NUMERO: 1425A - ANNO: 2015

A P P U N T I

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MATERIA: Radio Frequency Integrated Circuits. Prof.Pirola

RADIO FREQUENCY INTEGRATED CIRCUITS

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(2014 - 2015)
(S202483)

AMPLIFIERS

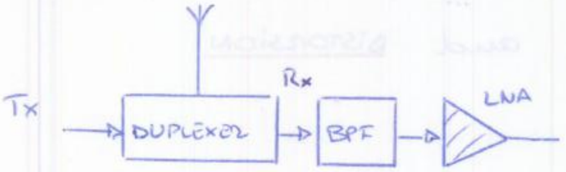
TIME DOMAIN ANALYSIS: generates a transient solution but I am not interested on it because AT MICROWAVE LEVEL WHAT I'M INTERESTED IN IS THE REGIME CONDITION!

RF TRANCEIVER SYSTEMS REVIEW

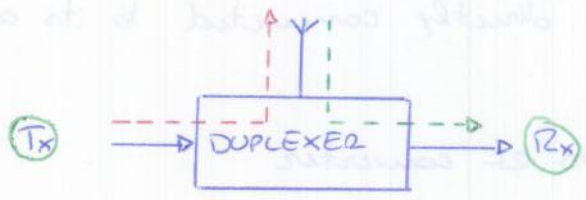
Very often, communication channels have to be shared among several users and, for this reason, FDD and TDD are used

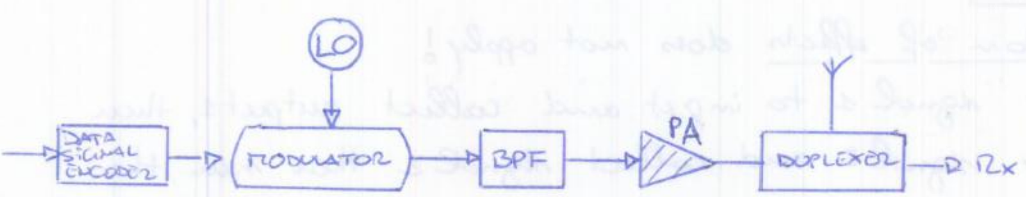
SYSTEM ARCHITECTURE: → HANDSET: only one channel is required
→ BASESTATION: can accept MULTIUSER AND MULTI-STANDARD conditions

SOFTWARE DEFINED RADIO:
In this system, terminals may be dynamically reconfigured. The basestation sends software for the new configuration needed by the terminal.



DUPLEXER:
- it is the only common element both for Rx and Tx, why?
  because there is only one antenna available
- basically it is a very selective filter
- it has to isolate high from low power circuits!

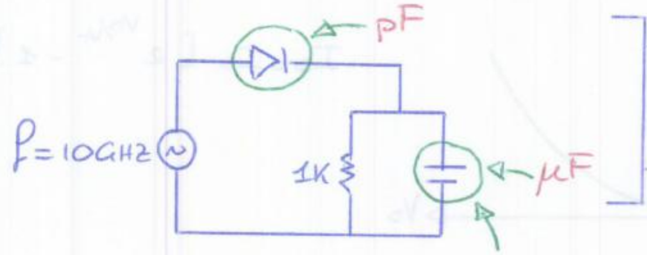




transmission chain

DRAWBACK:

Spice integrates step by step and at each one it performs some kind of approximations → introduces errors that ACCUMULATE!



→ VERY FAST DIODE TIME CONSTANT

→ VERY SLOW CAPACITOR TIME CONSTANT

$\tau = RC \simeq (1K)(1\mu) \simeq 1\,msec$

R and C values are established upon modulated signal frequency (~100KHz)

I GET TO REGIME ($t \to \infty$) AFTER 4÷5 TIMES THE MAXIMUM TIME CONSTANT

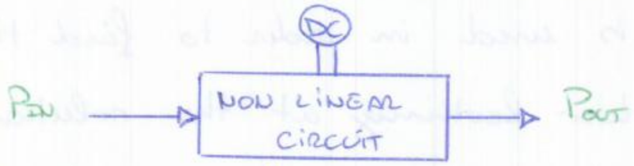
INTEGRATION STEP IS DEFINED BY THE MINIMUM TIME CONSTANT

THIS PROCESS WORKS but it is VERY MUCH INEFFICIENT

I spend too much time to get to regime if I have a too large variety of time constants (the largest and the smallest are too far apart)

IS IT POSSIBLE TO CREATE AN AMPLIFIER WITHOUT USING NON-LINEAR CIRCUITS?

No because an amplifier takes energy from the power supply and its output is, by definition, at larger energy. Ideally, the input signal is "zoomed" K times without adding any disturbance.



Where does the power comes from? Yes, from the DC through the bias circuit but, A LINEAR CIRCUIT CANNOT PRODUCE A SPECTRAL COMPONENT WHICH IS NOT PRESENT AT ITS INPUT!

The non linear circuit is the one that generates the desired output frequency component.

For the linear circuit I can surely use $\underline{Y}$ and $\underline{Z}$ matrices

$$\underline{I_L} = \underline{Y}\ \underline{V_L}$$

$$\boxed{I_L \Big|_{1\ldots M} - I_{NL}\Big|_{1\ldots M} = 0}$$

Kirckhoff law holds for each wire so $I_L + I_{NL} = 0$ for $\forall$ of them.
Also here, I_L and I_{NL} can be written as summations by mean of fourier expansion.

THERE MUST BE BALANCE OF EACH HARMONIC BETWEEN THE 1^{ST} OF THE LINEAR AND 1^{ST} OF THE NON LINEAR EXPRESSION, OF THE 2^{ND} WITH THE 2^{ND}, AND SO ON, IN ORDER TO ENSURE KIRKHOFF'S LAW

$\Rightarrow$ I CAN WRITE DOWN A SYSTEM OF EQUATIONS HAVING
(n° wires) $\times$ (n° harmonics) EQUATIONS
P
$(2N+1) \times$ (n° of ports) UNKNOWNS

$2N+1$ comes out of the $2\times\left(\sum_{n=1}^{N}\right)$ and the DC term

$P(2N+1)$ equations
$(2N+1)\,P$ unknowns $\Big\}$ $\Rightarrow$ I can solve the system.

Carefull: the equations are non linear $\to$ so I have a non linear system $\to$ I USE NEWTON RAPHSON METHOD.

———————— o ————————

Each physical wire can be seen as a $(2N+1)$ wire, where each one is a single HARMONIC that links linear to non linear circuit. then I can write I_L as:

$$\underline{I_L} = \underline{Y_L}\ \underline{V_L} \quad \text{where} \quad \underline{I_L} = \begin{bmatrix} I_{1a1} \\ I_{1a2} \\ \vdots \\ I_{1aN} \\ I_{1b1} \\ \vdots \end{bmatrix}$$ $2N+1$

$\uparrow$ NUMBER OF HARMONICS $\to$ much complex problem to solve $\to$ algorithm less efficient.

$$\boxed{\Delta T = \frac{T}{(2N+2)-1} = \frac{T}{2N+1}}$$

OPTIMAL INTERVAL CHOICE $\Rightarrow$ same concept of sampling theorem.

$S(0) = a_0 + a_1 + \dots + a_N + 0$

$S(\Delta T) = a_0 + a_1 \cos(\omega_0 \Delta T) + a_2 \cos(2\omega_0 \Delta T) + \dots + a_N \cos(N\omega_0 \Delta T)$
$+ b_1 \sin(\omega_0 \Delta T) + \dots + b_N \sin(N\omega_0 \Delta T)$

$S(2\Delta T) = \dots$

$\vdots$

$S((2N+1)\Delta T) = \dots$

I can write it in matrix form as

$$\underline{S}_{\Delta T} = \underline{\underline{F2T}} \begin{bmatrix} a_0 \\ a_1 \\ b_1 \\ b_2 \\ \vdots \end{bmatrix}$$

$\underline{\underline{F2T}}$: frequency to time matrix $(2N+1)\times(2N+1)$

$$\underline{\underline{F2T}} = \begin{bmatrix} 1 & 1 & 0 & 1 & 0 \\ 1 & \cos(\omega_0 \Delta T) & \sin(\omega_0 \Delta T) & \cos(2\omega_0 \Delta T) & \dots \\ 1 & \cos(2\omega_0 \Delta T) & \sin(2\omega_0 \Delta T) & \cos(4\omega_0 \Delta T) & \dots \\ \vdots & & & & \end{bmatrix}$$

I can also have that

$$\begin{bmatrix} a_0 \\ a_1 \\ b_1 \\ a_2 \\ b_2 \\ \vdots \end{bmatrix} = \underbrace{\underline{\underline{F2T}}^{-1}}_{\underline{\underline{T2F}}} \underline{S}_{\Delta T}$$

THE MATRIX DEPENDS ONLY ON THE TRUNCATION N AND ON THE FREQUENCY ω_0 NOT ON THE CIRCUIT !

CHOOSING AS TIME STEP ΔT I HAVE THAT $F2T \perp T2F$

$$i_D(t) = I_S \left[e^{\frac{v_1(t) - v_2(t)}{V_T}} - 1 \right]$$

How do I isolate the unknown? I write:

$$\underline{i_D}(\Delta T) = I_S \left[e^{\frac{\underline{v_1}(\Delta T) - \underline{v_2}(\Delta T)}{V_T}} - 1 \right]$$

$$\underline{i_D}(\Delta T) = I_S \left[e^{\frac{\underline{\underline{F2T}}\, \underline{V_1} - \underline{\underline{F2T}}\, \underline{V_2}}{V_T}} - 1 \right]$$

Finally

$$\underline{I_D} = \underline{\underline{T2F}}\, \underline{i_D}$$

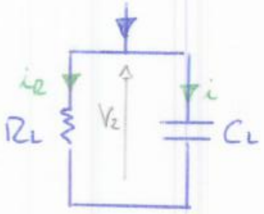
and : $\underline{I_{RS}} + \underline{I_D} = 0$

$$\frac{\underline{V_1} - \underline{V_S}}{\underline{\underline{R_S}}} + I_S\, \underline{\underline{T2F}} \left[e^{\frac{\underline{\underline{F2T}}(\underline{V_1} - \underline{V_2})}{V_T}} - 1 \right] = \varnothing$$

BALANCE EQUATION FOR NODE 1

$4N+2$ unknowns because I have $(2N+1)$ for each port

BALANCE EQUATION AT NODE 2



$i_R = V_2 / R_L$

$i = C_L\, dv/dt$

PROBLEM: dv/dt derivative operator should be applied to all the harmonics of v

generic cosine expression: $a_n \cos(n\omega_0 t)$
(of the voltage)

$$a_n \cos(n\omega_0 t) \rightarrow \boxed{d/dt} \rightarrow -C_L\, a_n\, n\omega_0 \sin(n\omega_0 t)$$

THE FIRST HARMONIC CONTRIBUTION OF THE VOLTAGE INFLUENCES THE FIRST SINE CONTRIBUTION OF THE CURRENT

another example: $b_n \sin(n\omega_0 t)$

$$b_n \sin(n\omega_0 t) \rightarrow \boxed{d/dt} \rightarrow b_n\, n\omega_0 \cos(n\omega_0 t)$$

First component of current signal (i_{DC}) must be zero since no DC passes through a capacitor.

NEWTON'S METHOD

I'm looking at the zero of a function;

I start by guessing a point x_1;

I approximate the function to its derivative in that point;

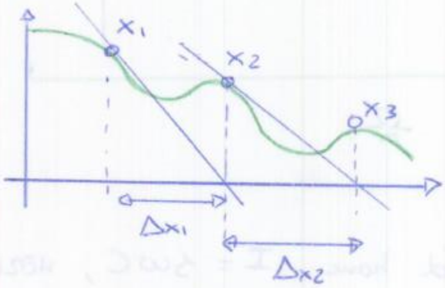
I compute Δx;

I take the following point as $x_1 + \Delta x$;

I iterate until $f(x_n) \hat{=} 0$

$$f(x) = f(x_1) + \left.\frac{df}{dx}\right|_{x_1} \Delta x \qquad \text{if } f(x) = 0 \Rightarrow -\Delta x = \frac{f(x_1)}{\left.\frac{df}{dx}\right|_{x_1}}$$

$$x_2 = x_1 + \Delta x$$



In my problem the function is a VECTOR FUNCTION where each entry is a $(2N+1)$ unknown equation; so I have a vector function of vectors of multiple independent variables.

I NEED TO EXTEND THE NEWTON'S METHOD TO MY PROBLEM

$\underline{f}(\underline{V_1}, \underline{V_2})$

I start guessing $\underline{V_1}$ and $\underline{V_2}$

I don't get a zero $\Rightarrow$ I get a RESIDUAL $\Rightarrow$ this will be the second guess of the iteration

I select Δx

In matrix form I have:

$$\Delta \begin{bmatrix} \underline{V_1} \\ \underline{V_2} \end{bmatrix} = \underline{\underline{J}} \begin{bmatrix} \underline{V_1} \\ \underline{V_2} \end{bmatrix}^{-1} \underline{R} \begin{bmatrix} \underline{V_1} \\ \underline{V_2} \end{bmatrix}$$

$\rightarrow$ residual

$\rightarrow$ Jacobian (represents the derivative)

Jacobian: matrix of all first order partial derivatives of a vector valued function.

