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A P P U N T I

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MATERIA: Analog and telecommunication electronics - Prof. Camarchia

ANALOG AND TELECOMMUNICATION ELECTRONICS

(VITTORIO CAMARCHIA)

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2016/17

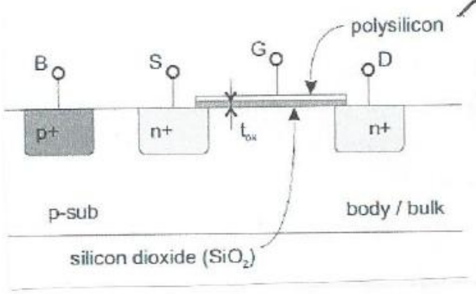
A1 - MOS AND BIPOLAR TRANSISTOR

06/03/17

MOS AND BJT STRUCTURES AND CHARACTERISTIC

N- AND P-MOS DEVICES → HAVE THE FOLLOWING STRUCTURE

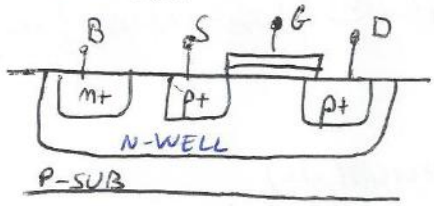
- N-MOS



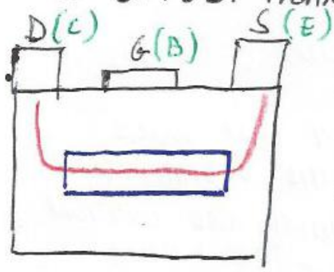
BEFORE THERE WAS METAL, NOW IT'S POLYSILICON

IT'S A SOURCE-GATE-DRAIN + BULK

- P-MOS



A FIELD EFFECT TRANSISTOR = IS A VARIABLE RESISTOR



THERE'S A CONDUCTIVE CHANNEL ▭ WHICH CARRIERS CROSS

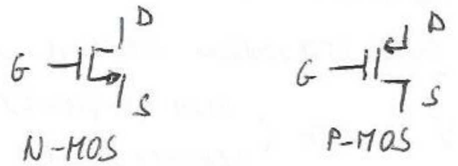
THERE'S AN INJECTION OF CURRENT IN THE BASE AND IT'S POSSIBLE TO MODULATE IT

THERE'S A CORRESPONDENCE WITH THE BJT (EVEN THOUGH THE PHYSICAL PROCEDURE IS DIFFERENT)

DRAIN → COLLECTOR
GATE → BASE
SOURCE → EMITTER

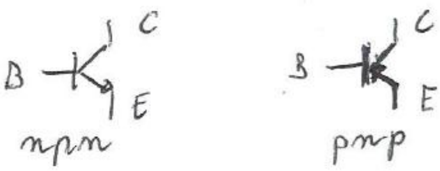
WE ADOPT THEM FOR OUR ELECTRONICS CIRCUITS
WE'RE JUST INTERESTED IN ITS EXTERNAL BEHAVIOUR

MOS:



(→ CURRENT FLOWS IN THE ARROW DIRECTION)

BJT:



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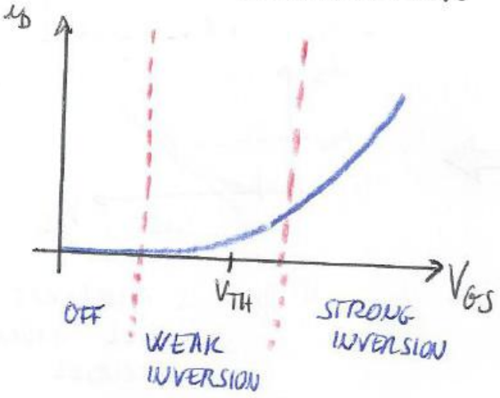
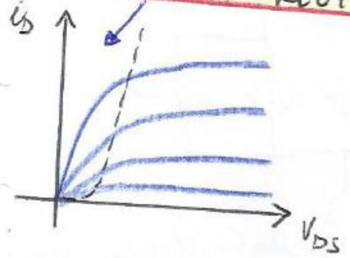
CONCERNING A MOS → IT'S NECESSARY TO ADOPT A MORE SOFISTICATED DESCRIPTION



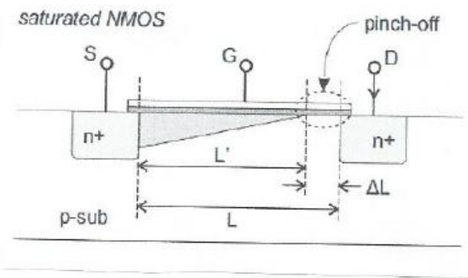
- WEAK INVERSION → CLOSE TO THE THRESHOLD
  ↓
  FOLLOWS A LOW EXPONENTIAL:
  $$i_D = I_{D0} \frac{W}{L} e^{\frac{v_{GS}}{\eta V_T}}$$
- STRONG INVERSION
  ↓
  QUADRATIC BEHAVIOUR
  $$i_D = \mu_n \frac{C_{ox}}{2} \frac{W}{L'} (v_{OD})^2$$

MOS TRIODE REGION → WHERE V_{DS} IS VERY SMALL:



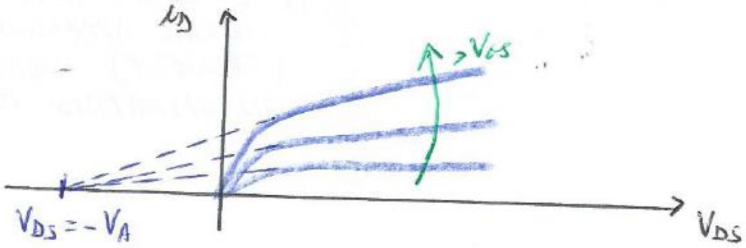
IT'S POSSIBLE TO CHANGE THE SLOPE OF THE CHARACTERISTIC
↓
INDEED IT'S A VARIABLE RESISTANCE, WHICH IS EMPLOYED FOR SPECIFIC APPLICATIONS
(BJT IS NOT GOOD AS A VARIABLE RESISTOR BECAUSE IT HAS FIXED dV_{CE}/dI_C)

CHANNEL PINCH-OFF



AFTER A GIVING VOLTAGE ON DRAIN AND SOURCE
↓
IT'S A REDUCTION ON THE CHANNEL LENGTH → WHOSE EFFECT CONSISTS IN CHANGING THE SLOPE OF THE CHARACTERISTIC

$$i_D = \mu_n C_{ox} \frac{W}{L} \frac{1}{2} v_{OD}^2 \longrightarrow i_D = \mu_n C_{ox} \frac{W}{L} \frac{1}{2} v_{OD}^2 (1 + \lambda v_{DS})$$



IT'S SIMILAR TO EARLY EFFECT IN BJT (WHERE AS V_{CE} INCREASES → THE BASE REGION WIDTH DECREASES)

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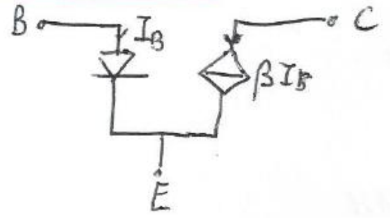
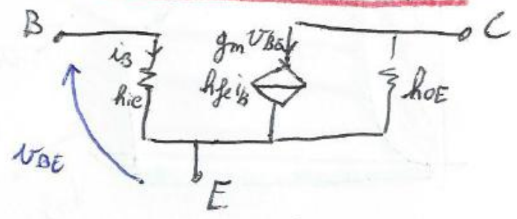
DIFFERENCE MOS/BJT

- **BJT SIMPLIFIED MODELS:**

BIAS MODEL



SMALL SIGNAL ANALYSIS



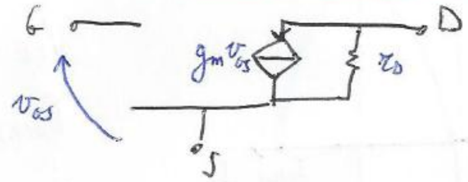
$$g_m = \frac{I_C}{V_T}$$

$$h_{ie} = V_T \cdot \frac{h_{fe}}{I_C}$$

- **MOS SIMPLIFIED MODELS**

BIAS POINT
↓
USE $I_d(V_{GS})$ EQUATION
(ORDER 2 OR BETTER MODEL)
↓
BECAUSE THERE ARE DIFFERENT REGIONS

SMALL SIGNAL ANALYSIS



$g_m \Rightarrow$ II ORDER EQUATION

$$r_D = \frac{dV_{DS}}{dI_D}$$

WITH THE < BJT / MOS MODEL → WE NEED TO TAKE INTO ACCOUNT DIFFERENT PARAMETERS
↓
LIKE PARASSITIC PHENOMENA DUE TO < CAPACITANCES / INDUCTANCES

IN HIGH FREQUENCY MODEL → INDUCTANCES MODEL INTERCONNECTIONS
C_π → STANDS FOR FEEDBACK CAPACITANCE

⇒ AS A CONSEQUENCE, CONCERNING
- SMALL SIGNAL → MOS AND BJT HAVE THE SAME LINEAR MODEL
- LARGE SIGNAL → SAME METHODS BUT DIFFERENT MODELS
  - BJT ↓ EXPONENTIAL LARGE SIGNAL MODEL (RATHER SIMPLE)
  - MOS ↓ LINEAR/QUADRATIC/EXP LARGE SIGNAL MODEL (MORE COMPLEX)

SIMILAR EFFECTS → DISTORSION, HARMONICS, FEEDBACK ...

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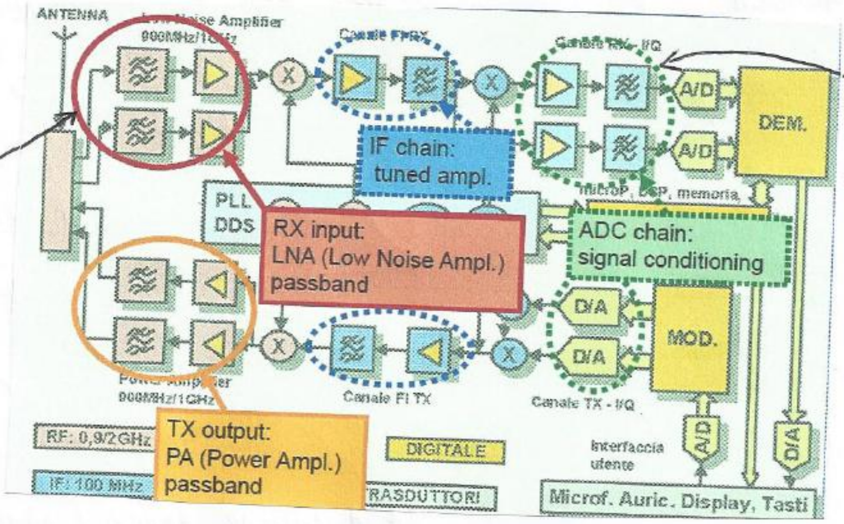
A2 - AMPLIFIERS

TYPES OF AMPLIFIERS

TRANSISTOR - BASED AMPLIFIERS → ARE LOCATED ALMOST AT ANY LEVEL:
- HF AMPLIFIERS
- FOCUS ON POWER AND HOW EFFICIENTLY IT'S GENERATED (POWER AMPLIFIERS)



THE MOST IMPORTANT CHARACTERISTICS ARE
- GAIN (POWER)
- BANDWIDTH
- LINEARITY → GOAL: AMPLIFY THE SIGNAL AND CREATE POWER. BUT LINEARITY NEEDS TO BE KEPT LARGE ENOUGH TO THE RECEIVER
- EFFICIENCY → MUST BE MAXIMIZED WITH LINEARITY ENSURED
- NOISE → WITH LOW NOISE AMPLIFIERS IT MUST BE KEPT AT THE MINIMUM POSSIBLE VALUE

! ALL THE SYSTEMS ARE NON-LINEAR
↓
WE TRY TO WORK WITH THEM CONSIDERING THEM REAL.
WHEN POSSIBLE WE TRY TO MAKE THEM AS SIMPLE AS POSSIBLE

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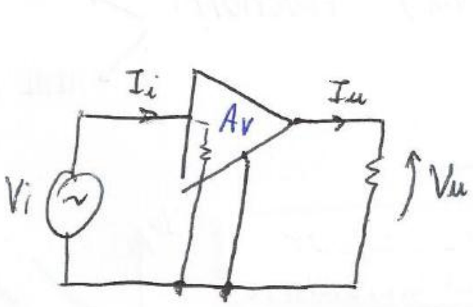
POWER GAIN

$$\boxed{G|_{dB} = 10 \log_{10}(K_p)} \quad [dB] \qquad K_p = \frac{P_u}{P_i}$$



THERE'S A RELATION POWER-VOLTAGE RESISTANCE

$$P = \frac{V^2}{R} \qquad A_v = \left(\frac{V_u}{V_i}\right)^2 \frac{R_i}{R_u}$$

IF $R_i = R_u$

$$\boxed{G_p = \frac{P_u}{P_i}\bigg|_{dB} = 20 \log_{10}\left(\frac{V_u}{V_i}\right)} \quad [dB]$$

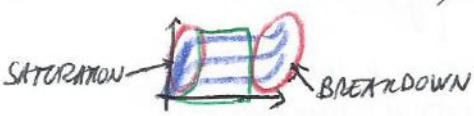
! POWER GAIN $\neq$ VOLTAGE GAIN $\Rightarrow$ $A_v = \sqrt{K_p}$ IN LINEAR BEHAVIOR

$\frac{P_{out}}{P_{in}}$ → STILL RULES → IF VOLTAGES YOU NEED TO CONSIDER THE RATIO $\frac{R_i}{R_u}$

AMPLIFIER PARAMETERS

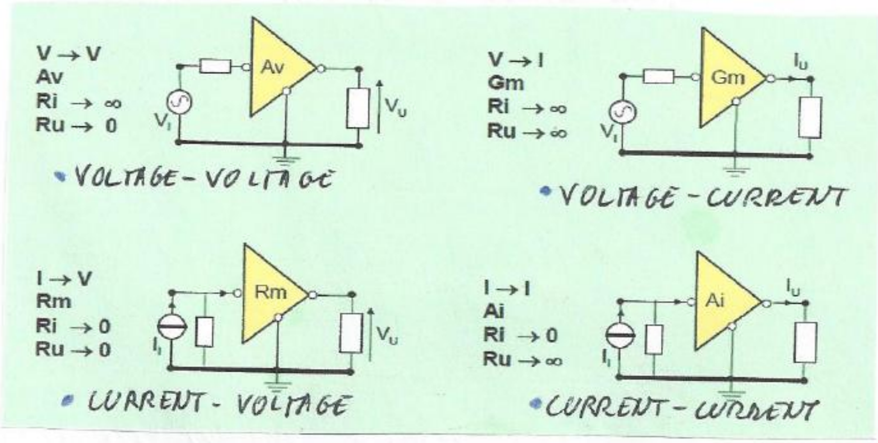
- LINEAR BEHAVIOR →
  - DC → GAIN (IF ANY) / (POSSIBLE) OFFSET
  - AC → $f \neq 0$

    A DC SIGNAL IS ASSOCIATED TO A VARIATION OF A QUANTITY
    ↓
    IT'S A SOURCE OF INFORMATION IF IT'S TIME VARIANT
  - NOISE → IS ALWAYS PRESENT → NEEDS TO BE CONSIDERED FOR A GOOD DESIGN
- NON-LINEAR BEHAVIOR →
  - STATIC NON-LINEARITY → CAN BE SATURATION / BREAKDOWN

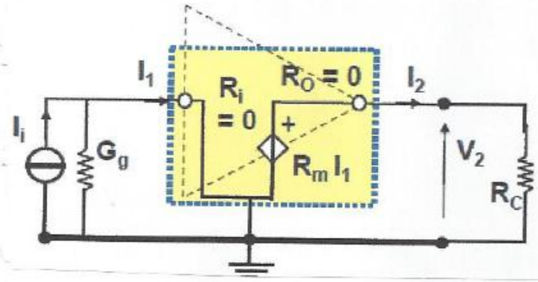


  - DYNAMIC SATURATION → WITH THE SLEW-RATE WE'RE MOVING TOO FAST AND AMPLIFICATION CAN'T FOLLOW THE SIGNAL
- OTHER → EMC (ELETROMAGNETIC COMPATIBILITY)

THERE ARE 4 TYPES OF AMPLIFIERS:

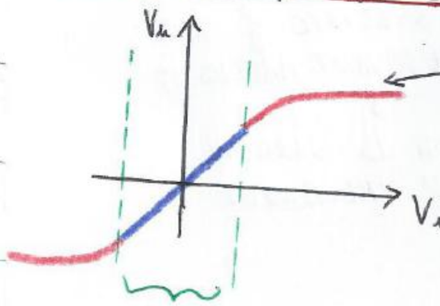


DEPENDING ON THE KIND OF THE AMPLIFIER → YOU CAN ADOPT A VOLTAGE OR CURRENT SOURCE



IN THIS CASE IT'S EASIER TO USE A CURRENT GENERATOR

NON LINEARITY AND SATURATION



LINEAR MODEL
(GOOD FOR SMALL CHANGES)

LARGE CHANGES → CAUSE SATURATION

↓

NON LINEARITY

⇓

DIFFERENT POSSIBILITIES

K-ORDER POLYNOMIAL — FOURIER SERIES

$$\boxed{V_u(V_{in}) = k + A V_i + B V_i^2 + C V_i^3 + \dots}$$

"GAIN" FOR THE FOUNDAMENTAL COMPONENT

DUE TO NON-LINEARITY (COMPONENTS DUE TO THE HARMONICS IF 1 FREQUENCY IS INCLUDED)

⇒ IN THIS CASE IT'S USEFUL TO EVALUATE THE DEGREE OF DISTORSION

$$THD = \frac{\sum_{n=2}^{\infty} P_n}{P_1}$$

P_n ← POWER OF ALL HARMONICS

P_1 ← POWER OF FOUNDAMENTAL

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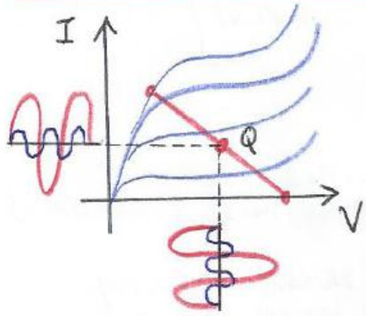
A3 - BJT AMPLIFIERS

08/03/17

TRANSISTOR AMPLIFIERS



IF WE WORK WITH A SIGNAL, AROUND A BIAS POINT:
IDEA: SET THE WORKING POINT $Q(I_C, V_{CE})$
[$Q(V_{GS}, V_{DS})$ IF WE'RE TALKING ABOUT MOS]

THEN DEAL WITH
- SMALL SIGNAL MODEL
- LARGE SIGNAL MODEL

THE FOLLOWING STEPS HAVE TO BE PERFORMED:
1) SET THE BIAS POINT
2) HOW TO VERIFY IF IT WORKS WELL?
EVALUATION OF SMALL SIGNAL PERFORMANCES (IN ORDER TO EVALUATE IF IT'S USEFUL TO ACHIEVE MY GOAL)
3) EVALUATION OF < THEORETICAL / PRACTICAL LIMITS

IT'S ALSO POSSIBLE TO < MODIFY / CHANGE THE STRUCTURES → BUT FIRST IT'S IMPORTANT TO UNDERSTAND HOW IT WORKS
DEPENDING ON THE APPLICATION

LAB1: STARTING FROM THE SPECIFICATIONS → DESIGN AN AMPLIFIER AND VERIFY IF IT WORKS AS EXPECTED

MOST IMPORTANT CHARACTERISTICS OF AN AMPLIFIER:
- GAIN → ALONE IT'S NOT SO USEFUL
  → WILL BE ALWAYS FREQUENCY DEPENDENT! ⇒ GAIN (f)
- BANDWIDTH → COVERS A SPECIFIC RANGE OF FREQUENCIES ($f_1 \div f_2$)
- LINEARITY → IF LINEAR: $V_u = K V_i$ (WITH K = NUMBER)
  IN ORDER TO BETTER DEFINE IT WE NEED TO TAKE INTO ACCOUNT THE FREQUENCY:

$$V_u(f) = \underbrace{K(f_0)}_{\text{FOUNDAMENTAL FREQUENCY}} V_i + \underbrace{\sum_{2}^{\infty} K_i(f) \cdot V_i}_{\text{SERIES OF SPURIOUS COMPONENTS DUE TO THE PRESENCE OF THE HARMONICS}}$$

WHICH DISSIPATE POWER

[THERE ARE SPECIFIC COMPONENTS WITH A STRONG NON-LINEARITY (EG TRAIN OF PULSES).
YOU NEED TO AMPLIFY THE FOUNDAMENTAL COMPONENT]

$$\boxed{\eta = \frac{P(f_0)}{P_{DC}}}$$

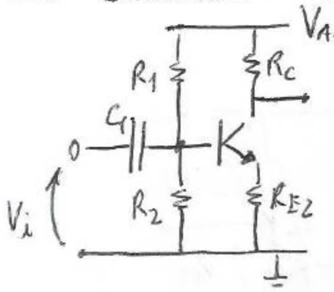
$P(f_0)$ ← WANTED POWER
P_{DC} ← POWER DISSIPATED IN DC

EFFICIENCY

$$\boxed{PAE = \frac{P_{out}(f_0) - P_{in}(f_0)}{P_{DC}}}$$

POWER ADDED EFFICIENCY
USEFUL → @ HIGH FREQUENCY [100 MHz - 1 GHz]
→ ESPECIALLY WHEN GAIN IS LOW

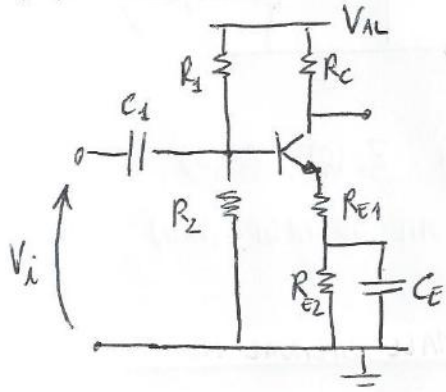
THE STANDARD SCHEME → ADDS AN OTHER RESISTOR R_2



FOR STANDARD VARIATIONS OF < PARAMETERS / TEMPERATURE

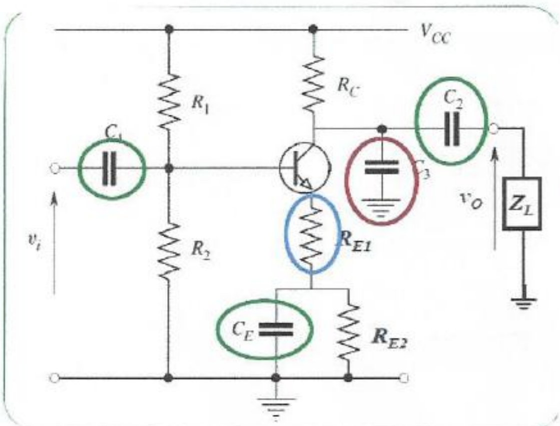
THE CIRCUIT IS ROBUST IN SETTING THE BIAS POINT
(IT'S ALSO IMPORTANT TO VERIFY IF YOUR ASSUMPTIONS ARE CORRECT!)

