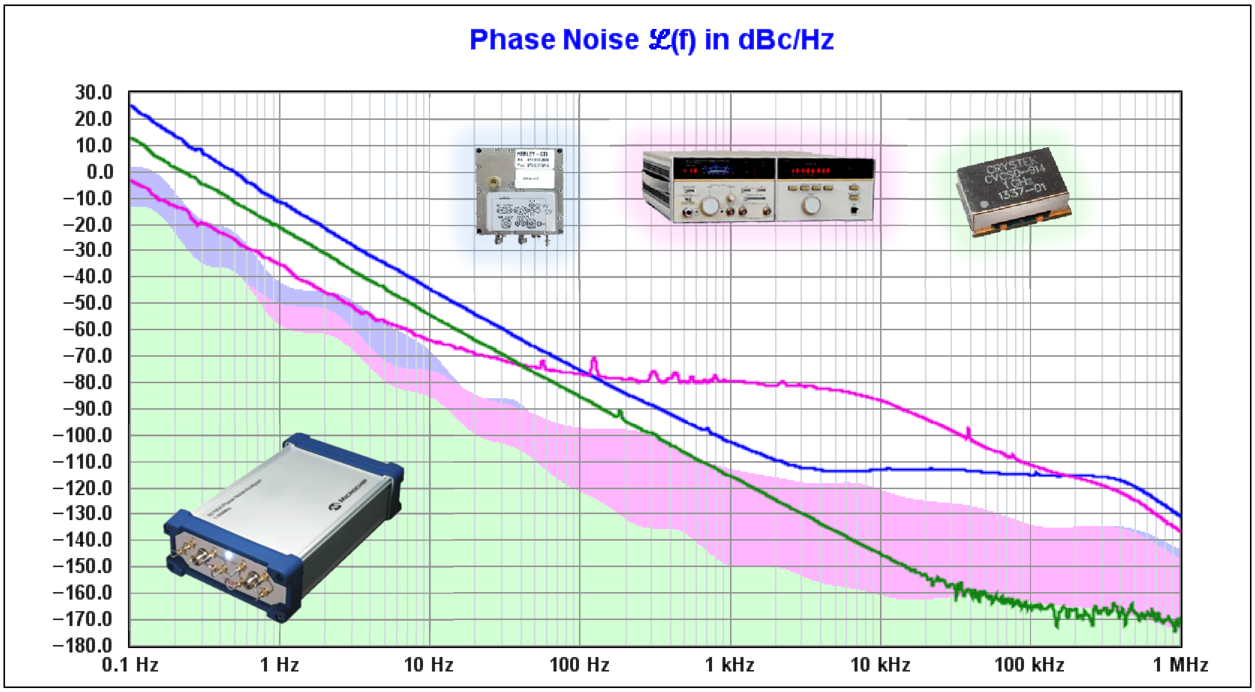
AN3899

UHF and Microwave Measurements with the 53100A Phase Noise Analyzer

1.0 INTRODUCTION

While the Microchip 53100A Phase Noise Analyzer is engineered to characterize the most demanding HF and VHF signal sources, many applications require similar measurement capabilities at higher carrier frequencies. This application note discusses how to take advantage of the 53100A's class-leading performance and ease-of-use at carrier frequencies well beyond its 200 MHz specification limit.

Three signal sources will be characterized by the 53100A in this demonstration: a voltage-controlled SAW oscillator (VCSO) at 1 GHz, a dielectric resonant oscillator (DRO) at 11.4 GHz, and a commercial signal generator at 18 GHz.[1] Parameters such as phase noise, AM noise, integrated jitter, and frequency stability will be measured with minimal additional equipment and operator workload.



To measure signals at carrier frequencies beyond the 53100A's 200 MHz specification limit, it's necessary to transform UHF and microwave signals from the device under test (DUT) to a frequency within the 1 MHz to 200 MHz range. Two different strategies for accomplishing the necessary frequency transformation will be discussed in this application note: frequency division and heterodyne downconversion.

Note 1: Third-party manufacturers and part numbers referenced in this note are cited for informational purposes and do not imply endorsement or recommendation by Microchip Technology Inc. All trademarks are property of their respective holders.

2.0 FREQUENCY DIVISION

The first technique is also the simplest one, in which a frequency divider IC is used to establish a division factor (N) that's high enough to yield a frequency within the 53100A's coverage range when driven by the DUT.

As with other 53100A applications, the frequency at the 53100A's own DUT input jack is largely arbitrary since it does not need to be phase-locked to the reference or other sources. However, measurement frequencies in the 53100A's lowest band (1 MHz to 50 MHz) are recommended for best performance. The 5 MHz to 15 MHz range works especially well with commonly available 10 MHz reference sources. Using the frequency division method as depicted in Figure 1 below, the source to be measured is connected to the input of a frequency divider whose division factor is chosen to yield an output in the target frequency range. The 53100A measures the divider's output signal against the user-supplied reference source.

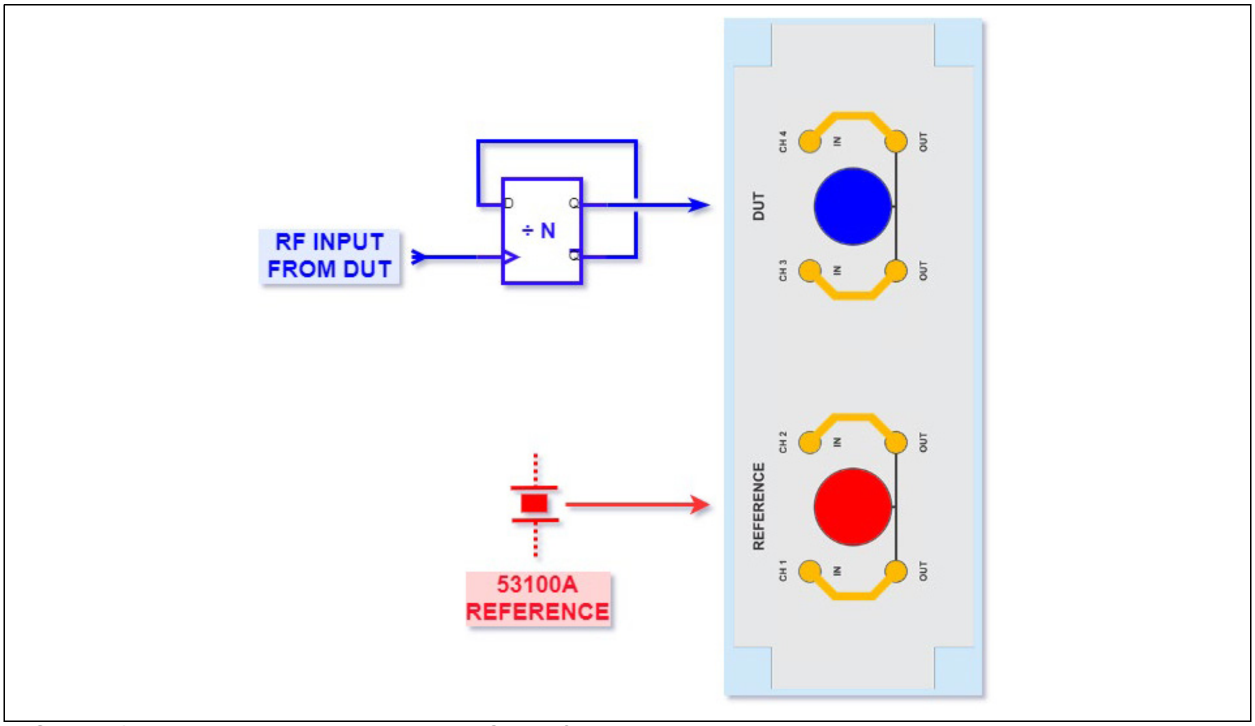


FIGURE 1: Basic Measurement Setup for Frequency Division. The D Flip-Flop Represents an Arbitrary Division of the DUT Frequency by N.

The required control settings in the 53100A's acquisition dialog box are shown in Figure 2. Select Frequency divider, then enter the division ratio in the N divisor field at right. As the measurement runs, TimeLab will automatically adjust its phase noise and jitter measurements as necessary to compensate for the ratio of the specified DUT frequency and the observed frequency at the 53100A's DUT input jack. Stability measurements including ADEV and other statistics, the frequency count chart, and the phase- and frequency-difference plots will be corrected as well.

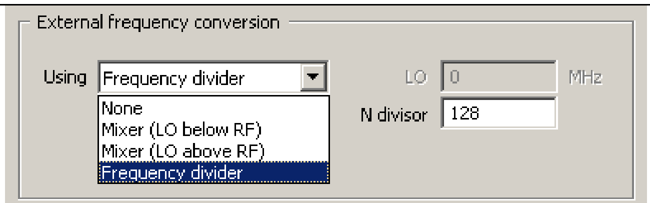


FIGURE 2: Acquisition Settings for Frequency Division.

