A +10-dBm IIP_3 SiGe Mixer With IM_3 Cancellation Technique

Shoji Otaka, *Member, IEEE*, Mitsuyuki Ashida, Masato Ishii, and Tetsuro Itakura, *Member, IEEE*

***Abstract*—A third-order intermodulation (IM_3) cancellation technique using a submixer is proposed for a low-power low-distortion mixer. The IM_3 cancellation is achieved by summing IM_3s generated in a main mixer and the submixer, which are almost the same amplitude and opposite phase. The mixer was designed to operate at 870 MHz. The proposed technique reduces IM_3 by 18 dB with a current increase of about 15% and is suitable for low-power applications. The mixer achieved an input-referred third-order intercept point (IIP_3) of 10 dBm, a gain of 8.7 dB, and an NF of 9.8 dB and dissipates 30 mW from 2.9 V. The IC is fabricated in a SiGe bipolar transistor with $f_T = 30$ GHz.The IC occupies 1.44 mm × 1.44 mm.**

***Index Terms*—Active mixer, code-division multiple access (CDMA), intermodulation, linearization, phase shifter, third-order distortion.**

I. Introduction

SINGLE-CHIP receiver ICs have been recently developed with silicon technology for low-cost terminals [1]–[3]. The single-chip receiver usually adopts a differential circuit topology regardless of receiver architectures such as direct-conversion architecture, sliding-IF architecture, and low-IF architecture, because front-end circuits including baseband circuits need to suppress common-mode undesired signals and second-order nonlinearity. Thus, a double-balanced mixer (DBM) is adopted as a front-end RF mixer for the single-chip receiver ICs. Furthermore, the DBM has advantages for suppressing spurious at the local oscillator (LO) frequency which would degrade the IF and baseband dynamic range.

Wireless terminals need to have low-noise performance and suppress odd-order nonlinearities as well as second-order nonlinearity and the spurious at the LO frequency so as to receive small desired signals even with strong interference. So, a low-noise amplifier (LNA) and the RF mixer must be designed to achieve low noise and high linearity. In particular, the RF mixer must satisfy a high third-order intercept point (IP_3) in code-division multiple-access (CDMA) systems, where the intermodulation specifications are challenging [4].

The RF mixer is also required to operate with low power for extending the phone talk time. However, the power consumption cannot be easily reduced, because there is a tradeoff between linearity and power consumption. Thus, low-power linearization techniques such as a low-impedance termination technique [5]–[7] and a feedforward linearization technique [8]–[10] have been proposed so far.

The low-impedance termination technique reduces third-order intermodulation (IM_3), which is caused by a product of a fundamental signal and a second-order distortion component in this case. IM_3 is reduced by setting an input terminal to low impedance at around dc and at two times of RF $(2f_{RF})$ to suppress the second-order distortion at the input terminal [7]. This technique is simple and works well. However, this requires on-chip LC resonators at $2f_{RF}$, which need large chip area.

The feedforward technique has a main path and an IM_3 cancellation path and cancels IM_3 by summing two outputs which have the same amplitude and the opposite phase to each other in IM_3s. In the following, IM_3 cancellation path is referred to as a subpath. Among the feedforward techniques, a derivative superposition technique is well known and uses two FETs operated in different bias conditions, where third-order derivatives of the two FETs are opposite phase to each other for cancellation of third-order distortion [8], [9]. However, this technique is sensitive to FETs' characteristics and requires a precise bias tuning to achieve sufficient cancellation of IM_3.

The other feedforward technique uses amplifiers and a divider in the subpath without making use of a FET's characteristics [10]. A proper bias current can be easily set by this technique. A sufficient improvement of input-referred IP_3 (IIP_3) has been experimentally confirmed. However, the technique consumes large power in the subpath. So, a power reduction scheme is required for the feedforward technique of this type.

This paper describes an IM_3 cancellation technique suitable for a low-power mixer, which is one of the feedforward techniques. The proposed technique is applied to DBM with SiGe BiCMOS technology. In Section II, a tradeoff between linearity and power consumption is reviewed for a differential amplifier used in the DBM. In Section III, we discuss a basic idea for a low-power IM_3 cancellation technique and show the required dc current for the subpath can be made sufficiently small. A circuit design of the IM_3 cancellation mixer in the receiver chain is described in Section IV, where a low-noise technique is introduced to the IM_3 cancellation mixer. Section V shows the measured results of the IM_3 cancellation mixer mounted on a circuit board. Section VI summarizes the proposed IM_3 cancellation technique.

Manuscript received April 15, 2004; revised July 8, 2004.

S. Otaka and T. Itakura are with the Corporate Research and Development Center, Toshiba Corporation, Kawasaki 212-8582, Japan (e-mail: shoji.ootaka@toshiba.co.jp).

M. Ashida is with the Semiconductor Company, Toshiba Corporation, Kawasaki 212–8520, Japan.

M. Ishii is with the Semiconductor Company, Toshiba Corporation, Yokohama 247–8585, Japan.

Digital Object Identifier 10.1109/JSSC.2004.836331

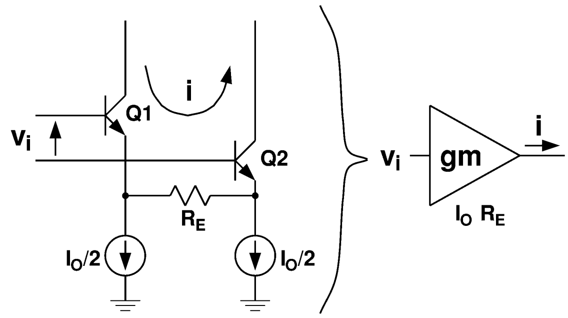


Fig. 1. Emitter-degenerated bipolar differential pair.

II. Tradeoff Between Linearity and Power Consumption

The mixers are categorized into passive mixers and active mixers. This paper focuses on active mixers or Gilbert-type DBMs, because the active mixers have a gain of larger than 0 dB and relax noise performance of the following analog circuit blocks. The DBM comprises a transconductance (gm) amplifier as an input stage and a current-switching stage. The linearity of the DBM mainly depends on that of the gm amplifier [11]. So, the linearity of the gm amplifier is considered. In a bipolar transistor (BJT) differential pair with degeneration resistor R_E, as shown in Fig. 1, an output current i is given by the following expression [12], [13]:

$$i = \frac{I_O/2}{\left(1+\frac{I_O R_E}{4V_T}\right)}\left(\frac{v_i}{2V_T}\right) - \frac{I_O/6}{\left(1+\frac{I_O R_E}{4V_T}\right)^4}\left(\frac{v_i}{2V_T}\right)^3 + \cdots \tag{1}$$

where I_O is a tail current of the differential pair, V_T is thermal voltage, and v_i is an input voltage. From (1), a current ratio of IM_3 (i_3) to fundamental element (i_1) is given as follows:

$$\frac{i_3}{i_1} \propto \frac{v_i^2}{I_O^3(4V_T/I_O + R_E)^3} \tag{2}$$

where i_1 is the first term and i_3 is proportional to the second term of (1), respectively. $4V_T/I_O + R_E$ means an equivalent resistance for emitter degeneration. Assuming that $R_E \gg 4V_T/I_O$, from (2), a current ratio of i_3/i_1 is approximately given as follows:

$$\frac{i_3}{i_1} \propto \frac{v_i^2}{(I_O R_E)^3}. \tag{3}$$

Equation (3) indicates that a ratio of i_3/i_1 decreases by 3 dB as I_O increases by 1 dB. Therefore, an IIP_3 is improved by 1.5 dB as the tail current I_O is increased by 1 dB, i.e., $\{(\mathrm{IIP}_3 + \Delta\mathrm{IIP}_3)/\mathrm{IIP}_3\}$ in dB$/\{(I_O + \Delta I_O)/I_O\}$ in dB $= 1.5$. Here, the figure of merit for IIP_3 is defined as follows:

$$\mathrm{FOM}_{\mathrm{IIP3}} \equiv \frac{\{(\mathrm{IIP}_3 + \Delta\mathrm{IIP}_3)/\mathrm{IIP}_3\} \text{ in dB}}{\{(\mathrm{I_O} + \Delta I_O)/I_O\} \text{ in dB}}. \tag{4}$$

To obtain higher $\mathrm{FOM}_{\mathrm{IIP3}}$, the IM_3 cancellation technique of the feedforward type has been studied for LNA as shown in Fig. 2 [10]. The subpath comprises the identical gm amplifier used in the main path, an additional amplifier having gain of 2, and a $1/8$ current divider. An IM_3 current of the subpath, i_{3S}, becomes almost the same as that of the main path, i_{3M}, from (1). The IM_3 cancellation is achieved by summing two IM_3s which are almost the same amplitude and opposite phase. An $\mathrm{FOM}_{\mathrm{IIP3}}$ is improved by 2.2 in [10], and this technique achieves lower power than the simple technique with a current increase only.

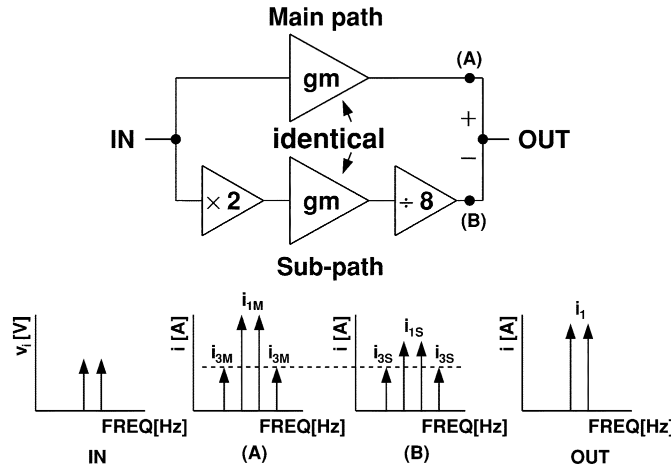


Fig. 2. Conventional IM_3 cancellation technique.

A fundamental signal is reduced by about 2.5 dB, because an amplitude of a fundamental signal in the subpath is a quarter of that in the main path. However, the reduction of the fundamental signal should be small for low power.

An input-referred 1-dB compression point $(\mathrm{IP}_{1\,\mathrm{dB}})$ is also an important specification as is the IIP_3 from the perspective of linearity. $\mathrm{IP}_{1\,\mathrm{dB}}$ expresses large signal nonlinearity. However, an equation for weak signal nonlinearity such as (1) is commonly used for an approximate analysis of $\mathrm{IP}_{1\,\mathrm{dB}}$. $\mathrm{IP}_{1\,\mathrm{dB}}$ is approximately expressed as a function of IIP_3 [13] as follows:

$$\mathrm{IP}_{1\,\mathrm{dB}} \approx \mathrm{IIP}_3 - 9.64\,(\mathrm{dB}) \tag{5}$$

where $\mathrm{IP}_{1\,\mathrm{dB}}$ is estimated by subtracting a third-order distortion from a linearly amplified signal at an applied frequency. Thus, $\mathrm{IP}_{1\,\mathrm{dB}}$ is approximately proportional to IIP_3 because both IIP_3 and $\mathrm{IP}_{1\,\mathrm{dB}}$ are mainly determined by third-order distortion.

The conventional feedforward technique has $\mathrm{IP}_{1\,\mathrm{dB}}$ that is 9 dB smaller compared to the simple technique with a current increase when the same currents are consumed in the two techniques, because the conventional feedforward technique consumes twice the dc current of a single LNA at least. This result in an unacceptable power loss.

Our goal is to reduce power consumption for improving IIP_3 with no degradation of $\mathrm{IP}_{1\,\mathrm{dB}}$, considering that the mixer is applied to CDMA systems for which the IIP_3 requirement is more severe than $\mathrm{IP}_{1\,\mathrm{dB}}$. In this case, $\mathrm{IP}_{1\,\mathrm{dB}}$ of the mixer of our goal approaches that of the simple technique when the same currents are consumed in the two techniques.

III. IM_3 Cancellation Technique

The issue concerning the conventional feedforward technique is overcome by reducing dc current consumed in the subpath. Fig. 3 shows the proposed IM_3 cancellation concept

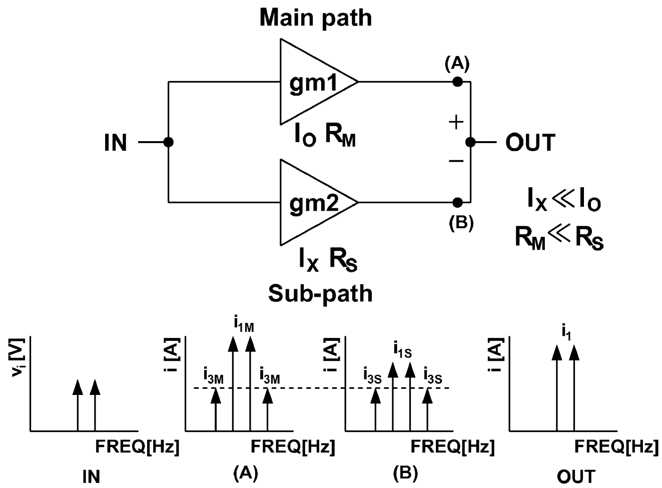


Fig. 3. Concept of the proposed mixer with IM$_3$ compensation.

