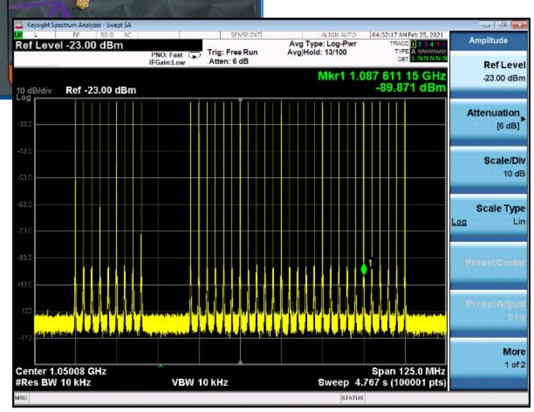
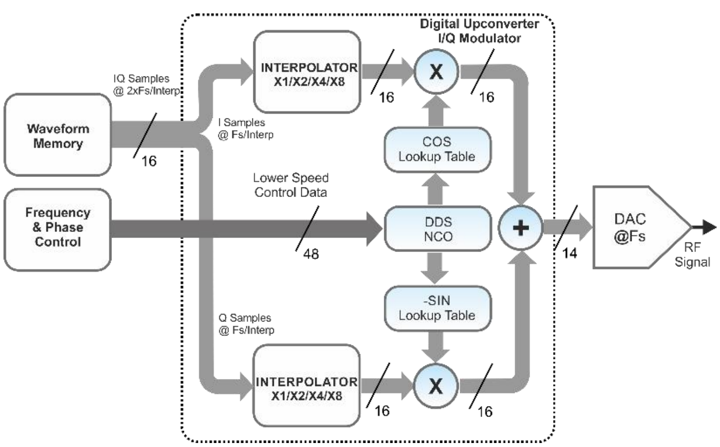
RF Signal Generation with Digital Up-Converters in AWGs

White Paper

Rev. 1.0





1 RF Signal Generation Using AWGs

Arbitrary Waveform Generators (AWG) have always been incorporated in RF signal generation systems to generate complex modulations, analog or digital. Traditionally, AWGs generated real or complex (I/Q) baseband signals to feed modulators. In particular, quadrature (IQ) modulators combined with 2-channel AWGs can generate any analog or digital modulation, provided the modulation bandwidth of the modulator and the bandwidth/sampling rate of the AWG are sufficient to faithfully generate the desired signal (fig. 1.1a). IQ modulators are very sensitive to differential responses for the I and Q signal path, no matter if they come from the AWG or the modulator. Any imbalance, quadrature, I/Q skew, etc. reduces the modulation accuracy, the available noise floor, and the usability of the generated signals. This issue grows exponentially with modulation bandwidth, so it is sometimes the most critical and costly factor for Vector Signal Generators.

As AWGs grew in bandwidth, linearity, and accuracy, a new approach was possible. Instead of generating the baseband signals, it was possible to generate an already modulated IF signal. The final RF frequency was then achieved through a mixer. Mixers require an additional component to work, a Local Oscillator Generator, and produce two sidebands. Most times one of them must be selected using a suitable BPF. As modulations are implemented mathematically, all the I/Q differential response issues disappear from the equation. However, mixers and L.O. add their own impairments such as intermodulation, conversion losses, flatness, and available modulation BW (i.e. connected to IF frequency).

The continuous advances in DAC and memory technologies have increased bandwidths and sampling rates for AWGs to the 10GHz range and beyond. This allows for the direct generation of modulated RF signals in the UHF, L, S, C and X Bands (fig. 1.1b). This approach can support extremely high modulation BW, well beyond 2GHz, and reduce the complexity and cost while improving flexibility and channel density, which is especially useful for today's radar (i.e. AESA radars) and wireless communication systems (i.e. Massive MIMO). In any case, high-quality direct RF signal generation requires a careful AWG design and waveform calculation. Proper continuous generation of a modulated RF signal requires seamless looping or sequencing of one or multiple waveforms. Modulation signals must be consistent at all levels (symbol, baseband filtering, modulation scheme) when looped and sequenced, so the modulation keeps its integrity and effects like spectral growth are avoided. For direct IF/RF generation, the integrity of the carrier must be kept as well. For a given time window (TW) there must be an integer number of cycles so the signal can be looped without any phase hit. Generally speaking, the number of cycles of the carrier must be an integer. In other words, carrier frequency must be quantized to multiples of 1/TW Hertz. This may be acceptable in some applications but not in others. "True arb" architecture AWGs can change their sampling rate with high resolution and accuracy so the carrier frequency can be adjusted further, by setting a slightly different sampling rate. However, modifying the sampling rate will result in a modification of the modulation signal as well (modulation frequency, baud rate, frequency, and phase deviation, etc). Again, this may be not acceptable in some applications. The timing and frequency accuracy for carrier and modulating signals can be improved by increasing TW (thus the number of samples for the waveform), but this leads to consumption of more waveform memory and increases the calculation and transfer times of those waveforms. An additional issue is the sampling rate requirements. For baseband signal generation, sampling rate must be higher than the modulation bandwidth (MBW). For direct IF/RF generation, sampling rate must be at least twice Fc + MBW/2, or Fc for small MBW compared to carrier frequency. A modulated RF signal, even for low modulation bandwidths, may require a huge number of samples to keep the required Time Window. Generating the same modulation at a different carrier frequency requires calculating and downloading a new waveform so the new carrier frequency (properly quantized) can be implemented.