⇒ A FINAL POSSIBLE TRICK → WILL BE TO SPLIT R_{E1} AND R_{E2} ↳ IN PARALLEL WITH A CAPACITOR



IN SMALL SIGNAL I'LL HAVE JUST THE CONTRIBUTION OF R_{E1} BECAUSE R_{E2} WILL BE SHORTCIRCUITED BY THE CAPACITOR

AFTER THE ANALYSIS OF THE BIAS POINT WE NEED TO ANALYZE THE SIGNAL PATH → WITH SOME MODIFICATIONS COMING FROM THE SMALL SIGNAL:



- DC GROUNDED V_{CC} ⊥
  V_{CC} MUST BE A CONSTANT VALUE
- WE NEED TO CONSIDER COMPONENTS LIKE < L / C
  IF THEIR VALUES ARE SIGNIFICANT IN THE BANDWIDTH WE'RE INTERESTED IN.

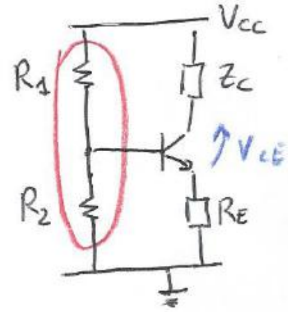
(C) PARASSITIC COMPONENTS OF THE TRANSISTOR

IN HIGH FREQUENCY → BJT / FET } WILL STOP WORKING BECAUSE THE OUTPUT WILL BE SHORTCIRCUITED

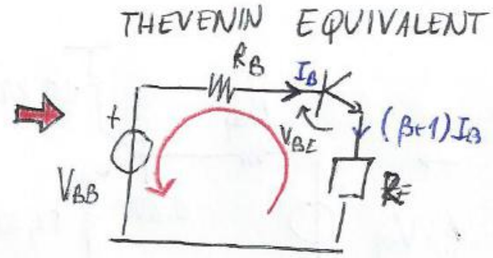
C_E → WILL BE SET ACCORDING TO THE BANDWIDTH

C_1 → WILL BE SET ACCORDING TO OUR SPECIFIC NEEDS

IN ORDER TO FIND I_C:



THEVENIN EQUIVALENT



$$V_{BB} = V_{CC}\,\frac{R_2}{R_1+R_2}$$

$$R_B = R_1 /\!/ R_2$$

$$(\beta = h_{fe})$$

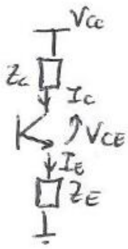
$$V_{BB} - V_{BE} = I_B R_B + (\beta+1) I_B Z_E$$

$$I_B = \frac{V_{BB}-V_{BE}}{R_B + (\beta+1) R_E}$$

$$I_E = I_B(\beta+1) = \frac{V_{BB}-V_{BE}}{R_E + \frac{R_B}{\beta+1}}$$

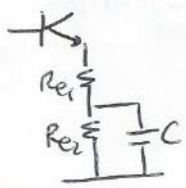
WE CAN ALSO REPLACE $(\beta+1)$ BY β WITH THE ASSUMPTION $I_C \approx I_E$

THEN → VERIFICATION → EVALUATING V_{CE}



$$V_{CE} = V_{CC} - I_C R_C - I_E R_E$$

! IF



$\Rightarrow Z_E(0) = R_{e1} + R_{e2}$

THERE ARE SOME SEMPLIFICATIONS, BUT THE VERIFICATION WITH THE REAL VALUES IS FAST, TOO.

! BEFORE DOING THE CALCULATION, IT'S USEFUL TO HAVE AN IDEA OF THE ORDER OF UNIT.

HOW TO CHOOSE $\boxed{R_1, R_2}$?

THE HIGHER R_1, R_2 ⇒ THE LOWER I_B

(IF WE HAVE A SMALL R_B → HIGH POWER CONSUMPTION)

→ COST: WE'RE NOT WORKING WITH A ROBUST CIRCUIT

WE'RE REDUCING THE DYNAMIC RANGE

! THE PERFECT BIAS POINT ISN'T THE PERFECT SOLUTION IN CASE OF SMALL SIGNAL MODEL

TYPICAL VALUES OF $R_1, R_2 < 10 \div 100\ M\Omega$

$\sim 1\ M\Omega,\ 100\ k\Omega$

(NOT TOO LOW IN ORDER TO NOT LOOSE POWER)

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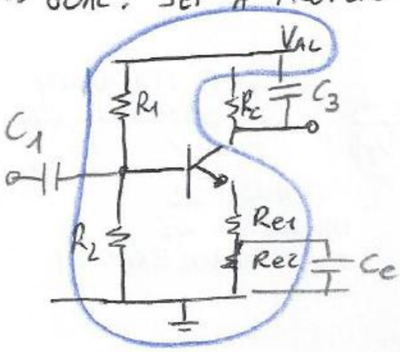
09/03/17

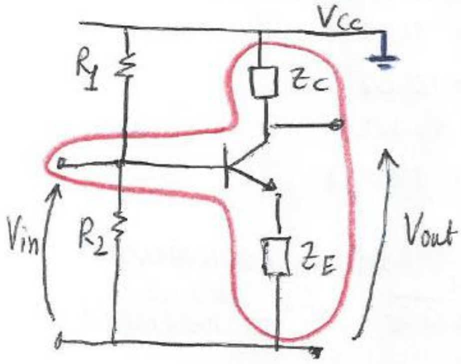
BIASING → GOAL: SET A PROPER I_C IN ORDER TO WORK IN A GOOD WAY



IT'S POSSIBLE TO ADOPT SIMPLER SOLUTIONS WHICH DEPEND ON TRANSISTOR PARAMETERS

IF YOU'RE ABLE TO DESIGN SOME COMPONENTS THAT YOU KNOW VERY WELL IT'S A MORE ROBUST BUT ALSO A MORE COMPLEX SOLUTION.

SMALL SIGNAL ANALYSIS



! YOU NEED TO TAKE INTO ACCOUNT THAT THE CIRCUIT CONTAINS A CAPACITOR (C_1) IN ORDER TO DECOUPLE < DC / SIGNAL

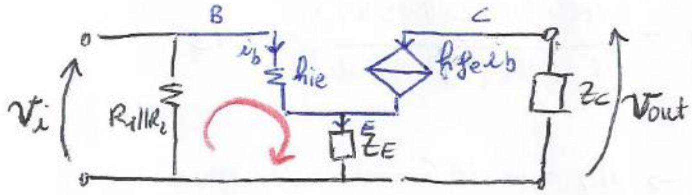
V_{CC} = GROUNDED BECAUSE WE'RE LOOKING FOR A VARIATION (IT MUST BE JUST A NUMBER ... WHY NOT GROUND?)

↓

AS A CONSEQUENCE $R_1 // R_2$

1st OPERATION → REPLACE THE BJT WITH THE SMALL SIGNAL EQUIVALENT CIRCUIT.

WE CAN REARRANGE THE CIRCUIT IN THIS WAY



$$v_{out} = -Z_C i_C$$

$$i_C = h_{fe} i_b$$

SOLVE THE MESH IN ORDER TO HAVE THE FINAL RELATIONSHIP

$$v_i = i_b h_{ie} + i_b (h_{fe}+1) Z_E \rightarrow i_b = \frac{v_i}{h_{ie} + (h_{fe}+1) Z_E}$$

$$v_{out} = -\frac{Z_C h_{fe} v_i}{h_{ie} + (1+h_{fe}) Z_E} \rightarrow A_V = \frac{v_{out}}{v_i} = -\frac{Z_C h_{fe}}{h_{ie} + (1+h_{fe}) Z_E}$$

WITHOUT ANY SEMPLIFICATIONS

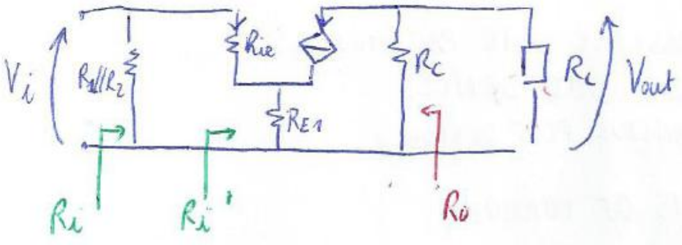
! IF $h_{fe} \gg 1 \Rightarrow \boxed{A_V = \frac{v_{out}}{v_i} \simeq -\frac{Z_C}{Z_E}}$

⇓

IN THIS WAY THE AMPLIFICATION IS COMPLETELY <u>INDEPENDENT</u> FROM THE VALUE OF THE CURRENT GAIN (THAT IS THE PARAMETER WE KNOW WITH LESS PRECISION)

THE RESULT WILL HAVE VERY SMALL VARIATIONS!

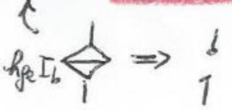
THEN EVALUATE R_{in} AND R_{out}



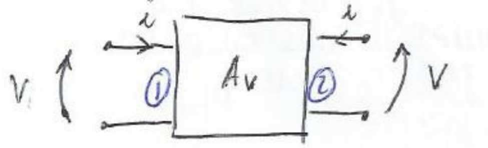
$R_i = (R_1 // R_2) // R_i'$

$$\boxed{R_i = (R_1 // R_2) // [h_{ie} + R_{E1}(1+h_{fe})]}$$

$R_o = R_c // \infty \rightarrow \boxed{R_o = R_c}$



AN IDEAL VOLTAGE AMPLIFIER NEEDS



$$\boxed{\begin{matrix} R_i \rightarrow +\infty \\ R_o \rightarrow 0 \end{matrix}}$$

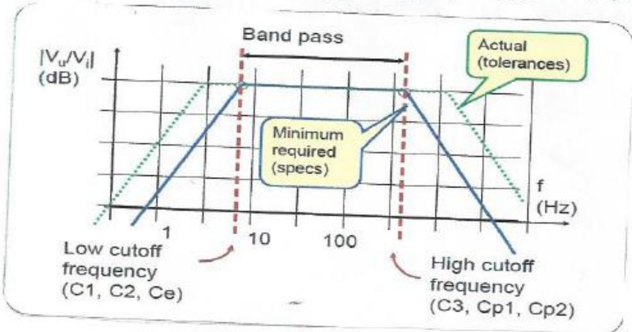
THAT MEANS < $R_{out} \ll R_L$, $R_{in} \gg R_s$

FOR WHICH FREQUENCY BANDS CAN I USE THIS CONFIGURATION OF COMMON EMITTER?
THERE ARE SOME INTRINSIC LIMITS IN MY DESIGN

LOWER BAND LIMIT:
THE SERIES CAPACITANCE INSERTED BY THE DESIGNER IN ORDER TO DECOUPLE < DC, SIGNAL
⇓
LIMITS THE LOWER FREQUENCY
(THERE'S NO DC AMPLIFICATION REGARDLESS THE VALUE OF C)

HIGHER BAND LIMIT:
- PARALLEL CAPACITORS TOWARDS GROUND
- INDUCTANCES DUE TO WIRES AND PACKAGE, ...

IT'S BETTER TO SET UP THESE LIMITS → THE IDEA OF A PASS-BAND MUST BE FOLLOWED



HOW LARGE IS THE BAND PASS STRONGLY DEPENDS ON THE APPLICATION
(eg: NARROW BANDS WITH RESPECT TO A CENTER FREQUENCY)

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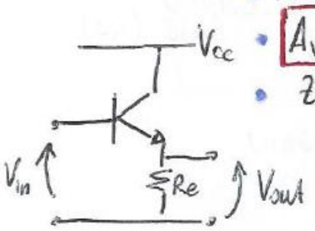
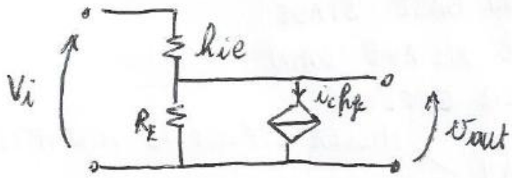
AN OTHER POSSIBLE CONFIGURATION IS THE **COMMON COLLECTOR**

- THE OUTPUT IS ON THE EMITTER
- IT SOLVES, AT LEAST IN PART, OUR PROBLEMS: NO MILLER EFFECT
- $\boxed{A_V \approx 1}$
- $Z_i \to HIGH$, $Z_{out} \to LOW$



LET'S EVALUATE A_V



$$v_{out} = i_E R_E = i_B (1+h_{fe}) R_E$$

$$v_i = i_B h_{ie} + (1+h_{fe}) R_E i_B$$

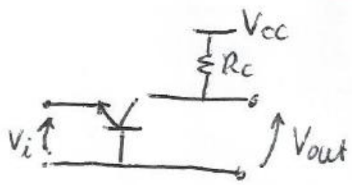
$$A_V = \frac{v_{out}}{v_i} = \frac{(1+h_{fe}) R_E}{h_{ie} + (1+h_{fe}) R_E} \sim 1$$

$\Rightarrow$ COMMON COLLECTOR
- WORKS WELL @HF (NO MILLER EFFECT)
- NO VOLTAGE GAIN $\to$ IT CANNOT BE USED IN ORDER TO AMPLIFY

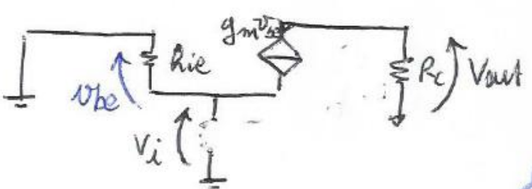
YOU CAN USE IT IN ORDER TO DECOUPLE A STAGE
IT'S ALMOST AN IDEAL AMPLIFIER WITHOUT AMPLIFICATION

ALSO THE **COMMON BASE** $\to$ CAN SOLVE THE MILLER PROBLEM WITH $\boxed{A_V \approx g_m R_C}$

$A_i \approx 1$



LET'S FIND A_V



$$\begin{cases} V_{out} = -R_C g_m v_{be} \\ v_{be} = -v_i \end{cases}$$

$$A_V = \frac{v_{out}}{v_{in}} = +R_C g_m$$

BUT IT'S NOT A GOOD AMPLIFIER BECAUSE OF THE IMPEDANCES
- LOW R_{in}
- HIGH R_o

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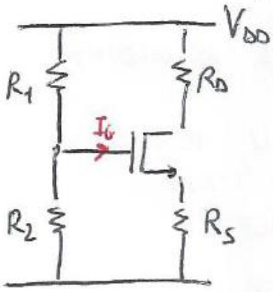
A4 - MOS AMPLIFIER STAGES

WITH MOS → C, B, E ARE RESPECTIVELY REPLACED BY D, G, S

IT'S POSSIBLE TO ADOPT A CASCODE CONFIGURATION, TOO



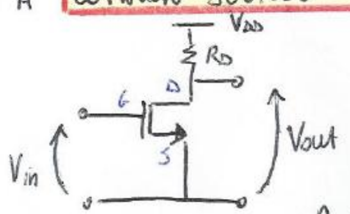
BIAS POINT → CONCERNING MOS, THE MAIN DIFFERENCE WITH BJT IS THAT WE NEED TO SET 2 VOLTAGES

MOS ARE MORE USED BECAUSE OF STATIC CONSUMPTION WHILE IN BJT YOU NEED A CURRENT WHICH CAUSES POWER LOSS, IN A MOS NO CURRENT FLOWS IN THE TRANSISTOR ($I_G \sim 0$)

ONE POSSIBLE ISSUE IS THAT THERE ISN'T A MODEL THAT IS MORE SUITED IN ORDER TO BETTER DESCRIBE IT

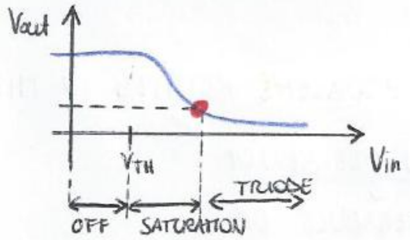
THE BEHAVIOR CAN BE
- LINEAR
- QUADRATIC → SATURATION

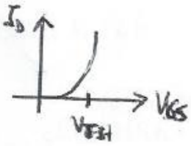
IN A **COMMON-SOURCE CONFIGURATION**



$A_V = g_m R_D$

YOU HAVE THE FOLLOWING TRANSCHARACTERISTIC



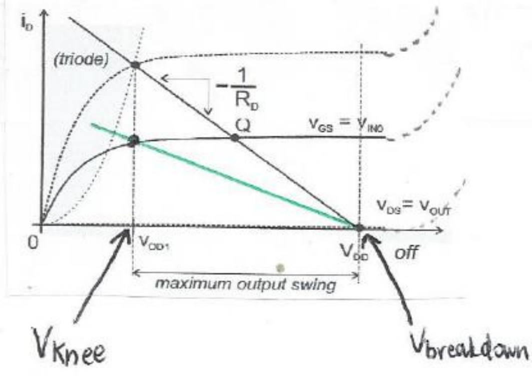


$$\boxed{V_{out} = V_{DD} - R_D I_D = V_{DD} - R_D \left[\frac{1}{2} \mu_n C_{ox} (V_{in} - V_{TH})^2 \right]}$$

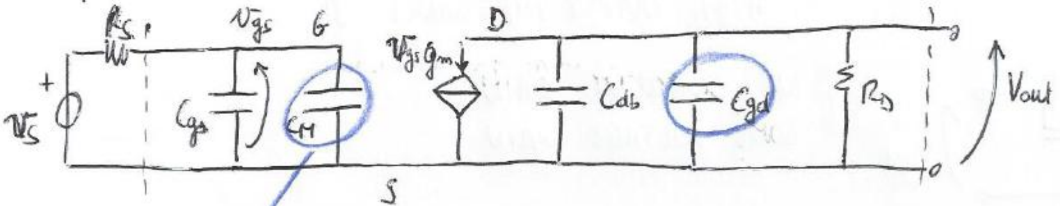
$$A_V = \frac{\partial V_{out}}{\partial V_{in}} = -R_D \underbrace{\left(\frac{1}{2} \mu_n C_{ox} \frac{W}{L} V_{OD} \right)}_{g_m \to \text{TRANSDUCTANCE}} \Rightarrow \boxed{A_V = -g_m R_D}$$

