



PRIMER

IQ/IR/SSB Mixer



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Wireless communication and radar systems are under continuous pressure to reduce size, weight, and power while increasing dynamic range and bandwidth. The quest for higher performance in a smaller package motivates the use of IQ mixers: mixers that can simultaneously mix ‘in-phase’ and ‘quadrature’ components. System designers use IQ mixers in RF and microwave systems to eliminate or relax the requirements for filters, which are typically large and expensive components. IQ mixers (sometimes called complex mixers) use phase manipulation to suppress signals rather than filters.

The goal of this white paper is to introduce IQ, single sideband (SSB), and image reject (IR) mixers in both theory and practice. It will discuss the differences between IQ, SSB, and IR mixers in addition to the operation, merits, specifications, limitations, linearity, and compensation of these mixers. Marki Microwave® uses passive diode mixer technology. The concepts presented in this paper are therefore focused primarily on passive diodes but are generally applicable to other IQ mixers and modulators.

I. MERITS OF IQ MIXERS

Communication systems look different using a traditional mixer, an SSB mixer, and an IQ mixer. The task is to upconvert two baseband signals $a(t)$ and $b(t)$ to be transmitted within limited available bandwidth and without transmitting in adjacent bands. The two signals could be separate channels or the same information broken into two streams. A key mixer challenge is that a standard mixer will create both upper and lower sidebands with redundant information in them. To create a single sideband signal using a normal mixer, users upconvert the signal and then use an image rejection filter to remove the redundant sideband from the signal, as shown in **Figure 1a**.

The separation between the desired signal and the unwanted signal is equal to the input signal frequency, requiring an expensive filter with a very sharp cutoff. Either increasing the input signal frequency at the cost of a more expensive and power-hungry baseband circuitry, or using a superheterodyne architecture, can ease filter requirements.

An upconversion using a single sideband mixer (**Figure 1b**) achieves the same result, but with no filter. There are many complications to the actual implementation, but as shown, it is possible to accomplish the same task by manipulating the phase of the signals that would otherwise require a filter. Consider a more complicated example, where users fill the available bandwidth directly using an IQ mixer, with no filtering or suppression (**Figure 1c**).

Using the IQ mixer has some important differences over the previous examples. Instead of putting $b(t)$ into a higher bandwidth channel on the same IF input, it is put in the same channel bandwidth, but on a separate IF input “Q”. The local oscillator (LO) is moved from the edge to the middle of the available bandwidth. In this case, both signals are converted using a double sideband upconversion into the same RF bandwidth, and neither sideband is suppressed. However, the two data signals have a phase difference, so they can be separated at the receiver provided both sidebands are transmitted. Again, no filter is required, and in this case, the IF circuitry has half the bandwidth as in the single sideband upconversion example. There is an additional, similar example to consider. A single sideband mixer incorporates an IF hybrid coupler (implemented digitally or in analog circuitry). If the ports of this

coupler are used simultaneously, then the upper and lower sideband can be generated simultaneously, as shown in **Figure 1d**. This has the benefit of low bandwidth IF circuitry and does not require phase discrimination in the receiver.

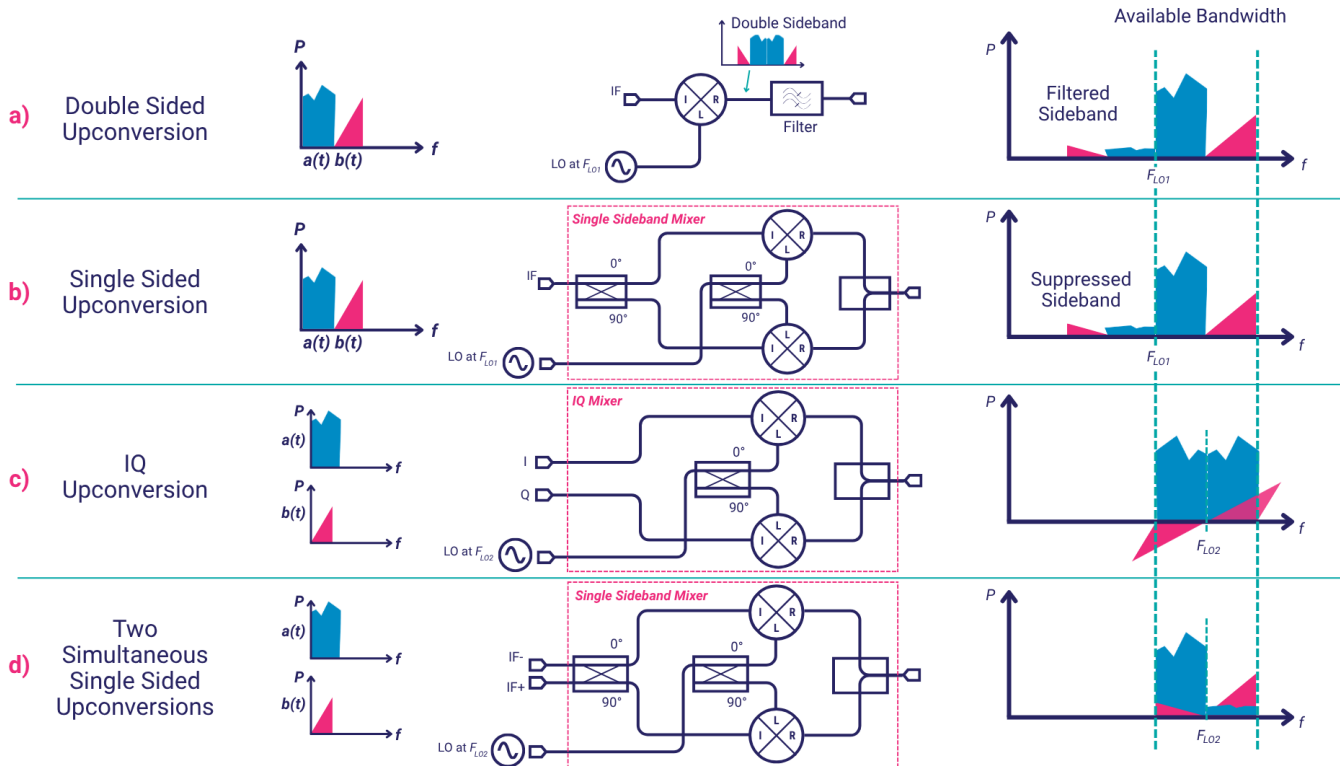


Fig. 1: Circuit structures of different types of mixers, including a) double sided upconversion, b) single sided upconversion, c) IQ upconversion, and d) two simultaneous single sided upconversions.

The following sections examine the internal circuitry of the IQ and single sideband mixer, and their downconverting equivalent, the image reject mixer. This section discusses the practical implementations of this circuitry and the implications to the performance of a microwave system.

