A High IIP2 Downconversion Mixer Using Dynamic Matching

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***Abstract*—This paper presents an RF downconversion mixer with improved rejection to second-order intermodulation for application within a direct-conversion receiver requiring high blocking performance. The mixer, implemented in a 2.7-V 0.35-μm BiCMOS process, achieves a second-order input intercept point of at least +72 dBm for a BiCMOS design and at least +66 dBm for an all-CMOS design. The design utilizes dynamic matching to enhance the balance of a fully differential mixer through mitigation of both component and device mismatches. In addition, dynamic matching is shown to improve the mixer's $1/f$ noise performance. For an all-CMOS mixer design, a 30-dB improvement in the mixer's noise floor at 1 kHz has been observed compared to conventional fully differential CMOS Gilbert-cell mixer. Additionally, background is given on second-order intermodulation and on system IIP2 requirements for a direct-conversion receiver.**

***Index Terms*—Differential RF mixer, direct-conversion receivers, dynamic matching, intercept point, second-order intermodulation.**

I. Introduction

WITH THE increasing demand for multiband/multimode wireless devices, there is a need for a cost-effective receiver solution which circumvents the overhead associated with a traditional superheterodyne receiver. The direct-conversion receiver (DCR) shown in Fig. 1 is one possible solution. This architecture represents a fully integratable receiver solution that eliminates the need for both IF and image reject filtering and requires only a single local oscillator (LO) source. Unfortunately, there are some limitations regarding this architecture that affect the reception of the desired signal [1], [2]. These limitations include: 1) dynamic dc offsets generated primarily from the self-mixing of RF or LO signals through undesirable leakage paths within the RF front-end; 2) $1/f$ noise at the RF front-end which affects receiver sensitivity for narrow-band systems; and 3) second-order intermodulation which introduces undesirable spectral components at baseband which degrades receiver sensitivity. While issues 1 and 2 have existing solutions which meet most system requirements [2]–[6], issue 3 remains the most challenging to overcome for systems requiring high blocking performance.

This paper introduces the use of dynamic matching to achieve a downconversion mixer with a high second-order input intercept point (IIP2). Section II introduces the concept of second-order intermodulation (IM2) distortion and its effect on DCRs. In Section III, the need for increased rejection of IM2 for high-spec DCR applications is discussed. Section IV introduces dynamic matching, covering theory, and implementation. Finally, measured results are presented in Section V, and conclusions are made in Section VI.

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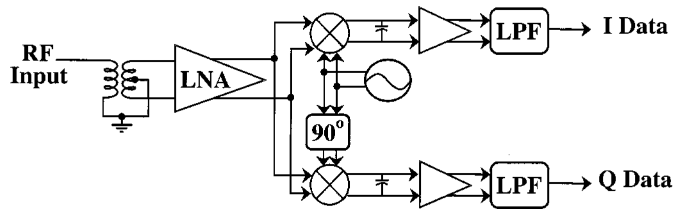

Fig. 1. Direct-conversion receiver architecture.

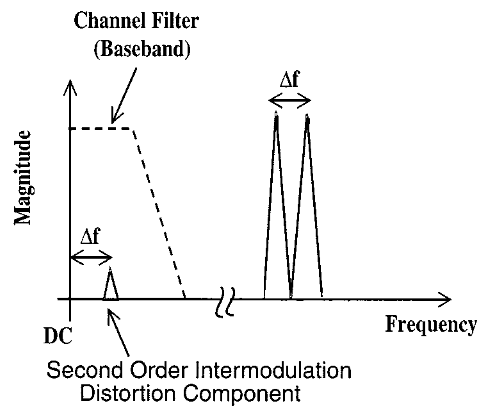

Fig. 2. Introduction of undesirable baseband spectral componet due to second-order intermodultion of two strong interferers.

II. Second-Order Intermodulation Distortion

For the purpose of distortion analysis, the transfer characteristic of a circuit may be expressed by a Taylor series expansion:

$$f[x(t)] = K_0[x(t)] + K_1[x(t)]^2 + K_2[x(t)]^3 + \cdots. \quad (1)$$

The term $K_1[x(t)]^2$ represents a second-order nonlinearity, which when exposed to strong signals, generates undesirable distortion products. For a DCR, where the desired signal of interest is downconverted from RF to baseband through a single modulation process, this source of interference is extremely deterimental to system performance. Consider the following scenario (Fig. 2) whereby two strong signals, $\cos\omega_1 t$ and $\cos\omega_2 t$, are within the bandwidth of the receiver's preselection filter and differ in frequency by an amount less than or equal to the signal bandwidth of interest as defined by the receiver's channel select filter. When these signals are exposed to a

second-order nonlinear circuit behavior, undesirable baseband spectral components are generated. These include a dc component and a baseband spectral component located at a frequency with which the two strong signals differ in frequency. Mathematically, this scenario is detailed by the following:

$$(\cos\omega_1 t + \cos\omega_2 t)^2 = 1 + 0.5\cos 2\omega_1 t + 0.5\cos 2\omega_2 t + \cos(\omega_1+\omega_2)t + \cos(\omega_2-\omega_1)t. \tag{2}$$

$$\text{DC term: } 1$$

$$\text{Baseband Component: } \cos(\omega_2-\omega_1)t \quad \omega_2-\omega_1 <= \omega_{\text{chan}}.$$

