A Digital Calibration Technique for Charge Pumps in Phase-Locked Systems

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Abstract—A digital technique is adopted to calibrate the current mismatch of the charge pump (CP) in phase-locked systems. In this digital calibration technique, there is no extra replica CP needed. In addition, it can calibrate the CP under different control voltages on the loop filter to be immune to the channel-length modulation. Due to the digital nature, the additional power consumption and digital switching noise from the calibration circuits are turned off once the calibration is finished. A 5 GHz frequency synthesizer is used to justify the proposed calibration technique. The measured output spur is suppressed by 5.35 dB at 5.2 GHz after the calibration circuits are active. The measured output spur levels are less than -68.5 dBc throughout the whole output frequency range. The measured phase noise is -110 dBc/Hz at an offset frequency of 1 MHz.

Index Terms—Charge pump (CP) calibration, digital calibration, frequency synthesizer, phase-locked loop (PLL).

I. INTRODUCTION

CHARGE pumps (CPs) are widely used in analog phase-locked systems, such as phase-locked loops (PLLs), delay-locked loops (DLLs), and clock and data recovery (CDR) circuits. The CP converts the phase difference to a proportional voltage to adjust the voltage-controlled oscillator (VCO) or voltage-controlled delay line (VCDL). However, the mismatch between the up and down currents in a conventional CP introduces several problems in phase-locked systems, such as reference spur [1]–[5], static phase error, and jitter. For the short-channel devices in advanced CMOS technology, the device mismatch and channel-length modulation becomes worse. In frequency synthesizers, the mismatch in a CP results in unwanted spurs and deteriorates the spectral performance [1]. In multiphase DLLs [6]–[8], the unbalanced CP creates the static phase error to generate the fixed pattern jitter or degrade the timing margin. Similarly, in conventional CDR circuits, the mismatch in a CP [9] makes the sampling clock away from the optimal position to increase the bit error rate.

Several methods [10]–[12] have been presented to improve the CP mismatch. In [10], a replica CP and a digital controller are used to calibrate the CP mismatch. However, the mismatch between the two CPs still exists. Since two CPs are calibrated on different controlled voltages, it is hard to compensate the current mismatch due to the channel-length modulation. In [11]

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Fig. 1. Traditional PFD and CP.

and [12], the common-mode feedback circuit is used. However, an additional reference voltage and the operational amplifier are needed. The nonideal effects of the operational amplifier have to be considered; such as the stability and offset voltage.

In this paper, a digital calibration technique is presented to solve these problems. It is realized without an additional replica CP. The power consumption and digital switching noise from the calibration circuit are turned off once the calibration is finished. This technique increases the settling time a little. This paper is organized as follows. Section II introduces the digital CP calibration technique. The system architecture, building blocks, and design considerations are presented in Section III. The experimental results are presented in Section IV and a conclusion is presented in Section V.

Fig. 2. Static phase error, current mismatch, and control line ripple: (a) for a normal reset delay and (b) for a enlarged reset delay.

II. DIGITAL CALIBRATION TECHNIQUE

Fig. 1 shows a conventional phase-frequency detector (PFD) and a CP. This PFD consists of two resettable D flip-flops (DFFs) and an AND gate. Assuming all the digital gates are ideal, we may model their propagation delays in the reset path of a PFD into a reset delay T_d , which is appropriately designed to avoid the dead zone. In Fig. 1, the PFD generates two signals, up and down, to control the CP. Assume that the PFD and CP in Fig. 1 are locked in a PLL. Their timing diagrams are shown in Fig. 1 where the high-level pulsewidths are T_{pup} and T_{pdn} for the up and down controlling signals, respectively. The current pulses corresponding for the up and down signals are also generated, respectively. Assume these two current pulses are perfectly matched, i.e., $I_{up} = I_{dn} = I$ and $T_{pup} = T_{pdn} = T_d$. When the PLL is locked, the voltage on the loop filter is fixed since the net charge provided by the CP should be zero. To maintain the locking condition, the following equation is always satisfied:

$$I_{up} \cdot T_{pup} = I_{dn} \cdot T_{pdn}. \quad (1)$$

When the CP mismatch exists, it is hard to detect the current mismatch directly. Hence, we have to find a signal, which may represent the amount of this current mismatch. For example, assume that the down current is only 80% of the up current, i.e., $I_{dn} = 0.8 \cdot I_{up}$ in Fig. 2(a). Since the up and down currents are not identical, the high-level pulsewidth of the up and down controlling signals should differ as well to satisfy (1). To consider Fig. 2(a), one may derive $T_{pdn} = 1.25 \cdot T_{pup} = 1.25 \cdot T_d$ and $I_{dn} = 0.8 \cdot I_{up} = 0.8 \cdot I$. Since both the falling edges of the up and down controlling signals are aligned by the PFD's resetting nature, the rising edge of the down controlling signal leads that of the up controlling signal; i.e., the static phase error exists. Assuming the reset delay $T_d = 1$ ns in a PFD of Fig. 2(a), it results in the static phase error of 0.25 ns. Similarly, if $I_{dn} > I_{up}$, the rising edge of the up controlling signal leads that of the down controlling signal. It inspires us that a bang-bang phase detector

Fig. 3. Flowchart for the digital calibration technique.

(BBPD) can detect this static phase error to know the current mismatch in a CP.

However, this static phase error imposes a hard limitation since it is much smaller than the reset delay T_d of a PFD, which is only several nanoseconds. In Fig. 2(a), assume that there is 20% current mismatch in a CP and $T_d = 1$ ns, the static phase error is only 0.25 ns. To consider a conventional BBPD in a typical $0.18 \mu\text{m}$ CMOS technology, its minimum detectable timing error $\Delta t_{\min}$ is hard to be less than 50 ps. It means that the calibration resolution is only 5% for the reset delay $T_d = 1$ ns.

In Fig. 2(b), if one can multiply the normal reset delay T_d by a factor of 10, the static phase error is also multiplied by the same factor. The calibration resolution in Fig. 2(b) is also enhanced by 10. The resolution enhancement factor A_{res} is defined as

$$A_{\text{res}} = \frac{T_{d_en}}{T_d} \quad (2)$$

where T_{d_en} represents the enlarged reset delay of a PFD. However, T_{d_en} cannot be increased indefinitely. According to the analysis in [13], the maximum reset delay for a PFD should be less than half of the period T_{ref} of the reference clock to maintain a conventional phase-locked system. Hence, the maximum achievable enhance factor $A_{\text{res_max}}$ is determined as

$$A_{\text{res_max}} = 0.5 \cdot \frac{T_{\text{ref}}}{T_d}. \quad (3)$$

Fig. 6. 4-bit digital-controlled CP.

200 kHz. Once the synthesizer experiences a frequency-hopping command, the LD resets the 4-bit SAR controller to switch the reset delay of a switched-delay PFD to the enlarged one, i.e., T_{d_en} . The 4-bit SAR controller then performs the digital calibration procedure of Fig. 3. Finally, when the calibration is finished, the reset delay of this PFD is switched to the normal one, i.e., T_d . As a result, the calibration technique is not limited by the channel-length modulation effect under different control voltages and no replica CP is needed. Furthermore, since the digital calibration circuits are inactive after the calibration is complete, the additional power consumption and digital switching noise from the calibration circuits are turned off. All the building blocks will be illustrated in the following.

A. Switched-Delay PFD

The switched-delay PFD is shown in Fig. 5 with two reset delays. Two inverter chains are used to realize the normal reset delay of 2 ns and the enlarged one of 20 ns. Once the LD resets the control bits in the SAR controller, the signal “SAR_stop” will be low to select the enlarged reset delay of 20 ns. When the SAR controller finishes, “SAR_stop” will be high to select the normal reset delay of 2 ns. Based on (4) and $\Delta t_{\min} = 50$ ps, the calibration resolution of 0.25% is achieved in our frequency synthesizer.

