A RF CMOS GNSS receiver with a passive mixer for GPS L1/Galileo E1/Compass B1 band

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Abstract: This paper presents the design and implementation of a 180 nm CMOS reconfigurable global navigation satellite system receiver, supporting GPS L1, Galileo E1 and Compass B1 bands. The low-IF receiver incorporates a pseudo-differential low-noise amplifier, a double-balanced passive mixer, a pair of trans-impedance amplifiers, a complex band-pass filter, an analog-to-digital converter and an automatic gain control. A phase-locked loop is integrated to provide 25% duty-cycle quadrature clock signals. The RF front-end achieves a maximum gain of 107.2 dB with a dynamic range of 78 dB, a noise figure of 1.8 dB and an image rejection ratio of 39.1 dB.

Keywords: global navigation satellite system, RF receiver, passive mixer, trans-impedance amplifier

Classification: Integrated circuits

References

1 Introduction

In recent years, global navigation satellite systems (GNSSs) have received noticeable research attention due to its ever-increasing influence on location-based services in a variety of civilian applications [1, 2, 3, 4, 5, 6]. After the deployment of Global Position System (GPS) by the U.S., many other countries have engaged in developing their own GNSS systems, such as global navigation satellite system (GLONASS) from Russia, Galileo from the European Union, and Compass (also known as “BeiDou-2”, or “BD-2”) from China.

As summarized in Table I, part of the GPS, Galileo and Compass signals are located at the 1.57 GHz RF band with a bandwidth of 2 or 4 MHz. For low cost applications where dual frequency ionospheric correction is not necessary, a single receiver can be reconfigured to support GPS L1, Galileo E1 and Compass B1, achieving sufficient flexibility with area efficiency. Without techniques such as differential GNSSs, it is essential to design the RF front-end with low noise figure (NF) for the single-band receiver to guarantee a fairly good positioning accuracy.

Traditionally, low-IF architectures were adopted in GNSS RF front-end [1, 2, 3, 4, 5, 7, 8] to relax image-rejection requirements and deal with problems such as dc offset and flicker noise. To further achieve even higher image rejection ratio (IMRR) performance, digital calibration techniques were employed to improve the matching between I and Q branches [4, 5, 9]. A typical architecture employs a...
pseudo-differential low noise amplifier (LNA) to perform single-ended-to-differential transformation [1, 2, 3] and an active mixer to realize down-conversion [1, 2, 3, 4, 5]. Receivers with a single-balanced passive mixer were implemented in [7] and [8] and good system linearity was observed. In [7] and [8], switched-on MOSFETs of the mixer serve as low impedance branches and direct the output current from the LNA into the trans-impedance amplifier (TIA). Passive mixers generally exhibit superior linearity, power consumption and $1/f$ noise performances, which makes them an alternative choice in receiver design despite of their gain loss [10, 11]. GNSS receivers with passive mixing techniques are of special importance for applications in which large dynamic range is desired, e.g., in an environment affected by jamming. Despite of its advantages in certain application scenarios, down conversion exploiting double-balanced passive mixing techniques haven’t received much research attention in GNSS receiver front-ends.

In this paper, a GNSS RF receiver with a double-balanced passive mixer is proposed. The architecture of the receiver and the design considerations are briefly discussed in section II. Section III presents the detailed circuit implementations of the receiver. The measured results of the fabricated RF receiver are shown in section IV and section V concludes this paper.

2 System architecture

2.1 Receiver architecture

Since the GPS L1, Galileo E1 and Compass B1 signals are all located at the 1.57 GHz RF band, only one receiver is needed to receive and process the three types of signals. Fig. 1 illustrates the block diagram of the proposed receiver, which adopts the low-IF architecture with an LNA, a passive quadrature mixer, a pair of TIAs, a complex band-pass filter (CBPF) and a programmable gain amplifier (PGA) followed by a 4-bit flash ADC and an automatic gain control (AGC) loop. The received RF signal is firstly fed to the pseudo-differential LNA to be amplified and converted to differential mode. The magnified RF signal is then down-converted to the IF band by the current-mode double-balanced passive mixer which dissipates no dc current and has less $1/f$ noise problem. The TIAs then offer extra gain and suppress the out-of-band noise by playing the role of an active first-order BPF. The active sixth-order CBPF provides moderate variable voltage gain, spectrum anti-aliasing and image rejection for the IF signal. The CBPF has a
center frequency of 4.092 MHz and a reconfigurable 2 or 4 MHz bandwidth. The three-stage PGA along with the digital AGC loop ensures desired magnitude of IF signal at the ADC input.

Besides, the quadrature LO signals with 25% duty cycle are generated by a Fractional-N phase-locked loop (PLL) with the output frequency of 1571.328 MHz for GPS L1 or Galileo E1 signal and 1557.006 MHz for Compass B1 signal. As a consequence, three types of RF signals are down-converted to the same IF band with a center frequency of 4.092 MHz.

