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はじめに
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- Shockleyの電流方程式

基礎編：
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- デバイスの基本原理
- 解析式の導出

応用編：
- 先端デバイスの特性
  - 短チャネル効果
  - Non-Quasi-Static効果
  - ノイズ
  - 高調波ひずみ
  - Y-パラメタ特性

基本方程式

-Poisson: $$\nabla^2\phi = -\frac{q}{\varepsilon_{Si}}(N_D - N_A + p - n)$$

$$n = n_i \exp\frac{q(\phi - \phi_n)}{kT}$$

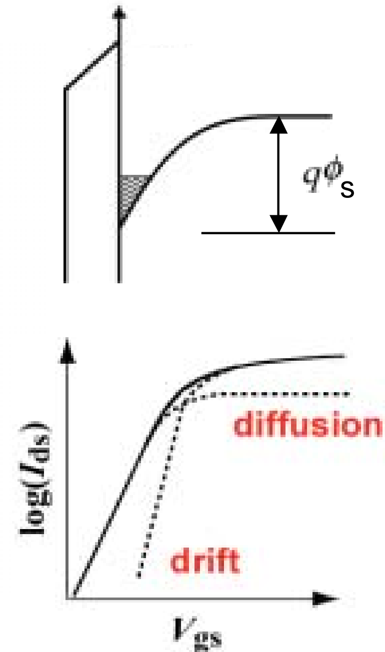
$$p = n_i \exp\frac{q(\phi_p - \phi)}{kT}$$

-Current Density: $$j_n = q\mu_n n\frac{\phi}{y} + qD_n\nabla n$$

$$j_p = q\mu_p p\frac{\phi}{y} - qD_p\nabla p$$

-Continuity: $$I(t) = I_0(t) + \frac{dQ}{dt}$$

(solved by SPICE)



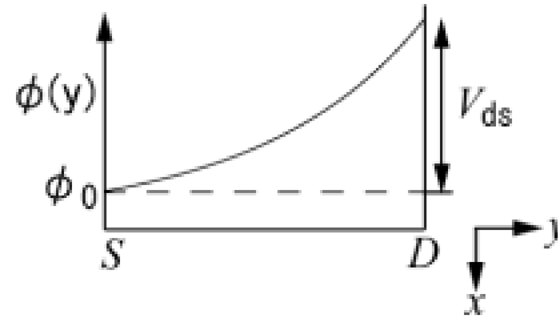
すべてのデバイス特性はポテンシャルの関数

ドリフト近似

$$I_{\mathrm{ds}} = W\mu q n \frac{d\phi}{dy}$$

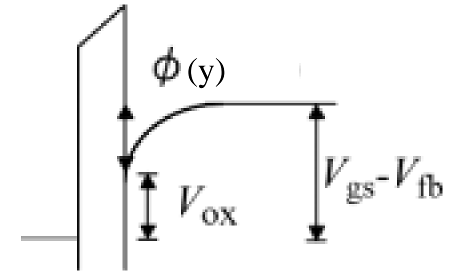
$$I_{\mathrm{ds}} dy = W\mu q n d\phi$$

$$\phi(y) = \phi_0 + V(y)$$



Gaussの法則：$E_{\mathrm{ox}}\varepsilon_{\mathrm{ox}} = E_{\mathrm{Si}}\varepsilon_{\mathrm{Si}} = -Q_{\mathrm{S}}$

$$\frac{V_{\mathrm{gs}} - V_{\mathrm{fb}} - \phi(\mathrm{y})}{T_{\mathrm{ox}}}\varepsilon_{\mathrm{ox}} = -Q_{\mathrm{n}} - Q_{\mathrm{b}}$$



$$Q_{\mathrm{n}} = qn = -\left[V_{\mathrm{gs}} - V_{\mathrm{fb}} - \phi_0 - V(y)\right]C_{\mathrm{ox}} + \sqrt{2\varepsilon_{\mathrm{Si}} q N_{\mathrm{sub}}\left[\phi_0 + V(y)\right]}$$

$$\int_0^L I_{\mathrm{ds}} dy = W\mu \int_0^{V_{\mathrm{ds}}} Q_{\mathrm{n}} dV \Rightarrow I_{\mathrm{ds}} = \mu \frac{W}{L} C_{\mathrm{ox}}\left[\left(V_{gs} - V_{\mathrm{th}}\right)V_{\mathrm{ds}} - \frac{1}{2}V_{\mathrm{ds}}{}^2\right]$$

$$V_{\mathrm{th}} = 2\Phi_{\mathrm{B}} + \frac{\sqrt{2E_{\mathrm{Si}} q N_{\mathrm{sub}} 2\Phi_{\mathrm{B}}}}{C_{\mathrm{ox}}}$$

しきい値条件

Poisson's Equation + Gauss's Law

$$C_{ox}(V'_G - \phi_S(y)) = \sqrt{\frac{2\varepsilon_s q N_{sub}}{\beta}} \left[\exp\{-\beta(\phi_S(y) - V_{bs})\} + \beta(\phi_S(y) - V_{bs}) - 1 + \frac{n_{p0}}{p_{p0}}\{\exp(\beta(\phi_S(y) - \phi_f(y))) - \exp(\beta(V_{bs} - \phi_f(y)))\}\right]^{-\frac{1}{2}}$$

しきい値条件：Carrier濃度が基板濃度と等しくなった時

$$n = n_i \exp\frac{q(\phi - \phi_n)}{kT} = n_{p0}\exp\left(\frac{q\phi_S}{kT}\right)$$

$$n_{p0}p_{p0} = n_i^2 \text{ : Moss-Action Law}$$

$$\Rightarrow \phi_S = \frac{2kT}{q}\ln\left(\frac{N_{sub}}{n_i}\right) = 2\Phi_B$$

しきい値条件下：n_{p0}/p_{p0} の項は無視できる

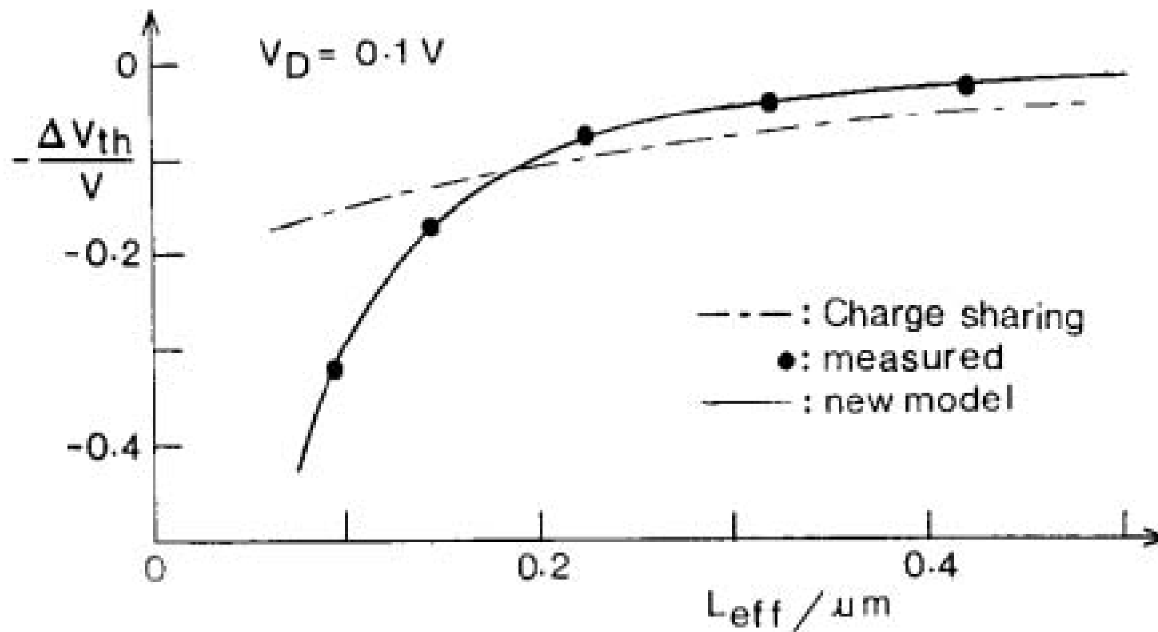
$$C_{ox}(V'_G - 2\Phi_B) = \sqrt{\frac{2\varepsilon_s q N_{sub}}{\beta}}\left[\beta(2\Phi_B) - V_{bs}) - 1\right]^{\frac{1}{2}}$$

$$\Rightarrow V_{th} = V_{fb} + 2\Phi_B + \gamma\left[\beta(2\Phi_B) - V_{bs}) - 1\right]^{\frac{1}{2}}$$

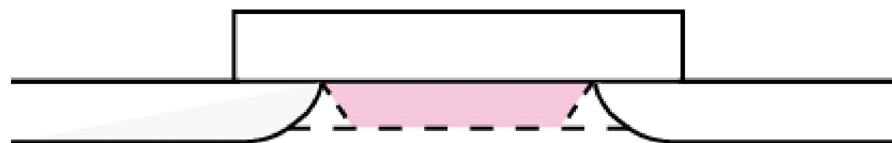
$$\gamma = \sqrt{\frac{2\varepsilon_s q N_{sub}}{\beta}} / C_{ox}$$

: **Body-Effect Coefficient**

しきい値電圧



charge sharingモデル

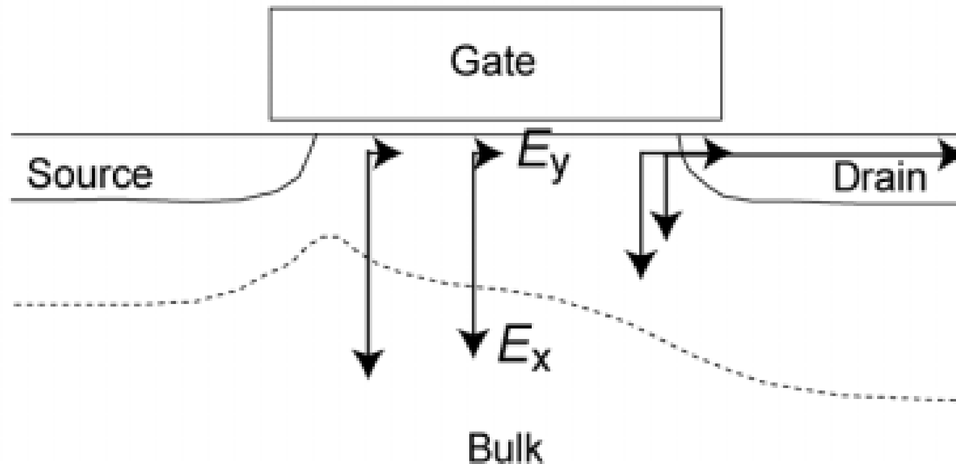


空乏層　チャネルが支配する領域

短チャネル効果

2次元効果：charge-sheet近似の破綻

2次元デバイスシミュレーション結果

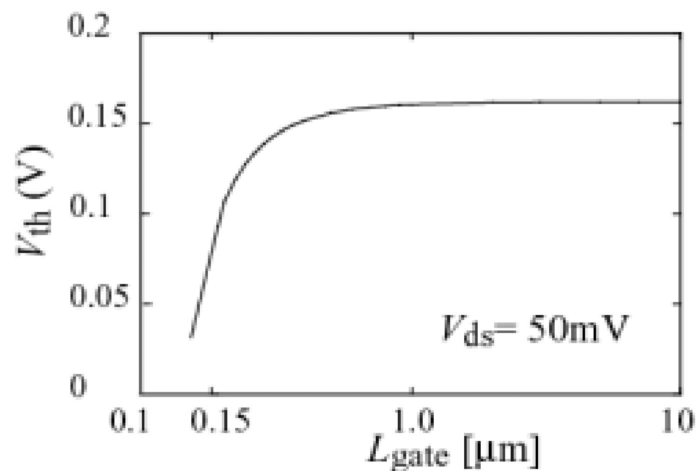


Poisson's Equation

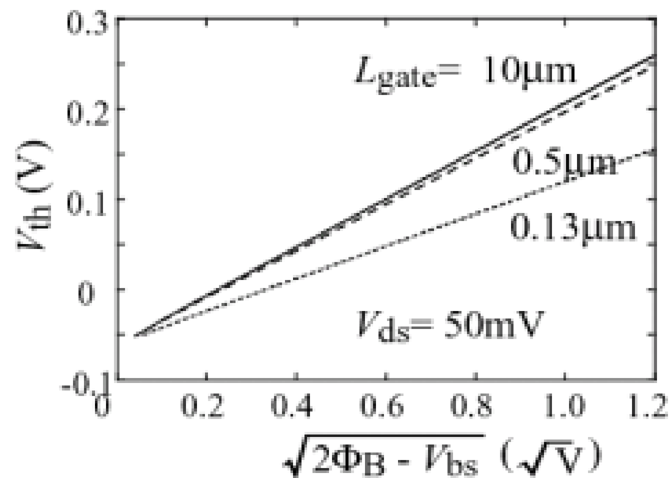
$$\nabla^2\phi = \frac{d^2\phi}{dx^2} + \frac{d^2\phi}{dy^2}$$

観測される短チャネル効果

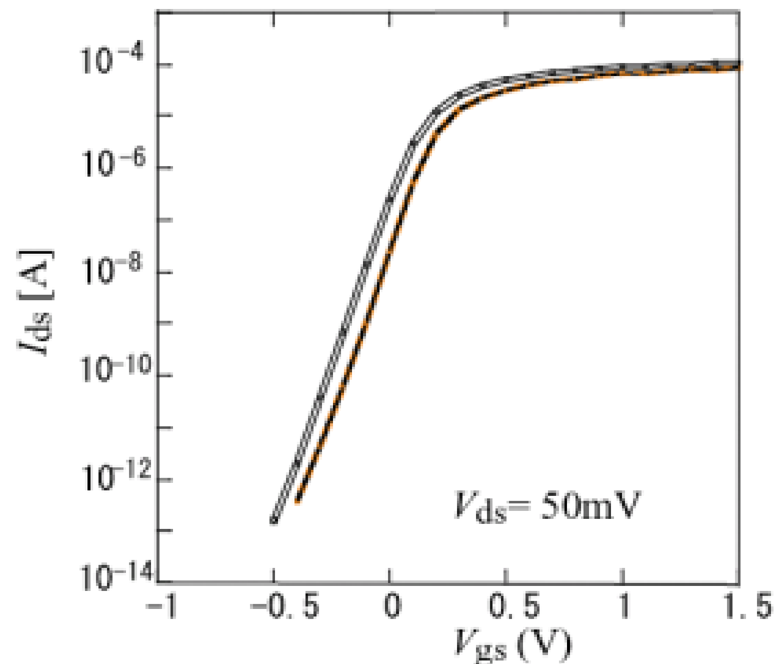
(I) V_{th} - L_{gate} 特性



(II) Body-Effect Coefficient



(III) Subthreshold Swing



(IV) ドレイン電圧依存

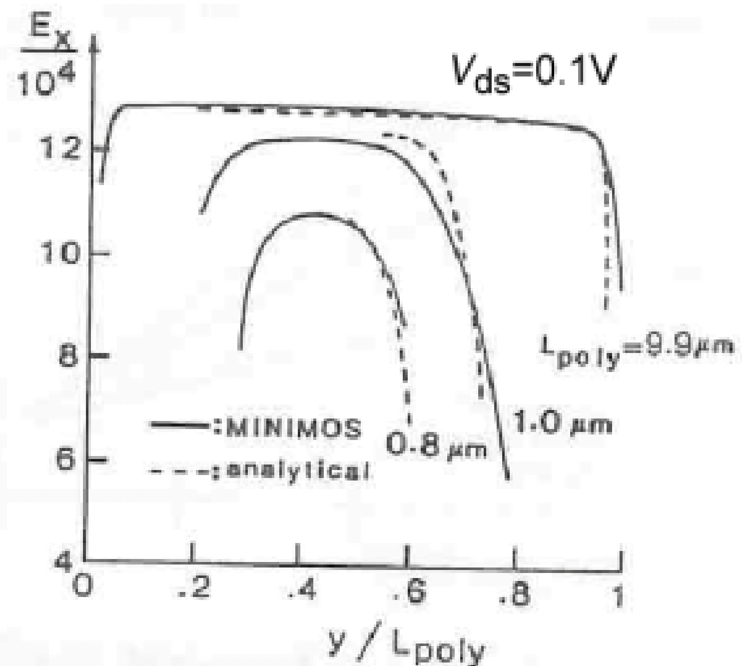
長チャネルの場合：

$$V_{th} = V_{fb} + 2\Phi_B + \sqrt{\frac{2\varepsilon_s qN_{sub}}{\beta}} / C_{ox} \left[\beta\ (2\Phi_B) - V_{bs}) - 1\right]^{\frac{1}{2}}$$

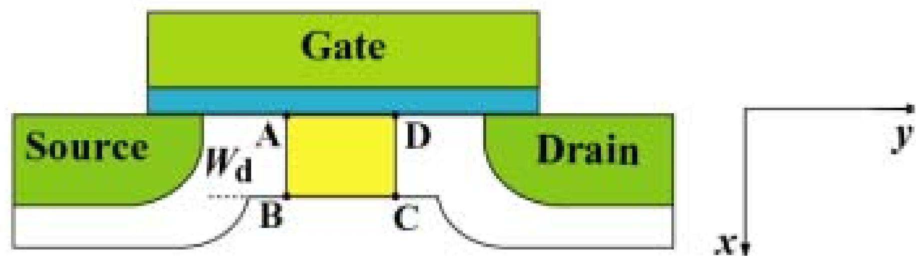
- テクノロジーの指標
- Short-Channel効果の見積もり

しきい値状態での
ゲート・コントロール

Normalizeした電流値が一定



M.Miura-Mattaush and H. Jacobs, Japn J. Appl. phys., vol. 29, p. L2279, 1990.



Gauss's Law:
$$-\int_A^B Ey1\,dx + \int_C^D Ey2\,dx + \int_D^A Ex\,dy = -\int \frac{Q_s}{\varepsilon_{si}} dy$$

$$Ex\,dy + W_d\,(Ey2 - Ey1) = \frac{(Q_b + Q_i)}{\varepsilon_{si}} dy; \quad Ex + W_d \frac{dEy}{dy} = \frac{(Q_b + Q_i)}{\varepsilon_{si}}$$

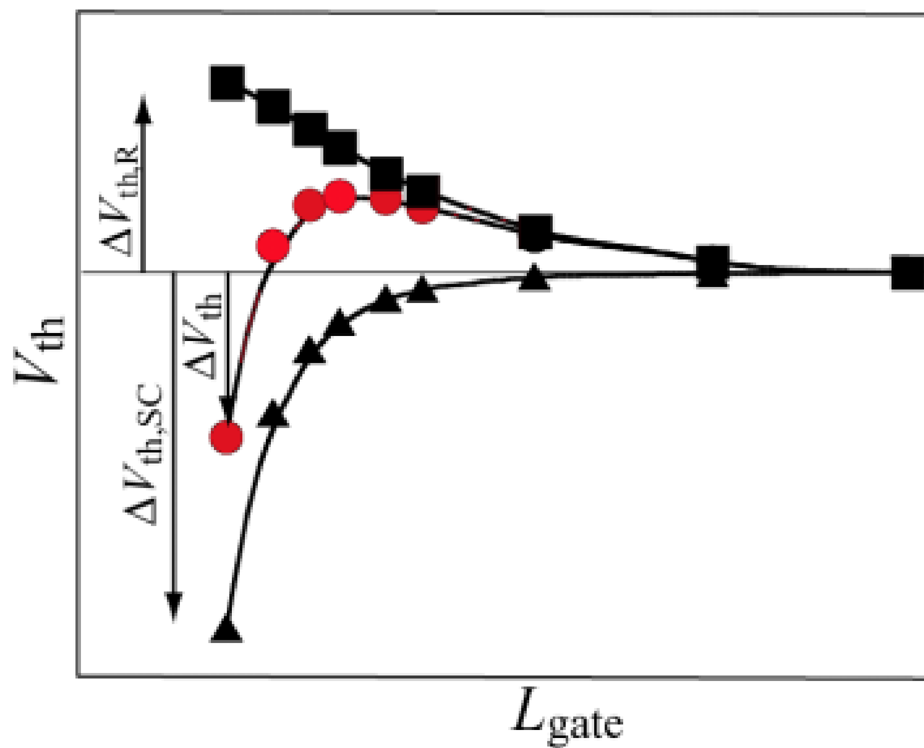
$$\Delta V_{th,SC} = \frac{\varepsilon_{si}}{C_{ox}} W_d \frac{dE_y}{dy}$$

ポテンシャル分布を２次関数近似

$$\frac{dE_y}{dy} = \frac{2(V_{bi} - 2\Phi_B)}{(L_{gate} - PARL2)^2} \left(SC1 + SC2 \times V_{ds} + SC3 \frac{2\Phi_B}{L_{gate}} \right)$$

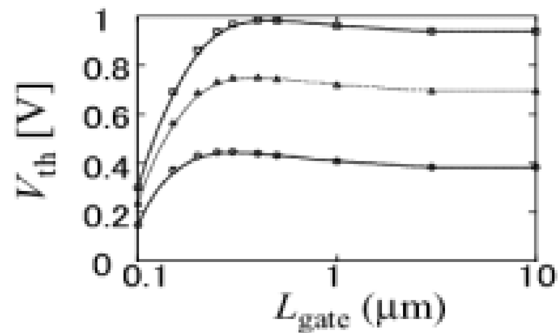
PARL2、*SC1*、*SC2*、*SC3*：モデルパラメタ

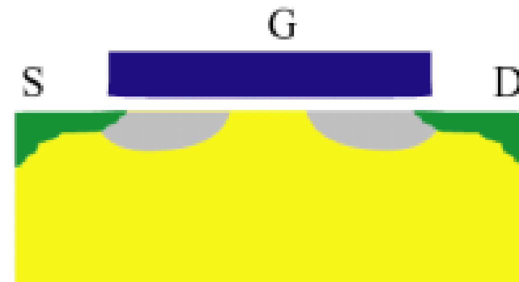
逆短チャネル効果



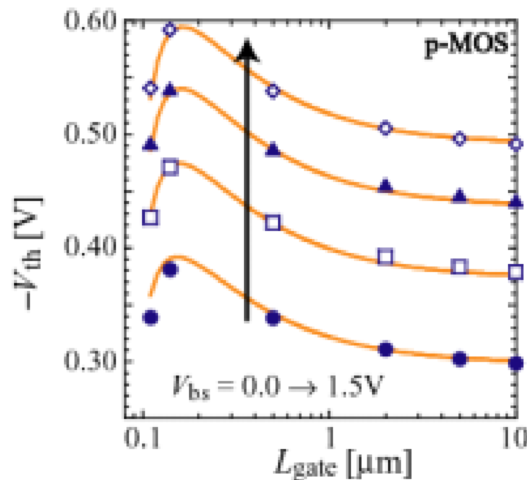
原因：チャネルエンジニアリング

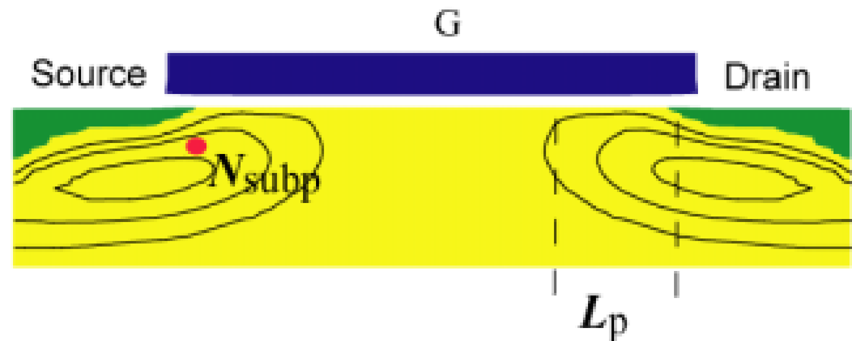
- **Retrograded Case: Impurity Pile-Up**