These spectral components degrade the reception of the desired signal. Another signal condition which results in the generation of undesirable baseband components is the exposure of a single strong interferer to a second-order nonlinearity. Under this condition, an undesirable baseband beat is created as shown in (3) and (4) which directly interfers with the desired signal.

$$(a(t)\cos(\omega t + \phi(t)))^2 = a^2(t)/2(1+\cos(2\omega t + 2\phi(t))) \tag{3}$$

$$==> a^2(t)/2 \quad \text{Undesirable baseband beat.} \tag{4}$$

Depending on the envelope $a(t)$ of this single strong interferer, IM2 will generate specific baseband components affecting the performance of a DCR. For a constant envelope, $a(t) = A_c$, IM2 will generate an undesirable dc component which can be addressed by various reported methods of dc offset correction [2]–[6]. For a nonconstant envelope, $a(t) = A_c(1+m(t))$ where $m(t)$ is a function of time for amplitude modulation, the baseband beat [(4)] generated from IM2 will be composed of several undesirable spectral components as given by (5).

$$a^2(t)/2 = 1/2[A_c(1+m(t))]^2 = A_c^2/2[1+2m(t)+m^2(t)]. \tag{5}$$

These components include a dc component [(6)], which again can be addressed by known mitigation methods, and other more troublesome baseband spectral components [(7)] for which no mitigation methods exist.

$$\text{DC Component: } A_c^2/2 \tag{6}$$

$$\text{Baseband Components: } A_c^2/2[2m(t)+m(t)^2]. \tag{7}$$

To protect a DCR from such undesirable spurious responses requires a high second-order intermodulation rejection ratio (IMR2) between the amplitude level of the interferer(s) and resulting IM2 component. Typically such figure of merit is specified by a IIP2. This IIP2 specification represents a fictitious input amplitude at which the desired signal becomes equal in amplitude to the spectral component generated from IM2 (Fig. 3). Thus, a high IIP2 downconversion mixer is required to minimize the effect of IM2 within a DCR.

III. System Requirements

In many of today's wireless standards, there are no system level receiver specifications for IIP2 since most existing systems have receivers based on the conventional superheterodyne architecture. The superheterodyne architecture, unlike the DCR,

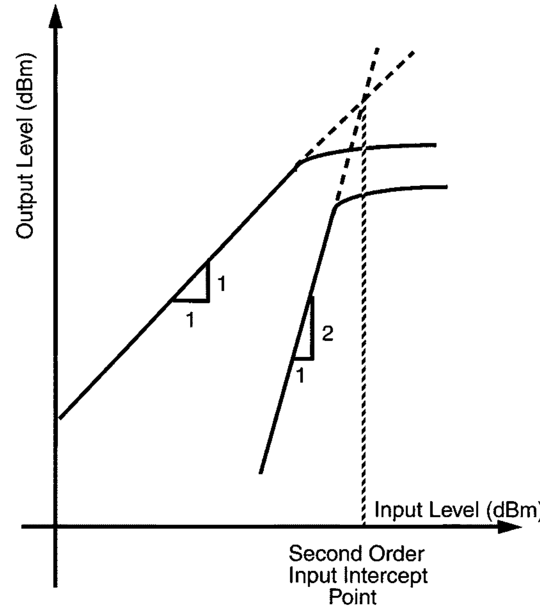

Fig. 3. Second-order input intercept point concept.

is not susceptible to IM2 since the IM2 baseband products are greatly attenuated by bandpass filters. Hence, there is a need to define such a requirement for DCRs to insure a reliable communication link. With the background given in Section II, it can be seen that in order to avoid any susceptibility to IM2 related blocking in a DCR, the system IIP2 should be chosen high enough to meet the blocking specification of the receiver when the blocker has nonconstant envelope modulation. As a result, classical IM2 interference from two interferers is mitigated since their effect on the receiver becomes equal to the single interferer [double sideband suppressed carrier (DSBSC)] blocking performance.

The concept of intermodulation and intercept point is defined based on two-tone interference. Two tones spaced by a frequency difference of Δf are equivalent to a DSBSC signal modulated with a sine wave of frequency $\Delta f/2$. The total power in such a signal is 3 dB more than the power in each tone. Equations (8)–(11) relate the IMR2, the IIP2, the input power of each interfering tone for the two-tone interference (P_{iu}), and the resulting interference (P_{id}) relative to the receiver's input port. All powers and intercept points are in dBm and all intermodulation ratios are in dB [7].

$$\text{IMR2} \equiv P_{iu} - P_{id} \text{ dB} \quad \text{(By definition)} \tag{8}$$

$$\text{IMR2} = \text{IIP2} - P_{iu} \text{ dB} \tag{9}$$

$$\text{IMR2} = \tfrac{1}{2}(\text{IIP2} - P_{id}) \text{ dB} \tag{10}$$

$$\text{IIP2} = 2P_{iu} - P_{id} \text{ dB} \tag{11}$$

where

- IMR2 second-order IM rejection ratio;
- P_{iu} input power of each interfering tone for the two-tone interference;
- P_{id} resulting interference generated relative to the input;
- IIP2 second-order input intercept point.