II. OPERATION OF IQ, IR, AND SSB MIXERS

The fundamental principle behind IQ, IR, and SSB mixers is signal cancellation through phase manipulation. These mixers create two copies of the desired signal that are in phase with each other and two copies of the undesired signal that are out of phase with each other. When these are combined, the undesired signal is canceled, leaving only the desired signal. This is analogous to how a double balanced mixer creates isolation by creating in-phase and out-of-phase copies of the LO and then combining them; however, it is done with quadrature hybrids instead of baluns. The path of the desired sideband in a single sideband mixer demonstrates how this concept works. The structure of a single sideband mixer is shown in **Figure 2a**, and the resulting relative signals are shown in **Figure 2b**.

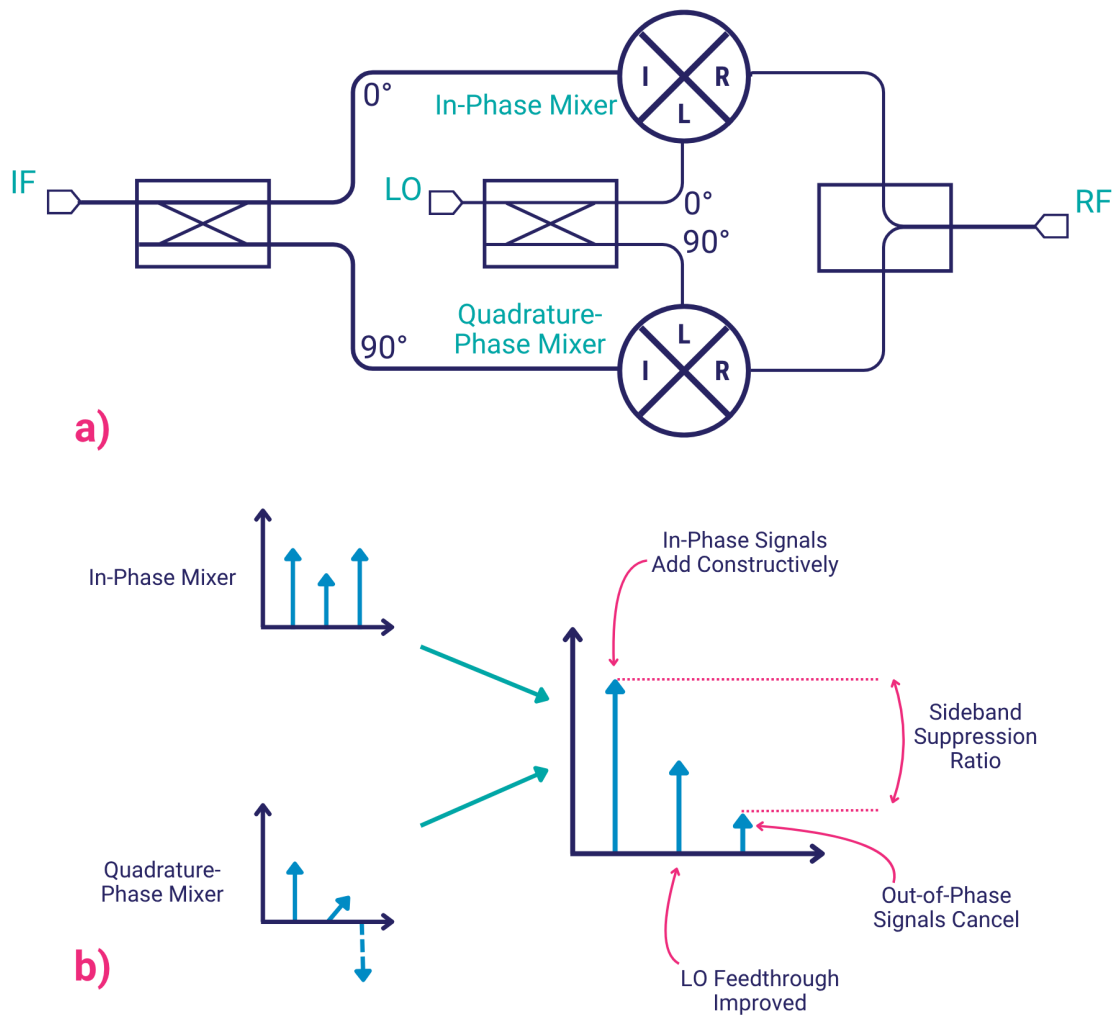


Fig. 2: a) Circuit structure of a single sideband mixer and b) resulting relative signals.

To create a single sideband, first the IF data signal is split into two copies with a 90° phase shift between them. The LO is also split into two copies with a 90° phase shift between them; one copy is applied to the “in-phase” mixer and the other copy to the “quadrature phase” mixer. This is the critical step since the two sidebands inherit the phase of the LO differently. The frequency (and hence the phase) of the upper sideband is the sum of LO and IF frequencies, while the frequency (and phase) of the lower sideband is the difference of the LO and IF frequencies. Thus, the phases add for the upper sideband and subtract for the lower sideband. In the in-phase mixer, both sidebands have a 0° phase shift. In the quadrature mixer, the two 90° phase shifts add to become a 180° phase shift on the upper sideband and subtract to become a 0° phase shift on the lower sideband. When the signals are recombined in the in-phase power combiner, the upper sidebands will cancel, and the lower sidebands will add together.

Several important things to note:

- 1) Two quadrature shifts can be applied in many combinations to create signal cancellation, leading to IQ and IR mixers.
- 2) Dynamic range of the circuit is limited not only by image rejection or sideband suppression, but also by LO isolation and other spurious products defined by other figures of merit. These other isolations and $m \times n$ spurious products are all affected (typically improved) by the way they inherit 0° , 90° , or 180° phase from the structure.
- 3) Each of the functional blocks can be realized in various ways (including digitally). This leads to numerous permutations on the basic concept depending on desired system tradeoffs.
- 4) Bandwidth can theoretically be limited by the quadrature hybrid on the IF, the quadrature hybrid on the LO, the power divider, or the mixer. Practically, it is almost always limited by the quadrature hybrid.
- 5) Image rejection is limited by the phase and amplitude balance of the circuits.

III. DIFFERENCES BETWEEN IQ, IR, AND SSB MIXERS

The previous section described the operation of a single sideband mixer, which upconverts a signal to a single sideband while suppressing the other sideband. The downconverting equivalent is an image reject mixer, which downconverts one sideband while rejecting the undesired (or 'image') sideband. The circuit structure (identical but operated reciprocally from the single sideband mixer) is shown in **Figure 3**.

