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SAND2002-1316
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Printed May 2002

Quadrature Mixer LO Leakage Suppression Through Quadrature DC Bias

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Quadrature Mixer LO Leakage Suppression Through Quadrature DC Bias

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ABSTRACT

A new concept has been developed which allows direct-to-RF conversion of digitally synthesized waveforms. The concept named Quadrature Error Corrected Digital Waveform Synthesis (QECDWS) employs quadrature amplitude and phase predistortion to the complex waveform to reduce the undesirable quadrature image. Another undesirable product of QECDWS-based RF conversion is the Local Oscillator (LO) leakage through the quadrature upconverter (mixer). A common technique for reducing this LO leakage is to apply a quadrature bias to the mixer I and Q inputs. This report analyzes this technique through theory, lab measurement, and data analysis for a candidate quadrature mixer for Synthetic Aperture Radar (SAR) applications.

ACKNOWLEDGEMENTS

This work was funded under the Sandia National Laboratories LDRD (Laboratory Directed Research and Development) program. The project is entitled "Low Cost Digital Radar for Fuzing, Tags, SAR Imaging, and Targeting" and is funded under the Emerging Threats SBU.

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1 Introduction

Sandia National Laboratories Synthetic Aperture Radar (SAR) Department is currently investigating the concepts of next-generation radar systems, which employ state-of-the-art high-speed digital hardware and digital signal processing (DSP) techniques. An integral part of this effort is the exploration of a novel technique, which can directly generate single-sideband (SSB) representations of the transmitted waveform with very low spurious signal content (spectrally pure). The technique, coined Quadrature Error Corrected Digital Waveform Synthesis [1] (QECDWS), employs a quadrature corrected direct-digital synthesizer in conjunction with a quadrature mixer. The mixer converts the synthesized linear-FM waveform (chirp) to either an intermediate frequency or directly to the desired RF frequency. A conceptual design for a next generation SAR RF and High-Speed Digital (HSD) subsystem, which employs the QECDWS is given in [3]. A simplified block diagram of a proposed wideband radar exciter subsystem is shown in Figure 1.

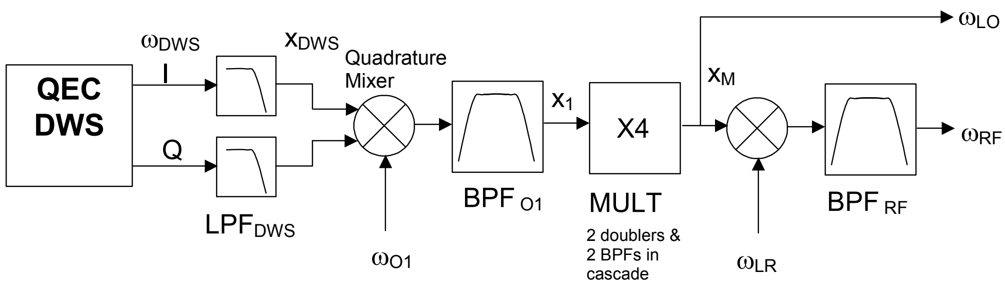

Figure 1: Proposed radar exciter [3] which utilizes a QECDWS.

A single-ended synthesizer has a theoretical maximum signal bandwidth of $f_s/2$, where f_s is the sample rate of the synthesizer. With quadrature synthesis and SSB conversion, the theoretical bandwidth is equal to f_s. This provides a huge advantage in radar systems, which require wide bandwidth signal generation, such as the Sandia developed SARs.

The biggest problem, which plagues quadrature synthesis and SSB conversion, is the presence of unwanted spurious signals. The two primary problematic spurs are the SSB image and the quadrature mixer LO signal leakage. The image is due to gain and phase imbalances between the In-phase (I) and Quadrature (Q) channels of the synthesizer/mixer combination. The QECDWS technique [2] uses quadrature waveform predistortion to cancel the effect of the quadrature error, thus suppressing the unwanted image. The mixer LO leakage, on the other had, can be suppressed by precisely adjusting the DC bias level of the mixer I and Q ports. A calibration algorithm, which suppresses both the image and LO leakage terms in a SAR exciter, has been analyzed and simulated [2].

This report documents the basic theory, laboratory testing, and analysis of a candidate quadrature mixer for future application in a next-generation SAR system employing a

QECDWS-based signal generator. The purpose of the testing was to thoroughly characterize the relationship between DC bias level and LO leakage level. In addition, algorithms for automated nulling of the LO leakage spur have been implemented and tested. Other parameters considered are sensitivity to input frequency and power level as well as time and temperature stability.

2. Theory

The model used to analyze the quadrature mixer LO leakage is shown in Figure 2.

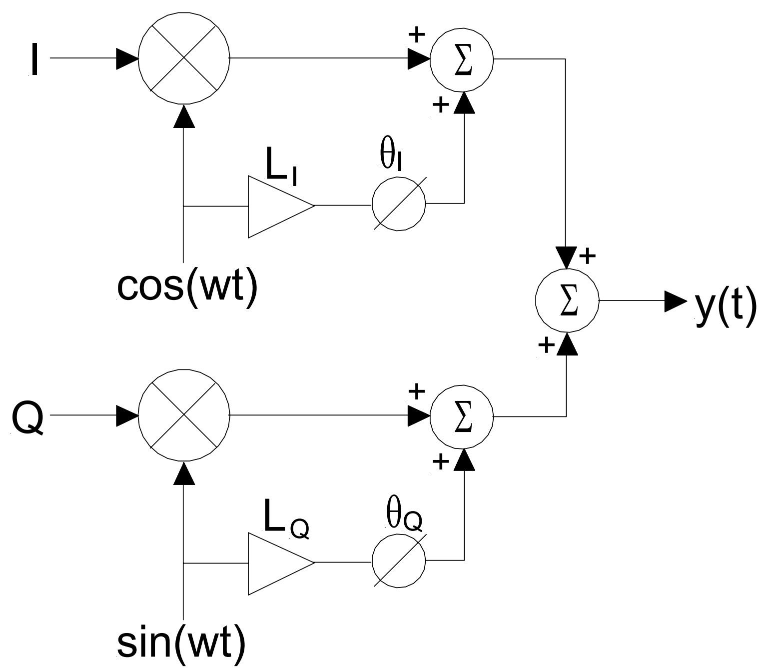

Figure 2: Model for analysis of the quadrature mixer LO leakage.

The mixer inputs are the DC bias I and Q. The leakage for each mixer is modeled as a scaled (L_I and L_Q) and phase shifted (θ_I and θ_Q) version of each LO input ($\cos(\omega t)$ and $\sin(\omega t)$). The goal is to determine I and Q such that y(t)=0.

We can write the mixer output y(t) as

$$y(t) = I\cos(\omega t) + L_I\cos(\omega t + \theta_I) + Q\sin(\omega t) + L_Q\sin(\omega t + \theta_Q)\,. \tag{1}$$

We can rewrite y(t) as the sum of a cosine and a sine term:

$$y(t) = A\cos(\omega t + \alpha) + B\sin(\omega t + \beta)\,, \tag{2}$$

where the amplitude and phase of the cosine and sine terms are

$$A = \sqrt{(I + L_I\cos\theta_I)^2 + (L_I\sin\theta_I)^2}\,, \tag{3}$$

$$\alpha = \tan^{-1}\left(\frac{L_I\sin\theta_I}{I + L_I\cos\theta_I}\right), \tag{4}$$

$$B = \sqrt{(Q + L_Q \cos\theta_Q)^2 + (L_Q \sin\theta_Q)^2} \text{, and} \tag{5}$$

$$\beta = \tan^{-1}\left(\frac{L_Q \sin\theta_Q}{Q + L_Q \cos\theta_Q}\right). \tag{6}$$

For y(t)=0, we must satisfy

$$\alpha - \beta = \frac{\pi}{2}\text{, and} \tag{7}$$

$$\text{A=B.} \tag{8}$$

To satisfy equation (7), we first note that if α and β differ by 90°, then their tangents are reciprocals, i.e.

$$(L_I \sin\theta_I)(L_Q \sin\theta_Q) = (I + L_I \cos\theta_I)(Q + L_Q \cos\theta_Q)\,, \tag{9}$$

or

$$\boxed{IQ + IL_Q \cos\theta_Q + QL_I \cos\theta_I + L_I L_Q \cos(\theta_I - \theta_Q) = 0}. \tag{10}$$

As for the amplitude, solving equation (8) and rearranging terms yields

$$\boxed{I^2 - Q^2 + 2IL_I \cos\theta_I - 2QL_Q \cos\theta_Q + L_I{}^2 - L_Q{}^2 = 0}. \tag{11}$$

Both equation (10) and (11) define hyperbolas on the I,Q plane. The asymptotes of equation (10) are the I and Q axes, with the asymptotes of equation (11) at 45° to the I and Q axes. Both asymptotes intersect at [$-L_I\cos\theta_I$, $-L_Q\cos\theta_Q$]. For arbitrary values of L_I, L_Q, θ_I, and θ_Q we can plot those values of I and Q which satisfy (or nearly satisfy) equation (10) and (11). Figure 3 shows an example plot where solutions to equations (10) and (11) below a threshold value (-60 dB in this case) are plotted.

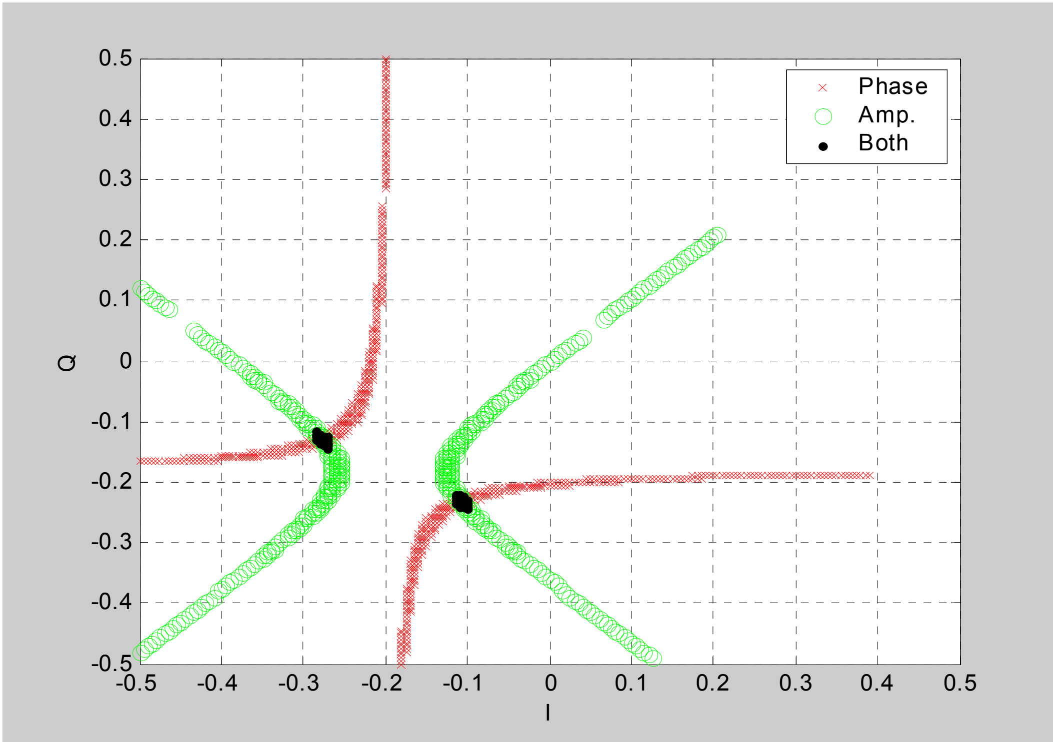

Figure 3: Location of minima for the phase and amplitude criteria for of L_I=L_Q=-14dB, and θ_I=25° and θ_Q=15°.