Proteus, the new family of high-performance AWG and AWT by Tabor Electronics, has been designed to support the generation and acquisition of high-quality RF and Microwave signals using high bandwidth DACs and ADC (up to 9GS/s and more than 9 GHz usable bandwidth). This document will cover in depth how real-time digital up-conversion (or DUC) and Digital down-conversion (or DDC) is applied to improve the usability, accuracy and RF performance while offering the best-in-class modulation and analysis bandwidths while supporting full coherence and phase control over tens and even hundreds of channels.

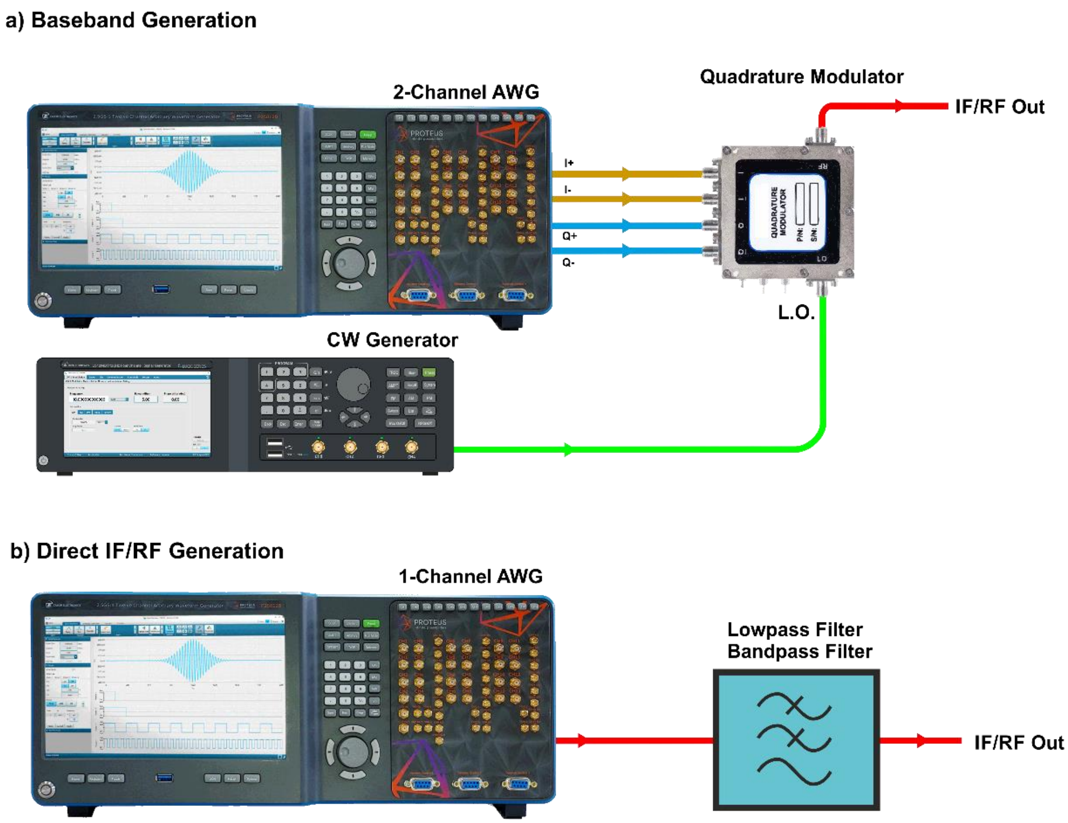


Figure 1.1: Modulated RF signal generation can be performed using AWGs. In a), the traditional IQ baseband signal generation is shown. A two channel AWG generates the two baseband components (as differential outputs in this case) to feed a Quadrature Modulator. This method requires an additional L.O generator to supply the carrier. In b) a high-speed AWG directly generates the modulated carrier using a single channel. There is no need for additional components other than filters and amplifiers.

