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LOW-POWER HIGH-PRECISION CRYSTAL OSCILLATORS

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- Crystal resonator.
- General theory of crystal oscillators
  - splitting for analysis
  - oscillation: condition and frequency
  - amplitude limitation
  - start-up time.
- Theory of the 3-point oscillator
  - linear analysis with and without losses
  - nonlinear behaviour
  - amplitude and energy of oscillation
  - frequency stability, frequency tuning
  - phase noise
  - elimination of unwanted modes
  - loading by amplifier.
- Practical implementations
  - grounded drain oscillator
  - grounded source oscillators
  - amplitude regulator
  - circuit examples.

APPLICATIONS OF CRYSTAL OSCILLATORS

- Time-keeping (real-time clocks, RTC)
  - precision and stability: 10^{-6}
    ($\cong$ 30s/year)
  - low power
    watches: $\leq 0.5\ \mu W$

- Radio communication
  - precision and stability: 10^{-6} to 10^{-5}
    (1 to 10kHz at 1GHz)
  - low power («1mW)
  - phase noise if :
    - no VCO (direct RF generation)
    - injection synchronized VCO
    - wide-band PLL loop

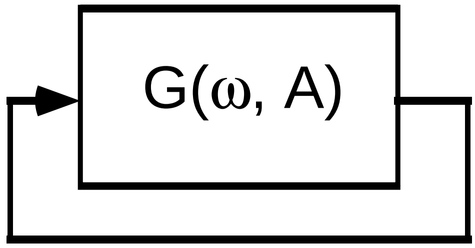
    otherwise negligible ($\sim 1/Q^2$)

- Clock of analog systems (filters)
  - precision and stability : 10^{-4}
    (better than component mismatch)

- Clock of digital systems
  - no high precision required
  - beware of overtone oscillation!

BASICS ON OSCILLATORS

- Frequency-dependent nonlinear loop.



ω = (angular) frequency
A = amplitude

- strongly nonlinear: relaxation oscillator
- weakly nonlinear: harmonic oscillator

- Stable oscillation at frequency ω_0: $\boxed{G(\omega_0, A_0) = 1}$

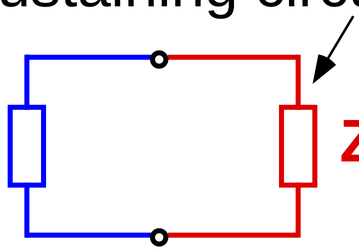
with: $\boxed{\left.\dfrac{d(\mathrm{Arg}(G))}{d\omega}\right|_{\substack{\omega=\omega_0 \\ A=A_0}} < 0}$ (phase stability)

and: $\boxed{\left.\dfrac{d|G|}{dA}\right|_{\substack{\omega=\omega_0 \\ A=A_0}} < 0}$ (amplitude stability)

Alternative representation:

- Resonator and sustaining circuit

complex resonator impedance $Z_r(\underline{\omega}, A)$

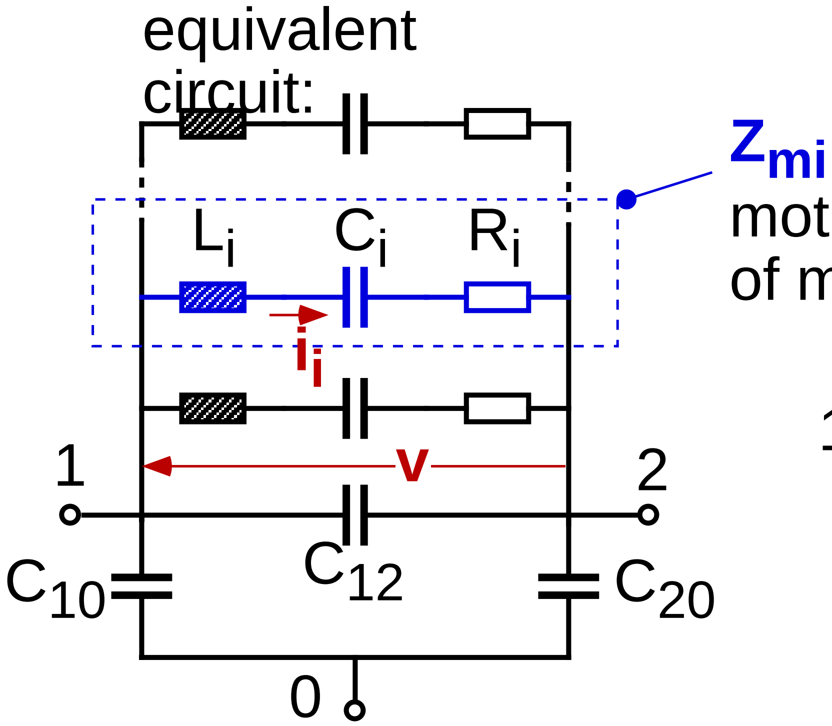


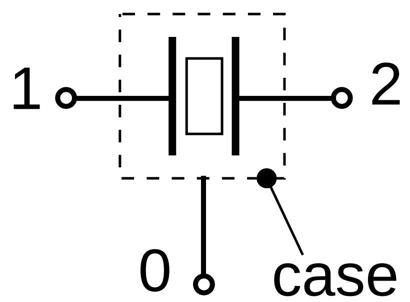
$Z_c(\omega, \underline{A})$ complex circuit impedance

- Stable oscillation at frequency ω_0:

$$\boxed{Z_r(\omega_0, A_0) + Z_c(\omega_0, A_0) = 0}$$

CRYSTAL RESONATOR





- Mechanical resonant frequency : $\omega_{mi} = \dfrac{1}{\sqrt{L_i C_i}}$
- Mechanical quality factor: $Q_i = \dfrac{1}{\omega_{mi} C_i R_i} \gg 1$
- i_i ~ velocity ~amplitude of mode *i*
- C_i ~ electromech. coupling of mode *i* , $C_i \ll C_{12}$
- ω_{mi} of different modes are not exact multiples of each other, and $Q_i \gg 1$; therefore:

even if v(t) is strongly distorted

all branches other than Z_{mi} are negligible

(for a single mode *i* of oscillation), and

the motional current **i_i is always sinusoidal**

MECHANICAL POWER AND ENERGY

- Mechanical energy of oscillation:

$$E_m = \frac{L\hat{I}^2}{2} = \frac{\hat{I}^2}{2\omega_m^2 C}$$

must be limited to avoid destruction, aging, and nonlinear effects.

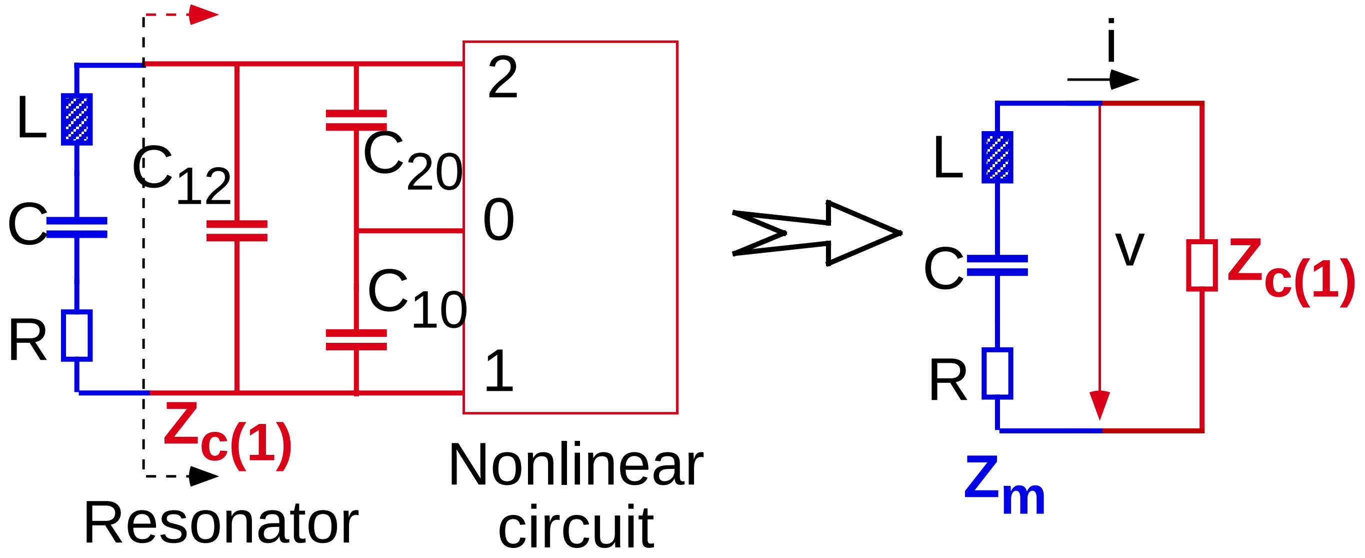
- Mechanical power dissipation:

$$P_m = \frac{R\hat{I}^2}{2} = \frac{\hat{I}^2}{2\omega_m Q C}$$

- Can be calculated as soon as the peak value $\hat{I}$ of sinusoidal current i(t) is known.

GENERAL FORM OF CRYSTAL OSCILLATOR

- The resonator is combined with an active circuit to sustain oscillation (compensate R)



- Frequency of oscillation ω slightly different of ω_m (effect of circuit)
- Frequency pulling $p = \frac{\omega - \omega_m}{\omega_m} \ll 1$
- The system must be conceptually split into:

Motional impedance Z_m $= R + j\frac{2p}{\omega C}$

(linear, strongly dependent on p)

Circuit impedance $Z_{c(1)}$, independent of p

Since no energy can be exchanged at harmonic frequencies (i sinusoidal), nonlinear effects are included by defining the circuit impedance at fundamental frequency:

$$\boxed{Z_{c(1)} = \frac{V_{(1)}}{I}}$$

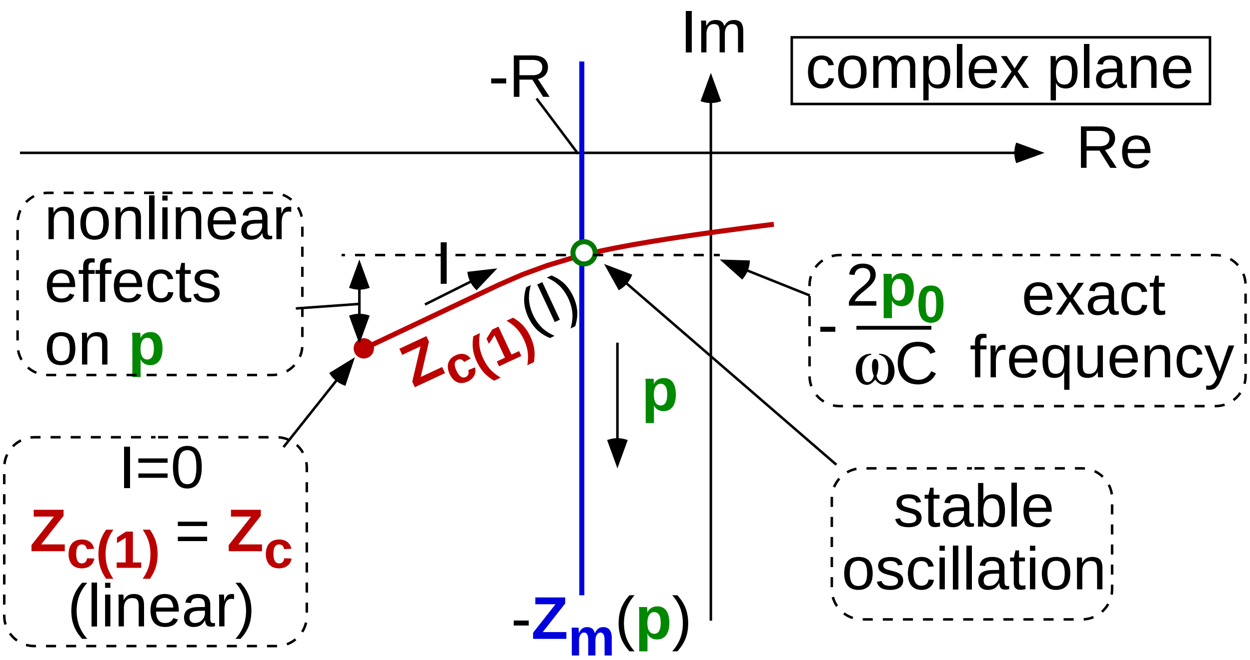
where $V_{(1)}$ is the complex value of the fundamental of v for complex value I of sinusoidal current i.

OSCILLATION

- Stable oscillation:

$$\boxed{Z_m + Z_{c(1)} = 0} \quad \begin{cases} \mathrm{Re}|Z_{c(1)}| = -R \\ \mathrm{Im}\,|Z_{c(1)}| = -\dfrac{2p_0}{\omega C} \end{cases}$$

where $Z_{c(1)} = \dfrac{V_{(1)}}{I}$ circuit imped. at fundam. frequency



- Growth of oscillation:

exponential, with time constant $\tau = \dfrac{2L}{-\mathrm{Re}|Z_{c(1)}| - R}$

- Critical condition for start-up of oscillation:

$$Z_m + Z_c = 0$$

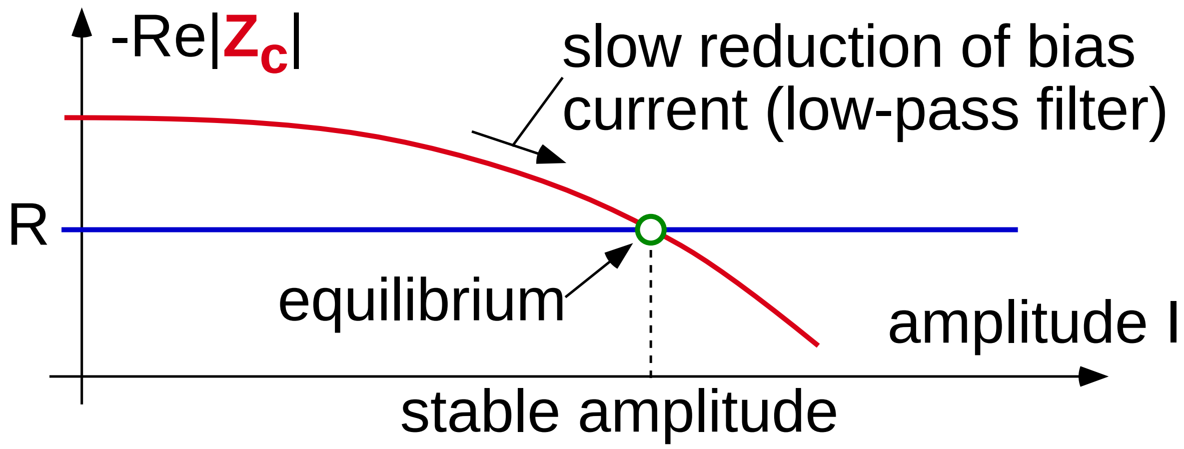
(linear circuit)

AMPLITUDE LIMITATION

- Necessary to define amplitude
  (an oscillator is always non linear)

- Instantaneous limitation by distortion:
  → creates harmonics
    → inter-modulation
      → addit. fundam. component of current
        → frequency change
          → poor stability
    → power dissipated in harmonics.

