MOSFET Optimization in Deep Submicron Technology for Charge Amplifiers

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Abstract—The optimization of the input MOSFET for charge amplifiers in deep submicron technologies is discussed. After a review of the traditional approach, the impact of properly modeling the equivalent series noise and gate capacitance of the MOSFET is presented. It is shown that the minimum channel length and the maximum available power are not always the best choice in terms of resolution. Also, in an optimized front-end, the low frequency noise contribution to the Equivalent Noise Charge may depend on the time constant of the filter. As an example, results from the commercial TSMC 0.25µm CMOS technology are reported.

I. INTRODUCTION

APPLICATION Specific Integrated Circuits (ASICs) can be a valuable solution in front-end electronics for radiation sensors. Detection systems with high sensor pixelation can benefit from low power, low parasitics, high front-end channel density, and low cost per channel. In addition the ASICs are characterized by good radiation tolerance [1-3] and can integrate large amount of additional signal processing and functions in analog, mixed-signal and digital domains, offering further advantage in terms of power and real estate.

In a properly designed front-end the resolution is limited by the noise from the input transistor. Consequently a relevant phase of the design consists of optimizing the input MOSFET with respect to sensor, interconnects and the specific application. The choice of the polarity (n- or p-channel), size (length L and width W) and operating point (drain current density Jd and drain-to-source voltage Vds) are deterministic in achieving the best performance. The optimization process relies on equations, models and parameters that can be strongly dependent on the technology. As deep submicron CMOS technologies are developed and characterized for digital design, the process of optimizing the input MOSFET can become very challenging. This contribution would like to provide the low-noise front-end designers with techniques to keep pace with the rapid evolution of CMOS technology.

After a review of the traditional optimization process, the impact of properly modeling the series noise and gate capacitance of the MOSFET is discussed. As an example, the commercial TSMC 0.25µm CMOS technology is characterized and investigated for the best achievable resolution.

II. THE ENC EQUATION

The resolution of a front-end can be measured in terms of Equivalent Noise Charge (ENC) [4-7]. The ENC corresponds to the charge that must be delivered to the front-end in order to achieve a signal to noise ratio equal to the unity.

In the following we will assume that the resolution of the front-end is dominated by the noise from the input MOSFET, characterized by a gate length L, a gate width W and a drain current density Jd=J/W. We will also assume that the front-end implements a time invariant filter with overall pulse response to a charge Q equal to Q·h(t), and characterized by a maximum Q·h(t)max.

In Fig. 1 a schematic of the front-end for the evaluation of the ENC is shown. The parallel noise contribution \( \overline{V_n^2} \), characterized by a unilateral power spectral density \( S_{in} \), is typically related to the sensor leakage current, to its bias circuitry and to the charge amplifier feedback circuitry. In the following we will initially assume a white density \( S_{in} \). Then, since this contribution is not related to the input MOSFET, this term will be neglected. The series noise contribution \( \overline{V_n^2} \), characterized by a unilateral power spectral density \( S_{vn} \), is dominated by noise processes in the input MOSFET. The low frequency term, known as 1/f noise, is strongly technology related. The white term is strongly related to the transconductance \( g_m \) of the MOSFET. The input capacitance \( C_i \) includes the sensor capacitance, the feedback capacitance and any parasitic connected to the input node and not dependent on the size (W,L) of the input MOSFET.

The ENC can be expressed as follows:

\[
\text{ENC}^2 = \frac{1}{2\tau} \int_{0}^{\infty} \left[ \frac{S_{vn}}{H(j\omega)^2} + \left( C_i + C_g \right) \omega^2 S_{vn} |H(j\omega)|^2 \right] d\omega
\]

where \( \tau \) is the time constant of the filter and \( H(s) \) is the transfer function.

This work was supported by the US Department of Energy, Contract No. DE-AC0298CH10886.

