DESIGN NOTES

Optimizing the Performance of Very Wideband Direct Conversion Receivers

Design Note 1027

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Introduction

Zero-IF receivers are not new; they have been around for some time and are prominently used in cell phone handsets. However their use in high performance wireless base stations has had limited success. This is due primarily to their limited dynamic range and that they are less well understood. A new wide bandwidth zero-IF IQ demodulator helps relieve the dynamic range and bandwidth shortcomings for main as well as DPD (digital predistortion) receivers, and enables 4G base stations to cost effectively address the ever-increasing bandwidth needs of mobile access. This article discusses how to optimize performance by minimizing IM2 nonlinearity and DC offset that reduce the dynamic range of zero-IF receivers, thus offering a viable alternative to an otherwise challenging design.

Pushing Ever Wider Bandwidth

Until recently, most base stations needed to only deal with a 20MHz wide channel bandwidth, typically allocated to various wireless carriers. Associated with this 20MHz channel is a companion 100MHz bandwidth DPD receiver to measure intermodulation distortion spurs up to 5th order for effective distortion cancellation. These requirements can generally be met effectively with high-IF (heterodyne) receivers. Nowadays though, such designs are more challenging, with industry trends pushing for base stations to support operation over the entire 60MHz bands. Accomplishing this feat has significant cost savings implications for the entire wireless manufacturing, installation and deployment business model.

To accommodate the three times increase in bandwidth, the DPD receiver bandwidth must increase from 100MHz to 300MHz. In 75MHz bands, the DPD bandwidth grows to a staggering 375MHz. The design of receivers that can support this bandwidth is not trivial. Noise increases due to the wider bandwidth, gain flatness becomes more difficult to achieve, and the required sampling rate of A/D converters increases dramatically. Furthermore, the cost of such higher bandwidth components is appreciably higher.

The modest bandwidth of a traditional high-IF receiver is no longer sufficient to support the 300MHz or higher DPD signals with typically ±0.5dB gain flatness. The 300MHz baseband bandwidth would require choosing an IF frequency of 150MHz at a minimum. It is not trivial to find an A/D converter capable of a sampling rate upward of 600Msps that is reasonably priced, even at 12-bit resolution. One may have to compromise and resort to a 10-bit converter.

New IQ Demodulator Eases Bandwidth Constraints

The LTC5585 IQ demodulator is designed to support direct conversion, thus allowing a receiver to demodulate the aforementioned 300MHz wide RF signal directly to baseband (see Sidebar: Theory of Operation of a Zero-IF Receiver). The I and Q outputs are demodulated to a 150MHz wide signal, only half the bandwidth of a high-IF receiver. In order to attain a passband gain flatness of ±0.5dB, the device's –3dB corner must extend well above 500MHz.

The LTC5585 supports this wide bandwidth with a tunable baseband output stage. The differential I and Q output ports have a 100Ω pull-up to V_{CC} in parallel with a filter capacitance of about 6pF (see Figure 1). This simple R-C network allows for the formation of off-chip lowpass or bandpass filter networks to remove high level out-of-band blockers and equalization of gain roll-off the baseband amplifier chain that follows the demodulator. With a 100Ω differential output loading resistance in addition to the external 100Ω pull-up resistors, the –3dB bandwidth reaches 850MHz.

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Baseband Bandwidth Extension

A single L-C filter section can be used to further extend the bandwidth of the baseband output. Figure 1 shows the chip's baseband equivalent circuit with baseband bandwidth extension. With 200Ω loading, the –0.5dB bandwidth can be extended from 250MHz to 630MHz using a series inductance of 18nH and a shunt capacitance of 4.7pF. Figure 2 shows the variety of output responses that are possible with different loading. One response is with differential loading resistances of 200Ω and 10kΩ. For 10kΩ loading, the –0.5dB bandwidth can be extended from 150MHz to 360MHz using a series inductance of 47nH and a shunt capacitance of 4.7pF.

Second-Order Intermodulation Distortion Spurs Matter

In a direct conversion receiver, the second order intermodulation distortion products (IM2) fall directly in-band at the baseband frequencies. Take, for example, two equal power RF signals, f1 and f2, spaced 1MHz apart at 2140MHz and 2141MHz, respectively, while the LO is spaced 10MHz apart at 2130MHz. The resultant IM2 spur would fall at f2 – f1, or 1MHz. The LTC5585 has the unique ability to adjust for minimum IM2 spurs independently on the I and Q channels by using external control voltages. Figure 3 shows a typical setup for IIP2 measurement and calibration. The differential baseband