V_{OUT} → DEPENDS ON I_D WHICH IN SATURATION REGION HAS A QUADRATIC DEPENDENCE ON V_{TH}

A_V → REDUCES TO $g_m R_D$ WHERE R_D CHARACTERIZE THE SLOPE OF THE LOAD LINE

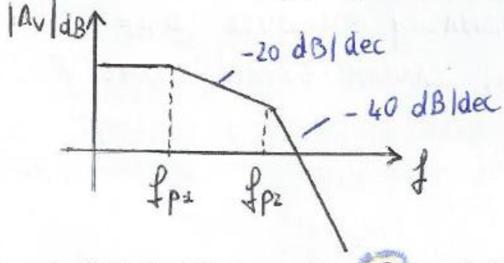


APPLYING MILLER



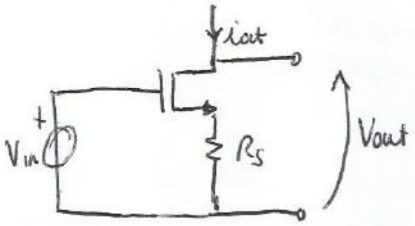
MULTIPLIED BY MILLER EFFECT

Z_{in} <
- WITHOUT MILLER → JUST DEPENDS ON THE INPUT PORT
- WITH MILLER → DEPENDS ALSO ON MILLER CAPACITOR WHICH MAKES FASTER THE DECAY AS WE CAN SEE FROM THE BODE DIAGRAM



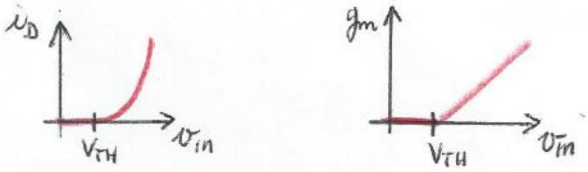
IF GAIN IS REASONABLY HIGH C_M IS HIGHER THAN C_{gs}
↓
THE 1st POLE SHIFTES TO LOWER FREQUENCIES

A POSSIBILITY CONSISTS IN ADDING R_S → LIMITS g_m AND ALLOWS TO CONTROL IT

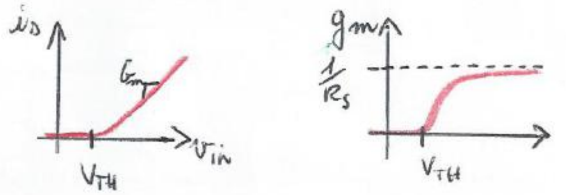


$$G_m \triangleq \frac{i_d}{v_{in}} = \frac{g_m}{g_m R_S + 1}$$

$R_S = 0$

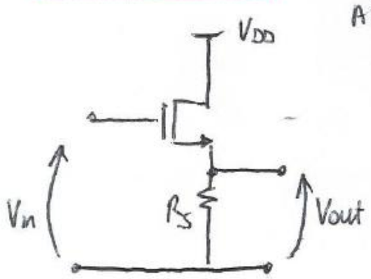


(SOURCE FEEDBACK) $R_S \neq 0$



IT'S POSSIBLE TO EMPLOY:

- COMMON DRAIN → EQUIVALENT TO COMMON COLLECTOR CONFIGURATION ADOPTED IN ORDER TO BUILD UP THE CASCODE CONFIGURATION



- $A_V \sim 1$ (SORT OF VOLTAGE FOLLOWER) NON INVERTING
- WITH GOOD CHARACTERISTICS IN TERMS OF INPUT AND OUTPUT IMPEDANCES

IT CAN BE USED AS A BUFFER AND NOT AS A VOLTAGE AMPLIFIER

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10/03/17

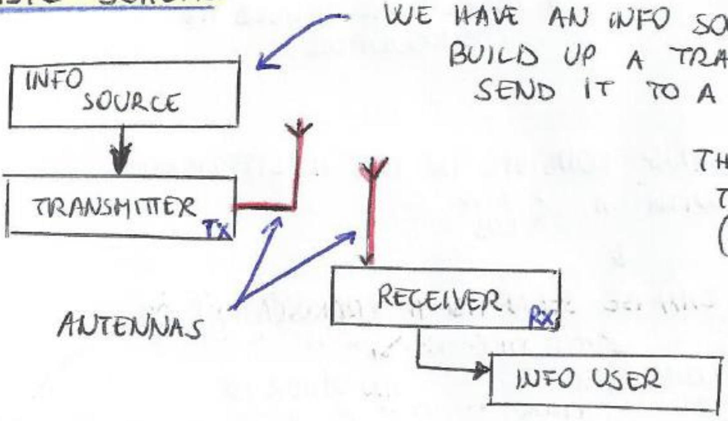
B1 - RADIO SYSTEMS ARCHITECTURE

STARTING FROM A SIGNAL → WE NEED TO < MANIPULATE / TRANSMIT IT

OUR FOCUS → IS ON TRANSMISSION THROUGH AIR (= OUR COMMUNICATION CHANNEL)

COMPONENTS OF RADIO FREQUENCY ARE MAINLY IN THE ANALOG DOMAIN BECAUSE DIGITAL SOLUTIONS ARE < EITHER NO AVAILABLE / OR TOO EXPENSIVE

BASIC SCHEME

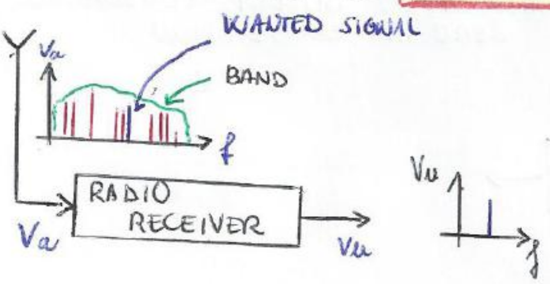


WE HAVE AN INFO SOURCE AND WE NEED TO BUILD UP A TRANSMITTER IN ORDER TO SEND IT TO A CUSTOMER
↓
THEN WE TRANSFER IT THROUGH ANTENNAS (IN AIR SPACE)
↓
THE INFO REACHES THE RECEIVER AND THE SIGNAL IS NOW AVAILABLE TO THE FINAL USER

THE SYSTEM TX-RX IS CALLED TRANSCEIVER

RX, TX } SHARE MOST OF THE CONCEPTS
THERE ARE SLIGHTLY DIFFERENCES ON HOW THE SIGNAL IS HANDLED

LET'S FOCUS ON THE $\boxed{\text{RECEIVER}}$ → HAS ADDITIONAL PROBLEMS IN MANIPULATING SIGNALS



IS USED TO DETECT A SIGNAL WHICH REACHES THE ANTENNA WITH MANY OTHERS INTERFERENCES

IT MUST BE ABLE TO CORRECTLY SELECT THE SIGNAL
↓
SHOULD BE A GOOD REPLICA OF THE SIGNAL THAT STARTS TRAVELLING IN THE AIR.

[GOOD SPECTRAL PURITY
HIGH POWER]

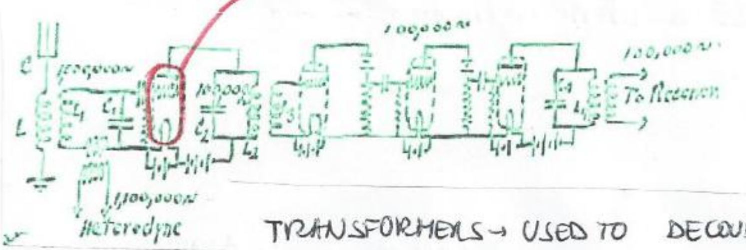
IF IT CANNOT BE RECOVERED IT'S NOT USEFUL

THE SIGNAL IS REDUCED IN AMPLITUDE AFTER IT HAS TRAVELLED THROUGH AIR (IT ARRIVES WITH AN AMPLITUDE OF (−40 ÷ −30) dBm)

RX OPERATIONS: → • DOWNCONVERSION
- DETECTION (→ SIGNAL MUST BE SELECTED AMONG MANY OTHERS)
- AMPLIFICATION (→ SIGNAL IS REDUCED IN AMPLITUDE)
- DEMODULATION (→ SIGNAL ISN'T A PURE TONE, IT'S AM/FM MODULATED [QAM, PSK..])
- DOWNCONVERSION (→ MEANS TO MIX THE SIGNAL WITH ANOTHER WITH A CERTAIN FREQUENCY IN ORDER TO MOVE THE CARRIER TO LOWER f)

ARMSTRONG SCHEME → FROM 1 MHz TO 100 kHz

TRIODE

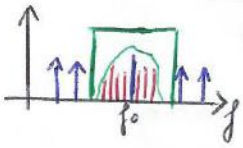


TRANSFORMERS → USED TO DECOUPLE → IMPEDANCE OF THE ANTENNA / IMPEDANCE OF THE SIGNAL

THIS SCHEME IS FAR FROM THE DC

(DIRECT CONVERSION TO DC → HAS SOME POSSIBLE COUNTEREFFECTS WHICH CAUSE SOME PROBLEMS)

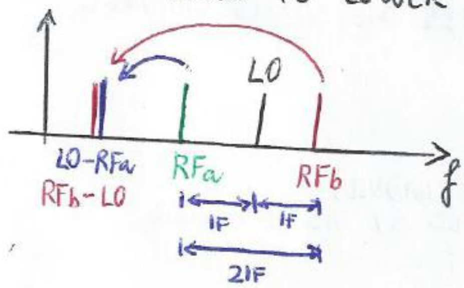
THE IDEA IS GOOD → THE SYSTEM WORKS WELL FOR A SINGLE CHANNEL.
WHAT HAPPENS IF MANY OTHER USERS ARE SHARING THE SAME BAND?
YOU NEED TO PROPERLY SELECT THE WANTED SIGNAL
THE IDEA OF SELECTION IS NOT SO TRIVIAL



→ YOUR TASK IS TO SELECT THE BAND BECAUSE SELECTING JUST THE CARRIER IS STILL ALMOST IMPOSSIBLE

⇓

AFTER HAVING SELECTED IT YOU HAVE TO MOVE THE BAND TO LOWER FREQUENCIES



RFa IS YOUR WANTED SIGNAL
@THE SAME FREQUENCY OF THE DOWNCONVERSION YOU'LL ALSO FIND RFb×LO (=IMAGE SIGNAL)
↓
IT'S A PROBLEM!
HOW CAN IT BE SOLVED?

- FILTER THE SIGNAL BEFORE MIXING
- ZERO-IF
- IMAGE REJECTION MIXER

WHAT ARE THE CONSTRAINTS ON FILTERS WHICH SELECTS ONLY RFa?

THANKS TO THE MIXER → MIXING THE DISTANCE MEANS HAVING LESS RESTRICTIONS ON THE BAND BUT AS A COUNTER-EFFECT THIS BRINGS THE RFa TO LOW FREQUENCIES AND IT'S VERY DIFFICULT TO DESIGN A FILTER LIKE THIS

ESPECIALLY @LF [IT NEEDS AN HIGH Q FOR THE IF FILTER]

⇓

! INCREASING/DECREASING THE FREQUENCY OF THE LOCAL OSCILLATOR MEANS TO RELAX THE CONSTRAINTS ON THE FILTER FOR THE SELECTION

(OR USING A MIXER WHICH TAKES INTO ACCOUNT THE FREQUENCY FROM WHICH THE SIGNAL IS COMING)

THE PROBLEM IS PRESENT BOTH IN THE HETERODYNE AND IN THE DIRECT] CONVERSION

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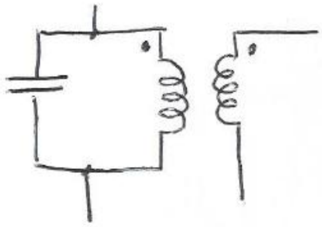
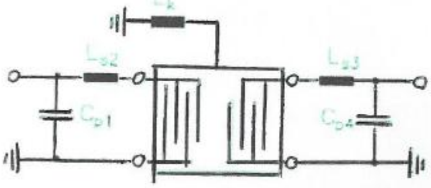
IF FILTERS:

- LC TUNED CIRCUIT → USED @ VERY HIGH FREQUENCY (NUMERICAL METHODS FAIL @ HF)



BASED ON LUMPED ELEMENTS
REQUIRES TUNING (→ VERY EXPENSIVE)
$\mathcal{L}$ → VERY DIFFICULT TO INTEGRATE (BECAUSE THEY'RE VERY LARGE INSIDE)
NOT PRECISE

- SAW (= SURFACE ACOUSTIC WAVES) → REPLACES LC
HIGH Q AND LOW COST



- QUARTZ → EMPLOYED AS A GOOD REFERENCE FOR < TEMPERATURE / TIME
→ HIGH-Q MECHANIC RESONATOR
- ACTIVE FILTERS, DIGITAL FILTERS → FOR LOW IF CHANNELS

COMPLEX MIXERS

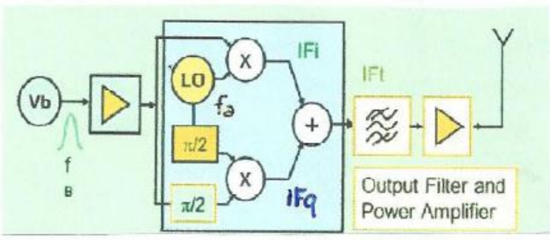
→ WORK WITH SIN AND COS COMPONENTS
→ SMART SOLUTION FOR FLEXIBILITY, GIVING THE POSSIBILITY TO SOLVE THE IMAGE FREQUENCY PROBLEM
→ OPERATE USING < IN-PHASE (I) / QUADRATURE (Q) } COMPONENTS

HOW DO THEY WORK? THERE ARE 2 MIXERS WHICH → ARE AS SIMILAR AS POSSIBLE
→ GIVE 2 REPLICA OF THE SIGNAL WITH THE SAME COHERENCE A PART FROM A 90° PHASE SHIFT

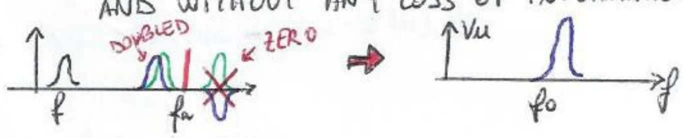
THESE 2 REPLICA ARE THE I AND Q COMPONENTS

2 TYPES OF THESE MIXERS: (NOW LET'S FOCUS ON THE TX)

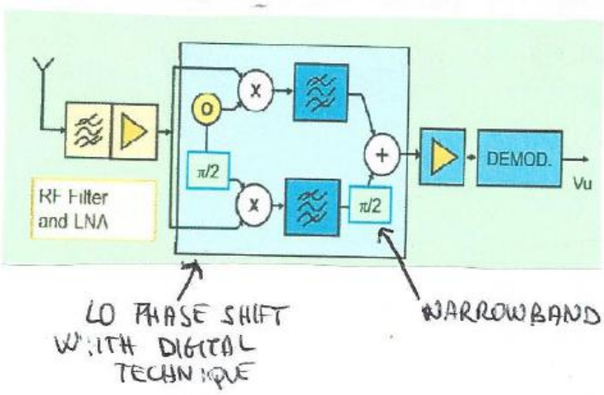
- SINGLE SIDE BAND (SSB) MIXER



WITH THE I/Q FREQUENCY TRANSLATION
↓
CANCELATION OF ONE SIDEBAND WITH A CLEAR GAIN FROM THE POINT OF VIEW OF SPECTRAL EFFICIENCY AND WITHOUT ANY LOSS OF INFORMATION

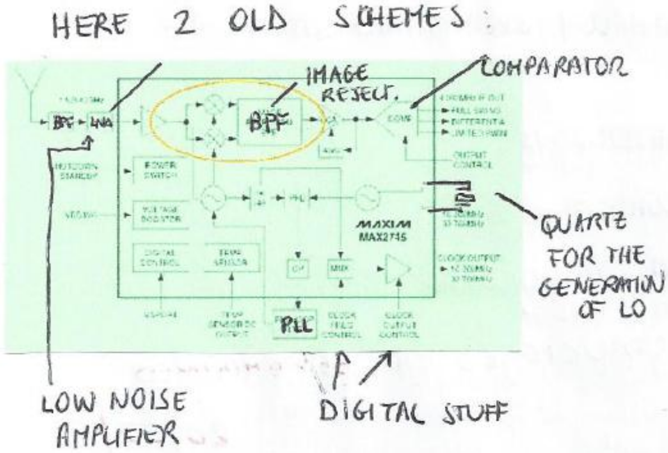


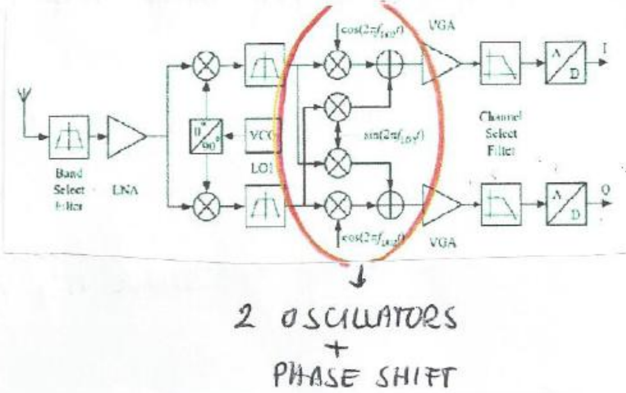
- IMAGE REJECTION MIXER



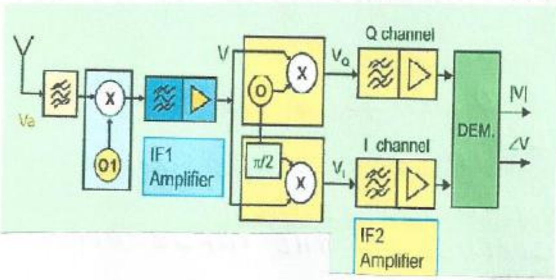
SPLITTING THE SIGNAL
↓
IT'S POSSIBLE TO GET RID OF THE IMAGE FREQUENCY MATHEMATICALLY

HERE 2 OLD SCHEMES:





HOW ARE I AND Q COMPONENTS GENERATED? **I-Q CHANNELS**:

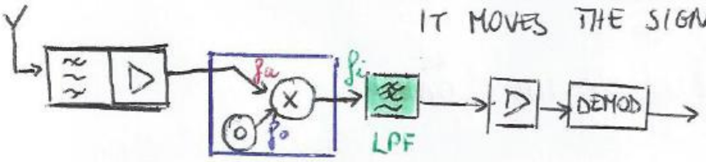


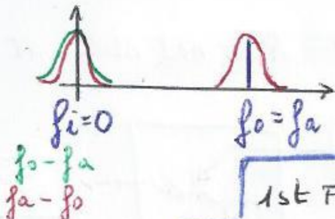
IF YOU ARE ABLE TO SPLIT THE IN-PHASE QUADRATURE COMPONENTS, YOU HAVE ALL THE INFORMATIONS ABOUT < AMPLITUDE |V| / PHASE ∠V

⇓

TECHNIQUE WHICH IS USED FOR COMPLEX PHASE / AMPLITUDE MODULATIONS

ZERO-IF RECEIVER → ≠ FROM HETERODYNE BUT SAME PRINCIPLE: INSTEAD OF MOVING RF TO IF, LOWER THAN RF, IT MOVES THE SIGNAL SPECTRUM TO THE DC





1st FILTER → TO SELECT THE BAND
MIXER → TAKES A SIGNAL AND MOVES IT TO LOWER FREQUENCIES
LOW-PASS FILTER → TO REMOVE THE REPLICA

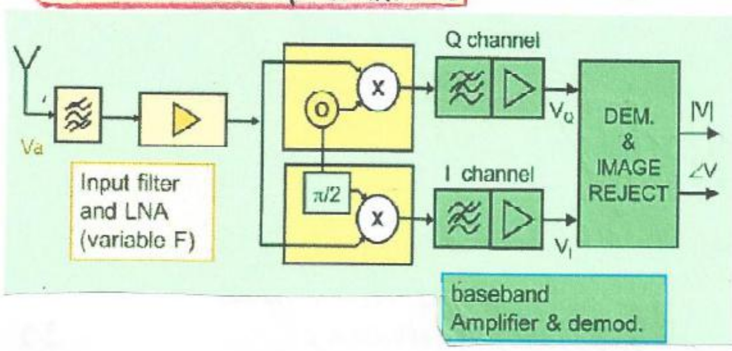
ADVANTAGES:

✓ JUST ONE MIXER, BETTER
✓ LPF → INSTEAD OF BPF
SIMPLER, MORE ROBUST WITH VERY GOOD PERFORMANCES

DISADVANTAGES

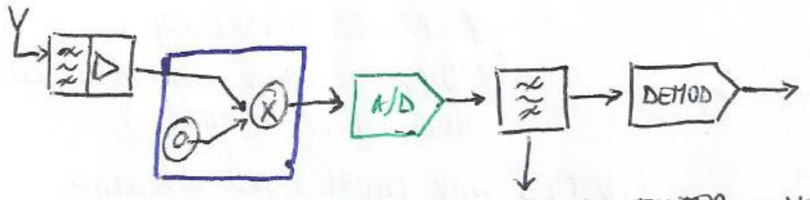
✗ IT'S IMPOSSIBLE TO GET RID OF THE IMAGE
✗ OFFSET AND A SPURIOUS EFFECT (NOISE) ARE PRESENT

ZIF WITH I-Q CHANNELS



ALL THE PART AFTER THE MIXER IS IN BASE-BAND AND IT CAN BE IMPLEMENTED IN DIGITAL DOMAIN (IF THE 1st PART IS DONE PROPERLY, WITH A VERY SELECTIVE FILTER)
BUT IT NEEDS AN ADC WHICH REQUIRES A CERTAIN ACCURACY