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2.1 Measuring the 1 GHz VCSO with a Frequency Divider

Figure 3 illustrates the results of our first attempt at characterizing the phase noise of the 1000 MHz voltage-controlled SAW oscillator. To help evaluate the results, the manufacturer's typical performance figures for the part have been copied to a mask definition in TimeLab.

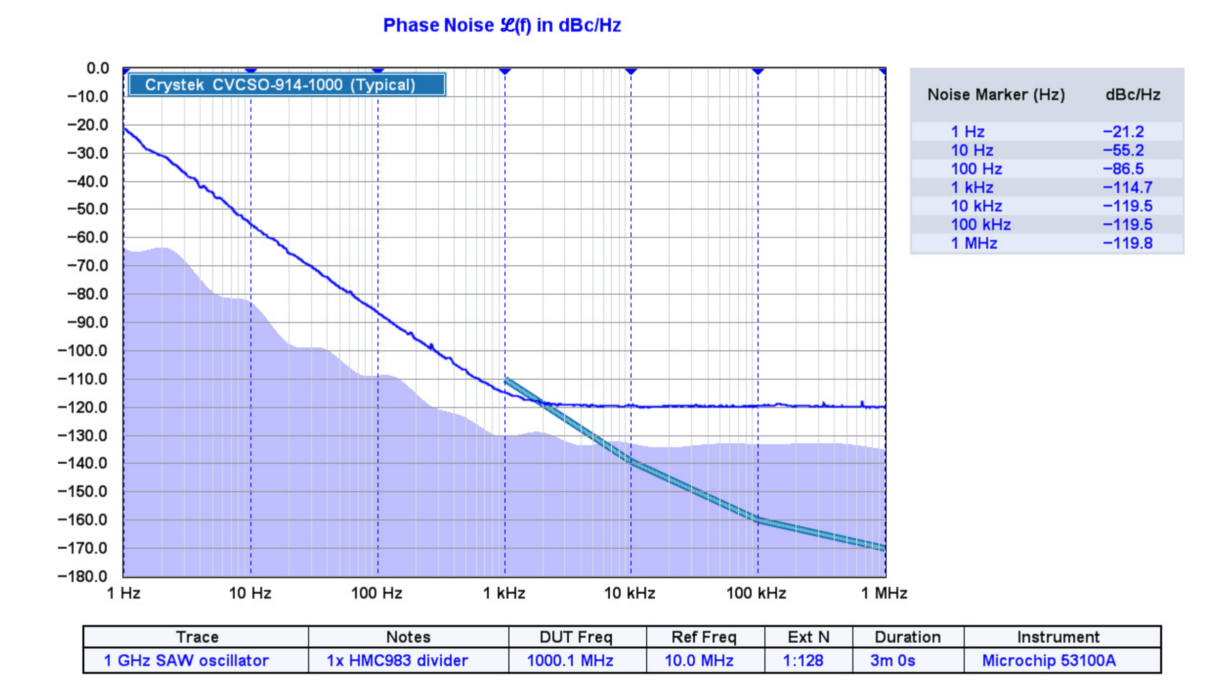

FIGURE 3: *First Attempt at Measuring the VCSO.*

The divider IC used for this experiment was the HMC983 from Analog Devices, capable of operation at input frequencies up to 7 GHz. The HMC983's division ratio is programmable in either integer-N or fractional-N steps ranging from 32 to 1,048,575. Its typical phase noise floor of –160 dBc/Hz is among the lowest available in an off-the-shelf part. It is good enough to measure the SAW oscillator at offsets approaching 100 kHz. However, our measurement fails to reach the oscillator's expected performance level at offsets as low as 2 kHz. It never goes below –120 dBc/Hz at any offset.

The reason for the inadequate performance seen above is relatively easy to understand: when a signal undergoes frequency division by a factor N, its phase noise can be expected to fall by the well-known 20 * log10(N) dB relation, or approximately 42 dB in the case where N=128. TimeLab must add this figure to the phase noise that the 53100A measures at its DUT input jack in order to display the correct phase noise level at 1000 MHz. In other words, to measure noise levels near –160 dBc/Hz at 1000 MHz with the frequency division technique, the 53100A would need to be able to measure the frequency divider's 7.8 MHz output signal at levels below –202 dBc/Hz.

Not only is this figure lower than the 53100A's typical phase noise performance floor near –180 dBc/Hz, it's also lower than the –177 dBm/Hz thermal phase noise floor, at least at reasonable carrier power levels. More importantly, it's much lower than the HMC983 itself permits us to reach. Essentially, the HMC983's banner specification of –160 dBc/Hz is an output-referred figure, not an input-referred one: no matter how quiet the DUT is, we will not be able to use the HMC983 to measure its noise at levels that would fall below –160 dBc/Hz after the 20 * log10(N) correction factor is subtracted.

At first glance, it appears that no commercially available frequency divider IC will provide the level of performance required to make this measurement.

2.2 Measuring the 1 GHz VCSO with Two Frequency Dividers

In order to measure phase noise and AM noise at extremely low levels, the 53100A uses cross spectrum averaging to suppress the noise contributed by its own internal analog-to-digital converters. In effect, the measurement is performed by two identical instruments at once, allowing uncorrelated noise to be removed over time in favor of the common-mode noise from the DUT. Because the 53100A provides individual access to all four of its ADC channels through front-panel jacks, **the cross correlation principle can also be used to extract the best possible noise-measurement performance from external devices** including oscillators, mixers, amplifiers, and frequency dividers.

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Figure 4 shows how to enhance the frequency division technique with cross correlation, along with some other recommended optimizations.

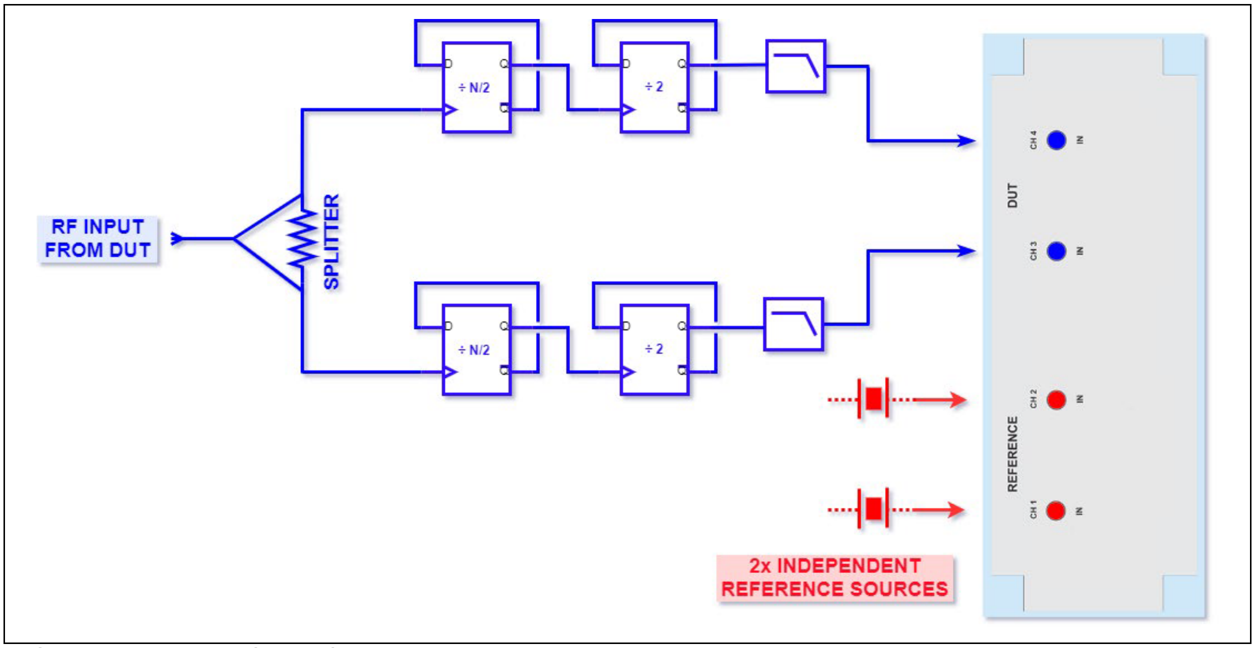

FIGURE 4: *Cross Correlated Frequency Division.*

In Figure 4, the single divider stage from Figure 1 has been replaced with two identical divider signal chains driven from an RF power splitter rated for operation at the required microwave input frequency. While a resistive or reactive splitter can be used, bear in mind that some dividers may exhibit unwanted feedthrough artifacts at their input ports, degrading the independence of the two signal paths. Wilkinson splitters have the advantage of maintaining better isolation between the two signal paths than other types can provide, and they impose less insertion loss than resistive splitters.

