A Wide-Tunable Translinear Second-Order Oscillator

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***Abstract*— This paper describes the design and measurement of a translinear second-order oscillator. The circuit is a direct implementation of a nonlinear second-order differential equation and follows from a recently developed synthesis method for dynamic translinear circuits. It comprises only two capacitors and a handful of bipolar transistors and can be instantaneously controlled over a very wide frequency range by only one control current, which indicates its suitability for spread-spectrum communications. Its total harmonic distortion can be made small by the design, which enables fully integrated transmitters. A semicustom test chip, fabricated in a standard 2-μm, 7-GHz, bipolar IC process, operates from a single supply voltage, which can be as low as 2 V and oscillates over six decades of frequency with −31 dB total harmonic distortion.**

***Index Terms*— Dynamic translinear, low power, oscillators, translinear.**

I. Introduction

RECENTLY, both an analysis method and a synthesis method for dynamic translinear circuits were proposed by the authors [1], [2]. The dynamic translinear principle can be regarded as a generalization of the well-known "static" translinear principle, formulated by Gilbert in 1975 [3].

An important subclass of dynamic translinear circuits is the class of "translinear filters," also called "log-domain" or "exponential state-space" filters, which were originally introduced by Adams in 1979 [4]. Although not recognized then, this was actually the first time a first-order linear differential equation was implemented using translinear circuit techniques. In 1990, Seevinck introduced a "companding current-mode integrator" [5], and since then the principle of translinear filtering has been extensively studied by Frey, see, e.g., [6], Punzenberger and Enz [7], Toumazou and Lande [8], Perry and Roberts [9], and Mulder and Serdijn, see, e.g., [10], [11].

However, the dynamic translinear principle is not limited to filters, i.e., linear differential equations. By using the dynamic translinear principle, it is possible to implement every linear or nonlinear differential equation, using transistors and capacitors only; see, e.g., [12]. Hence, a high functional density can be obtained, and the absence of large resistors makes them especially interesting for ultra-low-power applications [13].

Apart from this, dynamic translinear circuits also exhibit other interesting properties.

1) Owing to the exponential behavior of a bipolar transistor or an MOS transistor in its subthreshold region, the voltages in dynamic translinear circuits are logarithmically related to the currents. As a result, the voltage excursions are small, typically only a few tens of millivolts. This is beneficial in a low-voltage environment.
2) Due to these small voltage swings, the effects of parasitic capacitances are reduced. This facilitates relatively wide bandwidth operation [14], [15].
3) Dynamic translinear circuits are easily controlled over a wide range of several parameters, such as gain, frequency, or threshold. This increases their designability and makes them attractive to be implemented as standard cells or programmable building blocks.
4) Dynamic translinear circuits are easily implemented in class AB, which enables the signal currents to be much larger than the quiescent currents. This, in turn, entails a larger dynamic range and a reduced average current consumption [16].
5) In dynamic translinear circuits, transistors are used either as elements of the translinear loops or as nullors, to provide additional loop gain. Hence, in an IC process, only three types of components are required:
   - transistors that are well matched and have an accurate exponential transfer over a wide range of transistor current;
   - transistors with a large gain, also at higher frequencies;
   - capacitors.

Second-order oscillators are important building blocks in electronic systems to generate a periodic signal from dc power. They implement a second-order differential equation, which ideally equals

$$\ddot{x}(t) + \omega^2 x(t) = 0 \tag{1}$$

$x(t)$ and ω being the oscillator signal and the angular frequency, respectively. The dot represents differentiation with respect to time.

In order to compensate for the effects of nonidealities in the oscillator circuitry which cause the oscillator amplitude to be unstable, such as noise and drift, all practical second-order oscillators somehow implement the (nonlinear) second-order differential equation

$$\ddot{x}(t) + f(x)\dot{x}(t) + \omega^2 x(t) = 0 \tag{2}$$

where $f(x)$ is an arbitrary (nonlinear) even-symmetry function of x. When $f(x) > 0$, the oscillator is damped and the amplitude decreases. When $f(x) < 0$, the oscillator is undamped and the amplitude increases.

