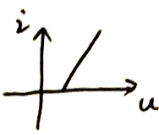


非线性电

一. 概述

四大数学模型: ①指数特性: $i = I_s e^{\frac{u}{U_T}}$, $U_T = 26 \text{ mV}$; ②折线特性: 

③双曲正切: $i = \frac{I_k}{1 + e^{\frac{u}{U_p}}} = \frac{I_k}{2} (1 + \tanh \frac{u}{U_p})$; ④平方特性: $i_D = I_{DSS} (1 - \frac{U_{GS}}{U_p})^2$

直流电导: $G = \frac{I_Q}{U_Q}$ 直流功耗: $P = G U_Q^2$ 交流电导: $g = \frac{\partial i}{\partial u} |_{u=U_Q}$

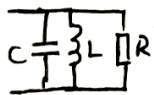
$G_{mn} \triangleq \frac{I_n}{U_i}$ ($n=1,2,3,\dots$) 为n次谐波等效电导 $\begin{cases} u = U_Q + U_i \cos \omega t \\ i(t) = I_0 + I_1 \cos \omega t + I_2 \cos 2\omega t + \dots + I_n \cos n\omega t + \dots \end{cases}$

时变电导: $u(t) = U_Q + U_1 \cos \omega t + U_2 \cos 2\omega t = U_Q(t) + U_2 \cos 2\omega t$ 条件: $U_1 \gg U_2$
 $i(t) \approx I_Q(t) + g(\omega t) U_2 \cos 2\omega t$ 其中 $g(\omega t) = \frac{\partial i}{\partial u} |_{u=U_Q(t)}$

变容二极管: $\frac{1}{C} \propto u(t)$ $C_j = C_0 (1 + \frac{u(t)}{U_\phi})^{-\gamma}$ 若 $u(t) = U_Q + U_1 \cos \omega t$, 则 $C_j(t) = C_{jQ} (1 + m \cos \omega t)^{-\gamma}$
 其中 $C_{jQ} = C_0 (1 + \frac{U_Q}{U_\phi})^{-\gamma}$, $m = \frac{U_1}{U_\phi + U_Q}$

并联谐振回路:

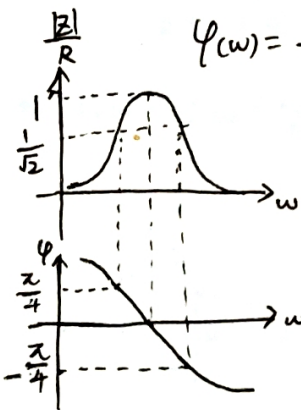
$$Z(\omega) = \frac{R}{1 + j(\omega CR - \frac{R}{\omega L})} = \frac{R}{1 + j(\frac{\omega}{\omega_0} - \frac{\omega_0}{\omega}) \frac{R}{\omega_0 L}} = \frac{R}{1 + jQ(\frac{\omega}{\omega_0} - \frac{\omega_0}{\omega})}$$



其中 $\omega_0 = \frac{1}{\sqrt{LC}}$ 为谐振频率, $Q = R \sqrt{\frac{C}{L}} = \frac{R}{\omega_0 L} = R \omega_0 C$ 为品质因数 ($Q = \frac{\omega_0}{\omega_k - \omega_L}$)

半功率点处 $|Z_{3dB}| = \frac{1}{\sqrt{2}} R$, $BW = \omega_h - \omega_l = \frac{\omega_0}{Q} = \frac{1}{RC}$ 半带宽 $\alpha = \frac{1}{2} BW = \frac{1}{2RC}$

$\varphi(\omega) = -\text{tg}^{-1} Q(\frac{\omega}{\omega_0} - \frac{\omega_0}{\omega}) \Rightarrow \varphi_{3dB} = \pm \frac{\pi}{4}$ $\omega > \omega_0$ 时呈容性; $\omega < \omega_0$ 时呈感性



$$Z(\omega) = \frac{R}{1 + jQ(\frac{\omega}{\omega_0} - \frac{\omega_0}{\omega})} \approx \frac{R}{1 + jQ \frac{2\Delta\omega}{\omega_0}} = \frac{R}{1 + j \frac{\Delta\omega}{\alpha}} \quad (\Delta\omega = \omega - \omega_0, \Delta\omega \rightarrow 0 \text{ 时成立})$$

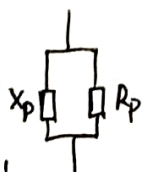
$$\left| \frac{Z(n\omega_0)}{Z(\omega_0)} \right| = \left| \frac{n}{n + jQ(n^2 - 1)} \right| \approx \frac{n}{Q(n^2 - 1)} \quad \text{谐波阻抗与基波阻抗之比}$$

若回路调谐于 $p\omega_0$, 则 $\left| \frac{Z(p\omega_0)}{Z(\omega_0)} \right| = \frac{1}{\sqrt{1 + Q^2(\frac{n^2 - p^2}{pn})^2}} \approx \frac{pn}{Q(n^2 - p^2)}$

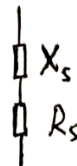
固有品质因数 Q_0 $\begin{cases} \text{并联谐振回路: } Q_0 = \frac{\text{储能}}{\text{损耗}} = \frac{U^2/\omega_0 L}{U^2/R} = \frac{R}{\omega_0 L} = R \omega_0 C \\ \text{串联谐振回路: } Q_0 = \frac{\text{储能}}{\text{损耗}} = \frac{I^2 \omega_0 L}{I^2 R} = \frac{\omega_0 L}{R} = \frac{1}{R \omega_0 C} \end{cases}$ (电容) 电感是储能元件, 电阻是耗能元件

串并联等效 (外导纳不变, Q 不变)

$$Q_{ep} = Q_{es}$$



$$Q_{ep} = \frac{U^2/X_p}{U^2/R_p} = \frac{R_p}{|X_p|}$$



$$Q_{es} = \frac{I^2/X_s}{I^2/R} = \frac{|X_s|}{R}$$

这里 X_p 和 X_s 为 ωL 或 $-\frac{1}{\omega C}$

$$\frac{R_p}{R_s} = 1 + Q^2 \quad \frac{|X_p|}{|X_s|} = 1 + \frac{1}{Q^2}$$

若 $Q \gg 1$, 则 $\frac{R_p}{R_s} \approx Q^2, \frac{|X_p|}{|X_s|} \approx 1$
 (窄带点频)

二. 非线性器件分析方法

1. 指数律特性分析 $B \begin{matrix} C \\ | \\ E \end{matrix} i_E \approx I_S e^{\frac{u_{BE}}{U_T}}$, 当 $u_{BE} = U_Q + U_1 \cos \omega t$ 时, $i_E(t) = I_{EP} \frac{e^{x \cos \omega t}}{e^x} = I_{EP} V_x(x)$

其中 $x = \frac{U_1}{U_T}$, 为归一化输入正弦幅度; $I_{EP} = I_S e^{\frac{U_Q}{U_T}} e^x$, 为电流峰值幅度

Fourier 级数展开: $i_E(t) = \sum_{n=0}^{\infty} I_{En} \cos n \omega t$, $I_{E0} = \frac{1}{2\pi} \int_{-\pi}^{\pi} i_E(t) d\theta = \frac{I_{EP}}{e^x} I_0(x) = I_{EQ} I_0(x)$

$$I_{En} = 2 \cdot \frac{1}{2\pi} \int_{-\pi}^{\pi} i_E \cos n \omega t d\theta = 2 \frac{I_{EP}}{e^x} I_n(x) = 2 I_{EQ} I_n(x)$$

$I_{EQ} = I_S e^{\frac{U_Q}{U_T}}$, 为静态工作电流; I_{E0} 为均值电流分量 (注: Fourier 展开中, 均值分量为其他分量的 $\frac{1}{2}$)

另一种表示形式: $i_E(t) = I_{E0} \left[1 + \sum_{n=2}^{\infty} \frac{2I_n(x)}{I_0(x)} \cos n \omega t \right]$

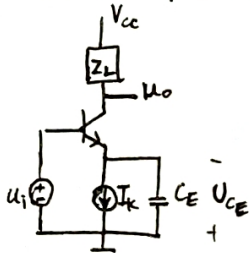
等效基波跨导: $G_{m1}(x) = \frac{\alpha I_{E1}}{U_1} = \frac{\alpha I_{E0}}{U_T} \frac{2I_1(x)}{x I_0(x)}$, 只与 x 和 U_Q 有关 $i_{c1} = G_{m1}(x) \cdot U_1 \cos \omega t = I_{c1} \cos \omega t$

总谐波失真 THD: 使用 RLC 并联谐振回路提取基波或 k 次谐波时衡量失真度

提取基波时, $THD_1 = \sqrt{\sum_{n=2}^{\infty} \left(\frac{U_n}{U_1}\right)^2}$ $THD_k = \sqrt{\sum_{n \neq k} \left(\frac{U_n}{U_k}\right)^2}$
 提取 k 次谐波时:

使用高 Q_T 的窄带 RLC 回路提取基波: $THD_{1,LPF} = \sqrt{\sum_{n=2}^{\infty} \left(\frac{I_n |Z(n\omega_0)|}{I_0 |Z(\omega_0)|}\right)^2} \approx \sqrt{\sum_{n=2}^{\infty} \left[\frac{I_n(x)}{I_0(x)} \cdot \frac{n}{Q(n^2-1)}\right]^2}$
 当 Q_T 足够大时, 可以为无失真, $u_o(t) \approx I_1 \cos \omega t \cdot R_L = \frac{1}{Q} D(x)$ (注: 只适用于指数律)

恒流偏置的晶体管放大器



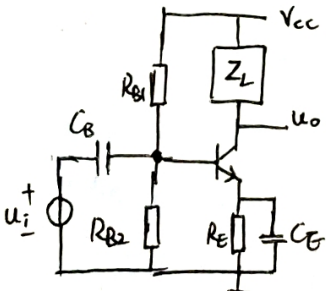
由于 I_k 作用, 使得 $I_{E0} = I_k$, 故 $G_{m1}(x) = \frac{\alpha I_k}{U_T} \frac{2I_1(x)}{x I_0(x)} = g_{m1} \frac{2I_1(x)}{x I_0(x)}$

静态时, I_k 给 C_E 充电, $I_k = I_S e^{\frac{U_{CE}}{U_T}}$

交流稳态时, C_E 起到提供 U_a 的作用: $I_{E0} = I_k = I_S e^{\frac{U_{CE}}{U_T}} I_0(x)$

静态和稳态时 C_E 上的电压不同, 为自生负偏压效应

电阻分压偏置的晶体管放大器

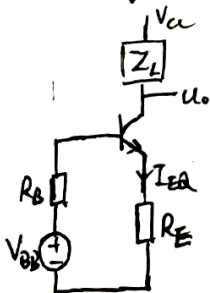


$$\frac{G_{m1}(x)}{g_{m1}} = \left[1 + \frac{I_n I_0(x)}{x I_1} \right] \frac{2I_1(x)}{x I_0(x)} \quad g_{m1} = \frac{\alpha I_{E0}}{U_T} = \frac{\alpha I_{EQ}}{U_T}$$

$$x_1 = \frac{U_1}{U_T}, \quad U_1 = V_{BB} - U_{BEQ} = \frac{R_{B2}}{R_{B1} + R_{B2}} V_{CC} - 0.7V \quad (\text{注: } U_1 \neq U_{EQ})$$

$$R_B = R_{B1} // R_{B2}, \quad I_{EQ} = \frac{V_{BB} - U_{BEQ}}{R_E + (1-\alpha)R_B} = \frac{U_1}{R_E + (1-\alpha)R_B}$$

