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A 60GHz VCO with 25.8% Tuning Range by Switching Return-Path in 65nm CMOS

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Abstract— This paper presents a novel inductive tuning method for 60GHz voltage controlled oscillator (VCO) design. A new inductor-loaded transformer is proposed by configuring different current return-paths in the secondary coil of a transformer. Different from published inductive tuning methods, the proposed topology can achieve a multi-band, wide tuning range within compact area using only one transformer. A 60GHz VCO is implemented at 65nm CMOS process for demonstration. The obtained oscillation frequency can vary from 51.9GHz to 67.3GHz, which covers the full 60GHz band and provides a wide frequency tuning range of 25.8%.

I. INTRODUCTION

Phase noise and tuning range are the two primary design targets for VCO designs. For 60GHz wireless communication systems, the unlicensed 7GHz band with 4 sub-bands calls for a wide tuning range VCO with full-band coverage requirement. A single varactor cannot cover such a wide range with good phase noise performance. With further consideration of large process-voltage-temperature (PVT) variations in 65nm process, a frequency tuning range (FTR) wider than 20% is typically required [1].

Multi-band operation is normally adopted to enhance the total FTR with reduced VCO gain (K_{VCO}), which can improve phase noise performance [2] and facilitate phase locked loop (PLL) design for wide frequency band. Conventionally, multi-band operation is implemented with capacitive tuning by switched capacitor banks. When frequency scales up to 60GHz, the large parasitic capacitance from the capacitor bank becomes too large and the quality factor of capacitor decreases significantly [2, 3]. Recently, inductive tuning has become a promising substitute to replace the capacitive tuning, and is normally implemented by a loaded transformer [2-4]. Besides a wide FTR, inductive tuning can also provide the benefit of isolated DC noise from the tuning element.

The load on transformer for inductive tuning can be roughly categorized into 3 types: resistor [2], capacitor [4], and inductor [3]. Wide tuning range is then achieved by controlling the value of the load. For example, resistor-loaded transformer implements the variable resistance with a transistor in triode region [2]. By tuning the bias voltage of the transistor, a wide FTR can be achieved. However, the

generated tuning-curve is very nonlinear. For a wide FTR, the effective K_{VCO} can become very large at certain biasing points, which can make PLL difficult to lock. In addition, the VCO performance degrades when the resistance is tuned away from its two boundary values [2]. Capacitor-loaded transformer, similarly, implements the variable capacitance with a varactor. It suffers from a narrow FTR due to the limited tuning range of the varactor. Inductor-loaded transformer implements a variable inductance with switches. Multi-band operation is achieved by switching on various combinations of inductors [3]. Fine-tuning at each sub-band is then achieved with another small varactor. Due to the need of multiple transformers working as coupled inductors, the number of sub-bands that can be achieved is often limited to below 4, which is constrained by layout size and complexity.

To overcome the above limitations for inductive tuning, a new inductor-loaded transformer is proposed in this paper. Instead of using multiple transformers, only one transformer is employed with switches at various locations on the transformer. By turning on different combinations of switches, different current return-paths are formed in the second coil of the transformer, leading to variable inductance. Compared with conventional inductive tuning methods, the proposed structure shows several advantages. It realizes a much lower K_{VCO} than the resistor-loaded transformer, a much wider FTR than the capacitor-loaded transformer, and achieves more sub-bands with more compact size than conventional inductor-loaded transformer. What is more, since transistors are turned either fully on or fully off as switches in the proposed structure, its performance degradation on VCO is relaxed.

The rest of paper is organized below. Section II shows a circuit analysis for loaded-transformer based inductive tuning, upon which a new inductive tuning structure by switching return-paths is proposed in Section III. A 60GHz VCO is then designed in Section IV to leverage the proposed new inductive tuning method. Experiment results are presented in Section V and the paper is concluded in Section VI.

II. LOADED TRANSFORMER FOR INDUCTIVE TUNING

The mechanism of loaded transformers applied for inductive tuning can be explained by Fig. 1. Since transistor (for resistor-loaded), varactor (for capacitor-loaded), and

Figure 1. Equivalent circuit model for transformer loaded inductive tuning.

switch (for inductor-loaded) can all be equalized to a RC tank, the 3 types of loaded transformers can be analyzed with the same equivalent circuit.