One way to implement (2) is to use a (passive) resonator and a nonlinear time-invariant circuit or component to undamp

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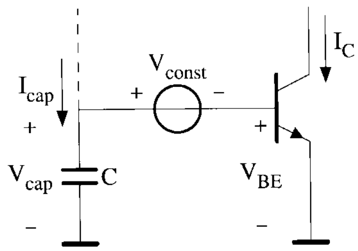


Fig. 1. Principle of dynamic translinear circuits.

the resonator. Usually, this type of oscillator is used to achieve a low phase noise. Another approach is the use of two active integrators in a two-integrator oscillator. This paves the way to a good (frequency) tunability.

The first translinear oscillator was proposed not earlier than 1995 by Pookaiyaudom and Mahattanakul [17]. The circuit, basically, comprises a cascade of an inverter (a current mirror) and two first-order all-pass filters that each have a transfer function

$$y(t) + \omega \dot{y}(t) = x(t) - \omega \dot{x}(t) \tag{3}$$

$x(t)$ and $y(t)$ being the filter input and output signal, respectively. The undamping circuit, an automatic gain control (AGC), was not discussed in the paper.

Although not verified experimentally in [17], simulations predicted a wide tuning range and the ability to operate from low supply voltages. These attractive properties are characteristic of translinear circuits [1].

Here, we present the design and experimental results of a wide-tunable translinear second-order oscillator. The circuit, which comprises only two capacitors and a handful of transistors, is a direct implementation of a nonlinear second-order differential equation by means of the synthesis method proposed in [2] and is tuned by only one control current. To the authors' knowledge, this is the first time a translinear oscillator has actually been implemented.

The organization of the paper is as follows. Section II introduces the dynamic translinear principle, which subsequently is elaborated into the design of a second-order oscillator in Section III, following a systematic approach. The experimental results of a semicustom implementation of the translinear oscillator are discussed in Section IV. Finally, Section V deals with the conclusions.

II. Dynamic Translinear Principle

The key to the dynamic translinear principle, from a current-mode point of view, are the capacitance currents. We, therefore, concentrate on the simple substructure, depicted in Fig. 1. Assuming a bipolar transistor, it follows [1]:

$$CV_T \dot{I}_C = I_C I_{\mathrm{cap}} \tag{4}$$

C, V_T, I_C, and I_{cap} being the capacitance value, the thermal voltage kT/q, the collector current, and the capacitance current, respectively. The dot again represents differentiation with respect to time.

From this equation, it can be seen that a time derivative in a differential equation can be replaced by a product of two currents. This product of currents can be elegantly realized by means of the translinear principle [3].

III. Oscillator Design

The design of a second-order dynamic translinear oscillator starts with a dimensionless differential equation that describes the oscillator behavior in the time domain

$$\frac{d^2x}{d\tau^2} + wf(x)\frac{dx}{d\tau} + w^2x = 0. \tag{5}$$

This equation describes the oscillator signal x as a function of a control signal w. $f(x)$ is again an arbitrary (nonlinear) even-symmetry function of x. τ is the dimensionless time of the oscillator.

A. Transformations

The first synthesis step is the transformation of the above differential equation such that the dimensions of the resulting equation are suitable for a translinear realization. In dynamic translinear circuits, all signals are currents. Hence, the signals w and x can be transformed into the currents I_F and I_{osc} through the equations

$$w = \frac{I_F}{I_O} \tag{6}$$

and

$$x = \frac{I_{\mathrm{osc}}}{I_O} \tag{7}$$

I_O being a dc bias current that determines the absolute current swings. It must be noted that the normalization of both w and x to I_O causes all currents to fall in the same order of magnitude. As a result, the oscillator output current I_{osc} becomes proportional to the control current I_F and hence, if I_F varies over a relatively wide range, so does I_{osc}. If this effect is undesirable, one can either normalize w and x to two different currents I_{O1} and I_{O2}, or let the oscillator follow by a variable gain amplifier with a gain that is proportional to I_O/I_F.

