Using a complex-baseband architecture in FMCW radar systems

Texas Instruments

Karthik Ramasubramanian
Radar Systems Manager

Texas Instruments

Abstract

This white paper explains the advantages of a complex-baseband architecture in frequency-modulated continuous wave (FMCW) radar systems. Typical radar front-end implementations use a real mixer with a real baseband and analog-to digital converter (ADC) chain. However, there are performance advantages that can be leveraged with the use of a quadrature mixer and complex-baseband architecture in the context of FMCW radar. This architecture has been implemented in the 76–81-GHz fully integrated complementary metal-oxide semiconductor (CMOS) millimeter wave (mmWave) sensors from Texas Instruments.

The first part of this white paper describes the complex-baseband architecture in the context of FMCW radar and the advantages of this architecture. The second part explains how complex baseband does not increase the memory requirements or computational burden on the digital signal processing (DSP) side.

Introduction

Historically, radar implementations used discrete components (power amplifiers [PAs], low-noise amplifiers [LNAs], voltage-controlled oscillators [VCOs], analog-to-digital converters [ADCs]), but more integrated solutions are now becoming available. A complementary metal-oxide semiconductor (CMOS)-based radar that integrates all radio-frequency (RF) and analog functionality as well as digital signal processing (DSP) capability into a single chip represents the ultimate radar system-on-chip solution. Such a highly integrated device significantly simplifies radar sensor implementations, enables a compact form factor for the sensor, and makes the solution cost effective. Texas Instruments (TI) offers a family of highly integrated 76–81-GHz radar devices for the rapidly growing automotive and industrial radar markets.

This white paper focuses on a particular aspect of Texas Instruments 76–81-GHz radar devices: the use of a quadrature mixer and complex-baseband architecture instead of the traditional real mixer and real baseband architecture.

FMCW radar concept

Let's review the operating principle behind FMCW radars. In FMCW radar solutions, the transmitted signal is a linear frequency-modulated continuous wave (L-FMCW) chirp sequence, whose frequency vs. time characteristic follows the sawtooth pattern shown in **Figure 1** on the following page. The frequency $f_T(t)$ and phase $\Phi_T(t)$ of the linear FMCW transmit chirp are expressed as linear and quadratic functions of time, as shown in **Figure 1**.

In a typical FMCW radar implementation (**Figure 2** on the following page), the local oscillator (LO) module generates a linear frequency-modulated continuous wave signal, $\cos(\Phi_T(t))$, which is amplified by the PA and transmitted from the antenna.

Any object(s) present in the region of interest illuminated by the radar reflect the transmitted

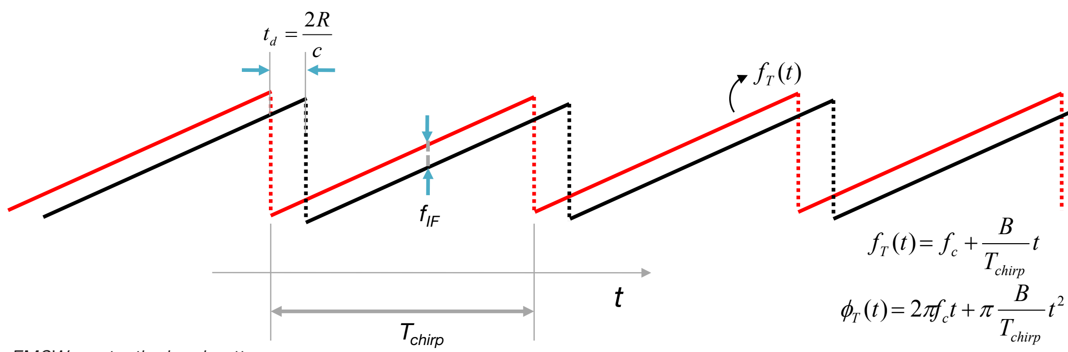

***Figure 1**. FMCW sawtooth signal pattern.*

signal. The receive antenna receives the reflected signal and the LNA amplifies it. This received signal mixes with the LO signal to produce the beat-frequency (intermediate-frequency [IF]) output, which the ADC digitizes and the DSP subsequently processes.