Note that for y(t)=0, both the phase and amplitude criteria must be satisfied (intersection of the hyperbolas shown in Figure 3). By plotting the sum of the absolute values (in dB) of the left sides of equations (10) and (11) above the I,Q plane, we get a three-dimensional representation of the relationship between the mixer output LO leakage (error) and the DC bias offsets I and Q. This is shown in Figure 4.

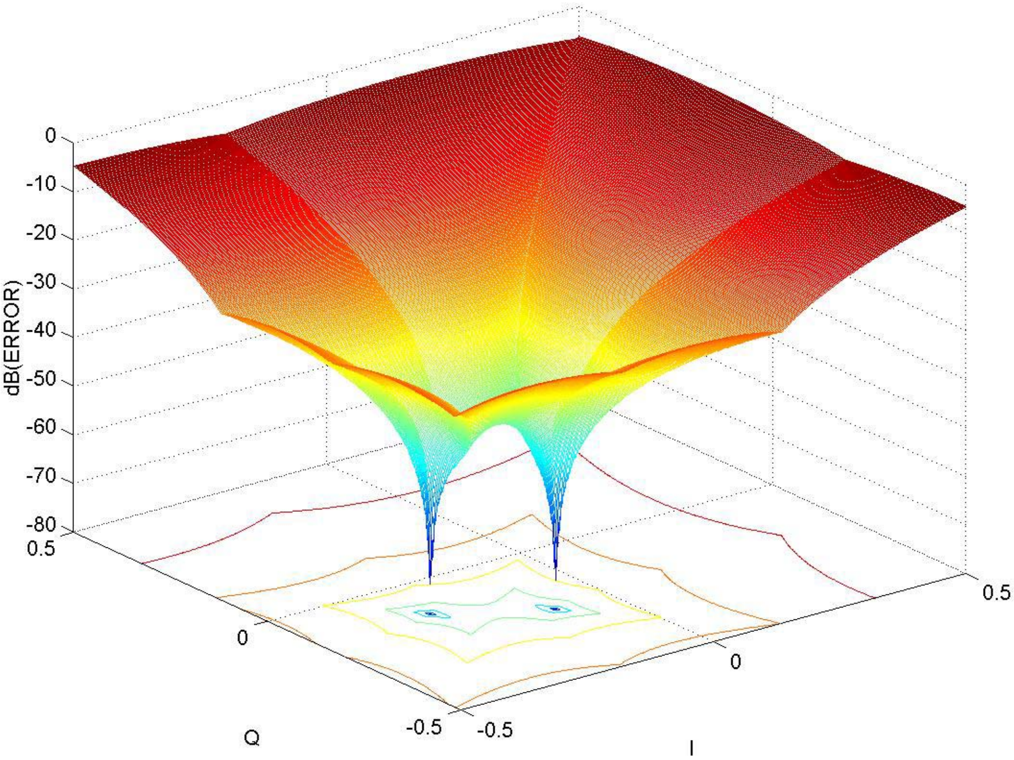

Figure 4: Quadrature mixer output LO leakage (error) vs. I and Q bias.

Note that for this case, the leakage amplitudes L_I and L_Q are rather large (-14 dB) which is not typical for a quadrature mixer. This was done to illustrate the existence of two minima. In reality, commercial mixers have LO leakage on the order of –25 to –35 dB, depending on frequency and implementation. Also, the leakage phase shift is expected to be very close to zero. If we recalculate the error for small leakage amplitudes and zero phase offset, we get a single minima as shown in Figure 5.

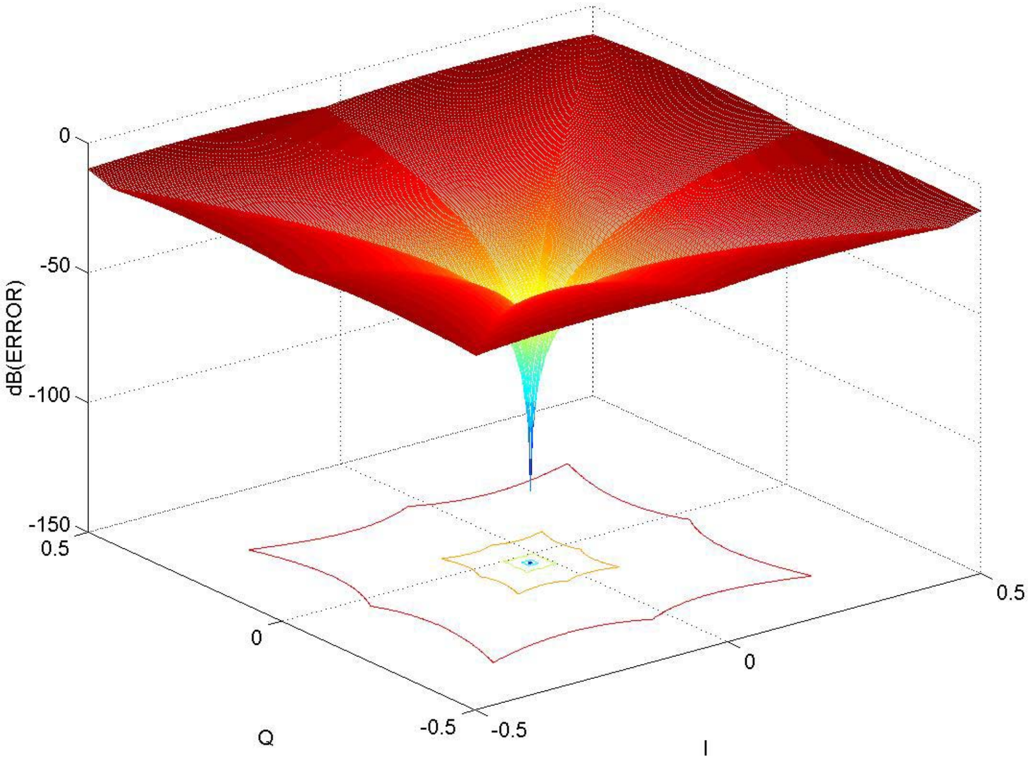

Figure 5: Single error minima for L_I=L_Q=-30dB, and θ_I=θ_Q=0.

For our commercial mixer (Remec MIQ24MS-2), the LO leakage is expected to be approximately –30 dB, which should result in a single null on the I,Q plane. The general nature of this null (stability over time and temperature, relationship to frequency, null depth, etc.) and the best search method for finding the null are the main points of the measurements and analysis to follow.

3. Quadrature Mixer Test Setup

The MIQ24MS-2 is a quadrature RF mixer from Remec. It is designed to combine the base-band in-phase (I) and quadrature-phase (Q) components (which may be between DC and 500 MHz) of a signal and multiply it with a local oscillator signal (between 1.9 GHz and 4.2 GHz). This particular model of mixer requires an LO power level of +10 dBm. The goal of this evaluation has been to determine the feasibility of rejecting the LO signal in the mixer output by adding small DC offsets to the I and Q inputs. Ideally, the process of rejecting the LO signal should be automated and should display a degree of stability with respect to frequency, temperature, and time.

Figure 6 illustrates the setup used to evaluate the mixer performance. All devices were controlled remotely through MATLAB [4], with the exception of the quadrature synthesizer, which was controlled manually via software provided with the AD9854 evaluation board.

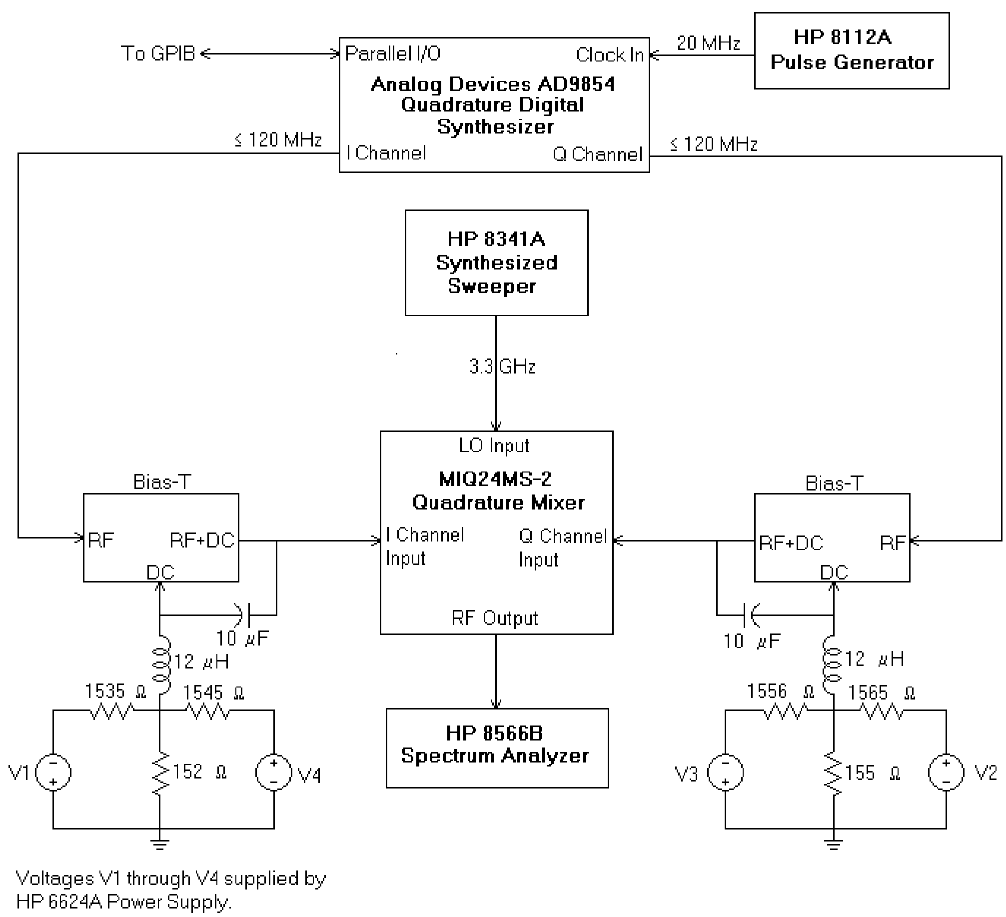

Figure 6: Test setup for the quadrature mixer.

4. Performance of the Quadrature DDS

An Analog Devices AD9854 Quadrature DDS was used to provide the RF test signal for the mixer. Figure 7 is a plot of the power envelope of the in-phase component of the DDS signal. The DDS output here has been mixed up to 3.3 GHz.