These equations hold true for the signal power regime where the receiver is not in compression and the IM2 interference

versus input power follows a true 2 : 1 slope (Fig. 3). Now, to know the actual IIP2 required for a given P_{iu} for the standard two-tone intermodulation measurement, a modified version of (11) is required:

$$\mathrm{IIP2} = 2P_{iu} - P_{sens} + C/I \text{ dBm} \tag{12}$$

where P_{sens} is the sensitivity or reference signal level in dBm, and C/I is the required (desired signal power)/(on-channel interfering power) ratio (in dB) to maintain the signal quality (e.g., bit error rate) at the level defined for sensitivity (P_{sens}).

For blocking due to a single DSBSC sinusoidally modulated blocker, which has a total power of P_{block}, the power in each tone of the blocker (P_{iu}) will be $P_{\text{block}} - 3$ dB. So the required IIP2 required to protect against a DSBSC signal with a blocking power of P_{block} is

$$\begin{aligned}\mathrm{IIP2} =& 2(P_{\text{block}} - 3 \text{ dB}) - P_{sens} \\ &+ C/I \text{ dBm} \quad \text{(DSBSC Blocker).}\end{aligned} \tag{13}$$

To establish how much interference is generated due to second-order intermodulation for other nonconstant-envelope modulation requires system-level simulations. A new parameter, DIIP2, is defined as the difference in IIP2 needed for a given modulation compared to the IIP2 required to protect against a DSBSC blocker. As a result, the IIP2 required for protection against a given blocking power can now be specified as

$$\mathrm{IIP2} = 2(P_{\text{block}} - 3 \text{ dB}) - P_{sens} + C/I + \mathrm{DIIP2} \text{ dBm.} \tag{14}$$

As an example, system simulations are performed on $\pi/4$-DQPSK with rolloff factor of $\alpha = 0.5$. Relative to a DSBSC blocker, this modulation requires a DIIP2 of -7 dB. This number is for a single interferer. There will be no interference, other than dc due to IM2, from a single FM or other constant envelope modulated interferer. There will be interference from two constant envelope interferers when the absolute value of the difference of the differences of their carrier frequencies from the desired carrier frequency falls within the IF passband of the receiver.

$$abs[(f_{\text{RF}} - f_1) - (f_{\text{RF}} - f_2)] <= \mathrm{BW}_{\text{IF}}/2. \tag{15}$$

Assuming the interferer has the same type of modulation as the desired on-channel signal, the required system IIP2 for wide-band $\pi/4$-DQPSK with $\alpha = 0.5$, $P_{\text{sens}} = -104$ dBm [noise figure (NF) $\sim= 7$ dB, BW $= 220$ kHz], $C/I = 10.4$ dB, $P_{\text{block}} = -25$ dBm (79 dB above sensitivity), is

$$\mathrm{IIP2} = 2(-25 - 3) - (-104) + 10.4 + (-7.0) = +51.4 \text{ dBm.} \tag{16}$$

If we assume that there is 15 dB net gain to the downconversion mixer, the IIP2 requirements of the mixer becomes +66.4 dBm. With added margin, the downconversion mixer requires +70 dBm IIP2. Similar analysis and simulation applied to other types of wireless standards also confirm the need for a downconversion mixer IIP2 of at least +70 dBm for application within a DCR. This level of IIP2 performance is significantly higher than what has been reported to date [8], [9].

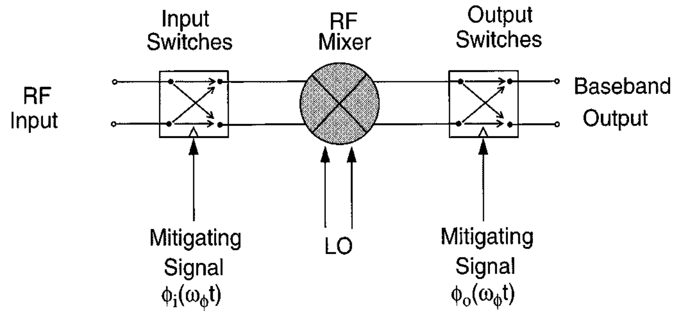

Fig. 4. General RF downconversion mixer topology utilizing dynamic matching.