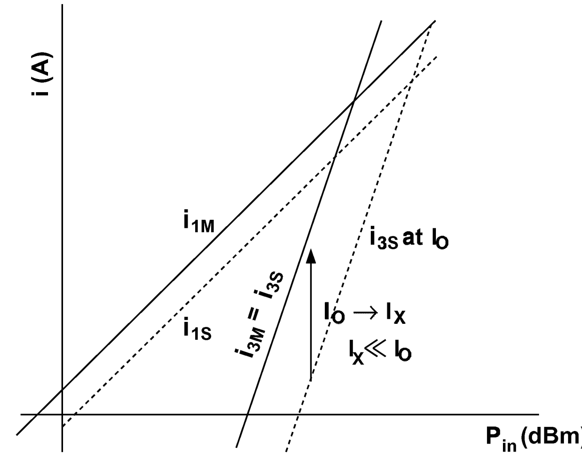


Fig. 4. Input power dependence of fundamental signals and IM$_3$ for both the main path and the subpath.

for achieving both low power and high linearity. A $gm2$ amplifier is equipped to remove IM$_3$ generated in a $gm1$ amplifier so as to keep IM$_3$ of the $gm2$ amplifier at the same amplitude with opposite phase compared to that of the $gm1$ amplifier, where a dc current of the $gm2$ amplifier is set to much less than that of the $gm1$ amplifier for low-power operation. The proposed technique improves FOM$_{\mathrm{IIP3}}$ compared to the conventional technique, in principle. The following discusses the design of the IM$_3$ cancellation mixers.

Fig. 4 shows input power dependence of the fundamental signal and IM$_3$ both for the main path and the subpath. For the main path, the fundamental signal and IM$_3$ are depicted by the solid line as i_{1M} and i_{3M}, respectively. Here, $R_M \gg 4V_T/I_O$ is assumed to avoid cumbersome analysis for the IM$_3$ cancellation concept. A current ratio of i_3/i_1 is approximately expressed by $K/(I_O R_M)^3$ from (3), where K is constant and R_M is an emitter degeneration resistance in the main path.

A current ratio i_3/i_1 is considered for the $gm2$ amplifier with the emitter degeneration resistor R_S set to αR_M, where $\alpha \geq 1$. In the case that the $gm2$ amplifier flows the same current I_O as the $gm1$ amplifier, the current ratio i_3/i_1 is $K/(\alpha I_O R_M)^3$. Thus, in the subpath of this case, the fundamental signal and IM$_3$ are depicted by the dotted line as i_{1S} and i_{3S} at I_O, respectively. IM$_3$ in the subpath needs to be increased so that the IM$_3$s in both the main path and the subpath are the same in order to cancel IM$_3$, i.e., i_{3S} at I_O should be increased up to i_{3M}. For this purpose, a dc current of the $gm2$ amplifier is reduced. This is effective for achieving lower power.

For proper IM$_3$ cancellation, i_3/i_1 of the subpath is set to a larger ratio than that of the main path by a ratio of the transconductance of the main path to the transconductance of the subpath, i.e., α in this case. Thus, the dc current I_X of the $gm2$ amplifier should be set as follows:

$$\frac{K}{(I_O R_M)^3} = \frac{1}{\alpha} \times \frac{K}{(\alpha I_X R_M)^3} \tag{6}$$

where $K/(I_O R_M)^3$ and $K/(\alpha I_X R_M)^3$ are i_3/i_1 of the main path and the subpath, respectively. $1/\alpha$ is a ratio of the subpath transconductance to that of the main path. From (6), we have

$$I_X = \alpha^{-4/3} I_O. \tag{7}$$

An accurate bias current setting can be determined by a ratio of the emitter degeneration resistances.

The transconductance in the subpath should be smaller than that in the main path for a small gain reduction. The prior approach shown in Fig. 2 has a gain reduction of 2.5 dB. In our design objective, the gain reduction is within 1 dB. For this purpose, a ratio of the transconductances, α, is set to about 10. In this case, I_X is reduced to approximately 0.05 I_O from (7).

An input range of the gm amplifiers in both paths is proportional to a product of the emitter degeneration resistance and the dc current. From (7), the product of the $gm2$ amplifier, $R_S I_X$, is given as follows:

$$R_S I_X = \alpha^{-1/3} R_M I_O. \tag{8}$$

The $gm2$ amplifier has an input range $\alpha^{-1/3}$ lower than the $gm1$ amplifier. So, in the case of $\alpha = 10$, the input range of the $gm2$ amplifier is half that of the $gm1$ amplifier, because $R_S I_X$ is approximately $0.5 R_M I_O$.

Fig. 5 shows an RF mixer for applying the IM$_3$ cancellation technique. The IM$_3$ cancellation mixer comprises a main gm path and a sub-gm path which are connected in parallel before common current switches Q10–Q13. The main gm path comprises a differential amplifier Q1–Q2 with an emitter degeneration of R_M and two tail current sources of $I_O/2$. The sub-gm path is composed of a differential amplifier Q3–Q4 with an emitter degeneration of R_S and two tail current sources of $I_X/2$. The tail current I_X should be set to $0.05 I_O$ for $R_S = 10 R_M$ as described above.

IV. Circuit Design for Low-Noise IM$_3$ Cancellation Mixer

Low-noise performance is required for the RF mixer in a receiver chain. In this section, a low-noise technique is described for the proposed IM$_3$ cancellation mixer.

A. IM$_3$ Cancellation Mixer With Emitter Degeneration Inductor

For low-noise performance, the emitter degeneration resistors of the main DBM and the sub-DBM would be replaced by

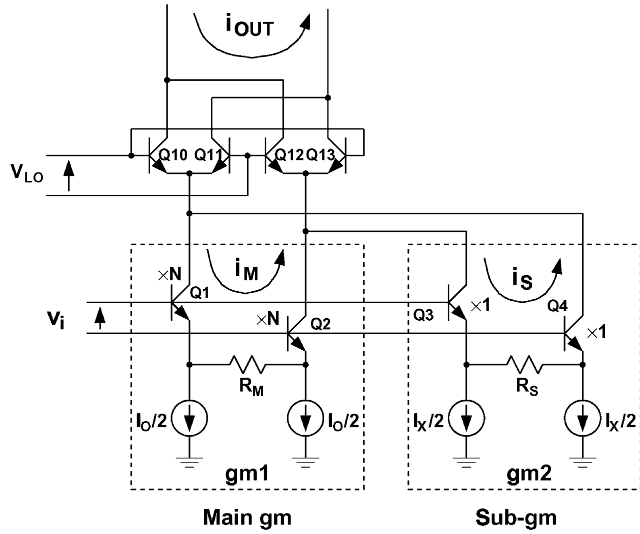


Fig. 5. IM_3 cancellation mixer with linearized transconductance amplifier.

