

Codesign of Mixer-VGA Downconverter Blocks

Co-design de blocs abaisseurs pour des mélangeurs VGA

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Abstract—Two radio frequency integrated circuit (RFIC) downconverters are presented, each consisting of an active mixer and a variable gain amplifier (VGA). The downconverter blocks (DCBs) have identical mixing stages but different VGA topologies. The mixers use a commutating switching network and a low-noise RF transconductance stage to reduce the system noise figure (NF). VGA gain control in one DCB (DCB-1) is done using a current-steering approach, while the second DCB (DCB-2) employs a transistor-based attenuator. Both RFICs were designed and fabricated using a standard 130-nm CMOS process and they were tested over the RF band covering 1–6 GHz and at an IF of 140 MHz. Measurements show that the conversion gain (CG) of DCB-1 ranges from 9.6 to 25.3 dB, and for DCB-2, the CG ranges from 10.7 to 23.7 dB. The minimum double-sideband NF for both the downconverters is 3.8 dB but their maximum NFs are different: 1) 14.2 dB for DCB-1 and 2) 11.5 dB for DCB-2.

Résumé—Deux abaisseurs implémentés sur des circuits intégrés-radiofréquence (CI-RF) sont présentés dans cet article. Chacun comporte un mélangeur actif et en un amplificateur à gain variable (AGV). Les blocs d'abaisseurs (BA) ont des modules de mélangeurs identiques, mais avec différentes topologies d'AGV. Les mélangeurs utilisent un réseau de commutation et un étage de transconductance RF à faible bruit pour réduire le bruit du système. La commande de gain d'AGV dans un BA (BA-1) se base sur un contrôle de courant, tandis que le second BA (BA-2) utilise un atténuateur à transistor. Les CI-RF ont été conçus et fabriqués en utilisant un procédé standard CMOS 130 nm et ils ont été testés avec la bande RF 1-6 GHz et à une fréquence intermédiaire (FI) de 140 MHz. Les mesures montrent que le gain de conversion (GC) du BA-1 se trouve dans la plage 9.6 à 25.3 dB, et pour BA-2 de 10.7 à 23.7 dB. La double bande latérale minimale du bruit pour les deux convertisseurs abaisseurs est de 3.8 dB, mais leurs bruits maximums sont différents: 1) 14.2 dB pour BA-1 et 2) 11.5 dB pour BA-2.

Index Terms—Amplifier, attenuator, CMOS, conversion gain (CG), current bleeding, digitally assisted, downconverter, Gilbert cell, IIP₃, low noise, microwaves, mixer, noise canceling, radio frequency integrated circuit (RFIC), receiver, steering, variable gain, variable gain amplifier (VGA).

I. INTRODUCTION

FREQUENCY conversion and amplification are fundamental types of signal conditioning needed in the radio front-end of mobile wireless transceivers. Gain control circuits for high-frequency systems include variable gain amplifiers (VGAs) [1], [2], voltage variable attenuators [3], and, more recently, variable conversion gain (CG) mixers [4]–[8]. It is well known that gain control blocks produce a significant amount of noise because their gain variation is usually accomplished through changes in resistance inside the circuit. Ideas on how to mitigate the impact of gain variations on the noise figure (NF) of VGA circuits are described in [9] and [10].

This paper reports on the design and experimental verification of two mixer-VGA downconverter blocks (DCBs) (Fig. 1) for front-end receiver modules. The downconverters

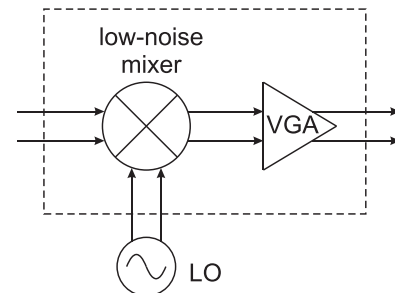


Fig. 1. DCBs in this paper consist of a low-noise mixer and a VGA.

have identical mixing stages but different VGAs at the IF path for gain tuning. To reduce the impact of the VGA gain variation on the NF of the downconverters, a low-noise mixer is used. This approach leads to a minimum double-sideband (DSB) NF of 3.8 dB for each block at the highest gain setting and a maximum DSB NF of 14.2 dB for one downconverter and 12 dB for the downconverter at their lowest gain setting. The DCBs are designed for RF input frequencies in the 1–6 GHz band.