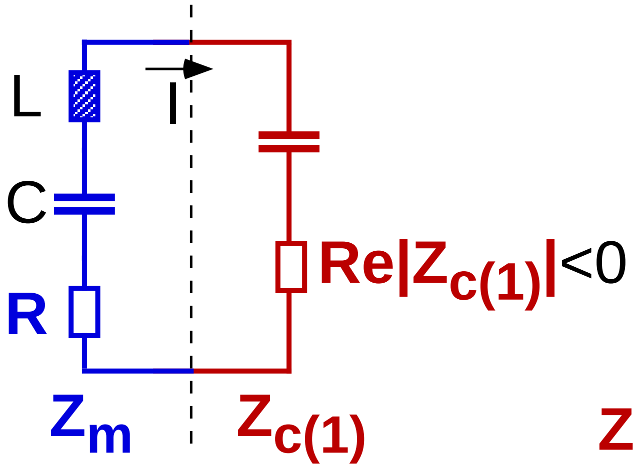
- Non-distorting amplitude limitation:
  - Z_c linear but slowly variable.

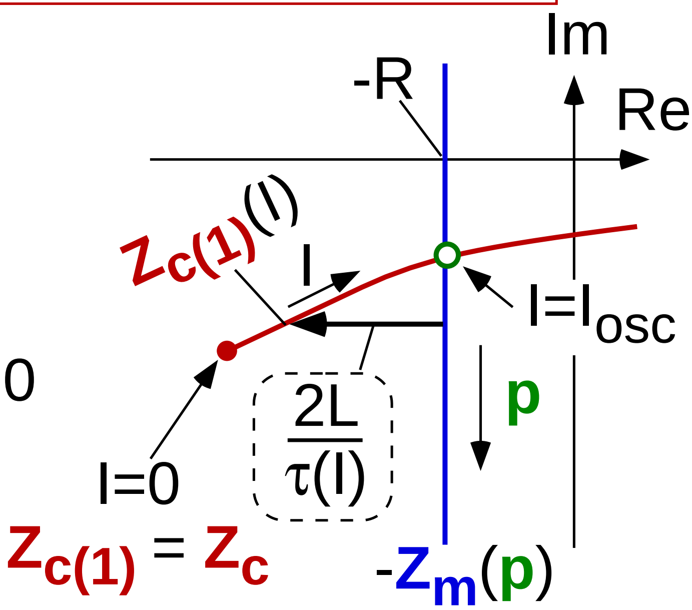


+ improved stability
+ reduced power dissipation in circuit
  (just above critical condition).

START-UP TIME OF OSCILLATOR

- Equivalent circuit:



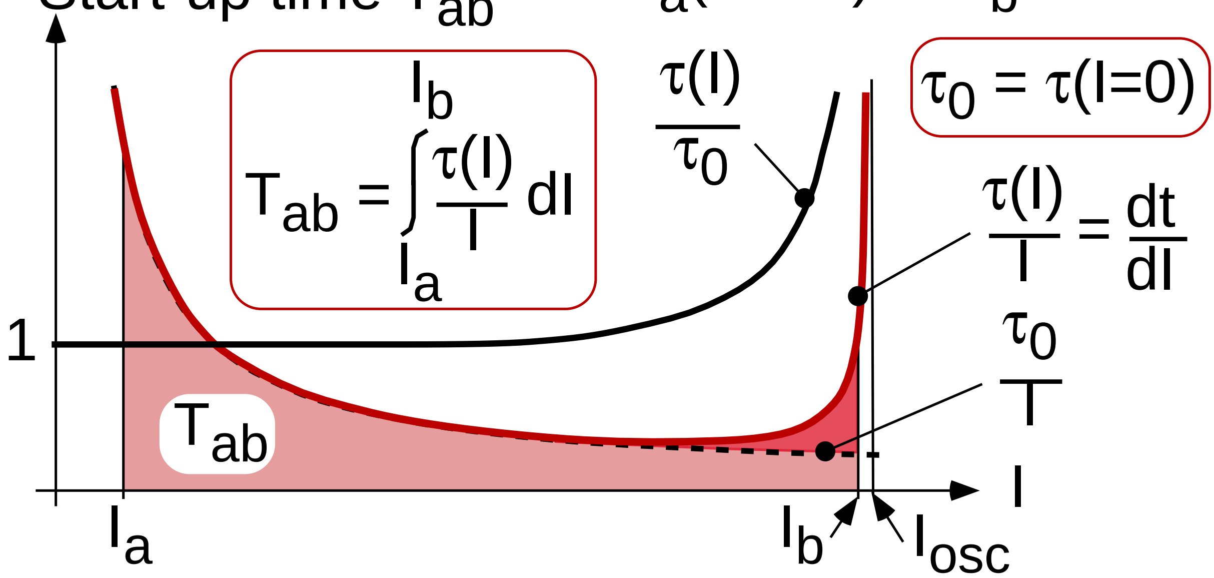


- Growth of oscillation:

$$I = I_0\, e^{t/\tau(I)} \qquad \text{with } \tau(I) = \frac{2L}{-\mathrm{Re}|Z_{c(1)}| - R}$$

$$\frac{dI}{dt} = \frac{I}{\tau(I)} \quad \longrightarrow \quad \boxed{\frac{dt}{dI} = \frac{\tau(I)}{I}}$$

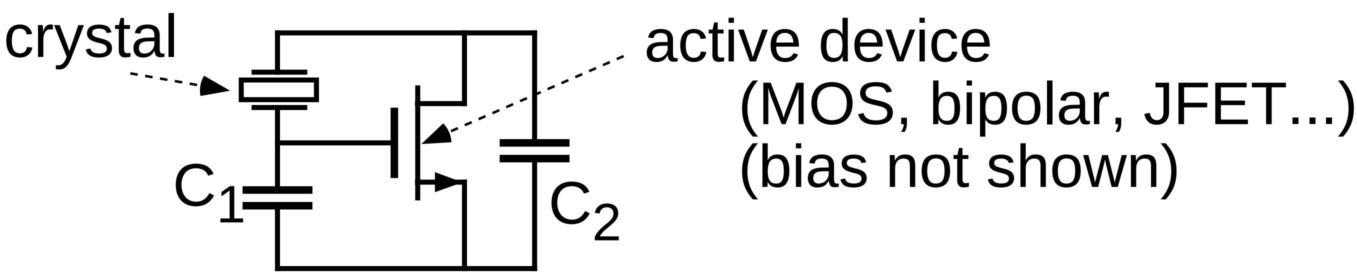
- Start-up time T_{ab} from I_a(noise) to I_b



- Approximation: $T_{ab} = \tau_0 \ln(I_b/I_a)$

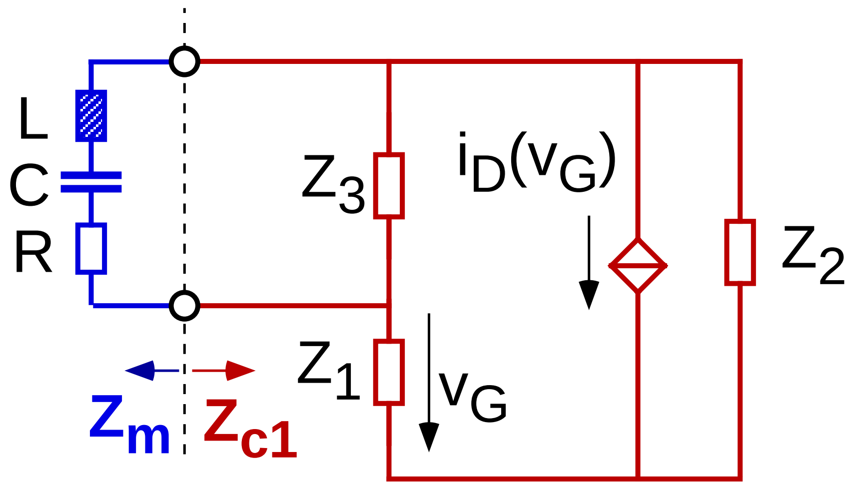
BASIC 3-POINT OSCILLATOR

- Only possibility to use a single active device when no inductance is available:



C_1,C_2 are necessary functional capacitors.

- General equivalent circuit:

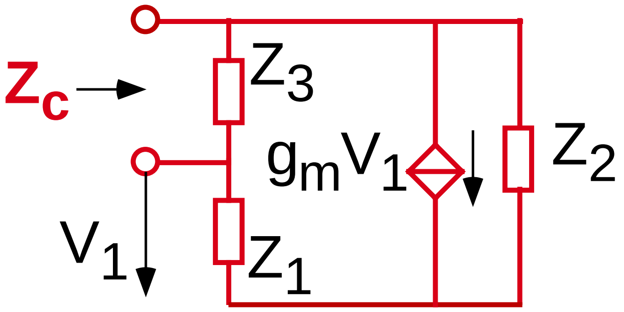


where $Z_i^{-1} = G_i + j\omega C_i$ (independent of **p**)

- include all circuit losses
- may be nonlinear.

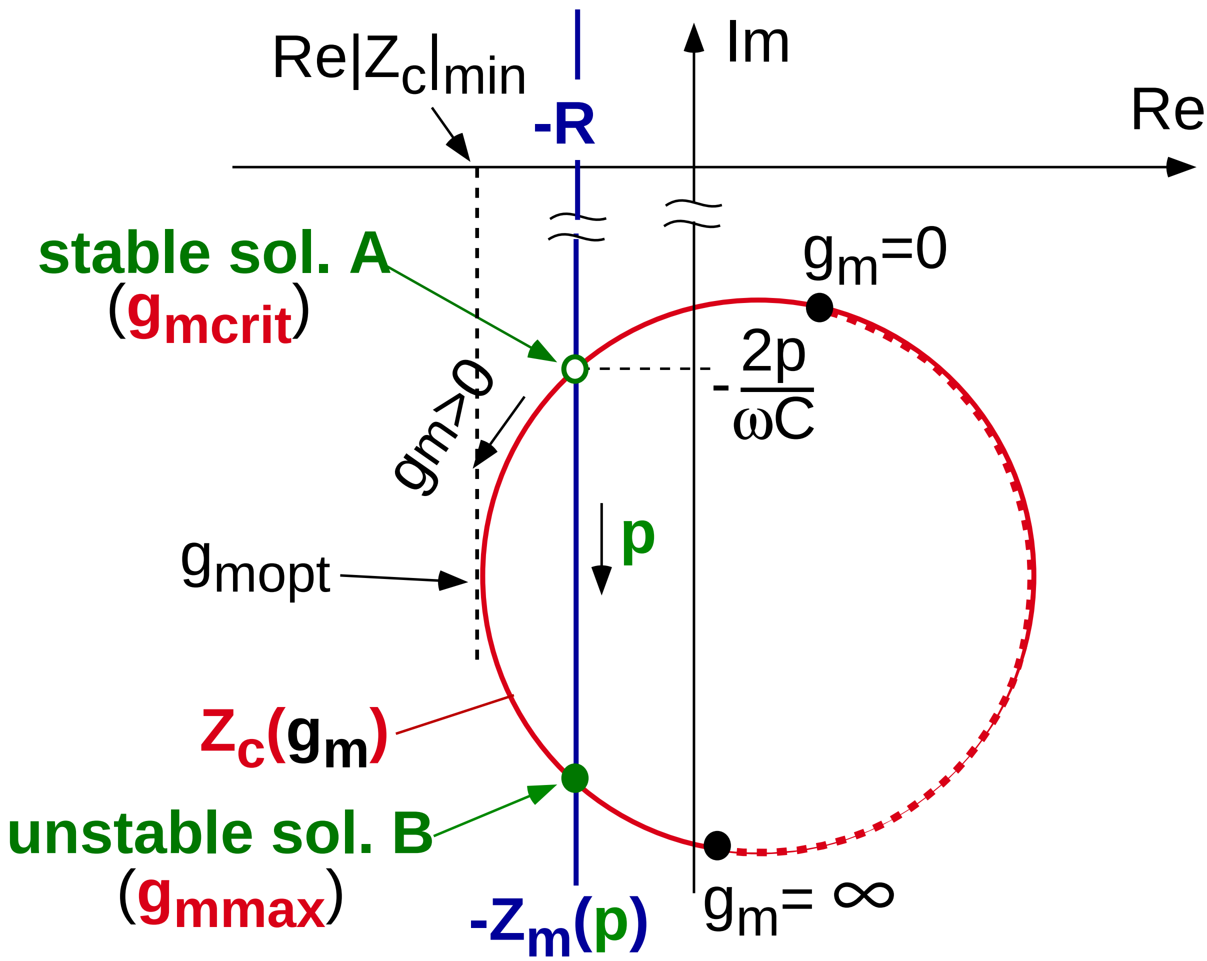
LINEAR ANALYSIS

- $Z_{c1} \rightarrow Z_c$ and $i_D(v_1) \rightarrow g_m V_1$
- General form of circuit (all losses included):



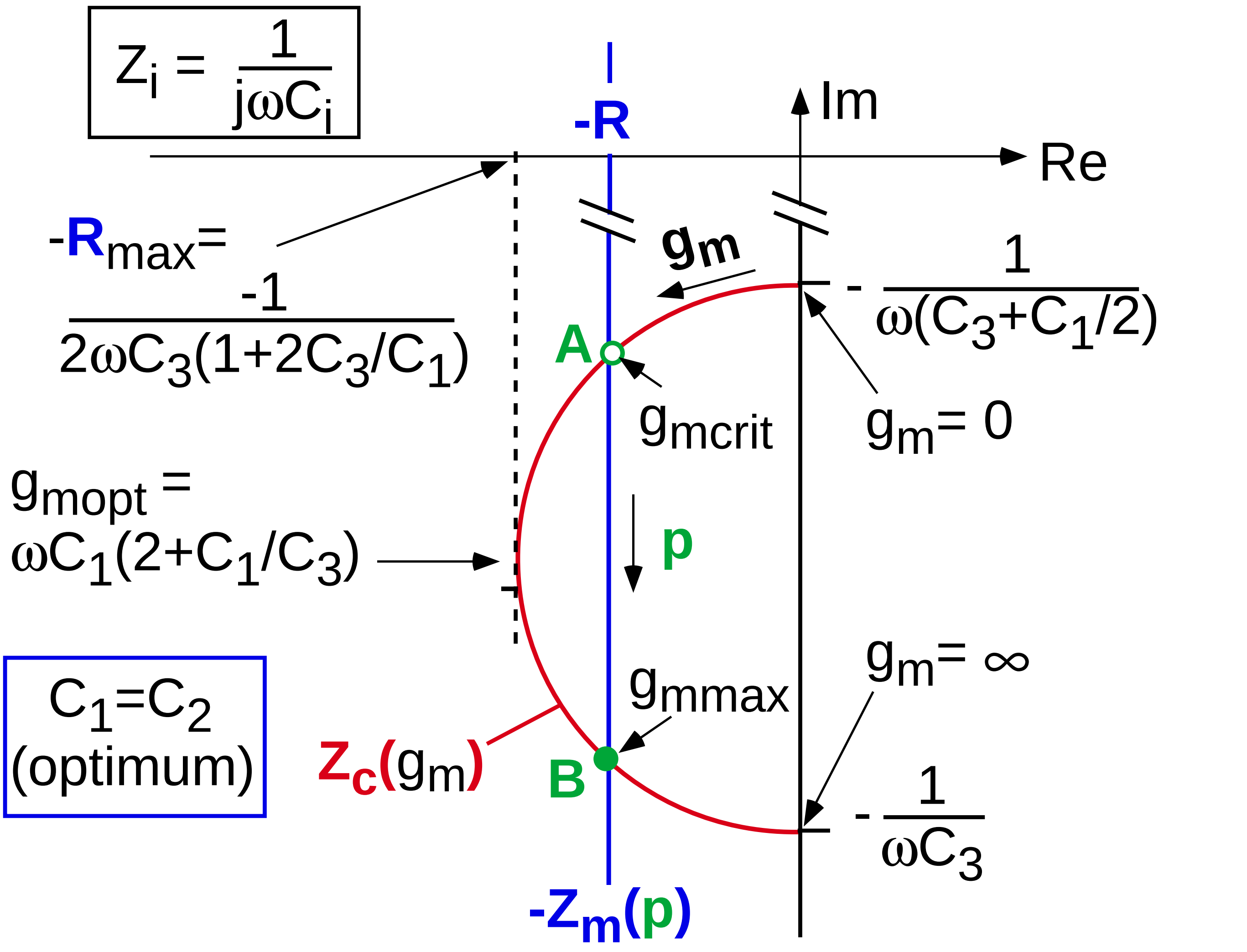
$$Z_c = \frac{Z_1Z_3 + Z_2Z_3 + g_mZ_1Z_2Z_3}{Z_1 + Z_2 + Z_3 + g_mZ_1Z_2}$$

bilinear function of g_m



- No oscillation if g_m too small or **too large**.