↓ 直流负载线等效



先算 U_1 , 再算 I_{EQ} , 最后得 $I_{c1} \Rightarrow u_o$

$$\alpha = \frac{\beta}{1+\beta}$$

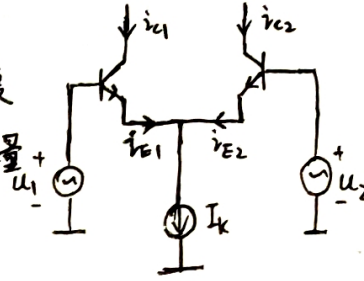
$$\beta = \frac{\alpha}{1-\alpha}$$

2. 双曲正切特性

$$\begin{cases} i_{E1} = \frac{I_k}{1+e^z} = \frac{I_k}{2}(1 + \tanh \frac{z}{2}) = \frac{I_k}{2} + i \\ i_{E2} = \frac{I_k}{1+e^{-z}} = \frac{I_k}{2}(1 - \tanh \frac{z}{2}) = \frac{I_k}{2} - i \end{cases}$$

其中 $z = \frac{u_1 - u_2}{U_r}$, 为归一化差分输入幅度

$i = \frac{I_k}{2} \tanh \frac{z}{2}$, 为发射极电流交流分量



(1) 当 $z \rightarrow \infty$ 时, 差分对有开关特性 (通常 $z \geq 4$ 即可视为 ∞)

$$z \rightarrow +\infty \text{ 时, } i_{E1} = I_k, i_{E2} = 0$$

$$z \rightarrow -\infty \text{ 时, } i_{E1} = 0, i_{E2} = I_k$$

(2) 当 $z \rightarrow 0$ 时, $\tanh \frac{z}{2} \approx \frac{z}{2}$

$$i_{E1,2} = \alpha i_{E1,2} = \frac{\alpha I_k}{2} (1 \pm \frac{z}{2}) = \frac{\alpha I_k}{2} \pm \frac{\alpha I_k}{4 U_r} (u_1 - u_2) = I_{CQ} \pm g_{mdQ} (u_1 - u_2)$$

其中 I_{CQ} 为集电极静态电流, $g_{mdQ} = \frac{1}{2} \frac{\alpha I_k}{2 U_r} = \frac{1}{2} g_{mQ}$ 为差分对小信号跨导, $g_{mQ} = \frac{\alpha I_{EQ}}{U_r} = \frac{\alpha I_k}{2 U_r}$ 为单管跨导
因此差分对小信号跨导为单管跨导的一半。

(3) 正弦激励: $u_1 - u_2 = U \cos \omega t$ 时

$$\text{令 } x = \frac{U}{U_r}, \text{ 则 } z = x \cos \omega t, \quad i = \frac{I_k}{2} \tanh(\frac{x}{2} \cos \omega t) = I_k \sum_{n=1}^{\infty} a_{2n-1}(x) \cos(2n-1)\omega t, \text{ 其中仅含奇次余弦成分}$$

$$\text{等效基波跨导: } G_{m1}(x) = \frac{\alpha I_{E1}}{U} = \frac{\alpha I_k a_1(x)}{x U_r} = \frac{\alpha I_k}{4 U_r} \cdot \frac{4 a_1(x)}{x} = g_{mdQ} \cdot \frac{4 a_1(x)}{x} \Rightarrow \frac{G_{m1}(x)}{g_{mdQ}} = \frac{4 a_1(x)}{x}$$

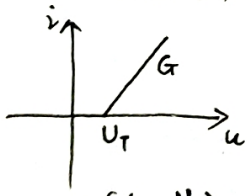
高 Q_T 时, $\text{THD}_1 = \frac{1}{Q_T} D(x) = \frac{1}{Q_T} \sqrt{\sum_{n=2}^{\infty} [\frac{a_{2n-1}(x)}{a_1(x)} \frac{2n-1}{(2n-1)^2-1}]^2}$, 差分对 $\text{THD} \ll$ 单管 THD , 因为无偶次谐波

动态限幅特性: $\frac{G_{m1}(x)}{g_{mdQ}}$ 随 x 的增大而减小, 可消除调频信号中的寄生调幅

负反馈差分电路: 略

3. 折线特性

大信号激励下, 二极管的电流电压关系、晶体管的输入特性或转移特性可用折线特性表征



$$i = \begin{cases} G(u - U_T), & u > U_T \\ 0, & u < U_T \end{cases}$$

对于正弦激励: $u = U_Q + U_1 \cos \omega t$

(1) $U_Q - U_1 > U_T$ 时, 工作在完全线性区

$$i = G(U_Q + U_1 \cos \omega t - U_T) = I_Q + I_1 \cos \omega t$$

(2) 当 $U_Q - U_1 < U_T$ 时, 工作在不完全线性区

$$\text{导通角 } \varphi = \arccos \frac{U_T - U_Q}{U_1} = \arccos \frac{\text{折点电压} - \text{工作点电压}}{\text{交流电压幅度}}$$

$$\text{电流峰值 } I_p = G(U_Q + U_1 - U_T) = G U_1 (1 - \cos \varphi)$$

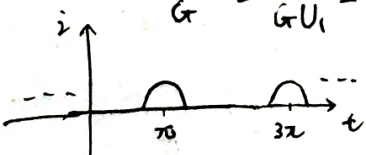
$$\text{电流 } i = G(U_Q + U_1 \cos \omega t - U_T) = G U_1 (\cos \omega t - \cos \varphi) = I_p \frac{\cos \omega t - \cos \varphi}{1 - \cos \varphi}$$

Fourier 展开: $i = I_p \sum_{n=0}^{\infty} \alpha_n(\varphi) \cos n \omega t$ α_n 是 φ 的函数, 可查表

$$\frac{G_{m1}(\varphi)}{G} = \frac{I_1}{G U_1} = \frac{1}{\pi} (\varphi - \sin \varphi \cos \varphi)$$

$$D(\varphi) = \sqrt{\sum_{n=2}^{\infty} (\frac{\alpha_n}{\alpha_1} \frac{n}{n^2-1})^2}$$

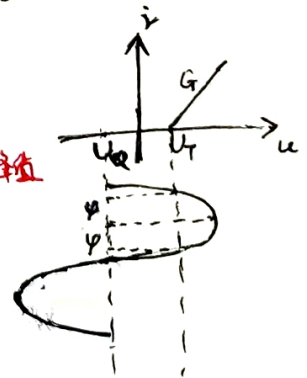
$$\text{THD}_1 = \frac{1}{Q_T} D(\varphi)$$



$$\Rightarrow i = I_p \sum_{n=2}^{\infty} \alpha_n(\varphi) \cos n(\omega t - \pi)$$

本质: 正弦脉冲峰值
导通角

可用于其他分析



4. 平方律特性

U_p : 夹断电压

FET:
$$i_D = \begin{cases} I_{DSS} \left(1 - \frac{u_{GS}}{U_p}\right)^2, & U_p < u_{GS} < 0 \\ 0, & u_{GS} < U_p \end{cases}$$

MOS:
$$i_D = \begin{cases} I_{D0} \left(1 - \frac{u_{GS}}{U_T}\right)^2, & u_{GS} > U_T \\ 0, & u_{GS} < U_T \end{cases}$$
 U_T : 开启电压

以 FET 为例分析: $u_{GS} = U_Q + U_1 \cos \omega t$

(1) 完全平方区: $U_Q + U_1 \leq 0$ 且 $U_Q - U_1 \geq U_p$

$$i_D = I_{DSS} \left(1 - \frac{U_Q + U_1 \cos \omega t}{U_p}\right)^2 = \frac{I_{DSS}}{U_p^2} \left[(U_p - U_Q)^2 + U_1^2 \cos^2 \omega t - 2(U_p - U_Q)U_1 \cos \omega t \right]$$

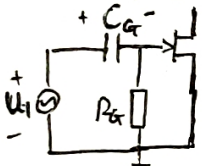
$$G_{m1} = \frac{I_{D1}}{U_1} = -\frac{2I_{DSS}}{U_p^2} (U_p - U_Q)$$

(2) 非完全平方区: $U_p < U_Q + U_1 \leq 0, U_Q - U_1 < U_p$

导通角: $\varphi = \arccos \frac{U_p - U_Q}{U_1}$ 电流峰值: $I_{Dp} = I_{DSS} \left(1 - \frac{U_Q + U_1}{U_p}\right)^2$

Fourier 展开: $i_D = I_{Dp} \sum_{n=0}^{\infty} \alpha_n(\varphi) \cos n\omega t$ $G_{m1}(\varphi) = \frac{I_{Dp} \alpha_1(\varphi)}{U_1}$ $D(\varphi) = \sqrt{\sum_{n=2}^{\infty} \left(\frac{\alpha_n}{\alpha_1} \frac{n}{n^2-1}\right)^2}$ $THD_1 = \frac{1}{Q_T} D(\varphi)$

自生负偏压电路: 充电时间常数: $r_{GS} C_G$, 放电时间常数: $R_G C_G$, 只要 $\begin{cases} R_G C_G \gg T \text{ (输入信号周期)} \\ R_G C_G \gg r_{GS} C_G \end{cases}$ 则 C_G 上将积累电荷, 产生自生负偏压: $U_{C_G} = U_1$



$u_1 = U_1 \cos \omega t$

则 $u_{GS} = -U_1 + U_1 \cos \omega t$ u_{GS} 的峰值被钳位于 0, 故为钳位平方律特性

5. 时变参量分析法

$u = U_Q + U_1 \cos \omega_1 t + U_2 \cos \omega_2 t = U_Q + u_1 + u_2, i = f(u)$

若 $U_1 \gg U_2$, 则可化为 $u = U_Q(t) + U_2 \cos \omega_2 t, i = I_Q(\omega_1 t) + g(\omega_1 t) U_2$

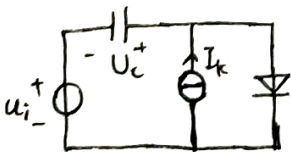
其中 $I_Q(\omega_1 t) = f(U_Q + U_1 \cos \omega_1 t) = I_{Q0} + I_{Q1} \cos \omega_1 t + \dots$

$g(\omega_1 t) = \frac{di}{du} \Big|_{u=U_Q+U_1 \cos \omega_1 t} = g_0 + g_1 \cos \omega_1 t + \dots$ $j(\omega_1 t) U_2 = g_0 U_2 \cos \omega_2 t + \frac{1}{2} g_1 U_2 [\cos(\omega_1 + \omega_2)t + \cos(\omega_1 - \omega_2)t] + \dots$

i 中的频率成分: $n\omega_1, \omega_2, n\omega_1 \pm \omega_2$

幅度分别为: $I_{Qn}, g_0 U_2, \frac{1}{2} g_n U_2$

电容自生负偏压效应



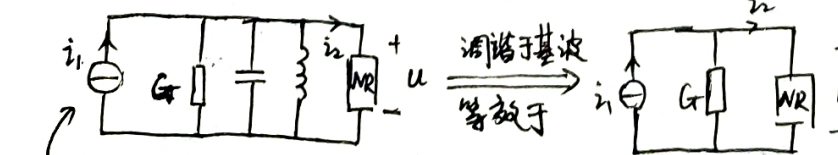
静态时: $I_k = I_s e^{\frac{U_{C0}}{U_r}} \Rightarrow U_{C0} = U_r \ln \frac{I_k}{I_s}$

