

A 12MHz Switched-Capacitor Relaxation Oscillator with a Nearly Minimal FoM of -161dBc/Hz

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Abstract— In this work the phase noise performance of relaxation oscillators has been analyzed resulting in simple though precise phase noise expressions. These expressions have lead to a new relaxation oscillator topology, which exploits a noise filtering technique implemented with a switched-capacitor circuit to minimize phase noise. Measurements on a 65nm CMOS design show a sawtooth waveform, a frequency tuning range between 1 and 12MHz and a rather constant frequency tuning gain. At 12MHz oscillation frequency it consumes 90 μ W while the phase noise is -109dBc/Hz at 100KHz offset frequency. By minimizing and balancing noise contributions of charge and discharge mechanisms, a nearly minimal FoM of -161dBc/Hz has been achieved, which is a 6dB improvement over state-of-the-art.

Index Terms— figure of merit, phase noise, relaxation oscillators, thermodynamics

I. INTRODUCTION

CLOCK generation is an important area in integrated circuit design. Both LC oscillators and RC oscillators are frequently applied in this area. Ring oscillators and relaxation oscillators are both subsets of RC oscillators featuring large tuning ranges and small areas. Fig. 1 shows a typical relaxation oscillator with a capacitor and two switched current sources.

Such relaxation oscillators have two advantages with respect to ring oscillators: 1) they have a constant frequency tuning gain; and 2) their phase can be read out continuously due to their triangular (or sawtooth) waveform. A major disadvantage of practical relaxation oscillators is their poor phase noise performance compared to ring oscillators [1], [2], [4]. This phase noise performance is the main focus of this paper.

First Section II will show typical phase noise performance of ring oscillators and relaxation oscillators. Section III will then present a new relaxation oscillator topology which shows very good phase noise performance. Section IV discusses the transistor implementation and simulation results and Section V discusses the IC implementation and measurement results. Section VI finally draws conclusions.

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II. OSCILLATOR FIGURE OF MERIT

The $1/f^2$ phase noise performance of oscillators can be compared using the figure of merit definition [1].

$$\text{FoM} = 10 \log \left(\mathcal{L}(f_m) \left(\frac{f_m}{f_{osc}} \right)^2 \frac{P_{core}}{1 \text{mW}} \right) \text{ [dBc/Hz]} \quad (1)$$

$\mathcal{L}(f_m)$ is the well-known single-sideband phase noise measure. f_{osc} is the oscillation frequency and f_m is the offset frequency with respect to this oscillation frequency. P_{core} is the power consumed by the oscillator core.

Navid *et al.* have shown that at 290K thermodynamics limits the FoM of ring oscillators and relaxation oscillators to -165.3dBc/Hz and -169.1dBc/Hz, respectively [2]. Interestingly, they have also shown that the FoM of practical ring oscillators is generally better than about -160dBc/Hz, while the FoM of practical relaxation oscillators is about 10dB worse. So in theory relaxation oscillators can be better, but in practice they are not.

Part of the explanation is given in [2]; the noise added by the comparator, which is present in relaxation oscillators (cmp_{osc} in Fig. 1) but not in ring oscillators, increases the phase noise. We will now show that by filtering this noise by exploiting a switched-capacitor discharge mechanism, the FoM of a practical relaxation oscillator can be as good as the FoM of ring oscillators.

III. SWITCHED-CAPACITOR RELAXATION OSCILLATOR

A. Operation

Fig. 2 shows the new relaxation oscillator. As in Fig. 1, I_1 charges capacitor C_1 . However, C_1 is not grounded, but connected across an *OTA*, and the discharge process exploits a switched capacitor, C_2 , which is reversed periodically. The operation of the circuit is described in the next sentences.

The initial voltage across C_2 is $V_{ref,OTA}$. At t_0 , I_1 is charging C_1 at a constant rate via the *OTA*, resulting in a linearly decreasing voltage V . At t_1 , V crosses $V_{ref,osc}$ and cmp_{osc} reverses C_2 . C_2 is then being charged from $-V_{ref,OTA}$ to $+V_{ref,OTA}$ by I_1 and the *OTA*. At t_2 , C_2 is charged to $+V_{ref,OTA}$ and, as a

