Frequency multiplier response to spurious signals and its effect on local oscillator systems in millimeter and submillimeter wavelengths

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ABSTRACT

Broadband fixed-tuned frequency multipliers in conjunction with broadband power amplifiers driven by frequency synthesizers are often used as local oscillator (LO) sources in the millimeter and submillimeter wave heterodyne instruments. At these frequencies the multipliers use Gallium Arsenide (GaAs) based Schottky varactor diodes as the nonlinear element, and like most other harmonic generators are susceptible to spurious signal interference. The state-of-the-art LO sources in the millimeter and submillimeter wavelengths use MMIC power amplifiers producing in excess of 250 mW of output power in the 100 GHz range, and they are used to drive the subsequent multiplier stages. Because of the high input power environment and the presence of noise in the system, the multipliers become vulnerable to spurious signal interference, either through the bias lines or through the RF port. As the spurious signals propagate through the multiplier chain, they generate inter-modulation products which might fall in the passband of the heterodyne instrument and seriously degrade its performance. The issues of frequency multiplier response to spurious signal interference and its effect on local oscillator performance in millimeter and submillimeter wave heterodyne instruments are investigated. Results of numerical harmonic balance simulations and laboratory experiments are presented here, and are found to show good agreement.

Keywords: spurious signals, frequency multipliers, millimeter and submillimeter waves

1. INTRODUCTION

Currently many new instruments, which are primarily intended for astrophysical observations, are being built at millimeter and submillimeter wavelengths. Heterodyne receivers (coherent detectors) are the receivers of choice for many of these instruments. In a heterodyne system, the incoming radio frequency signal (ν_{RF}) is down-converted to an intermediate frequency signal (ν_{IF}) in a mixer pumped by a local oscillator signal (ν_{LO}), as shown in the simplified block diagram of a heterodyne receiver in Figure 1. The noise performance of a heterodyne system depends primarily on the noise temperature of these front-end components. At millimeter and submillimeter wavelengths, the first front-end component is a mixer, often cryogenic, such as superconductor insulator superconductor (SIS) or hot electron bolometer (HEB) mixers. Although the noise temperature of the mixer very often determines the overall noise temperature of a heterodyne receiver at these frequencies, noise contributions from other front-end components, such as the local oscillator (LO) and the intermediate frequency (IF) amplifiers cannot be ignored. At millimeter and submillimeter wavelengths, LO injection into the mixer is usually accomplished using a waveguide coupler or optical beamsplitter. However, due to low available LO power, the coupling is fairly large, -10 dB or greater. This allows a significant amount of LO thermal noise to be injected into the receiver along with the LO signal, significantly increasing the receiver noise temperature. Apart from the LO thermal noise, the spectral purity of the LO signal also plays a major role in receiver performance. Often LO amplitude or phase noise show up in the passband of the IF signal, which increases the noise factor

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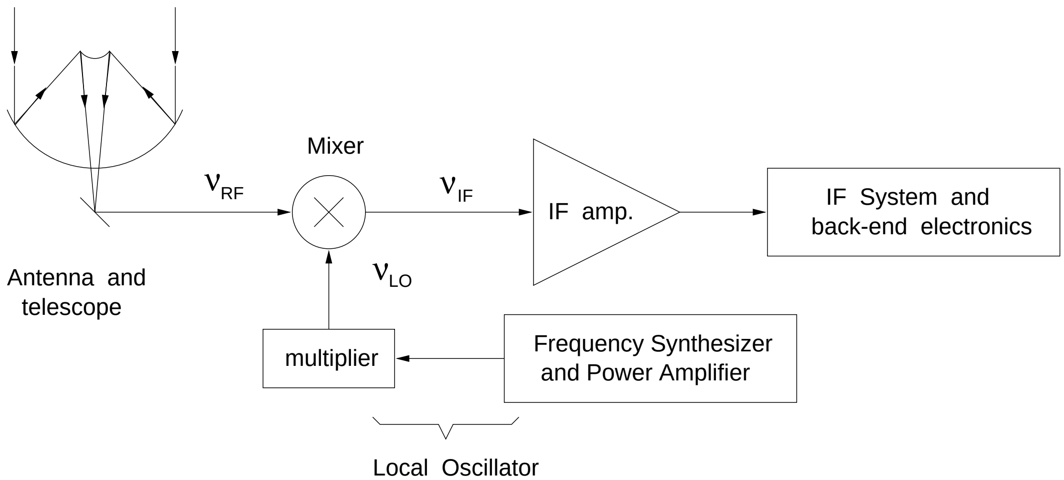