动态时: U_C 作为静态偏置, 由指数律特性可得: $I_k = I_0 = I_s e^{\frac{U_C}{U_r}} I_0(x)$ 基波电流分量

解得 $U_C = U_r \ln \frac{I_k}{I_s} - U_r \ln I_0(x) < U_{C0}$

自生负偏压系数: $\xi = \frac{dU_C}{dU_1}$, 先写出 U_C 表达式, 再计算

高Q RLC回路的非线性元件加载



可以是集电极等效而来

$u = U_0 \cos \omega t$, 而 i_1 和 i_2 均含多次谐波分量

基波分量满足以下方程:

$I_{11} = U_0 G + I_{21}$ 为线性负载方程

同时: $I_{21} = F(U_0)$, 为非线性负载方程 由非线性器件决定

$THD = \frac{1}{Q_T} \sqrt{\sum_{n=2}^{\infty} \left(\frac{I_{2n}}{I_{11}} \frac{n}{n^2-1}\right)^2}$ 流入回路的电流

第3章 功率放大器 (输入为小信号, 在集电极只有基波)

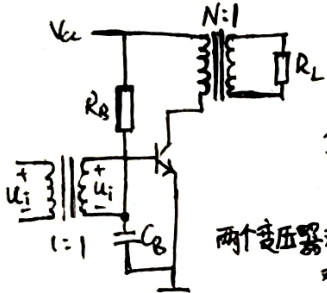
效率 { 晶体管: 集电极效率
场效应管: 漏极效率

$$\eta_c = \frac{P_o}{P_{DC}} = \frac{P_o}{P_o + P_c}$$

P_o : 功率管输出信号功率
 P_{DC} : 电源提供的功率 $P_{DC} = P_o + P_c$
 P_c : 集电极耗散功率

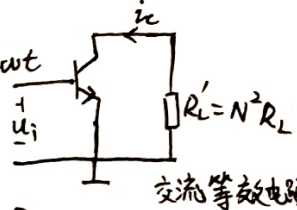
分类 (按导通角 φ): $\varphi = 180^\circ$: 甲类; $90^\circ < \varphi < 180^\circ$: 甲乙类; $\varphi = 90^\circ$: 乙类; $\varphi < 90^\circ$: 丙类

甲类功放:



静态分析: $i_c = I_{CQ}$
 $u_{CE} = V_{CC}$

交流分析: $i_c = I_c \cos \omega t$
 $u_{CE} = -i_c R'_L = -U_c \cos \omega t$

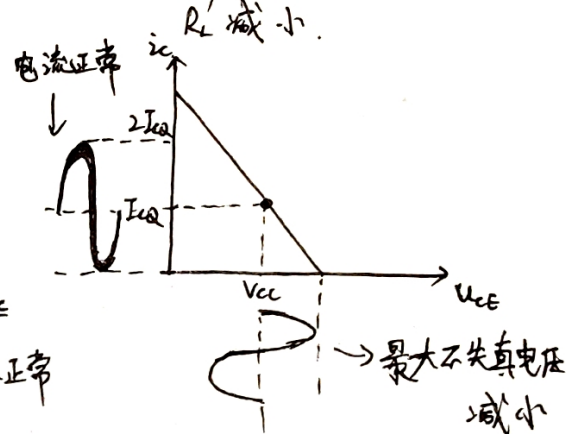
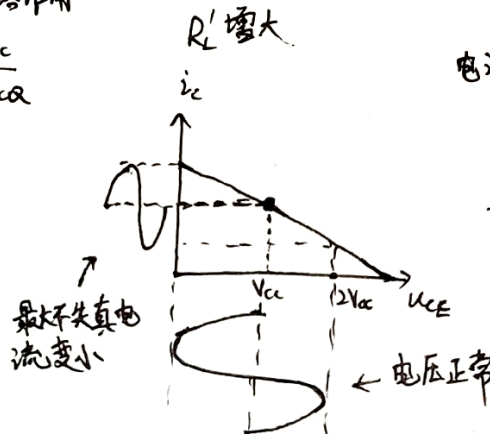
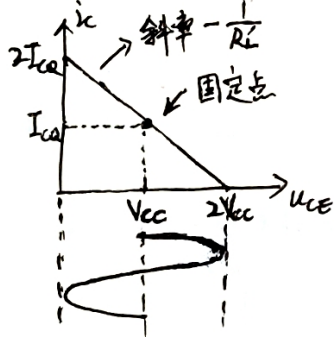


集电极耗散功率 $P_c = \frac{1}{2\pi} \int_0^{2\pi} i_c(t) \cdot u_{CE}(t) dt = V_{CC} I_{CQ} - \frac{1}{2} I_c U_c = P_{DC} - P_o$

输出功率 $P_o = \frac{1}{2} I_c U_c = \frac{1}{2} I_c^2 R'_L = \frac{1}{2} U_c^2 / R'_L$

两个变压器对交流起阻抗变换作用
对直流起短路作用

理想情况: 取合适 $R'_L = \frac{V_{CC}}{I_{CQ}}$



此时功率达到最大: $P_o = \frac{1}{2} I_{CQ} V_{CC} = \frac{1}{2} I_{CQ}^2 R'_L = \frac{1}{2} V_{CC}^2 / R'_L$

$P_o = \frac{1}{2} \frac{V_{CC}^2}{R'_L}$

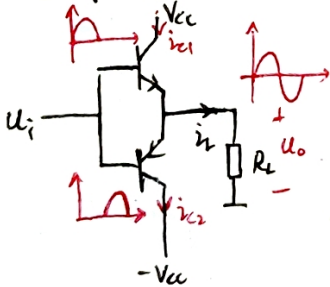
$P_o = \frac{1}{2} I_{CQ}^2 R'_L$

最大理想效率: $\eta = \frac{P_o}{P_{DC}} = \frac{\frac{1}{2} V_{CC} I_{CQ}}{V_{CC} I_{CQ}} = 50\%$

乙类功放:

乙类互补推挽功放

变压器耦合型
推挽功放: 略



u_i 正半周时, $i_{c1} = I_c \sin \omega t = i_L$; u_i 负半周时, $i_{c2} = -I_c \sin \omega t = -i_L$

故在一个周期内, $i_L = I_c \sin \omega t$, $P_o = \frac{1}{2} I_c^2 R_L = \frac{1}{2} I_c U_c$

$$P_{DC} = 2 P_{DC1} = 2 \frac{1}{2\pi} \int_0^\pi V_{CC} I_c \sin \omega t d\omega t = \frac{2}{\pi} I_c V_{CC}$$

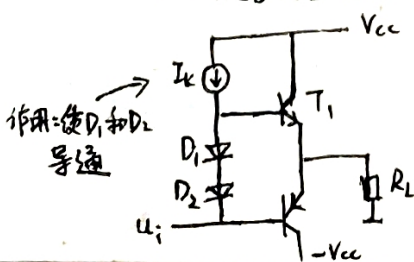
$$\eta_c = \frac{P_o}{P_{DC}} = \frac{\pi}{4} \frac{U_c}{V_{CC}} = \frac{\pi}{4} \xi$$

ξ 为电源电压利用系数

非理想情况

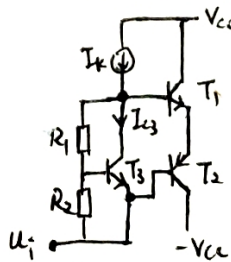
1. 交越失真: 零偏置条件下, 要考虑导通电压影响. 解决方法: 加偏置电路.

① 二极管偏置电路



作用: 使 D1 和 D2 导通

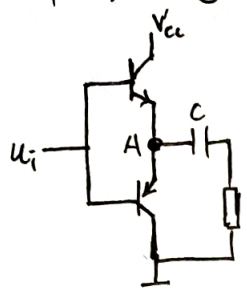
② 偏置倍增电路 R_1 较大, 使 $I_{C3} \approx I_k$



$$(U_{BE1} + U_{BE2}) \frac{R_2}{R_1 + R_2} = U_{BE3} = U_T \ln \frac{I_k}{I_s}$$

$$U_{BE1} + U_{BE2} = (1 + \frac{R_1}{R_2}) U_{BE3} = U_T (1 + \frac{R_1}{R_2}) \ln \frac{I_k}{I_s}$$

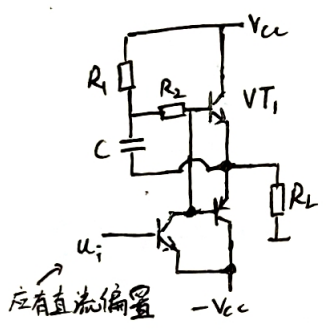
2. 单电源供电



$V_A = \frac{V_{CC}}{2}$, 电路相当于以 $\frac{V_{CC}}{2}$ 和 $-\frac{V_{CC}}{2}$ 供电的双电源电路

3. 复合管技术解决互补管配对问题 (略)

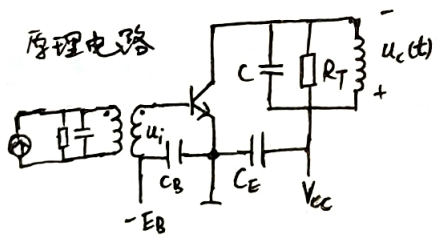
4. 自举电路: 提高效率



直流负载为 $R = R_1 + R_2$
 交流负载大于直流负载, 最大不失真电压幅度取决于直流负载线
 使电压幅度接近 V_{CC} , 效率提高

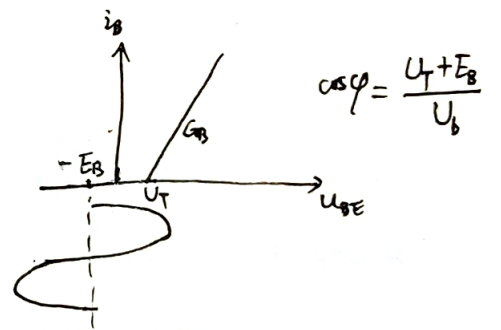
5. 保护电路 (略)

谐振功放 (以谐振系统作为匹配网络的功率放大器)

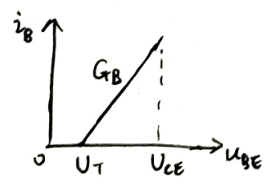


$u_i = U_b \cos \omega t, u_c = U_c \cos \omega t$

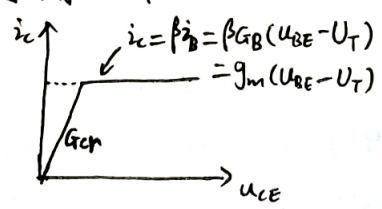
$$\begin{cases} U_{BE} = -E_B + U_b \cos \omega t \\ U_{CE} = V_{CC} - U_c \cos \omega t \end{cases}$$



