UWB Pulse Oscillator at 24 GHz with 2.1 GHz Bandwidth for Industrial Radar Sensor Applications

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***Abstract*— A fast pulse oscillator in thin film technique at 24 GHz for high resolution low power industrial radar sensor applications is presented. Both linear and nonlinear simulation techniques as well as a simple mathematical model are employed for design and optimization. The 3 dB-bandwidth of the pulse envelope is more than 2.1 GHz, the time duration is less than 480 ps. The average power of the pulses is compliant with FCC regulations for a pulse repetition frequency between 1 MHz and 4 MHz. The oscillator is mounted in a low-cost commercial package for SMT without significant performance degradation. Application in a radar sensor results in a range resolution of 10 cm.**

***Index Terms*— UWB, pulse oscillator, pulse radar.**

I. Introduction

THE large bandwidth of several gigahertz that ultra-wideband (UWB) radar devices are allowed to allocate, enables them to the high resolution that industrial sensors require. Not only to distinguish between two targets close to each other, but also to reduce the blind close-up range from which most radar systems suffer due to the unavoidable impedance mismatching of the antenna. Large bandwidth can be achieved in several ways, but only pulsed systems provide a low transmitted average power density, low complexity and compliance with the demanded regulation rules at the same time. But creating very short pulses is a challenge. One possibility is to convert short baseband pulses up to the desired radio frequency (RF) using a continuous-wave (CW) local oscillator (LO) [2]. Although generation of baseband pulses is easier than of RF-pulses, the approach is complex due to extra mixing stages. The power consumption is increased by the continuously running LO. Another common approach is to switch a CW-oscillator using RF-switches [3]. These switches need to be fast with a high switching contrast. Again the CW-oscillator increases power consumption. A third approach is the usage of pulse oscillators (PO) [4],[5]. Due to the low duty cycle the average current consumption of the oscillator can be three orders of magnitude smaller than in a switched CW-system. The system is less complex since it does not require RF-switches or frequency converters for signal generation. A challenge is the oscillator that powers up at fundamental frequency in a short time. In this paper we present a transmission-type pulse oscillator that shows a short rise time and therefore a short total width of the pulses which helps to improve resolution and decrease the blind close-up range of a UWB pulse radar system.

II. Technology Concept

The main goal of the design of pulse oscillators is to minimize the rise time of the oscillation at a given peak amplitude. As a first approximation the time T_{slope} it takes the oscillation to start is

$$T_{\text{slope}} \approx Q \cdot T \tag{1}$$

where Q is the quality factor (also: figure of merrit) of the tank circuit including the transistor and T the periodic time of the resonant frequency. Since T is predefined by the desired carrier frequency, only Q can be minimized. The most general definition for Q is

$$Q = 2\pi \cdot \frac{\text{Average energy stored in circuit}}{\text{Energy dissipated during one period } T} \tag{2}$$

Increasing the losses would help to lower the quality of the circuit, but also lowers the peak power of the resulting pulse. A better choice is to decrease the energy storage capability. This is done by avoiding capacitive and inductive components, circuitry and parasitics both in the passive feedback circuit and in the transistor's package. The passive circuit should be as short as possible since the quality factor of an oscillator grows with the electric length of the loop. One precondition for oscillation is an electric length of the loop that equals $n \cdot \lambda$, where $n \in \mathbb{N}$ and λ is the wavelength, so the optimum is achieved if $n = 1$. The transistor's transmission phase should be as small as possible. Otherwise it is difficult for the passive feedback network to close the loop geometrically and keep the total electric length short at the same time. The chosen RF-transistor is available in a ceramic package (NEC32584C) and as a bare die (NEC32500). The absolute phase lengths of the packaged chip and of the bare chip are 330° and 80° , respectively. The feedback loop must contain a DC-block in order to separate drain and gate bias. This is done with a quarter-wave coupler, which adds another electric length. For this reason the loop of the oscillator using the bare chip can be shorter by one wavelength compared to the packaged chip. But even with the bare die the design of the broadband feedback network is a challenge. For this reason we chose a coplanar design in thin film technique which allows more degrees of freedom with regard to line impedances and tolerances. It also allows standard flip-chip mounting technique which is reliable and causes few parasitics. The permittivity ϵ_r of the substrate has to be chosen carefully. If ϵ_r is too large it becomes impossible to limit the electric length of the passiv feedback network to the next multiple of λ at a given minimum geometric length. We chose SiO_2 with $\epsilon_r = 3.8$ and 525 m thickness.