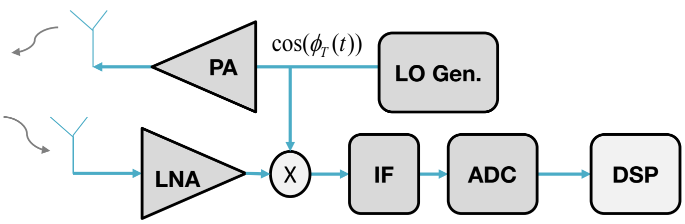

***Figure 2**. High-level block diagram of an FMCW radar.*

Figure 3 depicts the nature of the received FMCW signal, which comprises different delayed and attenuated copies of the transmit signal corresponding to various objects. You can see that the beat-frequency signal corresponding to each object is a tone (ignoring the edge effects at the start and end of the chirp), whose frequency, f_b, is proportional to the distance of the object from the radar, *R*. Therefore, the process of detecting objects (targets) and their respective distances from the radar involves taking a fast Fourier transform (FFT) of the beat-frequency signal and identifying peaks that stand out from the noise floor.

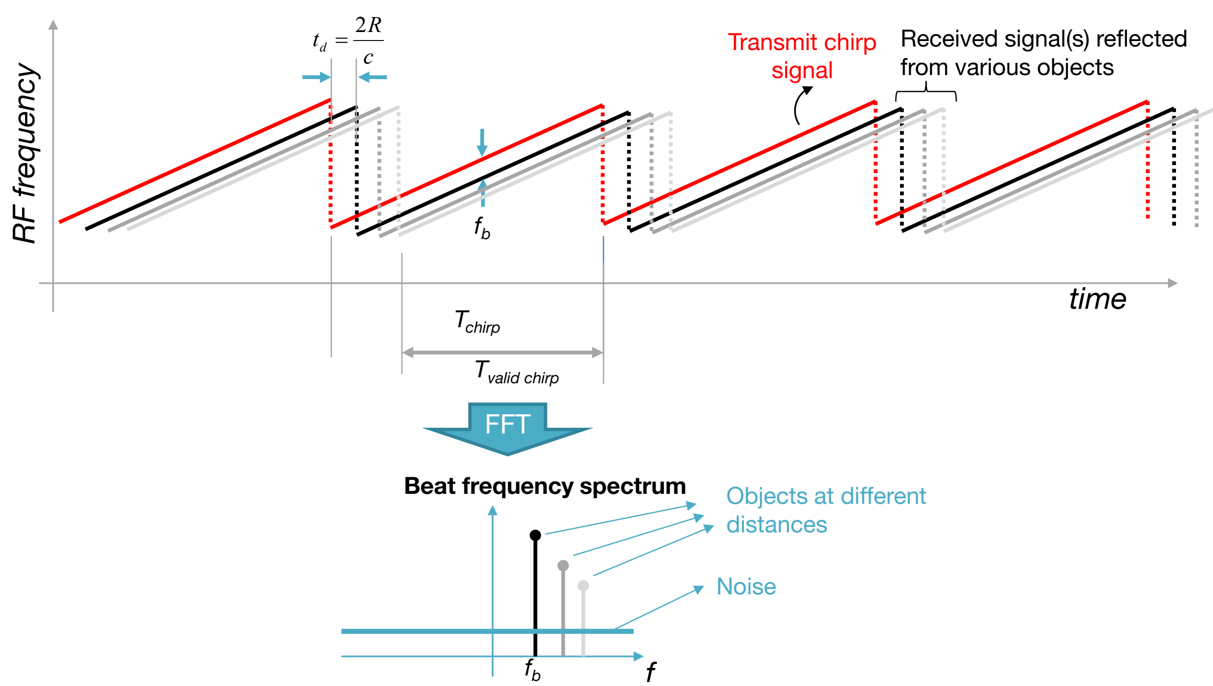

***Figure 3**. Received FMCW radar signal and beat-frequency spectrum.*

In reality, there are many details beyond this very simplified explanation of FMCW radar detection, particularly when considering moving objects. For moving objects, the beat-frequency signal also has a Doppler component that depends on the relative velocity between the radar and the target. You can estimate the Doppler component—and hence the relative velocity—by performing a second FFT across chirps and looking at the phase shift of the beat signal from one chirp to the next. To summarize, the detection process involves performing a first-dimension FFT of the received samples corresponding to each chirp and then a second-dimension FFT of this output across chirps. The result of the two-dimensional FFT procedure is an image of the target(s) in the range-velocity grid. The detection process occurs on the 2-D FFT output and involves detecting peaks amid the noise floor or surrounding clutter.

In most implementations, there is also an angle-estimation process based on beamforming with multiple antennas, although I won't get into those details in this white paper.

FMCW radar implementation using real baseband

Most radar implementations today use a real mixer and real (I-only) baseband and ADC chain. This type of implementation is partly motivated by the cost advantages obtained by not having to double the number of ADCs and variable gain amplifiers (VGAs) in discrete-solution-based radar implementations.

Figure 4 illustrates the instantaneous spectrum of the transmit (LO), receive (RX) and beat-frequency (IF) signals. **Figure 4a** shows the LO signal cos(Φ_T*(t))* spectrum representing the instantaneous frequency of the ramping LO. The RX signal spectrum in **Figure 4b** contains delayed and