A. Tuning of Oscillation Frequency

The transformer is assumed to be ideal with coupling factor k , and with L_1 and L_2 as the primary inductance and secondary inductance, respectively. The equivalent circuit with L_{eq} and R_{eq} can then be calculated as

$$\begin{cases} L_{eq} = \frac{R^2 L_1 [1 - \omega^2 C L_2 (1 - k^2)]^2 + \omega^2 L_1 L_2^2 (1 - k^2)^2}{R^2 (1 - \omega^2 C L_2) [1 - \omega^2 C L_2 (1 - k^2)] + \omega^2 L_2^2 (1 - k^2)} \\ R_{eq} = \frac{R^2 L_1 [1 - \omega^2 C L_2 (1 - k^2)]^2 + \omega^2 L_1 L_2^2 (1 - k^2)^2}{R k^2 L_2} \end{cases} \quad (1)$$

By defining the total capacitance in the oscillation LC-tank as C_t , the oscillation frequency becomes

$$\omega = \frac{1}{\sqrt{L_{eq} C_t}} \quad (2)$$

B. Effect of Coupling Factor k on FTR

For resistor or inductor-loaded transformer, the tuning range of the equivalent circuit depends on the equivalent inductance, and can be approximated as

$$\begin{cases} L_{eq_max} = L_{eq}(R \rightarrow \infty) = L_1 \times \frac{1 - \omega^2 C L_2 (1 - k^2)}{1 - \omega^2 C L_2} > L_1 \\ L_{eq_min} = L_{eq}(R \rightarrow 0) = L_1 (1 - k^2) \end{cases} \quad (3)$$

where a small C is assumed such that $\omega^2 C L_2 < 1$.

Assume $\omega^2 C L_2 \ll 1$, and L_{eq} has the tuning range from $L_1(1 - k^2)$ to L_1 , FTR can be calculated by

$$FTR = \frac{\frac{1}{\sqrt{1 - k^2}} - 1}{\frac{1}{\sqrt{1 - k^2}} + 1} \times 2 = \frac{1 - \sqrt{1 - k^2}}{1 + \sqrt{1 - k^2}} \times 2. \quad (4)$$

According to (4), to achieve a large FTR, a large k is required.

C. Implementation of Large Equivalent Resistance R_{eq}

As shown in Fig. 1, R_{eq} is in parallel to L_{eq} , thus should be designed as large as possible to minimize the loss and phase noise. According to (1), R_{eq} approaches infinity when R approaches 0 or infinity, but drops when R moves from the two boundaries. This can explain the performance degradation for resistor-loaded transformer.

III. INDUCTOR-LOADED TRANSFORMER BY SWITCHING RETURN-PATH

A. Concept

According to Fig. 1 and (1), there are 4 variables: R , C , L_2 , and k on the secondary side of the transformer, which control the oscillation frequency. Resistor-loaded transformer tunes R , capacitor-loaded transformer tunes C , while inductor-loaded transformer tunes both L_2 and k (Fig. 2(a)).

Figure 2. Proposed new inductor-loaded transformer by switching return-paths with one transformer: (a) concept, (b) layout implementation.

Conventionally, L_2 and k are tuned by switching on different combinations of transformers [3]. However, the large layout size of loop inductor and strong magnetic coupling with adjacent devices limit the number of transformers. Moreover, as more transformers are used, magnetic coupling from different transformers tend to cancel each other, and hence make the tuning less effective. As a result, the number of sub-bands achieved with conventional inductor-loaded transformer topologies often limits to 4 (with 2 transformers used).

In this paper, a new inductor-loaded transformer topology is proposed which breaks through the limit of the conventional inductive tuning. The concept of the proposed topology can be explained in Fig. 2(a). Only one transformer is used for the new inductor-loaded transformer, with switches placed at various locations on the secondary coil. When all switches are off, the structure works similarly to the resistor-loaded transformer with the maximum resistor value applied ($R \rightarrow \infty$). When some switches are turned on and a close-loop is constructed in the secondary coil, a current return-path forms. Different sized return-paths generate different magnetic fluxes, which are fed back to the primary coil. In this case, the structure works similarly to a resistor-loaded transformer with the minimum resistor value applied ($R \rightarrow 0$), but with tuned L_2 and k values defined by the size of the according return-path.