LOSSLESS LINEAR ANALYSIS



- Oscillation **only possible** if: $\dfrac{QC}{C_3} > 2(1+2C_3/C_1)$

If large margin to minimize d**p**/dR, then:

- $p_0 = \dfrac{\omega_0 - \omega_m}{\omega_m} \cong \dfrac{C}{2C_3+C_1}$
- $g_{mcrit} \cong \dfrac{\omega}{QC}(C_1+2C_3)^2 = \dfrac{\omega C}{Q p_0^2}$ **trade-off**
- $g_{mmax} \cong \omega C\, C_1^2 Q/C_3^2$

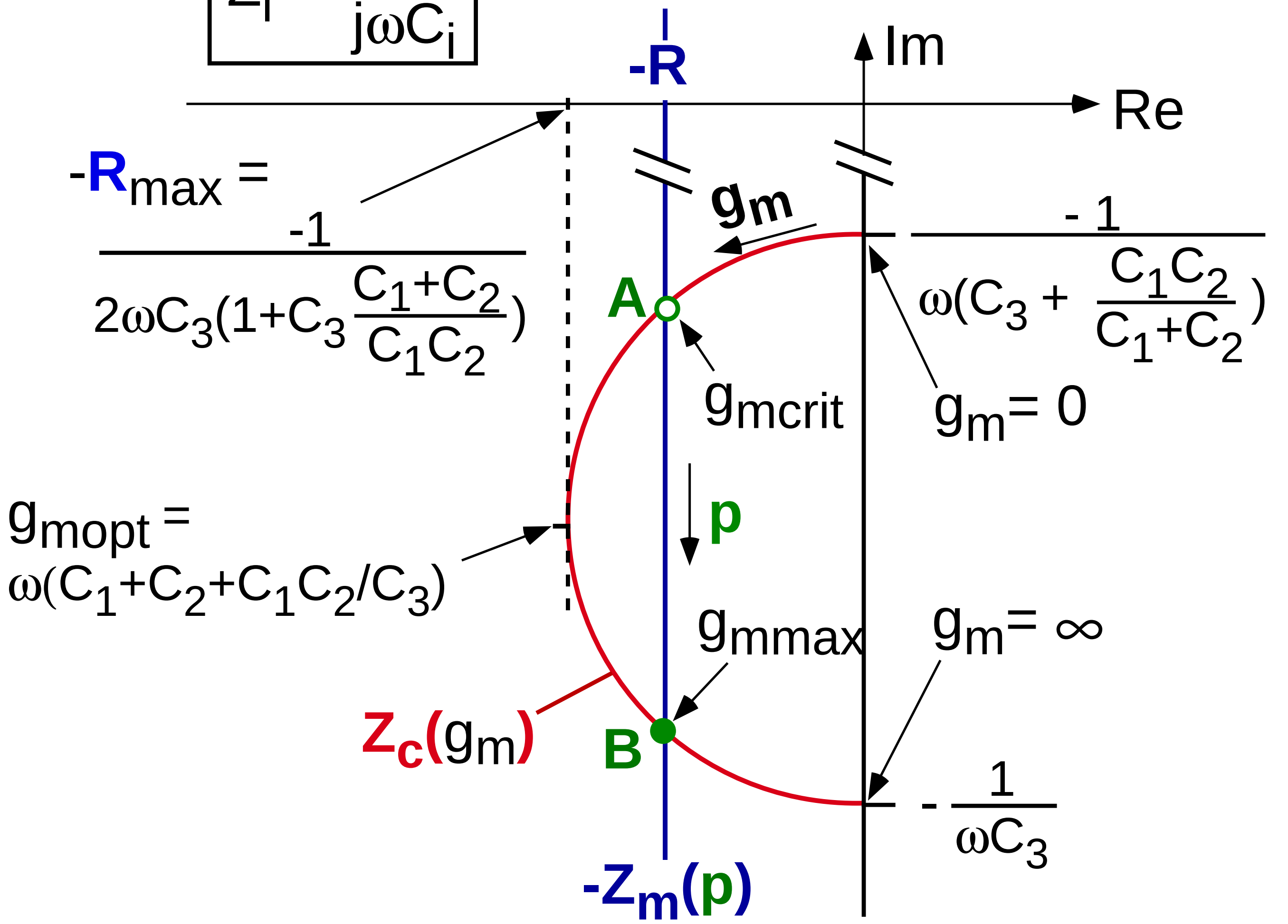
for $C_3 \ll C_1$

thus: $\dfrac{g_{mmax}}{g_{mcrit}} \cong \left[\dfrac{C\, C_1\, Q}{C_3(C_1+2C_3)}\right]^2 \cong \left[\dfrac{QC}{C_3}\right]^2$

LOSSLESS LINEAR ANALYSIS

OSC-8a

$$Z_i = \frac{1}{j\omega C_i}$$



- Oscillation **only possible** if: $\dfrac{QC}{C_3} > 2\left(1+C_3\dfrac{C_1+C_2}{C_1C_2}\right)$

If large margin to minimize d**p**/dR, then:

- $p_0 = \dfrac{\omega_0-\omega_m}{\omega_m} \cong \dfrac{C}{2\left(C_3+\dfrac{C_1C_2}{C_1+C_2}\right)}$

trade-off **min.** for $C_1=C_2$

- $g_{mcrit} \cong \dfrac{\omega}{QC}\dfrac{(C_1C_2+C_2C_3+C_3C_1)^2}{C_1C_2} = \dfrac{\omega C}{Qp_0^2}\dfrac{(C_1+C_2)^2}{4C_1C_2}$

- $g_{mmax} \cong \omega C\, C_1C_2Q/C_3^2$ for $C_3 \ll C_1$ and C_2

thus: $\dfrac{g_{mmax}}{g_{mcrit}} \cong \left[\dfrac{C\,C_1C_2\,Q}{C_3(C_1C_2+C_2C_3+C_3C_1)}\right]^2 \cong \left[\dfrac{QC}{C_3}\right]^2$

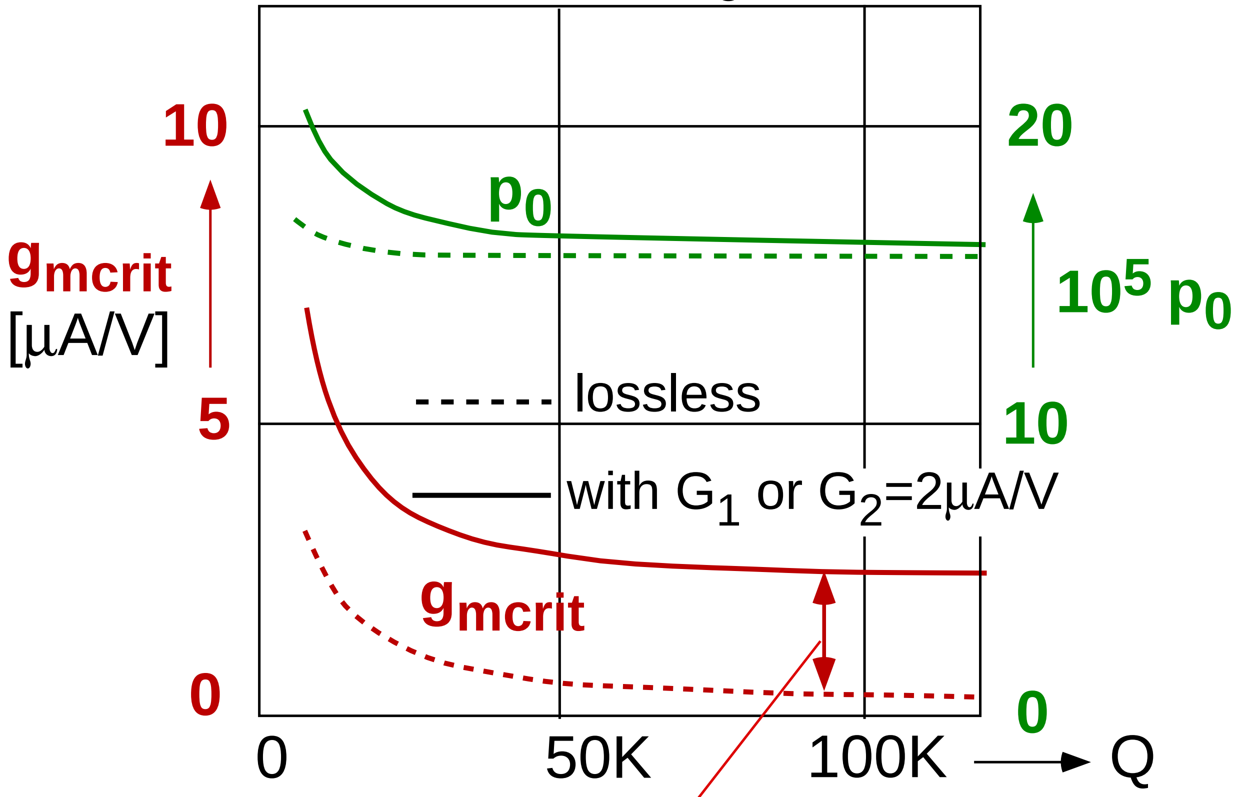
OSC-9

EFFECT OF LOSSES

- Example of linear analysis with:

$C = 3fF$ $\quad$ $C_1 = C_2 = 15pF$

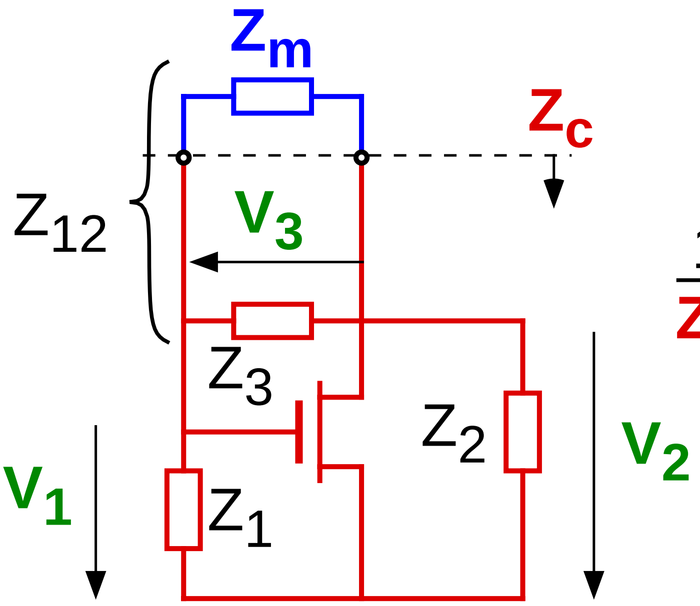
$f = 32KHz$ $\quad$ $C_3 = 2pF$



$$\Delta g_{mcrit} = \frac{C_2}{C_1} G_1 + \frac{C_1}{C_2} G_2$$

- Causes of losses:
  - biasing circuitry
  - loading by amplifier
  - series resistance of capacitors (HF)
  - output conductance of active device
  - input conductance of active device (bipolar)
  - external load (moisture)

RELATIVE VOLTAGE AMPLITUDES



$$\frac{1}{Z_c} = \frac{1}{Z_3} + \frac{1}{Z_1+Z_2+g_m Z_1 Z_2}$$

- Critical oscillation: $Z_m = -Z_c$

thus: $1/Z_{12} = 1/Z_m + 1/Z_3 = -1/Z_c + 1/Z_3$

yields: $Z_{12} = -(Z_1+Z_2+g_{mcrit}Z_1Z_2)$

$1+Z_{12}/Z_1 = \boxed{V_2/V_1 = -Z_2(1/Z_1 + g_{mcrit})}$

$$\boxed{V_3/V_1 = V_2/V_1 - 1}$$

- Losselss circuit ($Z_i = 1/j\omega C_i$) with $R \ll R_{max}$

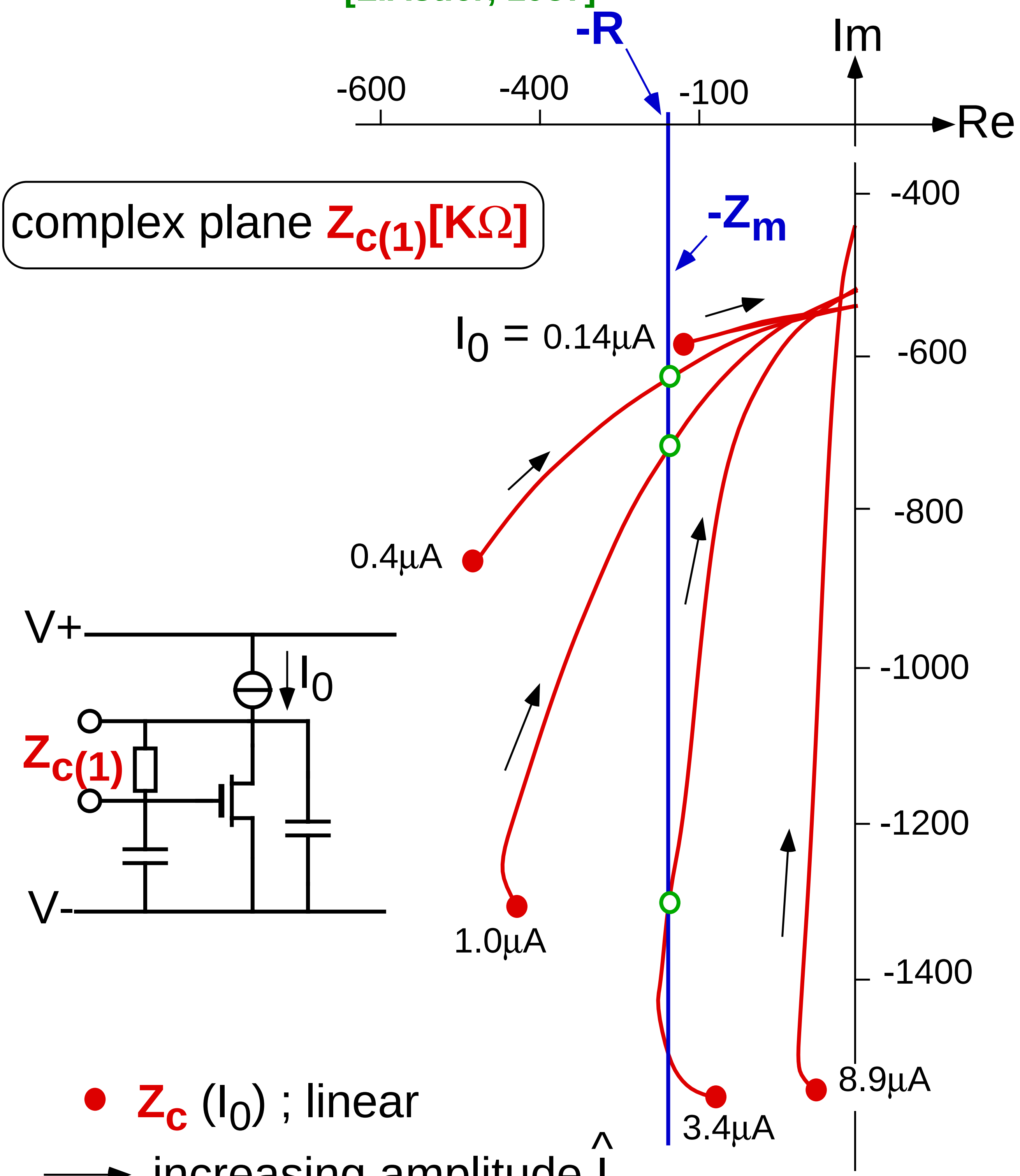
$$g_{mcrit} = \frac{\omega}{QC}\,\frac{(C_1C_2+C_2C_3+C_3C_1)^2}{C_1C_2}$$

yields: $$\boxed{\frac{V_2}{V_1} = -\frac{C_1}{C_2} + j\underbrace{\frac{C_1}{QC}\left(1+\frac{C_3}{C_1}+\frac{C_3}{C_2}\right)^2}_{\text{usually} < \text{ or } \ll 1}}$$