The dimensionless time τ can be transformed into the time t with dimension $[s]$ through [2]

$$\frac{d}{d\tau} = \frac{CV_T}{I_O}\frac{d}{dt}. \tag{8}$$

From (8), two important characteristics of dynamic translinear oscillators can be derived. First, time (t) is inversely proportional to current I_O. This means that the oscillator will be linearly frequency tunable by means of only one control current. Second, this control current must be proportional to the absolute temperature to eliminate the influence of the temperature on the oscillator.

Applying the above transformations, the resulting differential equation becomes

$$C^2V_T^2\ddot{I}_{\mathrm{osc}} + CV_TI_Ff(I_{\mathrm{osc}}, I_F)\dot{I}_{\mathrm{osc}} + I_F^2I_{\mathrm{osc}} = 0 \tag{9}$$

or, alternatively written, using $f(I_{\mathrm{osc}}, I_F)\dot{I}_{\mathrm{osc}} = \dot{F}(I_{\mathrm{osc}}, I_F)$

$$C^2V_T^2\ddot{I}_{\mathrm{osc}} + CV_TI_F\dot{F}(I_{\mathrm{osc}}, I_F) + I_F^2I_{\mathrm{osc}} = 0. \tag{10}$$

Since the currents in this differential equation will equal the currents in the final oscillator circuit, at this point it is already possible to determine the most important oscillator characteristics, which are its oscillation frequency ω_{osc} and its amplitude $\hat{I}_{\text{osc}}$. If we assume that the oscillator current is sinusoidal, thus $I_{\text{osc}} = \hat{I}_{\text{osc}} \sin(\omega_{\text{osc}} t + \theta_{\text{osc}})$, ω_{osc}, and $\hat{I}_{\text{osc}}$ follow from

$$\omega_{\text{osc}} = \frac{I_F}{V_T C} \tag{11}$$

and

$$\int_0^T f(I_{\text{osc}}, I_F)\dot{I}_{\text{osc}} dt = 0 \tag{12}$$

T being $2\pi/\omega_{\text{osc}}$.

B. Definition of the Capacitance Currents

The next synthesis step is the elimination of the derivatives. In the previous section, we saw that a derivative can be replaced by a product of a capacitance current and a collector current according to (4). The capacitor currents can be introduced one by one. Each capacitance current reduces the order of a differential equation by one, until finally a current-mode multivariant polynomial results. Since the differential equation to be implemented is second-order in I_{osc}, the definition of the first capacitance current should include I_{osc}, while the definition of the second capacitance current, apart from I_{osc}, should also include its derivative $\dot{I}_{\text{osc}}$. The function $F(I_{\text{osc}}, I_F)$ should be present in either the first or the second capacitance current.

Defining I_{cap1} and I_{cap2} as

$$I_{\text{cap1}} = CV_T \frac{\dot{I}_{\text{osc}}}{I_{\text{osc}} + I_F} \tag{13}$$

$$I_{\text{cap2}} = CV_T \frac{\dot{I}_Q}{I_Q + I_F} \tag{14}$$

$$I_Q = I_{\text{osc}} - F(I_{\text{osc}}, I_F) - \frac{CV_T \dot{I}_{\text{osc}}}{I_F}. \tag{15}$$

The above differential equation transforms into

$$F(I_{\text{osc}}, I_F)I_{\text{cap2}}I_F + I_{\text{cap1}}(I_{\text{cap2}} + I_F)(I_F + I_{\text{osc}}) - I_F(I_{\text{cap2}}(I_F + I_{\text{osc}}) - I_F I_{\text{osc}}) = 0. \tag{16}$$