B. 4-bit Digital-Controlled CP

To be compliant with the digital calibration technique, the digital-controlled CP is adopted in Fig. 6. The up current has a nominal value of 200 μ A and changes only with the control voltage on the loop filter due to the channel-length modulation. The down current is digitally controlled and it may vary from 180 to 210 μ A when the control voltage is pinned at half of the supply voltage, $V_{DD} = 1.8$ V. The current step is chosen as 2 μ A to relieve the worst case current mismatch to 1% in this digital-controlled CP. In this calibration technique, the control codes of the CP vary only during the calibration process. After the calibration is complete, the control codes are fixed to avoid extra digital noise. Furthermore, the cascode topology in this 4-bit digital-controlled CP minimizes the digital coupling.

C. VCO and Dividers

To have a low sensitivity to the gain of a VCO and a wide tunable frequency range to overcome the process variations, a

Fig. 7. VCO.

VCO with coarse and fine tunings are usually preferred [2], [3]. However, it needs an additional frequency acquisition circuit. In this work, only one set of the accumulation-mode varactor is used because the calibrated CP will have a small control voltage ripple. Fig. 7 shows our VCO, which adopts a complementary topology without a tail current source to improve phase noise performance and deliver constant oscillation amplitude [15]. This VCO provides a tuning range from 5 to 5.36 GHz and has a gain of over 300 MHz/V. A symmetric inductor is also used to enhance the phase noise performance. This inductor has a value of 2.2 nH and exhibits a quality factor of 10 at 5 GHz. The current consumption of this VCO is 1.5 mA.

The divider chain is composed of a fully differential divide-by-two circuit and a multimodulus divider. The divide-by-two circuit is realized by a current-mode logic (CML) DFF [16] to achieve high-speed operation and relax the speed requirement for the following dividers. In our frequency synthesizer, the multimodulus divider in [17] is adopted and it provides a division ratio from 257 to 266. With a reference clock of 10 MHz,

Fig. 8. (a) LD. (b) BBPD.

Fig. 9. 4-bit SAR controller.

this frequency synthesizer generates the output frequencies from 5.14 to 5.32 GHz in a step of 20 MHz.

D. LD and BBPD

The LD and BBPD are two important circuits in this digital calibration technique. It is because the LD determines when the digital calibration technique should be reset or turned on. In addition, the BBPD performs the phase comparison between the up and down controlling signals to provide the information for the SAR controller.

The proposed LD is shown in Fig. 8(a) and it is composed of a conventional LD with a window of $\pm T$ ($= 4$ ns) and a deglitch unit. Since the conventional LD has troublesome glitches during the locking transient, the glitches should be eliminated to trigger the digital calibration circuits correctly. The deglitch unit consists of one AND gate and two DFFs (DFF3 and DFF4). These DFFs are triggered by a clock, which is divided by 32 from the input reference clock. During the locking transient, the output of the conventional LD goes high. It is sampled to the output of DFF3 if the divide-by-32 clock is high. If the output of the conventional LD goes low before the next rising edge of the divide-by-32

clock, DFF3 will be reset and the output of the deglitch unit remains low. Conversely, if the output of the conventional LD remains high until the next rising edge of the divide-by-32 clock, both the outputs of DFF3 and DFF4 go high and the proposed LD will output high to indicate a locked condition. The period of the divide-by-32 clock is set to $3.2 \mu\text{s}$, which is long enough to ensure the output of the conventional LD to reach the steady state.

Ideally, a DFF can serve as a simple BBPD to determine the phase relation between the up and down controlling signals. However, in a DFF, there are a finite sampling offset and the unbalanced capacitive load for two controlling signals. To alleviate these problems, the proposed BBPD is shown in Fig. 8(b). The additional DFF not only provides a balanced load for the up and down controlling signals, but also freezes the control codes of the SAR controller when the two controlling signals are very close.

E. 4-bit SAR Controller

The 4-bit SAR controller in [14] and [18] is adopted as shown in Fig. 9. The operation principle is described as follows. In the beginning, the LD will clear all the cells in the SAR controller. When the first clock arrives, the most significant bit (MSB), i.e.,

bit3, is set to high. The MSB will then be maintained or changed to low according to the comparison result from the BBPD at the next clock edge and the next bit, i.e., bit2, is set to high. Finally, the process will repeat until the least significant bit (LSB), i.e., bit0, is determined. The SAR controller also generates a signal, i.e., "SAR_Stop," to indicate the completeness of the calibration procedure and switch the reset delay of a PFD to the normal one.

