Technical Article

How to Select the Best ADC for Radar Phased Array Applications—Part 1

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Abstract

Many papers discuss the system trade-offs and relative merits of digital vs. RF beamforming, and the hybrid blend thereof.[1] Building on prior work, this article uses RF-to-ADC cascade modeling to show dynamic range (linearity and noise) and sample rate trade-offs against DC power consumption in a multichannel system with varying channel summation in both the RF and digital realms. The optimal selection of sample rate, ADC ENOB, and RF vs. digital channel combining is weighed against DC power consumption. The popular Schreier and Walden ADC figures of merit (FOMs) are proposed as extensible to a multichannel system to express a single system FOM portraying optimal dynamic range normalized for DC power expense. The article has two parts. Part 1 explains the method of modeling the system, and "How to Select the Best ADC for Radar Phased Array Applications—Part 2" analyzes the results and draws conclusions from system FOMS.

Part 1: System Model Introduction

In the radar phased array community, the quest for every element digital beamforming gets a lot of attention. Significant government funding is leading industry to rapidly mature the capability, and the realm is a hotbed of research. The promise of omnidirectional physical arrays that are digitally beamformed into any combination of simultaneous beams allows software-defined multimission apertures. In other words, the one array that fits a lot of missions. The appeal is multiple independent, simultaneous, software-configurable digital beams and shapes that improve detection and enable multifunction.[1]

In practice, there are multiple challenges to overcome—the biggest being DC power consumption. In elemental digital beamforming the digitizer node (DAC/ADC) is distributed behind every single element and so must be lower power than what is available today, without sacrificing performance. Digitizer power burn is a direct function of dynamic range capability (ENOB) and sample rate. Choosing the optimal ADC bit resolution (ENOB) and sample rate, in the system context of RF performance, power consumption, and digital vs. analog beamforming capability is a complicated multidimensional puzzle for phased array system designers.

Thermal limits and the size footprint are also big factors. Larger phased arrays often employ blades that put the electronics orthogonal to the antenna face, offering lots of thermal and size flexibility. However, some systems especially those at higher frequencies require the electronics to be planar with the antenna and fit within the element lattice spacing. This creates very challenging thermal and size problems. The electronics need to shrink and shed power—without giving up performance.

Every phased array system has different required beam attributes, sensitivity, and dynamic range for a given DC power limit. So, a continuum of optimal performance vs. DC power exists, depending on the mission requirements.

System Figures of Merit

Dynamic range, or spurious-free dynamic range (SFDR), is the most common receiver FOM and is a function of linearity and sensitivity. Do not confuse receiver SFDR with ADC SFDR, they mean completely different things. ADC SFDR is a metric for the max ADC spur among harmonics, interleaving spurs, clock leakage, intermodulation products thereof, etc., and is not a direct representation of two-tone linearity. This article refers only to receiver SFDR. Receiver sensitivity is the minimum signal level that can be detected at some offset threshold from the noise floor (lower is better). Several considerations like waveform type and probability of detection determine the offset threshold, which is set to zero in this paper. Sensitivity does not consider linearity; it is solely a noise metric. A point of emphasis: radar and EW systems operate in blocker environments, so linearity (two-tone intermodulation) is as equally important as noise. Phased array is generally suboctave and cares mostly about IMD3 distortion, whereas EW is multi-octave and cares about IMD2 and IMD3 distortions. Usually, a radar or EW receiver is not optimized just for sensitivity. Linearity (that is, IP2 and IP3) is an important design goal, and receiver SFDR is a handy FOM because it considers both sensitivity and linearity.

SFDR is a single-point FOM that expresses the best-case signal to noise and distortion at a singular best-case RF input power. This occurs when the IMD spurs are at the same level as the noise.[2]

$$\begin{aligned} &\textit{spurious-free dynamic range (SFDR)}\ \text{dB} = \\ &\quad \frac{2}{3}\left(\textit{IIP3}\ \text{dBm} - \textit{Sensitivity}\ \text{dBm}\right) \\ &\textit{Sensitivity}\ \text{dBm} = -174\,\frac{\text{dBm}}{\text{Hz}} + NF + 10Log(IFBW) \end{aligned} \tag{1}$$

Where thermal noise spectral density is –174 dBm/Hz at T = 290 K and IFBW is the bandwidth of the noise channel, often set using a combination of IF and digital filtering.