IV. Dynamic Matching

A. Principle of Operation

Ideally, a perfectly balanced fully differential mixer should be immune to IM2 since the resulting products are common-mode. However, due to component and device mismatches, asymmetry is introduced into the mixer which gives rise to a finite IM2 response. Depending on the application of the mixer, this response may or may not be acceptable. For systems requiring high block performance, the best reported IIP2 of +60 dBm [9] using fully differential circuitry alone is inadequate. To resolve the generation of IM2 distortion and thus attain a high level of IIP2 performance, dynamic matching is employed. The concept of dynamic matching has been used for several years to design precision op-amps known as chopper stabilized op-amps [10] and to design linear multibit DACs for application within $\Sigma\Delta$ analog-to-digital converters (ADCs) [11]. Similar to those applications, dynamic matching is utilized here to mitigate the effects of both component and device mismatches. Within the context of a fully-differential RF downconversion mixer, this enhances overall circuit balance. Fig. 4 illustrates the conceptual mixer topology. The time-domain transfer function of this mixer to the baseband output from the RF input is given by (17):

$$h_1(t) = \phi_i(\omega_\phi t)^* \cos(\omega_{\text{LO}} t)^* \phi_o(\omega_\phi t) \tag{17}$$

where $\phi_i(\omega_\phi t)$ and $\phi_o(\omega_\phi t)$ are strong square wave signals. This choice of signal type for $\phi_i(\omega_\phi t)$ and $\phi_i(\omega_\phi t)$ facilitates accurate synchronization of both input and output switches and allows the switches to behave more like signal routing switches. Ideally, with $\phi_i(\omega_\phi t)$ in phase with $\phi_o(\omega_\phi t)$, their product is equal to unity and the time-domain transfer function reduces to that of simple ideal mixer [(18)].

$$h_1(t) = \cos(\omega_{\text{LO}} t). \tag{18}$$

The time-domain transfer function to the baseband output from the source of IM2 distortion originating from the RF mixer is given by

$$h_2(t) = \phi_o(\omega_\phi t). \tag{19}$$

From this perspective, the effect of IM2 is mitigated. Depending on the explicit choice of signal for $\phi_i(\omega_\phi t)$ and $\phi_o(\omega_\phi t)$, the IM2 distortion can be: 1) frequency translated out of the signal band of interest by choosing a periodic signal; or 2) frequency

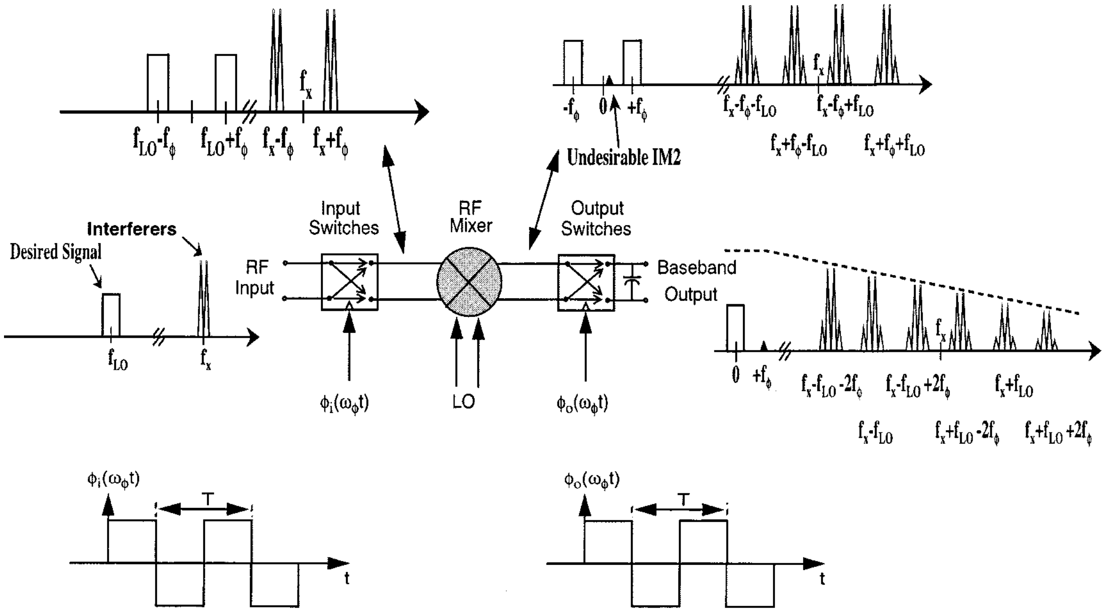

Fig. 5. Dynamic matching principle applied to an RF downconversion mixer.