II. RFIC DESIGNS

The mixers and VGAs used to carry out signal conditioning in transceivers are often optimized separately to meet their specifications and they are eventually interconnected in the final stages of the system design. Optimizing the performance

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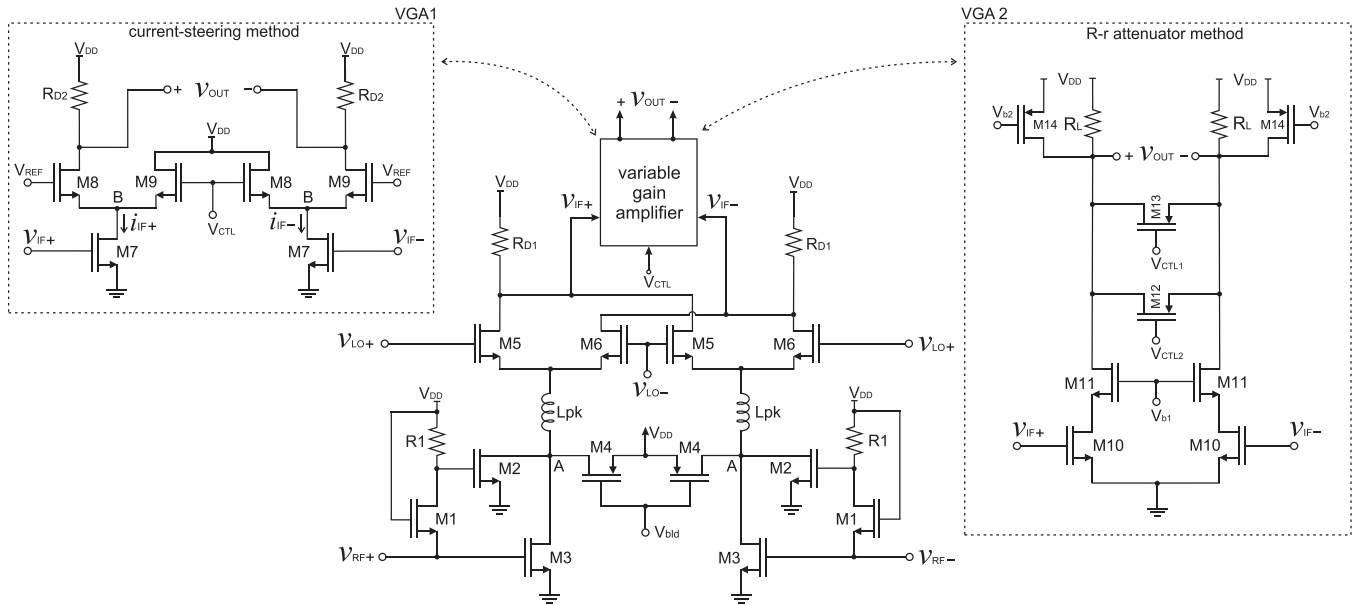


Fig. 2. Schematic of the DCBs.

TABLE I
DEVICE DIMENSIONS AND COMPONENT VALUES FOR THE
DOWNCONVERTER CIRCUIT SHOWN IN FIG. 2

Device	W/L	Device	W/L	Comp.	value
M_1	$\frac{40}{0.12}$	M_8	$\frac{26}{0.12}$	R_1	150 Ω
M_2	$\frac{42}{0.12}$	M_9	$\frac{88}{0.12}$	R_{D1}	150 Ω
M_3	$\frac{132}{0.12}$	M_{10}	$\frac{36}{0.12}$	R_{D2}	200 Ω
M_4	$\frac{80}{0.12}$	M_{11}	$\frac{38}{0.12}$	R_L	300 Ω
M_5	$\frac{80}{0.12}$	M_{12}	$\frac{30}{0.48}$	L_{PK}	238 pH
M_6	$\frac{80}{0.12}$	M_{13}	$\frac{30}{0.48}$		
M_7	$\frac{56}{0.12}$	M_{14}	$\frac{60}{0.12}$		

of the two blocks concurrently can lead to a net overall improvement for some of their metrics when measured as a unified block. This is what we call in this paper the codesign approach. Because certain metrics have an inverse relationship with one another, a tradeoff often has to be made as to which metrics are optimized and which are compromised on. The choice made here is to optimize first and foremost the NF and the gain of the mixer-VGA block.