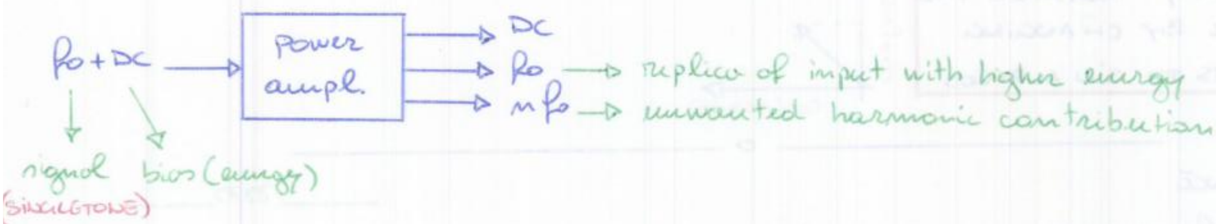
The system is solved if the norm of $\underline{R}$ is ≈ 0

14/10/14 **RF AMPLIFIERS I**

In the transmitter section I have power amplification but this is divided in different stages. The first are the PRE-AMPLIFIERS while the last is the POWER AMPLIFIER.
Generally, at the receiver I need more gain (since I deal with small signals) whereas on the transmitter I need high gain (8, 10, 12) dB and I must consider DISTORSION and NON-LINEARITIES

	Rx	Tx
MOST IMPORTANT STAGE	FIRST	LAST
	due to noise	due to power

POWER AMPLIFIER



GAINS

- TRANSDUCER GAIN $G_{TR} = \dfrac{P_{OUT}}{P_{DISP\_IN}}$
  P_{DISP_IN} → power that could enter the amp. if no mismatch would be present
- OPERATIONAL GAIN $G_{OP} = \dfrac{P_{OUT}}{P_{IN}}$
  P_{IN} → effective input power
  → $P_{IN} \cong P_{DISP\_IN}$ if $|\Gamma| \cong (-15 \div -20)$ dB
- CONVERSION GAIN $P_{OUT}(mf_0) = K_m \left[P_{IN}^m(f_0) \right]^{m}$
  $P_{OUT}(mf_0)$ → output power at the m^{th} harmonic
  K_m → conversion gain ← IT IS A NON LINEAR GAIN SINCE LINKS f_0 TO mf_0

FIGURES OF MERIT

- 3^{rd} ORDER HARMONIC FACTOR INTERCEPT POINT
  - ↳ higher this point, more linear the amplifier
  - ↳ it is not something I can measure
    - ↳ I need to get expressions for the two dotted red lines and equate them

output power @ m^{th} harmonic equals fundamental output power.

$$\boxed{\underbrace{\underbrace{[G_{OP,SS}]}_{\text{slope of the line.}} \cdot P_{IN\_m}}_{\text{first line}} = \underbrace{K_m [P_{IN\_m}]^m}_{m^{th}\text{ line}}}$$

↳ basically the small signal gain (when input power is small and relation $P_{OUT} \div P_{IN}$ is linear)

$$\boxed{P_{IN\_m} = \left[\frac{G_{OP,SS}}{K_m}\right]^{\frac{1}{m-1}}}$$

- CORRESPONDING INPUT INTERCEPT POWER

- EFFICIENCY

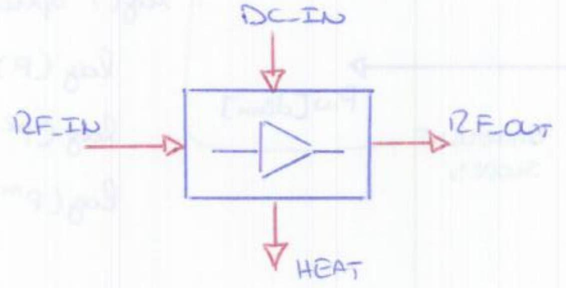
$$\boxed{\eta = \frac{P_{OUT}(f_0)}{P_{DC}}} \qquad \eta \leq 1$$

↳ power delivered by the bias $P_{DC} = V_{DC} I_{DC}$

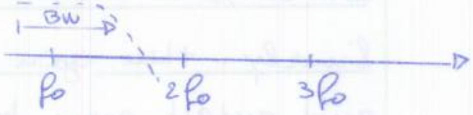
- POWER ADDED EFFICIENCY

$$\boxed{PAE = \frac{\overbrace{P_{OUT}(f_0) - P_{IN}(f_0)}^{\text{NET RF POWER}}}{P_{DC}} = \eta\left[1 - \frac{1}{G_{OP}}\right]}$$

if G_{OP} is large → $PAE \cong \eta$



However, II° harmonic is usually outside my bandwidth of interest, WHAT I'M CONCERNED ABOUT ARE THE MIXED TERMS



$$\begin{cases} 2f_1 - f_2 \\ 2f_2 - f_1 \end{cases}$$

IMP

ODD INTERMODULATION PRODUCTS ARE THE PROBLEMS BECAUSE PRODUCE CONTRIBUTIONS INSIDE THE BANDWIDTH

if I have a non linear system I surely have intermodulation problems.

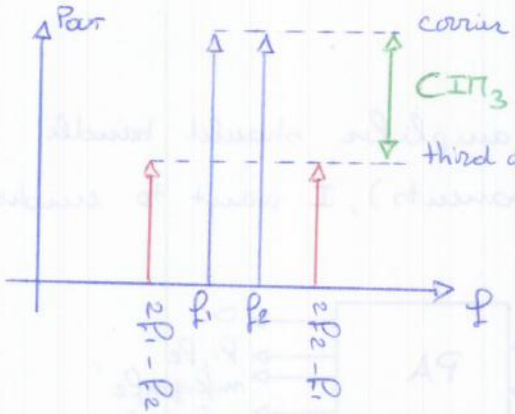
BIGGER m SMALLER is a_m (SINCE THE SERIE MUST CONVERGE) SO THE WORST INTERMODULATION IS FOR THE SMALLEST ODD m → 3

Intermodulation produces spectral regrowth (cannot be filtered out) this deteriorates the SNR

CIM3 carrier to IM3 ratio

$$CIM3 = \frac{P_{OUT}(f_0)}{P_{INT3}}$$

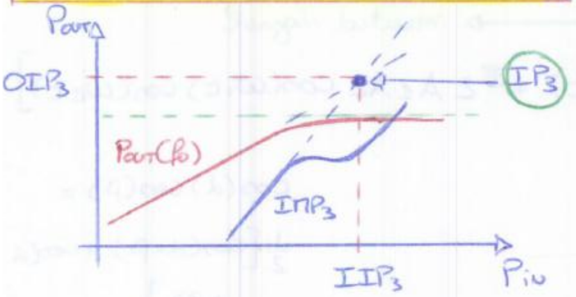
third harmonic contribution grows according to $()^3$ of input



if ↑ P_{IN} by 1dB ⇒ ↑ IMP_3 by 3dB ⇒ ↓ CIM_3 by 3dB / ↑ P_{IN} by 1dB

since I have $\log[P^3]$

3rd ORDER IMP INTERCEPT - IP_3 (another method to estimate non-linearity)

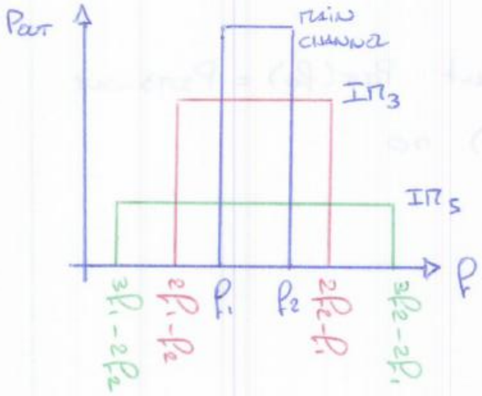


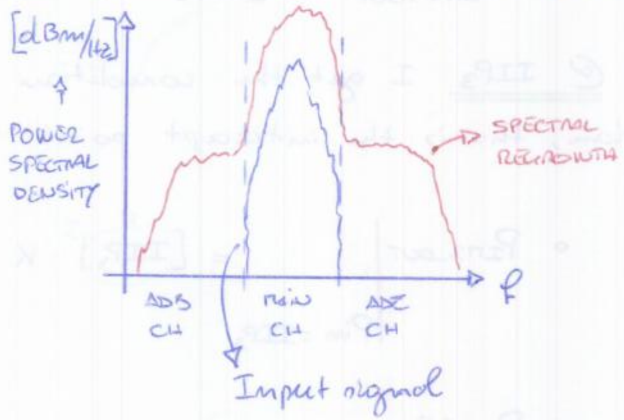
→ point where III° order intermodulation power intersects with fundamental output power

Again, this point cannot be evaluated but by using the small signal condition I can write the equations of the two dotted lines and find where they meet

$$IP_3 = (IIP_3 \, ; \, OIP_3)$$

Base stations handle more than one channel and, for each one, a portion of spectrum is reserved. I must limit the spectral regrowth in order to avoid a channel to interfere the others. I need to rate a way to regulate the spectral regrowth.





In adjacent channels I'd have a degradation of SNR due to the spectral regrowth of the main channel in them

ADJACENT CHANNEL POWER RATIO

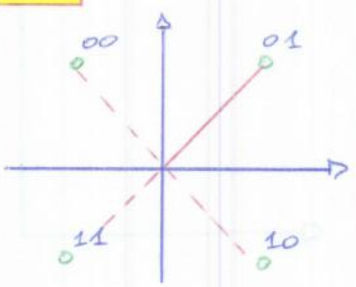
quantitative measure of spectral regrowth as an interference to nearby channels.

$$\boxed{ACPR = \frac{\int_{MC} P_{OUT}(f)\,df}{\int_{C_K} P_{OUT}(f)\,df}}$$

$\uparrow ACPR \Rightarrow \uparrow$ NON-LINEARITY

- of course the most dangerous are the most adjacent channels to the main channel
- depends also on the format of the my signal. (if constant envelop it is more sensitive)

GSM



I can transmit 2 bits at a time by mean of a CONSTANT ENVELOP signal. this means that the amplifier does not have to change power from one channel to the other!

I can use my class A amplifier always at its maximum efficiency (designed to be in correspondence of the constant input power)

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PA DYNAMIC RANGE ≡ SFDR

For dynamic range I define the input power range that let the amplifier work within the specs.
The MINIMUM signal is influenced and defined by the NOISE FLOOR of the amplifier. If in input I have a signal of which intensity is below noise floor, at output I've an undefined signal.
The MAXIMUM signal is defined by the DISTORSION that my amplifier introduces. (In other words, it is the maximum input power such that the output is non-distorted or with acceptable distorsion).

NOISE FIGURE

$$\boxed{NF = \frac{SNR_{IN}}{SNR_{OUT}}}$$

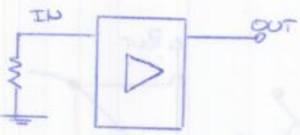
Noise figure can be thought also as

$$NF = \frac{N_L}{N_I \cdot G} =$$

N_L: total noise @ output

$N_I \cdot G$: noise @ the output due to the noise @ the input

$$= \frac{N_L}{N_I \cdot G} \cdot \frac{S_I}{S_I} = \frac{SNR_{IN}}{SNR_{OUT}}$$



SENSITIVITY

Minimum input signal yeelding a given SNR on the load

$$S = NF \cdot SNR_L \cdot \underbrace{KTB}_{\text{thermal noise}}$$

for thermal noise @ 300°K I have

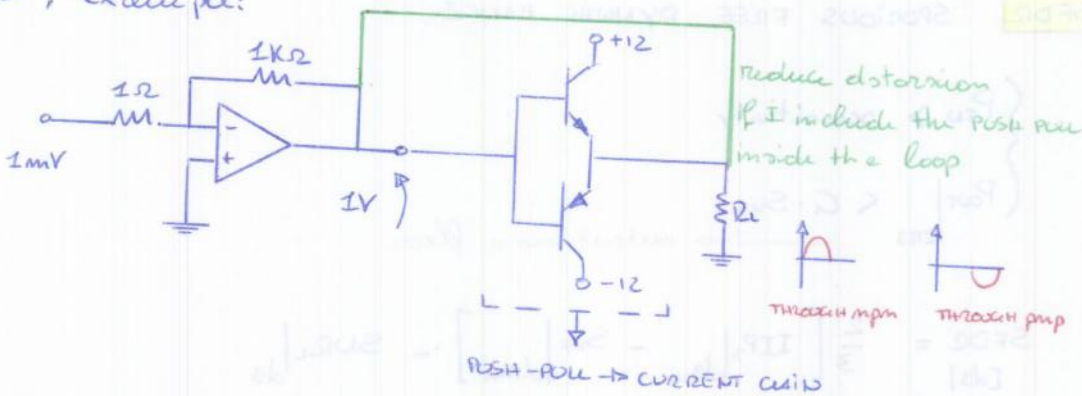
$$S\Big|_{dBm} = NF\Big|_{dB} + SNR_L\Big|_{dB} + 10 \log_{10} B - 173.81$$

PA CLASSES

I MUST CONSIDER MESFET DEVICES

CLASS A: My device is quasi linear and IS ALWAYS ON; the BIAS POINT IS IN THE CENTER OF THE CHARACTERISTIC (I_C, V_{BE}) OR (I_D, V_{DS})
EVEN IF I DON'T APPLY INPUT SIGNAL I HAVE I_C OR $I_D \neq 0$