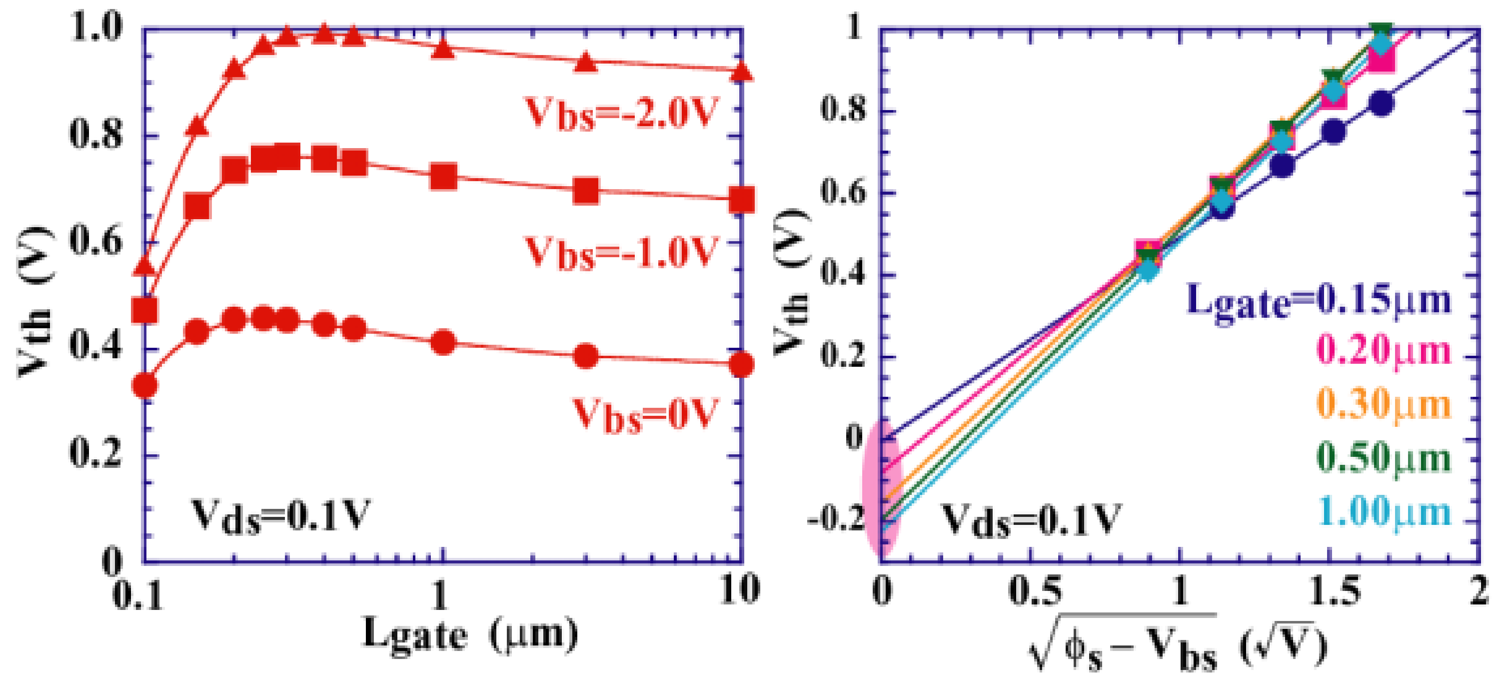


- **Pocket-Implanted Case**



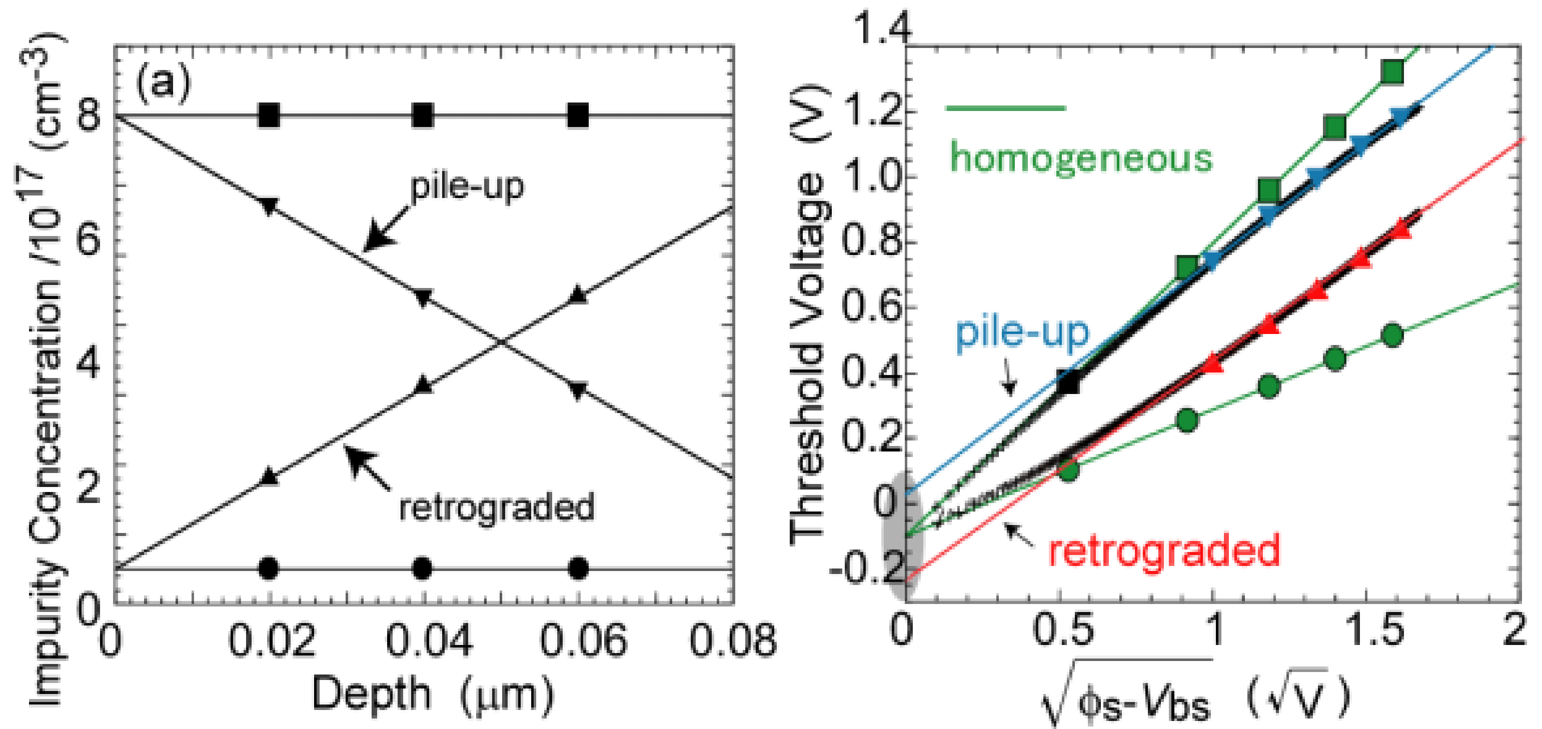


Retrogradedの場合



M. Suetake et al. Proc. SISPAD, p.207, 1999.

Vertical Profile

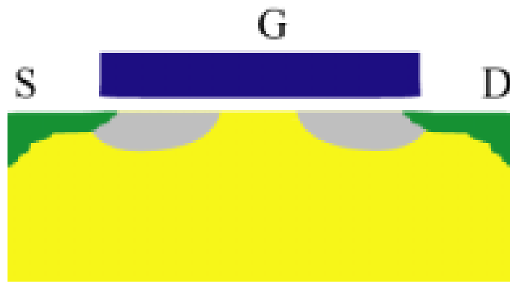


$$V_{th} = V_{fb} + 2\,\Phi_B + \gamma\ [\beta\ (2\,\Phi_B) - V_{bs}) - 1]^{\frac{1}{2}}$$

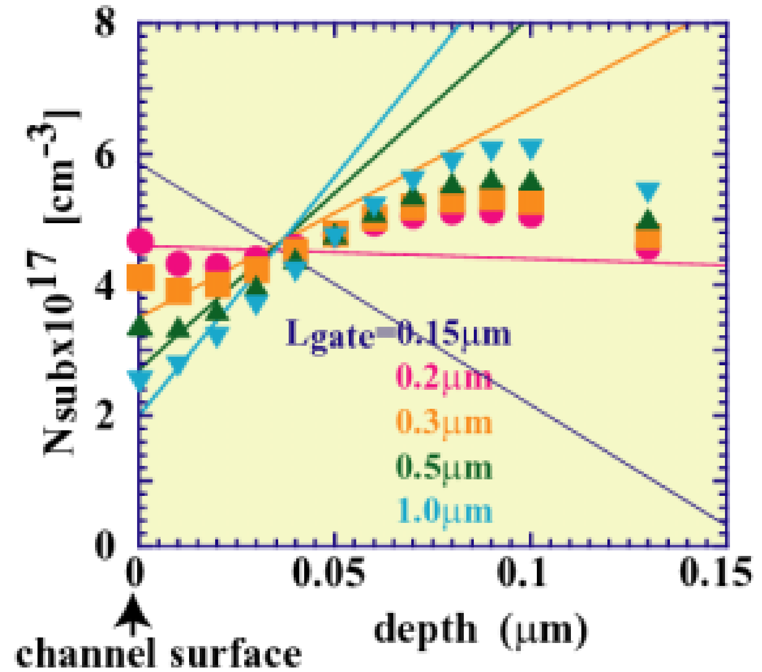
$$\gamma = \sqrt{\frac{2\varepsilon_s\, q N_{sub}}{\beta}} \,/\, C_{ox}$$: Body-Effect Coefficient

H. J. Mattausch et al., Appl. Phys. Lett., 80, p. 2994, 2002.

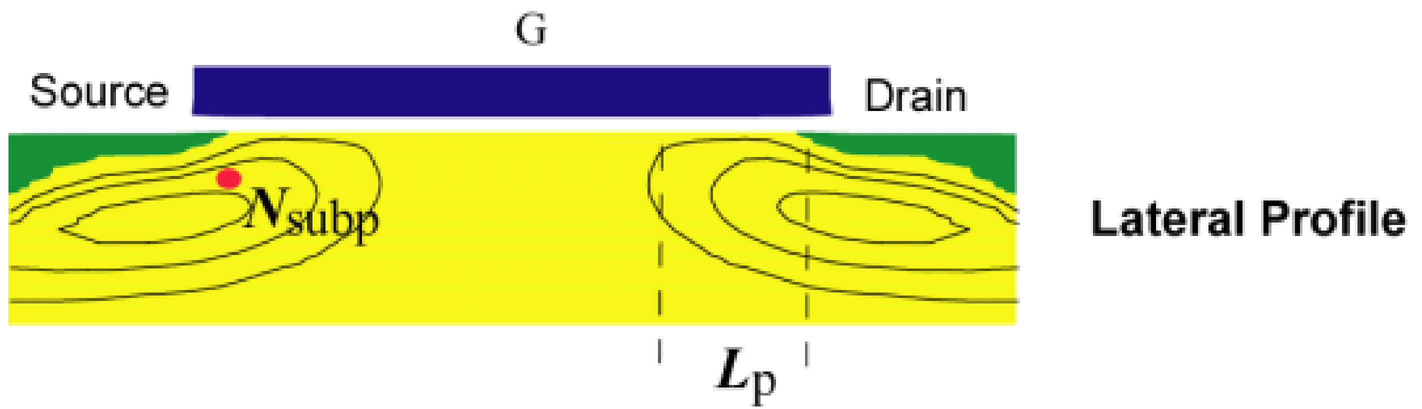
近似の検証



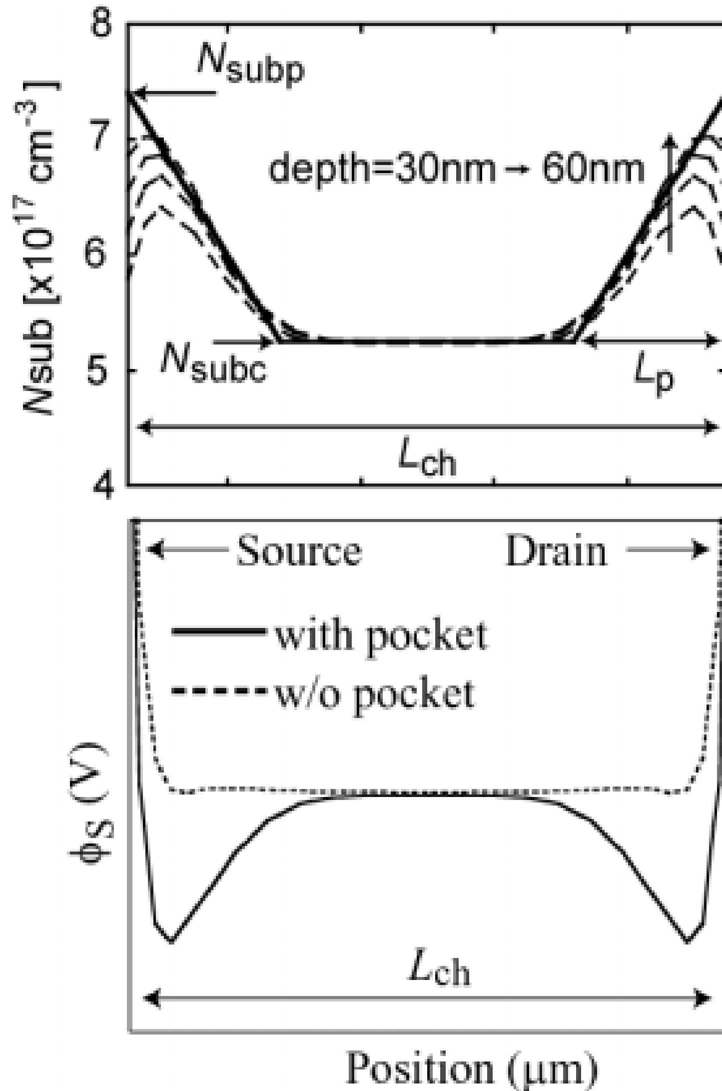
不純物のPileup現象

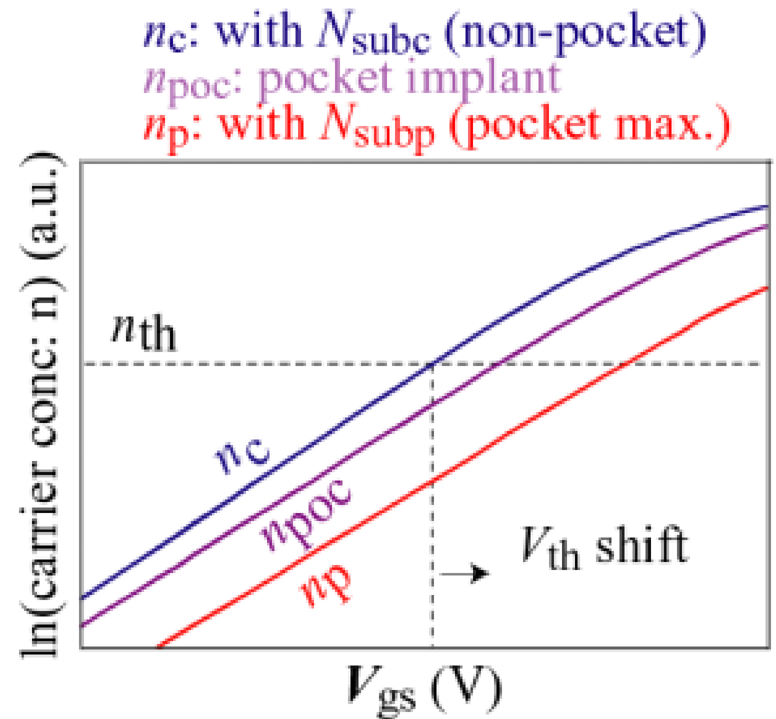


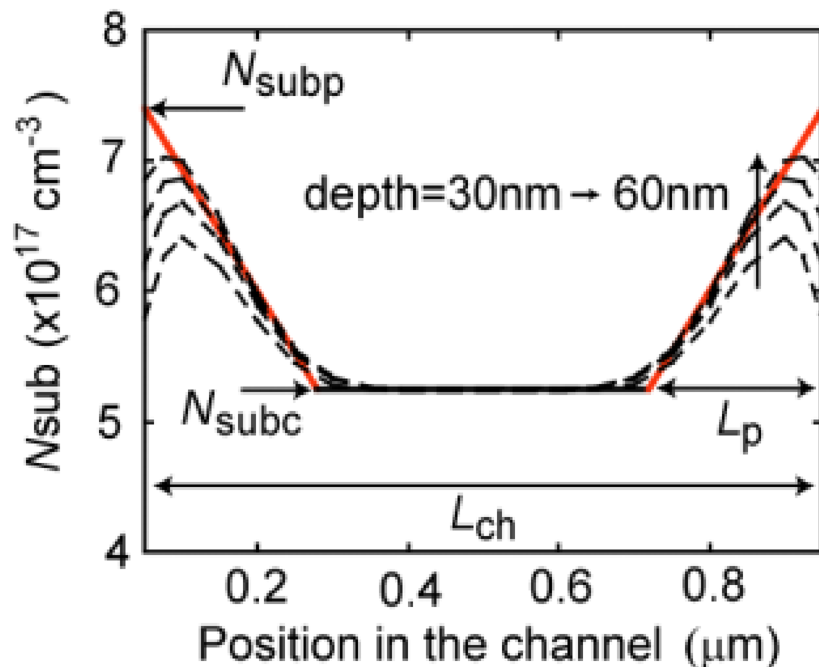
Pocket-Implantの場合



２次元デバイスシミュレーション結果







Simplified Pocket Profile

しきい値条件は全チャネル内キャリア濃度で決まる。

$$n_{av} = \frac{L_{ch}}{\dfrac{L_{ch}-2L_p}{n_c} + \dfrac{2L_p}{n_p}} = n_{th}$$

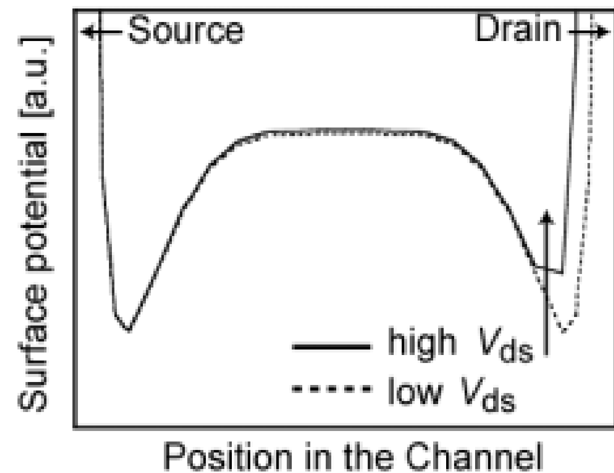
D. Kitamaru et al., Proc. SISPAD, p. 392, 2001.

解析式の導出

$$\phi_{th} = \frac{1}{\beta} \ln \left(\frac{n_{c,\,th}\sqrt{2\beta(2\phi_{Bc} - V_{bs}) - 2}}{\sqrt{\frac{N_{\mathrm{subc}}\epsilon_{\mathrm{Si}}}{q\beta}} \left(\frac{n_i}{N_{\mathrm{subc}}}\right)^2} \right)$$

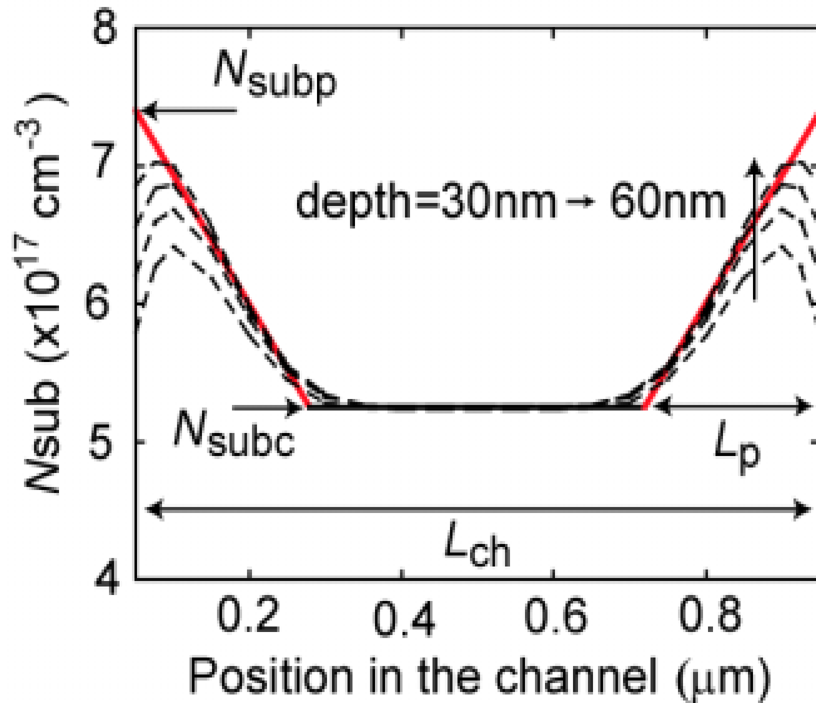
$$n_{c,\,th} = \left(L_{ch} + 2L_p \left(\frac{n_c}{n_p'} - 1 \right) \right) \frac{n_{th}}{L_{ch}}$$