III. Oscillator Simulation

Literature provides many approaches for oscillator design. Most of them can be grouped as small signal linear or large signal nonlinear techniques. Small signal calculations are able to predict loop matching, loop phase, quality and loop gain at the steady state of the oscillation. A simulation that wants to describe the onset and the peak amplitude of the oscillation in pulsed mode has to take into consideration the large signal behavior of the nonlinear semiconductor. In this paper both simulation types were used. In a first step small signal calculations for a first order approximation, in a second step large signal calculations for more accurate results and optimization in pulsed mode. Small signal design of oscillators is well known today. The oscillator consisting of the transistor, whose drain output is fed back to the input gate by a linear passive network, is split at an arbitrary point, as given in Fig. (1). The impedances of the input and output branch must be complex-conjugately matched ($\Gamma_g = \Gamma_d^*$), the electric length of the loop must be a multiple of λ, as described in Sec. II. At this point a trade-off has to be made. Maximizing the loop gain requires an exact matching, but this becomes narrow-band due to the low-impedance gate and therefore susceptible to tolerances. On the other hand a small gain does not allow the oscillator to reach the desired peak amplitude.

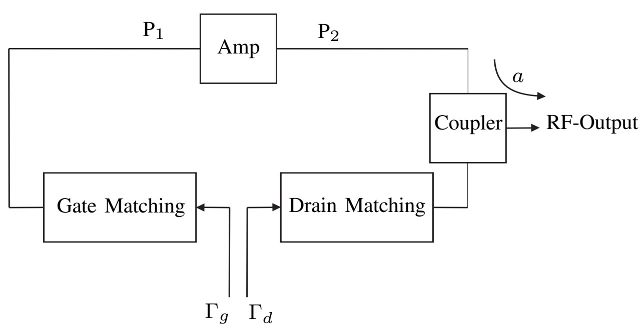

Fig. 1. Small-signal simulation setup.

Additionally the fraction of power coupled out of the loop must not be too large since this power is missing at the transistor input for feedback. To get a starting point for simulation the following simplified consideration is helpful. Let the circuit from Fig. (1) be perfectly matched and the loop be closed. The constant amplification until the transistor reaches saturation shall be $G = 8$ (9 dB), which is a typical small signal value of the applied transistor at 24 GHz. For the transistor output power P_2 follows

$$P_2 = G \cdot P_1 \qquad \text{and} \qquad P_{\text{out}} = a \cdot P_2 \tag{3}$$

with factor $0 < a < 1$ which represents the fraction of power P_2 that is coupled out. Finally

$$P_1 = P_2 - P_{\text{out}} \tag{4}$$

closes the iterative loop. Let the quality of the loop be $Q = 10$. Due to Equ. (1) it describes the approximate number of cycles it takes the oscillator to reach its maximum amplitude. Fig. (2) shows the logarithmic output power normalized to the maximum after 10 cycles versus logarithmic coupling factor of the output. The optimum output power is reached at $a = -10$ dB. A looser or tighter coupling degrades the output power. Small signal simulation is performed with a 2 1/2-D field

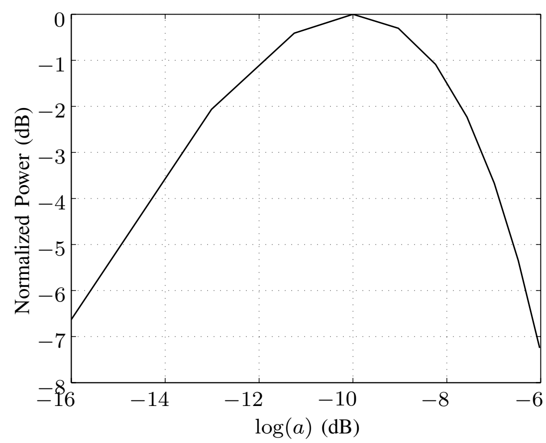

Fig. 2. Normalized logarithmic output power versus logarithmic coupling factor of the output.