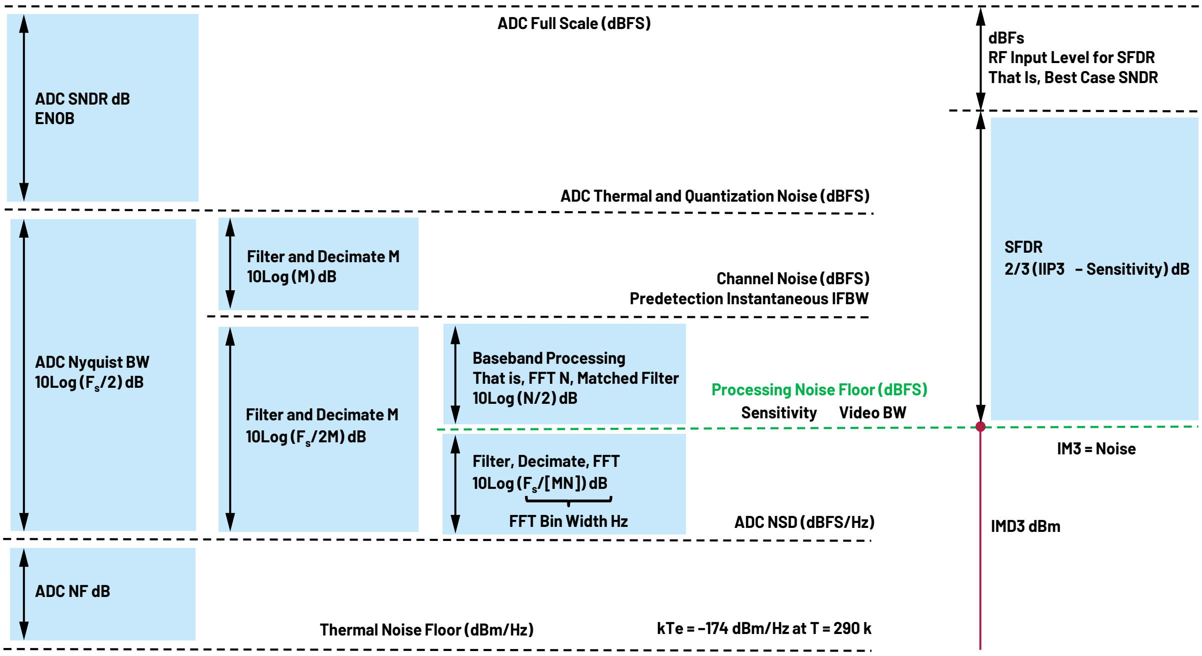

Figure 1. Relationship of ADC SNDR, NF, processing bandwidth, IMD3, and SFDR.[3]

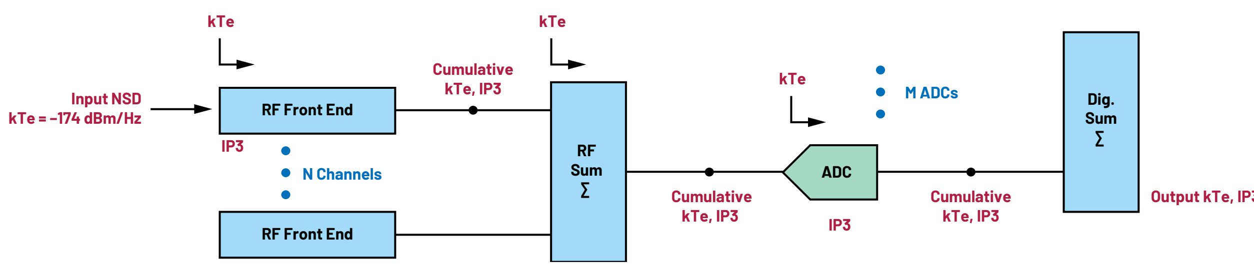

Figure 2. Cascade model of phased array using a combination of analog and digital beamforming and summation.

SFDR and sensitivity are the system performance FOMs analyzed herein. Both SFDR and sensitivity are processing bandwidth dependent. To generalize, the processing bandwidth is set to IFBW = 1 Hz. To adjust for the specific processing bandwidth:

- Add 10log IFBW to sensitivity
- Subtract 2/3 × 10log IFBW from SFDR

A receiver often has simultaneous requirements for both sensitivity and SFDR. Low front-end NF helps both, and high IP3 helps SFDR, but gain helps sensitivity and hurts SFDR. So, a Goldilocks RF front-end gain must be high enough to meet sensitivity but low enough to meet SFDR. A general rule of thumb is a receiver never wants excess RF gain; it wants just enough.

System Cascade Model: Overall

The objective herein is to build a simple MS Excel model that allows swept RF: digital beamforming ratios, swept ADC ENOB, and DC power drawn from Murmann survey data, and to decide the best combination of performance and DC power burn. The cascade model includes the RF front end, RF channel summation, ADC, and digital channel summation. Figure 2 shows the modeled blocks and the notable cascade metrics at each node. The model uses the method for summed-RF channel cascade analysis described by Delos et al.[4] The key to the model is to track device-additive and cumulative noise spectral density kTe at each node, and account for signal gain and noise gain separately (very important).

Using this manner of accounting it is possible to get a negative NF in systems with summed RF channels, which is exactly the desired advantage of coherent summation.

$$NF_{overall}\ \text{dB} = SNR_{in} - SNR_{out}\ \text{dB} = [signal_{in} - noise_{in}] - [signal_{out} - noise_{out}]\ \text{dB}$$

$$NF_{overall}\ \text{dB} = Gain_{signal} - Gain_{Noise}\ \text{dB}$$

$$Gain_{Noise}\ \text{dB} = NSD_{out} - NSD_{in}\ \frac{\text{dBm}}{\text{Hz}}$$

$$NSD\ \frac{\text{dBm}}{\text{Hz}} = 10log_{10}(kTe) \qquad (2)$$

$$k = 1.38 \times 10^{-23}$$

$$T_e = k\,(F-1)$$

Where F is noise figure in linear (not dB) and NSD is noise spectral density.

The following examples use a subarray size of 64 channels. In many of the plots, the horizontal axis shows the model sweep from all-digital summation on the left (64-channel digital sum, no RF sums) to all-RF summation on the right (no digital sum, 64-channel RF sum). In between is a digital and analog summation blend, sometimes called hybrid beamforming with increasing RF sum from left to right. These plots are in the next section. ADC ENOB is swept in the analysis and presented in the plots. Trends in DC power and performance (SFDR and SENS) are analyzed as these parameters are swept.

Modeling the RF Front End

The RF front-end model is an RF black box with attributes gain, NF, IP3, and DC power that are functions of swept attributes. As the system model sweeps the RF: digital sum ratio and ADC ENOB, the RF front-end attributes tune for the best cascade performance. Table 1 provides the equations behind each attribute function.