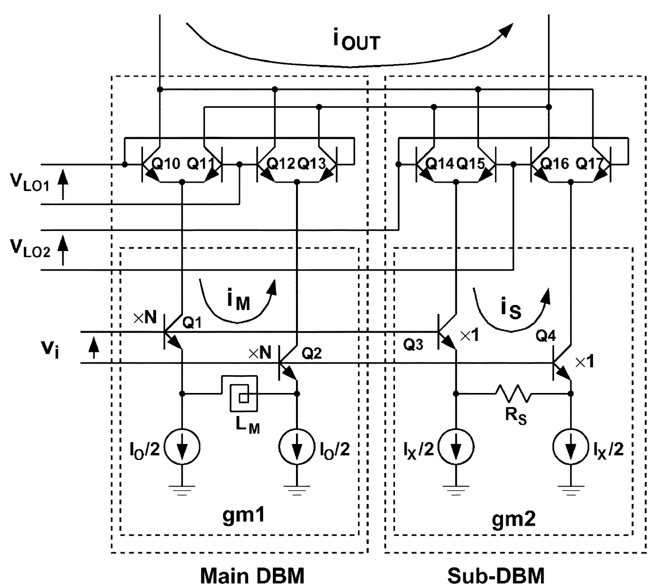


Fig. 6. Low-noise IM_3 cancellation mixer.

inductors. However, the emitter degeneration inductor L_S occupies a large area for the sub-DBM, because L_S is required to be ten times an inductance in the main DBM for the approximate analysis described in Section III. So, the emitter degeneration resistor R_M is only replaced by the emitter degeneration inductor L_M for the main DBM, while the emitter degeneration resistor R_S is used in the sub-DBM to avoid occupying a large area. This change of the emitter degeneration inductor results in a phase shift of the IM_3 in the main DBM. LO signals with different phases are applied to the main and the sub-DBM as described in Section IV-B, in order to compensate for the phase shift of IM_3. Fig. 6 shows a low-noise version for the proposed IM_3 cancellation mixer.

The phase shift is estimated from (1). An output current i_M of the differential pair for the main DBM is rewritten as follows:

$$i_M = i_{1M} - a_{3M} i_{1M}^3 + \cdots \tag{9}$$

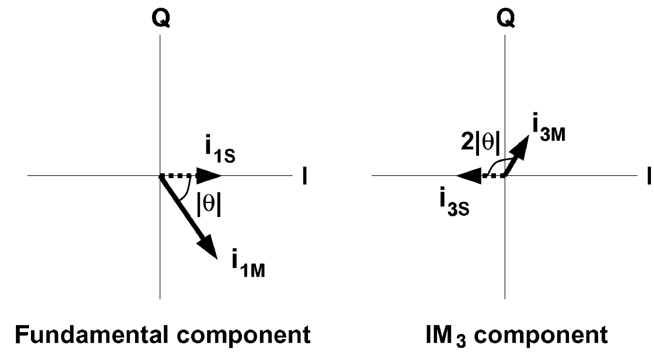


Fig. 7. Phases of fundamental and IM_3 currents in the main DBM and sub-DBM.

where $i_{1M} = v_i \times (4/gm_M + j\omega L_M)^{-1}, a_{3M} = (16V_T/3I_O^3) \times (4/gm_M + j\omega L_M)^{-1}, gm_M = I_O/V_T$. $(4/gm_M + j\omega L_M)^{-1}$ means a transconductance of the differential amplifier, and i_{1M} is a fundamental current for the differential amplifier. When two-tone voltage signals at angular frequencies ω_1, ω_2 are applied, fundamental signal currents are phase-shifted by $\theta_1 = -\tan^{-1}(gm_M\omega_1 L/4)$ and $\theta_2 = -\tan^{-1}(gm_M\omega_2 L/4)$ at ω_1 and ω_2, respectively. For a usual condition of $\omega_1 \approx \omega_2$, θ_1 is approximately θ_2. The phase shift of IM_3 at $\omega_3 = 2\omega_1 - \omega_2$ is $\theta_3 \approx \theta_1 - \tan^{-1}(gm_M\omega_3 L/4)$. The first term is due to third-order distortion for the current i_{1M} having two-tone signals, and a second term is due to third-order coefficient a_{3M}. θ_3 is approximately expressed by 2θ for $\omega_1 \approx \omega_3, \theta = \theta_1 \approx \theta_3$.

On the other hand, an output current i_S of the differential amplifier for the sub-DBM is expressed by

$$i_S = i_{1S} - a_{3S} i_{1S}^3 + \cdots \tag{10}$$

where $i_{1S} = v_i \times (4/gm_S + R_S)^{-1}, a_{3S} = (16V_T/3I_X^3) \times (4/gm_S + R_S)^{-1}, gm_S = V_T/I_X$. $(4/gm_S + R_S)^{-1}$ means a transconductance of the differential amplifier, and i_{1S} is a fundamental current for the differential amplifier. In this case, fundamental currents are not phase-shifted and the phase of the IM_3 is $-\pi$. Fig. 7 depicts phases of fundamental signal currents i_{1M}, i_{1S}, and IM_3 currents and i_{3M}, i_{3S} for the main DBM and the sub-DBM, respectively.

In our design, the emitter degeneration inductance L_M of 9.6 nH and the tail current I_O of 7 mA are set in the main DBM for low noise and high linearity, where a resistance of 15 Ω is parasitic to the emitter degeneration inductor in series. The emitter degeneration resistance R_S of 190 Ω and the tail current I_X of 600 μA are set in the sub-DBM to give an IM_3 cancellation. From (9) with a series resistance of L_M, the fundamental current i_{1M} of the main differential amplifier has a phase lag of -60° compared to the fundamental current i_{1S} of the subdifferential amplifier. The IM_3 current of the main differential amplifier has a phase lag 2θ of -120° compared to that of the subdifferential amplifier. Fig. 8(a) shows simulated results of the input power dependences of the phase difference $\theta_S - \theta_M$ between the main differential amplifier and the subdifferential amplifier, where θ_S and θ_M are phases of currents of the subdifferential amplifier and the main differential amplifier, respectively. The main differential amplifier has phase lags of -50° and -100°