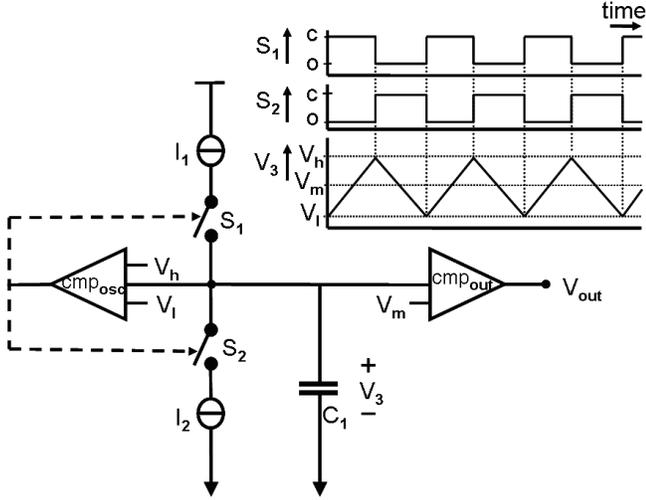


Fig. 1. Relaxation oscillator based on current sources (which is part of the general class of relaxation oscillators based on resistors, like in [2]).

result, a fixed charge packet equal to $2C_2V_{ref,OTA}$ has been subtracted from C_1 . $V_3 = V_+ - V_-$ is a sawtooth waveform. Subtracting this fixed charge packet filters out the noise of the oscillator comparator. The operation is illustrated in Fig. 3, which shows the control signal X , V_3 and the output of the comparator cmp_{out} , V_{out} , is also shown, which produces an edge whenever voltage V_3 reverses polarity. Suppose now that cmp_{osc} is noisy and C_2 is reversed at t_4 instead of at t_3 . Although the duty cycle of V_{out} is changed (at t_5), the active edge of V_{out} at t_6 is unaffected and so is the phase noise.

This filter technique is similar to the anti-jitter circuit (AJC) technique used in open-loop jitter filters [3]; note that we apply a switched-capacitor circuit to subtract the charge packet, which is very power-efficient.

B. Phase Noise Performance

Filtering out the noise of the oscillator comparator has two consequences: 1) the power dissipated by the oscillator comparator and its reference can be reduced without deteriorating the phase noise; and 2) the two remaining

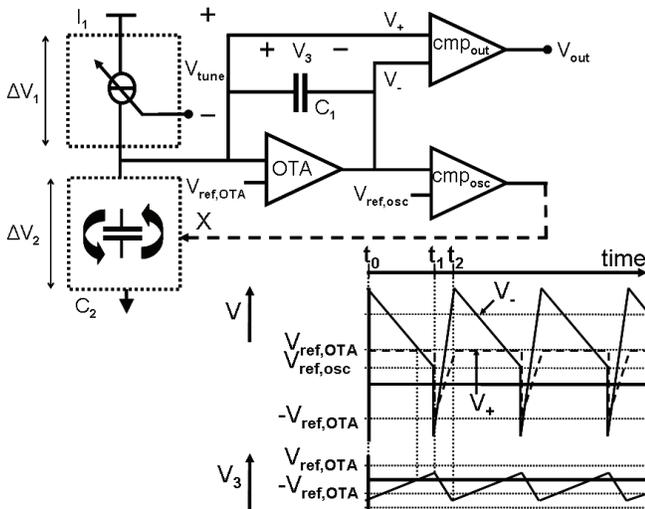


Fig. 2. Block-level schematic of a switched-capacitor relaxation oscillator.

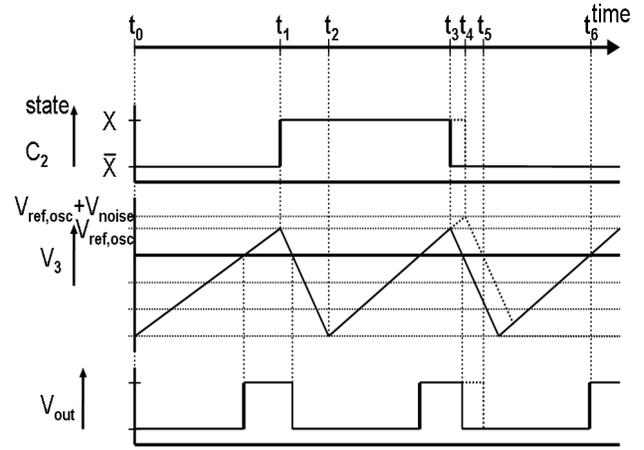


Fig. 3. Technique to filter out the noise of the oscillator comparator (cmp_{osc}).

contributions to the $1/f^2$ phase noise are the white noise of the charging and discharging mechanisms. It can be shown that the resulting FoM of such a relaxation oscillator is given by (2).