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Laplace transform of the pulse response $h(t)$. Introducing the typical spectral densities previously discussed it follows:

$$
\text{ENC}^2 = \left( C_{in} + C_d \right) \left[ \int_0^{\infty} |H(j\omega t)|^2 \omega^2 d\omega + A_i \int_0^{\infty} |H(j\omega t)| \omega d\omega \right] + \frac{S_p}{2 \pi \cdot h(t/\tau)_{\text{max}}} + \frac{a}{\tau^4 \cdot h(t/\tau)_{\text{max}}} \tag{2}
$$

where the time is normalized to an arbitrary time constant $\tau$, and $\tau H(\omega t)$ is the Laplace transform of $h(t)$. Assuming as time constant the pulse peaking time $\tau_p$, calculated from 1% to the peak, and transforming the integrals variable, the (2) becomes:

$$
\text{ENC}^2 = \left( C_{in} + C_d \right) \left[ \int_0^{\infty} |H(j\omega t)|^2 \omega^2 d\omega + A_i \int_0^{\infty} |H(j\omega t)| \omega d\omega \right] + \frac{S_p}{2 \pi \cdot h(t/\tau)_{\text{max}}} + \frac{a}{\tau^4 \cdot h(t/\tau)_{\text{max}}} \tag{3}
$$

where the three ENC coefficients:

$$
a_w = \frac{1}{2 \pi \cdot h(t/\tau)_{\text{max}}} \quad a_f = \frac{1}{2 \pi \cdot h(t/\tau)_{\text{max}}} \quad a_p = \frac{1}{2 \pi \cdot h(t/\tau)_{\text{max}}} \tag{4}
$$

depend only on the type of filter, and $H(j\omega)$ is obtained replacing $\omega$ with $x/\tau_p$ in $H(j\omega)$.

In Table I the coefficients for some commonly adopted time invariant filters are reported, along with the ratio between the pulse width $\tau_{in}$ calculated from 1% to 1% of the curve, and the peaking time $\tau_p$, calculated from 1% to the peak. The R-type filters have real poles only while the C-type filters, commonly adopted in commercial discrete shapers, have complex conjugate poles [8]. The order of the filters in Table I ranges from 2$^nd$ to 7$th$.

It can be observed that, as the order increases, the ratio $\tau_p/\tau_{in}$ decreases making the high order filters more suitable for high rate applications. Also, at equal order, the C-type filters offer a better $\tau_p/\tau_{in}$ ratio than the R-type filters. Finally, the advantage of zero area for bipolar filters, which in absence of effective baseline stabilization at high rate can be relevant, is compensated by worse values in the series coefficients $a_w$ and $a_f$, especially for high orders.

Without loss of generality we will simplify our analysis by assuming $a_w = 1$, $a_f = 0.5$, and $a_p = 0.5$, which are close to the typical values for a good time invariant unipolar shaper.

In the rest of the analysis the parallel noise contribution $S_p$ will be neglected, and the general expression for the ENC in (3) will be simplified as follows:

$$
\text{ENC}^2 = \left( C_{in} + C_d \right) \left[ \int_0^{\infty} |H(j\omega t)|^2 \omega^2 d\omega + A_i \int_0^{\infty} |H(j\omega t)| \omega d\omega \right] \tag{5}
$$

