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DOI 10.1109/ISCAS51556.2021.9401788

Publication date 2021 Document Version Final published version

**Published in** 2021 IEEE International Symposium on Circuits and Systems (ISCAS)

# Citation (APA)

Kumaran, A. K., Pashaeifar, M., D'Avino, M., de Vreede, L. C. N., & Alavi, M. S. (2021). On-chip output stage design for a continuous class-F power amplifier. In *2021 IEEE International Symposium on Circuits and Systems (ISCAS)* Article 9401788 IEEE. https://doi.org/10.1109/ISCAS51556.2021.9401788

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# On-Chip Output Stage Design for a Continuous **Class-F Power Amplifier** Anil Kumar Kumaran\*, Masoud Pashaeifar\*, Marco D'Avino<sup>†</sup>, Leo C.N. de Vreede\*, and Morteza S. Alavi\* \* Electronic Circuits and Architecture (ELCA) Research Group, Delft University of Technology Abstract—Continuous Class F (CCF) power amplifiers (PAs) overcome Class-F PA's disadvantage of narrow bandwidth by relaxing the short-circuit requirement at the 2<sup>nd</sup> harmonic while still maintaining 90.7% peak efficiency over the band of interest. This paper proposes four different CCF output networks, with their design procedure, suitable for on-chip implementation in the 2.1-2.7GHz band. The output stage with 2nd harmonic trap and no RF choke is favoured due to its flat real impedance,

structure is in reach. Index Terms-Continuous class F (CCF), Output matching network, Power amplifier (PA), Harmonic termination, Differentialmode analysis, Common-mode analysis.

low fundamental reactance, and compact layout. Using a 40nm

CMOS process, a passive efficiency of 68% at 2.4GHz for this

## I. INTRODUCTION

Today, there is an increased demand for high-speed, lowcost transceivers in cellular/WLAN wireless networks. A CMOS-based solution is considered to be a viable option to address these requirements since it can facilitate the implementation of a System-on-Chip (SoC) solution at a low cost. However, the TX power amplifier (PA) is generally considered as the most challenging block in a CMOS SoC since it must offer high-efficiency operation while meeting the stringent spectral mask requirements of wireless standards. Currently, most TXs employ linear class A/B PAs. Nevertheless, their ideal peak efficiencies are constrained to 50/78.5%, respectively, due to the relatively large overlap in their drain voltage/current waveforms (Fig. 1a/b).

On the other hand, class-F PAs ideally achieve a peak efficiency of 100% by utilizing harmonic-frequency resonators to short-circuit at even harmonics and open-circuit at odd harmonics leading to non-overlapping drain voltage/current waveforms (Fig. 1c). Nonetheless, in reality, controlling all harmonics is challenging and increases the network complexity and component losses, as such degrading its passive efficiency. Therefore, practical implementation of class-F PAs employ up to three harmonics and can ideally achieve a peak efficiency up to 90.7% (Fig. 1d) [1]. The generalized drain-source-voltage  $(V_{DS})$  containing all frequencies up to  $3^{rd}$  harmonic [2] is given bv:

$$V_{DS} = \underbrace{1}_{\text{DC}} - \underbrace{\frac{2}{\sqrt{3}}\cos\theta}_{1^{st} \text{ harmonic}} + \underbrace{\frac{1}{3\sqrt{3}}\cos3\theta}_{3^{rd} \text{ harmonic}}$$
(1)



Fig. 1: Class-A/B/F drain voltage/current (V/I) waveforms.

The drain current  $(I_D)$ , which is a half-sine wave, is given by

$$I_D = \underbrace{\frac{1}{\pi}}_{\text{DC}} + \underbrace{\frac{1}{2}\cos\theta}_{1^{st} \text{ harmonic}} + \underbrace{\frac{2}{3\pi}\cos2\theta}_{2^{nd} \text{ harmonic}} - \underbrace{\frac{2}{15\pi}\cos4\theta}_{4^{th} \text{ harmonic}}$$
(2)

where  $\theta$  is the fundamental phase. The load impedances at the 1<sup>st</sup>, 2<sup>nd</sup>, and 3<sup>rd</sup> harmonic are represented by  $Z_{1f} = \frac{4}{\sqrt{3}}$ ,  $Z_{2f} = 0$ , and  $Z_{3f} = \infty$ , respectively. As depicted in Fig. Id, one of the key advantages of class-F PAs is that its normalised peak drain voltage is two. Nonetheless, these traditional class-F PAs suffer from bandwidth limitations (typically up to 10%) relative bandwidth) due to their dependence on ideal short and open circuit harmonic terminations. To realize larger bandwidths, the continuous class-F (CCF) PA has been proposed [3]. In this work, we present and evaluate four different output network topologies, all suited to implement CCF operation.