simulator [6]. The calculated S-parameters of the passive circuit are combined with the measured S-parameters of the transistor in a circuit simulator [7]. The matching of the transistor ports should be as broadband as possible in order to minimize the quality factor. So low-impedance couplers and cascaded quarter-wave transformers [8] are preferred to parallel stubs since they are easier to predict. Parallel stubs suffer from higher radiation losses, more parasitics and a larger size. The broadband output connection is realized by a third parallel branch at the coupler. The coupling factor can be adjusted in accordance to Equ. (3)-(4) by the distance to the middle branch. At the gate and the drain port of the transistor a sequence of inductive coils serves as a low-pass access for DC-supply. The gain and the phase of the loop are determined as in

Fig. (3) and (4). The Q-factor, measured from the phase slope, meets exactly the simulated value of 8.8.

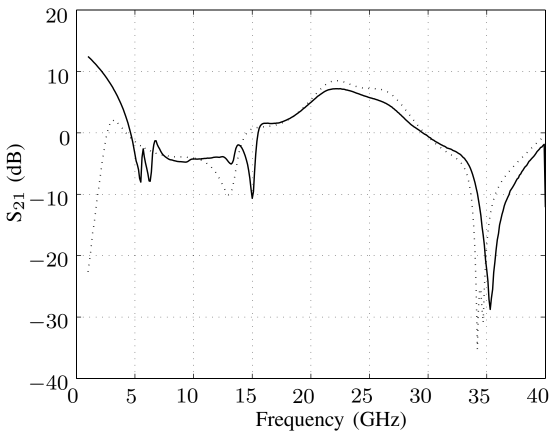

Fig. 3. Gain of the oscillator loop (solid: measured, dotted: simulated).

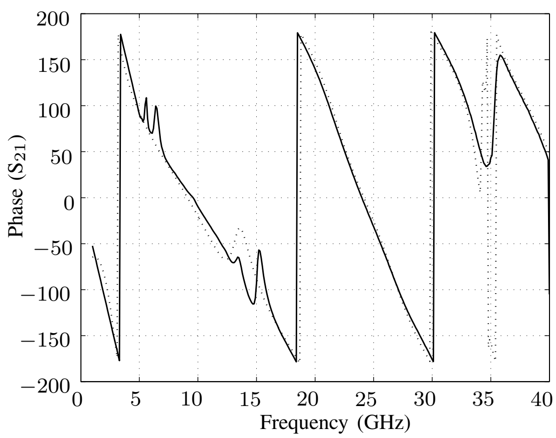

Fig. 4. Phase of the oscillator loop (solid: measured, dotted: simulated).

Large-signal simulation is done in a circuit simulator that provides a transient solver [7]. The transistor is modeled on the basis of SPICE parameters provided by the manufacturer. The passive linear network is included as small-signal parameters. The transistor is switched on at the drain-port by a baseband pulse of $3.3\,\mathrm{V}$ amplitude, $700\,\mathrm{ps}$ width and rise times of less than $100\,\mathrm{ps}$. The gate is tied to a fixed potential of zero volts by a $50\,\Omega$ resistor to ground. The resulting RF-pulse in Fig. (5) shows a peak amplitude of $1.15\,\mathrm{Vpp}$ and a width of $450\,\mathrm{ps}$ at half positive amplitude. Fig. (6) shows the measured attenuation from the transistor drain to the output port of the optimized circuit. The measurement is done without the transistor. Its value of $-9.2\,\mathrm{dB}$ at $24\,\mathrm{GHz}$ shows a good agreement with the predicted optimum in Fig. (2).

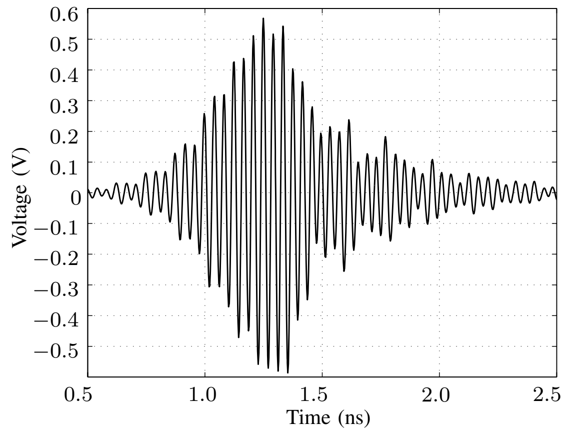

Fig. 5. Large-signal simulation of the oscillator.