### TABLE I

| Filter | Shape | $a_w$ | $a_f$ | $a_p$ | $\tau_p/\tau_{in}$ |
|--------|-------|-------|-------|-------|------------------|
| Triang. |       | 1.00  | 0.50  | 0.50  | 2.00             |
| RU-2   |       | 0.92  | 0.59  | 0.92  | 7.66             |
| RU-3   |       | 0.82  | 0.54  | 0.86  | 5.04             |
| RU-4   |       | 0.85  | 0.53  | 0.57  | 4.17             |
| RU-5   |       | 0.89  | 0.52  | 0.52  | 3.73             |
| RU-6   |       | 0.92  | 0.52  | 0.48  | 3.46             |
| RU-7   |       | 0.94  | 0.51  | 0.46  | 3.27             |
| CU-2   |       | 0.83  | 0.59  | 0.88  | 6.31             |
| CU-3   |       | 0.85  | 0.54  | 0.61  | 3.92             |
| CU-4   |       | 0.91  | 0.55  | 0.51  | 3.16             |
| CU-5   |       | 0.96  | 0.52  | 0.46  | 2.84             |
| CU-6   |       | 1.01  | 0.52  | 0.42  | 2.66             |
| CU-7   |       | 1.04  | 0.51  | 0.40  | 2.55             |
| RB-2   |       | 1.03  | 0.75  | 1.01  | 16.6             |
| RB-3   |       | 1.11  | 0.77  | 0.76  | 9.67             |
| RB-4   |       | 1.30  | 0.81  | 0.66  | 7.68             |
| RB-5   |       | 1.47  | 0.84  | 0.62  | 6.60             |
| RB-6   |       | 1.61  | 0.87  | 0.59  | 5.94             |
| RB-7   |       | 1.74  | 0.89  | 0.57  | 5.53             |
| CB-2   |       | 1.08  | 0.79  | 1.02  | 12.9             |
| CB-3   |       | 1.30  | 0.86  | 0.76  | 7.29             |
| CB-4   |       | 1.58  | 0.93  | 0.67  | 5.60             |
| CB-5   |       | 1.86  | 0.98  | 0.63  | 4.81             |
| CB-6   |       | 2.11  | 1.02  | 0.59  | 4.37             |
| CB-7   |       | 2.31  | 1.06  | 0.58  | 4.11             |

### III. CONSTRAINTS, VARIABLES AND MODELS

The input MOSFET must be optimized for the minimum ENC. The ENC depends on some parameters not related to the input MOSFET, specifically the input capacitance $C_{in}$, the peaking time $\tau_p$, and the maximum power $P_{d_{max}}$ available to the input MOSFET. Typically the input capacitance $C_{in}$ is set by the sensor and interconnects, the maximum power $P_{d_{max}}$ depends on system level constraints, and the peaking time $\tau_p$ is set by the rate or ballistic deficit. In presence of non-negligible parallel noise contribution a further constraint on the peaking time may occur.

The typical optimization process consequently assumes $C_{in}$, $\tau_p$, and $P_{d_{max}}$ as constraints. One further constraint derives from the use of a single supply $V_{dd}$:

$$
I_d = \frac{P_d}{V_{dd}} \Rightarrow J_d = \frac{I_d}{W} \tag{6}
$$

where $J_d$ is the MOSFET drain current density. The (6) establishes a constraint on the $J_dXW$ product. It follows that, once $C_{in}$, $\tau_p$, and $P_d$ are defined, the ENC depends only on two variables: $W$ (or $J_d$) and $L$. The optimization process will consequently return the two optimum values $W_{opt}$ (or $J_{d_{opt}}$) and $L_{opt}$ that give the minimum ENC ($\text{ENC}_{opt}$).

In order to proceed with the optimization of the input MOSFET, the parameters $C_{in}$, $S_p$, and $A_i$ must be expressed as functions of $W$ (or $J_d$) and $L$. In the following sections the modeling of these parameters will be discussed, starting from the solution that the CMOS designers frequently adopted in the past, before the advent of the deep submi-
cron CMOS technologies.

IV. CLASSICAL (C) MODEL FOR $C_v$, $S_w$ AND $A_f$

In the past the front-end designers frequently adopted the following models [9-13]:

$$C_v = C_{ox} W/L, \quad S_w = \frac{2}{3} \frac{4kT}{g_m(J_d,L)}, \quad A_f = \frac{K_f}{C_{ox} W/L} \quad (7)$$

where $C_{ox}$ is the oxide capacitance, $k$ is the Boltzmann constant, $T$ is the absolute temperature, $g_m$ is the MOSFET transconductance dependent on $J_d$ and $L$, and $K_f$ is the $1/f$ noise coefficient. We will refer to this model as C-model.

Introducing (7) in (5) it follows:

$$ENC^2 = (C_{ox} + C_{ox} W/L)^2 \left[ \frac{1}{2} \frac{4kT}{3g_m(J_d,L)W} + \frac{K_f}{C_{ox} W/L} \right] \quad (8)$$

where $g_m(J_d,L)$ is the transconductance per unit of $W$.

