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A Carrier Aggregation Transmitter Front End for 5-GHz WLAN 802.11ax Application in 40-nm CMOS

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Abstract—Carrier aggregation will be a key feature of the emerging 802.11ax system to boost data rate. Voltage-controlled oscillator (VCO) pulling and crosstalk between RF channels are two main problems related to carrier aggregation, which introduce severe impacts on adjacent channel leakage ratio (ACLR) and error vector magnitude (EVM). To address these two problems, high isolation between transmitter channels is required, but may not be realizable in some practical cases due to the constraint of the layout. This article presents a transmitter front end that can support two-carrier aggregation for 802.11ax without high on-chip isolation. The proposed transmitter front end employs a parallel direct conversion and double conversion configuration, and two carriers are combined before the power amplifier (PA). In this way, the intrinsic problems of VCO pulling and crosstalk can be mitigated. The proposed transmitter front end was implemented in a 40-nm CMOS technology. For contiguous intra-band carrier aggregation with VHT80, MCS9 signals [80-MHz bandwidth, 256-quadrature amplitude modulation (QAM), and 11.25-dB peak-to-average power ratio (PAPR)], the transmitter front end delivers an average output power of $5.3 + 4.8$ dBm for two carriers with an EVM of ≤ -32 dB, and the EVM can reach -36.1 dB.

Index Terms—Carrier aggregation, CMOS transmitter, crosstalk, 802.11ax, error vector magnitude (EVM), signal-to-noise ratio (SNR), voltage-controlled oscillator (VCO) pulling, wireless local area network (WLAN).

I. INTRODUCTION

RECENTLY, the accelerating growth of smartphones and mobile devices is propelling wireless communication to evolve toward higher data rate. Currently, as the main standard for broadband wireless local area network (WLAN)

Fig. 1. Carrier aggregation scenarios.

in the commercial market, the IEEE 802.11ac system cannot satisfy the ever-increasing demand on higher data rate. The development of new IEEE 802.11 standards is drawing much effort in academia and industry. As the successor to the IEEE 802.11ac, the emerging IEEE 802.11ax will increase the efficiency of WLAN networks, especially in high-density WLAN deployment scenarios, and the data rate is expected to reach 10 Gb/s. Carrier aggregation originating from LTE-A is a key feature of the IEEE 802.11ax standard that combines multiple channels to support a higher data rate. There are three modes of carrier aggregation, as shown in Fig. 1. The first mode is contiguous intraband carrier aggregation in which neighboring channels within one frequency band are used to transmit signal. The second mode is noncontiguous intraband carrier aggregation, and, in this mode, channels with frequency separation within one frequency band are combined. The third mode is interband carrier aggregation, in which signals are transmitted in channels located in different frequency bands.

Some transmitter architectures have been introduced and discussed in [1] and [2]. The main difference between these proposed architectures lies in where the carriers are combined in the transmitter. To support contiguous carrier aggregation, a single transmitter chain with all carriers aggregated at digital baseband is feasible. Since the aggregated signals are processed in only one RF transmitter chain after generated

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Fig. 2. (a) Transmitter architecture in [6]. (b) New transmitter architecture without using an external 4-GHz reference signal.

from a digital-to-analog convertor (DAC), this architecture is the least complex architecture in terms of the number of RF blocks required. In practice, this architecture can further support non-contiguous carrier aggregation with small frequency spacing. However, in the scenario where the frequency spacing between carriers is quite large, the sampling rate of DAC has to be extremely high accordingly, causing a considerable power consumption and intensifying the complexity of DAC. In [3] and [4], the sampling rates of DAC are as high as 2.4 and 1.14 GS/s, respectively.

To support interband carrier aggregation, the architecture with carriers combined in the RF part can be used. However, the conventional architecture for carrier aggregation in RF part has an inherent problem of crosstalk and voltage-controlled oscillator (VCO) pulling caused by the interaction between the transmitter chains in RFIC implementation, which has been briefly introduced in [5] and [6]. To increase the isolation and reduce the interaction between the transmitter chains, a sufficiently large physical separation between transmitter chains is required. However, in practice, a large physical separation between transmitter chains may not be achieved due to the constraint of the layout.