Looking downstream from the splitter, each divider chain consists of one stage configured for half of the desired N division ratio followed by another stage configured for division by 2. The latter guarantees a 50% output duty cycle for all supported division factors, important for obtaining the best available signal-to-noise ratio. In this example, each divider chain is implemented with the same HMC983 part already described, followed by an HMC394 5-bit counter that's hardwired for division by two.[2]

While even-order harmonics are also attenuated by the ÷2 stage in each path, most modern digital ICs produce output signals consisting of a broad spectrum of odd harmonics extending well into the GHz range. Frequency dividers are no exception. Because the 53100A is intended for use with sinusoidal input signals, additional lowpass filtering between the divider outputs and the 53100A input jacks is required for good spur performance. When strong harmonics are present in multiple Nyquist zones, the 53100A may fail to recognize the input signal altogether. Output filters used with microwave frequency divider ICs should be free of reentrant modes below 3 GHz for best results.

Once the –160 dBc/Hz noise floor of the divider ICs has been improved through cross correlation, the next major performance limitation is likely to come from the reference source. Anticipating the need for the lowest-possible white noise measurement floor, the single 10 MHz reference used earlier has been replaced with a pair of 100 MHz crystal oscillators in the test platform depicted in Figure 5.

2: Note that the rising edge of the HMC983 output is synchronized to the input clock, while the falling edge is not. As a result, the HMC983 behaves like a ripple counter when driving an analog phase detector, mixer, or ADC, introducing unnecessary jitter. Since it responds only to rising edges from the HMC983, the HMC394 has the additional benefit of reclocking the HMC983 output.

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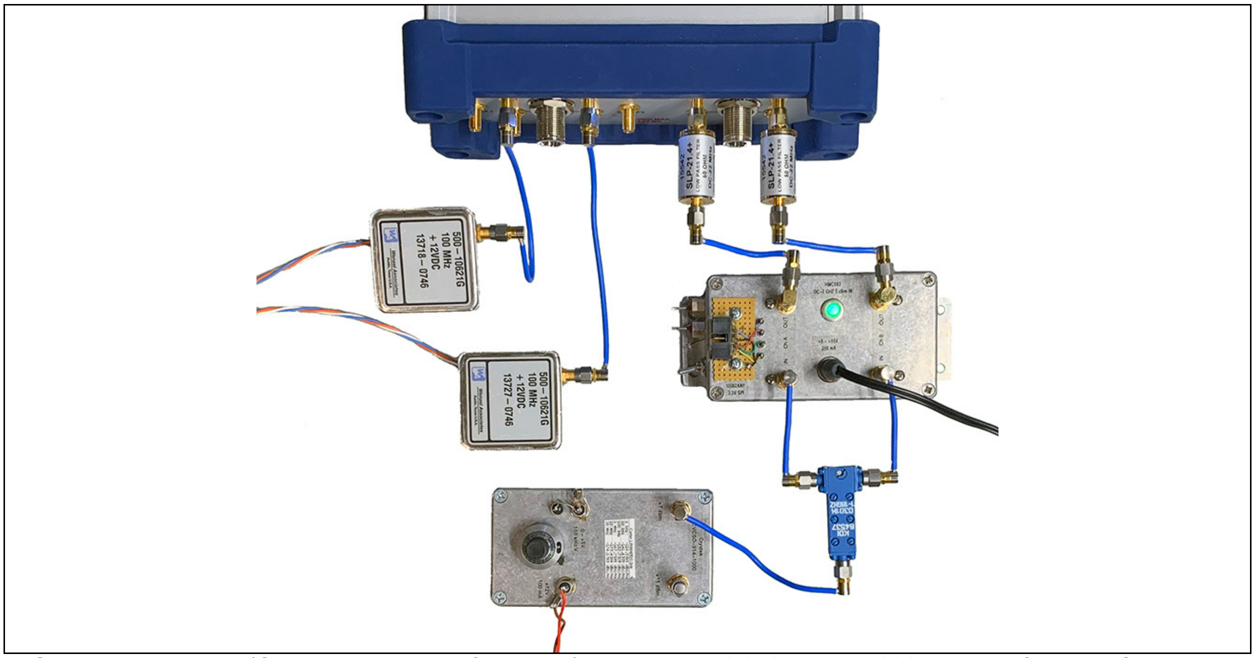

FIGURE 5: *(Counterclockwise from Top) Microchip 53100A, Dual 100 MHz Reference Oscillators, 1000 MHz SAW Oscillator Module (DUT), Microwave Power Splitter, Dual HMC983 Divider Module, 21.4 MHz Lowpass Filters (Mini-Circuits SLP-21.4).*

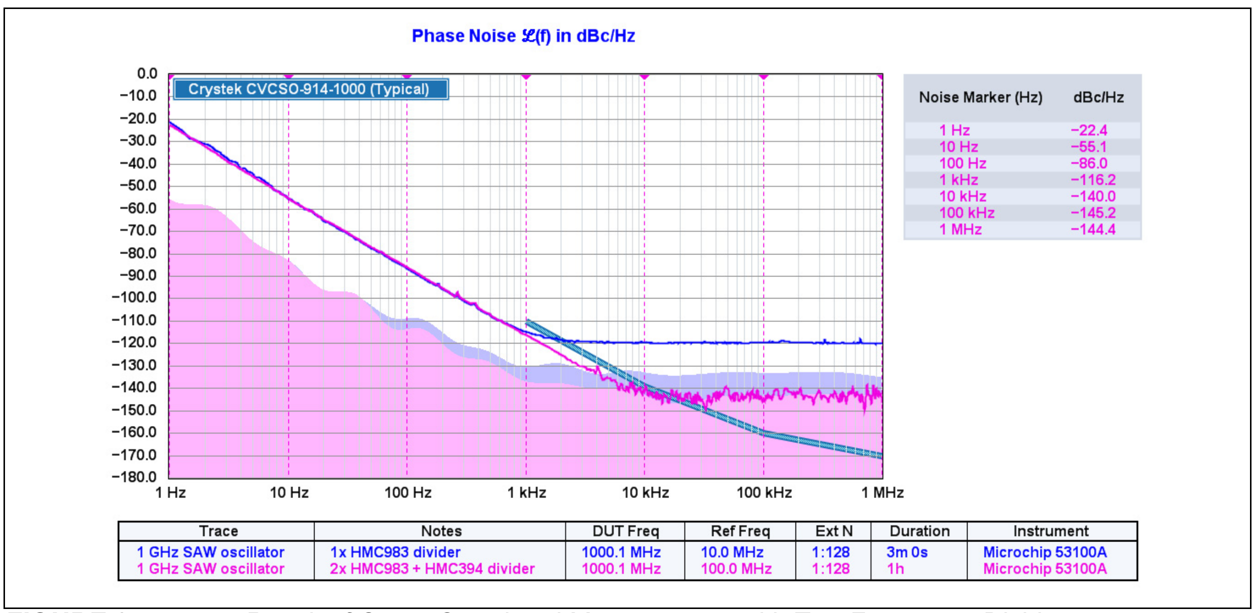

FIGURE 6: *Result of Cross-Correlated Measurement with Two Frequency Dividers.*

Figure 6 demonstrates a substantial improvement in the noise floor due to cross correlation—over 20 dB at offsets beyond 10 kHz—but the measurement once again falls short of the performance level that the manufacturer's data sheet leads us to expect from the SAW oscillator. Although the test was permitted to run for a full hour, the trace still exhibits substantial variance with smoothing turned on, and it remains close to the estimated noise floor. These are both indications that the measurement has not yet converged on its final value.

As noted earlier, this is no surprise considering that a noise floor well below –200 dBc/Hz at the divider's 7.8 MHz output frequency would be needed to give the SAW oscillator a fair trial. In this experiment, the 53100A was able to remove the additive phase noise from the two independent frequency dividers as well as the absolute phase noise of the two independent 100 MHz reference oscillators, yielding a result near –186 dBc/Hz. However, in order to display the DUT noise at the 1 GHz divider input frequency, TimeLab must apply a correction of 20 * log10(128) or 42 dB, elevating the result to approximately –144 dBc/Hz. While this is sufficient to characterize many microwave sources, the SAW oscillator isn't one of them. Devices that are expected to perform at levels below those depicted in Figure 6 must be measured using the second frequency-transformation method: downconversion.

3.0 FREQUENCY DOWNCONVERSION

Heterodyne frequency conversion or “mixing” is well known as the architectural keystone of almost every radio built since the 1920s. In a receiver or transmitter, the use of a frequency mixer and local oscillator (LO) to translate the RF signal to or from a constant intermediate frequency (IF) has the beneficial effect of separating the wideband tuned circuits near the antenna from the stages where most of the amplification and signal processing takes place. Subsequently, these stages can focus on maximizing performance at the IF rather than dealing with the entire RF coverage range.