This paper is organized as follows. Section II briefly explains the equations governing the operation of CCF. Section III discusses the design procedure of different CCF output networks. Section IV compares the proposed four output stages, and presents layout of the chosen design. Section V concludes this paper.

# II. CONTINUOUS CLASS-F

Compared to class-F, the CCF operation allows an extra imaginary part at both its  $1^{st}/2^{nd}$  harmonic of the voltage

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Fig. 2: (a)  $V_{DS}$  and  $I_D$  waveform for CCF with -1.5 <  $\gamma$  < 1.5, and (b) Variation of  $Z_{1f}$ ,  $Z_{2f}$ ,  $Z_{3f}$  for  $-1 < \gamma < 1$ .

waveform. Thus, the generalized  $V_{DS}$  for CCF is given by [4]:

$$V_{DS} = \underbrace{1}_{\text{DC}} - \underbrace{\frac{2}{\sqrt{3}}\cos\theta - \gamma\sin\theta}_{1^{st} \text{ harmonic}} + \underbrace{\frac{7\gamma}{6\sqrt{3}}\sin2\theta}_{2^{nd} \text{ harmonic}} + \underbrace{\frac{1}{3\sqrt{3}}\cos3\theta}_{3^{rd} \text{ harmonic}}$$
(3)

where  $\gamma$  is an empirical parameter. As in class-F,  $I_D$  is a half sinusoid [given in (2)]. These equations indicate that  $V_{DS}/I_D$  are chosen so that no power is dissipated at higher harmonics. Also, mathematically,  $\gamma$  in (3) does not affect the drain efficiency  $(\eta_D)$ , but it does affect the peak drain voltage, as depicted in Fig. 2a. However,  $I_D$  depends on  $V_{DS}$ , degrading  $\eta_D$  as  $\gamma$  increases.

Fig. 2a implies that at  $\gamma = 0$ , the waveform is like class– F. For the  $\gamma$  values between -1 and 1,  $V_{DS}$  remains positive, enabling CCF PA to meet linearity requirements. However, CCF PAs suffer from large peak drain voltage swings, which can be as high as 3.34 times the supply  $(V_{DD})$  when  $\gamma = -$ 1 or 1. Using (3) and (2), the load impedances of CCF are  $Z_{1f} = \frac{4}{\sqrt{3}} + j2\gamma$ ,  $Z_{2f} = 0 - j\frac{\pi}{2}\frac{7\sqrt{3}}{6}\gamma$ , and  $Z_{3f} = \infty$  [5].

In CCF,  $Z_{3f}$  remains open-circuited similar to class-F. Meanwhile,  $Z_{1f}$  and  $Z_{2f}$  have a reactive part, unlike class-F. From CCF load impedances and Fig. 2b, it is perceived that if the reactive part of  $Z_{1f}$  changes from inductive (+) to capacitive (-), then the reactive part of  $Z_{2f}$  needs to change from capacitive (-) to inductive (+) or vice-versa across the bandwidth, to achieve CCF operation. In the next section, the design procedure for the proposed four output networks is meticulously presented.