2.2 Gain and noise
The received RF signals from the satellites are far below the thermal noise level when they arrive at the antenna. As a result, the thermal noise always dominates the input power level [10]. For a signal with a bandwidth of 2 or 4 MHz, the thermal noise floor is calculated to be around $-110$ dBm. To get a full scale input range of 500 mV for the ADC, a typical 100 dB cascaded gain is needed for the receiver.

As mentioned in [1, 2] and [5], it is more convenient to quantify noise performance for the receiver by the carrier-to-noise density ratio ($C/N_0$) than signal-to-noise ratio (SNR). Once the minimum required $C/N_0$ at the ADC output is given by the baseband requirements, the receiver sensitivity $P_{Sen}$ is uniquely determined as

$$P_{Sen} = \left(\frac{C}{N_0}\right)_{\text{min}} \text{[dB-Hz]} + P_{n,\text{in}} \text{[dBm/Hz]} + NF\text{[dB]}$$

where $P_{n,\text{in}}$ is the thermal noise power spectral density at the antenna port which equals to $-174$ dBm/Hz and NF is the noise figure of the receiver. With an assumed minimum $C/N_0$ of 25 dB·Hz and a target NF of 2 dB, a sensitivity of $-147$ dBm can be achieved.

3 Circuit implementation
3.1 LNA
The pseudo-differential LNA, as shown in Fig. 2, is utilized to perform single-ended-to-differential transformation with high voltage gain and low noise. The LNA adopts cascode topology to reduce Miller effects of the transistors and provides high reverse isolation. The single-ended-to-differential conversion is realized by incorporating the compensation capacitor $C_{c}$, which avoids a lossy balun and provides an extra 6 dB voltage gain. The package bonding wire $L_s$ connecting the source of $M_1$ and the PCB board plays the role of degenerative inductor to achieve low-noise input matching.

$L_L$ and tunable $C_L$ resonate at 1.57 GHz, exhibiting a high loading impedance with a high quality factor. The gain of LNA can be controlled by gate bias of $M_5$ and $M_6$. The LNA brings a voltage gain ranging from 12 to 28 dB in order to be adapted to active or passive antennas. The bandwidth is set to be 30 MHz to ensure that it could linearly amplify the RF signals from GPS L1, Galileo E1 or Compass B1 band.

When loaded by a current-mode passive mixer terminated into low input impedance TIAs, the LNA acts as a trans-conductance amplifier. Since the mixer
branch exhibits much lower impedance than the LC branch at the resonating frequency, most of the amplified signal current will go through the mixer branch and further form a voltage output via the TIAs.

3.2 Passive mixer and TIA

Traditionally, active mixers are always employed in GNSS receivers to achieve moderate conversion gain and to suppress LO leakage. On the other hand, passive mixers can achieve superior linearity, power efficiency and 1/f noise compared to active Gilbert mixers and are widely used in many implementations [12, 13]. As shown in Fig. 3, a current-mode double-balanced passive mixer followed by a pair of TIAs is implemented. The mixer works with 25% duty-cycle LO signals and suppresses more spurious product terms of the LO and RF signals than its single-balanced counterpart.
The TIAs have low input impedances, which allows most of the LNA output current to flow through the passive mixer branches. The TIAs in I/Q paths are cross-connected to act as a first-order active CBPF. The bandwidth of TIAs can be configured to different mode by tuning $C_x$ and $R_c$. To accommodate for different applications, 2 MHz bandwidth is for the GPS L1 signal and 4 MHz bandwidth is for the Galileo E1 or the Compass B1 signal. In order to achieve a large dynamic range, part of the TIA gain is also designed to be adjustable by tuning $R_x$.

### 3.3 CBPF

Fig. 4 shows the sixth-order Chebyshev-I CBPF based on active-RC integrators to select the desired signals with image and out-of-band spurious signals rejection. The leap-frog CBPF is implemented by means of two real low-pass filters (LPFs), quadrature coupled through the cross-coupling paths to achieve the central frequency shift at the desired IF band [2, 5]. The center frequency is set to 4.092 MHz and the bandwidth can be configured to 2 MHz or 4 MHz by tuning capacitance of the feedback capacitors.

The OPA cell is realized by utilizing a two-stage fully differential op-amp to achieve low noise, high gain and high linearity. A common-mode feedback (CMFB) circuitry is added to make sure the output common-mode voltage stay within the desired range. Besides, since the frequency response characteristics of the CBPF will be affected by the PVT variations, a tuning circuit is included for real-time calibration of the CBPFs center frequency and bandwidth.

### 3.4 Synthesizer

Fig. 5 shows the schematic of the synthesizer realized by a Fractional-N PLL which consists of a phase frequency detector (PFD), a charge pump (CP), an on-chip LPF, a voltage controlled oscillator (VCO), a divider and an automatic frequency calibration (AFC) module. The divider includes a 4/5 prescaler, a programmable...
counter and a delta-sigma (ΔΣ) module. The output signals of the PLL are converted to 25% duty-cycle quadrature LO signals through the divide-by-2 divider and the AND gate, and are then fed to the mixer for the down-conversion.