Figure 1. Simplified block diagram of a millimeter and submillimeter wave heterodyne receiver system. The signal from a telescope at a frequency ν_{RF} is combined with a local oscillator at frequency ν_{LO} in a "mixer," which is a nonlinear device, to yield the difference or "intermediate" frequency $\nu_{\mathrm{IF}} = |\nu_{\mathrm{RF}} - \nu_{\mathrm{LO}}|$, which is then amplified and spectrally analyzed.

of the receiver, resulting in degradation of the overall performance of heterodyne instruments. As a result, one requires a spectrally pure and low thermal noise LO system for a low noise heterodyne receiver. Moreover, at millimeter and submillimeter wavelengths, heterodyne instruments require fixed-tuned broadband LO sources which are robust, easy to implement, and reliable, specifically for space-borne instruments.

LO systems at gigahertz (GHz) and terahertz (THz) frequencies have a variety of noise types associated with them. Apart from the thermal noise mentioned before, they have amplitude and phase noise which is due to the short time scale (< 1 s) zero-mean fluctuations in signal amplitude and phase, primarily caused by up-converted low frequency device noise. They have drift noise – which is due to the long time scale (> 1 s) fluctuations in amplitude and phase caused by temperature and other parameter changes. They also have spurious signal interference – which is due to pick up and propagation of spurious signals in a noisy environment, aided primarily by the nonlinear devices in the system. In this paper we will look in closely at the spurious signal interference in LO systems in millimeter and submillimeter wavelengths, specifically at the frequency multiplier response to these spurious signals in a LO chain.

2. FREQUENCY MULTIPLIERS AND SPURIOUS SIGNALS

State-of-the-art solid state LO sources in the millimeter and submillimeter wavelengths are constructed from chains of cascaded Schottky barrier varactor diode frequency multipliers driven by frequency synthesizers and broadband monolithic millimeter wave power amplifiers[1,2] . Figure 2 shows the block diagram of such a LO system working at 800 GHz. The signal from a frequency synthesizer is multiplied using a commercially available frequency multiplier module. This signal around 100 GHz is amplified using a MMIC power amplifier module producing in excess of 250 mW[3] . This high power signal is then multiplied with a series of solid state frequency multipliers to produce sufficient power at THz frequencies to adequately pump SIS or HEB mixers[4] .

The frequency multipliers at these frequencies use a novel planar substrateless technology in the fabrication of the Schottky barrier varactor diodes and various other on-chip matching circuits[5] . Using conventional optical lithographic techniques and back-side wafer processing, free-standing metal beam leads are realized for dc and RF contacts and input and output coupling elements[6] . The units are constructed using split waveguide blocks where the multiplier chip rests on the split waveguide and the input signal is directly coupled to the diodes and the output signal is coupled to the output waveguide by means of an E-field probe[7] .

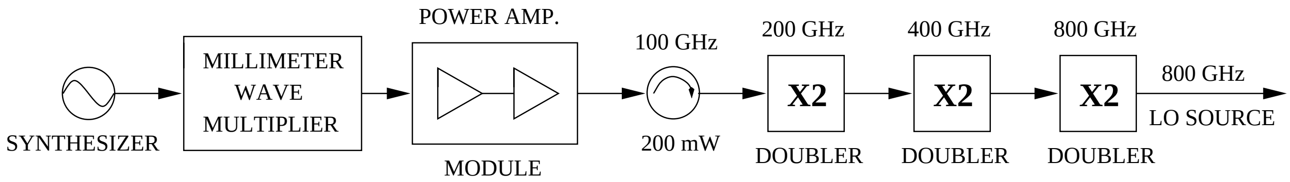


Figure 2. Block diagram of a solid state local oscillator chain using GaAs Schottky barrier varactor diode frequency multipliers. This specific chain is for 800 GHz output frequency, producing about 1 mW of output power at 800 GHz[7] .

