Solution Written Examination EITP01 2019-03-21

1. Ballistic Transport in QWFET
   a. (2p) $E_n = \frac{\hbar^2\pi^2 n^2}{2m^* t_w}$ → E_1 = 0.38 eV, E_2 = 1.51 eV
   b. (3p) C_{ox} = 0.035 F/m^2, C_C = 0.069 F/m^2, C_q = 0.027 F/m^2

      In on-state $C_G = \left(\frac{1}{C_{ox}} + \frac{1}{C_C} + \frac{1}{C_q}\right)^{-1}$ = <u>0.013 F/m^2.</u>

      However, in saturation $C_G = \left(\frac{1}{C_{ox}} + \frac{1}{C_C} + \frac{2}{C_q}\right)^{-1}$ =0.0085 F/m^2. The question was not clear on what was asked for, so both answers are correct. However, the result in saturation should be used in d-f.
   c. (3p) As the drain bias increases, the drain fermi level is pulled down and less and less carriers will be injected from the drain into the channel. The reason is that the top of the band in the channel ends up further and further above the drain fermi level where in the limit a negligible amount of electrons reside. Then the only carriers in the channel will be the carriers injected from the source traveling from source to drain. Assuming no short channel effects, this amount of electrons is not affected by further change in drain bias, thus the current saturates.
   d. (3p). V_T = 3.0 - 3.2 + 0.38 V = 0.18 V. $V_{ds,sat} = \frac{V_{gs}-V_T}{1+\frac{C_q}{2C'_{ox}}}$= 0.21 V < V_{DS} → Yes.
   e. (2p) $I_{DS} = T * I_{DS,ballistic}$ = 0.6*15.7mA = <u>10 mA</u>
   f. (2p) $g_m = \frac{\partial I_{DS}}{\partial V_{GS}} = \frac{\partial}{\partial V_{GS}}\left(T * I_{DS,ballistic}\right)$ = 44 mS

2. High frequency operation
   a. (4p) Implicitly assumes operation in saturation → Use $C_q/2$ → C_G = 0.0082 F/m^2→ C_{gg} = C_G*W*L_G = 7.4 fF. $f_T = \frac{C_{gg}}{2\pi g_m}$ = 1.2 THz (1190 GHz)
   b. (6p) $R_S = R_D$ = 62 Ohm. ($R_{contact}$ = 16 Ω, R_{access} = 3.1 Ω, R_{spacer} = 43 Ω). The undoped spacer region is the main contributor to the resistance.

$$\rho_{s,access} = \frac{1}{\sigma_{s,access}} = \frac{1}{q n_{access} \mu_{access}} = 1.25 * 10^{-5}\ \Omega m$$
$$\rho_{s,spacer} = \frac{1}{\sigma_{s,spacer}} = \frac{1}{q n_{spacer} \mu_{spacer}} = 5.2 * 10^{-4}\ \Omega m$$

$$R_{SH} = \frac{\rho_{s,access}}{t_w} = 2080\ \Omega, L_T = \sqrt{\frac{\rho_\sigma}{R_{SH}}} = 69\ nm$$

$$R_{contact} = \frac{\sqrt{R_{SH}\rho_\sigma}}{W}\coth\left(\frac{L_C = 100\ nm}{L_T = 69\ nm}\right) = 16\ \Omega$$

$$R_{access} = \frac{\rho_{s,access} L_{access}}{t_w W} = \frac{1.25E-5*15nm}{6nm*10\mu m} = 3.1\ \Omega$$
$$R_{spacer} = \frac{\rho_{s,spacer} L_{spacer}}{t_w W} = \frac{5.2E-4*5nm}{6nm*10\mu m} = 43$$

c. (5p) $C_{gg} = C_{gg,0}$ (from 2a) + $C_{gs,p}$ + $C_{gd,p}$ = 17.4 fF. Add R_S=R_D= 62 Ohm → f_T = 138 GHz.

