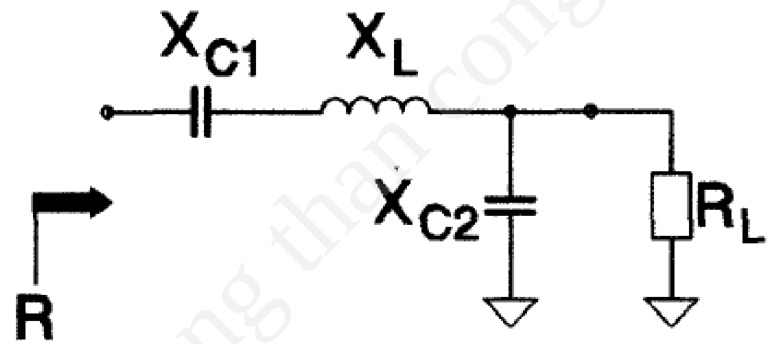
$$X_{L1} = QR$$

$$X_{L2} = R_L\sqrt{\frac{R}{R_L}\left(1+Q^2\right)-1} \qquad X_C = \frac{R\left(1+Q^2\right)}{Q+\sqrt{\frac{R}{R_L}\left(1+Q^2\right)-1}}$$

Lumped Element Narrowband Matching Networks (8)

- **Three-reactance matching networks: Three-reactance L matching network**

  This network is also very useful in solid-state RF Power Amplifiers because it yields practical components for low values of matched resistances.

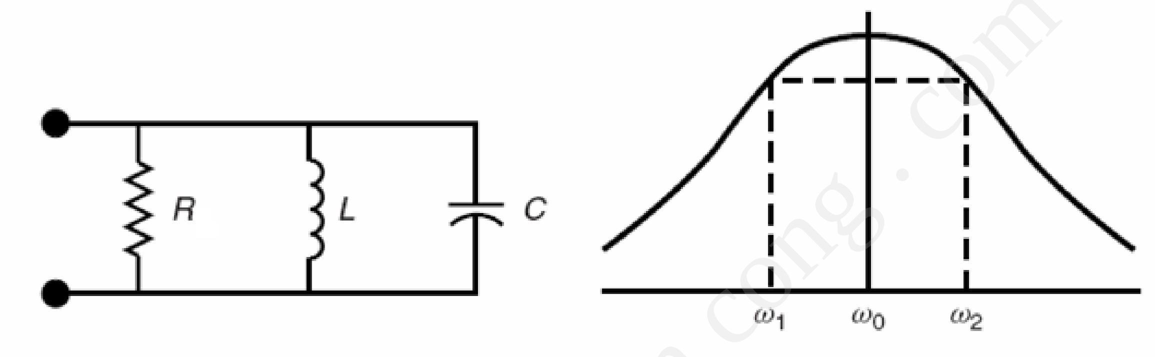


The design equations are:

$$X_{C1} = QR \qquad X_{C2} = R_L\sqrt{\frac{R}{R_L - R}} \qquad X_L = QR + \sqrt{R(R_L - R)}$$

Resonant Circuits (1)

❑ Parallel resonant circuit:



When this circuit is excited by a current source, and the output is terminated with an open circuit, the transfer function is

$$\frac{V_{\text{out}}}{I_{\text{in}}} = \frac{1}{(1/R) + j\omega C - (j/\omega L)}$$

The output voltage, V_{out}, drops from the resonant value by √2 (or 3 dB)

$$\left| \frac{1}{R} + j\omega C - j\omega L \right| = \frac{\sqrt{2}}{R}$$

Resonant Circuits (2)

The two 3 dB frequencies of the resonant circuit:

$$\omega_{1,2} = \omega_0 \left\{ \sqrt{1 + \frac{1}{4Q^2}} \pm \frac{1}{2Q} \right\}$$

The 3 dB bandwidth of the resonant circuit is the difference between the two 3 dB frequencies:

$$\Delta f = \omega_2 - \omega_1 = \frac{1}{RC} \quad \text{rad/s}$$

The resonant frequency is:

$$\omega_0 = 1/\sqrt{LC}$$

and the value of Q given by:

$$Q = \frac{\omega C}{G} \qquad \text{where } G = 1/R$$

or

$$Q = R\sqrt{C/L}$$

Resonant Circuits (3)

<u>(Ideal) parallel-tuned circuit</u>: An ideal parallel-tuned circuit is a paralleled LC circuit that provides **zero conductance** (that is, **infinite impedance**) at the **tuning frequency**, f_0, and **infinite conductance** (**zero impedance**) for **any other frequency**. When connected in parallel to a load resistor, R, the ideal parallel-tuned circuit only allows a **sinusoidal current** (with frequency f_0) to flow through the load. Therefore, the **voltage** across the RLC parallel group is **sinusoidal**, while the total current (that is, the sum of the current through load and the current through the LC circuit) may have any waveform.

A good approximation for the **ideal** parallel-tuned circuit is a circuit with a **very high loaded *Q*** (the higher the Q, the closer the approximation). Note that a high-Q parallel-tuned circuit uses small inductors and large capacitors, which may be a serious limitation in practical applications.

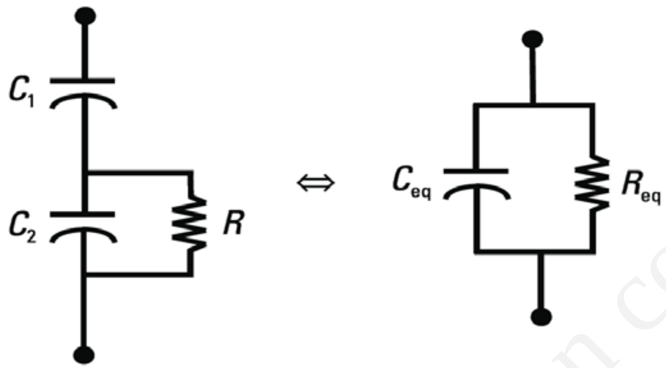
Resonant Circuits (4)

(Ideal) series-tuned circuit: An ideal series-tuned circuit is a series LC circuit that provides **zero impedance** at the **tuning frequency**, f_0, and **infinite impedance** for **any other frequency**. When connected in series to a load resistor, *R*, the ideal series tuned circuit only allows a sinusoidal current with frequency f_0 to flow through the load. Therefore, the **current** through the series RLC group is **sinusoidal**, while the voltage across the RLC group may have any waveform.

A good approximation for the **ideal** series-tuned circuit is a circuit with a **very high loaded *Q*** (the higher the *Q*, the closer the approximation). Note that a high-Q series-tuned circuit must use large inductors and small capacitors, which may be a serious limitation in practical applications.

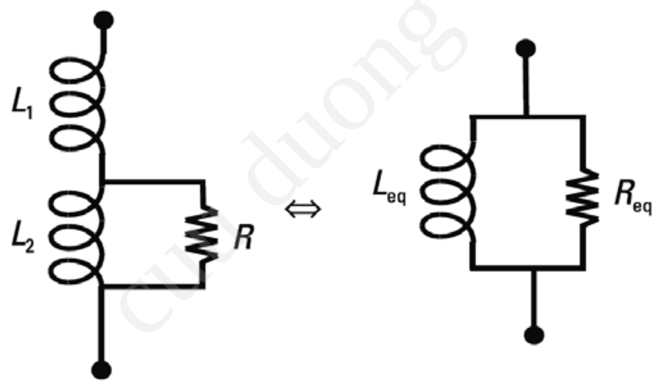
Resonant Circuits (5)