B. Comparison

Since the proposed topology can increase the number of sub-bands by simply adding more switches, the sub-band number can be easily designed to be larger than 4 with small layout overhead. As a result, the proposed topology can achieve more sub-bands within more compact size than the conventional inductor-loaded transformer.

Note that the lowest oscillation frequency occurs when all switches are off, and the highest frequency occurs when the largest return-path is formed. According to (3) and (4), with the same parameters provided for the transformer, the proposed topology achieves the same tuning range as resistor-loaded transformer. However, different from the resistor-loaded transformer which has a highly nonlinear tuning curve and large K_{VCO} , the new inductor-loaded transformer achieves a much smaller K_{VCO} through multi-band operation. As a result, the phase noise performance can be improved with no PLL locking difficulty. The small K_{VCO} may also be used to trade for a wider tuning range, which can be easily realized by

implementing a larger coupling factor, as explained in (4). What is more, with a small varactor implemented for fine tuning in each band, the total FTR is further increased. As a result, the proposed structure can also achieve a wider FTR than the resistor-loaded transformer. On the other hand, as switches are either fully turned on ($R \rightarrow 0$) or fully turned off ($R \rightarrow \infty$), R_{eq} in the proposed structure will not be degraded as the case for the resistor-loaded transformer.

In addition, the total parasitic capacitance from both switches and transformer forms C in Fig. 1. According to (3), a large C forces L_{eq_max} to approach L_{eq_min} , thus degrading the tuning range. However, this degradation can be minimized when keeping $\omega^2 CL_2 < 1$ in the proposed structure. In contrast, for capacitor-loaded transformer, the parasitic capacitance adds directly to the varactor capacitance, thus reducing the FTR significantly. As a result, the proposed structure achieves a much wider FTR than the capacitor-loaded transformer.

In conclusion, proposed new inductor-loaded transformer shows advantages over traditional inductive tuning methods. It realizes a much lower K_{VCO} than resistor-loaded transformer, obtains a much wider FTR than capacitor-loaded transformer, and can achieve more sub-bands with more compact size than the conventional inductor-loaded transformer.

C. Layout Implementation

Figure 3. Schematic of the 60GHz inductive-tuning VCO with the new inductor-loaded transformer implemented by switching return-paths.

One layout implementation is shown in Fig. 2(b). A transformer is loaded with 4 switches (S1~S4) at different locations. The inner loop is the primary coil, which serves as the inductor for LC-tank. The outer loop is the secondary coil, and is loaded with switches to control the return-path. Lengths of the 4 sections in secondary coil are marked as $a \sim d$. Different combinations of switches and corresponding effective lengths of return paths are summarized in Table I. For example, by turning on switches S1 and S2, selection mode 3 is invoked with a return-path formed by “ $a + b$ ”.

There are 7 bands selected from Table I with better phase noise performance. By adjusting section lengths to be

TABLE I. EFFECTIVE RETURN PATH LENGTHS IN SECONDARY COIL FOR DIFFERENT SUB-BAND SELECTION MODES

Sub-band Selection Mode	0	1	2	X	X	3	4	5	6
On Switches	Nil	S1	S2	S3	S4	S1+S2	S1+S3	S2+S4	S3+S4
Effective Length of Return Path	0	a	b	b+c	a+d	a+b	a+b+c	a+b+d	a+b+c+d

“ $b=d=2a=2c$ ”, the effective length of return-path in secondary coil varies from “0” to “ $6a$ ” linearly, resulting in evenly distributed frequency bands. Evenly distributed bands can help reduce the varactor size required and facilitates PLL design.

Note that more sub-bands can be realized by implementing more switches but may also degrade the phase noise. In addition, to achieve a high R_{eq} , a small R is desired according to (1). Therefore, the number of switches connected in serial in the active return-path should be minimized. The proposed structure limits this number to 2 switches.

IV. VCO CIRCUIT REALIZATION

To verify the function of the proposed inductive tuning, a VCO is designed at 60GHz. The new inductor-loaded transformer provides multiple frequency sub-bands to cover the wide frequency band at 60GHz. As shown in Fig.3, power supply is fed on its central tap. A varactor-pair is used for fine tuning within each sub-band. Its capacitance and the parasitic capacitance form C_i in (2). The LC-tank loss is compensated by a cross-coupled NMOS pair. Two output buffers are used for the power gain and isolation.

