TEXAS INSTRUMENTS

RF Sampling for Multi-band Radios

ABSTRACT

Recent advances in high-speed data converters now enable digitization and generation of signals directly at frequencies up to several gigahertz, replacing traditional radio-frequency (RF) components like mixers, local oscillators (LOs), and amplifiers. In addition, the inherent wide-bandwidth capabilities of RF-sampling data converters with multiple gigasample-per-second (GSPS) speeds are enabling radios that combine multiple bands for cellular infrastructure applications. The resulting smaller and lower-power systems can ultimately reduce the number of remote radio head (RRH) boxes at each cell site. This application report compares a RF sampling multi-band radio architecture using the AFE7686 with other traditional radio architectures, and calculates expected performance relative to the requirements from the 3GPP standards. The approach for the performance estimation is to generate a simple mathematical expression for each degradation source and uses an Excel spreadsheet to calculate the total degradation in different conditions.

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1   Introduction

Recent advances in high-speed data converters now enable digitization and generation of signals directly at frequencies up to several gigahertz, replacing traditional radio-frequency (RF) components like mixers, local oscillators (LOs), and amplifiers. In addition, the inherent wide-bandwidth capabilities of RF-sampling data converters with multiple gigasample-per-second (GSPS) speeds are enabling radios that combine multiple bands for cellular infrastructure applications. The resulting smaller and lower-power systems can ultimately reduce the number of remote radio head (RRH) boxes at each cell site.

This application report compares a RF sampling multi-band radio architecture using the AFE7686 with other traditional radio architectures, and calculates expected performance relative to the requirements from the 3GPP standards. The approach for the performance estimation is to generate a simple mathematical expression for each degradation source and use an Excel spreadsheet to calculate the total degradation in different conditions. While this may not be as accurate as a simulation model built in a tool such as Keysight ADS, it provides understanding for each degradation source and which sources are the largest contributors in each scenario.

2   Radio Architectures

The function of the radio transmitter is to convert a complex digital baseband signal to a modulated radio frequency carrier and output the antenna. The radio receiver receives a modulated radio frequency carrier and converts it to a complex digital baseband signal. This section provides a brief overview of different radio architectures used in cellular infrastructure radios. Each radio architecture has the same basic functions: quadrature modulation, upconversion, and filtering. The difference is the location of the split between digital and analog functions, which determines whether the function is done in the analog or digital domain.

2.1   0IF Radio Architecture

Figure 1 and Figure 2 show a zero-intermediate frequency (0IF) architecture for a transmitter and a receiver. For the transmitter, a dual channel digital-to-analog converter (DAC) generates a complex analog signal. The baseband signal is filtered to removed images of the signal that are generated at higher frequencies in the DAC, and are then mixed with a local oscillator (LO) in a quadrature modulator to the RF carrier frequency. In the analog quadrature modulator, the LO synthesizer output is split into two carriers 90 degrees out of phase, and mixed with the real and imaginary components of the complex analog baseband signal. The mixed signals are summed, filtered to remove unwanted spurious signals, and then amplified before being output at the antenna.
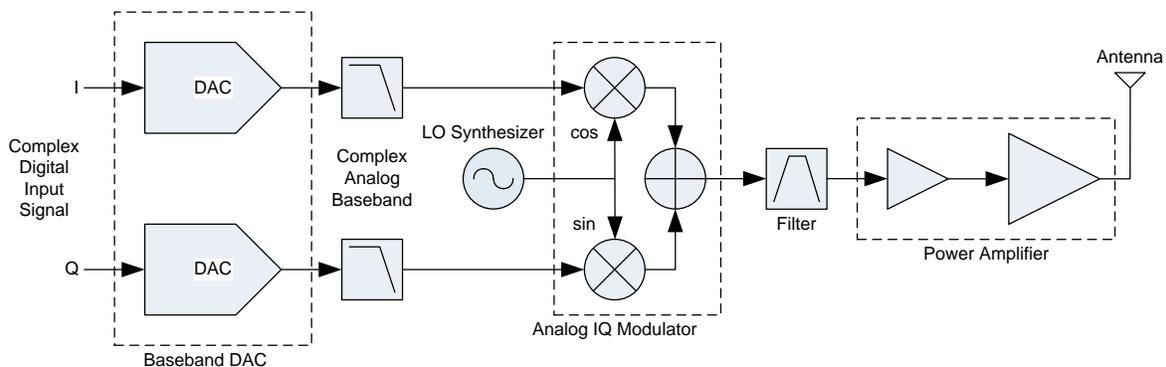
Figure 1. 0IF Transmitter Diagram
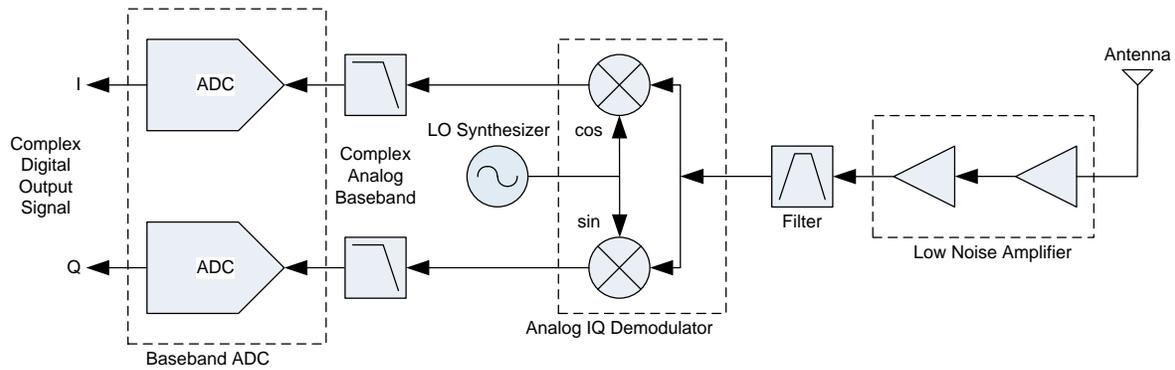
Figure 2. 0IF Receiver Diagram

The function of the receiver is opposite of the transmitter. The RF signal at the antenna is first amplified by a low noise amplifier before being filtered to suppress large unwanted signals at frequencies outside of the band of interest. The RF signal is down converted to an analog baseband signal by mixing with an LO that is split into two carriers 90 degrees out of phase. The analog baseband signal can be filtered to further suppress out-of-band energy and then sampled with a dual channel analog-to-digital converter (ADC).

The advantage of the 0IF architecture is the ease in filtering at baseband (only low pass filtering is required) and at RF. There are no mixing images as discussed in Section 2.2. The disadvantage is that the analog IQ mismatch creates sideband images and LO feedthrough in-band, which typically needs to be corrected digitally to meet the spectral mask and blocker sensitivity requirements. Figure 4 shows the frequency spectrum at baseband for the receiver with impairments including the blocker image due to IQ mismatch, IMD2 distortion centered at DC, and DC offset. Figure 3 shows the frequency spectrum at RF for the transmitter with impairments including the images due to IQ mismatch and LO feedthrough.
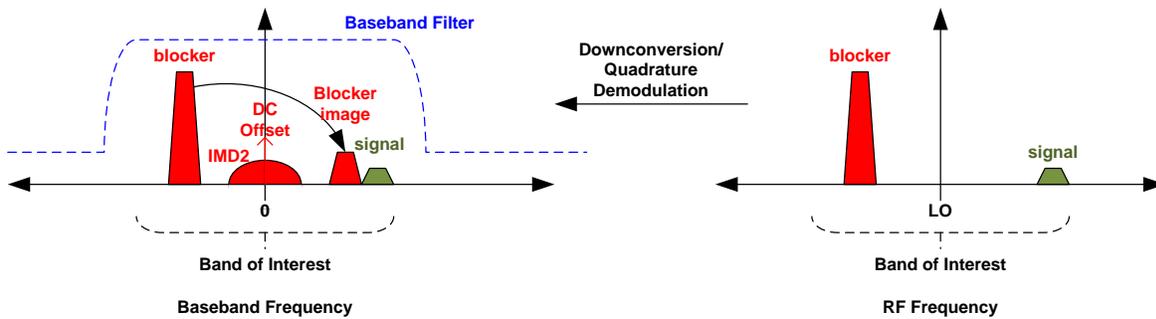
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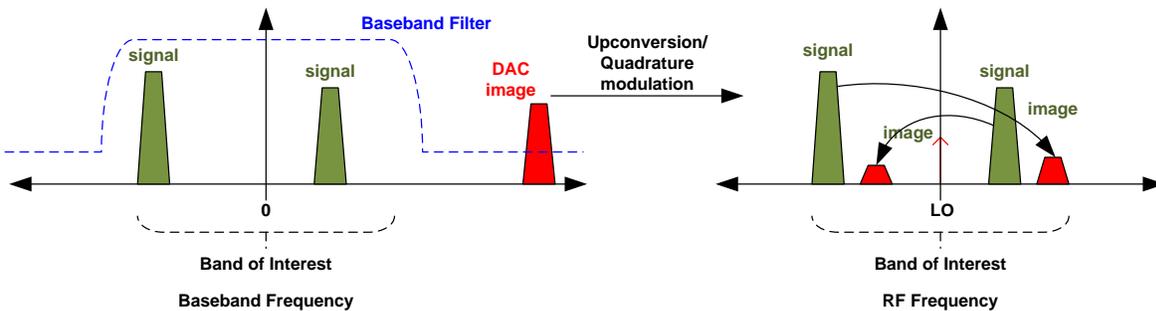
Figure 4. Frequency Spectrum with Impairments for 0IF Receiver

2.2   IF Radio Architecture

Figure 5 and Figure 6 show traditional signal chains for an IF radio architecture. In an IF architecture, quadrature modulation and demodulation is done digitally using an NCO and digital mixers. For the transmitter, the digital signal is output as an analog signal by a DAC at a frequency typically ~10% of the final RF frequency. The signal is filtered and upconverted using a mixer with LO to the final RF frequency, where it is filtered and amplified before being output at the antenna.

On the receiver side, the RF signal is received by the antenna, and then filtered before being mixed with an LO and down converted to an IF frequency ~10% of the RF frequency. At IF, the signal is filtered and digitized with an ADC, and then quadrature demodulated digitally with an NCO and digital mixers.
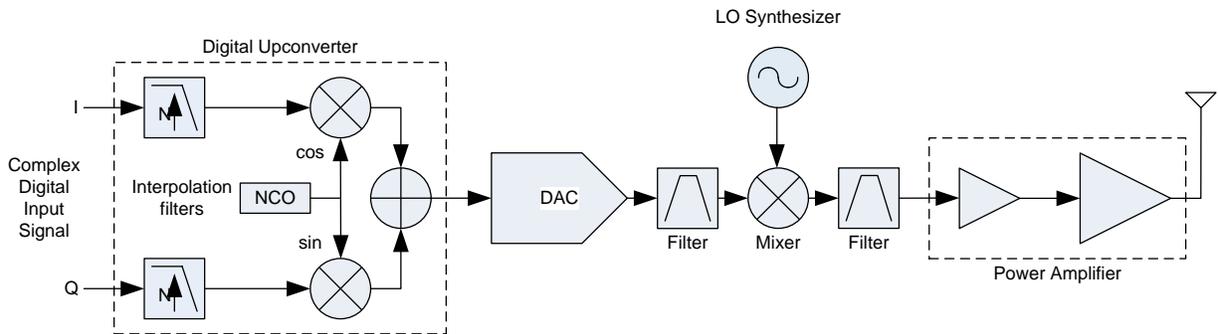
Figure 5. IF Transmitter
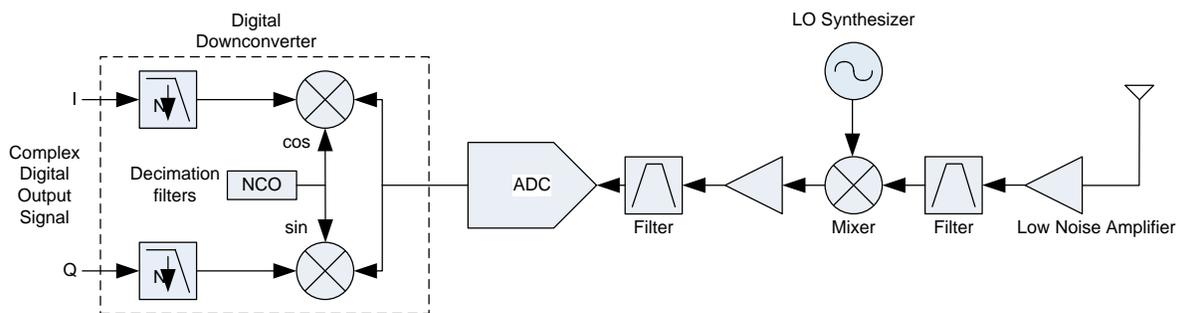
Figure 6. IF Receiver

The advantage of the IF architecture over 0IF architecture is the perfect IQ balance provided by the digital quadrature modulation and demodulation, and that second order intermodulation products do not fall in-band. The disadvantages are the higher number of components (and corresponding size and power) and the more difficult filtering required at IF and RF. Figure 7 and Figure 8 show the frequency spectrum and impairments for an IF transmitter and receiver. For the transmitter at RF, there is an image offset by 2 × IF that has the same power as the wanted signal, and an LO feedthrough at the LO frequency that is typically suppressed by ~30 dB from the wanted signal power.

On the receiver side, any energy on the other side of the LO at an offset of IF directly down convert to the same IF frequency, and needs to be suppressed by the difference between the power of the interfering energy and the wanted signal (which can be 100 dB or more). The significant filtering required makes it difficult to have a bandwidth greater than a third of the IF frequency.
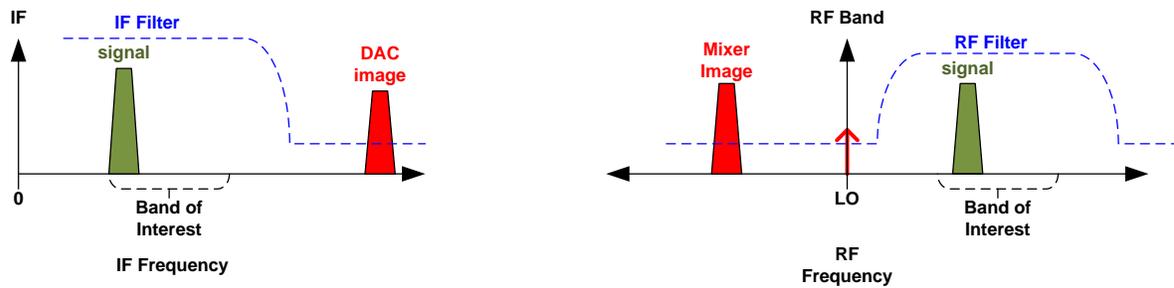
Figure 7. Frequency Spectrum with Impairments for an IF Transmitter
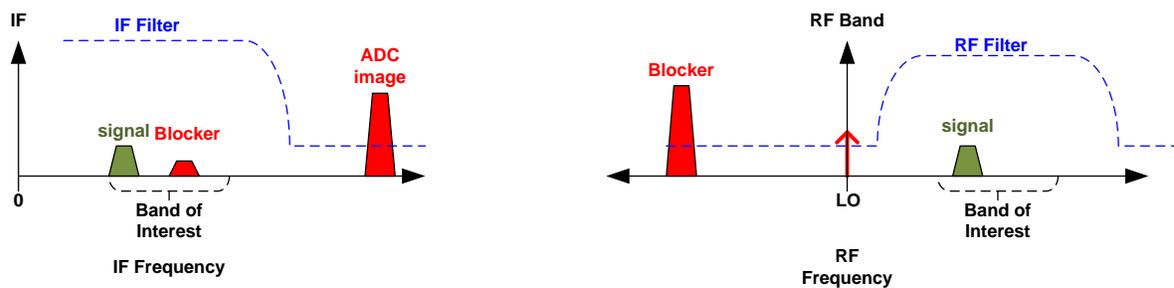
Figure 8. Frequency Spectrum with Impairments for an IF Receiver

2.3 Complex IF Architecture

The complex IF architecture is a variation of the IF architecture that uses analog quadrature modulators and demodulators rather than a single mixer to an up and down converter from IF to RF and vice versa (Figure 9 and Figure 10). The analog quadrature modulation rejects the mixing images on the other side of the LO, to the extent of the balance of the analog IQ matching (typically > 40 dB). As shown in the frequency spectrum for the complex IF transmitter in Figure 11, the image rejection reduces the filtering required at RF, enabling wider bandwidths up to ~ ½ the IF frequency.

