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Citation for published version (APA):

Bechthum, E., Radulov, G. I., Briaire, J., Geelen, G., & Roermund, van, A. H. M. (2013). A novel output transformer based highly linear RF-DAC architecture. In *Proceedings of the 21st European Conference on Circuit Theory and Design (ECCTD 2013), 8-12 September 2013, Dresden, Germany* (pp. 1-4)
<https://doi.org/10.1109/ECCTD.2013.6662314>

DOI:

[10.1109/ECCTD.2013.6662314](https://doi.org/10.1109/ECCTD.2013.6662314)

Document status and date:

Published: 01/01/2013

Document Version:

Accepted manuscript including changes made at the peer-review stage

Please check the document version of this publication:

- A submitted manuscript is the version of the article upon submission and before peer-review. There can be important differences between the submitted version and the official published version of record. People interested in the research are advised to contact the author for the final version of the publication, or visit the DOI to the publisher's website.
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- The final published version features the final layout of the paper including the volume, issue and page numbers.

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A novel output transformer based highly linear RF-DAC architecture

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Abstract—A major limitation of the linearity of Current Steering (CS) RF-DACs is the large output voltage swing (typically $1V_{pp}$), which couples to sensitive internal nodes and thereby causes non-linear distortion. This paper proposes a novel approach for the linearization of the CS RF-DAC. An output transformer decouples the output from the circuit core and attenuates the voltage swing seen by the RF-DAC current cells.

A lumped element model of a transformer is used in calculations and simulations to analyze the performance of the transformer in the Mixing-DAC application, and to select optimal design parameters for high linearity. Verification with a simulation model of an RF-DAC shows that the output related non-linearity (IMD3) of the CS RF-DAC improves with about 14dB when the proposed transformer parameters are used.

I. INTRODUCTION

Highly linear wideband transmitters are needed for the transmission of multicarrier GSM signals[1]. A transmitter with an RF-DAC is shown in Fig. 1. The advantages of using a Mixing-DAC as an RF-DAC are discussed in [2].

Fig. 1. Transmitter with RF-DAC

An RF-DAC based on the Current Steering DAC approach can be very linear when the 1-bit switched current sources are independent and identical. However, among the main causes of harmonic distortion are the output related error mechanisms. The large signal swing at the output (typically $1V_{pp}$) modulates the non-linear output capacitance and couples to internal nodes due to the finite output impedance. When the output signal swing is reduced, the linearity of the converter improves.

Without loss of generality, in this paper a specific RF-DAC implementation is assumed: a Current Steering Mixing-DAC based on [2]. The presented analysis also applies to other Current Steering (RF)-DACs.

An exemplary application, requiring high linearity, is multicarrier GSM. The required IMD3 is -80dBc [2] at $f_{out}=0.7-4.0GHz$ to account for current and future GSM bands. For a simple simulation, Fig. 2 shows the IMD3 (i.e. linearity) versus the load resistance at $I_{out,pp}=20mA$ and $f_{out}=4GHz$. At the typical load resistance of 50Ω , the IMD3 is -71dBc, which is not sufficient for the exemplary application. Since the output related non-linearity mainly originates from the switching output transistors, only those transistors are realistically modeled in the simulation of Fig. 2.

In [2], local output cascode transistors are used to shield the internal switching transistors from the output, and hence

Fig. 2. Current Steering Mixing-DAC output linearity dependence on the load resistor (i.e. output voltage swing)

Fig. 3. Current Steering Mixing-DAC with output transformer

improve the linearity. This paper proposes another solution: the use of an output transformer, which is shown in Fig. 3. When the turn ratio of the transformer n is larger than one, the load impedance seen from the Mixing-DAC current cells is roughly R_L/n^2 . Therefore the voltage swing at the output of the Mixing-DAC current cells is lower than at the load, which reduces the output related non-linearities. A RF-DAC with output transformer has been presented before [3], but it has never been proposed for linearity reasons.

The next section compares these two linearity enhancement methods. In section III, implementation options for the output transformer are discussed. Section IV presents calculations of the specific transformer setup. Section V presents simulation results of the transformer and Mixing-DAC with transformer.