Noise performance of the frequency multipliers at millimeter and submillimeter wavelengths is certainly an important issue, as we have mentioned earlier. Almost all studies concerning noise properties of frequency multipliers relates to close-in noise of the LO carrier[8, 9] . Most of the power of a noisy frequency multiplied LO signal is generally concentrated in the close vicinity of the carrier, whose frequency slowly oscillates around an expected average value. This leads to the concept of fuzzy carrier and gives rise to noise sidebands. Phase noise measurement techniques are used to measure the close-in phase noise and noise sidebands. Allan variance measurements are done to determine the long term phase and amplitude drift of LO signals[9–11] . Moreover, linear analysis techniques of noise conversion have been developed over the years which can adequately predict the near-carrier noise properties, such as the amplitude and the phase noise, of typical frequency multipliers[12] .

Less is reported about the far-from-carrier spurious interference properties of frequency multipliers in a LO system. To produce enough pump power at the highest frequency stage of a multiplier chain, the first multiplier in the chain is often pumped with a high power input signal. Because of this high power environment and the fact that the multipliers and the driver amplifiers are very broadband, the frequency multipliers become vulnerable to spurious signal interference and can propagate out of band signals through the multiplier chain. As the spurious signals propagate through a chain such as shown in Figure 2, they generate inter-modulation products which cannot be filtered out easily and may show up in the IF pass band. Frequency multipliers can pick up spurious signals either through the dc bias lines or through the RF input port. The power amplifiers are also vulnerable to spurious signal interference which will propagate through to the input of the frequency multipliers, which in turn generate more spurious sidebands. These frequency multipliers often have on-chip bias filters which reject very high frequency noise signals on the bias lines, and also have external bias filters to reject low frequency noise signals from the multiplier bias. However, if not filtered properly, noise signals in the low GHz range can still leak through the bias lines of the multipliers and propagate to the output through inter-modulation products. These low GHz spurious signals are the major cause of concern because most of the heterodyne instruments currently being built in the millimeter and submillimeter wavelengths have IF frequencies in the low GHz range. We report here the results of studies that we undertook to investigate the effects of these spurious signals on a frequency multiplier. We carried out numerical simulations as well as experimental studies to understand how the spurious signals propagate through the frequency multiplier. We confined our studies to spurious signals in the 1–5 GHz range and spurious signal interference through the multiplier bias line and through the RF input port.

3. NUMERICAL ANALYSIS AND EXPERIMENTAL SET-UP

We used Agilent's[13] nonlinear harmonic balance simulator (ADS) to simulate the effect of spurious signal interference through the bias line and RF input port of our 200 GHz frequency multiplier[5] . The frequency multiplier in question is a six-anode balanced doubler with 72 fF diode junction capacitance at zero bias. The schematic of the harmonic balance simulation is shown in Figure 3. The high frequency structure simulator (HFSS)[14] – a finite element 3-D electromagnetic simulator was used to design the doubler along with the ADS nonlinear simulator. The S-parameters obtained from HFSS simulations along with other circuit elements such as lowpass, highpass, and bandpass filters and Schottky varactor diode models are used in the spurious signal interference simulations.

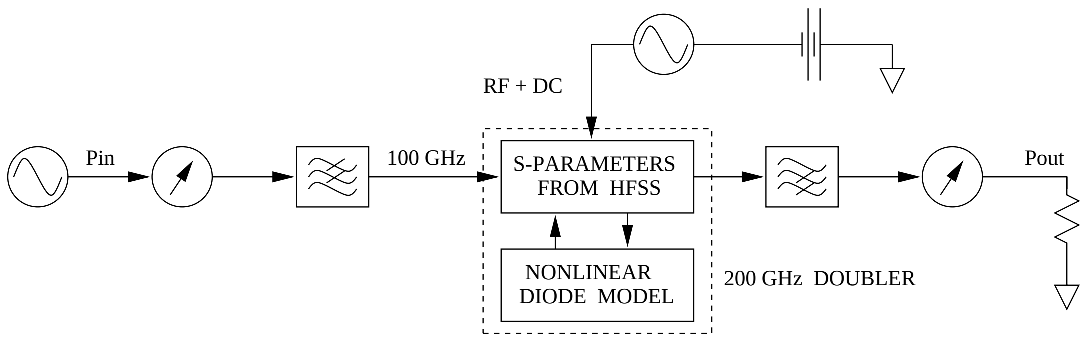


Figure 3. Schematic diagram of the harmonic balance simulation setup used in ADS simulation.