On the receiver side, the RF signal is received by the antenna, and then filtered before being mixed with an LO and down converted with an analog quadrature demodulator to a complex IF frequency ~10% of the RF frequency. The complex IF signal is filtered and digitized with a dual ADC, and then shifted to 0 Hz digitally by fully complex mixing with a complex tone at -IF. As shown in the frequency spectrum for the complex IF receiver in Figure 12, the complex IF architecture provides rejection of the sideband image at RF, reducing the RF filtering requirement.
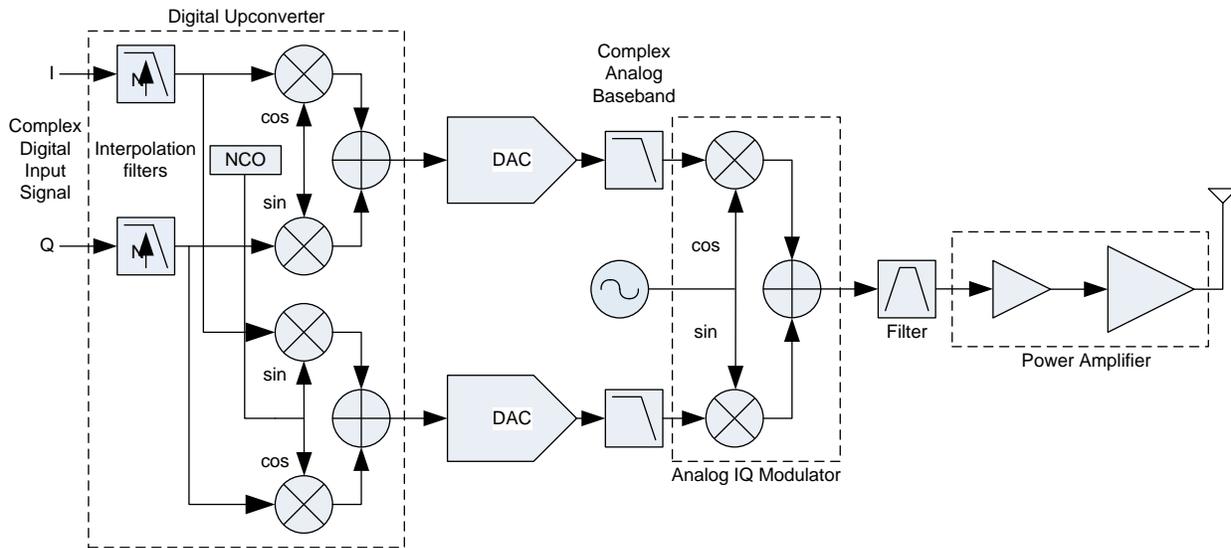
Figure 9. Complex IF Transmitter Diagram
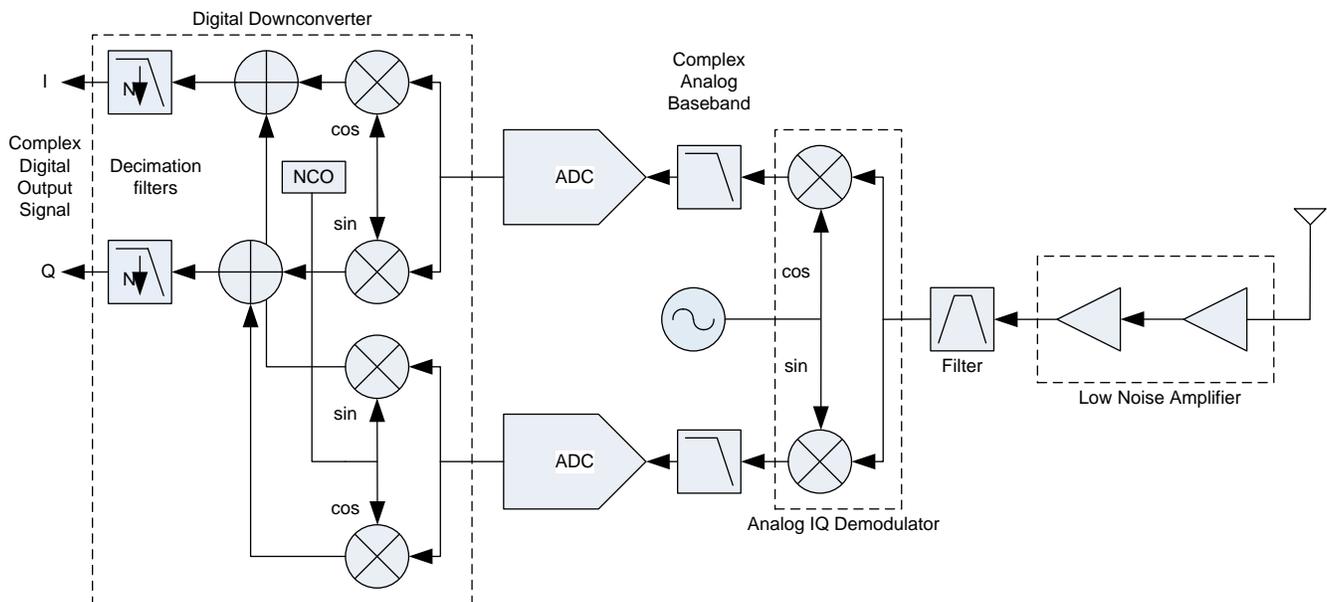
Figure 10. Complex IF Receiver Diagram
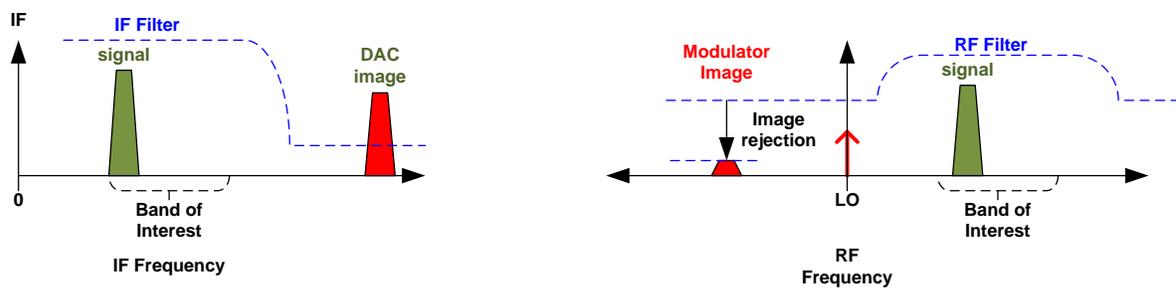
Figure 11. Frequency Spectrum for the Complex IF Transmitter
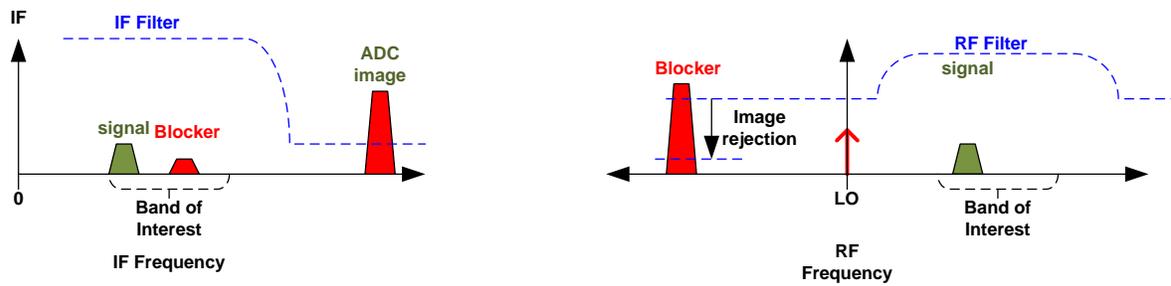
Figure 12. Frequency Spectrum for the Complex IF Receiver

2.4 RF Sampling Architecture

With the RF sampling radio architecture, all quadrature modulation and up and down conversion is done digitally, and the signal is converted to and from analog directly at the RF frequency.

Figure 13 shows a diagram of the RF sampling transmitter, and Figure 14 shows the transmitter frequency spectrum with impairments. The DAC sample rate is typically greater than twice the RF output frequency, so the output signal in the first Nyquist zone is used. A band or low pass RF filter is then needed to remove higher Nyquist zone images, which are offset by at least a GHz. Since the DAC sample rate is very high to support the RF output frequency, an RF sampling transmitter is inherently very wide bandwidth. A 9 GSPS DAC, common for this application, can generate signals from 0 GHz to 4 GHz, covering all 3G and 4G cellular bands.
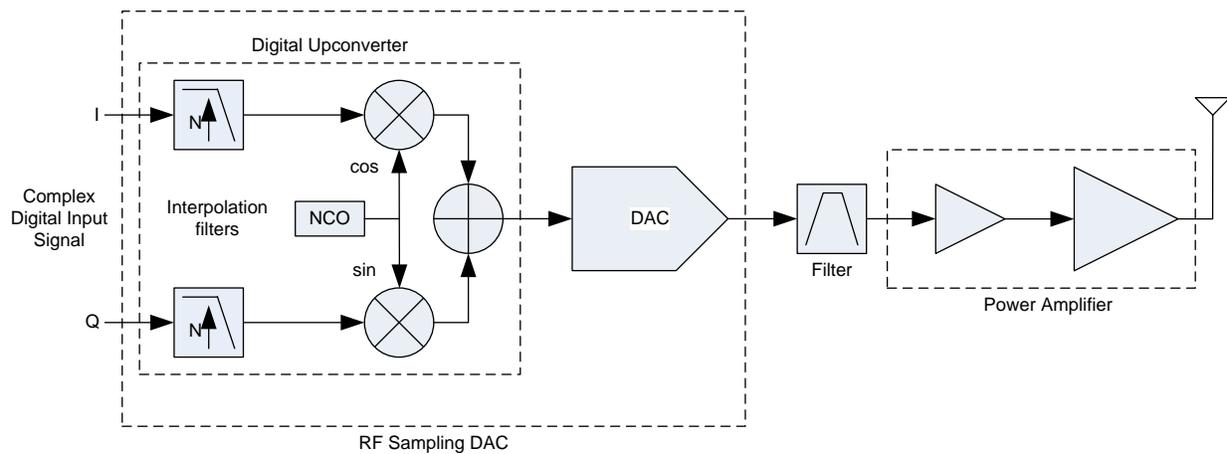
Figure 13. RF Sampling Transmitter Diagram
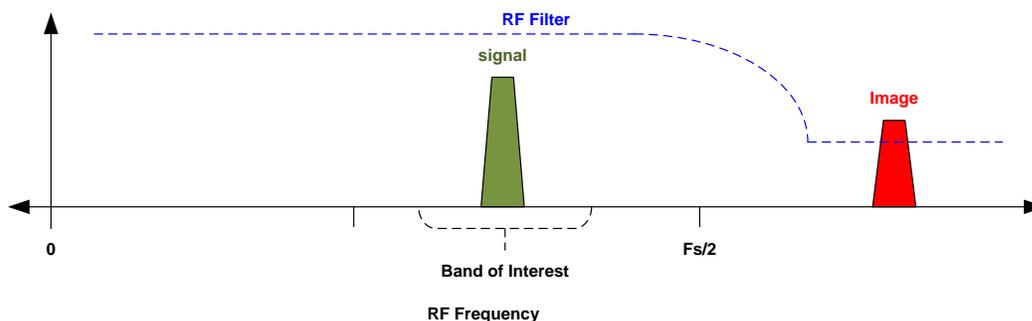
Figure 14. RF Sampling Transmitter Frequency Spectrum

Figure 15 shows the RF sampling receiver, and Figure 16 shows the receiver frequency spectrum with impairments. The RF sampling ADC is usually used in undersampling, where the input signal is above the first Nyquist zone (> f_{ADC} / 2). The sample rate only needs to be high enough to sample the signal bandwidth and provide enough frequency offset to the Nyquist images to enable filtering. For most 3GPP bands, a 2.4576 or 2.94912 GSPS sample is sufficient.
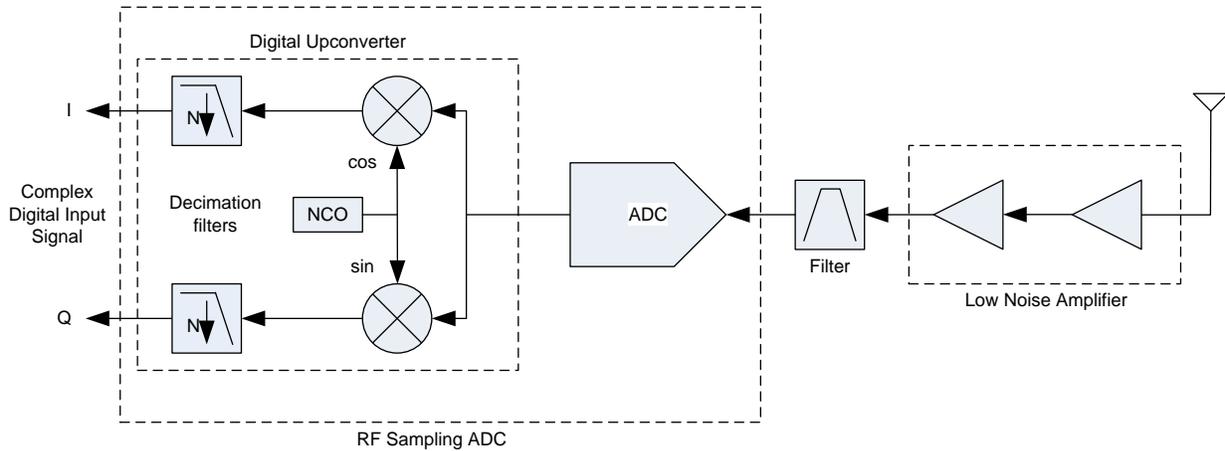
Figure 15. RF Sampling Receiver Diagram
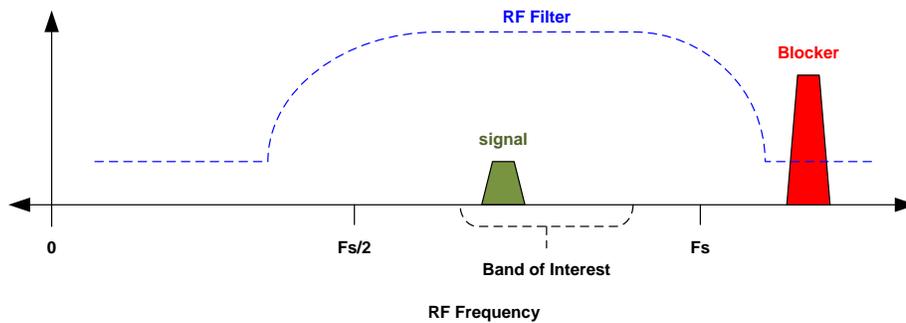
Figure 16. RF Sampling Receiver Frequency Spectrum

The advantages for RF sampling are fewer components, smaller PCB area, and wide bandwidth. As quadrature modulation and conversion are done digitally, RF sampling enables multi-band radios where multiple bands are up-converted and down-converted to and from RF through the same DAC or ADC. This is not be practical with the 0IF, IF, or complex IF architectures, unless the bands are very close together and can be treated as a single band in the radio.

2.5   Integration

The IF filters used for IF radios are typically surface acoustic wave (SAW) technology, which is not compatible with monolithic integration with the silicon processes typically used for the data converters. Integration for IF therefore relies on multi-chip modules (MCM), which are typically larger and more expensive than monolithic integration, that is integration on a single die. Compared to a radio with discretely packed components, MCM integration also tends to be higher cost. Therefore, MCM integration for IF radio architectures is less common than integration of architectures (MCM or monolithic).

The 0-IF radio architecture lends itself better to integration due to the ability to relax the analog baseband filter by increasing the data converter sample rate. Sample rates 5x or higher the baseband bandwidth are common, which allow a filter with 2 to 3 poles to provide significant rejection of the data converters Nyquist image. 0-IF radios are therefore compatible with monolithic integration, and offer attractive advantages in power, size, and cost for signal bandwidths up to 200 MHz.

Integrated 0-IF radios can also be used for a complex IF architecture by shifting the signal from 0 Hz so that the signal is on one side of the baseband spectrum. This limits the bandwidth of the signal to less than ½ the baseband bandwidth, and is typically not used for the transmitter where wider bandwidths are required. For the receiver, it relaxes the receiver IQ matching requirement for in band blockers, as their sideband image falls on the opposite side of the baseband spectrum.

RF sampling also lends itself well to integration. Also all mixing is done digitally, there are no analog filters to integrate. The data converters only require a single clock, or as is common, separate DAC and ADC clocks where the ADC clock is a integer divided version of the DAC clock. This reduces the potential of mixing between different LO frequencies for frequency division duplex bands where the transmitter and receiver operate at different frequencies.

3   Receiver Specifications and Analysis

3.1   *System Specifications*

This section discusses receiver system specifications for cellular radios, specifically 3GPP 2nd Generation GSM and 4th Generation LTE (3rd Generation WCDMA requirements are similar to the requirements for LTE).

3.1.1   How the Tests Are Defined

The 3GPP standards for basestation radios are defined to ensure the cellular network is robust in real world scenarios. For receivers, the standard tests specify the minimum received signal power from the user equipment (UE) that can meet the required bit error rate (BER) at the BTS reciever.

The most challenging scenarios from the radio point of view are when the wanted UE signal is far away from the BTS and the received power is at a minimum while a large interfering or blocking signal, is also received by the antenna.

The output power of the UE is controlled by the service provider network controller to only output as much power as needed for a reliable connection – transmitting higher power can needlessly drain the UE battery. When the UE is near the BTS antenna for its service provider, only a low output power is used, and when the UE is far from the BTS antenna for its service provider (or indoors which significantly attenuates the signal), the UE transmits at its maximum output power.