$$n_p' = \frac{3n_c n_p}{n_c + \sqrt{n_c n_p} + n_p}$$

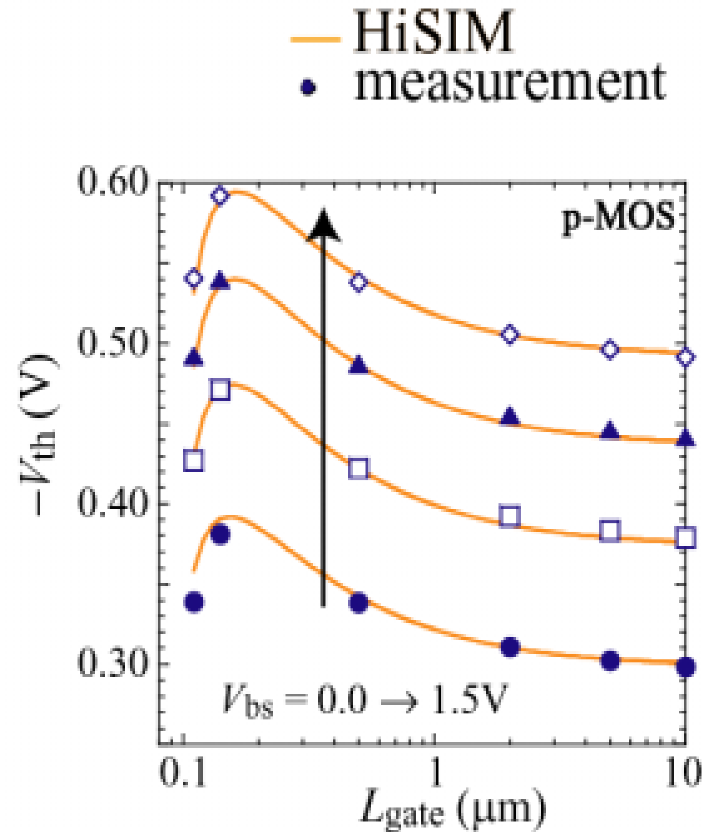


長チャネルで短チャネル効果が現れる

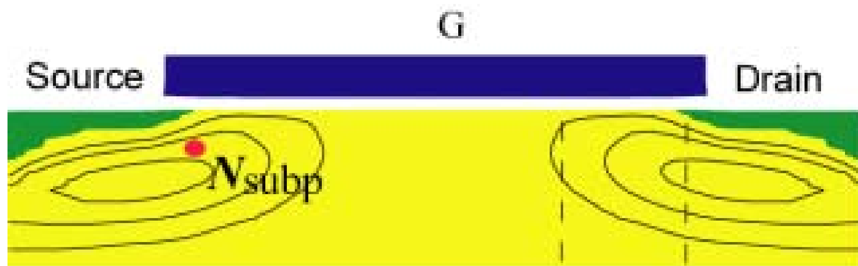
実測値との比較

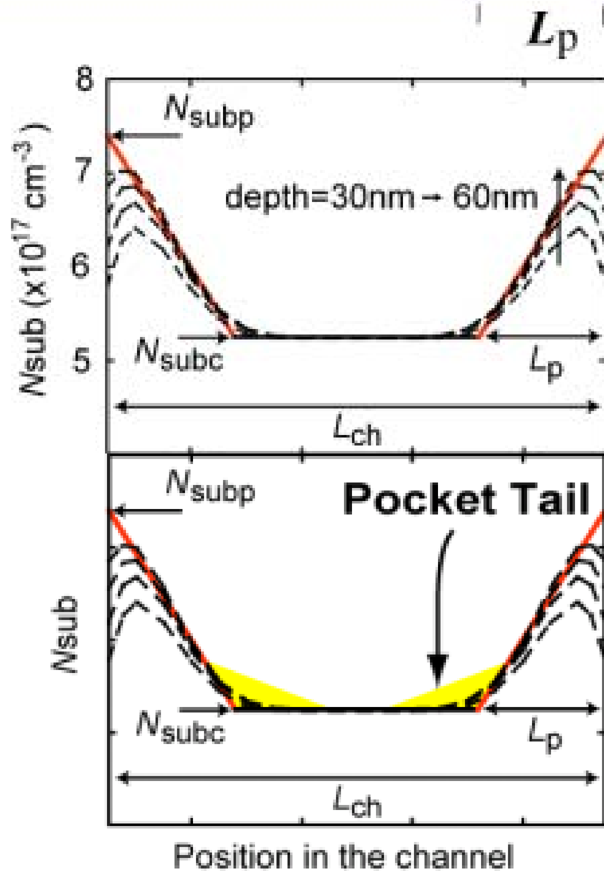


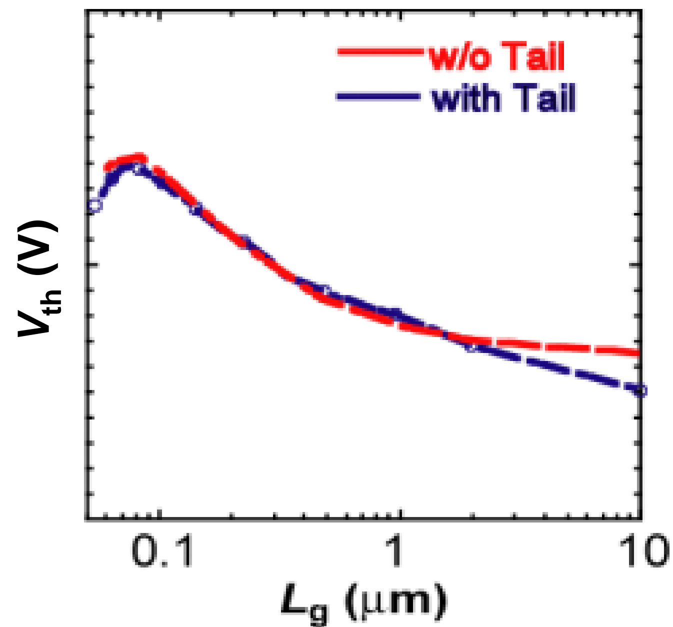
Extracted Pocket Profile



H. Ueno et al., IEEE Trans. Electron Devices, vol. 49, p. 1783, 2002.

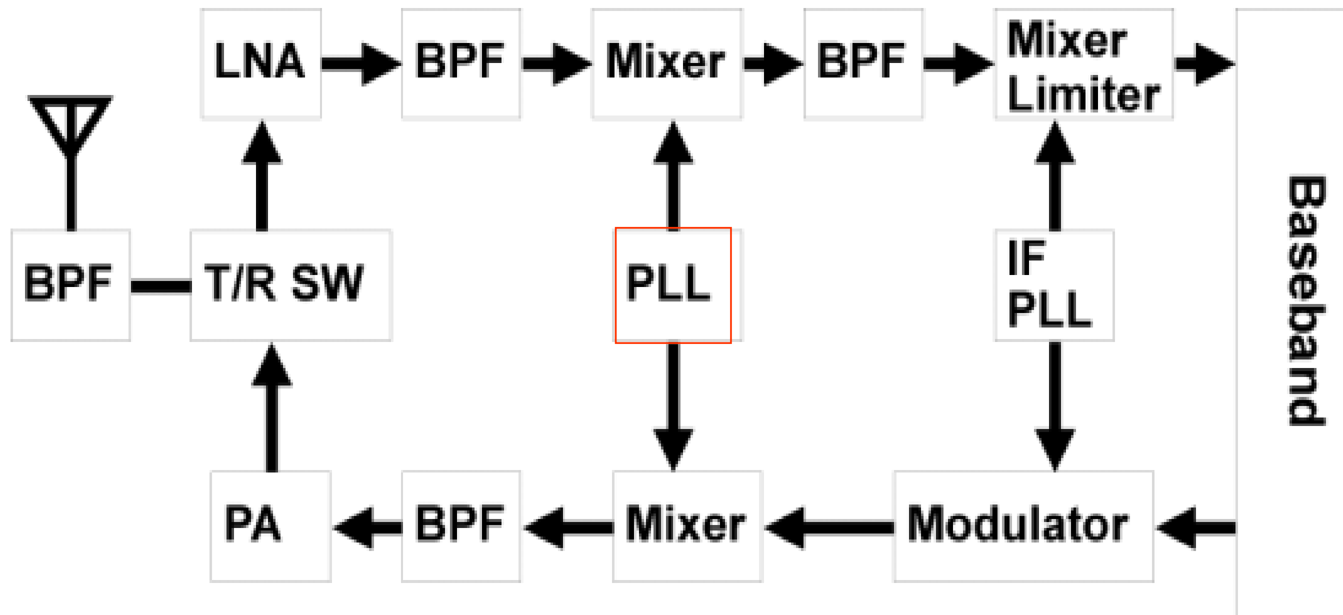






H. Ueno et al., IEEE T ED, 49, p. 1783, 2002.

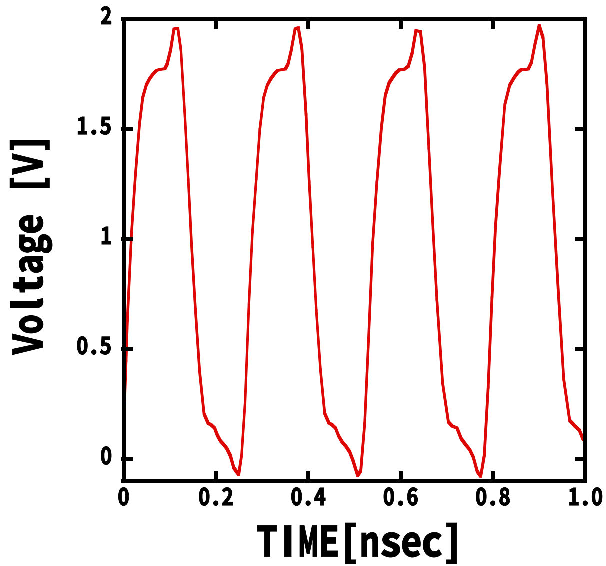
RF回路



1.High Frequency
2.Noise Influence
3.Non-Linearity

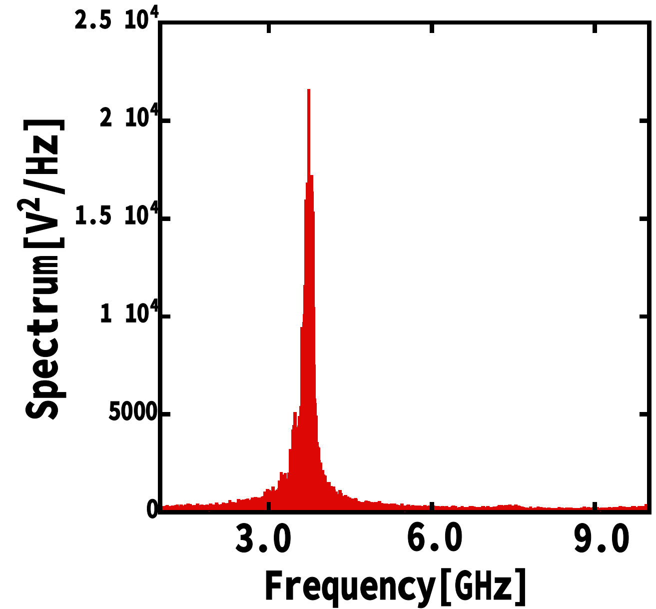
PLLの特性

Time-Domain Analysis

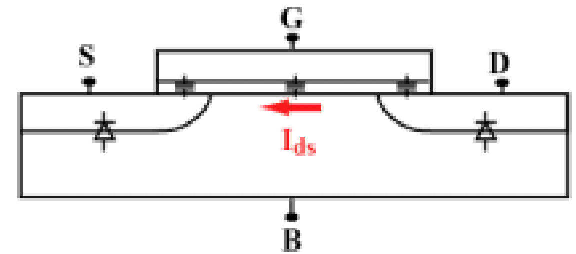


Fourier Transform

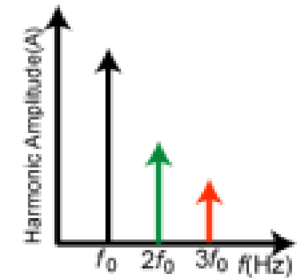
Frequency-Domain Analysis



モデルへの要求



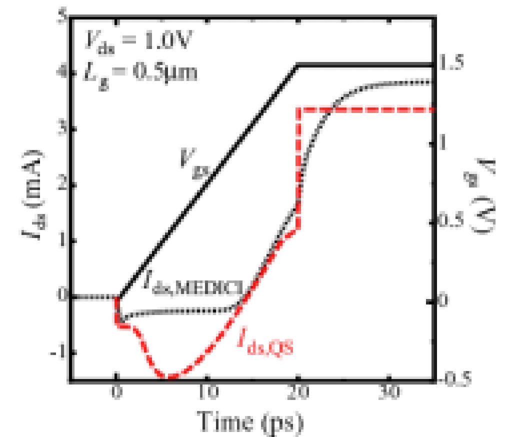
✓キャリア応答遅延：Non-Qausi-Static効果



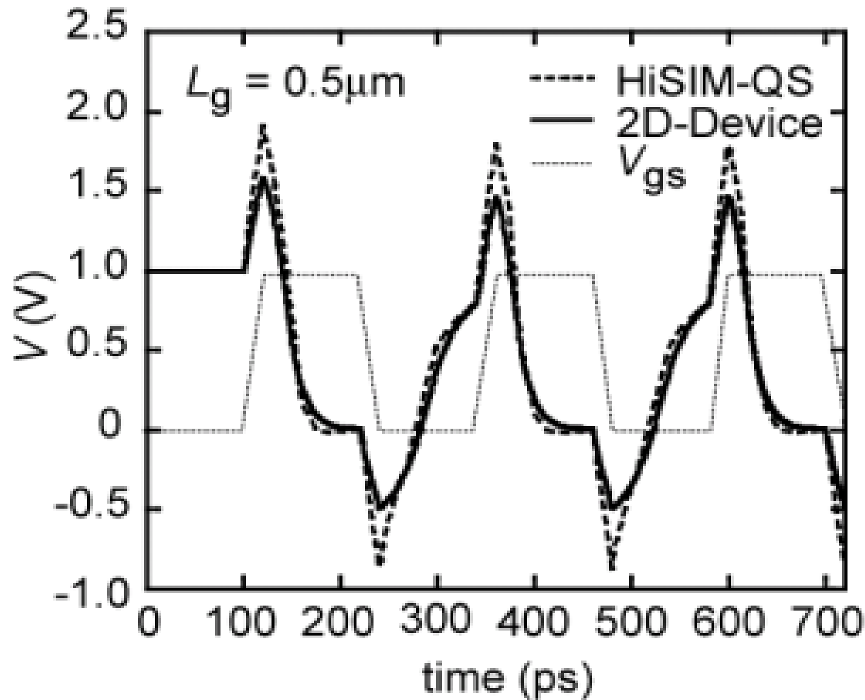
✓ノイズ特性

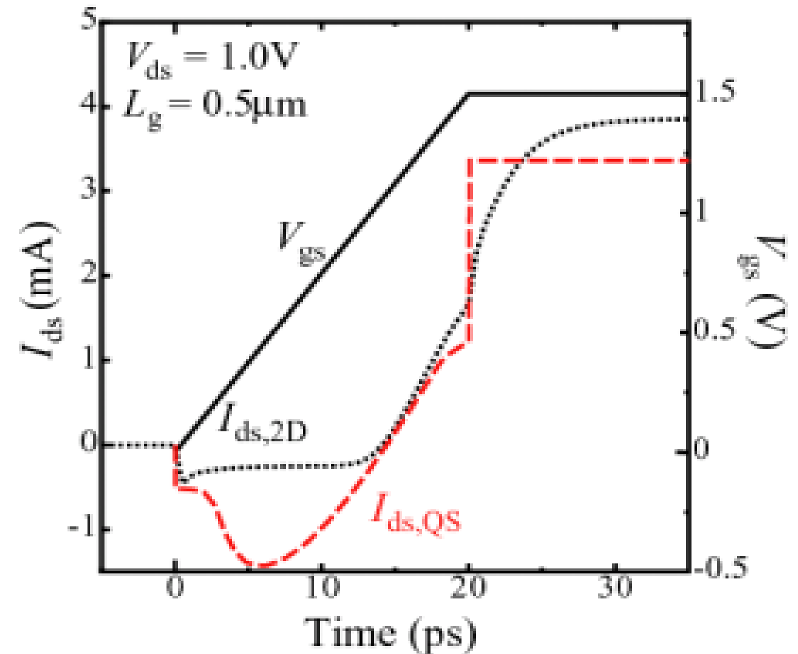
✓高調波ひずみ

✓Y-パラメタ特性



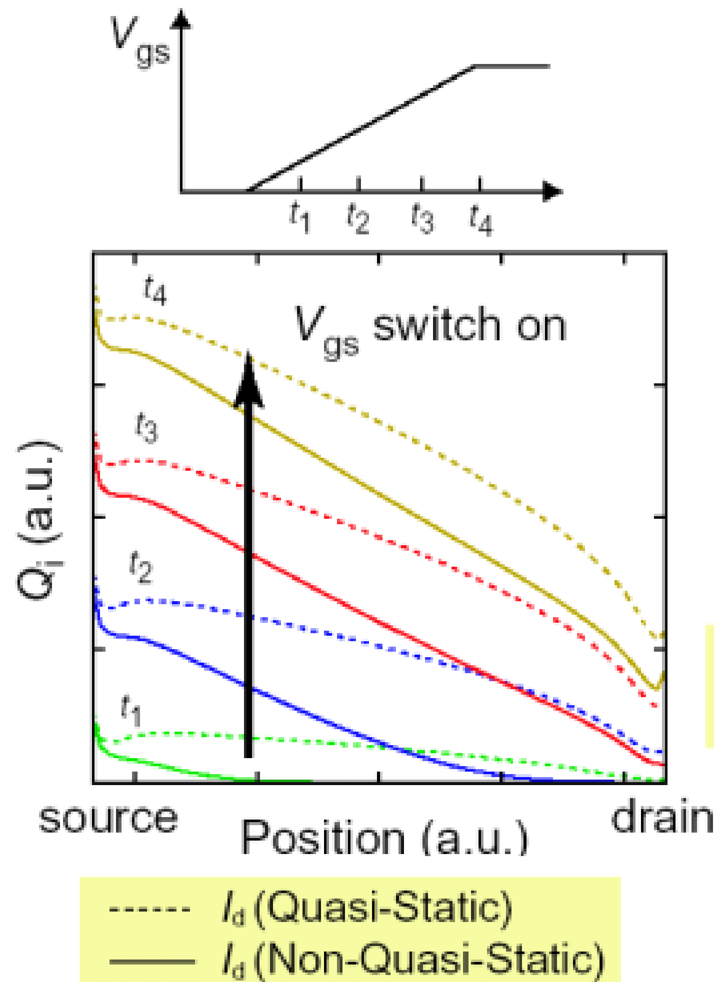
Non-Quasi-Static効果



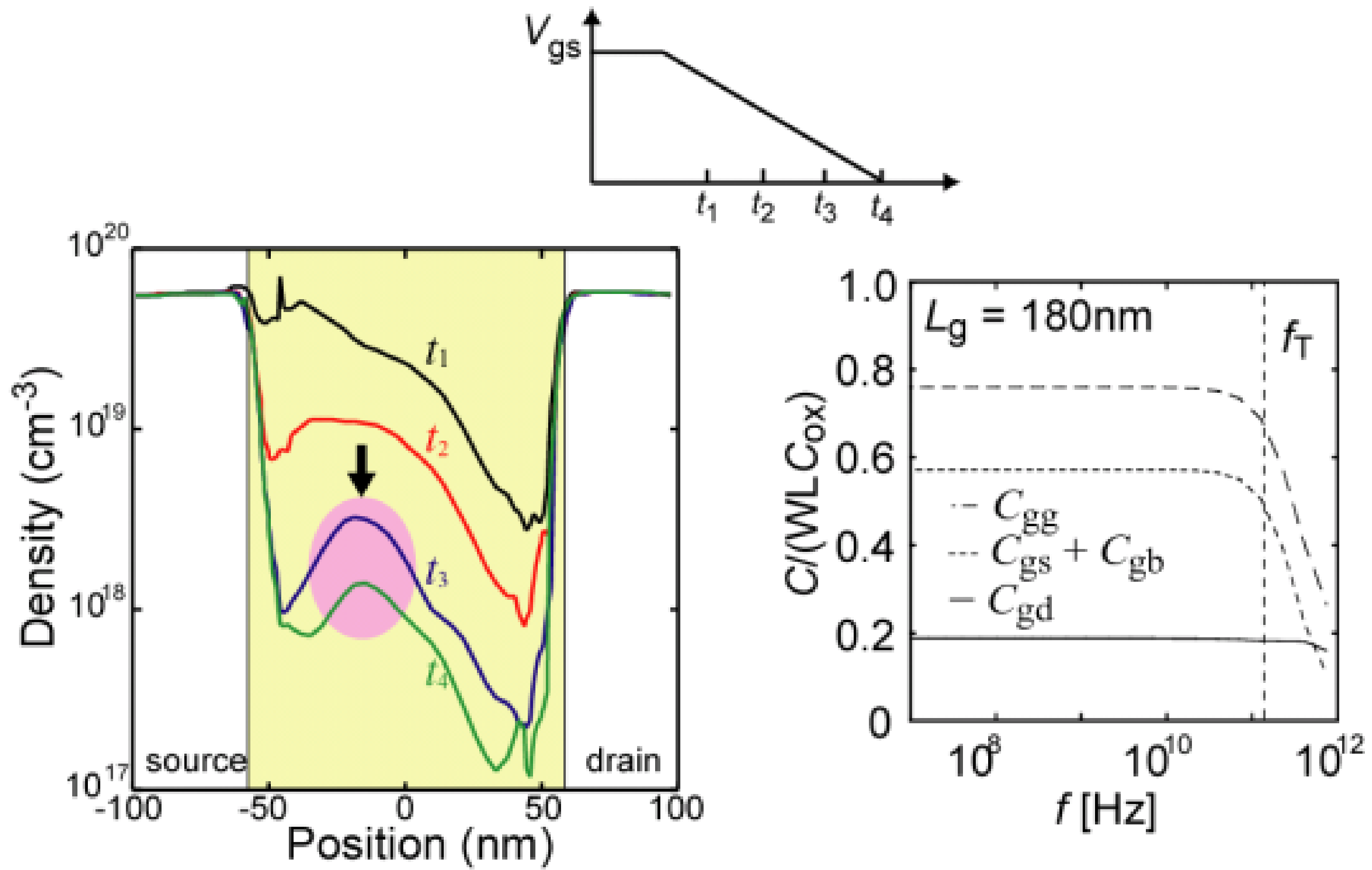


Quasi-Static Approximation: spontaneous carrier response

キャリア応答遅延モデル＝NQSモデル



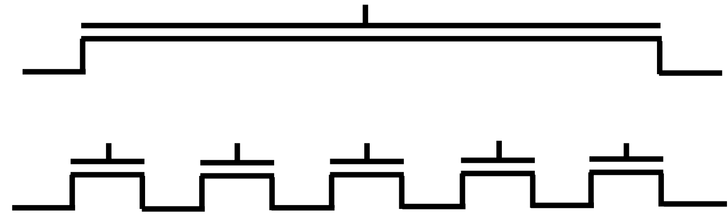
NQS:キャリアが応答できない



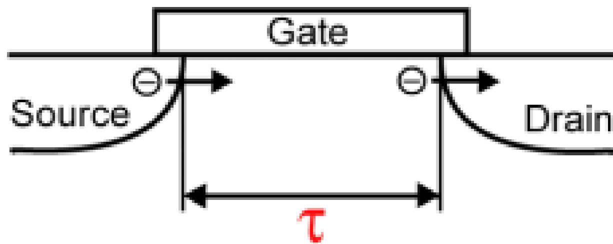
T.Okagaki et al., IEEE EDL, vol. 23, pp. 154, 2002

NQS効果のモデリング

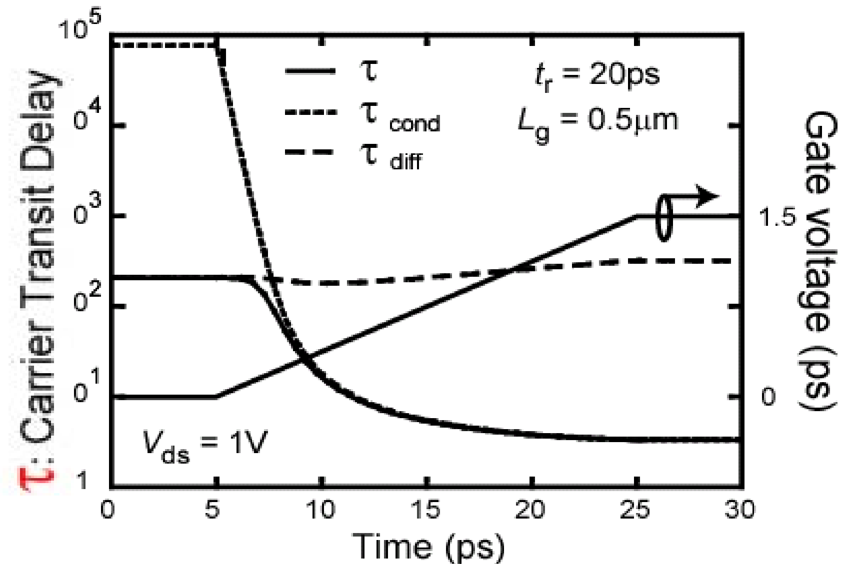
PSP: Channel Segmentation
(10以上必要)



HiSIM: キャリア走行時間を考慮



τ: Carrier Transit Delay
(function of surface potentials)

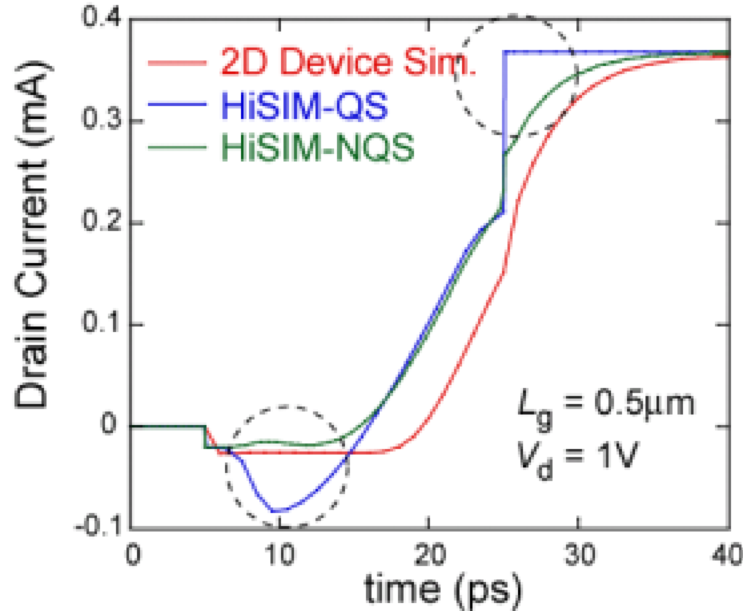


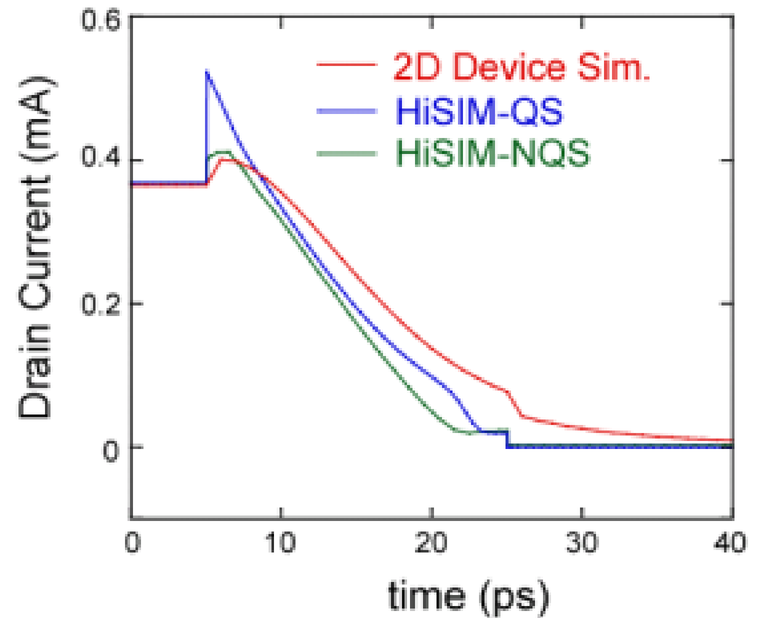
N. Nakayama et al., IEE Electron. Lett., 40, p. 276, 04.