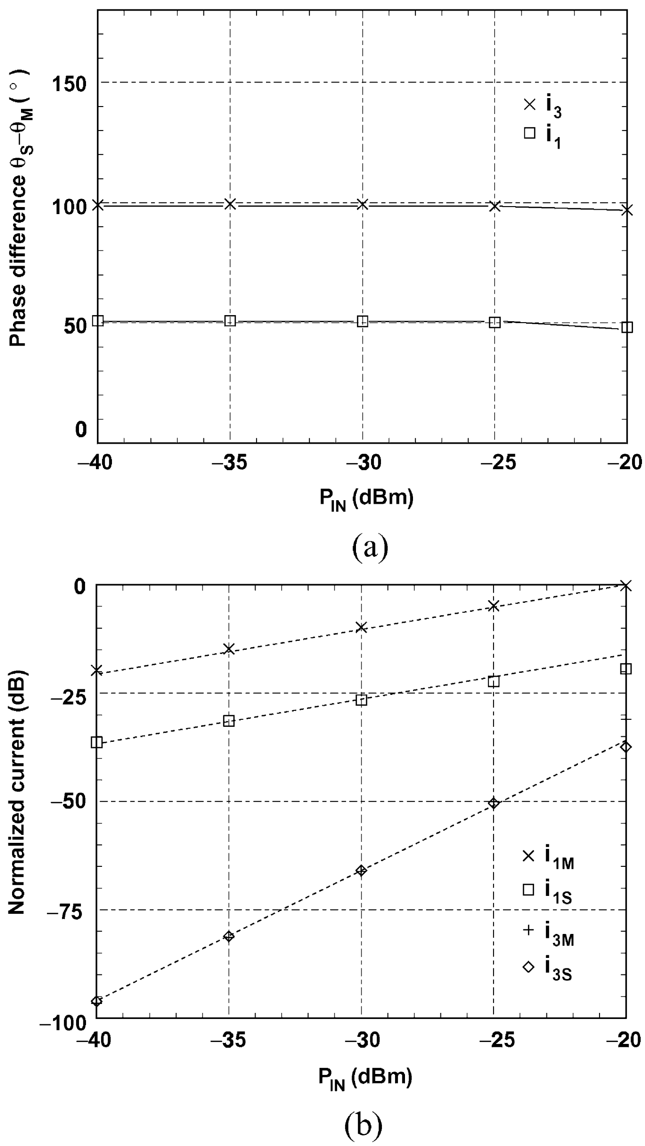


Fig. 8. (a) Simulation results of input power dependence on the phase difference $\theta_S - \theta_M$. θ_M is the phase of current in the main path. θ_S is the phase of current in the subpath. (b) Simulated results of input power dependence on current amplitudes of fundamental component and IM$_3$ in the main path and the subpath.

in the fundamental and IM$_3$ currents, respectively, compared to the subdifferential amplifier.

Fig. 8(b) shows simulated results of the input power dependence of normalized current amplitudes of the fundamental signal and IM$_3$ for both the main and the subdifferential amplifiers. Amplitudes of IM$_3$ are almost the same in the main and the subdifferential amplifiers. A ratio of i_{1M}/i_{1S} is about 7 or 17 dB. These results indicate that a phase shift of $-100°$ is needed for i_{3S} to achieve the IM$_3$ cancellation without a large reduction of the gain, considering that outputs of the two DBMs are connected with opposite polarity.

B. Phase Shifter for Low-Noise IM$_3$ Cancellation

There are two methods to achieve a phase shift IM$_3$ which facilitates an IM$_3$ cancellation. One is to phase shift the RF signal, while the other is to phase shift the LO signal. The LO phase-shift method is chosen because a signal path loss due to a phase shifter should be avoided to achieve low-noise performance [14]. Fig. 9 shows the proposed low-noise IM$_3$ cancellation mixer with a phase shifter equipped in the LO signal path. Fig. 10(a) shows the phase shifter using *RC* bridge circuits, where amplitudes of $V_{\rm LO1}$ and $V_{\rm LO2}$ are the same for all frequencies, but a phase difference between those depends on the frequency and is expressed by $-2\tan^{-1}\{1/(\omega RC)\}$ [15]. The *RC* phase shifter in the LO signal path makes the IM$_3$ current i_{3S} of the sub-DBM phase-shifted by $-100°$, as shown in Fig. 10(b) as well as fundamental current i_{1S}. Therefore, downconverted components of i_{3M} and i_{3S} are summed in opposite phase to each other. On the other hand, a phase difference θ between downconverted fundamental signals i_{1M} and i_{1S} is $-50°$. This effect is described in the next section.

The LO buffer is comprised of a differential amplifier and loads in which on-chip inductors and the *RC*-type phase shifters are connected in parallel, as shown in Fig. 10(a). The drive current of the LO buffer needs to be increased for the phase shifters and the sub-DBM. The LO buffer consumes 2 mA for driving the main DBM, the sub-DBM, and the phase shifters. In our estimation, the additional current of 0.8 mA is required for driving the phase shifter and the sub-DBM to keep a sufficient amplitude for the current switch of the DBMs.

C. Estimation of the Gain Reduction

In our design objective, the gain reduction is less than 1 dB. In the actual design described above, a ratio of the transconductance of the main differential amplifier to that of the subdifferential amplifier is about 7 in simulation. This means that the gain reduction is 1.3 dB if the fundamental signals of both the main DBM and sub-DBM are of opposite phase after phase shifting. However, the gain reduction is within 1 dB since the phase difference θ between downconverted fundamental signals i_{1M} and i_{1S} is $-50°$. Hence, a ratio of the fundamental currents becomes approximately $1 : (1/7 \times \cos 50°) \approx 1 : 0.09$ in this case.

D. Estimation of the Process Variation

Process variation affects the IM$_3$ cancellation. The relation between process variation and the expected IM$_3$ cancellation is described in the following.

The absolute resistance variation causes an amplitude difference between the IM$_3$s of the main DBM and the IM$_3$s of the sub-DBM. When an absolute resistance increases to α_R times by process variation, dc currents, I_O and I_X, are assumed to be $(\alpha_R)^{-1}$ times. In this case, the IM$_3$s of the main DBM and sub-DBM, i_{3M} and i_{3S}, are expressed from (9) and (10) as follows:

$$i_{3M} = k\left(\frac{\alpha_R}{I_O}\right)^3\left(\frac{4\alpha_R V_T}{I_O} + R_{ML} + j\omega L_M\right)^{-4} \quad (11)$$

$$i_{3S} = k\alpha_R^{-1} I_X^{-3}\left(\frac{4V_T}{I_X} + R_S\right)^{-4} \quad (12)$$

where R_{ML} is the parasitic resistance of the inductor. In our design, i_{3M} approximately increases to α_R^3 times and i_{3S} decreases to α_R^{-1} times. A residual IM$_3$ becomes $(\alpha_R^3 - \alpha_R^{-1})i_{3M,O}$, where $i_{3M,O}$ is i_{3M} without absolute