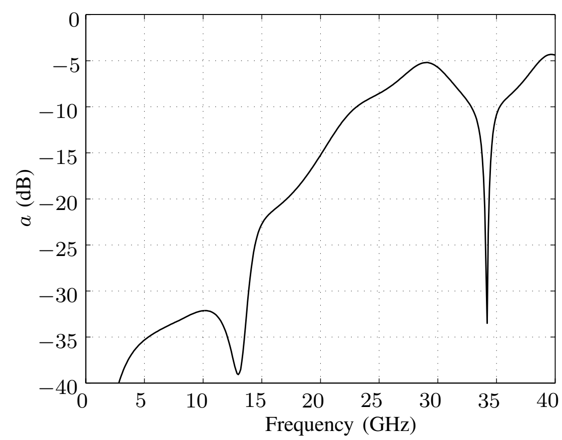

Fig. 6. Measured attenuation from drain-port to the output, which equals factor a (measured without transistor).

IV. Results

The pulse oscillator is manufactured as shown in Fig. (7). The bond wires across the coplanar lines serve to suppress slot modes. The gate-resistor can be placed outside the package. Fig. (8) shows the measured voltage of the RF-pulse at $50\,\Omega$ in time domain. The peak amplitude with $1\,\mathrm{Vpp}$ is only little smaller than predicted $1.15\,\mathrm{Vpp}$ from the time-domain simulation of Fig. (5). The width of the pulse is measured as $480\,\mathrm{ps}$, which almost meets the simulated $450\,\mathrm{ps}$. The ringing of the pulse is less that in the simulation. This is caused by the deficient consideration of losses in the simulation, which attenuate the oscillator loop when it is switched off. Fig. (9) shows the measured corresponding power spectrum density. The $3\,\mathrm{dB}$-bandwidth is determined as $2.1\,\mathrm{GHz}$.

The oscillator is mounted in a low-cost commercial package, covered by a metal lid and soldered to a microstrip footprint on a RO4003 PCB. The module is

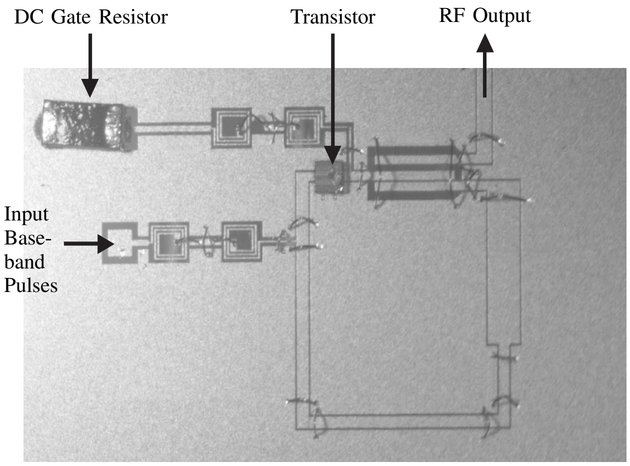

Fig. 7. Photograph of the oscillator including flip-chip mounted transistor, gate bias resistor and bond wires to suppress slot modes. The size of the module is 3.9×5.7mm

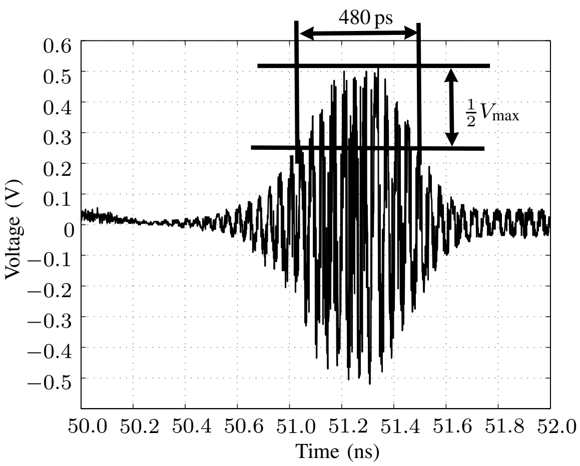

Fig. 8. Measured voltage of RF-pulse at $50\,\Omega$ in time domain.