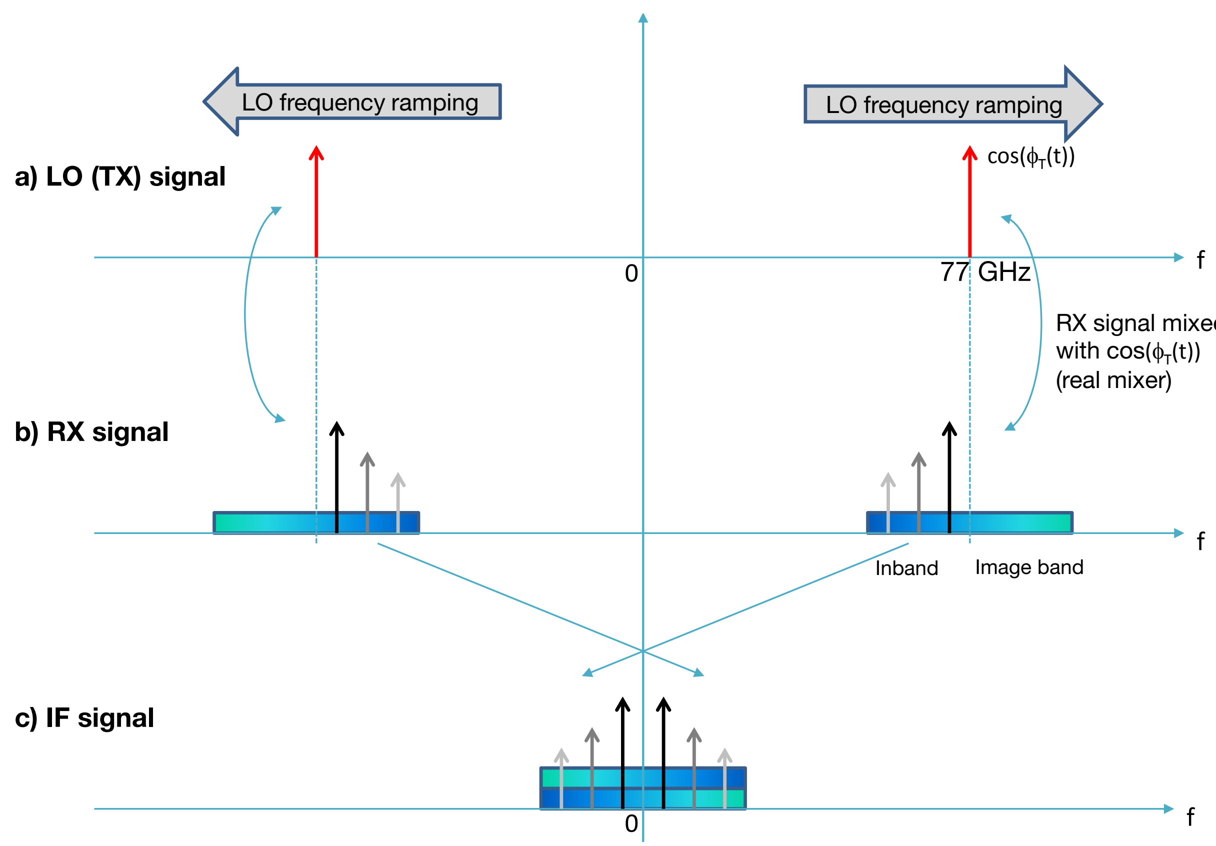

Figure 4*. A real mixer and real baseband suffers from image-band noise foldback. (a) Instantaneous spectrum of LO (TX) signal showing ramping frequency, (b) RX signal after reflection from various objects, (c) IF signal after real mixer.*

attenuated versions of the LO signal representing different targets. The signal of interest is contained in the "in-band" portion of the RX signal spectrum, while the "image band" portion of the spectrum is devoid of any signal of interest. This is because the received signal is always "delayed" with respect to the transmit LO signal. Therefore, the beat frequency corresponding to different objects always falls on one side of the complex-baseband spectrum. The thermal noise floor, shown as a blue horizontal bar, is spread across both the in-band and image band.

Looking at **Figure 4c**, when using a real mixer and real baseband chain, the IF signal spectrum after the mixer suffers from image-band noise foldback. In other words, the IF