To address this issue, we have proposed a transmitter architecture using the parallel direct-conversion and double-conversion configuration in [6], as shown in Fig. 2(a). This transmitter architecture is the first attempt to simultaneously solve the problems of crosstalk and VCO pulling without a large physical isolation requirement. In this architecture, five local oscillator (LO) signals are used to drive the mixers. To generate these five LO signals, three internal phase-locked loops (PLL) and one more 4-GHz external reference signal were adopted, which increase the design complexity and the power consumption. In addition, in practice, this 4-GHz reference signal with such a high frequency cannot be provided

by a crystal oscillator; thus, one more PLL is required, which further increases the design complexity. In order to address these problems, we propose a new architecture, as shown in Fig. 2(b). In the new architecture, only three PLLs are needed without the 4-GHz reference, which reduces the power consumption and the design complexity. In this article, the proposed transmitter front end employs this new architecture. Since 802.11ax will employ two-carrier aggregation, only two paths (Path1 and Path2) are used in the proposed transmitter front end. In addition, the transmitter in [6] was measured by using 802.11ac signal, and the error vector magnitude (EVM) performance is poor. While, in this article, we redesign circuit blocks, mainly including the mixers and power amplifier (PA), and the EVM performance has been improved when the transmitter was tested with practical 802.11ax signals. In addition, we make more analysis on crosstalk and VCO pulling, and the mitigation of crosstalk is analyzed in detail in this article.

In this article, we first study how the signal-to-noise ratio (SNR) affects EVM for WLAN signals, and then analyze the effects of crosstalk and VCO pulling in Section II. This section proves that the EVM for 802.11ax signal is vulnerable to the effect of crosstalk, and adjacent channel leakage ratio (ACLR) may degrade due to VCO pulling for contiguous carrier aggregation. In Section III, the proposed transmitter front end is introduced, and the mechanism of crosstalk and VCO pulling mitigation is investigated. The design of the transmitter front end is described in Section IV, and the measurement results are presented in Section V. Section VI concludes this article.

II. CROSSTALK AND VCO COUPLING

A. SNR Degradation Due To Crosstalk Caused By Baseband Coupling And LO Coupling

When more carrier signals are concurrently processed in one CMOS chip, due to the low-resistance substrate and the electromagnetic radiation, coupling or leakage between different RF signal processing paths is unavoidable, leading to two main problems, namely crosstalk and VCO pulling. Crosstalk is mainly caused by the baseband coupling and the LO coupling [5]. Fig. 3(a) demonstrates the effect of baseband coupling on the transmitted signals. In Fig. 3(a), two RF paths use a direct-conversion mixer, and two carriers are combined after the mixer. The coupling signal from BB1 falls on top of BB2 and vice versa. As a result, before injected into the mixer, both the signals BB1 and BB2 are interfered by the coupling signal from the other path. Consequently, the SNR of each baseband signal is degraded. Fig. 3(b) demonstrates the effect of crosstalk caused by the LO coupling on the transmitted signals, and the LO coupling has the similar impact on the SNR of the RF signals as the baseband coupling does. In Fig. 3, the bandwidth of the signal BB2 is assumed to be smaller than that of the signal BB1; therefore, for the signal BB2, not only the SNR but also the ACLR is degraded. However, in the practical carrier aggregation scenario, all carriers have the same bandwidth; thus, in this article, the degradation of the ACLR caused by the baseband coupling or the LO coupling is

Fig. 3. Effect of crosstalk on SNR of RF transmitted signals. (a) Baseband coupling. (b) LO coupling.

only demonstrated in figures, but not considered in the further analysis.

The degradation of the SNR mainly affects the EVM performance, which is the most widely adopted standard in the industry to quantify the performance of the signal in the transmitter. Regardless of the nonlinearity of transmitter and phase noise of LO, EVM is a function of SNR [7], expressed by

$$\text{EVM} = 100\% \cdot \sqrt{\frac{\sqrt{M} + 1}{3 \cdot \text{SNR} \cdot (\sqrt{M} - 1)}} \quad (1)$$

where M is the order of modulation. While, in practice, $\text{EVM}_{\text{rms,avg}}$ is measured through a vector signal analyzer. As an alternative definition of EVM, $\text{EVM}_{\text{rms,avg}}$ normalizes the error vector to the average symbol energy [8], only depending on SNR, expressed by

$$\text{EVM}_{\text{rms,avg}} = 100\% \cdot \sqrt{\frac{1}{\text{SNR}}} \quad (2)$$

In the following part of this article, EVM refers to $\text{EVM}_{\text{rms,avg}}$. In order to explore the impact of SNR on EVM for practical WLAN signals, a simulation has been performed to calculate EVM using 802.11ac signals with different SNRs. Since 802.11ac signal has the same modulation as the 802.11ax signal does, 802.11ac signal can be an alternative to 802.11ax signal when simulating constellation and EVM. In the simulation, 80-MHz 802.11ac signals with 64-quadrature amplitude modulation (QAM) and 256-QAM modulation are used. The simulated results are illustrated in Fig. 4. In Fig. 4, two EVM curves for 64-QAM and 256-QAM almost overlap, showing that EVM ($\text{EVM}_{\text{rms,avg}}$) is independent of the modulation. In addition, the simulated EVM for 802.11ac

Fig. 4. Simulated EVM for 802.11ac signals with different values of SNR.

signal is slightly higher than that predicted from (2) but has the same tendency with SNR. From Fig. 4, the 50-dB SNR results in the EVM of 0.42%. Considering the stringent EVM requirement of 2.5% (−32 dB) and 1.8% (−35 dB) for 256-QAM and 1024-QAM, the deterioration of the EVM by 0.42% has a severe impact on system performance, which will intensify the noise and linearity requirement for other blocks in the transmitter. Therefore, for the 802.11ax system employing 256-QAM and 1024-QAM modulations, an SNR of better than 50 dB is required. Hence, it deserves much effort to increase the isolation between signal paths to achieve a high SNR for the 802.11ax system operating in the carrier aggregation mode.