C. Translinear Decomposition

The above polynomial is the basis of the next synthesis step, which is translinear decomposition. That is, the polynomial has to be mapped onto a set of translinear loop equations that are each characterized by the general equation

$$\prod_{\text{CW}} J_{C,i} = \prod_{\text{CCW}} J_{C,i} \tag{17}$$

$J_{C,i}$ being the transistor collector current densities in clockwise (CW) or counterclockwise (CCW) direction. To this end, the synthesis methods for static translinear circuits expounded in [18] can be used. One possible solution is achieved by "parametric" decomposition of (16). Two intermediate currents, I_P and I_Q, the latter already suggested by the above definition of I_{cap2}, are introduced, resulting in

$$(I_F + I_P - I_Q)I_F = (I_F + I_{\text{cap1}})(I_{\text{osc}} + I_F) \tag{18}$$

$$(I_F + I_P)I_F = (I_F + I_{\text{cap2}})(I_Q + I_F) \tag{19}$$

$$I_P = 2I_{\text{osc}} - F(I_{\text{osc}}, I_F). \tag{20}$$

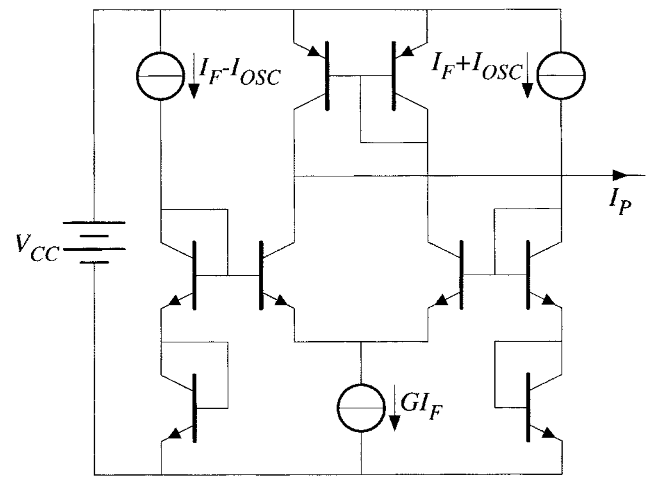


Fig. 2. Implementation of the translinear function F: $I_P = 2GI_{\text{osc}}I_F^2/(I_{\text{osc}}^2 + I_F^2)$.

From its definition, it follows that $F(I_{\text{osc}}, I_F)$ must be a nonlinear time-invariant odd-symmetry function of I_{osc} and I_F, whose derivative $f(I_{\text{osc}}, I_F)$ with respect to I_{osc} is negative for small values of I_{osc} and positive for large values of I_{osc}. A suitable choice is

$$F(I_{\text{osc}}, I_F) = 2I_{\text{osc}} - \frac{2GI_{\text{osc}}I_F^2}{I_{\text{osc}}^2 + I_F^2} \tag{21}$$

G being a constant, which must be larger than one. Hence, for intermediate current I_P, it follows

$$I_P = \frac{2GI_{\text{osc}}I_F^2}{I_{\text{osc}}^2 + I_F^2}. \tag{22}$$

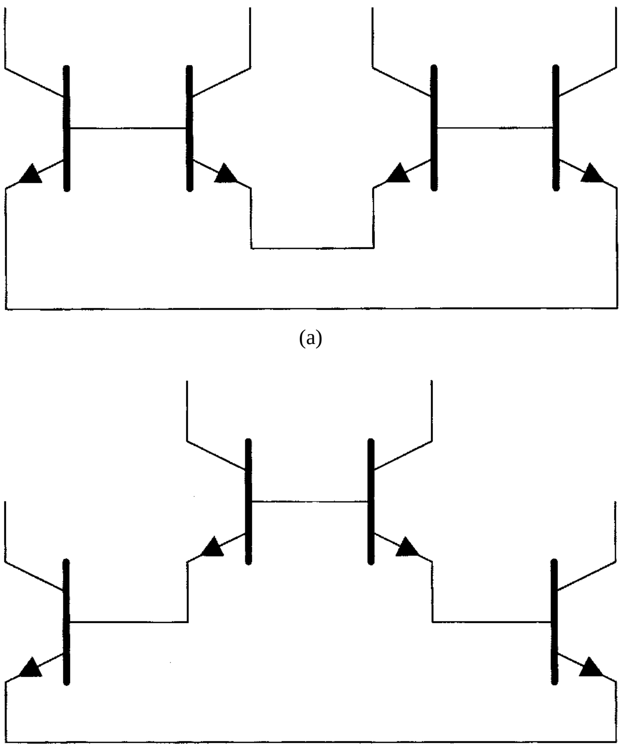
This function is easily implemented in a translinear circuit using the generic principle described in [15]; see Fig. 2. Using this translinear undamping circuit and assuming the oscillator output current to be sinusoidal, the oscillator amplitude $\hat{I}_{\text{osc}}$ follows from (12), which yields