The period of the clock in the 4-bit SAR controller is an important parameter. If the clock period is too short, the synthesizer remains unsteady and the BBPD may fail to provide the correct information. Conversely, if the clock period is too long, the total calibration time increases dramatically. As a result, it is necessary to choose an appropriate clock period for our calibration system. Since the CP is switched during the calibration transient, the synthesizer may experience the phase acquisition. As shown in Fig. 6, the largest current step in the down current is 16 μA . In the following analysis, the appropriate clock period for our calibration technique is derived.

Assume the synthesizer is locked before the down current is switched and it is modeled as a linear system. A sudden current change I_{incr} of the down current is modeled as a phase step θ_{step} as

$$\theta_{\text{step}} = \left(\frac{I_{\text{up}}}{I_{\text{dn_min}}} \cdot T_{d_en} - \frac{I_{\text{up}}}{I_{\text{dn_min}} + I_{\text{incr}}} \cdot T_{d_en} \right) \cdot 2\pi \quad (6)$$

where $I_{\text{dn_min}}$ denotes the smallest down current in the beginning of the calibration process. Let I_{up} , $I_{\text{dn_min}}$, I_{incr} , T_{d_en} , and T_{ref} be 200, 180, and 16 μA , and 20 and 100 ns, respectively. According to (6), the phase step is calculated as $\theta_{\text{step}} = 0.114$ rad or 6.5° .

The phase transfer function $H(s)$ of the frequency synthesizer on phase domain is shown as

$$H(s) = \frac{\theta_{\text{out}}(s)}{\theta_{\text{in}}(s)} = \frac{N \cdot \omega_C \cdot (s + \omega_Z)}{\frac{s^3}{\omega_P} + s^2 + \omega_C \cdot s + \omega_C \cdot \omega_Z}$$

where

$$\begin{aligned} \omega_P &= \frac{C_1 + C_2}{C_1 \cdot C_2 \cdot R_2} \\ \omega_Z &= \frac{1}{R_2 \cdot C_2} \\ \omega_C &= \frac{I_{\text{cp}} \cdot K_{\text{vco}} \cdot R_2 \cdot C_2}{N \cdot (C_1 + C_2)} \end{aligned} \quad (7)$$

and I_{cp} , K_{vco} , and N denote the nominal CP current, VCO gain, and division ratio, respectively. To simplify the analysis, the system is designed with the maximum phase margin at the unity gain frequency, i.e., $\gamma \equiv (\omega_C / \omega_Z) = (\omega_P / \omega_C)$. The phase error transfer function $H_e(s)$ between the phase error and input phase is then expressed as

$$H_e(s) = 1 - \frac{H(s)}{N} = \frac{s^3 + \omega_C \cdot \gamma \cdot s^2}{s^3 + \omega_C \cdot \gamma \cdot s^2 + \omega_C^2 \cdot \gamma \cdot s + \omega_C^3} \quad (8)$$

Finally, the step response of the phase error $\theta_{e_sr}(s)$ can be derived as

$$\begin{aligned} \theta_{e_sr}(s) &= \frac{\theta_{\text{step}}}{s} \cdot H_e(s) \\ &= \frac{\theta_{\text{step}} \cdot (s^2 + \omega_C \cdot \gamma \cdot s)}{s^3 + \omega_C \cdot \gamma \cdot s^2 + \omega_C^2 \cdot \gamma \cdot s + \omega_C^3} \end{aligned} \quad (9)$$