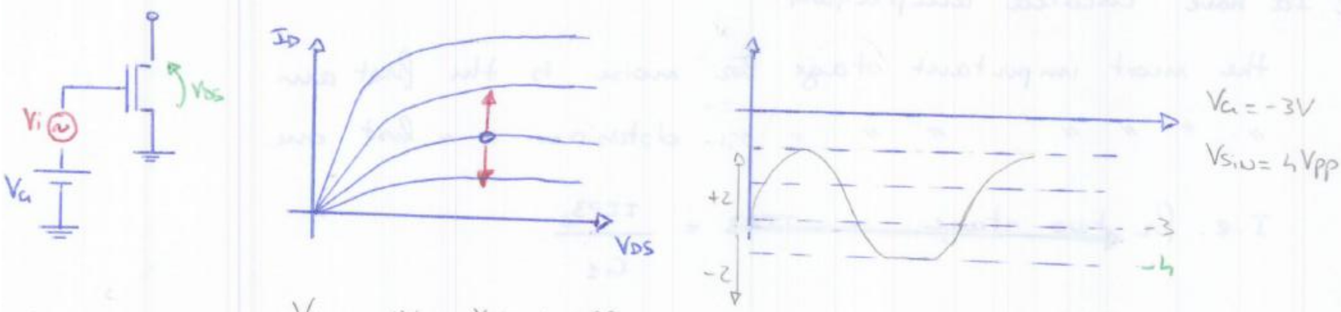
CLASS B: My device IS ON EXACTLY FOR HALF PERIOD OF MY SINUSOIDAL INPUT, example:



CLASS AB The behavior is in between A and B since CURRENT FLOWS FOR MORE THAN 1/2 PERIOD BUT LESS THAN AN ENTIRE ONE

CLASS C Current FLOWS FOR LESS THAN 1/2 PERIOD

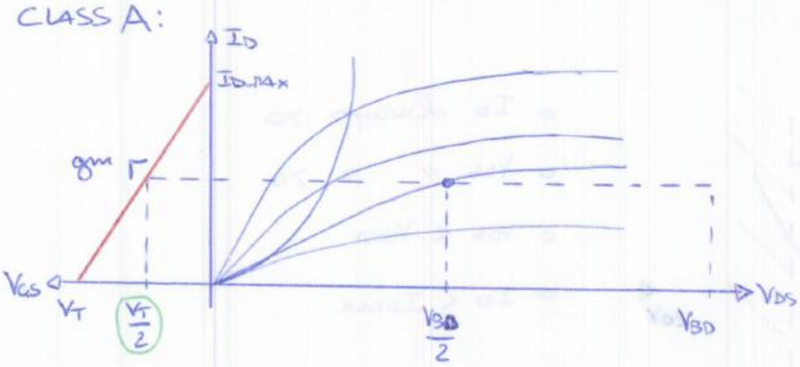
• CLASS C

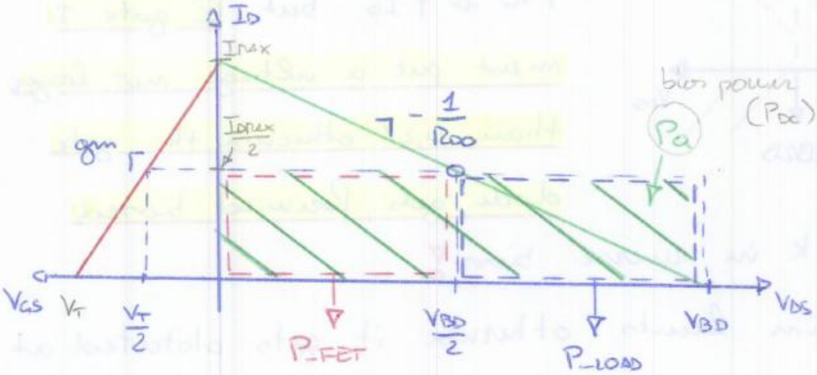


$V_G = -4V \to V_{pinch\ off}$

If V_G is such that my signal goes below V_{PO} it gets clipped

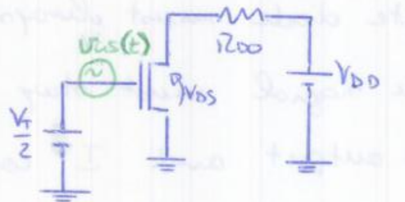
CLASS A:





Consider $V_{BD} = V_{DD}$

$$\boxed{R_{DO} = \frac{V_{BD}}{I_{MAX}} = \frac{V_{DD}}{I_{MAX}}}$$



I bias my circuit such that I have $I_{D\_MAX}/2$
for $V_{GS} = V_T$ → pinch off

What is the delivered power?

$$\boxed{P_a = V_{BD}\left(\frac{I_{D\_MAX}}{2}\right)} \rightarrow \text{quieshent power}$$

If I apply a signal in input $v_{GS}(t)$ which is synusoidal, then I can express the drain current as $I_D(t) = I_{DQ} + i_D(t)$

Now, I MUST BE CAREFULL SINCE I HAVE TWO POWERS, THE ONE DISSIPATED BY THE FET AND THE ONE DISSIPATED BY THE LOAD

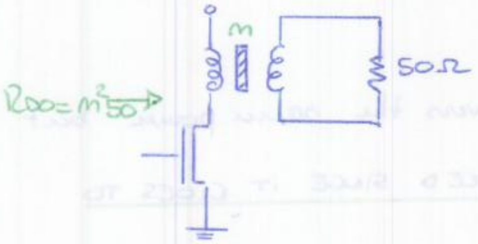
What is the average power given by the battery when I apply the signal?:

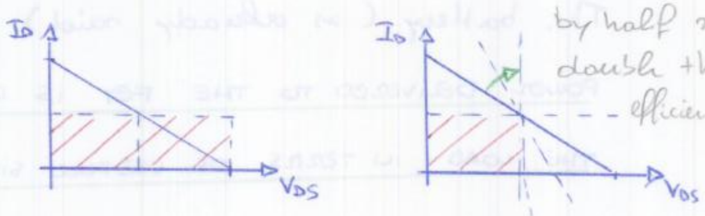
$$\frac{1}{T}\int_0^T I_D(t)\, V_{BD}\, dt = \frac{V_{BD}}{T}\int_0^T \underbrace{\left[I_{DQ} + i_D(t)\right]}\, dt = V_{BD}\, I_{DQ}$$

For a class A amplifier, the power delivery by the battery is the same whereas there is input signal or not

TUNED LOAD

For $R_D = 0$ I'd have a vertical load line (∞ slope)
My problem is that I DON'T WANT R_D IN DC BUT I NEED IT WHEN I HAVE A SIGNAL → I need a load which is 0 in DC and ≠ 0
I could use a TRANSFORMER → STATIC AND DYNAMIC LOAD LINES ARE DECOUPLED

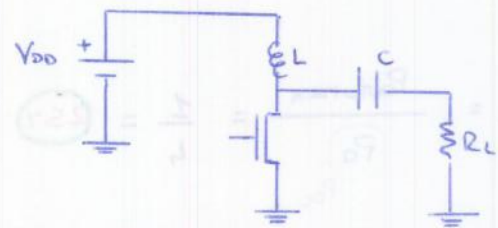




I can reduce P_{DC} by half so double the efficiency

$$\eta = \frac{P_{OUT\_MAX}}{P_{DC}} = \frac{1/8}{1/4} = \frac{1}{2}$$ → TUNED LOAD CLASS A AMP → $\eta = 50\%$

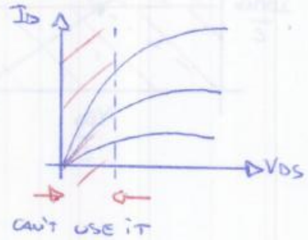
Problem: @ high frequency I cannot use a transformer so I use a configuration as:



BIAS CIRCUIT FOR HF AMPLIFIERS

Tuned load class A amp.

Remark: 50% is the maximum theoretical limit for class A but generally no more than (30/35)% is achieved because I cannot exploit the complete V_{DS} I_D graph.



Class A amplifier is at the worst temperature condition when I apply no signal to it because all the power is dissipated by the FET!

no signal → class A heats-up since it dissipates all power itself

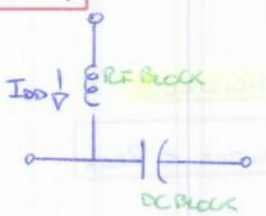
22/10/14

To decrease distorsion at the output of my amplifier I can insert a resonator at the output such that I have

- OPEN CIRCUIT @ f_0
- CLOSE CIRCUIT @ $2f_0, 3f_0$ → remove from the load the harmonic components.

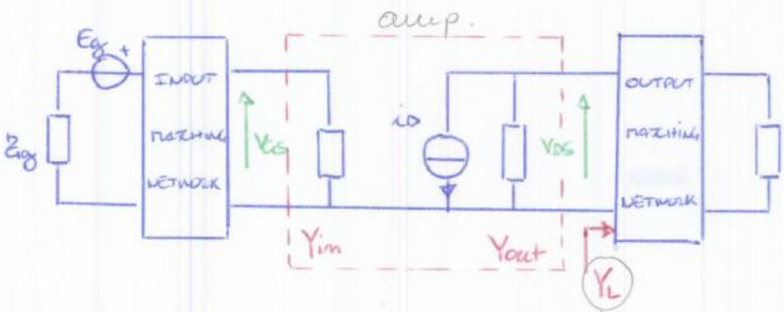
The drawback is that I reduce the bandwidth since the resonator works just @ its resonant frequency

BIAS T



working in the range of GHz what matters most is the reactance of the capacitor (it is, in practice, enough to place a wire of few pH value since, for an inductor reactance is $\omega \cdot L$)

CLASS A DESIGN



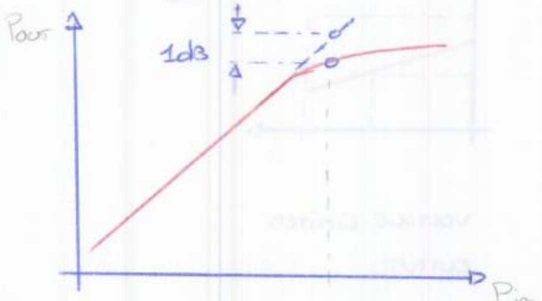
In an amplifier for gain it is possible to identify an optimum load for gain maximization

TO MAXIMIZE POWER I MUST PRESENT A PROPER TERMINATION (that is the one that let the load line pass through the diagonal exactly) this is the real part

The general expression is

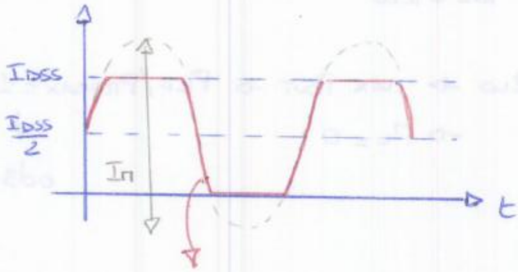
$$\boxed{Y_L = \frac{1}{R_{lo}} - jB_{out}}$$

QUASI LINEAR REGIME



If I'm above the 1dB compression point (I've more compression) my device cannot be anymore considered linear.

CURRENT CLIPPING



$\frac{I_{DSS}}{2} + I_M \cos(\omega t)$

Current clips @ $P_{IN} = P_{IN\_LIM}$

" " where $I_M > I_{DSS}/2$

ξ is the angle for which $I_M \cos(\xi) = \frac{I_{DSS}}{2}$

$$\cos(\xi) = \frac{I_{DSS}}{2 I_M} \longrightarrow \boxed{\xi = \cos^{-1}\left[\frac{I_{DSS}}{2 I_M}\right]}$$

for $\xi = 0 \Longleftrightarrow$ NO CLIPPING

Analitically (from slides)

$$i_D(t) = a_0 + \sum_{n=1,3,5}^{n=\infty} a_n \cos(n\omega t)$$

$$a_0 = \frac{I_{DSS}}{2} \quad \text{(dc component)}$$

$$a_1 = I_M \left\{ 1 - \frac{2}{\pi} \cos^{-1}(\xi) + \frac{1}{\pi} \sin\left[2\cos^{-1}(\xi)\right] \right\}$$

power saturation threshold: $P_{IN} = P_{IN\_LIM}$

$P_{OUT} = G_{OP} P_{IN}$ for $P_{IN} < P_{IN\_LIM}$

$P_{OUT} = G_{OP} P_{IN} \left\{ 1 - \frac{2}{\pi} \cos^{-1}(\xi) + \frac{1}{\pi} \sin\left[2\cos^{-1}(\xi)\right] \right\}^2$ for $P_{IN} > P_{IN\_LIM}$

SATURATION POWER $\quad \boxed{P_{SAT} = \frac{16}{\pi^2} G_{OP} P_{IN\_LIM}}$

1,6

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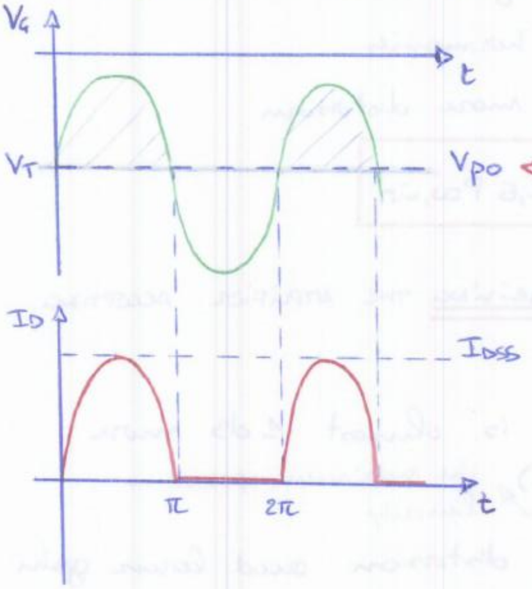
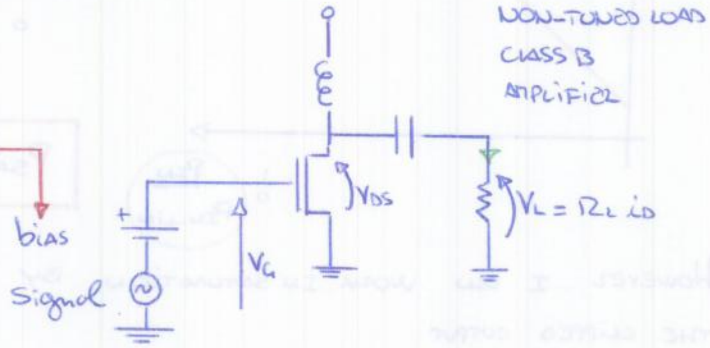
CLASS B AMPLIFIER