# III. DESIGN OF OUTPUT NETWORK FOR CCF

In this work, the push-pull (differential) structure is chosen for the PA, mainly because it decouples the odd harmonics  $(1^{st}/3^{rd})$  from even harmonic  $(2^{nd})$  impedance. Moreover, it suppresses supply/substrate noise, second-order nonlinearities and doubles the RF output power for a given breakdown voltage. In this work, the targeted peak RF power is 27dBm while its operational bandwidth is 2.1-2.7GHz with  $V_{DD}$  = 2.7V. To achieve this, a fundamental differential impedance of 38.7 $\Omega$  should be presented to the drains of the transistors across the 2.1-2.7GHz band, which can be obtained as follows:

#### TABLE I: OUTPUT NETWORK SPECIFICATIONS.

| Class of   | First Harmonic                 | Second Harmonic               | Third Harmonic              |
|--|--------------------------------|-------------------------------|-----------------------------|
| Operation  | (ω)                            | (2ω)                          | (3 <i>w</i> )               |
| Class F  | $\Re(Z_D) = 38.7\Omega$        | $\Re(Z_D) = 0\Omega$          | $ 7_{-}  \sim 10000$        |
| (2.4GHz)   | $\Im(Z_D) = 0\Omega$           | $\Im(Z_D) = 0\Omega$          | $ Z_D  \approx 100052$      |
|  | $\Re(Z_D) = 38.7\Omega$        | $\Re(Z_D) = 0\Omega$          |                             |
| CCF  | $\mathfrak{I}(Z_D) \to +$ to - | $\Im(Z_D) \to -$ to +         | $ Z_D  \approx 1000 \Omega$ |
| (2.1-2.7GHz)   | OR                             | OR                            |                             |
|  | $\mathfrak{I}(Z_D) \to -$ to + | $\mathfrak{I}(Z_D) 	o +$ to - |                             |
| $V_{FUND-diff} = 2 \times V_{FUND-single} = 2 \times \frac{2}{\sqrt{3}} V_{DD} = 6.24 V$ |                                |                               |                             |
| $R_{D-diff} = \frac{V_{FUND-diff}^2}{2} = 38.7 \ \Omega$                                 |                                |                               |                             |

$$P_{D-diff} = rac{V_{FUND-diff}}{2 \times P_{OUT_{max}}} = 38.7 \ \Omega$$

The requirements of the output network for CCF are exhibited in Tab. I. The PA operates in class-F mode at the center frequency  $\omega_0$  (2.4GHz) with a short at  $2\omega_0$  (4.8GHz) and an open at  $3\omega_0$  (7.2GHz). But, for all other frequencies, the PA performs in the CCF mode.

The four different output networks proposed are based on lumped passive components, see Fig. 3a/b/c/d. All the designs include a balun and load capacitance  $(C_L)$ . The balun converts the balanced (differential) signal to its single-ended companion, whereas  $C_L$  adjusts the 3<sup>rd</sup> harmonic impedance. The power transistor's drain-source capacitance (assumed  $C_{DS} = 1.87 \text{pF}$ ) is absorbed into the output network to reduce its impact on the PA performance. The balun is modeled using an ideal transformer, consisting of magnetizing inductance  $(L_m)$ , leakage inductance  $(L_k)$ , primary inductance  $(L_P)$ , and coupling coefficient  $(k_m)$  [6].

# A. Design A (no RF Choke & with $L_2C_2$ )

Design A (Fig. 3a) includes a  $2^{nd}$  harmonic trap  $(L_2C_2)$ , which provides short at  $2\omega_0$ . The  $V_{DD}$  is supplied through the balun's center tap, and  $L_{BND}$  is used to model the bond-wire inductance ( $\approx$ 1nH). Analysis of the schematics is performed in differential-/common-mode scenarios by utilizing the equivalent circuits depicted in Fig. 3a.1/a.2, respectively, to calculate the unknown parameters: transformer's coupling factor  $(k_m)$ and turn ratio (n) along with  $L_P$ ,  $C_2$ ,  $L_2$ , and  $C_L$ . Fig. 3a.1 shows that the drain impedance  $(Z_D)$  is given by

$$Z_D = \left(\frac{1}{\frac{j\omega C_{DS}}{2}} + \frac{1}{Z_{SB}}\right)^{-1} = 38.7 \ \Omega \tag{5}$$

The value of  $Z_{SB}$  [impedance of  $L_2C_2$  and balun given by (6)] that will provide  $Z_D$  of 38.7 $\Omega$  can be calculated from (5) and the value is  $\Re(Z_{SB})(\omega_0) = 29.8\Omega$  and  $\Im(Z_{SB})(\omega_0) =$ 16.6Ω.