τ :channel長 / キャリア速度

シミュレーション結果

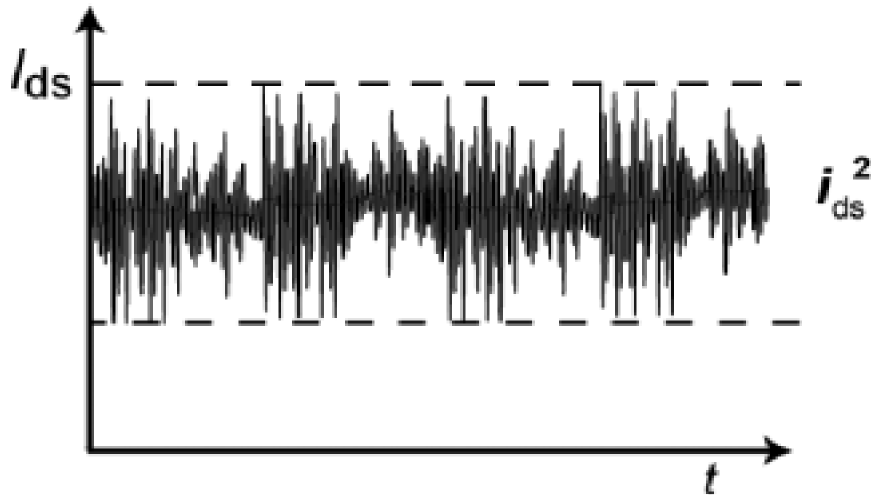
V_{gs}=1.5V　rise/fall time = 20ps



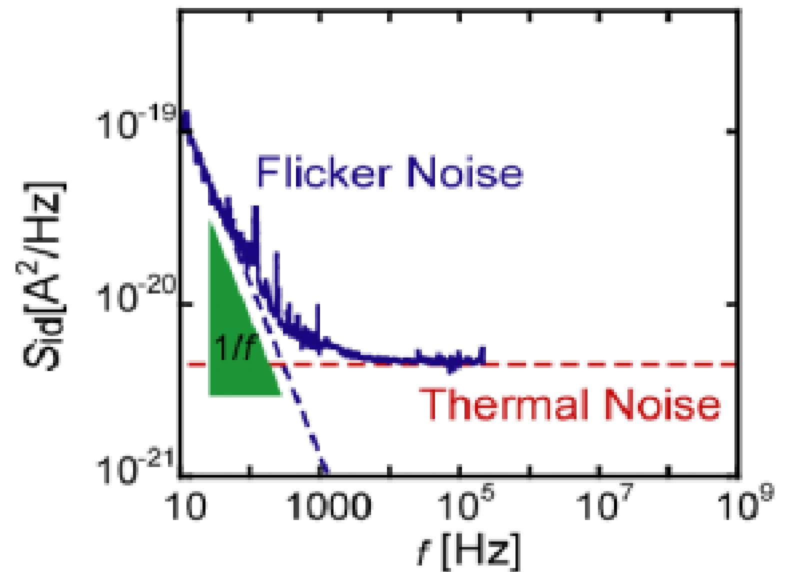


D. Navarro et al., IEEE TED, p. 2025, 2006.

MOSFETにおけるノイズ

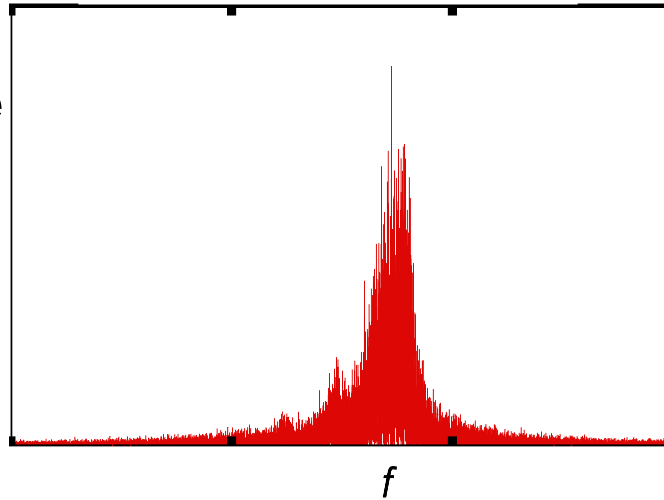


$$S_{id} = i_{ds}^2 / f$$

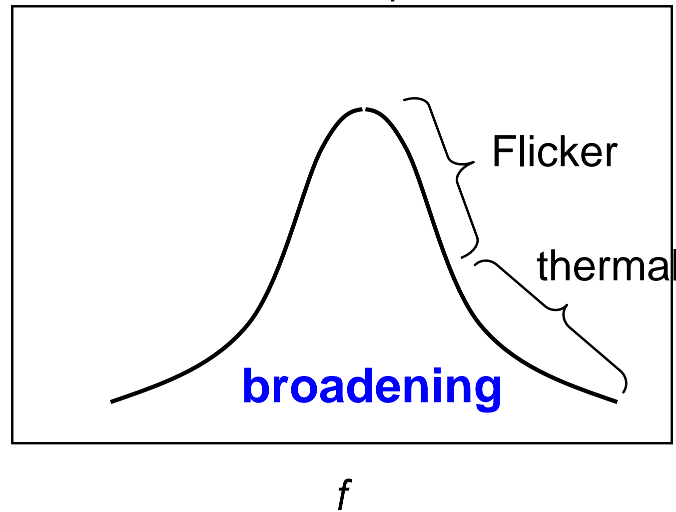


ノイズの問題

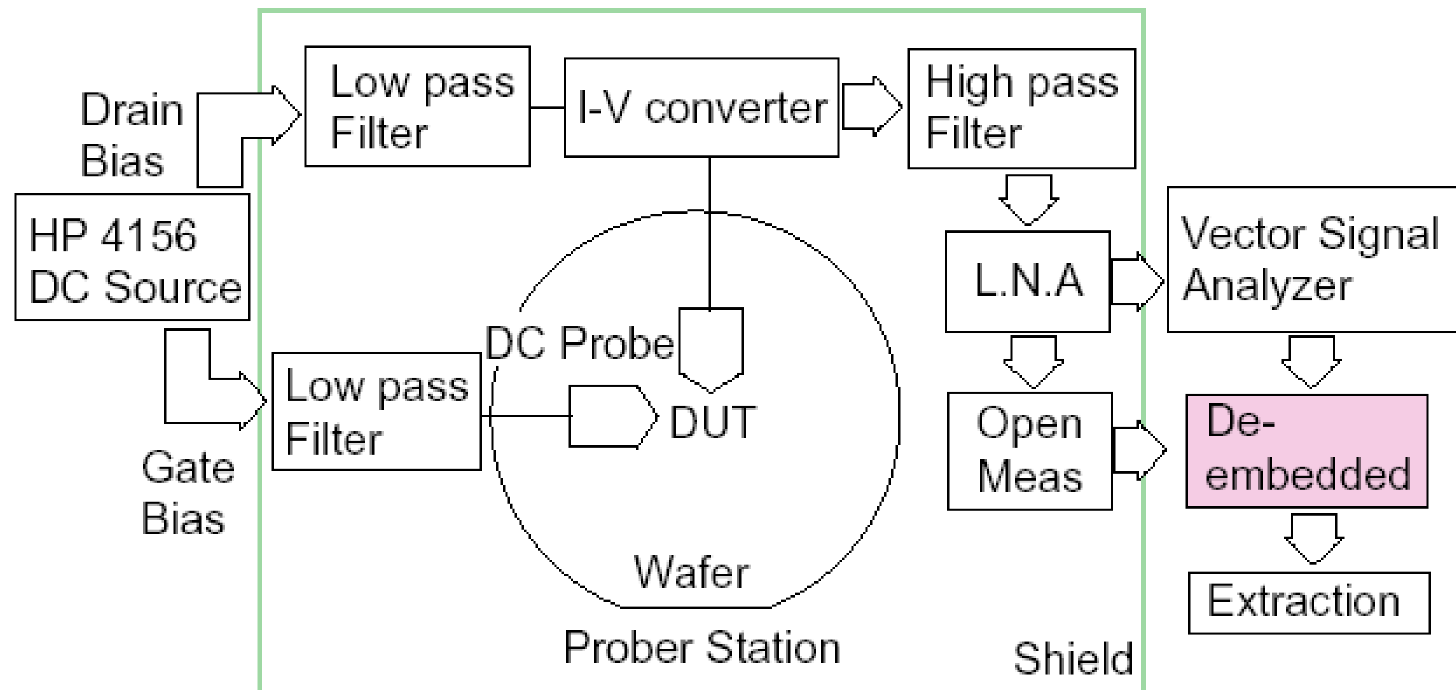
Without Noise



With Noise

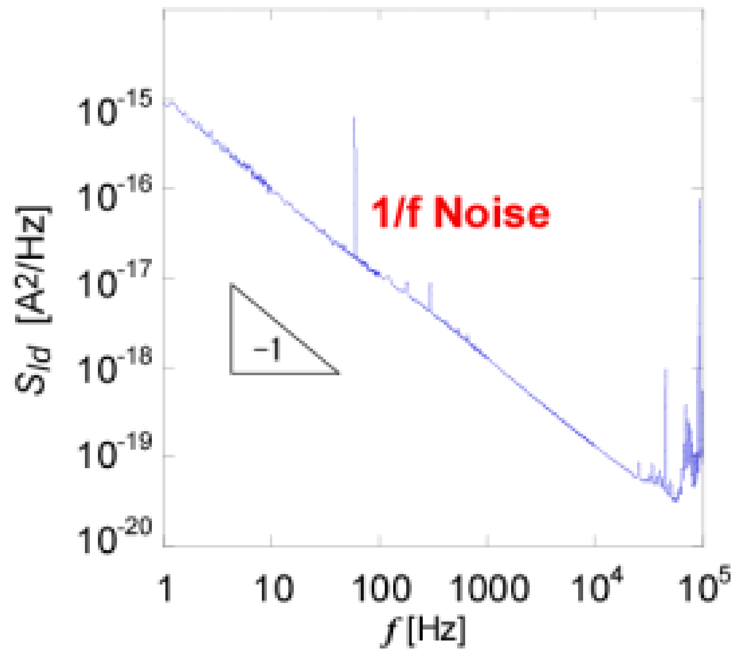


1/f ノイズの測定系

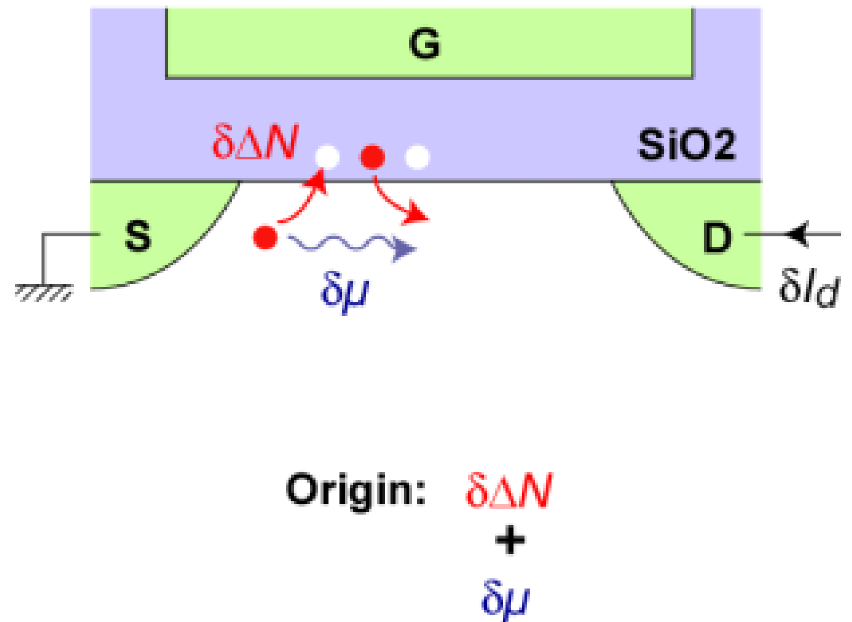


1/*f*ノイズの起源

［測定値］

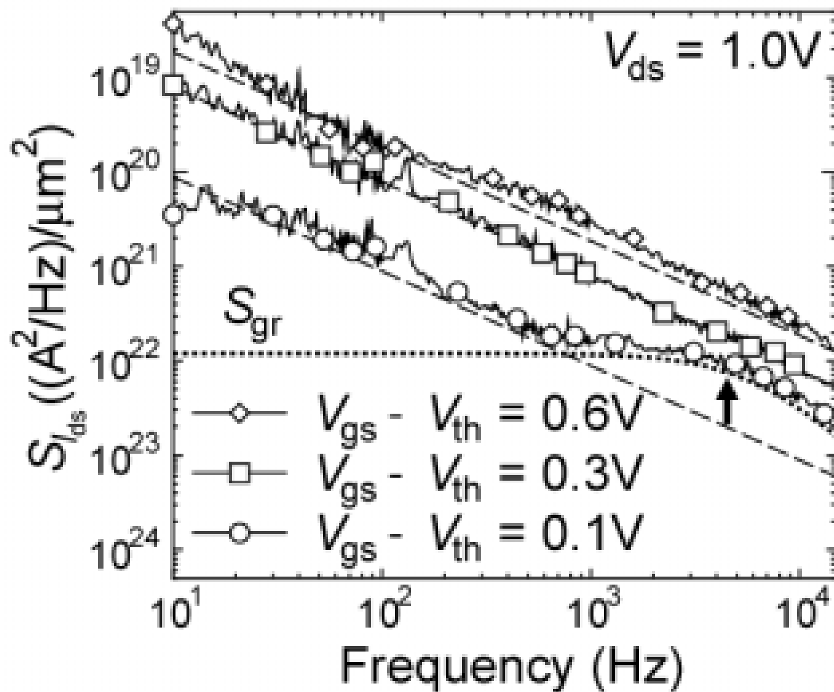


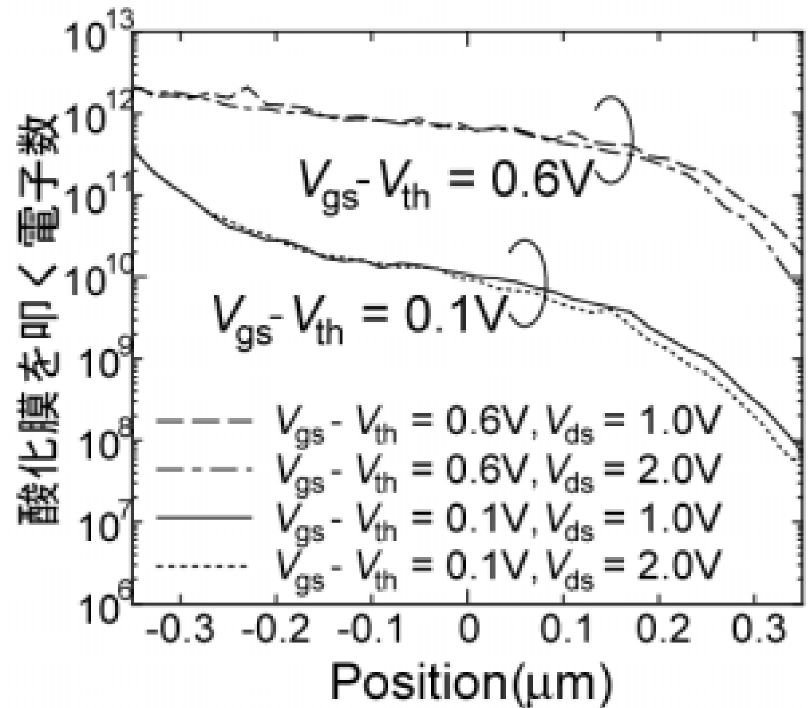
［原理］



たとえば：K. K. Hung et al. IEEE Trans. ED, 37, p.1323, 1990.

Monte Carloシミュレーション

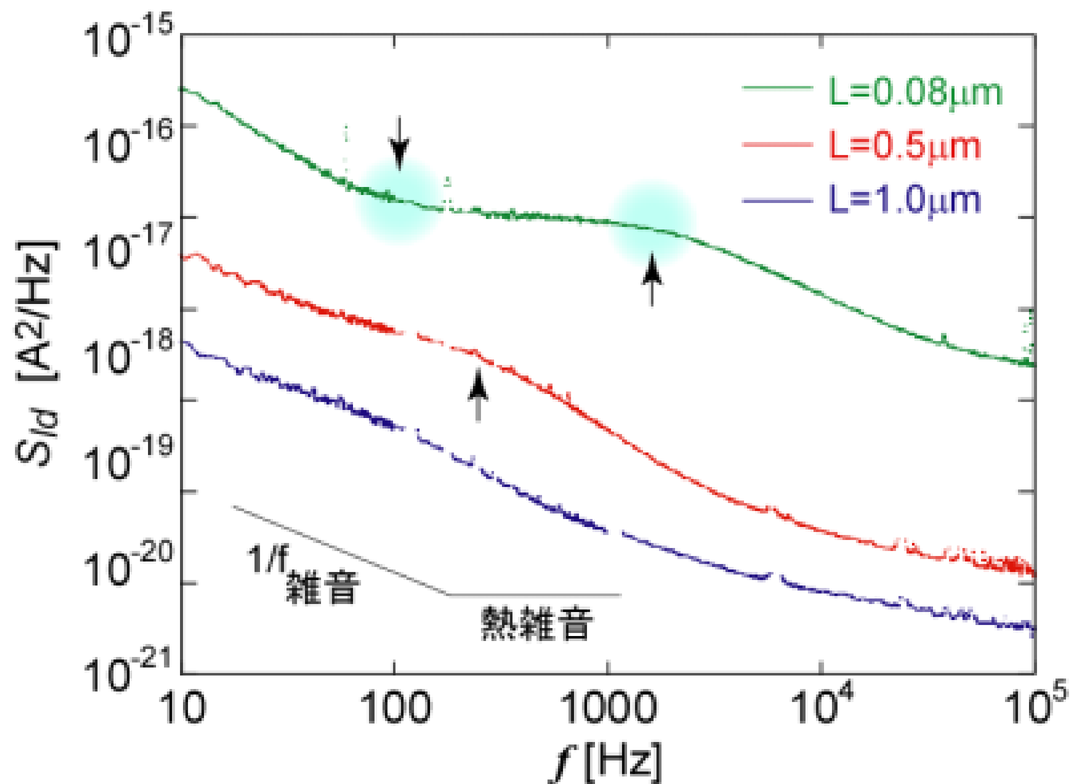




ノイズは主に酸化膜でのTrap/Detrapによって生じる。

H. Ueno et al., Appl. Phys. Let., 78, p.380-382, 2001.

測定結果



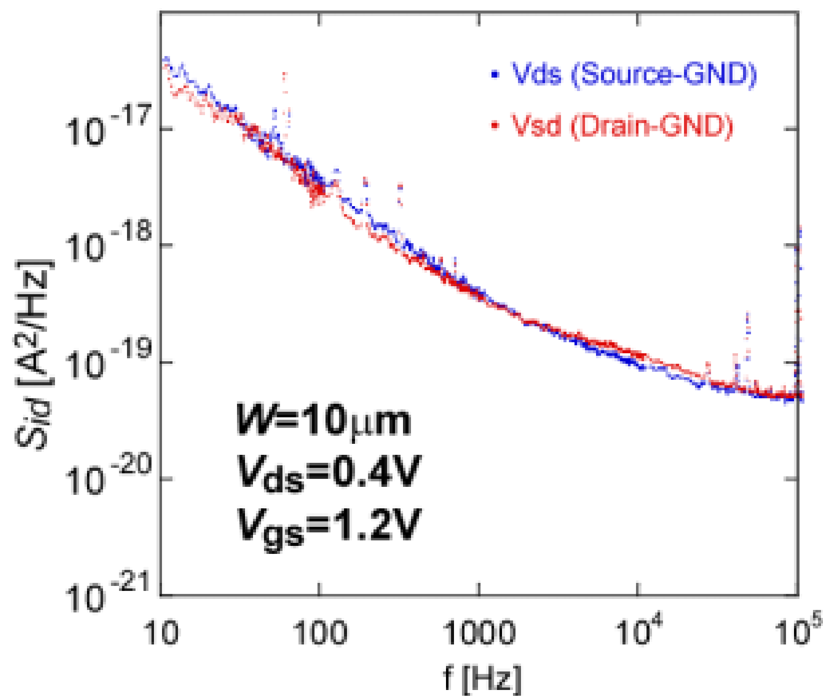
nMOSFET
W=10μm
V_{gs}=1.2V
V_{ds}=0.4V

測定特性は必ずしも1/fではない。

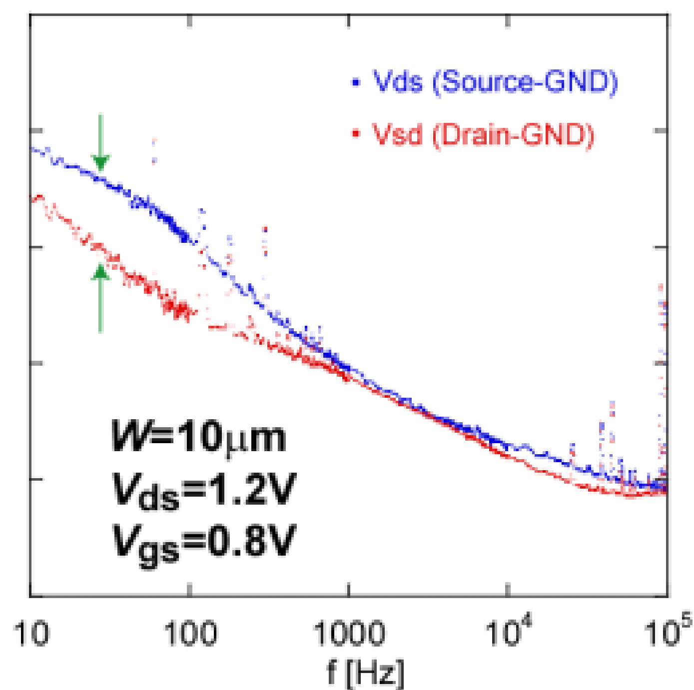
原因を解明：ソースとドレイン電極を入れ替えた測定

L_g=1μm (nMOSFET)