→ GATE BIASED @ PINCH OFF AND CURRENT FLOWS ONLY FOR 1/2 PERIOD

Configuration used for increasing efficiency





For having P_{max} I need $I_D = I_{DSS}$; I expand in Fourier $I_D(t)$

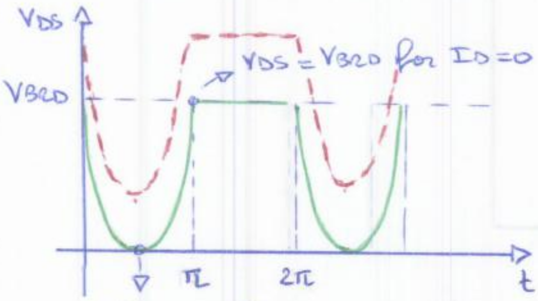
$$I_D(t) = \frac{1}{2\pi}\int_0^{\pi} I_{DSS}\sin(t)\,dt = \frac{1}{2\pi}\cos(t)\Big|_0^{T} I_{DSS} = \boxed{\frac{I_{DSS}}{\pi}} \quad \text{(DC comp.)}$$

$$\frac{2}{2\pi}\int_0^{\pi}\sin^2(t) I_{DSS}\,dt = \frac{I_{DSS}}{\pi}\left[\frac{1}{2}t - \frac{1}{4}\sin(2t)\right]_0^{\pi} = \boxed{\frac{I_{DSS}}{2}} \quad \text{(fundam.)}$$

$$\left[\; I_D(t) = \underbrace{\frac{I_{DSS}}{\pi}}_{DC} + \underbrace{\frac{I_{DSS}}{2}}_{fund.}\sin(\omega t) + \ldots \;\right]$$

What about V_{DS}? → for $I_D = 0$ it is constant

the average value must be the battery value ($V_{BRD}/2$)



$V_{DS} = 0$ for $I_D = I_{DSS}$

--- original V_{DS} waveform

— V_{DS} I waveform I want

(I always consider the "best swing" condition) (working with the battery and R_L I should be able to get $0 < V_{DS} < V_{BRD}$)

MAX POWER CONDITION
$$\begin{bmatrix} V_{DS\_min} = 0 \\ V_{DS\_max} = V_{BRD} \end{bmatrix} \quad \begin{bmatrix} I_{DS\_min} = 0 \\ I_{DS\_max} = I_{DSS} \end{bmatrix}$$

MAXIMUM EFFICIENCY

$$\boxed{P_{OUT}} = \frac{1}{2}(\text{max swing of current})^2 \cdot R_L$$

$$= \frac{1}{2}\left(\frac{I_{DSS}}{2}\right)^2 \cdot R_L$$

$$= \frac{I_{DSS}^2}{8}\,\frac{V_{BD}}{\cancel{I_{DSS}}} = \boxed{\frac{V_{BD}\,I_{DSS}}{8}} \quad (\text{same as class A})$$

$$\boxed{P_{DC} = V_{BD}\left(1-\frac{1}{\pi}\right)\left[\frac{I_{DSS}}{\pi}\right]} \quad V_{BATT}\cdot I_{DC}$$

$$\boxed{\eta} = \frac{P_{OUT\_max}}{P_{DC}} = \frac{1}{8}\,\frac{\pi^2}{\pi-1} \cong 58\%$$

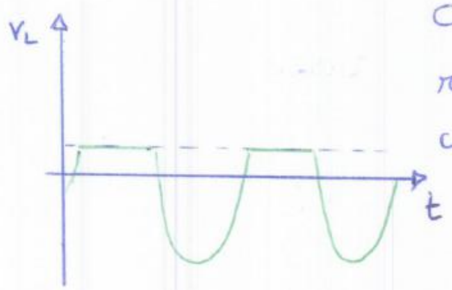
The improvement is not drastic since I still have harmonics reaching my load and I reduced the power of the battery. Moreover, here, the battery delivers power according to the signal since this depends on the DC component of drain current

→ no signal applied → no dissipation because no current

→ signal applied → power dissipated

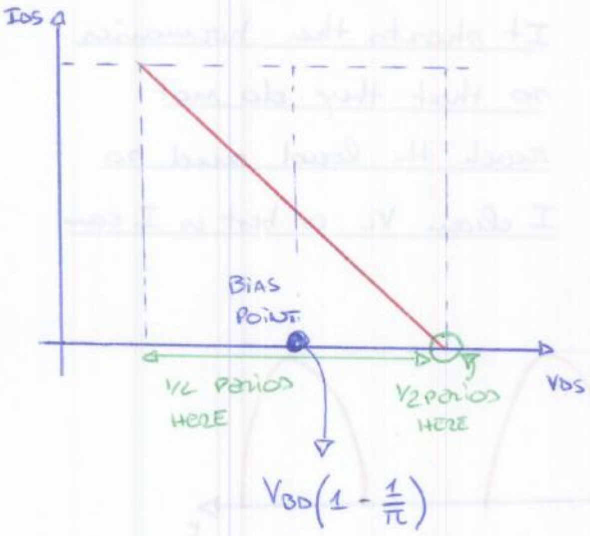
η accounts only for the fundamental

harmonics destroy the efficiency, how to improve? → RESONATOR @ OUTPUT

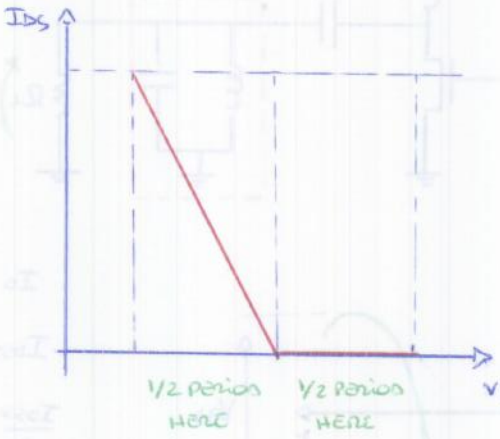


Capacitor @ the output removes the DC compont to V_{DS} waveform.

LOAD LINE UNTUNED CLASS B



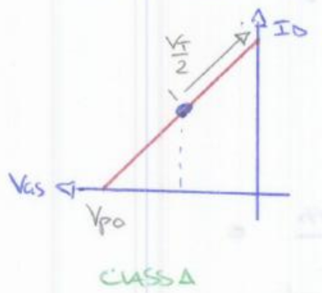
LOAD LINE TUNED CLASS B

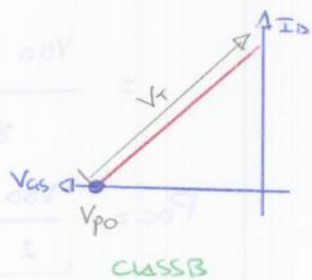


What about gain?

if $P_{OUT}\Big|_{CLASS A} = P_{OUT}\Big|_{CLASS B} \not\Leftrightarrow P_{IN}\Big|_{CLASS A} = P_{IN}\Big|_{CLASS B}$

This is due to the different bias point





In class B I need a doubled input to get the same output as class A

This means that I need 4x more the input power to get the same output power.

To get the same output power as a class A, a class B amplifier requires 4x the input power so +6dB of input power (and, as a consequence requires ↓ 6dB of gain) o CLASS B : η DECREASES AS $\sqrt[2]{\ }$ OF BACK-OFF

example: 10 dB PAPR → 1/3 η

↳ $\sqrt{10} \cong 1/3$ ↑

Summary

o CLASS A:

→ best linearity because the bias is at the center of the characteristics

→ largest bandwidth since there are no resonators

→ η decreases as back off (linearly)

If I have a 3dB PAPR signal I must work in 3dB back off condition (P_{avg} is 3dB less than P_{max})

o CLASS B:

→ WITH NO RESONATOR voltage and current are similar to 1/2 a sinusoid and $\eta = 58\%$

→ WITH RESONANT CIRCUIT the harmonics of the voltage are filtered and do not reach the load, $\eta = 78\%$ and η decreases as the $\sqrt{\ }$ of back off → reacts better at same PAPR

Drawback is that the system is less linear and the bandwidth is reduced

From the previous definition I get: $I_q + I_p \cos\left(\frac{\alpha}{2}\right) = 0$
combining this with the general expression of i_o I get a system

$$\begin{cases} I_q = -I_p \cos\left(\frac{\alpha}{2}\right) \\ I_P = I_q + I_p = -I_p \cos\left(\frac{\alpha}{2}\right) + I_p \end{cases}$$

$$\Rightarrow I_P = \left[1 - \cos\left(\frac{\alpha}{2}\right)\right] I_p$$

$$\Rightarrow I_p = \frac{I_P}{\left[1 - \cos\left(\frac{\alpha}{2}\right)\right]} \Rightarrow I_q = -\frac{I_P}{\left[1 - \cos\left(\frac{\alpha}{2}\right)\right]} \cos\left(\frac{\alpha}{2}\right)$$

Finally

$$i_o(\vartheta(t)) = -\frac{I_P}{\left[1 - \cos\left(\frac{\alpha}{2}\right)\right]} \cdot \cos\left(\frac{\alpha}{2}\right) + \frac{I_P}{\left[1 - \cos\left(\frac{\alpha}{2}\right)\right]} \cos(\vartheta)$$

$$= \frac{I_P}{\left[1 - \cos\left(\frac{\alpha}{2}\right)\right]} \left[\cos(\vartheta) - \cos\left(\frac{\alpha}{2}\right)\right]$$

What about fourier coefficients?

○ AVERAGE VALUE: I_o

$$I_o = \frac{1}{T}\left[\int_0^{\alpha/2} i_o(\vartheta)\, d\vartheta\right] \cdot \underset{\substack{\uparrow \\ \text{because I have} \\ \text{two regions } (0;\frac{\alpha}{2}) \text{ and} \\ (\pi - \frac{\alpha}{2};\, \pi)}}{2} = \frac{1}{\pi} \int_0^{\alpha/2} \frac{I_P}{\left[1 - \cos\left(\frac{\alpha}{2}\right)\right]} \left[\cos(\vartheta) - \cos\left(\frac{\alpha}{2}\right)\right] d\vartheta =$$

$$= \frac{I_P}{\pi} \frac{1}{\left[1 - \cos\left(\frac{\alpha}{2}\right)\right]} \left[\sin(\vartheta)\Big|_0^{\alpha/2} - \cos\left(\frac{\alpha}{2}\right)\left(\frac{\alpha}{2}\right)\right] =$$

$$= \frac{I_P}{\pi} \frac{1}{\left[1 - \cos\left(\frac{\alpha}{2}\right)\right]} \left[\sin\left(\frac{\alpha}{2}\right) - \cos\left(\frac{\alpha}{2}\right)\left(\frac{\alpha}{2}\right)\right]$$

Same as slide (sust group a 2 from [1])

For maximum swing I must place the bias point @ $\frac{V_{BD}}{2}$

$V_{pp} = V_{BD}$ $\quad V_p = V_{BD}/2 \quad I_{peak} = I_1$

I can define

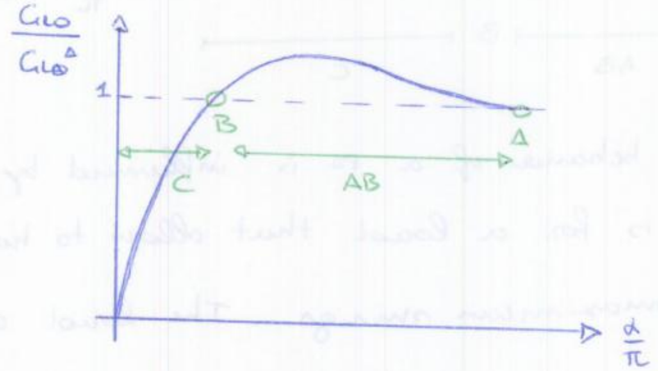
$\boxed{\text{OPTIMUM LOAD CONDUCTANCE}}$

$$G_{LO}(\alpha) = \overbrace{\left[\frac{I_P}{2\pi}\,\frac{\alpha - \sin\alpha}{\left[1-\cos\left(\frac{\alpha}{2}\right)\right]}\right]}^{\text{peak current}}\overbrace{\left[\frac{V_{BD}}{2}\right]^{-1}}^{\text{peak voltage}} = \qquad G = \frac{I}{V}$$

$$= \frac{I_{DSS}}{V_{BD}}\left[\frac{1}{\pi}\,\frac{\alpha - \sin(\alpha)}{\left[1-\cos\left(\frac{\alpha}{2}\right)\right]}\right] =$$

$$= G_{LO}^{A}\left[\frac{1}{\pi}\,\frac{\alpha - \sin(\alpha)}{\left[1-\cos\left(\frac{\alpha}{2}\right)\right]}\right]$$