Figure 17 shows how such a blocking condition can arise. Service provider A with BTS A is communicating with UE A, which is at the limit of the distance from BTS A at the cell edge. Near UE A is a second BTS for service provider B, which is communicating with UE B, which is also far away from BTS B. UE A and UE B are transmitting at maximum power to reach their respective BTS, but since UE A is much closer to BTS B than UE B, the received power from UE A at BTS B is much higher than the actual wanted signal from MS B. MS A is therefore an interferer or blocker for BTS B, and the radio for BTS B must have enough dynamic range (how small a signal can be received compared to the largest signal) to allow reception of the signal from MS B.
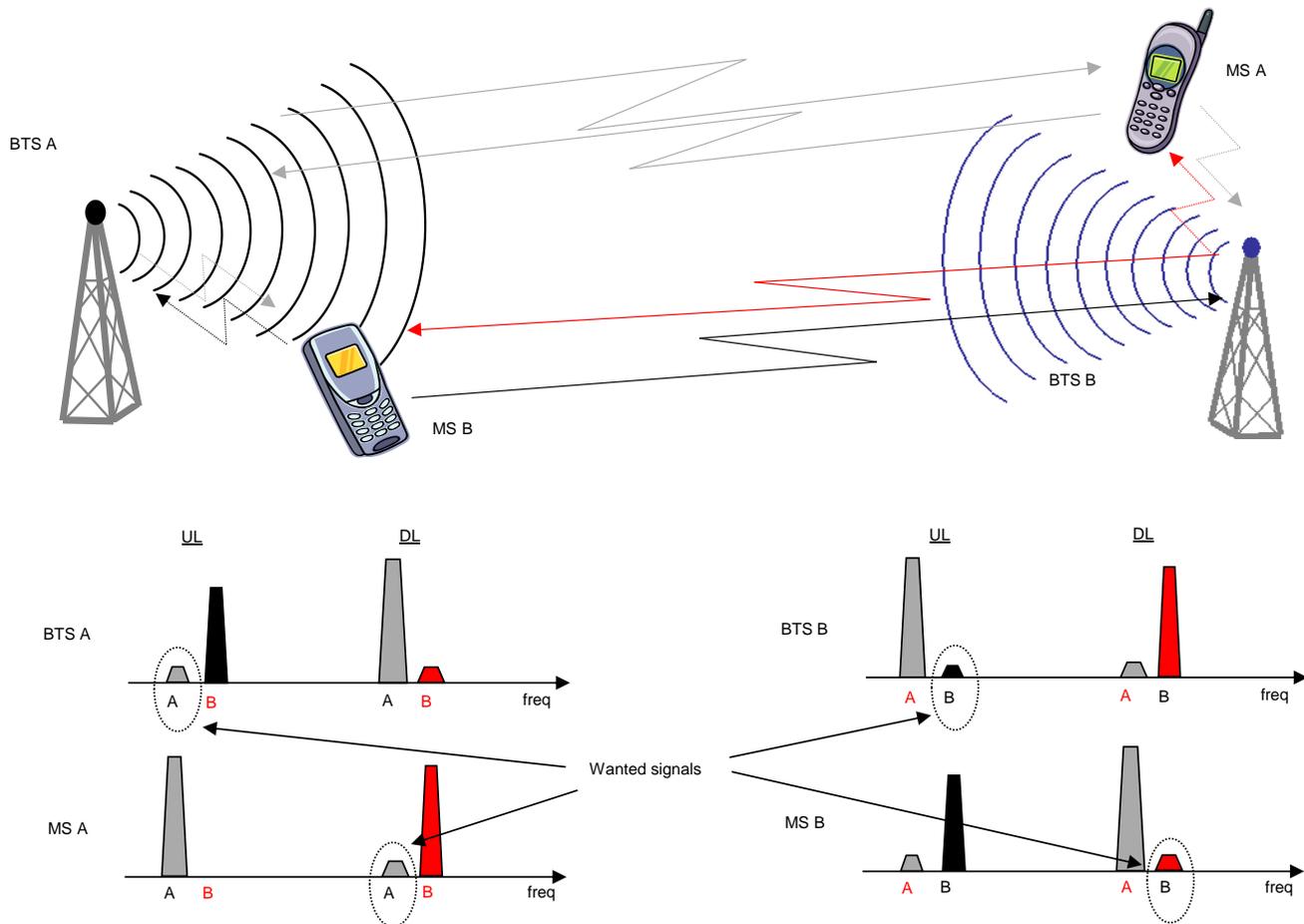
Figure 17. Blocker Scenario for BTS Receivers

3.1.2   Requirements for Single-band vs. Multi-band Radios

The 3GPP specifications are written for single band radios, and how to extend the requirements to multi-band radios is not specified. For example, when a blocker is in-band, typically the specifications allow for a relaxation in the required wanted signal sensitivity (although in practice, LTE Macro basestation radios are significantly over designed and often do not need the relaxation). For a radio covering two bands, what is the requirement for band 1 when there is a blocker in band 2? Is there a relaxation in sensitivity for band 1 as there would be in band 2? Should the requirement in band 1 be the same sensitivity as without a blocker?

A 3GPP technical report (3GPP) makes a recommendation that the relaxation for in-band blockers in the other band be reduced from 6 dB to 1.4 dB. When doing the multiband analysis, choose the conservative approach of requiring the non-blocker sensitivity when the blocker is in the other band, that is, no relaxation. This is equivalent to having two separate single band radios co-located. The exception is the out-of-band blocker, which is out-of-band for both bands and therefore the out-of-band requirements hold.

4   Typical Radio components for RF Sampling Receiver Front End

In order to estimate the performance requirements for the RF sampling ADC, you need to make some assumptions on the surrounding components, especially those in the signal chain between the antenna and ADC input.
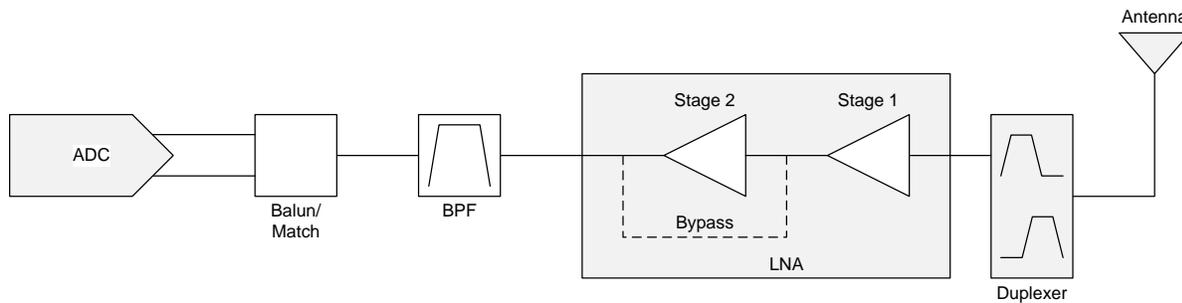
Figure 18. FDD RF Sampling Receiver Front End

Figure 18 shows a front-end lineup for the RX between the antenna and ADC input. For FDD systems, a duplexer following the antenna is required to suppress the large TX signal to avoid saturating the RX components. The duplexer can also suppress out-of-band blockers, at least to the level of the in-band blockers. The passband of the duplexer must have very low loss (< 1 dB) not to significantly impact the RX sensitivity. For this analysis, assume a 1 dB loss prior to the LNA for both FDD and TDD.

After the duplex filter, the received signal is amplified by a low noise amplifier (LNA). For the following analysis, the LNA device QPL9065 from Qorvo is considered. Over temperature, the gain, noise figure (NF), and third order intercept point (IP3) vary, so the parameters for the minimum gain (high temperature) and maximum gain (low temperature) need to be considered.

After the LNA, a band pass filter (BPF) for anti-alias image rejection is required, followed by a balun for single to differential conversion and components to improve the matching to the ADC input impedance. The combined BPF, balun and matching components is assumed to have 2 dB of loss within the receive band. The stopband rejection requirements of the BPF is discussed later.

Table 1 lists the cascaded parameters from the antenna to ADC input.

Table 1. Front-end Characteristics with QPL9065

TEMPERATURE (°C)	SECOND STAGE ENABLED		SECOND STAGE BYPASSED	
	85	-40	85	-40
Gain (dB)	28	32	10	17
NF (dB)	2.05	1.65	2.15	1.67
IIP$_3$ (dBm)	-0.7	-3.2	15	9

4.1 Duplex Filter for FDD

For FDD systems where TX and RX are simultaneously active and connected to the same antenna, the duplex filter provides three functions:

- The duplexer suppresses the high power TX signal to prevent saturation of the receiver.
- The duplexer suppresses any energy generated by noise or distortion in the RX band so it does not degrade the receiver performance significantly.
- The duplexer suppresses out-of-band signals from the antenna.
- Table 2 lists the TX path and duplexer characteristics for the receiver analysis.

Table 2. Duplexer Characteristics Assumed for FDD Analysis

TX PATH CHARACTERISTICS		
Pout (@ TX per 25RB)	40	dBm
TX Noise in RX band (@ PAout)	-86.5	dBm/Hz
DUPLEXER REJECTION		

Table 2. Duplexer Characteristics Assumed for FDD Analysis (continued)

TX PATH CHARACTERISTICS		
TX-RX isolation (TX freq)	85	dB
TX-RX isolation (RX freq)	95	dB
Rej @ F<=FRX_band_min – 20 MHz	15	dB
Rej @ F>=FRX_band_max + 20 MHz	15	dB

5 ADC Characteristics

The ADC parameters (listed in Table 3) used in the analysis are from the AFE7686, a 4 channel RF sampling transceiver (TRX) with 4 RF sampling DACs, and 4 RF sampling ADCs from the .

The parameters are typical values at ~2.6 GHz for analysis of system performance for the LTE bands 7 or 41.

Table 3. Key ADC Parameters from the AFE7686

PARAMETER	VALUE	UNIT
Sampling Clock	2949.12	MHz
FS (@ Atten = 4)	-3.6	dBm
Maximum Input Power	15.4	dBm
NF @ Attenuation >= 4 dB @ 85C	15.1 + Attenuation	dB
NF @ Attenuation >= 4 dB @ -40C	13.6 + Attenuation	dB
IP3 @ Attenuation = 0 dB	17	dBm
SFDR excluding HD2/3 and IL spur	-90	dBFs
Additive Phase noise @ 1.91 GHz and 800 kHz offset	140.7	dBc/Hz

Referring to the spectral plots at ~2.6 GHz with different DSA settings in the AFE7686 datasheet, notice the non-HD2/3 spurs are largely independent of DSA setting.

5.1 ADC Frequency Planning

For the ADC, a sample rate of 2949.12 MSPS is used for the analysis. This sample rate has the beneficial property that in-band blockers, the second harmonic (HD2), third harmonic (HD3), and interleaving spurs (IL FSdiv2/4) fall outside of band 41 and do not interfere with the wanted signal. This is shown in Figure 19. As a consequence, tests with in-band blockers can exclude HD2, HD3, and IL FSdiv2/4 from the analysis. HD2 and HD3 from out-of-band blocker can fall in-band, which ultimately impacts the BPF filter requirements to reduce the amplitude of the out-of-band blockers.
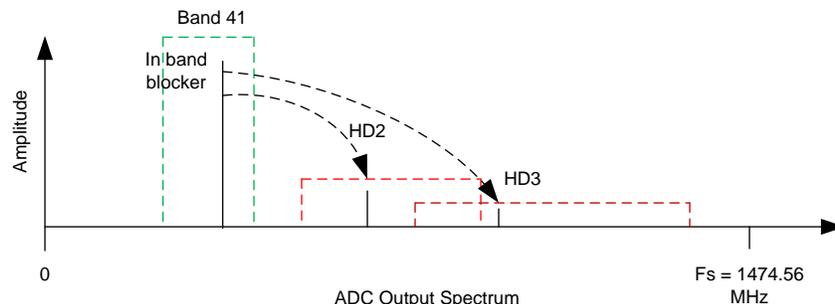
Figure 19. HD2 and HD3 for an In-band Blocker for Band 2 and f_{ADC} = 2949.12 MSPS

Table 4 lists the spurs from in-band signals that fall back in-band at different sample rates for each 3GPP band. Since duplex filter does not immediately suppress out-of-band blockers, include a guard band of 20 MHz on either side of the receive band. A sample rate of 2949.12 MSPS avoids the in-band blocker's HD2 and HD3 from falling back in-band for most bands, except bands 1, 36, and 43. For bands 1 and 36, a sample rate of 2457.6 MSPS prevents HD2 and HD3 from in-band blockers from falling back in-band. For band 43, either HD2 or HD3 from in-band blockers fall back in-band. However, the AFE7686 HD3 of > 74 dBc is sufficient to prevent HD3 from limiting the receiver performance.

Table 4. HD2, HD3, and IL Spurs falling In-band for In-band Blockers

BAND #	RX BAND (MHz)		SAMPLE RATE (MSPS)		
	LOW	HIGH	2211.84	2457.6	2949.12
1	1920	1980	IL FSdiv4		HD2
2	1850	1910		HD3	
3	1710	1785			
4	1710	1755			
5	824	849	IL FSdiv4	HD2	
7	2500	2570		HD2	
8	880	915			
9	1749.9	1784.9			
10	1710	1770			
11	1427.9	1447.9			
12	699	716			
13	777	787			
14	788	798			
17	704	716			
18	815	830	IL FSdiv4	HD2	
19	830	845	IL FSdiv4		
20	832	862	IL FSdiv4		
21	1447.9	1462.9			HD3
22	3410	3490			
23	2000	2020			
24	1626.5	1660.5	HD3	HD2	
25	1850	1915		HD3	
26	814	849	IL FSdiv4	HD2	
27	807	824		HD2	
28	703	748	HD2		HD3
29	717	728			
30	2305	2315			
31	452.5	457.5			
32	2575	2635			IL FSdiv4
33	1900	1920			
34	2010	2025			
35	1850	1910		HD3	
36	1930	1990	IL FSdiv4		HD2
37	1910	1930			
38	2570	2620			IL FSdiv4
39	1880	1920			
40	2300	2400	HD2	HD2	
41	2496	2690		HD2	IL FSdiv4
42	3400	3600	HD3	HD3	

Table 4. HD2, HD3, and IL Spurs falling In-band for In-band Blockers (continued)

BAND #	RX BAND (MHz)		SAMPLE RATE (MSPS)		
	LOW	HIGH	2211.84	2457.6	2949.12
43	3600	3800	HD2	HD3	HD3
44	703	803	HD2		HD3
45	1447	1467			HD3
46	5150	5925	HD2	HD2	HD2
65	1920	2010	IL FSdiv4		HD2
66	1710	1780			
67	738	758	HD2		HD3
68	698	728			

For multi-band use cases, you need to look at the in-band blockers from each band and whether they fall in any of the receive bands. Of course, it is more difficult to always be able to plan around HD2, HD3, and the IL spur. For example, consider a band 1 and band 3 combination. At 2949.12 MSPS, HD2 from band 1 falls back in band 1. A better rate is 2457.6 MSPS. Figure 20 shows the frequencies where HD2 or HD3 fall in either band 1 or band 3. Since all the frequencies are outside the bands, this is a good choice of sample rates as the blockers can be sufficiently suppressed to lower the energy due to HD2 or HD3 that fall in-band.
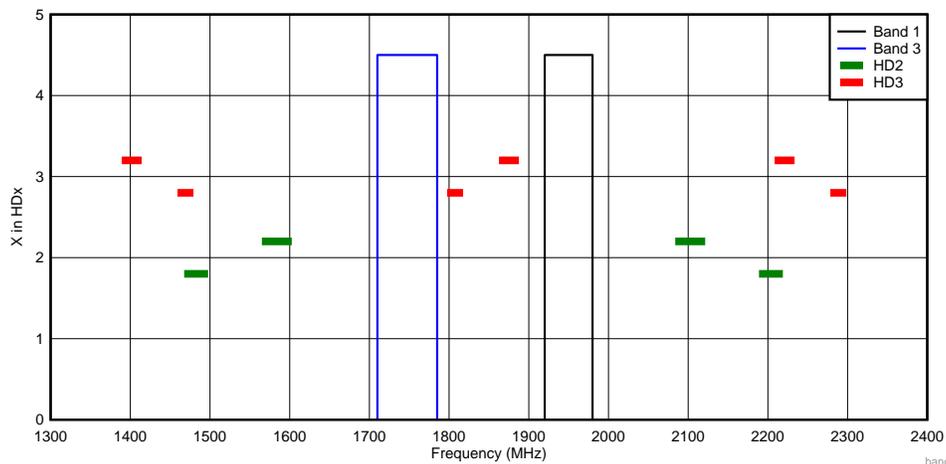
Figure 20. HDx Frequencies That Fall In-Band for Band 1 and Band 3 with f$_{ADC}$ = 2457.6 MSPS

5.2 *Out-of-band Filter Requirements*

Out-of-band blockers are specified across a frequency range of 1 MHz to 12750 MHz, excluding a guard band of 20 MHz around each wanted band. Filter in required between the antenna and the ADC for three effects:

- Blockers at the ADC image location in other Nyquist zones fall in band, and therefore need to be suppressed below the wanted signal sensitivity level. Typically more than 90 dB is required.