2 Numerical Up-Conversion Using DUCs

A Quadrature (IQ) Modulator (fig. 2.1) takes a complex baseband signal (In-Phase or real part, I, and Imaginary or quadrature part, Q) and translates it from 0Hz up to Fc, or carrier frequency. Mathematically speaking, it does it by multiplying the complex baseband signal by a complex carrier:

$$SRF(t) = Real\{ (I(t) + j\,Q(t)) \times e^{j\omega t} \} = Real\{(I(t) + j\,Q(t)) \times (\cos \omega t + j \sin \omega t)\}$$

$$SRF(t) = I(t) \times \cos \omega t - Q(t) \times \sin \omega t, \qquad \omega = 2 * \pi * Fc \qquad (1)$$

In a traditional IQ modulator, a Local Oscillator produces two sinewaves with a nominal 90º phase difference (quadrature carriers) and those are supplied to two mixers along with the corresponding I and Q baseband signals. Finally, the outputs are combined, and the quadrature modulated RF signal is obtained.

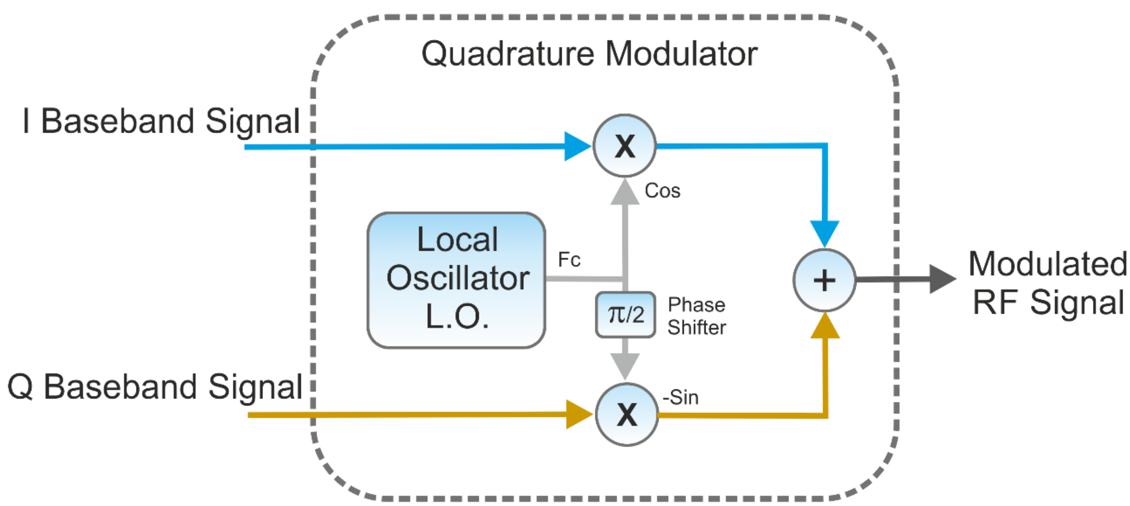


Figure 2.1: Block Diagram of an IQ Modulator. The Local Oscillator (or L.O.) can be external to the modulator itself. The quality of the signal output is influenced by the accuracy and alignment of all the signals and components.

The above process is simple to define in mathematical terms but quite difficult to implement in a practical way, especially when high carrier frequencies and modulation bandwidths are involved. A series of impairments may show up reducing the accuracy and quality of the modulation and the RF signal:

- **Quadrature Imbalance**: It occurs when the I and Q components at the mixer outputs have different amplitudes (fig. 2.2c).
- **Quadrature Error:** This impairment is caused by the lack of orthogonality between the L.O. signals applied to the I and Q mixers respectively (fig. 2.2b).

- **Carrier Feed-Through:** Part of the carrier goes directly, unmodulated, to the final RF signal, interfering with it and wasting power. It can be caused by DC offsets in the I and Q signals, the L.O. signals or by an incorrect working point of the mixers (fig. 2.2d).
- **I/Q Skew:** Differential delay between the I and Q signals becomes more important as modulation bandwidth grows.