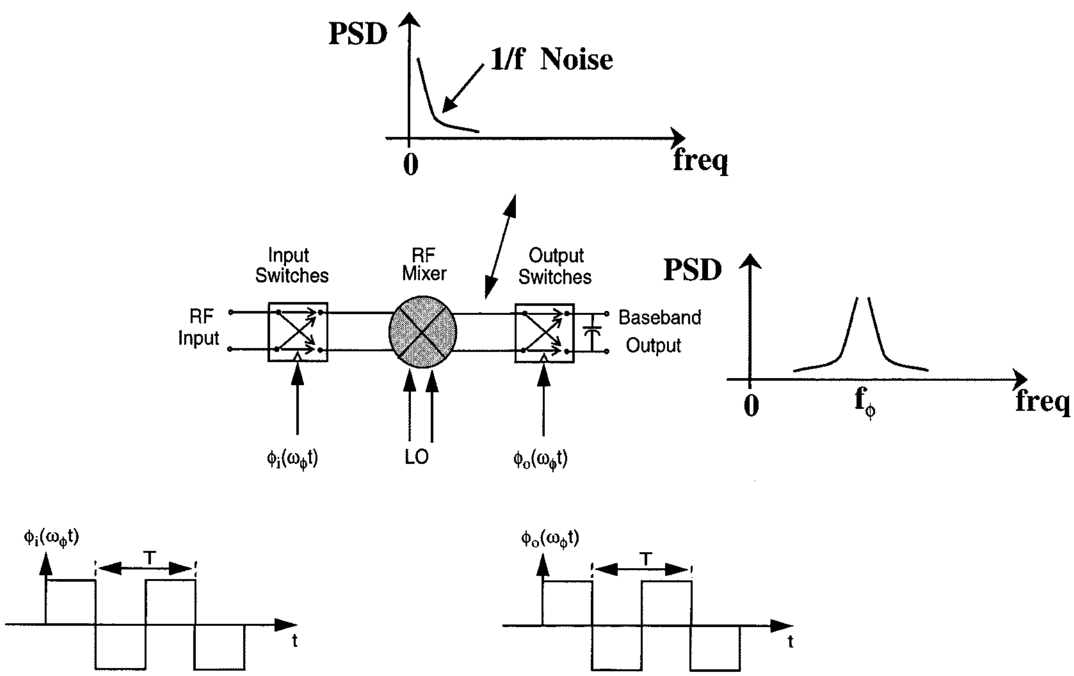

Fig. 6. Mitigation of flicker noise originating from RF downconversion mixer through the use of dynamic matching.

spread within the signal band of interest by choosing a pseudorandom signal. Fig. 5 illustrates the dynamic matching principle where $\phi(\omega_\phi t) = \phi_i(\omega_\phi t) = \phi_o(\omega_\phi t)$ has been chosen to be periodic.

To reduce the aliasing of undesirable signals into the signal band of interest, the frequency of $\phi(\omega_\phi t)$ is chosen to exceed the bandwidth of the receiver's preselection filter. The input switches modulate the band limited input RF signal to frequencies $f_{rf} \pm f_\varphi$. A second modulation, within the main mixer core occurs at the LO frequency to translate the signal at $f_{rf} \pm f_\varphi$ to $f_{rf} \pm f_\varphi \pm f_{\mathrm{LO}}$. The output switches, synchronously commutating with input switches, then demodulate the signal at $f_{rf} \pm f_\varphi \pm f_{\mathrm{LO}}$ to the desired frequency of $f_{rf} \pm f_{\mathrm{LO}}$ which for a DCR is baseband centered about dc. During this process, any undesirable IM2 products generated in the mixer are modulated to f_φ where they can easily be filtered off. Simulations show that this method can yield a 20-dB IIP2 improvement. With a pseudorandom signal for $\phi(\omega_\phi t)$, the undesirable IM2 products generated by the mixer are spread over a range of frequencies. By properly designing the spreading sequence to possess a certain power spectral density, the required mixer IIP2 performance can be achieved.

Now an additional benefit of dynamic matching is the reduction of $1/f$ noise [12]. For a narrow-band DCR, this source of close-in noise degrades the noise performance of the downconversion mixer and hence the system NF. Through

