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2009

Diao, S., Zheng, Y., & Heng, C. H. (2009). A CMOS Ultra Low-Power and Highly Efficient UWB-IR Transmitter for WPAN Applications. IEEE Transactions On Circuits And Systems—II. 56(3), 200-204.

<https://hdl.handle.net/10356/92057>

<https://doi.org/10.1109/TCSII.2009.2015369>

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A CMOS Ultra Low-Power and Highly Efficient UWB-IR Transmitter for WPAN Applications

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Abstract—This brief presents an on-off LC oscillator-based ultrawideband impulse radio (UWB-IR) transmitter for long-range application. A thorough theoretical analysis of the pulse generation is provided. Implemented in a $0.18\text{-}\mu\text{m}$ CMOS, the transmitter works in the UWB lower band of 3–5 GHz and consumes an ultralow average power of $236\text{ }\mu\text{W}$ at 1.8-V power supply. UWB pulses with a bandwidth of 2 GHz and 10-dB sidelobe suppression are generated. The transmitter can deliver a large differential output swing of 4.9 V under $100\text{-}\Omega$ load with the highest power efficiency of 25.4% to date. It is targeted for wireless sensor network (WSNs) and wireless personal area network (WPAN) applications.

Index Terms—CMOS IC, FCC spectral mask, impulse radio (IR), low data rate, low duty cycle, low power, pulse generator, ultrawideband (UWB), wireless.

I. INTRODUCTION

MULTIBAND ultrawideband impulse radio (UWB-IR) is a promising technology for low-data-rate communication (e.g., IEEE 802.15.4a) and offers additional capabilities such as precision localization and positioning besides data communication. Research on this technology is spurred by emerging applications such as the wireless sensor network (WSN) and wireless personal area network (WPAN), which require a low-power and low-cost communication system.

Several works have been published for the UWB-IR transmitter. In [1], although a simple pulse generator is used, it requires a complicated pulse modulator for pulse shaping to meet the FCC spectral mask requirement, and many driving amplifiers to achieve the desired output power, which results in a lot of inductor usage and more power. In [2], an on-off LC oscillator-based pulse generator is reported, which demonstrates both low complexity and low power consumption. However, the reported transmitter output amplitude is limited to 180 mV peak-to-peak and may be suitable for very-short-range communication.

To date, most of the reported transmitters have limited peak output power and, thus, limit the communication distance. Section II shows that low pulse repetition frequency (PRF) and short pulse duration can be traded for high peak output power to achieve an overall UWB output spectrum that still meets the FCC requirement. Therefore, long-range communication

is possible for indoor/outdoor low-data-rate application if the transmitter is capable of achieving high peak output power given the same receiver sensitivity. In this brief, we adopt an on-off LC oscillator-based pulse generator and target at long-range application, which poses different design challenges due to the large differential output voltage swing requirement. The proposed transmitter is capable of delivering high peak output power at high efficiency. Fast startup is achieved, which is critical in getting low duty cycle for ultralow average power consumption.

Section II presents the theoretical analysis of an on-off LC oscillator-based UWB pulse transmitter suitable for long-range application. The overall CMOS implementation of the proposed transmitter is shown in Section III, and the experimental results are discussed in Section IV. This is then followed by a conclusion in Section V.

II. THEORETICAL ANALYSIS

A. UWB Pulse

A typical UWB pulse train waveform in time domain and its corresponding output spectrum are shown in Fig. 1. As illustrated, the output spectrum dependence on some controlled variables such as PRF, pulse duration (t_{pd}), and signal amplitude (V_{pk}) is shown. The three parameters mentioned are not the only parameters that will affect the spectrum, but are chosen because they can readily be changed. In general, a lower PRF and a shorter pulse width can be exchanged for a larger signal amplitude without violating the FCC spectral mask requirement [3]. The resulting larger signal amplitude, which has a higher peak output power, could then be used for long-range application.