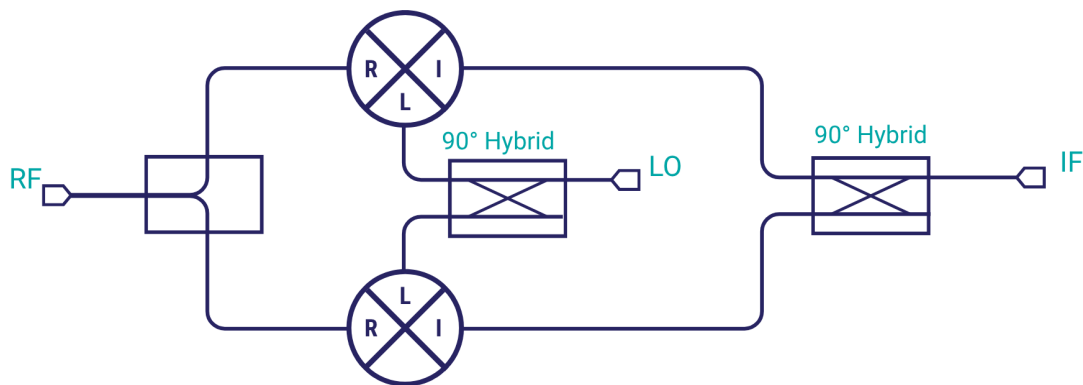


Fig. 3: The circuit structure of an image reject mixer.

Similar to the single sideband mixer, the IR mixer functions on the principle that the phases add differently when converting RF to LO and converting LO to RF. Even if no signal is present at the image frequency, the image reject mixer can improve system performance by rejecting the noise power in the image frequencies.

The IQ mixer is similar in structure but eliminates the IF hybrid, as shown in **Figure 4**. Each phase of the RF signal can be recovered separately from the I and Q ports at the receiver, or a different signal can be applied to the I and Q ports for transmission.

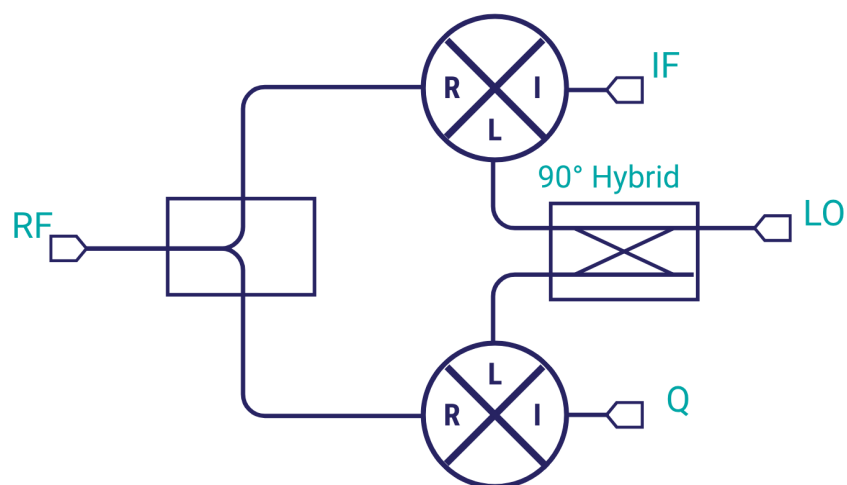


Fig. 4: Circuit structure of an IQ mixer.

One input is converted with an in-phase LO (the I input) and one with a quadrature phase LO (the Q input). These are combined at the RF port for broadcasting. At the receiver, a second IQ mixer can demodulate the I and Q signals separately. This only works, however, if both sidebands are transmitted and the LO has the same relative phase as the transmitting LO. If one sideband is filtered, cancellation will not occur. If the LO is not phase locked, the undesired signal will not be suppressed. This illustrates a weakness of IQ modulation relative to single sideband modulation, in that the phase of the carrier must be recovered in order to separate the I and Q signals.

Figure 5 shows that two quadrature phase shifts are combined to get signal cancellation through a positive and negative version of the undesired signal. There are other ways to achieve this goal (through a quadrature hybrid on the RF port instead of the LO, for example), but there are practical reasons related to the subcomponents of the IQ mixer that make these the most common. It is important to understand key optimizations for an IQ/IR/SSB design before discussing these practical reasons.

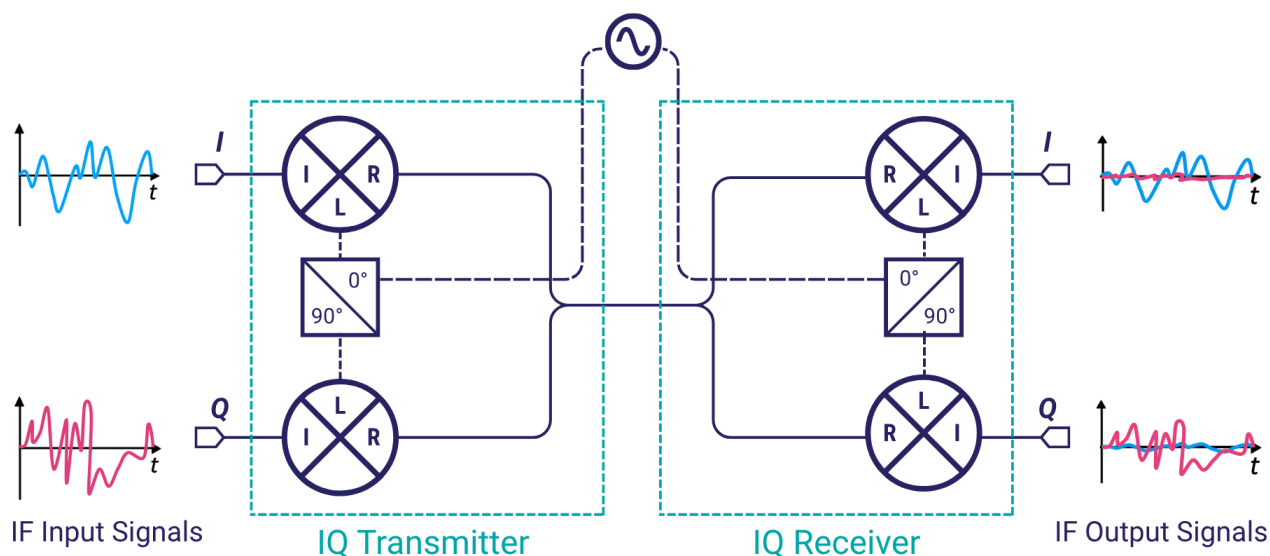


Fig. 5: Combined quadrature phase shifts.

IV. FIGURES OF MERIT

An IQ, IR, or SSB mixer is subject to the linear and nonlinear mixer figures of merit discussed in the [Marki Microwave Mixer Basics Primer](#) including conversion loss, single tone intermodulation distortion, multi-tone intermodulation distortion, VSWR/return loss, P1dB, and isolations (LO to I/Q, RF to I/Q, and most importantly LO to RF).¹

Additional figures of merit measure how close the matched quadrature components are to being ideal. There are two ways to express this. One is to measure the I/Q amplitude and phase balance independently. This is measured by downconverting a signal from the RF port and measuring how close the I/Q ports are to being equal amplitude (amplitude balance) and 90° out of phase (phase balance) across the bandwidth of each of the ports.