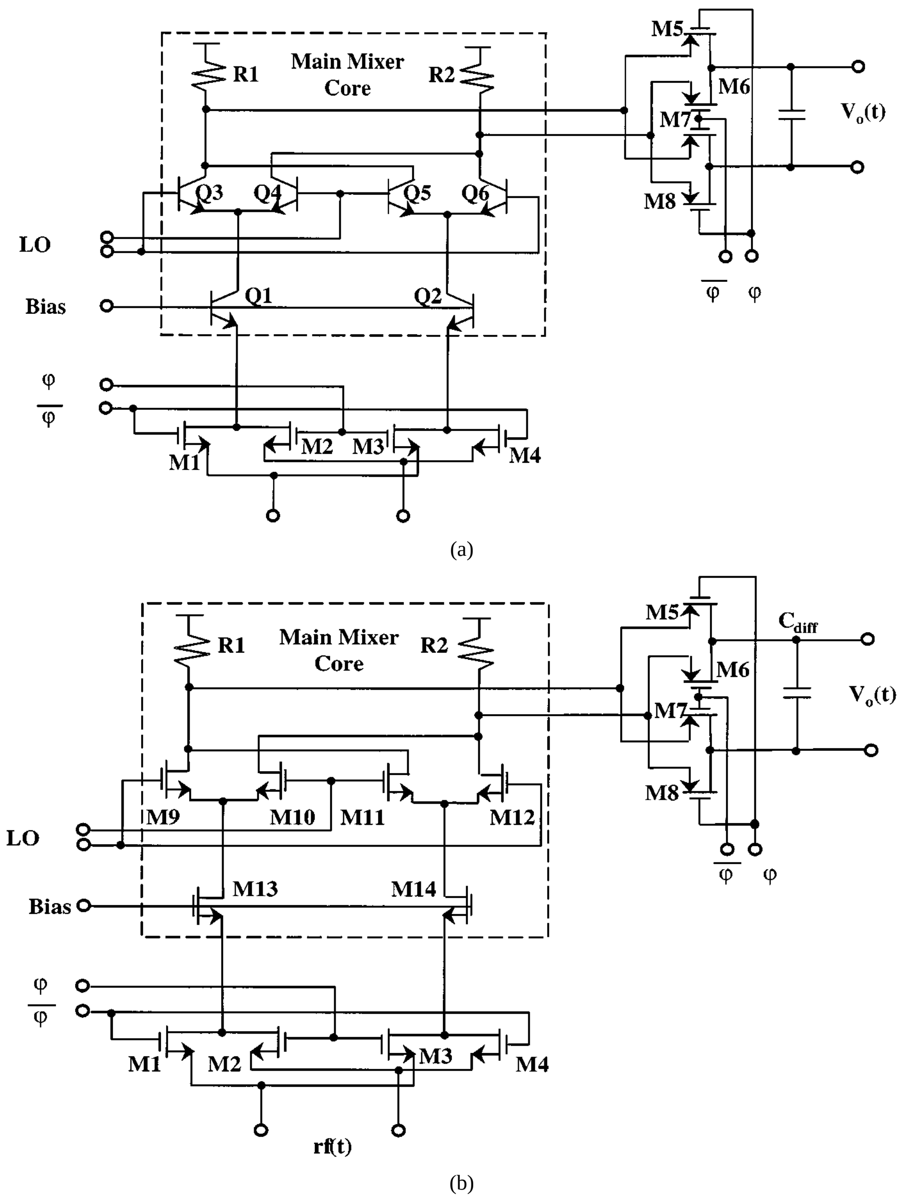

Fig. 7. (a) BiCMOS dynamically matched RF downconversion mixer. (b) All-CMOS dynamically matched RF downconversion mixer.

the use of dynamic matching, the flicker noise can also be frequency translated and subsequently filtered out or it can be spread over a range of frequencies depending on whether one chooses a periodic or pseudorandom signal for $\phi(\omega_\phi t)$. Fig. 6 details the mitigation of this noise source using a periodic signal for $\phi(\omega_\phi t) = \phi_i(\omega_\phi t) = \phi_o(\omega_\phi t)$. By improving the mixer's NF, less takeover gain is required from the low noise amplifier (LNA) which correspondingly relaxes the mixer's intermodulation performance requirement. Dynamic matching is especially useful in an all-CMOS downconversion mixer where the contribution of $1/f$ noise is high. Note that to prevent the aliasing of wideband thermal noise coming into the dynamically matched mixer, the frequency of $\phi(\omega_\phi t)$ should be chosen to be at least twice the bandwidth of preselection filter. With this choice of frequency, a measured degradation in mixer's NF is only about 0.5 to 1 dB.

Concerns regarding this method of IM2 mitigation are undesirable spurious responses at the frequency of $\phi(\omega_\phi t)$ and harmonics of it. The mixer alone provides 40 dB of rejection to these spurs. This limitation is due to limited on-chip matching and the synchronization of the input and output switches. To obtain additional attenuation of these spurs requires careful selection of the frequency of $\phi(\omega_\phi t)$. Specifically, choosing the frequency of $\phi(\omega_\phi t)$ to be greater than the bandwidth of the preselection filter will provide additional attenuation as determined by the roll-off of the preselection filter. Hence, from a receiver system point-of-view, the frequency of $\phi(\omega_\phi t)$ and the preselection filter bandwidth can be chosen accordingly to meet requirements placed on the spurious response of the receiver.

B. Circuit Implementation

Fig. 7(a) and (b) illustrates the circuit details of the dynamically matched downconversion mixers. The downconversion mixers consist of input switches M1–M4, the main mixer core composed of an input amplifier stage and a Gilbert cell, and output switches M5–M8. The input stage of the mixer is im-

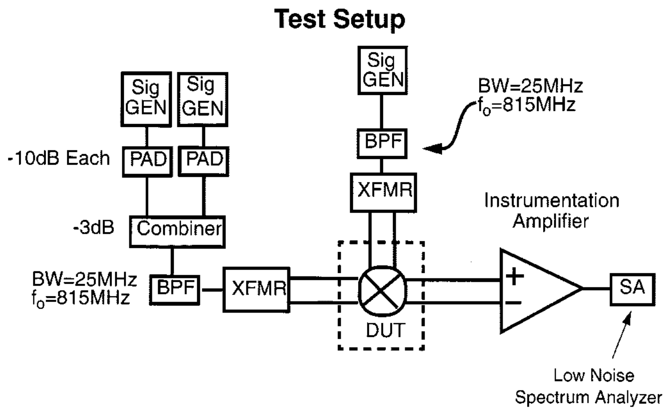

Fig. 8. Experimental test setup.