- At frequencies where low order harmonics such as HD2 or HD3 fall in-band, the out-of-band blocker must be suppressed so the spurious energy is below the in-band SFDR level. Using a HD2 model that improves 2 dB per dB and a HD3 model that improved 3dB per dB (similar to the model shown in Figure 26) and assuming 70 dBc for a fullscale tone, the blocker amplitude needs to be -5 dBFS for frequencies where HD3 falls in band and -7.5 dBFS where HD2 falls in-band. Include suppression of F$_S$ / 2 and F$_S$ / 4 interleaving spurs, which image 1 dB per dB with amplitude.

- At other frequencies, the out-of-band blocker is suppressed to increase the overall rejection of the out-of-band blocker. In the analysis, a minimum value of 15 dB suppression is used. For a typical filter, this is defined by the filter suppression at the edge of the guard band, as at a larger frequency offset the

suppression increases.

The filter suppression can be a combination of the antenna filter or duplexer and frequency responses of the LNA, balun, and Nyquist filter. Figure 21 and Figure 22 (zoomed) show an example of the filter requirements for Band 41 with a sample rate of 2949.12 MSPS. The third Nyquist zone image between approximately 3200 MHz and 3400 MHz is the closest point requiring 100 dB of suppression, which then repeats at larger offset frequencies in higher Nyquist zones.

Comparing to other radio architectures, the filtering requirements near the band are similar to an IF receiver with an IF of at 340 MHz. The mixer can also have MxN spurious products that require suppression of the out-of-band blocker at other frequencies, similar to HD2 and HD3 for RF sampling. A 0IF receiver has lower requirements on the large suppression requirement as analog baseband filtering before the ADC can provide 40 to 50 dB of suppression. Also, the frequency offset of the Nyquist image typically is several GHz. Therefore, the out-of-band filtering requirement is an advantage for 0IF radios.
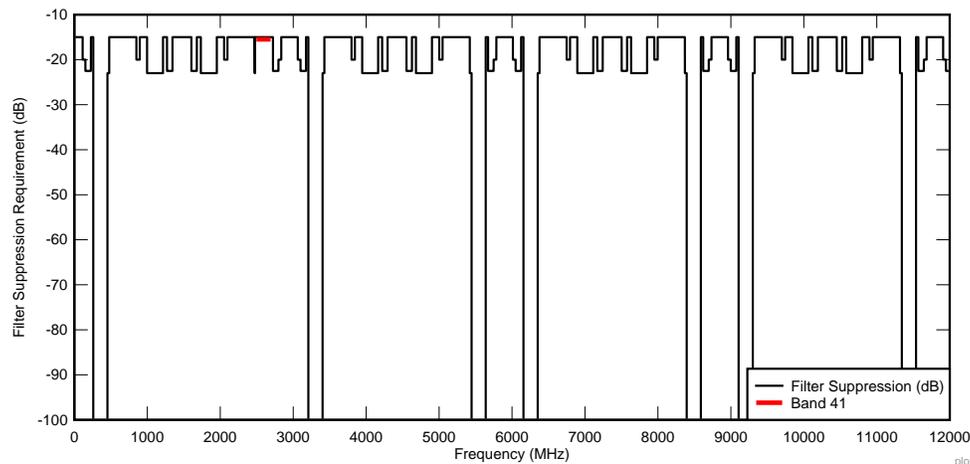
Figure 21. Filter Suppression Requirements for Band 41
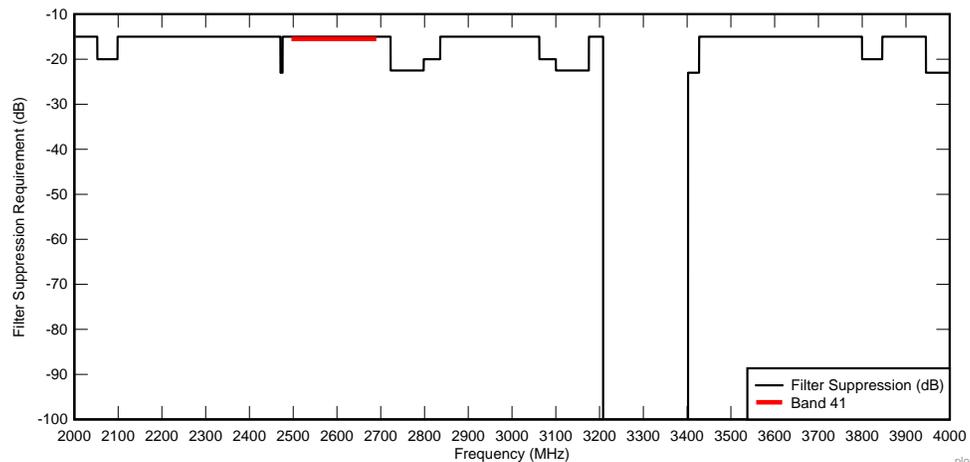
Figure 22. Filter Suppression Requirements for Band 41 (Zoomed)

5.3 Radio Parameters Versus DSA Setting

There are two limits for the maximum input peak power for the AFE7686:

1. The power at the input pins before the DSA must be below the maximum input power for reliability. This is independent of DSA setting, although the having the DSA at maximum attenuation improves the lifetime when the reliability limit is exceeded.

2. The power at the ADC after the DSA must be less than the ADC fullscale. The power at the input pins

increases with increasing DSA setting.

The AFE7686 has an automatic gain controller (AGC) that monitors the peak power at the ADC and adjusts the DSA setting so that the ADC does not saturate. Using the front-end parameters in Table 1, the input fullscale (FS) at the antenna versus DSA setting is shown in Figure 23. Both the minimum and maximum gain values are shown, as it is important to verify that the input power to the AFE7686 is not above the maximum input power.
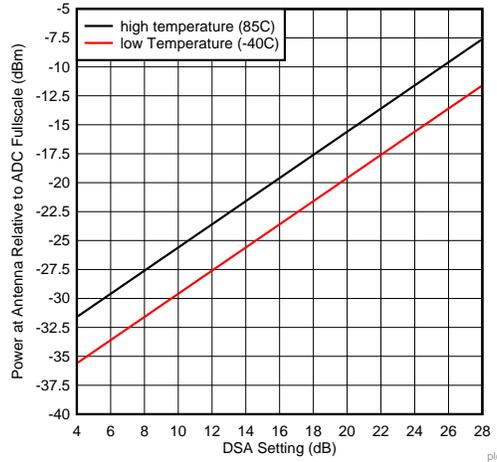
Figure 23. Power at the Antenna Corresponding to ADC Fullscale

Between 0 dB and 4 dB, the DSA has no effect on NF and IIP_3. Above 4 dB, NF, and IIP_3 increase approximately by 1 dB per dB of attenuation. Figure 24 shows the effect on the total cascaded NF for the radio, and Figure 25 shows the total cascaded IIP_3 for the radio.
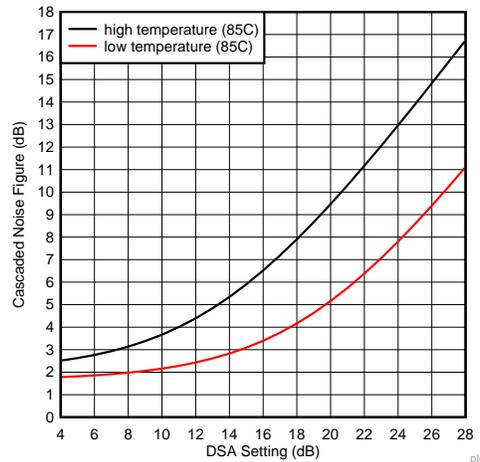
Figure 24. Cascaded NF Versus DSA Setting
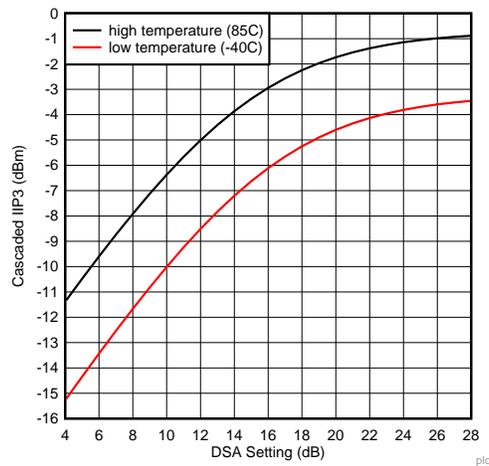
Figure 25. Cascaded IIP$_3$ Versus DSA Setting

5.4 ADC Parameters Versus Input Amplitude

It is also important to look at how the amplitude of distortion products like the second harmonic (HD2) and third harmonic changes with the amplitude of the signal at the ADC input (after the DSA). Figure 26 show HD3, 4, and 5 for the AFE7686 versus the input amplitude with a fixed DSA attenuation of 4 dB. Each simplified worst case model is generated using the expected slope for each order of distortion that saturates at a level of -85 dBFS.
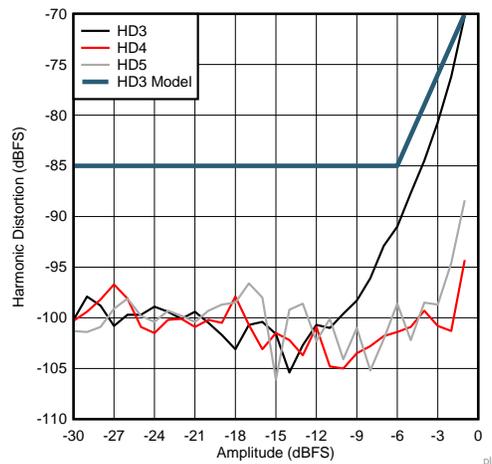
Figure 26. HD3, HD4, and HD5 Versus Input Amplitude

5.5 ADC Characteristics for Modulated Signals

In most receiver specifications, the blocking signals in the receiver tests are modulated signals. ADC datasheet parameters, however, use tones (also known as continuous waves - CW). There are two effects to be considered for applying the CW harmonic distortion parameters to a modulated signal.

First, the power of a CW signal is constant, while a modulated signal such as LTE, has instantaneous peaks in power that are significantly higher than the average power. This is characterized by the peak-to-average ratio (PAR). For a single LTE carrier, an estimate of the PAR is 10 dB. To avoid saturation of the ADC, which can cause severe distortion, the DSA attenuation can be set so the peak power never exceeds -1 dBFS and the average power -11 dBFS.

If the power of the modulated blocker signal is varying, how should the distortion level be estimated from the cascaded IIP$_3$ be used? There are three different cases to consider:

1. A single modulated signal, measuring the power in the adjacent channel
2. A modulated signal and a CW tone, measuring the power at the intermodulation products $2 \times f_{MOD} - f_{CW}$ and $2 \times f_{CW} - f_{MOD}$
3. Two modulated signals, Measuring the power at the intermodulation products at $2 \times f_{MOD1} - f_{MOD2}$ and $2 \times f_{MOD1} - f_{MOD2}$

Using a simple polynomial model for IMD3, you can estimate the equivalent 2-tone power that produces the same power in the measured bandwidth. See for a more thorough discussion of this topic.

Table 5 summarizes the results.

Table 5. IMD3 Frequency and Equivalent 2-tone Power

SIGNAL 1	SIGNAL 2	IMD3 MEASUREMENT FREQUENCY	2-TONE EQUIVALENT TOTAL POWER(dBm)	AADJUSTMENT FACTOR AF (dB)
5 MHz LTE	NA	ACPR = fLTE +/- 2.25 to 4.75 MHz	P_{LTE} + AF	0
5 MHz LTE	CW	$2 \times f_{LTE} - f_{CW}$		2
		$2 \times f_{CW} - f_{LTE}$		4
5 MHz LTE	5 MHz LTE	$2 \times f_{LTE1} - f_{LTE2}$		-1
		$2 \times f_{LTE2} - f_{LTE1}$		-1
5 MHz LTE	Narrowband LTE	$2 \times f_{LTE1} - f_{LTE2}$		-1
		$2 \times f_{LTE2} - f_{LTE1}$		0

Once there is an equivalent 2-tone power, you can estimate the IMD3 product by the cascaded IIP_3.

A second difference between a modulated and CW signal is that a CW signal has no bandwidth and the modulated signal has bandwidth. While the distortion product of a CW signal also has no bandwidth, a modulated signal spreads in frequency by the order of distortion (2x for HD2 for example). If the blocker and wanted signal are both the same bandwidth, then even if the blocker distortion product falls directly on the wanted signal, some of the energy of the distortion product falls outside the wanted signal and, therefore, not degrade the wanted signal.

For reasons discussed in Section 3.1.1, this second difference is not included in our analysis.

6 System Degradation Mechanisms

There are several degradation sources that limit the minimum wanted signal sensitivity in the receiver tests. Depending on the specific receiver test, sum the sources in different ways to estimate the total degradation. All equations below are the degradation power in dBm at the antenna input.

6.1 System Noise

The total system noise, characterized by the cascaded NF, summed across the wanted signal bandwidth is calculated by Equation 1.

$$P_{noise}\left(dBm\right) = -174\frac{dBm}{Hz} + NF(DSA_{setting}) + 10 * \log_{10}\left[BW\left(Hz\right)\right]$$

where
- NF is the cascaded Noise Figure which is a function of the DSA setting (Figure 24)
- BW is the bandwidth of the wanted signal   (1)

6.2 TX Noise Leakage (FDD Only)

For FDD systems, the high power TX is on at the same time as the RX. In the RX band frequency, the TX has noise and residual distortion power that leaks through the duplex filter in the RX. The degradation power is calculated by Equation 2.

$$P_{TXinRX}(dBm) = P_{TXnoise}\left(\frac{dBm}{Hz}\right) - Iso_{TXtoRX}(dB) + 10 * \log_{10}\left[BW(Hz)\right]$$

where

- $P_{TXnoise}$ is the power at the TX output in the RX band
- Iso_{TXtoRX} is the duplex filter rejection between TX and RX in the RX band
- BW is the bandwidth of the wanted signal   (2)

$P_{TXnoise}$ is fixed regardless of the DSA setting. For TDD systems, TX is not active during the RX time, and so this degradation source is not included.

6.3 Phase Noise

The phase noise of the clock for the AFE7686 ADC mixes with any signal at the receiver and generate energy at the wanted signal. This is equivalent to reciprocal mixing with LO phase noise in a down converting mixer. The phase noise must include the phase noise from the clock source and any additive phase noise in the clock path. The AFE7686 has an internal PLL and VCOs to generate the high frequency sampling clock for the DAC and ADCs, which is typically good enough for LTE. For multi-carrier GSM, a lower phase noise external clock needs to meet the large blocker requirements.

Approximate the phase noise due to a blocker by using the spot phase noise at the offset frequency that is the frequency difference between center of the blocker and center of the wanted signal. This essentially treats the phase noise as constant across the blocker and wanted signal combined bandwidths.

During the ADC sampling, the phase noise at the ADC output from mixing with a blocker is calculated by Equation 3.

$$P_{pn}(dBm) = P_{blk}(dBm) + PN_{blkoffset}\left(\frac{dBm}{Hz}\right) + 10 * \log_{10}\left[BW(Hz)\right]$$

where

- P_{blk} is the RMS power of the blocker
- $PN_{blkoffset}$ is the phase noise at the offset equal to the spacing between the center frequencies of the blocker and wanted signal
- BW is the bandwidth of the wanted signal   (3)

The phase noise is at the blocker input frequency, not the ADC sample rate, and can be adjusted from the sample rate of the measurement is calculated by Equation 4.

$$PN_{blk}(dBm) = PN_{meas}\left(\frac{dBc}{Hz}\right) + 20 * \log_{10}\left[\frac{f_{blk}}{f_{meas}}\right](dB)$$

where

- PN_{meas} is the measured phase noise
- f_{meas} is the frequency of the phase noise measurement
- f_{blk} is the frequency of the blocker   (4)

The phase noise for the internal PLL and VCO is shown in the datasheet in section 5.11.3 Typical PLL Characteristics using the TX output clock measured at either 1966.08 MHz for the VCO at 5898.24 MHz or 2211.84 MHz for the VCO at 8847.36 MHz.

For the MCGSM use case with external clock, the additive phase noise is listed in Table 3 and needs to be power summed to the input clock adjusted by the clock frequency in Equation 4.

6.4 Adjacent Channel Intermodulation

When a large blocker is adjacent to the wanted signal channel and has the same BW, third order intermodulation causes some of the blocker power to leak into the wanted signal.