$$\hat{I}_{\text{osc}} = I_F\sqrt{G-1}. \tag{23}$$

Note that $\hat{I}_{\text{osc}}$ is indeed proportional to I_F as was discussed previously.

D. Biasing

The final synthesis step is biasing. In other words, the translinear decomposition that was found during the previous synthesis step has to be mapped onto a correct translinear circuit topology and the correct currents must be supplied to



(a)

(b)

Fig. 3. Two basic translinear topologies.

this topology. During this process, attention must be paid to the following [2].

- *The choice between a stacked and an up-down topology* [19], as depicted in Fig. 3. In a stacked topology, there are both current summation and subtraction nodes, whereas in an up-down topology, there are only summation points. This choice is also important in view of possible low-voltage operation.
- *The use of current mirrors*, used as "inverters," to generate subtraction points.
- *The use of complementary transistors*.
- *The use of compound transistors* [9], which, owing to their fourth terminal, provide an additional degree of freedom and thus facilitate biasing.
- *The use of the MOS bulk terminal as a second gate* [20].
- *The use of MOS transistors in the triode region* [21]. The MOST is symmetrical with respect to source and drain, and hence, when operating the MOST in its triode region, the drain current is also exponentially or quadratically related to the drain voltage. Consequently, the gate-drain voltage can be employed as part of a translinear loop.
- *The use of nullors to force the correct transistor currents* [22]. Again, additional terminals add some degrees of freedom. Nullors also serve as buffers for base currents and often, one-transistor implementations already significantly reduce the influence of base currents.

A possible biasing arrangement for the translinear oscillator, assuming ideal bipolar transistors, current mirrors, voltage, and current sources, is depicted in Fig. 4. Transistors $QN1$–$QN4$ and $QN11$–$QN14$ implement (19) and (18), respectively. Note that current source $G \cdot I_F$ is current controlled. $QN31$ delivers the oscillator output current I_{osc}.

Replacing all the ideal sources by practical ones and compensating for the Early effect by adding a number of common-base stages yields the circuit diagram depicted in Fig. 5.

A PNP current mirror with multiple outputs (transistors $QP41$–$QP48$) produces replicas of I_F. $QP40$ enlarges the loop gain, thereby reducing the influence of the base currents. The constant factor G is set by the ratio of the emitter areas of $QN48$ and $QN46$.

IV. Experimental Results

The circuit shown in Fig. 5 was simulated using SPICE and realistic (IC) capacitor and (minimum-size) transistor models. The results indicate the correct operation of the translinear oscillator for various temperatures and values of I_F, G (>1) and cap_1 ($= cap_2$), yielding oscillations from 70 mHz ($cap_1 = cap_2 = 1$ nF, $I_F = 10$ pA) up to 20 MHz ($cap_1 = cap_2 = 1$ pF, $I_F = 1$ mA), in accordance with (11) and (23). The supply voltage equaled 3 V. For $G = 1.1$, the total harmonic distortion was below 2%.

To verify the circuit operation in practice, the active circuitry of the oscillator was integrated onto a semicustom version of our in-house 2-μm bipolar IC process. Typical transistor parameters are: $h_{\text{fe, NPN}} \approx 100$, $f_{T,\text{NPN}} \approx 7$ GHz, $h_{\text{fe, LPNP}} \approx 80$, and $f_{T,\text{LPNP}} \approx 40$ MHz. G equals 5/4 by design.