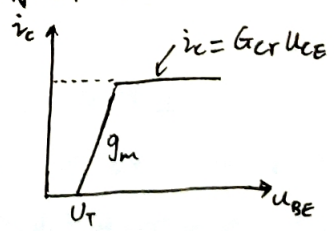
静态输入特性



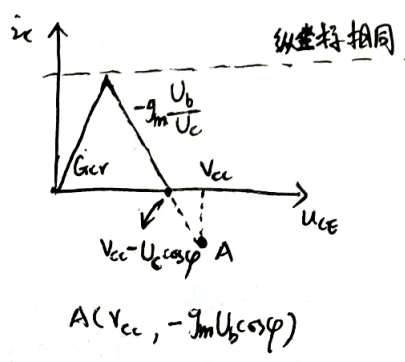
静态输出特性



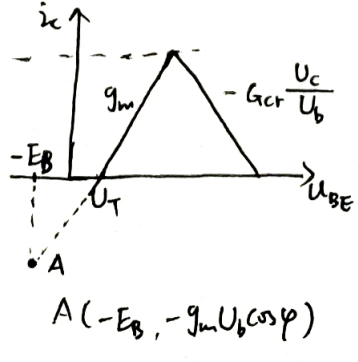
静态转移特性



动态输出特性



动态转移特性



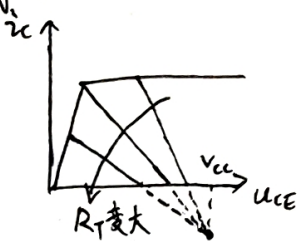
谐振功放工作状态：欠压 → 临界 → 过压， $U_b \uparrow$ ，图见书 P44

欠压或临界时，按折线律分析： $I_{cp} = g_m(-E_b + U_b - U_T)$ ， $I_{co} = I_{cp} \alpha_o(\varphi) \rightarrow P_{DC} = V_{cc} I_{co}$

$$I_{c1} = I_{cp} \alpha_1(\varphi) \rightarrow P_o = \frac{1}{2} I_{c1}^2 R_T = \frac{1}{2} I_{c1} U_c$$

$$\eta = \frac{P_o}{P_{DC}} = \frac{1}{2} \frac{\alpha_1(\varphi)}{\alpha_o(\varphi)} \cdot \frac{U_c}{V_{cc}} = \frac{1}{2} \frac{\alpha_1(\varphi)}{\alpha_o(\varphi)} \xi \quad \leftarrow \text{源电压利用系数}$$

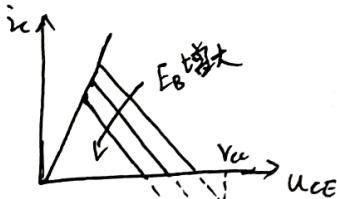
负载特性 (E_b, V_{cc}, U_b 不变, 改变 R_T)



随 R_T 变大，工作状态：欠压 → 临界 → 过压

当达到临界状态， P_o 达到最大， η 也较大，此时 R_T 为最佳负载

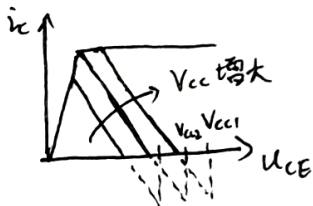
基极调制特性 (V_{cc}, U_b, R_T 不变, 改变 E_b)



随 E_b 变大，工作状态：过压 → 临界 → 欠压

在欠压区， I_{c1} 随 E_b 增大而几乎线性减小（产生调幅波）

集电极调制特性 (E_b, U_b, R_T 不变, 改变 V_{cc})



随 V_{cc} 增大，工作状态：过压 → 临界 → 欠压

在过压区， I_{c1} 随 V_{cc} 增大而几乎线性增大（产生调幅波）

放大特性 (E_b, R_T, V_{cc} 不变, 改变 U_b)

图略

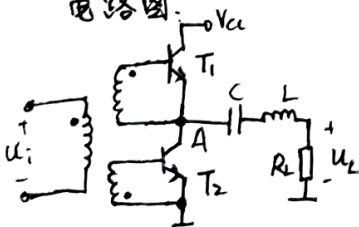
随 U_b 增大，工作状态：欠压 → 临界 → 过压

在欠压区， I_{c1} 和 U_c 随 U_b 增大近似线性增大：线性放大特性

（欠压区）放大调幅波时，应使 $\varphi = 90^\circ$ ；临界状态用于放大调频波；过压区有动态限幅特性，可抑制寄生调幅

丁类谐振功放

电路图



$$u_i = U_i \cos \omega t$$



$$\text{基波分量 } U_L = \frac{2}{\pi} (V_{cc} - 2U_{ces})$$

$$I_L = \frac{2}{\pi R_L} (V_{cc} - 2U_{ces})$$

$$P_o = \frac{1}{2} U_L I_L = \frac{2(V_{cc} - 2U_{ces})^2}{\pi^2 R_L}$$

U_i 幅度大，使二极管饱和导通， $U_{CE} = U_{CES}$

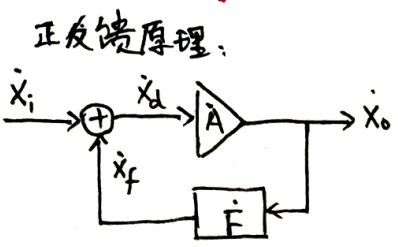
LC 串联谐振于基波频率

T_1 和 T_2 均半周期导通， I_{c1} 和 I_{c2} 的幅值均为 I_L

$$I_{c1,0} \text{ 的均值分量 } I_{c1,0} = \frac{1}{2\pi} \int_0^{2\pi} I_L \cos \omega t \, d\omega t = \frac{I_L}{\pi}$$

$$P_{DC} = 2 P_{DC1} = 2 \cdot \frac{V_{cc}}{2} \cdot \frac{I_L}{\pi} = \frac{V_{cc}}{\pi} I_L = \frac{2V_{cc}}{\pi^2 R_L} (V_{cc} - 2U_{ces})$$

四、正弦振荡器



开环增益: $A = \frac{X_o}{X_d}$ 反馈系数: $F = \frac{X_f}{X_o}$ 环路增益: $T = AF = \frac{X_f}{X_d}$

起振条件: $T = AF > 1 \Rightarrow \begin{cases} |AF| > 1 \\ \varphi_{AF} = \varphi_A + \varphi_F = 2n\pi \end{cases}$

平衡条件: $T = AF = 1 \Rightarrow \begin{cases} |AF| = 1 \\ \varphi_{AF} = \varphi_A + \varphi_F = 2n\pi \end{cases}$

实际工作中, $\varphi_T(\omega) = \varphi_A(\omega) + \varphi_F(\omega) = \varphi_f(\omega) + \varphi_z(\omega) + \varphi_F(\omega) = 0$ 为相位条件

谐振回路相位 $\varphi_z(\omega) = -[\varphi_f(\omega) + \varphi_F(\omega)] = -\varphi_F(\omega)$ 为基准

若 $\varphi_F(\omega)$ 足够小, 则 $\varphi_z(\omega) = 0 \Rightarrow \omega_{osc} = \omega_0$ (谐振回路振荡频率)

稳定条件 (电路平衡后不被干扰而破坏平衡)

① 稳幅: $\frac{\partial T}{\partial U_i} = A \frac{\partial F}{\partial U_i} + F \frac{\partial A}{\partial U_i} < 0$ 若反馈网络为线性 $F \frac{\partial A}{\partial U_i} < 0 \Rightarrow \frac{\partial A}{\partial U_i} = \frac{\partial G_{m1}(U_i)}{\partial U_i} < 0$

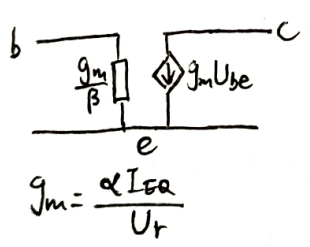
非线性器件一般均自带天然稳幅能力

② 稳频: $\frac{\partial \varphi_T}{\partial \omega} \Big|_{\omega=\omega_{osc}} < 0$ 因此以 RLC 作定时网络: $\frac{\partial \varphi_T}{\partial \omega} \approx \frac{\partial \varphi_z}{\partial \omega} \approx -\frac{2Q_T}{\omega_0} < 0$

晶体管简化模型

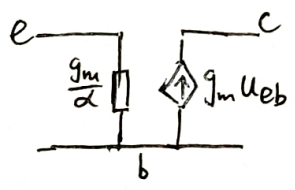
(一般为 NPN 管)

共射极简化模型



(一般为 PNP 管)

共基简化模型

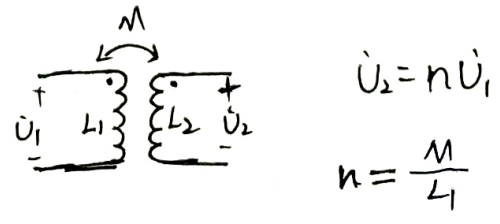
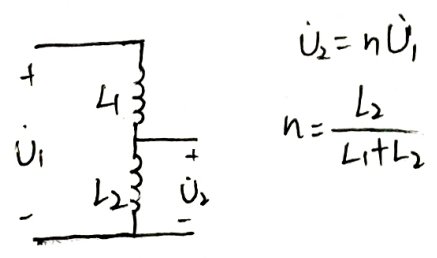
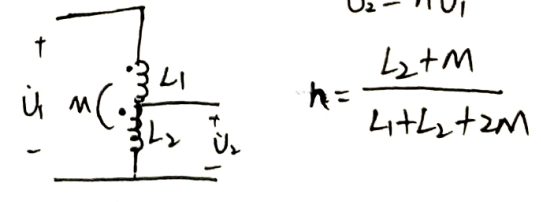
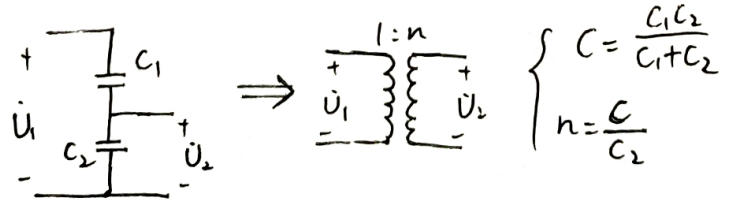


共集简化模型

略

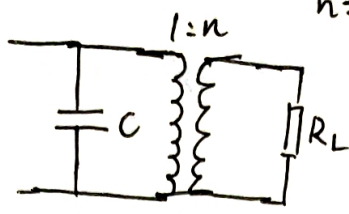
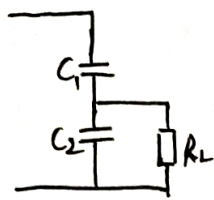
注: 各简化模型间在应用中应无本质区别

部分接入 (n 为接入系数) 电压关系

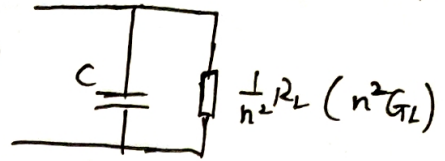


$k = \frac{M}{\sqrt{L_1 L_2}}$ 为耦合系数

电容分压式的等效模型变换

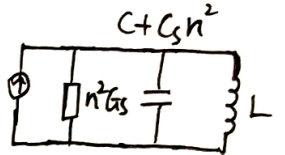
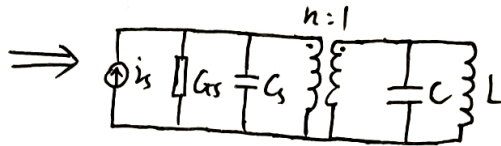
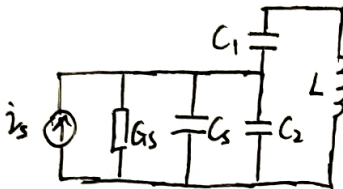


$n = \frac{C}{C_2}$ $C = \frac{C_1 C_2}{C_1 + C_2}$ (条件: $Q^* = \frac{R_L}{|X_{C_2}|} > 10$)

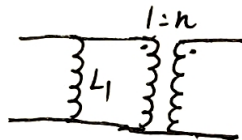
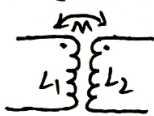