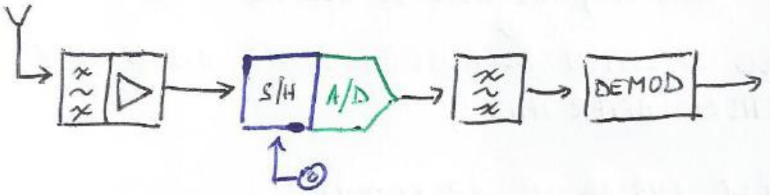
- 3rd STEP → ADC BEFORE IF FILTER



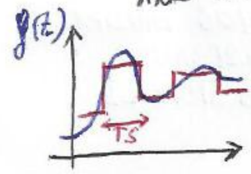
IF DIGITAL FILTER → MORE EASY TO BUILD/MODIFY THAN ANALOG FILTER

YOU ONLY PAY WITH AN HIGHER REQUIREMENT OF POWER
(BECAUSE OF COMPLEXITY OF CPU)

- LET'S REPLACE THE MIXER WITH A SAMPLE & HOLD + ADC



S/H → IS A CIRCUIT WHICH TAKES THE VALUE OF YOUR REAL SIGNAL AND KEEPS IT



SAMPLING → CONSISTS IN MULTIPLYING A SIGNAL BY THE SPECTRUM OF A TRAIN OF δ

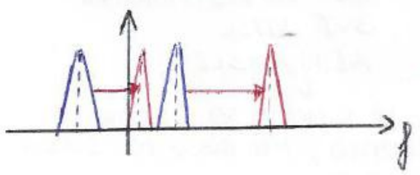
$$x[m] = f(t)\,\delta(t + nT_s)$$

$$x(t) = f(t) \cdot \sum_{k=-\infty}^{+\infty} \delta(t + kT_s)$$

$\Updownarrow$ IN FREQUENCY DOMAIN MULTIPLY MEANS CONVOLUTION

$$\boxed{X(\omega) = F(\omega) * \sum_{k=-\infty}^{+\infty} \delta\left(\omega + \frac{k}{T_s}\right)}$$

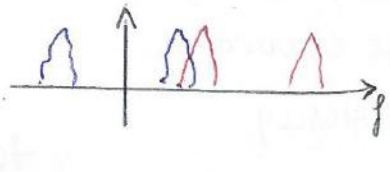
INSTEAD OF HAVING MY ORIGINAL SIGNAL I HAVE A REPETITION OF MANY REPLICA OF MY SIGNAL. THE COPIES ARE SEPARATED BY $\frac{1}{T_s}$
DUE TO THE PRESENCE OF HOLD,
GIVEN ALL THESE REPLICA → I JUST NEED TO TAKE THE ONE NEAR TO Ø SO I NEED A FILTER TO REMOVE ALL THE UNWANTED COPIES



— ORIGINAL SIGNAL
— SIGNAL OBTAINED BY MULTIPLYING THE ORIGINAL ONE WITH $\delta(t + kT_s)$ [OR CONVOLUTION IF f DOMAIN]

IF K IS LARGE ENOUGH I CAN MOVE THE SIGNAL IN THE PART OF POSITIVE f

(THE NEGATIVE PART IS THE MIRROR OF THE POSITIVE ONE THAT'S WHY SOME COPIES ARE FLIPPED)



THE OVERLAP OF THE SIGNAL WITH ITS REPLICA IS UNWANTED
I NEED TO BE SURE THAT THE SHIFT IS LARGE ENOUGH TO AVOID OVERLAPPING.

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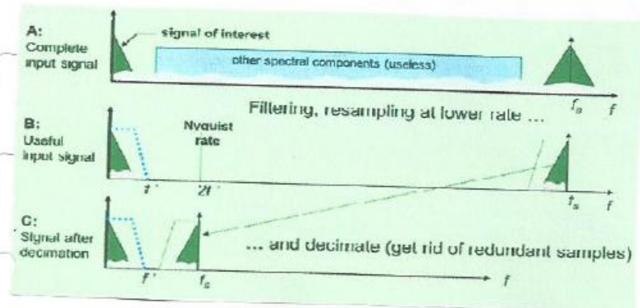
! EACH TIME I CONVERT AN ANALOG SIGNAL INTO DIGITAL WORLD
↓
I HAVE THE PROBLEM OF JITTER → IF I'M NOT PRECISE WITH JITTER I'VE AN ERROR WITH THE SIGNAL
↓
BECAUSE OF FLUCTUATION OF THE CLOCK
IT RESULTS INTO AN AMPLITUDE ERROR THROUGH THE SIGNAL SLEW RATE
↓
IT DEPENDS ON < CARRIER / SAMPLING FREQUENCY

$$\text{Amplitude error} = t_J \cdot SR \quad , \quad SR = \text{SLEW-RATE} = \frac{dv(t)}{dt}$$

→ PEAK AMPLITUDE MULTIPLIED BY ω

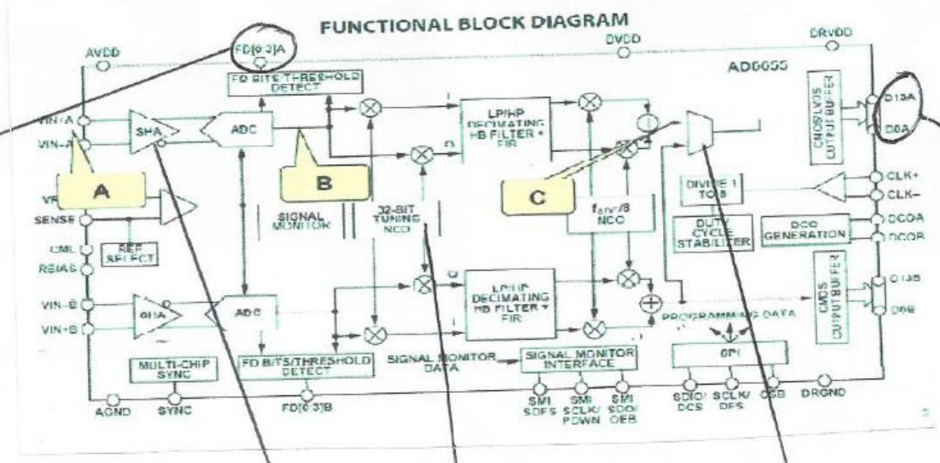
↳ DUE TO NOT EXACT TIME IN SAMPLING OF THE SIGNAL BECAUSE OF JITTER

OVERSAMPLING → LET'S CONSIDER A GROUP OF 5 CHANNELS
I JUST WANT THE 1st CHANNEL



OVERSAMPLING ALLOWS TO USE WIDE BW ANTI-ALIAS FILTERS BUT IT INCREASES THE DIGITAL PROCESSING REQUIREMENTS
↓
DECIMATION OF THE SIGNAL → IN ORDER TO TAKE INTO ACCOUNT THAT MY WANTED SIGNAL HAS A LOWER FREQUENCY (DIGITAL FILTER)
↓
NOT TOO MUCH BECAUSE OF PROBLEM OF ALIASING
✓ REDUCED COMPLEXITY OF THE ANALOG PART

GSM RECEIVER



THRESHOLD DETECTOR
↓
IT'S NOT VERY EASY TO DECODE IF N_b IS HIGH
↓
IF I HAVE A CONVERSION RANGE ~ μV I JUST WANT SOME BITS

— GOOD
NOT SATURATING AND NOT TOO SMALL

AMPLIFIER
↓
ADAPT DYNAMICS TO ONE OF THE ADC

FREQUENCY OF LO
↓
INTERNALLY GENERATED

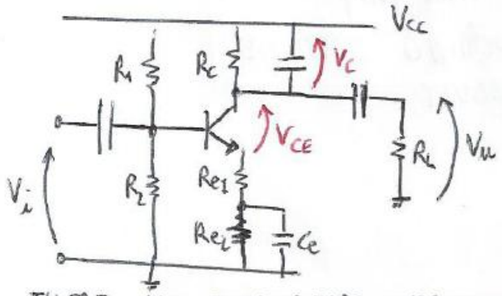
MUX → ALLOWS TO INTERLEAVE THE 2 CHANNELS < A / B

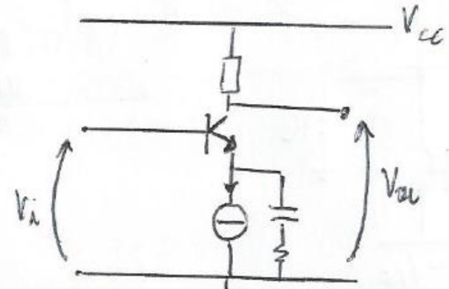
14 BITS OUTPUT

B2 - AMPLIFIERS NON LINEARITY

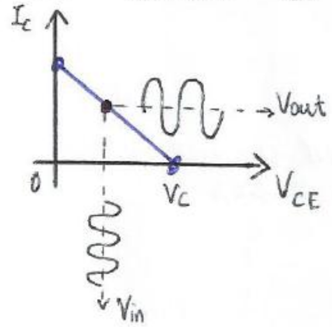
22/03/17

GIVEN THE COMMON EMITTER REFERENCE CIRCUIT

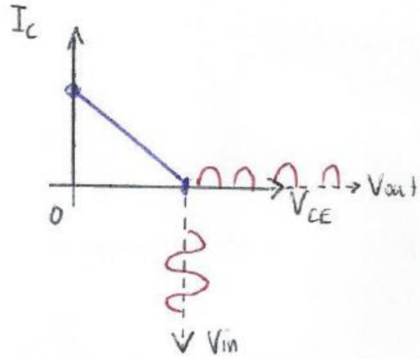




THERE ARE DIFFERENT WORKING CLASSES DEPENDING ON THE BIAS POINT
I CAN RELATE V_{CE} WITH R_C ($V_{CE} > 0.2\,V$ IN ORDER TO AVOID SATURATION)



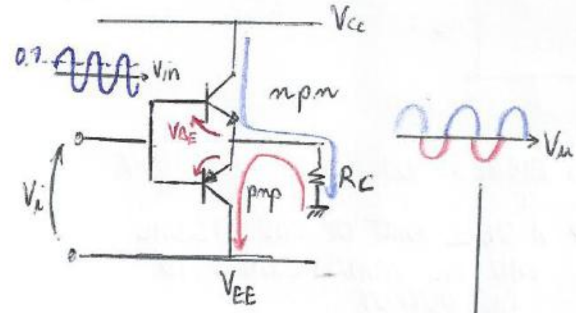
LET'S PLACE IN THE MIDDLE OF THIS STRAIGHT LINE
↓
IT'S AN AMPLIFIER OF $\boxed{\text{CLASS A}}$ [GOOD FOR SOUND / HIGH POWER DISSIPATION]
↓
THE SIGNAL IS AMPLIFIED WITH A LIMITED DYNAMIC → THERE'S A GOOD REPRESENTATION OF THE SIGNAL



$\boxed{\text{CLASS B}}$ AMPLIFIER
↓
V_{out} IS ONLY THE POSITIVE PART OF MY SIGNAL

LOWER DISSIPATION BUT DISTORTED SIGNAL

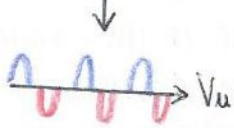
THERE ARE ALSO CLASS B AMPLIFIERS WHICH ARE ABLE TO AMPLIFY BOTH THE NEGATIVE AND THE POSITIVE PART.



$\boxed{\text{PUSH-PULL CONFIGURATION}}$

IF NO INPUT → NO CURRENT FLOWS TO THE LOAD
IF I INCREASE V_{in} TO A POSITIVE VALUE → WITH V_{in} LARGE ENOUGH ($V_{in} > 0.7\,V$),
⇓
ACTIVE REGION → CURRENT FLOWS FROM V_{cc} TO THE LOAD —

IF V_{in} LOW ENOUGH → CURRENT FLOWS IN pnp AND PROVIDES A NEGATIVE SIGNAL —



IN REALITY THE OUTPUT SIGNAL IS DISTORTED

[POWER DISSIPATION]

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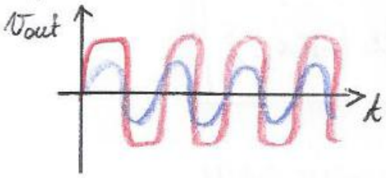
NON LINEAR MODEL

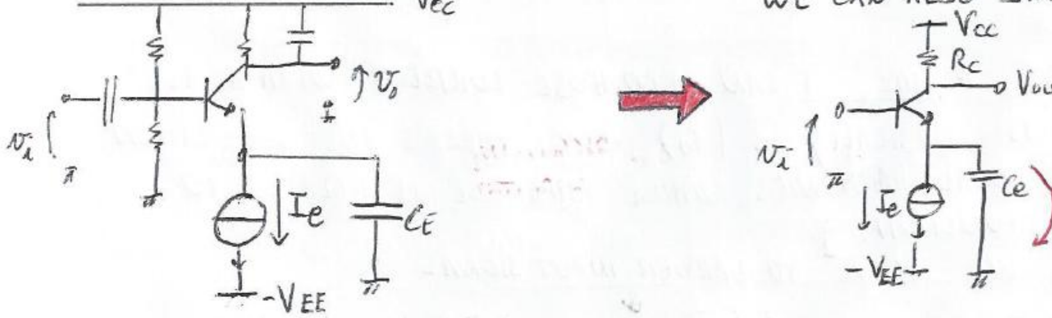
BY AMPLIFYING THE INPUT SIGNAL → IT SATURATES THE DYNAMIC RANGE



WHAT CAUSES SATURATION AND NON LINEAR BEHAVIOR?
THE SIGNAL ISN'T A SINE ANY MORE
BECAUSE THERE ARE OTHER COMPONENTS
(HARMONIC COMPONENTS)

LET'S REDUCE THE CIRCUIT

WE CAN ALSO SIMPLIFY MORE



WE CAN DEFINE A POSITIVE V_E BY REVERSING THE DIRECTION OF THE ARROW

WITH A PROPER SELECTION OF $\begin{cases} V_i \\ I_e \end{cases}$ I HAVE THE ORIGINAL CIRCUIT

WE'RE ASSUMING THAT THERE'S NO RESISTANCE IN THE EMITTER, SO WE'RE CONSIDERING R_e SHORTCIRCUITED.

$$v_i = V_i \cos(\omega_i t) = V_{BE} - V_E$$

IN THE PREVIOUS CIRCUIT WE'RE FIXING I THANKS TO THE EMITTER
IN THIS CASE IF I CHANGE THE VOLTAGE THERE'S NOTHING FIXING I_E
AND THE ONLY THING I CAN WRITE IS:

$$I = I_E = I_C = I_S e^{\frac{V_{BE}}{V_T}}$$

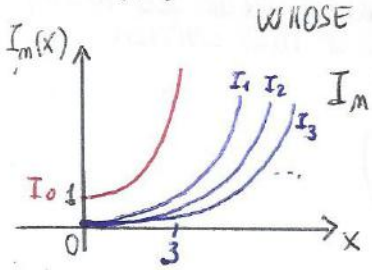
$$\Downarrow$$

$$i_c = I_S e^{\frac{V_E + v_i}{V_T}} = I_S e^{\frac{V_E}{V_T}} e^{\frac{V_i \cos(\omega_i t)}{V_T}}$$

LET'S COMPUTE THE FOURIER SERIES:

$$e^{x\cos(\omega_i t)} = I_0(x) + 2\sum_{n=1}^{\infty} I_n(x)\cos(n\omega_i t) \quad , \quad x = \frac{V_i}{V_T}$$

$I_n(x)$ => ARE THE BESSEL MODIFIED 1st ORDER FUNCTIONS WHOSE MATHEMATIC DEFINITION:



$$I_n(z) = \frac{1}{2\pi i} \oint e^{\frac{z}{2}} e^{\frac{t+1}{t}} e^{-n-1} dz$$

[Matlab command Besseli(n,x)]

$$\lim_{x \to 0} I_n(x) = \frac{\left(\frac{x}{2}\right)^n}{n!}$$

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EFFECTS OF NON LINEARITY

- → HARMONICS EFFECT
  - SIGNAL DISTORSION
  - GENERATION OF CONTENTS: $\cos(2\omega_i t)$, $\cos(3\omega_i t)$ ...
  - INTERMODULATION
- → GAIN CHANGES → IT DEPENDS ON V_i

A SINUSOIDAL SIGNAL WITH AMPLITUDE 13 [AFTER 52] mV IS APPLIED IN INPUT

LET'S DRAW THE SPECTRUM OF V_{out} ASSUMING $V_o(\omega_i) = 0$ dB

ASSUME Z_C REAL $(Z_C = R_L)$

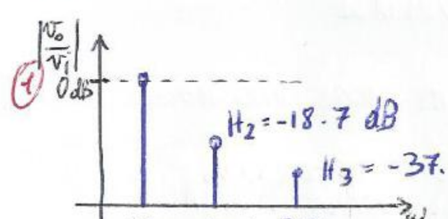
$$V_o = -R_C I - R_C I \frac{I_1(x)}{I_0(x)} \cos(\omega_i t) - R_C I \frac{I_2(x)}{I_0(x)} \cos(2\omega_i t)$$

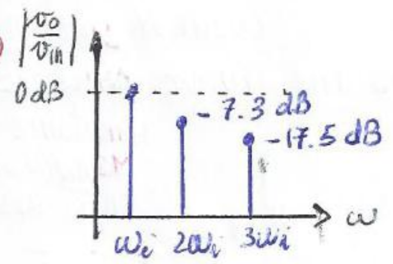
① $X = \dfrac{13\ mV}{26\ mV} = 0.5$ 1st CASE

② $X = \dfrac{52\ mV}{26\ mV} = 2$ 2nd CASE

NOW LOOK AT TABLES (OR COMPUTE THOSE VALUES WITH THE FUNCTION Besseli in Matlab)

X	$2I_1(x)/I_0(x)$	$2I_2(x)/I_0(x)$	$2I_3(x)/I_0(x)$	
0.0	0	0	0	] SMALL SIGNAL
0.1	0.0999	0.0024	0.0000	
0.5	0.4850	0.0600	0.005	
⋮	⋮			
2	1.3955	0.6045	0.1866	
⋮				
10	1.8872	1.6206	1.2490	] SATURATION

①


②


$$H_2 = \left.\frac{2I_2/I_0}{2I_1/I_0}\right|_{dB}$$

SIGNAL IS MORE DISTORTED BECAUSE OF AN HIGHLY CONTRIBUTE OF THE HARMONICS

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INTERMODULATION

↓

IS AN EFFECT DUE TO THE APPLICATION OF 2 INPUT SIGNALS (OR A SIGNAL WITH 2 TONES WHICH ARE VERY CLOSE)

DUAL-TONE INPUT: $V_i = V_a + V_b$

$$V_i^2 = (V_a + V_b)^2 = V_a^2 + 2 V_a V_b + V_b^2$$ 2nd ORDER COMPONENTS

LOCATED AT $2F_a$, $F_a - F_b$, $F_a + F_b$, $2F_b$

↓

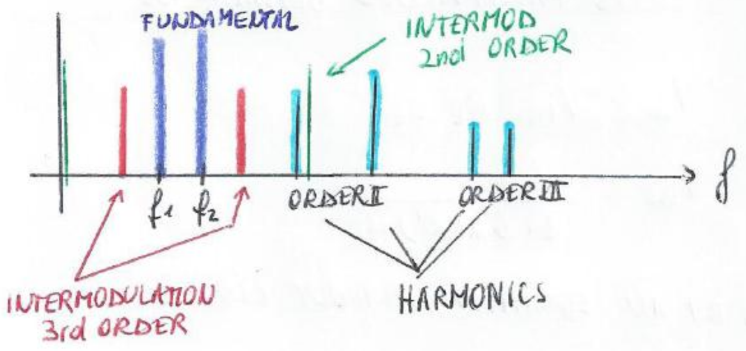
THESE COMPONENTS ARE CLOSE TO THE DC OR FAR FROM THE FUNDAMENTALS, SO THERE'S NO PROBLEM, BECAUSE THEY CAN BE EASILY FILTERED OUT (OUTBAND)

CONSIDERING 3rd ORDER COMPONENTS

$$V_i^3 = (V_a + V_b)^3 = V_a^3 + 3 V_a^2 V_b + 3 V_a V_b^2 + V_b^3$$

@ $3F_a$; $2F_a - F_b$, $2F_a$; $2F_b - F_a$, $2F_b$; $3F_b$

! THESE COMPONENTS ARE VERY CLOSE TO THE FUNDAMENTALS AND CANNOT BE FILTERED OUT (IN BAND COMPONENTS)



THESE SPURIOUS SIGNALS ALTERATE THE INFORMATION!