❑ Tapped capacitors and inductors



$$R_{eq} \approx R\left(\frac{C_1 + C_2}{C_1}\right)^2$$

$$C_{eq} \approx \left(\frac{1}{C_1} + \frac{1}{C_2}\right)^{-1}$$



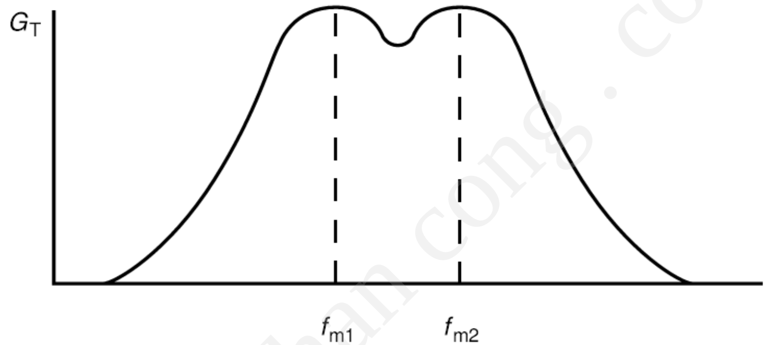
$$R_{eq} \approx R\left(\frac{L_1 + L_2}{L_2}\right)^2$$

$$L_{eq} \approx L_1 + L_2$$

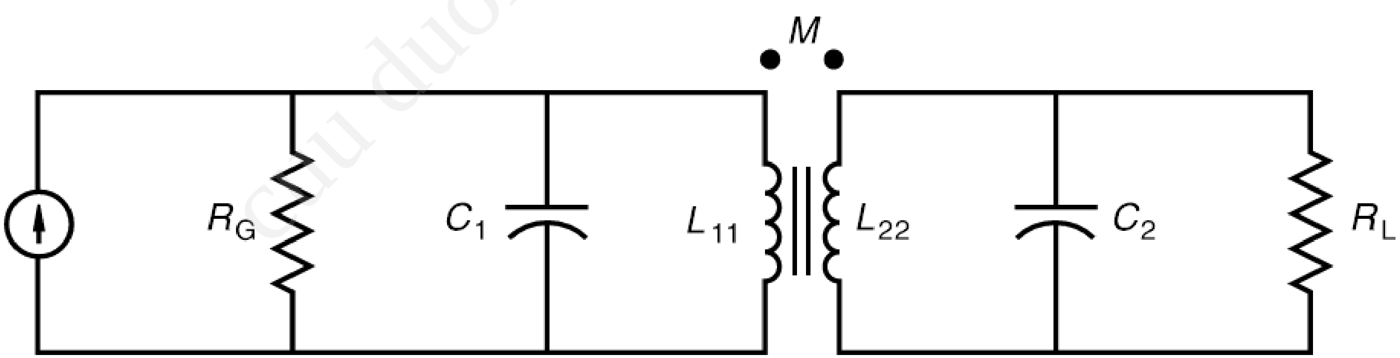
Resonant Circuits (6)

❑ **Double-tuned matching circuits:**

Specify the bandwidth by two frequencies ω_{m1} and ω_{m2}



The construction of a double-tuned circuit typically includes a real transformer and two resonating capacitor

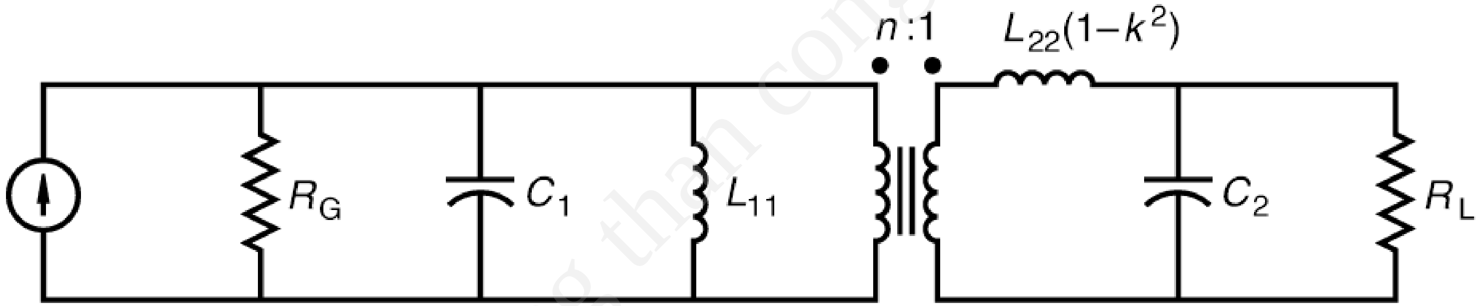


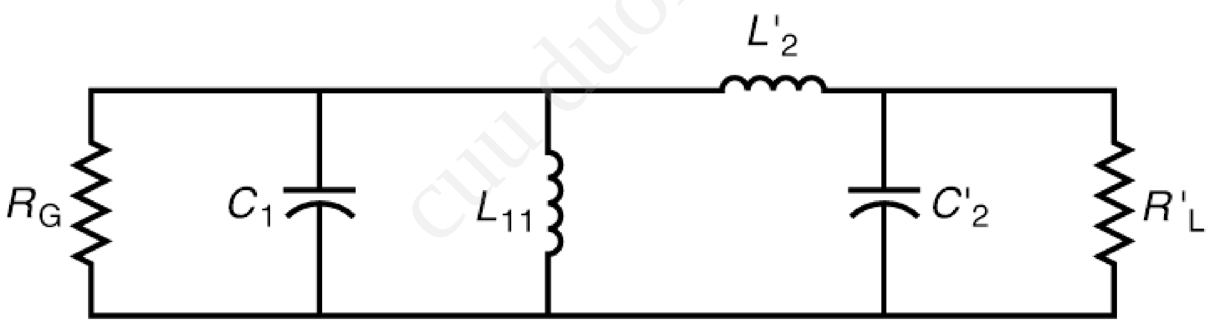
Resonant Circuits (7)

Transformer turn ratio n and coupling coefficient k are related by

$$n = \sqrt{\frac{L_{11}}{k^2 L_{22}}}$$

Equiqualent model:





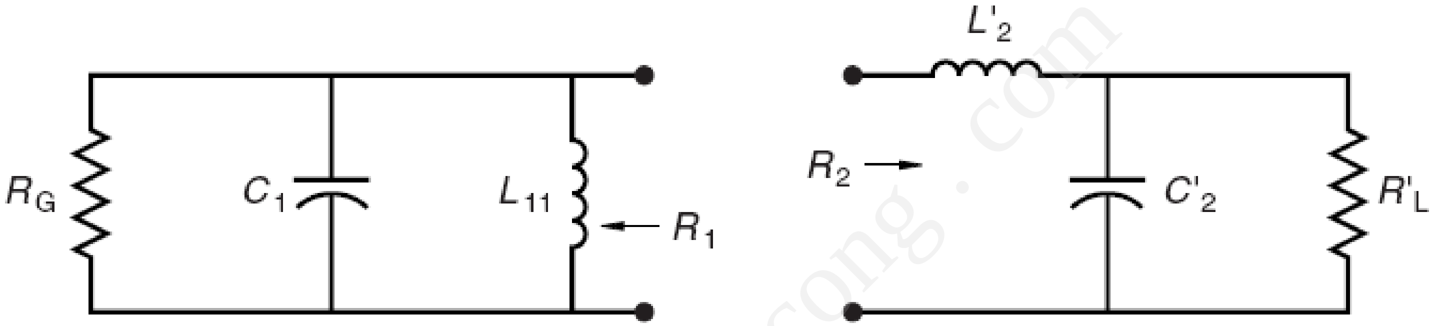
$$L_2' = L_{11}\left(\frac{1}{k^2} - 1\right)$$

$$C_2' = \left(\frac{L_{11}}{k^2 L_{22}}\right) C_2$$

$$R_L' = \frac{L_{11}}{k^2 L_{22}} R_L$$

Resonant Circuits (8)

Exact match is to be achieved at two given frequencies f_{m1} and f_{m2}



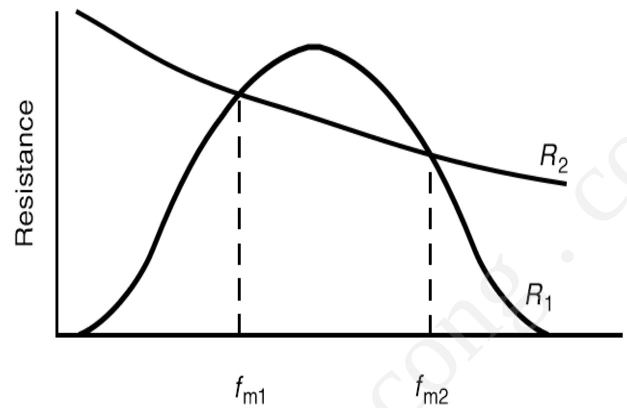
Observe that

- R_1 resonates at certain frequency, but it is usually less than R_G
- R_2 decreases monotonically with frequency