II. CASCODE VS. TRANSFORMER

Two methods for RF-DAC linearization, discussed in this section, are: local output cascode and output transformer. Note

that the two methods are not mutually exclusive, but can be combined. Table I gives an overview of the comparison.

TABLE I
COMPARISON OF CASCODE AND TRANSFORMER FOR RF-DAC

	Cascode	Transformer
Area	++	--
Power consumption	++	--
Common mode isolation	--	++
Voltage headroom	--	++
High frequency output filtering	-	+
DC output filtering	o	o
Added non-linear elements	-	+
High frequency capability	-	+
Noise	-	+

Main disadvantages of the output transformer method include power consumption and chip area usage. For the transformer, a turn ratio $n > 1$ is required to decrease the output related non-linear distortion. However, the Mixing-DAC output current needs to be increased by n to maintain equal signal power at the load, increasing the power consumption and area of the Mixing-DAC core. In addition, inductors usually consume significant chip area and create power losses.

Main advantages of the transformer are as follows. The output transformer introduces common mode isolation between the Mixing-DAC and the load, resulting in a completely independent common mode level at the load. The output cascode requires additional voltage headroom. Especially in advanced process technology nodes, the voltage headroom is severely limited. The output transformer also filters the low frequency offset and higher frequency undesired output components. Low frequency filtering can be a disadvantage of the transformer when a low frequency offset at the RF-DAC output is required. A core-less transformer (i.e. passive element) is highly suitable for high frequencies and does not introduce significant additional non-linear distortion. A transistor (active element) is inherently non-linear and introduces additional parasitic capacitances, limiting the maximum signal frequency. The wideband output noise of the Mixing-DAC is mainly generated by the current source transistors. Since the output current is increased in the transformer method, the SNR will also improve for equal current density.

III. TRANSFORMER IMPLEMENTATION OPTIONS

Three different transformer implementation options are considered: on-chip (i.e. on-die), on-laminate and on-PCB; see Fig. 4. In principle, multiple cascaded transformers with different implementation can also be used. For simplicity, in this paper only a single transformer is considered.

On-chip transformers are very popular in literature [4]–[6], but not yet proposed for linearity enhancement of RF-DACs. For the design and simulation of a fully customized transformer, numerous tools are available. The interface between the Mixing-DAC and the transformer is very good, since both reside on the same substrate. However, the transformer coils can be bulky for the low target frequencies, which can make the on-chip transformer expensive.

Fig. 4. Three implementation options for the Mixing-DAC output transformer

A discrete on-PCB transformer is the most inexpensive and flexible option, but is only available in a limited number of configurations. The interface between the transformer and the Mixing-DAC contains one or multiple bondwires, which can degrade the transformer performance.

With on-laminate transformers, large inductors are possible due to the low cost of laminate area compared to standard CMOS chip area, and custom transformer design is possible. However, a bondwire interface is present between the Mixing-DAC and the transformer. Flip-chip assembly can minimize the inductance and resistance of the chip-to-laminate interface.

Simulations of the transformer model including bondwires reveal the most suitable implementation, see section V-B.

IV. TRANSFORMER THEORY AND MODELING

A. Calculation and simulation model

For calculations and simulations, the *Direct form model* of [4] is used. Fig. 5 shows this transformer model with the source and load configuration, where k is the transformer coupling factor ($k < 1$), and n is the turn ratio $n = \sqrt{L_2/L_1}$.

Fig. 5. Simple transformer model for calculations

The signal source is the Mixing-DAC, which can be modeled as a current source (I_{in}) with finite output impedance. In practice, the output resistance is much larger than the load resistance, hence it is not modeled. The output capacitance C_{par} is typically 1pF. The load impedance is ideally only a resistor (R_L), but typically also contains some capacitance (C_L).

In reality, transformers also contain parasitic capacitances and resistances. Bondwires are usually not taken into account in transformer analysis, but can severely limit the transformer performance. Therefore, also an extended model of the transformer with the aforementioned elements is used in simulations, see Fig. 6.

Calculations on transformer characteristics in open literature usually discuss the transformer performance in the case of matched source and load impedances and only discuss S-parameters. In the Mixing-DAC output transformer, the source impedance is not matched and the S-parameters do not give the desired information. Instead, the voltage attenuation between input and output, $H_{VV}(\omega)$, is of main importance. The power

Fig. 6. Extended model of a transformer in a Mixing-DAC

transfer is analyzed using the input current to output voltage transfer of the transformer $H_{VI}(\omega)$. Both performance metrics are discussed in the next subsections. For the first order approximation, the simple model of Fig. 5 is used.