Then: $V_2/V_1 \cong -C_1/C_2$ and $V_3/V_1 \cong -(1+C_1/C_2)$

EXAMPLE OF NONLINEAR ANALYSIS

[L.Astier, 1987]



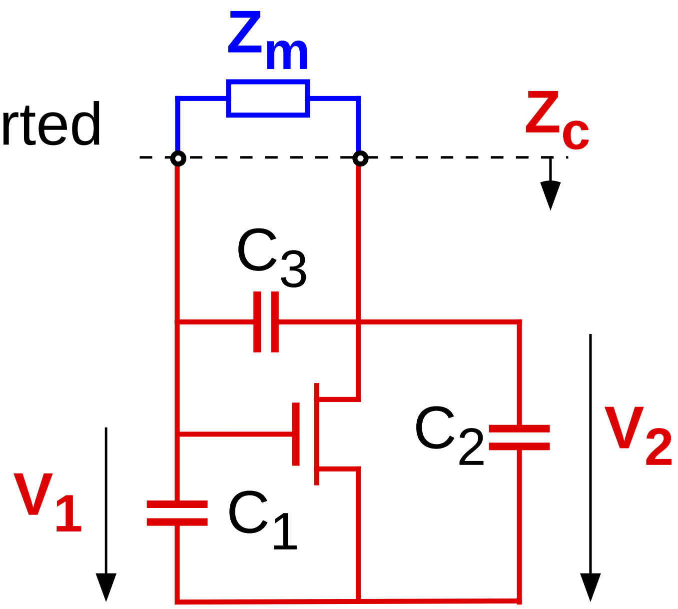
● Z_c (I_0) ; linear

→ increasing amplitude $\hat{I}$

o stable oscillation for particular value of **R**

OSC-10a

DISTORTION OF GATE SIGNAL



- Drain current is distorted

  ↓

  V_2 is distorted

- Drain to gate attenuation:
  - for fundamental frequency: $F = \left.\frac{V_1}{V_2}\right|_{fund.} \cong -\frac{C_2}{C_1}$

    (as shown before)
  - for harmonic components: $H = \left.\frac{V_1}{V_2}\right|_{harm.} \cong \frac{C_3}{C_1+C_3}$

    ($Z_m = \infty$)
  - relative attenuation $\left|\frac{H}{F}\right| = \frac{C_3 C_1}{(C_1+C_3)C_2}$

  usually « 1, thus V_1 approximately **sinusoidal**

- Effect of residual distortion of V_1:

  intermodulation of harmonics in transistor

  creation of out -of-phase fund. in drain current

  change in p, ⇒ frequency instability

- **C_3** must therefore be **minimized** (Z_3 large)

AMPLITUDE LIMITATION BY $i_D(v_G)$

- Small effect on **p** (none if $Z_3 = \infty$).
- Assumption: AC signal at gate sinusoidal:

$$v_G = V_0 + V_1 \sin\omega t$$

Results in: $i_D = f(v_G) = I_0 + I_1 \sin\omega t$ +harmonics

then: $I_1/V_1 = \mathbf{g_{m1}}(V_1) = g_{mcrit}$

transconductance for fundamental frequency

for **stable oscillation**

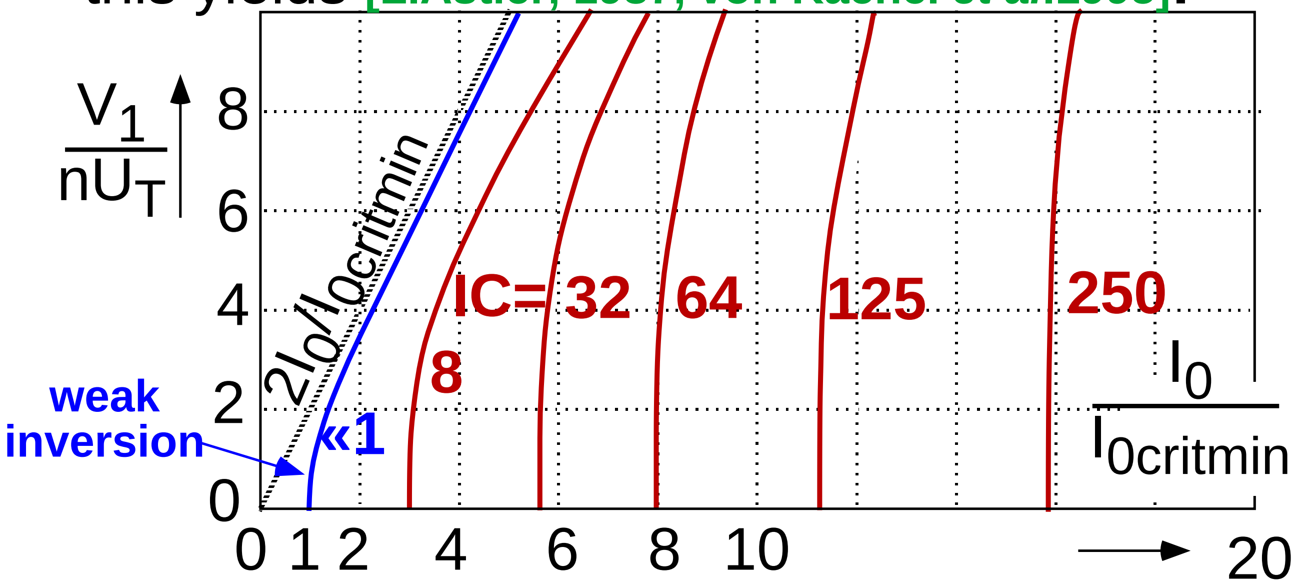
- Using continuous model for saturated MOS

$$\boxed{i_D = I_S \ln^2(1 + e^{v/2})}$$

[Vittoz, 1994]
[Enz *et al*., 1995]

where $v = (v_{GS} - V_T)/(nU_T)$ and $I_S = 2n\beta U_T^2$

this yields **[L.Astier, 1987, von Kaenel *et al*.1995]**:



V_1 = peak voltage amplitude at gate of transistor

I_0 = bias current of transistor

$I_{0critmin} = nU_T g_{mcrit}$

IC $= I_{0crit}/I_S$ = inversion coefficient at I_{0crit}

- $I_0(V_1)$ is minimum in weak inversion ($I_{0critmin}$).

AMPLITUDE OF OSCILLATION

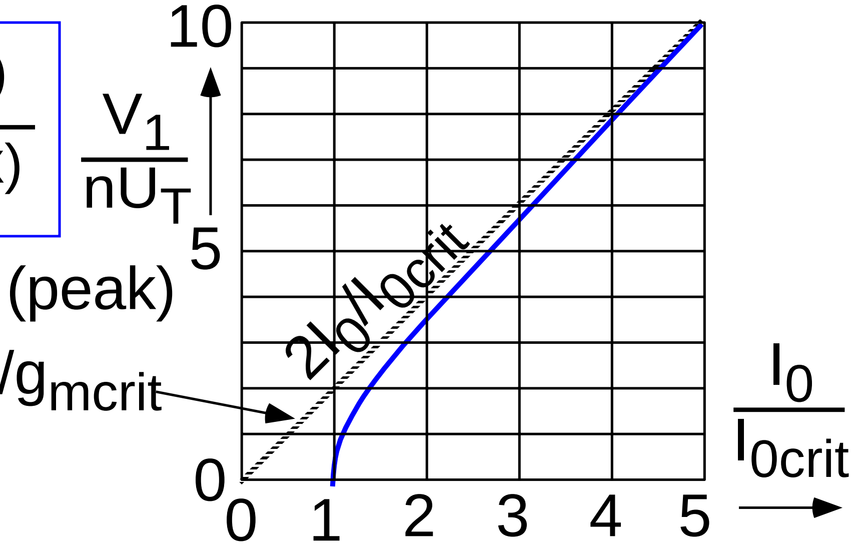
- Limitation by nonlinear $i_D(v_G)$ only
  $\rightarrow$ Very small effect on frequency pulling **p**
  (none if $Z_3 = \infty$).
- Goal: minimum current to produce g_{mcrit}:
  $\rightarrow$ transistor operated in **weak inversion**:
  $$i_D = A \exp(v_G/nU_T) \quad \text{(with n=1.4...1.6)}$$
- Assumption: AC signal at gate sinusoidal:
  $$v_G = V_0 + V_1 \sin(\omega t)$$
  results in: $i_D = I_0 + I_1 \sin(\omega t) + \text{harmonics}$

  where: $$I_1 = I_0 \frac{2I_{B1}(x)}{I_{B0}(x)} \quad \text{with} \quad \boxed{x = V_1/nU_T}$$

  and $I_{Bk}(x)$ are modified Bessel functions of order k
- Transonductance $\mathbf{g_{m1}}$ for the fundamental:
  $$\mathbf{g_{m1}} = I_1/V_1 \underset{\Uparrow}{=} g_{mcrit} = I_{0crit}/(nU_T)$$
  stable oscillation

Yields bias current I_0 as a function of amplitude V_1:

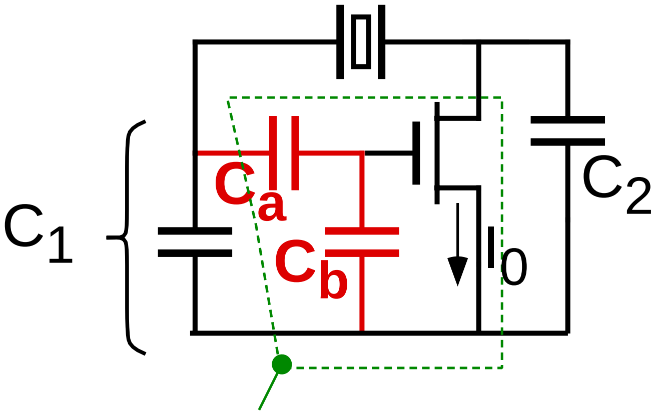
$$\boxed{\frac{I_0}{I_{0crit}} = x \frac{I_{B0}(x)}{2I_{B1}(x)}}$$



OSC-11b

LARGER AMPLITUDES

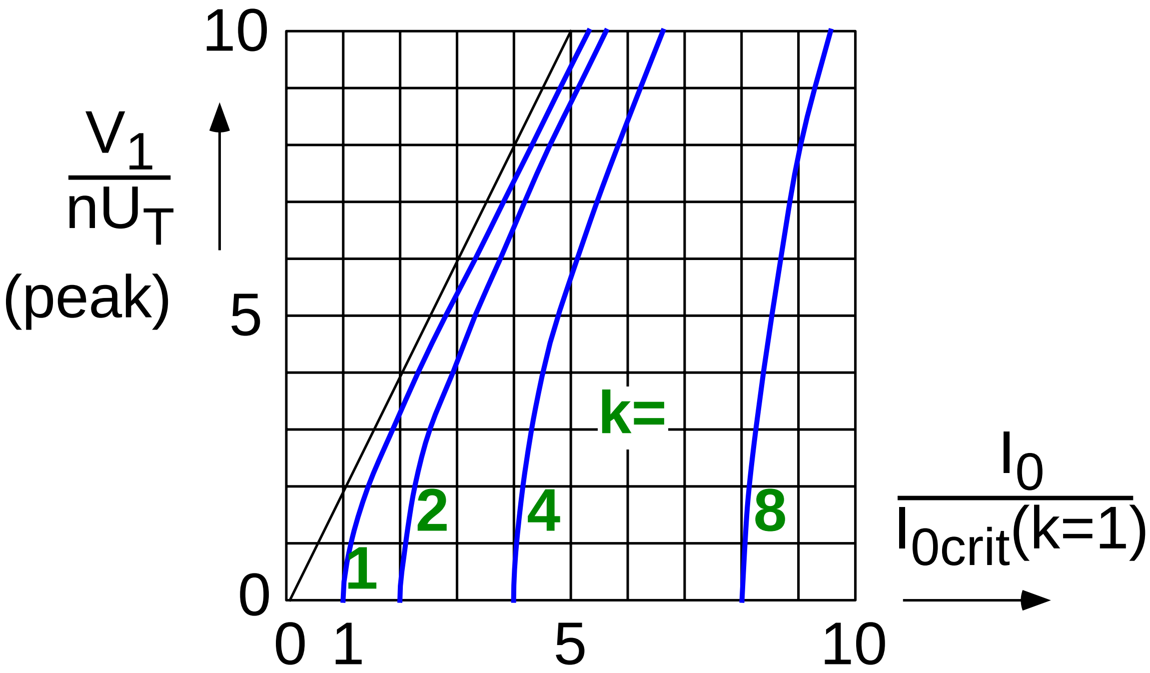
- Limited overdrive to avoid excessive distortion
- Use capacitive input attenuator C_a-C_b:



- Attenuation $1/k = \frac{C_a}{C_b + C_a}$
- C_b includes C_G
- In weak inversion:
  $C_G = C_{ox}(1-1/n)$

equivalent to transistor with $U_T \Rightarrow kU_T$

- Result for transistor in **weak** inversion:
  $g_{mequ.} = I_0/(knU_T)$, thus V_1 and I_0 amplified by **k**.