B. Coupling Factor in Layout

We have thus far analyzed the impact of SNR on EVM, and now turn to study how SNR is affected by the coupling in layout. In Fig. 3(a), the SNR of BB1 can be simply defined as follows:

$$\text{SNR} = 20 \log \frac{A_1}{kA_2} \quad (3)$$

where k is the coupling factor between signal paths. The coupling exists between the devices and the metal signal traces, respectively, and the coupling strength or isolation can be roughly estimated in the simulation.

First, the coupling factor between the two devices through the substrate has been simulated in the EM simulation software HFSS V15. Two cases are considered in the simulation: one case where there is no guard ring, and the other case where one active area is surrounded by a guard ring. The physical models used for the simulation in the TSMC 40-nm CMOS process are depicted in Fig. 5, and the simulated results are shown in Fig. 6, showing that the coupling factor is dependent on the area of the active device and the distance between the devices [9]. Comparing Fig. 6(b) and Fig. 6(a), the guard ring can provide 7–10 dB isolation. Although adding a guard ring can help to increase isolation, it cannot ensure sufficient high isolation in some practical cases.

Then, the coupling factor between the single-ended metal signal traces is simulated. The physical model used to simulate the coupling factor is depicted in Fig. 7(a), in which the signal traces use top-layer metal with a width of 2 μm and the thickness of 3.5 μm . In a practical layout, to reduce

Fig. 5. (a) and (b) Top view and (c) cross-sectional view of two active devices in the TSMC 40-nm CMOS process.

Fig. 6. Simulated coupling factor between the two active devices through the substrate (unit: micrometer). (a) No guard ring. (b) One device with a guard ring.

the parasitic capacitance as much as possible, no ground metal is placed under the metal signal traces in most cases. Hence, there is no ground under the simulated signal traces in the model shown in Fig. 7. Here, we only consider the

Fig. 7. Model to simulate the coupling factor. (a) Single ended. (b) Differential.

Fig. 8. Simulated coupling factor with different length and spacing for single-ended traces (unit: micrometer).

coupling between two simple signal traces that are mostly used in layout, without considering the coupling between T-lines or CPWs. Fig. 8 illustrates the simulated coupling factor with frequencies for different length L and spacing S . Obviously, a longer coupling path and smaller spacing produce a higher coupling factor. More importantly, the coupling factor is proportional to the frequency, since the electromagnetic coupling becomes strong and dominates at high frequency. While, in most cases, the signals are differential, instead of single-ended. Thus, we simulated the coupling factor between two differential traces, and the model used for simulation is shown in Fig. 7(b). The simulation results are illustrated in Fig. 9. Compared with the simulated results for the single-ended traces, the differential signal traces have a remarkable anti-interference ability. The reason is that the polarity of the two metal lines in one differential pair is opposite, their external electromagnetic fields can cancel each other out, and much less electromagnetic energy is leaked to the outside. Therefore, using differential signal traces can effectively minimize the interference between signal traces.

From Figs. 7 and 8, the coupling factor in the frequency range of <100 MHz is much smaller than the values at the

Fig. 9. Simulated coupling factor with different length and spacing for differential traces (unit: micrometer).

frequencies of >5 GHz. For the 80-MHz 802.11ax signal, since the maximum analog frequency of the baseband signal is only 40 MHz and the coupling factor is lower than -80 dB for the baseband within 40 MHz even with a long metal signal trace and small spacing between the traces, the baseband coupling has a negligible effect on SNR in comparison with the LO coupling. According to (2) and (3), the -80 -dB coupling factor only leads to the deterioration of EVM by 0.01%, which can be ignored in a practical measurement. Hence, the baseband coupling is not considered in the transmitter front-end design in this article.

To achieve good EVM performance for the 802.11ax application, the coupling factor is required to be lower than -50 dB [5], [10], and the physical separation of hundreds of micrometers is required based on the simulation results. Although increasing physical separation in layout can effectively reduce the coupling and improve the isolation [10], [11], if done without an elaborate layout, or in some cases where the large physical separation cannot be achieved, the coupling factor cannot be ensured to be kept below -50 dB.