Experiments proved the correct operation of the translinear oscillator for supply voltages from 5 down to 2 V. The current consumption approximately equals 18 times I_F. Fig. 6 depicts the oscillation frequency f_{osc} (in Hz) as a function of control current I_F for four different capacitor values, all easily integratable: 560, 56, 15 pF, and the "intrinsic" capacitance, stemming from a bond flap, a bond wire, and a pin. From this plot, it can be deduced that this particular translinear oscillator can be controlled over a very wide frequency range of six (!) decades. Deviations from the theoretically linear relationship between I_F and f_{osc} are caused by offset currents, introduced by the measurement set-up, at the low end of the frequency range and by high-level injection and the relatively low transit frequencies of the lateral PNP's at the high end of the frequency range. This was confirmed by simulations. Nevertheless, oscillations higher than half the $f_{T,\text{LPNP}}$ were measured.

The favorable property of a very wide frequency range makes the translinear oscillator an interesting candidate for frequency synthesizers, such as those needed in, e.g., spread-spectrum receivers and transmitters.

Since F is a time-invariant function of I_F and I_{osc}, the output current waveform is independent of the oscillation frequency. This has been verified by means of a dynamic signal analyzer and proved to be true for the complete "linear" current range, i.e., between 2 nA and 200 μA. Fig. 7 depicts the output frequency spectrum of the oscillator running at 1.7 MHz. The supply voltage and the current consumption equal 3 V and 2.8 mA, respectively. Both external capacitors equal 560 pF. The total harmonic distortion is mainly determined by the

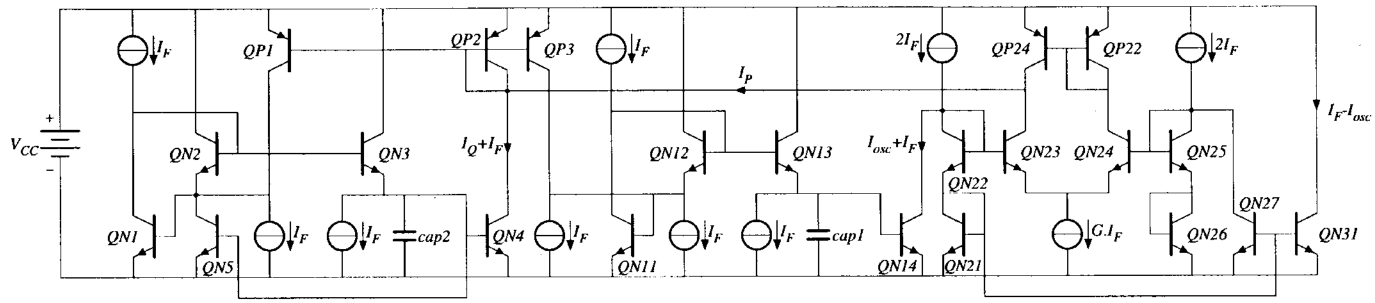


Fig. 4. Possible biasing arrangement for the translinear oscillator. Ideal transistors, current mirrors, voltage, and current sources are assumed.