B. Current-to-voltage transfer $H_{VI}(\omega)$

It is assumed that the output current of the Mixing-DAC current cells is scaled with n , such that the output power at the load is constant. The output impedance of the Mixing-DAC current cells also changes accordingly: $I_{in} = I_{in,0} \cdot n$, $C_{par} = C_{par,0} \cdot n$. The calculated transfer characteristic is:

$$H_{VI}(\omega) = \frac{V_{out}(\omega)}{I_{in}(\omega)} = H_0 \cdot \left(\frac{1}{1 - \left(\frac{\omega}{\omega_{res}} \right)^2} \right) \cdot \left(\frac{1}{1 + j \frac{\omega}{\omega_H} + \frac{1}{j \frac{\omega}{\omega_L}} + \frac{1}{j \frac{\omega}{\omega_A} + j \frac{1}{\omega_B}}} \right), \quad (1)$$

where:

$$H_0 = R_L \cdot k \cdot \frac{1}{n}, \quad \omega_{res} = \sqrt{\frac{1}{(1 - k^2) \frac{L_2}{n} C_{par,0}}}, \quad (2)$$

$$\omega_L = \frac{R_L}{L_2}, \quad \omega_A = \frac{R_L k^2}{L_2 (1 - k^2)}, \quad (3)$$

$$\omega_H = \frac{1}{R_L C_L}, \quad \omega_B = \frac{n}{k^2 R_L C_{par,0}}. \quad (4)$$

An approximation of the bode plot of the $H_{VI}(\omega)$ amplitude is shown in Fig. 7. The passband amplitude H_0 is determined by R_L , converted by the transformer with ratio k/n . The lower frequency corner ω_L is mainly due to the R_L/L_2 time constant. A resonance peak and second order low-pass filter corner is present at ω_{res} . This resonance is caused by an LC tank, formed by the leakage inductance $(1 - k^2)L_2$ and the Mixing-DAC output capacitance seen at the transformer output $C_{par,0}/n$. The other high frequency pole ω_H is mainly due to the $R_L C_L$ time constant. The other two roots at ω_A and ω_B only slightly influence the H_{VI} transfer characteristic around the resonance frequency.

Fig. 7. Amplitude of the $H_{VI}(\omega)$ of the simplified transformer model

C. Voltage gain $H_{VV}(\omega)$

The voltage gain $H_{VV}(\omega)$ of the transformer is a measure for the linearity improvement which can be attained when adding the transformer to the output of the Mixing-DAC:

$$H_{VV}(\omega) = \frac{V_{out}(\omega)}{V_{in}(\omega)} = k \cdot n \cdot \frac{1}{1 + (1 - k^2)j\omega \frac{L_2}{R_L} - (1 - k^2)\omega^2 C_L L_2} \quad (5)$$

As expected, the low frequency value of $H_{VV}(\omega)$ is determined by $k \cdot n$. The bandwidth is determined by the leakage inductance due to imperfect transformer coupling $L_2(1 - k^2)$ together with R_L and C_L .

V. SIMULATION RESULTS

In this section, simulation results of the separate transformer and simulations of the Mixing-DAC with output transformer are presented.

A. Transformer

The configuration of the simulation setup is: $R_L = 50\Omega$, $C_L = 500fF$ and $C_{par,0} = 1pF$. Simulations have shown that an optimal transformer configuration for the exemplary application (0.7-4.0GHz) is: $L_2 = 8nH$, $k = 0.85$ and $n = 4$ (hence, $C_{par} = 4pF$ and $L_1 = 0.5nH$), where k and n are limited by what is practically achievable. The resulting simulated transfer characteristic $H_{VI}(\omega)$ and voltage gain $H_{VV}(\omega)$ are shown in Fig. 8.