$$Z_{SB} = \left(\frac{1}{Z_B} + \frac{1}{Z_S}\right)^{-1} \text{ where, } Z_S = 2j\omega L_2 + \frac{1}{\frac{j\omega C_2}{2}},$$
$$Z_B = \left(\frac{1}{R_P} + \frac{1}{2j\omega L_m} + j\omega C_P\right)^{-1} + 2j\omega L_k, \qquad (6)$$
$$R_P = R_L \left(\frac{k_m}{n}\right)^2, C_P = C_L \left(\frac{n}{k_m}\right)^2$$



Fig. 3: (a) Design A (Balun,  $L_2C_2$  and  $C_L$ ), (a.1) odd-mode, (a.2) even-mode equivalent circuit (b) Design B (Balun and  $C_L$ ), (c) Design C (Balun, RF choke,  $L_2C_2 \& C_L$ ), and (d) Design D (Balun, RF choke  $\& C_L$ ).

It is evident from (5) that  $C_{DS}$  should resonate with  $\Im(Z_{SB})$  to attain high drain impedance at  $\Im\omega_0$ . This implies  $\Im(Z_{SB})(\Im\omega_0) = 24.96\Omega$ . Ideally,  $\Re(Z_{SB})(\Im\omega_0)$  should be  $\Omega\Omega$  to achieve high  $\Im^{rd}$  harmonic impedance. However, Fig. 4c depicts that having a larger  $\Re(Z_{SB})(\Im\omega_0)$  contributes to a constant  $P_{OUT}$  across the operational bandwidth by flattening the real part at the fundamental. Another benefit is that it has a more linear reactive part at the fundamental, which is essential in CCF operation. However, this leads to a lower  $\Im^{rd}$  harmonic impedance, as showcased in Fig. 4b. This also emphasizes the significance of  $C_L$ . Thus, to achieve an open circuit at the  $\Im^{rd}$  harmonic, an impedance of approximately  $1000\Omega$  is required



Fig. 4: (a) Magnitude of  $Z_D$  vs real part of  $Z_{SB}$  ( $\Re(Z_{SB})$ ) at  $3\omega_0$ , (d) Magnitude of  $Z_D$  at  $3^{rd}$  harmonic for various  $R_{SB}(3\omega_0)$ , and (c)  $Z_D$  at fundamental for various  $R_{SB}(3\omega_0)$ .

which is obtained with  $\Re(Z_{SB})(3\omega_0) = 0.5\Omega$  (Fig. 4a). Fig. 3a.2 proves that  $L_2$  should have a series resonance with  $C_2$  to get short at  $2\omega_0$ .

$$L_2 = \frac{1}{4 \times \omega_0^2 \times C_2} \tag{7}$$

The five unknowns in the circuit:  $k_m$ , n,  $L_P$ ,  $C_2$ , and  $C_L$  can be calculated by assuming one of them and using four equations  $[\Re(Z_{SB})(\omega_0) = 29.8\Omega, \Im(Z_{SB})(\omega_0) = 16.6\Omega, \Re(Z_{SB})(3\omega_0 = 0.5\Omega)$  and  $\Im(Z_{SB})(3\omega_0) = 24.96\Omega$ ]. In this work,  $k_m = 0.69$  is assumed and the remaining unknowns (n = 1,  $L_P = 0.8$ nH,  $C_L = 2.4$ pF,  $C_2 = 0.8$ pF,  $L_2 = 1.4$ nH) are calculated.  $k_m$  can be varied to get different sets of results in which  $L_P$  is minimal and thereby making the layout friendly.

# B. Design B (no RF Choke & no $L_2C_2$ )

In this design (Fig. 3b),  $L_2C_2$  is removed, reducing the number of unknowns. Like the previous case, differential mode analysis yields four equations, and there are four unknowns, thus leading to a single set of values for  $k_m = 0.72$ , n = 0.9,  $L_P = 0.63$ nH,  $C_L = 3.96$ pF, unlike design A.  $C_{DDP}$  is tuned to provide a short at  $2\omega_0$  such that  $C_{DDP}$  resonates with  $L_P$ ,  $L_{BND}$ , and  $C_{DS}$ , which is obtained from the common-mode analysis. Moreover,  $C_{DDP}$  provides RF ground and blocks DC.