3. Power gain
   a. (2 p) The finite transconductance of the channel in a ballistic transistor induces a phase shift between input (gate) and output (drain) as well as limiting the power transfer. This is modelled by a channel resistance R_i.
   b. (5 p) Parameters excluding R_G, R_D, and R_S:

$$y_{11} = \left(\frac{1}{j\omega C_{gs}} + R_i\right)^{-1} + j\omega C_{gd}$$
$$y_{12} = -j\omega C_{gd}$$
$$y_{21} = \frac{g_m}{1 + j\omega C_{gs} R_i} - j\omega C_{gd}$$
$$y_{22} = g_d + j\omega C_{gd}$$

$$\rightarrow Y = \begin{bmatrix} 0.0010 + 0.011j & -0.0031j \\ 0.040 - 0.0084j & 0.0050 + 0.0031 \end{bmatrix}$$

$$\rightarrow Z = \begin{bmatrix} 17.26 - 28.1j & 17.6 + 0.16j \\ 49.1 + 219j & 58.8 - 6.1j \end{bmatrix}$$

Add $Z_{res} = \begin{bmatrix} 70 & 50 \\ 50 & 100 \end{bmatrix} \rightarrow Z' = Z + Z_{res} = \begin{bmatrix} 87 - 28j & 68 + 0.16j \\ 99 + 219j & 159 - 6.1j \end{bmatrix}$

   c. (2 p) Using z-parameters including R_S, R_D and R_G → k = 1.29 > 1. Yes, it is stable.
   d. (2 p) For a stable device, use MAG. Use z_{21} and z_{12} → MAG = 1.68 = 2.25 dB.
   e. (4 p) f_T = 140 GHz → f_{max} = 195 GHz using model including R_i (197 GHz with standard model). Except for increasing f_T more, one could work on reducing R_G and g_d as well as C_{gd}.

4. Designing a Low noise amplifier
   a. (3 p) $\omega_T = \frac{g_m}{C_{gs}} = 1.9\, THz, \omega = 314\, GHz \ll \omega_T, Z_{opt} = 351 + j303 \rightarrow z_{opt} = 7 + j6$
   $F\_\min = 1.25 \rightarrow NF = 0.96\, dB$
   b. (5 p) The distance from Z_0 (center) to Z_{opt} is 6.5 cm in the Smith chart. Draw a circle with this radius and find the intercept going from Z_{opt} "towards generator" (i.e. clockwise) with the "admittance = 1" at z_1 = 0.09 - 0.28j. The angle between these corresponds to a phase shift (read scales) from 0.239λ to 0.457 λ = 0.218λ. The length of the microstrip TRL should thus be $l_{phys} = \frac{0.218}{\beta} = 0.218 * \frac{c}{\sqrt{\varepsilon_{r,eff}}\omega}$.
   $\varepsilon_{r,eff} = 2.18$ → λ = 4.06 mm → $\beta = \frac{2\pi}{\lambda} = 1.55\, mm^{-1}$ → length of TRL = 141 µm.
   A shorted stub is used to shift to from z_1 to the center (z_0), since z_1 → y_1 = 1 + 3.24j this requires a normalized susceptance of -j3.24 → normalized impedance of j0.31. The characteristic impedance of the stub is Z_{stub} = $\approx \frac{60}{\epsilon_r} \ln\left(\frac{8H}{W} + \frac{W}{4H}\right)$ = 60 Ω → a changing $l_{stub} = \frac{1}{\beta} \text{atan}\left(\frac{Z_0}{Z_{stub}}\right)$ = 160 µm. (one can also use an open stub of length 1.0 mm).
   c. (5 p) The λ/4 transformer should transform from Z_0 = 50 Ω to 351 Ω. → Z_{trans} = 132 Ω. This means W = 54 nm → $\varepsilon_{r,eff} = 2.09$ → λ = 4.1 mm → $\beta = 1.52\, mm^{-1}$ →

$l_{phys} = \frac{0.25}{\beta}$ = <u>165 µm</u>.

Then a positive normalized reactance of 6j is needed to move from z = 351/50 to Z_{opt}. An inductor with L = 0.95 nH in series can be used to shift to Z_{opt}. Alternatively one can view the process in reverse, but have then to start at Z_{opt}^{*}.

d. (2 p) The network in c is more preferable from a robustness perspective since the path to Z_{opt} stays close to the center line of the Smith chart → low Q factor. This means that the design works for a larger frequency band around f = 100 GHz. (However, the narrow width of this line ~ 50 nm means that it may be problematic to make in practice).