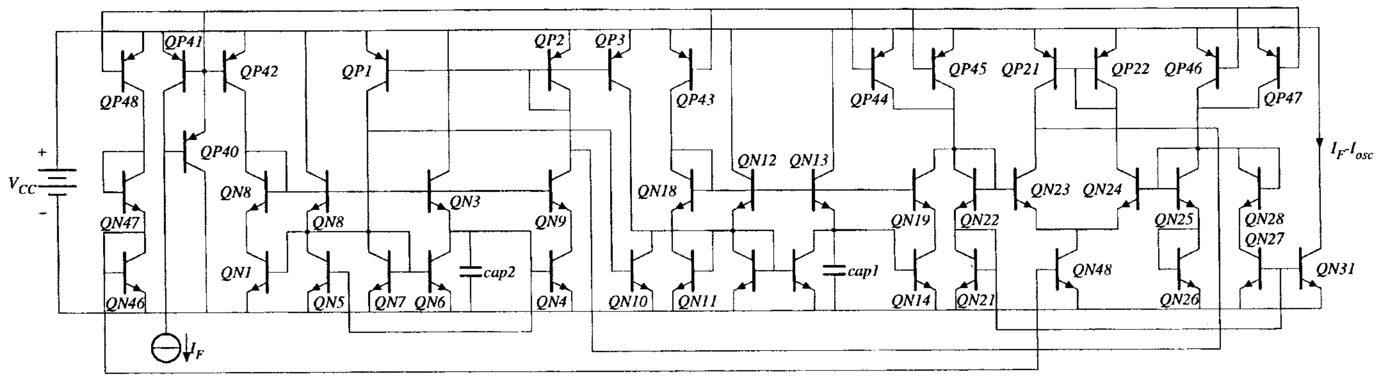


Fig. 5. Complete circuit diagram of the translinear oscillator.

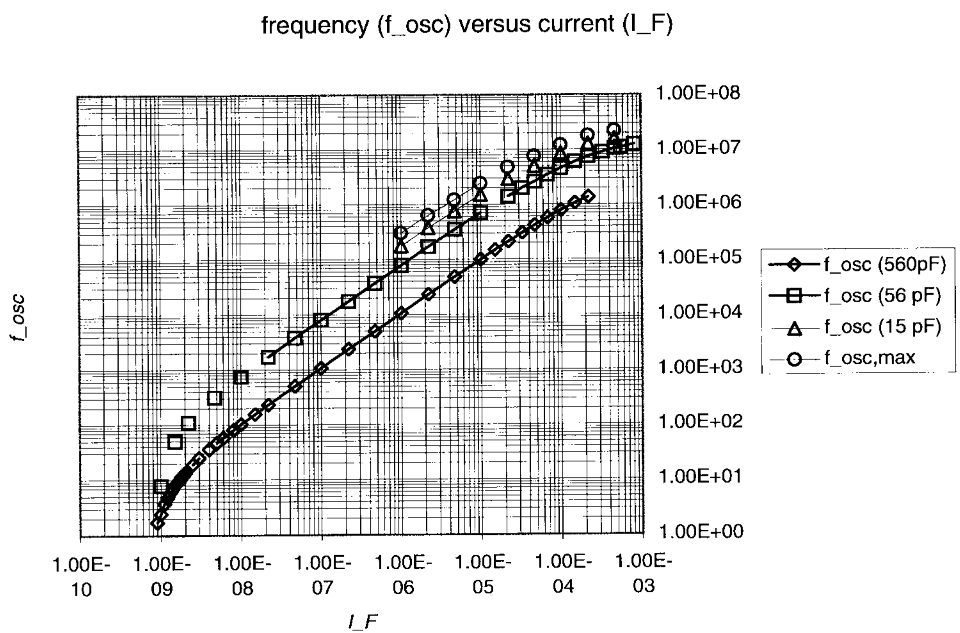


Fig. 6. Measured oscillation frequency as a function of control current I_F for four different capacitor values.

second and the third harmonic and equals 2.7% or −31 dB. As indicated by the above simulations, a smaller G will lower the distortion even further.

The small harmonic distortion feasible with translinear oscillators makes them especially interesting for the development of fully integrated transmitters, since bulky and expensive off-chip filtering may be no longer necessary.