Therefore, if R_L is sufficiently small, there will be two frequency values where $R_1 = R_2$

<u>Our objective</u> is to match R_G and R_L over a bandwidth B centered at f_0, usually with alowable ripple in the pass band. A design procedure for the parallel double-tuned circuit is summarized below

Resonant Circuits (9)



1. Determine f_{m1} and f_{m2}:

$$\Delta f \approx \sqrt{2}(f_{m2} - f_{m1})$$
$$f_0 \approx \sqrt{f_{m1} f_{m2}}$$

The minimum pass band gain for the filter is dependent on the difference between the match frequencies:

$$G_{\text{Tmin}} = \frac{4 f_{m2}/f_{m1}}{(f_{m2}/f_{m1})^2 + 2 f_{m2}/f_{m1} + 1}$$

Resonant Circuits (10)

2. Determine the actual transducer gain for the given ripple factor:

$$G_T = 10^{-\text{ripple factor (dB)}/10}$$

3. Find the resistance ratio if $G_T > G_{T\min}$, the pass band ripple specification can be met:

$$r = \frac{1 + |1 - G_T|^{1/2}}{1 - |1 - G_T|^{1/2}}$$

4. Calculate the Q_2 at the two matching frequencies:

$$Q_{2-m1}^2 = r\frac{f_{m1}}{f_{m2}} - 1$$

$$Q_{2-m2}^2 = r\frac{f_{m2}}{f_{m1}} - 1$$

Resonant Circuits (11)

5. Solve the following simultaneous equations for L'_2 and C'_2:

$$-\omega_{m1}L'_2 + \frac{1}{\omega_{m1}C'_2} = |Q_{2-m1}|\frac{R_G}{1+Q^2_{2-m1}}$$

$$+\omega_{m2}L'_2 + \frac{1}{\omega_{m2}C'_2} = |Q_{2-m2}|\frac{R_G}{1+Q^2_{2-m2}}$$

6. Find the value for R'_L:

$$R'_L = \frac{1+Q^2_{2-m1}}{\omega^2_{m1}C'^2_2 R_G}$$

7. Calculate the input susceptance of the right-hand side where $G'_L = 1/R'_L$:

$$B_{m1} = Im\left\{\frac{1}{j\omega_{m1}L'_2 + (1/G'_L + j\omega_{m1}C'_2)}\right\}$$

$$B_{m2} = Im\left\{\frac{1}{j\omega_{m2}L'_2 + (1/G'_L + j\omega_{m2}C'_2)}\right\}$$

Resonant Circuits (12)

8. Solve the following simultaneous equations for L_{11} and C_1:

$$\frac{1}{\omega_{m1}L_{11}} - \omega_{m1}C_1 = |B_{m1}|$$

$$\frac{1}{\omega_{m2}L_{11}} - \omega_{m2}C_1 = |B_{m2}|$$

9. Find the transformer coupling coefficient, and hence L_{22} and C_2:

$$k = \frac{1}{\sqrt{1 + L_2'/L_{11}}}$$

$$L_{22} = \frac{L_{11}R_L}{k^2 R_L'}$$

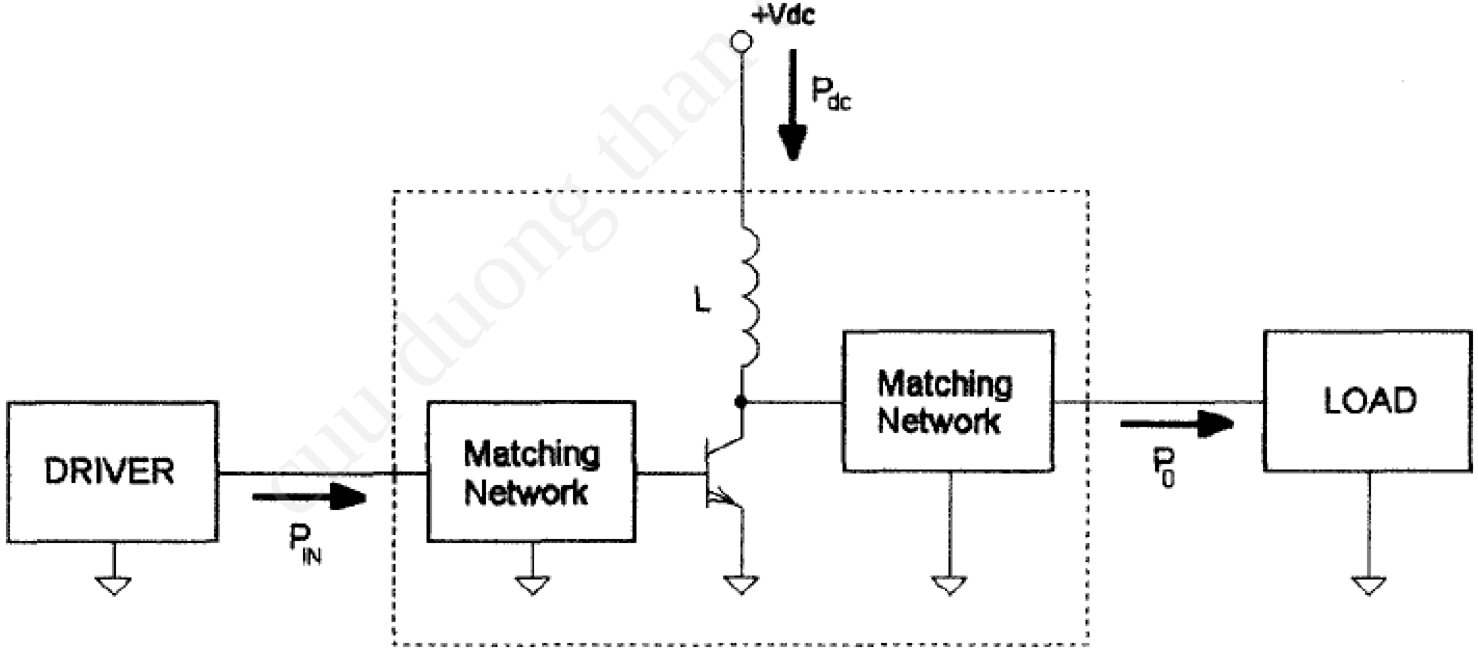
$$C_2 = \frac{L_{11}}{k^2 L_{22}} C_2'$$

Chapter 2:

RF Power Amplifiers

Definitions (1)

- **<u>Efficiency</u>**: Efficiency is a crucial parameter for RF power amplifiers. It is important when the available input power is limited, such as in battery-powered portable or mobile equipment. It is also important for high-power equipment where the cost of the electric power over the lifetime of the equipment and the cost of the cooling systems can be significant compared to the purchase price of the equipment.



Definitions (2)

Collector efficiency: Collector efficiency is a term more appropriate for amplifiers using bipolar transistors (BJTs), although it is often used for any RF power amplifiers. Some authors prefer to use **plate efficiency** for amplifiers using vacuum tubes or **drain efficiency** for amplifiers using MOSFETs or, simply refer to it as **efficiency**. Collector efficiency is defined as

$$\eta = \frac{P_0}{P_{dc}}$$

where P_0 is the RF output power (dissipated into the load) and $P_{dc} = V_{dc}I_{dc}$ is the input power supplied by the dc supply to the collector (or drain /plate) circuit of the power amplifier. P_0 usually includes both the RF fundamental power and the harmonics power. In many applications, harmonic suppression filters are included in the output-matching network. Because the harmonic power is negligible, the RF fundamental power is a very good approximation for P_0.