The second method is to measure the image rejection or sideband suppression ratio (**Figure 6a**), expressed in dBc. This is the ratio of the desired sideband power to the undesired sideband power, and it combines the effect of amplitude balance, phase balance, and conversion loss simultaneously. This is ultimately the main concern of the system designer; it expresses the ratio of the desired signal to the next highest nearby signal. A sideband suppression of 20 dBc is easy to achieve at narrow bandwidths, and 40 to 50 dBc is routinely achieved for tuned/calibrated single frequency designs. It is very difficult to achieve sideband suppression of greater than 15 dBc in an untuned, wideband (>5:1 ratio) analog design. The sideband rejection (sideband suppression minus the conversion loss) is determined by the amplitude and phase balance of the system, as illustrated in **Figure 6b**.

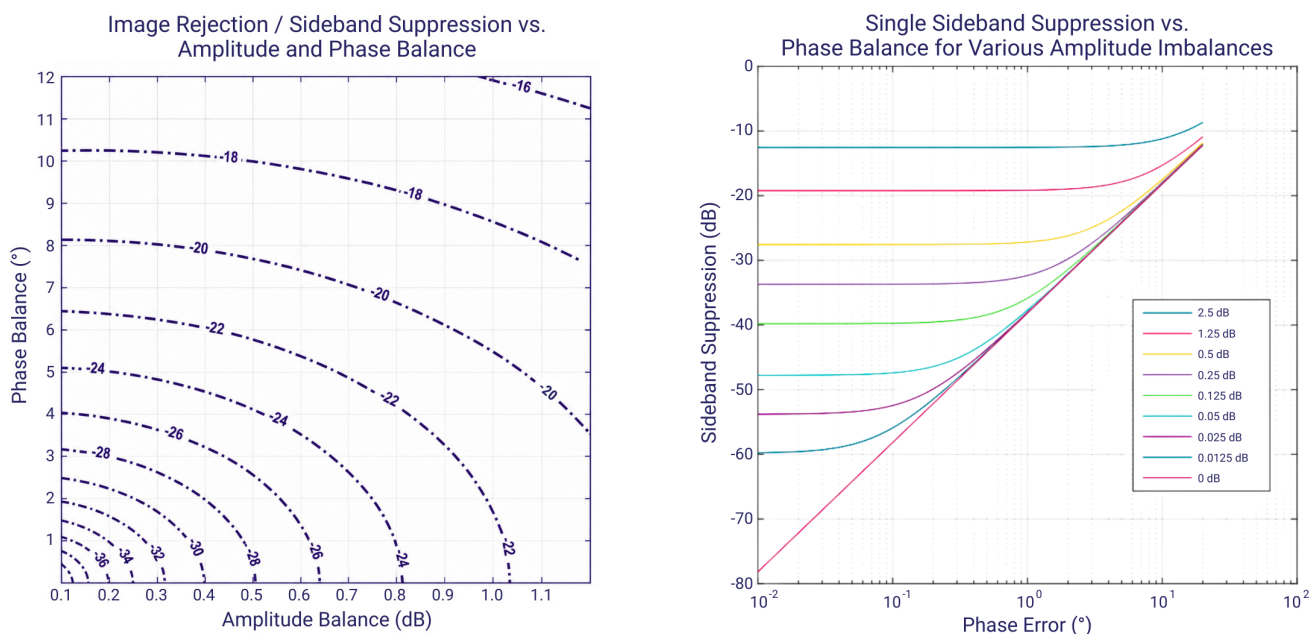


Fig. 6: Graph of a) sideband suppression ratio and b) sideband rejection.

An interesting aspect of sideband suppression (or image rejection) is that it results from the net phase or amplitude balance. This means that if the phase balance on the LO quadrature signal generator is off by some error θ , it can be corrected by applying a phase differential of $-\theta$ between the I and Q ports of a single sideband mixer. This technique is used frequently in IQ communications systems, as errors in the IQ mixer can be corrected digitally.

In addition to the sideband suppression, the dynamic range of a system is limited by the isolation or spurious suppression. The $2 \text{ IF} \times 1 \text{ LO}$ spur and the LO feedthrough will both be separated from the desired tone by the IF frequency while the sideband is separated by twice the IF frequency; therefore, both can limit the dynamic range even if the sideband is adequately suppressed. The bleed through of the LO signal is frequently higher in power than the suppressed sideband in a single sideband upconversion scheme, so filter requirements may not be reduced with a single sideband architecture without excellent LO to RF isolation, as demonstrated in **Figure 7**.

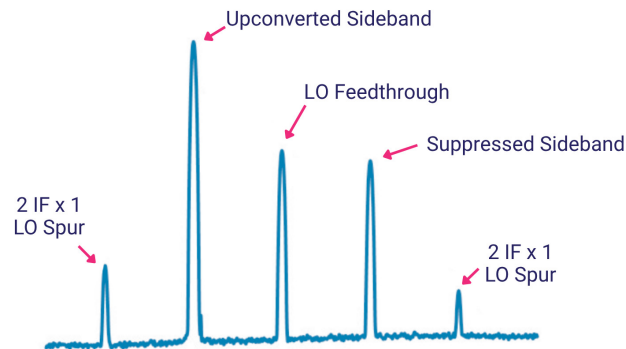


Fig. 7: Realistic single sideband upconversion spectrum.

Each subcomponent in the IQ/IR/SSB architecture can be implemented and impact the system dynamic range differently.

V. HARDWARE LIMITATIONS

The optimal implementation of an IQ/IR/SSB mixer depends on the desired system goals, particularly with respect to size, bandwidth performance, and dynamic range (usually inversely related to each other and cost/size). It also depends on what form factor or production method is available for realization. Potential form factors include connectorized modules, surface mount, chip and wire assembly, and integration in an integrated circuit. The optimal implementation also depends on system economics, including required system cost and production volume. All functionality of an IQ/IR/SSB mixer can be realized for extremely low cost in a tiny package using either digital signal processing or analog CMOS circuits. This typically comes at a cost of lower operation frequency, narrower bandwidth, lower linearity, and higher development costs. For broadband, high dynamic range microwave frequency application, the standard approach uses passive diode mixers. **Figure 8** and the following discussion focuses on the subcomponent tradeoffs for achieving broadband, high dynamic range IQ/IR/SSB mixers in a small, repeatable, planar package. **Table 1** summarizes the information.