Table 1. RF Front-End Attribute Equations

Block	Model Attribute	Equation	Note
RF Front End	Gain	−4.2 × ADC_ENOB + 50 dB	
	NF	5 dB	Nominal
	OIP3	ADC_IP3 + 8 − 7 LOG10(N)	N = RF sum ports
	DC power	(N = 1) m × OIP3 + b (mW) (N > 1) 2× (N = 1 case)	M = 0.14, b = 0.02 (model fit)

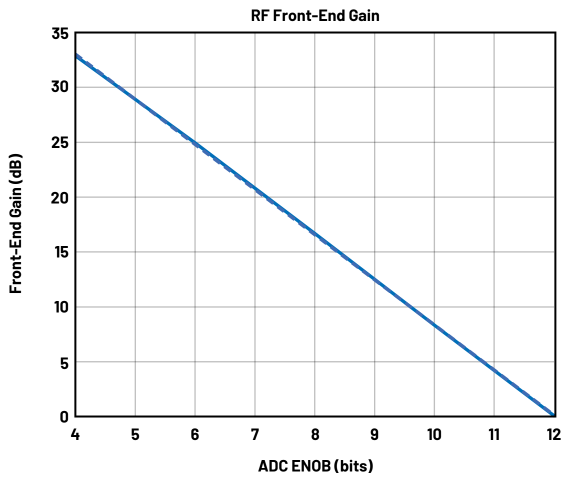

Figure 3. RF front-end gain vs. ADC ENOB.

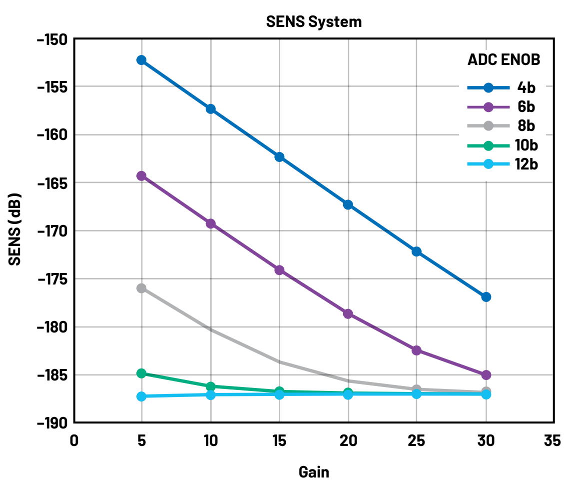

Figure 4. Overall sensitivity vs. front end gain for varying ENOB.

The model sets RF front-end gain as a function of ADC_ENOB, a swept parameter, as shown in Figure 3. The model uses a linear equation to set a minimum viable gain for reasonable system kTe while seeking to maximize SFDR. RF gain is bad for SFDR so it should be minimized just enough to meet NF requirement. Lower-resolution ADCs have a much higher noise figure and require more RF gain in front (at low NF) to set acceptable system NF. In contrast, an ADC with ENOB = 12 has excellent NF and requires no front-end gain, a huge dynamic range benefit albeit at ADC DC power penalty. This is shown in Figure 5 and Figure 6. The ENOB = 8 case sees improving sensitivity and neutral SFDR impact as the gain is increased to ~15 dB. Above 15 dB, SFDR begins to degrade steadily. In contrast, the ENOB = 12 case has a superior ADC NF and as such wants no gain in front. Putting the same 15 dB gain block in front would have a net negative impact on ENOB = 12 performance.

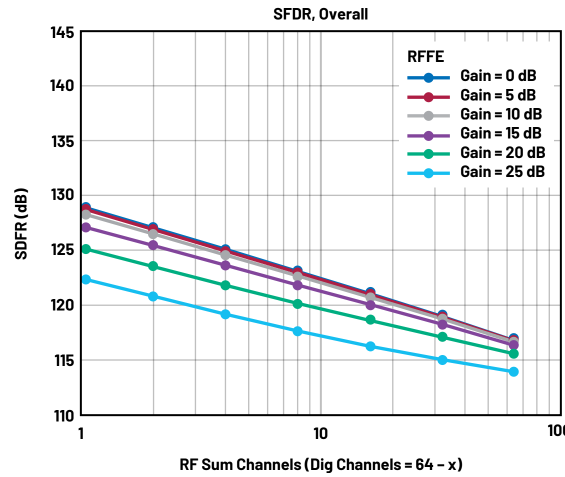

Figure 5. RF front-end gain impact on SFDR for ADC ENOB = 8.

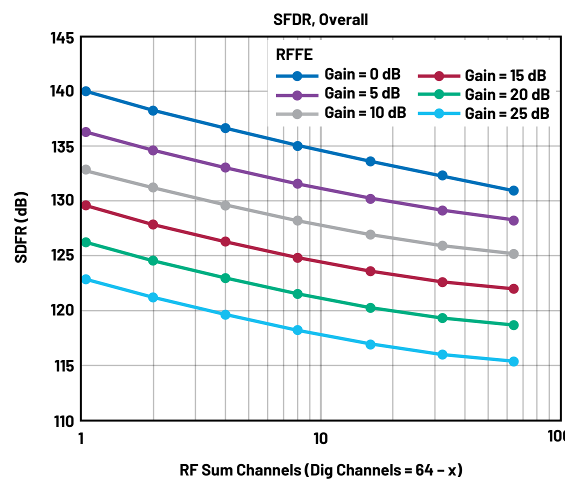

Figure 6. RF front-end gain impact on SFDR for ADC ENOB = 12.

RF front-end noise figure is set at 5 dB and held constant across all simulations. This is a middle of the road value that acknowledges the need for:

- Trade off with high linearity. A front-end block with NF = 5 dB and OIP3 = 30 dBm to 40 dBm is realistic. Assuming NF is much less than that while maintaining OIP3 isn't realistic.
- Front-end RF filtering, RF switching, RF limiter, or other loss elements.

The model sets RF front-end OIP3 as a function of ADC_IP3 and the number of summed RF ports. Every element digital summation (RF = 1, left endpoint in Figure 7) requires the highest OIP3 RF front end, 8 dB higher than the ADC IP3 to result in a reasonable ~0.8 dB degradation to system cascaded IP3. As RF sum channels increase, the single-channel OIP3 requirement eases. This in turn eases the DC power required in the RF front end.