Definitions (3)

<u>Overall efficiency:</u> Although it is a very convenient measure of a circuit's performance, collector efficiency does not account for the **drive power** required, which may be quite substantial in a power amplifier. **Power gains** (that is the **ratio of output power to drive power**) of 10 dB or less are common at high RF frequencies. In general, RF power amplifiers designed for high collector efficiency tend to achieve a low power gain, which is a disadvantage for the overall power budget.

From a practical standpoint, a designer's goal is to minimize the total dc power required to obtain a certain RF output power. The **overall efficiency** is defined as

$$\eta_{OVERALL} = \frac{P_0}{P_{dc} + P_{IN}} = \frac{P_0}{P_{dc} + \frac{P_0}{G_P}}$$

where the power gain is

$$G_P = \frac{P_0}{P_{IN}}$$

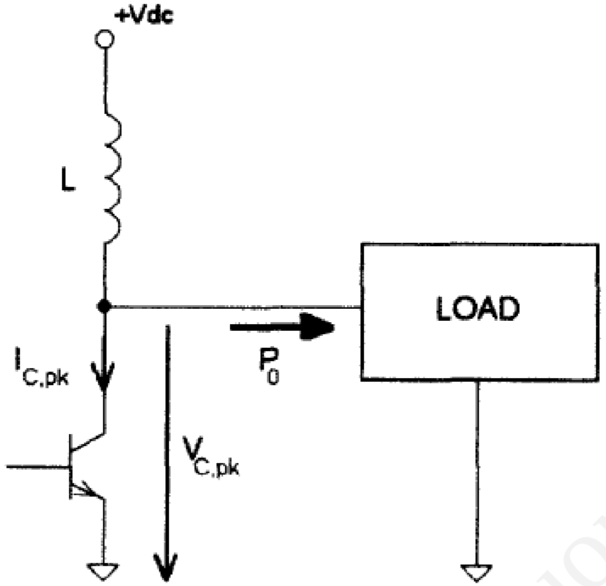
Definitions (4)

<u>**Power-added efficiency**</u>**:** Power-added efficiency is an alternative definition that includes the effect of the drive power used frequently at RF/microwave frequencies and is defined as

$$\eta_{POWER-ADDED} = \frac{P_0 - P_{IN}}{P_{dc}} = \frac{P_0 - \dfrac{P_0}{G_P}}{P_{dc}}$$

Definitions (5)

Power output capability: The power output capability, C_P, provides a means of comparing different types of power amplifiers or amplifier designs.



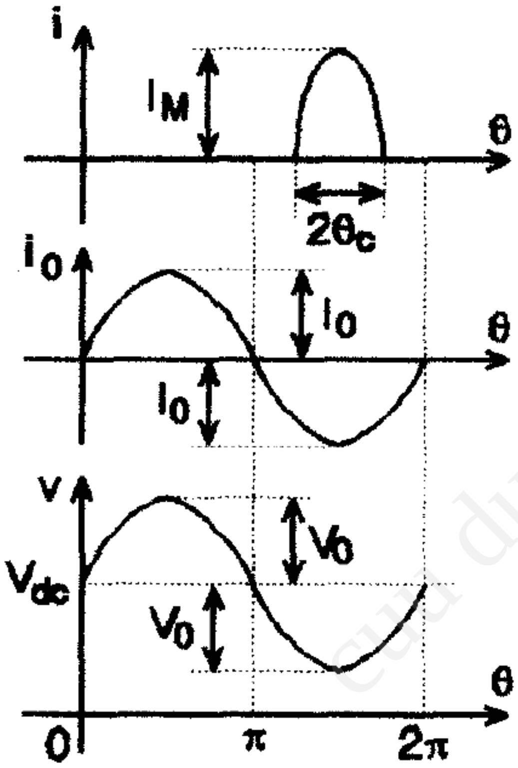
If P_0 is the RF output power, $I_{C,pk}$ is the peak collector current, $V_{C,pk}$ is the peak collector voltage, and N is the number of transistors in circuit, then the **power output capability** is given by

$$C_P = \frac{P_0}{N I_{C,pk} V_{C,pk}}$$

Power transistors are the most expensive components in power amplifiers. Designers are constrained to use the lowest cost transistors. This means the devices have to be used as close as possible to their maximum voltage and current ratings. Therefore, the larger the power output capability of the circuit, the cheaper its practical implementation.

Class C RF Power Amplifier (1)

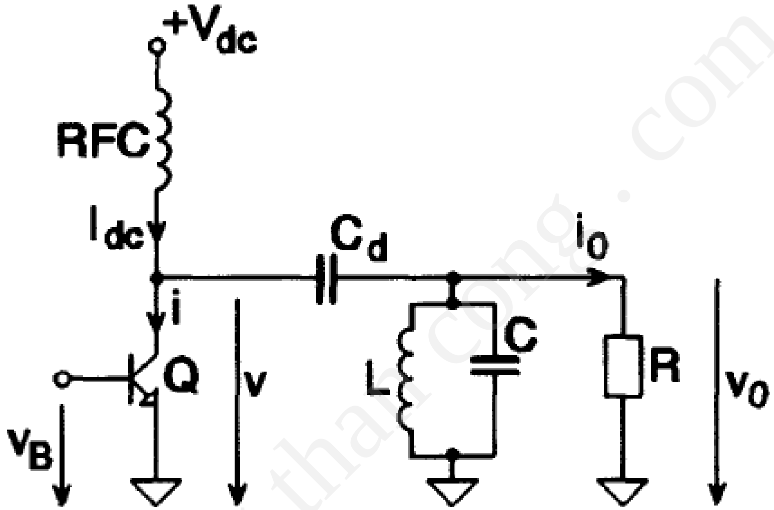
- **<u>Conductance angle</u>**: The portion of the RF cycle the device spends in its active region is the **conduction angle** and is denoted by $2\theta_C$. Based on the conduction angle, the amplifiers are generally classified as:



- **Class A amplifiers**, if $2\theta_C = 360°$. The active device is in its active region during the entire RF cycle.
- **Class AB amplifiers**, if $180° < 2\theta_C < 360°$.
- **Class B amplifiers**, if $2\theta_C = 180°$.
- **Class C amplifiers**, if $2\theta_C < 180°$.

Class C RF Power Amplifier (2)

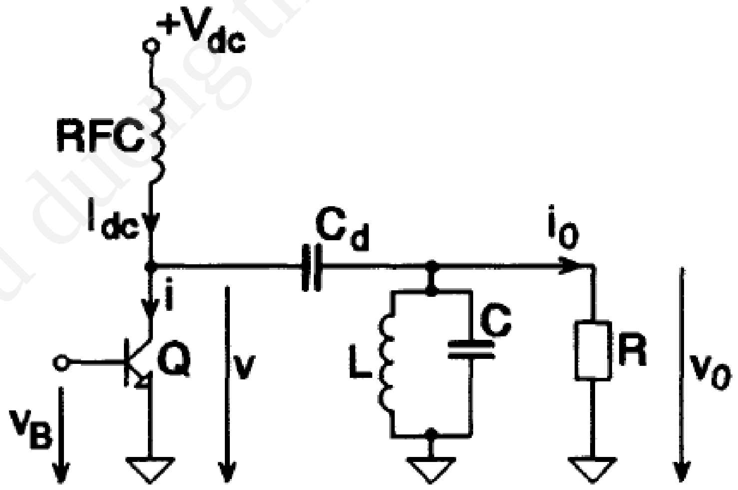
- **<u>Basic circuit of single-ended class A, AB, B, or C amplifier</u>**:



This is a single-ended circuit, and the transistor operates in the common emitter (CE) configuration (common-base configurations are also possible). Variations among practical circuits operating in **different classes may occur in the base-bias or drive circuits**. The collector circuit includes an RF choke (RFC) that provides a DC input current, I_{dc}, a DC blocking capacitor, C_d (short-circuit at the operating frequency and its harmonics), the load resistor, R, and a parallel resonant LC circuit tuned to the operating frequency ω_0.