In this application note, the heterodyne principle is employed to mix the higher-frequency signal from the DUT down to an IF that the 53100A can measure directly. The required additional stages are shown in Figure 7, where the 53100A’s front panel is depicted in its factory default configuration with all four SMA jumpers in place. A user-supplied frequency reference standard is connected to the REFERENCE input jack, while the downconverter’s IF amplifier drives the DUT input jack.

Figure 8 illustrates the actual components that will be used in our first attempt to characterize the 1 GHz VCSO with the downconversion method. These components and stages are discussed below.

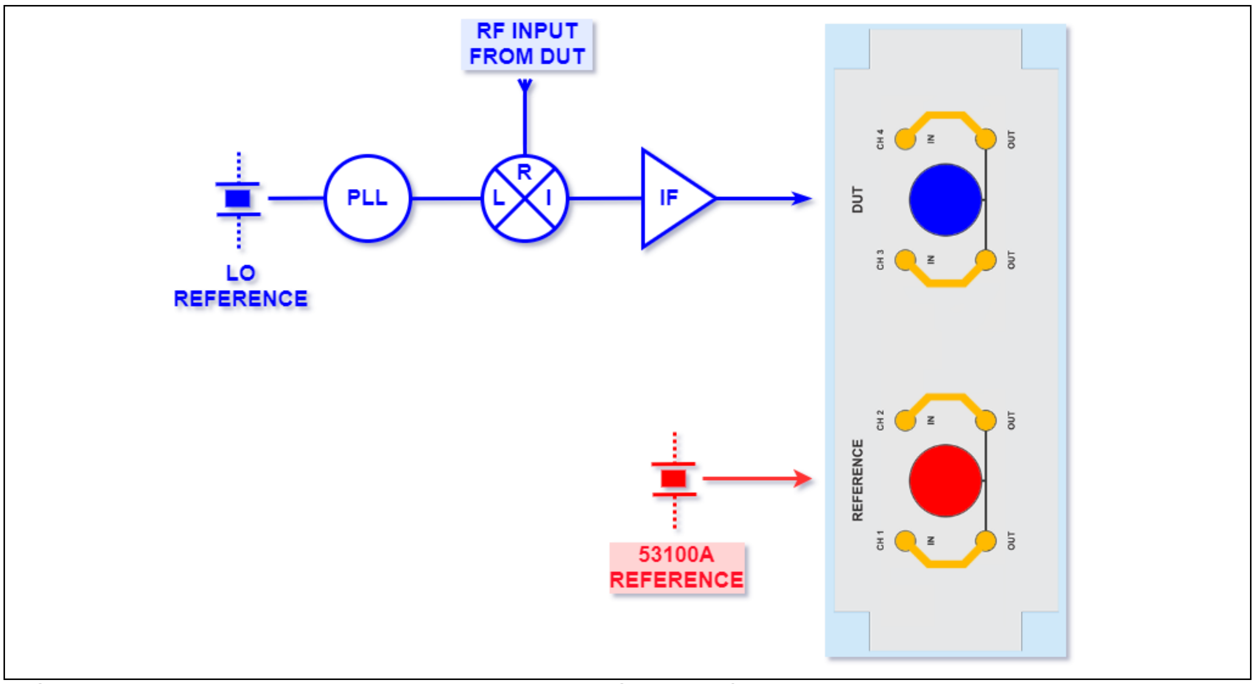

FIGURE 7: Downconverter Block Diagram Showing Connections to Microchip 53100A Front Panel.

3.1 Frequency Downconversion Hardware Setup

3.1.1 DUT

As in our earlier experiments with frequency division, the specific device under test in Figure 8 is a voltage-controlled SAW oscillator which has been mounted in a small enclosure with low noise voltage regulation and a 10-turn precision potentiometer for frequency adjustment. In addition to verifying that the SAW oscillator is capable of delivering the expected phase noise performance when packaged with real-world power supplies and other connections in place, we’ll also take the opportunity to measure its frequency stability characteristics.

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3.1.2 MIXER AND RELATED COMPONENTS

Both active and passive mixers work well in this application, but active mixers have some definite advantages. Depending on specifications, they can offer superior port isolation, relative insensitivity to termination impedance, reduced LO drive requirements, and the ability to provide conversion gain rather than loss.

The effective noise figure at the 53100A's front panel is typically 20 dB or more, determined largely by the instrument's intended use with carrier signals near the +7 dBm to +15 dBm range. Because most real-world mixers have already exceeded their input compression points when optimum signal levels for phase noise measurement are applied to their RF ports, it's often impractical to simply increase the RF input power level to make up for mixer conversion loss. Instead, the additional IF amplifier stage indicated in Figure 7 should be included, allowing sufficient gain to deliver +7 dBm to +15 dBm to the 53100A's DUT input jack after accounting for conversion loss as well as any additional attenuation that may be needed at the mixer's RF port.

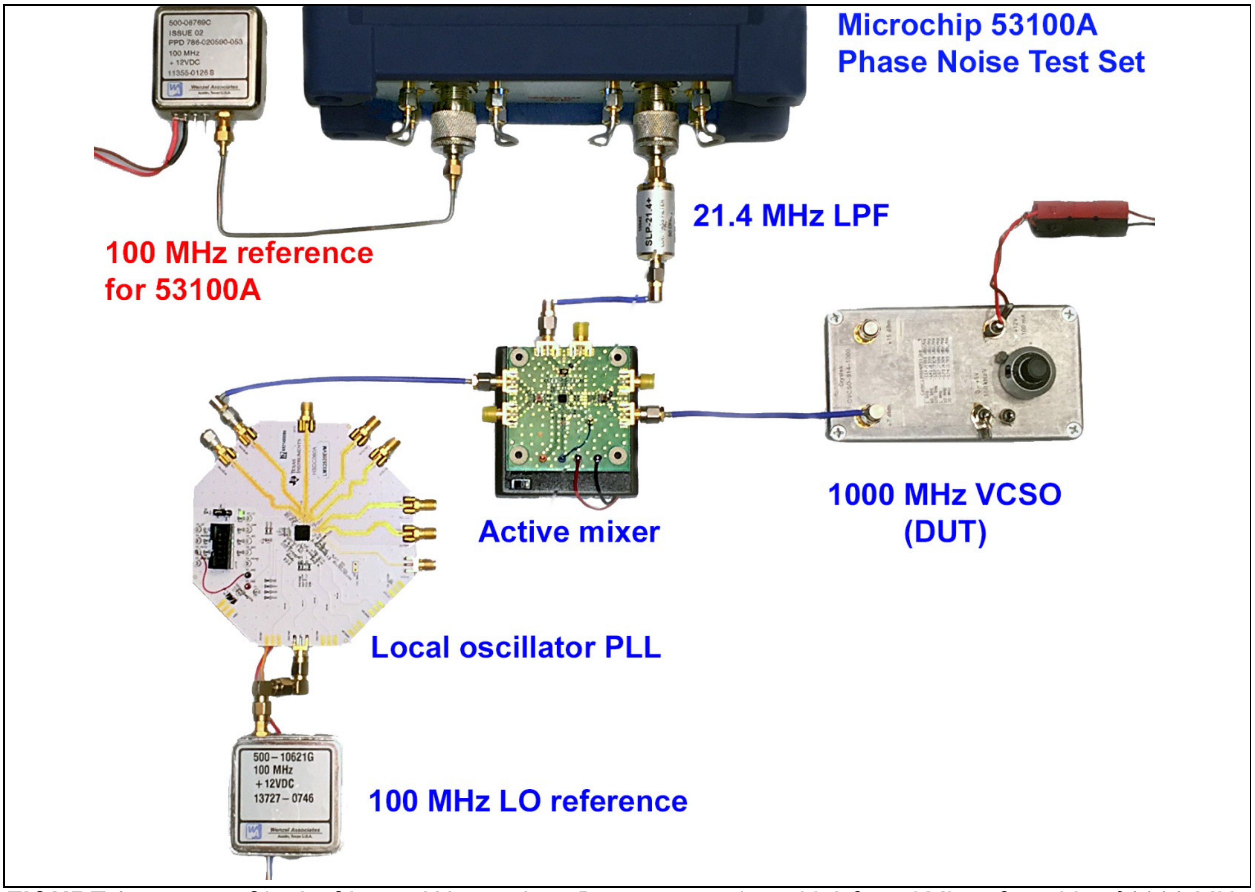

FIGURE 8: *Single-Channel Heterodyne Downconversion with LO and Mixer Capable of 6000 MHz Operation.*

The mixer used in this example is the ADL5801 from Analog Devices, Inc., installed on its customer evaluation board (ADL5801-EVALZ) and operating from four 1.2V rechargeable 'AA' cells in the battery compartment beneath the PCB. When the ADL5801-EVALZ's default single-ended connections are used in a 50Ω system, the mixer's net conversion gain approaches unity. In combination with its P1dB compression point of +13.3 dBm, this means that the ADL5801 can be used for 53100A measurements with no added RF attenuation or IF gain at all.[3]

3: For optimum dynamic range, the ADL5801-EVALZ has been modified in accordance with the manufacturer's instructions to connect the DETO and VSET terminals together.