signal experiences a signal-to-noise (SNR) ratio loss caused by noise from both the in-band and image band. This leads to a performance loss of up to 3 dB that is avoidable with a complex-baseband chain, as you'll see.

Complex-baseband implementation

The block diagram in **Figure 5** shows the use of a quadrature mixer and complex-baseband architecture. In this case, the received signal mixes with the cos() and sin() versions of the LO, with a duplicated IF chain and ADC for the in-phase (I) and quadrature (Q) channels.

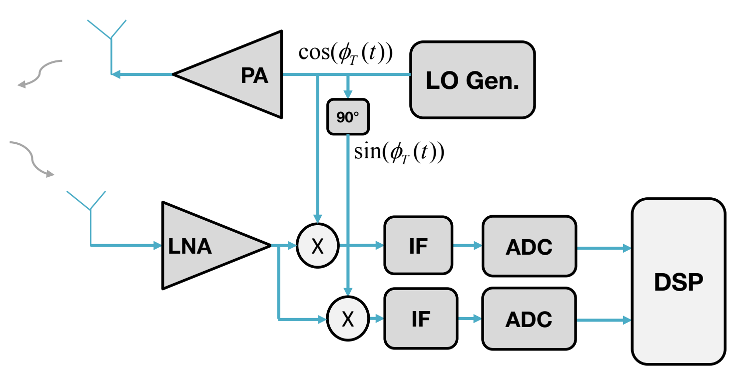

***Figure 5**. The complex-baseband architecture.*

Figure 6 on the following page shows the spectrum of various signals in a quadrature mixer and complex-baseband implementation. Since the RX signal mixes with cos($\Phi_T(t)$) + jsin($\Phi_T(t)$) in a quadrature mixer, the in-band and image bands remain separate, and there is no noise increase due to image-band noise foldback. Thus, there is an overall noise-figure advantage possible with this architecture.