受控源阻抗变换

若负载为小电容 C_L , 则变换后为 $n^2 C_L$



全耦合变压器变换



$n = \frac{M}{L_1}$

正弦振荡器分析步骤:

① 判断电路有无错误

② 画直流电路, 计算 I_{EQ} 和 g_{mQ} . $g_{mQ} = \frac{\alpha I_{EQ}}{U_T}$

③ 画交流等效电路, 判断相位平衡条件和起振幅值条件. $T > 1 \Rightarrow g_m > g_{m, min}$, 若 $g_{mQ} > g_{m, min}$ 则可以起振

④ 计算 f_{osc} 和 U_{osc} . 平衡状态时, $G_{in}(x) = g_{m, min} \Rightarrow$ 可得 $\frac{G_{in}(x)}{g_{mQ}} = \frac{g_{m, min}}{g_{mQ}} \Rightarrow$ 查表得 x

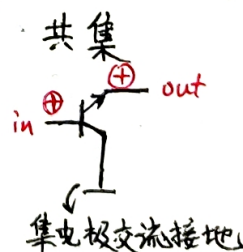
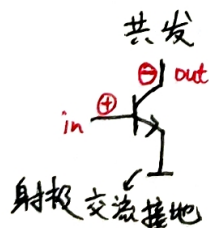
回路两端电压

注意 $x = \frac{U_{be}}{U_T}$, 可由交流等效电路求得 U_{osc}

$\omega_{osc} = \frac{1}{\sqrt{LC}}$

瞬时极性法 (判断相位是否符合起振条件)

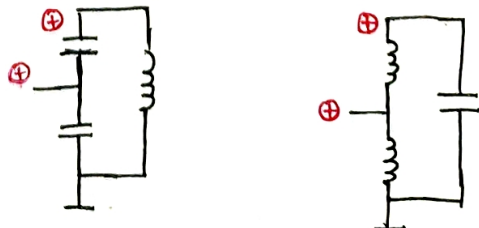
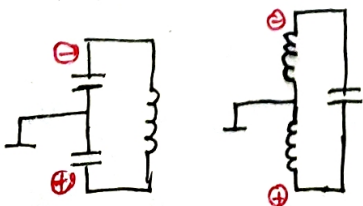
晶体管相位关系 (只有共发极性相反, 其余两种组态极性相同)



电容、电感三电式相位关系

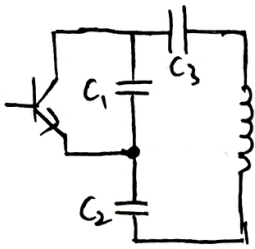
① 中间交流接地, 则两头相位相反

② 两头其中一头交流到地, 另两点相位相同



三点式振荡电路

相位平衡条件: 共射组态: 射同基反; 共基组态: 射同基反; 共集组态: 射同基反集反
改进型电容三点式电路.

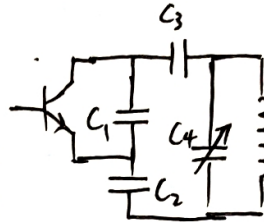


Clapp 电路

取 $C_3 \ll C_1, C_2$
极间电容接入效应为 $n^2 C_{ce}, n^2 C_{be}$
使极间电容影响变小 (作用)

$$f_0 \approx \frac{1}{2\pi\sqrt{LC_3}}$$

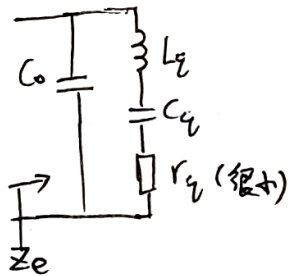
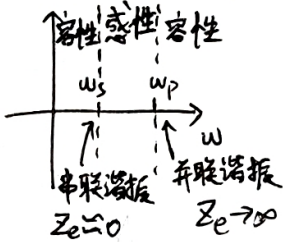
缺点: 改变 C_3 可能使 Q_{min} 变大而不能起振, 不方便通过改变 C_3 来改变频率



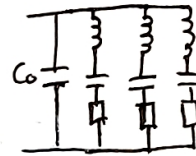
$C_3, C_4 \ll C_1, C_2$
优点: 通过改变 C_4 调节频率, 同时不影响任何其它接入系数

$$f_0 \approx \frac{1}{2\pi\sqrt{L(C_3+C_4)}}$$

石英晶体振荡器

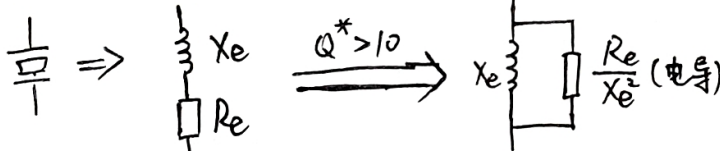


泛音 (与基频成奇数倍关系)

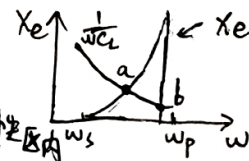


工作模式:

① 并联型 (L 模式): 当电感使用, 代替电容三点式中的电感



可求得谐振时 X_c 满足的方程



a 处 $\frac{dX_c}{d\omega} < 0$
稳定

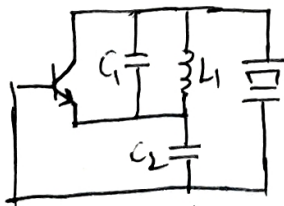
结论: 并联型晶体振荡器的 ω_{osc} 在串联谐振频率 ω_s 附近的感性区内

② 串联型 (r 模式): 代替高频旁路电容

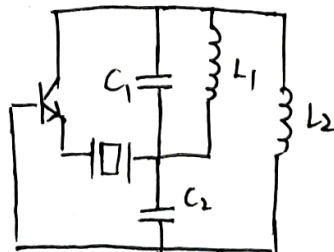
串联型晶振的振荡频率就是晶体的串联谐振频率, 即工作时 $\omega_{osc} = \omega_s$

③ 泛音模式 (使晶体工作在泛音频率)

相比基音模式, 泛音模式需加入抑制电路 ($L_1 C_1$ 回路)



并联型



串联型

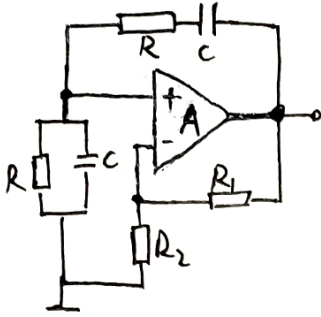
$\omega > \frac{1}{\sqrt{LC_1}}$ 时, 回路呈容性, 满足射同基反

$\omega < \frac{1}{\sqrt{LC_1}}$ 时, 回路呈感性, 不能起振

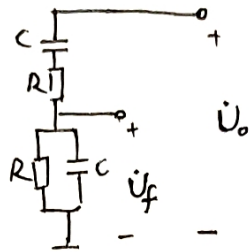
要得到 5 次泛音频率, 需调节 $L_1 C_1$ 回路使其谐振频率位于 3~5 次泛音频率之间, 对 5 次泛音可以起振。对 7 次以上泛音, 回路容性太大, 幅值平衡条件难以满足。

阻容振荡器

文氏桥振荡器



A, R₁, R₂ 构成同相比倒运算电路: $A_f = 1 + \frac{R_1}{R_2}$
RC串并联网络起选频和反馈作用, $\omega_0 = \frac{1}{RC}$



$$\dot{F} = \frac{\dot{U}_f}{\dot{U}_0} = \frac{R // \frac{1}{j\omega C}}{R + \frac{1}{j\omega C} + R // \frac{1}{j\omega C}} = \frac{1}{3 + j(\frac{\omega}{\omega_0} - \frac{\omega_0}{\omega})}$$

其对 $\omega = \omega_0$ 的信号相移为 0, 且 |F| 最大, 故对 ω_0 有选频作用

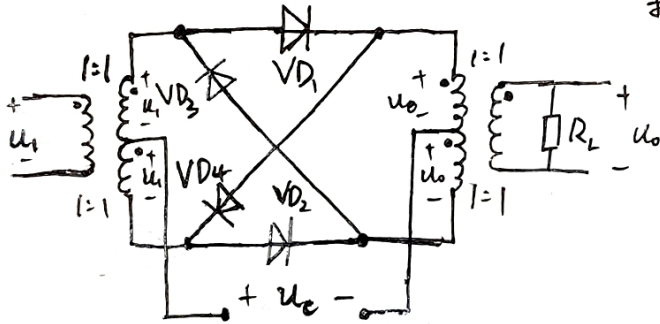
故 $\omega_{osc} = \omega_0 = \frac{1}{RC}$, 此时 $\dot{F} = \frac{1}{3}$ $\dot{T} = A_f \dot{F} = \frac{1}{3}(1 + \frac{R_1}{R_2})$
起振幅度条件 $\dot{T} > 1 \Rightarrow R_1 > 2R_2$

五. 模拟乘法器

变跨导式乘法器: 以双平衡差分电路为核心电路 (略)

指数、对数乘法器: 略

双平衡 = 极管准模拟乘法器



条件: $U_c \gg U_1, U_0$, 即 4 个二极管通断受制于 u_c

① u_c 正半周时, VD_1, VD_2 导通, 导通电导为 g

$$\begin{cases} u_{D1} = u_1 + u_c - u_0 \Rightarrow i_{D1} = g(u_1 + u_c - u_0) = i_1 \\ u_{D2} = u_c - u_1 + u_0 \Rightarrow i_{D2} = g(u_c - u_1 + u_0) = i_2 \end{cases}$$

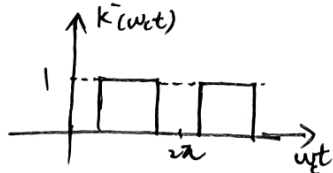
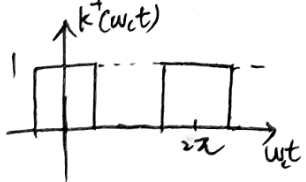
$$u_0 = i_0 R_L = (i_1 - i_2) R_L = 2gR_L(u_1 - u_0)$$

② u_c 负半周时, VD_3, VD_4 导通

$$\begin{cases} u_{D3} = -u_c - u_0 - u_1 \Rightarrow i_{D3} = g(-u_c - u_0 - u_1) \\ u_{D4} = -u_c + u_0 + u_1 \Rightarrow i_{D4} = g(-u_c + u_0 + u_1) \end{cases}$$

$$u_0 = R_L(i_3 - i_4) = -2gR_L(u_1 + u_0)$$

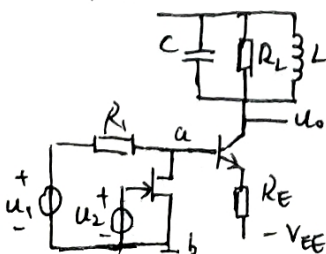
全周期表达式: $u_0 = 2gR_L(u_1 - u_0)k^+(u_c) + 2gR_L(-u_1 - u_0)k^-(u_c)$ k^+ 和 k^- 为开关函数



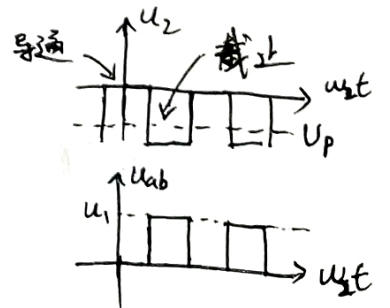
$$\begin{aligned} k^+(wt) + k^-(wt) &= 1 \\ k^+(wt) - k^-(wt) &= \frac{4}{\pi} \cos wt - \frac{4}{3\pi} \cos 3wt + \dots \end{aligned}$$

$$k^+(wt) = \frac{1}{2} + \frac{2}{\pi} \cos wt - \frac{2}{3\pi} \cos 3wt + \dots \quad k^-(wt) = \frac{1}{2} - \frac{2}{\pi} \cos wt + \frac{2}{3\pi} \cos 3wt + \dots$$