IN ORDER TO QUANTIFY AND TO EVALUATE THIS EFFECT (OF 3rd ORDER TERMS)

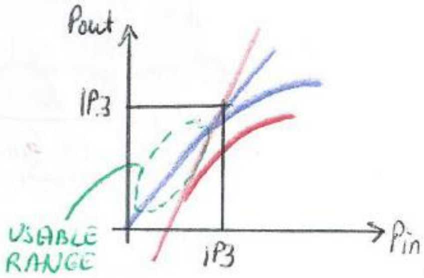
↓

$\boxed{\text{INTERCEPT POINT}}$ = POINT WHERE

LINEAR CHARACTERISTIC OF LINEAR MODEL

LINE CHARACTERISTIC OF AN HARMONIC

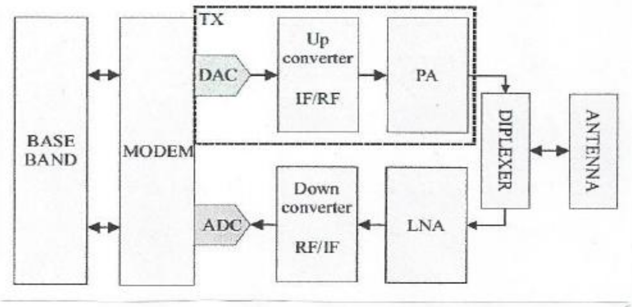
INTERCEPT THEMSELVES



(YOU CAN PUSH YOUR AMPLIFIER UP TO THIS POINT BEFORE RISKING TO TRANSFORM IT INTO A MIXER WITH HIGH DISTORSION)

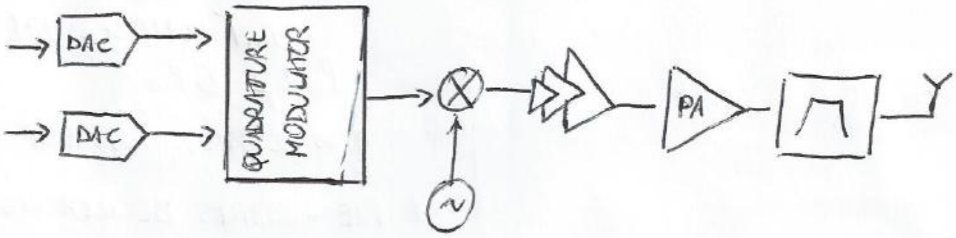
IP → IS ALWAYS A PROJECTION
BEYOND THIS POINT WE DON'T KNOW EXACTLY HOW AMPLIFIER WORKS

NOTES ON TRANSCEIVER



IN A TRANSMITTER THE DIGITAL BASEBAND SIGNAL IS
↓
CONVERTED TO ANALOG (DAC)
↓
MODULATED INTO < IN-PHASE (I) / QUADRATURE COMPONENT
↓
UP-CONVERTED TO THE RF CARRIER FREQUENCY
↓
AMPLIFIED

TX:



PA = POWER AMPLIFIER → GOAL: BOOST THE ENERGY OF A SIGNAL WITHOUT MODIFYING THE INFORMATION IT IS CARRYING

→ MOST IMPORTANT FIGURES OF MERIT ARE

LINEARITY
↓
DISTORTION, DUE TO NON LINEARITY, CAN BE OBSERVED IN

- FREQUENCY DOMAIN
  ↓
  FREQUENCY COMPONENTS NOT PRESENT IN THE INPUT SIGNAL
- TIME DOMAIN
  ↓
  THE WAVEFORM SHOWS CHANGES IN SHAPE

EFFICIENCY

- GAIN

$$G_p = \frac{P_{out}(f_0)}{P_{in}(f_0)}$$

$$P_{out}(f_0) = \frac{1}{2}\mathrm{Re}\{V_{out}(f_0) \cdot I_{out}^*(f_0)\}$$

G_p IS OPERATIVE GAIN
↓
LINKED TO EFFICIENCY

$$\boxed{\eta = \frac{P_{out}(f)}{P_{DC}}}$$

OUTPUT POWER OF THE FUNDAMENTAL

POWER DISSIPATED IN DC

- PAE = POWER ADDED EFFICIENCY

$$\boxed{PAE = \frac{P_{out}(f_0) - P_{in}(f_0)}{P_{DC}} = \eta\left(1 - \frac{1}{G_p}\right)}$$

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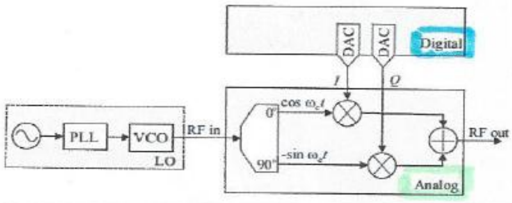
UPCONVERSION (TX)



THERE ARE I/Q < PHASE / AMPLITUDE ERRORS

DC OFFSET → IS SUPERIMPOSED TO THE PART OF THE CONVERSION

THERE'S NOT AN IDEAL MIXER → SOME PROBLEMS OCCUR:

OSCILLATOR PULLING → PHENOMENON OF DIRECT-UP CONVERSION

IT COMES FROM THE FACT THAT < LO / RF } HAVE THE SAME FREQUENCY

↓

IT'S POSSIBLE THAT, AFTER THE OUTPUT SIGNAL HAS BEEN MODULATED BY THE SIGNAL COMING FROM THE DAC, AN UNWANTED SIGNAL IS SUPERIMPOSED

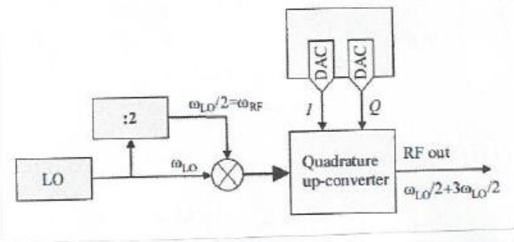
↓

A PERIODICAL MODULATION OF THE LO OUTPUT PHASE OCCURS

THIS PROBLEM CAN BE AVOIDED IF $\omega_{LO} \neq \omega_{RF}$ (MORE THAN 20%)

⇓

AS A SOLUTION IN THE FOLLOWING SCHEME



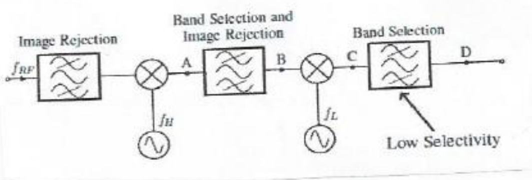
AN OSCILLATOR WHICH HAS A FREQUENCY $2\omega_{RF}$ IS ADOPTED

↓

THEN IT'S FREQUENCY IS DIVIDED BY 2 SO THAT THE INJECTION PULLING IS STRONGLY REDUCED

IN RX THIS PROBLEM IS LESS PRONOUNCED

DOWN CONVERSION (RX)



INTERMEDIATE SOLUTION BETWEEN SUPER-HETERODYNE / DIRECT CONVERSION

IS LOW IF APPROACH

↓

BASICALLY AN HETERODYNE CIRCUIT, BUT WITH THE LAST INTERMEDIATE FREQUENCY CHOSEN TO BE AS LOW AS ONE OR 2 CHANNEL BANDWIDTHS.

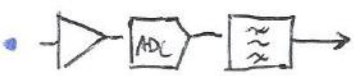
IN DOWN-CONVERSION THE CHAIN IS MADE OF 3 COMPONENTS <
- LPF → FOR CHANNEL SELECTION
- AMPLIFIER → NEEDS TO BE IN ANALOG DOMAIN, AMPLIFIES THE SIGNAL
- ADC

↓

DIFFERENT ADVANTAGES/PROBLEMS DEPENDING ON THE WAY YOU PUT THESE 3 BLOCKS

- 

LPF ALLOWS TO ATTENUATE < NOISE / OUTBAND INTERFERERS
  ↳ MOST CRITICAL COMPONENT

- 

THERE'S A WIDE BAND SIGNAL < MAKES THE WORK EASIER FOR THE LPF / LINEARITY OF AMPLIFIER IS IMPORTANT

- 

A LESS PRECISE ADC CAN BE ADOPTED

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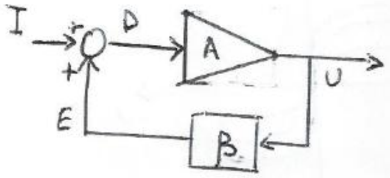
HOW IS IT POSSIBLE TO REALIZE AN OSCILLATOR?

THANKS TO POSITIVE FEEDBACK → BECAUSE WE HAVE TO SATISFY THE BARKHAUSEN CRITERION



2 CONDITIONS MUST BE RESPECTED:

$$\begin{cases} |A\beta| = 1 \\ \angle A\beta = 0 \end{cases}$$

LET'S COMPUTE THE GAIN

$U = A \cdot D$

$D = E + I = I + U\beta \rightarrow U = A(I + U\beta) \rightarrow U = \frac{A}{1-\beta A} I \Rightarrow \boxed{\frac{U}{I} = \frac{A}{1-A\beta}}$

THE BARKHAUSEN CONDITIONS COULD BE CONTROLLED → IF I'M ABLE TO START IN A CONDITION WHERE $A\beta > 1$ KEEPING THIS PHASE RELATION ($\angle A \cdot \beta = 0$) MY AMPLITUDE WILL RISE UP TO THE PHYSICAL LIMITS

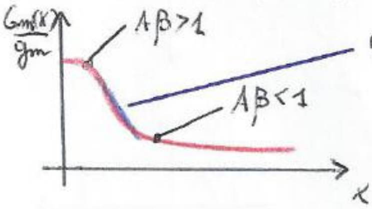
⇓

NON LINEAR AMPLIFIER : INCREASING AMPLITUDE ⇒ GAIN REDUCES

A = OPEN LOOP GAIN OF THE AMPLIFIER → IN ORDER TO SATISFY $|A \cdot \beta| = 1$

β → TO HAVE SOMETHING THAT CAN BE FREQUENCY DISPERSIVE
THANKS TO β WE ARE ABLE TO SET THE FREQUENCY ($\angle A\beta = 0$)
TO SELECT f_0

HOW CAN WE KEEP A GAIN THAT CAN BE AUTOMATICALLY CONTROLLED



YOU TURN ON THE SIGNAL WITH NOISE
THE SYSTEM STARTS CREATING OSCILLATIONS
UNTIL IT REACHES AN EQUILIBRIUM POINT WHERE $A \cdot \beta = 1$

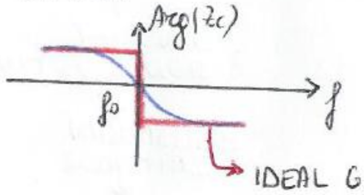
CONCERNING FREQUENCY → WE NEED TO KEEP THE PHASE ROTATION OF OUR RESONANT CIRCUIT

β → IF IT NEEDS TO BE FREQUENCY SELECTIVE

⇓

LC CIRCUIT IS NECESSARY

Q AS HIGH AS POSSIBLE IN ORDER TO HAVE A SHARP PHASE ROTATION



IDEAL GOAL → NOT POSSIBLE
IT MEANS $Q \rightarrow \infty$

⇓

WE NEED TO REACH THIS CONDITION AS CLOSE AS POSSIBLE

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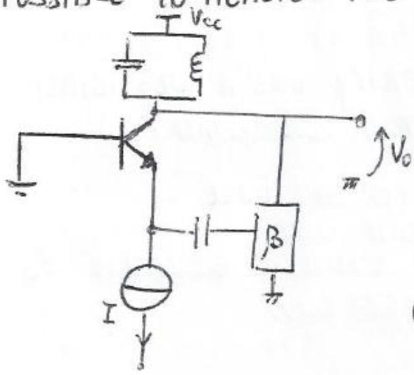
IT'S POSSIBLE TO REALIZE THE OSCILLATOR → THANKS TO A TRANSISTOR, IN COMMON BASE CONFIGURATION



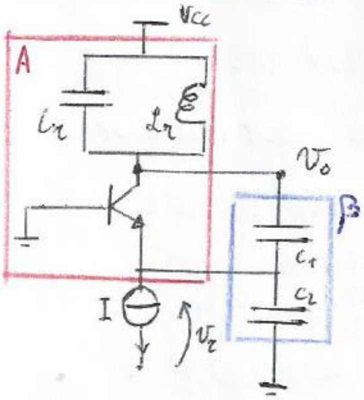
COMMON BASE → BECAUSE GAIN IS POSITIVE. THE TASK IS TO HAVE A β WHICH FULLFILLS THE REQUIREMENT OF PHASE SHIFT = 0°

OSCILLATOR IS: TRANSISTOR AMPLIFIER + LC CIRCUIT

LC CIRCUIT IS THE LOAD
GAIN IS CONTROLLED THROUGH THE NON LINEARITY

KINDS OF OSCILLATORS:

- COLLPITTS OSCILLATOR

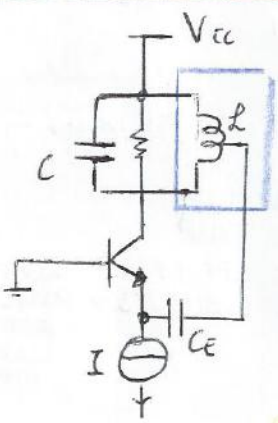


$$\boxed{V_r = V_o \frac{C_1}{C_1 + C_2}}$$

CAPACITIVE VOLTAGE DIVIDER

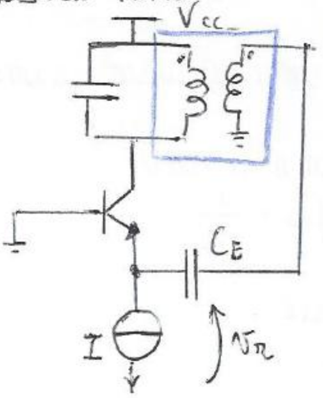
ADDING CAPACITORS → WE'RE MOVING FREQUENCY BUT WE'RE NOT INFLUENCING THE Q FACTOR (FROM THE EMITTER WE CAN SEE $1/g_m$)

- HARTLEY OSCILLATOR

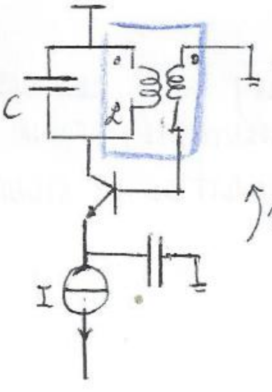


WE CAN REPLACE L TO C
INDUCTIVE VOLTAGE DIVIDER, WITH A FEEDBACK LOOP

- FEEDBACK THROUGH TRANSFORMER (→ MEISNER OSCILLATOR)
NOT GOOD @HF

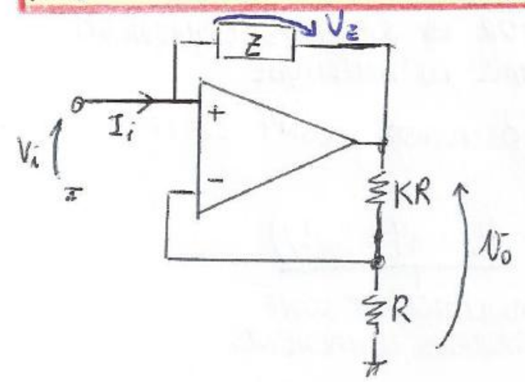


COMMON BASE CONFIGURATION



COMMON EMITTER CONFIGURATION

NIC = NEGATIVE IMPEDANCE CONVERTER

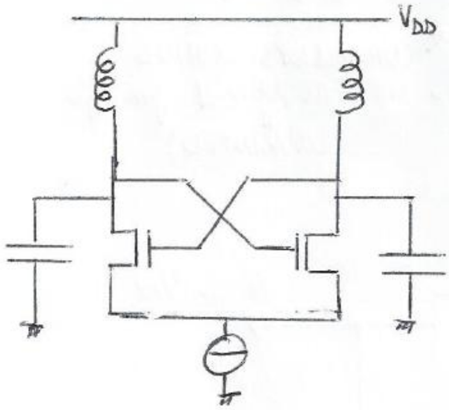


$V_u = (1+K)V_i$

$-\dfrac{V_u - V_i}{Z} = I_i \qquad Z_i = \dfrac{V_i}{I_i} = -\dfrac{Z}{K}$

GOOD ACCURACY → INCREASING INPUT AMPLITUDE THERE'S GAIN SATURATION
↓
$|Z_i|$ DECREASES AS THE SIGNAL INCREASES

A MORE COMPLEX CONFIGURATION → FOLLOWS THE SAME IDEA BUT IT'S BASED ON FET (MOS)



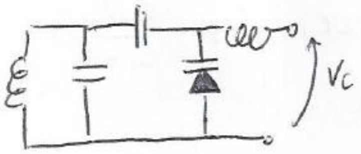
DIFFERENTIAL STRUCTURE WITH 2 COMMON SOURCE CONFIGURATIONS

$R_{eq} = -\dfrac{2}{g_m}$ SMALL SIGNAL

$R_{eq} = -\dfrac{Z}{G_m(x)}$ LARGE SIGNAL ($V_m >$ ⟹ $< G_m(x)$)

DIFFERENTIAL STRUCTURES <
- HIGH ROBUSTNESS
- INCREASE THE NUMBER OF COMPONENTS (NEGLIGIBLE WITH SILICON)
- DECREASE LOSSES

HOW TO MODIFY THE FREQUENCY OF OSCILLATION: IT'S NECESSARY TO ADD SOME PROPER COMPONENTS
↓
SOMETHING THAT CHANGES (inductor) AND (capacitor)
⇓
LET'S USE AS VARICAP A DIODE



Q → HOW CAN WE INCREASE IT? REDUCING LOSSES!

Q DEPENDS ON < SERIES / PARALLEL LOSSES

IF WE'RE ABLE TO → INCREASE PARALLEL RESISTANCES
→ ADD SEPARATION BUFFER BETWEEN < LOAD / FEEDBACK
(IN ORDER TO GET RID OF THE EMITTER IMPEDANCE)

→ USE DIELECTRIC RESONATORS (MECHANICAL)
↓
1st LEVEL OSCILLATORS → FOR A STABLE FREQUENCY REFERENCE

→ USE QUARTZ OSCILLATOR
↓
PIEZOELECTRIC MATERIAL THAT HAS A PROPER OSCILLATION FREQUENCY

ADOPTED BECAUSE <
- QUITE CHEAP
- VERY HIGH Q FACTOR
- STABLE AND PRECISE

→ OTHER RESONATORS (CERAMIC FILTERS, SAW)

29/03/17

B5 - MULTIPLIERS / MIXER CIRCUITS

MIXER = MODULE WITH WHICH IT'S POSSIBLE TO MOVE A SIGNAL FROM A FREQUENCY TO AN OTHER (UP OR DOWN)

↓

THIS TASK CANNOT BE ACHIEVED WITH JUST A LINEAR DEVICE BUT FOR EXAMPLE WITH A RESISTANCE + NON LINEAR DEVICE

THERE ARE DIFFERENT KINDS OF MIXERS IN A TRANSCEIVER

BASING ON WERNER'S RELATIONS

$$\sin A \sin B = \frac{1}{2}\left[\cos(A-B) - \cos(A+B)\right]$$

$$\sin A \cdot \cos B = \frac{1}{2}\left[\sin(A-B) - \sin(A+B)\right]$$

$$\cos A \cos B = \frac{1}{2}\left[\cos(A-B) + \cos(A+B)\right]$$

→ THE MULTIPLICATION RESULTS INTO SOMETHING WHICH IS SHIFTED IN FREQUENCY

↓

FROM A,B → TO < SUM A+B / DIFFERENCE A-B

MAIN TASKS OF MIXERS: < FREQUENCY TRANSLATION / PHASE DETECTION < PLL / DEMODULATORS

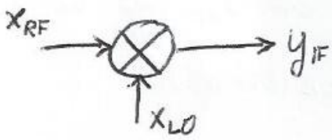
WORKING AT MEDIUM FREQUENCIES → IT'S NOT EASY TO DISTINGUISH BETWEEN TRANSLATION AND DEMODULATION

THERE ARE → IMAGE REJECTION MIXERS

→ MIXER II → MULTIPLIER OF DIFFERENT SIGNALS

IDEAL MIXER OPERATION = 3 PORT DEVICE WITH < 2 INPUT / 1 OUTPUT

MULTIPLICATION OF 2 SIGNALS < RF SIGNAL / SIGNAL COMING FROM LO

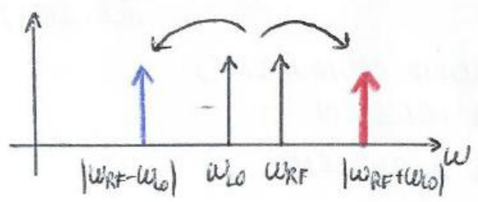


$$x_{RF} = A_{RF}\cos(\omega_{RF}t + \varphi_{RF})$$

$$x_{LO} = A_{LO}\cos(\omega_{LO}t + \varphi_{LO})$$

$$y_{IF} = \frac{A_{RF}A_{LO}}{2}\left\{\underbrace{\cos\left[(\omega_{RF}-\omega_{LO})t + \varphi_{RF} - \varphi_{LO}\right]}_{\text{DOWNCONVERSION}} + \underbrace{\cos\left[(\omega_{RF}+\omega_{LO})t + \varphi_{RF} + \varphi_{LO}\right]}_{\text{UPCONVERSION}}\right\}$$

