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<td>Diao, Shengxi; Zheng, Yuanjin; Heng, Chun-Huat</td>
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A CMOS Ultra Low-Power and Highly Efficient UWB-IR Transmitter for WPAN Applications

Shengxi Diao, Yuanjin Zheng, Member, IEEE, and Chun-Huat Heng, Member, IEEE

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 generator is provided. Implemented in a 0.18-μm CMOS, the transmitter works in the UWB lower band of 3–5 GHz and consumes an ultralow average power of 236 μW 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-Ω load with the highest power efficiency of 25.4% to date. It is targeted for wireless sensor network (WSN) 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 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.
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 $RC$ 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:

$$V_{G, tail} = V_{DD} \left(1 - e^{-t/RC_{gate}}\right) \approx V_{DD} \times t/RC_{gate}$$  \hspace{1cm} (1)

$$t_{off} \approx V_{THN}/V_{DD} \times RC_{gate} \times t$$  \hspace{1cm} (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]:

$$V_o(t) = f[V_i(t)] + A_1 \cdot e^{-t/(1-g_m R_T)} \cos(\omega_0 t)$$  \hspace{1cm} (3)

where $A_1$ is a constant that depends on the initial condition, $\omega_0$ is the desired oscillation frequency, $\omega'_0$ 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:

$$I_{tail} = K_{tail}(V_{G, tail} - V_{THN})^2 \approx K_{tail}(\alpha t - V_{THN})^2$$  \hspace{1cm} (4)

$$g_m = \sqrt{4K_{cp}I_{tail}/2} \approx \sqrt{2K_{cp}K_{tail}\alpha^2 \times t_1} = \beta t_1$$  \hspace{1cm} (5)
where \( t_1 = t - t_{\text{off}} \) and \( K_{\text{tail}} \) and \( K_{\text{cp}} \) 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:

\[
V_o(t_1) \approx f [V_i(t_1)] + A_1 \cdot e^{\frac{-g_m R_T}{2} t_1^2} \cos (\omega_0 t_1). \tag{6}
\]

After the off period, it will take some time \((t_{\text{st}})\) 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

\[
t_{\text{st}} \approx \sqrt{\ln(10) \times \frac{2Q}{\omega_0 R_T}}. \tag{7}
\]

The sum of \( t_{\text{off}} \) and \( t_{\text{st}} \) would be the required setup time \((t_{\text{setup}})\), 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 \(t_{\text{on}}\) can be obtained as follows:

\[
t_{\text{on}} \approx \sqrt{\ln(100\gamma) \times \frac{2Q}{\omega_0 R_T}}. \tag{8}
\]

\(\gamma\) 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:

\[
G_m \approx \frac{(V_{GS} - V_{THN})}{V_o} g_m \tag{9}
\]

where \((V_{GS} - V_{THN})\) 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

\[
V_o(t_1) = f [V_i(t_1)] + A_1 \times e^{\frac{-g_m (V_{GS} - V_{THN})}{V_{THN} R_T} t_1} \cos (\omega_0 t_1). \tag{10}
\]

Unfortunately, no closed-form solution can be derived for \(t_{\text{nl}}\) 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 \(\gamma\) factor used in the linear startup analysis can approach unity, and the \(t_{\text{nl}}\) can be assumed negligible. In fact, this assumption is corroborated by both the calculation and the simulation result shown in the subsequent sections.

III. UWB-IR TRANSMITTER DESIGN

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

![Image](image-url)
(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 $t_{\text{on}}$ and the constant gate capacitance model adopted in the theoretical calculation, which validates our earlier assumption of ignoring $t_{\text{on}}$ in $t_{\text{pd}}$. 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 accumulation-mode MOS varactor employed in this design which is acceptable for an energy-detection UWB system. The design is traded off with the poorer phase noise performance, and the constant gate capacitance model adopted in the theoretical calculation, which validates our earlier assumption of ignoring $t_{\text{on}}$ and the constant gate capacitance model adopted in the theoretical calculation, which validates our earlier assumption of ignoring $t_{\text{on}}$ in $t_{\text{pd}}$. This gives rise to the effect of reducing $t_{\text{on}}$ pulse duration is mixed. It will result in little net change in $t_{\text{pd}}$. The differential output is converted to a single-ended 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 $t_{\text{on}}$ but decrease $t_{\text{off}}$ and result in little net change on $t_{\text{pd}}$. 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 $\mu$W. The performance is summarized in Table I.

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

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

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

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 × 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 transmitted 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 $t_{\text{on}}$ but decrease $t_{\text{off}}$ and result in little net change on $t_{\text{pd}}$. 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 $\mu$W. The performance is summarized in Table I.

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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

<table>
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<th>Reference</th>
<th>Technology</th>
<th>( \eta /\text{(pulse)} )</th>
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<td>[2], 2007</td>
<td>CMOS 0.18( \mu \text{m} )</td>
<td>1.575%</td>
<td>1.823m\text{W} at 100M</td>
<td>0.3944</td>
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<td>[6], 2007</td>
<td>CMOS 90( \mu \text{m} )</td>
<td>5.745%</td>
<td>0.452m\text{W} at 10M</td>
<td>0.08</td>
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<tr>
<td>[7], 2007</td>
<td>CMOS 0.18( \mu \text{m} )</td>
<td>0.22%</td>
<td>29.7m\text{W} at 36M</td>
<td>0.4</td>
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<td>[8], 2006</td>
<td>CMOS 0.13( \mu \text{m} )</td>
<td>0.67%</td>
<td>10m\text{W} at 160M</td>
<td>1.56</td>
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<td>[9], 2006</td>
<td>CMOS 0.18( \mu \text{m} )</td>
<td>&gt;1%</td>
<td>1.8m\text{W} at 10M</td>
<td>0.57</td>
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<td>[10], 2008</td>
<td>BiCMOS -</td>
<td>-</td>
<td>-184M</td>
<td>0.017</td>
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</table>

This Work | CMOS 0.18\( \mu \text{m} \) | 25.4% | 0.236m\text{W} at 2M | 0.188 |

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 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-\( \mu \text{m} \) 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 efficiency has been achieved. It is well suited for WSN and WPAN applications.

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REFERENCES