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It should be noted that operation at input levels near P1dB is not recommended in most RF mixer applications, as intermodulation distortion often reaches objectionable levels before the mixer's compression point is reached. However, because our application has the luxury of dealing with a single CW carrier signal, operating the mixer with high-level signals at the RF port is permissible as long as the possible risks are taken into consideration. Apart from the usual IMD concerns, these risks may include AM-PM conversion, inadvertent suppression of AM noise that's intended to be measured, and reduction in long-term phase stability due to heating.

Of course, the 'single carrier signal' model is somewhat oversimplified. For example, thermal noise at the image frequency will elevate the DUT's apparent phase noise and AM noise unless a mixer topology capable of image rejection is used. While this isn't usually a concern given the 53100A's specified input signal levels and performance limits, higher IF frequencies and/or additional RF bandpass filtering may be called for when measuring synthesizers with high broadband noise floors.

Finally, to help guard against broadband spurious responses, it's a good idea to include an additional low-pass filter at the 53100A DUT input jack when working with mixers that have wideband IF outputs. The ADL5801's IF output range extends beyond 600 MHz, so a 21 MHz lowpass filter (Mini-Circuits SLP-21.4) is used here to enhance the stopband attenuation of the 53100A's internal anti-aliasing filters. External filtering isn't as important with mixer-based block diagrams as it was in the earlier frequency-division test, but it won't do any harm.

3.1.3 LOCAL OSCILLATOR PLL AND REFERENCE

When we attempt to measure the phase noise of the VCSO with the single-mixer downconverter, the local oscillator's phase noise contribution is not only capable of swamping the phase noise of a quiet DUT, it's very likely to do so. A key reason is that the LO in a general-purpose downconverter must cover a wide range of frequencies compared to devices such as SAW oscillators, crystal oscillator/multiplier stages, or other high-performance sources that target a relatively narrow frequency range. The situation is essentially the same one that crops up when attempting to measure the phase noise of a high-quality source with a traditional spectrum analyzer.

Consequently, in a single-mixer downconverter, the choice of local oscillator is determined by the types of devices to be measured. The best LO is one that's quieter than the anticipated DUT at all offsets of interest. The second-best one is a copy of the DUT itself, retuned or otherwise modified to provide a suitable IF offset from the actual DUT frequency. When neither option is feasible, the designer must weigh frequency coverage against economics in the search for the best overall compromise.

Fortunately, integrated PLL/VCO synthesizer ICs have become available with substantial improvements over previous-generation ICs from the same manufacturers. Continuous coverage from VHF to K-band microwave frequencies is afforded by parts such as the LMX2820 from Texas Instruments, ADF4371 from Analog Devices, and 8V97003 from Renesas. The PLL synthesizer employed in this note is based on the LMX2820EVM evaluation kit, but the Analog Devices or Renesas devices should serve equally well. All of these parts are capable of phase noise performance previously obtainable only from discrete synthesizers. They are well worth considering when assembling a test platform for downconversion measurements.

Some guidelines for best performance in the LO and mixer stages are listed below.

- Passive mixers capable of handling RF input signals at the +10 dBm to +15 dBm level fall into the "Level 17" class. This nomenclature refers to the mixer's minimum LO drive requirement. Such mixers will require additional LO amplification regardless of the choice of synthesizer IC. Vendors such as Mini-Circuits offer appropriate connectorized amplifier modules, noting that the LO amplifier alone will often cost more than an active mixer.
- Although modern PLL/VCO ICs include on-die voltage regulation, clean power supply rails are still a vitally important factor in noise and spur performance. Data sheets don't always acknowledge this. LDOs used with these parts should be selected with attention to both RMS noise and ripple rejection, particularly with regard to their performance at offsets above 10 kHz. If your plot includes unexplained spurs or spur clusters that seem to defy all efforts at remediation, try some additional power supply filtering.
- Use bypass capacitors at the external pins of any OCXOs employed in an open layout such as the demonstration platform described here. Only oscillators with connectorized RF outputs should be considered. This is equally true for the LO reference oscillator(s) and the 53100A's measurement reference source itself.
- If the LMX2820EVM is used for LO generation as in this example, a jumper wire should be installed between the TP_MUTE and TP_GROUND test points. The jumper will allow the board to maintain its output signal with the USB programming cable disconnected, eliminating a potential source of spurs.
- As noted earlier, the downconversion IF—meaning the signal that the 53100A actually measures—should be chosen within the 53100A's first Nyquist zone for best results. In practice, this means selecting an IF below 50 MHz, where the 53100A's internal anti-aliasing filters are most effective. Intermediate frequencies between 5 MHz and 15 MHz are suggested under most circumstances, but avoid choosing IFs near 10 MHz or other frequencies

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where interference from strong local signals may cause beat notes or other spurs.

- Also in the interest of spur performance, choose LO frequencies that allow the synthesizer IC to operate in its integer-N mode when possible. The IF used in this test was 11 MHz, with the LO synthesizer arbitrarily programmed on the high side of the RF input (1011 MHz).
- In the dual-channel downconverter variation described later, LO frequencies of 1009 MHz and 1011 MHz were used to help ensure decorrelation of synthesizer spurs. Because the 53100A's frequency stability measurements are based on the frequency of the signal at the Channel 3 input by default, the frequency of the LO associated with Channel 3 is the one that should be entered in the acquisition dialog. The Channel 4 LO frequency should be on the same side of the RF input frequency as the Channel 3 LO, but the two LO frequencies do not need to match (and generally should not match).

3.1.4 53100A REFERENCE SOURCE

As with any other measurement, the 53100A should be provided with either one or two clean, stable reference frequency sources. In this example, a single free-running 100 MHz OCXO is used as the measurement reference. This reference provides an excellent broadband noise floor for IFs in the lower HF range thanks to its 20 * log10(N) advantage over IF frequencies in the recommended 5 MHz to 15 MHz range. At offsets closer to the carrier, its stability characteristics will be adequate for testing SAW oscillators and most other free-running microwave sources.

If you need to make low-level phase noise measurements with a measurement reference in the 10 MHz range, or if your measurement is otherwise likely to approach the phase noise performance of the 53100A's reference source, consider using two independent references to allow their influence to be averaged out of the plot over time. This powerful technique, described in Application Note AN3526, works as well with downconversion measurements as it does with conventional ones. Dual references aren't as important as they are with the frequency divider method, however, because downconverted microwave sources don't usually approach the limits of a good reference OCXO.

3.2 Measuring the 1 GHz VCSO with an External Mixer and Local Oscillator

After selecting *Acquire>Microchip 53100A…* in TimeLab, the measurement parameters may be adjusted as necessary prior to data acquisition. Begin by selecting the required option for frequency downconversion with an external mixer and LO, and enter the frequency of the local oscillator as shown in Figure 9. This step allows TimeLab to calculate the DUT signal frequency at the mixer RF input and adjust the measurement results accordingly. The DUT frequency will be saved with the `.TIM` file, eliminating the potential for confusion when the measurement is reloaded later.

Most of the other settings in the 53100A acquisition dialog can be left at their default values when performing frequency downconversion measurements. However, the stability characteristics of a SAW oscillator cannot be measured in the 53100A's default 50 Hz measurement bandwidth because the DUT can be expected to drift significantly more than this over the course of the measurement. Either 5 kHz or 50 kHz is appropriate for this DUT (Figure 10), although 50 kHz was selected here to permit stability measurements to be made shortly after warmup.

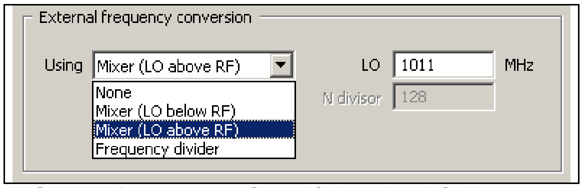

FIGURE 9: *Specifying the LO Frequency and Sideband.*

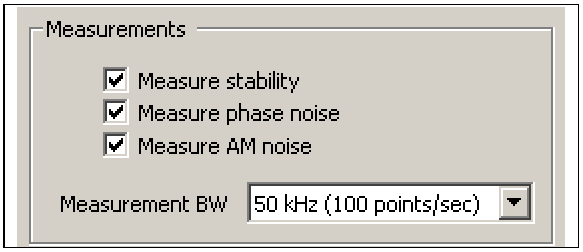

FIGURE 10: *Increasing the Stability Measurement Bandwidth.*

An additional change is made to the Duration field, increasing the test time from its default of 3 minutes to a duration sufficient to measure the short-term Allan deviation (ADEV) of the SAW oscillator over the desired tau range. A one-hour run will return meaningful ADEV results at intervals up to t=900 seconds, although in this case we'll also change the *Trace History* parameter in the *Additional Options* dialog to 10 in order to visualize the ADEV improvement over time as the device temperature stabilizes (Figure 11).