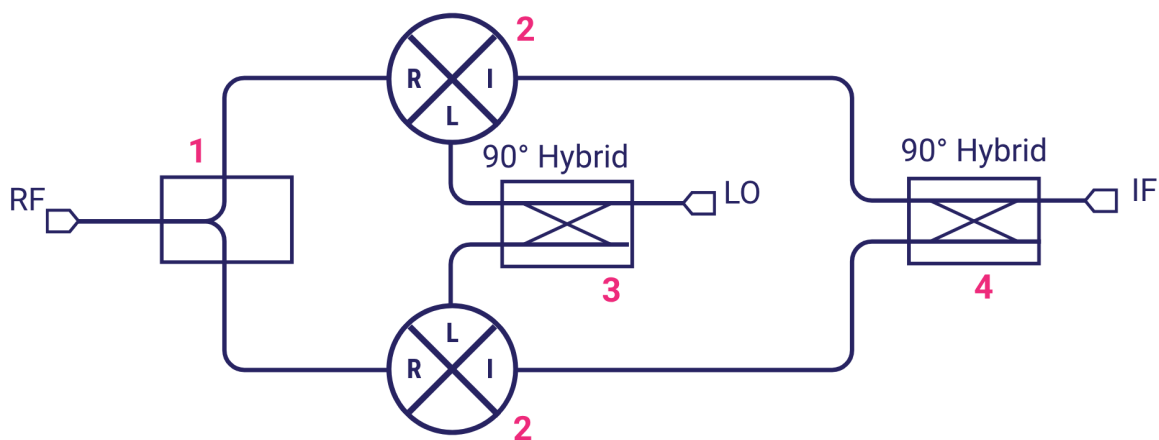


Fig. 8: Circuit design for high dynamic range IQ/IR/SSB mixers.

1. High Frequency Power Divider: The power divider is the most straightforward subcomponent to realize. IQ mixer operation requires excellent phase match and low loss across the RF operating band, and isolation is desirable to reduce spurious products. A resistive power divider is possible but has high loss and no isolation. A reactive tee can be used as a splitter; however, it limits the bandwidth and provides no isolation in an upconversion (which can create challenges). The [Wilkinson](#) power divider provides the highest-performance solution with low loss, excellent phase matching, and good isolation, but requires more circuit area.

2. Matched Mixers: Options for mixers are abundant (single diode, balanced FET, Gilbert Cell, triple balanced, etc.), but for most microwave applications, the best choice is the [double balanced diode mixer](#). This mixer construction offers high isolation and spurious rejection, high P1dB, excellent phase delay and amplitude balance repeatability, and single ended operation on all three ports. It also has a DC IF capability that allows compensation for the LO-RF isolation. The more closely the mixers are matched, the better the sideband suppression and balance will be. Thus, MMIC mixers with nearly identical diodes and passive structures are strongly preferred for this application. Furthermore, this mixer can be planarized to provide a compact form factor.

As mentioned above, the output of an IQ or single sideband mixer will be limited by the LO-RF isolation and the 2 IF x 1 LO (and possibly 3 IF x 1 LO) spurious products. The dynamic range (or bandwidth) of an image reject mixer can be limited by the 2 LO x 2 RF spur, which will be at twice the IF frequency. These performance limitations are inherited directly from the mixer core used in the IQ structure, so it is critically important to select the best possible mixer for this application. Additionally, by selecting an appropriate diode level, it is possible to tradeoff linearity and LO feedthrough. Therefore, it is important to understand whether spurious products, LO feedthrough, or the signal sideband will be the limiting factor to dynamic range. For more on linear specifications, see [Section 6](#): Linearity in IQ/IR/SSB Mixers.

3. LO Quadrature Signal Generation: There are several options for LO signal generation, and the preferred option depends on the application. Phase balance is the critical parameter for LO signal generation,

as it directly affects sideband suppression. Mixer performance metrics are only weakly affected by LO power, so the LO drive can be significantly unbalanced without a direct penalty to the sideband suppression. Uneven drive, however, will lead to duty cycle (i.e. phase) variations that will degrade sideband suppression.

For a single frequency LO, a power divider with a simple phase delay (either a length of transmission line or a phase shifter) can be used to create a nearly perfect 90° phase delay, provided the frequency of operation is known. In this case, the image rejection is limited by the amplitude and phase balance of the IF (which can be accounted for digitally). Other narrowband applications can use a simple branchline coupler.

In digital/CMOS implementations, a small size polyphase quadrature splitter can be made to reduce chip area, at the cost of very high loss for broadband designs as well as sensitivity to temperature and LO harmonic content. Alternatively, a digital divider can be used to create broadband quadrature LO signals, but with some noise addition from the digital logic, increased DC power consumption, low LO output power, and potential additive phase noise.

For broadband analog designs in a quasi-planar form, a [quadrature hybrid coupler](#) is the best option. A quadrature hybrid coupler uses edge or broadside coupling to implement a backwards wave coupler with an equal power split. Marki Microwave is an expert in the design and realization of planar quadrature hybrids in small form factors.

There are many electromagnetic effects that make broadband coupler design difficult, including dispersion, weak coupling, impedance control, etc. These effects must be carefully controlled for and added to the simulation of broadband designs. Marki Microwave uses these advanced quadrature hybrid design techniques in all broadband MMIC IQ mixers and quadrature hybrids, which are suitable for applications from 1.5 to 120 GHz.

4. IF Quadrature Signal Divider/Combiner: The IF quadrature splitter/combiner is more challenging to realize in an integrated circuit than the LO quadrature splitter for two main reasons. First, the IF is at a lower frequency than the LO, meaning the quarter wavelength is significantly longer. Second, several of the techniques useful for LO quadrature signal splitting are not useful for splitting or combining broadband signals.

Fortunately, there are two off-chip options available for performing the IF quadrature splitting/combining function. Analog splitting/combining can be accomplished with a large, low frequency surface mount quadrature hybrid. Fortunately, there are many of these devices available at lower frequencies due to the popularity of high power balanced amplifiers. For high IF applications, Marki Microwave offers MMIC quadrature hybrids from 2 to 42 GHz in [bare die](#) form factors and 2 to 55 GHz in [surface mount](#) form factors. Additionally, Marki offers [connectorized](#) quadrature hybrid couplers from 0.5 to 110 GHz.

A more powerful but complicated method is to combine or divide the signal digitally. In this case, two ADCs/DACs are used to read/write the signal to/from the I and Q ports individually. This has the unique distinction of working from DC to the bandwidth of the ADC/DAC. Additionally, impairments and

imbalance in the IQ mixer itself can be digitally compensated. This is an incredibly powerful technique that is popular as a result.