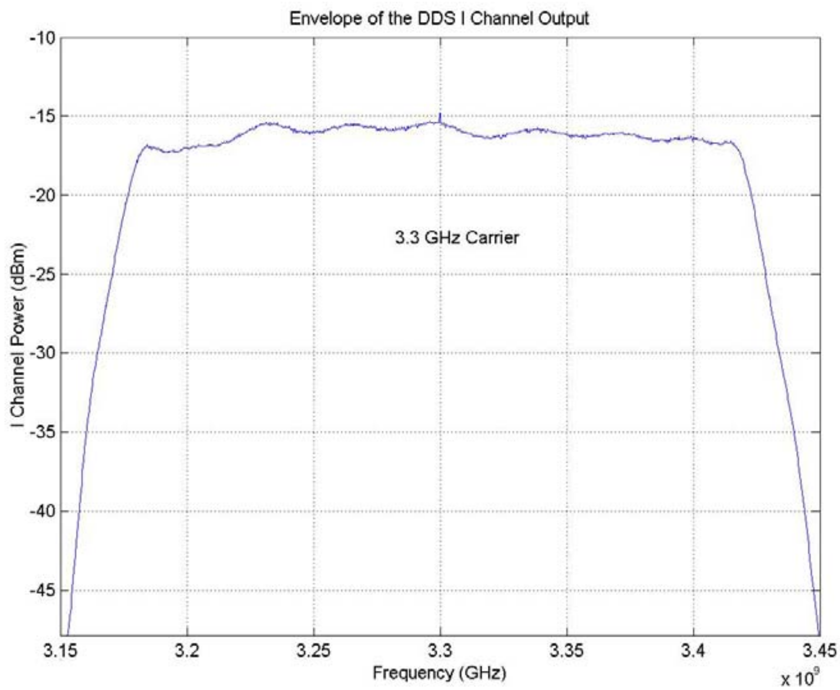

Figure 7: DDS output power vs. frequency.

The DDS amplitude here has been set to its full-scale value of FFF_{Hex}, and unless otherwise noted this is true for all the following tests. Within the passband of the DDS (120 MHz) the signal power varies by no more then 2 dB, and is typically –16 dBm. The roll off which begins at approximately ±120 MHz about the LO frequency is due to filters integral to the AD9854 evaluation board.

Figure 8 is the spectrum of a constant 20 MHz DDS output in a 100 MHz window, as it appears at the mixer output. The I and Q channels have no added DC offset so the LO feed through is at its maximum, unmodified level of approximately –10 dBm (+6 dBc). Typically the Q channel rejection is on the order of –20 dBc.

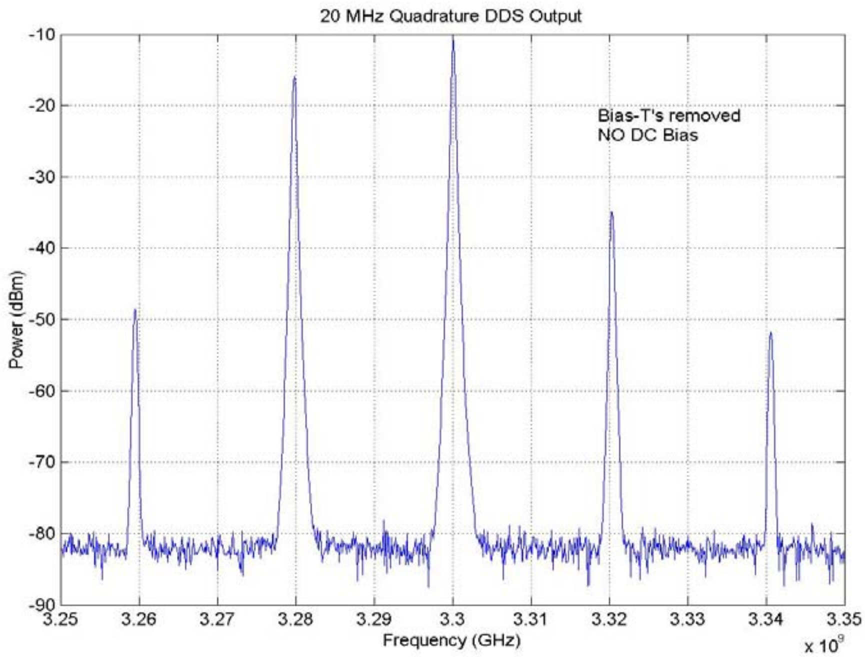

Figure 8: Spectrum of a 20 MHz DDS output.

5. Bias-T Characteristics

A “bias-T” network, capable of adding a DC offset to an RF signal, was added to both the in-phase and quadrature channels. After some LO rejection tests had already been conducted there was some concern that the RF input to the bias-T was having an effect on the DC offset applied to the network (which is undesirable). The resistance between the DC and DC/RF ports was measured to be 3 Ω for each network. For a given DC value (which suppressed the LO signal to –74 dBc) the RF input was swept over its entire range and the DC voltage between the DC input and DC/RF output (to the mixer) was measured. The voltage drop varied by no more than 300 μV for the I channel network and 190 μV for the Q channel network.

There appears to be an approximately 20 dB/decade relationship between the LO power and the offset voltage (determined from the following measurement).

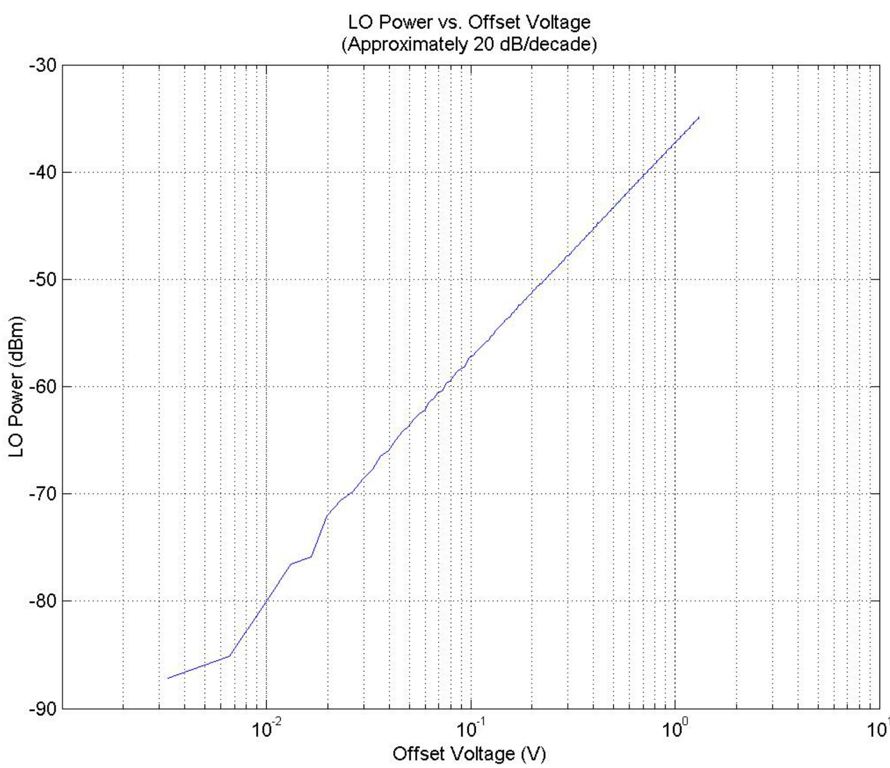

Figure 9: Variation in LO power vs. Q-channel bias.

Figure 9 shows the variation in LO power with respect to Q channel voltage, with the I channel voltage fixed at that value which allows optimal LO suppression (these data sets were taken from the 2 dimensional null search plots). In this case the LO power is approximately –37 dBm at an offset of 1 V. For offsets in the mV range, the LO power is less than –80 dBm and very near the lowest point measured. In this range, a DC change of 0.3 mV (the maximum RF induced voltage change) would produce a change of only 0.6 dB. This suggests that the effects of the RF signal on the DC offset voltages are negligible.

6. DC Power Supply Characteristics

The original mixer test configuration used two independent HP 6236B triple output power supplies to provide the DC bias for the I and Q channels. A potentiometer was used to manually fine-tune the offset voltage for each channel. By carefully adjusting the offsets it was possible to consistently achieve an LO rejection on the order of –85 dBc (-100 dBm) as shown in Figure 10.

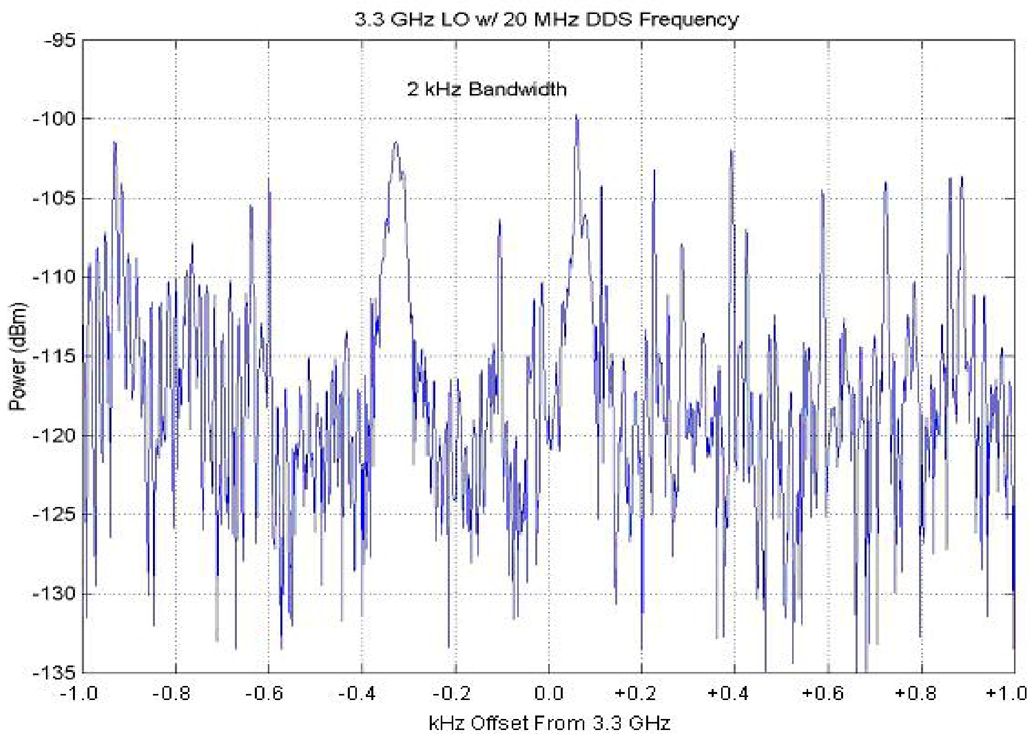

Figure 10: LO leakage vs. offset from carrier.

Unfortunately, this setup was very vulnerable to minor disturbances. Measured values indicated an LO power to voltage sensitivity of approximately 20 dB per decade of voltage, that is, a differential change of 10 mV was enough to increase the LO leakage level from –100 dBm to –80 dBm. Temperature, RF frequency, and LO frequency all degrade the LO rejection for a given voltage offset as well, although the offset can be adjusted to compensate for any changes of those values. Through the course of evaluation it was noted that the maximum (approximate) LO rejection of –85 dBc was impractical because it was nearly impossible to maintain. Temporal changes alone were often cause enough to increase the LO leakage level.

The manually adjusted potentiometers were replaced with a 4 channel DC power supply (HP 6624A) to enable the automation of the LO rejection process. The four channels of the power supply were hooked up to a voltage divider network as shown in Figure 6. The scaled difference of the two input voltages becomes the DC offset input to the bias-T network. Unfortunately the precision of the DC power supply is an issue. Figure 11 illustrates the difference between the desired voltage and the measured voltage for a single channel. As illustrated in Figure 11, the DC power supply exhibits peak errors as on the order of 7 mV. These measurements were repeated for all four channels with very similar results.