The function $g_m(J_d,L)$ can be easily extracted for specific values of $L$ from a Spice simulation using the model and parameters available for the technology. The coefficient $K_f$ can be extracted from a measurement of the input noise spectral density, it is assumed to be independent of $L$ and it may differ for n-channels ($K_{fn}$) and p-channels ($K_{fp}$).

For the TSMC 0.25µm technology $C_{ox}=6.1fF/µm^2$, $g_m(J_d,L)$ can be obtained from BSIM3v3.1 simulations (see Fig. 2) and, from measurements on samples with minimum $L$ at 1kHz, it follows $K_{fn}=6.10^{-24}$ J and $K_{fp}=0.30.10^{-24}$ J. In Fig. 3 the optimum ENC for $C_v=1pF$ and $\tau=1µs$, calculated for n- and p-channels with different $L$ as function of $P_d$ are shown. For minimum $L$ the white and $1/f$ components and the optimum $W$ are also reported. It can be observed that in this case the p-channel offers a better resolution than the n-channel (lower $K_v$) and this is generally true. It also indicates that the choice $L_{opt}=L_{min}$ and $P_{dopt}=P_{dmin}$ offers better results in an amount that depends on the relative ratio between the white and $1/f$ contributions.

A slower interpolation function is reported in [15], along with the companion model for the gate capacitance. A comparison to $C_v$ simulations and to noise measurements [2] suggests that (11), here adopted, might be more accurate. In
Fig. 4 the gamma coefficient $\gamma(J_d,L)$ versus the inversion coefficient $u(J_d,L)$ is shown. The regions of operation for the MOSFET are also indicated. In most cases the optimum input MOSFET operates in moderate inversion.

Introducing the new model (10)-(11) in (5) it follows:

$$\text{ENC}^2 = (C_m + C_{in}) W L T_{p} \left[ \frac{1}{\tau_p} \left( \frac{\alpha_{in} (J_d, L)}{C_{in}} \right)^{2} + 2 \pi \frac{K_f}{C_{in} W L} \right]$$

For the TSMC 0.25$\mu$m technology typical values for $\alpha_{in}$ are in the range 300-600 fF/mm, where the lower values are reported by TSMC and the higher values are reported by the MOSIS Service [21].

In Fig. 5 the optimum ENC from (13) for $C_{in}=1pF$ and $\tau_p=1\mu$s, calculated for n- and p-channels with minimum L as function of $P_d$, is shown, compared to C-model results. The optimum W is also reported. It can be observed that the new model provides a better estimate especially at low power, where the white noise component dominates. A relatively small difference in terms of optimization ($W_{opt}$) can also be observed. The $L_{opt}=L_{min}$ and $P_{d_{opt}}=P_{d_{max}}$ still appears the best choice.

VI. ENHANCED (E) MODEL FOR $C_g$

The gate capacitance $C_g$ in (5) includes any term dependent on the size (W,L) of the input MOSFET. A better estimate of $C_g$ should consider the gate-to-source and gate-to-drain overlap components [15]. These extrinsic terms are proportional to the width W of the MOSFET and, in deep submicron technologies, are typically not negligible. In addition, the intrinsic component of $C_g$ in saturation is a fraction $=2/3$ of $C_{ox} W L$ [15]. An equation that improves the estimate of $C_g$, here referred as Ci-model, is:

$$C_g = 2C_{in} W + \frac{2}{3} C_{in} W L$$  \hspace{1cm} (14)

where $C_{in}$ is the overlap capacitance density, equal for drain and source. Introducing (14) in (13) it follows:

$$\text{ENC}^2 = \left( C_m + 2C_{in} W + \frac{2}{3} C_{in} W L \right)^2 \times \left( 1 - \frac{\pi \frac{K_f}{C_{in} W L}}{\tau_p \left( \frac{\alpha_{in} (J_d, L)}{C_{in}} \right)^{2}} \right)$$  \hspace{1cm} (15)

For the TSMC 0.25$\mu$m technology typical values for $C_{ov}$ are in the range 300-600 fF/mm, where the lower values are reported by TSMC and the higher values are reported by the MOSIS Service [21].

