LETTER

An ultra-wideband injection-locked frequency divider with a multi-order resonator and the passive injection-boosted technique

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Abstract This paper presents an ultra-wideband injection-locked frequency divider (ILFD) design which is applicable to a multi-applications system. A transformer-based multi-order resonator is utilized in the ILFD to attain a flat phase response. The passive injection-boosted technique is proposed to increase the injection signal amplitude and thus widen the locking range. With the injection-boosted technique and the multi-order resonator, the proposed ILFD demonstrates a measured locking range of 50 GHz, covering 48 GHz to 98 GHz. The divider designed in a 40-nm process consumes a DC power of 11.2 mW excluding buffers and occupies a core size of 0.38 × 0.63 mm².

key words: injection-locked frequency divider, ultra-wideband, passive injection-boosted technique, multi-order resonator, locking range

Classification: Microwave and millimeter wave devices, circuits, and hardware

1. Introduction

The applications in the millimeter-wave regime commonly require that the RF front-end works in a wide band [1, 2, 3, 4, 5, 6, 7, 8]. For example, the IEEE 802.11ad standard covers from 57 GHz to 63 GHz, and the bandwidth requirement of the automotive radars in the 77-GHz band is 4 GHz. The frequency synthesizer is one of the most critical blocks in a wireless system since it, meanwhile, serves the Rx and Tx chain. A phase-locked loop (PLL) based frequency synthesizer diagram is presented in Fig. 1(a). Usually, the voltage-controlled oscillator (VCO) and the millimeter-wave frequency divider need to work in high frequencies and a wide band. Recently, some wideband VCOs [9, 10, 11] and multi-mode VCOs [12, 13, 14, 15, 16, 17] have been demonstrated with ultra-wide tuning bands. For example, [9] exhibits a wideband VCO with a 39.5% relative tuning range. The VCO in [10] realizes a tuning range of 42 GHz. The multi-mode VCO in [12] demonstrates a 32.6-GHz bandwidth. The above results imply the possibility of a multi-band system with a single frequency synthesizer. The frequency synthesizer in Fig. 1(b) using the dual-mode VCO is possible to cover more than one application in a stand-alone system. Accordingly, the first stage divider, i.e., the mm-wave divider, is expected to realize an ultra-wideband locking range (LR) to cover the two tuning bands of the dual-mode VCO.

Fig. 1. The PLL architectures with the (a)single-mode VCO, (b) dual-mode VCO.

It is well known that the injection-locked frequency dividers (ILFDs) with the highest operation frequencies are the more suitable choice for the mm-wave divider compared with static dividers and miller dividers. Nevertheless, the traditional injection-locked divider has a narrow operation bandwidth. Much research [18, 19, 20, 21, 22, 23, 24, 25, 26, 27, 28] has been conducted to enhance the LR of the injection-locked frequency dividers and multipliers. [18] reduces the Q value of the tank to attain a flat phase response. [19] proposes the frequency-tracking technique achieving 39% fractional bandwidth. [20] using the current-synthesis technique realizes a LR from 26.1 GHz to 38.2 GHz. In [23], the design employing the transformer-based high-order resonator exhibits a LR from 32.4 to 61.9 GHz. [26, 27, 28] prove the effectiveness of the transformer cou-
pling resonator.

However, there are less discussion and methods about the injection boosting technique, and the presented LRs are not enough to cover multiple applications. In this paper, a passive injection-boosted technique is proposed, which widens the LR by 14 GHz. Besides, the proposed divider uses a transformer-based 4th-order resonator to get a flatter phase response. With the above techniques, our circuit has a measured LR from 48 GHz to 98 GHz, which makes it possible to cover multiple applications using one frequency synthesizer. We note that the earlier work is presented in a conference paper [29]. This paper adds detailed LR analysis of the 4th-order network and the simulated LR increment is demonstrated due to the passive injection-boosted technique. The experiment results of the 4th-order network and the simulated LR increment is demonstrated due to the passive injection-boosted technique.

As illustrated in Fig. 2, three current components i.e., the injection current, the tank current, and the oscillation current, contribute to the maximum phase deviation $\theta_{\text{max}}$ in the phase response. With the above techniques, our circuit has a measured LR from 48 GHz to 98 GHz, which makes it possible to cover multiple applications using one frequency synthesizer. We note that the earlier work is presented in a conference paper [29]. This paper adds detailed LR analysis of the 4th-order network and the simulated LR increment is demonstrated due to the passive injection-boosted technique. The experiment results of the 4th-order network and the simulated LR increment is demonstrated due to the passive injection-boosted technique. The experiment results and discussions are demonstrated in section 3. Finally, the conclusions are presented in section 4.