# C. Design C (with RF Choke & with $L_2C_2$ )

In this design (Fig. 3c),  $V_{DD}$  is supplied through a RF choke, unlike the previous designs. RF chokes are assumed to have a fixed value of 5nH. The differential mode analysis yields four equations similar to design A. Assuming  $k_m = 0.73$ , the other unknowns are calculated as n = 0.93,  $L_P = 0.95$ nH,  $C_L =$ 2.5pF,  $C_2 = 0.61$ pF. Like design A,  $C_2$  should resonate out with  $L_2$  to get short at  $2\omega_0$ . Thus,  $L_2 = 1.8$ nH.



Fig. 5: (a) Impedance  $(Z_D)$  at 1<sup>st</sup> harmonic, (b) Reactive part of  $Z_D$  ( $\Im(Z_D)$ ) at 2<sup>nd</sup> harmonic, and (c) Magnitude of  $Z_D$  ( $|Z_D|$ ) at 3<sup>rd</sup> harmonic.

## D. Design D (with RF Choke & no $L_2C_2$ )

Unlike design C,  $L_2C_2$  is removed in this design (Fig. 3d). Also, the RF choke should be calculated, assuming  $C_{DS}$  resonates with it at  $\omega_0$  (RF choke = 2.35nH). The differential mode analysis yields four equations and the four unknowns (n = 0.84,  $L_P = 0.86$ nH,  $C_L = 3.95$ pF,  $k_m = 0.77$ ) can be calculated.  $C_{DDP}$  can be tuned to provide a short at  $2\omega_0$ .

# **IV. RESULTS**

Fig. 5 shows that all the four designs satisfy the main CCF requirement, which is a decreasing trend of the reactive part at the fundamental and increasing trend of the reactive part at  $2^{nd}$  harmonic. As illustrated in Fig. 5a, the real part in the desired frequency band is flatter for design A/C, which leads to constant  $P_{OUT}$ , unlike design B/D. Moreover, to fully accomplish CCF operation, the series  $L_2C_2$ , which resonates at the  $2^{nd}$  harmonic, acts as a capacitor at the desired band. The designs B/D have a higher reactive part at the fundamental than other designs, contributing to a larger  $\gamma$  value and, thus, higher peak drain voltage (Fig. 2a). Fig. 5b/c show that all designs have a similar response at  $2^{nd}/3^{rd}$  harmonics. Nevertheless, design A generates a more constant  $P_{OUT}$  in the band of interest and allows the use of compact passive components.

To verify the proposed output passive stage design approach, design A is laid out in an LP 40nm CMOS process (Fig. 6) with a  $600\mu m$  balun. Subsequently, design A is simulated in three different ways: as a lossless network, with losses using a quality factor of 7/13/100 for the inductors/balun/capacitors, respectively, and as the proposed layout. The ADS (Momentum) simulation results are depicted in Fig. 7, which indicates a good agreement between the lossy and the proposed layout, reaching 68% passive efficiency at 2.4GHz. Moreover, using an ideal push-pull PA, the proposed design achieves peak



Fig. 6: Layout of the design A (balun,  $L_2C_2$  and  $C_L$ ).



Fig. 7: (a) Passive efficiency, and (b) Maximum  $P_{OUT}$  and drain efficiency versus frequency band.

RF output power/drain efficiency of more than 25dBm/62%, respectively, over the desired frequency band.

# V. CONCLUSION

In this paper, the primary advantage of CCF PAs over its class–F companion, as well as the critical requirement needed for the output network to operate in CCF mode are presented. Furthermore, the design procedure of various output networks for the 2.1–2.7GHz band is presented, and the proposed passive output stages are synthesized. The  $2^{nd}$  harmonic trap, which acts as a varying capacitor at  $1^{st}$  harmonic, helps in maintaining constant output RF power in the desired frequency band for the design A/C. Consequently, design A, with no RF choke and a  $2^{nd}$  harmonic trap is chosen since it has minimal on-chip passive components. This design is laid out in a 40nm CMOS while achieving a 68% passive efficiency at 2.4GHz.

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