C. Effect of VCO Pulling

In transmitter supporting multiple carriers, two or more PLLs are integrated on the same substrate and operate concurrently. Due to the limited isolation, the oscillators undergo a mutual interaction between each other through the substrate or electromagnetic coupling, resulting in unwanted sidebands or spurs at the oscillators' outputs [12]–[14]. These resulted sidebands or spurs are at an offset equal to the frequency spacing between the PLLs' outputs, as shown in Fig. 10. As a result, for the mixer in each path, there are three LO signals, one desired LO signal, and two interferences caused by VCO pulling, as illustrated in Fig. 11. The requirement of an emission mask may not be satisfied due to these two interferences. In addition, for contiguous carrier aggregation, ACLR will degrade, since the resulted interferences are adjacent to the desired carriers.

The sideband magnitude can be approximately predicted by [14]

$$M_{SB}(\Delta\omega) = \frac{k\omega_0}{2Q} \left| \frac{1}{s + \frac{I_{CP}}{2\pi} \frac{K_{VCO}}{N} H_{LPF}(s)} \right|_{s=j\Delta\omega} \quad (4)$$

Fig. 10. VCO pulling between PLLs.

Fig. 11. Effect of VCO mutual pulling.

where k is the coupling factor between the VCOs, Q is the LC tank's quality factor in the VCO, K_{VCO} is the gain of the VCO, I_{CP} is the magnitude of the charge-pump current, N is the division ratio of the divider, $H_{LPF}(s)$ is the transfer function of the loop filter, and $\Delta\omega$ is the frequency spacing between the VCOs. According to (4), increasing the quality factor Q of the LC tank can reduce the VCO pulling effect. However, in a bulk CMOS technique, the quality factor of the inductor cannot be very high. Thus, it is not feasible to alleviate VCO pulling mainly through increasing the quality factor of the VCO. In addition, reducing the coupling factor between two VCOs can effectively lower the unwanted sideband, which can be done by increasing the physical spacing between two VCOs, placing ground shielding around the LC tank, and using separated dc supply and ground for each VCO. Moreover, the sideband magnitude becomes lower with the increase of the frequency spacing between the VCOs.

III. PROPOSED TRANSMITTER FRONT END

The block diagram of the transmitter front end is shown in Fig. 12, including the LO generator, mixers, drivers, and PA. In this transmitter, the two carriers are combined after the driver and before injected into the PA. While in [5] and [6], the carriers are directly combined after the up-conversion mixer, and the output ports of the mixers are connected together, leading to no isolation between the mixers in different paths. In this design, the driver is added before the combination

Fig. 12. Proposed transmitter front-end architecture.

of carriers to increase the isolation between the mixers in two paths. Three LO signals (LO_1 , LO_{2A} , and LO_{2B}) are generated from two corresponding LO input signals (LO_{1_IN} and LO_{2_IN}) to feed the mixers.

A. Mitigation of VCO Pulling

According to (4), the effect of VCO pulling can be reduced by increasing the frequency spacing between VCOs. In [5], the division ratio of the divider in one path is the double of the division ratio of the divider in the other path. Consequently, one VCO operates at twice the operating frequency of the VCO in the other path. In this way, sufficient frequency spacing can be obtained when supporting contiguous carrier aggregation. Although the approach used in [5] to achieve large frequency spacing is feasible for LTE-A with an operating frequency of <3 GHz, it is not suitable for 802.11ax in a 5-GHz WLAN band of 5.17–5.835 GHz [15]. Such an approach will require a PLL with frequency as high as 23.34 GHz, which drastically increases the PLL's design complexity and power consumption.

To provide sufficient large VCOs' frequency spacing without the cost of PLL's design complexity, the parallel direct conversion and double conversion configuration have been proposed. In the proposed transmitter front end, since

$$\begin{aligned} f_{LO1} &= \frac{f_{LO1_IN}}{2} \\ f_{LO2} &= \frac{3f_{LO2_IN}}{2} \end{aligned} \quad (5)$$

the frequency of LO_{1_IN} is from 10.36 to 11.65 GHz, and the frequency LO_{2_IN} is from 3.4533 to 3.8833 GHz, respectively.

If two PLLs are integrated to provide LO_{1_IN} and LO_{2_IN} , the frequencies of the two VCOs will be 10.36–11.65 GHz and 3.4533–3.8833 GHz, respectively. Since the frequency ranges of these two VCOs are far apart from each other, the effect of VCO pulling is negligible if two PLLs are on chip. In this article, these two LO input signals are provided by external PLLs from signal sources instead of internal PLLs. However, it is undeniable that the proposed transmitter front end is immune to VCO pulling.