B. On-Off LC Oscillator-Based Transmitter

Although a Gaussian pulse has all the desirable characteristics suitable for UWB transmission, its generation requires a complicated filtering circuit and is thus less power and area efficient. A typical on-off LC oscillator-based transmitter has been proposed [2] as an alternative for UWB transmission. Return-to-zero (RZ) data with very short pulse duration (a few nanosecond) can be input to the transmitter to power up/down the LC oscillator. Its main attractiveness lies in its simplicity and high efficiency in generating a short oscillating output waveform that closely approximates a triangularly shaped pulse. However, previous analysis has been limited to small output amplitude ($< 180\text{ mV}$), and the LC oscillator is only allowed to operate within the early startup linear regime. The design equations presented in [2] are revisited in this brief to address the high-output-swing requirement.

Manuscript received July 11, 2008; revised October 17, 2008. Current version published March 20, 2009. This paper was recommended by Associate Editor M. Ghovanloo.

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Digital Object Identifier 10.1109/TCSII.2009.2015369

Fig. 1. (a) Time-domain UWB pulse. (b) UWB output spectrum for different PRF, t_{pd} , and V_{pk} .

The whole on-off LC oscillator pulsed waveform can be split into a few regions to ease the analysis, as illustrated in Fig. 2. They are the off period (t_{off}), the linear startup period (t_{st}), the linear on period ($t_{l,on}$), the nonlinear period (t_{nl}), and the linear turnoff period ($t_{l,off}$). Together, $t_{l,on}$, t_{nl} , and $t_{l,off}$ form the total pulse duration (t_{pd}).

1) *Off Period*: To speed up the startup transient of the LC oscillator and maximize the output swing, a huge tail current transistor is used. This presents a large gate capacitance (C_{gate}) to the digital driving logic gates, which control the on/off of the tail current transistor. Therefore, the gate voltage of the tail current transistor charges up linearly following the RC

Fig. 2. Oscillation process of a switchable LC oscillator.

Fig. 3. (Left) General LC VCO and (right) its small-signal equivalent circuit.

time constant. The off period would last until the gate voltage ($V_{G,tail}$) of the tail current exceeds the threshold voltage (V_{THN}) and can be shown as follows:

$$V_{G,tail} = V_{DD} \left(1 - e^{-\frac{t}{RC_{gate}}} \right) \approx V_{DD} \times \frac{t}{RC_{gate}} = \alpha t \quad (1)$$

$$t_{off} \approx \frac{V_{THN}}{V_{DD}} \times RC_{gate}. \quad (2)$$

2) *Linear Startup Period*: Using the small-signal equivalent circuit illustrated in Fig. 3, the startup transient of the oscillator output ($V_o(t)$), given an input excitation of $V_i(t)$, can be derived as follows [4]:

$$V_o(t) = f[V_i(t)] + A_1 \cdot e^{-\frac{\omega_0}{2Q}(1-g_m R_T)t} \cos(\omega'_0 t) \quad (3)$$

where A_1 is a constant that depends on the initial condition, ω_0 is the desired oscillation frequency, ω'_0 is the frequency of the zero crossings during the oscillation buildup, g_m is the transconductance of the cross-coupled pair, and R_T is the total output resistance, including the LC tank equivalent resistance, as well as the output resistance of the cross-coupled pair.

As it takes time for the gate voltage of the tail transistor to charge up, the transconductance of the cross-coupled pair, which depends on the tail current, would also vary with time. To the first order, both the tail current and the transconductance can be derived based on the gate voltage in (1) as follows:

$$I_{tail} = K_{tail}(V_{G,tail} - V_{THN})^2 \approx K_{tail}(\alpha t - V_{THN})^2 = K_{tail}(\alpha t_1)^2 \quad (4)$$

$$g_m = \sqrt{4K_{cp} \frac{I_{tail}}{2}} \approx \sqrt{2K_{cp} K_{tail} \alpha^2 \times t_1} = \beta t_1 \quad (5)$$

where $t_1 = t - t_{\text{off}}$, and K_{tail} and K_{cp} are the MOSFET device parameters associated with the tail current transistor and cross-coupled pair transistor, respectively.