Let's discuss some of the key advantages of a complex-baseband architecture in FMCW radars.

Improved noise figure

The most straightforward benefit of the complex-baseband architecture is the noise-figure improvement achievable by eliminating the image-band noise foldback. Compared to the single-sideband (SSB) noise figure representative of a real-only implementation, here the effective noise figure corresponds to the improved double-sideband (DSB) noise figure.

In theory, the noise-figure improvement can be as much as 3 dB; in practice, the noise-figure improvement will be somewhat smaller and implementation specific, due to the signal power loss associated with splitting the received signal into the I and Q paths after the LNA, and the resulting higher contribution of IF noise to the overall noise figure. Nevertheless, there is an effective noise-figure improvement in a complex-baseband implementation.

This improvement is particularly important when considering radar systems dominated by a TX noise (amplitude noise or uncorrelated phase noise) skirt. In these systems, the noise skirt from antenna coupling or bumper reflection dominates the RX thermal noise floor. Under such conditions, a complex-baseband architecture realizes the full 3-dB noise-figure benefit.

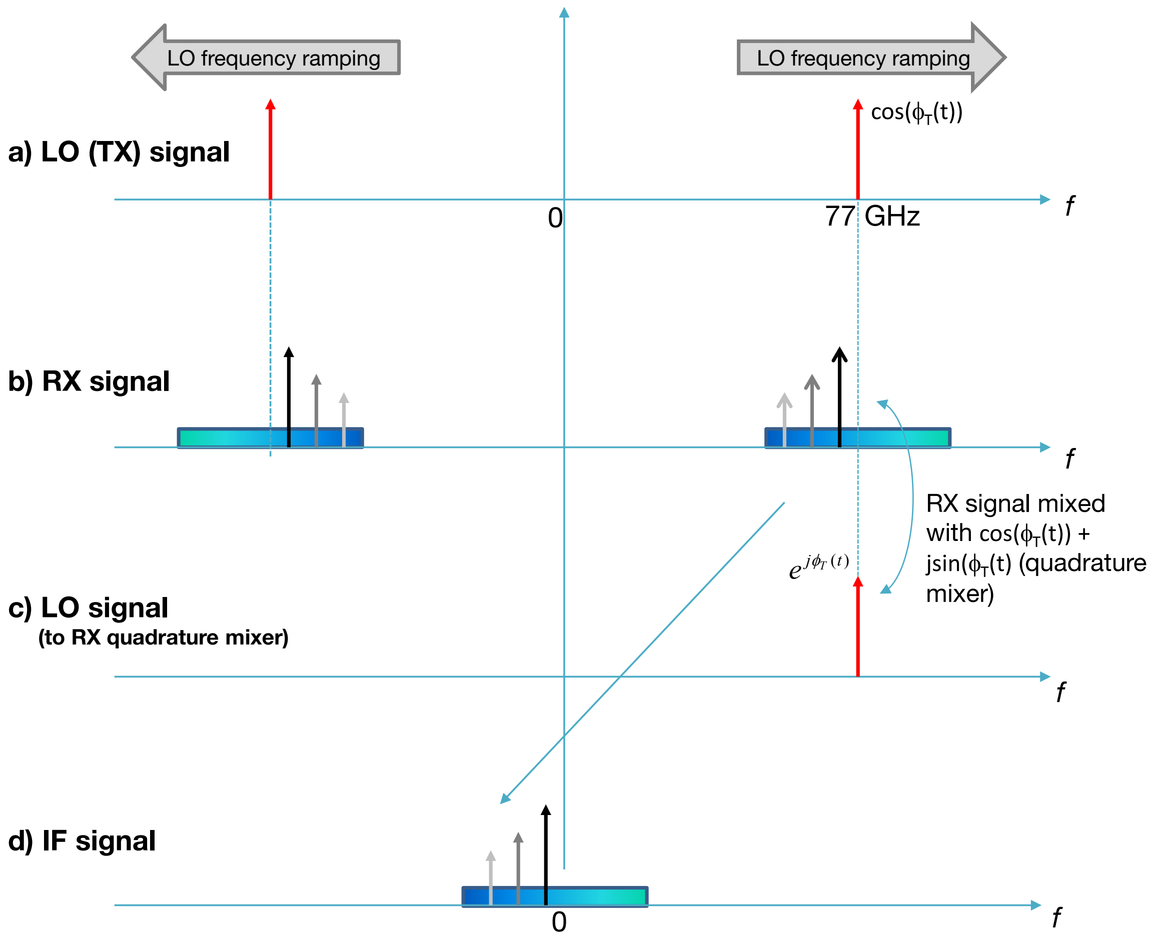

***Figure 6**. Quadrature mixer and complex baseband with no image-band noise foldback. (a) Instantaneous spectrum of LO (TX) signal showing ramping frequency, (b) RX signal after reflection from various objects, (c) Quadrature LO signal, (d) IF signal after quadrature mixer.*

Improved interference tolerance

In an FMCW radar, the image band contains only noise and is free of any desired signal. Therefore, with a complex-baseband implementation, you can monitor the image-band spectrum to detect interference and/or estimate the thermal noise level accurately without clutter.

For example, you can easily identify the presence of a tone or energy spike in the image band as coming from an interfering radar device without any ambiguity over whether it could be a genuine object of interest—in other words, detecting and mitigating interference from jamming radars without any ambiguity over genuine objects.

Also, since a complex-baseband architecture prevents image-band foldback, it enjoys greater robustness against any interference present in the image band. In a real baseband architecture, any interference present in or sweeping through the image band will also fold back in-band, thereby becoming more susceptible to performance loss.