场效应管开关准模拟乘法器



① u_2 为高电平时, 场效应管导通, $u_{ab} = 0$
② u_2 为低电平时, 场效应管截止, $u_{ab} = u_1$ ($u_2 < U_p$)



六、调幅与检波

一般调幅波

$u_c = U_c \cos \omega_c t$ 为高频载波

$s(t)$ 为归一化基带信号

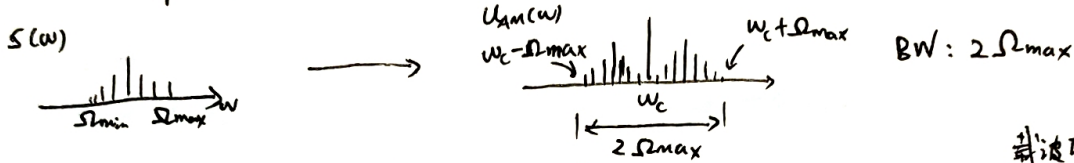
$$\begin{cases} \overline{s(t)} = 0 \text{ (均值为0)} \\ s(t)_{\max} = 1 \text{ (归一化处理)} \\ f_{\min} < f < f_{\max} \text{ (带限信号)} \end{cases}$$

调制后: $u_{AM} = U_c [1 + m s(t)] \cos \omega_c t = A(t) \cos \omega_c t$

其中 m 为调幅指数(调制度), $m \leq 1$

$$m = \frac{U_c(1+m) - U_c(1-m)}{U_c(1+m) + U_c(1-m)} = \frac{A_{\max} - A_{\min}}{A_{\max} + A_{\min}}$$

设 $s(t) = \sum_{n=1}^q a_n \cos n \Omega_d t$, 则 $u_{AM} = U_c \cos \omega_c t + \frac{1}{2} m U_c \sum_{n=1}^q a_n \cos(\omega_c + n \Omega_d) t + \frac{1}{2} m U_c \sum_{n=1}^q a_n \cos(\omega_c - n \Omega_d) t$



调幅波在单位等效电阻上消耗的平均功率

$$P_{AM} = \frac{\overline{u_{AM}^2}}{R} = \frac{U_c^2 [1 + 2m s(t) + m^2 s^2(t)] \cos^2 \omega_c t}{R} = \frac{1}{2} U_c^2 + \frac{1}{2} m^2 U_c^2 \overline{s^2(t)} = P_c + P_m$$

效率 $\eta = \frac{P_m}{P_c + P_m}$, 效率低

抑制载波的双边带调幅 (DSB): 没有载波频率, 提高 η

$$u_{DSB} = U_c \cdot s(t) = U_c s(t) \cos \omega_c t \quad BW = 2 \Omega_{\max}$$

波形特点: ① 在 $s(t) = 0$ 两旁, $s(t)$ 正负变化, 使 u_{DSB} 相位发生 180° 突变

② 上下包络形状并非 $s(t)$ 形状, 正峰值包络: $|s(t)|$, 负峰值包络: $-|s(t)|$

DSB特点: 解调复杂, 需要同步检波

抑制载波的单边带调幅 (SSB): 减小带宽

$$\begin{aligned} u_{SSB} &= \sum_{n=1}^q a_n \cos(\omega_c + n \Omega_d) t = \cos \omega_c t \sum_{n=1}^q a_n \cos n \Omega_d t - \sin \omega_c t \sum_{n=1}^q a_n \sin n \Omega_d t \\ &= s(t) \cos \omega_c t - x(t) \sin \omega_c t \end{aligned}$$

$x(t)$ 为 $s(t)$ 经过 90° 延迟的信号

调幅电路

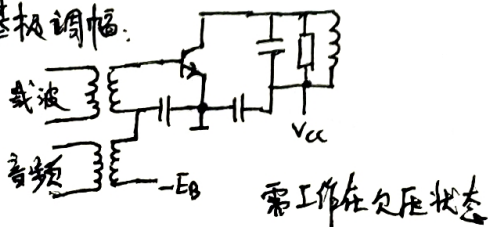
(1) 低电平调幅 (调幅在系统低电平级进行)

用乘法器实现: 变跨导式乘法器、双平衡二极管准模拟乘法器 略。

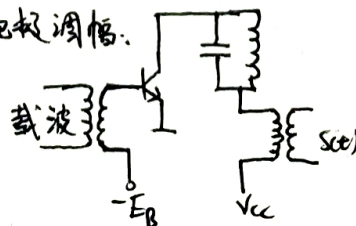
(2) 高电平调幅 (调幅在系统高电平级进行)

利用谐振放大器的基极调制特性和集电极调制

基极调幅:



集电极调幅:



需工作在过压状态。

单边带信号的产生

(1) 滤波法：由DSB信号经过单边带滤波器得到，由于上、下边带距离近，需多级滤波

(2) 移相法：
 $U_{SSB} = S(t)\cos\omega_c t - X(t)\sin\omega_c t$ 余弦载波，上边带
 $U_{SSB} = S(t)\sin\omega_c t + X(t)\cos\omega_c t$ 正弦载波，上边带
 X(t)为S(t)经90°相移得到

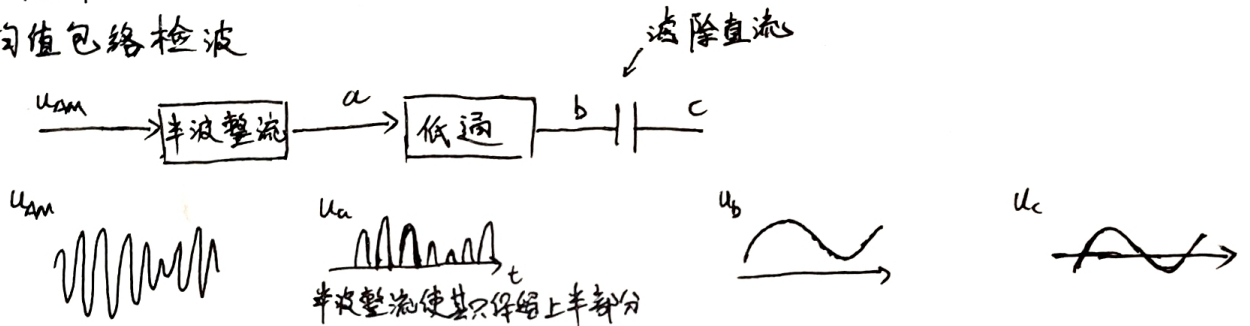
注：90°相移器为延迟90°相位（或超前90°） $\sin \rightarrow \cos$, $\cos \rightarrow \sin$?

(3) 移相滤波法：略

调幅波的解调（检波）：包络检波（普通调幅波）、同步检波（普通调幅波、DSB、SSB）

包络检波

(1) 均值包络检波



$$u_{AM} = U_c [1 + m S(t)] \cos \omega_c t$$

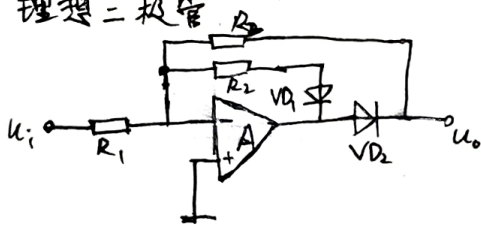
$$u_a = U_c [1 + m S(t)] \cos \omega_c t \cdot k(\omega_c t)$$

$$= U_c [1 + m S(t)] \cos \omega_c t \left[\frac{1}{2} + \frac{2}{\pi} \cos \omega_c t - \frac{2}{3\pi} \cos 3\omega_c t + \dots \right]$$

低频信号由 $U_c [1 + m S(t)] \cos \omega_c t \cdot \frac{2}{\pi} \cos \omega_c t$ 产生， $u_b = \frac{1}{\pi} U_c [1 + m S(t)]$

$$u_c = \frac{U_c m S(t)}{\pi}$$

(2) 理想二极管

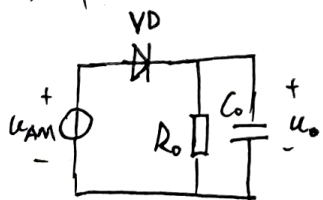


R_2, VD_1 支路：防止A在 u_i 负半周出现开环状态

u_i 正半周： VD_1 导通， VD_2 截止

u_i 负半周： VD_1 截止， VD_2 导通

(3) 峰值包络检波



$\tau_{充} \gg T_{充}$

输出电压有锯齿状纹波

假定二极管理想

$$u_{AM} = U_c [1 + m S(t)] \cos \omega_c t = b(t) \cos \omega_c t$$

$$u_o(t) = b(t)$$

现考虑各谐波对 $u_o(t)$ 的影响。

窄脉冲 $\rightarrow i_0 = I_{D0}(t) \left(1 + 2 \sum_{n=1}^{\infty} \cos n\omega_c t \right)$

$$I_{D0} = \frac{b(t)}{R_0}$$

$$R_{in} = \frac{b(t)}{I_{D1}} = \frac{b(t)}{2I_{D0}} = \frac{R_0}{2} \text{ 为基波阻抗}$$

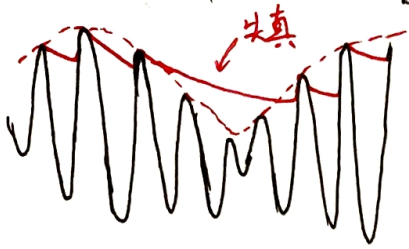
$$|Z(n\omega_c)| \approx \frac{1}{n\omega_c C_0}$$

$$U_n = |Z(n\omega_c)| \cdot I_{Dn} = \frac{2I_{D0}}{n\omega_c C_0} = \frac{2b(t)}{n\omega_c C_0 R_0}$$

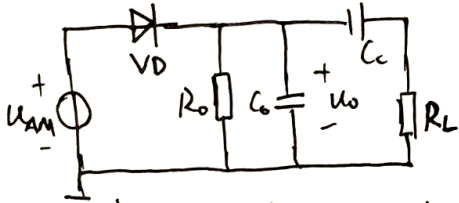
$$\text{纹波系数 } \gamma = \frac{\sqrt{\sum_{n=1}^{\infty} U_n^2}}{U_0} \approx \frac{U_1}{U_0} = \frac{2}{\omega_c R_0 C_0}$$

γ 越大，锯齿越大

(4) 检波失真(峰值包络检波) —— 失随失真



原因: 包络下降速率大于输出电压下降速率
不失真要求: $|\frac{du_o(t)}{dt}| \geq |\frac{d|b(t)|}{dt}|$, 即输出下降大于包络下降速率
调制度 m 越大, ω_{ct} 频率越高, 越容易失随失真