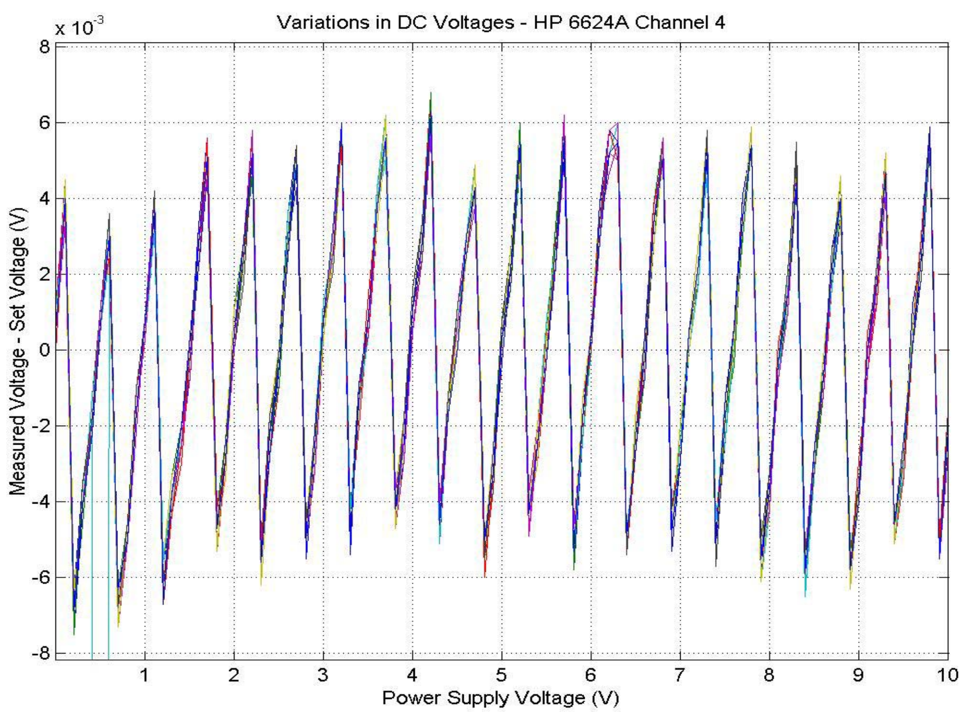

Figure 11: DC bias supply error.

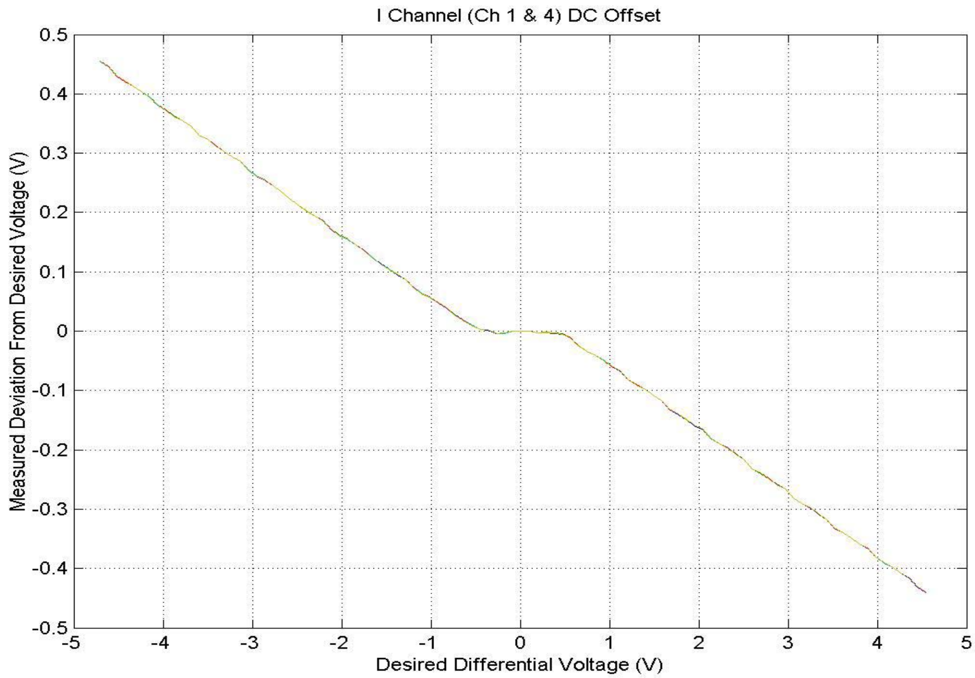

Figure 12: Differential bias error.

The precision of the power supply is an issue because it directly affects the ability of a search algorithm (which controls the power supply output) to consistently locate an LO minima. Assuming that the deviation from the desired voltage is no greater than 8 mV, then the effect on the LO power could be on the order of 16 dB near the null. This applies only to one channel in error. Because the location of the null in the voltage field is not fixed with time, temperature, and frequency it is not practical to record the offsets required to reach the null and then attempt to reuse those same values blindly at future times. A search algorithm is used to locate the null from "scratch" each time so the absolute accuracy of the DC power supply is not very important. The search algorithm should find the null regardless of any offsets so long as the offsets are continuous. Unfortunately, the deviation of a single channel is not continuous. An extra 7 mV error added to a voltage step in the search algorithm would ruin the chances of an acceptable null value being located. Fortunately, when two DC channels are combined to produce a differential voltage, the cyclical error from each channel (Figure 11) cancels resulting in a much more linear error characteristic as shown in Figure 12. Over the range of approximately ±500 mV the measured output is equal to the desired output. Above or below 500 mV the deviation of the measured output from the desired output becomes linear with a slope of approximately –100 mV/V. In the linear search algorithm used where the final voltage step is 3 mV, the actual deviation from this should be about 0.3 mV, which equates to roughly 0.6 dB variation in the LO power (potentially stepping "across" the absolute minima). This can be considered a relatively minor source error, so long as a minima below –90 dBm is not needed. Ideally, a far more precise power supply should be used to provide the DC offset to the I and Q channels.

7. LO Minima Over The Entire Voltage Field

By searching over a continuous range of I and Q offset voltages, it became clear that there is a single LO minima rather than multiple local nulls (Figure 13). This agrees with the previous calculation of a single LO minima assuming a relatively small LO leakage amplitude and an LO leakage phase of approximately zero.

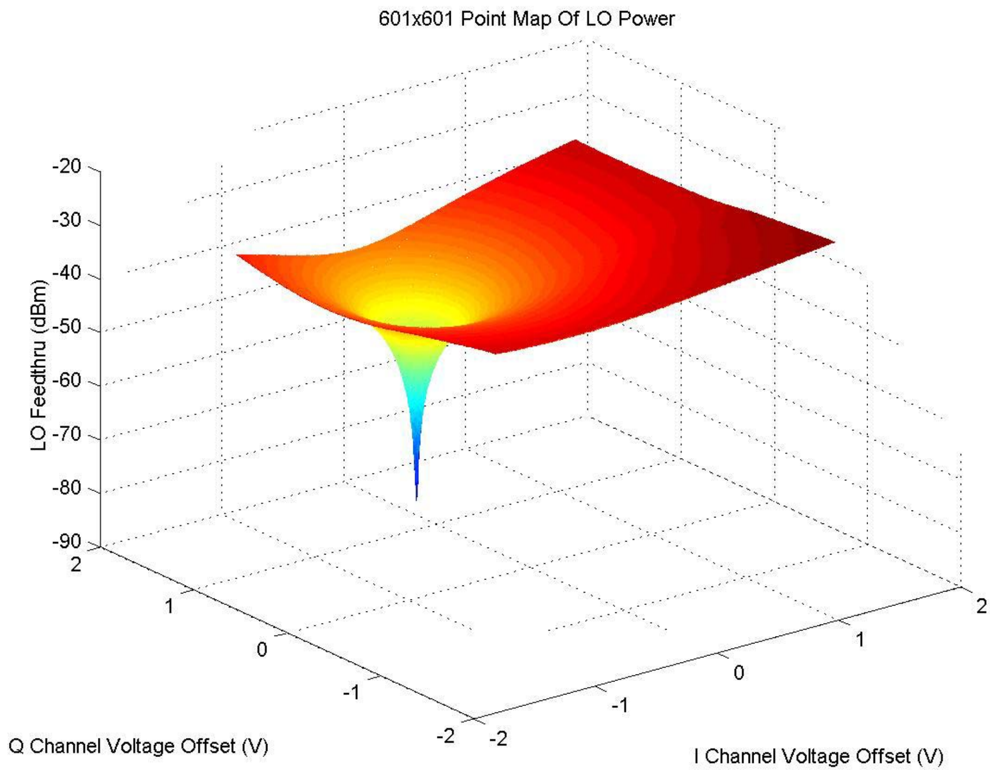

Figure 13: LO feed-through vs. I and Q bias showing single minima.

The exact location of this null in the voltage field is approximately fixed assuming a generally constant temperature and RF frequency. There is some minimal drift in time. Figure 14 and Figure 15 illustrate the LO null measured on three different days.

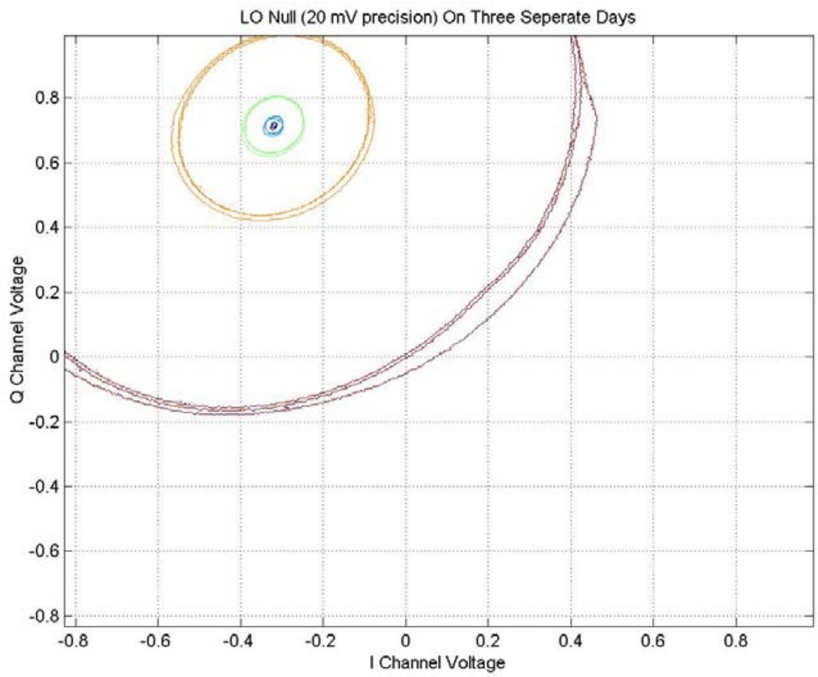

Figure 14: Drift in LO null over time.