You can estimate this using the cascaded IIP_3 and the equivalent 2-tone equivalent power from Table 5 by Equation 5.

$$P_{IMD3acpr}(dBm) = 3 * P_{BLK}(dBm) - 2 * IIP_3(dBm)$$

where

- P_{blk} is the RMS power of the blocker
- IIP_3 is the cascaded IP_3 at the antenna input (5)

6.5 ADC SFDR

For bands where HD2, HD3, and the IL spurs fall out-of-band for the blocker location, you can use the specification in Table 3 for the SFDR excluding HD2, HD3, and IL spurs. The value of -85 dBFS at the ADC is independent of DSA setting and input amplitude, but referred to the antenna input. You need to account for the gain between the antenna and ADC, including the DSA. The estimation is calculated by Equation 6.

$$P_{sfdr}(dBm) = SFDR_{adc}(dBFS) + FS_{DSA=MIN}(dBm) - DSA_{setting}(dB) - FEGAIN(dB)$$

where

- $SFDR_{adc}$ is the in-band ADC SFDR in dBFS
- $FS_{DSA=MIN}$ is the input fullscale at the minimum DSA setting
- FEGAIN is the front-end gain to the input of the AFE7686 (6)

For the case where HD2, HD3, or the IL spurs fall in band, the appropriate SFDR estimation depends on the PAR of the blocker signal. For modulated signals with a PAR of 10 dB or greater, the average power blocker is always < 10 dBFS and HD2, HD3, and IL components are all limited by the -85 dBFS floor for SFDR shown in .

For a CW blocker, as the ADC nears fullscale, the SFDR for HD2, HD3, or IL can be worse than the -85 dBFS floor. To account for this, for the CW out-of-band blocker, change the SFDR term above to Equation 7.

$$SFDR_{adc}(dBFS) = MAX\left[-85, -75 + A_{in}(dBFS)\right]$$

where

- A_{in} is the amplitude of the tone in dBFS (7)

6.6 Intermodulation Between the Blocker and TX Leakage in TX Band (FDD Only)

For FDD systems, the large TX signal power in the TX band leaks through the duplex filter into the RX. This mixes with the blocker due to third order intermodulation, with the product possibly falling on the wanted signal. Using the equivalent 2-tone equivalent power from Table 5, the estimate is calculated by Equation 8.

$$P_{tx*blk}(dBm) = 3*\left\{\frac{\left[P_{blk}(dBm) + P_{tx}(dBm) - ISO_{txband}(dBm)\right]}{2} + AF(dB)\right\} - 2*IIP_3(dBm)$$

where

- P_{blk} is the RMS power of the blocker
- P_{tx} is the TX output power in the BW of the wanted signal
- Iso_{txband} is the duplex filter isolation between the TX and RX at the TX band frequency
- AFE is the adjustment factor from Table 5
- IIP_3 is the cascaded IP_3 at the antenna input (8)

AF depends on the characteristics of the blocker – modulated wideband, modulated narrowband, or CW as listed in Table 5.

7 Receiver Tests for LTE

7.1 3GPP Standards Radio Performance

The 3GPP specifications covering the base transceiver station (BTS) are in the following documents available for download from the 3GPP website.

Table 6. 3GPP Standard Specifications for Radio Performance

STANDARD	DOCUMENT	VERSION USED
Multi-carrier GSM (MC-GSM)	3GPP TS 45.005	V13.1.0 (2016-05)
LTE	3GPP TS 36.104	V15.1.0 (2017-12)
NR (5G)	3GPP TS 38.104	V15.3.0 (2018-9)

In addition, there are other useful specification documents discussing conformance testing and system scenarios.

Table 7. Other Useful Standard Documents for LTE

DOCUMENT	TITLE
3GPP TS 36.141	E-UTRA Base Station (BS) conformance testing
3GPP TR 36.942	Radio Frequency (RF) system scenarios

In the subsequent analyses, the performance of the specifications with blockers is estimated in two ways:

1. The sensitivity level margin at the specified blocker power
2. The maximum blocker power at which the sensitivity level specification is met

7.2 LTE Tests – Single Band at High Temperature

7.2.1 Specification Description

The reference sensitivity test for LTE is a described in section 7.2 of 36.104, and measures the minimum LTE signal power at the antenna connector that can be received in the absence of any interfering signals (other than the TX output leaking into the receiver for FDD systems) and meets the minimum throughput requirement (a measure of the bit error rate).

The required reference sensitivity power level (PREFSENS), taken from table 7.2.1-1, are listed versus channel bandwidth in Table 8. The required signal PSD, assuming a flat noise and interference level across the bandwidth, is listed in the last column.

Table 8. Reference Sensitivity Power Levels for LTE

E-UTRA CHANNEL BANDWIDTH [MHz]	REFERENCE SENSITIVITY POWER LEVEL, PREFSENS [dBm]	PREFSENS PSD LEVEL dBm/Hz
1.4	-106.8	-167.8
3	-103.0	-167.3
5	-101.5	-168.0
10	-101.5*	-168.0
15	-101.5*	-168.0
20	-101.5*	-168.0

A specific LTE signal is defined in Annex A.1 for the reference sensitivity test, which is QPSK with a code rate of 1/3. The signal-to-interference and noise ratio (SINR) required to meet the throughput requirement is 0.25 dB (Ghosh, 2011, p. 98). Since the PSD requirement for the 5 MHz channel is the lowest (most stringent), as long as the noise and interference level is flat across frequency, you only need to analyze the 5 MHz channel bandwidth test case to be assured the other cases pass.

7.2.2   Specification Analysis

For the receiver sensitivity test, the key impairments are the system noise and for FDD the TX noise leakage through the duplexer into the receiver in the RX band. Since the only input is the small wanted signal, there is no large input signal to generate spurs that fall into the wanted signal and degrade the sensitivity. However, any fixed spur (those that are independent of amplitude) need to be considered if they fall into the RX band. Similarly, phase noise can also be ignored as it scales with the input signal.

Table 9 lists the results for high temperature, showing the system must have 3.2 dB margin for TDD LTE and 2.79 dB margin for FDD LTE. The results for lower temperature are included in the summary in section 2.1.7.2, as they are qualitatively similar to high temperature.

Table 9. Receiver Sensitivity Test Parameters and Estimation

SENSITIVITY TEST		
Signal BW	4.5	MHz
TX Noise in RX band (@ PAout)	-86.5	dBm/Hz
TX-RX isolation	95	dB
REQUIREMENTS		
PSENS	-101.5	dBm
SNR	0.25	dB
RESULTS		

Conditions	Impairment (dBm)		Margin (dB)	
DSA Setting (dB)	Noise	FDD TX	TDD	FDD
4	-104.95	-114.97	3.20	2.79

7.2.3   Adjacent Channel Selectivity Test

7.2.3.1   Specification Description

The adjacent channel selectivity test is a described in section 7.5 of 36.104 and is a measure of the ability for the receiver to receive a wanted signal in the presence of an adjacent blocking signal. The blocker can be a full LTE signal (wideband) or a narrow-band LTE signal with one resource block (RB) in the adjacent channel. Table 10 lists the blocker power levels.

Table 10. Adjacent Channel Selectivity Test Power Levels for Wide Area BTS

WANTED RECEIVED SIGNAL		NARROWBAND	WIDEBAND
BANDWITH (MHz)	POWER (dBm)	POWER (dBm)	POWER (dBm)
1.4	-100.8	-49	-52
3	-97.0	-49	-52
5	-95.5	-49	-52
10	-95.5*	-49	-52
15	-95.5*	-49	-52
20	-95.5*	-49	-52

7.2.3.2 Specification Analysis – Narrowband ACS

For the narrowband adjacent channel selectivity test, in addition to the degradation from the noise and TX leakage for FDD, you need to add the degradations due to the ADC phase noise and ADC SFDR. Table 11 lists the test parameters and results. At the blocker level in the specification, both TDD and FDD have more than 8 dB of margin for the sensitivity level, meaning that even the sensitivity level for the sensitivity test can be met with the blocker at the specified power - no relaxation is needed. The maximum blocker level where the sensitivity level can still be met is -29 dBm (20 dB above the specification).

Table 11. Narrowband Adjacent Channel Selectivity Test Parameters and Results for High Temperature

NARROWBAND BLOCKER TEST		
Signal BW	4.5	MHz
TX Output Power	40	dBm
TX "ACPR" RX band (@ PAout)	-60	dBc
TX-RX isolation	95	dB
Spec Blocker Amplitude	-49.0	dBm
Blocker PAR	10.0	dB
Phase noise @ 2.8 MHz offset	-138.4	dBc/Hz
AGC Target	-1.0	dBFS
Min DSA Setting	4.0	dB
Max DSA Setting	24.0	dB

REQUIREMENTS		
PSENS	-95.5	dBm
SNR	0.25	dB

RESULTS								
Conditions				Degradation Sources (dBm)				Margin (dB)	
Blocker Level (dBm)	DSA (dB)	Peak Amp (dBFS)	Peak Input (dBm)	Noise	FDD TX	PN	SFDR	TDD	FDD
-49	4.0	-7.4	-11.0	-105.0	-115.0	-120.9	-116.6	8.8	8.4
-29	18.0	-1.4	9.0	-99.6	-115.0	-100.9	-102.6	0.3	0.3

Figure 27 shows how the DSA setting and degradation power from different sources (referred to the antenna input) change with blocker level. The TX leakage in the RX band (FDD TX) is independent of DSA setting and constant for all blocker levels. The phase noise (PN), which is constant in dBc, increases 1 dB/dB with blocker level. The noise is constant until the DSA attenuation begins to increase at -42 dBm, although slowly at first, as the front-end noise dominates. The ADC SFDR is constant until the DSA attenuation increases, after which it increases 1 dB/dB. At the maximum blocker level of -29 dBm, the largest degradation is noise, followed by phase noise and ADC SFDR.
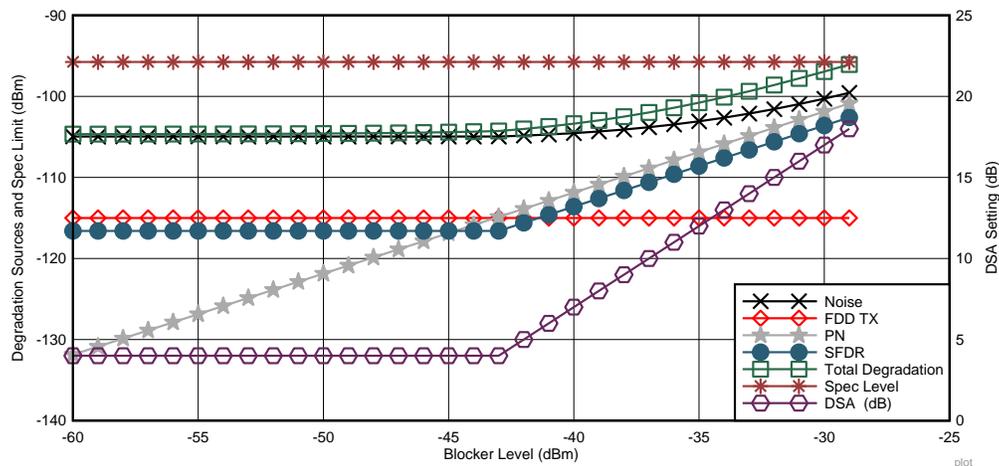
Figure 27. DSA and Degradation Power Versus Blocker Level for the Narrowband Adjacent Blocker Sensitivity Test

7.2.3.3   Specification Analysis – Wideband ACS

For the wideband adjacent channel selectivity test, in addition to the degradation sources listed above for the narrowband blocker, you need to add the IMD3 of the adjacent blocker. Table 12 lists the test parameters and results. At the blocker level in the specification, both TDD and FDD have more than 8 dB of margin for the sensitivity level. The maximum blocker level where the sensitivity level can still be met is -37 dBm (15 dB above the specification).

Table 12. Wideband Adjacent Channel Selectivity Test Parameters and Results for High Temperature

ADJACENT CHANNEL SELECTIVITY TEST			
Signal BW	4.5	MHz	
TX Output Power	40	dBm	
TX "ACPR" RX band (@ PAout)	-60	dBc	
TX-RX isolation	95	dB	
Spec Blocker Amplitude	-52.0	dBm	
Blocker PAR	10.0	dB	
Phase noise @ 2.8 MHz offset	-138.4	dBc/Hz	
AGC Target	-1.0	dBFS	
Min DSA Setting	4.0	dB	
Max DSA Setting	24.0	dB	

REQUIREMENTS			
PSENS	-95.5	dBm	
SNR	0.25	dB	

RESULTS										
Conditions				Degradation Source (dBm)					Margin (dB)	
Blocker Level (dBm)	DSA Setting (dB)	Peak Amp (dBFS)	Peak Input (dBm)	Noise	FDD TX	PN	IMD3	SFDR	TDD	FDD
-52	4	-10.4	-14.0	-105.0	-115.0	-123.9	-133.2	-116.6	8.86	8.48
-37	10	-1.4	1.0	-103.8	-115.0	-108.9	-98.3	-110.6	0.97	0.91

Figure 28 shows how the DSA setting and degradation sources referred to the antenna input change with blocker level. At the maximum blocker level, the largest degradation source is the IMD3 for the blocker falling into the adjacent wanted signal.
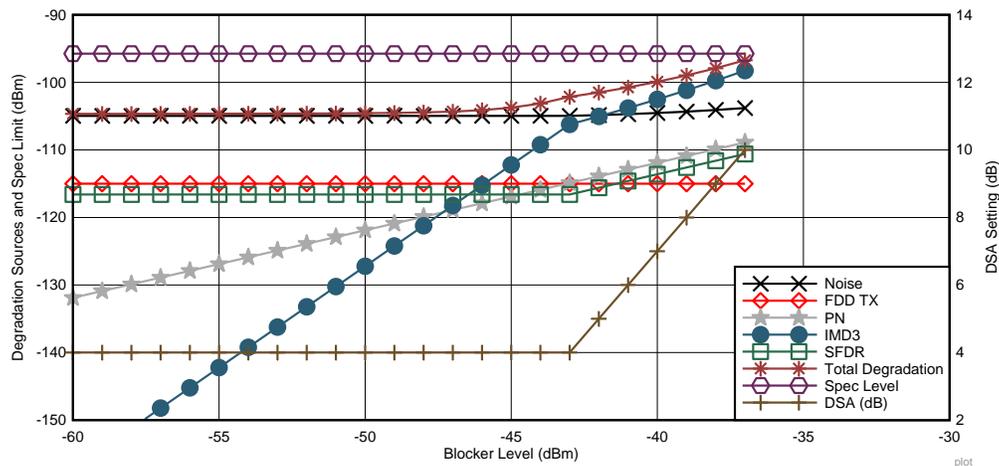
Figure 28. DSA and Degradation Power Versus Blocker Level for the Wideband Adjacent Channel Selectivity Test

7.2.4 Blocker Tests

7.2.4.1 Specification Description

The blocker test is a measure of the ability for the receiver to receive a wanted signal in the presence of the blocker, which is not adjacent to the wanted signal. There are different blocker requirements for in-band and out-of-band. The in-band blocker test modulated blocker located inside the uplink band plus a guard band (usually 20 MHz on each side of the band). The modulated blocker is a full LTE signal of the same bandwidth as the wanted signal up to 5 MHz, and a 5 MHz LTE signal for 10 and 20 MHz wanted signals.

The in-band blocker can be located anywhere in the uplink band (plus guard band), to a minimum offset of 1.5x the wanted signal bandwidth. The power level is either -43 dBm or -38 dBm, depending on the band.

The out-of-band blocker is CW tone located outside the uplink band plus the guard band.

Table 13. Blocker Test Power Levels for Wide Area BTS

WANTED SIGNAL		IN-BAND BLOCKER	OUT-OF-BAND BLOCKER
Bandwidth (MHz)	Power (dBm)	Power (dBm)	Power (dBm)
1.4	-100.8	-43 (-38)	-13
3	-97.0	-43 (-38)	-13
5	-95.5	-43 (-38)	-13
10	-95.5*	-43 (-38)	-13
15	-95.5*	-43 (-38)	-13
20	-95.5*	-43 (-38)	-13

7.2.4.2   *Specification Analysis – Inband Blocker*

For the in-band blocker test, in addition to the degradation from the noise and TX leakage for FDD, you need to add the degradations due to the ADC phase noise, ADC SFDR and for FDD, the intermodulation distortion resulting from the blocker mixing with the TX band leakage. Table 14 lists the test parameters and results. At the blocker level in the specification, both TDD and FDD have more than 8 dB of margin for the sensitivity level. The maximum blocker level where the sensitivity level can still be met is -29 dBm (14 dB above the specification).