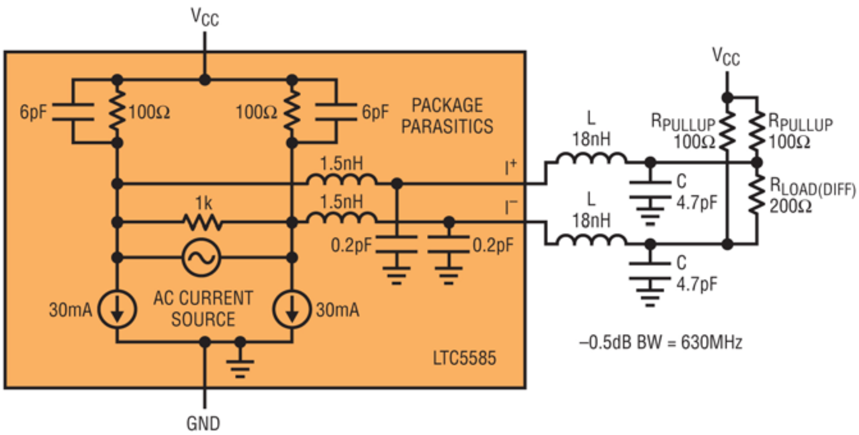

Figure 1. Baseband Output Equivalent Circuit for Bandwidth Extension with L = 18nH and C = 4.7pF

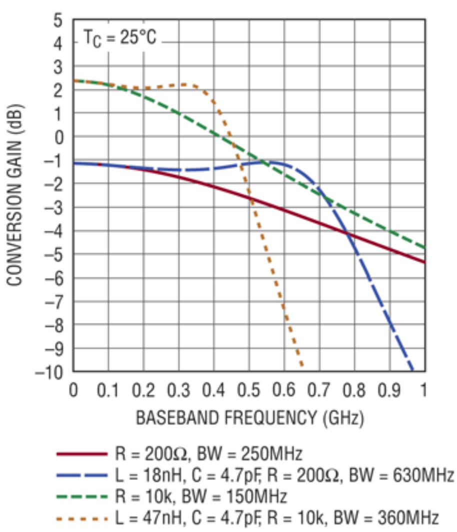

Figure 2. Conversion Gain vs Baseband Frequency with Differential Loading Resistance and L-C Bandwidth Extension

outputs are combined using a balun and the 1MHz IM2 difference frequency component is selected with a lowpass filter to prevent the strong main tones at 10MHz and 11MHz from compressing the spectrum analyzer front end. Without the lowpass filter, 20dB to 30dB of attenuation and long average measurement times are necessary on the spectrum analyzer to attain a good measurement. As shown in the output spectrum of Figure 4, the IM2 component predictably falls in-band at 1MHz. The plot also shows the IM2 product before and after adjustment, reducing the spur level by approximately 20dB by adjusting the control voltages on the IP2I and IP2Q pins. This adjustment reduces the IM2 spur down to a level of –81.37dBc.

With this IIP2 optimization capability, two possible strategies of IP2 calibration are possible. One option is a set-and-forget calibration step performed at the factory. In this case, a simple trim potentiometer for each adjustment pin suffices, as illustrated in Figure 3. Alternatively, an automatic, closed loop calibration algorithm can be implemented in software, allowing the equipment to be calibrated on a periodic basis. For DPD receivers that are already monitoring their transmitters' output, this is trivial as the transmitters can easily generate the two test tones. For main receivers, this calibration may involve additional hardware to loop back the two test tones to the receiver channel. In any event these can all be performed during an off-line calibration cycle. Such an approach would take into account the actual operating environmental factors that may affect base station performance.

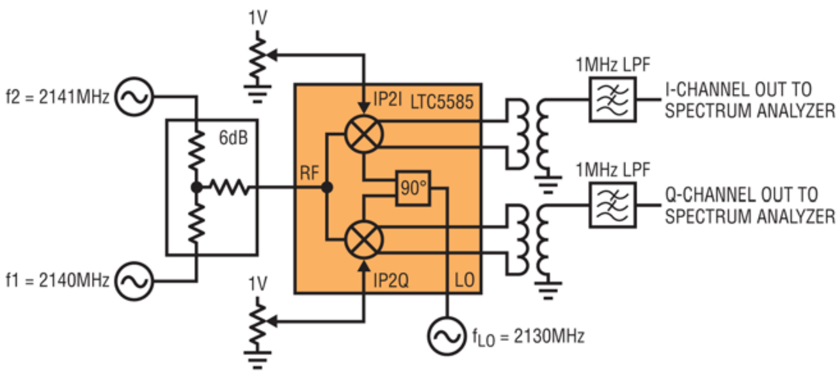

Figure 3. Test Setup for IIP2 Calibration with 1MHz Lowpass Filters to Select the IM2 Component