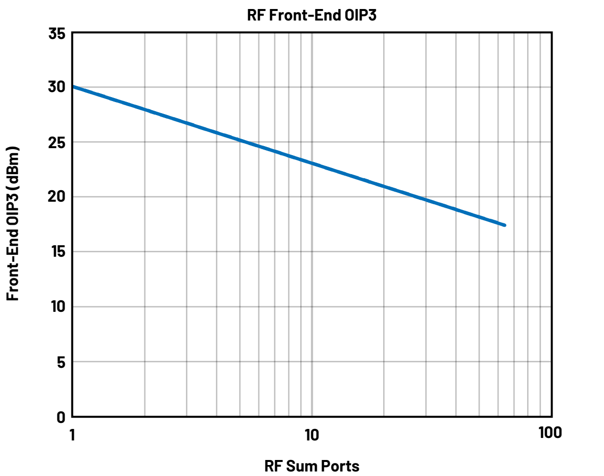

Figure 7. RF front-end single-channel OIP3 vs. the number of RF sum ports and ADC IP3 = 22 dBm.

The model sets RF front-end DC power as a function of RF front-end OIP3, which is a function of ADC IP3 and RF sum ports (described previously), shown in Figure 10. The RF front-end function is different in the RF = 1 vs. RF > 1 cases.

- RF = 1 is the every-element digital case where the RF front end does no beamforming. It is primarily a variable gain stage with signal filtering.
- RF > 1 requires time or phase delay and attenuation control to accomplish RF beamforming. The assumption here is a time delay unit and a variable attenuator sit between two gain stages. The extra gain stage vs. the RF = 1 case doubles the DC power and is required to overcome the TDU and DATT loss. This is what causes the DC power to jump up from RF = 1 to RF = 2 in Figure 10.

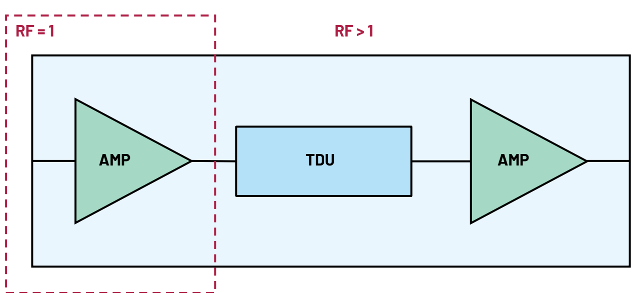

Figure 8. RF front end for RF = 1 and RF > 1 cases.

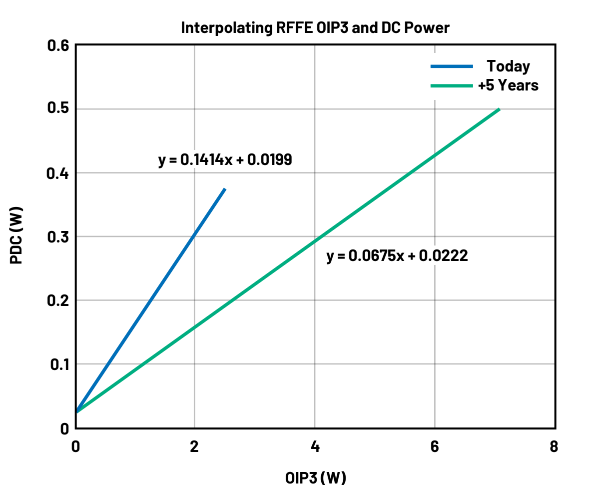

Figure 9. RFFE DC power model relation to OIP3.

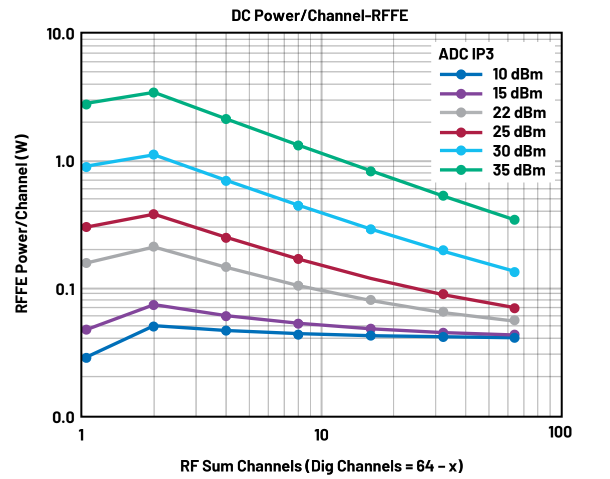

Figure 10. RF front-end DC power/channel.

System IP3 drives DC power, so do not overdesign or worse, throw away IP3 performance. The RF front end must have high enough IP3 so that the ADC is the linearity bottleneck, driving up the RF front-end DC power. Figure 10 shows the increasing DC power of the RF front end as the ADC IP3 increases.

Another notable relationship is the required front-end single-channel OIP3 decreases as the RF sum channels increase. Summing RF channels improves the signal-to-noise ratio (SNR) by coherent addition of the signal and noncoherent combination of Gaussian white noise. This better SNR is great, but now there is a stronger signal before the nonlinear ADC vs. the alternative of digitally summing channels after the ADC. The spur level resulting from multitone intermodulation is a function of ADC IP3 and the signal level into the ADC. For two tones at the same level:

$$P_{IN,IM3}\ \text{dBm} = 3 \times P_{IN}\ \text{dBm} - 2 \times IIP3\ \text{dBm} \tag{3}$$

In short, RF summing before the ADC improves SNR but degrades the two-tone spurs due to ADC nonlinearity compared to the same N channels summing digitally after the ADC.

You can now see why digital summing after the ADC, where the signal gain is computed in the digital realm after the ADC, has the SFDR advantage. The ADC is not asked to handle a larger signal; the SNR benefit is realized through the combination of digital data streams at the price of bit growth.