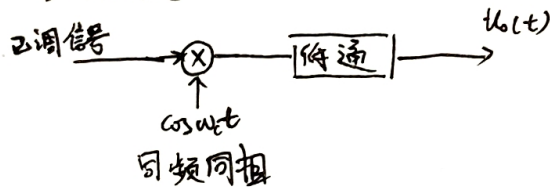


实际的带负载峰值包络检波器

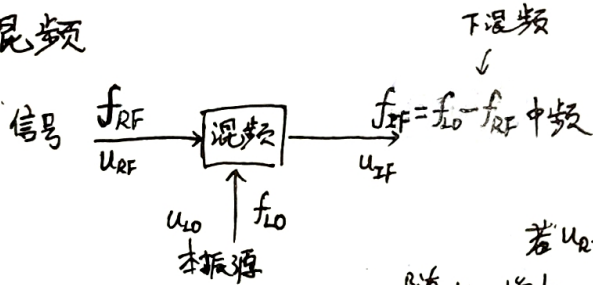
不失真条件: $m \leq \frac{k}{\sqrt{1 + (\Omega_{max} \tau)^2}}$

其中 $k = \frac{R_0 // R_L}{R_0}$ 为检波器交流负载与直流负载之比
 Ω_{max} 为 ω_{ct} 的最高频率, $\tau = (R_0 // R_L) C_0$, 为 C_0 放电时间常数

同步检波



七. 混频



混频电压增益 $A_u = \frac{U_{IF}}{U_{RF}}$

混频功率增益 $A_p = \frac{U_{IF}^2 / 2R_L}{U_{RF}^2 / 2R_S} = A_u^2 \frac{R_S}{R_L}$

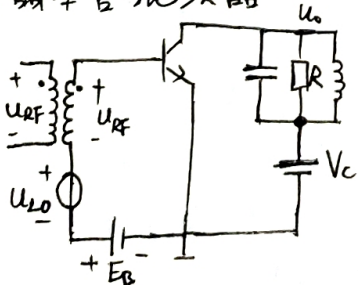
$A_p > 1$ 为有源混频器; $A_p < 1$ 为无源混频器

若 u_{RF} 幅度太大, 则中频电压将不与 u_{RF} 成正比, 产生非线性失真
随 u_{RF} 增大, A_p 将减小, 把减小 1dB 时的 P_{RF} 称为 -1dB 压缩点

噪声系数 $NF = \frac{SNR_i}{SNR_o} = \frac{P_i / N_i}{P_o / N_o} = \frac{\text{输入信噪比}}{\text{输出信噪比}}$

有源混频器

1. 晶体管混频器



条件: 满足 $U_{LO} \gg U_{RF} \Rightarrow$ 时变参量法

$\omega_{IF} = \omega_{LO} - \omega_{RF}$

$u_{BE} = E_B + U_{LO} \cos \omega_{LO} t + U_{RF} \cos \omega_{RF} t$

$i_c(t) = I_{c0}(\omega_{LO} t) + g(\omega_{LO} t) \cdot U_{RF} \cos \omega_{RF} t$

$g(\omega_{LO} t) = \frac{\partial i_c}{\partial u_{BE}} \Big|_{u_{BE} = E_B + U_{LO} \cos \omega_{LO} t} = \frac{\alpha I_{ES}}{U_T} e^{\frac{E_B}{U_T}} e^{\frac{U_{LO}}{U_T} \cos \omega_{LO} t}$

$g_0 = \frac{\alpha}{U_T} \cdot I_{ES} e^{\frac{E_B}{U_T}} I_0(x)$ $g_1 = \frac{2\alpha}{U_T} I_{ES} e^{\frac{E_B}{U_T}} I_1(x)$ $g_1 = g_0 \frac{2I_1(x)}{I_0(x)}$

$i_c(t) = I_{c0} + I_{c1} \cos \omega_{LO} t + I_{c2} \cos 2\omega_{LO} t + \dots + (g_0 + g_1 \cos \omega_{LO} t + g_2 \cos 2\omega_{LO} t + \dots) U_{RF} \cos \omega_{RF} t$

$I_{c,IF} = \frac{1}{2} g_1 U_{RF}$ $u_o = V_{CC} - I_{c,IF} R \cos \omega_{IF} t$ 混频跨导: $g_c = \frac{I_{c,IF}}{U_{RF}} = \frac{1}{2} g_1$

输入电抗 $g_{in} = \frac{I_{B,RF}}{U_{RF}} = \frac{g_0 U_{RF}}{\beta U_{RF}} = \frac{g_0}{\beta}$

3. 交叉调制干扰：外界干扰信号 $V_n(t)$ 和有用信号 $V_s(t)$ 在混频时相互作用产生的干扰

$$v = a_0 + a_1 v + a_2 v^2 + a_3 v^3 + \dots \quad v = V_L \cos \omega_L t + V_S(t) \cos \omega_S t + V_n(t) \cos \omega_n t$$

中频项 $3a_4 V_L V_n^2(t) V_S(t)$ 为交叉干扰项

产生干扰时同时听到有用信号和干扰声音，有用信号停止播音时，干扰信号也随之消失。

4. 互调干扰：两个或多个外界干扰信号 $V_n(t)$ 同时进入混频器而产生的干扰

$$\text{混频器输出频率 } f_{out} = p f_{LO} + q f_{RF} + r f_{n1} + s f_{n2}$$

$$\left. \begin{aligned} \text{干扰条件: } r f_{n1} + s f_{n2} &= f_{RF} \\ \text{又 } f_{n1} &\approx f_{n2} \approx f_{RF} \end{aligned} \right\} \Rightarrow r + s = 1 \quad \text{幅度最大的为3阶互调干扰: } |r| + |s| = 3$$

八、角度调制与解调

调相波

归一化基带信号

$$\text{高频正弦载波: } u(t) = U \cos \omega_0 t \quad \text{基带信号 } u_{\Omega}(t) = U_{\Omega} s(t)$$

$$\text{调相波表达式 } u_{PM} = U \cos [\omega_0 t + k_p U_{\Omega} s(t)] = U \cos [\omega_0 t + m_p s(t)] \quad \varphi(t) = \omega_0 t + m_p s(t)$$

其中 k_p 为调相增益，表示单位电压引起的相位变化； m_p 为调相指数，表示基带信号引起的最大相移

$$\omega = \frac{d\varphi(t)}{dt} = \omega_0 + m_p \frac{ds(t)}{dt}, \text{ 即 } \frac{ds(t)}{dt} \text{ 越大, } u_{PM} \text{ 频率越高}$$

调频波

$$\text{高频正弦载波: } u(t) = U \cos \omega_0 t \quad \text{基带信号 } u_{\Omega}(t) = U_{\Omega} s(t)$$

$$\text{调频波瞬时频率 } \omega(t) = \omega_0 + k_f U_{\Omega} s(t) = \omega_0 + \Delta \omega s(t) \quad D = \frac{\Delta \omega}{\omega_0} \text{ 为相对频偏}$$

其中 k_f 为调频增益，表示单位电压引起的频率变化； $\Delta \omega$ 为最大频率偏移

$$\text{调频波表达式: } u_{FM}(t) = U \cos \left[\int_0^t \omega(\tau) d\tau \right] = U \cos \left[\omega_0 t + \Delta \omega \int_0^t s(\tau) d\tau \right]$$

$$\text{当 } s(t) = \cos \Omega_{max} t \text{ 时, } u_{FM}(t) = U \cos \left[\omega_0 t + \frac{\Delta \omega}{\Omega_{max}} \sin \Omega_{max} t \right] = U \cos \left[\omega_0 t + m_f \sin \Omega_{max} t \right]$$

其中 m_f 为调频指数，表示基带信号引起的最大相移

调频波频谱

$$u_{FM} = U \cos \omega_0 t \cos(m_f \sin \Omega_{max} t) - U \sin \omega_0 t \sin(m_f \sin \Omega_{max} t)$$

$$\cos(m_f \sin \Omega_{max} t) = J_0(m_f) + 2 \sum_{n=1}^{\infty} J_{2n}(m_f) \cos 2n \Omega_{max} t, \text{ 只含偶次项, 为 } \cos$$

$$\sin(m_f \sin \Omega_{max} t) = 2 \sum_{n=1}^{\infty} J_{2n+1}(m_f) \sin(2n+1) \Omega_{max} t, \text{ 只含奇次项, 为 } \sin$$

$$\text{由 } J_{-n}(m_f) = (-1)^n J_n(m_f) \text{ 可得: } u_{FM} = U \sum_{n=-\infty}^{\infty} J_n(m_f) \cos(\omega_0 + n \Omega_{max}) t$$

m_f 一定时， $J_n(m_f)$ 随着 $|n|$ 增大，总的趋势减小，太小时可忽略，故可度量 u_{FM} 的 BW

定义 Carson 带宽： $BW_{CR} = 2(m_f + 1) \Omega_{max}$

亦可估计调相波带宽： $BW_{CR} = 2(m_p + 1) \Omega_{max}$

调频波功率 (单位电阻上)

$$u_{FM} = U \sum_{n=-\infty}^{\infty} J_n(m_f) \cos(\omega_0 + n\Omega_{max})t$$

$$P_{FM} = \sum_{n=-\infty}^{\infty} \frac{U^2 J_n^2(m_f)}{2} = \frac{U^2}{2} \sum_{n=-\infty}^{\infty} J_n^2(m_f) = \frac{U^2}{2} \quad \text{与载波功率相等}$$

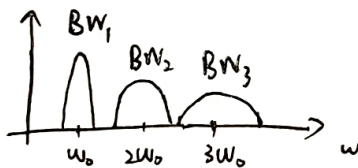
调频信号通过非线性电路

1. 倍频网络

$$u_{FM} = U \cos[\omega_0 t + \Delta\omega \int_0^t s(\tau) d\tau] = U \cos[\omega_0 t'] \quad t' = t + \frac{\Delta\omega}{\omega_0} \int_0^t s(\tau) d\tau = t + D \int_0^t s(\tau) d\tau$$

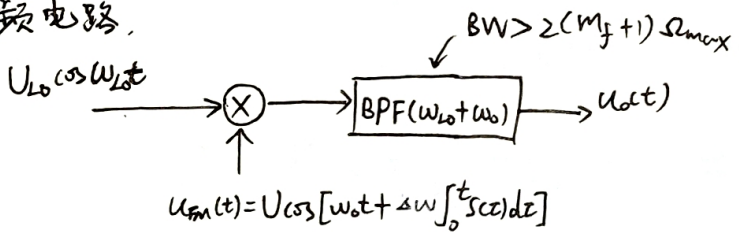
n 倍频后 $\rightarrow u_{FM} = U \cos[n\omega_0 t'] = U \cos[n\omega_0 t + n\Delta\omega \int_0^t s(\tau) d\tau]$, 调频波参数发生变化

$$\text{带宽 } BW_n = 2(m_{fn} + 1)\Omega_{max} = 2(nm_f + 1)\Omega_{max} \quad \text{变大}$$



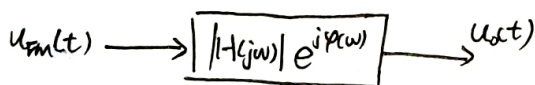
$$n \text{ 次谐波可分离条件: } \frac{BW_n}{2} + \frac{BW_{n+1}}{2} < \omega_0$$

2. 混频电路



$$u_o(t) = \frac{1}{2} U_{L0} U_{FM} \cos[(\omega_{L0} + \omega_0)t + \Delta\omega \int_0^t s(\tau) d\tau]$$

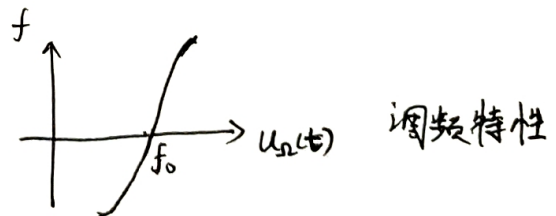
调频信号通过线性网络



$$\text{满足准静态条件时 (略), } u_o(t) = U \cdot |H(j\omega(t))| \cos\left\{ \omega_0 t + \Delta\omega \int_0^t s(\tau) d\tau + \arg[H(j\omega(t))] \right\}$$