Class C RF Power Amplifier (3)

The DC component of the collector current $i(\theta)$ flows through the RFC and then through the DC-power supply. The variable component of $i(\theta)$ flows through DC-blocking capacitor C_d and through the parallel RLC tuned circuit. The tuned circuit provides a zero impedance path to ground for the harmonic currents contained in $i(\theta)$ and only the **fundamental component** of $i(\theta)$ flows through the load resistance. As a result, the **output voltage** is a **sinusoidal waveform**. This requires the use of a parallel resonant circuit (or an equivalent band-or low-pass filter).



Class C RF Power Amplifier (4)

The collector current is a periodical waveform described by

$$i(\theta) = \begin{cases} \dfrac{I_M(\cos\theta - \cos\theta_c)}{1 - \cos\theta_c} & -\theta_c + 2k\pi \le \theta \le \theta_c + 2k\pi \quad k \in Z \\ 0 & \text{otherwise} \end{cases}$$

Its Fourier analysis results:

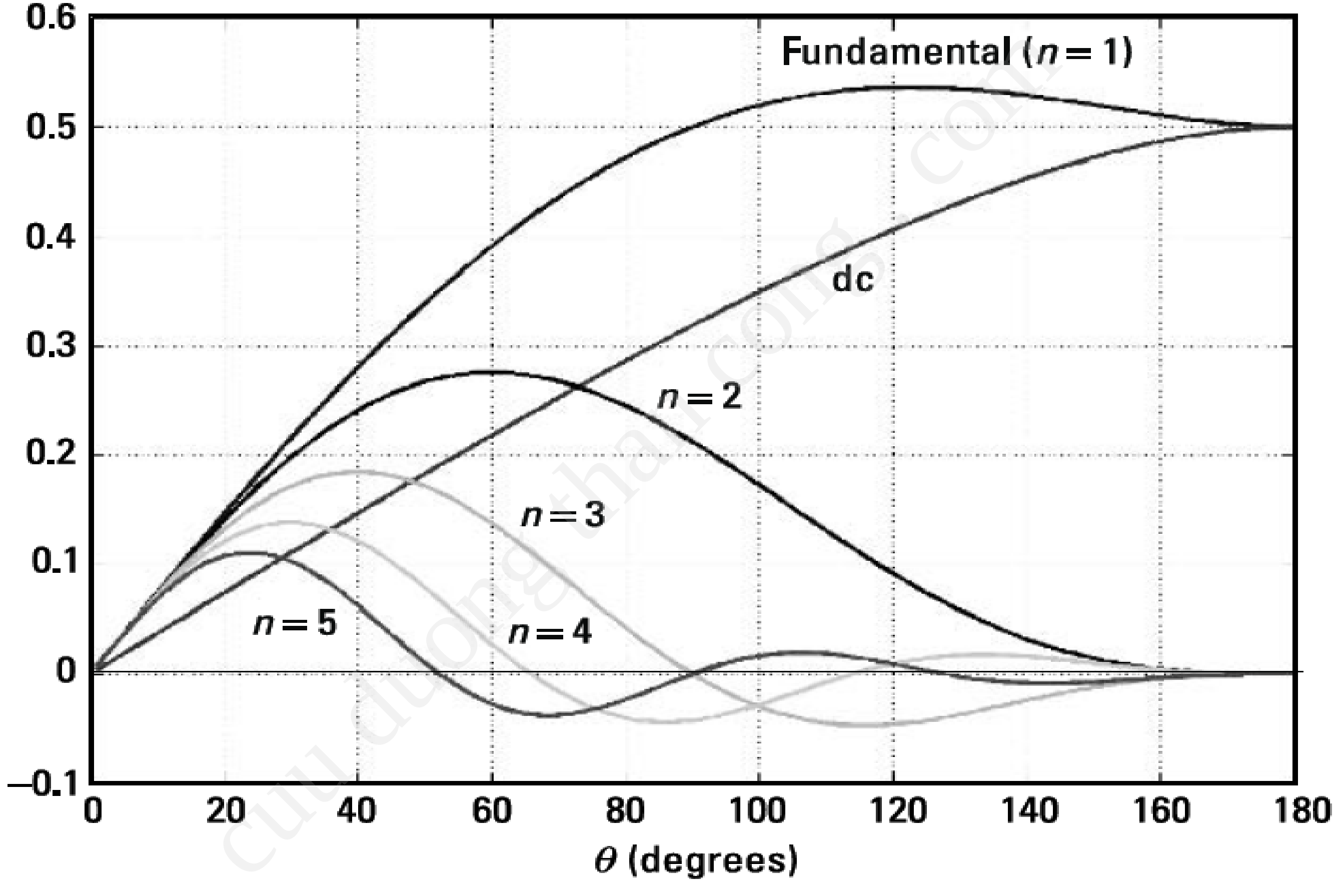
$$i(\theta) = I_M \sum_{n=0}^{\infty} \alpha_n(\theta_c)\cos n\theta$$

where

$$\alpha_0(\theta_c) = \frac{\sin\theta_c - \theta_c\cos\theta_c}{\pi(1 - \cos\theta_c)} \qquad \alpha_1(\theta_c) = \frac{\theta_c - \sin\theta_c\cos\theta_c}{\pi(1 - \cos\theta_c)}$$

$$\alpha_n(\theta_c) = \frac{\dfrac{\sin(n-1)\theta_c}{n-1} - \dfrac{\sin(n+1)\theta_c}{n+1}}{n\pi(1 - \cos\theta_c)} \qquad n = 2, 3, \ldots$$

Class C RF Power Amplifier (5)



Fourier series coefficients α_n versus the conduction angle

Class C RF Power Amplifier (6)

Due to the ideal tuned circuit, the output current (flowing through the load resistance *R*) is sinusoidal and its amplitude is given by

$$I_0 = I_M \alpha_1(\theta_c)$$

As a result, the output voltage is also sinusoidal, with the amplitude $V_0 = RI_0$. The collector voltage is

$$v(\theta) = V_{dc} + V_0 \cos\theta = V_{dc} + RI_M \alpha_1(\theta_c)\cos\theta$$

The DC input power P_{dc} and the collector efficiency η are

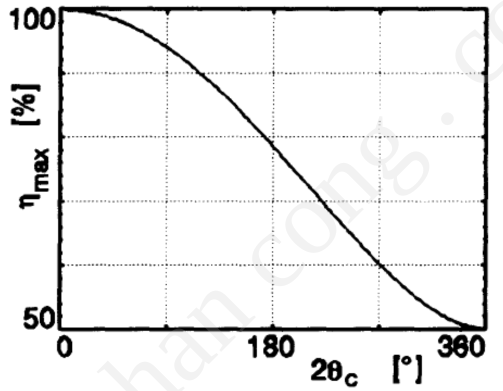
$$P_{dc} = V_{dc} I_{dc} = V_{dc} I_M \alpha_0(\theta_c)$$

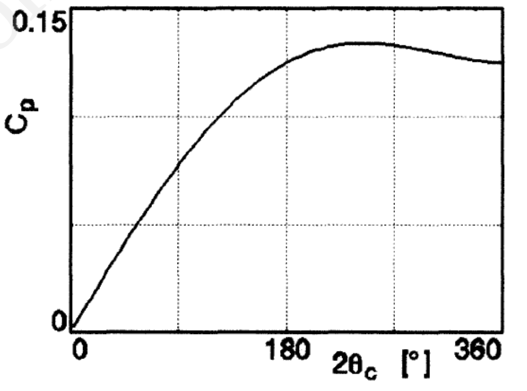
$$\eta = \frac{P_0}{P_{dc}} = \frac{V_0^2}{2RV_{dc} I_M \alpha_0(\theta_c)} = \frac{V_0}{V_{dc}} \frac{\alpha_1(\theta_c)}{2\alpha_0(\theta_c)} = \frac{V_0}{V_{dc}} \frac{\theta_c - \sin\theta_c \cos\theta_c}{2(\sin\theta_c - \theta_c \cos\theta_c)}$$

Class C RF Power Amplifier (7)

The maximum theoretical collector efficiency (obtained for $V_0 = V_{dc}$) varies with the conduction angle as

$$\eta_{max} = \frac{\theta_c - \sin\theta_c \cos\theta_c}{2(\sin\theta_c - \theta_c \cos\theta_c)}$$





If $V_0 = V_{dc}$, the peak collector voltage is $v_{max} = 2\,V_{dc}$ and the peak collector current is given by

$$i_{max} = i_M = \frac{V_{dc}}{R\alpha_1(\theta_c)}$$

The output power P_0 and the power output capability C_P are

$$P_0 = \frac{V_0^2}{2R} = \frac{V_{dc}^2}{2R} \qquad C_P = \frac{P_0}{v_{max} i_{max}} = \frac{\alpha_1(\theta_c)}{4}$$

Class C RF Power Amplifier (8)

Comments:

a. The **collector efficiency** is higher in Class C amplifiers than in Class A, AB, or B amplifiers, and it increases as the conduction angle decreases. If $\theta_C \to 0$, then $\eta \to 100\%$.