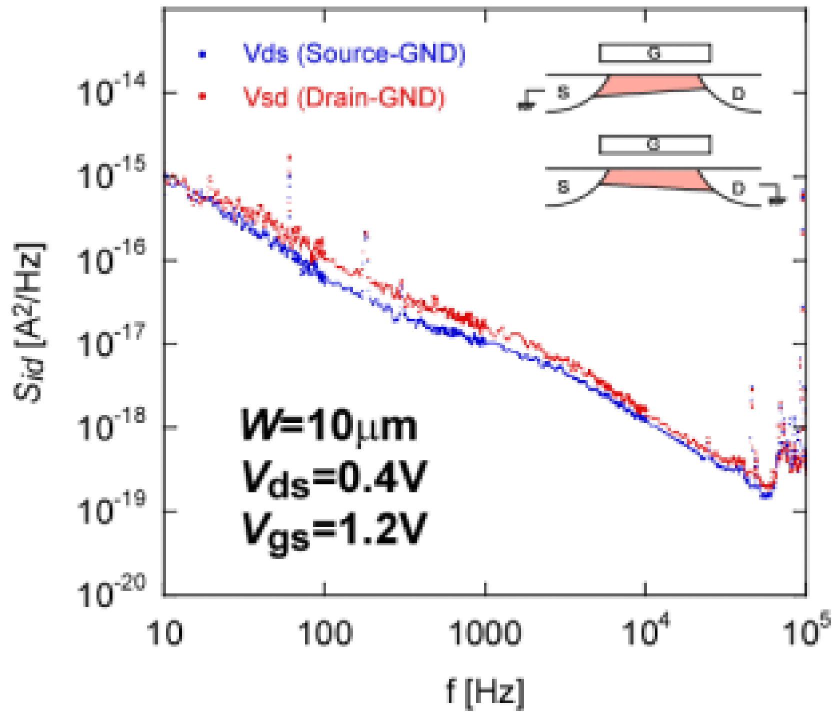
Linear Condition

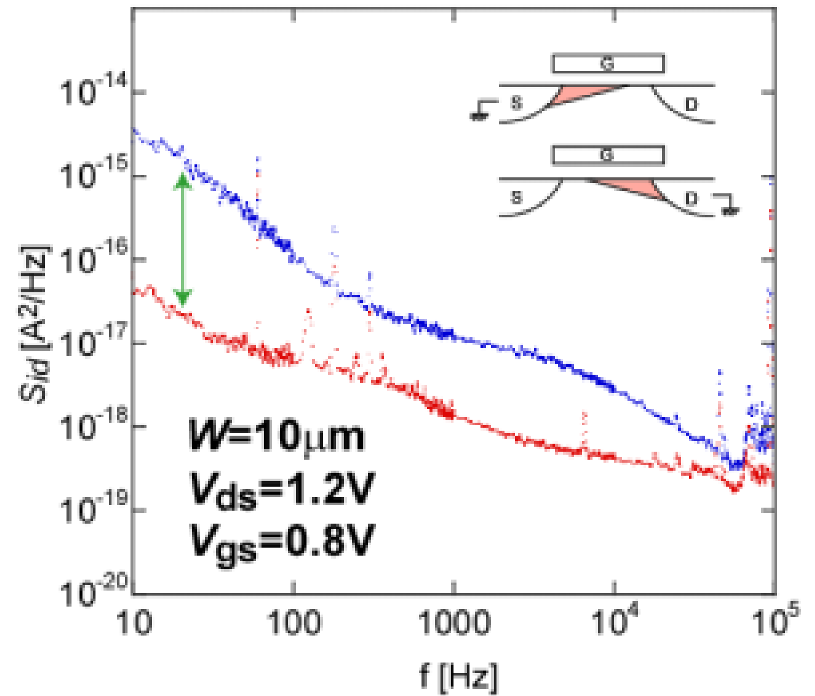


Saturation Condition



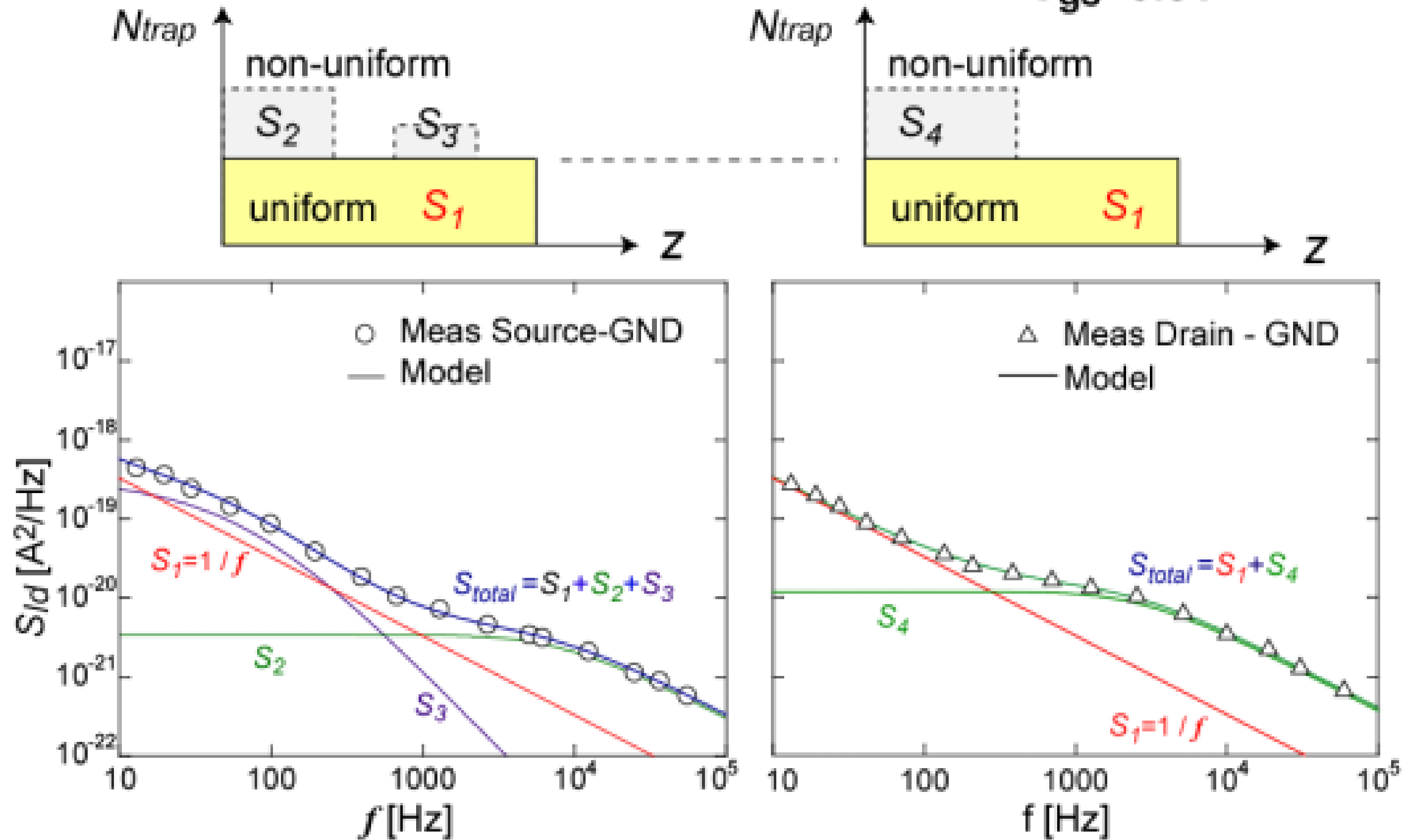
L_g=0.13μm (nMOSFET)





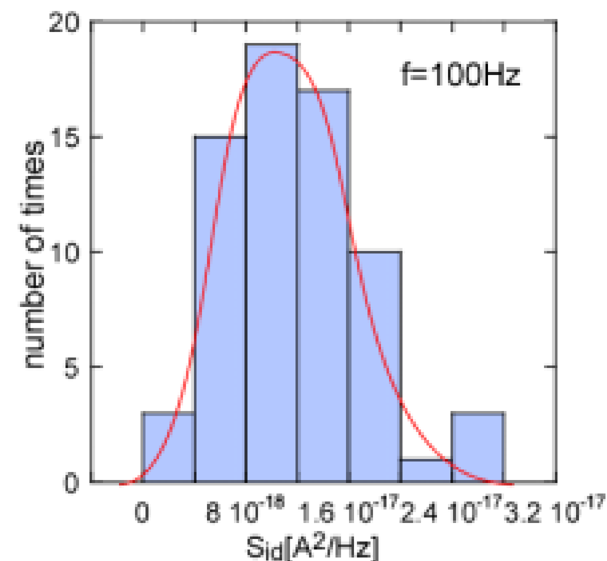
L_g=1μm (nMOSFET)

Saturation Condition: W_g=10μm
V_{ds}=0.4V
V_{gs}=0.5V



不均一なTrap-Siteが原因

Waferの測定結果



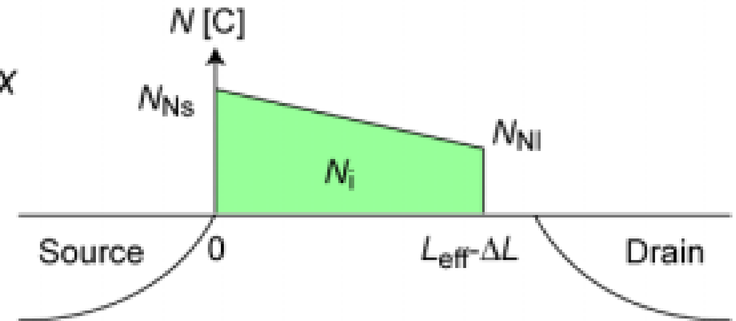
Trap-Siteはランダムに分布

ウエファ上の平均は1/*f*

S. Matsumoto et al., IEICE, Trans. Electron., vol. E88 - C, Feb, 2005.

表面ポテンシャルモデルHiSIMでのモデル化

$$S_{I_{ds}}(f) = \frac{W_g N_t}{q L_g^2 \eta f} kT \int_0^{L-\Delta L} \left(\frac{I_{ds}}{W_g}\right)^2 \left(\frac{1}{N(x)} \pm \alpha\mu\right)^2 dx$$



$$S_{I_d}(f) = \frac{(L-\Delta L)}{L^2} \frac{I_{ds}^2}{W} \frac{N_t(E_f)}{q\eta f} kT \left\{ \frac{1}{(N_s+N^*)(N_l+N^*)} + \frac{2\alpha\mu}{N_l - N_s} \log\left(\frac{N_l+N^*}{N_s+N^*}\right) + (\alpha\mu)^2 \right\}$$

$\alpha = CIT \simeq 0$; $N^* = \dfrac{C_{ox}+C_{dep}+CIT}{q\beta}$; N_s: ソース端のキャリア数
N_l: ドレイン端のキャリア数

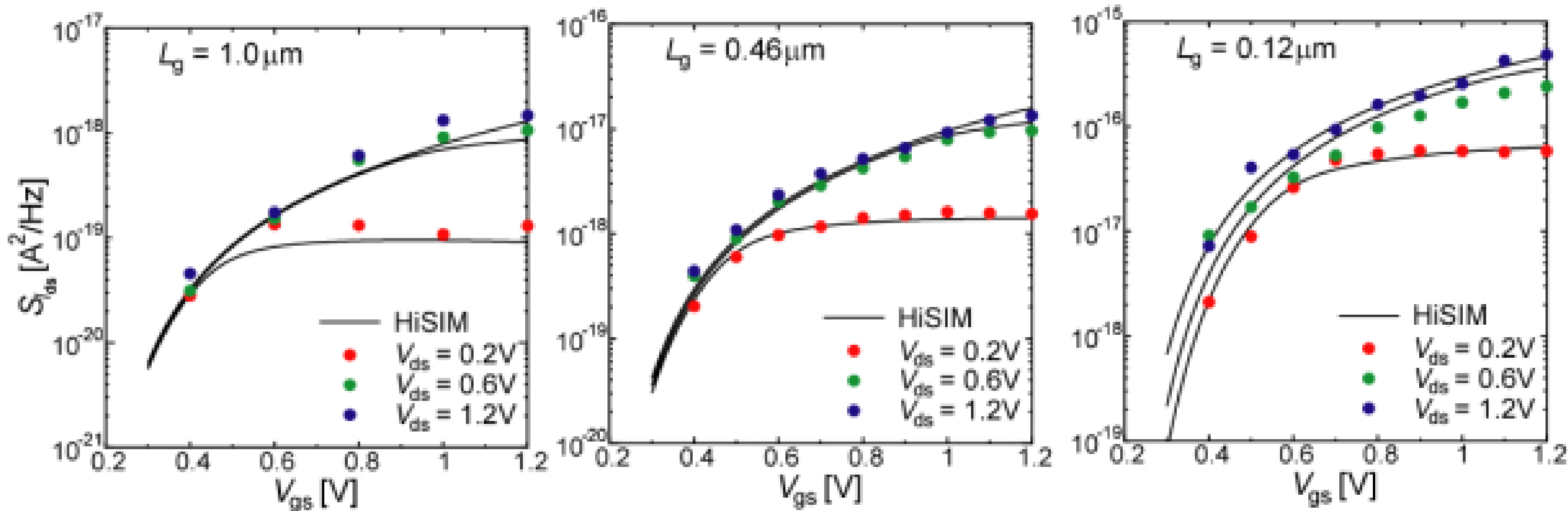
[Model Parameters]

Trap Density: $N_{trap} = N_t(E_f)/\eta \;\; [eV^{-1}cm^{-3}][cm] = [eV^{-1}cm^{-2}]$

Scattering Coeff.: α [Vs]

Flicker Noiseはキャリアのチャネル内分布によって決まる。

1/*f*ノイズの測定値との比較

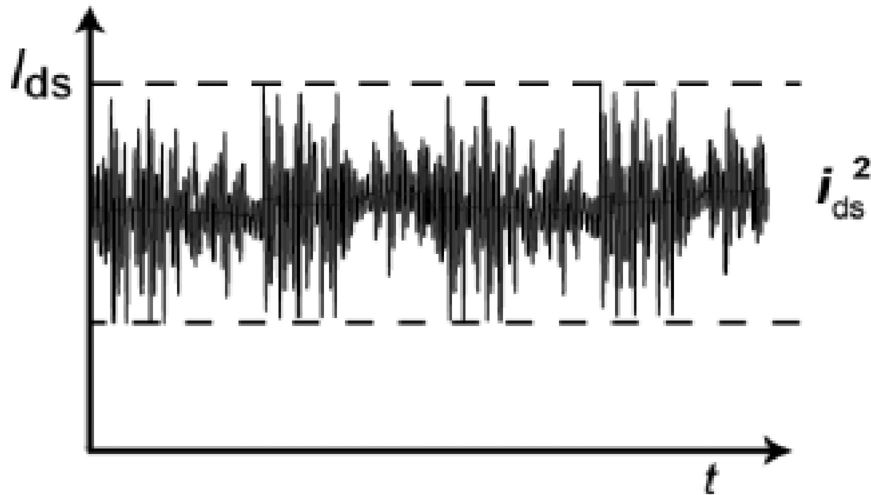


- N_{trap} と α がモデルパラメタ
- 安定した技術であればdefaultで十分

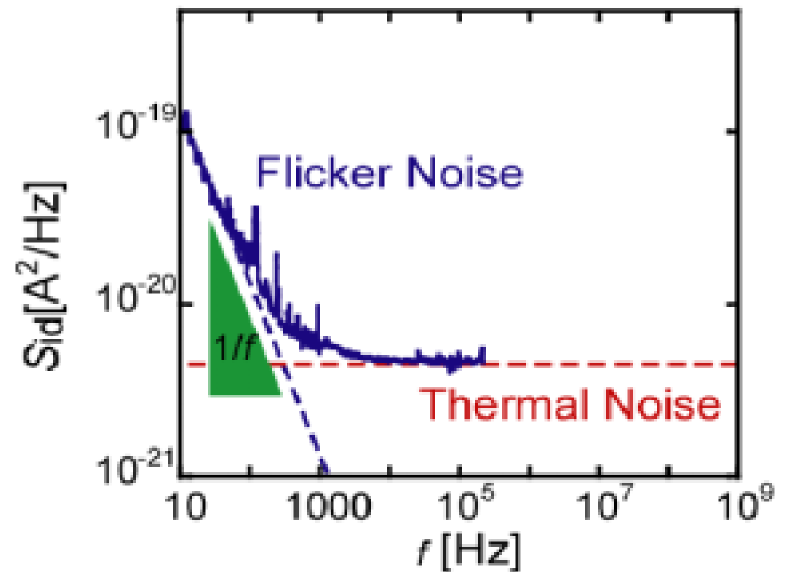
➢ チャネル内キャリアの分布が1/*f*ノイズ特性を決定している。

S. Matsumoto et al., IEIEC T E, E88-C, p. 247, 2005.

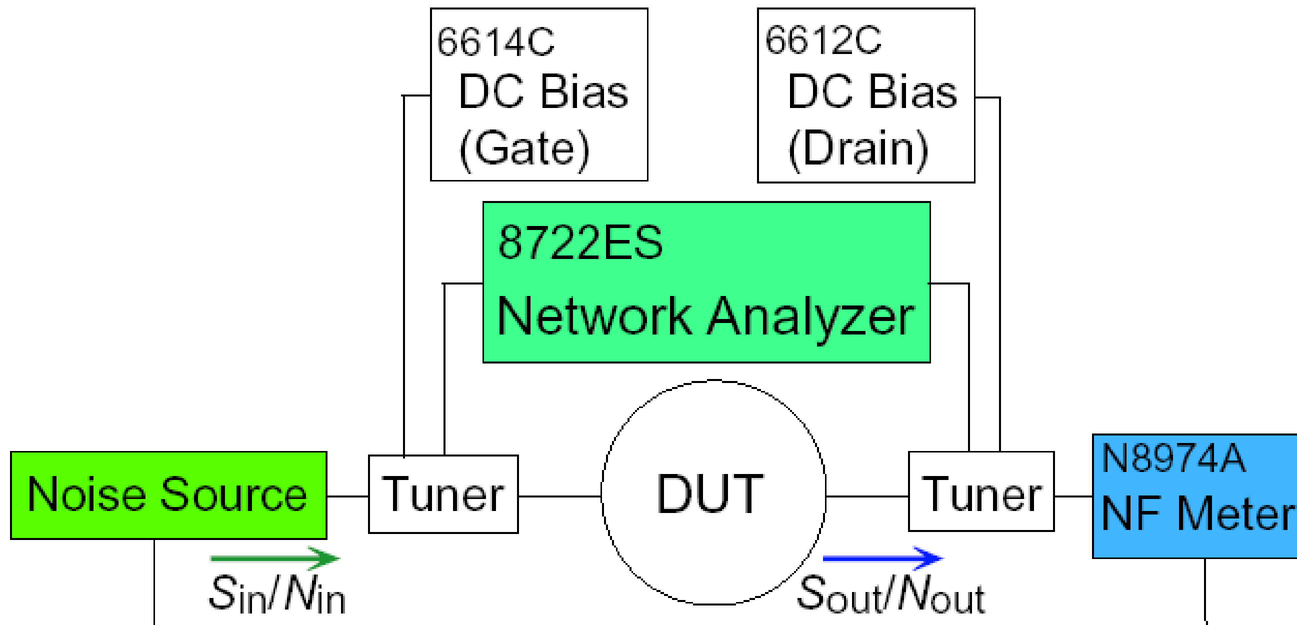
MOSFETにおけるノイズ



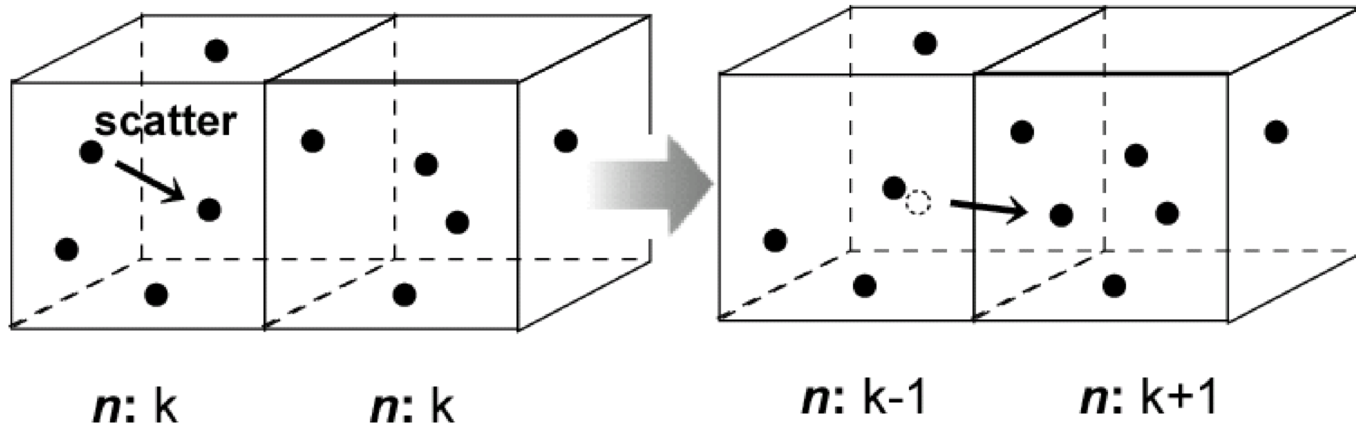
$$S_{id} = i_{ds}^2 / f$$



Noise Figure測定



Thermalノイズの起源



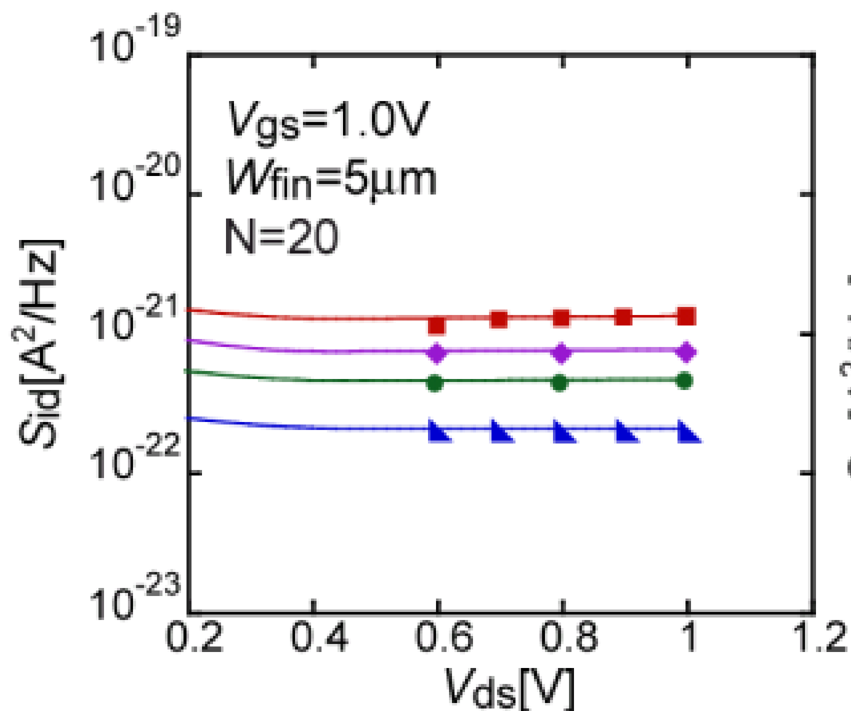
- van der Ziel Description based on Nyquist's Theorem