Combining (5) and (3) and assuming the loop gain ($g_m R_T$) to be much larger than unity, a resulting exponential envelope is obtained during the linear startup region as follows:

$$V_o(t_1) \approx f[V_i(t_1)] + A_1 \cdot e^{\frac{\omega_0}{2Q} \beta R_T t_1^2} \cos(\omega_0' t_1). \quad (6)$$

After the off period, it will take some time (t_{st}) for the oscillation to build up from the initial off condition ($\sim 0.001 \times V_{\text{peak}}$) to about $0.01 \times V_{\text{peak}}$. The equation is shown as

$$t_{\text{st}} \approx \sqrt{\ln(10) \times \frac{2Q}{\omega_0 \beta R_T}}. \quad (7)$$

The sum of t_{off} and t_{st} would be the required setup time (t_{setup}), which limits the minimum duty cycle as well as the overall energy efficiency. If we assume that the oscillator is operating at the linear startup region when $V_o(t_1)$ rises from $0.01 \times V_{\text{peak}}$ to $\gamma \times V_{\text{peak}}$, then $t_{l,\text{on}}$ can be obtained as follows:

$$t_{l,\text{on}} \approx \sqrt{\ln(100\gamma) \times \frac{2Q}{\omega_0 \beta R_T}}. \quad (8)$$

γ is a coefficient to define the boundary condition between linear and nonlinear regions.

3) *Nonlinear Period:* When the output voltage grows larger, the cross-coupled pair no longer operates at the saturation region and causes the oscillator to enter the nonlinear startup region. From [5], it was shown that when a large output voltage is applied across the transistor, the ratio of the large signal transconductance (G_m) to the small signal transconductance (g_m) is inversely proportional to the output voltage amplitude as follows:

$$\frac{G_m}{g_m} \approx \frac{(V_{\text{GS}} - V_{\text{THN}})}{V_o} \quad (9)$$

where $(V_{\text{GS}} - V_{\text{THN}})$ is the gate-to-source overdrive of the cross-coupled pair transistor, and the relationship holds if V_o is relatively large compared to the gate-to-source overdrive. Given this information, the nonlinear transient in (3) can be modified and becomes

$$V_o(t_1) = f[V_i(t_1)] + A_1 \times e^{-\frac{\omega_0}{2Q} \left[1 - g_m \frac{(V_{\text{GS}} - V_{\text{THN}})}{V_o(t)} R_T\right] t_1} \cos(\omega_0' t_1). \quad (10)$$

Unfortunately, no closed-form solution can be derived for t_{nl} based on (10). However, the huge tail current has resulted in a large cross-coupled-pair gate-to-source overdrive given a relatively smaller sizing of the cross-coupled pair transistor. This implies that the linear startup analysis can be valid even for relatively large $V_o(t)$. Therefore, the γ factor used in the linear startup analysis can approach unity, and the t_{nl} can be assumed negligible. In fact, this assumption is corroborated by both the calculation and the simulation result shown in the subsequent sections.

Fig. 4. t_{pd} , t_{setup} , and loop gain for various tail current transistor sizing.

4) *Linear Turnoff Period:* When the tail current transistor turns off, the oscillator dies down following the $C_T R_T$ time constant, where C_T is the total output capacitance that helps define the oscillation frequency. The time taken for it to discharge from V_{peak} to $0.01 \times V_{\text{peak}}$ can be derived as follows:

$$t_{l,\text{off}} = C_T R_T \ln(100). \quad (11)$$

5) *Design Consideration:* The Q of the tank circuit is less than 2 due to the small load resistance of $\sim 50 \Omega$. It will affect both the phase noise performance and the startup characteristics. The output peak voltage is mainly determined by the maximum tail current and the effective R_T . Although a larger Q is desirable for better phase noise performance, larger amplitude, and longer pulse duration, and can be improved by using a larger inductance, it is ultimately limited by the desired tuning range and the parasitics.

A series of plots versus the tail current transistor sizing is obtained based on the equations derived earlier, as illustrated in Fig. 4. To generate a 2-GHz bandwidth (-10 dB) UWB pulse, a larger pulse duration (t_{pd}) is desired. Although this can be achieved by choosing a larger tail current transistor, it has the detrimental effect of reducing the energy efficiency due to the increasing t_{setup} , as well as lowering the initial loop gain needed to guarantee oscillation startup. A sizing of $400 \mu\text{m}$ is chosen as a compromise among various performance considerations, which will generate a UWB signal with pulse duration of 910 ps and an initial loop gain of 2.