In Fig. 6 the optimum ENC from (15) for $C_{in}=1pF$ and $\tau_p=1\mu$s, calculated for n- and p-channels with minimum L as function of $P_d$, is shown. When compared to Figs. 3 and 5 it can be observed that the choice $L=\min$ does not apply anymore. Both n-channel and p-channel can offer better resolution for L higher than $L_{min}$.

This conclusion is valid whenever 1/f noise dominates, and can be understood by rewriting (15) for this case:

$$\text{ENC}^2 = \left( C_m + 2C_{in} W + \frac{2}{3} C_{in} W L \right)^2 \frac{K_f}{C_{in} W L}$$  \hspace{1cm} (16)

By calculating the differential of (15) with respect to $W$,
the optimum width for the minimum ENC can be calculated for each L:

\[ W_{opt} = \frac{C_m}{2C_{ov} + \frac{2}{3}C_{ov}L}. \]  

(17)

Capacitive matching applies \((C_e = C_m)\) and, by superposition, the \(ENC_{opt}\) can be written:

\[ ENC_{opt}^2 = \frac{8}{3}C_m \pi \lambda \left(1 + \frac{3C_{ov}}{C_{ov}L}\right). \]  

(18)

The optimum width for the minimum ENC can be calculated for each L:

\[ W_{opt} = \frac{C_m}{2C_{ov} + \frac{2}{3}C_{ov}L}. \]  

(17)

The plot in Fig. 7, derived from (18) for n-channel MOSFETs, illustrates the impact of L on the optimum ENC for the case of dominant 1/f noise. Intuitively, as L increases the 1/f noise contribution decreases more rapidly than the increase in gate capacitance.

\[ ENC_{opt}^2 = \frac{8}{3}C_m \pi \lambda \left(1 + \frac{3C_{ov}}{C_{ov}L}\right). \]  

(18)

The Ci-model (14) for \(C_g\) doesn’t take into account the dependence of the intrinsic gate capacitance on the drain current density \(J_d\). This dependence in deep submicron technologies can be not negligible, especially in the transition from weak, thorough moderate, to strong inversion. An enhanced (E) model for \(C_g\) can be written as:

\[ C_g = C_{gw}(J_d, L)W. \]  

(19)

where \(C_{gw}(J_d, L)\) is the gate capacitance density and includes the bias dependence, intrinsic and extrinsic components.

As for \(g_m(J_d, L)\) the function \(C_{gw}(J_d, L)\) can be extracted for specific values of L from a Spice simulation using the model and parameters available for the technology. For the TSMC 0.25\(\mu\)m \(C_{gw}(J_d, L)\) can be obtained from BSIM3v3.1 simulations as shown in Fig. 8. The results for C- and Ci-models for the n-channel at minimum L are reported for comparison. It is worth noting that the cutoff frequency of the MOSFET, given by \(g_m/(2\pi C_g)\), remains an increasing function of \(J_d\) and, in the regions of interest, typically exceeds the tens of MHz.

\[ ENC_{opt}^2 = \frac{8}{3}C_m \pi \lambda \left(1 + \frac{3C_{ov}}{C_{ov}L}\right). \]  

(18)

Introducing the (19) in (13) it follows:

\[ ENC_{opt}^2 = [C_m + C_{ov}(J_d, L)]W\left\{\frac{1}{\tau_p} \frac{\alpha_{m\text{f}}(J_d, L)kT}{B_{m\text{f}}(J_d, L)} + \frac{K_f}{C_{ov}WL}\right\}. \]  

(20)

In Fig. 9 the optimum ENC from (20) for \(C_e = 1\)pF and \(\tau_p = 1\)\(\mu\)s, calculated for n- and p-channels with different L as function of \(P_d\), is shown. For two cases the 1/f component and the optimum W are also reported. It can be observed again that the choice \(L_{opt} = L_{min}\) does not apply. In addition, for n-channel MOSFETs, the choice \(P_{dopt} = P_{dmax}\) does not apply, since they can offer better resolution at \(P_d\) lower than the maximum available.