Structure	Bandwidth	Data or Single Frequency	Size	Benefits	Limitation
Digital Phase Shift	DC to GHz	Data	Tiny	Can be nearly perfectly implemented with software	Limited by the DAC or ADC used to convert to the analog domain
3 dB Quad Hybrid Coupler	Up to 10:1	Data	Large	Multioctave bandwidth, phase balance, power handling	Design and fabrication are challenging
Digital Phase Splitter	DC to 10s of GHz	Single frequency	Small	Balance limited only by clock jitter	Requires 2 x LO frequency, low power handling
Lange Coupler	Up to 3:1	Data	Medium	Quasi-planar, suitable for MMICs	Requires wire bonds or crossovers, limited power
Polyphase Quad Splitter	Suboctave	Data	Small	Can be made in CMOS	Loss, requires differential input, low isolation, sensitive to temperature and LO harmonics
Schiffman Phase Shifter	Multioctave	Data	Large	Power handling	Difficult to design, esoteric
Branchline Coupler	30 to 40%	Data	Large	East to implement	Size and bandwidth
Power Divider - Delay Line	Single frequency	Single frequency	Large	East to implement and tune	Single frequency

Table 1: Benefits and limitations of different structures.

VI. LINEARITY IN IQ, IR, AND SSB MIXERS

Linearity is improved in IQ, IR, and SSB mixers over their standalone mixer counterparts in two ways:

1. In some configurations, signal splitting reduces the signal power seen by each mixer. This can improve overall power compression (P1dB), two tone intermodulation (IP3), and higher order multitone intermodulation distortion (spurious). For more details, see **Table 2**.
2. By dividing signals in quadrature and combining them in phase, some isolations and spurious signals either add in quadrature phase (3 dB improvement) or out of phase (~20 dB or more improvement). For example, the LO-RF isolation of an IQ mixer improves by at least 3 dB over the equivalent standalone mixer because the LO is split in quadrature and recombined in phase. For some higher order spurs, the suppression improvement can be dramatic.

Conversion	Input IP3 and P1dB	Output IP3 and P1dB
IQ Mixer Upconversion	No effect	Degrades by 3 dB plus the excess insertion loss of the power divider
IQ Mixer Downconversion	Improves by 3 dB plus the excess insertion loss of the power divider	No effect
Image Reject Downconversion	Improves by 3 dB plus the excess insertion loss of the power divider	Improves by 3 dB minus the excess insertion loss of the IF hybrid
Single Sideband Upconversion	Improves by 3 dB plus the excess insertion loss of the IF hybrid	Improves by 3 dB minus the excess insertion loss of the power divider

Table 2: Linearity metrics of IQ, IR, and SSB mixers versus standalone mixers.

As with all claims about linearity in mixers, these theoretical improvements may or may not work in practice. Whether the linearity improves as expected depends on packaging, port impedances, harmonic content on the input signals, and other factors. Calculating the enhancement to spur suppression is complicated by the fact that the phase and amplitude error is multiplied for higher order terms. Each spur can be the vectorial sum of several products, leading to difficulty in calculating the expected spur suppression. For these reasons, precise simulations are preferred to analytic solutions. The Marki Microwave [PDKs](#) are invaluable tools for predicting the spurious suppression of IQ, SSB, and IR mixers in realistic environments.

As an example of spur suppression in a single sideband mixer, **Table 3** compares a simulation of the spurious suppression of the [MM1-0312H](#) as a standalone mixer and as a single sideband mixer. For each measurement, the inputs are the same: LO = 15 (18) dBm @ 3 GHz and IF = 0 dBm @ 60 MHz. On the left is the standalone suppression, and on the right in parentheses is the suppression of the mixer when configured to suppress the lower sideband with ideal quad hybrids and an ideal power combiner.

In addition to the sideband suppression, the dynamic range of a system is limited by the isolation or spurious suppression as well. The 2 IF x 1 LO spur and the LO feedthrough will both be separated from the desired tone by the IF frequency while the sideband is separated by twice the IF frequency, so both can limit the dynamic range even if the sideband is adequately suppressed. The bleedthrough of the LO signal is frequently higher in power than the suppressed sideband in a single sideband upconversion scheme, so filter requirements may not be reduced with a single sideband architecture without excellent LO to RF isolation.

	LO Order				
		1	2	3	4
IF Order	-4	-78 (-85)	-57 (-99)	-71 (-76)	-59 (-67)
	-3	-35 (-40)	-40 (-50)	-31 (-81)	-55 (-70)
	-2	-58 (-64)	-43 (-45)	-57 (-60)	-49 (-106)
	-1	0 (-60)	-13 (-16)	-6 (-5)	-34 (-36)
	0	-35 (-35)	-22 (-75)	-51 (-54)	-36 (-31)
	1	Reference	-13 (-16)	-6 (-62)	-35 (-38)
	2	-57 (-63)	-42 (-45)	-59 (-63)	-48 (-96)
	3	-33 (-81)	-40 (-50)	-32 (-40)	-57 (-79)
	4	-80 (-80)	-57 (-98)	-73 (-83)	-59 (-66)

Table 3: Simulated spur suppression of the MM1-0312H. The left value is the suppression when configured as a standalone mixer, and the right value in parentheses is the suppression when configured as a single sideband upconverter with the upper sideband selected.

As shown in Table 3, tones show a minor improvement in suppression (such as the 1 LO x 2 IF) and some show dramatic improvement (such as the 1 LO x 3 IF) depending on whether the constituent spurs add in quadrature, in phase, or out of phase. All tones but one (the 3 LO x -1 IF) show improvement.

In comparison, **Table 4** shows the experimentally measured spurious suppression of the [MM1-1044H](#) compared to the [MMIQ-1037H](#) with an external IF hybrid configured as a single sideband upconverter. The MM1-1044H is essentially a standalone version of the mixer inside the MMIQ-1037H, so users can expect similar results to Table 3. In this case, LO = 15 (18) dBm @ 3 GHz and IF = 0 dBm @ 60 MHz.

	LO Order				
		1	2	3	4
IF Order	-4	-67 (-90)	-67 (-80)	-71 (-76)	-59 (-67)
	-3	-23 (-45)	-45 (-70)	-31 (-81)	-55 (-70)
	-2	-55 (-66)	-64 (-57)	-57 (-60)	-49 (-106)
	-1	0 (-13)	-28 (-43)	-6 (-5)	-34 (-36)
	0	-35 (-37)	-27 (-21)	-51 (-54)	-36 (-31)
	1	Reference	-29 (-36)	-6 (-62)	-35 (-38)
	2	-57 (-66)	-60 (-55)	-59 (-63)	-48 (-96)
	3	-23 (-65)	-46 (-86)	-32 (-40)	-57 (-79)
	4	-68 (-93)	-70 (-79)	-73 (-83)	-59 (-66)

Table 4: Measured spur suppression of the MM1-1044H. The left value is the suppression when configured as a standalone mixer, and the right value in parentheses is the suppression when configured as a single sideband upconverter with the upper sideband selected.