G_{LO}^{A} optimum load transconductance for class A

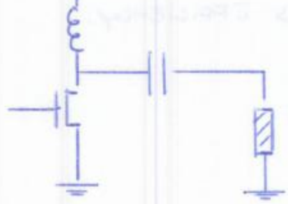


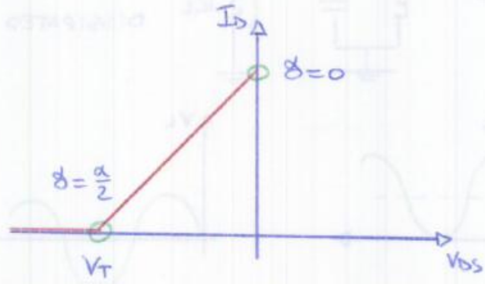
RF POWER : $P_{RF} = \frac{1}{2} I_{peak} V_{peak} = \frac{1}{2} I_1 \frac{V_{BD}}{2} = \frac{1}{2} I_1 V_{DC}$ $\qquad V_{DC} = \frac{V_{BD}}{2}$

DC POWER : $P_{DC} = V_{DC} I_0$

↳ DC COMPONENT OF CURRENT

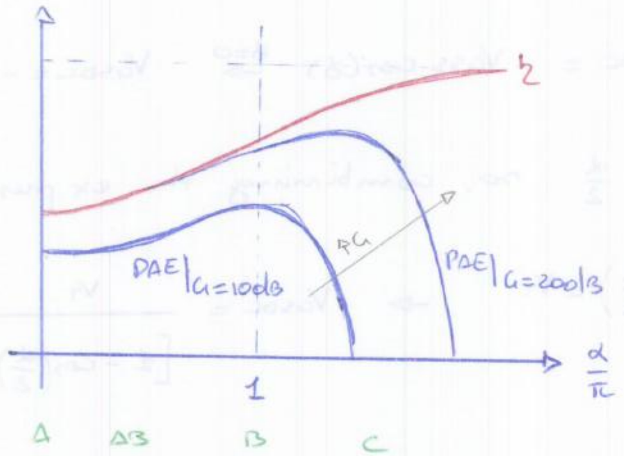
How can I force the proper α? → by V_{GS} because it is the bias that defines when I have $I_D = 0$ or not





$$PAE = \eta\left(1 - \frac{1}{G}\right) =$$ PAE as function of α

$$= \frac{1}{2} \frac{\alpha - \sin(\alpha)}{2\sin\left(\frac{\alpha}{2}\right) - \alpha\cos\left(\frac{\alpha}{2}\right)} \left\{ 1 - \frac{1}{G_A} \frac{4\pi}{[\alpha - \sin(\alpha)][1 - \cos\left(\frac{\alpha}{2}\right)]} \right\}$$



higher the gain the more PAE ≈ η

$\uparrow \eta \Rightarrow \downarrow \alpha$

but $\downarrow \alpha \Rightarrow \downarrow G$

So $PAE - \eta \uparrow$

The higher the gain, the more PAE factor approaches η but, the two blue curves do not reach the efficiency because, as η grows α decreases and this implies that gain decreases too and as a consequence PAE lines do not approach η even if I boost gain.

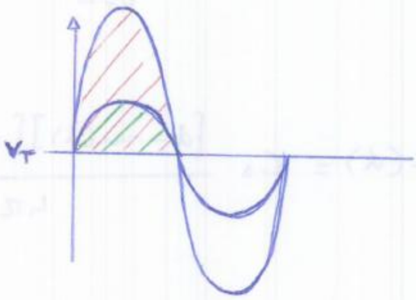
CLASS A and CLASS B are the only on which CIRCULATION ANGLE DOES NOT DEPEND ON THE APPLIED SIGNAL So the gain is not affected by the amplitude of input signal.

In class C $\Rightarrow \uparrow \alpha \Rightarrow$ DC voltage changes

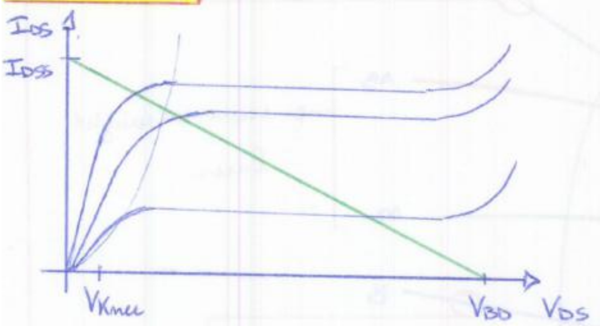
THE DEVICE IS SELF BIASING SINCE GATE SIGNAL MOVES THE DC VOLTAGE

CHANGE DC $\Rightarrow$ CHANGE g_m $\Rightarrow$ CHANGE GAIN

$\Rightarrow$ STRONGLY NON-LINEAR CIRCUIT



Exercise

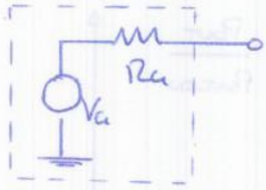


$I_{DSS} = 1A$

$V_{BD} = 20V$

$V_{Knee} = 1V$

$V_{PO} = -2V$



generator

$V_{GSpp} = 2V \quad R_G = 50\Omega$

Evaluate all parameters of interest both for a class A and a class B amplifier

Output Power

$$P_{OUT\_MAX} = \frac{V_{BD}\, I_{DSS}}{8} = \frac{1}{8}\, 20 \cdot 1 = 2{,}5W \longrightarrow 10\log_{10}(2{,}5/1\cdot 10^{-3}) = 34\,dBm$$

for class B, maximum output power is the same

Load Resistance

$$R_{OD} = \frac{V_{BD}}{I_{DSS}} = 20\Omega \quad (19\Omega \text{ considering } V_{Knee})$$

Bias Point

$$\left(\frac{I_{DSS}}{2}\;;\;\frac{V_{BD}-V_{Knee}}{2}\right) = (0{,}5A;\; 9{,}5V)$$

Gate Bias

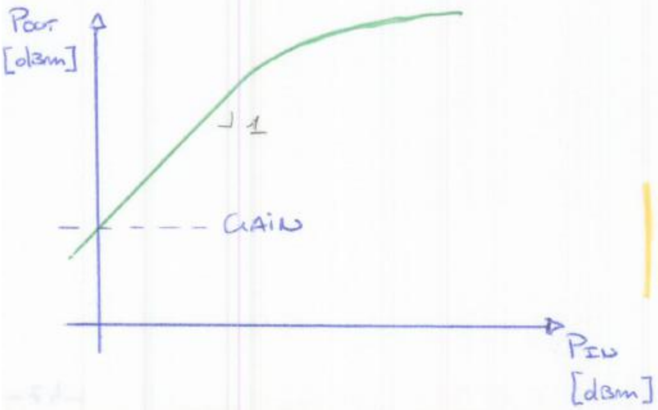
$$V_G\Big|_{CLASS\,A} = \frac{V_{PO}}{2} = -1V$$

$$V_G\Big|_{CLASS\,B} = V_{PO} = -2V$$

Efficiency:

CLASS A: $\eta = 50\%$

CLASS B: $\eta = 78\%$ (assuming tuned load)



When I have $P_{IN} = 0\,dBm$, the value of P_{OUT} is actually the gain in dBm because

$$G = \frac{P_{OUT}}{P_{IN}} \qquad P_{OUT} = G\,P_{IN} \quad \text{So}$$

$$P_{OUT}\Big|_{[dBm]} = G\Big|_{[dBm]} + P_{IN}\Big|_{[dBm]}$$

HIGH EFFICIENCY POWER AMPLIFIERS

I can say that some non linear effect is taking place if I have an output spectrum $\neq 0$ in the condition of input spectrum $= 0$ (generation of harmonic components other than the fundamental)

ACPR DEPENDS ON THE TYPE OF MODULATION

FM modulation is used instead of AM because, for an external effect, to change frequency it would be a non-linear one. (a cloud can reduce the amplitude of the signal introducing attenuation but cannot modify the frequency of it)

modulation turns a baseband signal into a passband signal centered around the carrier frequency f_c

$$B_{FM} \simeq \left[2 \frac{m A_m}{\omega_m} + 1 \right] B_{BB}$$

So frequency modulation increases the bandwidth and so the SNR.

demodulators for FM:

- PLL
- 

A modulated signal can be represented by mean of base functions (which are orthogonal among them)

$$x(t) = \alpha_1 \phi_1(t) + \alpha_2 \phi_2(t) + \dots + \alpha_n \phi_n(t)$$

ASK $x(t) = \alpha_1 \cos(\omega_c t)$ $\quad \alpha_1 = (0, A_c)$

PSK $x(t) = \alpha_1 \cos(\omega_c t)$ $\quad \alpha_1 = (-A_c, A_c)$

FSK $x(t) = \alpha_1 \cos(\omega_1 t) + \alpha_2 \cos(\omega_2 t)$

$$[\alpha_1, \alpha_2] = [(0, A_c), (A_c, 0)]$$

Spectral regrowth vs constant/variable envelope

$$x(t) \longrightarrow \boxed{} \longrightarrow y(t) = \alpha x^3(t)$$

constant amplitude signals:

$$x(t) = A \cos[\omega t + \varphi(t)]$$

$$y(t) = \frac{\alpha A^3}{4} \cos[3\omega t + 3\varphi(t)] + \frac{3\alpha A^3}{4} \cos[\omega t + \varphi(t)]$$

variable amplitude signals

$$x(t) = A_I(t) \cos(\omega t) - A_Q(t) \sin(\omega t)$$

$$y(t) = \dots + \frac{3\alpha A_I^3(t)}{4} \cos(\omega t) - \frac{3\alpha A_Q^3(t)}{4} \sin(\omega t) + \dots$$

SPECTRAL CONTENTS OF IN-PHASE AND QUADRATURE IS WIDER → SPECTRUM SHOWS GROWTH.

($A_I(t)$ and $A_Q(t)$ are baseband signals → if I cube them I get spectral regrowth)

This implies that requirements for constant envelop signal amplifiers are different from non-constant envelop ones.

Constraints for mobile applications and base stations are different.
For MOBILE I can have max 30dBm power (1W); high efficiency means longher battery lifetime and lower power dissipation. For mobile, linearity is not a critical issue I work just on one channel (I don't care about spectral regrowth since there are no other channels)

For BASESTATIONS, I need high output power (10 ÷ 350)W and linearity (since I have multiple channels) and efficiency

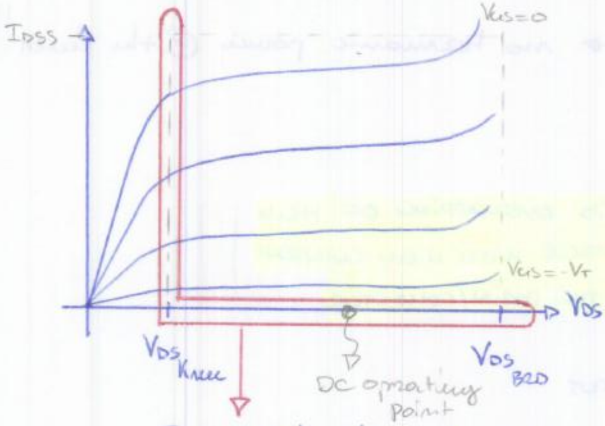
Problem is: η and linearity are trade-off conditions.
High η requires to work @ max power while High linearity @ low power.

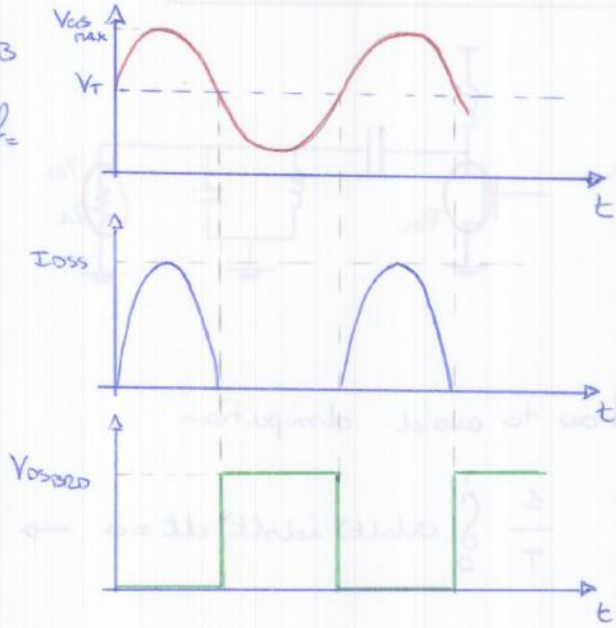
CLASS F AMPLIFIER

Input waveform is the same as class B

The load at different harmonics is different.

I want to control the harmonics





I want the device to work just at the border of the active region → no power dissipation

Fourier expressions for v_{ds} and i_D

$$v_{ds}(t) = V_{DS_{BRD}}\left[\underbrace{\frac{1}{2}}_{DC} - \frac{2}{\pi}\underbrace{\sum_{m=1,3,5,\dots}\frac{1}{m}\sin(m\omega t)}_{ODD}\right]$$

$$i_{DS}(t) = I_{DSS}\left[\underbrace{\frac{1}{\pi}}_{DC} + \frac{1}{2}\sin(\omega t) - \frac{2}{\pi}\underbrace{\sum_{m=2,4,6,\dots}\frac{1}{m^2-1}\cos(m\omega t)}_{EVEN}\right]$$

@ fundamental:

$$R_{L0} = \frac{4 V_{DS_{BRD}}}{\pi I_{DSS}}$$

$$P_{RF} = \frac{1}{2}\,\frac{2V_{DS_{BRD}}}{\pi}\,\frac{I_{DSS}}{2} = \frac{I_{DSS} V_{DS_{BRD}}}{2\pi}$$

$$P_{DC} = \frac{I_{DSS}}{\pi}\,\frac{V_{DS_{BRD}}}{2} = \frac{I_{DSS} V_{BRD}}{2\pi}$$

$$\eta = \frac{P_{RF}}{P_{DC}} = 100\%$$

ALL DC POWER GOES TO RF

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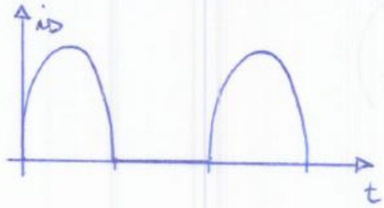
REMARK: class F amplifier is based on two main concepts

1) AVOID THE OVERLAP OF VOLTAGE AND CURRENT (guarantees no P_{DC} and so no self heating)

2) NO POWER DISSIPATED BY HARMONIC COMPONENTS @ THE LOAD

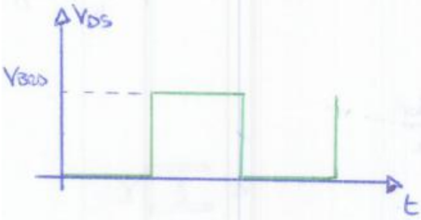
These two conditions are essential to get $\eta = 100\%$.