\[ ENC_{opt}^2 = [C_m + C_{ov}(J_d, L)]W\left\{\frac{1}{\tau_p} \frac{\alpha_{m\text{f}}(J_d, L)kT}{B_{m\text{f}}(J_d, L)} + \frac{K_f}{C_{ov}WL}\right\}. \]  

(20)

This conclusion is valid whenever 1/f noise dominates, and can be understood by rewriting (20) for this case:

\[ ENC_{opt}^2 = [C_m + C_{gw}(J_d, L)]W\left\{\frac{1}{\tau_p} \frac{\alpha_{m\text{f}}(J_d, L)kT}{B_{m\text{f}}(J_d, L)} + \frac{K_f}{C_{gw}WL}\right\}. \]  

(21)

By taking into account (6) and calculating the differential of (21) with respect to \(W\), the optimum width for the minimum ENC can be calculated for each L:

\[ ENC_{opt}^2 = [C_m + C_{gw}(J_d, L)]W\left\{\frac{1}{\tau_p} \frac{\alpha_{m\text{f}}(J_d, L)kT}{B_{m\text{f}}(J_d, L)} + \frac{K_f}{C_{gw}WL}\right\}. \]  

(20)
\[ W_{\text{opt}} = \frac{C_m}{C_{gw}(J_d, L) \left[ 1 - 2 \frac{J_d}{C_{gw}(J_d, L)} \frac{\partial C_{gw}(J_d, L)}{\partial J_d} \right]} . \]  

(22)

Capacitive matching does not apply except for the regions where \( \frac{\partial C_{gw}}{\partial J_d} = 0 \) and, by superposition, the ENC_{opt} turns out to be:

\[ \text{ENC}_{\text{opt}}^2 = \left\{ \begin{array}{ll} 1 + \frac{1}{\eta(J_d, L)} C_m \pi K_f \frac{C_{gw}(J_d, L)}{C_{in}} L & \text{if } \frac{\partial C_{gw}}{\partial J_d} \neq 0 \\
1 - 2 \frac{J_d}{C_{gw}(J_d, L)} \frac{\partial C_{gw}(J_d, L)}{\partial J_d} & \text{if } \frac{\partial C_{gw}}{\partial J_d} = 0 \end{array} \right. \]  

(23)

The plot in Fig. 10, derived from (23) for n-channel with minimum \( L \) and \( C_{\text{in}} = 1 \text{pF} \), illustrates the impact of \( P_d \) on the optimum ENC for cases of dominant 1/\( f \) noise. The ratio \( C_{\text{opt}} / C_{\text{in}} \) is also reported. Intuitively, as \( P_d \) increases the gate capacitance increases while the 1/\( f \) noise contribution does not change.

VII. ENHANCED (E) MODEL FOR LOW-FREQUENCY NOISE

In Fig. 11 the typical equivalent input noise spectral densities measured on n- and p-channels in TSMC 0.25\( \mu \text{m} \) with different \( L \) are shown, and are compared to the 1/\( f \) slope. Two relevant details should be observed. The first concerns the slope, which differs from 1/\( f \), being in this case higher for p-channels and lower for n-channels. The second concerns the ratio between spectra, which is higher \(( \approx 2.4 \text{ for } L=0.24\mu \text{m} \text{ vs } L=0.48\mu \text{m}) \) compared to the square root of the ratio between \( L \) \(( \approx 1.4 \text{ for } L=0.24\mu \text{m} \text{ vs } L=0.48\mu \text{m}) \). A similar trend for short channels in deep submicron technologies was reported by other authors [2,22-25].

In Fig. 12 the typical spectra for n- and p-channels with \( L=0.24\mu \text{m} \) at different drain current densities \( J_d \) are shown. The dependence of the low-frequency component on the bias point appears negligible. This result seems in agreement with others in the literature for MOSFETs operating from weak inversion up to the border between moderate and strong inversion [3,26-29], which is the region of interest for our applications. Some authors have reported an increase moving towards very strong inversion [28-30].