The stability of the system is heavily related to the value of γ . In order to have a well-controlled settling behavior, the phase margin of 64° and γ of 4.5 are chosen. It ensures that there is no under-damping settling behavior. If $\gamma > 3$, (9) can be further decomposed into (10) as

$$\theta_{e_sr}(s) = \frac{\theta_{\text{step}} \cdot \frac{(1-\gamma)}{(3-\gamma)}}{s + \alpha_1} + \frac{\frac{\theta_{\text{step}}}{(3-\gamma)}}{s + \alpha_2} + \frac{\frac{\theta_{\text{step}}}{(3-\gamma)}}{s + \alpha_3} \quad (10)$$

where $\alpha_1 = \omega_C$, $\alpha_2 = ((\gamma - 1) - \sqrt{\gamma^2 - 2\gamma - 3})/2 \cdot \omega_C$, and $\alpha_3 = ((\gamma - 1) + \sqrt{\gamma^2 - 2\gamma - 3})/2 \cdot \omega_C$.

We also know that α_1 , α_2 , and α_3 are positive real numbers for $\gamma > 3$. Now we can derive the step response of the phase error in the time domain

$$\begin{aligned} \theta_{e_sr}(t) &= \frac{\theta_{\text{step}} \cdot (1-\gamma)}{(3-\gamma)} \cdot e^{-\alpha_1 \cdot t} + \frac{\theta_{\text{step}}}{(3-\gamma)} \cdot e^{-\alpha_2 \cdot t} \\ &\quad + \frac{\theta_{\text{step}}}{(3-\gamma)} \cdot e^{-\alpha_3 \cdot t} \end{aligned} \quad (11)$$

Substituting the corresponding values of this synthesizer into (11), the step response of the phase error is shown in Fig. 10(a). Referring to Fig. 6 and (6), the smallest phase step $\theta_{\text{step_min}}$ is 0.015 rad when the system experiences a 2 μA down current change in the CP. From Fig. 10(a), a calibration period larger than 4 μs is enough for a phase error smaller than $\theta_{\text{step_min}}$ to ensure the calibration resolution.

In this work, assume that the down current is initially less than the up current in the CP. We detect the polarity of the static phase error to increase the down current in the CP current. The binary-weighted current is updated by the SAR controller. The updated process is from the MSB to LSB to minimize the current mismatch in the CP. The period of the calibration clock for the 4-bit SAR controller is chosen as 12.8 μs . Furthermore, as described in Section III-D, the switched-delay PFD enables a detecting resolution of 0.25%, which is enough to detect the 1% current step (2 μA) in the CP. As a result, the calibration process may guarantee a minimum current mismatch in the CP.

A behavior simulation using the parameters in Table I is performed and the result is shown in Fig. 10(b). This frequency synthesizer experiences a frequency change from 5.18 to 5.26 GHz. It takes about 22 μs for the LD to acquire the lock condition. The calibration procedure spends 51.2 μs under the calibration period of 12.8 μs . Finally, the total settling time is about 80 μs for this frequency synthesizer to settle within 20 ppm of the target frequency. It should be noted that the calibration procedure is performed once the frequency changes. In fact, the digital codes do not vary much across the full VCO tuning range. To overcome the PVT variations, this technique can be refreshed in the idle time, or the 4-bit SAR controller is executed only once and then transformed into a 4-bit counter [19] to overcome the PVT variations.

Fig. 10. (a) Step response for 16 μ A current change, (b) Behavior simulation from 5.18 to 5.26 GHz.

F. Theoretical Spur Level and Nonidealities

Once the parameters of the PLL are determined, the reference spur can be calculated approximately [20], [21] by assuming a rectangular current pulse and neglecting the PFD's delay time. The reference spur P_{spur} is given as [20], [21]

$$P_{\text{spur}} = 20 \log \left[\frac{I_{\text{cp}} \cdot R_2 \cdot K_{\text{vco}} \cdot T_d}{\sqrt{2}} \cdot \frac{\Delta I}{I_{\text{cp}}} \right] - 20 \log \left(\frac{f_{\text{ref}}}{f_{p1}} \right) \text{ (dBc)} \quad (12)$$

where f_{ref} is the input reference frequency, I_{cp} is the nominal CP current, ΔI is the CP current mismatch, and f_{p1} ($= (C_1 + C_2) / (2\pi \cdot R_2 \cdot C_1 \cdot C_2)$) is the pole of the loop filter. Assume the initial current mismatch is 10% and the calibrated current mismatch is 1% for the CP. By using (12) and the parameters in Table I, the calculated reference spur without and with calibration is -51 and -71 dBc, respectively. However, some effects will degrade the reference spur further, such as undesired coupling from the substrate and supply voltage and the reference clock feedthrough from the parasitic capacitor and routed metal lines [22]. The practical improvement may not be as good as expected.