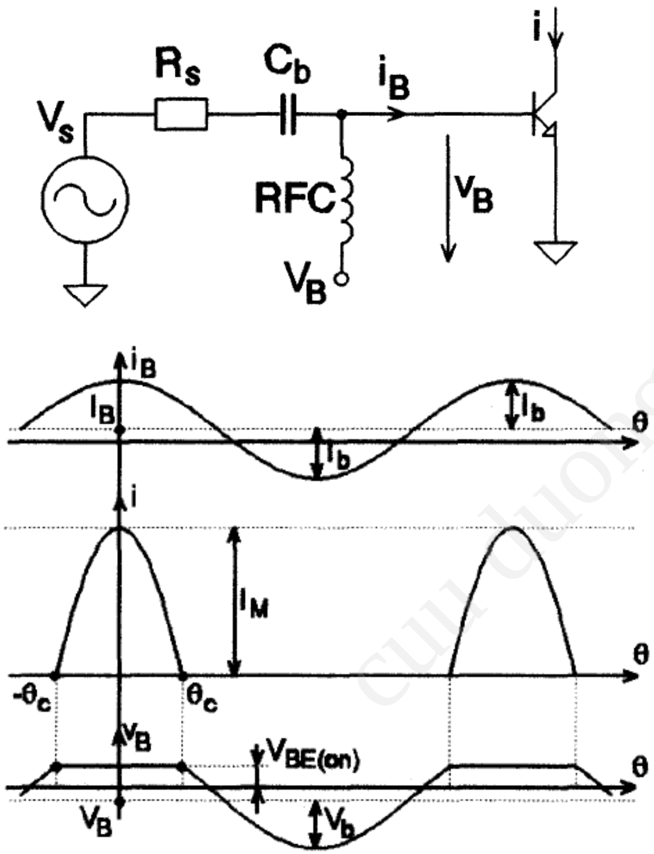
b. The **power output capability** of Class C amplifiers is lower than 0.125 (as obtained in Class A or B circuits) and decreases as the conduction angle decreases.

c. As a result, the **choice of conduction angle** would be a tradeoff among **collector efficiency**, **peak value of the collector current**, and **power gain**.

Class	η_{max}	$\frac{P_0}{V_{dc}^2 / R}$	$\frac{v_{max}}{V_{dc}}$	$\frac{i_{max}}{I_{dc}}$	C_P
A	50%	0.5	2	2	0.125
B	78.5%	0.5	2	π	0.125
C	$\frac{\alpha_1(\theta_c)}{2\alpha_0(\theta_c)}$	0.5	2	$\frac{1}{\alpha_0(\theta_c)}$	$\frac{\alpha_1(\theta_c)}{4}$

Class C RF Power Amplifier (9)

- **DC bias**: The conduction angle in a Class C amplifier is controlled by a DC-bias voltage V_B applied to the base, and an amplitude V_b of the signal across the base-emitter junction.



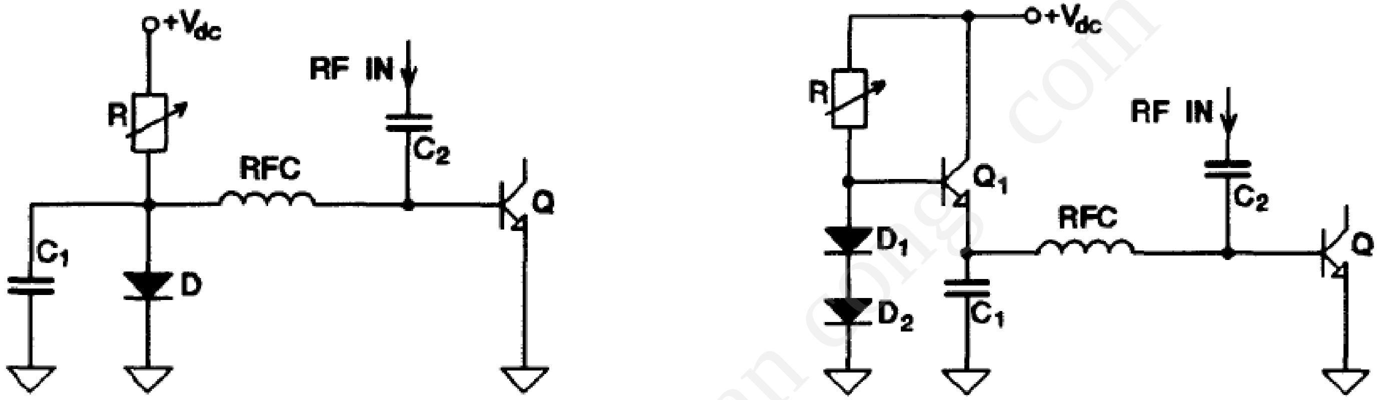
For $-\theta_c < \theta < \theta_c$, the transistor is in its active region. Consequently, the voltage across its base-emitter junction is $V_{BE(on)} \approx 0.7$ and

$$V_{BE(on)} = v_B(\theta_c) = V_B + V_b \cos\theta_c$$

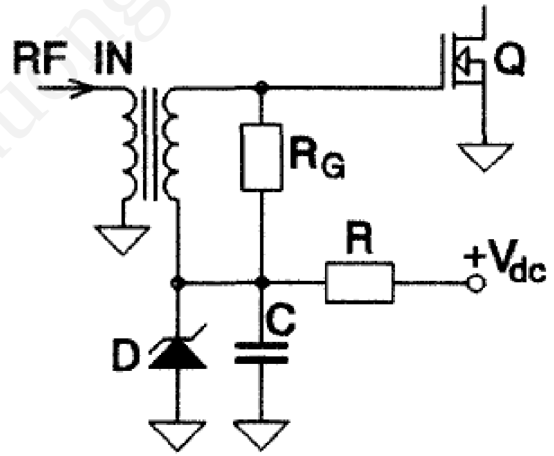
This equation allows calculation of the required bias voltage, V_B, in the base circuit.