B. Mitigation of Crosstalk

To explicitly explain the mechanism of mitigation of crosstalk caused by LO coupling for the proposed transmitter front end, we first consider LO coupling from Path2 to Path1, as shown in Fig. 13(a). To facilitate the analysis, two baseband signals and three LO signals are assumed to be

$$\begin{aligned} BB_1 &= A_1 \cos(\omega_{BB1}t + \varphi_{BB1}) \\ BB_2 &= A_2 \cos(\omega_{BB2}t + \varphi_{BB2}) \\ LO_1 &= \cos(\omega_{LO1}t + \varphi_1) \\ LO_{2A} &= \cos(\omega_{LO2A}t + \varphi_{2A}) \\ LO_{2B} &= \cos(\omega_{LO2B}t + \varphi_{2B}). \end{aligned} \quad (6)$$

Then, the upconverted signal in Path1 can be written as follows:

$$\begin{aligned} x_{out1}(t) &= G_1 A_1 \cos((\omega_{BB1} + \omega_{LO1})t + \varphi_1 + \varphi_{BB1}) \\ &+ kG_1 A_1 \cos((\omega_{BB1} + \omega_{LO2A})t + \varphi_{2A} + \varphi_{BB1}) \\ &+ kG_1 A_1 \cos((\omega_{BB1} + \omega_{LO2B})t + \varphi_{2B} + \varphi_{BB1}) \\ &+ kG_1 A_1 \cos((\omega_{BB1} + \omega_{LO1} \pm \omega_{LO2A})t \\ &\quad + \varphi_1 \pm \varphi_{2A} + \varphi_{BB2}) \\ &+ kG_1 A_1 \cos((\omega_{BB1} + \omega_{LO1} \pm \omega_{LO2B})t \\ &\quad + \varphi_1 \pm \varphi_{2B} + \varphi_{BB2}) \\ &+ k^2 G_1 A_1 \cos((\omega_{BB1} + \omega_{LO2A} \pm \omega_{LO2B})t \\ &\quad + \varphi_{2A} \pm \varphi_{2B} + \varphi_{BB2}). \end{aligned}$$

Among $x_{out1}(t)$, the interferences caused by the unwanted LO signals' coupling from Path2 exist at frequencies: $f_{LO1} \pm f_{LO2A}$, $f_{LO1} \pm f_{LO2B}$, and $f_{LO2B} \pm f_{LO2A}$, respectively.

Since

$$\begin{cases} 5.17G \leq f_{LO1} \leq 5.835G \\ 1.723G \leq f_{LO2A} \leq 1.945G \\ 3.447G \leq f_{LO2B} \leq 3.89G \\ 6.893G \leq f_{LO1} + f_{LO2A} \leq 7.78G \\ 3.225G \leq f_{LO1} - f_{LO2A} \leq 4.112G \\ 8.617G \leq f_{LO1} + f_{LO2B} \leq 9.725G \\ 1.28G \leq f_{LO1} - f_{LO2B} \leq 2.388G. \end{cases}$$

Obviously, these interferences at $f_{LO1} \pm f_{LO2A}$ and $f_{LO1} \pm f_{LO2B}$ will fall into out-of-band, thus having no impact on the SNR of the transmitted signal in Path2. Moreover, these interferences will be further suppressed by the following resonant tank used in the driver and PA and can be easily filtered out by the external SAW filter. Among the interferences, one in-band interference at the frequency of $f_{LO2}(f_{LO2A} + f_{LO2B})$, which is the mixing product of LO_{2A} and LO_{2B} , will potentially affect the SNR of the transmitted signal in Path2. For this case, the SNR of the transmitted signal in Path2 can be defined as follows:

$$SNR_2 = 20 \log \frac{A_2}{k^2 A_1}. \quad (7)$$

Since k is <1 , k^2 becomes much smaller than 1, and SNR_2 becomes very large. Hence, LO coupling has a negligible impact on the SNR of the transmitted signal in Path2.

Fig. 13. Mitigation of crosstalk between Path1 and Path2 in the proposed transmitter architecture.

Now, let us consider the case of LO coupling from Path1 to Path2. Using the same above-mentioned analysis, all the interferences caused by LO coupling from Path1 fall into out-of-band, as illustrated in Fig. 13(b).

In conclusion, the proposed parallel direct-conversion and double-conversion configuration can push the interferences caused by LO coupling into out-of-band; consequently, the transmitter front end is essentially immune to the problem of crosstalk caused by LO coupling.

IV. TRANSMITTER FRONT-END DESIGN

A. Mixer Design

Mixers are generally classified as active and passive mixers, which are both used in the state-of-the-art transmitters. Active mixers can provide higher conversion gain and higher port isolation, and require lower LO power, but have higher noise than their passive counterparts. Passive mixers, on the other hand, typically show conversion loss but exhibit high linearity and low noise at the expense of high LO power requirement and poor port isolation [16], [17]. Mixer design requires many compromises among different figures of merit such as conversion gain, LO power, linearity, noise figure, port-to-port isolation, and total power dissipation.