- Alternative solution: **strong** inversion
  - might be necessary for f»10MHz
    (large W/L ⟶ large C_1 and C_2)

AMPL. LIM. BY $i_D(v_G)$ IN STRONG INVERSION

- Assumption: AC signal at gate sinusoidal:

$$v_G = V_0 + V_1 \sin(\omega t)$$

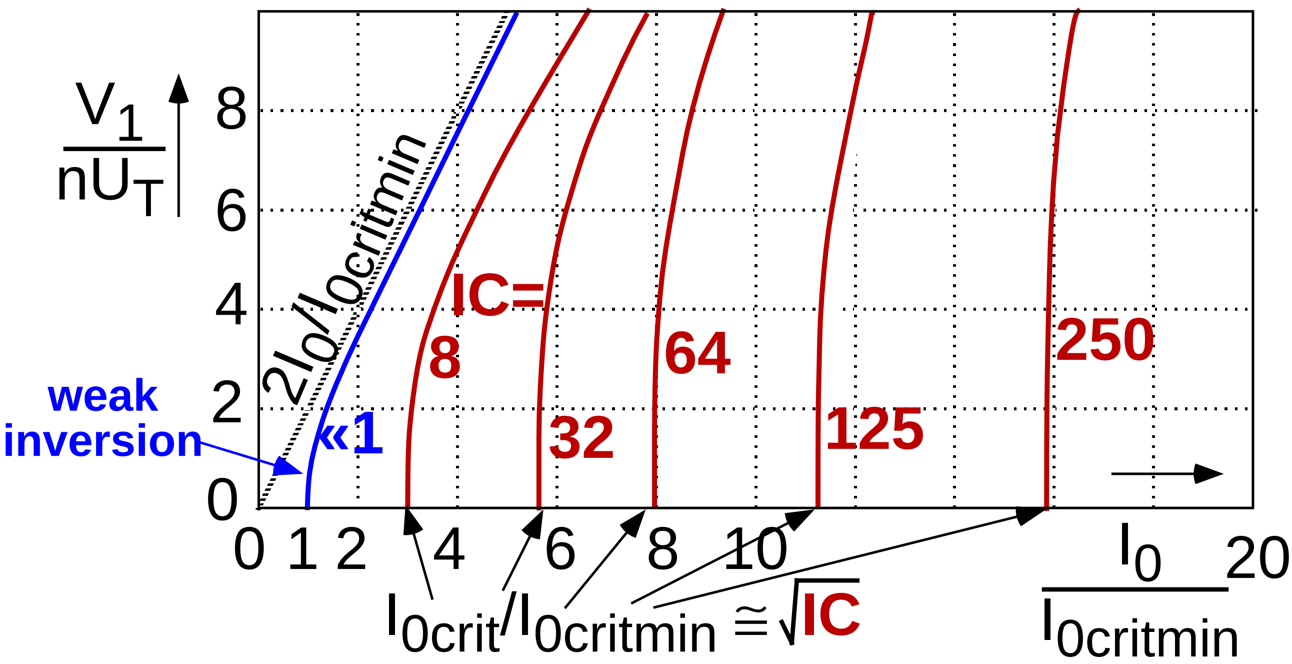
results in: $i_D = I_0 + I_1 \sin(\omega t)$ + harmonics

- $I_1/V_1 = \mathbf{g_{m1}}(V_1) = g_{mcrit}$ can be calculated numerically by using a continuous model for saturated MOS transistor :

$$\boxed{i_D = I_S \ln^2(1 + e^{v/2})}$$

where $v = (v_{GS} - V_T)/(nU_T)$ and $I_S = 2n\beta U_T^2$

this yields **[L.Astier, 1987]**:



V_1 = peak voltage amplitude at gate of transistor

I_0 = DC bias current of transistor

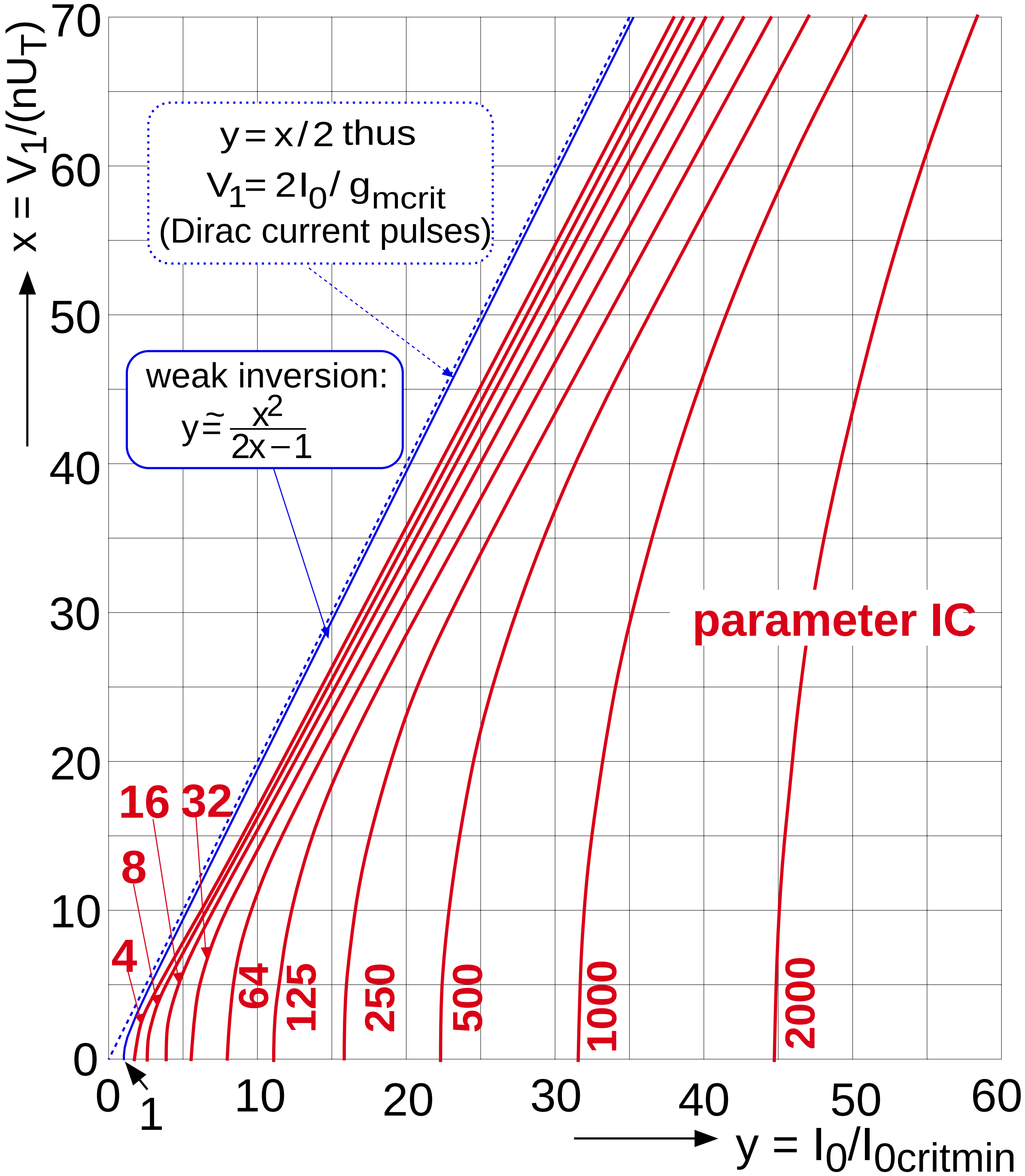
$I_{0critmin} = nU_T g_{mcrit}$ (weak inversion)

IC = I_{0crit}/I_S = inversion coefficient at I_{0crit}

CHART FOR LARGE AMPLITUDES

[L.Astier, 1987]

- Assumptions:
  - gate voltage sinusoidal (peak V_1)
  - transistor always saturated
  - constant mobility
  - constant Q, linear Z_1,Z_2 and Z_3
- Definitions:
  - $I_{0critmin}$=I_{0crit} in weak inversion
  - **IC** = I_{0crit}/I_S with $I_S = 2n\beta U_T^2$



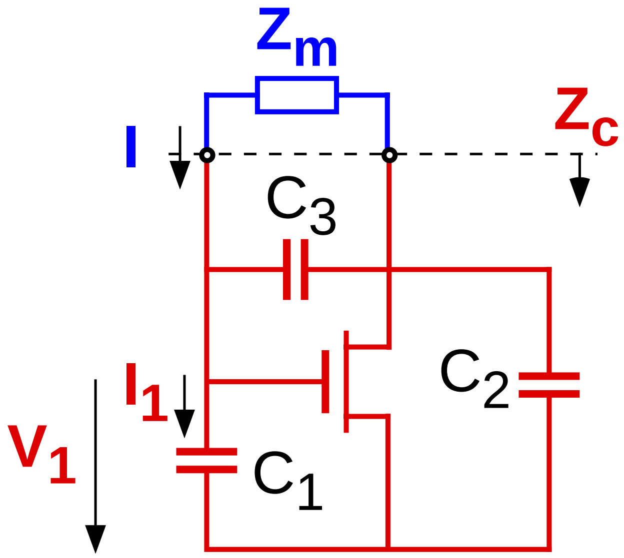
BASIC DESIGN PROCEDURE

	main criteria	ref.OSC
1. Select crystal resonator	frequency	
	temp. stability	
	cost, size	
2. Choose value of $C_1=C_2$	precision	12
	power	8a
	"circle"	8a
3. Calculate p_0 and $\omega_m=\omega_0(1-p_0)$		8a
4. Calculate g_{mcrit}		8a, 9
$I_{0critmin}$		11c
5. Fix ampl. of oscill. V_1	too small:	
	- phase noise	13b
	- amplification	14a
	- jitter of amplif.	
	too large:	
	- power	
	- aging	
6. Fix amount of overdrive	too small:	11d
	- large I_0/V_1	
	too large:	
	- poor stability	12
	- risk of overtone	14
7. Select $IC=I_{0crit}/I_S$	from 5. and 6.	11d
8. Calculate I_{0crit}, I_0, $I_S=2n\beta U_T^2$, β,W/L		11d
9. Calculate energy E_m and phase noise		13a,b
10. Return to 2, 5 or 6, or detailed design.		

ENERGY OF MECHANICAL OSCILLATION

$$I = I_1 \frac{Z_{C3}}{Z_m + Z_{C3}}$$

$$Z_m = R + j\frac{2p}{\omega C}$$



(I , I_1 and V_1 are complex **RMS** values)

Thus:

$$I = I_1 \frac{-j/\omega C_3}{R + 2pj/\omega C - j/\omega C_3}$$

Assumptions:

$2p \gg \omega CR = 1/Q$ — **negligible**

$p = \dfrac{C}{2(C_3 + C_1C_2/(C_1+C_2))} \ll \dfrac{C}{2C_3}$ — **dominates**

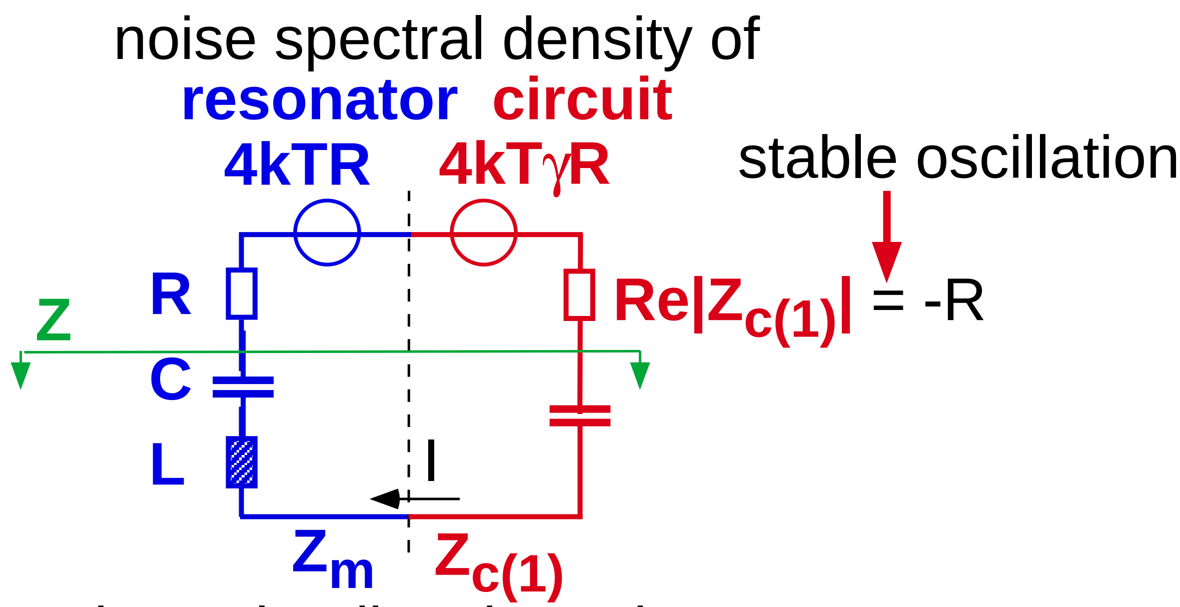
Then: $\boxed{I \cong I_1\left(1 + \frac{C_3}{C_s}\right)}$ where $C_s = \dfrac{C_1C_2}{C_1 + C_2}$

$\boxed{I \cong j\omega C_1 V_1\left(1 + \frac{C_3}{C_s}\right)}$ current through motional impedance Z_m

- Mechanical energy: $\boxed{E_m \cong \frac{|I|^2}{\omega^2 C} = \frac{C_1^2|V_1|^2}{C}\left(1 + \frac{C_3}{C_s}\right)^2}$

PHASE NOISE

- Simple model **[Leeson, 1966]** (linear, time invariant)
- Equivalent circuit at stable oscillation



- Impedance loading the noise sources:

$$Z = 2j\omega L\left(\frac{f-f_0}{f_0}\right) = 2jQR\left(\frac{f-f_0}{f_0}\right) \quad \text{for } |f-f_0| \ll f_0$$

where f_0 is the frequency of oscillation

- Noise current I_N circulating in the loop:

$$\frac{dI_N^2}{df} = \frac{4kT(1+\gamma)R}{|Z|^2} = \frac{(1+\gamma)kT}{Q^2R}\left(\frac{f_0}{f-f_0}\right)^2$$

- Phase noise spectral density:

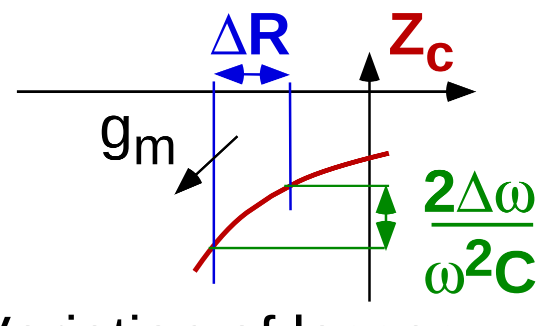
$$\frac{d\varphi_N^2}{df} = \frac{1}{2}\,\frac{dI_N^2/df}{I^2}$$ (half phase noise, half amplitude noise)

$$\boxed{\frac{d\varphi_N^2}{df} = \frac{(1+\gamma)kT}{2Q^2P_m}\left(\frac{f_0}{f-f_0}\right)^2 = \frac{(1+\gamma)kT}{2\omega QE_m}\left(\frac{f_0}{f-f_0}\right)^2}$$

- Nonlinear, time-variant:
  noise may added, including 1/f **[Hajimiri-Lee,1999]**

FREQUENCY INSTABILITY

Cause	**Remedy**
a. Crystal resonator	
• Aging	• Pre-aging.
• Temperature	• Better cut • analog or digital compensation.
b. Nonlinear effects in circuit (variations with V_B, V_T, T)	
• Nonlinear Z_1, Z_2 or Z_3	• Keep trans.in saturation • avoid C(V) effects.
• Nonlinear $I_D(V_G)$	• Reduce overdrive • stabilize amplitude • increase $\lvert Z_3 \rvert$.
c. Variation of linear effects	
• Variation of R~1/Q	• Increase Q • reduce losses in circuit. • increase $\frac{C_1 C_2}{C_3(C_1+C_2)}$
• Variation of losses	• Reduce losses
• Variation of C_1, C_2, C_3	• Decrease pulling p • avoid C(V) effects • stabilize V_B.



FREQUENCY TUNING

- On the resonator:

  precision limited to a few 10^{-5}.

- By C_1 and/or C_2 in the circuit:

  tuning range:

$$\frac{f_{max} - f_{min}}{f} = p_{max} - p_{min}$$

p_{max} → **C_1,C_2 small** — **large effect of circuit** → **degrades stability**

p_{min} → **g_m large** → **increases power consumption**

- Digital tuning:
  - adjust the ratio of subsequent counters
  - inhibit an adequate percentage of pulses

    requires a few bits of memory:

    pad bondings

    RAM

    E^2PROM.