Modeling the RF Channel Summation

The model uses the method detailed in the article "Hybrid Beamforming Receiver Dynamic Range Theory to Practice" to cascade noise and linearity. There is a thorough explanation in that paper that isn't repeated here. Passive RF summation is used, so there is insertion loss but no additive noise and no impact on IP3. RF summation benefits SNR and has no impact on IP3. That is, intermodulation spurs add coherently across combined channels like the signal.

Table 2. RF Summation Attribute Equations

Block	Model Attribute	Equation
RF Sum	Gain, signal	F {N, 'on' ports}
	Gain, noise	-sqrt(N) dB

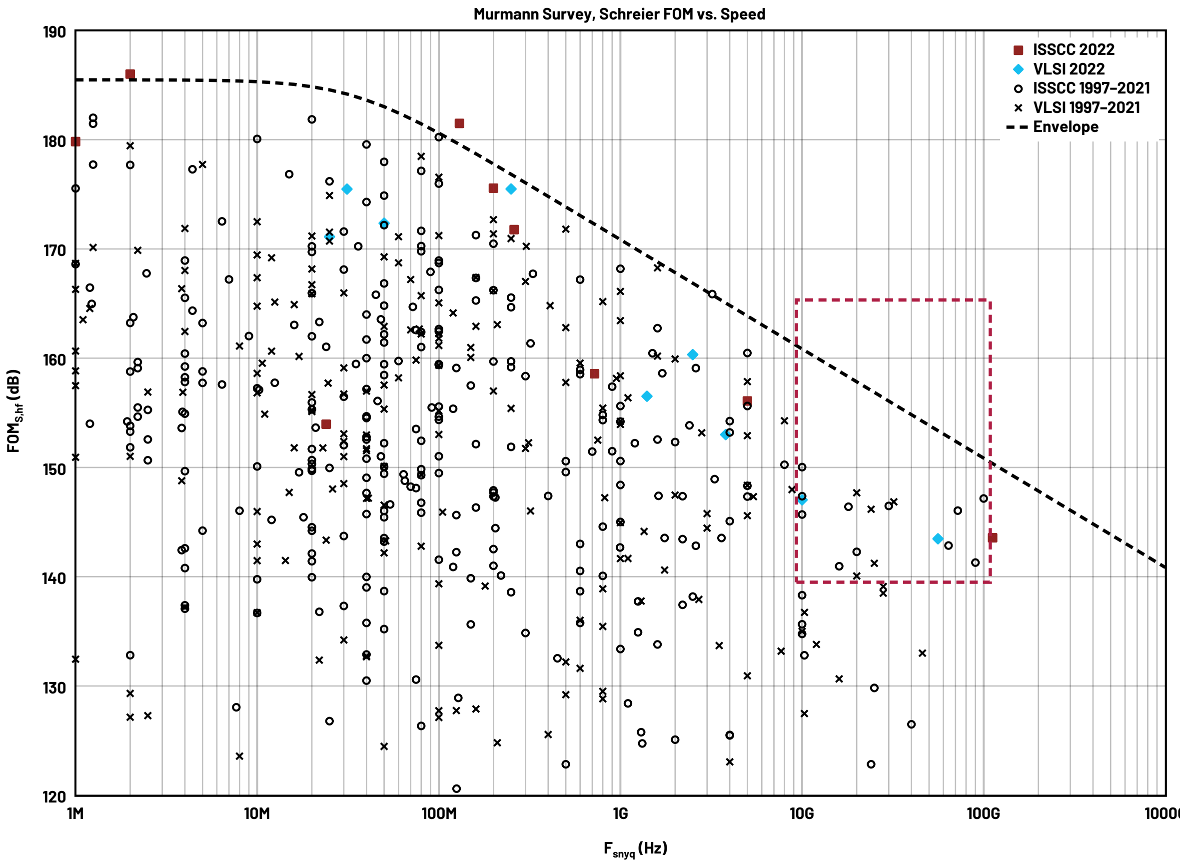

Figure 11. Schreier FOM from Murmann survey.

Sensitivity degrades as RF sum channels increase due to the RF-combiner insertion loss, modeled as:

$$\text{the Insertion Loss}_{RF\ Sum}\ \text{dB} = \sqrt{\text{Number of RF Sum Channels}}\ \text{dB} \quad (4)$$

Modeling the ADC

The ADC model uses behavioral equations derived from population data from Boris Murmann's ADC survey.[5] In an attempt to compare apples-to-apples merit, near-peer data points from similar-class ADCs are selected using two separate (but similar) FOMs. A black box ADC model allowing swept attributes is created from a fit to population data points.

The analysis uses the same two FOMs Murmann uses. The Walden FOM favors low resolution ADCs as it moves 2× per bit. A lower FOM is better.

$$FOM_W = \frac{Power}{2^{ENOB} \times f_{s,Nyq}} \left(\frac{fJ}{conv - step}\right) \quad (5)$$

$$ENOB = \frac{SNDR - 1.76}{6}$$

The Schreier FOM favors high resolution ADCs as it moves 4× per bit. Higher is better.

$$FOM_S = SNDR + 10Log\left[\frac{f_{s,Nyq}/2}{\text{Power}}\right] \text{dB} \quad (6)$$

In both cases, for a fixed FOM value, DC power moves proportional to the sample rate and exponential with dynamic range. This is an important rule of thumb to remember.

To meet the high dynamic range and direct sampling requirements in radar, EW, and MILCOM, new ADCs will pop up with FOMs in the boxes annotated on Murmann's ADC survey shown in Figure 11 and Figure 12. Since the Murmann survey frontier (sloping fit line on the right bound of data) depicts results published early in maturity, the frontier translates to commercially available devices several years out. Published data showing up in these boxes, however, isn't automatically a good fit for radar phased array. The most successful ADCs will balance adequate dynamic range and a high enough sample rate at the lowest possible DC power.

What are some example ADC cases on the state-of-the-art frontier that are highly interesting to phased array? Within the Figure 12 box of interest, start with (f_s, Nyq, FOM) points off the plot frontier.