2. Circuit Design

The classic injection-locked frequency divider is shown in Fig. 2. The principles of the injection-locked technique are illustrated in [30]. It is clear that the LR is determined by the maximum phase deviation $\theta_{\text{max}}$ and the phase response. As illustrated in Fig. 2, three current components i.e., the injection current, the tank current, and the oscillation current, follow the vector superposition principle. Therefore, the $\theta_{\text{max}}$ between $I_1$ and $I_{\text{osc}}$ is attained when it is perpendicular to $I_{\text{inj}}$. Therefore, $\theta_{\text{max}}$ is derived in Eq. (1). A larger $\theta_{\text{max}}$ and a flatter phase response would contribute to a larger LR.

$$\theta_{\text{max}} \approx \arcsin \left( \frac{I_{\text{inj}}}{I_{\text{osc}}} \right)$$  (1)

![Fig. 2. The classic ILFD structure and principles.](image)

The proposed ILFD circuit is exhibited in Fig. 3. $C_{\text{gs, inj}}$ represents the parasitic capacitance between the gate and the source of the injection MOS transistor, while $C_{\text{gd, inj}}$ is the parasitic capacitance between the gate and drain. At first, we study the key parameters in the transformer-based 4th-order resonator determining the LR. Fig. 4 shows the model of the resonator and the corresponding phase response. A general 4th-order resonator has three zero-crossing points in the phase response trace resulting in larger LR than the 2nd-order LC resonator. The three zero-crossing points correspond to the pole frequencies of the tank impedance $Z_{11}$. Here, $L_1$ and $L_2$ denote the primary inductor and the secondary inductor, respectively. $C_1$ represents the parasitic capacitance of the cross-coupled MOS transistors. $C_2$ includes the parasitic capacitance of the buffer and a standalone capacitor. $C_x, C_c$ is the total capacitance of the coupling capacitance between two windings and the parasitic capacitance between the gate and drain of $M_2$. The coupling factor of the transformer is represented by $k$. To simplify the analysis, it is assumed that $L_1 = L_2 = L$ and $C_1 = C_2 = C$.

![Fig. 3. The proposed ILFD schematic.](image)

Apply the basic Kirchhoff principle to the network in Fig. 4, we get the Eq. (2)-Eq. (6), where the complex angular frequency $s$ equals $j\omega$.

$$V_1 = i_p s L + i_s s M = i_p s L + i_s s k L$$  (2)
$$V_2 = i_s s L + i_p s M = i_s s L + i_p s k L$$  (3)
$$i_c = (V_1 - V_2) s C_c$$  (4)
$$V_1 = \left( i_1 - i_c - i_p \right) \frac{R_1}{1 + s C R_1}$$
$$= \left[ i_1 - (V_1 - V_2) s C_c - i_p \right] \frac{R_1}{1 + s C R_1}$$  (5)
$$V_2 = (i_c - i_s) \frac{R_2}{1 + s C R_2}$$
$$= \left[ (V_1 - V_2) s C_c - i_s \right] \frac{R_2}{1 + s C R_2}$$  (6)

Through the above equations, we get the input impedance of the 4th-order resonator $Z_{11}$ shown in Eq. (7). The zero-crossing point represents that $Z_{11}$ is a resistance at the corresponding frequency. The numerator and denominator both include imaginary parts. So, the ratio of the
imaginary part and the real part in the numerator equals that in the denominator at the pole frequencies. Then we get Eq. (8).

In Eq. (8), there is a special condition in which the numerator is zero, as shown in Eq. (9). Resolving Eq. (9), we can deduce one pole frequency expressed in Eq. (10).

\[
1 - \left(1-k^2\right) \omega^2 L (C+C_c) = 0
\]

\[
\omega_1^2 = \omega_0^2 \frac{C}{(C+C_c)(1-k^2)}
\]

From Eq. (8), we also derive Eq. (11). Compared with the second item, the third item is so smaller that it can be ignored, as shown in Eq. (12). Accordingly, the other two pole frequencies are derived as Eq. (13) and Eq. (14).

\[
2 \omega^2 L \left[(1-k)C_c + C\right] \gg \frac{(R_1 + R_2)(1-k^2)}{R_1 R_2^2} \omega^2 L^2
\]

\[
\omega_2^2 \approx \frac{\omega_0^2}{1+k}
\]

\[
\omega_3^2 \approx \frac{\omega_0^2 C}{1-k C + 2C_c}
\]