Considering both advantage and disadvantage of active and passive mixers, the proposed double-conversion mixer in Path2 employs a passive mixer as the first stage and an active

mixer as the second stage based on the tradeoff between the conversion gain and the noise performance. While, the mixer in Path1 uses an active mixer mainly based on the considerations of conversion gain and port isolation. The proposed mixers are shown in Fig. 14. The passive mixer is driven by a 25% duty-cycle clock, while the active mixers are driven by a 50% duty-cycle clock. A 5-bit current-steering gain control unit is used in each path, and the range of the control gain is ~ 24 dB. In order to keep sufficient voltage headroom, the supply voltage for the mixers is chosen to be 1.8 V.

Since the second-stage active mixer in Path2 is not IQ mixer, there are two identical mixing products at the output of the second-stage active mixer: one required signal at the frequency of $f_{LO2B} + f_{LO2A}$ (5.17–5.835 GHz) and one unwanted image at the frequency of $f_{LO2B} - f_{LO2A}$ (1.7266–1.9417 GHz). This image increases the power consumption of the second-stage active mixer, which is the main impact on the second-mixing stage. In the transmitter, the mixer is followed by the driver and the PA, and there are three transformers after the mixer, forming three resonant tanks, which can suppress the magnitude of the unwanted image.

B. LO Input Balun and Divider Design

Since LO input signal is provided from an external signal source, a balun is required to convert the single-ended LO input signal to differential signal. Passive baluns are the

Fig. 14. Schematic of mixers. (a) Mixer in Path1. (b) Double-conversion mixer in Path2.

most popular choice, which are often implemented by passive transformer or coupler. However, in the frequency range of 10.34–11.67 GHz, passive baluns may not have small size. Moreover, passive baluns suffer from high loss, thus necessitating high power level of the input signal. While, active baluns are superior to passive baluns in terms of size and loss. Therefore, an active balun is used in this design, mainly considering small size and low loss, as shown in Fig. 15(a) [18]. The simulated gain imbalance is <0.2 dB, and the simulated phase difference is $180^\circ \pm 1.5^\circ$ from 0.5 to 20 GHz. On the other hand, the phase noise of this active balun should be low, such that the phase noise of the input signal should not be degraded by this balun. The simulated phase noises of the balun are -137.1 and -144.7 at 0.1- and 1-MHz offsets, showing the balun exhibits quite low phase noise.

The divider to generate 50% duty-cycle clock for LO1 uses current-mode logic (CML) latch, which has the advantage of high speed. The simulated phase noises of this divider are -143.56 and -151.97 dBc/Hz at 0.1- and 1-MHz offsets.

The other divider to generate 25% duty-cycle clock for LO2 originates from the divider presented in [19], and the NMOS crosscouple in the divider in [19] is replaced with a CMOS crosscouple to reduce power consumption and improve phase noise. The schematic of the proposed divider is shown in Fig. 15(b). The simulated phase noises are -147.7 and -155.24 dBc/Hz at 0.1- and 1-MHz offsets, respectively, and they are ~ 9 dB lower than those in [19]. The simulated power consumption of the structure in [19] and the proposed structure are 2.1 and 0.38 mW, respectively.

Fig. 15. (a) LO input balun. (b) 25% duty-cycle divider.

The LO phase noise causes a frequency spreading of the carrier frequency, thus impacting EVM. The EVM degradation is defined as follows [8]:

$$\text{EVM}_{\text{rms}} = 100\% \cdot \sqrt{2 - 2e^{-\sigma^2/2}} \quad (8)$$

where σ is the phase error in radians and is derived from the integrated phase noise. In practice, jitter is often used to quantify integrated LO phase noise in broadband communication systems. Since the relationship between jitter and phase error is

$$J = \frac{\sigma}{2\pi f_c} \quad (9)$$

EVM can be derived as follows:

$$\text{EVM} = 100\% \cdot \sqrt{2 - 2e^{-(2\pi f_c J)^2/2}} \quad (10)$$

To achieve $<0.5\%$ (-46 dB) EVM, a jitter of <136.4 fs is required. Excluding the phase noise contribution from the input balun and divider, the jitter of the external signal source is required to be lower than 134.2 and 127.4 fs for LO1_IN and LO2_IN, respectively. In the measurement, the jitters of the two signal sources to be used to provide LO1_IN and LO2_IN are 13.5 and 35.2 fs, respectively, which are much lower than the required value.

C. Driver and PA Design

Fig. 16 shows the schematic of the drivers and the PA, consisting of a pseudodifferential pair and cascode devices to provide a high output power under a supply voltage of 2.5 V. Thick-oxide devices are used in the common gate device to sustain gate-drain voltage stress in a high output power level. In our design, the PA is targeted to deliver an average output power of 10 dBm for the VHT80, MCS9, 802.11ax signal. Based on the high PAPR of the VHT80, MCS9 signal, which is at least 10 dB, to meet the requirement of an average output power of 10 dBm, the saturated output power of the PA should be >22 dBm. Accordingly, the transistor sizes of the common

Fig. 16. Schematic of the driver and the PA.