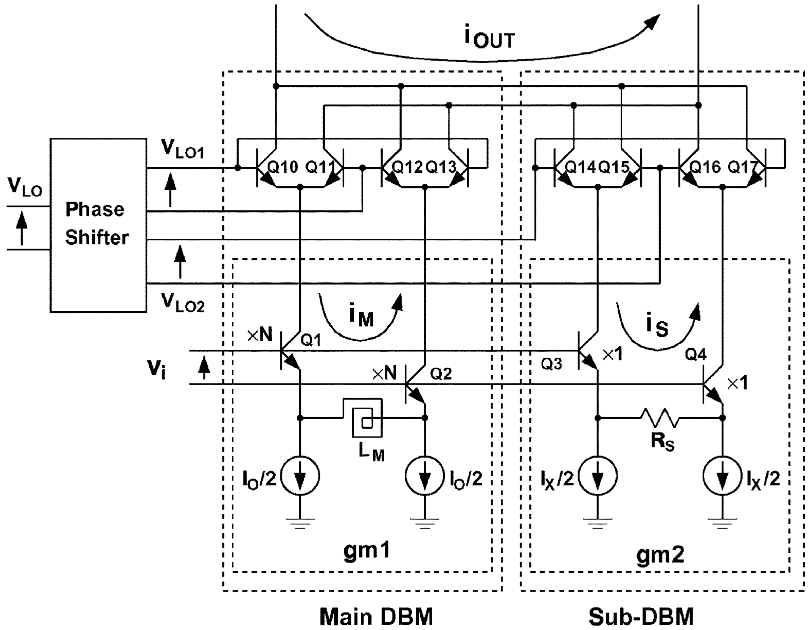
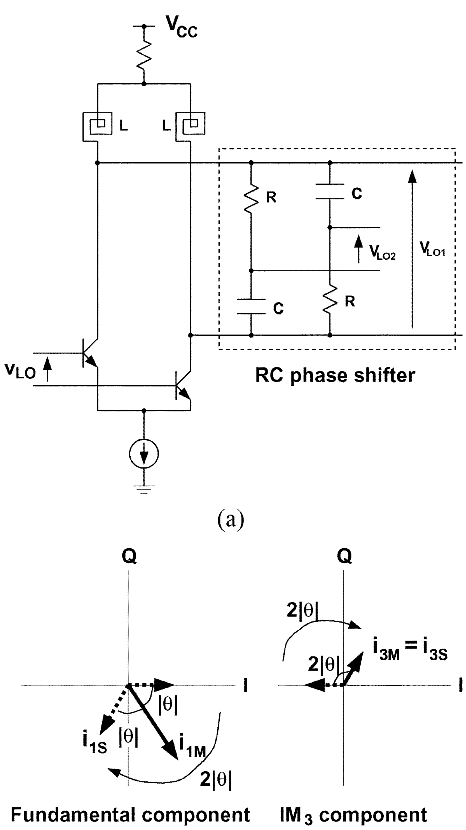


Fig. 9. Low-noise IM_3 cancellation mixer with an LO phase shifter.



(b)

Fig. 10. (a) Phase shifter with the LO buffer. (b) Phase shift for a fundamental component and IM_3 in the subpath.

resistance variation, i.e., at $\alpha_R = 1$. The IM_3 cancellation is limited by $20\log(\alpha_R^3 - \alpha_R^{-1})$ in dB. For IM_3 cancellation of over 20 dB, $0.97 \le \alpha_R \le 1.03$ is required. For $\alpha_R = 1.1$, IM_3 cancellation is limited by 7.5 dB. Fig. 11(a) shows i_{3M}, i_{3S} in this case.

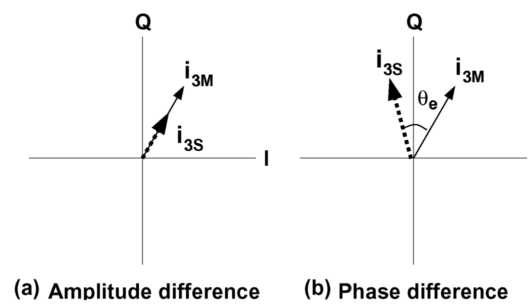


Fig. 11. Amplitude and phase error due to process variation. (a) Amplitude difference. (b) Phase difference.

The variation of the absolute resistance or the absolute capacitance causes a phase difference of the IM_3 between the main DBM and sub-DBM, because the phase shift of the phase shifter depends on the *RC* product. When the absolute resistance and capacitance increase to α_R, α_C times, respectively, the phase difference θ_e is $2|\tan^{-1}(\omega RC)^{-1} - \tan^{-1}(\omega\alpha_R\alpha_C RC)^{-1}|$. Fig. 11(b) shows i_{3M}, i_{3S} at the phase difference θ_e. IM_3 cancellation is limited by $20\log(\sin\theta_e)$ in dB. In the case that $\alpha_R\alpha_C = 1.2, \theta_e = 10.6°$, where $\omega RC = 0.84$ in our design. The IM_3 cancellation is limited by 15 dB in this case.

V. Measured Results

The low-noise IM_3 cancellation mixer was fabricated in a SiGe BJT with f_T =30 GHz. A microphotograph of the IM_3 cancellation mixer is shown in Fig. 12. The die area is 1.44×1.44 mm^2, where the sub-DBM and the phase shifter occupies a small portion of the total area. A plastic package was mounted on a glass-epoxy circuit board for evaluation. The characteristics of the low-noise IM_3 cancellation mixer were measured with an external LO frequency set to 1230 MHz and an IF frequency of 360 MHz with an off-chip matching circuit.

Fig. 13 shows characteristics of fundamental components and IM_3 with and without the IM_3 cancellation technique when

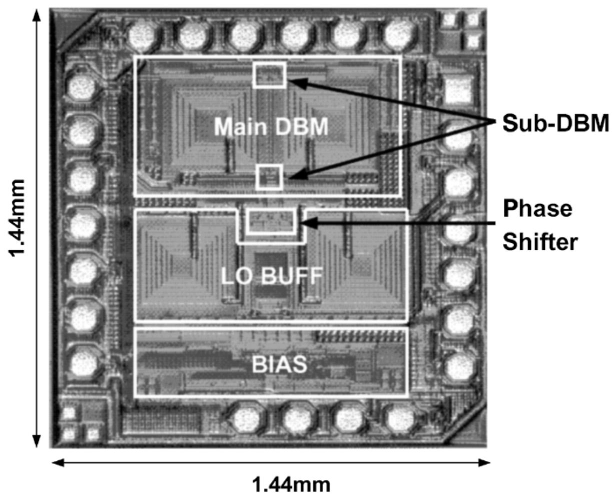


Fig. 12. Die photograph.

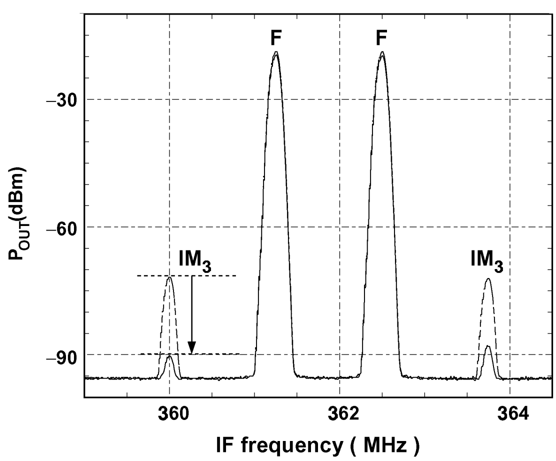


Fig. 13. The fundamental components and IM_3 at mixer output when two tones are applied at 871.25 and 872.5 MHz.

two-tone RF signals of 871.25 and 872.5 MHz are applied to the input. This figure indicates that the proposed IM_3 cancellation mixer successfully cancels the IM_3 by 18 dB. Fig. 14 shows input power dependence of fundamental components and IM_3 with and without the IM_3 cancellation technique. The IM_3 cancellation technique improves IIP_3 by 9 dB compared to the mixer without IM_3 cancellation, while a bias current of the mixer is increased from 10.2 to 10.8 mA. FOM_{IIP3} of 7.5 is obtained, considering the incremental current of the LO buffer for driving the phase shifter and the sub-DBM. The proposed IM_3 cancellation technique has FOM_{IIP3} superior to that of the conventional IM_3 cancellation technique applied to the LNA, of $FOM_{IIP3} = 2.2$ [10]. Table I summarizes a comparison of FOM_{IIP3} for the proposed technique with previously proposed methods. This indicates that the proposed technique improves IIP_3 sufficiently with a minimal current increase.