Digital frequency/phase shift for RF delay compensation

In a typical radar implementation supporting multiple RX chains (for RX beamforming), the antenna routing delays and/or RF circuit delays of all RX chains must match in order to ensure proper beamforming functionality. This poses constraints on board routing, as well as RF component matching across channels.

In this context, from **Figure 3** you can see that in FMCW radar signals, a "delay" is equivalent to a "frequency shift." The beat frequency, f_b, is proportional to the round-trip delay, t_d, as shown in the figure. Based on this observation, it is possible to compensate for various delays in the radar system through the use of digital frequency and/or phase shift.

With the complex-baseband architecture, any delay mismatch and/or RF phase response mismatch across channels can be digitally compensated elegantly using a complex-baseband output, even before FFT processing, by using different digital frequency/phase de-rotations on the I and Q complex data samples corresponding to each RX channel.

Reduced impact of RF intermodulation products

It's a well-known fact that RF nonlinearity (for example, cubic nonlinearity) results in the creation of intermodulation products at $(2f_1–f_2)$ and $(2f_2–f_1)$ when there are two tones at the f_1 and f_2 input frequencies.

In FMCW radar receivers, the presence of a strong antenna coupling or bumper reflection signal (say, at power level P_1 and frequency f_1), together with a desired strong object (say, at power level P_2 and frequency f_2), can result in intermodulation products that result in ghost objects.

In most cases, the antenna coupling or bumper reflection signal (P_1) would be large and close to DC (with f_1 close to zero); therefore, the intermodulation product at $2f_1–f_2$ would be relatively large and fall in the image band (approximately at $–f_2$). In a real-only implementation, this intermodulation product will fold back in-band and degrade the SNR of the actual object at f_2. A complex-baseband implementation significantly mitigates this problem, since the image band does not fold back.

Redundancy for functional safety monitoring

The availability of dual (I and Q) IF and ADC channels indirectly provides a form of redundancy that can help with functional safety monitoring. Here again, in a fully functioning system, the image band is void of any desired signal; therefore, you can observe image-band energy in relation to in-band energy to detect failures in either the I or Q channel, which improves functional safety monitoring for the IF and ADC sections.