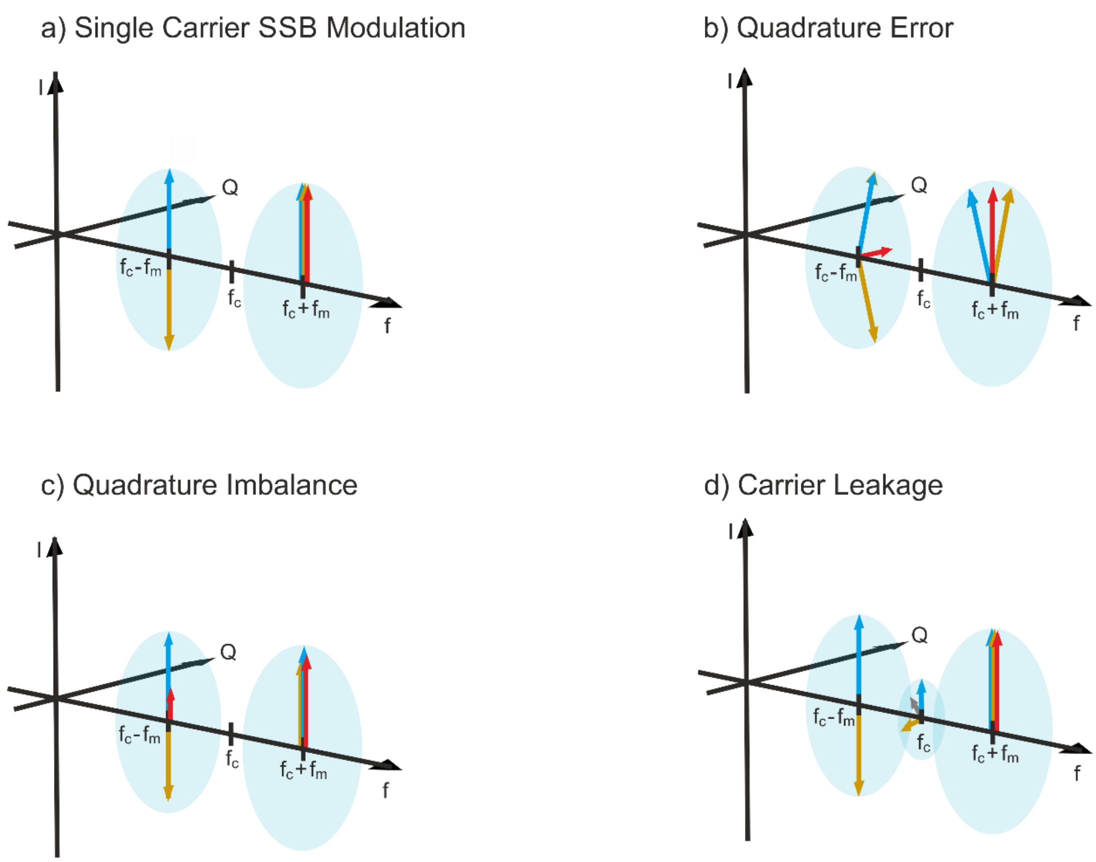


Figure 2.2: IQ Modulation can generate multiple impairments. In this case, a single sideband carrier is generated by supplying two Fm tones with 90º phase. In a perfect modulator, the right sideband is generated while the undesired sideband is nulled (a). If the relative phase of the carriers supplied to each multiplier is not 90º, the quadrature error is produced (b) and an unwanted residual carrier shows up in the opposite sideband. If the amplitude of the I and Q components is not the same, an unwanted sideband shows up as well as the nulling is not complete (c). Finally, any DC component in any of the I or Q components will show up as an unwanted tone at the Fc carrier frequency (d). Real modulators combine all the above impairments that can be a function of the Fm frequency. These are considered linear impairments. Other non-linear impairments are not shown here.

The above impairments, if moderated, can be compensated by a very careful alignment of the modulator and the I and Q signals sources. When the source is an AWG, the I and Q signals can be modified to correct, totally or partially, these impairments. However, both procedures are difficult, and impairments may drift over time, temperature, or frequency so applying them to test equipment, where conditions change from test to test, may be impractical or even not possible.