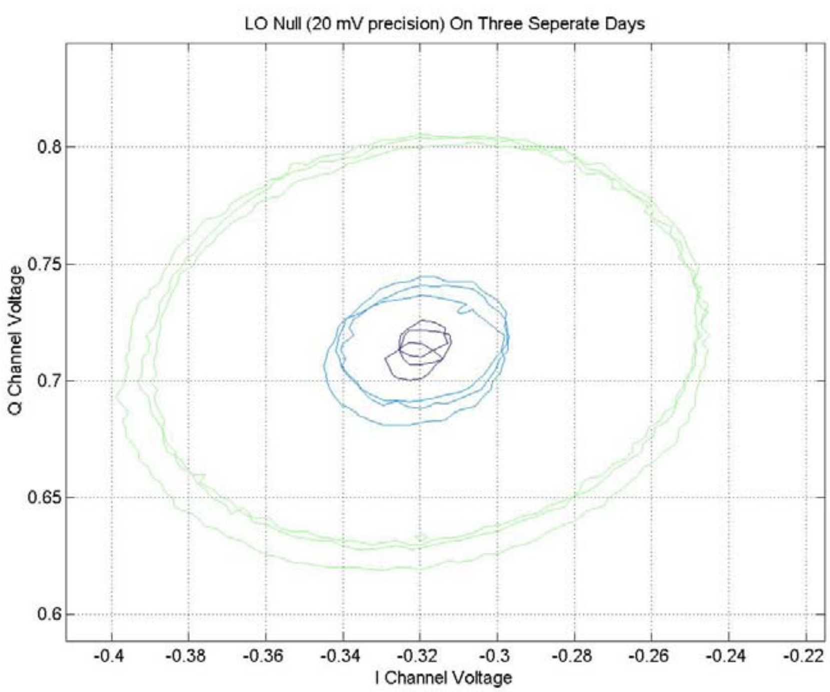

Figure 15: Drift in LO null over time (zoomed).

The exact location of the minima appears to vary by a few 10's of mV in terms of the I and Q channel voltages. Each of these data sets required several hours to gather so it must be assumed that some drift occurred during the data acquisition process (this may be one reason why a lower minima was not detected).

8. Automated LO Minima Search

Two automated methods of locating a minima in the LO leakage were tried. First, a gradient search, and then a linear search were used.

The ability of a gradient search to converge on a minima depended on two values,

$$V_{diff} = (\Delta V^- - \Delta V^+) / 2, \text{ and}$$

$$V_{new} = V_{old} + \mu * \text{sign}(V_{diff}) / |V_{diff}|.$$

Here, V_{diff} is determined by measuring the LO power at points on either the I or Q axis, before and after (ΔV^- and ΔV^+ voltages) the current voltage offset value. Because the voltage change is inversely related to the difference between the two measured points (the slope of the curve) the steps should be large when far away from the minima (where the derivative is small) and decrease as the minima is approached. The constant μ scales the voltage step to prevent excessively large changes which might pass over the minima entirely. Small values of μ require more time for the minima to be found, with larger values running the risk of passing over the minima. The value of ΔV is limited at both the high and low ends. If ΔV is too large then points may be taken on either side of the minima which could cause the search to move in the wrong direction momentarily and almost certainly would prevent the lowest possible value from being found. Because the resolution of the power levels taken from the spectrum analyzer is somewhat coarse (0.1 dBm), and if ΔV is too small, then V_{diff} may not accurately reflect the slope of the surface. This would cause an inappropriately large step to be taken, possibly skipping across the minima. Also, due to very small local minima in the surface, the derivative could point in the wrong direction if ΔV is too small. Unfortunately there is no ideal value for either ΔV or μ which holds over the entire surface. When starting far away from the minima and seeking a relatively low LO leakage level, it is necessary to set intermediate LO thresholds between which the values for ΔV and μ are changed.

In practice, the gradient search algorithm proved very inconsistent as illustrated Figure 16 and Figure 17.

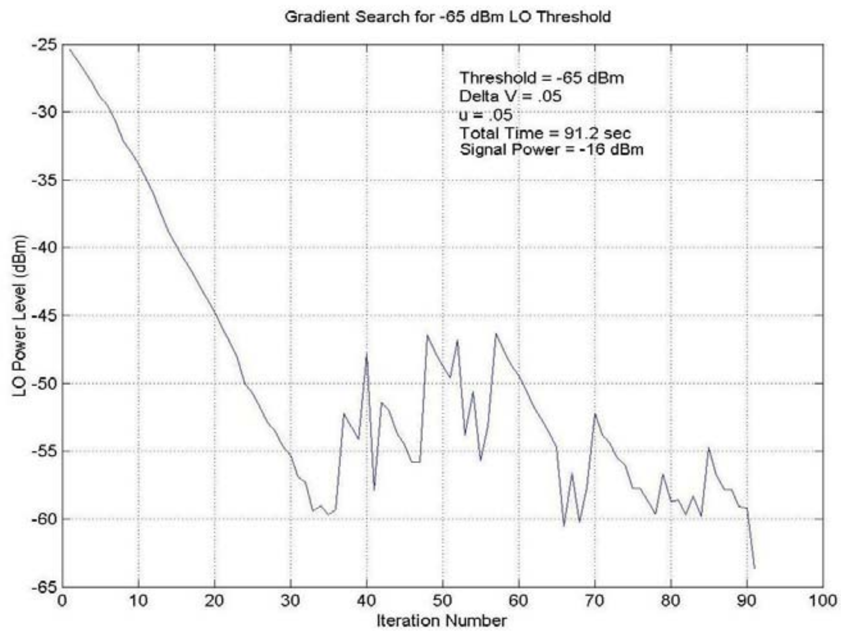

Figure 16: Gradient search example.

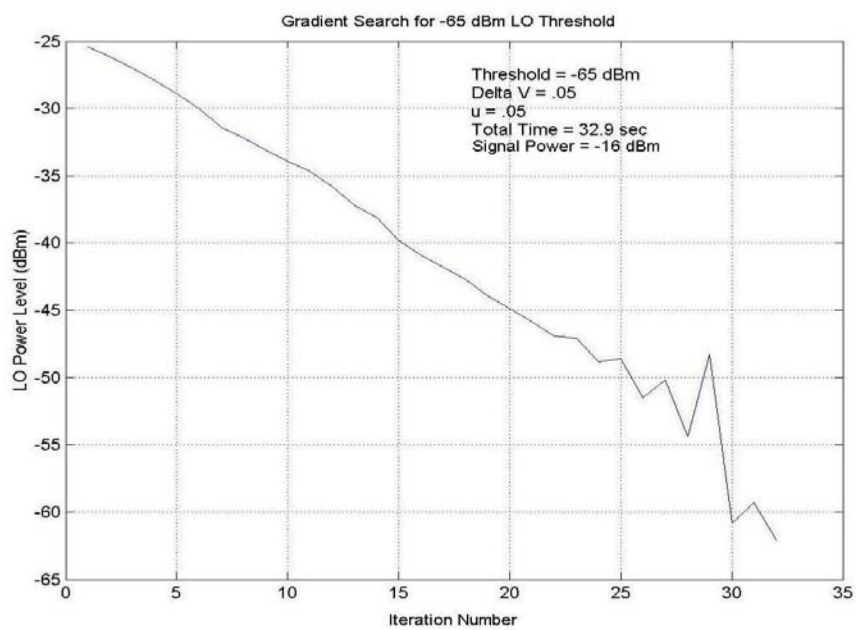

Figure 17: Gradient search example.

With the gradient search, one trial might reach the desired LO power level in fewer than 50 iterations while another would fail to locate it all together in over 300 iterations. It is likely that this was due to the algorithms dependence on the difference between the samples taken above and below the current offset. Because the two samples are not taken simultaneously there is a strong possibility that, when ΔV is small enough, there is enough noise in the system to change the LO power level enough to yield a slope in the wrong direction. The plots of Figure 18 were generated using the gradient search on a static data set (the field had already been recorded thus it was unchanging during the test), to determine if the algorithm itself was at fault. These tests yielded far more optimistic results than the performance of the algorithm on “real-time” data, which suggests that noise or other temporal changes are to blame for the poor performance of the gradient search.

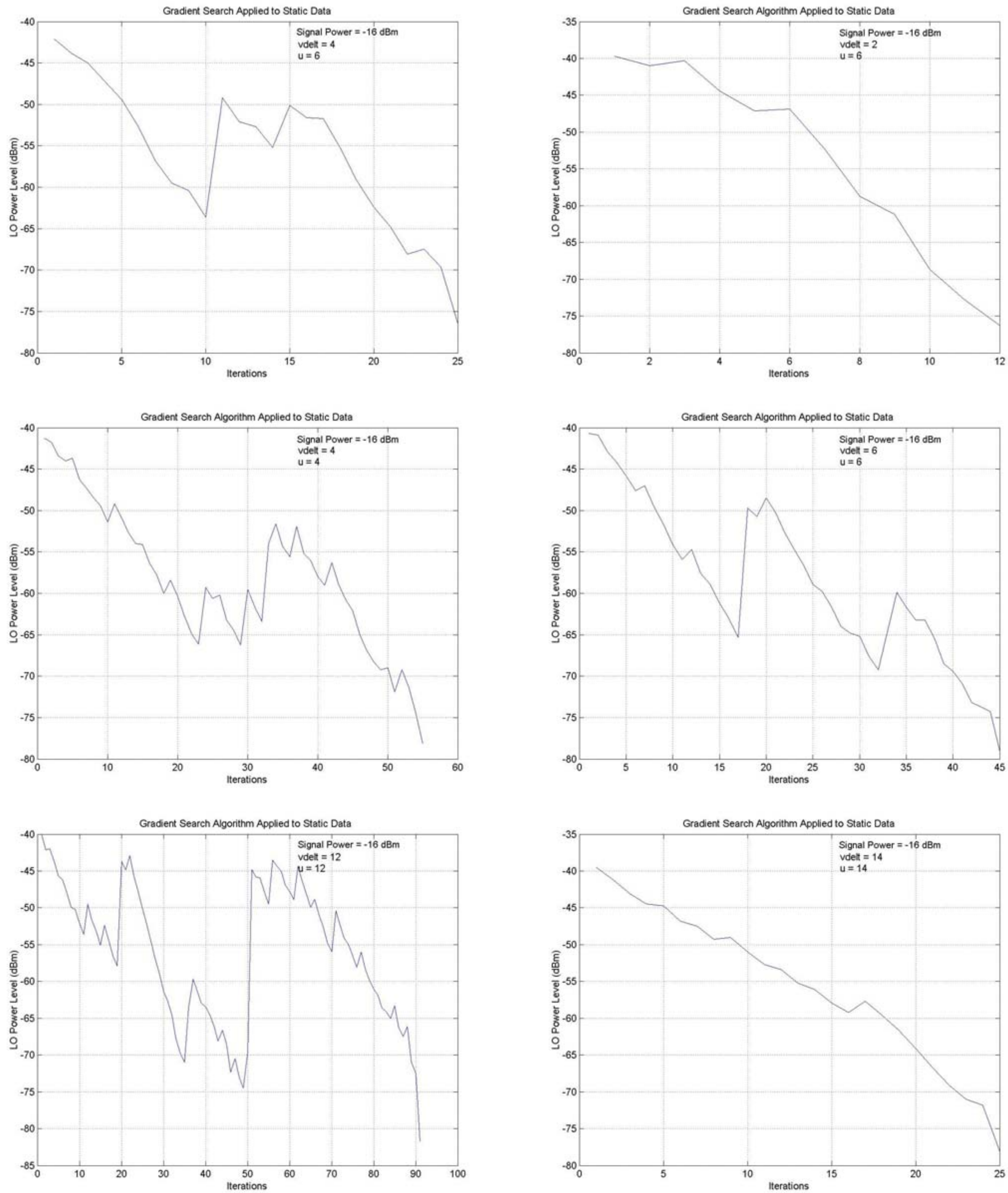

Figure 18: Gradient search results on static test data.

A second, simpler, linear search algorithm was also used with far more consistent results. The original idea was to conduct a gradient search down to a certain threshold then switch to some other kind of search. However, the gradient search was discarded altogether when the linear search proved more effective overall.