⇓



IT'S LIKE A FREQUENCY TRANSLATION WHERE YOU CAN FILTER OUT THE UNWANTED COMPONENT OF THE TWO

MIXING → MEANS TO GET A NON-LINEAR BEHAVIOR → IT'S NECESSARY TO HAVE A NON-LINEAR DEVICE

↘ PRODUCT, BEING A NON LINEAR EFFECT, WILL BE LOSSY IN TERMS OF ENERGY

WITH PASSIVE MIXER → OUTPUT LOWER THAN THE INPUT IN TERMS OF MAGNITUDE

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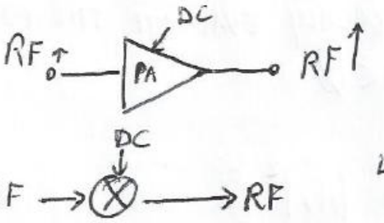
- POWER AMPLIFIER → IT INCREASES AMPLITUDE



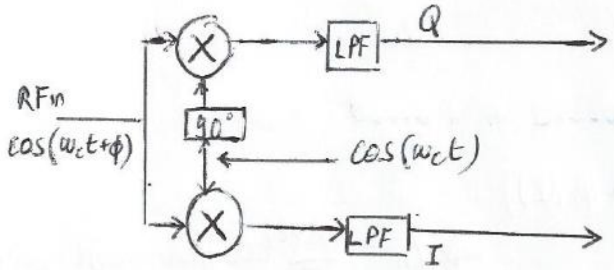
THINKING ABOUT IT FROM A DIFFERENT PERSPECTIVE IT'S A MIXER, TOO

AS FOR PA → THERE ARE SOME LIMITS
IT'S POSSIBLE TO DEFINE FOR LARGE SIGNAL
- IIP (= INTERSECT POINT)
- SATURATION (→ OCCURS WITH TOO HIGH POWER)
- GAIN

⇓

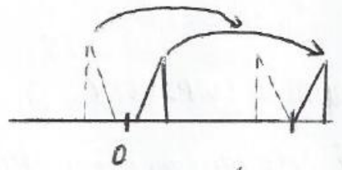
UNWANTED NON LINEARITIES ARISE

- HETERODYNE MIXER

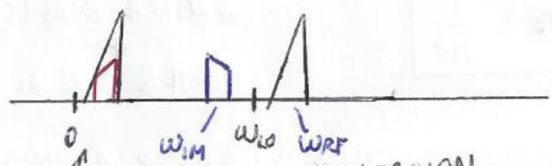


THANKS TO 2 MIXERS
WE CAN OBTAIN THE DOWNCONVERSION OF THE IN-PHASE / QUADRATURE COMPONENTS

! DURING DOWNCONVERSION → IMAGE REJECTION MUST BE TAKEN INTO ACCOUNT



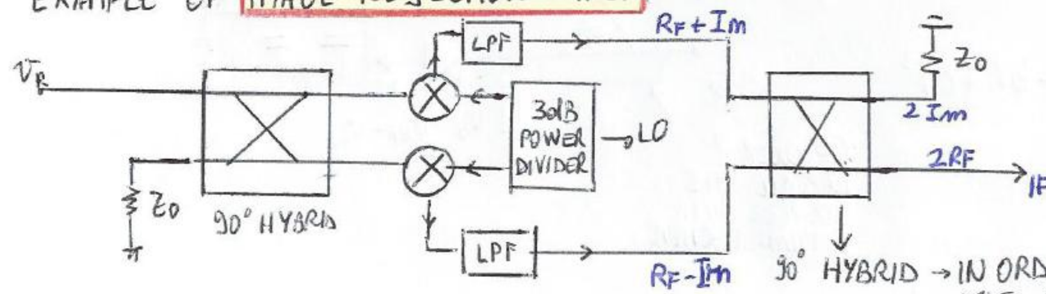
UPCONVERSION
NO PROBLEM



DOWNCONVERSION

IMPOSSIBLE TO DISCRIMINATE IF SIGNAL OR IMAGE

EXAMPLE OF IMAGE REJECTION MIXER



90° HYBRID → IN ORDER TO REMOVE THE IMAGE BY TRANSLATING IT TO THE OTHER BRANCH

WE { SPLIT / MERGE } SIGNALS → REMOVING THE IMAGE AUTOMATICALLY

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IN ORDER TO REMOVE THE v_{RF} COMPONENT, TOO

↓

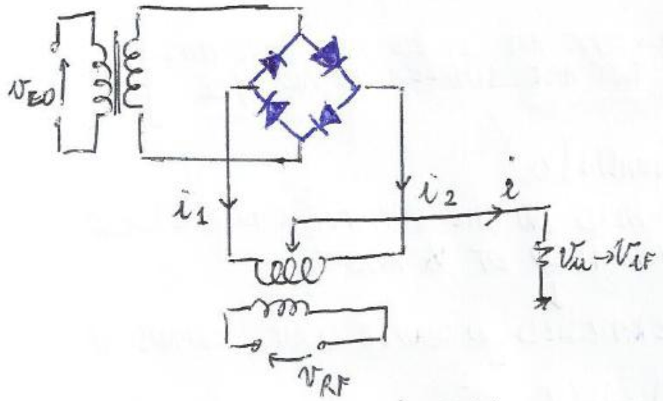
LET'S INCREASE A LITTLE BIT THE COMPLEXITY

↓

DOUBLE BALANCED MIXER → ACHIEVES AT THE SAME TIME MIXING WHAT WE WANT IN A ~~ALMOST~~ IDEAL WAY

↓

WITH LO/RF REJECTION



$$i = i_1 + i_2 = \boxed{8 b v_{LO} v_{RF}}$$

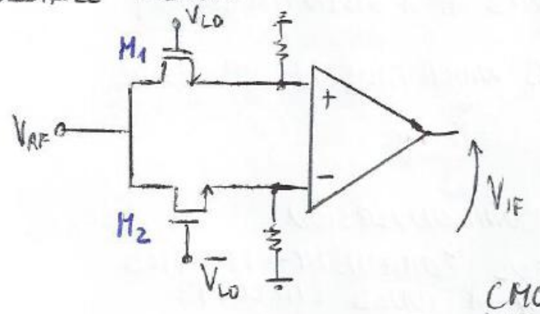
IT REMOVES PART OF NON LINEARITY, TOO (ODD-ORDER NON LINEAR TERMS)

WITH EQUAL DIODES PROCESSES ARE TRULY THE SAME VS TEMPERATURE VARIATIONS, ETC...

TRANSFORMERS → FOR LOW/MEDIUM FREQUENCIES

30/03/17

POSSIBLE PASSIVE IMPLEMENTATION → CMOS PASSIVE SINGLE BALANCED MIXER



↓

PASSIVE MIXER EVEN IF IT'S EMPLOYING ACTIVE ELEMENTS

SAME WORKING PRINCIPLE AS SINGLE-ENDED

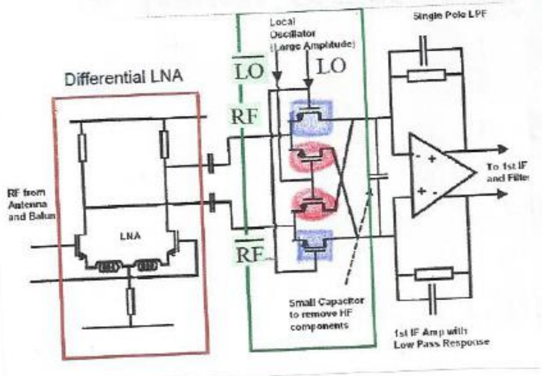
↓

CMOS, ACTING LIKE A SWITCH, MAKES IMPOSSIBLE FOR THE V_{RF} TO REACH V_{LO}

$$\boxed{G_t = \frac{4}{\pi^2}}$$

→ PAY 6 dB (IDEALLY 4 dB)

RF → IS MULTIPLIED WITH A PERIOD THAT COMES FROM THE LOCAL OSCILLATOR FREQUENCY



A DOUBLE BALANCE CONFIGURATION CAN BE MADE REPLACING DIODES WITH MOS

CONDUCTIVE → IN 1st HALF T

CONDUCTIVE → IN 2nd HALF T

MULTIPLICATION BY A SQUARE WAVE

↓

FEASABLE SOLUTION EMPLOYED FOR LOW COST MIXERS

PASSIVE MIXERS → ARE A SOLUTION FOR MANY APPLICATIONS WHEN YOU NEED TO KEEP NOISE VERY LOW

↓

IN OTHER CASES YOU DO CONVERSION KEEPING THE AMPLITUDE OF YOUR SIGNAL

↓

ACTIVE MIXERS → USE TRANSCONDUCTANCE OF ACTIVE ELEMENTS

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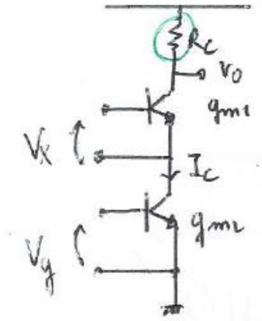
TRANSCONDUCTANCE MULTIPLIER → 1 QUADRANT → DC COMPONENTS: HIGH FEEDTHROUGH



$V_x, V_y \geq 0$

$V_0 = V_x g_{m1} R_c$ $\quad V_{cc}$ ↓ , $g_{m1} = \frac{I_c}{V_T}$

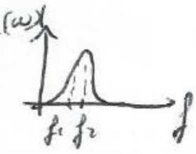
$I_c \simeq g_{m2} V_y$

$V_0 \simeq V_x \frac{V_y}{V_T} g_{m1} g_{m2} R_c$ K ➡ $V_0 = k V_x V_y$

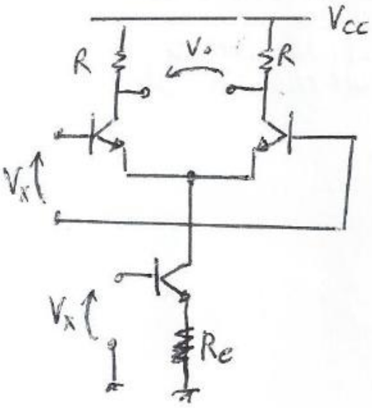
ACTING ON R_c → WE REPLACE A RESISTOR WITH A TUNED FILTER $Z_c(\omega)$
↓
TUNES THE BEHAVIOR OF OUR SIGNAL

⇓

SHAPING PROPERLY THE FILTER MEANS TO CLEAN FROM THE OUTPUT THE UNWANTED COMPONENTS
↓
THIS REDUCES THE PHASE NOISE



THIS CONFIGURATION → WORKS ON 1 QUADRANT
↓
IT CAN BE EXTENDED TO MORE QUADRANTS (2/4) BY MAKING IT DIFFERENTIAL

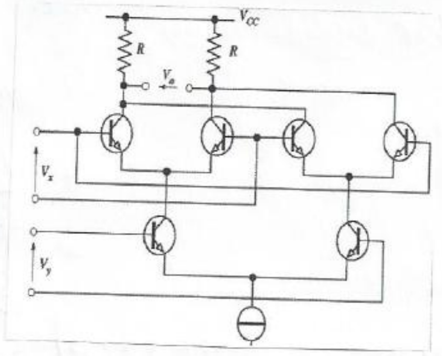


THIS REMOVES THE COMMON MODE (ROBUST CIRCUIT TO DC DISTURBANCES, ETC...)

WE'VE REDUCED (IN THEORY NULLED) THE FEEDTHROUGH FROM V_y

THERE IS STILL FEEDTHROUGH FROM V_x
↓
DOUBLE BALANCED MIXER
↓
TO GET RID OF DETRIMENTAL EFFECTS OF f_x, f_y
↓
IT'S A ROBUST SOLUTION

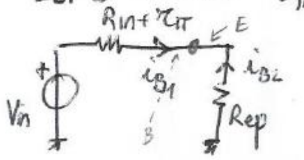
IN ORDER TO REMOVE FEEDTHROUGH FROM $< \begin{matrix} V_x \\ V_y \end{matrix}$



WE CAN EMPLOY MOS INSTEAD OF BJT (→ MOS GILBERT CELL)

! WE'RE KEEPING ALL CONSIDERATIONS UNDER THE ASSIGNMENT → SMALL SIGNAL

LET'S WRITE v_{in} AS A FUNCTION OF i_B → SOLVING THE MESH :



$$v_{in} = i_{B1}\left(R_{in}+r_\pi\right) + i_{B1}\,\frac{\left(R_{in}+r_\pi\right)}{\cancel{(\beta+1)}}\,\cancel{(1+\beta)}$$

R_{eq}

$$v_{in} = 2\left(R_{in}+r_\pi\right) i_{B1}$$

$$i_{B1} = \frac{v_{in}}{2\left(R_{in}+r_\pi\right)}$$

USING (1)

$$\begin{cases} V_{out_1} = -R_C\,\beta\, i_{B1} = -R_C\,\beta\,\dfrac{v_{in}}{2\left(R_{in}+r_\pi\right)} = -\dfrac{R_C\, g_m r_\pi}{2\left(R_{in}+r_\pi\right)} \\ V_{out_2} = -V_{out1} \end{cases}$$

$i_{B2} = -i_{B1}$

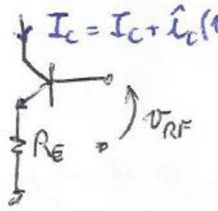
⇓

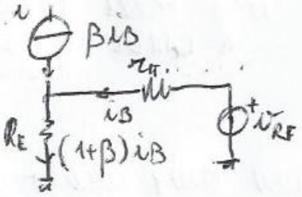
$\hat{\imath}(v_{RF})$

$$\begin{cases} V_{out_1}(t) = V_{CC} - \dfrac{R_C I_0}{2} - \dfrac{R_C g_m r_\pi}{2\left(R_{in}+r_\pi\right)}\, v_{LO}(t) \simeq V_{CC} - \dfrac{R_C I_0}{2} - \boxed{\dfrac{R_C I_0}{4V_T}\, v_{LO}(t)} \\ V_{out_2}(t) = V_{CC} - \dfrac{R_C I_0}{2} + \dfrac{R_C g_m r_\pi}{2\left(R_{in}+r_\pi\right)}\, v_{LO}(t) \simeq V_{CC} - \dfrac{R_C I_0}{2} + \dfrac{R_C I_0}{4V_T}\, v_{LO}(t) \end{cases}$$

NOW LET'S REPLACE I_0 WITH $I_0 + \hat{\imath}$
WITH $\hat{\imath}$ = CURRENT WHICH DEPENDS ON SMALL SIGNAL FROM A GIVEN VOLTAGE
$\hat{\imath}(v_{RF})$
⇓
IN THIS WAY IT DEPENDS ON $\left\langle \begin{matrix} v_{LO} \\ v_{RF} \end{matrix} \right.$ ⇒ ADDING THE STANDARD COMMON EMITTER

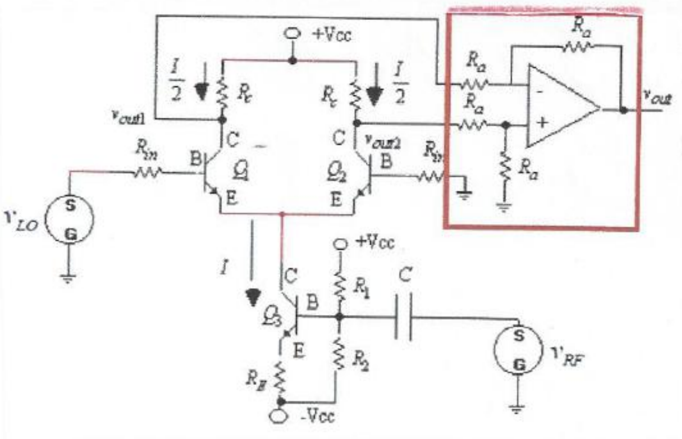
$I_C = I_C + \hat{\imath}_C(v_{RF})$





$$i_B = \frac{v_{RF}}{r_\pi + (1+\beta)R_E}$$

$$i = \hat{\imath}_C = i_B\beta = \frac{\beta v_{RF}}{r_\pi + (1+\beta)R_E} \overset{\beta\to\infty}{\simeq} \frac{v_{RF}}{R_E}$$



AT LAST → LET'S PUT A UNIT-GAIN AMPLIFIER

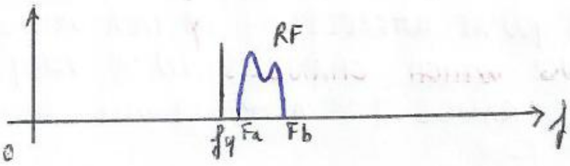
$$V_{out} = V_{out_2}(t) - V_{out_1}(t) =$$

$$= \frac{R_C I_0}{2V_T}\, v_{LO}(t) + \frac{R_C}{2V_T R_E}\, v_{RF}(t)\, v_{LO}(t)$$

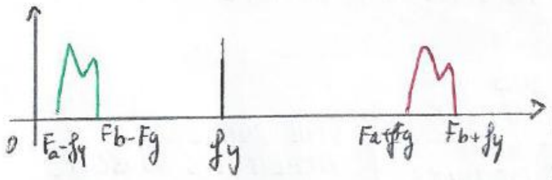
↑

SINGLE-BALANCED MIXER CONTAINS A COMPONENT COMING FROM THE LO

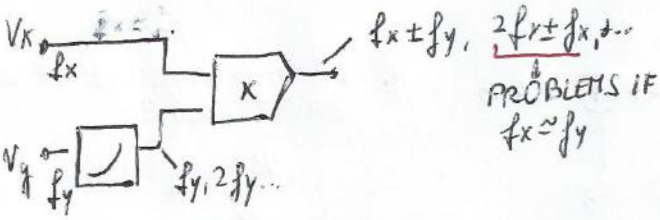
LET'S CONSIDER SOMETHING CLOSER TO REALITY
↓



SIGNAL IS DEFINED IN A BAND
↓
YOU TRANSFER A RANGE OF FREQUENCIES

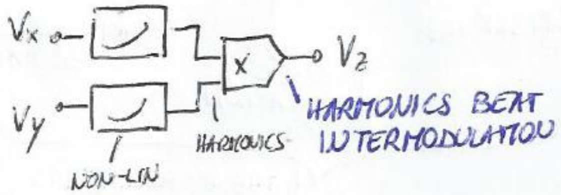


WHAT HAPPENS IF ONE SIGNAL IS NON LINEAR? IT CONTAINS SPURIOUS ELEMENTS

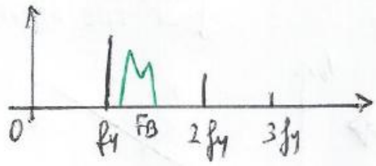


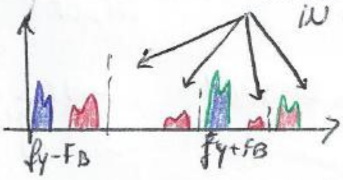
PROBLEMS IF $f_x \simeq f_y$ => INTERMODULATION TERMS (MAY) APPEAR

IF BOTH INPUTS ARE NON LINEAR => THE OUTPUT WILL **ALWAYS** HAVE INTERMODULATION COMPONENTS

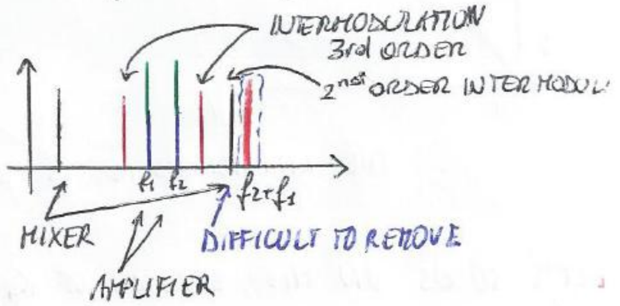


example:





SIGNAL LOCATED IN OTHER FREQ. BAND
↓
REDUCE THE SPECTRAL PURITY OF YOUR BAND



REDUCING NOISE → IT'S POSSIBLE TO HAVE

SSB
↓
NOISELESS FILTER HAS A +3dB NOISE FIGURE

DSB
↓
NOISELESS FILTER HAS A 0 dB NOISE FIGURE

NOISE SOURCES IN GILBERT MIXERS
- DUE TO LOADS
- LO SWITCHES
- THERMAL NOISES

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HOW TO REALIZE A LOGARITHMIC AMPLIFIER → WITH A PN JUNCTION (DIODE)