Direct generation of the modulated RF signal with an AWG removes the above impairments as waveforms are defined mathematically. Even more, impairments may be introduced in a controlled way for margin test purposes with a high level off accuracy and repeatability. Traditional AWGs can generate those signals by playing back waveforms from the waveform memory with the full modulation already implemented in it. As previously mentioned, sampling rate is linked to the carrier frequency more than to the baseband signal bandwidth. Some AWGs, though, can take a different path to solve the IF/RF generation issue. It consists in the implementation of a numerical, real-time IQ modulator, or Digital Up-Converter (DUC, fig. 2.3). In these devices, the waveform memory does not store the modulated RF signal but just the baseband waveforms, either real or complex (I/Q) depending on the modulation scheme. This architecture has important advantages over the traditional direct RF generation using AWGs:

- Carrier frequency is not set by the waveforms stored in the memory and it can be independently set without having to replace the waveforms by operating the digital quadrature L.O. (known as NCO, or Numerically Controlled Oscillator). The carrier frequency is not linked to the time window for the modulating signals anymore.
- As samples stored in the waveform memory carry just the baseband information, bandwidth and sampling rate requirements are set basically by the desired modulation bandwidth. The sampling rate for the baseband waveforms and the final sample rate for the DAC must be adapted, though. This operation can be performed using real-time interpolators.
- As traditional direct RF signal generation, this architecture does not suffer from any impairments as described above.
- Multiple DUC blocks can be combined into a single feed to any of the DACs in the AWG, so more than one carrier with any desired carrier frequency can be generated simultaneously.

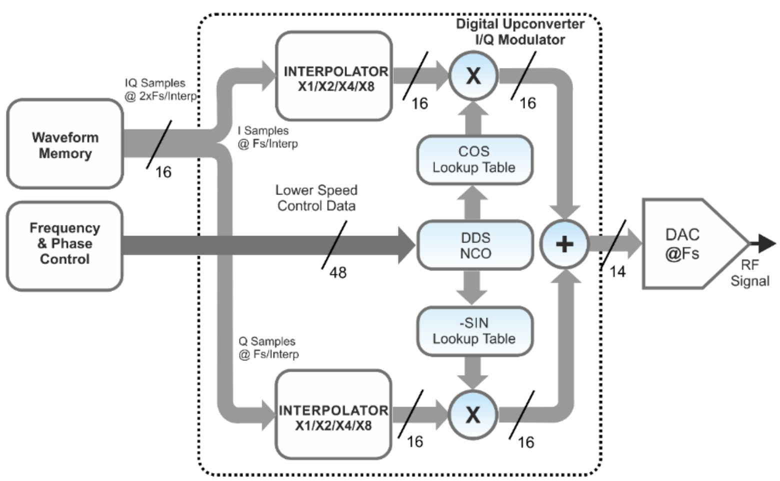


Figure 2.3: Block diagram of the DUC implementation in the Tabor Proteus AWG. Two of these blocks are implemented for each channel. The fully numerical operation removes all the sources for impairments while the usage of interpolators results in an important saving in terms of waveform memory and data transfer rate. The carrier frequency and phase can be changed without recalculating and downloading new waveform data.

The NCO block

A very important component of a DUC is the quadrature NCO (fig 2.4). It can be implemented in different ways so the final carrier frequency is synthesized. Analog L.O. may use a PLL based synthesizer to define carrier frequency. Such synthesizers offer a great deal of flexibility, accuracy, and resolution. However, frequency switching times are influenced by the bandwidth of the closed loop control in the PLL. There may be a trade-off between switching time and phase noise performance. The behavior of the L.O. during the switch may be difficult to predict and random.