The transformer is implemented with top metal layer for higher Q . According to (4), k limits the theoretical maximum FTR that can be achieved. To implement a large k , a small gap of 3.5um is designed between transformer primary and secondary coils and hence ensures a strong magnetic coupling. In addition, to achieve a large R_{eq} in (1), a large size should be implemented for each switch transistor. However, the parasitic capacitance from switches (C) would also be increased proportionally. According to (3), a large C forces L_{eq_max} to approach L_{eq_min} , thus degrading the tuning range. A size of 50um/60nm is adopted for switch transistors in final design.

V. IC FABRICATION AND MEASUREMENT

The designed 60GHz VCO is implemented in STM 65nm 1P7M CMOS process, with die photo shown in Fig. 4. Decoupling capacitors are implemented by MIM capacitors

Figure 4. VCO die photo, fabricated in STM 65nm CMOS process. and used to stabilize DC signals. The total area is

852×451μm², which is mainly constrained by PADs. The core area takes only 163 ×190 μm². The circuit is designed and verified by EM simulation (ADS-Momentum) before the fabrication. It is measured on CASCADE Microtech Elite-300 probe station, with Agilent PNA-X spectrum analyzer, E5052 source signal analyzer, and 11970V harmonic mixer. A 67GHz bias-T is used to provide load to buffer. Low output power was observed due to high loss from cable, bias-T, and harmonic mixer. Two IF amplifiers are connected after harmonic mixer to boost the output power.

With VDD equals to 1V and Vtune varies from 0.5 to 1.5V, the obtained tuning curves under different sub-band selection modes are plotted in Fig. 5 (a). The entire tuning range is divided into 7 sub-bands. Within each sub-band, a tuning range of 2.5~4.5 GHz is achieved by a small varactor. Evenly distributed sub-bands are preferred for the easy PLL implementation, which can be achieved by adjusting locations of switches. The obtained oscillation frequency varies from 51.9 to 67.3GHz, which covers the whole 60GHz band and provides a FTR of 25.8%. Note the tuning voltage (0.5~1.5V) is selected to provide maximum tuning range for varactor, and can be easily changed to 0~1V by adding a serial capacitor between varactor and power supply.

A sample of phase noise plot is shown in Fig. 6. At 60GHz, the phase noise is -106.7dBc/Hz at 10MHz offset. The

Figure 5. Measured performance. (a) Tuning curves under different band selection modes. (b) VCO phase noise performance for all selection modes.

Figure 6. Measured VCO phase noise at 60GHz.

TABLE II. PERFORMANCE COMPARISON OF 60GHz CMOS VCO

Ref.	Tech (CMOS)	Freq (GHz)	FTR (%)	Phase Noise (dBc/Hz)	Power (mw)	FOM (dBc/Hz)	FOMt (dBc/Hz)
[5]	130nm	55	11	-107@10MHz	10	-172	-172.8
[6]	90nm	58	9.3	-111@10MHz	8.1	-177.2	-176.6
[7]	90nm	72	2.8	-112.2@10MHz	19.2	-176.5	-165.4
[2]	90nm	57	14.3	-118.8@10MHz	8.7	-184.3	-187.4
This Work	65nm	60	25.8	-106.7@10MHz	5.4	-174.9	-183.1

measured phase noise performance for all modes is shown in Fig. 5 (b). Due to the low output power, there is a deviation in phase noise from simulation. In addition, one can observe degraded phase noise in modes with asymmetric return-paths.

A comparison with previously published 60GHz VCOs is summarized in Table II. The proposed VCO is able to achieve the widest tuning range (25.8%) with the moderate figure-of-merits (FOM and FOMt) despite of phase noise degradation.

VI. CONCLUSION

A new inductor-loaded transformer based inductive tuning is proposed in this paper for wide tuning range of 60GHz VCO. Different from previously published inductive tuning methods, by configuring different current return-paths in the secondary coil of one transformer, a wide range multi-band tuning can be achieved within compact area using only one transformer. One demonstrated 60GHz VCO shows the measured 25.8% tuning range using 65nm CMOS process.

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