The cross-coupled pair will affect the center frequency, the pulse duration, the loop gain, and the voltage headroom. For higher loop gain and voltage headroom, a larger sizing is preferable. However, the rule of thumb is to make it smaller than the tail current transistor. This not only ensures that the nonlinear period (t_{nl}) remains negligible due to the large gate-to-source overdrive of the cross-coupled pair transistors but also makes room for the wide tuning range that employs a varactor. Therefore, a sizing of $120 \mu\text{m}$ is chosen in this design.

III. UWB-IR TRANSMITTER DESIGN

Fig. 5 shows the proposed UWB transmitter, which includes a pulse generator and an LC voltage-controlled oscillator

Fig. 5. Proposed UWB-IR transmitter.

(VCO). The modulation adopted is on-off keying (OOK) with RZ input data. The pulse width of the RZ input data has 50% duty cycle and is not critical in determining the UWB output spectrum. An internal pulse generator with programmable control pulse width and shape would be used to generate the desired UWB pulse that controls the LC oscillator. The control pulse width and shape at the gate of the tail current transistor can be varied by adjusting C3–C6 and controlling gate bias of M4–M6, which behave like a linear-region resistor. From the simulation, the pulse generator is capable of generating nanosecond pulse (1 ns), whereas the theoretical calculation predicted a pulse duration of 910 ps. The slight discrepancies are due to the exclusion of t_{nl} and the constant gate capacitance model adopted in the theoretical calculation, which validates our earlier assumption of ignoring t_{nl} in t_{pd} . The VCO core consists of the cross-coupled NMOS pair (NM2 and NM3), the 1-nH inductors (L1 and L2), and the varactors (C1 and C2) of 0.5 pF, and its output directly drives the 100- Ω differential loads, as shown. This choice impacts the parallel load on the inductor and gives a Q lower than 2. Here, the simplicity in design is traded off with the poorer phase noise performance, which is acceptable for an energy-detection UWB system. The accumulation-mode MOS varactor employed in this design can achieve 20% tuning range, which can compensate the frequency drift due to process and temperature variations. From the simulation, t_{setup} , which includes t_{off} and the time taken for the oscillation to reach $0.01 \times V_{peak}$ (t_{st}), can be controlled to within 255 ps. Our theoretical prediction indicates 360 ps, which can serve as a low bound for prediction of the startup time, particularly during corner simulation. The discrepancy is due to the mismatch purposely introduced to the cross-coupled pairs to aid the start-up process, which is difficult to include in our theoretical analysis. The introduced mismatch is equivalent to injecting a larger noise amplitude into the resonant tank and thus reducing the t_{st} required to reach the $0.01 \times V_{peak}$.

The peak current consumption (I_{max}) during turn-on is around 70 mA, and the R_T is around 42 Ω . This gives rise to a differential output voltage amplitude of 5.88 V, as shown in Fig. 6, which matches quite closely to the measurement result (dash line). As the proposed transmitter operates as an open-loop VCO, the center frequency cannot accurately be controlled (within 1%) without phase-locked loop (PLL). However, it is sufficient to be used in a noncoherent receiver that employs energy detection and for achieving low-power consumption.

Fig. 6. Simulation results of the proposed UWB-IR transmitter.

Fig. 7. Die photograph of the UWB transmitter.