IM_3 cancellation is degraded for input power of over -22 dBm. This result is due to a limitation on an input range of the sub-DBM, because the product of the emitter degeneration impedance and dc current of the sub-DBM, $R_S I_X$, is set to a smaller value than that of the main DBM. There is a tradeoff between the gain reduction and the input range. In our design,

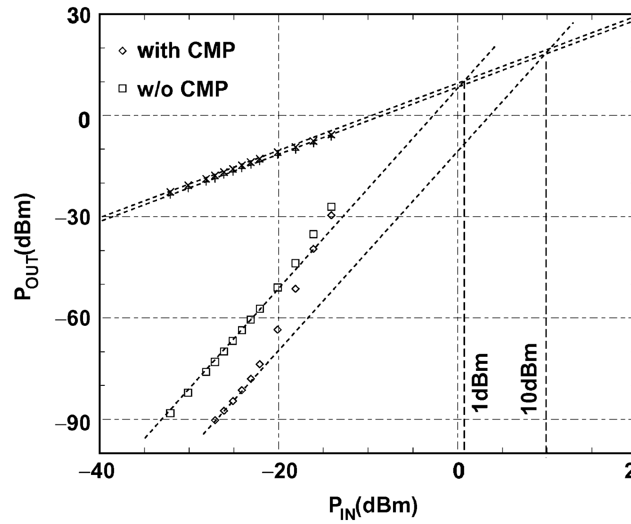


Fig. 14. Input power dependence of the fundamental components and IM_3 when two tones are applied at 871.25 and 872.5 MHz.

TABLE I
COMPARISON OF FOM_{IIP3}

ref.	FOM_{IIP3}
proposed technique	7.5
[10]	2.2
current increase	1.5

the input range of the sub-DBM is estimated to be about 3 dB smaller than that of the main DBM, while the gain reduction is within 1 dB.

A conversion gain of the low-noise IM_3 cancellation mixer is 8.7 dB, which is reduced by 0.9 dB compared to that without the IM_3 cancellation technique. This result agrees with our estimation as described in Section IV-C. The measured single-sideband (SSB) noise figure (NF) is 9.8 dB. The SSB NF is degraded by 1.3 dB compared to that without the IM_3 cancellation technique. The SSB NF degradation is mainly caused by the gain reduction of the IM_3 cancellation technique.

Equations (9) and (10) indicate that the IM_3 cancellation depends on frequency due to the frequency response of the emitter degeneration inductor of the main DBM. Fig. 15(a) shows frequency responses of the IM_3 with and without the IM_3 cancellation technique when two tones are applied at an input power of -26 dBm. For example, when two tones are applied at 880 and 890 MHz, the output of the IM_3 is plotted in Fig. 15(b) on the x axis at 880 MHz. IM_3 is reduced by more than 8 dB in the frequency range from 840 to 900 MHz. This demonstrates that the IM_3 cancellation is achieved over a wide frequency range.

The IC operates from a 2.9-V power supply. The measured $IP_{1\,dB}$ of the mixer with IM_3 cancellation is -10 dBm and is the same as that of the the mixer without IM_3 cancellation. Therefore, this proposed technique does not degrade $IP_{1\,dB}$. The measured results are summarized in Table II.

VI. CONCLUSION

An IM_3 cancellation technique is proposed for low power consumption. The IM_3 cancellation is achieved by summing

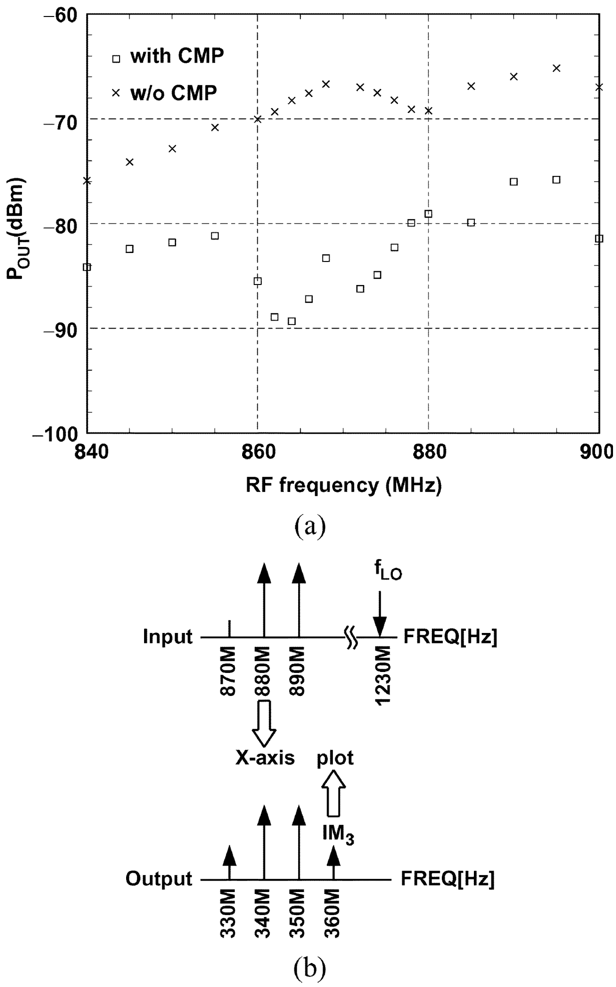


Fig. 15. (a) Frequency responses of IM_3 product with and without IM_3 cancellation. (b) Frequency setting for the frequency response.

TABLE II
SUMMARY OF MEASURED RESULTS

	with cancellation	without cancellation
Supply Voltage	2.9 V	2.9V
Current Consumption	10.8 mA	10.2 mA
RF Frequency	870 MHz	870 MHz
LO Frequency	1230 MHz	1230 MHz
Gain	8.7 dB	9.6 dB
SSB Noise Figure	9.8 dB	8.5 dB
IIP_3	10 dBm	1 dBm
IP_{1dB}	−10 dBm	−10 dBm

IM_3s of both the main gm amplifier and the sub-gm amplifier, which are opposite phase and equal amplitude. Furthermore, relations of the fundamental signal and IM_3 in phase and amplitude are analyzed for this IM_3 cancellation technique.