Table 14. Inband Blocker Test Parameters and Results for High Temperature

WIDEBAND BLOCKER TEST		
signal BW	4.5	MHz
TX Output Power	40	dBm
TX "ACPR" RX band (@ PAout)	-60	dBc
TX-RX isolation	95	dB
Spec Blocker Amplitude	-43.0	dBm
Blocker PAR	10.0	dB
Phase noise @ 2.8 MHz offset	-138.4	dBc/Hz
AGC Target	-1.0	dBFS
Min DSA Setting	4.0	dB
Max DSA Setting	24.0	dB

REQUIREMENTS		
PSENS	-95.5	dBm
SNR	0.25	dB

RESULTS									
Conditions				Degradation Sources (dBm)					Margin (dB)	
Blocker Level (dBm)	DSA Setting (dB)	Peak Amp (dBFS)	Peak Input (dBm)	Noise	FDD TX	PN	BLK*TX	SFDR	TDD	FDD
-43	4	-1.4	-5.0	-105.0	-115.0	-114.9	-118.2	-116.6	8.35	8.01
-29	18	-1.4	9.0	-99.6	-115.0	-100.9	-115.5	-102.6	0.28	0.22

shows how the DSA setting and degradation sources referred to the antenna input change with blocker level. At the maximum blocker level, the largest degradation source is the IMD3 for the blocker falling into the adjacent wanted signal.
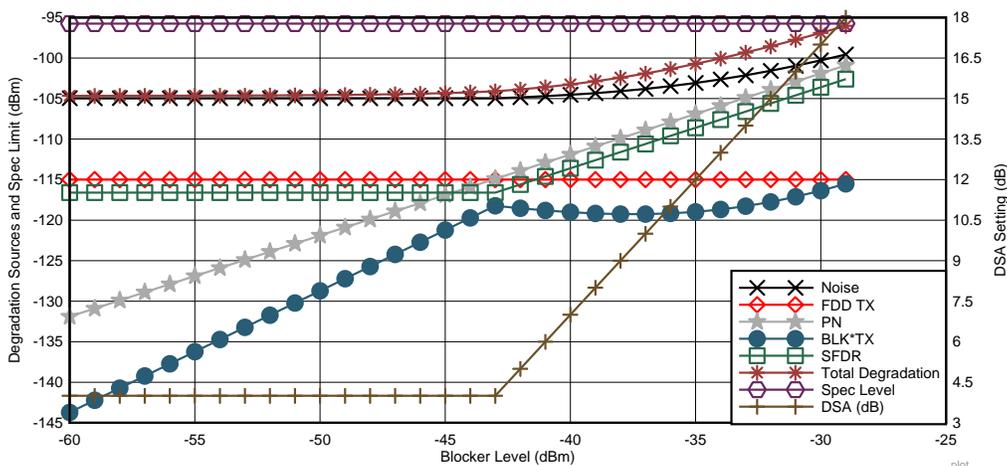

Figure 29. DSA and Degradation Power Versus Blocker Level for the Inband Blocker Test

7.2.4.3 Specification Analysis – Out-of-Band Blocker

For the out-of-band blocker test, in addition to the degradation from the noise and TX leakage for FDD, you need to add the degradations due to the ADC phase noise, ADC SFDR and for FDD, the intermodulation distortion of the blocker mixing with the TX band leakage. Table 15 lists the test parameters and results. At the blocker level in the specification, both TDD and FDD have more than 5 dB of margin for the sensitivity level. The maximum blocker level where the sensitivity level can still be met is -5 dBm (10 dB above the specification).

Table 15. Out-of-Band Blocker Test Parameters and Results for High Temperature

CW BLOCKER (OUT-OF-BAND)		
signal BW	4.5	MHz
TX Output Power	40	dBm
TX "ACPR" RX band (@ PAout)	-60	dBc
TX-RX isolation	95	dB
RX Filter @ 20 MHz out-of-band	15	dB
Spec Blocker Amplitude	-15.0	dBm
Blocker PAR	0.0	dB
Phase noise @ 22.5 MHz offset	-152.4	dBc/Hz
AGC Target	-1.0	dBFS
Min DSA Setting	4.0	dB
Max DSA Setting	24.0	dB
MIN ADC SFDR	-75.0	dBFS
REQUIREMENTS		
PSENS	-95.5	dBm
SNR	0.25	dB

RESULTS										
Conditions				Degradation Sources (dBm)					Margin (dB)	
Blocker Level (dBm)	DSA Setting (dB)	Peak Amp (dBFS)	Peak Input (dBm)	Noise	FDD TX	PN	BLK*TX	SFDR	TDD	FDD
-15	7	-1.4	-2.0	-104.5	-115.0	-115.9	-104.0	-105.0	3.87	3.75
-9	13	-1	4	-102.6	-115.0	-109.9	-103.7	-99.0	0.56	0.51

Figure 30 shows how the DSA setting and degradation sources referred to the antenna input change with blocker level. At the maximum blocker level, the largest degradation source is noise and IMD3 between the blocker and TX band leakage.
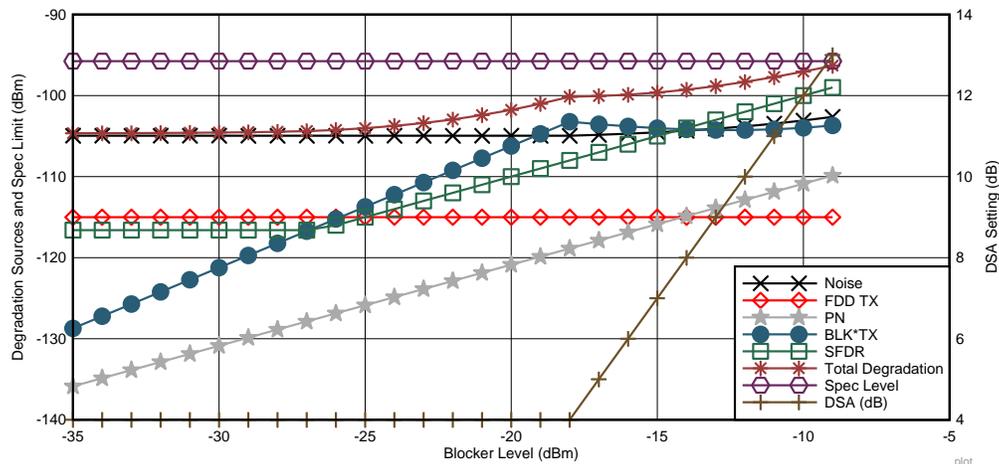

Figure 30. DSA and Degradation Power Versus Blocker Level for the Out-of-Band Blocker Test

7.2.5 IMD Test

7.2.5.1 2.1.7.1.4.1 Specification Description

The intermodulation test is a measure of the ability for the receiver to receive a wanted signal in the presence of two blocking signals. The wanted signal is located at the frequency of the third order intermodulation distortion generated by the blockers. The blockers are a modulated LTE signal and a CW tone, both with -52 dBm input power.

Table 16. Intermodulation Test Blocker Parameters

WANTED RECEIVED SIGNAL		LTE BLOCKER		CW BLOCKER
Bandwidth (MHz)	Power (dBm)	BW (MHz)	Power (dBm)	Power (dBm)
1.4	-100.8	1.4	-52	-52
3	-97.0	3	-52	-52
5	-95.5	5	-52	-52
10	-95.5*	5	-52	-52
15	-95.5*	5	-52	-52
20	-95.5*	5	-52	-52

7.2.5.2 Specification Analysis - IMD Test

For the IMD test, in addition to the degradation from the noise and TX leakage for FDD, you need to add the degradations due to the ADC phase noise, ADC SFDR, and the intermodulation product between the two blockers. Table 17 lists the test parameters and results. At the blocker level in the specification, both TDD and FDD have more than 9 dB of margin for the sensitivity level. The maximum blocker level where the sensitivity level can still be met is -38 dBm (14 dB above the specification).

Table 17. IMD Test Parameters and Results for High Temperature

RX IMD		
signal BW	4.5	MHz
TX Output Power	40	dBm
TX "ACPR" RX band (@ PAout)	-60	dBc
TX-RX isolation	95	dB
RX Filter @ 20 MHz out-of-band	15	dB
Spec Blocker Amplitude	-52.0	dBm
Blocker PAR	13.0	dB
Phase noise @ 22.5 MHz offset	-152.4	dBc/Hz
AGC Target	-1.0	dBFS
Min DSA Setting	4.0	dB
Max DSA Setting	24.0	dB

REQUIREMENTS		
PSENS	-95.5	dBm
SNR	0.25	dB

RESULTS										
Conditions				Degradation Sources (dBm)					Margin (dB)	
Blocker Level (dBm)	DSA Setting (dB)	Peak Amp (dBFS)	Peak Input (dBm)	Noise	FDD TX	PN	IMD3	SFDR	TDD	FDD
-52	4	-7.4	-11.0	-105.0	-115.0	-137.9	-127.2	-116.6	8.89	8.51
-38	12	-1.4	3.0	-103.1	-115.0	-123.9	-98.0	-108.6	0.77	0.71

Figure 31 shows how the DSA setting and degradation sources referred to the antenna input change with blocker level. At the maximum blocker level, the largest degradation source is the intermodulation product between the blockers.
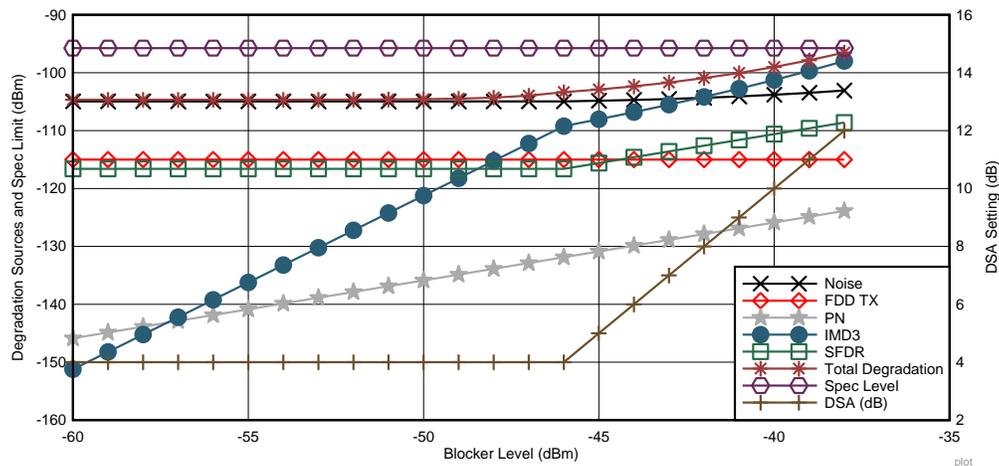

Figure 31. DSA and Degradation Power Versus Blocker Level for the IMD Test

7.3 LTE Tests – Single Band at Low Temperature

From Table 1 and Table 3, the differences for low temperature compared to high temperature are the following:

- Noise of the front end and AFE7686 is lower
- Front-end gain is higher
- Cascaded IIP$_3$ is lower

Tests dominated by noise improve versus high temperature and test dominated by distortion degrade.

The higher front-end gain at low temperature also results in the worst case for high power levels exceeding the operating range of the AFE7686. Between the maximum blocker level where the specification sensitivity is still met and the maximum blocker power for operation, the sensitivity is not met at one channel frequency (for a given blocker frequency), but at a higher wanted signal power is still properly received at that frequency and wanted signals at other channel frequencies still meet the specification requirement. When the blocker is high enough to the maximum input power for the AFE7686, then the ADC saturates and it is likely that wanted signals at most channel frequencies are not be properly received.

Table 18 shows a summary of the single band test results, including high and low temperature.

Table 18. Summary of Single Band LTE Tests

LTE STANDARD			HIGH TEMPERATURE			LOW TEMPERATURE			
Test	Wanted signal level (dBm)	Blocker level (dBm)	Margin at Spec Level (dB)		Maximum Blocker Amplitude (dBm)	Margin at Spec Level (dB)		Maximum Blocker Amplitude (dBm)	
			TDD	FDD	@Spec	TDD	FDD	@Spec	Operation
Sensitivity	-101.5	NA	3.20	2.79	NA	3.94	3.45	NA	NA
7.5 Narrow blocker	-95.5	-49	8.81	8.44	-29	9.67	9.22	-28	-11
7.5 ACS	-95.5	-52	8.86	8.48	-37	9.69	9.23	-38	-11
7.6 Blocking wideband	-95.5	-43	8.35	8.01	-29	9.67	9.22	-29	-11

Table 18. Summary of Single Band LTE Tests (continued)

LTE STANDARD			HIGH TEMPERATURE			LOW TEMPERATURE			
7.6 Blocking CW	-95.5	-15	5.12	4.96	-5	4.85	4.69	-6	14
7.8 Intermod	-95.5	-52	8.89	8.51	-38	9.62	9.17	-39	-14

7.4 LTE Tests – Multi-band Radios

As discussed in Section 3.1.2, for the multi-band analysis, use the non-blocker sensitivity level when the blocker is in a different band.

In the estimation, the noise degradation and FDD TX leakage in the RX band are the same between in-band and other band blockers. The ADC SFDR is also assumed to be the same, which is the case as long as HD2, HD3, and IL spurs from other band blockers do not fall in-band. This can be determined by multi-band frequency planning discussed in Section 5.1. The degradation due to IMD3 between the blocker and TX leakage or the two blockers in the IMD test are assumed to be the same.

summarizes the multiband performance when the blocker is in the other band. Since there is no blocker in the reference sensitivity test, there is no difference in the sensitivity level between a single and multi-band receiver. For the tests with a blocker, due to the lower wanted signal sensitivity for the tests, the margin at the specified blocker level is lower. For the ACS and Intermod blocker tests, the performance is better than when the blocker(s) are in-band. The reason for this is that the IMD contribution does not fall on the wanted signal when the blockers are in the other band. Also, the offset for the phase noise contribution is higher, so there is less degradation due to phase noise.

Table 19. Summary of Multi-band Performance when Blocker is in the Other Band

INTERPRETATION OF THE LTE STANDARD			HIGH TEMPERATURE			LOW TEMPERATURE		
Test	Wanted signal level (dBm)	Blocker level (dBm)	Margin at Spec Level (dB)		Maximum Blocker Amplitude (dBm)	Margin at Spec Level (dB)		Maximum Blocker Amplitude (dBm)
			TDD	FDD		TDD	FDD	
Sensitivity	-101.5	NA	3.20	2.79	NA	3.94	3.45	NA
7.5 Narrow blocker	-101.5	-49	2.91	2.53	-35 (-6 dB)	3.79	3.33	-34 (-6 dB)
7.5 ACS	-101.5	-52	2.91	2.52	-35 (+2 dB)	3.80	3.33	-34 (+4 dB)
7.6 Blocking wideband	-101.5	-43	2.72	2.35	-35 (-6 dB)	3.73	3.27	-34 (-5 dB)
7.6 Blocking CW	-95.5	-15	5.12	4.96	-5 (+4 dB)	4.90	4.75	-6
7.8 Intermod	-101.5	-52	2.91	2.53	-38	9.80	9.33	-29 (+10 dB)

8 Multi-carrier GSM Tests

8.1 Multi-carrier GSM Standard

GSM (Global System for Mobile communications) was developed in Europe as a second generation digital communications standard, first deployed in the early 1990's. While three new generations of cellular communications have since then been developed (roughly one per decade), many FDD networks still support GSM for voice and low data rate services.

Due to the stringent performance requirement in the GSM standard, all BTS radios were designed for a single carrier, relying on an IF SAW filter to reject blockers, until the mid-2000's when a relaxed standard for multi-carrier GSM (MCGSM) BTS was developed (3GPP, TS 45.005). This enabled multi-standards radios, covering GSM, WCDMA, LTE, and now 5G NR.

Table 20 lists a summary of the key MCGSM performance test and a comparison to LTE. In both cases, the most difficult test case (that is the case which sets the RX component performance requirements) is listed.

Table 20. Key MCGSM Performance Requirements

PARAMETER	MCGSM	LTE	UNIT
Carrier BW	*0.2*	5	MHz
Wanted signal SNR required	*6*	0.25	dB
IN-BAND FOR 1800 MHz			
Offset frequency	*0.8*	5	MHz
In-band blocker level	-25	-52	dBm
Sensitivity	-101	-95.5	dBm
Blocker/sensitivity	*76*	43.5	dB
Blocker type	*CW*	LTE5	
Exceptions allowed	*12*	0	
IN-BAND FOR 900 MHz			
Offset frequency	0.8	5	MHz
In-band blocker level	-16	-52	dBm
Sensitivity	-92	-95.5	dBm
Blocker/sensitivity	76	43.5	dB
Blocker type	CW	LTE5	
Exceptions allowed	*6*	0	
OUT-OF-BAND			
Out-of-band blocker	-15	-13	dBm
Sensitivity	-95	-95.5	dBm
Blocker/sensitivity	80	82.5	dB
IMD TEST			
Offset frequency	0.8	10	MHz
CW signal level	-43	-52	dBm
Modulated signal level	-43	-52	dBm
Sensitivity	-104	-95.5	dBm
Blocker/sensitivity	61	43.5	dB

The key specification that defines the performance requirements for the radio components is the in-band blocker test. The key differences to LTE are highlighted in the following list and the impact on the RF sampling AFE component is discussed below:

- **Carrier BW**: The narrower carrier bandwidth for GSM means that in the blocker tests, the radio noise figure is a much lower percentage of the overall degradation power than for LTE.
- **Required SNR and Blocker/Sensitivity Ratio**: The required SNR and blocker to sensitivity ratio combined are ~40 dB higher than for LTE. This places a much more stringent requirement of SFDR and phase noise for the RF sampling ADC.
- **Blocker Offset Frequency**: The lower offset frequency for the MCGSM blocker places a difficult requirement on the PLL/VCO phase noise. As a result, an external clock source is required for MCGSM radios.
- **Blocker Type**: Since the MCGSM blocker is defined as a CW source, there is no spreading in the harmonics that can lower the integrated power of the blocker harmonic in the wanted signal.
- **Exceptions**: MCGSM does allow exceptions for each in-band blocker frequency, where the blocker power is relaxed to -43 dBm. This can be used to excluded deterministic spurs or close in phase noise.
-

8.2 Calculation of Clock Phase Noise

To achieve the required phase noise at 800 kHz offset for the receiver, an external clock is required for the AFE7686. Even then, the sum of the external clock phase noise and the additive phase noise from the AFE7686 clock path still dominates the total degradation budget.