Fig. 17. (a) Input capacitance compensation. (b) Simulated AM-PM distortion of the PA.

source and common gate devices are set at 0.896 mm/40 nm and 1.536 mm/270 nm, respectively, such that the PA can deliver a maximum output power of 23.5 dBm. The simulated OIP3 of the PA is 38.89 dBm.

For CMOS PAs, AM-PM distortion is a major contributor to EVM. The nonlinear variation of the gate-source capacitance of input NMOS transistor with gate-source voltage results in AM-PM distortion, which can be compensated by adding a PMOS varactor or a PMOS transistor [20]–[22]. As evidence in [21], compared with the PMOS transistor, the PMOS varactor can more effectively compensate the nonlinear input capacitance of the input NMOS transistor in virtue of larger ratio of $C_{\max}/C_{\min}$. Therefore, the PMOS varactor is used in our design, and Fig. 17(a) shows the variation of input capacitance versus input voltage with and without PMOS varactor. In addition, the simulated AM-PM performance of the PA at different frequencies are plotted in Fig. 17(b), showing the AM-PM distortion of $< \pm 2.3^\circ$ across the 5-GHz WLAN band.

V. MEASUREMENT RESULTS

The proposed transmitter front end was fabricated in a TSMC 40-nm RF CMOS process with ten metal layers. The die microphotograph of the transmitter front end is shown in Fig. 18, and the die area is 1.1 mm $\times$ 1.3 mm.

A. Single-Carrier Performance

Before evaluating the transmitter front end's practical performance for carrier aggregation, each path was tested by using a VHT20, MCS5 (20-MHz bandwidth, 64-QAM modulation, and 3/4 code rate) signal and a VHT80, MCS9, 802.11ax signal (80-MHz bandwidth, 256-QAM modulation,

Fig. 18. Photograph of the transmitter front-end die.

Fig. 19. EVM measurement setup.

and 5/6 code rate) without digital predistortion (DPD) [23]. The measurement setup is shown in Fig. 19. The two baseband signals are first created in MATLAB, then downloaded to AWG (81180A) and VSG (E8267D). In the measurement, the RF output signal is directly sampled by using a Keysight DSOZ634A oscilloscope at a sampling rate of 160 Gsps, and demodulated in Keysight 89600. EVM was first measured by using the VHT20, MCS5 signal at 5.18, 5.6, and 5.825 GHz in Channels 36, 120, and 165, and then measured by using the VHT80, MCS9 signal at 5.29, 5.53, and 5.775 GHz in Channels 58, 106, and 155, respectively. The measured EVM in different average output power levels are reported in Figs. 20 and 21. For the VHT80, MCS9 signal, when satisfying the standard specification of the EVM of ≤ -32 dB, Path1 delivers an average output power of 9.75, 9.5, and 9 dBm, and Path2 delivers an average output power of 9.1, 8.3, and 8.37 dBm, at 5.29, 5.53, and 5.775 GHz, respectively. The best EVM achieved can reach -43.3 and -43.64 dB for Path1 and Path2, respectively. Since the PAPR of the VHT80, MCS9 signal is as high as 11.25 dB, large back-off is required for the PA to meet the stringent EVM requirement of -32 dB for 256-QAM modulation without using DPD. In this design, we do not focus on the PA design with a large output power; thus, the average output power of this transmitter front end for 802.11ax signal is lower than that of some other commercial transceivers with integrated PA.

TABLE I
MEASURED PERFORMANCE SUMMARY AND COMPARISON

		This work	ISSCC2018 [10]	ISSCC2017 [11]	JSSC2017 [24]
WLAN standards		11a/ac/ax	11abgn/ac/ax	11abgn/ac	11abgn/ac
Process (nm)		40	28	40	40
Integrated PA		Yes	No	Yes	Yes
Non-contiguous CA		Yes	Yes	Yes	No
Contiguous CA		Yes	No	Yes	No
Measured Signal		VHT80, MCS9, 11ax	VHT80, MCS10, 11ax	VHT80, MCS9, 11ac	VHT80, MCS9, 11ac
EVM Floor (dB)	Single Carrier	-42.5		-36.5	-38
	CA	-37.6 non-contiguous	-36.1 contiguous	-36.5 non-contiguous	N.A
P _{sat} (dBm)		21.9	N.A	27	29
Power Consumption (mW)		428@(5.3+4.8 dBm) (2 Streams without PLL)	832@(-5 dBm) (4 Streams +2 PLLs)	4164@(19+19 dBm) (2 Streams +2 PLLs)	1750@(17 dBm) (2 Streams +1 PLL)
Transmitter efficiency (%)		1.5	0.156	3.81	5.73

Fig. 20. Measured EVM with the VHT20, MSC5, 802.11ax signal for a single-carrier mode. (a) Path1. (b) Path2.