Fig. 5 also presents the detailed schematic of the LC-VCO utilizing the PMOS cross-coupled pair topology with a poly resistor $R_b$ rather than a current source to set the VCO current for good phase noise performance. A 4-bit binary-coded capacitor bank controlled by the AFC is integrated in the VCO for frequency coverage. To shield noise possibly coupled from the common power supply, the VCO power supply is regulated by a low drop-out regulator (LDO). Once the automatic calibration is done, the PLL loop will be closed and a normal locking process will start.

![Fractional-N PLL](image)

**Fig. 5.** Fractional-N PLL.

### 4 Measurement results

The proposed GNSS receiver is implemented in a 180 nm RF CMOS technology and a die microphotograph is shown in Fig. 6. The whole die area is about 1.36 mm² including ESD protection circuitry and pads. The receiver consumes a total current of 16 mA from a 1.8 V power supply.

![Chip microphotograph](image)

**Fig. 6.** Chip microphotograph.
Fig. 7 shows the measured phase noise performances of the PLL, in which the bandwidth is about 200 kHz. The measured in-band phase noise at 100 kHz frequency offset is around $-100\text{ dBc/Hz}$. The phase noise at 1 MHz frequency offset is about $-122.4\text{ dBc/Hz}$ for the GPS L1 or the Galileo E1 signal and $-124.17\text{ dBc/Hz}$ for the Compass B1 signal.

Fig. 8 shows the frequency spectrum at the IF output point when a $-85\text{ dBm}$ 1.57542 GHz single-tone input is applied at the input port. The output IF power $P_{s,\text{out}}$ measured at 4.092 MHz is $-1.616\text{ dBm}$ and the output noise power is $-39.223\text{ dBm}$ with a 91 kHz video bandwidth (VBW). As a result, an output noise power spectral density $P_{n,\text{out}}$ of $-88.813\text{ dBm/Hz}$ can be calculated. With input thermal noise power spectral density $P_{n,\text{in}}$ of $-174\text{ dBm/Hz}$, the NF of the GNSS RF front-end could be obtained as follows.
NF[dB] = \left( \frac{C}{N_0} \right)_{\text{in}} \text{[dB-Hz]} - \left( \frac{C}{N_0} \right)_{\text{out}} \text{[dB-Hz]}
= (P_{s,\text{in}}[\text{dBm}] - P_{n,\text{in}}\left[\frac{\text{dBm}}{\text{Hz}}\right]) - (P_{s,\text{out}}[\text{dBm}] - P_{n,\text{out}}\left[\frac{\text{dBm}}{\text{Hz}}\right])
= (-85 - (-174)) - (-1.616 - (-88.813))
= 1.803

Fig. 9 shows the IF output spectrum in a two-tone test with a voltage gain of 30 dB. The third-order input-referred intercept point (IIP3) of the RF front-end can be calculated by

\text{IIP3}[\text{dBm}] = P_{m}[\text{dBm}] + \frac{(P_f[\text{dBm}] - P_{IM3}[\text{dBm}])}{2} \tag{3}

where \(P_m\) = \(P_f\) - \(Gain[\text{dB}]\). When both the VBW and the resolution bandwidth (RBW) are set to 11 kHz, \(P_f\) and \(P_{IM3}\) are about \(-5.1\text{dBm}\) and \(-36.65\text{dBm}\) respectively, achieving an IIP3 of \(-19.325\text{dBm}\) with an input power \(P_{m}\) of \(-35.1\text{dBm}\).
The 1-dB compression point $P_{1\text{dB}}$ of the RF front-end can be measured by sweeping the input power and observing the output amplitude of the PGA. As shown in Fig. 10, the measured input-referred 1-dB compression point is $-29$ dBm with a 29.5 dB gain setting. Fig. 11 shows the measured IMRR with a single-tone input. With a 4.092 MHz IF center frequency, the IMRR is 39.12 dB for a bandwidth configuration of 2.5 MHz, and is 39.15 dB for a bandwidth configuration of 5 MHz.

The measured maximum gain is 107.2 dB with a dynamic range of 78 dB. The performance of the proposed receiver and comparisons with the-state-of-the-art GNSS receivers are listed in Table II.

![Fig. 10. Measured 1 dB compression point at minimum gain setting.](image1)

![Fig. 11. Measured IMRR.](image2)
5 Conclusion

A reconfigurable GNSS receiver has been designed and implemented in a 180 nm CMOS process for GPS L1, Galileo E1 and Compass B1 bands. A pseudo-differential LNA is utilized to achieve a single-ended-to-differential transformation. A current-mode double-balanced passive mixer along with a pair of TIAs is adopted to achieve good linearity, less power consumption, area efficiency and better 1/f noise. The three kinds of GNSS signals are down-converted to the same IF band with a central frequency of 4.092 MHz and a reconfigurable bandwidth of 2 or 4 MHz, with the assistance of a tunable LO frequency. This RF front-end achieves a NF of 1.8 dB, a maximum gain of 107.2 dB with a dynamic range of 78 dB, an IMRR of 39.1 dB, an input-referred 1-dB compression point of −29 dBm and an IIP3 of −19.3 dBm. The receiver occupies a die area of about 1.36 mm², drawing a total 16 mA current from a 1.8 V voltage supply.

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