Table 3. Frontier FOM Points

f_s, Nyq (GSPS)	FOMs dB	FOMw fj/conv-step
10	161	11
20	158	23
30	156	35
60	153	70

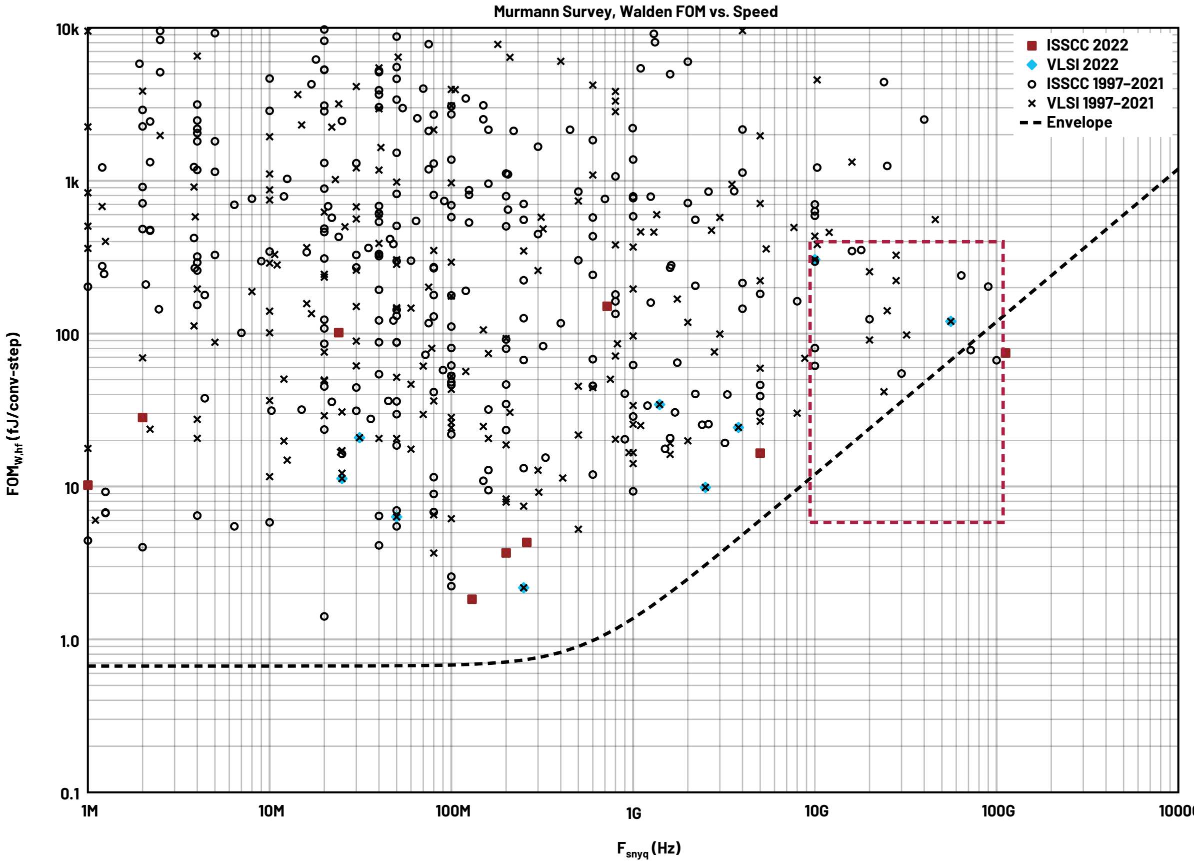

Figure 12. Walden FOM from Murmann survey.

Then, rearrange the FOM equations to plot the power vs. ENOB (dynamic range) relationship at that frontier point.

$$FOM_W \times f_{s,Nyq} \times 2^{ENOB} = Power\ \text{(Walden)}$$
$$\frac{f_{s,Nyq}/2}{10^{[FOMs - SNDR]/10}} = Power\ \text{(Schreier)} \qquad (7)$$

For a given sample rate on the frontier, Figure 13 and Figure 14 plot how a designer can trade off power and ENOB to arrive at the same FOM.

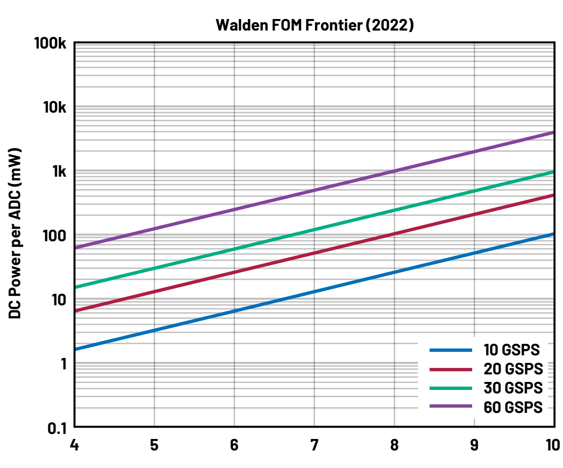

Figure 13. DC power vs. ENOB on the Walden frontier.

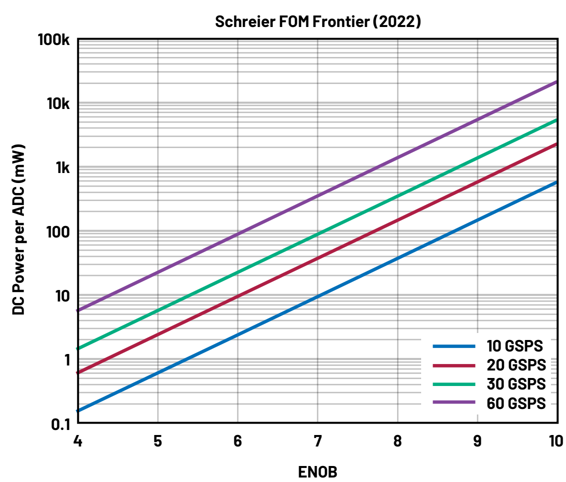

Figure 14. DC power vs. ENOB on the Schreier frontier.