The linear search algorithm steps along one of the two voltage axis (I or Q) measuring the new LO power with every iteration. Once a new LO power is detected that is greater than the previous value, indicating that a local minima has been crossed, the algorithm returns to the current minima then changes axis (90° turn on the I,Q plane) and repeats the process. This continues until the desired threshold is reached. It is necessary to introduce intermediate thresholds at which the step size is reduced to prevent the algorithm from bouncing endlessly around the minima once the step size becomes too large. Figure 19 shows a plot of LO power levels vs. iteration for the given step and threshold values (determined by trial and error) shown below. 20 trials were performed.

Step size = [.4 .2 .05 .01 .004 .002] Volts

Threshold = [-30 -45 -55 -68 -77 -80] dBm

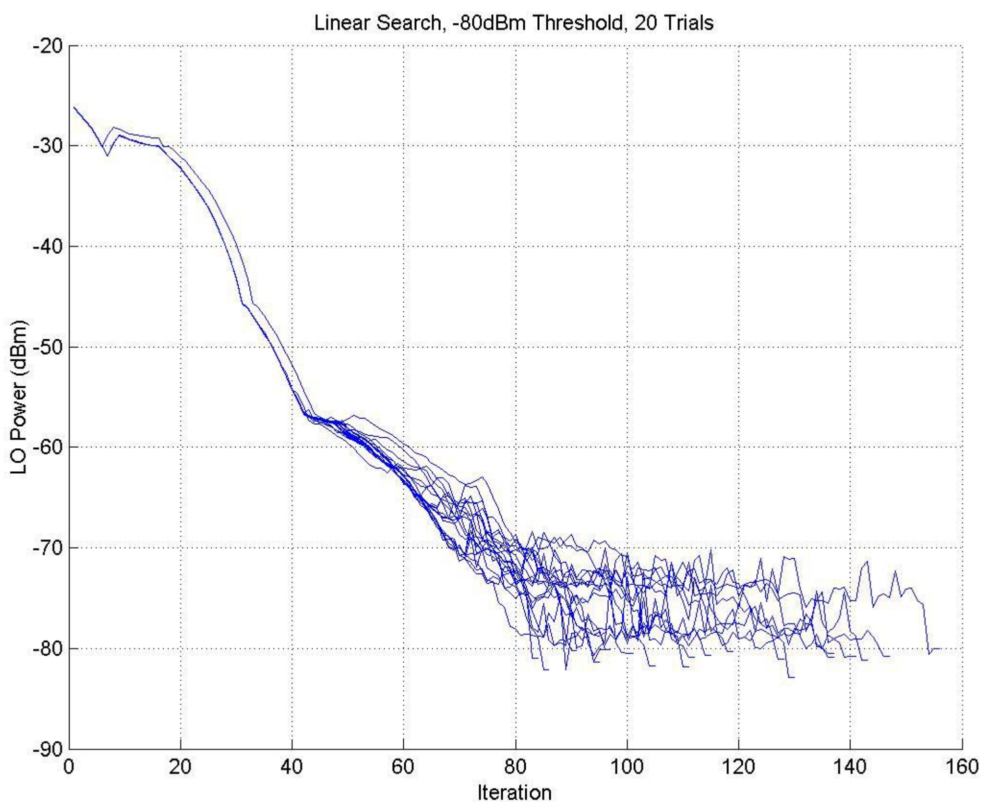

Figure 19: Linear search results for 20 trials.

Each test required between 80 and 160 iterations to reach the desired level (-80 dBm). For this particular setup a linear search method appears to be far more effective than a gradient search.

9. LO Minima Stability

The positional stability of the LO minima on the I,Q plane is sensitive to several factors. RF frequency, RF amplitude, and temperature changes can all degrade (increase) the LO leakage level. This strongly suggests that any system employing a quadrature mixer with LO nulling capability must be periodically calibrated to maintain the lowest possible LO leakage.

For any fixed RF frequency, the LO leakage level can be suppressed. However, any changes in that frequency will result in an increase in the LO power. Figure 20 through Figure 23 are a few plots of LO power levels vs. RF frequency. The plots were generated "asynchronously" (the samples are not directly related to frequency) because of the difficulty involved in automating the DDS. The spikes in the figures occur at the carrier LO frequency (3.3 GHz) and do not reflect a spike in the LO power level, they are actually caused by the RF signal as it passes through "DC" while it chirps through ±120 MHz.

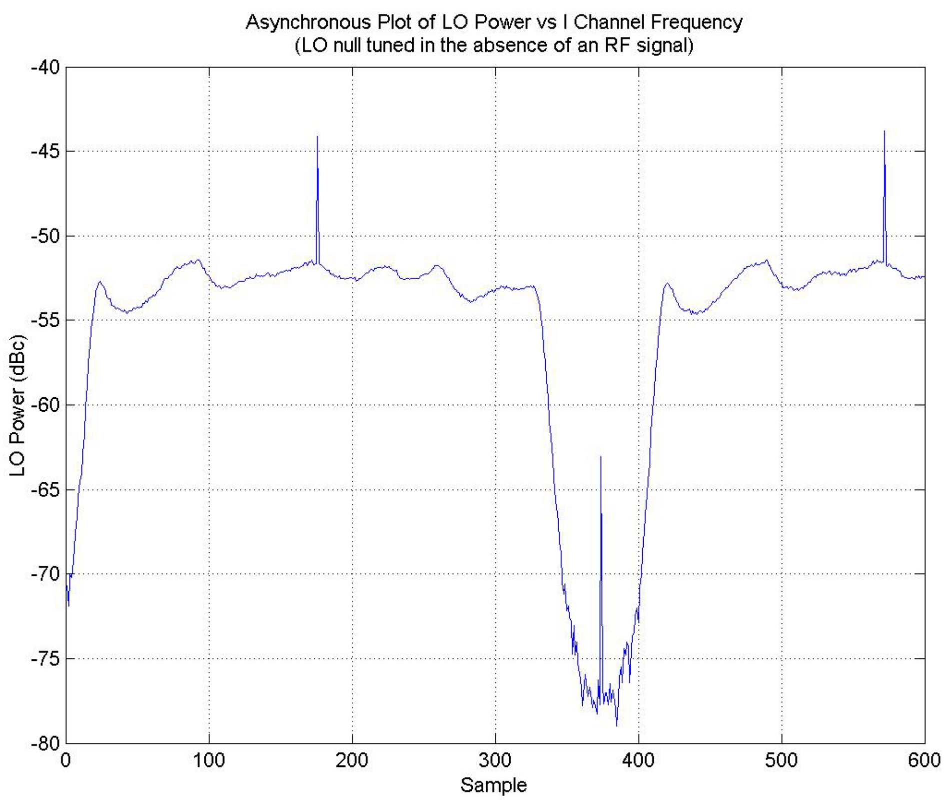

Figure 20: LO leakage power over a swept RF frequency.

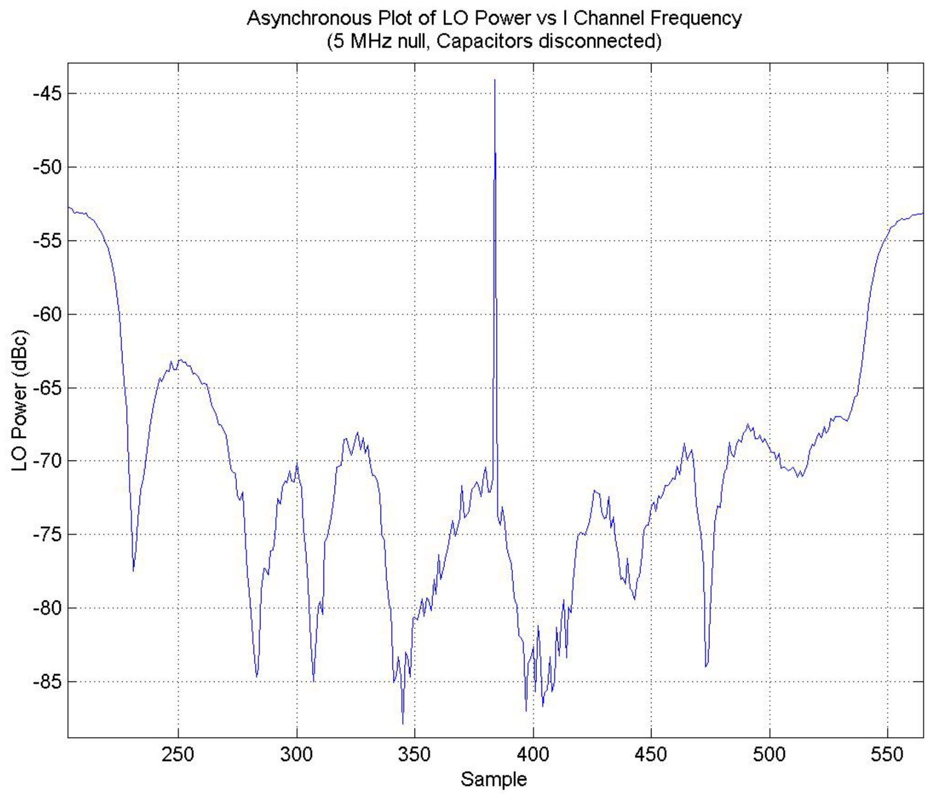

Figure 21: LO leakage power over a swept RF frequency.

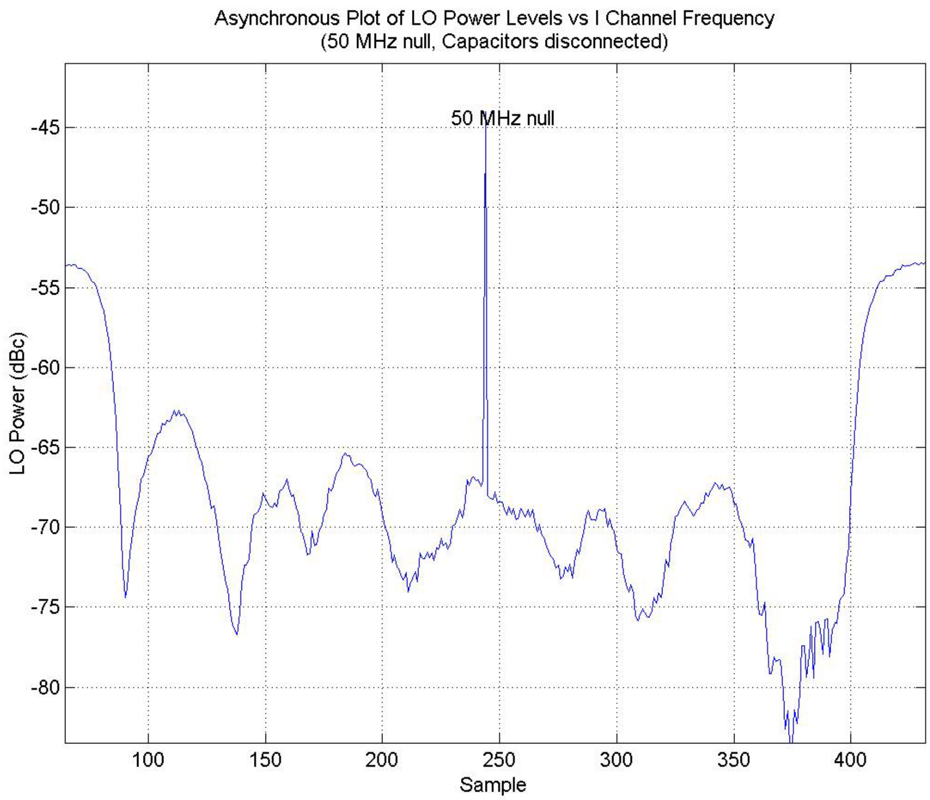

Figure 22: LO leakage power over a swept RF frequency.