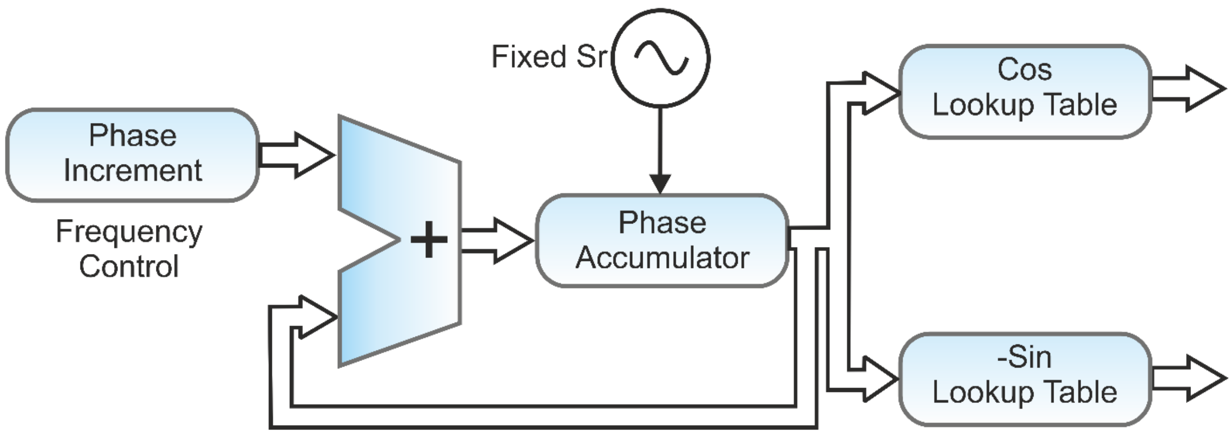


Figure 2.4: A quadrature NCO is the numerical Local Oscillator in a DUC. IQ modulation requires two carriers with a 90º phase difference. In a quadrature NCO, a perfect 90º phase can be obtained by using two lookup tables from the same DDS synthesizer output. Frequency is controlled by the phase increment added for every Sample clock period. Initial phase can be controlled by setting up the initial content of the Phase accumulator. Phase and frequency resolution depend on the size (in bits) of the Phase Accumulator.

Although such a synthesis scheme can be also implemented digitally, NCOs in DUC are typically based on the DDS (Direct Digital Synthesis) architecture. A DDS generates a numerical sinewave by using a phase accumulator and a lookup table. Basically, for every sampling clock, a given number is added to the phase accumulator controlling the frequency. The initial value for the accumulator controls the phase of the sinewave. The value in the accumulator represents the instantaneous phase of the synthesized sinewave, so the corresponding amplitude is read from the lookup table. In a quadrature NCO, two lookup tables are implemented and accessed by the same phase word from the accumulator.

One contains the amplitude values corresponding to the Cos signal and the other one the values corresponding to the -Sin signal. The frequency of the sinewave can be set with a very high resolution according to the size of the phase accumulator. Fc is set according to the following expression:

$F_c = CW * F_{DAC} / 2^{RES}$, CW = Control Word, RES = size of the accumulator in bits (2)

$F_{RES} = F_{DAC} / 2^{RES}$

Using the actual figures from the Tabor Proteus:

RES = 48 bits

FDAC = 9×10^9 Hz

Fc = 0 Hz ... 9×10^9 Hz

FRES = $9 \times 10^9 / 2^{48} = 32 \times 10^{-6} = 32\ \mu$Hz

One of the advantages of NCOs based in the DDS architecture is that frequency switch is instantaneous (from one sample to the next) in a phase-continuous manner so switching glitches are not generated. The size of the lookup tables is limited so some rounding takes place when converting the phase accumulator contents to a given entry in the table. The size (resolution) of the entries themselves in the table is also limited to the size of the multiplier or DAC attached to it. The resolution in the time and amplitude domains of the lookup table is chosen so any impairment (i.e. spurs) introduced by the rounding processes taking place (such as the phase noise coming from the limited number of entries), is negligible in respect to other sources of impairments, such as quantization noise, or the Sampling Clock (Sclk) phase noise.

The frequency of the output sinewaves can be chosen from DC up to the sampling rate. Traditional AWG generation can reproduce signals with frequency components between DC and half the sampling rate (Nyquist Sampling Theorem), called the first Nyquist Zone (or NZ). However, images are produced around multiples of the sampling clock. Each FDAC /2 wide section of the spectrum is called a Nyquist Zone and is numbered depending on its frequency location. Images located at these upper order Nyquist zones can be used, sometimes by filtering out the unwanted images including the one in the NZ #1. As frequency response of the DAC falls with frequency, not all the images can be effectively used for practical purposes. Typically, the second and sometime the third Nyquist Zones can be used if the analog bandwidth of the DAC and the output stage are sufficient, despite the zeroth-order hold response (sin Af / Af with zeros at all multiples of FDAC) of ideal DACs. One way to select the right carrier frequency for the NCO in a higher order Nyquist Zone is selecting the following Fc:

Fc = abs(Fc' – (n – 1) x FDAC), n = NZ #, Fc' = target Carrier Frequency, n >= 2

For even numbered Nyquist zones, the spectrum of the images will be reversed. If the application required preserving the original, non-inverted spectrum, then the complex baseband signal (I/Q) must be replaced by its complex conjugate (by reversing one of the components). However, if the DDS allows for Fc higher than FDAC / 2, it is better to set up that frequency directly in the CW. The subsampling of the NCO output in respect to the target Fc' results in the reversion of the sign of the sin(x) lookup table for the Fc' frequency, so the spectrum around the Fc frequency will be reversed, and the one located in the odd numbered NZs will be right. In this way, baseband data can be preserved unmodified, regardless of the NZ being targeted (fig. 2.5).