A schematic of the downconverters is shown in Fig. 2, and the device dimensions and component values are listed in Table I. The mixer circuit has a double-balanced topology and it uses a noise-canceling RF input stage for low-noise performance. Considering the symmetry of the circuit, the transistors and components on the left- and right-hand sides of the mixing stage are labeled identically for simplicity. The VGA in one DCB (DCB-1) employed a current-steering approach [11], while the VGA in second DCB (DCB-2) used the R-r attenuator method [12], [13]. In the remainder of this section, the mixer circuit and the two VGA circuits will be described.

The mixer uses the topology reported in [14] and here only a brief summary of its operation is given. The first device in

the RF stage of the mixer circuit is M_1 , which is biased in common gate mode. The RF signal at the drain of M_1 is in-phase with the input signal v_{RF} , but the noise voltages at the source and drain terminals of the device are out-of-phase. Therefore, as the RF signal propagates through M_2 and M_3 , the signal currents add in-phase at node A. The noise currents, however, subtract out at node A, because they are out-of-phase. Maximum noise cancellation at node A is obtained when the magnitude of the noise currents that emerge from the drains of M_2 and M_3 are as close to each other as possible, meaning that the transconductances of M_2 and M_3 must have the right values. The basic design equation that leads to optimal noise cancellation is $g_{m3} = g_{m2}(R_1/Z_0)$, where g_{m2} and g_{m3} are the transconductances of M_2 and M_3 , respectively, Z_0 is the characteristic impedance at the RF input terminal, and R_1 is the bias resistor at the gate of M_2 .

The noise-canceling circuit is in wide use at present because of its broadband performance and compact design. To supply the noise canceller with the dc bias it needs while keeping the current flow through M_5 and M_6 low enough to facilitate their ON/OFF switching, current bleeding is implemented using transistor M_4 . The inductor L_{pk} is used for gain peaking at high frequencies, thereby helping to extend the 3-dB bandwidth of the mixer.

The switching core's intermediate frequency (IF) load consists of resistor R_{D1} and the variable gain stage. Locating the variable gain circuitry on the mixer's IF path instead of its RF path reduces the degradation of its NF as its CG is lowered, because the RF stage is the most important contributor to the mixer NF. The reason for this is easily deduced from Friis' equation for the NF of a cascaded network, which shows that the gain and noise contribution of the first stage have a dominant impact on the total NF of the network.

Gain control in DCB-1 is implemented through current steering in a differential pair. The IF signal emerging from the mixer's switching core is fed to transistor M_7 , which

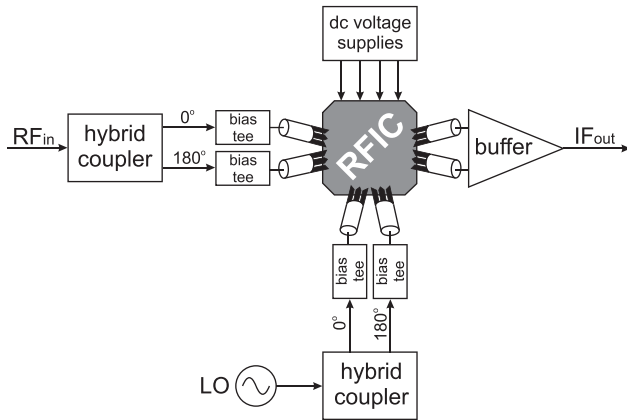


Fig. 3. Experimental test setup.

provides the tail current, i_{IF} , to the differential pair consisting of devices M_8 and M_9 . The voltage V_{ref} remains fixed at all times and is the reference voltage about which the gain control voltage, V_{ctl} , swings. As V_{ctl} changes, the current i_{IF} splits at node B and the proportion of the current flowing through M_8 and M_9 varies. When V_{ctl} decreases, a greater proportion of the current i_{IF} flows through M_8 and the CG of the down-converter increases. Conversely, when V_{ctl} increases, a lesser proportion of i_{IF} flows through M_8 and the CG decreases.

For DCB-2, the gain control is carried out using a cascode differential pair ($M_{10} - M_{11}$) and a variable load network. The load network consists of a fixed resistor R_L and pMOS devices M_{12} and M_{13} . In this method, gain variation is obtained by changing the channel resistance of M_{12} and M_{13} , which are biased in the triode region. The pMOS transistors are used as variable resistors so that negative gate voltages can be used to achieve a larger gain range. Furthermore, pMOS devices have superior low flicker noise character over nMOS devices due to the buried channel behavior [15].