Again, the spurs almost uniformly improve, sometimes dramatically. The 1 LO x 2 IF is suppressed by an additional 9 dB, and the 1 LO x 3 IF is suppressed by an additional 42 dB. However, the harmonic LO isolations are all degraded, as well as the 3 LO x ± 2 IF and the 2 LO x ± 2 IF spurs. Further investigation showed that this was a result of package resonances in the IQ mixer packaging. In conclusion, spurious suppression can be dramatically improved, and extremely precise modeling is required to determine what the exact impact will be.

VII. COMPENSATION OF IQ, IR, AND SSB MIXERS

It is extremely common in practice to compensate for the intrinsic imbalance of an IQ mixer, especially in applications where a DAC or ADC is connected directly to the I and Q ports. It can be shown that the phase and amplitude imbalance of the IQ mixer can be perfectly compensated for by applying phase and amplitude offsets to the I and Q signals at the DAC or ADC, and the LO feedthrough can be reduced by applying DC voltage offsets to the I and Q ports. A typical procedure is as follows:

1. Apply the LO signal to the mixer. Add a positive or negative DC voltage to the I and Q ports of the mixer while measuring the LO to RF feedthrough to find the minimum value (do not exceed the datasheet specifications for IF current on the mixer).
2. Transmit a signal. Digitally adjust the phase balance and amplitude balance of the I and Q ports until maximum sideband suppression is achieved.

3. Repeat this process for all desired LO frequencies and across temperatures to “train” the receiver/transmitter, and store the values determined.

In practice, this method is limited by the resolution of the DAC or ADC and the timescale of the drift of the circuit, so it always results in a limited improvement over the intrinsic rejection ratio of the mixer. Therefore, a high rejection mixer is desirable even when compensation techniques are used.

VIII. CONCLUSION

Integrating the complex functionality of IQ, image reject, and single sideband mixers into small, planar packages is extremely challenging. By combining precise fabrication with advanced mixer circuit design, Marki Microwave’s line of IQ mixers (the MMIQ series) and quadrature hybrids (the MQH series) offer superior performance. They can be used to create small form factor, high dynamic range transmitters and receivers that will enable significant architectural improvements for electronic warfare, radar, research, and test and measurement applications.

APPENDIX A: A PRACTICAL EXAMPLE

To illustrate the implications of various performance limits, consider a designer trying to create an X-Band synthesizer. The designer must generate a continuous wave tone between 8 and 12 GHz with the highest spur suppression possible. However, they only have

- A two-channel DAC capable of generating signals up to 2 GHz that acts as an arbitrary waveform generator
- An oscillator with an output frequency of their choice.

By changing from a positive 90° to a negative 90° phase shift between the DAC outputs, the designer can change from high side suppression to low side suppression. Therefore, they only need an LO at 10 GHz to cover this entire band. If they use the [MMIQ-0626](#), they will have an LO rejection of approximately 48 dB with a drive level of 19 dBm, creating a fixed output tone at 10 GHz of -29 dBm, a combined conversion loss of approximately 9 dB, and sideband suppression between 30 and 35 dBc untuned, as demonstrated in [Figure A1](#).

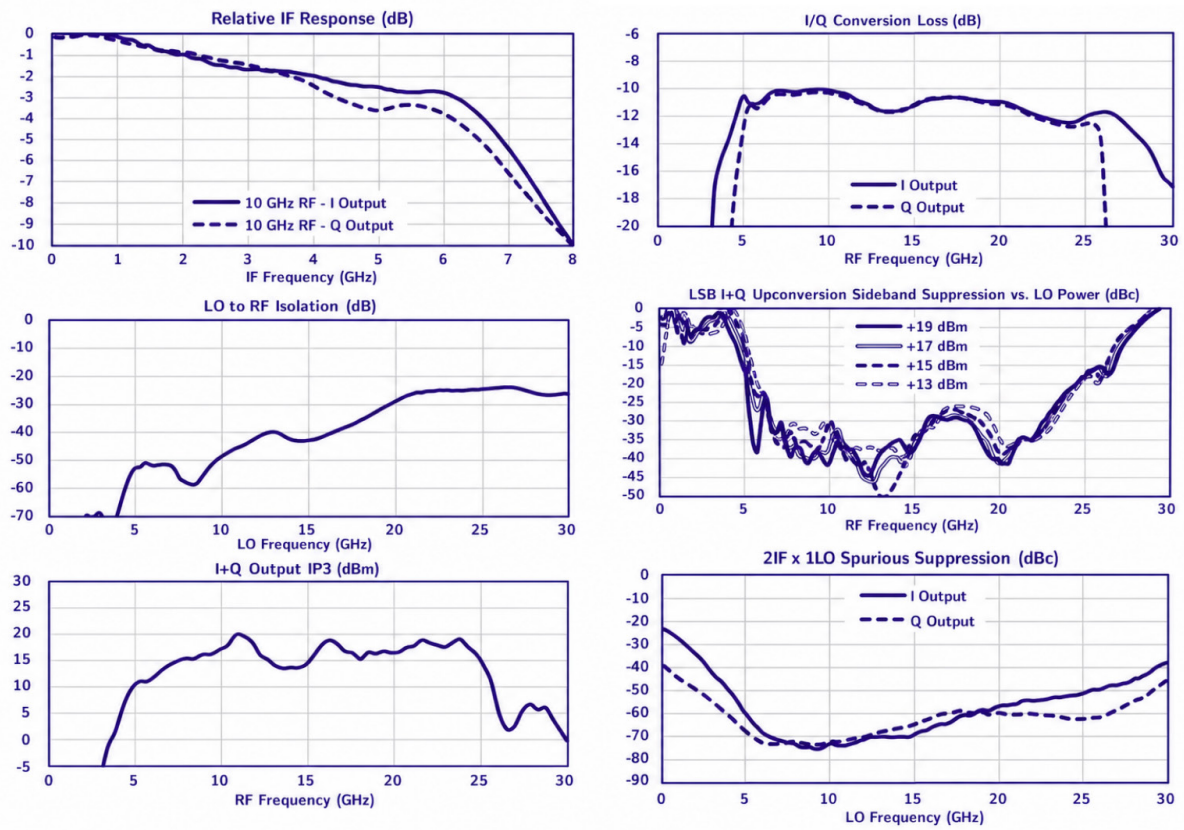


Figure A1. Performance of proposed set up with MMIQ-0626.

The last consideration is the 1 LO x 2 IF spur mentioned before, which is approximately 70 dBc for the MMIQ-0626H in the IQ configuration. However, since the designer is using it in the SSB configuration, the spurious suppression will likely improve. The only real way to know how much it will improve is to measure it. **Table A1** shows different tones and their associated power levels.