Fig. 22. Measured EVM performance for a contiguous carrier aggregation mode. (a) 5.21 + 5.29 GHz. (b) 5.53 + 5.61 GHz.

Fig. 21. Measured EVM with the VHT80, MSC9, 802.11ax signal for a single-carrier mode. (a) Path1. (b) Path2.

B. Carrier Aggregation Performance

We measured the EVM using two VHT80, MCS9, 802.11ax signals for intraband and interband carrier aggregation modes. The measured EVM of each carrier for two contiguous intraband carrier aggregation cases (5.21 + 5.29 and 5.53 + 5.61 GHz) are plotted in Fig. 22. For noncontiguous interband carrier aggregation, EVM was measured for three cases: 5.21 + 5.53, 5.61 + 5.775, and 5.21 + 5.775 GHz, and the measured results are shown in Fig. 23. Compared with the measured results for single-carrier operation, 2–3 dB more backoff is required to meet the linearity requirement in terms of the total output power of the two carriers. It is reasonable that the linearity of PA degrades when processing two wideband signals concurrently. If DPD is employed, the average output power will be increased by a large degree, and it is beyond the scope of this article. The measured

Fig. 23. Measured EVM performance for a noncontiguous carrier aggregation mode. (a) 5.21 + 5.53 GHz. (b) 5.21 + 5.775 GHz. (c) 5.61 + 5.775 GHz.

power consumptions of the two LO generators, two mixers, two drivers, and PA are 88.9, 54, 127.5, and 157.5 mW, respectively.

Meanwhile, the measured performance is summarized and compared with other recent works listed in Table I. From Table I, the EVM floor achieved in this article has the same level as that in other state-of-the-art works, and verifying a good EVM performance can be achieved even without high

Fig. 24. Spectrum for contiguous carrier aggregation with power difference between two carriers.

Fig. 25. Measured EVM versus power difference.

physical isolation between RF signal paths. Compared with some other works, the power consumption is high, and the transmitter efficiency is low. One reason for the high power consumption is that the circuit blocks are not optimal regarding the power consumption and efficiency. In addition, when the PA concurrently processes two carriers using VHT80, MCS9, 802.11ax signals, large backoff results in lower efficiency of PA and drivers.

During the measurement, the power level difference between the two carriers is kept within ± 1.5 dBm, aiming to avoid EVM degradation by the emission from the adjacent channel in the contiguous carrier aggregation mode. When the output power level has a large difference between the two carriers, the carrier having lower power will suffer from considerable adjacent channel emission from the other carrier having a larger output power, thus leading to SNR deterioration and EVM degradation, as illustrated in Fig. 24 [11]. In order to gain insight into the impact of power level difference on EVM for the contiguous carrier aggregation mode, we measured the EVM of Path2 at 5.61 GHz with a fixed output power, while tuning the output power of Path1 at 5.53 GHz. The measured EVM versus the power level difference is reported in Fig. 25. When the output power of Path2 is much larger than that of Path1, better EVM of Path2 is achieved and vice versa.

In addition, the measured spectrum for contiguous carrier aggregation (5.53 + 5.61 GHz) is depicted in Fig. 26(a), showing that the requirement of emission mask is satisfied in a 5-GHz band. From the wideband spectrum shown in Fig. 26(b), the two unwanted signals caused by

Fig. 26. Measured spectrum for contiguous carrier aggregation. (a) 1-GHz span. (b) 5-GHz span.

LO coupling fall into out-of-band at the frequencies of 3.66 and 7.4 GHz ($f_{LO1} \pm f_{LO2A}$), as previously predicted in Section III.B. In addition, there are two more unwanted signals at 3.74 and 7.48 GHz, which are caused by the double-conversion mixer in Path2. Although the resonant tanks in the transmitter front end can suppress the magnitude of these unwanted out-of-band signals, some signals are still quite strong and violate the emission mask. Therefore, in a practical application, one more bandpass filter is required to suppress these unwanted signals.

VI. CONCLUSION

A new transmitter architecture supporting two-carrier aggregation for a 5-GHz WLAN 802.11ax application is presented, and a prototype is implemented in a TSMC 40-nm CMOS process. For a single-carrier operation, the required EVM floor reaches -43.3 dB, and the transmitter delivers an average output power of 9.75 dBm with the EVM of -32 dB for the VHT80, MCS9, 802.11ax signal. By virtue of the proposed parallel direct conversion and double conversion configuration, the transmitter can mitigate the intrinsic problems of crosstalk and VCO pulling without high physical isolation between

RF channels. For the contiguous carrier aggregation, which is the worst case for carrier aggregation, the measured EVM reaches -36.1 dB for VHT80, MCS9 signals, verifying a high SNR is achieved in the proposed transmitter front end. In addition to the 802.11ax WLAN application, the proposed architecture is also suitable for LTE-A and emerging 5G communication applications.

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