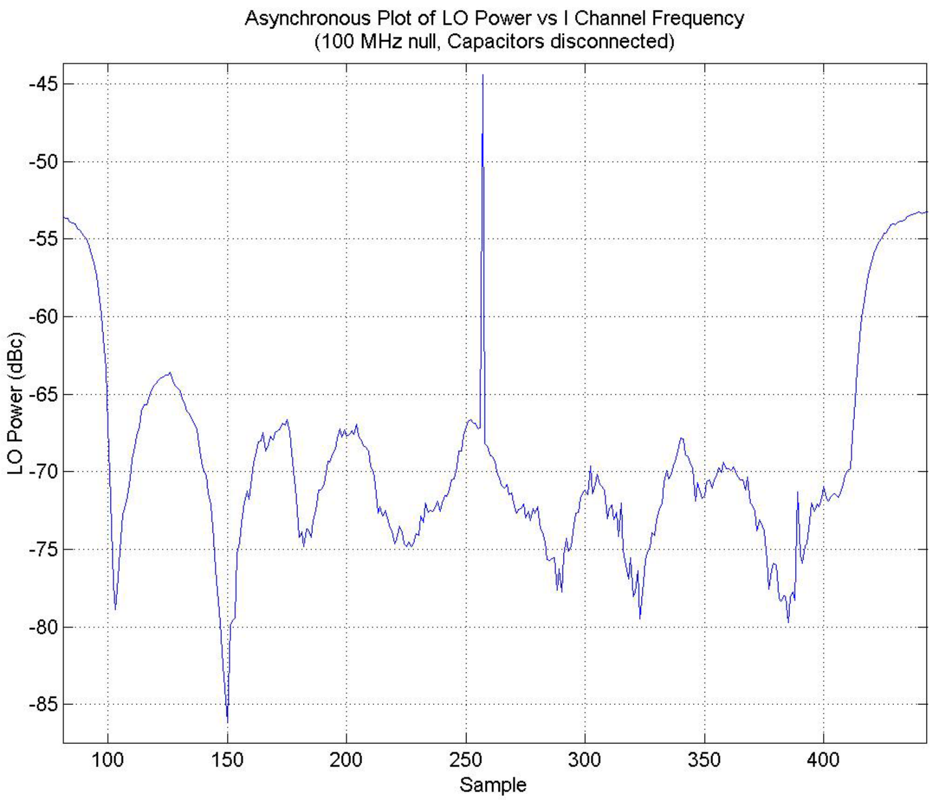

Figure 23: LO leakage power over a swept RF frequency.

It can be seen from these responses that as long as the LO null is tuned in the presence of some RF signal then the exact frequency of that signal is not very relevant. When the null is tuned in the absence of an RF signal then there is a very noticeable increase in LO power when an RF signal is introduced. However, even in the worst case the LO power level is still suppressed to below –50 dBc.

RF signal level is also shown to have an effect on the LO suppression. Figure 24 is an asynchronous plot of the LO power level vs. RF frequency for five different RF power levels. Apparently the RF amplitude is directly proportional to the level of LO suppression achievable. The LO null here has been tuned at the full-scale power of –15.9 dBm (corresponding to a hex amplitude of FFF), as the amplitude is reduced the LO power level increases.

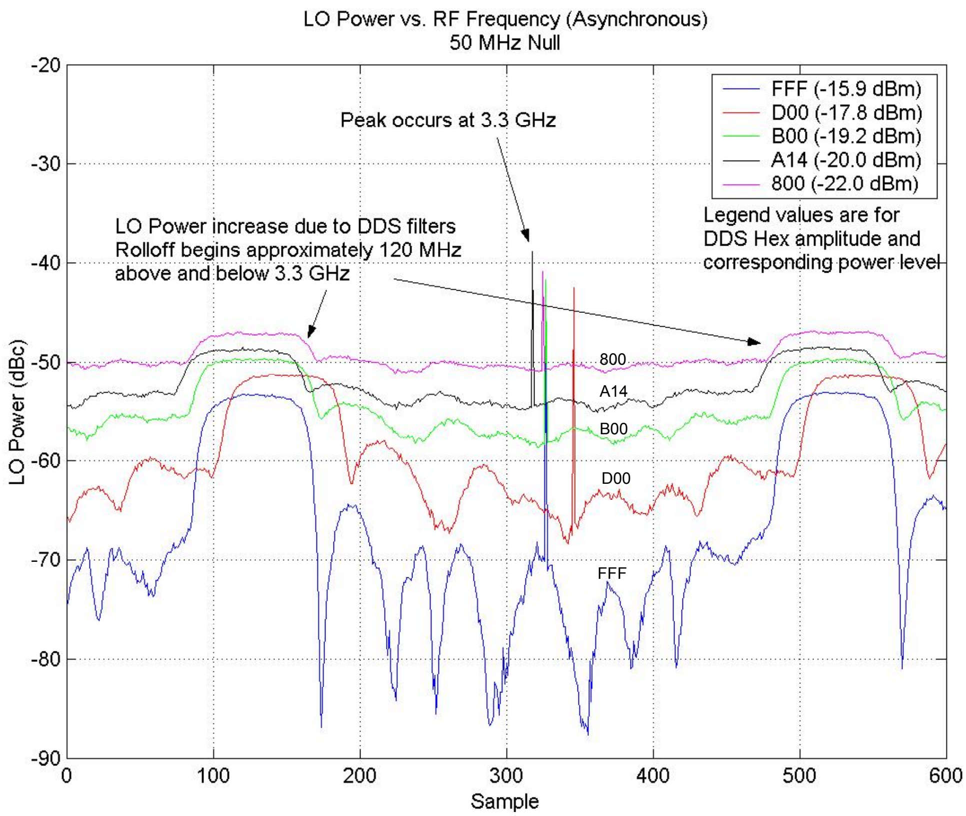

Figure 24: LO leakage power over a swept RF frequency for various mixer input power levels.

In order to test the temperature sensitivity of the LO minima the mixer was placed into a temperature chamber. With the LO feed through suppressed to approximately –58 dBc at 25 degrees C the temperature of the chamber was swept from 15° to 40° in 5° increments. Figure 25 illustrates the results. This limited sample set shows a linear increase in LO power of roughly 4 dB for every 5° offset from the initial temperature.

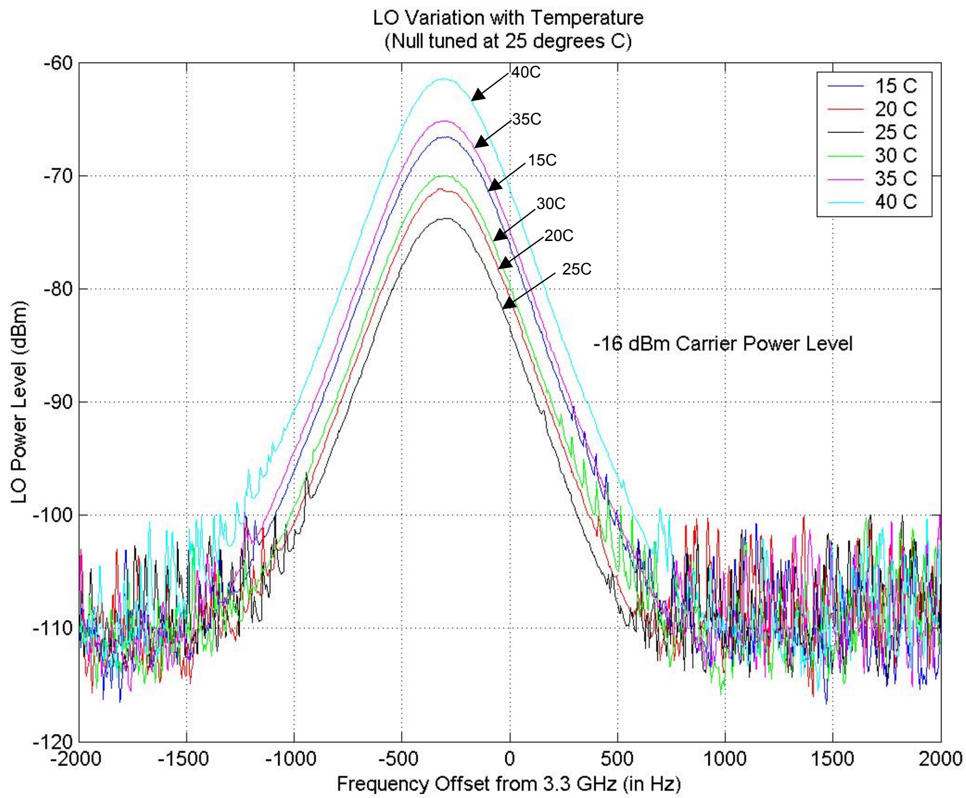

Figure 25: Temperature dependency of LO leakage.

10. Summary

By adding a DC offset to the in-phase (I) and quadrature-phase (Q) components of a synthesized signal prior to being mixed (converted to RF), it is possible to achieve and maintain LO rejection levels better than –60 dBc over a limited RF frequency and temperature range. By using a linear search algorithm to determine the DC offsets of the QECDWS outputs for the radar exciter shown in Figure 1, it is possible to consistently locate an LO minima of –80 dBm (-64 dBc) in fewer than 160 iterations. The sensitivity of the LO leakage level to changes in DC offset was found to be approximately 20 dB for every decade of voltage change (Figure 9). The power supply used to provide the DC offsets to the system was not precise enough to consistently achieve lower (<-64 dBc) LO power levels. However, manual (analog) tuning of the offsets resulted in LO suppression on the order of –85 dBc. Once the DC offsets were adjusted for an LO leakage minima, variations in the synthesized frequency over 120 MHz about the LO frequency at no point resulted in an LO leakage level greater than –50 dBc. This suggests that frequency dependent bias corrections may not be needed if –50 dBc worse-case LO leakage is acceptable.

A long-standing requirement for SAR systems built at Sandia National Laboratories is to have peak spurious levels at the RF upconverter output be ≤ -40 dBc. Referring to Figure 1, this level would have to be reduced by 12 dB at the mixer output due to the X4 frequency multiplier. This would require quadrature mixer output spurious levels, which include the LO leakage term, of ≤ -52 dBc. The results presented show that this is achievable, possibly without any synthesizer frequency dependent correction.