The merit of a high speed ADC boils down to how good is the dynamic range (ENOB), over what instantaneous sampling bandwidth, and at what DC power. Being the best at these three things simultaneously is very difficult. DC power increases proportionally with sample rate, and exponentially with ENOB. For example, a hard max DC power limit of 100 mW per ADC requires a compromise in sample rate and ENOB that is optimal for the phased array radar. At 60 GSPS, a 6 ENOB ADC is possible at the current state of the art. Lowering the sample rate to 10 GSPS allows an increase in dynamic range to ENOB = 8.7. This is a big improvement in dynamic range. Which ADC is better? Neither. They are both on the FOM state-of-the-art frontier, and so are equally good. The better ADC depends on what the system needs to prioritize.

The highest sample rate converters get folks excited about the benefits to frequency planning, instantaneous coverage, software-defined digital tuning, RF simplification, etc. But before drawing any conclusions, ask "Sample rate…at what DC power and ENOB." The dynamic range, sample rate, and DC power performance triangle determines overall ADC merit. In a radar application, for example, 60 GSPS at ENOB = 6 could be completely useless, and 10 GSPS at ENOB = 8.7 is required. The higher sampling might be nice to have but the ENOB and power limit are often higher priority system critical limits. So, a compromise uses the 10 GSPS ADC to achieve the required ENOB at the required DC power.

Now, on to modeling the ADC performance in this cascade. The ADC model used in the cascade here is an RF black box with attributes NF, IP3, and DC power that are functions of swept attributes. As the system model sweeps ADC ENOB, the ADC attributes tune.

Table 4. ADC Attribute Equations

Block	Model Attribute	Equation
ADC	NF	F {f_S, ENOB, full scale}
	IP3	22 dBm
	DC power	F {ENOB}

The equation for ADC NF is a function of the effective number of bits (ENOB), or SNR.

$$NF_{ADC}\,\text{dB} = kTe_{ADC}\,\frac{\text{dBm}}{\text{Hz}} - (-174)\,\frac{\text{dBm}}{\text{Hz}}$$

$$\begin{aligned}&Noise\ Spectral\ Density{:}\ kTe_{ADC}\,\frac{\text{dBm}}{\text{Hz}} = Full\ Scale_{ADC}\,\text{dBm} - \\ &\quad SNR_{ADC}\,\text{dB} - 10Log\left(\frac{f_S}{2}\right)\text{Hz}\end{aligned} \tag{8}$$

$$SNR_{ADC}\,\text{dB} = 6 \times ENOB_{ADC} + 1.76\ \text{dB}$$

Noise spectral density, kTe (dBm/Hz), is equivalent to sensitivity (dBm) in a 1 Hz bandwidth. To generalize, assume a 1 Hz bandwidth throughout, which can be adjusted to a specific bandwidth by adding 10LogBW.

Don't take ADC two-tone IP3 performance lightly, it should be treated with the same critical eye as noise.

The model for **ADC DC power** uses published data summarized in the Murmann survey. The Murmann ADC survey data (rev20220719) is reduced to 20 members, filtering on the following criteria:

- Analog Devices and industry peers
- CMOS <32 nm
- f_S> = 4 GSPS

Using the Murmann data, the model derives best fit, low bound, and high bound equations for DC power as a function of ENOB, shown in Figure 15.

$$\frac{Power_{DC}}{freq_{sample}} = ae^{k(ENOB)}\ pJ \tag{9}$$

Table 5. ADC DC Power Upper, Best, and Lower Fit Values

	a	k
Best fit	0.045	0.93
Low bound	0.025	0.93
High bound	0.08	0.93

Using the selected set of Murmann data, the model derives a relation for ENOB as a function of bits using linear fit, shown in Figure 16.

$$ENOB_{ADC} = 0.6 \times bits_{ADC} + 1 \tag{10}$$

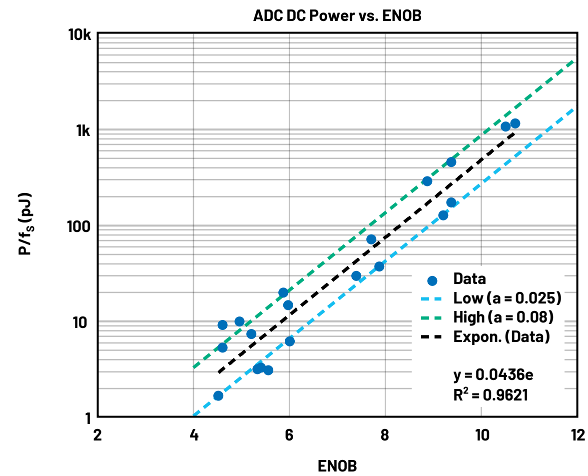

Figure 15. Fit curve for ADC DC Power vs. ENOB, using Murmann survey data.[5]

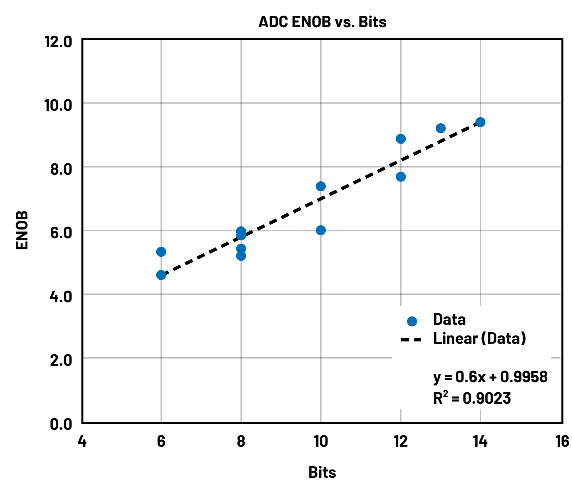

Figure 16. Fit curve for ADC ENOB vs. bits, using filtered Murmann survey data.[5]

These equations comprise a DC power and RF attribute model for the ADC as a function of ENOB as shown in Figure 17. Figure 18 shows DC power/channel vs. ENOB and RF: digital beamforming.