III. EXPERIMENTAL RESULTS

The radio frequency integrated circuits (RFICs) were fabricated using IBM's 130-nm CMOS process and they were tested over the RF band from 1 to 6 GHz and at an IF of 140 MHz. The local oscillator (LO) input frequency was swept to maintain the IF constant. All testing was carried out on-wafer using differential probes at the signal ports. Off-chip power splitters were used to generate the differential RF and LO signals, and a high-linearity active buffer was used at the IF output port to convert the signal from differential mode to single-ended for testing purposes. The experimental test setup is shown in Fig. 3 and microphotographs of the fabricated RFICs are shown in Fig. 4.

A. DCB-1 Results

The voltage CG for DCB-1 was measured as a function of RF input frequency and the gain control voltage, V_{ctl} , and the results are shown in Fig. 5. The highest CG is 25.3 dB, which is obtained at 1 GHz and $V_{ctl} = 0.7$ V. Keeping V_{ctl} fixed at 0.7 V and sweeping the frequency, the gain drops by 3–22.3 dB at 6 GHz. At $V_{ctl} = 1.25$ V, the CG is 12.6 dB

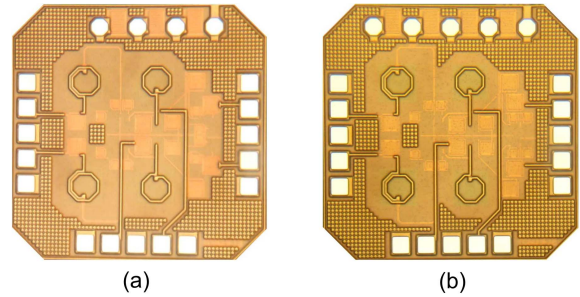
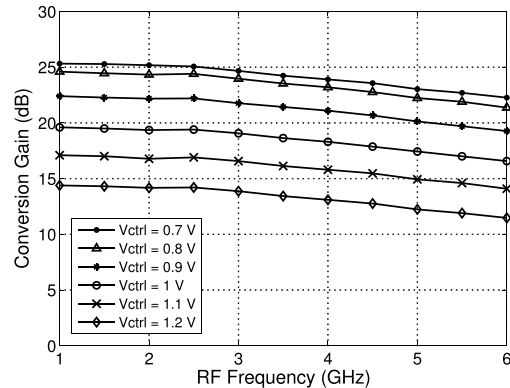
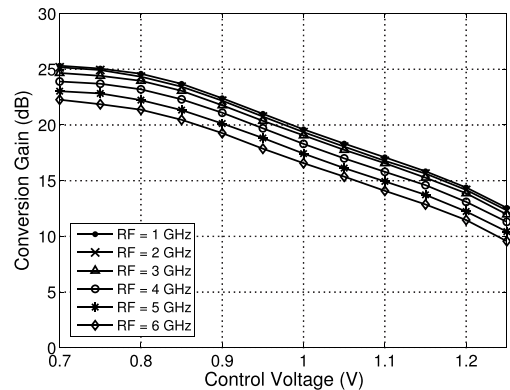


Fig. 4. Fabricated RFICs with (a) current-steering and (b) R-r attenuator gain control.

Fig. 5. Measured voltage CG versus RF and gain control voltage, V_{ctl} .Fig. 6. DCB-1 measured voltage CG versus control voltage, V_{ctl} .

at 1 GHz and drops to 9.6 dB at 6 GHz. Fig. 6 shows the measured voltage CG of the downconverter plotted versus V_{ctl} on the horizontal axis at different test frequencies. The circuit exhibits a smooth gain variation versus control voltage at all measured frequencies.

The downconverter's DSB NF measured as a function of RF input frequency for different values of V_{ctl} is shown in Fig. 7. The lowest NF is 3.8 dB at 1.5 GHz when V_{ctl} is 0.7 V. The NF climbs to 14.3 dB at 6 GHz when $V_{ctl} = 1.25$ V, which corresponds to the lowest CG point.