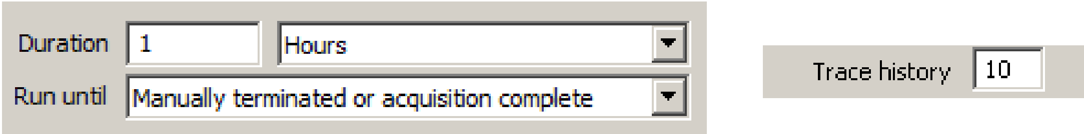

FIGURE 11: *Setting the Measurement Duration and Allan Deviation Trace History Count.*

Finally, before starting the measurement, it will be necessary to configure the PLL synthesizer or other signal source that provides the LO signal to the mixer. When the LMX2820EVM board is used as the LO source, for example, the TICS Pro software from Texas Instruments is used to initialize the PLL with its default register settings, followed by entry of a frequency that will yield a suitable IF at the mixer output (Figure 12).

FIGURE 12: *Setting the LO Frequency.*

After the desired parameter changes have been made in the acquisition dialog and the LO source has been enabled, the *Start Measurement* button will initiate data acquisition as requested.

3.3 Stability Measurement Results

Because the LO synthesizer is tuned to 1011 MHz in this example, the nominal RF and IF frequencies are 1000 MHz and 11 MHz respectively. Like many narrowband VCOs, the SAW oscillator under test achieves its best phase noise performance at higher frequencies where the varactor's contribution to the total Q factor is minimized, so these tests were conducted near the high-frequency end of its tuning range for demonstration purposes. As a result, the VCSO's output frequency was closer to 1000.120 MHz than 1000.000 MHz. After being rounded to the nearest 100 kHz for display, the reported DUT frequency in Figure 13 is 1000.1 MHz.

Frequency counts at various averaging times relative to the end of the trace are also shown in Figure 13. In this example, the frequency counts are inexact because neither the 53100A measurement reference nor the LO reference oscillator was calibrated or phase-locked at a known frequency. Given the 53100A's ability to work with free-running reference and DUT sources, measurements that don't require accurate frequency counts are much easier to set up than they are with traditional quadrature PLL instrumentation.

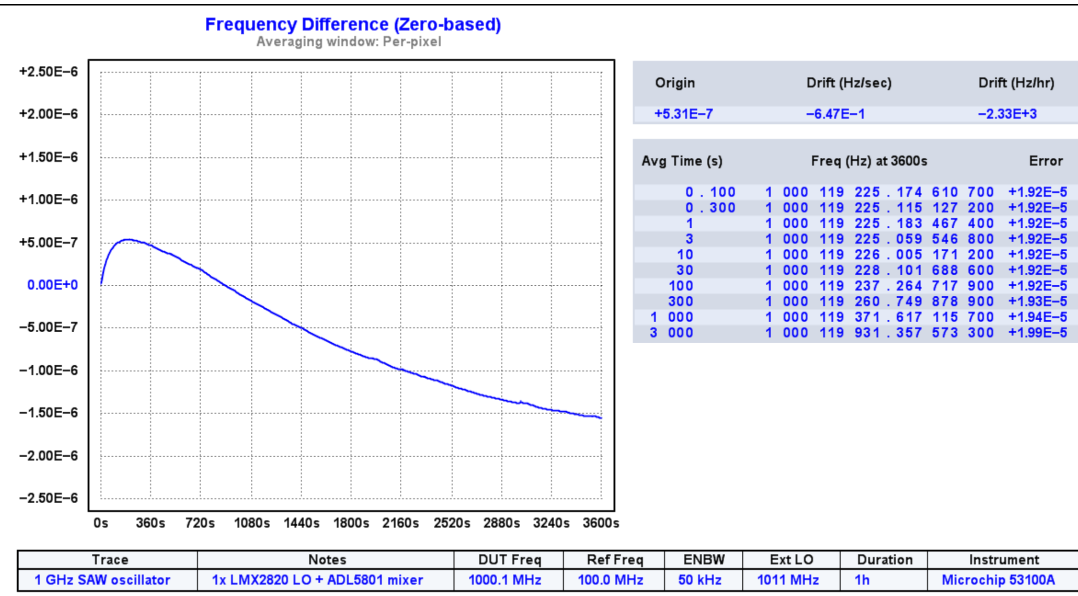

FIGURE 13: *1000 MHz VCSO Frequency Drift Shortly after Power-On.*

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The Allan deviation plot in Figure 14 is also subject to influence from the unlocked reference and LO sources. Of course, we're measuring a SAW oscillator against good-quality OCXOs, so the DUT's ADEV performance should be largely unbiased by reference instability.

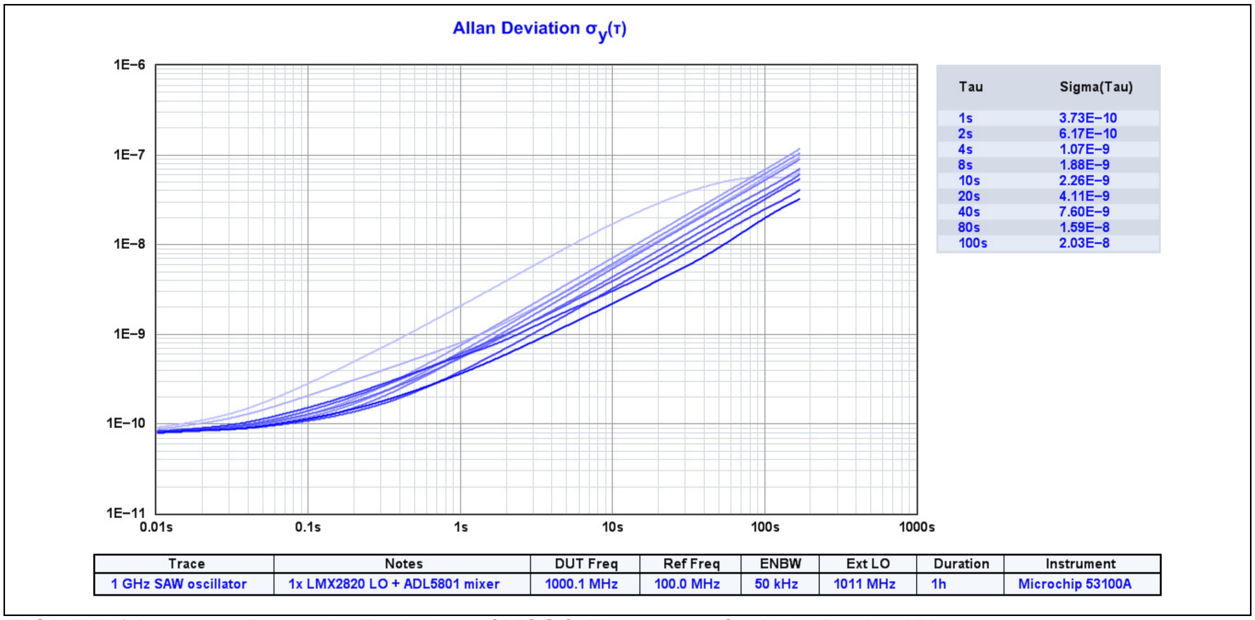

FIGURE 14: *Dynamic Evolution of VCSO Frequency Stability During Warmup.*

With *Trace History* set to 10 for a one-hour run, each trace in Figure 14 represents the DUT's stability over successive six-minute intervals. Darker traces represent data collected towards the end of the run.

3.4 Noise Measurement with Frequency Downconversion

We've obtained good stability measurement results from the single-channel downconverter, but as the green trace in Figure 15 reveals, the apparent phase noise (observed in a separate five-minute run after completion of warmup) lies about halfway between the two frequency divider measurements made earlier.