IF WE DRIVE A PROPER CURRENT I_E

WE CAN GET V_D AS A LOGARITHMIC FUNCTION OF I_E APART FROM $\eta \in (1 \div 2)$, V_T DEPENDENT ON TEMPERATURE

$$\boxed{V_D = \eta V_T \ln \frac{I_E}{I_S}}$$

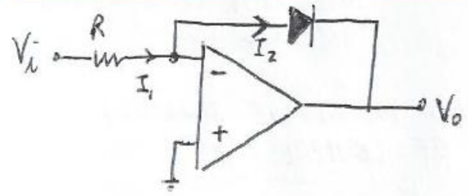
SET CURRENT → READ VOLTAGE

WE NEED TO GET RID OF THESE 2 PARAMETERS

⇓

ADOPTING A PROPER CONFIGURATION

LOGARITHMIC AMPLIFIER: 1st SCHEME



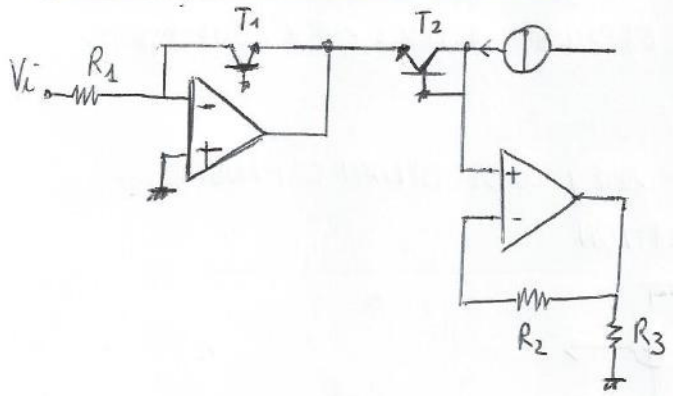
$$\begin{cases} V_0 = -V_D \\ I_1 = I_2 = \frac{V_i}{R} \end{cases} \rightarrow I_2 = I_S e^{\frac{V_D}{\eta V_T}}$$

$$V_u = -\eta V_T \ln \left(\frac{V_i}{I_S R} \right)$$

RESULT IS AFFECTED BY PROBLEMS RELATED TO TERMS WHICH CHANGE THE SOLUTION

IN ORDER TO SOLVE THE PROBLEM → 2 STANDARD CONFIGURATIONS:

- (INVERTING) LOGARITHMIC AMPLIFIER



$$\boxed{V_u = -V_T \ln \left(\frac{R_3 + R_2}{R_3} \right) \ln \left(\frac{V_i}{R_1 I} \right)}$$

T_1 = LOGARITHMIC JUNCTION → TO GET RID OF η

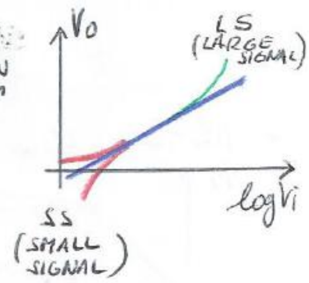
T_2 = REFERENCE JUNCTION

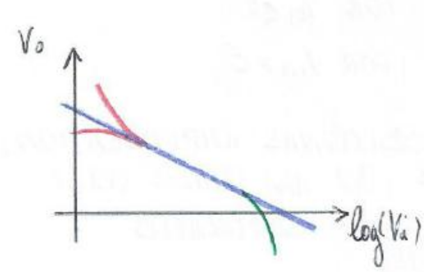
→ IN SOME WAY COUNTERACTS THE VARIATIONS DUE TO T EVEN THOUGH IT'S STILL SUBJECTED TO V_T

THERE ARE STILL ERRORS DUE TO
- LOW V ACCROSS R_1 → OFFSET OF OP-AMP
- LOW I IN log JUNCT → I_{off} AND I_{bias}

} LOW V_{in}

- ADDITIONAL VOLTAGE DROP ON THE SMALL RESISTANCE r_{BB} OF BJT } HIGH V_{in}

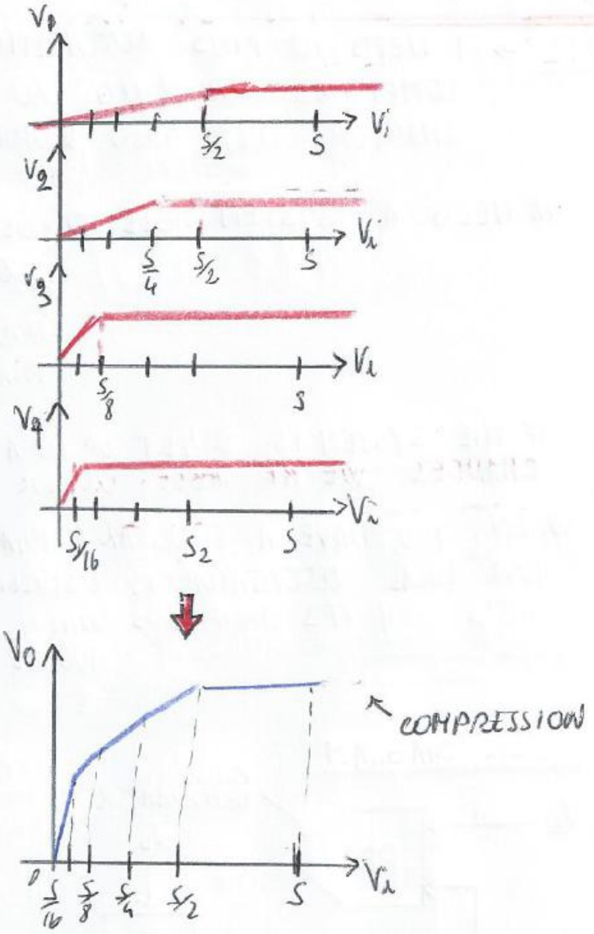


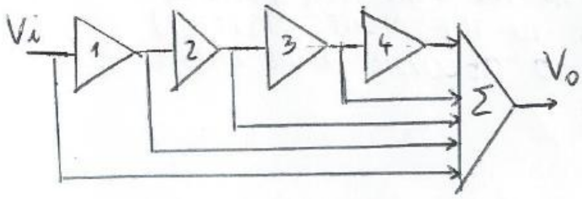


SAME CONDITIONS FOR THE INVERTING FUNCTION

SATURATING CHAIN ($S = V_0$ AT SATURATION)

- STAGE 1 → G = A FOR $V_i < \frac{S}{2}$
  SATURATED FOR $V_0 = 2V_i$
- STAGE 2 → GAIN = A FOR $V_i < \frac{S}{4}$
  SATURATED FOR $V_0 = 4V_i$
- STAGE 3 → GAIN = A FOR $V_i < \frac{S}{8}$
  SATURATED FOR $V_0 = 8V_i$
- STAGE 4 → GAIN A FOR $V_i < \frac{S}{16}$
  SATURATED FOR $V_0 = 16 V_{in}$





LOGARITHMIC AMPLIFIERS → CHARACTERISTICS

- WIDE DYNAMIC RANGE → GOOD FOR AC
- FOR RF POWER MEASUREMENTS
- RSSI (= RECEIVED SIGNAL STRENGTH INDICATOR) → MEASURES THE POWER AT THE OUTPUT OF AN ANTENNA

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LINEARITY → CONSIDERING LINEAR THE DIFFERENT BLOCKS
- PD → SOME OF THEM ARE NON-LINEAR
- FILTER → COULD BE < PASSIVE / ACTIVE
- VCO → MOST OF THEM → NON-LINEAR $\omega(V_c)$

↓

IS NOT TRIVIAL WORK

PLL TRANSFER FUNCTION

IDEA → CLEAR RELATIONSHIP < ISTANTANEOUS PHASE / TIME

$$v_i = V_i \sin(\omega_i t + \theta_i)$$
$$v_0 = V_0 \cos(\omega_0 t + \theta_0)$$
$$V_d = K_d(\theta_i - \theta_0)$$
$$V_c(s) = V_d(s) \cdot F(s)$$
→ WE NEED A FILTER BECAUSE WE AREN'T INTERESTED IN ALL POSSIBLE VARIATIONS BUT JUST A PART OF THE SIGNAL IS SIGNIFICANT
LPF IS USUALLY IMPLEMENTED

$$\Delta\omega_0 = K_0 V_c$$
$$\Delta\omega_0(t) = \frac{d\theta_0}{dt} \longrightarrow \Delta\omega_0(s) = s\,\theta_0(s)$$
→ THAT'S WHAT ARRIVES AT THE VCO

WE STARTED FROM 2 DIFFERENT SIGNALS
↓
WE EVALUATED THE PHASE DIFFERENCE
↓
WE FILTERED OUT THE UNWANTED (HIGH ORDER) COMPONENTS
↓
WE REALIZE A FREQUENCY VARIATION ON OUR LOCAL OSCILLATOR (INJECTION ON VCO) TILL 2 ω ARE EQUAL

⇒ THE OVERALL TRANSFER FUNCTION:
$$s\,\theta_0(s) = K_0 V_d(s) F(s) = K_0 K_d F(s)(\theta_i - \theta_0)$$
⇓

PLL TRANSFER FUNCTION:
$$\boxed{\frac{\theta_0(s)}{\theta_i(s)} = \frac{K_0 K_d F(s)}{s + K_0 K_d F(s)}}$$
RELATED TO GAINS THEN SELECTED BY THE FILTER

PLL → SENSES AND HANDLES PHASE → FOR GIVING US A REFERENCE f

LOCK CONDITION → MEANS $\omega_0 = \omega_i$ ⇒ PHASE DIFFERENCE DOESN'T CHANGE, IT'S KEPT CONSTANT ($\theta_e = \text{const}$)

↘ AS ω_i CHANGES
↓

LOCK KEEPING MECHANISM
- θ_e = ERROR IN TERMS OF DIFFERENCE → CHANGES.
- THE CORRESPONDING V_d IS MODIFIED AND FILTERED BY $F(s)$
- ↓ IT PULSES THE VCO TO MODIFY f UNTIL $\omega \neq \omega_0$

THERE ARE SOME POSSIBLE CASES → WHERE IT'S NOT POSSIBLE TO REACH THE LOCK CONDITION

⇒ PLL-ORDER → $\boxed{H(s) \text{ ORDER} = F(s) \text{ ORDER} + 1}$

PLL ORDER
- 1 → PARAMETER ω_0
- 2 → PARAMETERS ω_0, ξ

} DC GAIN ($F(0)$)

GAIN OF THE LOOP → FUNDAMENTAL EFFECT

IN ORDER TO CHANGE ω_0, IT'S NECESSARY TO MODIFY V_C
THAT IS A RESULTS OF MODIFICATION ON θ_e

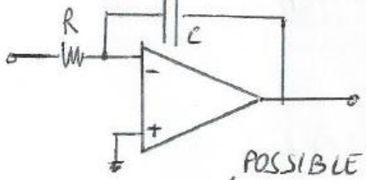
IF $F(0) \to +\infty$ ⇒ YOU NEED TO HANDLE THAT EVEN FOR A NULL PHASE $\theta_e = 0$ → YOU CAN HAVE $V_C \neq 0$

⇓

WITH AN INFITVITE GAIN (VERY HIGH IN REALITY) → IT'S POSSIBLE TO GET $\theta_e = 0$

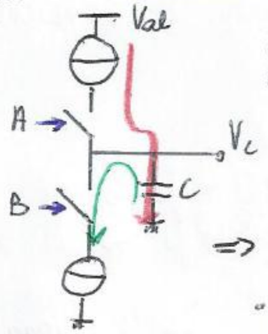
2 WAYS TO GET IT:

WITH AMPLIFIER (WITH HIGH GAIN)



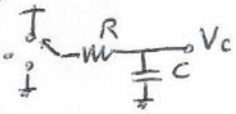
PROBLEMS
- POSSIBLE INSTABILITIES
- HOW TO DEAL WITH { NOISE, OFFSET }

WITH A CHARGE PUMP CIRCUIT



A → CHARGES C
B → DISCHARGES C

MORE REALISTIC SCHEME: ⇒



USED WITH C-MOS TECHNOLOGY

STEADY STATE ERROR → LIMITS IN TERMS OF PHASE ERROR

$$\theta_{er} = \lim_{t \to +\infty} \theta_e(t) \longrightarrow \theta_{er} = \lim_{s \to 0} s\theta_e(s)$$

IN TIME DOMAIN — IN LAPLACE DOMAIN

→ DEPENDS ON
- INPUT SIGNAL θ_i
- DC LOOP GAIN $K_d K_o F(0)$

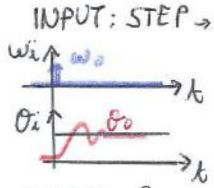
$$\frac{\theta_e}{\theta_i} = \frac{s}{s + K_d K_o F(s)}$$

⇓

$$\theta_{er}(s) = \lim_{s \to 0} \frac{s^2}{s + K_d K_o F(s)}$$

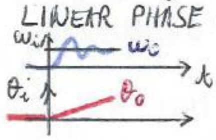
3 POSSIBLE CASES DEPENDING ON THE INPUT

INPUT: STEP → SINGLE f



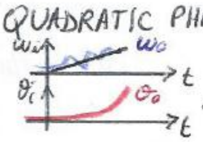
FINITE / INFINITE } LOOP GAIN
↓
SS ERROR = 0

LINEAR PHASE RAMP



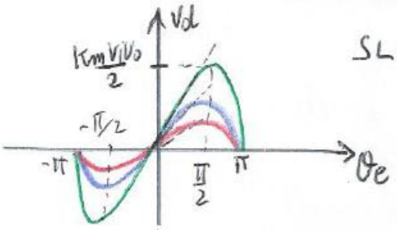
FINITE LOOP GAIN → θ_{er} = constant
INFINITE LOOP GAIN → $\theta_{er} = 0$

QUADRATIC PHASE RAMP



← VARIATION TOO FAST

FINITE LOOP GAIN → θ_{er} UNBOUNDED ($\to +\infty$)
INFINITE LOOP GAIN → θ_{er} = const



SLOPE: $K_m \frac{V_i V_0}{2}$

LET'S LIMIT THE POSSIBLE VARIATION OF V_i → WE'RE JUST INTERESTED IN THE PHASE DIFFERENCE AND PRESERVING THE INFORMATION ON IT, NOT ON DYNAMIC / SHAPE.

↓

A DYNAMIC RANGE COMPRESSION CAN HELP

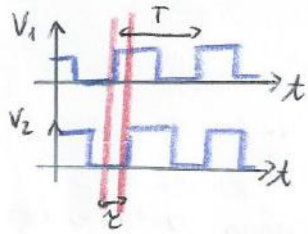
COMPRESSOR:
- V_0: CONSTANT, K_m } FIXED
- V_i: VARIABLE AMPLITUDE

ANY DEVICE THAT IS ABLE TO GENERATE NON-LINEARITY IS A GOOD CANDIDATE TO REALIZE A PHASE DETECTOR

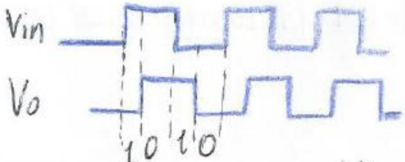
A FIGURE OF MERIT COULD BE: PHASE NOISE.

POSSIBLE $\boxed{\text{DIGITAL IMPLEMENTATIONS}}$

PHASE ERROR/DIFFERENCE → IN DIGITAL DOMAIN
GIVEN A PERIOD $T = 2\pi$
τ = DIFFERENCE BETWEEN EDGES

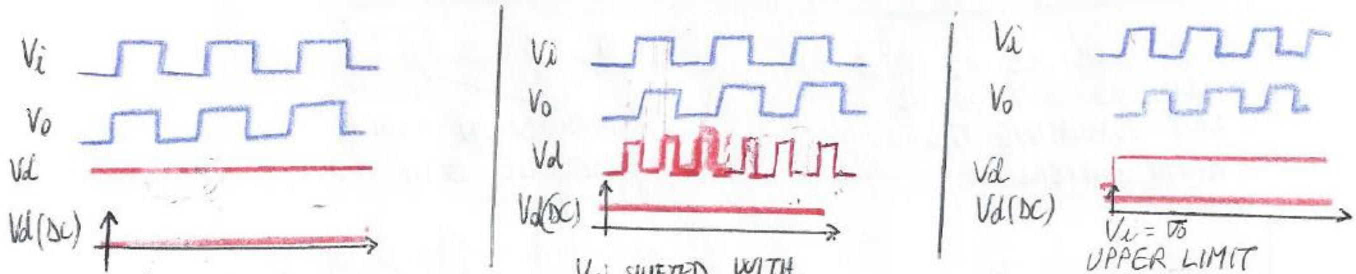


- $\boxed{\text{XOR}}$ → CAN IMPLEMENT THIS DIFFERENCE IF DC = 50%.



A	B	A⊕B
0	0	0
0	1	1
1	0	1
1	1	0

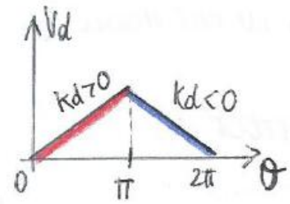
OUTPUT:
- $V_d = 0$ → IF 2 SIGNALS ARE SYNCHRONOUS
- AVERAGE OVER THE PERIOD → IF 2 SIGNALS AREN'T SYNCHRONOUS
  V_d → RELATED TO THE RATIO 1/0 OVER THE PERIOD



$V_i = V_0 \rightarrow \theta_e = 0$
$V_d = \emptyset$

V_i SHIFTED WITH RESPECT TO V_0

$V_i = \overline{V_0}$
UPPER LIMIT
$\theta_e = \pi$



$K_d = \frac{dV_d}{d\theta_e}$ → TWO BRANCHES: $K_d > 0$, $K_d < 0$

LOOP STABILITY ANALYSIS → SELF-SELECTED DURING THE OPERATION, THROUGH NEGATIVE FEEDBACK

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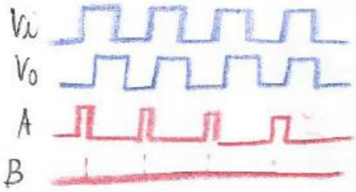
⇒ XOR AND SR-FF HAVE LIMITS ON THE DUTY CYCLE
↓
IF DC IS GENERIC → POSSIBLE SOLUTIONS

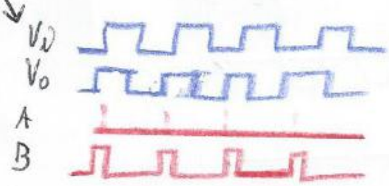
- CONDITIONATE THE SIGNAL ↓ GET SQUARE WAVE (DIVIDING BY 2)
- USE A FF WORKING ON PULSES AND SENSIBLE TO EDGES
- INDEPENDENT ON DUTY CYCLE
  - AD-HOC ASYNCHRONOUS CIRCUITS
  - EDGE SENSITIVE
- PHASE FREQUENCY DETECTORS

PFD = PHASE FREQUENCY DETECTOR → BASED ON CHARGE-PUMP

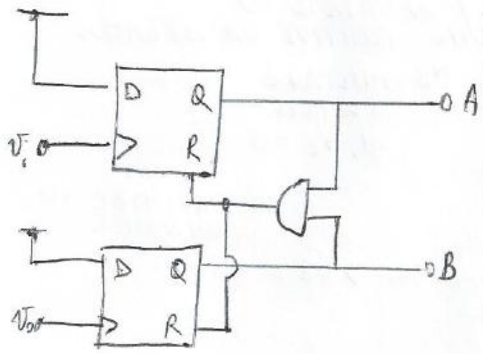


Vi LEADS V0
↓
PULSES ON A → OF SAME DURATION AS THE LEAD

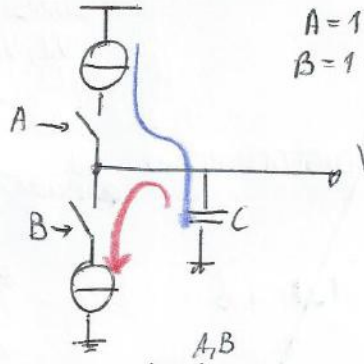
SYNCHRONIZED EDGES
↓
NO PULSE ON A, B



Vi LAGS V0
↓
PULSES ON B → OF SAME DURATION AS LAG

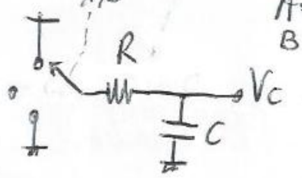


FROM SIGNAL A, B WE CAN RECOVER A SHIFT
↓
PFD IS MERGED TO THE FILTER



A = 1 → C CHARGES
B = 1 → C DISCHARGES

MORE REALISTIC IMPLEMENTATION
↓
LPF EMBEDDED THANKS TO A RESISTOR



RC INTEGRATOR

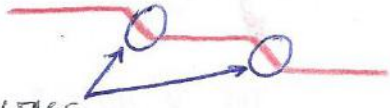
A = 1 → R TO VCC
B = 1 → R TO GND
A = B = 0 → V_C CONSTANT

CHARGE PUMP → INFINITE DC GAIN

A ACTIVE

VOLTAGE INCREASES EACH TIME A = 1

B - ACTIVE



VOLTAGE DECREASES EACH TIME B = 1

ANY LOAD/LEG CLOSES A/B SWITCHES → V_C CHANGES THE VCO FREQUENCY
ONLY STABLE STATE → $\theta_C = 0$ → EQUIVALENT TO A DC INFINITE GAIN — YOU'RE ABLE TO FOLLOW LINEAR VARIATION OF INPUT