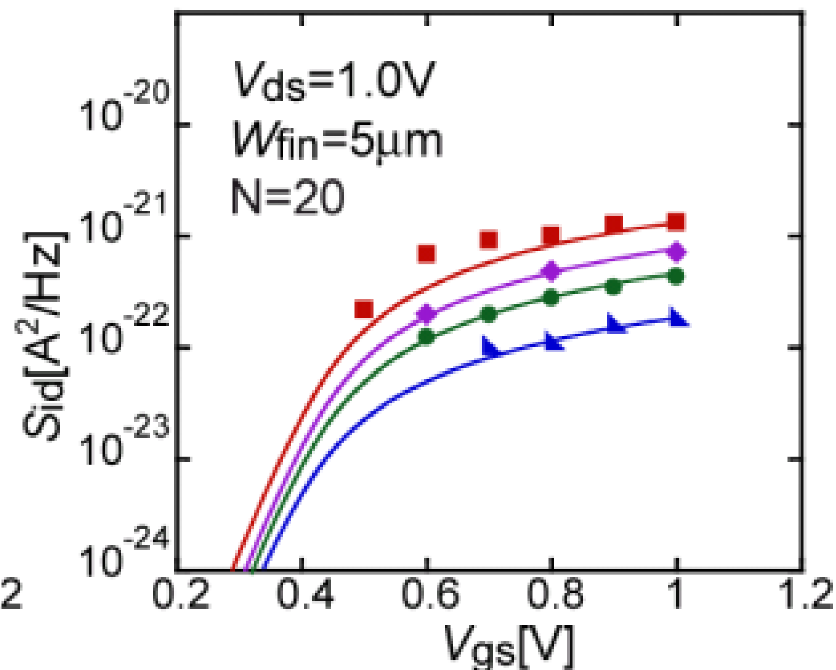
$$S_{\mathrm{id}} = \frac{4kT}{L_{\mathrm{eff}}{}^{2} I_{\mathrm{ds}}} \int g_{\mathrm{ds}}{}^{2}(y)dy$$

$$= 4kTg_{\mathrm{ds0}}\gamma$$

$g_{ds}(y)$: Channel Conductance
g_{ds0}: at V_{ds}=0
γ : Noise Coefficient

測定値とHiSIM計算結果の比較





*L*g=0.17μm
*L*g=0.3μm
*L*g=0.5μm
*L*g=1.0μm

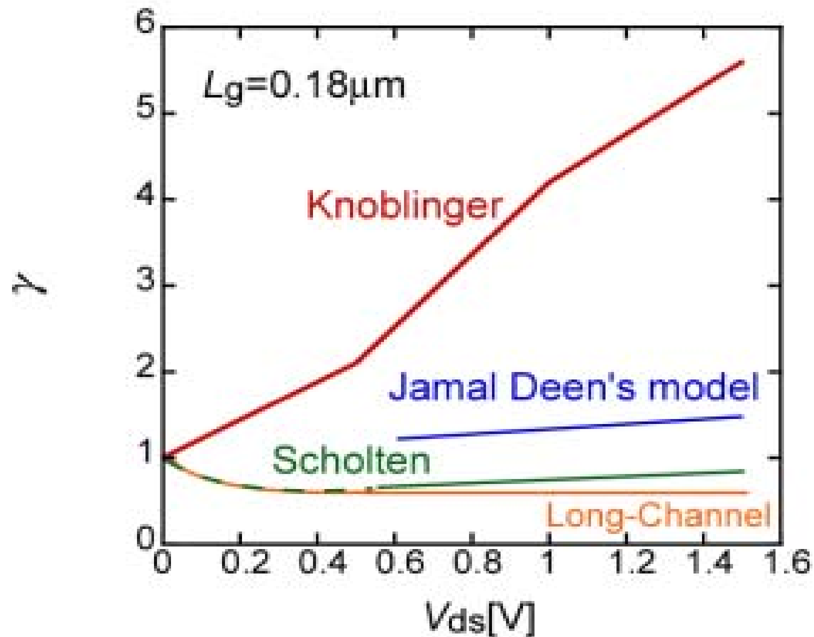
Lines: HiSIM

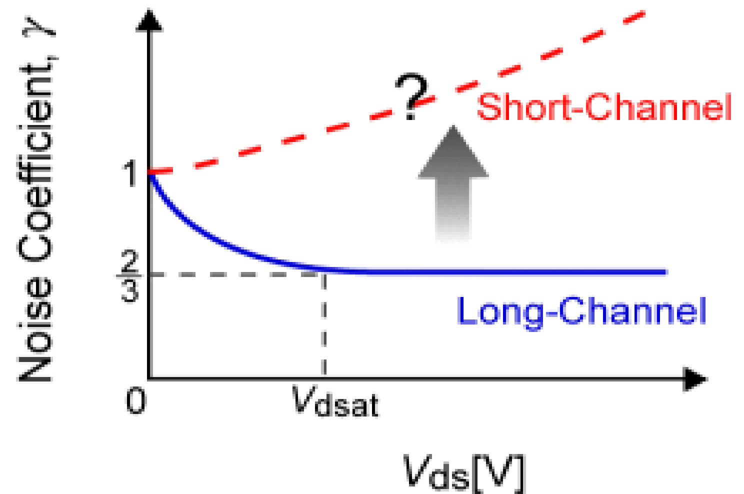
Symbols: Measurements

モデルパラメタはなし

S. Hosokawa et al., Appl. Phys. Lett., 87, 092104, 2005.

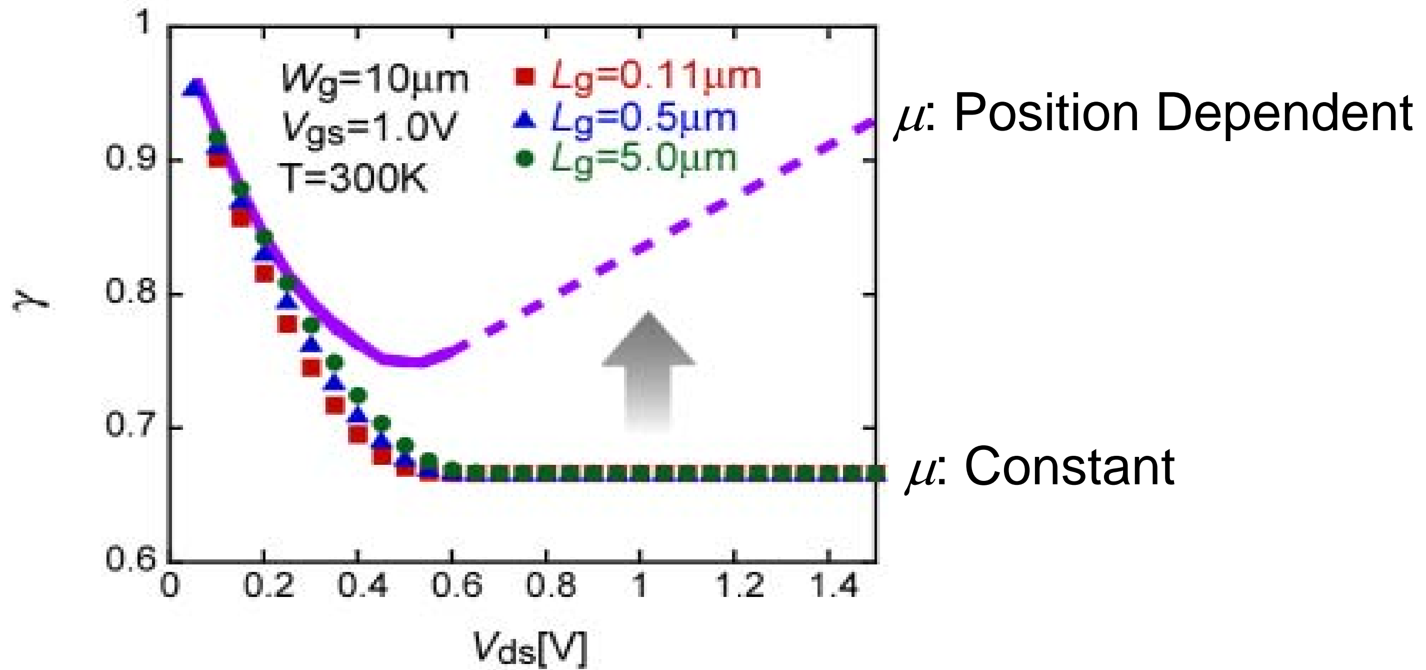
ノイズ係数の考察





- Knoblinger et al. (2001): Hot Electron Contribution
- Jamal Deen et al. (2002): Channel Length Modulation
- Scholten et al. (2002): Velocity Saturation

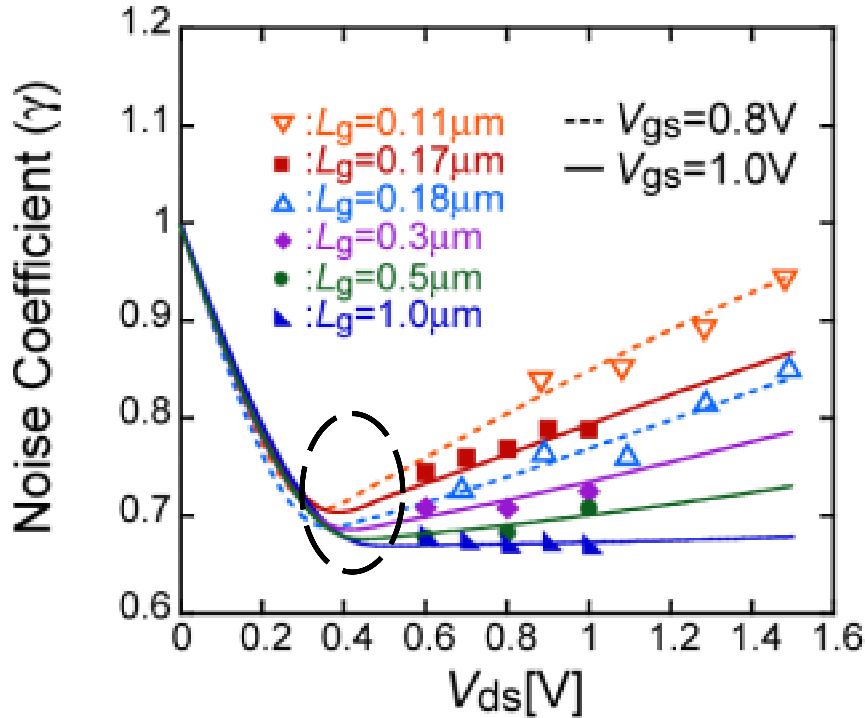
測定特性の解析



Origin of γ Increase → Mobility Reduction along Channel

γ Increase → Potential Increase

異なるテクノロジーの比較

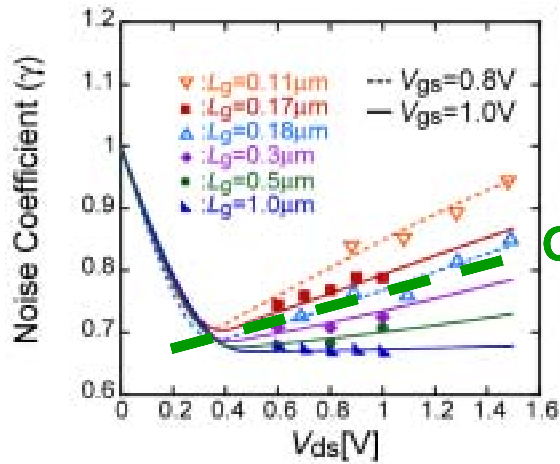


Lines: Simulation (HiSIM)

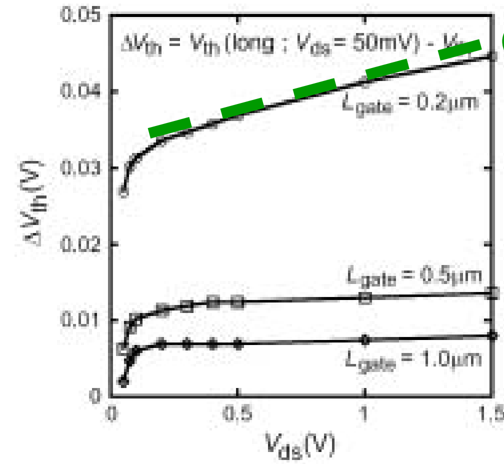
Symbols: Measurements

- **first γ reduction and increase in the saturation region**
- **no drastic increase of γ**
- **γ minimum increase from 2/3**

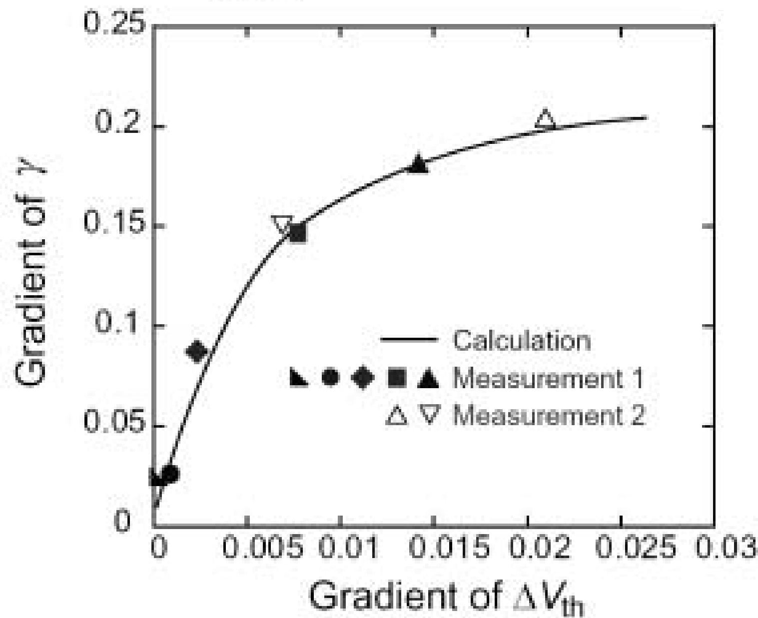
V_{th}shiftとの関係



Grad. γ



Grad. ΔV_{th}

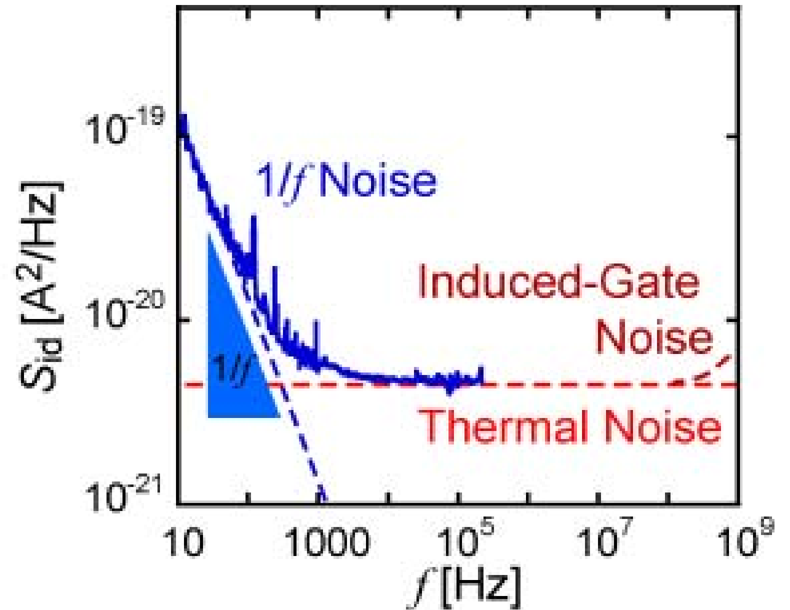


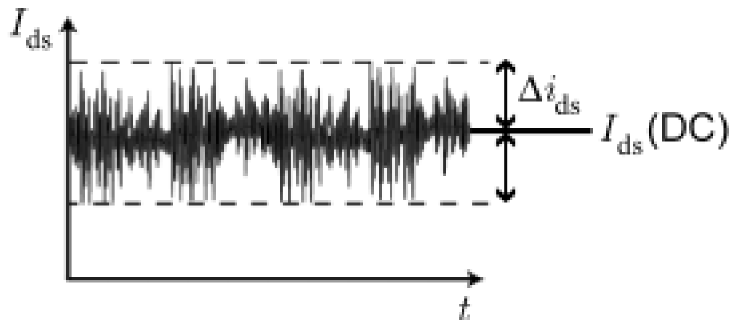
Universal Relationship

MOSFETにおける更なるノイズ

- 1/*f* Noise
- Thermal Noise
- Induced Gate Noise
- Cross-Correlation Noise

- Shot Noise
- Junction Noise

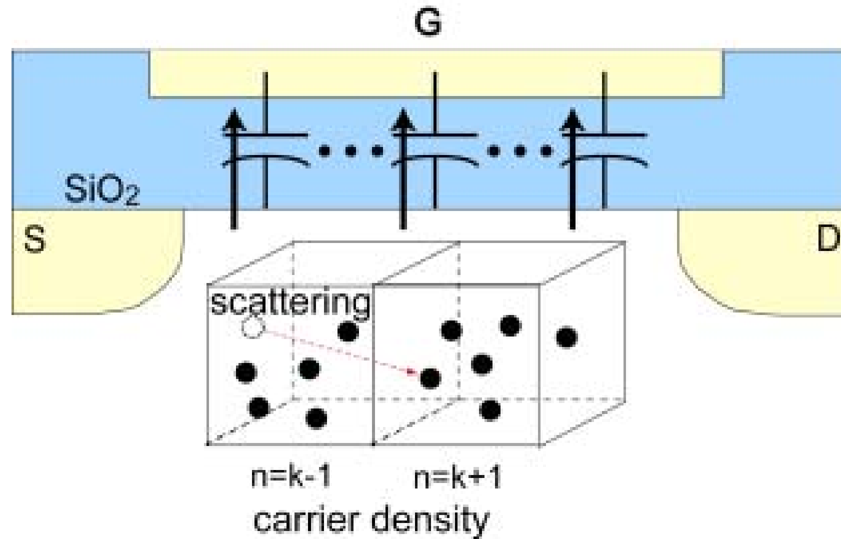




Noise Spectral Intensity

$$S_{i_d} = \frac{\overline{\Delta i_{ds}^{\ 2}}}{f}$$

Induced Gate & Cross Correlation Noise



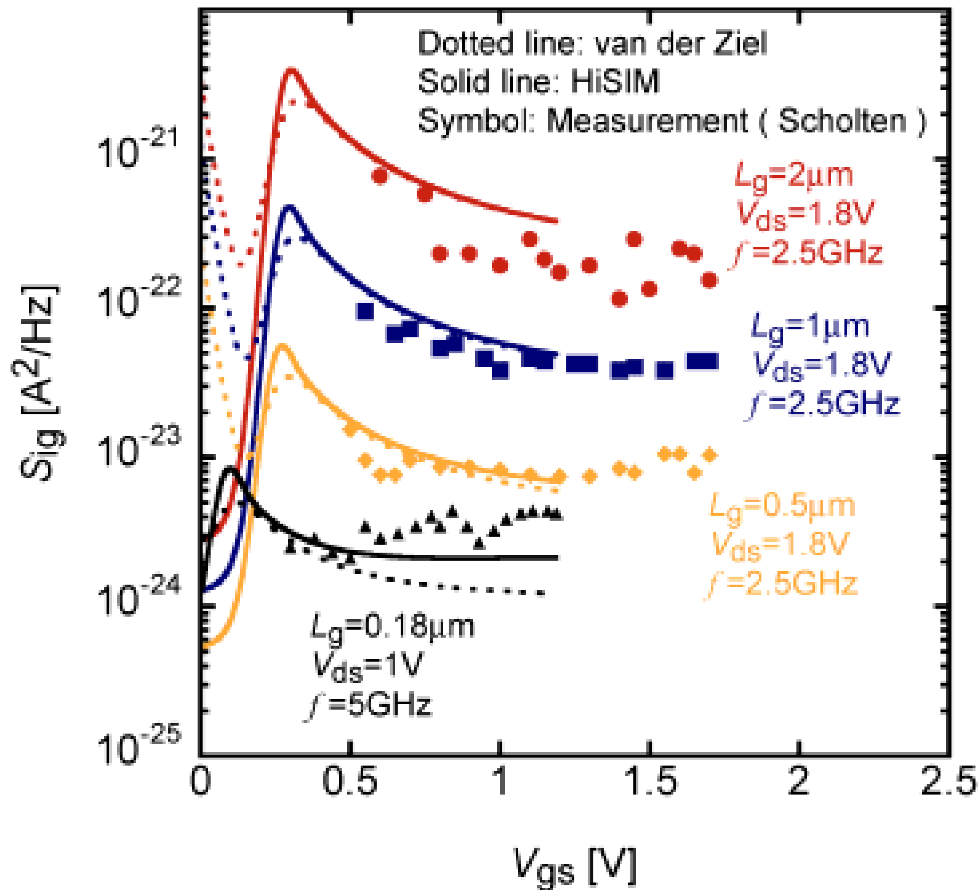
$$S_{i_g} = \int \left| \Delta i_g \right|^2 dx = \int \left(\Delta I_1' + \Delta I_2' \right)^2 \Delta v^2 dx \text{ : induced gate}$$

$$S_{i_g i_d} = \int \Delta i_g^{\ *} \Delta i_d dx = \int \left(\Delta I_1' + \Delta I_2' \right)^* \left(\Delta I_2' \right) \Delta v^2 dx \text{ : cross - correlation}$$

モデルパラメタなし

T. Warabino et al., Proc. SISPAD, p. 158, 2006.

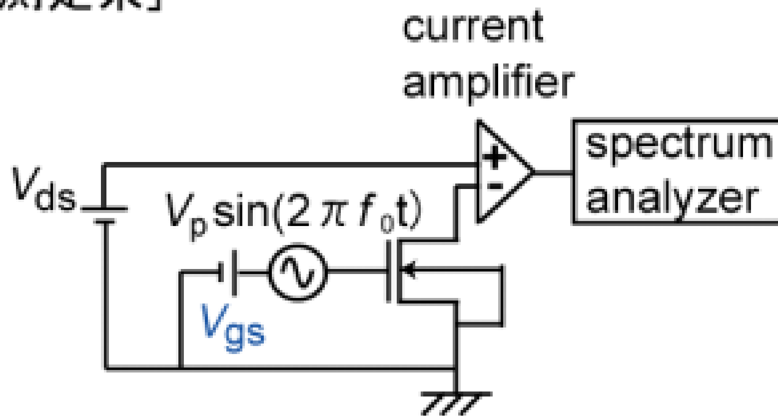
測定値の考察



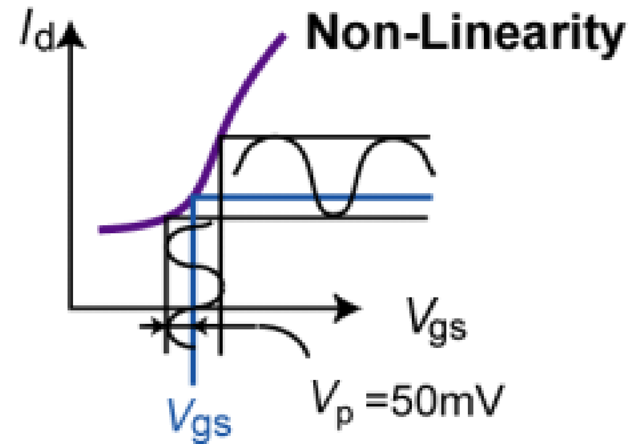
A. J. Scholten et al., Tech. Dig. IEDM, p. 867, 2003.
T. Warabino et al., Proc. SISPAD, p. 158, 2006.