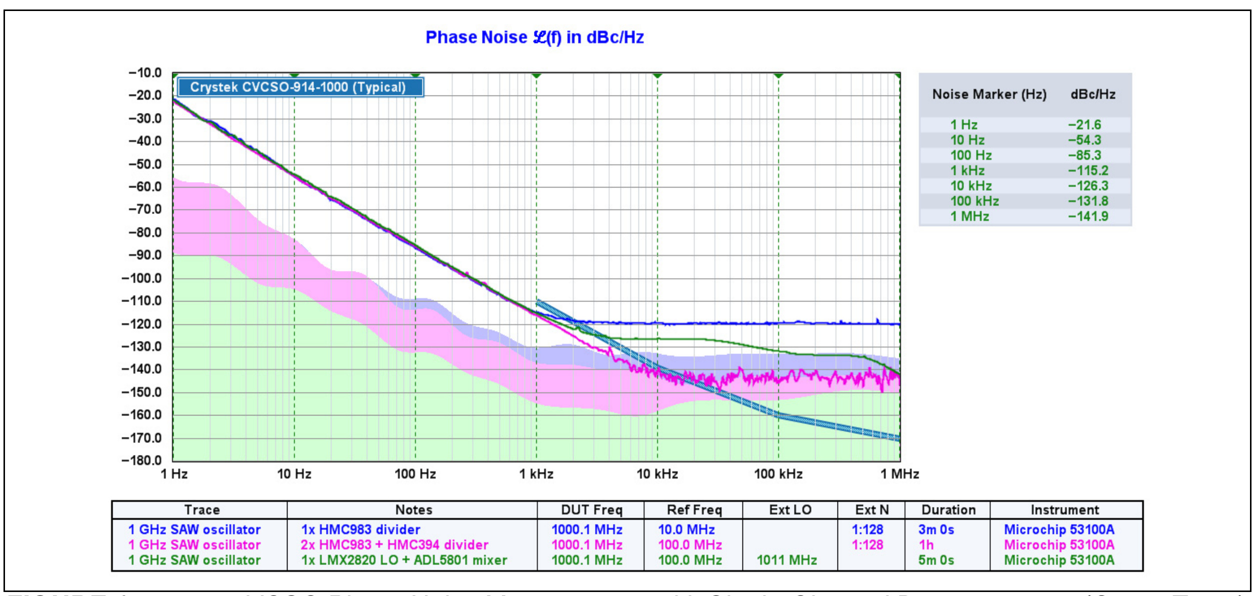

FIGURE 15: *VCSO Phase Noise Measurement with Single-Channel Downconverter (Green Trace).*

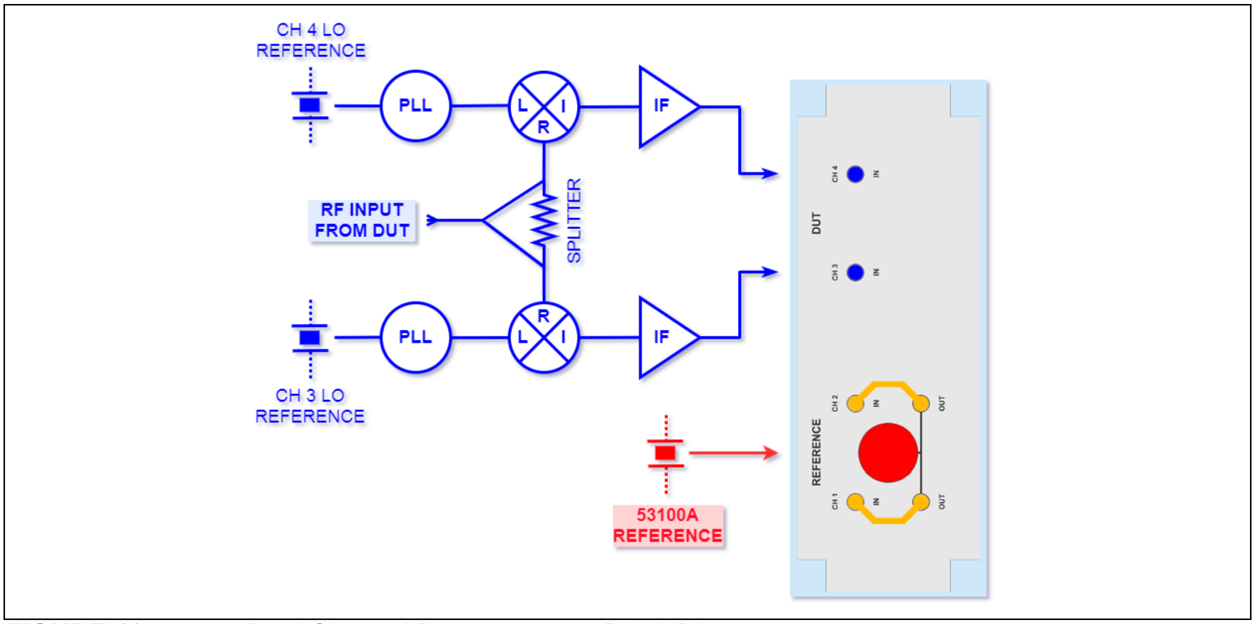

FIGURE 16: *Dual-Channel Downconverter Block Diagram.*

The measurement in Figure 15 has converged quickly, yielding a trace with low variance and plenty of headroom above the floor estimate. At offsets below 1 kHz there is good agreement with our earlier results. But beyond that point, the measured noise is coming from the LMX2820 LO synthesizer rather than the SAW oscillator under test.

Fortunately, the same strategy that was used to achieve a lower measurement floor with the frequency divider ICs can be applied with mixers and their accompanying local oscillators as well (Figure 16). To set up the test hardware in Figure 17, we've simply duplicated the existing downconverter components and added a power splitter. The SMA jumpers for the two DUT input channels (Ch 3 and Ch 4) have been removed from the 53100A's front panel, allowing each mixer to drive its own ADC channel.

Thanks to cross spectrum averaging, the instrument performance shown in Figure 18 is now good enough to confirm that the SAW oscillator achieves its typical phase noise performance at the 1 kHz, 10 kHz, 100 kHz, and 1 MHz offsets specified by its manufacturer.

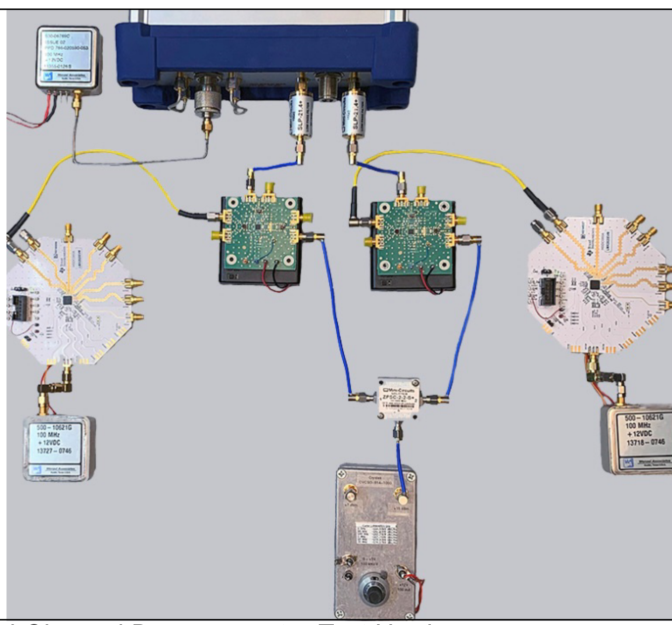

FIGURE 17: *Dual-Channel Downconverter Test Hardware.*

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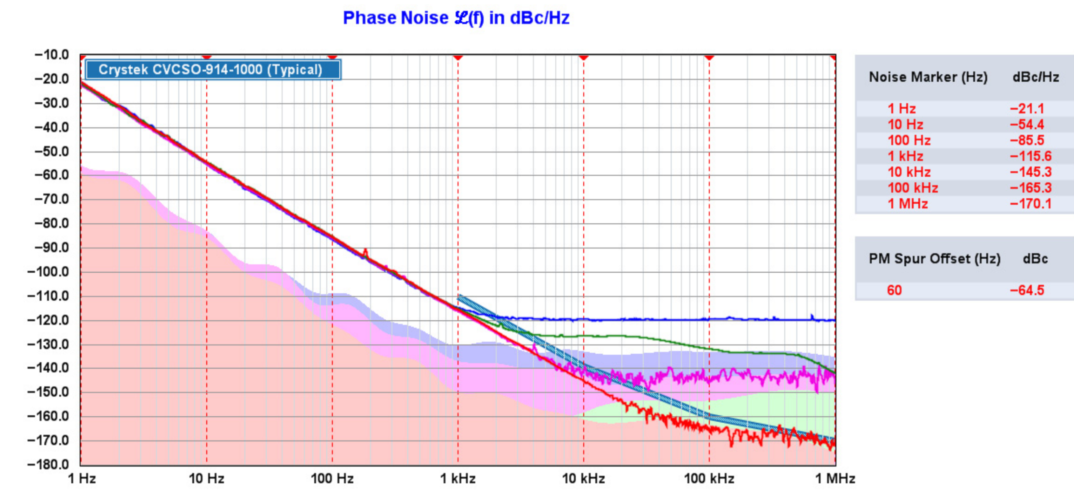


Trace	Notes	DUT Freq	Ref Freq	Ext LO	Ext N	Duration	Instrument
1 GHz SAW oscillator	1x HMC983 divider	1000.1 MHz	10.0 MHz		1:128	3m 0s	Microchip 53100A
1 GHz SAW oscillator	2x HMC983 + HMC394 divider	1000.1 MHz	100.0 MHz		1:128	1h	Microchip 53100A
1 GHz SAW oscillator	1x LMX2820 LO + ADL5801 mixer	1000.1 MHz	100.0 MHz	1011 MHz		5m 0s	Microchip 53100A
1 GHz SAW oscillator	2x LMX2820 LO + ADL5801 mixer	1000 MHz	100.0 MHz	1011 MHz		2h	Microchip 53100A

FIGURE 18: VCSO Measurement with Dual-Channel Downconversion (Red Trace) Compared to Previous Results.