The downconverter's P_{1dB} was measured as a function of V_{ctl} (i.e., CG) at a representative frequency of 5 GHz. Two-tone tests were also carried out to determine the IP_3 of the mixer. The tones used for the IP_3 measurements were 1 MHz apart and were centered at 5 GHz. The results of those measurements are presented in Fig. 8. The IP_{1dB} ranges

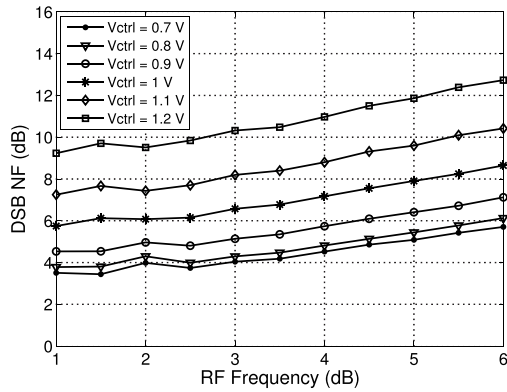


Fig. 7. DCB-1 measured DSB NF versus RF input frequency.

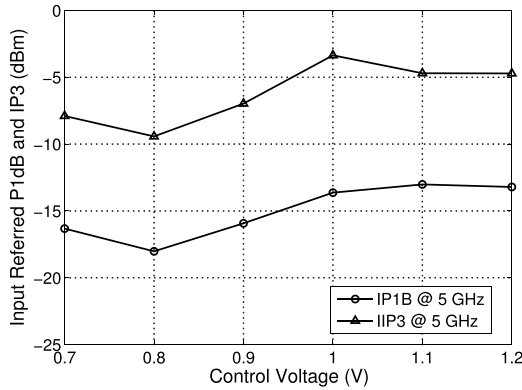
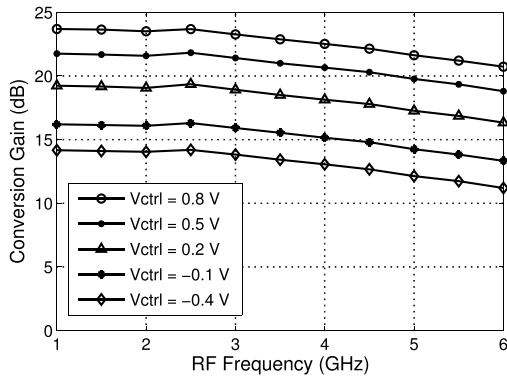
Fig. 8. DCB-1 measured IP_{1dB} and IIP_3 versus gain control voltage.

Fig. 9. DCB-2 measured voltage CG versus RF frequency.

from -17 to -13 dBm and the IIP_3 from -9 to -3 dBm as a function of CG.

The measured reflection coefficient at the downconverter's RF port is between -15 and -21 dB over the 1–6-GHz range. The port-to-port isolations were also measured. The LO-to-RF isolation ranges from -50 dB at 1 GHz to -30 dB at 6 GHz, while the LO-to-IF and the RF-to-IF isolations are better than -50 and -47 dB, respectively, over the band.

B. DCB-2 Results

The DCB-2 circuit has two gain control voltages: 1) V_{ctl1} and 2) V_{ctl2} . Throughout the testing only V_{ctl1} was varied, while V_{ctl2} was fixed at -0.5 V.

The measured voltage CG for DCB-2 as a function of RF input frequency and the gain control voltage, V_{ctl1} , are shown in Fig. 9. When $V_{ctl1} = 0.8$ V, the maximum gain

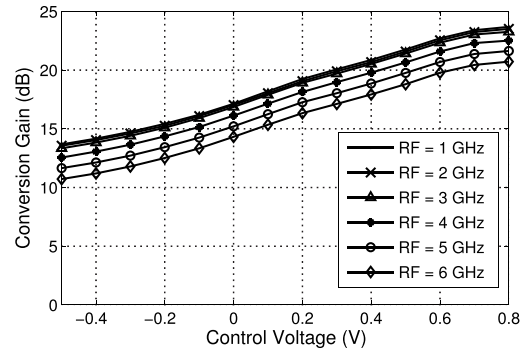


Fig. 10. DCB-2 measured voltage CG versus control voltage.

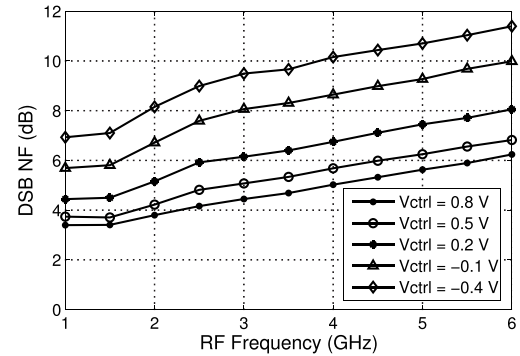


Fig. 11. DCB-2 measured DSB NF versus RF frequency.

is 23.7 dB at 1 GHz and drops gradually to 21 dB at 6 GHz. When $V_{ctl1} = -0.4$ V, the downconverter exhibits a gain of 14 dB at 1 GHz and gradually drops to 10.7 dB at 6 GHz. Fig. 10 shows the gain versus V_{ctl} and, like the previous case, the gain variation is smooth.