高調波ひずみ

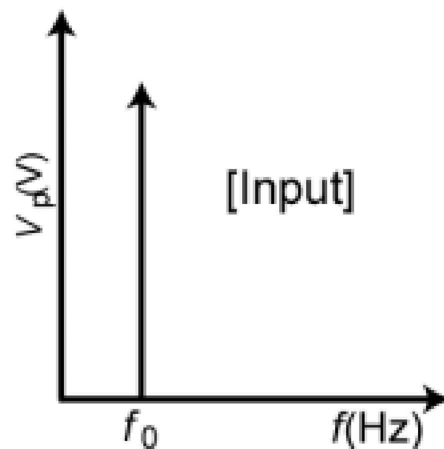
［測定系］

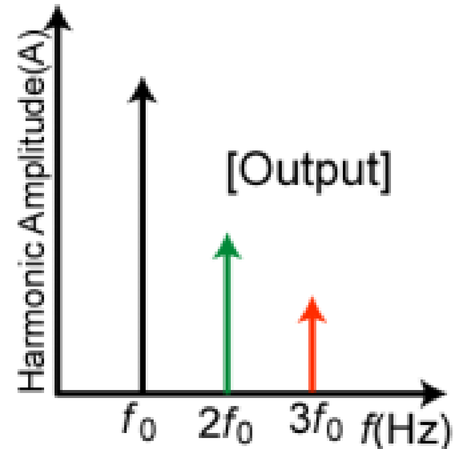


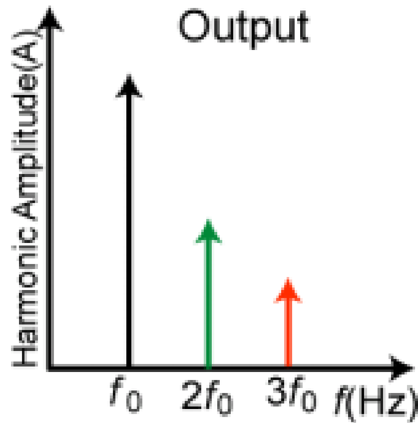
［原理］



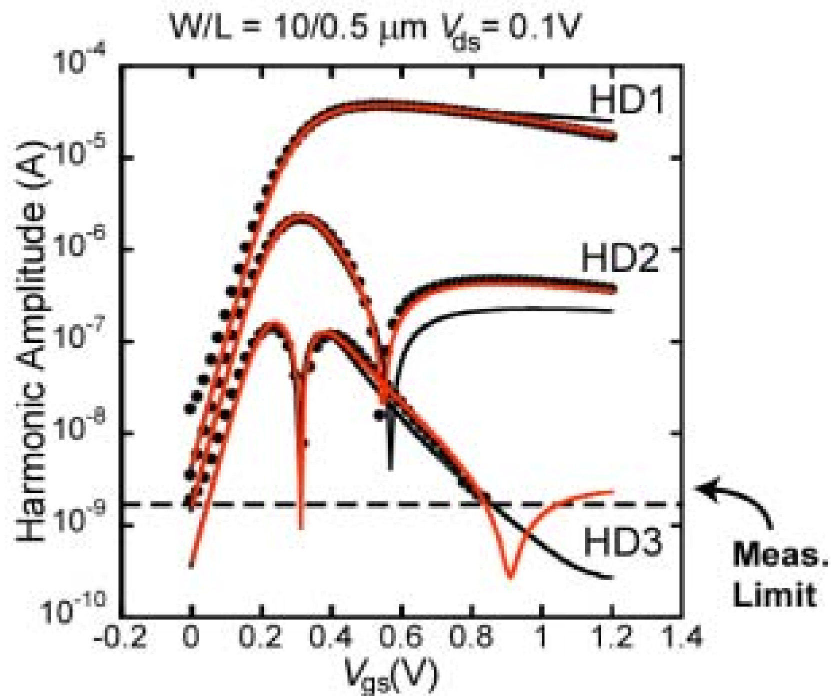
［測定結果］



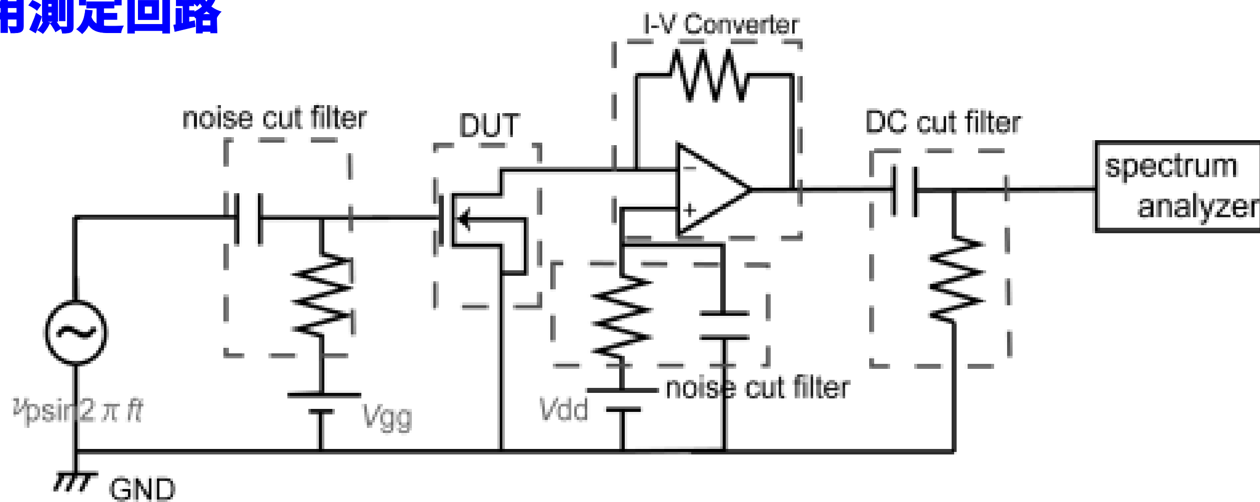




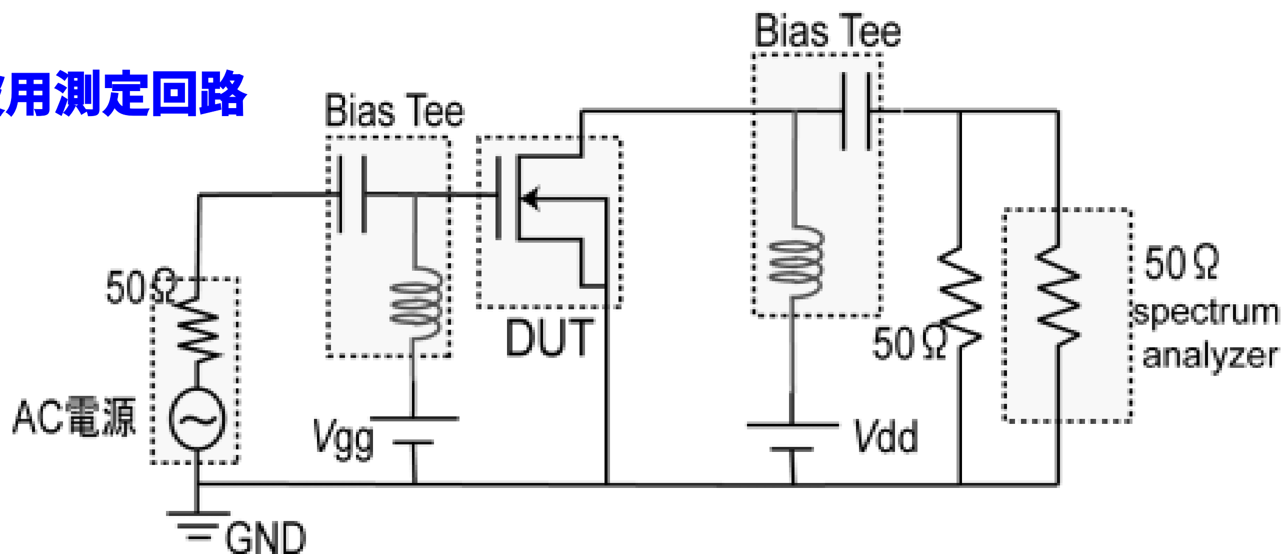
$$i = \frac{dI_{ds}}{dV_{gs}} V_p \cos(2\pi f_0 t) + \frac{1}{4}\frac{d^2 I_{ds}}{dV_{gs}^2} V_p^2 \cos(2\pi 2 f_0 t) + \frac{1}{24}\frac{d^3 I_{ds}}{dV_{gs}^3} V_p^3 \cos(2\pi 3 f_0 t) + \cdots$$



低周波用測定回路

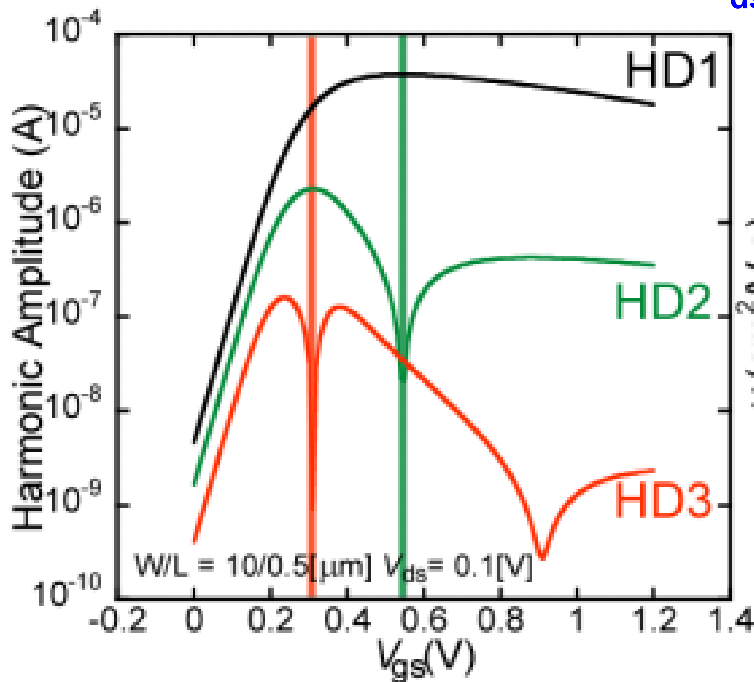


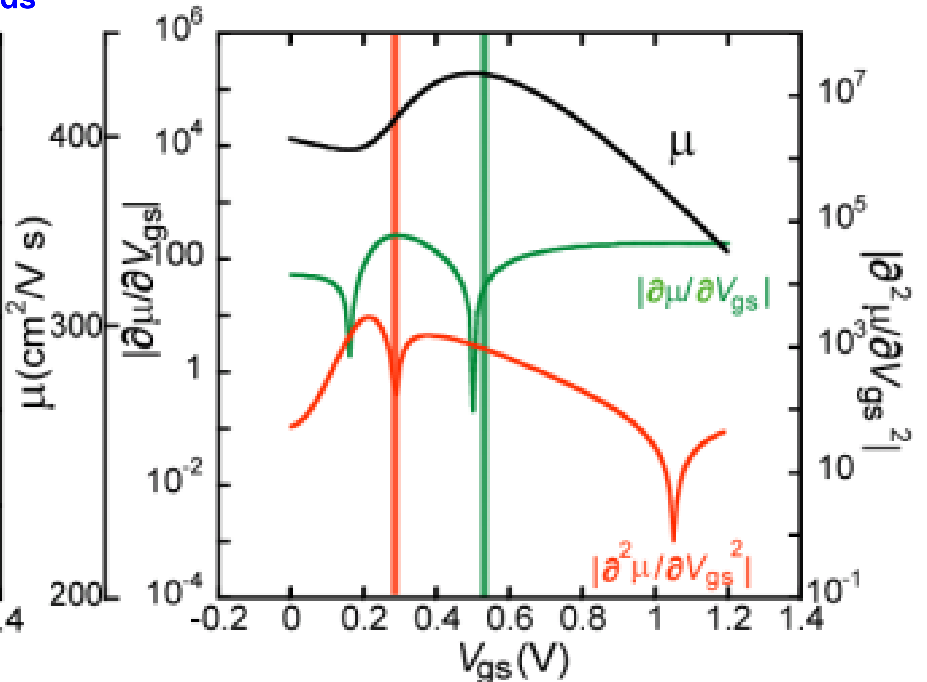
高周波用測定回路



低周波における高調波ひずみ

V_{ds}=0.1V

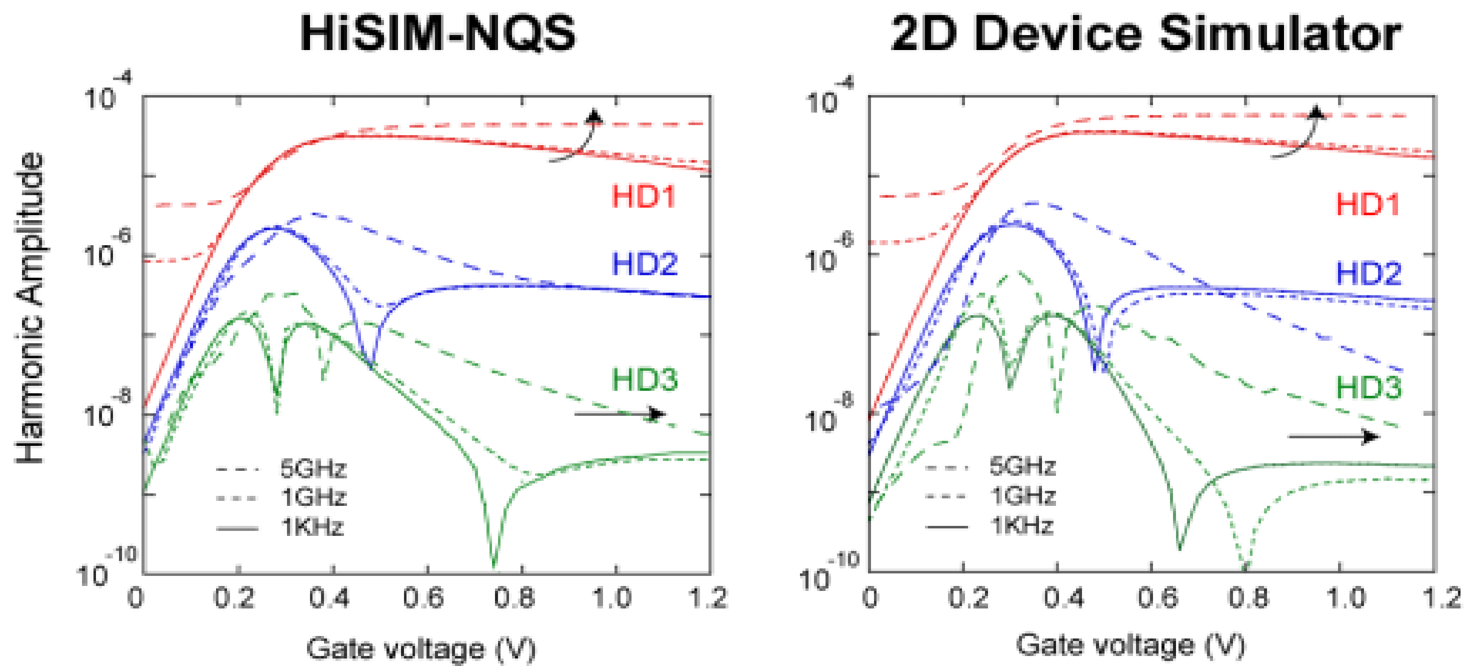




$$HD1 \approx \left| V_P \frac{\partial I_{ds}}{\partial V_{gs}} \right| \qquad HD2 \approx \left| -\frac{1}{4} V_P^2 \frac{\partial^2 I_{ds}}{\partial V_{gs}^2} \right| \qquad HD3 \approx \left| -\frac{1}{24} V_P^3 \frac{\partial^3 I_{ds}}{\partial V_{gs}^3} \right|$$

移動度特性がHD特性を支配している

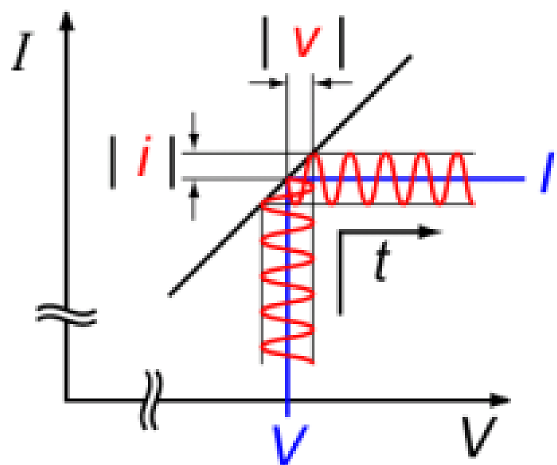
高周波における高調波ひずみ

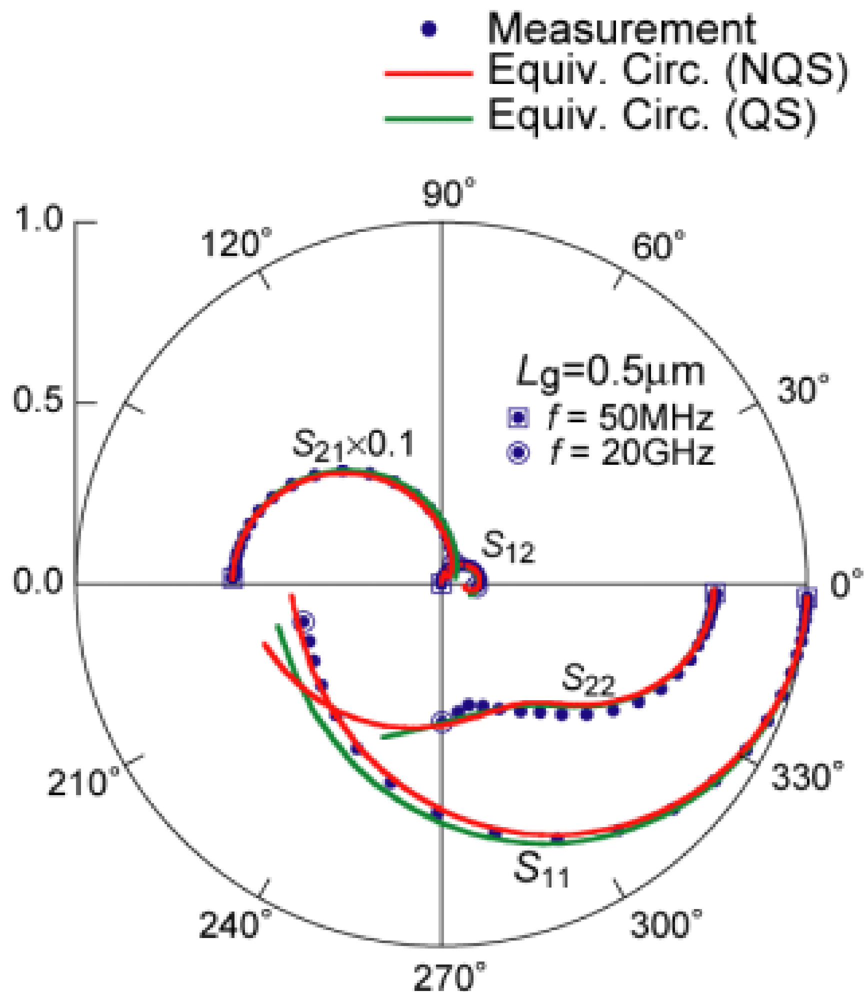


Carrier transit delay dominates the HD characteristics.