ELIMINATION OF UNWANTED MODES

- A resonator has always several mech. modes (parallel series resonator in model).
- One mode is wanted (**WM**)
- All other modes are unwanted (**UM**).

$$g_{mcrit} = \sim \underbrace{\frac{\omega}{QC}}_{=\,\omega^2 R} \underbrace{(C_1+2C_3)^2}_{\text{same for all modes}} \quad (\text{for } C_2 = C_1)$$

activity different for each mode

a. Non-distorting amplitude limitation

- No interaction between modes; g_m decreases until the most active reaches critical amplitude.
- **WM** ensured if $\omega^2 R|_{\mathbf{WM}} < \omega^2 R|_{\mathbf{UM}}$

b. Distorting amplitude limitation

- Possible interaction between modes (very complicated).
- Safe solution: $g_{mcrit}|_{\mathbf{WM}} < g_m < g_{mcrit}|_{\mathbf{UM}}$

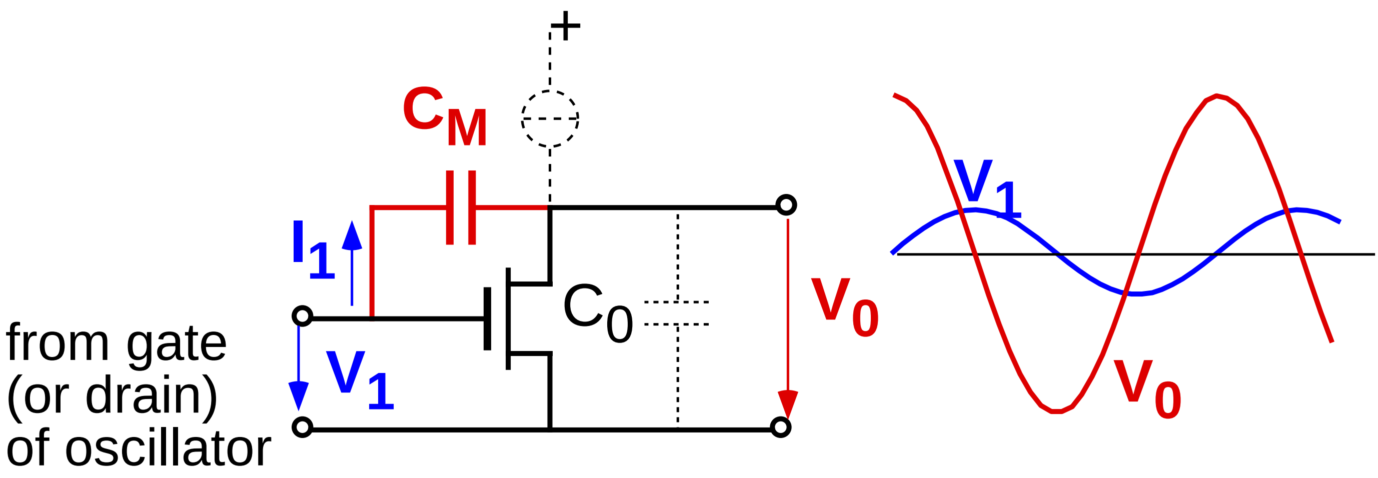
  requires $\omega^2 R|_{\mathbf{WM}} \ll \omega^2 R|_{\mathbf{UM}}$

c. Selection of **WM** of lesser activity

- frequency selective Z_c $\rightarrow$ degrades stability.
  - low- or high-pass for large difference in ω_m
  - LC bandpass to select a particular overtone

LOADING BY OUTPUT AMPLIFIER

- Elementary voltage amplifier, gain $A = |V_0|/|V_1|$



- Capacitive load: $V_0 = j\, A V_1$ (complex gain)

- Miller capacitance C_M, thus, for $A \gg 1$

$$I_1 = -j\omega C_M V_0 = \omega A C_M V_1, \text{ in phase with } V_1$$

- Input **conductance** $\boxed{G_1 = I_1/V_1 = \omega C_M A}$

  may be large if C_M and A large.

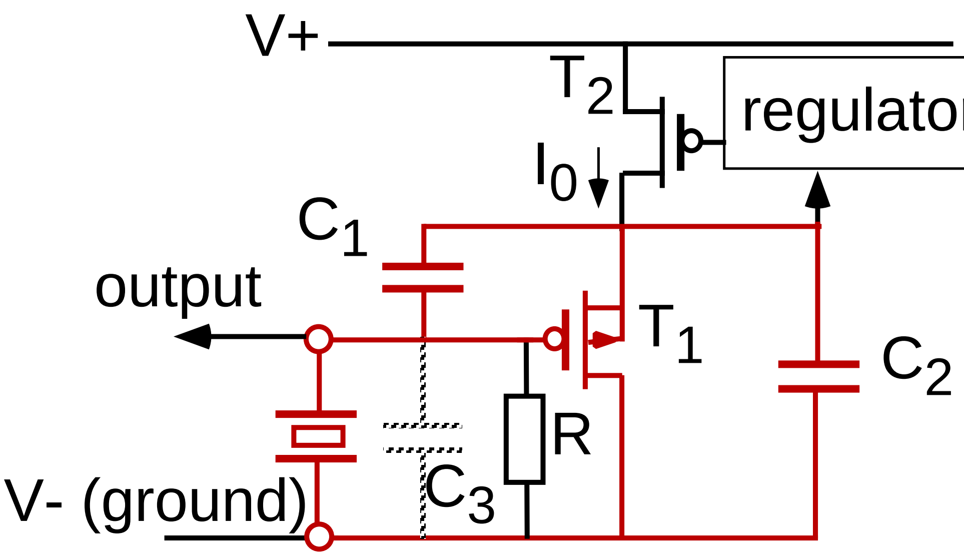
  ↓

  significant loss in oscillator

- Increasing V_1 requires more current in lossless oscillator, but reduces loss due to G_1: **trade-off**.

GROUNDED-DRAIN OSCILLATOR

[Luescher, 1968, Santos-Meyer, 1984]



T_1 active, biased by R and current source T_2 .

+ One single pin for resonator ("1-pin oscillator").
+ Doubled output amplitude.

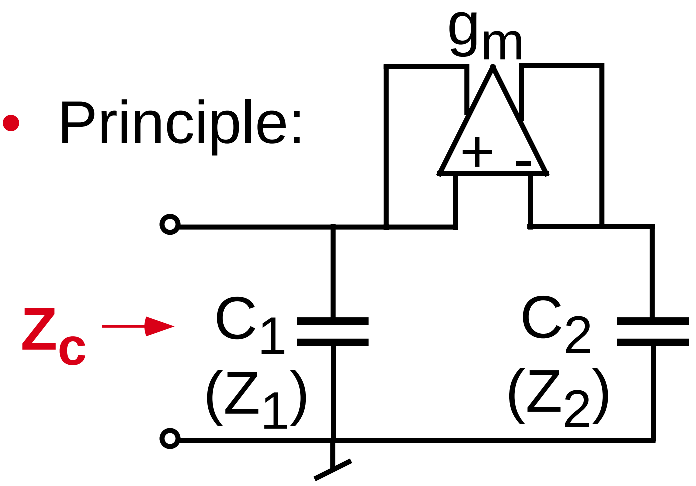
– Increased C_3 : decreases stability and/or increases power.

– T_1 must be put in a separate well connected to its source; otherwise an additional conductance $g_m(n-1)$ is added to G_2 (large increase of losses).

ONE-PIN OSCILLATOR WITH GROUNDED C's

[van den Homberg, 1998/99]

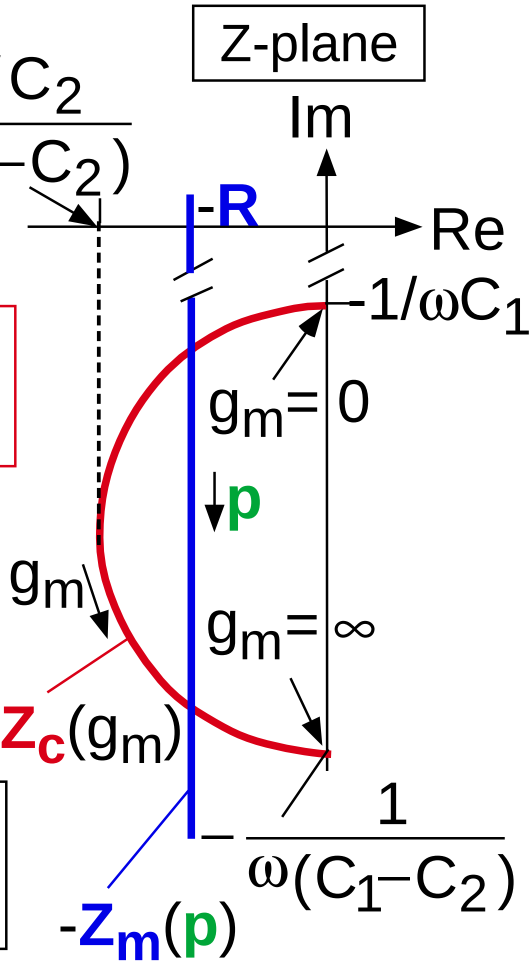
- Principle:



$$\boxed{Z_c = \frac{Z_1 + g_m Z_1 Z_2}{1 + g_m (Z_1 - Z_2)}}$$

bilinear function of g_m

- For $Z_i = \dfrac{1}{j\omega C_i}$:



- Necessary condition for oscill.:

$$R < \frac{C_1/C_2}{\omega(C_1-C_2)} \rightarrow \boxed{Q > \frac{C_1-C_2}{C \cdot C_1/C_2}}$$

- If realized with large margin:

$$\boxed{p_0 = \frac{C}{2C_1}}$$

$$\boxed{g_{mcrit} = \frac{\omega}{QC} C_1 C_2 = \frac{\omega C}{4Qp_0^2} \frac{C_2}{C_1}}$$

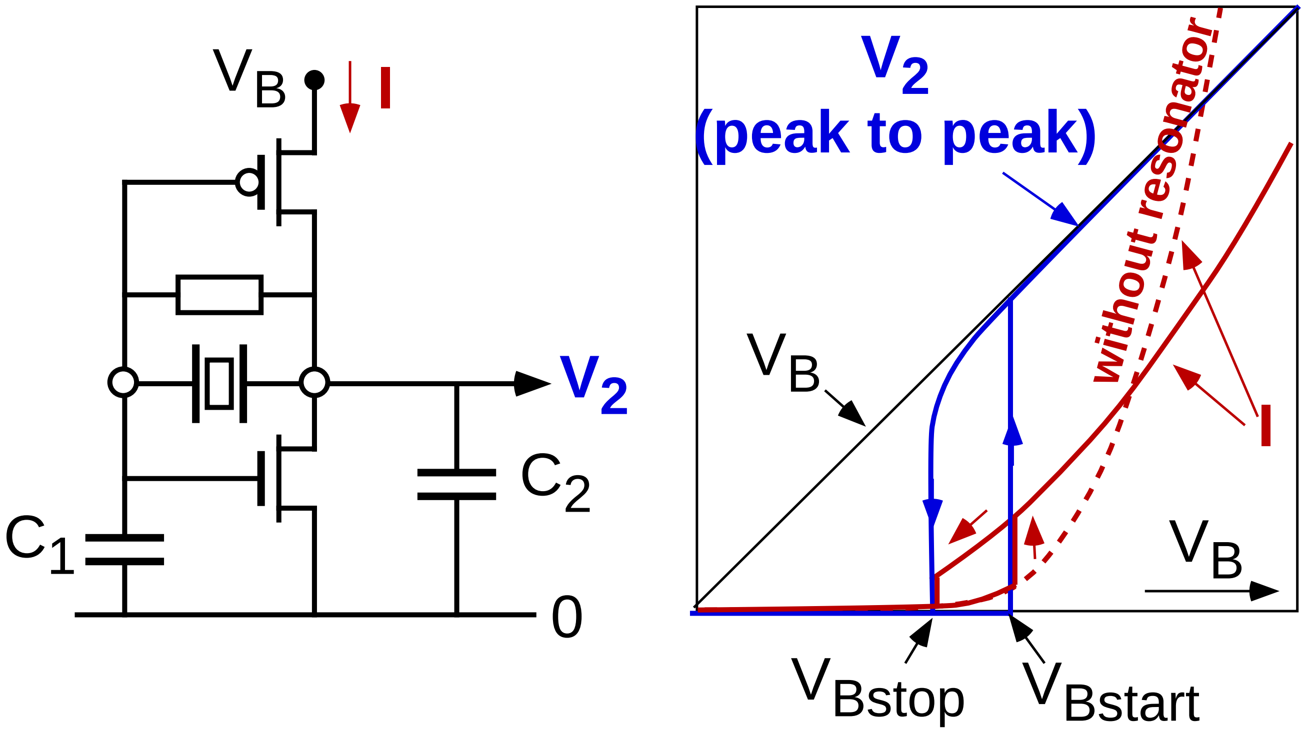
- Condition for stability:
  (pole of Zc with negative real part)

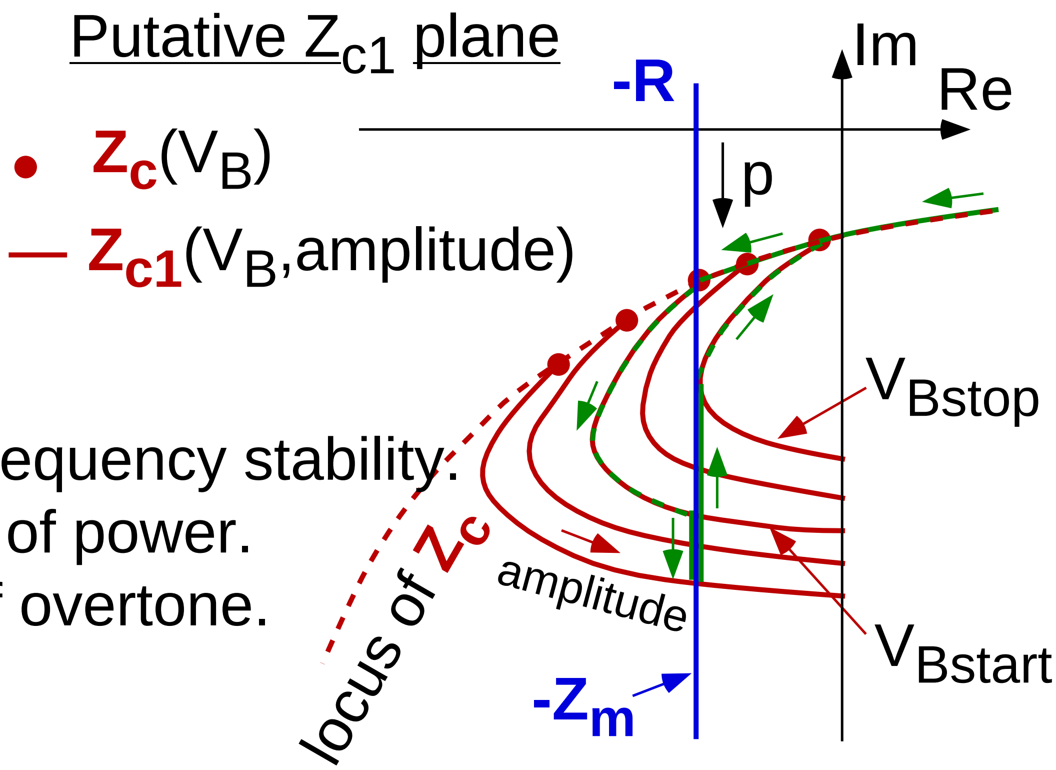
$\boxed{C_1 > C_2}$ radius of circle **reduced for** increased **margin**

OSC-16

CMOS-INVERTER OSCILLATOR

- a simple but poor solution -



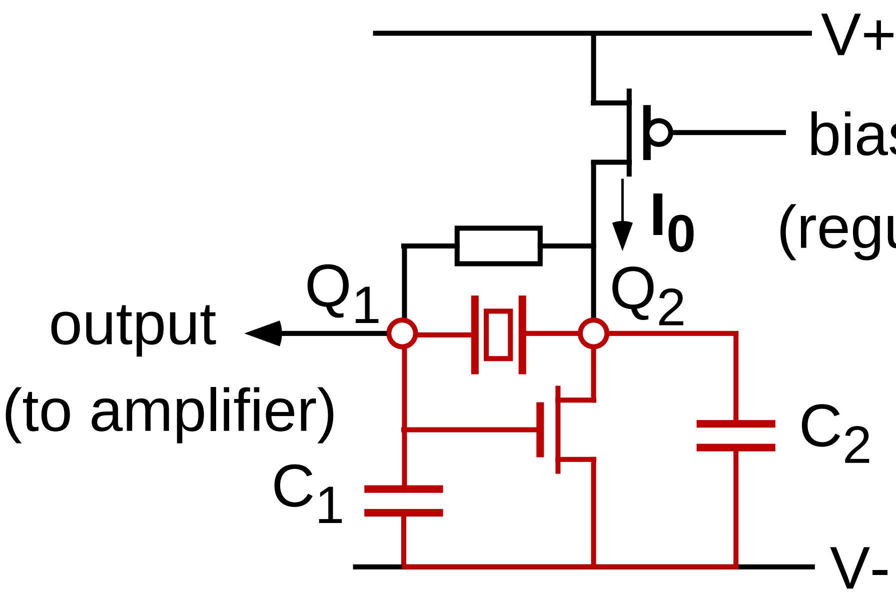


- Poor frequency stability.
- Waste of power.
- Risk of overtone.