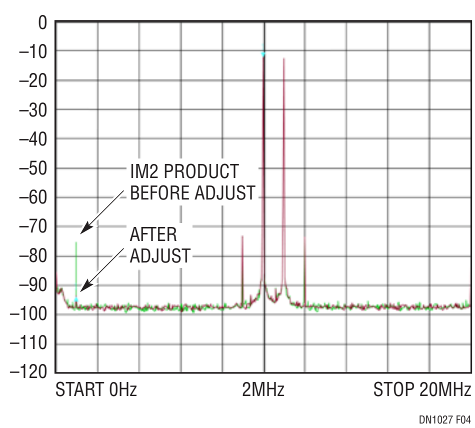

Figure 4. Output Spectrum without Lowpass

DC Offset Voltage Null Helps to Optimize A/D Converter Dynamic Range

A similar adjustment capability is also integrated into the chip to zero out the I and Q's DC output voltage. DC offset, a product arising from internal mismatch and self-mixing of the LO and RF input leakages, can diminish the ADC's dynamic range when the signal chain is DC coupled throughout. To illustrate, a modest 10mV of output DC offset voltage, when passed through a 20dB gain stage, would result in 100mV of DC offset at the input of the A/D converter. With $2V_{P-P}$ input range of a 12-bit ADC, this amount of DC offset represents 205 LSBs of headroom reduction, or effectively reducing the ADC's dynamic range by 0.9dB.

To minimize the leakage between the LO and RF inputs, care should be taken to isolate these two signals. In the PCB layout, separate these two signal traces from one another to prevent cross-coupling. The LO signal, even if there is measurable leakage to the RF port, will self-mix to form a DC offset term at the output. Fortunately the LO level is usually constant, so the DC offset voltage is also constant and can be easily canceled by the adjustment. More problematic is the RF input, which can vary over wide signal levels. Any signal leakage to the LO input would self-mix and produce a dynamic DC offset voltage as the signal varies. This will distort the demodulated signal. So keeping the leakage small helps reduce the DC offset to a minimum.

Potential Cost Benefits of Direct Conversion Receivers

A zero-IF receiver is particularly compelling due to its potential cost savings. As mentioned above, the RF signal demodulates to a low frequency baseband. At lower frequencies, the design of the filter becomes easier. Furthermore, zero-IF demodulation produces no image at the baseband, thus eliminating the need for a relatively expensive SAW filter. Perhaps most attractive is that the ADC sampling rate can be significantly reduced. In the example above, the 150MHz I and Q baseband bandwidth can be effectively addressed with a dual 310Msps ADC such as LTC2258-14, without resorting to a much more expensive higher sampling rate ADC.

Conclusion

As the bandwidth and performance of wireless receivers increase, a new wideband quadrature demodulator offers an alternative approach that helps to address architectural shortcomings and raises receiver performance while reducing systems costs.

SIDEBAR

THEORY OF OPERATION OF IQ DEMODULATION

IQ Demodulation

The operation of an IQ demodulator can be explained by representing its RF input signal $S_{RF}(t)$ as a combination of two double sideband modulated quadrature carriers:

$$S_{RF}(t) = S_I(t) + S_Q(t) = I(t)\cos\omega_{RF}t - Q(t)\sin\omega_{RF}t \quad (1)$$

As illustrated in Figure A, the in-phase component I(t) and quadrature component Q(t) are baseband signals that can be viewed as inputs to an ideal IQ modulator generating $S_{RF}(t)$.

An IQ demodulator achieves perfect reconstruction of I(t) and Q(t) by exploiting the quadrature phase relation between $S_I(t)$ and $S_Q(t)$. The frequency-domain representation of a –90° phase-shift corresponds to multiplication by the Hilbert transform:

$$H(j\omega) = j\,\mathrm{sgn}(\omega) \quad (2)$$

It converts a spectrum with even symmetry around $\omega = 0$ to a spectrum with odd symmetry and vice versa. The spectra of $S_I(t)$ and $S_Q(t)$ therefore exhibit different symmetry; $S_I(t)$ has even symmetry, $S_Q(t)$ has odd symmetry. Downconversion of the even RF input component $S_I(t)$ with the even LO (cosine) retrieves I(t), while $S_Q(t)$ with the odd LO (sine) retrieves Q(t). Cross-combinations of even and odd yield zero.