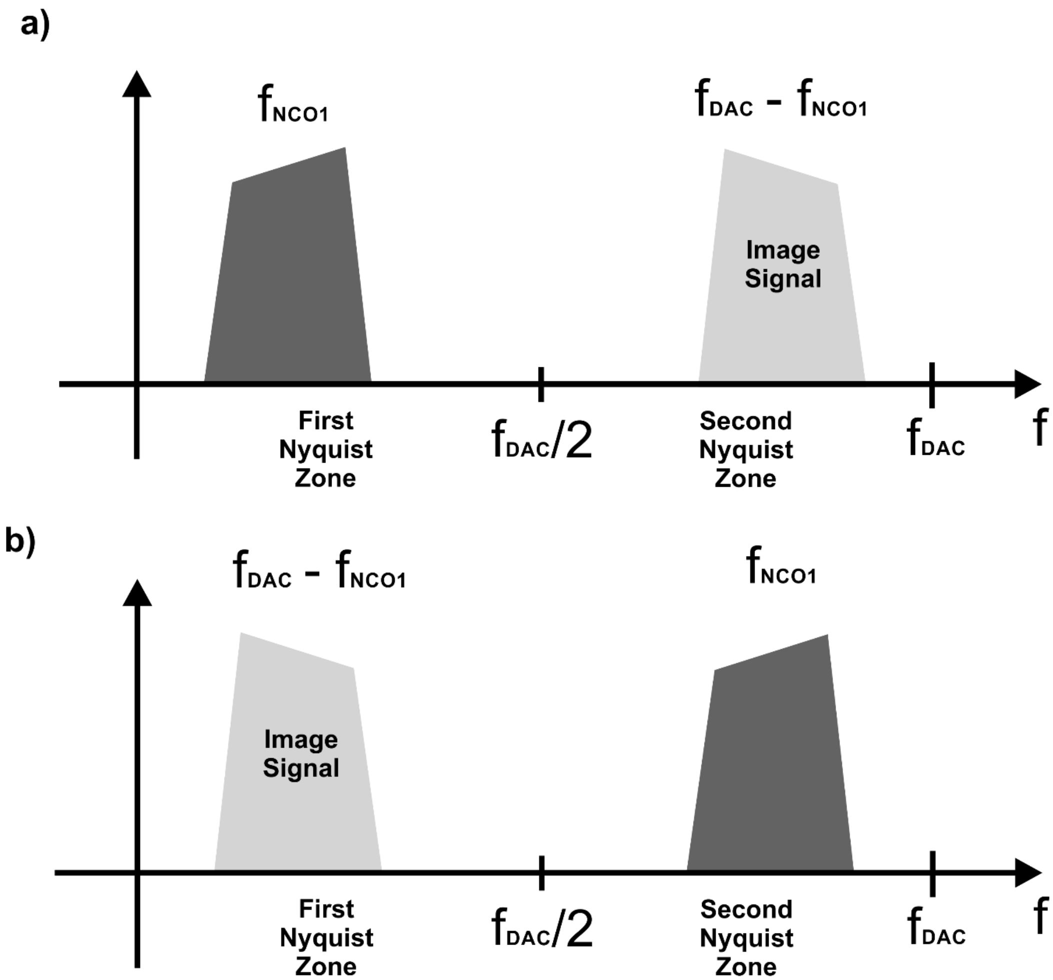


Figure 2.5: The NCO can be set from DC up to the sampling frequency. The resulting modulated signal will include the corresponding images resulting from the sampled nature of the waveforms. When generating a signal in the second Nyquist band, the NCO can be set to the image frequency in the first Nyquist band. However, the spectrum of the modulated signal will be reversed in the second Nyquist band. Although this problem can be fixed by reversing one of the baseband components, it is much better to set-up the NCO frequency at the carrier frequency in the second Nyquist band. In this way there is no need to reverse and update one of the IQ components.

Interpolation

Using the DUC architecture opens the door to separate the sampling rate of the baseband data from the final sampling rate of the DAC. The sampling rate of the complex baseband data must be higher than the modulation BW. A 100MHz modulation bandwidth complex baseband signal could be made of two 50MHz bandwidth signals that should be sampled, at least, at 100MS/s each. However, the DUC must operate at the final DAC sample rate. This means that the I and Q sampled waveforms must be resampled (typically upsampled) before reaching the multipliers. It is extremely convenient for practical purposes that the ratio between the sampling rate of the DAC, and the sampling rate of the baseband waveform is an integer, N. One simple way to upsample the baseband waveforms could be keeping the same sample value for N samples. However, this method would reproduce the images in the original baseband sample waveforms, and they will show up as unwanted sidebands in the modulated signal. In order to avoid that, a near ideal interpolation process must be applied before reaching the multiplier (fig. 2.6). Ideal interpolation cannot be carried out in the real World, especially if it has to be applied using real-time signal processing. Practical interpolation consists in the upsampling of the incoming signal using a zero-padding process first (by adding N-1 zero samples between actual samples) and then applying a powerful digital low-pass filter using a linear phase response LPF FIR to remove the unwanted sidebands from the interpolated waveform. Practical FIR filters show some roll-off, so the actual modulation BW supported depends on the size of it. 80 to 90% of the theoretical maximum modulation BW are typically supported in real implementations.