The IM_3 cancellation technique is applied to an RF mixer. The RF mixer comprises a main DBM and a sub-DBM which are connected in parallel, where the dc current of the sub-DBM is much smaller than that of the main DBM. For low-noise performance and small chip area, an emitter degeneration inductor is adopted in the main DBM, while an emitter degeneration resistor is used in the sub-DBM. An *RC* phase shifter in the LO signal path is introduced by compensating a phase difference of IM_3 between the main DBM and the sub-DBM. An additional area for the sub-DBM and the phase shifter is a small portion of the total area of the mixer.

The measured results indicate that the proposed mixer can reduce IM_3 by 18 dB and improves IIP_3 by 9 dB. The improvement of IIP_3 is confirmed over an applied frequency range of 60 MHz. The proposed technique is applicable toward active current steering balanced mixers.

ACKNOWLEDGMENT

The authors wish to thank T. Kuroda for technical support regarding chip layout and fabrication. They are also grateful to Dr. T. Yamaji and H. Yoshida for valuable discussion on the mixer and the specification.

REFERENCES

[1] J. Jussila, J. Ryynanen, K. Kivekas, L. Sumanen, A. Parssinen, and K. A. I. Halone, “A 22 mA 3.7 dB NF direct conversion receiver for 3 G WCDMA,” *IEEE J. Solid-State Circuits*, vol. 36, pp. 2025–2029, Dec. 2001.

[2] M. Zargari, D. K. Sue, C. P. Yue, S. Rabii, D. Weber, B. J. Kaczynski, S. S. Mehta, K. Singh, S. Mendis, and B. A. Wooly, “A 5-GHz CMOS transceiver for IEEE 802.11a wireless LAN systems,” *IEEE J. Solid-State Circuits*, vol. 37, pp. 1688–1694, Dec. 2002.

[3] F. O. Eynde, J.-J. Schmit, V. Charier, R. Alexandre, C. Sturman, K. Coffin, B. Mollekens, J. Craninckx, S. Terrijn, A. Monterastelli, S. Beerens, P. Goetchackx, M. Ingels, D. Joos, S. Guncer, and A. Pontioglu, “A fully-integrated single-chip SOC for bluetooth,” in *ISSCC 2001 Dig. Tech. Papers*, pp. 196–197.

[4] “C.S0011-B v1.0: Recommended Minimum Performance Standards for cdma2000 Spread Spectrum Mobile Stations,” TIA: TIA-98-E, http://www.3gpp2.com/Public_html/specs/index.cfm#tsgc, 2002/12.

[5] K. L. Fong, “High-frequency analysis of linearity improvement technique of common-emitter transconductance stage using a low-frequency trap network,” *IEEE J. Solid-State Circuits*, vol. 35, pp. 1249–1252, Aug. 2000.

[6] V. Aparin and L. E. Larson, “Linearization of monolithic LNA’s using low-frequency low-impedance input termination,” in *Proc. IEEE Eur. Solid-State Circuits Conf.*, 2003, pp. 137–140.

[7] L. Sheng and L. E. Larson, “An Si-SiGe BiCMOS direct-conversion mixer with second-order and third-order nonlinearity cancellation for WCDMA applications,” *IEEE Trans. Microwave Theory Tech.*, vol. 51, pp. 2211–2220, Nov. 2003.

[8] D. R. Webster, D. G. Haigh, J. B. Scott, and A. E. Parker, “Derivative superposition—A linearization technique for ultra broadband systems,” *Proc. IEE Colloq. Wideband Circuits, Modeling and Techniques*, pp. 3/1–3/14, May 1996.

[9] Y.-S. Youn, J.-H. Chang, K.-J. Koh, Y.-J. Lee, and H.-K. Yu, “A 2 GHz 16 dBm IIP3 low noise amplifier in 0.25 μm CMOS technology,” in *ISSCC 2003 Dig. Tech. Papers*, pp. 452–453.

[10] Y. Ding and R. Harjani, “An +18 dBm IIP3 LNA in 0.35 μm CMOS,” in *ISSCC 2001 Dig. Tech. Papers*, pp. 162–163.

[11] B. Leung, *VLSI for Wireless Communication*. Upper Saddle River, NJ: Prentice-Hall, 2002.

[12] D. Coffing and E. Main, “Effects of offsets on bipolar integrated circuit mixer even-order distortion terms,” *IEEE Trans. Microwave Theory Tech.*, vol. 49, pp. 23–30, Jan. 2001.

[13] W. Sansen, “Distortion in elementary transistor circuits,” *IEEE Trans. Circuits Syst. II*, vol. 46, pp. 315–325, Mar. 1999.

[14] T. Yamaji, D. Kurose, O. Watanabe, S. Obayashi, and T. Itakura, “A four-input beam-forming downconverter for adaptive antennas,” *IEEE J. Solid-State Circuits*, vol. 38, pp. 1619–1625, Oct. 2003.

[15] D. Pache, J. M. Fournier, G. Billoit, and P. Senn, “An improved 3 V 2 GHz BICMOS image reject mixer IC,” in *Proc. Custom Integrated Circuits Conf.*, 1995, pp. 95–98.

Shoji Otaka (M'04) received the B.E., M.E., and Ph.D. degrees in electrical engineering from Tohoku University, Sendai, Japan, in 1985, 1987, and 2001, respectively.

In 1987, he joined the Research and Development Center, Toshiba Corporation, Kawasaki, Japan. Since then, he has been engaged in the research and development of analog integrated circuits for wireless communications.

Mitsuyuki Ashida received the B.S. degree in physics from Science University of Tokyo, Tokyo, Japan, in 1999, and the M.S. degree in electronics from Tokyo Institute of Technology, Tokyo, in 2001.

In 2001, he joined the System LSI Group, Semiconductor Company, Toshiba Corporation, Kawasaki, Japan. His current research interests include the design and analysis of analog integrated circuits for wireless communications.

Masato Ishii received the B.S. degree in electronics from Waseda University, Tokyo, Japan, in 1998.

In 1998, he joined the System LSI Group, Semiconductor Company, Toshiba Corporation, Yokohama, Japan. Since then, he has been engaged in the research and development of analog integrated circuits for wireless communications.

Tetsuro Itakura (M'04) received the B.E. degree from Tokyo University of Agriculture and Technology, Tokyo, Japan, in 1981, the M.S. degree from Stanford University, Stanford, CA, in 1989, and the Doctor of Engineering degree from Tokyo Institute of Technology, Tokyo, in 2003.

In 1981, he joined Toshiba Corporation, Kawasaki, Japan, where he is with the Mobile Communication Laboratory, Corporate Research and Development Center. He has been involved in the research and development of analog integrated circuits for telecommunications and for LCD driver ICs. His current interests are analog LSI design and signal processing.