An error φ on the quadrature relation between the LO outputs causes crosstalk between the I- and Q-channels. Using the I-phase channel as reference, an even component is introduced in the Q-channel LO:

$$\sin(\omega_{RF}t + \varphi) = \sin(\omega_{RF}t)\cos\varphi + \cos(\omega_{RF}t)\sin\varphi \quad (3)$$

resulting in a contribution of I(t) to the Q-channel output $Q_{OUT}(t)$:

$$Q_{OUT}(t) = Q(t)\cos\varphi + I(t)\sin\varphi \quad (4)$$

Image Cancellation Receiver

Another IQ demodulator application is an image rejection/cancellation receiver with non-zero IF frequency, as shown in Figure B.

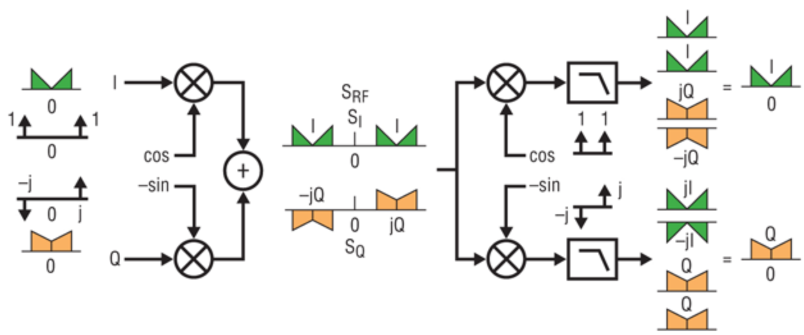

Figure A. Concept of IQ Modulation and IQ Demodulation

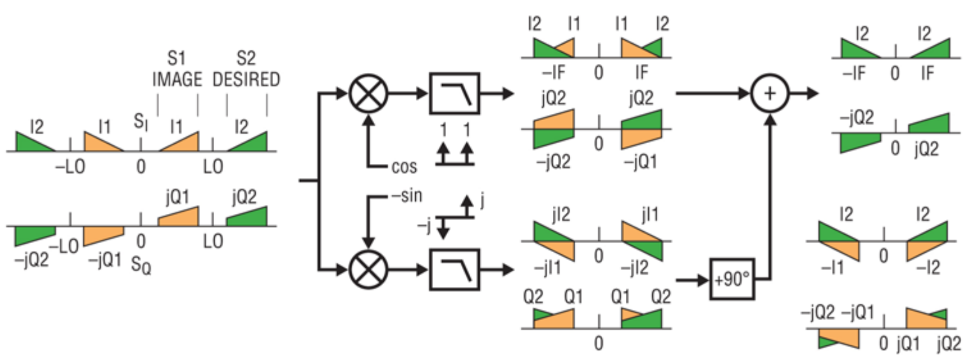

Figure B. Operation of the Hartley Image Rejection Receiver

The I-channel preserves the symmetry in the RF input signal, while the Q-channel converts even components to odd and vice versa. The extra 90° phase shifter restores the original symmetry in the Q-channel, but with opposite sign for the signals $S_1(t)$ and $S_2(t)$; the phase of $S_2(t)$ is ahead of the LO since its center frequency is higher, while the phase of $S_1(t)$ lags behind. Addition to the I-channel reconstructs the downconverted signal $S_2(t)$; subtraction reconstructs $S_1(t)$.

The image rejection (IR) is degraded in the presence of a quadrature phase error φ or gain mismatch α between I- and Q-channels. The phase error introduces crosstalk between the channels, while gain mismatch results in imperfect cancellation by the adder:

$$IR = 10\log\left(\frac{1+\alpha^2+2\alpha\cos\varphi}{1+\alpha^2-2\alpha\cos\varphi}\right) \quad (5)$$

Figure C depicts the result for different gain and phase error combinations. Small gain errors have a larger impact than small phase errors.

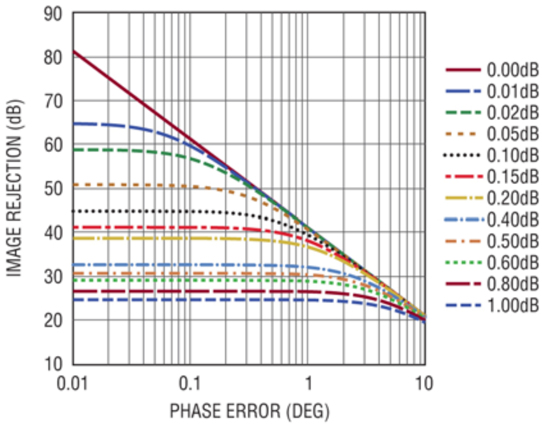

Figure C. Image Rejection vs Phase Error for Different I/Q Gain Mismatch

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