Improved bumper signature and nearby object detection

A complex-baseband architecture enables accurate estimations of the amplitude and phase of the bumper reflection and/or objects very nearby. Specifically, given that the beat frequencies from bumper reflection and very nearby objects are at a low frequency (close to DC), the availability of I and Q outputs enables more accurate estimations of the frequency and phase of these signals. Such estimates are much more difficult with a real-only chain given the low frequency of the signals and the short observation window available during a chirp.

TI's 76–81-GHz integrated mmWave sensing solutions implement the complex-baseband architecture and also include digital baseband circuitry that help leverage the advantages outlined here.

DSP requirements

The duplication of IF and ADC to support the complex-baseband architecture does not lead to an increased burden in memory or processing requirements on the DSP. Let's discuss why.

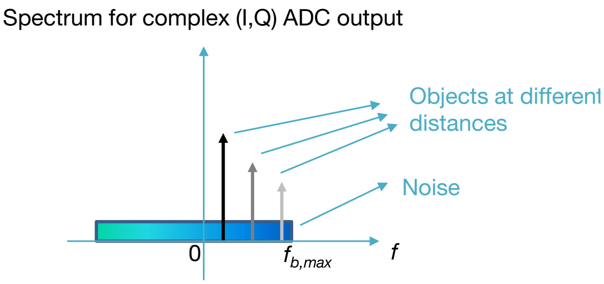

***Figure 7**. Beat-frequency spectrum in FMCW radar.*

Consider the beat-frequency spectrum for the complex baseband shown in **Figure 7**. This figure shows a flipped version of the spectrum from **Figure 6d**, just for the sake of convenience, so that all objects appear on the positive frequency side, with farther objects seen at a larger frequency. In **Figure 7**, $f_{b,max}$ denotes the maximum beat frequency corresponding to the farthest object of interest.

In a real-only traditional implementation, ADC samples need to go out to the DSP at a minimum (Nyquist) sampling rate of $2f_{b,max}$. See the left-hand side of **Figure 8**, which shows the real-only spectrum with the higher noise figure.

In a complex-baseband implementation, shown on the right-hand side of **Figure 8**, it is not necessary to double the ADC output interface rate. In fact, it is possible to frequency-shift the spectrum, perform image-reject filtering and send out decimated I and Q ADC samples to the DSP at $f_{b,max}$. Thus, the interface rate of the ADC samples going to the DSP does not really increase due to the use of the complex baseband—the real output at $2f_{b,max}$ changes to a complex output at $f_{b,max}$. The frequency shift to center the spectrum around DC helps simplify the implementation of image-reject filtering.

TI's radar chip includes a built-in digital frequency shifter to frequency-shift the samples, perform image-reject filtering and send out the complex-baseband output at the reduced interface rate (similar to a real-only implementation).