$$\text{FoM} = 10 \log \left(\frac{2kT \cdot P_{core}}{I_1 \cdot 1\text{mW}} \cdot \frac{\Delta V_1 + \Delta V_2}{\Delta V_1 \cdot \Delta V_2} \right) [\text{dBc/Hz}] \quad (2)$$

k is the Boltzmann constant, T is the absolute temperature and P_{core} is the power consumed by the oscillator core: $P_{core} = V_{DD}I_{core}$. I_1 is the charge current and ΔV_1 and ΔV_2 (also shown in Fig. 2) are the voltage headroom reserved for the charging and the discharging mechanisms, respectively. For a good FoM we want ΔV_1 and ΔV_2 to be high and about equal, while we also want a large V_3 swing to reduce the phase noise floor contribution of cmp_{out} . In Fig. 1 this is not possible, since the sum $\Delta V_1 + \Delta V_2 + \Delta V_3$ has to fit in the supply V_{DD} . In Fig. 2 the voltage swing of V_3 mainly occurs at the output of the OTA, leaving the full V_{DD} for $\Delta V_1 + \Delta V_2$.

Instead of reversing C_2 , C_2 could be discharged to ground before connecting it to V_+ , which would be easier to implement. Reversing C_2 has some advantages though: 1) ΔV_2 can be doubled without increasing power dissipation; and 2) the time allowed for settling is doubled (C_2 needs to settle only once instead of twice every period). By reversing C_2 , both a near optimal and a practical choice would be $\Delta V_1 = \Delta V_2 = 2V_{DD}/3$. In the case of a sawtooth waveform, the total core current, I_{core} , is at least $2I_1$ in steady state ($= I_1 + \bar{I}_2$); the discharge current has to be equal to the charge current. This implies a theoretical FoM of -163.2dBc/Hz at 290K , which is similar to that of ring oscillators.

C. Waveform

The waveform of the switched-capacitor relaxation oscillator can be a sawtooth waveform, a triangle waveform or anything in between. A sawtooth waveform arises when

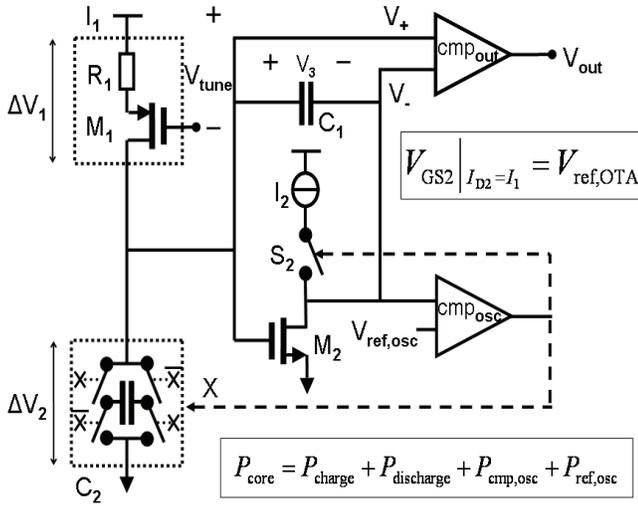


Fig. 4. Actual implementation of a switched-capacitor relaxation oscillator.

current source I_2 sources the full discharge packet instantly, i.e. when $I_2 = C_2 \Delta V_2 / t_{act} = I_1 t_{osc} / t_{act}$ and $t_{act} \rightarrow 0$. A triangle waveform arises when $I_2 = I_1$ and $t_{act} = t_{osc} / 2$, i.e. both are a function of frequency. Furthermore, in case of a triangle waveform C_1 is halved in size with respect to C_1 in case of a sawtooth waveform.

IV. TRANSISTOR IMPLEMENTATION AND SIMULATIONS

Figure 4 shows the actual implementation of the switched-capacitor relaxation oscillator. I_1 is implemented by a resistively degenerated PMOST to increase control linearity and decrease thermal noise. M_2 could be biased continuously by I_2 , but this would worsen the FoM dramatically. Only when C_2 is reversed, S_2 is closed briefly (during $t_{osc} \cdot I_1 / I_2$) and I_2 discharges C_1 through C_2 during this time. When S_2 is opened, C_2 starts settling to $V_{ref,OTA}$. Note that the accuracy of the charge packet with which C_1 is discharged is only a function of the settling of C_2 ; noise on I_2 does not affect the accuracy. $V_{ref,OTA}$ is implemented as the gate-source voltage of M_2 biased