IV. EXPERIMENTAL RESULTS

This chip has been fabricated in a 0.18 μ m CMOS 1P6M mixed-signal technology. A die photograph is shown in Fig. 11. The total chip area and active region occupy 1.4 mm $\times$ 1.2 mm and 0.8 mm $\times$ 0.8 mm, respectively. The measured tuning curve of the VCO is shown in Fig. 12. The VCO covers a frequency

Fig. 11. Die photograph.

Fig. 12. Measured tuning curve of the VCO.

range from 5 to 5.36 GHz and it is suitable for applications in lower U-NII band (5.15–5.35 GHz). The measured spectral performance without and with the CP calibration is shown in Fig. 13(a) and (b), respectively. The carrier frequency is chosen to be 5.22 GHz because the VCO has a maximum tuning gain here. The output spurs are further suppressed by 5.35 dB when the calibration circuits are activated. Fig. 14 shows the measured spurs versus the synthesizer's output frequencies. Our synthesizer has the output spur levels less than -68.5 dBc throughout the output frequency range. The measured phase noise is -110 dBc/Hz at an offset frequency of 1 MHz from the carrier, as shown in Fig. 15.

The digital calibration transient is measured by a modulation analyzer. The frequency hops from 5.18 to 5.26 GHz with an 80 MHz step the same as [2]. Limited by the maximum detectable frequency of 2.5 GHz in this modulation analyzer, the locked synthesizer's output had been down-converted by a 4 GHz tone using a Mini-Circuit ZMX-8 GH mixer in advance of the analyzer's input. The measured switching process in Fig. 16 shows that the synthesizer takes about 80 μ s to settle within 20 ppm of the target frequency, which meets the simulation in Section III.

Table II gives the performance summary of the proposed frequency synthesizer with those in literature. Compared with other works, our synthesizer has a very low f_{Ref} -bandwidth

Fig. 13. Synthesizer spectrum: (a) without calibration and (b) with calibration.

Fig. 14. Measured spur levels versus output frequency.

Fig. 16. Measured frequency switching behavior from 5.18 to 5.26 GHz.

TABLE II
SUMMARY AND COMPARISON

Ref.	[2]	[3]	[4]	[5]	This work
Freq. (GHz)	5.14-5.7	4.12-4.72	5.15-5.825	5.15-5.35	5.14-5.32
Tech.	0.25μm CMOS	0.25μm CMOS	0.18μm CMOS	0.18μm CMOS	0.18μm CMOS
f _{Ref} (MHz)	10	4	10	13.3	10
BW (kHz)	25	90	150	200	200
Settling time (μs)	100	<60	N/A	N/A	80
Phase noise @ 1MHz off. (dBc/Hz)	-116	-87	-115	-113	-110
Loop filter	3rd	2nd	3rd	2nd	2nd
Spurs (dBc)	-70	-45	<-65	<-66	<-68.5
Power (mW)	13.5	117.5	N/A	56	19.8

V. CONCLUSION

A digital calibration technique has been proposed to alleviate the mismatch in conventional CP circuits. Unlike the previous method with a replica CP, this proposed technique calibrates the CP circuit in the loop without a replica CP. For the wideband applications, the control voltage of the VCO may vary in a wide range. Our technique may provide a good way to compensate the CP current mismatch due to the channel-length modulation. The technique consumes negligible power with a cost of extra

Fig. 15. Measured synthesizer phase noise.

ratio, but still presents a comparable spur level and frequency switching time. Furthermore, only a second-order loop filter is used in this work.

calibration time. However, the extra calibration time can be further reduced by transforming the SAR controller into a counter once the SAR controller has completed.

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