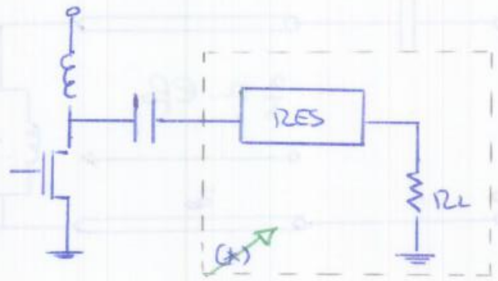


$$I_D = I_{DSS}\left[1 - \frac{V_{GS}}{V_P}\right] \quad \text{linear model}$$

where $V_{GS} = V_{GS}\big|_{DC} = V_P$

Using the linear model for $I_D(V_{GS})$ I can get a perfect ½ sinusoid shape of the current (whereas if I use the quadratic model I get a quadratic shape of I_D) The bias is the same as class B

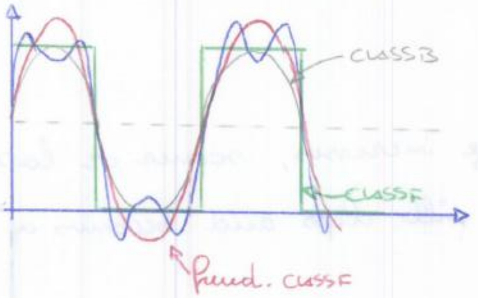




I would like to do the same in order to get a square wave as V_{DS} (Doing so I'd have in theory a null power dissipation) How?

I must properly chose the harmonic loading condition ⇒ the (*) block must produce the green squarewave moreover, the resonator must be purely reactive ⇒ no losses inside of it

The voltage is made of just odd harmonics; due to this fact I have a waveform centered with respect to the mean value.



Since I have the same current swing; the optimum load is $\frac{4}{\pi}$ the old one ⇒ I have a voltage swing $\frac{4}{\pi}$ times larger than the old one

In back off, current is reduced as well as voltage, so efficiency reduces, moreover

if $R_L > \frac{Z_0^2}{R_{L0}}$ from the device I "see" a lower value of R_L and this means that the voltage don't reach the maximum value possible (lower resistance higher current and lower voltage)

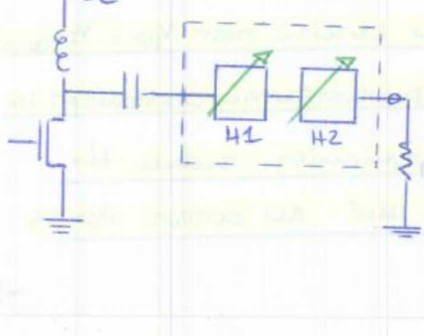
if $R_L < \frac{Z_0^2}{R_{L0}}$ from the device I "see" an higher value of R_L and this means that current does not reach its maximum value possible. (higher resistance lower current and higher voltage)

I have $\eta = 100\%$ but no maximum power output.

In real cases I consider just second and third harmonics (low order class F)

$\boxed{\text{LOAD PULL}}$

Empirical method that consists in measuring the behaviour of the device changing real and immaginary part of the load in



order to obtain, let say, a given output power; after having identified the best load for the case, I can design an OUTPUT MATCHING NETWORK which is basically a filter → linear → can be designed with the

The optimum small signal load is the one that optimizes the gain (no power concern)

Unconditional stability implies gain maximization

ECE 1352F
Analog Integrated Circuits I

Reading Assignment -

RF Power Amplifiers

CLASS-F

Prepared by: Poon, Alan Siu Kei

I. Introduction

The rapid growth of mobile telecommunications services created increasing demand for low-cost, low-power and reduced size and weight equipments. An increasingly higher level of integration is needed to meet these requirements. Thanks to the advancement in deep sub-micron CMOS technology, this is easily achievable for digital signal and low-frequency signal processing. However, in order to reach the final goal of System-on-a-Chip (SoC) solution, the final piece of puzzle is still missing – the RF front end. In fact, being the most power hungry component of the RF front end, it is widely known that the RF power amplifier (PA) is one of the most critical building blocks in low power SoC integration. Therefore, it is clear that RF PA deserves increased design research to remove the bottleneck of the development of mobile communication devices. Among different classes of PAs, the Class-F topology has been drawing more attention by researchers in the last decade. In this paper, the recent development of Class-F PAs for portable devices will be discussed. In section II, an introduction to Class-F PA operation will be presented. Also, previously developed state of the art Class-F PA systems will be discussed in this section. In section III, current development of Class-F PA will be presented. The future focus and challenges to the development of Class-F PA will be revealed in section IV followed by a conclusion in section V.

Class-F PA has drawn more attention for its easier implementation and better integration with sub-micron CMOS technology.

IIb. Class-F power amplifier operation

A Class-F PA uses a output filter to control the harmonic content of its drain-voltage or drain-current waveforms, thereby shaping them to reduce power dissipation by the transistor and thus to increase efficiency. An example of the output voltage and current waveform of an ideal Class-F PA is shown in Fig. 1 [3].

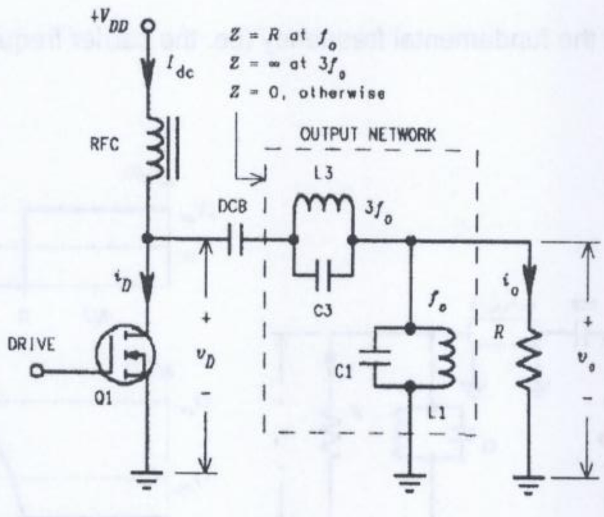


Fig. 1. Example of a Class-F power amplifier.

In the figure, it is noticed that the output voltage waveform is a square wave while the drain current is a half-rectified sinusoid. From the output waveforms it is also noticed that in the ideal case, there is no overlapping between the output voltage

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The impedance seen by the drain can be easily found from this simple equation. At fundamental frequency, the drain sees a pure resistive load of $R_L=Z_L$, since the output tank is open circuit. The tank is a short at all frequencies away from the fundamental. At even harmonics, the transmission line appears to be a half-wavelength line to the drain. We know that for half-wavelength line, the input impedance is $Z_{in}=Z_L$. Therefore, the drain sees a short at all even harmonics, which would result in a half-rectified sinusoid current output as desired. Conversely, at odd harmonics, since the output tank still appears as a short, according to equation (1) the drain sees an open circuit. That is,

$$Z_{in} = \frac{Z_o^2}{Short} = Open \, .$$

If the transistor is assumed to act as a switch, the output network will guarantee that all of the drain voltage will see an open and hence a square wave would be resulted as desired [4]. Therefore, the ideal Class-F output waveform can be achieved by this simple circuit. Note that the same ideal maximum efficiency of 100% can also be achieved by producing a square wave current and half-rectified sinusoid voltage at the output [3]. An example of such implementation using a quarter-wavelength transmission line with series-tuned tank is shown in Fig. 3 [3].

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SWITCHING AMPLIFIERS

Up to now I considered that $I_D(V_{GS})$, for switching amplifiers, this assumption is not anymore valid because I pass from a LOW VOLTAGE - HIGH CURRENT to a LOW CURRENT - HIGH VOLTAGE condition

So the device works either in triode or in saturation regions

They have poor gain since the input signal has to overdrive the device

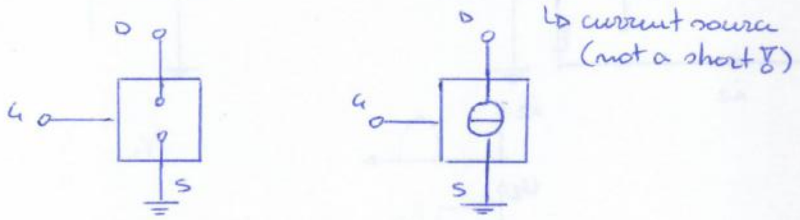
The principle to obtain $\eta = 100\%$ is:

- NO POWER DISSIPATION OF THE ACTIVE DEVICE
- NO HARMONIC POWER AT THE LOAD

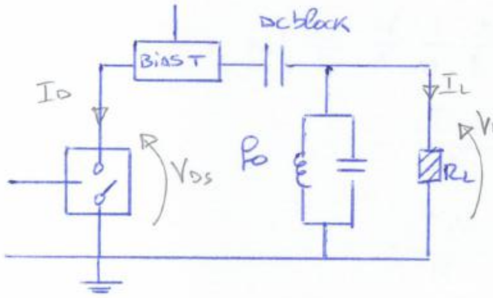
Important to point out is the fact that OUTPUT POWER IN SWITCHING AMPLIFIERS DOES NOT DEPEND LINEARLY ON INPUT POWER. This makes the sw. amp. not a true amplifier; INPUT POWER IS JUST NEEDED FOR MOVING THE DEVICE FROM CUT-OFF TO SAT (and viceversa). Independently on the input I'm always at maximum power condition.

DEVICE MODEL

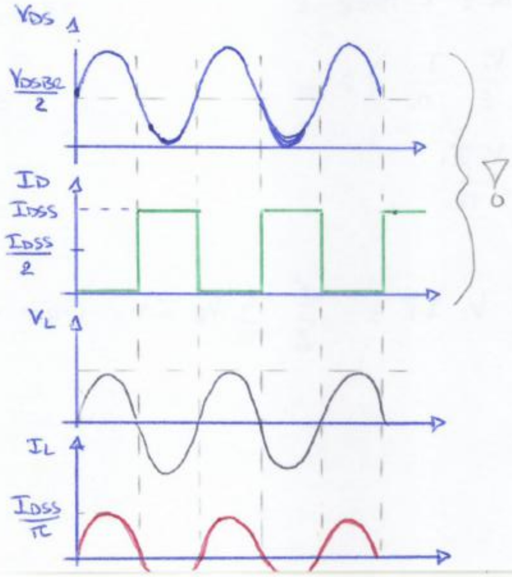
OPEN CONDITION | CLOSE CONDITION



The switch (device) has parasitics so it dissipates some power when it is flown by current.



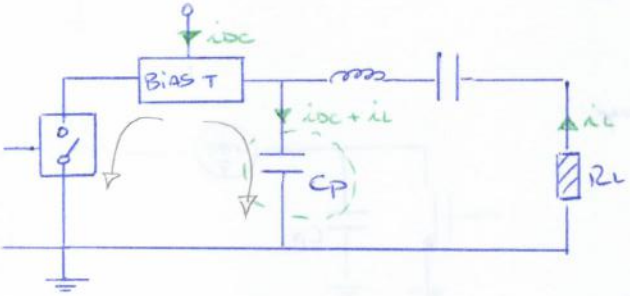
TUNED LOAD
SW. AMPLIFIER



SW ampl. do have to work only with PM or FM signals because amplitude mod is lost if sent through them.

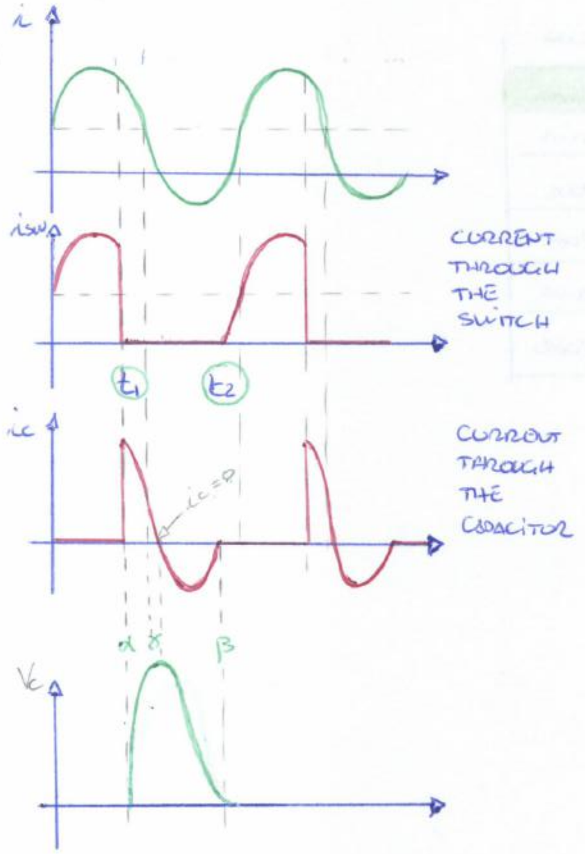
CLASS E AMPLIFIER

The working principle is still the removal of harmonics @ the load and avoid the overlapping of voltage and current waveforms.