Fig. 8. Simulation results of simple transformer model

The shape of the simulated $|H_{VI}(\omega)|$ and $|H_{VV}(\omega)|$ closely match the calculated shape. The passband amplitude of $|H_{VI}(\omega)|$ is 22dBΩ, resulting in approximately $1V_{pp}$ signal at R_L for $I_{in}=n \cdot 20mA=80mA$. The passband flatness is 6dB. At low frequencies $|H_{VV}(\omega)|$ is 10.7dB, lowering to 9.4dB at 4.0GHz. The $|H_{VI}(\omega)|$ in-band flatness of 6dB can be improved by improving k or C_{par} , which are both practically

limited. However, simple digital preprocessing can easily counteract the non-flatness of the power transfer. The 4GHz $|H_{VV}(\omega)|$ value can only be increased by decreasing the L_2 inductance, increasing the turn ratio n , or improving the coupling factor k , which are all impractical or undesirable.

B. Extended transformer model

One of the important extensions to the simple transformer model is bondwire inductance and resistance. The bondwire inductances act in the same way as imperfect transformer coupling k , lowering the resonance frequency in $|H_{VI}(\omega)|$ and lowering the bandwidth of $|H_{VV}(\omega)|$.

The transformer input inductance value L_1 is very low ($L_1 = L_2/n^2$), hence the transformer input is sensitive to additional bondwire inductance. Fig. 9 shows the $|H_{VI}(\omega)|$ flatness and the values of $|H_{VV}(\omega)|$ at the extremes of the band of interest, as a function of L_{bond1} . For higher L_{bond1} values, the $|H_{VV}(\omega)|$ bandwidth and amplitude degrades, and the resonance peak in $|H_{VI}(\omega)|$ shifts in the band of interest, causing a poor $|H_{VI}(\omega)|$ flatness. Therefore, L_{bond1} should be less than 0.15nH, which is not practical. Hence, transformers for high frequency and with high turn ratios (i.e. small inductance at the transformer input) can only be implemented on-chip, resulting in $L_{bond1} \approx 0$.

Fig. 9. Transformer performance degradation due to L_{bond1}

A bondwire at the transformer output with achievable values for the inductance and resistance ($L_{bond2}=0.5$ nH, $R_{bond2}=1\Omega$) does not degrade the transformer performance significantly.

The required parasitic coil resistance and capacitance for performance preservation are not critical. The following values are assumed: $C_1=100$ f, $C_2=200$ fF, $C_{ps}=200$ fF, $R_1=1\Omega$, $R_2=4\Omega$.

C. Transformer and Mixing-DAC

The Mixing-DAC model, mentioned in section I, together with the simple and extended transformer model of section V-A and section V-B is used to verify the linearity improvement when using the output transformer on a Mixing-DAC. The IMD3 dependence on n of the Mixing-DAC for the most critical output frequency (4GHz) is shown in Fig. 10.

The signal voltage swing at the load is approximately $1V_{pp}$ for all simulations. For an almost ideal transformer ‘Simple, $k=0.95$ ’, the IMD3 at $n=4$ is -89dBc, 18dB better than the conventional Mixing-DAC. However, the linearity

Fig. 10. Simulated IMD3 for Mixing-DAC with output transformer

degrades when the non-idealities of the transformer are added, specifically at lower values of k . Also k itself strongly influences the linearity. For the simulation setup with insufficient linearity improvement, it can be observed that signal power at the transformer input at multiples of the LO frequency is not attenuated by the transformer because of the limited transformer bandwidth. This unattenuated high frequency signal power causes non-linear distortion in the Mixing-DAC. A higher k increases the bandwidth of the transformer and hence improves the linearity. Using the extended transformer model at $n=4$ and assuming $k=0.9$ is achievable, the resulting IMD3 is -85dBc, 14dB better than the conventional Mixing-DAC.

VI. CONCLUSION

Output related non-linearity is a major concern for highly linear current steering RF-DACs. This paper proposes a novel approach for RF-DAC linearization based on a wideband impedance transformation. Simulations of the presented transformer model clearly indicate that for high-frequency wideband RF-DACs, on-chip transformer implementation is the only viable approach. Higher turn ratios result in a larger voltage gain, and hence a higher linearity improvement, but come at the cost of more power consumption. Simulations of a specific RF-DAC model with an output transformer show that the IMD3 improves by about 14dB. The proposed novel output-transformer based RF-DAC architecture is expected to enable the design of highly linear wideband transmitters.

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