As the AFE7686 requires an external clock at the DAC sample rate when the internal PLL/VCO is bypassed, a clock frequency of 5898.24 MHz or 8847.36 MHz is required. For this analysis, the LMX2594 at 8847.36 MHz is used as the clock source. In addition, two exceptions at +/- 800 kHz are used, so the offset frequency of concern expands to 1 MHz, reducing the phase noise. At 1 MHz offset and 8847.36 MHz, the LMX2594 phase noise is -127.4 dBc/Hz. At the highest blocker input frequency (3GPP band 2) of 1910 MHz, the phase noise at the wanted signal is -140.7 dBc/Hz.

The measured additive phase noise at 1 MHz offset for the AFE7686 is -140.5 dBc/Hz. Combined, the total phase noise at 1 MHz offset from the blocker is -137.6 dBc/Hz.

8.2.1 Summary of MCGSM In-band Blocker Test

Table 21 shows the summary and results for the MCGSM in-band blocker test. Unlike LTE where there is significant margin at the specified blocker level, in the MCGSM test, there is only 0.5 dB of margin. To improve the margin, another four exceptions can be used to increase the blocker offset frequency to 1.4 MHz, in which case the phase noise improves to -141.1 dBc/Hz, resulting in a total of 2.3 dB of margin.

Table 21. MCGSM In-band Blocker Test Results

IN-BAND BLOCKER TEST		
Signal BW	0.2	MHz
TX Output Power	40	dBm
TX "ACPR" RX Band (@ PAout)	-90	dBc
TX-RX Isolation	110	dB
Spec Blocker Amplitude	-25.0	dBm
Blocker PAR	0.0	dB
Phase Noise @ 0.8 MHz Offset	-137.6	dBc/Hz
AGC Target	-1.0	dBFS
Min DSA Setting	4.0	dB
Max DSA Setting	24.0	dB

REQUIREMENTS		
PSENS	-101	dBm
SNR	6	dB

| RESULTS								
Conditions			Degradation Sources (dBm)					Margin (dB)
Blocker Level (dBm)	DSA Setting (dB)	Peak Amp (dBFS)	Peak Input (dBm)	Noise	FDD TX	PN	BLK*TX	SFDR	FDD
-25	12	-1.4	3.0	-116.6	-160.0	-109.6	-126.5	-113.6	0.5
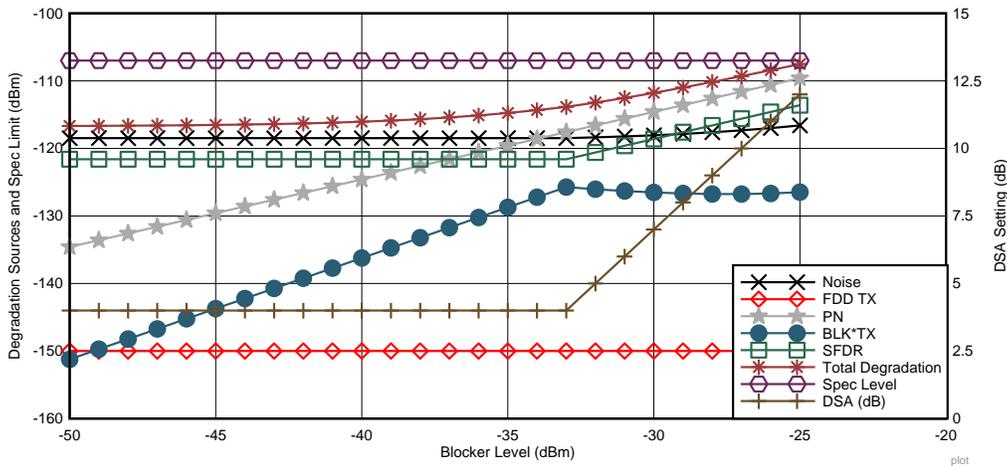
Figure 32. DSA and Degradation Power Versus Blocker Level for the MCGSM In-band Blocker Test

9   Transmitter Specifications and Analysis for LTE

9.1   TX System Architecture

The BTS transmitter requirements for RF sampling are easier to analyze and test than for the receiver. The requirements that affects the RF sampling component can be divided into the spectral mask and signal quality.

Figure 33 and Figure 34 show examples of the TX signal path. Multiple carriers are combined within a band in the digital upconverter (DUC). For a single carrier radio, the DUC is followed by a Crest Factor Reduction (CFR) block to reduce the peak to average ratio (PAR) of the signal. Following CFR, the signal bandwidth is expanded and pre-distorted to cancel the distortion in the power amplifier (PA).

For multiband systems, there are two options for how to combine the bands. Figure 33 shows a TX chain with composite DPD, where the bands are summed after the DUCs and before the CFR block. The advantage of this approach is that the interaction between the bands in the CFR and DPD blocks is accounted for, as the combined signal is used for both. For example, interband distortion cross terms that are within the expansion bandwidth can be corrected. The disadvantage is that the bandwidth of the composite signal can be significantly higher than the sum of the bandwidth for each individual band, and so more digital process resources can be needed.
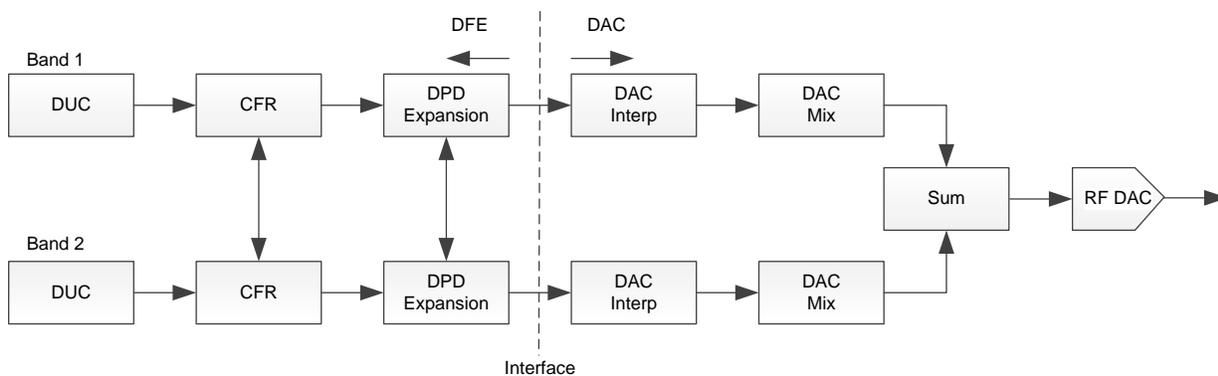

Figure 33. Multi-band TX Chain with Composite DPD

The second approach is the keep the two bands separate until after the DPD block. The advantage is that the bandwidth of the two individual bands can be less than for the composite spectrum, requiring fewer resources. However, the CFR block needs to be able to take into account the potential for peak regrowth after the signals are combined.
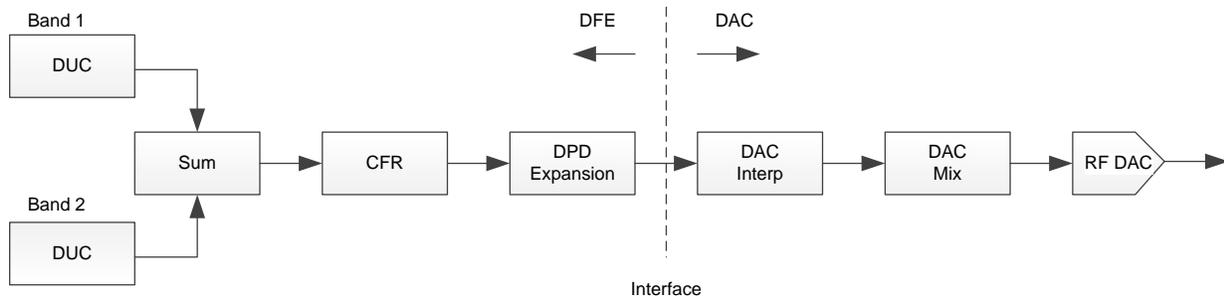
Figure 34. Multi-band TX Chain With Separate Band DPD

Generally, when bands are far apart relative to their bandwidth, the second approach is simpler to implement.

9.2 Spectral Mask Requirements

The TX chain spectral mask requirements are defined by Adjacent Channel Leakage Ratio (ACLR) for within the band and unwanted emissions outside of the band.

ACLR is defined as the ratio of carrier power to the power of the same bandwidth in either the adjacent carrier spacing for the same bandwidth or the next adjacent carrier spacing (sometimes called the alternate channel).

3GPP defines several requirements for different bands, but in general ~45 dBc for the adjacent and alternate channel is required for the entire signal chain. As most of the ACLR budget is used for the PA to be able to operate at the highest possible power, the requirement for the small signal transmitter is usually > 20 dB lower, or ~65 dBc.

While it is straightforward to simply measure the output spectrum from the small signal transmitter, it is important to understand the characteristics that can impact ACLR. Assuming an appropriate sample rate so that the second and third order harmonic distortion products fall out-of-band, the three components are output noise floor, phase noise, and intermodulation distortion.

Estimating the ACPR for a 20 MHz LTE signal from RF DAC parameters can be done as follows. The average power of the LTE signal used for testing is ~12 dB lower than the peak, so to prevent saturation the average power in the DAC would need to be < -12 dBFS. From the AFE7686 datasheet (Figure 119), the Noise Spectral Density (NSD) at 2.7 GHz for a -13 dBFS tone is -163 dBFS/Hz. For a single 20 MHz LTE signal at the same power, expect the same NSD. Integrating across the bandwidth of the adjacent (or alternate) channel, the noise floor contribution would be -90 dBFS, or -77 dBc.

The phase noise between 10 – 30 MHz offset can be estimated from the AFE7686 datasheet (Figure 146) – it is approximately -153 dBc/Hz at 2211 MHz. Adjusting for an output frequency of 2.7 GHz by 20 × log10(2700 / 2211) gives -151 dBc/Hz. Integrating across 18 MHz BW results in a degradation contribution of -78.5 dBc.

For the degradation due to IMD3, you can reuse the relationship between IMD3 and ACPR in Table 5 (the bandwidth of the LTE signal does not matter for the estimate). From the AFE7686 datasheet TX specification table, the IMD3 at 2.6 GHz with – 13 dBFS backoff (-10 dBFS total power) is -69 dBc, which is the same as the estimate of the IMD3 contribution to ACPR.

Combining all the contributions in Table 22, you can estimate an ACPR of -68 dBc at 2.7 GHz. The value listed in the AFE7686 TX specification table is -69 dBc, only 1 dB different than the estimate. Both values exceed the target of -65 dBc.

Table 22. Estimate of ACPR at 2.7 GHz from Basic Parameters

SOURCE	NOISE (dBFS/Hz)	NOISE (dBc/Hz)	INTEGRATION BW (MHz)	NOISE (dBFS/18MHz)	BACKOFF (dB)	ACPR CONTRIBUTION (dBc)
NSD	-163.0		18.0	-90.4	13.0	-77.4
Phase Noise		-151.3	18.0			-78.7
IMD3						-69.0
Total						-68.0

9.3 Signal Quality

The metric for signal quality is Error Vector Magnitude, or EVM. From 3GPP TS36.104, the most stringent EVM requirement is 3.5% for QAM256 modulation. In the system, there are other sources of EVM degradation, primarily in band noise from Crest Factor Reduction techniques for reducing the peak to average ratio.

EVM for the AFE768x is dominated by the integrated phase noise of the internal PLL/VCO. From Figure 5-153 in the AFE7686 datasheet, the integrated phase noise between 1 kHz and 40 MHz for the high VCO with CP = 5 (default settings) is -52 dBc/Hz. For a 20 MHz LTE carrier, only up to 20 MHz is needed, but most of the integrated phase noise comes from inside the PLL loop filter, so there is little error by including 20 to 40 MHz. Adjusting to 2.6 GHz by 20 × log10(2.6 / 2.211), the integrated phase noise at 2.6 GHz is 50.5 dBc, or ~0.29%.

A second source of EVM degradation is IMD3. From Table 5, the inband distortion is ~9 dB higher than the ACPR due to distortion. At 2.6 GHz, the inband distortion is therefore ~61 dBc, or 0.09%.

Combining the integrated phase noise and IMD3 degradation, the total degradation is ~0.31%. The actual measured value from AFE7686 datasheet Figure 7-3 is 0.496%.

10   5G NR cmW System Analysis

5G New Radio (NR) is the successor to LTE. 5G NR covers systems below 6 GHz with requirements similar to LTE discussed above, and a new frequency ranges called FR2 starting at 24 GHz and above (known as centimeter wave or cmW). The cmW frequency offer several GHz of new spectrum, more than all the sub-6 GHz spectrum available for cellular networks. Our system analysis is for a 800 MHz bandwidth radio at 28 GHz (specifically the US band from 27.5 GHz – 28.3 GHz).

10.1   5G NR cmW Radio Architecture

FR2 radios are two dimensional phased array antennas, and rely on beam forming to increase the radiated and received power in a specific direction to overcome the lower output power of components at cmW frequencies and the free space loss. The antenna spacing is typically one half of the wavelength, or approximately 5 mm at 28 GHz and 4 mm at 39 GHz. Arrays often use a hybrid beam forming architecture whereby the 256 Element antenna is split into a 4×64 element sub-arrays each with its with its own independent analog beam and supporting an orthogonal data stream. Figure 35 shows the system block diagram. In the diagram, the beam forming device (BFIC - Anokiwave AWMF 0157) allows phase control and provides gain for antennas, both RX and TX. Following the approach from 16 BFICs are connected by a series of four passive 2:1 Wilkinson splitters/combiners (TX/RX) which reduces to one signal, the supporting 64 antennas (following the architecture described in).

In Rx mode, the coherently combined signal from the (sub) array is amplified, filtered, and then mixed with the same low noise LO signal used for the Tx, and down-converted to an IF frequency, where it is passed through an anti-aliasing filter, and digitized by an ADC (in this case one of four in the AFE7686).

In TX mode, the DAC generates a signal at IF, which is filtered and up-converted to the RF frequency using the a low-noise LO synthesizer. The signal is then bandpass filtered at RF to remove spurious signals and used as the input signal for the coherent combining network.

To support beam forming by all 256 antennas, the same signal is generated by the AFE7686 for TX, or combined after digitization digitally for RX.
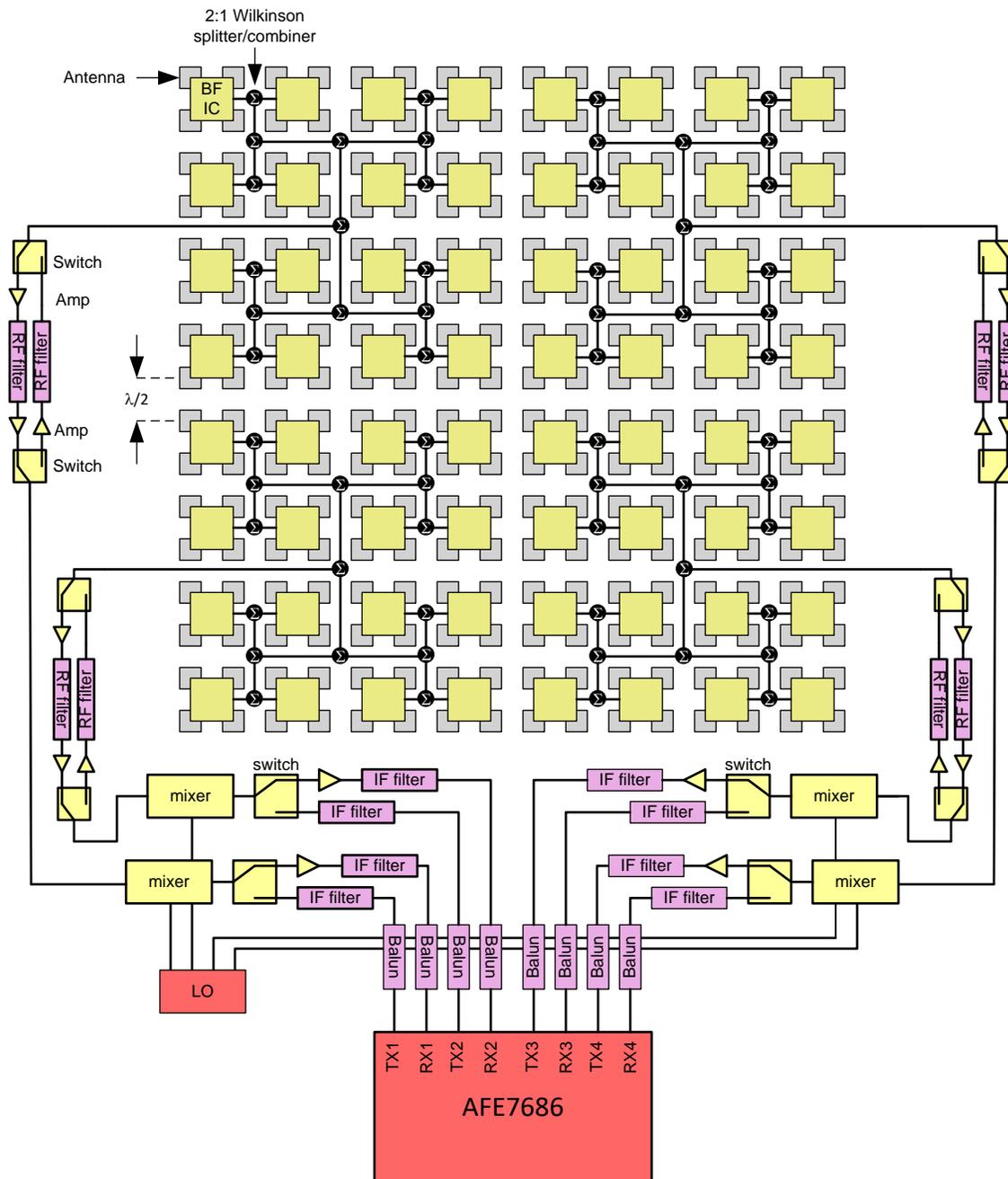
Figure 35. 5G NR cmW Phase Array Radio