- Possible improvement by resistors in the drains.

GROUNDED-SOURCE OSCILLATOR

(non-complementary)



- For fixed bias current I_0:

  Margin needed for variations of Q and process

  possible overdrive

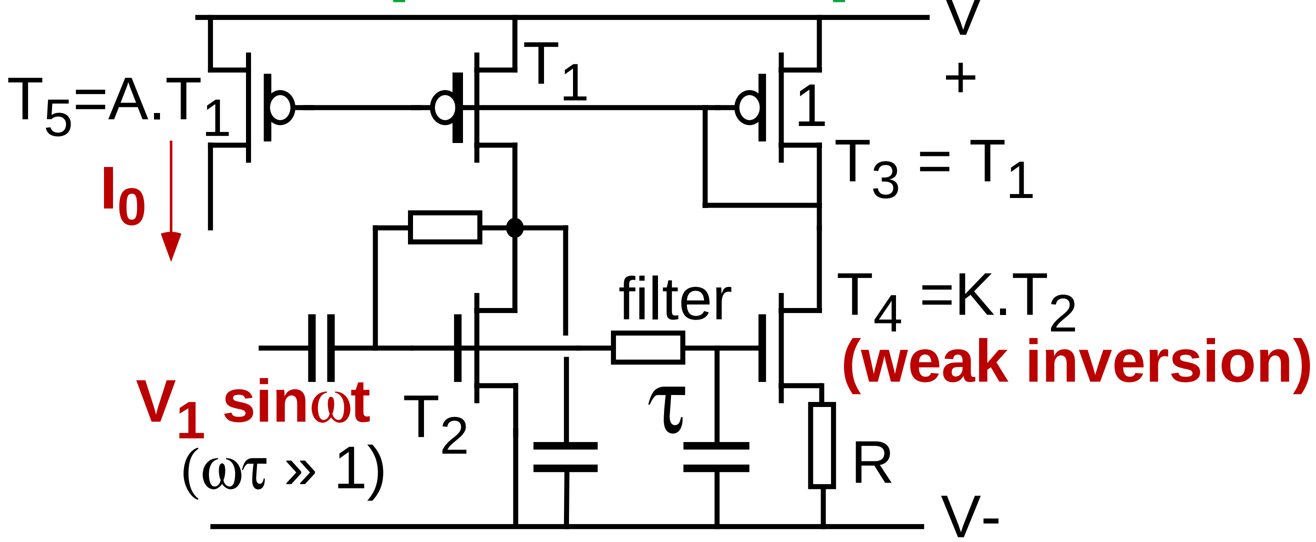
  waste of power   limitation by distortion

  Best:

- Low-level **amplitude regulation**+ output amplifier

AMPLITUDE REGULATOR

[Vittoz-Fellrath, 1977]



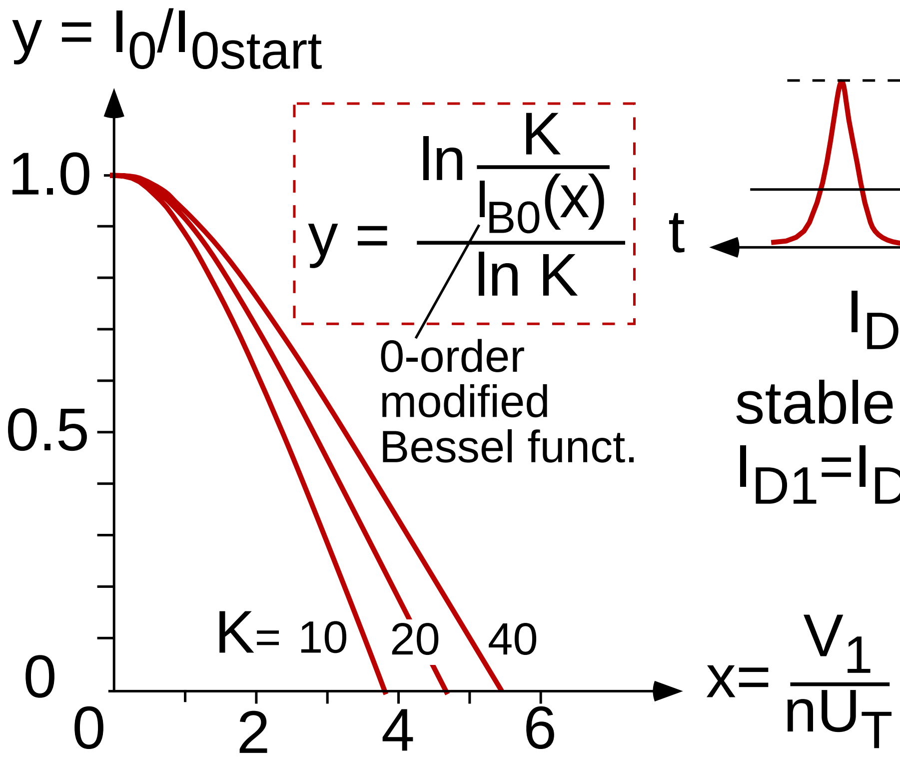
- No AC input voltage (V_1=0):

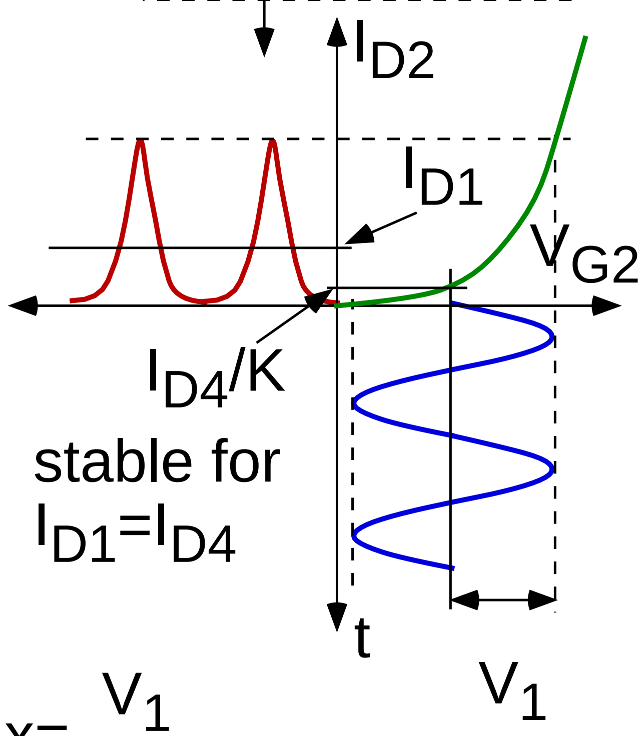
$$I_0 = I_{0start} = \frac{AU_T}{R}\ \ln K \quad \text{(start-up current)}$$

- For V_1>0:

General solution

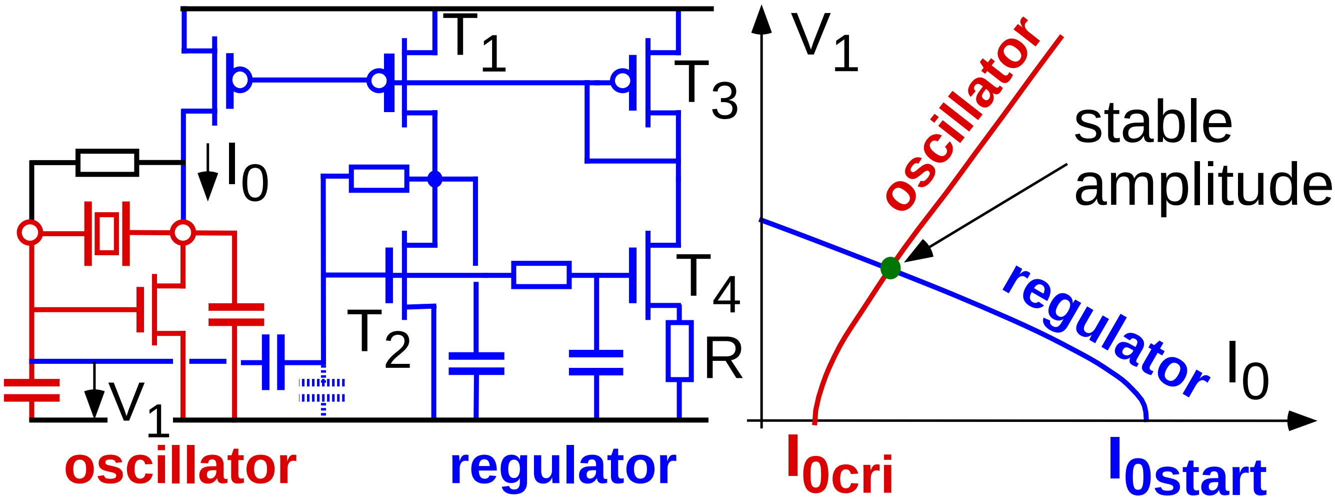
For R negligible



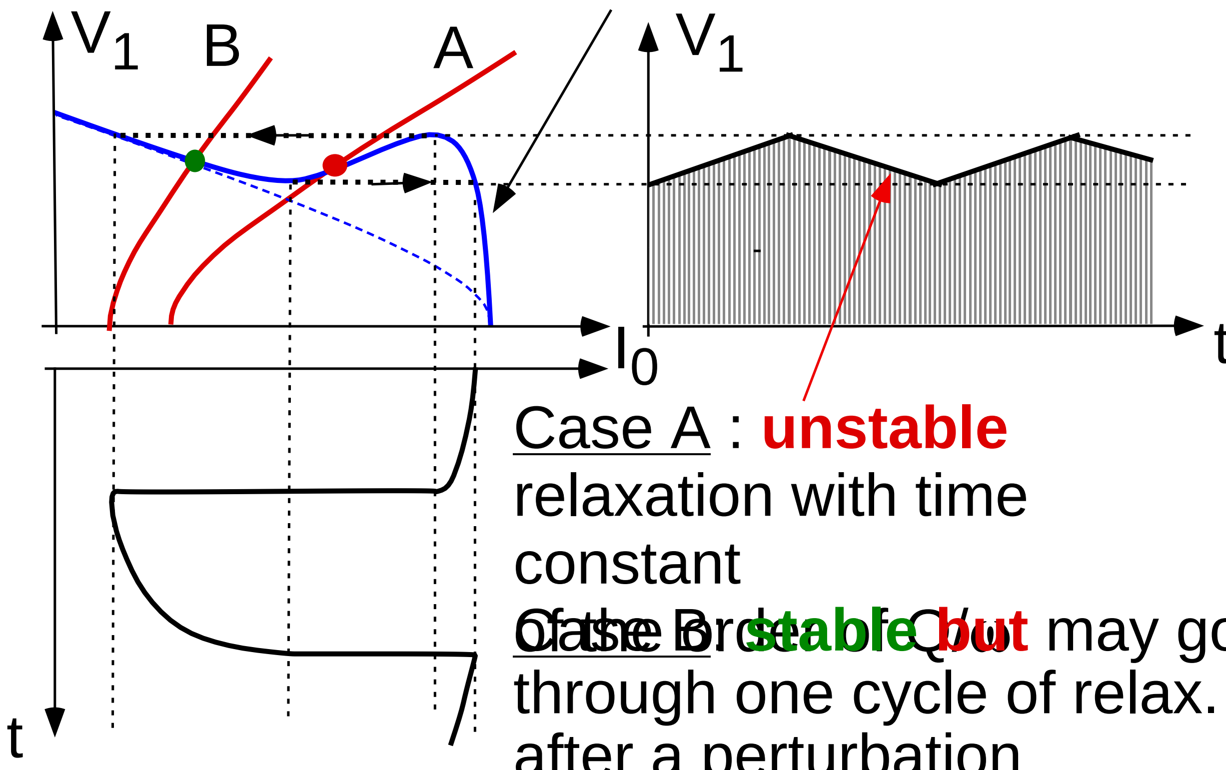


OSC-18a

AMPLITUDE REGULATING LOOP



- Effect of T_2-T_4 entering strong inversion: distortion of the regulator's characteristics



Case A : **unstable**
relaxation with time constant

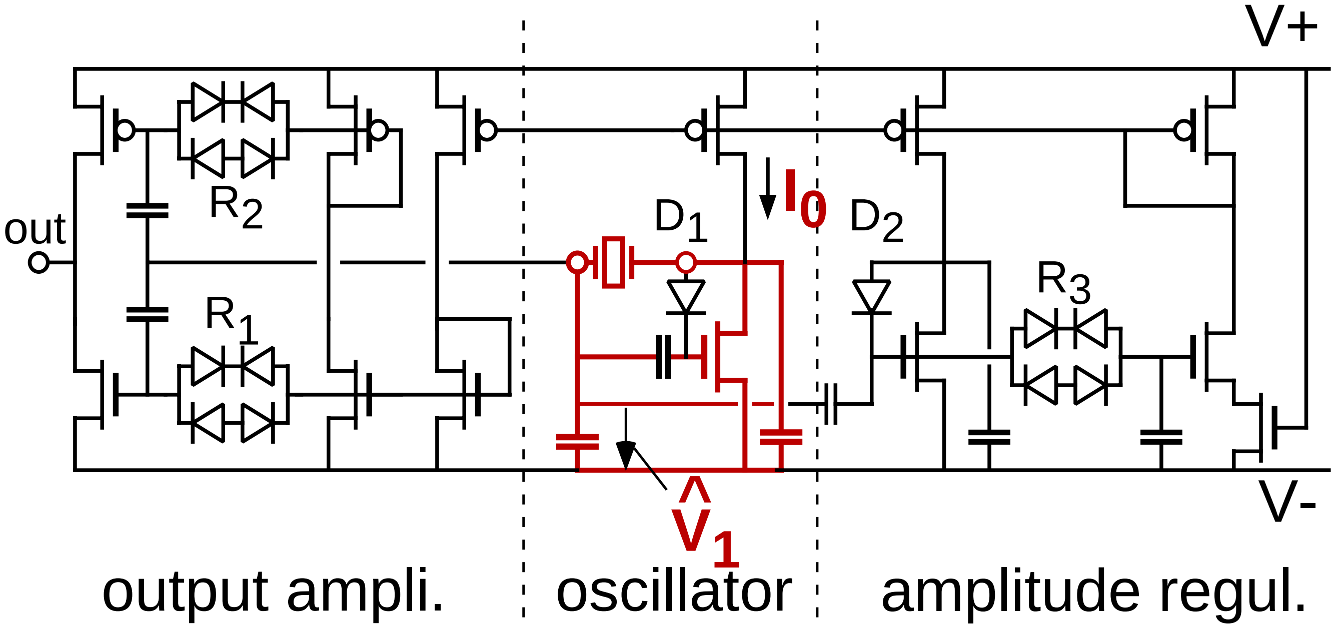
Case B : **stable but** may go through one cycle of relax. after a perturbation