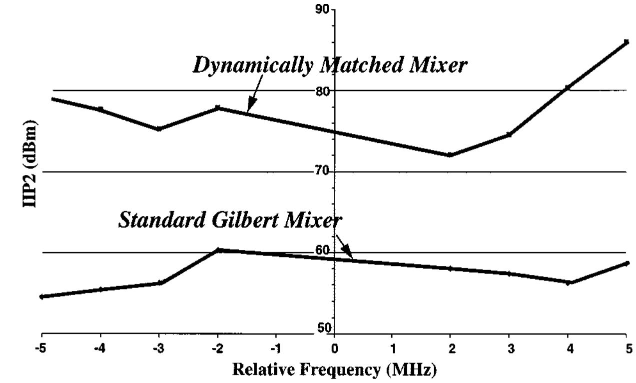

Fig. 9. Measured IIP2 performance of BiCMOS mixer design. IIP2 versus frequency location of two interferers relative to LO of 815 MHz.

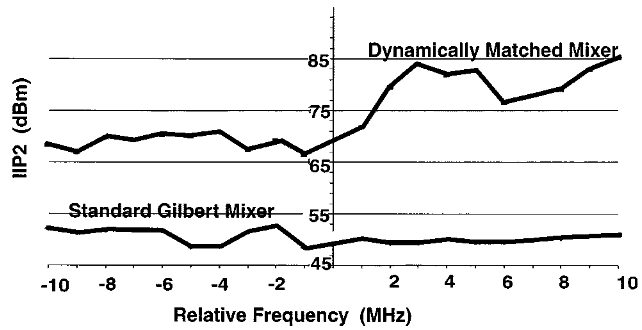

Fig. 10. Measured IIP2 performance of all-CMOS mixer design. IIP2 versus frequency location of two interferers relative to LO of 815 MHz.

plemented using a wide-band common-base or common-gate amplifier topology to provide low impedance at RF while achieving a high third-order intercept point (IIP3) with low current drain. This choice, however, is made at the expense of lower power gain. Added linearization of this stage has also been achieved through the input switches which provide a finite amount of emitter or source degeneration. The input and output switches have been chosen to be n and p MOSFETs, respectively, due to the inherent dc operating voltage of the mixer. To attain maximum IIP2 performance, synchronization between input and output switches is required. As such, the aspect ratio (W/L) of both nMOS and pMOS devices have been sized accordingly to attain equal time constant—$R^*_{\rm ON}C_{load}$. In addition, these switches are gated as fast as possible and the layout of each clock signal path is equalized. Because the switches are used in a differential mixer topology, the effects of clock feedthrough and channel charge injection is minimized since their contribution becomes common-mode. To minimize the impact of mismatches in clock duty cycle for the case where $\phi(\omega_\phi t)$ is chosen to be a periodic signal, a clock signal at twice the frequency of $\phi(\omega_\phi t)$ is divided by two through a master–slave flip-flop to obtain 50% duty cycle. Finally, the entire mixer has been laid out in a symmetrical fashion in order to get the best device and component matching for IIP2 performance.

From a noise performance point of view, the Gilbert cell [devices M9–M12 of Fig. 7(b)] within the main mixer core is the dominant source of noise which is primarily $1/f$. The input and output switches are driven by strong square wave signals such that their noise contribution is primarily thermal (due to $R_{\rm ON}$).

V. Measured Results

The BiCMOS and all-CMOS mixers of Fig. 7 were implemented to operate at 2.7 V in a 0.35-μm BiCMOS process. Each mixer was measured separately by using the experimental test setup of Fig. 8. Two signal generators are combined together through the use of RF attenuators and a 3-dB combiner to generate the two-tone interference. The combined signal is then bandpass filtered to eliminate any harmonics of the tones as well as any intermodulation distortion terms that may have originated from the signal generators. This signal is then fed to a transformer to convert the single-ended signal to a differential signal prior to its application to the mixer's RF port. The LO signal, from a separate signal generator, is also bandpass filtered and converted to a differential signal prior to its application to the LO ports of the mixer. The LO signal is fed at twice the frequency of the RF signal and its frequency is divided by two through the use of an emitter coupled logic (ECL) divide-by-two master–slave flip-flop. This improves the mixer's LO to RF isolation which minimizes the generation of dynamic dc offsets.

Fig. 9 shows the measured IIP2 of the BiCMOS mixer [Fig. 7(a)]. The x-axis represents the difference in frequency between the LO and two interferers which should generate the baseband IM2 product. With a periodic signal of 75 MHz, dynamic matching provides at least $+72$-dBm IIP2 performance. This is a 12-dB improvement over a standard BiCMOS Gilbert-cell mixer. Fig. 10 shows the measured IIP2 of the all-CMOS mixer of Fig. 7(b). Once again, with a periodic signal of 75 MHz, dynamic matching attains an IIP2 of at least $+66$ dBm. This represents an 11-dB improvement over a standard all-CMOS Gilbert-cell mixer. At some frequencies, the IIP2 was measured to be as high as $+80$ dBm, which is an outstanding performance for an all-CMOS mixer.