The LR has a direct relationship with the lowest and highest pole frequency, \(\omega_2\) and \(\omega_3\). With the phase of the resonator smaller than \(\theta_{\text{max}}\), a larger \(k\) contributes to a larger LR. On the other hand, the parasitic \(C_c\) makes \(\omega_3\) smaller and keeps \(\omega_2\) constant, which deteriorates the LR. Some simulations are conducted to get the results demonstrated in Fig. 5 and Fig. 6. The presented influence of \(k\) and \(C_c\) accords with the above derivations.
\[ Z_{11} = \frac{V_1}{i_1} = \frac{s L k R_1 \left[ R_2 + R_2 \left( 1 - k^2 \right) s L \left( s C + s C_c \right) + \left( 1 - k^2 \right) s L \right]}{s L k \left( R_1 + R_2 \right) \left[ 1 + s L \left( 1 - k^2 \right) \left( s C + s C_c \right) \right] + k R_1 R_2 \left\{ s^2 L^2 s C \left( s C + s C_c \right) + 2 s L \left[ \left( 1 - k \right) s C_c + s C \right] + 1 \right\} } \]  
\[ \frac{R_2 \left[ 1 - (1 - k^2) \omega^2 L \left( C + C_c \right) \right]}{(1 - k^2) \omega L} = \frac{(R_1 + R_2) \omega L \left[ 1 - (1 - k^2) \omega^2 L \left( C + C_c \right) \right]}{R_1 R_2 \left\{ \omega^4 L^2 \left( C^2 + C C_c \right) - 2 \omega^2 L \left[ \left( 1 - k \right) C_c + C \right] - \frac{(R_1 + R_2) \left( 1 - k^2 \right)}{R_1 R_2^2} \omega^2 L^2 \right\} + 1 = 0 \]  

Fig. 5. The influence of \( k \) on the phase response.  

Fig. 6. The influence of \( C_c \) on the phase response.  

Fig. 7. The amplitude gain simulation configuration.  

when the frequency is larger than 98 GHz. We also get the phase noise difference between the input and output signal, which is demonstrated in Fig. 11, which accords with the theoretical value, 6 dB. The ILFD consumes a DC power of 11.2 mW, excluding the buffers. The supply \( V_{DD1} \) providing a DC current of 16 mA is 0.7 V while \( V_{DD2} \) equals 1.1 V. The performance comparison with other ILFD designs is summarized in Table I. This work realizes the largest absolute LR. FoM [23] is defined in Eq. (21).

\[ \text{FoM(GHz/mW)} = \frac{\text{Locking Range(GHz)}}{\text{DC Power(mW)}} \]  

Fig. 8. The simulated amplitude gain and locking ranges.  

Fig. 9. The chip photograph of the proposed ILFD.  

4. Conclusions  

This paper proposes an ultra-wideband ILFD using the pas-
Table I. Performance comparison with other ILFD designs

| Reference | [19] | [20] | [23] | [25] | [26] | [27] | This Work |
|-----------|------|------|------|------|------|------|-----------|
| Technology(nm) | 65 | 55 | 65 | 65 | 40 | 90 | 40 |
| Operation Frequency(GHz) | 53.4-79.4 | 26.1-38.2 | 27.9-53.5 | 25-53.6 | 86-104 | 75.1-97 | 48-98 |
| Locking Range(GHz) | 26 | 12.1 | 25.6 | 28.6 | 18 | 21.9 | 50 |
| Fractional Bandwidth(%) | 39.2 | 37.6 | 62.9 | 72.8 | 19 | 25.4 | 68 |
| DC Power(mW) | 2.9 | 5.6 | 5.8 | 6.7 | 7.7 | 2.45 | 11.2 |
| FoM(GHz/mW) | 8.97 | 2.16 | 4.41 | 4.27 | 2.34 | 8.94 | 4.46 |

* Simulation results.

Fig. 10. The simulated and measured input sensitivity of the designed ILFD.

Fig. 11. The phase noise difference between the input and output signal.

The passive injection-boosted technique and the high-order resonator. First, the pole frequencies of the transformer-based 4th-order resonator are analyzed in detail. Then, the passive injection-boosted technique is demonstrated to offer a significant injection amplitude gain, which enhances the LR by about 14 GHz. The measurement results demonstrate a LR of 50 GHz, from 48 GHz to 98 GHz. This work makes it possible that one frequency synthesizer covers multiple applications with the proposed divider design.

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