C_P is an external component but in practice, it is a combination of external and parasitic capacitances.

In input I provide a SIGNAL WITH A GIVEN DUTY CYCLE so the device changes state at precise instants.
When the switch is open, current goes through the cap, when it is closed, through the device itself. so the current waveform can be split.



SW opens @ t_1
closes @ t_2

Current through the capacitor is integrated since

$$V_C = \frac{1}{C}\int i_C$$

the voltage has maximum value for

$$\frac{dV_C}{dt} = 0 \rightarrow \frac{1}{C} i_C = 0 \rightarrow i_C = 0$$

BY CONSTRUCTION CURRENT AND VOLTAGE OF THE SWITCH NEVER CROSS EACHOTHER + THANKS TO THE RESONATOR, NO HARMONIC POWER REACHES THE LOAD

$\eta = 100\%$

Drawback:
The output power can be only a fraction of the maximum power that the device can handle (example: to handle 10W I'd need a device designed for 50W capability)

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EFFICIENCY ENHANCEMENT TECHNIQUES

Handling NON-CONSTANT envelop signals I have two problems

- LINEARITY is limited by AM and PM distorsion
- EFICIENCY is good ONLY if my amp works @ maximum power (near saturation) because if it is backed-off → η↓

Back off is used in class A and B to get more linearity but there is a trade off to do between efficiency and linearity

When BACK-OFF happen?

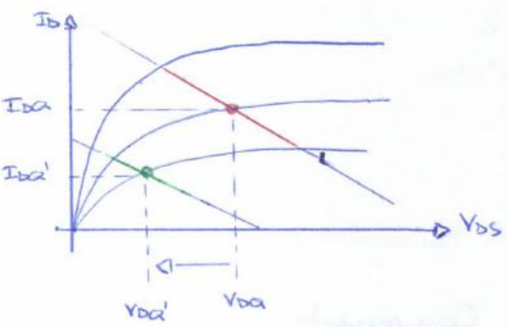
BACKOFF → ↑ LINEARITY, → ↓ EFFICIENCY — TRADE-OFF

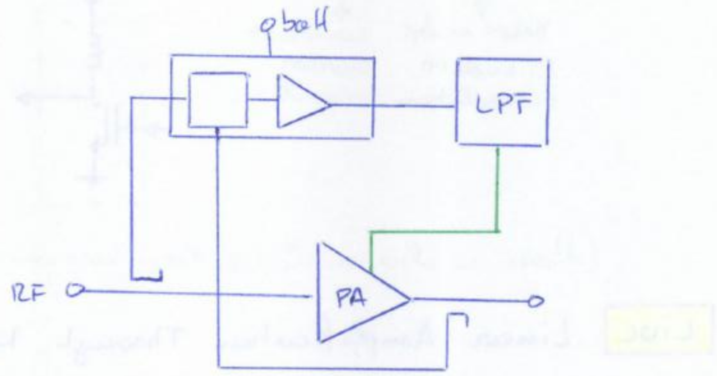
- In BASE STATIONS because of multichannels
- In HANDSETS due to the motion of the device wrt the base station

So backoff happens also with constant envelop signals

SUPPLY MODULATION

The key idea is to USE A BIAS POINT WHICH IS NOT FIXED BUT IT IS FOLLOWING THE ENVELOP CHANGE (for small dynamics I use a small V_{DS_BIAS}, for large dynamics a larger one) SO I CAN ALWAYS WORK WITH A SATURATED AMP.



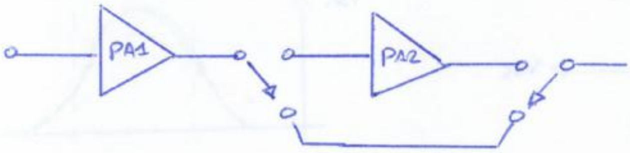


APPLY ONLY THE BIAS I NEED IN ORDER TO HAVE ALWAYS A MAXIMUM SWING AND A SATURATION CONDITION

The device emploied to achieve this objective must be able not to dirt the efficiency improvement it generates; this can be done with a common collector (emitter follower) but it is not a good solution since it has no high efficiency by itself; This block works @ ENVELOP FREQUENCY BAND (not at RF) so it must be

STAGE BYPASSING

Split the amplification in more stages and take output from the appropriate chain by means of switches.



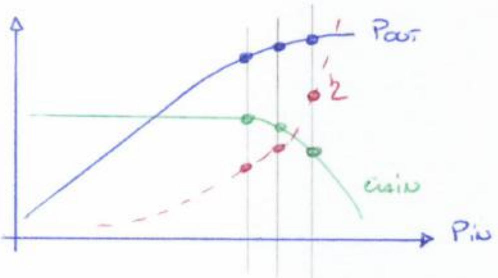
example:

PA1 : designed for 5W handling
PA2 : " " for 10W " "

the only problem is how to drive switches, digitally or directly from the envelop

DOHERTY PA

PA FOR BASE STATIONS

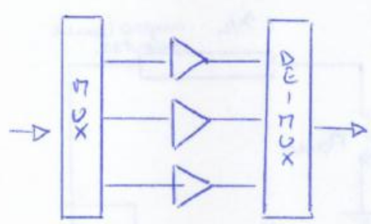


The more compression I have, the lower linearity I have.
Efficiency gets large when I approach saturation, if power is backed off, efficiency drops

Back off usually improves linearity though, so I need a trade off

MULTICHANNEL AMPLIFIER

Either I can have a single amp to which each signal (channel) is sent or a set of amps with MUX-DEMUX structures (but these are fixed)



I NEED TO IMPROVE AMPLIFIER LINEARITY AT HIGH INPUT POWER LEVELS

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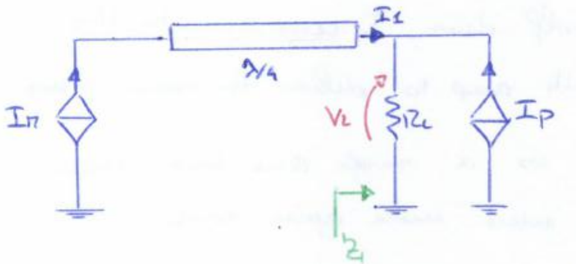
As a first approximation I consider the tr. as current sources.



$$V_L = R_L (I_1 + I_p)$$

$$Z = \frac{[R_L(I_1+I_p)]}{I_1}$$ → test current

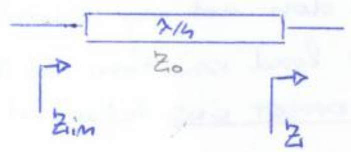
assume transistor n to be a linear class B so $I_n = g_m V_{in}$

for transistor p instead

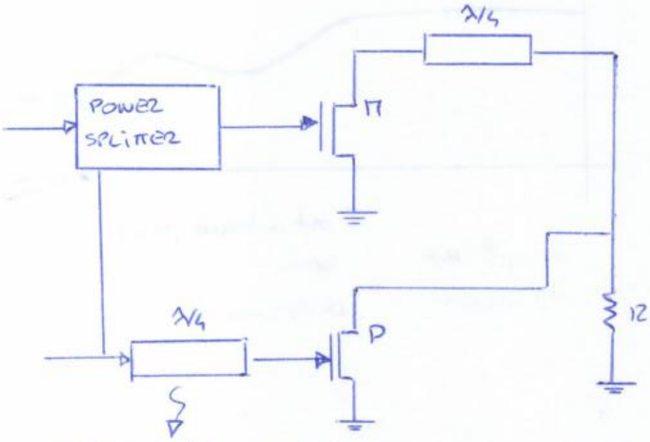
$$I_p = f(V_{in}) = \begin{cases} 0 & \text{for } V_{in} < \frac{V_{in\_SAT}}{2} \longrightarrow Z = R_L \\ g_m V_{in} & \text{for } V_{in} > \frac{V_{in\_SAT}}{2} \longrightarrow Z = 2R_L \end{cases}$$

How? → biasing transistor p in CLASS C

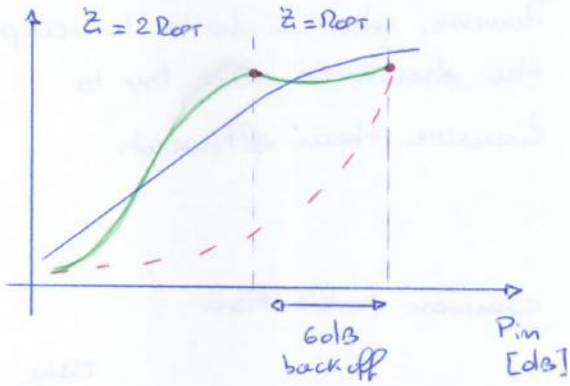
Remark



$$Z_{in} = \frac{Z_0^2}{Z}$$

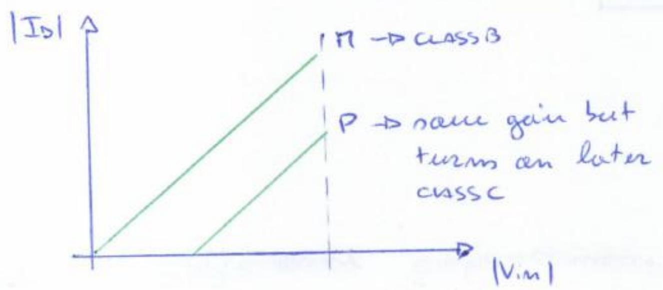


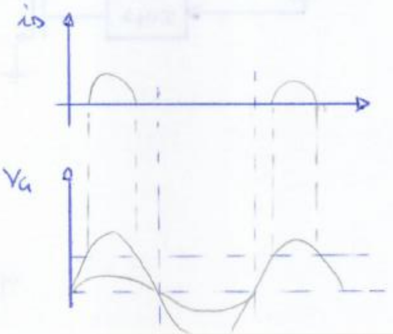
needed to keep currents in phase @ the load (so keep the load real)



1/2 drain voltage
⇓
-6dB power

THE RESULT IS AN IMPROVEMENT @ -6dB





$$\frac{Z_{0x}^2}{R_0} \parallel \frac{2}{3} R_0$$

$$\frac{\frac{Z_{0x}^2}{\cancel{R_0}} \frac{2}{3} \cancel{R_0}}{\frac{Z_{0x}^2}{R_0} + \frac{2}{3} R_0} \qquad \frac{\frac{2 Z_{0x}^2}{3}}{\frac{Z_{0x}^2 + \frac{2}{3} R_0^2}{R_0}} = \frac{Z_{0x}^2 \, 2 \, R_0}{3 Z_{0x}^2 + 2 R_0^2}$$

$$\frac{2 R_0}{\left(3 + 2 \frac{R_0^2}{Z_{0x}^2}\right)}$$

$$\frac{\frac{Z_{0x}^2}{\cancel{R_0}} \cancel{R_0}}{\frac{Z_{0x}^2}{R_0} + R_0} = \frac{Z_{0x}^2}{\frac{Z_{0x}^2 + R_0^2}{R_0}} \qquad \frac{Z_{0x}^2 R_0}{Z_{0x}^2 + R_0^2} \qquad 2R_0^2$$

$$\frac{Z_{0x}^2}{R_0} \parallel R_0 \qquad \frac{\frac{Z_{0x}^2}{R_0} R_0}{\frac{Z_{0x}^2}{R_0} + R_0} \qquad \frac{Z_{0x}^2}{\frac{Z_{0x}^2 + R_0^2}{R_0}} \qquad \frac{R_0 Z_{0x}^2}{Z_{0x}^2 + R_0}$$

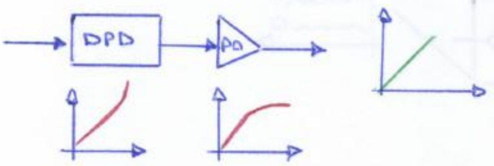
LINEARITY ENHANCEMENT TECHNIQUES

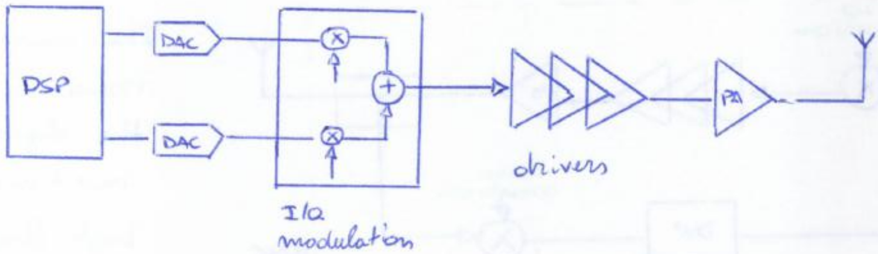
EVM : time domain figure of merit for linearity

$$EVM = \frac{\sum_i \| y(k) - x(k) \|^2}{\sum_i \| x(k) \|^2}$$

PREDISTORSION

The concept is to compensate at the input for the non-linearity of the PA usually it is done digitally ⇒ DIGITAL PRE DISTORSION ⇒ more flexible