Class C RF Power Amplifier (10)

Simple bias circuits for BJTs



Simple bias circuits for MOSFETs



Class C RF Power Amplifier (11)

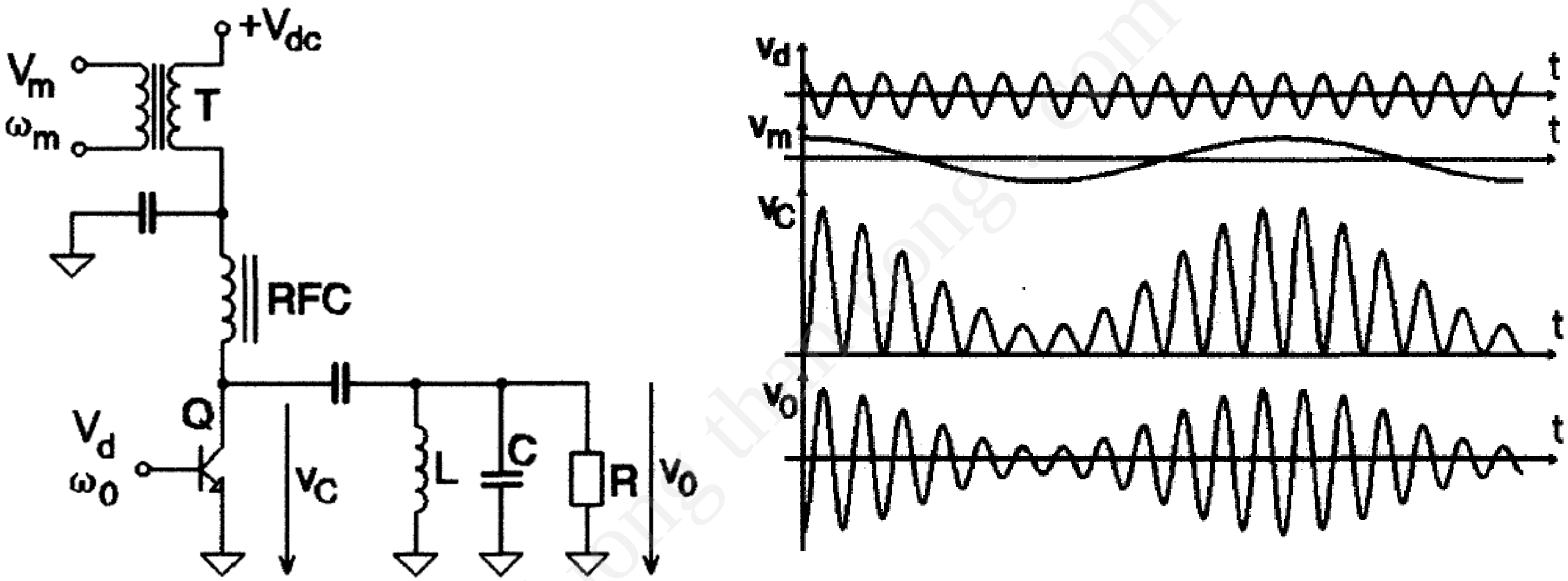
- **<u>Practical considerations</u>:** The effects of V_{sat} on the performance of Class C amplifiers are determined as

$$P_{0,max} = \frac{(V_{dc} - V_{sat})^2}{2R} \qquad \eta_{max} = \left(1 - \frac{V_{sat}}{V_{dc}}\right)\frac{\theta_c - \sin\theta_c \cos\theta_c}{2(\sin\theta_c - \theta_c \cos\theta_c)}$$

$$C_P = \frac{\alpha_1(\theta_c)}{2}\frac{V_{dc} - V_{sat}}{2V_{dc} - V_{sat}}$$

Amplitude Modulation (1)

❑ Amplitude modulation (AM) using collector-modulated RF amplifier



The modulating signal (information): $v_m(t) = V_m \cos\omega_m t$ is used to produce a **time-varying collector-supply voltage** for the RF amplifier:

$$v_C(t) = V_{dc} + V_m \cos\omega_m t$$

Amplitude Modulation (2)

The voltage across the load is an AM signal:

$$v_0(t) = V_0 \cos\omega_0 t = V_{dc}(1 + m\cos\omega_m t)\cos\omega_0 t$$

where m is the modulation depth (modulation index):

$$m = \frac{V_m}{V_{dc}} \leq 1$$

Ignoring V_{sat} and taking into account that the collector voltage must be positive, $v_C(t) \geq 0$, $m \leq 1$. Under the peak modulation condition, the maximum collector voltage is $2V_{\text{dc}}$. AM signals with $m > 1$ cannot be obtained using collector modulation.

Class C Frequency Multipliers (1)

- Frequency multipliers are often used to multiply the frequency of the master oscillator or to increase the modulation index in the case of phase or frequency modulation.

The Class C frequency multiplier has the **same schematic** as the Class C power amplifier and **operates in much the same way**. The only difference is that the collector **resonant circuit is tuned to the desired harmonic**, suppressing all other harmonics.

Assuming that the parallel LC output circuit is ideal, tuned to the *n*th harmonic, a sinusoidal output voltage is obtained:

$$v_0(\theta) = V_0 \cos\theta = RI_M \alpha_n(\theta_c)\cos\theta$$

The output power is given by

$$P_0 = \frac{V_0^2}{2R} = \frac{1}{2} RI_M^2 \alpha_n^2(\theta_c)$$

Class C Frequency Multipliers (2)

The DC power is

$$P_{dc} = V_{dc} I_{dc} = V_{dc} I_M \alpha_0(\theta_c)$$

The collector efficiency is

$$\eta = \frac{P_0}{P_{dc}} = \frac{V_0}{V_{dc}} \frac{|\alpha_n(\theta_c)|}{2\alpha_0(\theta_c)}$$

The collector efficiency is highest if $V_0 = V_{dc}$

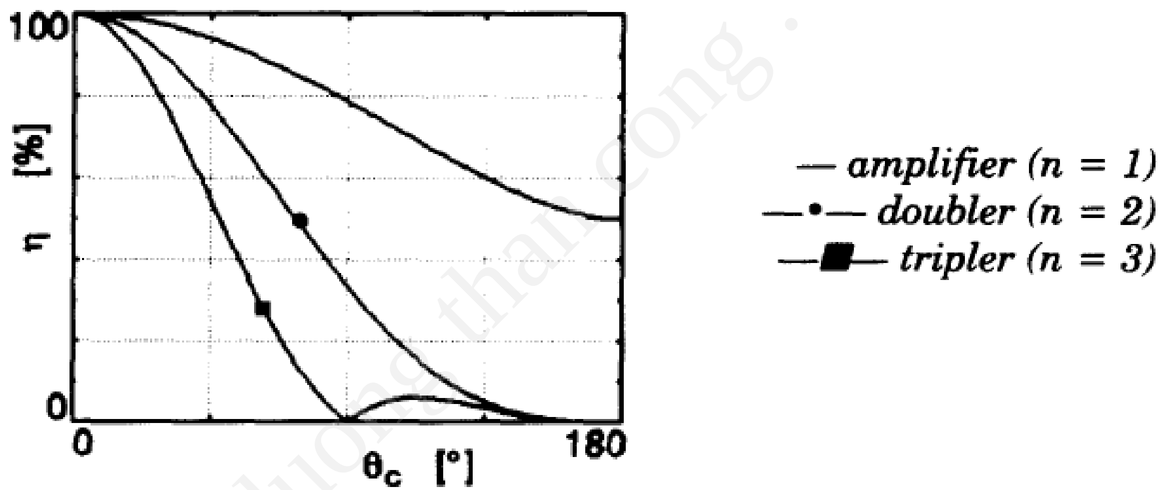
$$\eta_{max} = \frac{|\alpha_n(\theta_c)|}{2\alpha_0(\theta_c)}$$

Finally, the power output capability (for $V_0 = V_{dc}$) is given by

$$C_P = \frac{P_0}{v_{max} I_M} = \frac{|\alpha_n(\theta_c)|}{4}$$

Class C Frequency Multipliers (3)

- The variation of the maximum collector efficiency η_{max} with the conduction angle θ_C, for a Class C amplifier (n = 1), a doubler (n = 2), and a tripler (n = 3), is shown



Note that the collector efficiency decreases as the multiplying order n increases. Also note that a Class B circuit (θ_C = 90°) cannot be used as a frequency tripler, because a half-wave sinusoidal waveform does not contain the third harmonic

Class C Frequency Multipliers (4)

- Figure below shows the variation of power output capability C_P with the conduction angle θ_C . Optimum performance of frequency multipliers (i.e., maximum C_P) is obtained for
  a. frequency doubler: $\theta_C = 60°$, $C_P = 0.06892$, $\eta_{\max} = 63.23\%$
  b. frequency tripler: $\theta_C = 39.86°$, $C_P = 0.04613$, $\eta_{\max} = 63.01\%$

- As **multiplication factor *n* increases**, the **output power** (and also the **power gain** of the stage), the **collector efficiency**, and the **power output capability decrease**. On the other hand, if *n* increases, it becomes more difficult to filter out adjacent harmonics $n - 1$ and $n + 1$ because they lie closer to the desired harmonic, and the relative bandwidth becomes narrower. As a result, Class C frequency multipliers are not recommended for use at high power levels or for a multiplication factor exceeding $n = 3$.