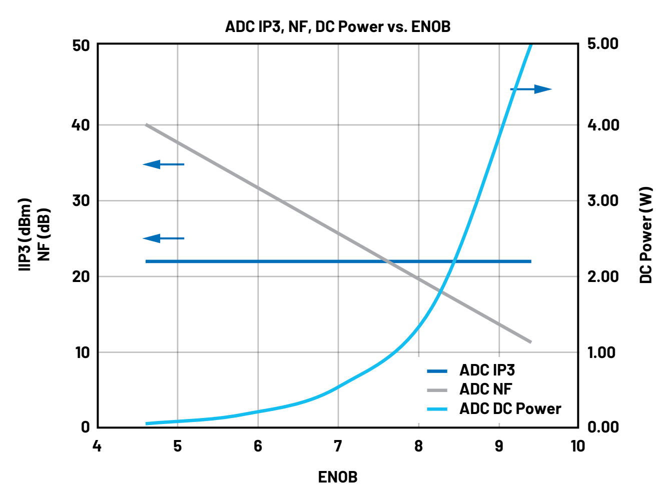

Figure 17. Generalized model for digital converter NF and DC power vs. ENOB.

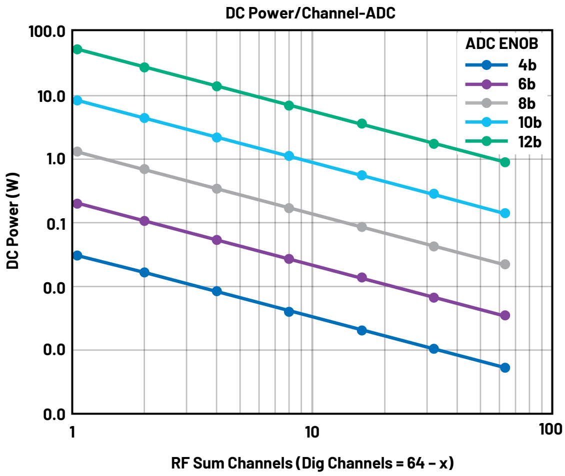

Figure 18. ADC DC power per element vs. number of RF sum channels and ADC ENOB.

Modeling the Digital Payload Interface and Summation

The high speed data payload and sum computation DC power are estimated from the transport energy per bit.[6]

Table 6. Table 6. Digital Payload DC Power Model

Block	Model Attribute	Equation
Digital Sum	DC power	F {bits, IBW, lanes, pJ/b}

The digital payload transport power associated with the ADC-to-digital sum node scales up as the number of digital sum channels and IBW increases. The following calculates the DC power burn for the transport of the physical links, that is, high speed interface.

$$\begin{aligned} &Power_{Digital\ Sum\ Interface}\ W = [Energy_{Serializer} + \\ &\quad Energy_{Deserializer}\ \ J/bit]\ \times Payload\ bits\ \times \\ &\quad Dig.Sum\ Channels \\ &Payload\ bits = Encode\ Rate\ \mathrm{Gbps}\ \times \\ &\quad Overhead_{JESD204C} \times bits_{ADC} \\ &Encode\ Rate\ \mathrm{Gbps} = IBW\ GHz \times 2\ \mathrm{bits} \times 1.2 \\ &Overhead_{JESD204C} = 66/64 \end{aligned} \tag{11}$$

A JESD link is assumed with

$$\begin{aligned} &Energy_{Serializer} = 3\,\frac{pJ}{bit} \\ &Energy_{Deserializer} = 4\,\frac{pJ}{bit} \\ &\mathrm{IBW} = 1\ \mathrm{GHz} \end{aligned} \tag{12}$$

The computational power burn for the complex multiply is assumed equal to the interface power. This is a crude approximation dependent on other factors like beam-bandwidth, but close enough. As a reasonable estimate, the interface power is doubled. Figure 19 shows the overall interface plus digital sum DC power/channel vs. RF: digital sum ratio.

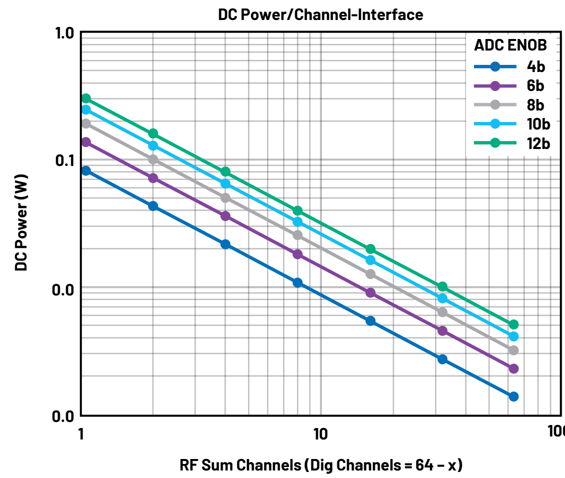

Figure 19. Digital payload interface and complex multiply DC power per element vs. number of RF sum channels and ADC ENOB.

Conclusion

Part 1 constructs an RF cascade model for a phased array receiver for the purpose of analyzing the impact of RF and digital channel summation on dynamic range and DC power. The model is built in excel and consists of:

- RF front end
- Variable RF channel summation
- Analog-to-digital converter
- High speed digital Interface and basic compute.
- Variable digital channel summation

Next, Part 2 presents and analyzes results from the model in Part 1, boils the results down to system FOMs, and draws conclusions. The optimal ADC bit resolution is recommended.

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About the Author

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