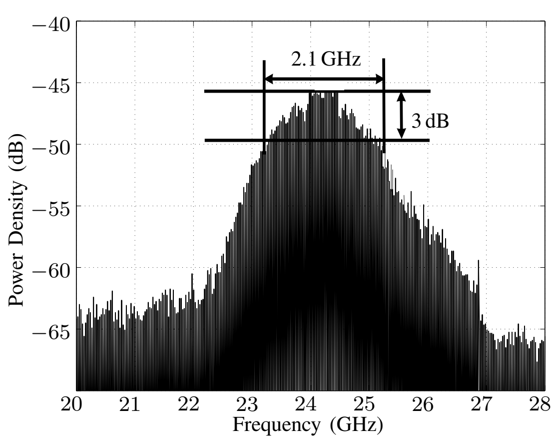

Fig. 9. Measured spectral power density of the RF-pulse at $50\,\Omega$.

employed in a incoherent superregenerative radar system as described in [5]. The pulse shape degradation is mainly caused by ringing due to package parasitics. The achieved resolution of the radar system of 10 cm was not affected by this effect.

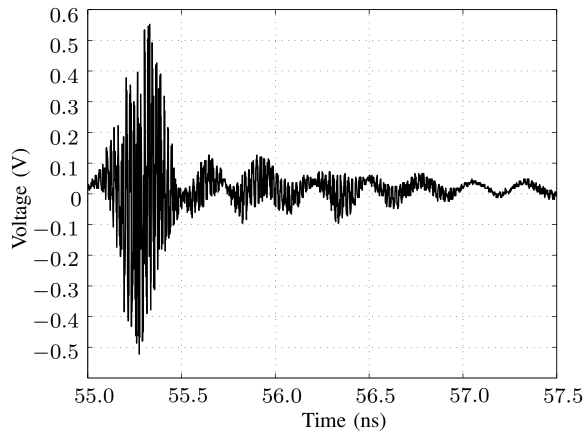

Fig. 10. Measured output signal of the oscillator mounted in a low-cost commercial package which is soldered to a microstrip footprint on a RO4003 PCB.

V. Conclusion

The paper shows the successful optimization of the time duration of a pulse oscillator at 24 GHz using using small-signal and large-signal simulation techniques as well as simple mathematical models. The oscillator is designed in thin film technique and proves a bandwidth of 2.1 GHz at 24 GHz, which corresponds to a time duration of 480 ps. To the author's best knowledge this is the shortest pulse duration reported at 24 GHz using switched oscillators. The module is packaged and employed in a incoherent radar system which achieves a resolution of 10 cm.

References

[1] Federal Communications Commission, FCC. *First Report and Order. Revision of Part 15 of the Commissions Rules Regarding Ultra-Wideband Transmission Systems*, 2002.

[2] D. Barras, F. Ellinger, H. Jäckel, W. Hirt. *A Robust Front-End Architecture for Low-Power UWB Radio Transceivers*. IEEE Transactions on Microwave Theory and Techniques, vol. 54, no. 4, 2006.

[3] T. Troll, J. Detlefsen. *Motion Monitoring with a Millimeter Wave Radar Sensor.* 27th European Microwave Conference and Exhibition, p. 39 - 44, 1997.

[4] A. Kryshtopin, G. Sevskiy, K. Markov, P. Heide, M. Nalezinski, R. Roskosch, M. Vossiek. Cost-Minimized 24 GHz Pulse Oscillator for Short-Range Automotive Radar Application. *European Microwave Conference*, p. 1131-1134, 2003.

[5] T. Wuchenauer, M. Nalezinski, W.Menzel. *Superregenerative Incoherent UWB Pulse Radar System*. IEEE MTT-S International Microwave Symposium Digest, p. 1410-1413, 2006.

[6] Agilent Technologies. *Momentum*. 2 1/2-D Field Simulator.

[7] Agilent Technologies. *Advanced Design System (ADS)*. Circuit Simulator.

[8] G. L. Matthaei, L. Young, E. M. T. Jones. *Microwave Filters, Impedance Matching Networks, and Coupling Structures*. Sec. 6.12, Arctech House, 1980.