Class D RF Power Amplifiers (1)

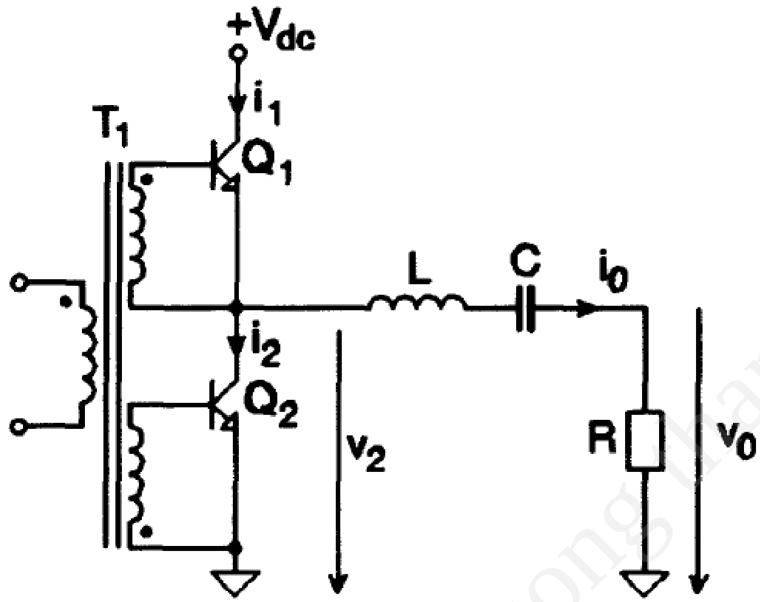
- Class D amplifier is a **switching-mode amplifier** that uses **two active devices** driven in a way that they are **alternately switched ON** and **OFF**.

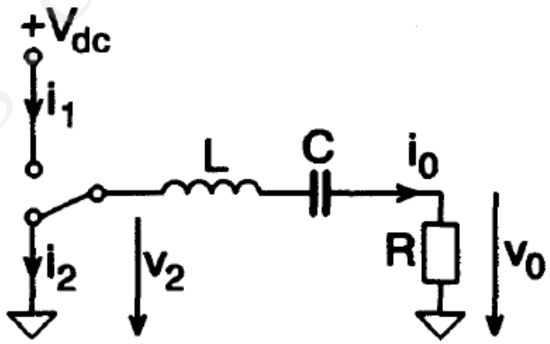
  The active devices form a two-pole switch that defines either a rectangular voltage or rectangular current waveform at the input of a load circuit. The load circuit contains a band- or low-pass filter that removes the harmonics of the rectangular waveform and results in a sinusoidal output.

  The load circuit can be a series or parallel resonant circuit tuned to the switching frequency. In practical applications, this circuit can be replaced by narrowband pi or T-matching circuits, or by band- or low-pass filters (in wideband amplifiers).

Class D RF Power Amplifiers (2)

- **<u>Complementary Voltage Switching</u> (CVS) <u>Circuit</u>**





Input transformer T_1 applies the drive signal to the bases of Q_1 and Q_2 in opposite polarities. If the drive is sufficient for the transistors to act as switches, Q_1 and Q_2 switch alternately between cut-off (OFF state) and saturation (ON state). The transistor pair forms a two-pole switch that connects the series-tuned circuit alternately to ground and V_{dc}.

Class D RF Power Amplifiers (3)

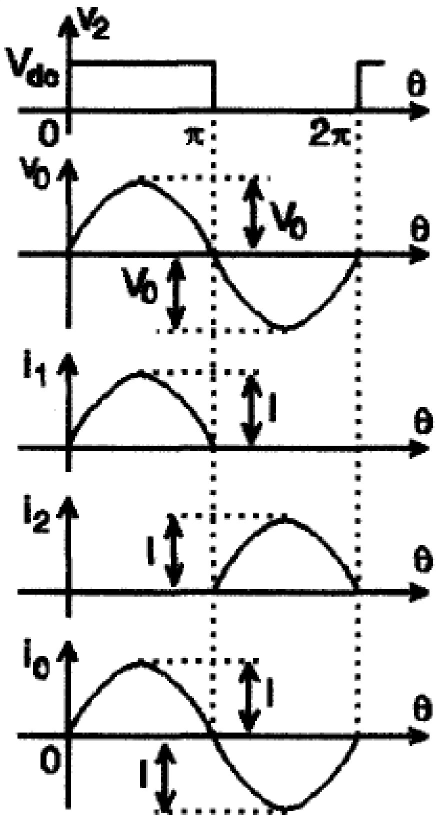
The analysis below is based on the following assumptions:

- The series resonant circuit, tuned to the switching frequency, *f*, is ideal, resulting in a sinusoidal load current. The CVS circuit requires a **series-tuned circuit** or an equivalent (that imposes a sinusoidal current), such as a **T-network**. A parallel-tuned circuit (or an equivalent, such as a pi-network) cannot be used in the CVS circuit.
- The active devices act as ideal switches: zero saturation voltage, zero saturation resistance, and infinite OFF resistance. The switching action is instantaneous and lossless.
- The active devices have null output capacitance.
- All components are ideal. (The possible parasitic resistances of *L* and *C* can be included in the load resistance *R;* the possible parasitic reactance of the load can be included in either *L* or *C*).

Class D RF Power Amplifiers (4)

Assuming a 50 percent duty cycle (that is, 180 degrees of saturation and 180 degrees of cut off for each transistor), voltage $v_2(\theta)$ applied to the output circuit is a periodical square wave:

$$v_2(\theta) = \begin{cases} V_{dc}, & 0 \le \theta \le \pi \\ 0, & \pi \le \theta \le 2\pi \end{cases}$$



where $\theta = \omega t = 2\pi ft$.

Decomposing $v_2(\theta)$ into a Fourier series yields:

$$v_2(\theta) = V_{dc}\left(\frac{1}{2} + \frac{2}{\pi}\sum_{n=1}^{\infty}\frac{\sin(2n-1)\theta}{2n-1}\right)$$

Class D RF Power Amplifiers (5)

Because the series-tuned circuit is ideal, the output current and output voltage are sinusoidals:

$$i_0(\theta) = I\sin\theta = \frac{2}{\pi}\frac{V_{dc}}{R}\sin\theta$$

$$v_0(\theta) = V_0\sin\theta = \frac{2}{\pi}V_{dc}\sin\theta$$

At one moment, the sinusoidal output current flows through either Q_1 or Q_2, depending on which device is ON. As a result, collector currents $i_1(\theta)$ and $i_2(\theta)$ are half sinusoid with the amplitude:

$$I = \frac{2}{\pi}\frac{V_{dc}}{R}$$

The output power (dissipated in the load resistance R) is given by

$$P_0 = \frac{I^2}{2}R = \frac{2}{\pi^2}\frac{V_{dc}^2}{R} \approx 0.2026\frac{V_{dc}^2}{R}$$

Class D RF Power Amplifiers (6)

The DC input current is the average value of $i_1(\theta)$:

$$I_{dc} = \overline{i_1(\theta)} = \frac{1}{2\pi}\int_0^{2\pi} i_1(\theta)\,d\theta = \frac{I}{\pi} = \frac{2}{\pi^2}\frac{V_{dc}}{R}$$

The DC input power is given by

$$P_{dc} = V_{dc} I_{dc} = \frac{2}{\pi^2}\frac{V_{dc}^2}{R} = P_0$$

and the collector efficiency (for the idealized operation) is **100 percent**:

$$\eta = \frac{P_0}{P_{dc}} = 1$$

The power output capability is obtained by normalizing the output power (P_0) by the number of active devices (two), the peak collector voltage (V_{dc}), and the peak collector current (I):

$$C_P = \frac{P_0}{2V_{dc}I} = \frac{1}{2\pi} \approx 0.1592$$