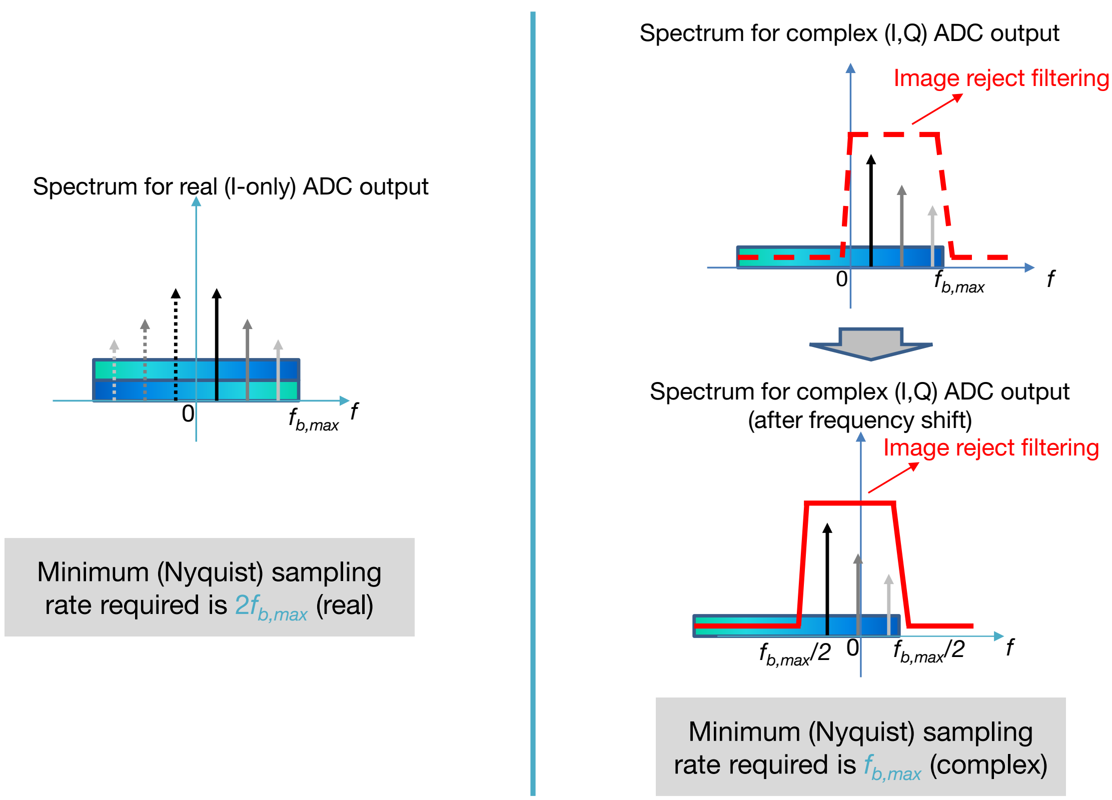

***Figure 8**. Beat-frequency spectrum for real and complex-baseband outputs.*

Another advantage is related to memory and million-instructions-per-second (MIPS) requirements for processing on the DSP. A real-only implementation needs to compute a 2N-point FFT with real samples, while a complex-baseband implementation needs to compute an N-point FFT with complex input samples. Most DSP architectures can achieve both with similar complexity. In fact, the N-point complex FFT would consume lower MIPS than the 2N-point real FFT, making it advantageous to use the complex-baseband output. Similarly, the memory requirement for M chirps/frame is the same for both options. **Table 1** summarizes the comparison between the complex-baseband and real-only options.

Summary

The use of a complex-baseband architecture in FMCW radar systems enables various performance benefits without any penalty in ADC interface rate or memory/MIPS requirements on the DSP. In highly integrated CMOS radar solutions this architecture is implemented efficiently, at low cost and with low power.

Considering the noise-figure improvement, there is no significant penalty in current consumption, since you can trade off the better noise figure against the on/off duty cycle of operation to effectively reduce current consumption. Thus, the complex-baseband architecture is a useful feature that you can leverage through TI's integrated radar solutions.

Comparison item	Complex-baseband option	Real-only option	Comments
ADC output data rate	Complex (I,Q) samples at $f_{b,max}$	Real (I-only) samples at $2f_{b,max}$	Both options are similar
FFT complexity ($N = T_c f_{b,max}$)	N-point FFT with complex input	2N-point FFT with real input	Both options are similar, with complex baseband having a slight advantage (2N-point FFT of real samples is possible using N-point complex FFT, plus a few additional operations)
Memory requirement (for M chirps/frame, for 1 RX)	NM complex samples to be stored	NM complex samples to be stored (negative frequency components discarded after 2N-point FFT of real input)	Both options are similar
Noise figure	Better than baseline by up to 3dB	Baseline	Advantage with complex baseband

Table 1. *Data rate, MIPS and memory-requirement comparison.*

Mailing Address: Texas Instruments, Post Office Box 655303, Dallas, Texas 75265