Fig. 8 depicts the same output frequency spectrum, but now zoomed in at the carrier frequency. At 20 kHz offset, the phase noise equals −99 dBc/Hz, which is reasonable for an oscillator with such a wide tuning range. Since the dominant noise sources in translinear oscillators, as in all (dynamic) translinear circuits, i.e., the collector shot noise sources, are proportional to the collector currents, a better noise performance may be obtained at the expense of a larger control current I_F and, for the same oscillation frequency, two larger capacitors. As a rule of thumb, every 3 dB improvement of the carrier-to-noise ratio costs twice as much current and twice as much capacitance. It must be noted, however, that for this design no attempt has been made to achieve a low phase noise. Further,

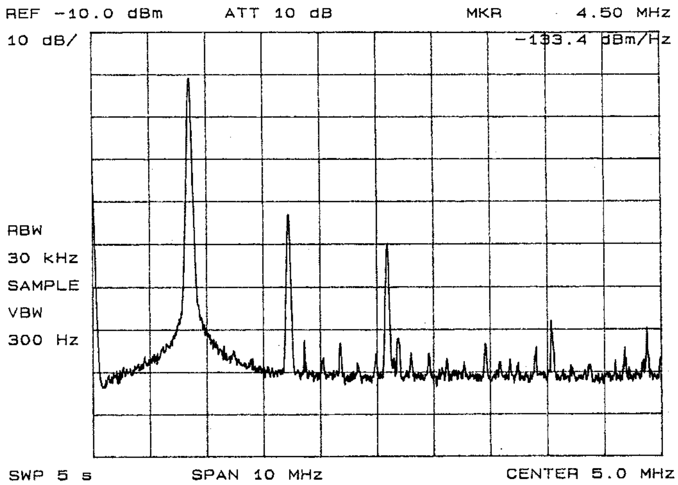


Fig. 7. Measured output spectrum for $f_{\rm osc} = 1.7$ MHz.

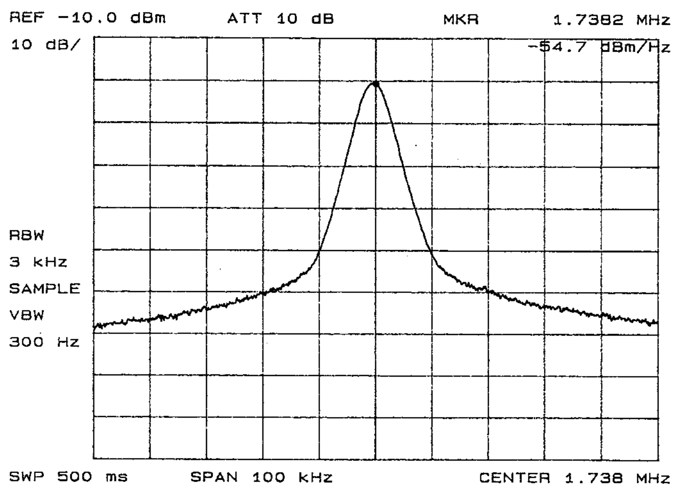


Fig. 8. Measured output spectrum around the carrier frequency $f_{\rm osc} = 1.7$ MHz.

since the noise originates from collector currents, the noise is modulated by the signal. Noise calculations in static and dynamic translinear circuits are far from trivial [16]. It is reasonable, though, to expect that a better noise performance is obtained by using class-AB circuitry, compound transistors, a different translinear decomposition, or a larger G, the latter, of course, at the expense of a larger harmonic distortion. Alternatively, the translinear oscillator may be embedded in a phase-locked loop whose closed-loop bandwidth is tailored to suppress oscillator noise.

V. Conclusions

A translinear second-order oscillator has been introduced. The circuit is a direct implementation of a nonlinear second-order differential equation. It comprises only two capacitors and a handful of transistors and can be instantaneously controlled over a very wide frequency range by only one control current (I_F), which indicates that the translinear oscillator is an interesting candidate for spread-spectrum frequency synthesizers. Its harmonic distortion is directly related to another parameter (G) and can be made small by design, thereby paving the way to fully integrated transmitters.

A semicustom version of the proposed circuit operates from a single supply voltage down to 2 V, oscillates over six decades of frequency with −31 dB total harmonic distortion. Oscillation frequencies up to half the $f_{T,\,\rm LPNP}$ were measured. At 1.7 MHz, using two 560 pF capacitors, the current consumption equals 2.8 mA, and the phase noise equals −99 dBc/Hz at a 20-kHz offset from the oscillation frequency.

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