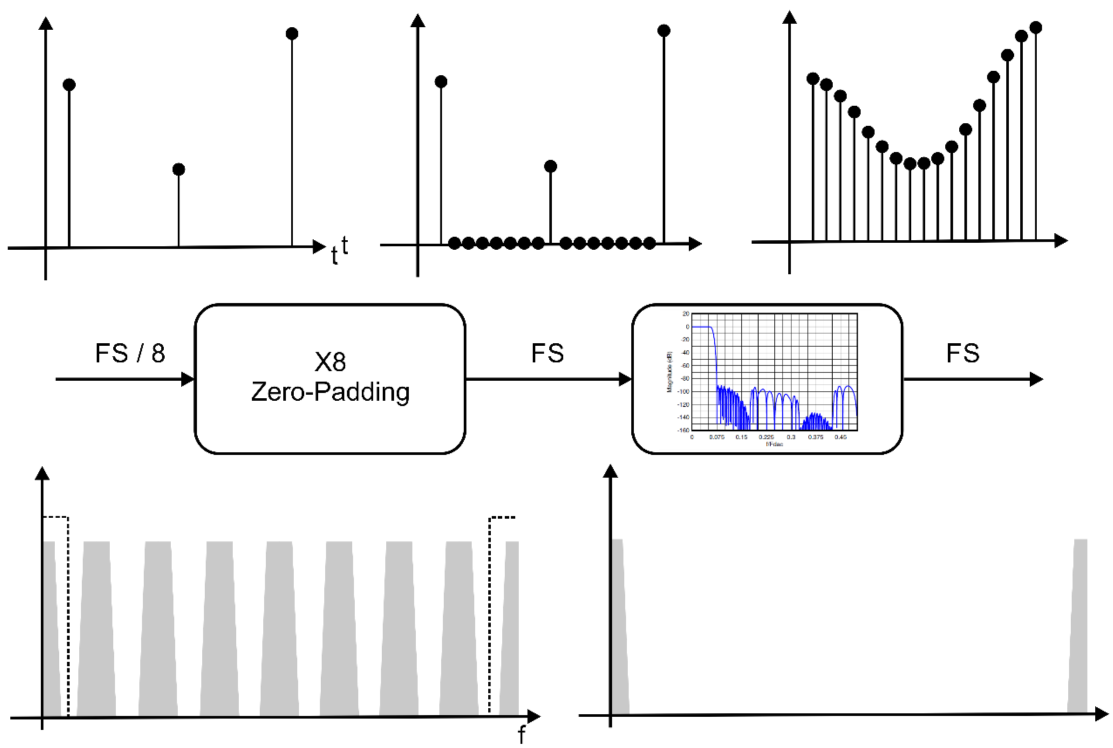


Figure 2.6: Interpolation is a very important factor for Digital Up-Conversion. Interpolators increase the sample rate through a zero-padding process. This process, though, keeps the unwanted images of the signal sampled at a lower speed. A real-time Low-Pass FIR, or interpolation filter, adds the intermediate samples while removing the images above the original first Nyquist zone. Here, the X8 interpolator implemented in the Tabor Proteus AWG is shown.

The Proteus family of AWGs supports multiple interpolation factors (x2, x4, x8) so the sampling rate of the incoming signal can be reduced according to the actual modulation bandwidth requirements and the final FDAC. The FIR filters applied are optimized for each factor as the number of available taps grow with the interpolation factor. Interpolation reduces the size of the waveform in a factor equal to half the oversampling as two samples (I & Q) per sampling period are required.

Resources & Contact

For more information on Microwave signal generation challenges and solutions, review the following resources:

- White Paper: Multi-Nyquist Zones Operation-Solution Note
- White Paper: Direct Generation/Acquisition of Microwave Signals
- White Paper: Effective Number of Bits for Arbitrary Waveform Generators
- White Paper: Multi-Tone Signal Generation with AWGs
- Solution Brief**:** Envelope Tracking – Solution Note
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