11. MATLAB Linear Search Algorithm

```
--------------------------------------------------------------------------------------------------
% This program locates a LO minima by tuning the I and Q channel DC offsets.
% A "linear" search algorithm is used to find this point.
% For each iteration of the loop either the I or Q offset is changed
% and the new LO power level is measured. If the new level is greater than
% the previous level then the loop breaks, and the direction of travel (+ or - V)
% is reversed. This process alternates between the I and Q channels until
% specified LO rejection level is achieved.

clear all;

load addr_list;
ud=gpib('find','gpib0');
ibsta=gpib('sic',0);
ibsta=gpib('clr',ud);
ibsta=gpib('enableremote',0);

% Power supply address is 05, Spectrum Analyzer address is 18.
%******Initialize Spectrum Analyzer
ibsta=gpib('send',0,18,'IP',2,1);        % Reset Instrument
ibsta=gpib('send',0,18,'SP100KZ',7,1);    % Frequency span to 100 kHz
ibsta=gpib('send',0,18,'S1',2,1);       % Continuous Sweep
ibsta=gpib('send',0,18,'CF3300MZ',8,1); % Center frequency to 3.3 GHz
ibsta=gpib('send',0,18,'RL0DM',5,1);    % 0 dBm reference
ibsta=gpib('send',0,18,'LG10DB',6,1);   % 10 dB/div
ibsta=gpib('send',0,18,'ST500MS',7,1);  % 500 ms sweep time
ibsta=gpib('send',0,18,'VB3KZ',5,1);    % 3 kHz video bandwidth
%******

vi=0;           % Starting I channel voltage
vq=0;           % Starting Q channel voltage
vmax=10;        % negative voltage is equal to vmax-vi or vq
lopow=0;        % LO feedthru power

tempout=[];
threshout=[];
viout=[];
vqout=[];
emcount=0;

tic;            % Start stopwatch

vistep=[.5 .4 .05 .02];         % DC power supply voltage steps
```

```
vqstep=[.5 .4 .05 .02];         % DC power supply voltage steps
lothresh=[-50 -62 -68 -72];     % LO thresholds (dBm)

for x=1:2   % after the 1st iteration the spectrum analyzer settings
         % are changed to lower the noise floor
for y=1:length(lothresh)

   while lopow > lothresh(y)
      powcheck=lopow;

      % Following loop tracks the Q axis
      loold=.1;
      qcount=0;
      while lopow < loold
         % following commands step down from current vq
         loold=lopow;        % saves to previous LO value
         vqold=vq;           % saves the previous voltage
         qcount=qcount+1;

         vq=vq+vqstep(y);
         dat2n=['vset 2,' num2str(vq)];    % Instructions for supply 2 - negative
         dat3n=['vset 3,' num2str(vmax-vq)];   % Instructions for supply 3 - negative
         ibsta=gpib('send',0,05,dat2n,length(dat2n),1);
         ibsta=gpib('send',0,05,dat3n,length(dat3n),1);
         ibsta=gpib('send',0,18,'TS',2,1);
         ibsta=gpib('send',0,18,'E1',2,1);
         ibsta=gpib('send',0,18,'MA',2,1);
         [lopow,ibsta]=gpib('receive',0,18,6,1); % vqneg is the value at the negative step
         lopow=str2num(lopow);

         tempout=[tempout lopow];
         vqout=[vqout vq];
         threshout=[threshout lothresh(y)];

         if lopow < lothresh(y)
            break
         end

      end
      %lopow=loold;
      %vq=vqold;

      %tempout=[tempout lopow];

      vqstep(y)=vqstep(y)*-1;
```

```
    % Following loop tracks the I axis
    loold=.1;
    icount=0;
    while lopow < loold
        % following commands step down from current vq
        loold=lopow;        % saves the previous LO value
        viold=vi;           % saves the previous voltage
        icount=icount+1;

        vi=vi+vistep(y);
        dat4n=['vset 1,' num2str(vi)];    % Instructions for supply 2 - negative
        dat1n=['vset 1,' num2str(vmax-vi)];   % Instructions for supply 3 - negative
        ibsta=gpib('send',0,05,dat4n,length(dat4n),1);
        ibsta=gpib('send',0,05,dat1n,length(dat1n),1);
        ibsta=gpib('send',0,18,'TS',2,1);
        ibsta=gpib('send',0,18,'E1',2,1);
        ibsta=gpib('send',0,18,'MA',2,1);
        [lopow,ibsta]=gpib('receive',0,18,6,1); % vqneg is the value at the negative step
        lopow=str2num(lopow);

        tempout=[tempout lopow];
        viout=[viout vi];
        threshout=[threshout lothresh(y)];

        if lopow < lothresh(y)
            break
        end

    end
    %lopow=loold;
    %vi=viold;

    vistep(y)=vistep(y)*-1;

    %tempout=[tempout lopow];

    % Checks should lift search out of 'ruts'
    if powcheck == lopow
        emcount=emcount+1;
        if emcount > 4
            vi=vi+vistep*2.5;
            vq=vq+vqstep*2.5;
            emcount=0;
        end
    end
end
```

```
end
tempout=[tempout lopow];
threshout=[threshout lothresh(y)];

% Spectrum analyzer is readjusted to lower noise floor
%*********
ibsta=gpib('send',0,18,'SP50KZ',6,1);    % Frequency span to 50 kHz
ibsta=gpib('send',0,18,'RL-60DM',7,1);    % -60 dBm reference
ibsta=gpib('send',0,18,'LG5DB',6,1);   % 5 dB/div
ibsta=gpib('send',0,18,'VB100HZ',7,1);   % 100 Hz video bandwidth
ibsta=gpib('send',0,18,'ST2SC',5,1);    % 2 sec sweep time
%*********

vistep=[.006 .003];         % New voltage steps
vqstep=[.006 .003];
lothresh=[-85 -90];         % New LO threshold

end

plot(tempout); grid on; hold on; plot(threshout,'r'); hold off;
xlabel('Iterations'); ylabel('LO Power (dBm)');
title('Liner Search for LO Null');
T=toc;
```
----

12. MATLAB Gradient Search Algorithm

```
----------------------------------------------------------------------------------------------------
% This program implements a gradient search to locate the minimum LO feedthru
% 6 August 01

clear all;

load addr_list;
ud=gpib('find','gpib0');
ibsta=gpib('sic',0);
ibsta=gpib('clr',ud);
ibsta=gpib('enableremote',0);

% Power supply address is 05, Spectrum Analyzer address is 18.
%******Initialize Spectrum Analyzer
ibsta=gpib('send',0,18,'SP300KZ',7,1);    % Frequency span to 3 MHz
ibsta=gpib('send',0,18,'S1',2,1);       % Continuous Sweep
ibsta=gpib('send',0,18,'CF3300MZ',8,1); % Center frequency to 3.3 GHz
ibsta=gpib('send',0,18,'RL0DM',5,1);    % 0 dBm reference
ibsta=gpib('send',0,18,'LG10DB',6,1);   % 10 dB/div
%******

vdelt=.05;      % Voltage delta
u=.08;          % Convergence Constant
zqstep=u;   % Voltage step for the case when there is no change in power
zistep=u;
vi=0;           % Starting I channel voltage
vq=0;           % Starting Q channel voltage
vmax=10;        % negative voltage is equal to vmax-vi or vq
lopow=0;        % LO feedthru power
lothresh=-65;   % Cutoff threshold for LO signal power

tempout=[];
viout=[];
vqout=[];

tic;            % Start stopwatch
while lopow > lothresh

   % Supplies 2 and 3 control the Q channel offset

   % following commands step down from current vq
   vsn=vq-vdelt/2;
   dat2n=['vset 2,' num2str(vsn)];    % Instructions for supply 2 - negative
```

```
dat3n=['vset 3,' num2str(vmax-vsn)];   % Instructions for supply 3 - negative
ibsta=gpib('send',0,05,dat2n,length(dat2n),1);
ibsta=gpib('send',0,05,dat3n,length(dat3n),1);
ibsta=gpib('send',0,18,'TS',2,1);
ibsta=gpib('send',0,18,'E1',2,1);
ibsta=gpib('send',0,18,'MA',2,1);
[vqneg,ibsta]=gpib('receive',0,18,6,1); % vqneg is the value at the negative step
vqneg=str2num(vqneg);

if vqneg <= lothresh
   break
end

% following commands step up from current vq
vsp=vq+vdelt/2;
dat2p=['vset 2,' num2str(vsp)];    % Instructions for supply 2 - positive
dat3p=['vset 3,' num2str(vmax-vsp)];   % Instructions for supply 3 - positive
ibsta=gpib('send',0,05,dat2p,length(dat2p),1);
ibsta=gpib('send',0,05,dat3p,length(dat3p),1);
ibsta=gpib('send',0,18,'TS',2,1);
ibsta=gpib('send',0,18,'E1',2,1);
ibsta=gpib('send',0,18,'MA',2,1);
[vqpos,ibsta]=gpib('receive',0,18,6,1); % vqpos is the value at the positive step
vqpos=str2num(vqpos);

if vqpos <= lothresh
   break
end

% following commands determine new value of vq
qdir=vqneg-vqpos;   % increase V for qdir>0, decrease V for qdir<0
if qdir ~= 0
   vq=vq+u*sign(qdir)/abs(qdir);
   zqstep=u*sign(qdir)/abs(qdir);
elseif (qdir==0) & (lopow >= lothresh)
   vq=vq+zqstep;       % Assumes the ideal voltage has not been passed yet
end

vqout=[vqout vq];

% Supplies 1 and 4 control the I channel offset
% following commands step down from current vi
vsn=vi-vdelt/2;
dat1n=['vset 1,' num2str(vmax-vsn)];    % Instructions for supply 1 - negative
dat4n=['vset 4,' num2str(vsn)];   % Instructions for supply 4 - negative
ibsta=gpib('send',0,05,dat1n,length(dat1n),1);
```

```
ibsta=gpib('send',0,05,dat4n,length(dat4n),1);
ibsta=gpib('send',0,18,'TS',2,1);
ibsta=gpib('send',0,18,'E1',2,1);
ibsta=gpib('send',0,18,'MA',2,1);
[vineg,ibsta]=gpib('receive',0,18,6,1); % vineg is the value at the negative step
vineg=str2num(vineg);

if vineg <= lothresh
   break
end

% following commands step up from current vi
vsp=vi+vdelt/2;
dat1p=['vset 1,' num2str(vsp)];    % Instructions for supply 1 - positive
dat4p=['vset 4,' num2str(vmax-vsp)];   % Instructions for supply 4 - positive
ibsta=gpib('send',0,05,dat1p,length(dat1p),1);
ibsta=gpib('send',0,05,dat4p,length(dat4p),1);
ibsta=gpib('send',0,18,'TS',2,1);
ibsta=gpib('send',0,18,'E1',2,1);
ibsta=gpib('send',0,18,'MA',2,1);
[vipos,ibsta]=gpib('receive',0,18,6,1); % vipos is the value at the positive step
vipos=str2num(vipos);

if vipos <= lothresh
   break
end

% following commands determine new value of vi
idir=vineg-vipos;   % increase V for qdir>0, decrease V for qdir<0
if idir ~= 0
   vi=vi+u*sign(idir)/abs(idir);
   zistep=u*sign(idir)/abs(idir);
elseif (idir==0) & (lopow >= lothresh)
   vi=vi+zistep;           % Assumes the ideal voltage has not been passed yet
end
viout=[viout vi];

% Collect the new value of the LO feedthru
dat1=['vset 1,' num2str(vmax-vi)];      % Instructions for supply 1
dat2=['vset 2,' num2str(vq)];           % Instructions for supply 2
dat3=['vset 3,' num2str(vmax-vq)];      % Instructions for supply 3
dat4=['vset 4,' num2str(vi)];           % Instructions for supply 4
ibsta=gpib('send',0,05,dat1,length(dat1),1);
ibsta=gpib('send',0,05,dat2,length(dat2),1);
ibsta=gpib('send',0,05,dat3,length(dat3),1);
ibsta=gpib('send',0,05,dat4,length(dat4),1);
```

```
    ibsta=gpib('send',0,18,'TS',2,1);
    ibsta=gpib('send',0,18,'E1',2,1);
    ibsta=gpib('send',0,18,'MA',2,1);
    [lopow,ibsta]=gpib('receive',0,18,6,1);
    lopow=str2num(lopow);

    tempout=[tempout lopow];

end
toc;            % Stop stopwatch
T=toc;

plot(tempout);
```

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