- Design criterion: $i_{D2max} < I_{S2} = 2n\beta_2 U_T^2$ (too strict)

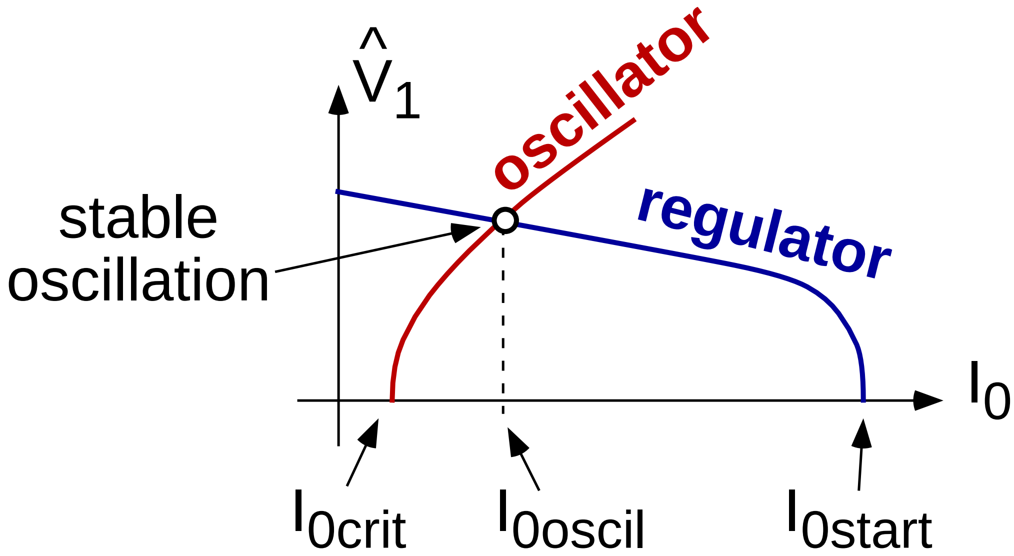
semi-empirical: $\boxed{R > \dfrac{2\beta_1/\beta_3}{\beta_2 n U_T}}$

MICROPOWER CRYSTAL OSCILLATOR

[Vittoz, 1979]



R_1,R_2,R_3,D_1,D_2 : lateral diodes in poly layer

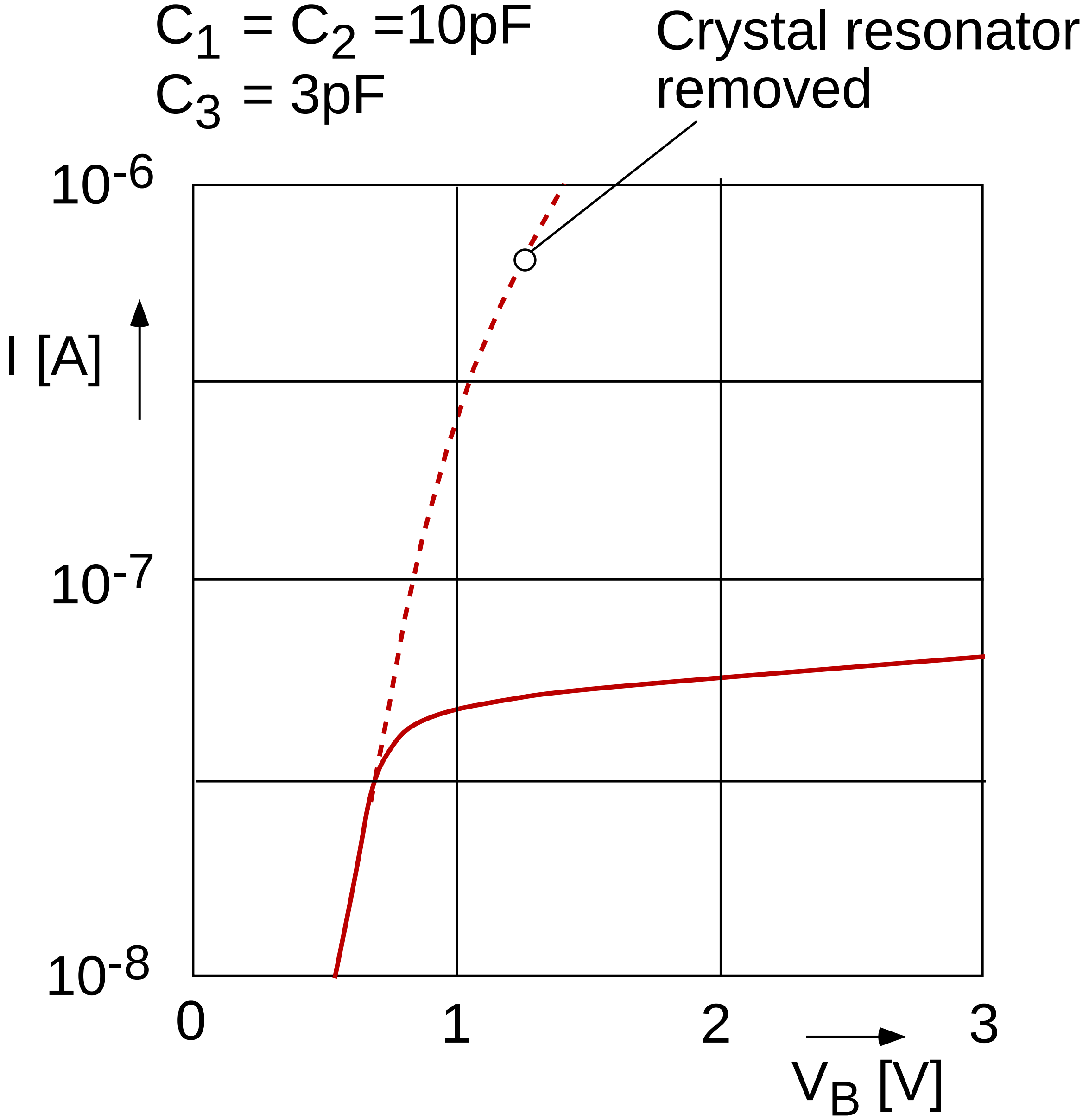


Example: f=32KHz
V_B=1 to 3 V
I_{tot}= 20 to 100nA (depends on C_1,C_2,Q)
no external component except crystal.

CURRENT DRAIN

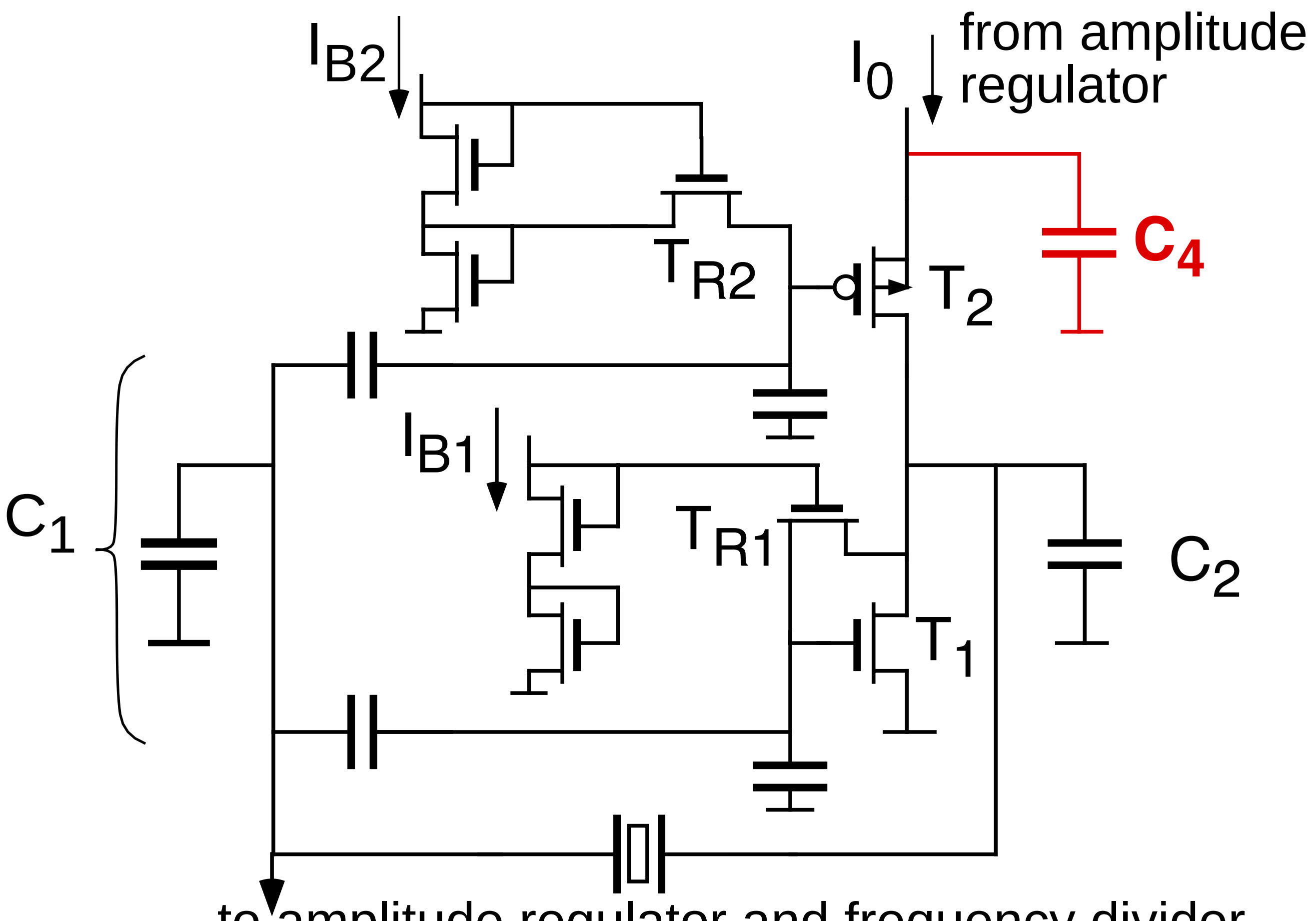
of micropower crystal oscillator

f = 32KHz
C = 2.9fF
Q = 80'000
C_1 = C_2 =10pF
C_3 = 3pF



VERY LOW-POWER 2MHZ OSCILLATOR

[Aebischer *et al*, 1997]

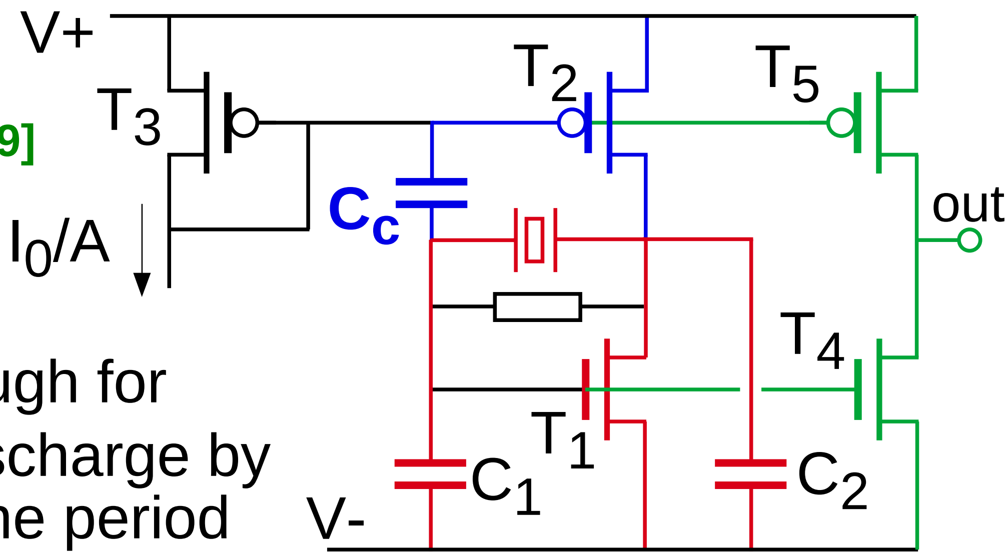


- Currrent controlled CMOS inverter T_1-T_2
  - gates separately biased by T_{R1}-T_{R2}
    - controlled by bias I_{B1}-I_{B2}
  - source of T_2 AC grounded by C_4, with $\omega C_4 \gg g_{ms2}$
- 2.1 MHz ZT cut quartz, C=0.5fF, Q=300-900K
  C_1=C_2=2.5pF,C_3=0.7pF, C_4= 10pF

I = 60-180nA (core oscillator only)
I < 500nA (oscill.+freq. divider+ dig. tuning)
@ 1.8 to 3.5V

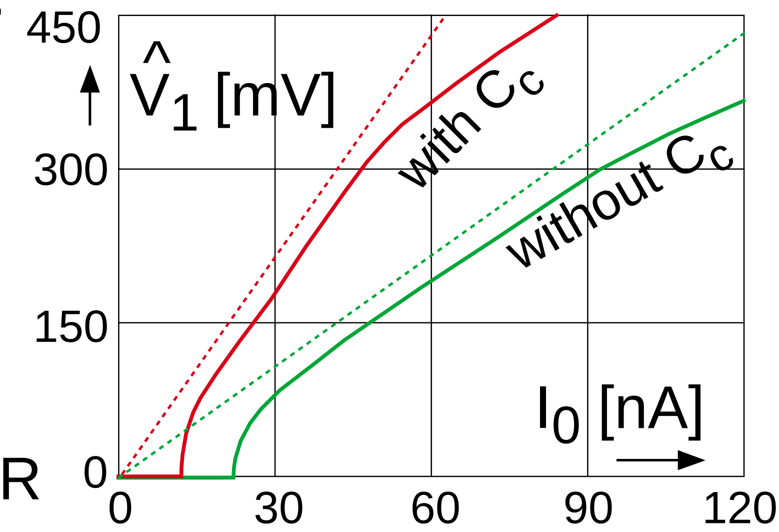
COMPACT PUSH-PULL OSCILLATOR

[Thommen, 1999]



- C_c large enough for negligible discharge by I_0/A during one period
- Average current I_0 through T_1 and T_3:
  - imposed by T_3
  - from amplitude regulator
- Instantaneous current in T_3
  - proportional to that in T_2
  - creates a loss conductance, thus:
  - effect. trans. of T_2 for fundamental: $g_{m2(1)}(1-1/A)$
- Output amplifier T_4-T_5 directly coupled to T_1-T_2
- Experimental results:

f	=32 kHz
R	=35 kΩ
A	=16
C_1	=12.3 pF
C_2	=24.6 pF
C_3	= 1pF
V_{Bmin}	= 0.7V



- DrawbacK:poor PSRR

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