The measured DSB NF is shown in Fig. 11 versus RF input frequency at different values of V_{ctl1} . The lowest NF is 3.8 dB when $V_{ctl1} = 0.8$ V and the worst case NF is 11.5 dB at 6 GHz for $V_{ctl} = -0.4$ V.

The IP_{1dB} and IIP_3 of DCB-2 as a function of V_{ctl1} at a test frequency of 5 GHz are shown in Fig. 12. The IP_{1dB} ranges from -6.5 to -13.8 dBm and the IIP_3 ranges from -4 to -5.6 dBm. The IIP_3 response of DCB-2 shows a smaller dependence on the gain than the response of DCB-1.

The measured reflection coefficient at the RF port for DCB-2 is -15 and -20 dB over the 1–6 GHz band. The LO-to-RF isolation for the circuit ranged from -50 dB at 1 GHz to -28 dB at 6 GHz, while the LO-to-IF and the RF-to-IF isolations are better than -54 and -52 dB, respectively, over the band.

C. Comparison Between DCB-1 and DCB-2

A performance comparison between the two DCBs and the other works is presented in Table II. The difference in the minimum CG of DCB-1 and DCB-2 largely explains why DCB-1 has a higher NF_{max} of 14.2 dB, while DCB-2 has an NF_{max} of 12 dB. While DCB-1 has a larger gain variation range of 15.7 dB versus a range of 13 dB for DCB-2, the fact that DCB-2 has a better IP_{1dB} performance and a lower variance in its IIP_3 is due to the fact that its gain control occurs at the resistive network and transistors M_{10} and M_{11} are always on. Meanwhile, in DCB-1, its gain control function

TABLE II
COMPARISON AND PERFORMANCE SUMMARY

	This Work			
	DCB-1	DCB-2	[5]	[6]
RF frequency (GHz)	1–6	1–6	1–12	2–10
IF frequency (MHz)	140	140	110	200
gain range (dB)	9.6–25.3	10.7–23.7	1.2–17	9, 18, 24
DSB NF (min) (dB)	3.8	3.8	11	8
IP1dB [†] (dBm)	-13	-6.5	-3.7	-4
IIP ₃ [†] (max) (dBm)	-3	-4	8.6	3.5
S ₁₁ (dB)	<-15	<-15	-	-
supply voltage (V)	1.5	1.5	1.2	1.2
core area (mm ²)	0.436	0.436	0.105	0.19

[†] measured at 5 GHz

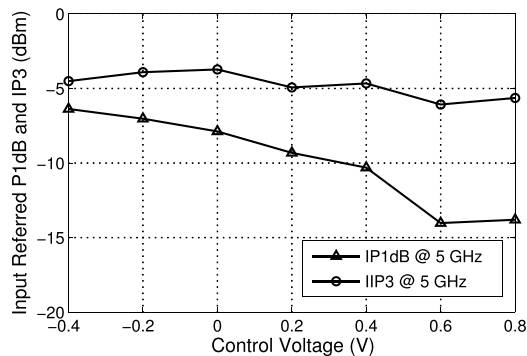


Fig. 12. DCB-2 measured IP_{1dB} and IIP₃ versus control voltage V_{ctl} .

requires devices M_8 and M_9 to swing back and forth from shutoff to triode to saturation to carry out the current steering operation and that has an adverse impact on the linearity of the circuit.

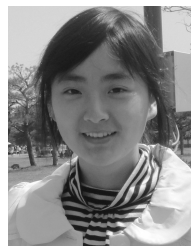
IV. CONCLUSION

The majority of receivers have some form of gain control to compensate for temperature, frequency, and aging effects. In this paper, we have designed two mixer-VGA DCBs in which a noise-canceling mixer is used to mitigate the noise contribution of the VGA. We tested two different VGA topologies, one based on current-steering and another on the R-r attenuator concept, and found that the DCB with the attenuator approach has a better NF and linearity performance than the other DCB.

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