10.2 5G NR cmW TX System Lineup

Figure 36 shows the TX system lineup for our analysis of a 5G NR cmW radio. The input to the AFE7686 TX are four streams, each with 12-bit resolution and a 983.04 MSPS complex sample rate. The digital signal is interpolated 9x to 8847.36 MSPS and mixed to a center frequency of 3.6 GHz. Figure 37 shows the DAC output spectrum with signal and second Nyquist zone image. The image is centered at 5.3 GHz and the edge to edge spacing between the signal and image is 900 MHz. A Nyquist filter (DEA203600BT-1240B2 from TDK) is required to reduce the image power to meet the TX spectral mask specification. The filtered IF signal is then mixed to RF. The mixer is a HMC264LC3B and the LO the LMX2594 from TI. At RF, the signal passes through a switch to allow separate TX and RX paths, is amplified and filtered with a stripline filter to remove the mixing image, and then distributed to the 64 antennas.
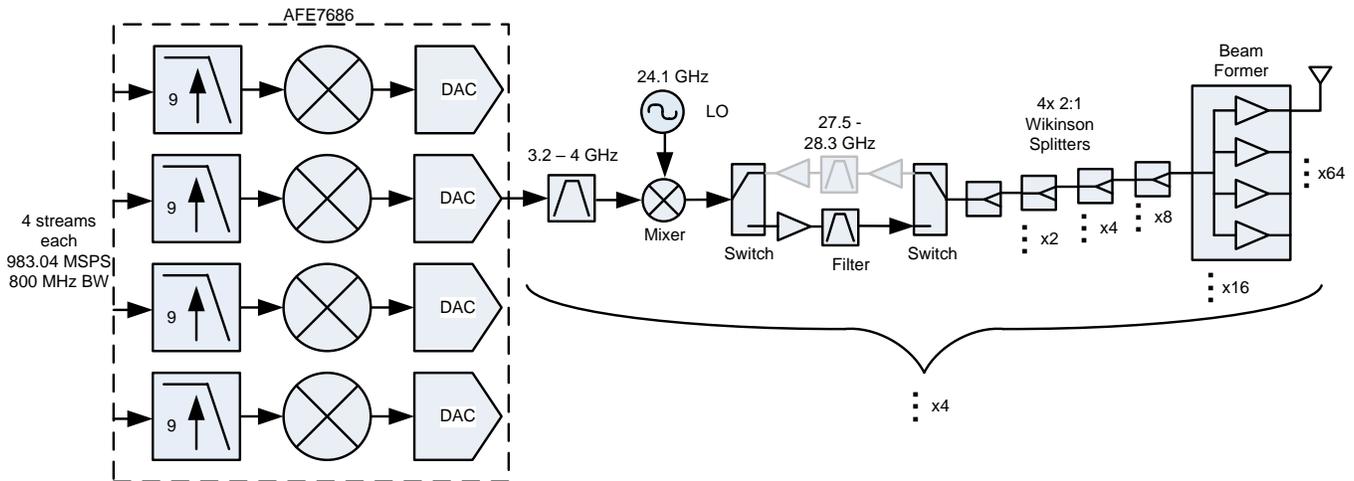
Figure 36. TX Lineup for 5G NR cmW Radio
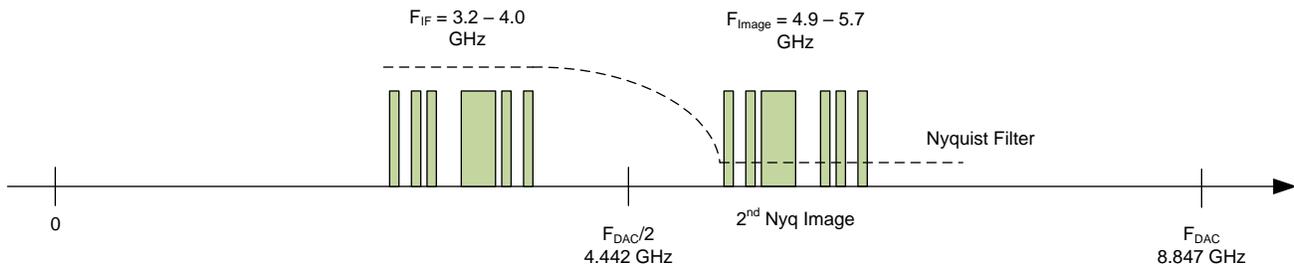
Figure 37. TX DAC IF Output Spectrum

10.3 5G NR cmW RX System Lineup

Figure 38 shows the RX system lineup got the analysis of a 5G NR cmW radio. The signal is received by the 64 antennas per sub-array. In the BFIC, the signal is amplified and the phase is adjusted so the signal from a particular direction combines coherently in the array. In the first BFIC, the signals from four antennas are summed and output as a single signal. 16 signals from 16 BFICs are combined passively in a series of four 2:1 Wilkinson combiners. The output of the combiner passes through a switch to separate a TX and RX path to allow amplification. The RX path after the switch has an amplifier, stripline filter, and second amplifier and then passes through a second switch so the mixer can be used for both TX and RX. The stripline filter provides rejection at the mixer image frequency at ~ 20 GHz. A stripline filter (see) covering 27.5 to 28.3 GHz with six elements, occupying 17×12 mm, already provides > 25 dB rejection, 1 GHz out-of-band, and 40 dB rejection at 1.5 GHz out-of-band (the closest location of the out-of-band blocker). The signal is then mixed to an IF frequency of 3.2 – 4.0 GHz. The same mixer and LO as for TX are used for RX. The IF signal is amplified, filtered and then digitized by the AFE7686 ADC with a sample rate of 2949.12 MSPS. The IF filter needs to provide noise rejection from the adjacent Nyquist zone to prevent degradation of the system NF from noise aliasing in band, and also some rejection of the out-of-band blocker 1.5 GHz offset from the band edge. The same filter used for TX (DEA203600BT-1240B2 from TDK) is used for RX, providing > 25 dB rejection at the Nyquist image and out-of-band blocker frequencies. The digitized signal is mixed to baseband and decimated by three and output at 983.04 MSPS complex with 12-bit resolution.
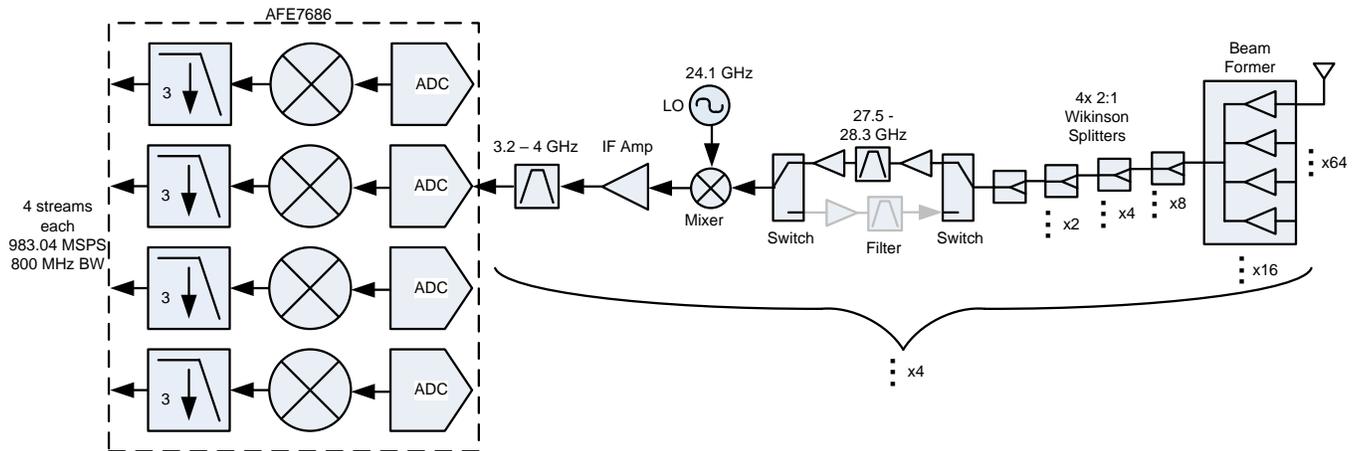
Figure 38. RX Lineup for 5G NR cmW Radio

Table 23 shows the cascaded IIP$_3$ and NF for the RX lineup. The first two columns are the coherent and electronic gain of the lineup. Due to the signal combining in the BFIC and Wilkinson combiners, signals that combine coherently (voltage sum) have an 18 dB higher gain than those that combine non-coherently (power sum). The cascaded IIP$_3$ uses the coherent gain, the cascaded NF the electronic gain. The noise figure is dominated by the antenna feed and BFIC, with the rest of the chain only adding 0.07 dB to the NF. The cascaded IIP$_3$ is dominated by the IF Amplifier.

Table 23. Cascaded IIP$_3$ and NF for 5G NR cmW RX Lineup

FUNCTION	DEVICE	CG (dB)	EG (dB)	NF (dB)	CASCADED NF (dB)	IIP3 (dBm)	CASCADED IIP3 (dBm)
	Ant Feed	-1	-1	1	1.00	100	100.00
Beam former	AWMF 0157	32	26.0	5.5	6.50	-21	-20.00
	PCB loss	-1	-1	1	6.50	100	-20.00
	2:1 Wilkinson Comb	2	-1	1	6.50	100	-20.00
	PCB loss	-1	-1	1	6.50	100	-20.00
	2:1 Wilkinson Comb	2	-1	1	6.50	100	-20.00
	PCB loss	-2	-2	2	6.51	100	-20.00
	2:1 Wilkinson Comb	2	-1	1	6.51	100	-20.00
	PCB loss	-3	-3	3	6.53	100	-20.00
	2:1 Wilkinson Comb	2	-1	1	6.54	100	-20.00
Switch	TGS4302 (Switch)	-1	-1	1	6.55	20	-20.64
Amplifier	QPA2628 (Amp)	23	23	1.6	6.57	4	-27.90
	stripline Filter	-2	-2	2	6.57	100	-27.90
Amplifier	QPA2628 (Amp)	23	23	1.6	6.57	4	-48.04
Switch	TGS4302 (Switch)	-1	-1	1	6.57	20	-55.80
Mixer	HMC264LC3B (Mixer)	-12	-12	12	6.57	12	-62.93
	IF Amp	17	17	3	6.57	13	-63.11
	Filter	-2	-2	2	6.57	100	-63.11
	Balun	-2	-2	2	6.57	100	-63.11
Data converter	AFE7686	0	0	19	6.57	17	-64.27
	Total	75	57				

10.4 RX Requirements and System Analysis for 5G NR cmW

The RX requirements for 5G cmW are in section 10 of TR 38.104 from 3GPP. The 5G NR cmW equipment is referred to as "BS type 2-O". Unlike prior specifications for 4G LTE and earlier standards that used conductive measurements (a cable can replace the antenna cable and antenna connect the equipment to the test equipment), the 5G NR cmW specifications are radiated and tested "over the air" (OTA). Therefore, the specifications apply to the entire array including antennas, not just one connection.

10.4.1 Reference Sensitivity Level

The reference sensitivity level for BS type 2-O does not have a specification, but is rather derived from a measurement of a 50 MHz channel reference measurement called $EIS_{REFSENS\_50M}$. EIS stands for equivalent isotropic sensitivity and is the sensitivity level for the array in the receiver target direction. Table 24 shows the calculation of $EIS_{REFSENS\_50M}$. The thermal noise density is integrated across 50 MHz BW. The SNR required for demodulation of the 50 MHz signal must be similar to that for LTE used above (0.25 dB). The cascaded NF of 6.57 dB is added. Finally, the array gain (log10 sum of the # of antennas) and the antenna gain and efficiency is subtracted (lowering the sensitivity level), resulting in an $EIS_{REFSENS\_50M}$ of -117.28 dBm.

Table 24. Calculation of $EIS_{REFSENS\_50M}$

Thermal Noise	-174	dBm/Hz
Log Bandwidth (Hz)	76.99	dB
SNR (QPSK, 1/3 coding)	0.25	dB
Cascaded NF	6.57	dB
Array gain	-24.08	dB
Antenna gain + Efficiency	-3	dB
$EIS_{REFSENS\_50M}$	-117.28	dBm

10.4.2 Adjacent Channel Selectivity

The adjacent channel selectivity test looks at the receiver sensitivity in the presence of an adjacent higher power interfering signal. The interferer is assumed to be in the same direction as the wanted signal, so the array and antenna gains for interferer and wanted signal are the same. The sensitivity level is allowed to increase 6 dB, and the power of the interferer is 27.7 dB above $EIS_{REFSENS\_50M}$.

The degradation mechanisms are the following:
- Noise
- IMD3
- LO phase noise
- AFE clock noise phase

As shown in Table 25, the degradation mechanisms other than noise contribute almost nothing to the overall degradation of the wanted signal sensitivity, resulting in nearly the same sensitivity level as for the reference sensitivity level.

Table 25. Adjacent Channel Selectivity System Analysis

REQUIREMENTS		
Wanted Signal Mean Power	-111.28	dBm
Interferer Power	-89.58	dBm
SNR	0.25	dBm
DEGRADATION SOURCES		
Noise	-117.53	dBm
IMD3	-140.18	dBm
LO Phase Noise (25 - 75MHz offset)	-159.21	dBm

Table 25. Adjacent Channel Selectivity System Analysis (continued)

AFE Clock Phase Noise (25 - 75MHz offset)	-164.37	dBm
Total Degradation	-117.50	dBm
Margin	5.976	dB

10.4.3   In-band Blocking

The in-band blocking test looks at the receiver sensitivity in the presence of an non-adjacent higher power interfering signal. The interferer is assumed to be in the same direction as the wanted signal, so the array and antenna gains for interferer and wanted signal are the same. The sensitivity level is allowed to increase 6 dB, and the power of the interferer is 33 dB above EIS$_{REFSENS\_50M}$ and the center frequency is offset 75 MHz from the edge of the wanted signal.

The degradation mechanisms are noise, LO phase noise, and AFE clock phase noise. As shown in Table 26, the degradation mechanisms other than noise contribute almost nothing to the overall degradation of the wanted signal sensitivity, resulting in nearly the same sensitivity level as for the reference sensitivity level.

Table 26. In-band Blocker System Analysis

REQUIREMENTS		
Wanted Signal Mean Power	-111.28	dBm
Interferer above EISRESENS_50M	33.0	dB
Interferer Power	-84.28	dBm
SNR	0.25	dBm
DEGRADATION SOURCES		
Noise	-117.53	dBm
LO Phase Noise (75 - 125 MHz offset)	-155.91	dBm
AFE Clock Phase Noise (75 - 125 MHz offset)	-160.07	dBm
Total Degradation	-117.58	dBm
Margin	5.999	dB

10.4.4   Out-of-band Blocking

The out-of-band blocking test looks at the receiver sensitivity in the presence of an out-of-band CW signal. The power of the signal is given an RMS field strength of 0.1 W/m and the sensitivity level of the wanted signal is increased by 6 dB. At 28 GHz, this translates to a power per antenna of 66.2 dBm, or 45.1 dB above the wanted signal mean power.

Given the significant frequency offset to the wanted signal, the only degradation mechanism of concern can be aliasing of the out-of-band blocker in the ADC. To prevent this, > 45 dB suppression of the output for the band blocker is required. Between the stripline RF filter and IF bandpass filter, the CW interferer at 1.5 GHz offset is suppressed by 65 dB (40 dB from the RF filter and 25 dB from the IF filter), easily meeting this requirement.

11   Summary

RF Sampling is a new radio architecture that can provide the benefit of wide bandwidth or multi-band operation with competitive performance for 2G, 4G, and 5G cellular basestations. In the future, new data converter architecture and process technology improvements will enable higher bandwidth, lower power, and higher density radios to provide smaller, lower cost and power basestations.

12 References

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