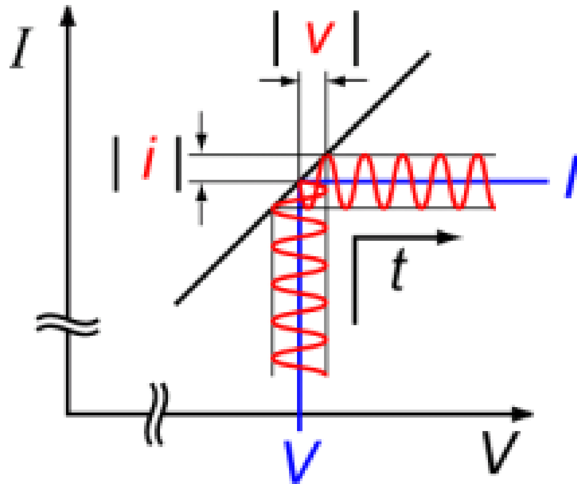
D. Navarro et al., SISPAD, p. 259, 05

Y-パラメタ特性



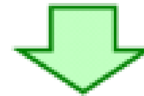


小信号解析



$$v = |v| \exp(j\omega t)$$
$$i = |i| \exp(j\omega t)$$

：小信号

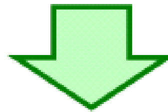


$$\begin{bmatrix} i_g \\ i_d \end{bmatrix} = \begin{bmatrix} Y_{gg}(\omega) & Y_{gd}(\omega) \\ Y_{dg}(\omega) & Y_{dd}(\omega) \end{bmatrix} \begin{bmatrix} v_g \\ v_d \end{bmatrix}$$

$$Y_{\alpha\beta} = \mathrm{Re}(Y_{\alpha\beta}) + \mathrm{Im}(Y_{\alpha\beta})$$

Admittance Matrix

S-Parameters : Power Measurement



Y-Parameters : Admittance

$$\begin{bmatrix} i_g \\ i_d \end{bmatrix} = \begin{bmatrix} Y_{11}(\omega) & Y_{12}(\omega) \\ Y_{21}(\omega) & Y_{22}(\omega) \end{bmatrix} \begin{bmatrix} v_g \\ v_d \end{bmatrix}$$

Admittance Matrix

$$Y_{\alpha\beta} = Re(Y_{\alpha\beta}) + Im(Y_{\alpha\beta})$$

Analytical Description:

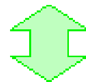
$$Y_{11}(\omega) = \omega^2 R_g C_{gg}{}^2 + j\omega C_{gg}$$

$$Y_{21}(\omega) = g_m - \omega^2 R_g C_{gg} (C_m + C_{gd})$$

HiSIM-SSA

$$Y_{dg} = \frac{g_m - \frac{2}{3} g_m \bar{s} \frac{\alpha(\alpha+2)}{(1+\alpha)^2} - \frac{2}{45} g_m \bar{s}^2 \frac{\alpha(1-\alpha)(2\alpha^2+8\alpha+5)}{(1+\alpha)^3} \cdots}{1 + \frac{4}{15} \bar{s} \frac{(1-\alpha)(1+3\alpha+\alpha^2)}{(1+\alpha)^3} + \frac{1}{45} \bar{s}^2 \frac{(1-\alpha)^2(1+4\alpha+\alpha^2)}{(1+\alpha)^4} \cdots}$$

$$\bar{s} = j\omega/\omega_0 = j\omega C_{gg}/g_m \qquad \alpha = U(L)/U(0)$$

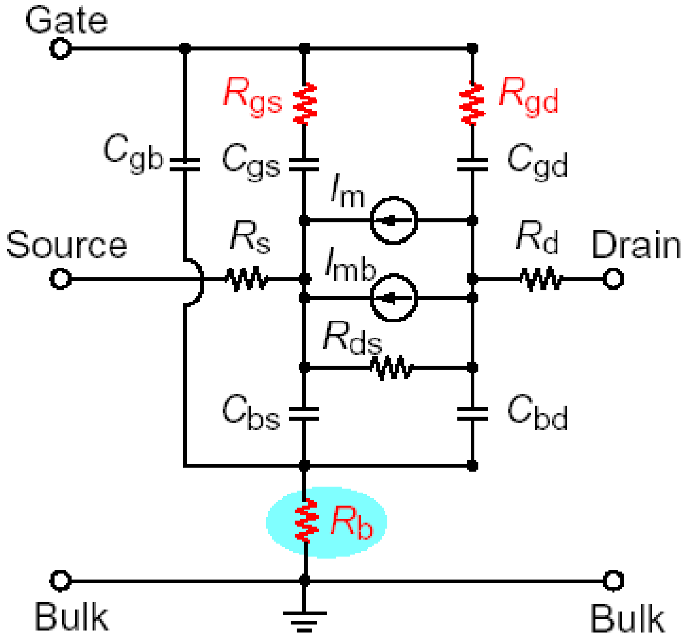


$$Y_{gg} = j\omega C_{gg}$$

$$Y_{dg} = g_m - j\omega(C_m - C_{gd})$$

S. Jinbou et al., Ext. Abs. SSDM, p. 26, 2002.
H. Ueno et al., Proc. SISPAD, p. 71, 2002.

等価回路



※$I_m = g_m + j\omega C_m$

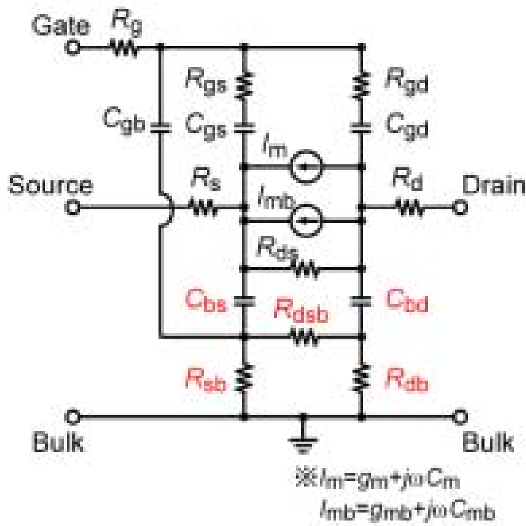
$I_{mb} = g_{mb} + j\omega C_{mb}$

R_b:基板ネットワーク

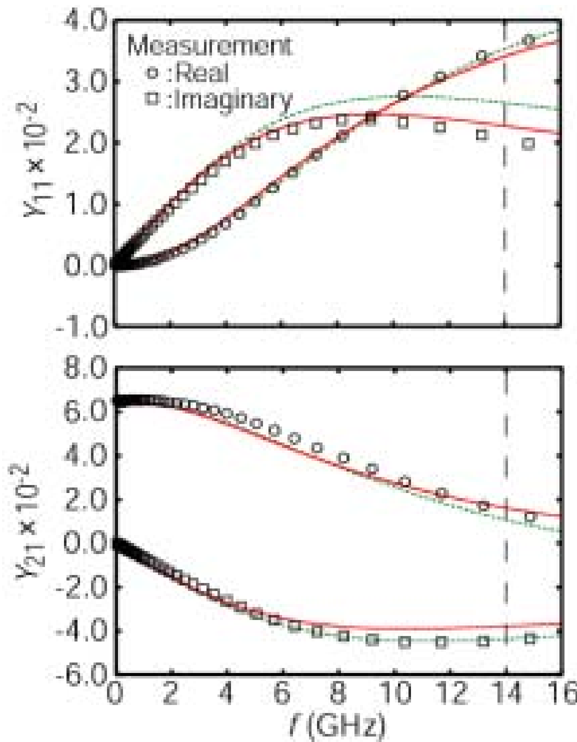
$$R_{gs} = \frac{2}{15} \frac{1}{\omega_0 C_{gs}}$$

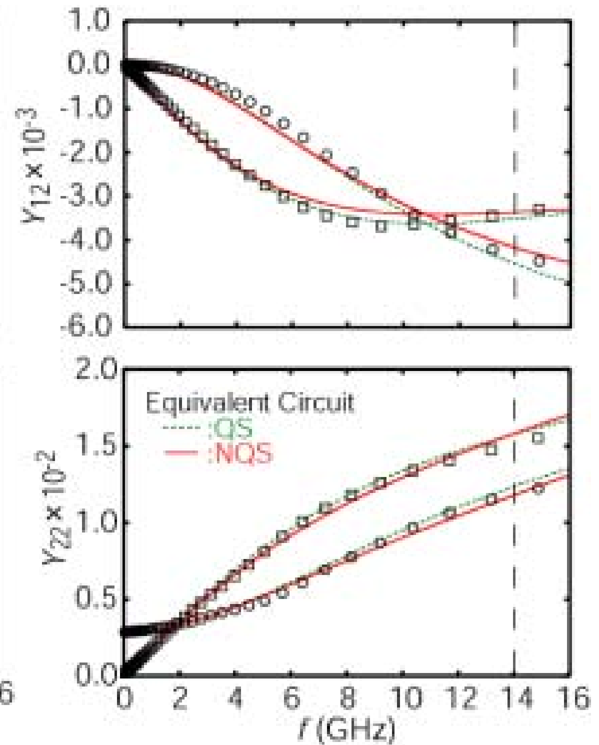
$$R_{gd} = \frac{1}{10} \frac{1}{\omega_0 C_{gd}}$$

$(\omega_0 = 2\pi f_t)$



パラメタ抽出が困難
NQSモデルが本質

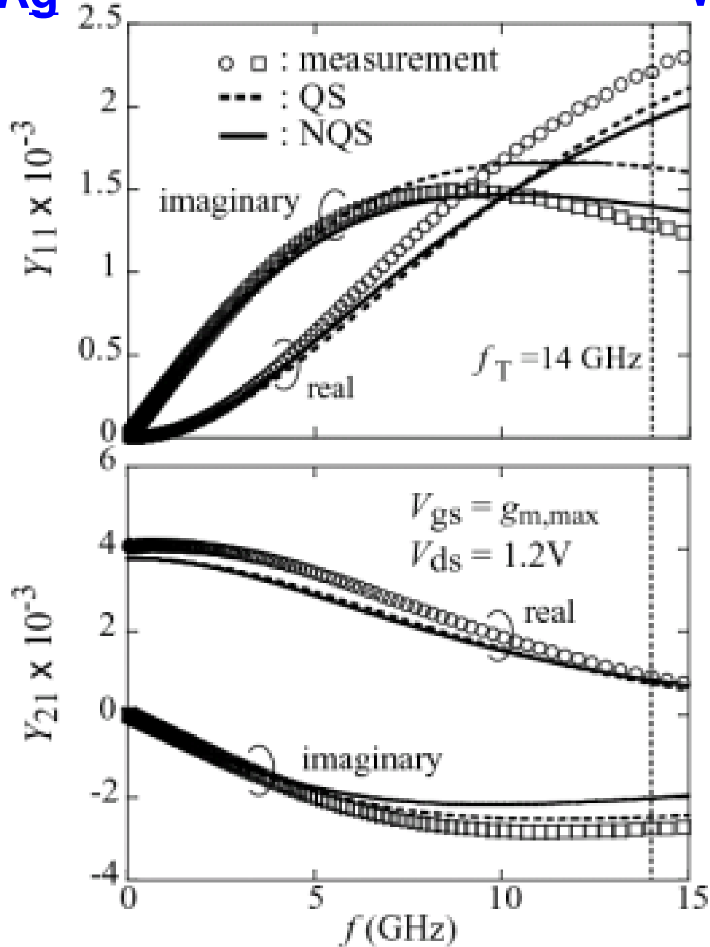




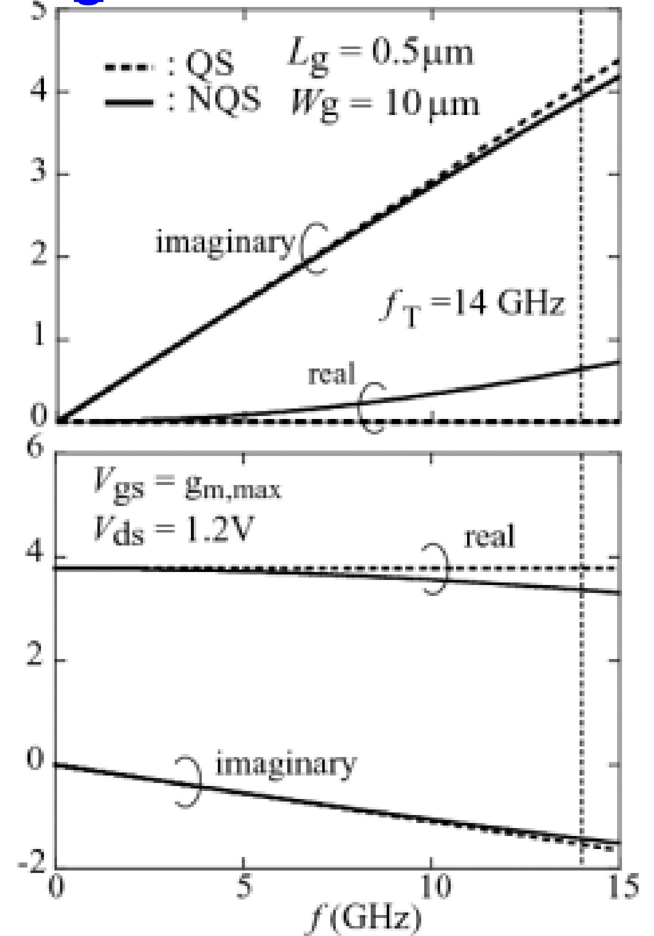
H. Kawano et al., Digest Microwave Symp., P. 2121, 2002.

NQS効果の検証

with *R*g



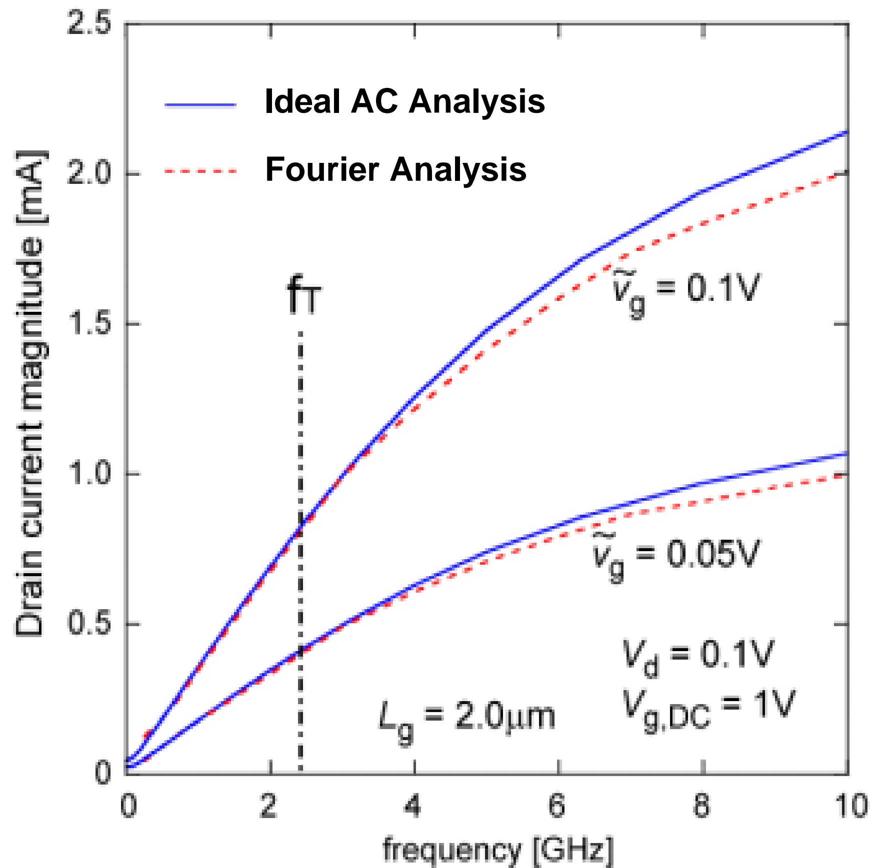
without *R*g



S. Jinbou et al., Ext. Abs. SSDM, P. 26, 2002.

Strong External Contribution

大信号モデルの小信号モデルへ拡張

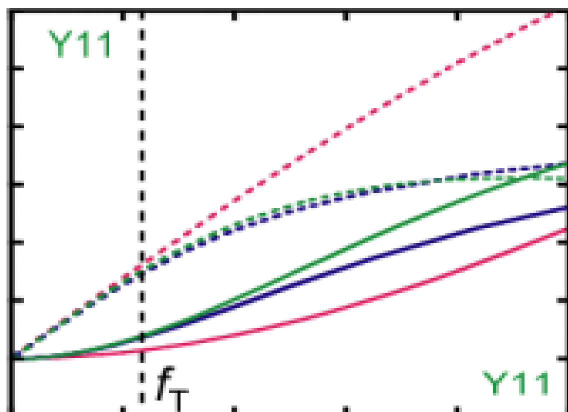


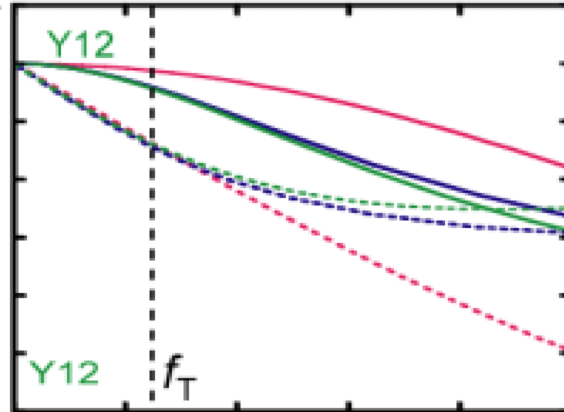
K. Machida et al., SiRF., p. 57, 2006.

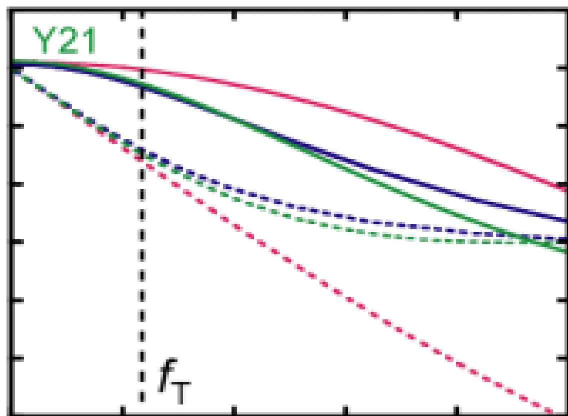
Y-パラメタ特性

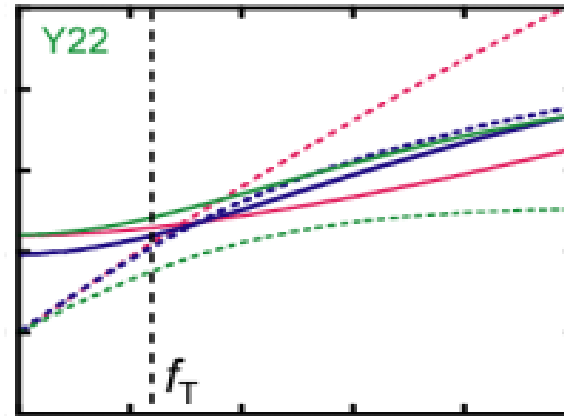
long small V_{ds}

Measurement Real
Measurement Imaginary
HiSIM-QS Real
HiSIM-QS Imaginary
HiSIM-NQS Real
HiSIM-NQS Imaginary







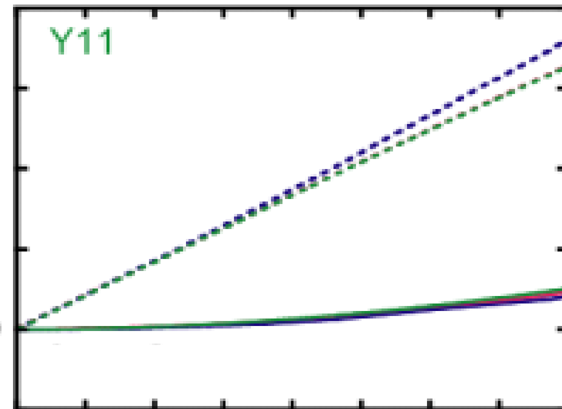


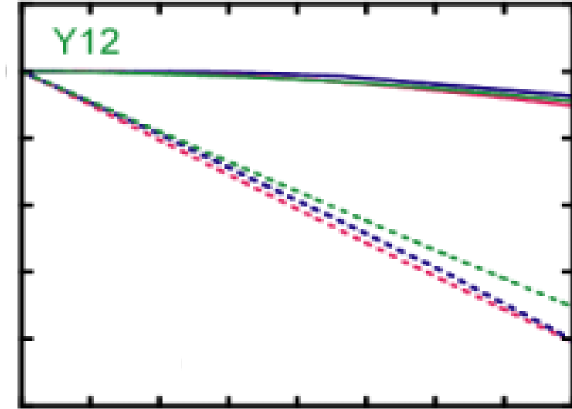
freq (GHz)

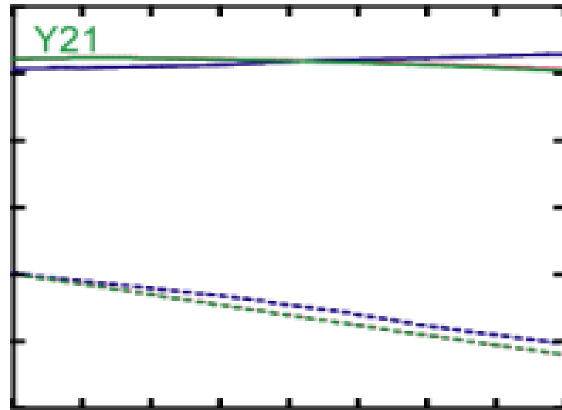
freq (GHz)

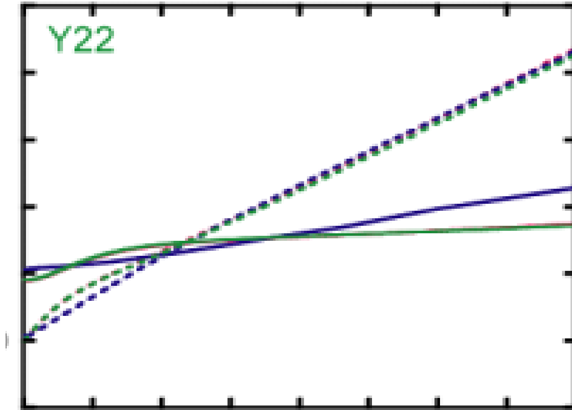
short large V_{ds}

— Measurement Real
---- Measurement Imaginary
— HiSIM-QS Real
---- HiSIM-QS Imaginary
— HiSIM-NQS Real
---- HiSIM-NQS Imaginary



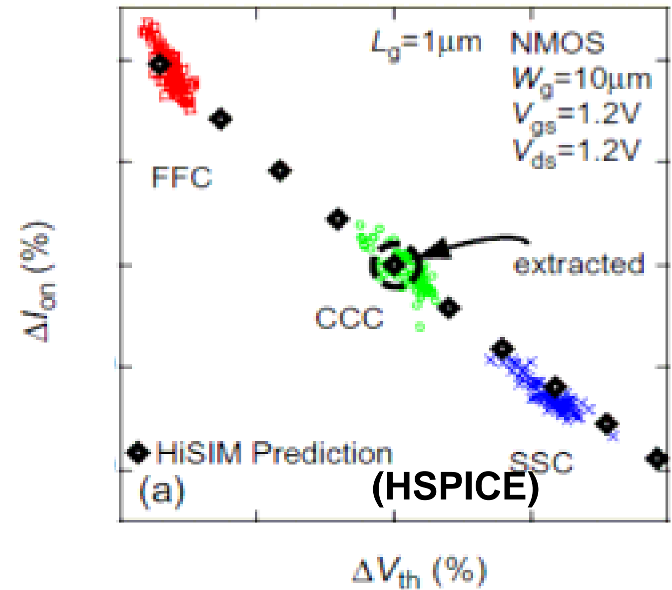




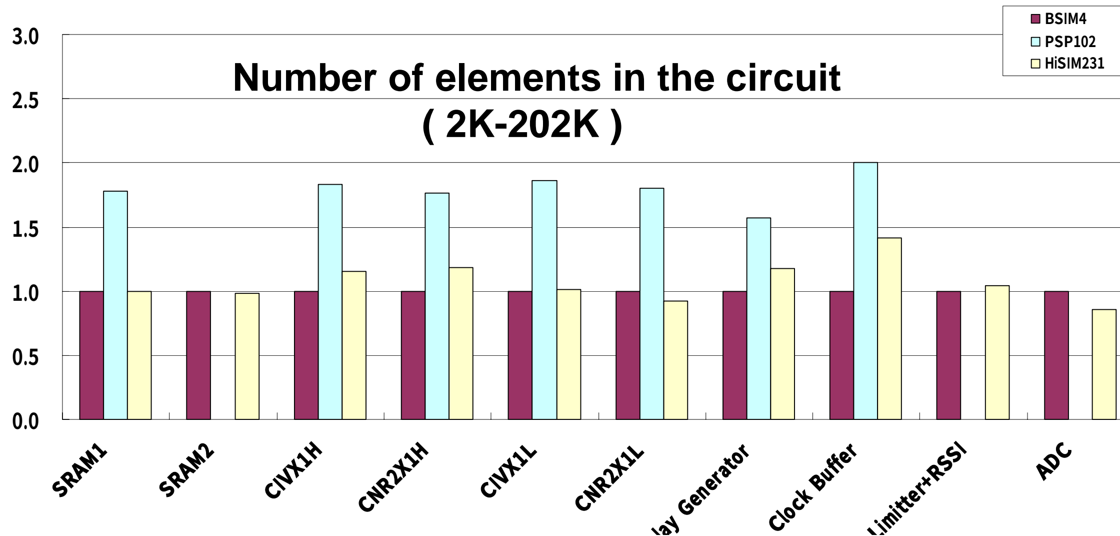


今後の課題

- ✓ ばらつきモデル
- ✓ Layout依存性の考慮
- ✓ 計算時間の高速化



CPU time Device: 90nm NMOS, PMOS

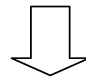


STARC, March'07

表面ポテンシャルモデルの特徴

$$I = qn\mu E$$

すべての物理量が同じルーツ(ポテンシャル) で計算されている



デバイス物理の基本

RFデバイスの特性評価

- ✓ ノイズ
- ✓ 高調波ひずみ
- ✓ Y-パラ

RFデバイス特性は電流‐電圧特性で決まる

回路モデルへの要求

- ✓ デバイス特性がコンパクトに記述されている
- ✓ すべての観測される現象がモデル化されている
- ✓ デバイス特性と回路性能の関係が予測できる
- ✓ 大規模回路がシミュレーションできる
- ✓ すべての市販しミュレータに組み込まれている

動向

- ✓ 先端MOSFETの性能向上
- ✓ これを十分に利用できる信頼性の高い回路モデル
- ✓ 回路モデルがBSIMからPSPあるいはHiSIMに移行