Tones	Frequencies	Parameter	Power Level at Maximum Output Power
Desired Conversion	8 to 10, 10 to 12 GHz	Conversion loss, IF response	0 to -2 dBm
LO Feedthrough	10 GHz	LO to RF isolation	-31 dBm
Undesired Sideband	8 to 10, 10 to 12 GHz	Upconversion sideband suppression	-30 dBm
Multitone Spurious	8 to 12 GHz	Output IP3	-30 dBm
Single Tone Spurious	8 to 10, 10 to 12 GHz	1 LO x 2 IF spurious suppression	-60 dBm

Table A1. Potential tones and their associated frequencies, parameters, and power levels.

Therefore, the signal is roughly equally degraded by the LO feedthrough and the sideband suppression, while the spurious products are at a much lower level. However, the designer can compensate for the LO feedthrough and sideband suppression, but not the spurious suppression. Thus, there are other examples where the dynamic range of the signal would be dominated by spurious suppression.

APPENDIX B: MATHEMATICAL FOUNDATIONS FOR IQ/IR/SSB MIXERS

Derivation of IQ Transmission Using Complex Exponentials

Assume the transmission of a signal $a(t)$ using an LO at frequency ω .

Use the identities

$$\cos(\omega t) = \frac{e^{j\omega t} + e^{-j\omega t}}{2}$$

$$\sin(\omega t) = \frac{-je^{j\omega t} + je^{-j\omega t}}{2}$$

Then the modulated signal is

$$a(t) \frac{e^{j\omega t} + e^{-j\omega t}}{2}$$

The demodulated signal with arbitrary phase offset is

$$\begin{aligned} a(t) \frac{(e^{j\omega t} + e^{-j\omega t})(e^{j\omega t + j\phi} + e^{-j\omega t - j\phi})}{4} \\ = a(t) \frac{e^{j\omega t + j\omega t + j\phi} + e^{j\omega t - j\omega t + j\phi} + e^{j\omega t - j\omega t - j\phi} + e^{-j\omega t - j\omega t - j\phi}}{4} \\ = a(t) \frac{e^{2j\omega t + j\phi} + e^{j\phi} + e^{-j\phi} + e^{-2j\omega t - j\phi}}{4} \end{aligned}$$

After lowpass filtering to remove the $2j\omega t$ terms, the result is

$$a(t) \frac{e^{j\phi} + e^{-j\phi}}{4}$$

When $\phi=0$, the transmitted signal is recovered, when $\phi=\pi/2$, nothing is recovered, and when ϕ is in between, $\cos(\phi)$ times the transmitted signal is recovered.

Derivation of Single Sided Downconversion

Assume that the input signal is:

$$\frac{a(t)}{2} e^{-j\omega t}$$

Multiply it by the LO signal with an arbitrary phase:

$$\cos(\omega t + \varphi) = \frac{e^{j\omega t + j\varphi} + e^{-j\omega t - j\varphi}}{2}$$

The resulting demodulated signal:

$$\frac{a(t)}{2} e^{-j\omega t} (e^{j\omega t + j\varphi} + e^{-j\omega t - j\varphi}) = \frac{a(t)}{2} e^{-j2\omega t - j\varphi} + e^{j\varphi}$$

Lowpass filtering removes the $e^{-j2\omega t - j\varphi}$ term, which leaves:

$$\frac{a(t)}{2} e^{j\varphi}$$

This is the original input signal $a(t)$ with an LO dependent phase shift φ applied to it.

Derivation of Single Sided Upconversion

I signal input: $a(t)$

LO drive for I mixer: $\frac{e^{j\omega t} + e^{-j\omega t}}{2}$

Upconverted signal from I mixer, with no phase shift applied: $a(t) \frac{e^{j\omega t} + e^{-j\omega t}}{2}$

Q signal input, after a phase shift of -90° ($-j$) (chosen to select the upper sideband) is applied: $j * a(t)$

LO drive for Q mixer: $\frac{-j * e^{j\omega t} + j * e^{-j\omega t}}{2}$

Upconverted signal from Q mixer, with -90° phase shift applied: $j * a(t) \frac{-j * e^{j\omega t} + j * e^{-j\omega t}}{2}$

Combined signal output from the power combiner:

$$a(t) \frac{e^{j\omega t} + e^{-j\omega t}}{2} + j * a(t) \frac{-j * e^{j\omega t} + j * e^{-j\omega t}}{2} = \frac{a(t)}{2} [1 + (j * -j)] e^{j\omega t} + [1 + (j * j)] e^{-j\omega t} = a(t) e^{j\omega t}$$

Derivation of Image Reject Downconversion

Assume an RF input signal consisting of sideband A and sideband B: $a(t)e^{j\omega t} + b(t)e^{-j\omega t}$

The I mixer is downconverted with the LO: $\frac{e^{j\omega t} + e^{-j\omega t}}{2}$

$$\begin{aligned} \text{Giving the output: } & [a(t)e^{j\omega t} + b(t)e^{-j\omega t}] \frac{e^{j\omega t} + e^{-j\omega t}}{2} \\ &= \frac{a(t)}{2} e^{j2\omega t} + \frac{b(t)}{2} e^{j\omega t - j\omega t} + \frac{a(t)}{2} e^{-j\omega t + j\omega t} + \frac{b(t)}{2} e^{-j2\omega t} \\ &= \frac{a(t)}{2} (1 + e^{j2\omega t}) + \frac{b(t)}{2} (1 + e^{-j2\omega t}) \end{aligned}$$

The Q mixer is downconverted with the LO: $\frac{-j * e^{j\omega t} + j * e^{-j\omega t}}{2}$

$$\begin{aligned} \text{Giving the output: } & [a(t)e^{j\omega t} + b(t)e^{-j\omega t}] \frac{-j * e^{j\omega t} + j * e^{-j\omega t}}{2} \\ &= \frac{-j * a(t)}{2} e^{j2\omega t} + \frac{-j * b(t)}{2} e^{j\omega t - j\omega t} + \frac{j * a(t)}{2} e^{-j\omega t + j\omega t} + \frac{j * b(t)}{2} e^{-j2\omega t} \\ &= \frac{j * a(t)}{2} (1 - e^{j2\omega t}) + \frac{j * b(t)}{2} (-1 + e^{-j2\omega t}) \end{aligned}$$

Applying a phase shift j to the I mixer output and combining:

$$\begin{aligned} & \frac{-j * a(t)}{2} e^{j2\omega t} + \frac{-j * b(t)}{2} e^{j\omega t - j\omega t} + \frac{j * a(t)}{2} e^{-j\omega t + j\omega t} + \frac{j * b(t)}{2} e^{-j2\omega t} \\ &= \frac{j * a(t)}{2} (1 - e^{j2\omega t}) + \frac{j * b(t)}{2} (-1 + e^{-j2\omega t}) \end{aligned}$$

After lowpass filtering, the result is:

$$j * a(t)$$

REFERENCES

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