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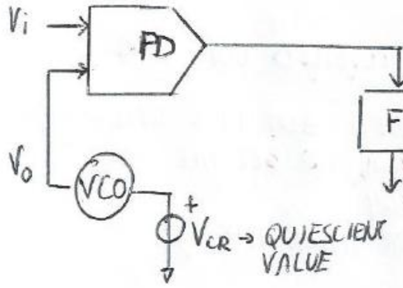
C3 - PHASE - FREQUENCY BEHAVIOR

PHASE DIFFERENCE → IS EVALUATED IN ORDER TO CREATE A REPLICA OF V_{in}

WE'LL ANALYZE WHEN AND HOW THE PLL CAN TRACK THE INPUT f AND KEEP THE INPUT SIGNAL

IN THE BUTTERFLY DIAGRAM → THERE ARE 2 POSSIBLE RANGES OF FREQUENCY
- CAPTURE REGION → PLL GETS THE LOCK (LOCKS f_{IN})
- LOCK IN REGION → PLL FOLLOWS THE INPUT FREQUENCY VARIATIONS WITHOUT LOOSING THE LOCKING

OPEN-LOOP BEHAVIOR

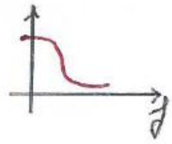


PD → COMPARISON BETWEEN
- V_i
- V_o → SIGNAL WHICH IS COMPLETELY SCORRELATED, COMING FROM THE V_{CO}, GIVEN BY A CONSTANT VOLTAGE

VCO → IS DRIVEN BY A FREE-RIDE FREQUENCY
LET'S SUPPOSE $V_{CR}: \omega_o = \omega_{or}$

AFTER PD → $\theta_e = \theta_i - \theta_o \rightarrow \omega_d = \omega_i - \omega_{or}$
WE OBTAIN V_D AND WE FILTER OUT THE UNWANTED COMPONENTS FROM IT

DIFFERENCE V_D → IS A SIGNAL PLACED @LOW FREQUENCY
IF IT'S IN THE BAND OF THE LPF WE OBTAIN V_c → STRONGLY DEPENDS ON THE FILTER

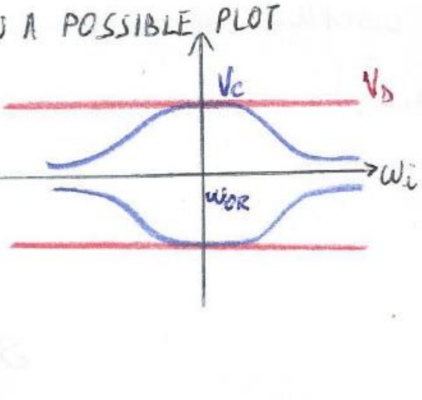


IF $\omega_i \neq \omega_{or} \rightarrow \omega_d \rightarrow$ HAS 2 BEATS $\left\langle \begin{matrix} 1 @ HF \\ 1 @ LF \end{matrix} \right\}$ BUT SURELY NOT IN DC

THE FILTER REMOVES ALMOST EVERYTHING FROM THE V_c INCREASING THE AMPLITUDE OF V_c
- V_c HIGH WITH LOW DIFFERENCE BETWEEN FREQUENCIES ($\omega_i \sim \omega_{or}$)
- V_c LOW WITH HIGH DIFFERENCE BETWEEN FREQUENCIES

IT'S DUE TO $|F(s)|$

IN A POSSIBLE PLOT



V_D → BEFORE THE FILTER
V_c → AFTER THE FILTER

ω_{OR} IS FIXED
WE SWEEP ω_i INCREASING ITS FREQUENCY

V_c
- STRONGLY ATTENUATED → IF $\omega_i \gg \omega_{or}$
- EQUAL TO V_D → IF $\omega_i \approx \omega_{or}$ (OR CLOSE TO ω_{or})

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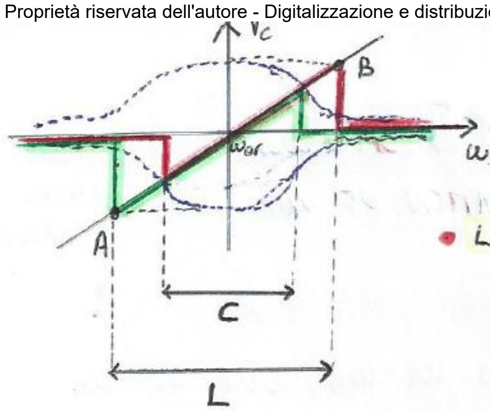


— INCREASING ω_i
— DECREASING ω_i

THERE ARE 2 POSSIBLE REGIONS:

- LOCK REGION (L) → THE PLL HAS LOCKED f_{IN} IF ω_i CHANGES WITHIN THIS RANGE THE LOCK IS MANTAINED
  ↓
  BEING SET BETWEEN $\{^A_B$, IT CLEARLY DEPENDS ON V_C WHICH DEPENDS ON DC LOOP GAIN: $K_d K_0 F(0)$

CONCERNING THE PHASE DETECTOR

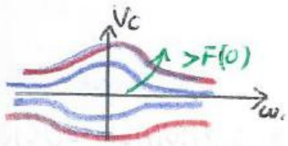
ANALOG → INFLUENCED BY V_o / V_i AMPLITUDE

DIGITAL → V_i, V_0 HAVE A FIXED AMPLITUDE (ϕ - Vdd) → K_d IS FIXED
↓
IT'S POSSIBLE TO REALIZE A $F(0) \to +\infty$ (INFINITE GAIN) s.t. THE LOCK RANGE IS LIMITED BY VCO NON LINEARITY

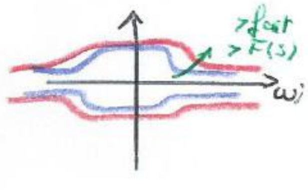
- CAPTURE REGION (C) → HOW THE VCO IS ABLE TO FOLLOW THE VARIATIONS DEPENDS ON THE FILTER CHARACTERISTIC
  ↓
  THE BAND OF THE LPF CHANGES THE WIDTH OF THE CAPTURE REGION
  WE NEED TO TAKE CARE OF ALL THE SHAPES OF THE FILTER $F(s)$
  WITH THE CHARGE-PUMP → THE CAPTURE RANGE CORRESPONDS TO THE LOCK RANGE

EFFECT OF CHANGE OF PARAMETERS:

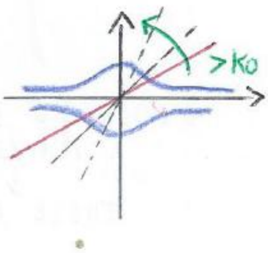
- VERTICAL CALIBRATION → DEPENDS ON DC GAIN OF THE FILTER
  $> F(0) \Rightarrow >$ PEAK VALUE OF V_C



- HORIZONTAL CALIBRATION → THE CAPTURE RANGE IS MODIFIED CHANGING THE CUT-OFF FREQUENCY (SO WITH A PROPER CHOICE OF THE SHAPE OF YOUR FILTER $F(s)$)



- SLOPE → OF THE VCO IS MODIFIED CHANGING K_0
  (SO $\{^{LOCK}_{CAPTURE}\}$ RANGES ARE MODIFIED)



C4- PLL AND NOISE

THE PLL CAN BE SEEN HAS A BAND PASS FILTER
THAT'S BECAUSE ACTUALLY YOU'RE GETTING AN INPUT SIGNAL WHICH HAS < NOISE / UNWANTED MODULATIONS
↓
AND THROUGH THE PLL YOU REMOVE THE UNWANTED COMPONENTS

WHY DON'T WE USE DIRECTLY A BPF? BECAUSE IN THE PLL YOU HAVE TO DESIGN AN LPF WHICH IS SIMPLER TO DESIGN THAN A BPF

WHAT IS THE BEHAVIOR OF A SIGNAL DEFINED IN A GIVEN BAND?
GIVEN AN INPUT SIGNAL WITH ADDITIVE NOISE
↓
LET'S TRY TO EVALUATE THE EQUIVALENT INPUT PHASE NOISE:

PLL AS A FILTER

PLL HAS A
↗ NOISY INPUT
↘ CLEAN OUTPUT, IT HAS CENTER FREQUENCY WHICH FOLLOWS THE SIGNAL
↓
SO IT ACTS AS A TUNABLE FILTER

THE BANDWIDTH DEPENDS ON $F(s)$

HOW TO EVALUATE ITS PERFORMANCES? LET'S CHARACTERIZE ITS BANDWIDTH
- APPLY WIDE-BAND NOISE AT INPUT (SELECT f_1, f_2 COMPUTE THEM)
- MEASURE NOISE POWER AT < IN / OUT
- EVALUATE THE POWER RATIO

EVALUATE THE SPECTRUM OF A GIVEN BANDWIDTH

↓
LET'S SUPPOSE < ANALOG PD / SINUSOIDAL VCO
↘ FOLLOW THESE STEPS:

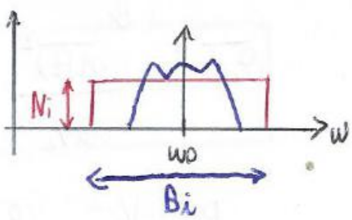
1) INPUT SIGNAL IS $\emptyset$: WE'RE JUST EVALUATING NOISE AND COMPUTING ITS POWER
2) INPUT SIGNAL = PURE SINE + SOME UNWANTED MODULATION (PHASE NOISE: eg jitter)
   LET'S COMPUTE ITS POWER
3) COMPUTE THE SPECTRAL DENSITY OF EQUIVALENT PHASE NOISE
4) EVALUATE THE BANDWIDTH
5) EVALUATE THE SNR

THE SYSTEM HAS THE FOLLOWING PARAMETERS
↗ INPUT: SINE V_i + NOISE $\quad v_i' = V_i \sin(\omega_i t + \theta_i) + n(t)$
↘ LOCK CONDITION: $F_{or} = F_i$

NOISE: $n(t) = n_c(t)\cos(\omega_0 t) + n_s(t)\sin(\omega_0 t)$

N_i : SPECTRAL POWER DENSITY

B_i : BANDWIDTH



CONSIDERING $N_i = \frac{\overline{n(t)^2}}{B_i}$

THE **PHASE NOISE** IS DEFINED AS $\phi = \frac{\overline{\theta_{in}(t)^2}}{\frac{B_i}{2}}$ REPLACING $\overline{\theta_{in}^2}$

$$\phi = \underbrace{\frac{\overline{n(t)^2}}{V_i^2} \cdot \frac{2}{B_i}}_{N_i} \Rightarrow \boxed{\phi = \frac{2N_i}{V_i^2}}$$

PHASE NOISE SPECTRAL POWER DENSITY

NOISE POWER AT THE OUTPUT

THE OUTPUT NOISE CAN BE EVALUATED FROM THE TRANSFER FUNCTION $H(s)$

SOMETHING PRESENT AT THE INPUT + EFFECT OF ADDITIONAL NOISE DUE TO THE FILTER

$$\theta_i \rightarrow \boxed{H(s)} \rightarrow \theta_o$$

$$\theta_o(s) = H(s)\,\theta_i(s)$$

$$\overline{\theta_{on}(t)^2} = \int_0^{\frac{B_i}{2}} \phi\,|H(j\omega)|^2 df = \phi \underbrace{\int_0^{\frac{B_i}{2}} |H(j\omega)|^2 df}_{B_L}$$

OUTPUT NOISE POWER

INPUT SPECTRAL DENSITY

B_L: EQUIVALENT BANDWIDTH

$$B_L = \frac{\overline{\theta_{on}(t)^2}}{\phi} \rightarrow \boxed{B_L = \int_0^{B_i/2} |H(j\omega)|^2 df}$$

$B_L \rightarrow$ DEPENDS ON B_i

$\Downarrow$

LET'S SUPPOSE $|H(j\omega)|$ NEGLIGIBLE AROUND $B_i/2$

$$B_L = \int_0^\infty |H(j\omega)|^2 df$$

$\rightarrow$ IT JUST DEPENDS ON $|H(j\omega)|$ FUNCTION OF $F(s)$

$\Downarrow$

! B_L DEPENDS ON THE FILTER

LOW $H(s)$ f_{cutoff}	HIGH $H(s)$ f_{cutoff}
NARROW B_L (HIGH FILTERING)	WIDE B_L
LOW NOISE	MORE NOISE
SLOW RESPONSE	FAST RESPONSE

SIGNAL-TO-NOISE RATIO

$$SNR = \frac{P_{signal}}{P_{noise}} = \frac{V_i^2/2}{V_i^2/\theta_i^2}$$

$$SNR_i = \frac{1}{2\overline{\theta_{in}^2}} \qquad SNR_o = \frac{1}{2\overline{\theta_{on}^2}}$$

$$\Rightarrow \boxed{SNR_o = SNR_i \frac{B_i}{2B_L}}$$

WITH NARROW B_L: $SNR_o \uparrow$; WIDE B_i: $SNR_o \uparrow$

$\Rightarrow$ AS A CONCLUSION: PLL ACTS AS A PASSBAND FILTER FROM V_i TO V_o WITH A BANDWIDTH B_L

IT FILTERS THE PHASE NOISE THANKS TO THE LOOP CONFIGURATION WITHOUT ANY EFFECT ON NOISE AMPLITUDE

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- AM DEMODULATION → IDEA → DIVIDED IN 2 STEPS → DURING:

(1) MODULATION → CARRIER IS MULTIPLIED BY $M(t)$ (MODULATING SIGNAL)

$$S = F_p \times M(t)$$

IN TIME DOMAIN: M IS THE ENVELOP OF $S(t)$

IN FREQUENCY DOMAIN: M IS THE SPECTRUM $M(0)$ TRANSLATED TO $M(f_p)$

(2) DEMODULATION → THERE ARE ENVELOPE-DETECTORS $\left\langle \begin{matrix} \text{HALF-WAVE} \\ \text{FULL-WAVE} \end{matrix} \right\}$ RECTIFIERS (DIODES) + LPF (C+R)

OR S IS MOVED BACK TO BASEBAND THROUGH A PLL] ⇒ COHERENT DEMODULATION

(1) ANALOG AMPLITUDE MODULATION

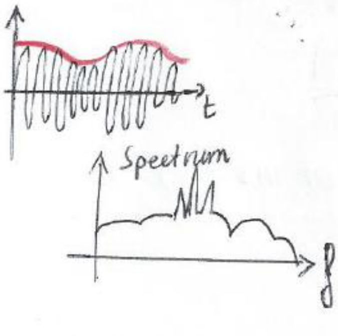
eg: GIVEN A MODULATING SIGNAL : $A\sin(\omega t) + B$ (VARYING VS TIME)

LET'S DEFINE THE MODULATION INDEX:

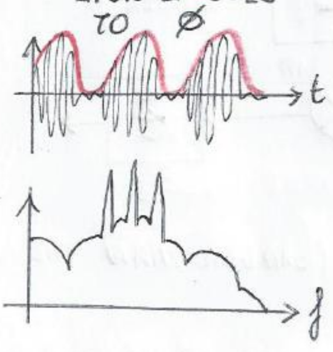
$$\boxed{m = \frac{Amp_{max} - Amp_{min}}{Amp_{max} + Amp_{min}}} \rightarrow \text{IN THIS CASE: } \begin{matrix} max: A+B \\ min: B-A \end{matrix} \quad m = \frac{A+\cancel{B}-\cancel{B}+A}{A+B+B-\cancel{A}} = \frac{A}{B}$$

INCREASING THE MODULATION INDEX m YOU CAN HAVE A SIGNAL WHICH IS VARYING WITH ENVELOP WHICH IS ALWAYS POSITIVE WITH RESPECT TO THE CARRIER

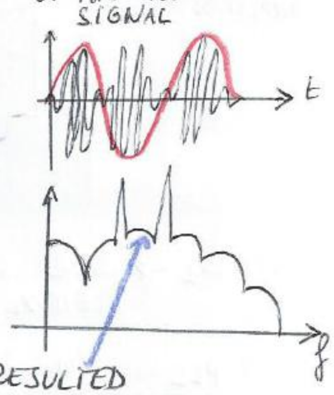
$m = 0{,}25 < 1$ (YOU'RE NOT CROSSING THE Ø V WITH THE ENVELOP DETECTOR)



$m = 1$ THE MODULATED SIGNAL GOES TO Ø



$m > 1$ THERE'S A SORT OF PHASE INVERSION OF THE MODULATED SIGNAL



THIS RESULTED INTO A SUPPRESSION OF THE CARRIER f

↓

ALWAYS EMPLOYED BECAUSE THIS SOLUTION ALLOWS A LOWER POWER CONSUMPTION

THE MAIN PROBLEM → f COMING OUT FROM THE VCO
WHICH IS IN THEORY CONSTANT
BUT IN PRACTICE AFFECTED BY PHASE NOISE

THERE'S NOT ANY CONTROL ON V_{in}
f_i HAS SOME FREQUENCY MODULATION
EVEN IF VCO IS PERFECT → ω_o IS COMPARED TO ω_i
ω_{co} IS MOVED
PHASE ERROR $\neq 0$

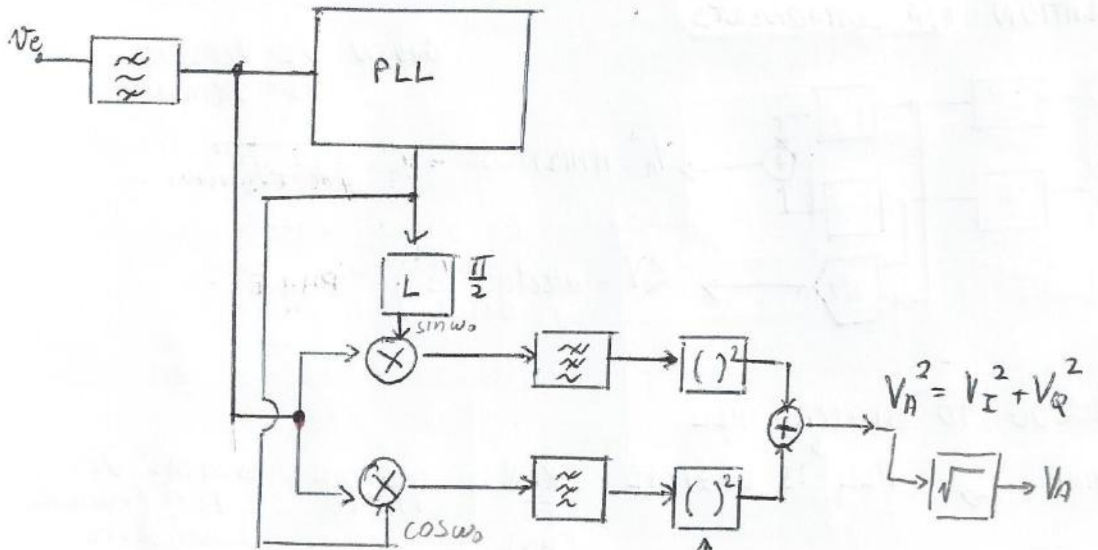
AN UNWANTED PHASE MODULATION MEANS TO HAVE AN UNWANTED AMPLITUDE MODULATION (CROSS-DEMODULATION)

IN ORDER TO HAVE A LOWER PHASE ERROR

INCREASE THE LOOP GAIN
ENLARGE THE FILTER $F(s)$
s.t. YOU REDUCE θ_e
BUT YOU'RE DECREASING SNR

→ I/Q DEMODULATOR

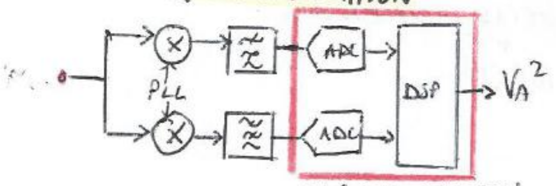
ANALOG I-Q DEMODULATOR → ALLOWS TO SEPARATE < PHASE / QUADRATURE COMPONENTS



BECAUSE WE'RE INTERESTED IN AMPLITUDE → $|V| = \sqrt{V_I^2 + V_Q^2}$

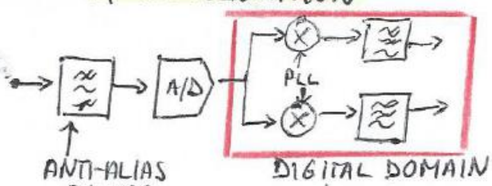
WE CAN ALSO REALIZE DIGITAL I-Q DEMODULATORS

IF DIGITAL I-Q DEMODULATION



DIGITAL DOMAIN
IT'S EASIER TO REALIZE THIS OPERATION
WE'RE NO MORE DEALING WITH THE CARRIER FREQUENCY → NYQUIST THEOREM IS RELATED TO THE BAND OF THE SIGNAL, NOT TO f_o

RF DIGITAL IQ DEMODULATION



ANTI-ALIAS FILTER

DIGITAL DOMAIN
SOMETHING THAT CANNOT BE DONE AT VERY HIGH FREQUENCY (BECAUSE OF THE COST OF THE ADC)