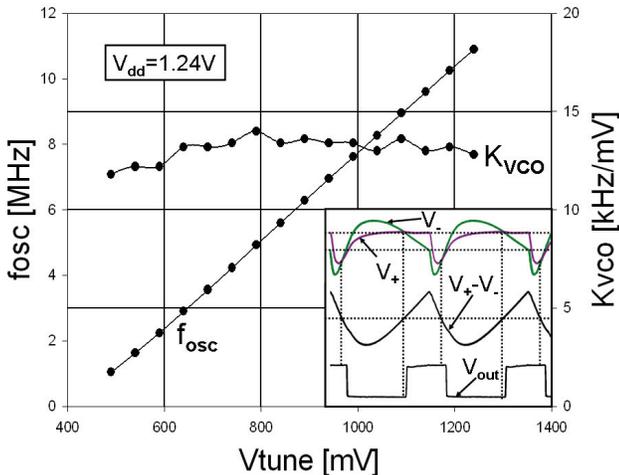


Fig. 5. Measured waveforms, frequency tuning range and frequency tuning gain (Agilent DSO6104A oscilloscope and LeCroy AP033 active differential probe, R&S FSP spectrum analyzer).

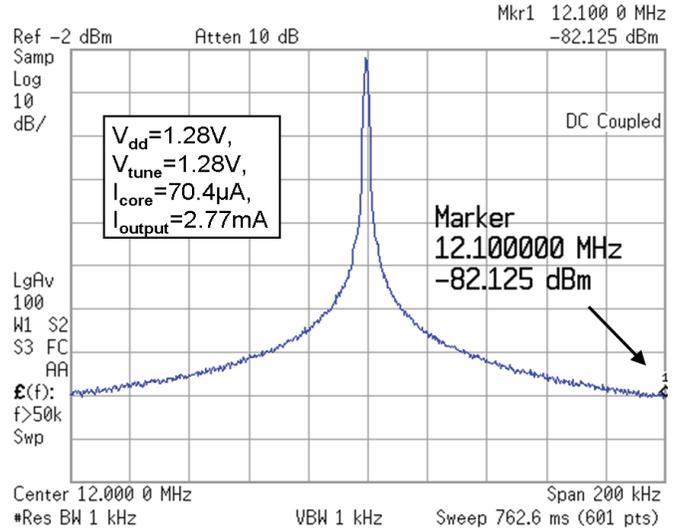


Fig. 6. Measured phase noise (Agilent E4440A spectrum analyzer, Keithley 2000 multimeter).

at current I_1 .

The relaxation oscillator of Fig. 4 has been designed in a standard 65nm CMOS process ($V_{DD} = 1.2V$). The main design choices are: $V_{ref,OTA} = V_{DD}/3$, $V_{ref,osc} = V_{DD}/6$, $\Delta V_1 = \Delta V_2 = \Delta V_3 = 2V_{DD}/3$, $I_1 = I_2 = I_2/4 = 25\mu A$, $C_1 = C_2 = 2.5pF$. The measurement buffers, oscillator comparator and its reference are designed to consume about 2.5mA, 10 μA and 5 μA , respectively. The circuitry to switch I_2 and reverse C_2 reliably consumes about 5 μA . As a result, $f_{osc} = 12.5MHz$, $I_{core} = 2.8I_1 = 70\mu A$.

According to (2), the FoM is expected to be -161.7dBc/Hz at 290K, which is similar to the -161.4dBc/Hz predicted by simulation. Simulation also predicts a lower oscillation frequency and lower peak voltages than calculated, mainly due to the gate-source and gate-drain capacitances of M_2 .

V. IC IMPLEMENTATION AND MEASUREMENTS

The active area of the 65nm CMOS design measures 200x150 μm^2 [5]. Fig. 5 and 6 show the measurement results. The circuit is measured using a battery supply. It is fully functional and the performance is similar for supply voltages between 1.0 and 1.3V. Unfortunately, S_2 is closed for somewhat less than $t_{osc} \cdot I_1 / I_2$, so the waveform is slightly deformed; this can be easily corrected in a re-design. The measured FoM is

-161dBc/Hz, which is similar to both analysis and simulation results. Ten samples have been measured and all have similar FoMs.

VI. CONCLUSION

Measurements illustrate the advantages of a relaxation oscillator: the phase can be read out continuously and the tuning range is both large and linear. Measurements also show an outstanding phase noise performance; the FoM is at least 6dB better than state-of-the-art relaxation oscillators [1], [2],

[4] and similar to state-of-the-art ring oscillators [2]. Note that we include all the core current consumption. As it seems to be increasingly more difficult to reach the minimal FoM for ring oscillators in smaller CMOS technologies [2], relaxation oscillators could well become the preferred choice of RC oscillators in low-power applications, like sensor networks.

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