IV. EXPERIMENTAL RESULT

The transmitter is realized in Chartered Semiconductor Manufacturing's (CSM) 0.18- μm CMOS process and occupies a die area of 0.47 mm $\times$ 0.4 mm (core circuit), as illustrated in Fig. 7. With 2-Mb/s RZ input data, the measured timing waveform is shown in Fig. 8(a). From the measurement, the transmitter requires a total setup time of 236 ps, which is quite close to the simulation results. It is capable of generating an output pulse with 1-ns pulse duration and peak-to-peak amplitude of 4.9 V. Both the total setup time and the pulse duration measurement match closely to the simulated result. However, the desired output swing is smaller than the simulated result, probably due to the lower quality factor of the spiral inductor, which results in a smaller R_T . On the other hand, the effect of reducing R_T on pulse duration is mixed. It will increase $t_{l,on}$ but decrease $t_{l,off}$ and result in little net change on t_{pd} . The differential output is converted to a single-ended output through balun for measuring the transmitter output spectrum shown in Fig. 8(b). It is centered at 4.3 GHz and exhibits a 2-GHz bandwidth (-10 dB). This meets the FCC requirement of both peak and average output power [3]. To our best knowledge, this is the highest output swing reported for the integrated CMOS UWB transmitters. At a data rate of 2 Mb/s (can be scalable), the whole transmitter consumes an average power of 236 μW . The performance is summarized in Table I.

To facilitate the comparison of different transmitters, the energy efficiency (η) is defined as follows:

$$\eta = \frac{\text{Output Peak Power} \times \text{Pulse Width (J)}}{\text{DC Energy Consumption per Pulse (J/pulse)}}. \quad (12)$$

η is defined as the ratio of the emission energy to the average power consumption per pulse. The pulse width is effective from

(a)

(b)

Fig. 8. (a) Chip output pulse sequence (x : 500 ns/div; y : 2 V/div) with 2-MHz square-wave digital input (x : 500 ns/div; y : 1 V/div). (b) Measured UWB pulse output spectrum in compliance with the FCC mask (dash line).

TABLE I
PERFORMANCE SUMMARY

Measurement results	
Process	CMOS 0.18 μ m
Supply Voltage	1.8V
Die Area	0.47mm $\times$ 0.4mm (core circuit)
Modulation	OOK
Maximum Data Rate	2 Mbps
Input data duty cycle	0.2%
Average Power consumption(@2Mbps)	236 μ W
Pulse duration	~1ns (tunable)
Differential output swing (V_{pp})	4.9V
Center frequency	4.3GHz
Frequency band	3.3~5.3GHz
Setup time	236ps
Peak current	64mA

5% of the peak amplitude. Due to different applications, the data rate and the output peak power vary a lot from different UWB transmitters. Therefore, it is only fair to compare the energy efficiency per UWB pulse, which is independent of the data rate. A comparison of the energy efficiency, power consumption, and area for different transmitters is listed in Table II. The dc energy consumption for all the blocks is considered for this work, whereas the others are based on their

TABLE II
PERFORMANCE COMPARISON ON UWB PULSE TRANSMITTERS

Reference	Technology	η (/pulse)	Power consu @Data Rate (W/bps)	Area (mm ²).
[2], 2007	CMOS 0.18 μ m	1.575%	1.823m@100M	0.3944
[6], 2007	CMOS 90nm	5.745%	0.452m@10M	0.08
[7], 2007	CMOS 0.18 μ m	0.22%	29.7m@36M	0.4
[8], 2006	CMOS 0.13 μ m	0.67%	10m@160M	1.56
[9], 2006	CMOS 0.18 μ m	>1%	1.8m@10M	0.57
[10], 2008	BiCMOS 0.35 μ m	-	-@184M	0.017
This Work	CMOS 0.18μm	25.4%	0.236m@2M	0.188

quoted value, and some have excluded the buffer's contribution. As illustrated, our proposed transmitter exhibits the best energy efficiency.

V. CONCLUSION

A UWB impulse transmitter based on a burst-controlled LC VCO has been presented in this brief. Theoretical equations that aid the selection of transistor sizing have been derived and are shown to predict the pulse duration quite well. The chip has been implemented in a 0.18- μ m CMOS process and demonstrated UWB pulse generation using OOK modulation with low power consumption. The highest output swing suitable for long-range application with high energy efficiency has been achieved. It is well suited for WSN and WPAN applications.

ACKNOWLEDGMENT

The authors would like to thank Annamalai, H. Dong from IME, and F. Ting from NUS for their kind suggestions and help on this circuit design and implementation.

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