The results reported in Figs. 11 and 12 indicate that the low-frequency component of the noise power spectral density could be better approximated by using the following Enhanced (E) equation:

\[ S_y = \frac{A_y}{f^{\alpha_y}} = \frac{K_y(L)}{C_{in} W L f^{\beta_y}} \]  

(24)

where \( K_y(L) \) now depends on \( L \), and the slope depends on the coefficient \( \alpha_y \).

The non-1/\( f \) slope requires a review of the low-frequency component ENC_{LF} of the ENC. The second term of (2) now becomes:

\[ \text{ENC}_{LF}^2 = \frac{\left( C_m + C_g \int J \left| h \right| (J \sigma) \right) d\omega d\omega}{2\pi \hbar (t / \tau)^\omega_{\max}} . \]  

(25)

and (3) can be rewritten:

\[ \text{ENC}^2 = \left( C_m + C_g \right)^2 S_p \tau_p + A_y \left( \frac{2\pi N_y}{\tau_p} - \frac{N_y}{\tau_p} \right) + S_p \tau_p N_p \]  

(26)
where the ENC coefficient \(a_n(\alpha_f)\) is given by:

\[
a_n(\alpha_f) = \frac{\int |H(j\omega)|^2 \omega^{\alpha_f-1} d\omega}{2\pi h(t/\tau_p)}.
\]  

Concerning the two other coefficients in (4) it is worth noting that \(a_w=a_n(0)\) and \(a_p=a_n(2)\). In Table 2 the ratio \(\rho_f=a_n(\alpha_f)/a_n(1)\) is calculated for the filters of Table 1.

### TABLE II

| Filter | Relative coeff. \(\rho_f\) |
|--------|---------------------------|
| RU-2   | 1.0                        |
| RU-3   | 0.9                        |
| RU-4   | 1.0                        |
| RU-5   | 0.9                        |
| RU-6   | 1.1                        |
| RU-7   | 1.2                        |
| CU-2   | 0.8                        |
| CU-3   | 0.9                        |
| CU-4   | 1.0                        |
| CU-5   | 0.9                        |
| CU-6   | 1.0                        |
| CU-7   | 0.8                        |
| RB-2   | 0.8                        |
| RB-3   | 0.9                        |
| RB-4   | 1.0                        |
| RB-5   | 0.9                        |
| RB-6   | 1.0                        |
| RB-7   | 0.9                        |
| CB-2   | 0.8                        |
| CB-3   | 0.9                        |
| CB-4   | 1.0                        |
| CB-5   | 0.9                        |
| CB-6   | 1.0                        |
| CB-7   | 0.8                        |

The final expression for the ENC related to series noise, adopting the E-model for \(S_w\), \(C_g\) and \(S_f\), becomes:

\[
\text{ENC}^2 = [C_{in} + C_{ow}(J_d,L)] \cdot W \times \left[ \frac{1}{\tau_p} \frac{\alpha_w(n+1)}{\tau_p} \frac{\omega}{\tau_p} \frac{K_f}{C_{ow}WL} \right] \frac{\alpha_f}{2\pi h(t/\tau_p)}.
\]  

For the TSMC 0.25\(\mu\)m technology typical values for \(K_f\) and \(\alpha_f\) are reported in Table 3. As in previous cases we will simplify our analysis by assuming \(\rho_w=1.05\) for n-channels and \(\rho_p=0.95\) for p-channels, values close to the typical for a good time invariant unipolar shaper.

### TABLE III

| Filter | Typical values of \(K_f\) and \(\alpha_f\) for the TSMC 0.25\(\mu\)m technology. |
|--------|--------------------------------------------------------------------------------|
| Nch    | Pch                                   |
| Noh-0.24\(\mu\)m | Pch-0.24\(\mu\)m | Pch-0.36\(\mu\)m | Noh-0.36\(\mu\)m | Pch-0.24\(\mu\)m | Pch-0.36\(\mu\)m |
| \(K_f\) | 2.71 \times 10^{-12} | 1.40 \times 10^{-12} | 9.97 \times 10^{-13} | 6.00 \times 10^{-13} | 6.50 \times 10^{-13} |
| \(\alpha_f\) | 0.85 | 0.85 | 0.85 | 0.95 | 1.08 | 1.08 |

As consequence of the non-1/\(f\) slope, a front-end optimized for a peaking time \(\tau_p^{opt}\) will show in the ENC a low-frequency noise component dependent on \(\tau_p\), as shown in Fig. 13 for the case \(C_{in}=1\)\(\mu\)F, \(P_d=200\)\(\mu\)W, \(\tau_p^{opt}=1\)\(\mu\)s.