Visual observation through a spectrum analyzer has shown dynamic matching effective in minimizing the impact of $1/f$ noise. For an all-CMOS mixer, the improvement of the noise floor was observed to be 30 dB at 1 kHz compared to that of a standard all-CMOS Gilbert-cell mixer [Fig. 11(a)]. Fig. 11(b) illustrates a similar plot but in a much wider scale. From this perspective, the effect of a single pole filter at 20 kHz on the predominant thermal noise of the dynamically matched mixer and

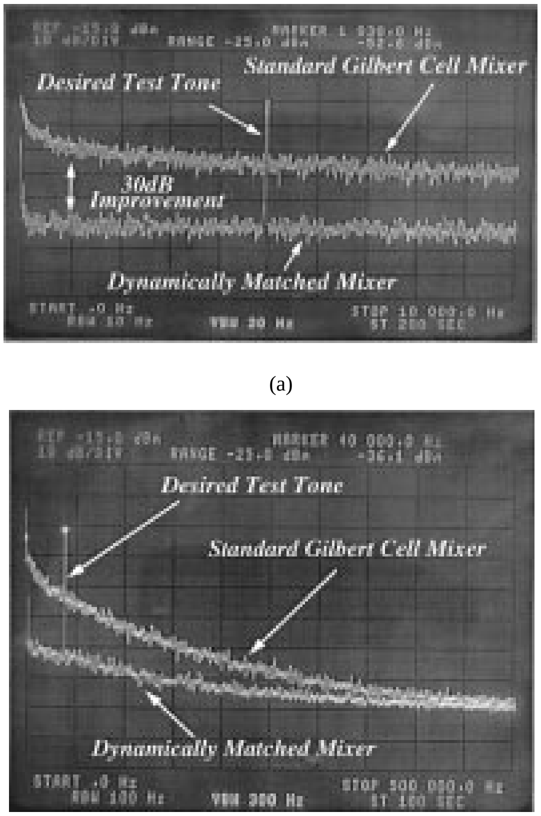

Fig. 11. (a) Close-in noise performance of all-CMOS mixer designs. (b) Far-out noise performance of all-CMOS mixer designs.

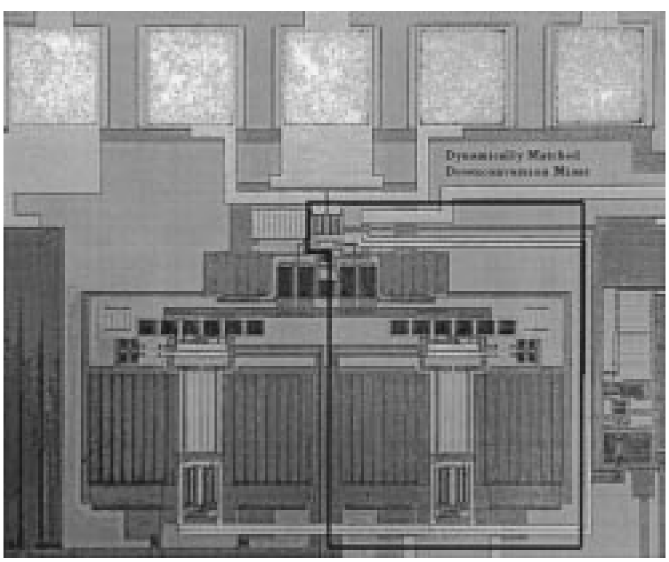

Fig. 12. Die micrograph.

on the $1/f$ noise of the standard all-CMOS Gilbert-cell mixer can be observed. Fig. 12 shows the die micrograph.

Tables I and II summarize the performances attained in a BiCMOS and an all-CMOS dynamically matched downconversion mixer.

TABLE I
MEASURED BiCMOS DYNAMICALLY MATCHED MIXER PERFORMANCE SUMMARY

Parameters	Values
IIP2	> +72dBm
IIP3	+4.9dBm
NF	11.0dB
Conversion Gain	19.0dB
Supply	2.7V
Current Dissipation	2.25mA
DC Offset Improvement	>20dB
Frequency of Periodic Dynamic Matching Signal	75MHz
Measured LO Signal at RF Port	-92dBm
RF/LO Frequency	815MHz

TABLE II
MEASURED ALL-CMOS DYNAMICALLY MATCHED MIXER PERFORMANCE SUMMARY

Parameters	Values
IIP2	> +66dBm
IIP3	+2.4dBm
NF	12.0dB
Conversion Gain	14.5dB
Supply	2.7V
Current Dissipation	4.0mA
DC Offset Improvement	>20dB
Frequency of Periodic Dynamic Matching Signal	75MHz
Measured LO Signal at RF Port	-92dBm
RF/LO Frequency	815MHz

VI. CONCLUSION

This paper presented an RF downconversion mixer which demonstrated a significant improvement in IIP2 performance. Through the application of dynamic matching, mismatches in both components and devices within a fully differential mixer have been effectively mitigated from disrupting overall circuit balance. Additionally, the use of dynamic matching has improved the mixer's close-in noise performance by greatly reducing the contribution of flicker or $1/f$ noise.

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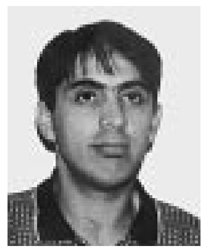

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