In particular, note the separation between the results from the single-channel downconverter (green trace) and dual-channel implementation (red trace). Recalling that the green trace was limited by the phase noise of the LO synthesizer IC, we've achieved more than 30 dB of effective noise floor improvement at offsets above 25 kHz. There's an obvious trade-off at work: on one hand, the measurement required two hours to perform almost 4,000,000 averages, but on the other hand, all of the equipment in Figure 17 is available off the shelf at minimal expense. Using a pair of high-performance microwave synthesizers, the measurement would have taken less time at greater cost.

3.5 Measurement at Higher Frequencies

All of the techniques reviewed in this note are applicable at arbitrarily high microwave frequencies when implemented with appropriate components. As an example, the dual-channel downconverter block diagram used to measure the 18 GHz RF synthesizer and 11.4 GHz DRO source is shown in Figure 19.

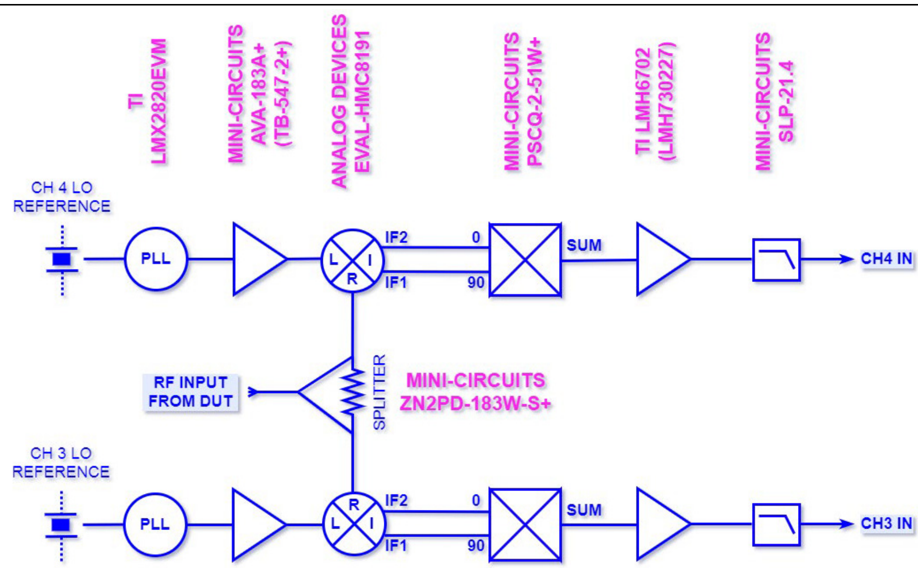


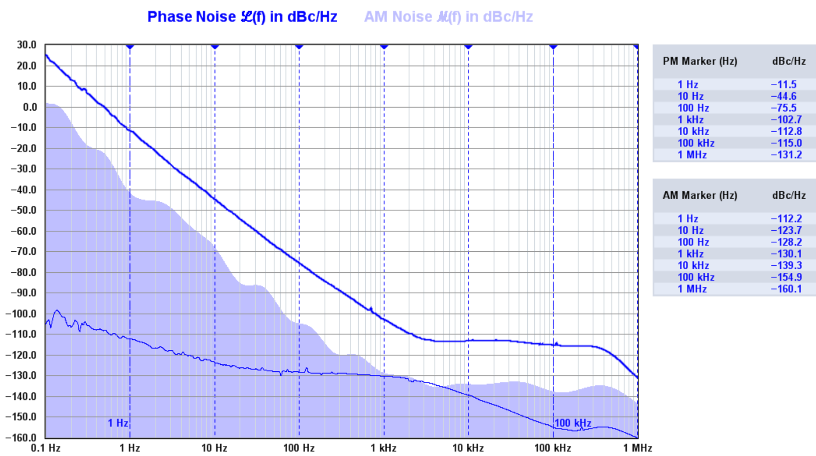
FIGURE 19: Example Block Diagram for 6 GHz to 18 GHz Dual-Channel Downconverter.

Component choices are generally not critical. The 6 GHz to 18 GHz downconverter can use the same LMX2820EVM synthesizer evaluation boards from the previous demonstration because they provide good performance through 22.6 GHz. Image-reject mixers based on Analog Devices EVAL-HMC8191 boards and Mini-Circuits PSCQ-2-51W+ quadrature hybrids were used to suppress unwanted broadband noise from the opposite sideband. The IF1 and IF2 connections shown in Figure 19 assume that a high-side LO is used to downconvert the RF signal, but can be reversed to select the other sideband if needed.

The HMC8191 mixers are passive devices with higher LO power requirements than the LMX2820EVM boards can supply. The required amplification was obtained from AVA-183A+ amplifiers, whose evaluation board can be ordered from Mini-Circuits under the part number TB-547-2+.

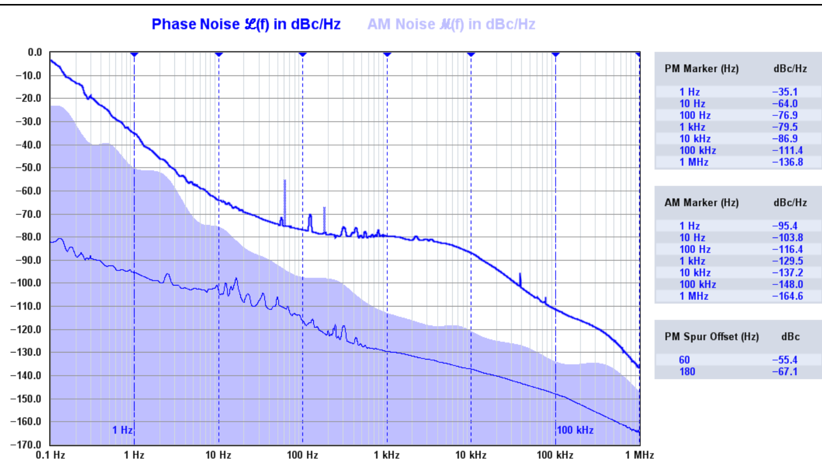
Approximately 10 dB of IF amplification is also needed in each channel to compensate for losses in the mixer and quadrature hybrid. LMH6702 amplifiers from TI were used in the test bed, but almost any low-noise gain stage that provides a good broadband termination for the mixer and hybrid should work well. The LMH6702s were used in a noninverting configuration with 50Ω series terminations at their outputs. Finally, SLP-21.4 filters from Mini-Circuits serve to keep distortion products out of the 53100A's front end.

The results of two typical measurements at 11.4 GHz and 18 GHz appear in Figure 20 and Figure 21 respectively, using the *<u>Trace>Show AM Noise in PN View</u>* <F8> function in TimeLab to render the AM noise alongside the phase noise traces. Integrated jitter between 1 Hz and 100 kHz is also shown.



Trace	Notes	DUT Freq	Ext LO	RMS Jitter	Duration	Instrument
11.4 GHz DRO	2x LMX2820 LO + HMC8191 mixer	11400 MHz	11409 MHz	3.5E−12 s	1h	Microchip 53100A

FIGURE 20: *Phase Noise, AM Noise, and Jitter of 11.4 GHz Dielectric Resonant Oscillator (DRO), Measured with Dual-Channel Microwave Downconverter.*



Trace	Notes	DUT Freq	Ext LO	RMS Jitter	Duration	Instrument
HP 8672A synthesizer	2x LMX2820 LO + HMC8191 mixer	18000 MHz	18011 MHz	1.8E−13 s	30m 0s	Microchip 53100A

FIGURE 21: *Phase Noise, AM Noise, and Jitter of Commercial Microwave Synthesizer, Measured with Dual-Channel Downconverter.*

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4.0 CONCLUSION

Using the Microchip 53100A Phase Noise Analyzer together with readily available external modules and components, we've demonstrated how to characterize high-quality signal sources at frequencies up to 18 GHz and beyond. Strategies include the use of frequency divider ICs for economical measurement of close-in noise and heterodyne downconversion for applications where low-level phase noise and AM noise must be measured at a wide range of offset frequencies.

For the most demanding applications, we've shown how to leverage the 53100A's multichannel cross correlation capabilities to take full advantage of both the frequency division and downconversion methodologies. With the techniques outlined here, all of the 53100A's capabilities—ranging from high precision frequency measurement, trend analysis, and deviation statistics to phase noise, AM noise, and integrated jitter—are now available in the UHF/microwave spectrum.

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ISBN: 978-1-5224-7863-8

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