### VIII. Achievable Resolution in TSMC 0.25\(\mu\)M

By applying (28) to the TSMC 0.25\(\mu\)m CMOS it is possible to estimate the ENC achievable with this technology. The results below will give a general idea of what to expect. An exhaustive analysis is beyond the scope of this work.

- **Fig. 13.** Simulated ENC vs \(\tau_p\) for a design optimized for \(C_{in}=1\)\(\mu\)F, \(P_d=200\)\(\mu\)W, \(\tau_p^{opt}=1\)\(\mu\)s.

- **Fig. 14.** Simulated (a) ENC \(\text{opt}\) vs \(P_d\) and (b) corresponding \(W_{opt}\) vs \(P_d\) for n- and p-channels with different \(L\) in TSMC 0.25\(\mu\)m adopting the E-model. Results from the C-model at minimum \(L\) are also shown.
In Figs. 14(a) and 14(b) the ENC_{opt} and W_{opt} for C_{in}=1pF and τ_{p}=1µs, calculated using the E-model for n- and p-channels with different L as functions of P_{d} are shown. For comparison the two minimum L cases with the C-model are reported. The difference between E- and C-model in the estimate of the ENC_{opt} for the n-channels is relevant, while for the p-channel may appear small. On the other hand the difference in terms of optimization (W_{opt}) is in both cases relevant.

In Figs. 15(a) and 15(b) the ENC_{opt} and P_{d_{opt}} for C_{in}=1pF vs τ_{p} and for τ_{p}=1µs vs C_{in}, calculated using the E-model for n- and p-channels with different L, are shown, assuming a power budget limit of P_{d_{max}}=1mW. In the case of Fig. 15(a) the reduction in P_{d_{opt}} can be observed for higher values of τ_{p}. In the case of Fig. 15(b) the reduction in P_{d_{opt}} can be observed for smaller values of C_{in}. In both cases this is consequence of the dependence of the gate capacitance on the drain current density, which pushes the input MOSFET towards the weak inversion.

In Figs. 16(a) and 16(b) we report some experimental results from two front-end ASICs recently developed in TSMC 0.25µm CMOS. In Fig. 16(a) The ENC vs C_{in} from a prototype for a Time Projection Chamber [31] is shown. The ASIC implements 32 front-end channels with n-MOS inputs, 16 with L=0.24µm and 16 with L=0.36µm. The gate capacitance was, in both cases, on the order of 1.4pF. We observed a ≈40% difference in ENC slope, a result that appears in line with the discussion here presented. In Fig. 16(b) the ENC vs τ_{p} from a prototype for a Coplanar Grid Sensor [32] is shown. The ASIC implements an n-MOS input with L=0.36µm. The gate and input capacitances were on the order of 12pF and 60pF respectively. A low-frequency noise component proportional to τ_{p}^{-0.15}, in agreement with α_{f}≈0.85, can be observed.

**IX. CONCLUSIONS**

The optimization of the input MOSFET for charge amplifiers in deep submicron technology requires a proper modeling of the series noise and gate capacitance, and a review of the ENC equation. The enhanced modeling and associated ENC equation presented here allow a better ENC estimate and a more accurate optimization. The analysis shows that the traditional choice of selecting the minimum channel length and the maximum of the available power do not always offer the best resolution. Also, for a defined front-end, the low-frequency noise contribution to the ENC may de-
pend on the time constant of the filter. The results here re-
ported are based on the commercial TSMC 0.25µm CMOS,
but can be easily extended to other deep submicron tech-
nologies.

X. ACKNOWLEDGMENT

The authors are grateful to Giovanni Anelli (CERN), Veljko Radeka (BNL), and Angelo Dragone (Bari Poly-
technic) for helpful discussions.

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