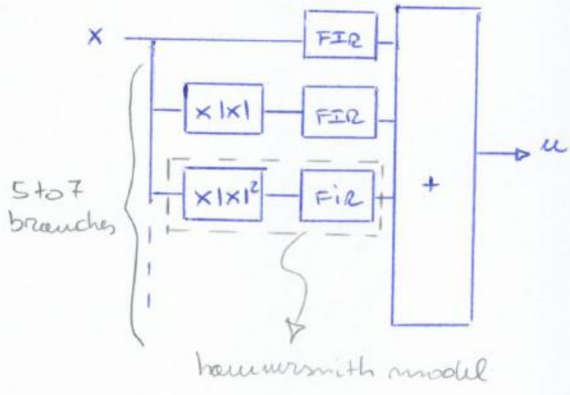




The complexity of the DSP depends on the non-linearity of the PA. Another problem is that, the nonlinearity of the PA is dynamic (due for example to temperature variation) so I must have a dynamic DPD

A model of the PA must be extracted, in particular, a base-band model is obtained for example described by I/Q components. Moreover there are two categories, models with memory and without. Some examples are

- LOOK UP TABLES
- MEMORY POLYNOMIAL →
- NEURAL NETWORKS

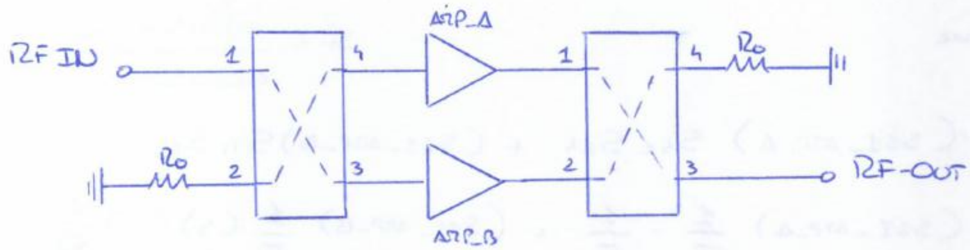


$$u(x) = \sum_{p=0}^{P} x|x|^{2p} b_p$$

$$= b_0 x + b_1 x |x|^2 + \dots$$

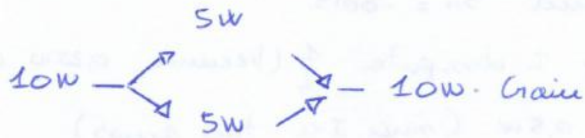
The predistorsion is stored in the coefficients $b_0, b_1, \dots$

BALANCE AMPLIFIER



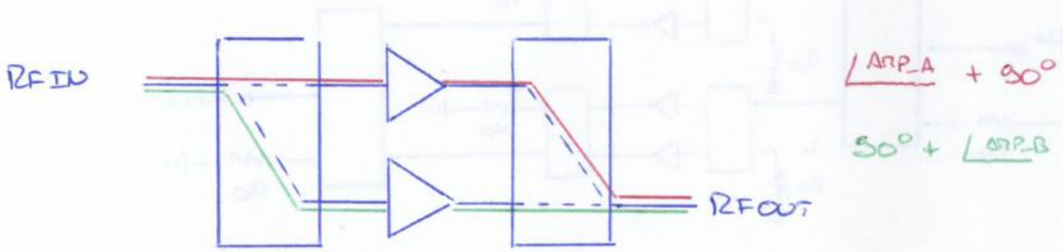
The basic concept is to combine more than one module of the same amplifier in order to increase power level. I want to rise the power level without redesigning the core.

POWER SPLITTERS:
implemented by hybrid modules or four port circuits able to split incoming power and direct it to the output ports. the last port is used for dissipating reflected power. They introduce a voltage phase variation of 90°.
The proposed configuration allows to handle a double amount of input power without modifying the amplifier.



This means that if the amp reaches compression @ 5W, the input is @ 10W

When I combine back the signals I've to take care of delays & but



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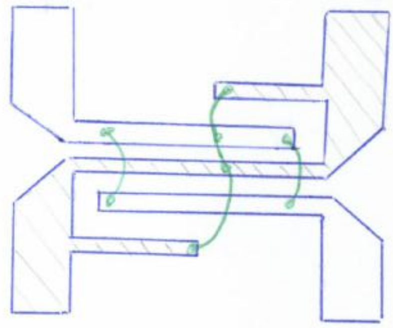
LANGE COUPLER

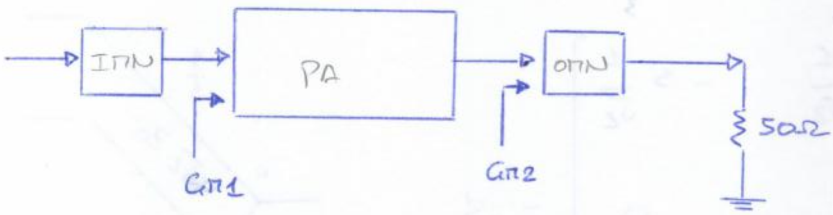


one octave bandwidth

- larger bandwidth
- symmetrical (good for layout)
- spacing between lines is not critical since are connected by wire bonding

FOLD LANGE is better suited since it provides a symmetric layout

Drawback: requires wire bonding.



I cannot have simultaneous matching at the two ports

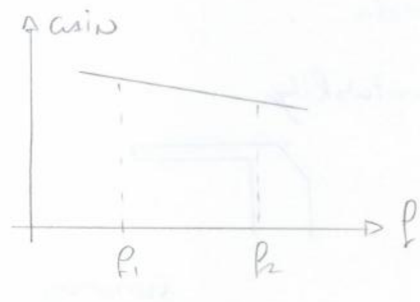
If my load is chosen accordingly to max power are it is different from GΓ2

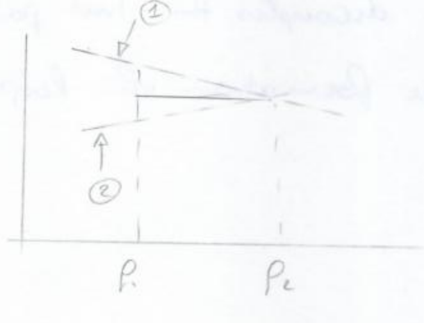
@ the input I want $S_{11} = 0$

at the output is impossible to have $S_{22} = 0$

$S_{11} = S_{22} = 0$ @ the same time → IMPOSSIBLE !

I can use a combined stage





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26/11/14 Donati: REALISTIC MODELS FOR DEVICE SIMULATION

DEVICES FOR RF APPLICATIONS

Not ∀ device is suited for RF applications so I've to select among a defined number of types. Moreover, for analog IC, I need, other than good active devices, also passive elements and not all technologies can provide good passive devices

For digital circuits, passive elements are not so crucial and the best technology for them is CMOS. Using the same for analog would be the best option for integration and get a true system on chip.

RF-CMOS processes are available and compatible with CMOS but can be used for "low" frequencies (hundreds of MHz)

Problem for Si devices is that power is small since small amount of currents flow on the devices; this limitation can be overcome by parallelizing devices (recently we got up to 1W power handling that is good for handsets applications for example)

Communication systems are dropping power of transmissions so CMOS could become a valid option for RF power applications in the near future. (Larger frequency, smaller antennas, smaller distance between Rx and Tx)

LDMOS

Nowadays, basestations, must handle hundreds of watts. This allow to work up to 5GHz, it is cheap but cannot be integrated with CMOS

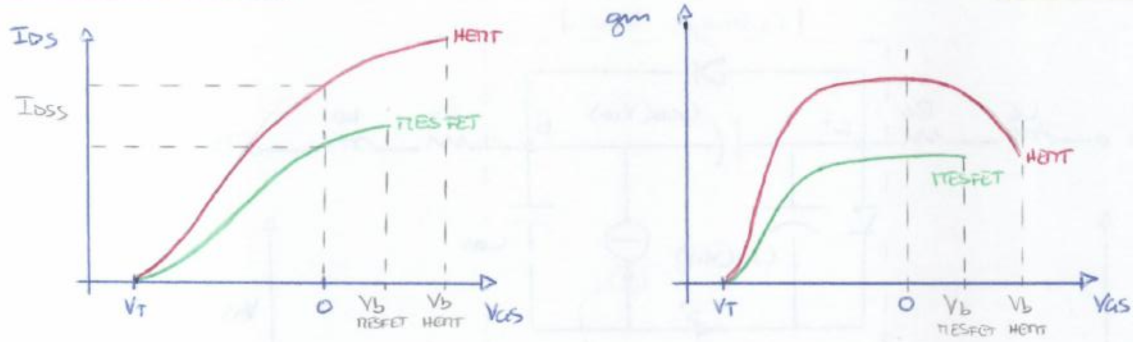
LDMOS is especially dedicated for extremely high power

However, with silicon, high power and high frequency handling cannot be reach simultaneously

BJT/HBT

Suited for low noise applications.

HEMT VS MESFET



HEMTs have ↑ Vbr and ↑ saturation value. !

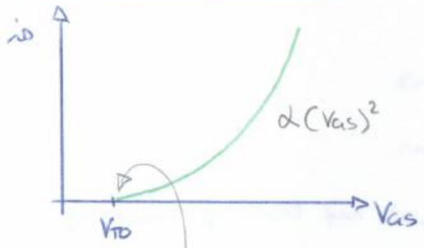
A good model is one that is C^∞ (continuous derivatives up to ∞ order) I need continuity of derivatives since I've to solve non-linear systems (many loops where I compute derivatives) and IMPs where I use taylor expansion up to 3rd 4th 5th order approximation

⇒ CIRCUIT ORIENTED MODELS

Good for circuit simulations but they lose the link with physical technological processes. However, physical models are too much complicated to be adopted in circuit simulators.

Si device circuits wants to keep the link with technology whereas, compound semiconductor circuits are more detached from the physics of semiconductors. Finally, the model must be FIT with measurements.

QUADRATIC MOS-LIKE TRANSCHARACTERISTIC



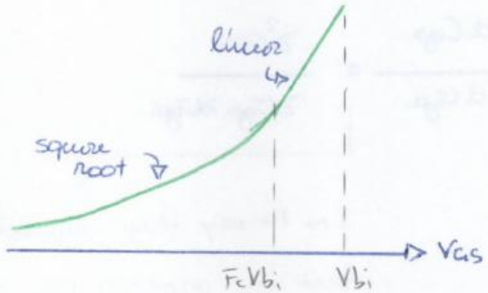
0 derivative -> smooth approach -> good

tanh is a transcendental function so it is a computer heavy computation; the model, then, is good for circuits with small number of elements

Capacitances:

$$C_{GS}(v_{GS}) = \begin{cases} \dfrac{C_{GS0}}{\sqrt{1 - \dfrac{v_{GS}}{V_{bi}}}} & v_{GS} < F_c V_{bi} \\[2ex] \dfrac{C_{GS0}}{\sqrt{1 - F_c}} \left[1 + \dfrac{v_{GS} - F_c V_{bi}}{2 V_{bi} (1 - F_c)} \right] & v_{GS} \gg F_c V_{bi} \end{cases}$$

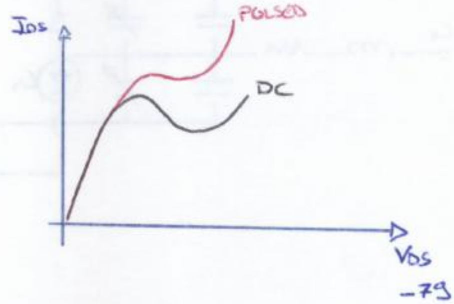
what is important to notice is that, close to V_{bi} I have an asymptot so I linearize in that region and solve the problem



All parameters are fitting parameters extracted based on measurements on a particular device

DRAWBACKS:

- DISPERSION effect is not considered
  performing static or pulsed V_{DS} stimulus measurements of the device current I get different results this is because (thermal effect for example) is a slow effect compared to the core of the model, this could be imputed to
  - -> THERMAL EFFECT
    ↑TEMP ↓MOBILITY -> LOWER CURRENT
  - -> CHARGE TRAPS



- DIODES

$$I_{gs} = I_S \left[\exp\left(\frac{v_{gs}}{NV_T}\right) - 1 \right]$$

however, I want to drive the diodes with current not with voltage so I put an external resistance (R_F) so:

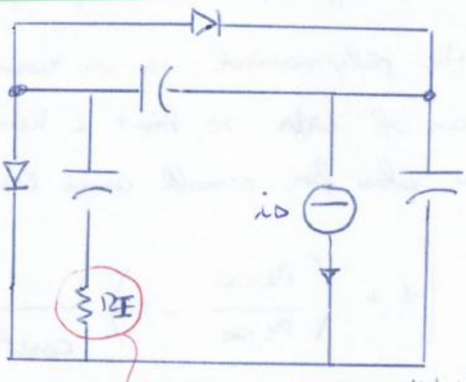
$$\begin{cases} I_{gs} = \dfrac{(v_{gs} - V_{bi})}{R_F} & V_{GS} > V_{bi} \\ I_{gs} = 0 & v_{gs} < V_{bi} \end{cases}$$

similarly for breakdown condition

$$I_{gs} = -I_R \left[\exp\left(\frac{v_{gs} - V_b}{\underbrace{V_{SR}}_{26mV}}\right) - 1 \right]$$

CUBIC CURTICE MODEL

quadratic transcharacteristic is too strict and so, current generator has been substituted with a cubic polynomial. The core of the model is then:



represents the distributed path that e^- must follow moving from source to drain.

$$i_D = \begin{cases} (A_0 + A_1 v_1 + A_2 v_1^2 + A_3 v_1^3)(1 + \lambda v_{DS}) \tanh(\alpha v_{ds}) & i_D > 0 \\ 0 & i_D < 0 \end{cases}$$

Another drawback: curtice model cannot be used for switching since cannot withstand polarity inversion of V_{DS}

OTHER MODELS: MATERKA YHLAND