调频信号产生

$$\text{调制灵敏度 } k_f = \left. \frac{df}{du_o(t)} \right|_{f=f_0}$$



调频方法

1. 直接调频

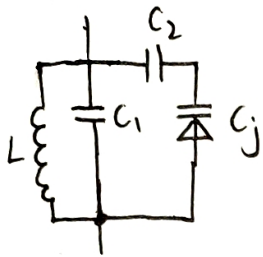
$s(t) \rightarrow \text{OSC}$ 使用 $s(t)$ 线性控制 OSC 的振荡频率 (控制变容=极管)

$$\text{元件: 变容=极管 } \frac{1}{C} \quad C_j = C_{j0} \left(1 + \frac{u}{U_\phi}\right)^{-\gamma} \quad \gamma \text{ 为变容指数}$$

$$u = U_Q + U_a \cos \Omega_{max} t \Rightarrow C_j = C_{jQ} (1 + M \cos \Omega_{max} t)^{-\gamma}$$

$$\text{工作点处结电容 } C_{jQ} = C_{j0} \left(1 + \frac{U_Q}{U_\phi}\right)^{-\gamma} \quad M = \frac{U_a}{U_Q + U_\phi} \text{ 为归一化交流电压幅度}$$

变容二极管谐振频率分析



$$\omega_0 = \frac{1}{\sqrt{L(C_1 + \frac{C_2 C_{jQ}}{C_2 + C_{jQ}})}} \quad A = \frac{C_{jQ}}{C_{jQ} + C_{open}} \quad B = \frac{C_{jQ}}{C_{jQ} + C_{close}}$$

C_{open} : 电感开路时, 变容管外部电容, 这里 $C_{open} = C_1 \text{ 串 } C_2$

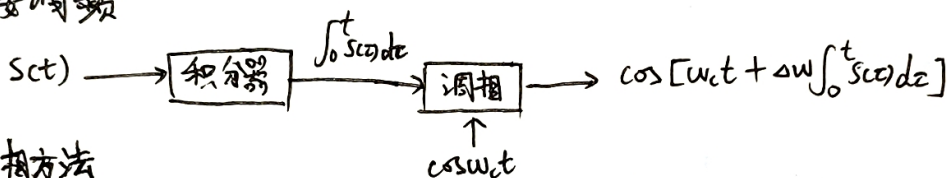
C_{close} : 电感短路时, 变容管外部电容, 这里 $C_{close} = C_2$

$$u = U_Q + U_{\Omega}(t)$$

$$\omega \approx \omega_0 + D_1 \omega_{\Omega}(t) \quad D_1 = \frac{1}{2} M \gamma (A - B) \quad \text{代表相对频偏} \quad \Delta \omega = D_1 \omega_{\Omega}$$

$$\text{中心频率漂移 } \varepsilon = \frac{1}{2} D_2, \quad \text{调频二次谐波失真 } \gamma = \frac{D_2}{2D_1}$$

2. 间接调频

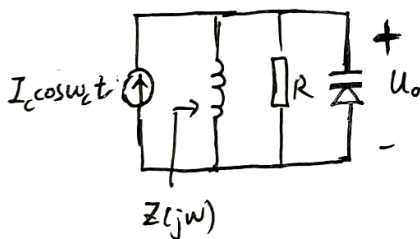


简单积分器



调相方法

(1) 谐振回路调相



$$u = U_c + U_{\Omega}(t)$$

谐振频率
 $\omega_i \approx \omega_c + \omega_c D_1 s(t)$

$$\omega_0 = \frac{1}{\sqrt{LC_{jQ}}}$$

$$Z(j\omega) = \frac{R}{1 + jQ(\frac{\omega_c}{\omega_i} - \frac{\omega_i}{\omega_c})} \approx \frac{R}{1 + 2jQ\frac{\omega_c - \omega_i}{\omega_0}} \approx \frac{R}{1 - 2jQD_1 s(t)}$$

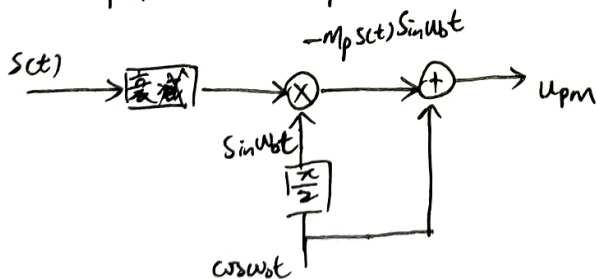
$$\arg Z(j\omega) = \arctan[2QD_1 s(t)] \approx 2QD_1 s(t)$$

$$m_p = 2QD_1 = QM\gamma(A-B) \quad U_o = IR \cos[\omega_0 t + m_p s(t)]$$

(2) 合成矢量调相

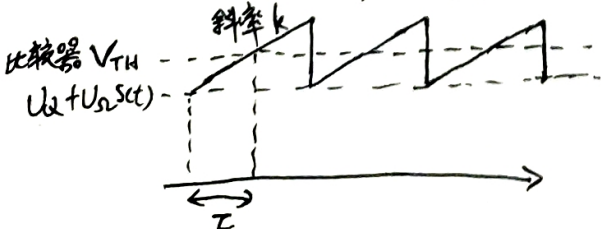
$$U_{pm} = U_{pm} \cos[\omega_0 t + m_p s(t)]$$

$$= U_{pm} \cos \omega_0 t \cos[m_p s(t)] - U_{pm} \sin \omega_0 t \sin[m_p s(t)] \xrightarrow{m_p \ll 1} U_{pm} \cos \omega_0 t - U_{pm} \sin \omega_0 t \cdot m_p s(t)$$



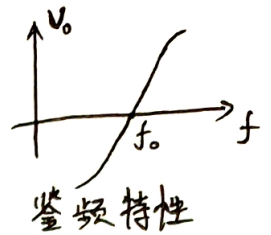
(3) 时延法调相

$$\cos \omega_0(t - \tau) = \cos(\omega_0 t - \tau \omega_0) \xrightarrow{\tau \propto s(t)} \cos[\omega_0 t - m_p s(t)]$$



$$\tau = \frac{V_{TH} - [U_Q + U_{\Omega}(t)]}{k} = \tau_0 - \frac{U_{\Omega}}{k} s(t)$$

鉴频原理与电路

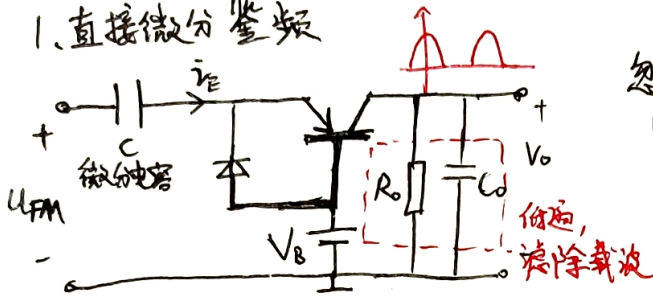


原理：频率变化 $\xrightarrow{\text{转化为}}$ 电压变化 \rightarrow 幅度解调

鉴频器指标：① 线性度： $V_o = a\omega + b$ ② 鉴频宽度： $BW_F > BW_{CR}$ ③ 鉴频灵敏度 $k = \left. \frac{dV_o}{df} \right|_{f=f_0}$

鉴频方法：

1. 直接微分鉴频



忽略三极管导通电压和二极管导通电压。
 U_{FM} 负半周时二极管导通使 C 放电，故不会产生负电压

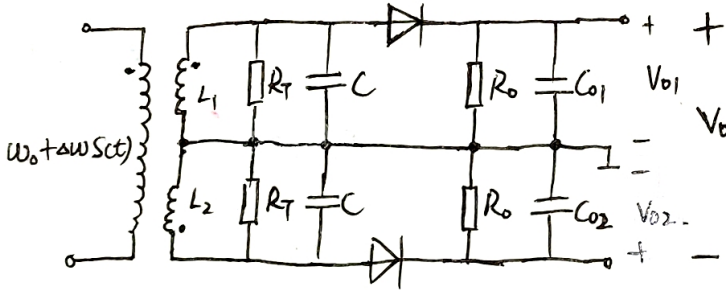
$$i_C = C \frac{dU_C}{dt} = C \frac{d(U_{FM} - V_B)}{dt}$$

$$= -C V_B \sin[\omega_0 t + \omega \int_0^t sct dz] [\omega_0 + \omega sct]$$

$$= -C V_B \omega_i(t) \sin[\omega_0 t + \omega \int_0^t sct dz]$$

输出端需要的是直流分量： $I_{C0} = \frac{I_{CP}}{\pi} = \frac{\alpha C V_B \omega_i(t)}{\pi}$ $U_o = I_{C0} R_o$

2. 斜率鉴频器鉴频 (利用线性网络的幅频特性完成 FM-AM)



$$\frac{1}{\sqrt{L_1 C}} = \omega_1 = \omega_0 + \delta$$

$$\frac{1}{\sqrt{L_2 C}} = \omega_2 = \omega_0 - \delta$$

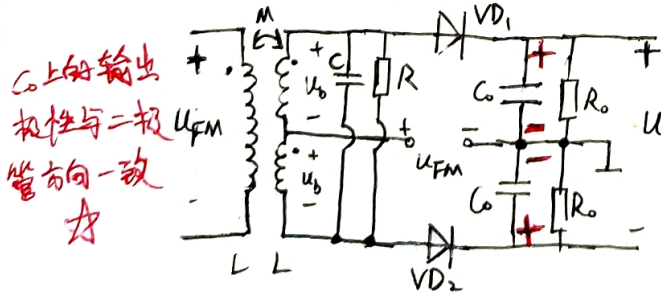
工作在双失谐状态

$$Z_1(j\omega) \approx \frac{R}{1 + j \frac{\omega - \omega_1}{\alpha}} = \frac{R}{1 + j [\frac{\Delta \omega sct}{\alpha} - \frac{\delta}{\alpha}]}$$

$$Z_2(j\omega) \approx \frac{R}{1 + j(\alpha + \omega)}$$

I 为流过回路的电流 $V_o = I|Z_1| - I|Z_2| = I(|Z_1| - |Z_2|)$ \rightarrow 展开求解

3. 相位鉴频器 (叠加型)



C_o 上的输出
 极性与二极
 管方向一致
 ☆

$$\dot{U}_{D1} = \dot{U}_{FM} + \dot{U}_b \quad \dot{U}_{D2} = \dot{U}_{FM} - \dot{U}_b \quad (|\dot{U}_b| = |\dot{U}_{FM}| = U)$$

$$|\dot{U}_{D1}| = \sqrt{2} U \sqrt{1 - \sin\theta}$$

$$|\dot{U}_{D2}| = \sqrt{2} U \sqrt{1 + \sin\theta}$$

$$u_o(t) = |\dot{U}_{D1}| - |\dot{U}_{D2}| = -2\sqrt{2} U \sin\frac{\theta}{2}$$

$$\approx -\sqrt{2} U \frac{\Delta \omega}{\alpha} sct$$

改进型相位鉴频器：略。

$$|H(j\omega)| = \frac{U_b}{U_{FM}}$$

$$|H(j\omega)| \approx 1$$

$$\arg H(j\omega) \approx -\frac{\pi}{2} - \frac{\Delta \omega}{\alpha} sct = -\frac{\pi}{2} - \theta$$