To simulate spurious signal interference through the bias line we introduced a single tone voltage source along with dc bias in the bias line of the frequency multiplier as shown in Figure 3. As the bias signal passes through the nonlinear diodes it interacts with the input pump signal at 100 GHz. We used nominal input pump power to generate about 20 mW of output power at 200 GHz. For the simulation we used 2 GHz and 5 GHz signals with different amplitude levels as the single tone voltage source on the bias line. The harmonic balance simulator was set-up as a two-tones simulator with a maximum of three harmonics on each tone and the total maximum order was set to eight, allowing eight different mixing products to show up at the output along with the three harmonics of each of the tones.

For the simulation of spurious signal interference through the input port of the doubler, we set-up the simulator as a three-tone harmonic balance simulator with the center frequency tone at 100 GHz and sideband tones at ±1 GHz and ±2 GHz. In this particular simulation, we used a maximum of two harmonics for each tone and again the maximum order was set to eight, allowing eight different mixing products and two harmonics of each of the tones to show up at the output of the doubler. Two harmonics were chosen for each tone rather than three to facilitate convergence of the harmonic balance simulator.

Figure 4 shows the schematic diagram of the experimental set-up for the spurious signal interference through the bias line experiment. We used a synthesizer at 16.67 GHz to pump a X 6 active multiplier to produce a pump signal of 100 GHz to drive the mm-wave power amplifier module[3] . A WR-10 coupler is used to measure the input power at 100 GHz. The WR-10 harmonic mixer connected to the coupled port of the coupler is not

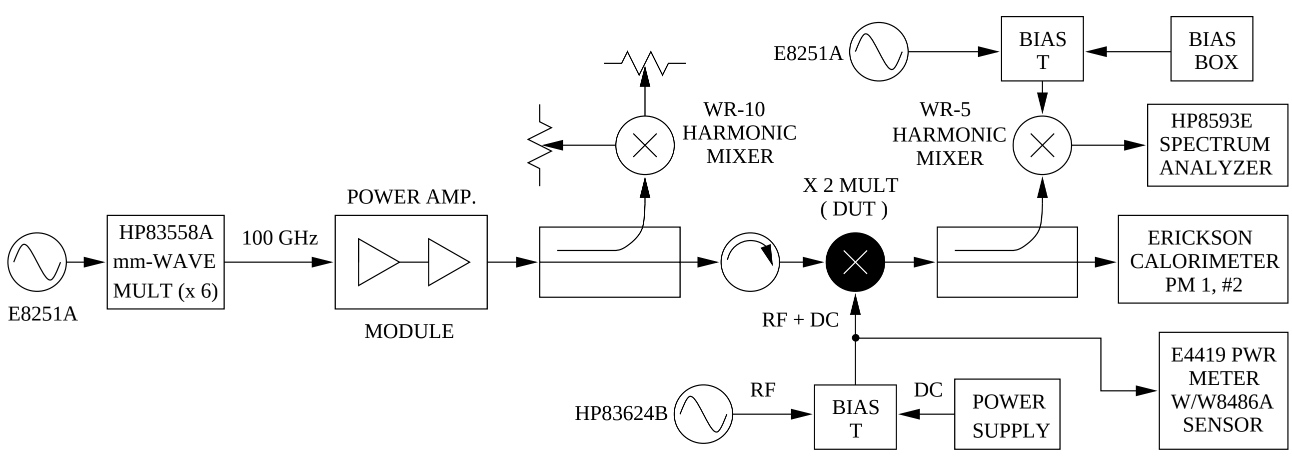


Figure 4. Schematic showing the experimental set-up used for measuring spurious signal interference through the bias line of the balanced doubler at 200 GHz. For this particular measurement the WR-10 harmonic mixer was not used. Spurious signals at 2 GHz and 5 GHz with different amplitude levels were injected in the bias line through the bias-T using a synthesizer shown at the top right of the figure.

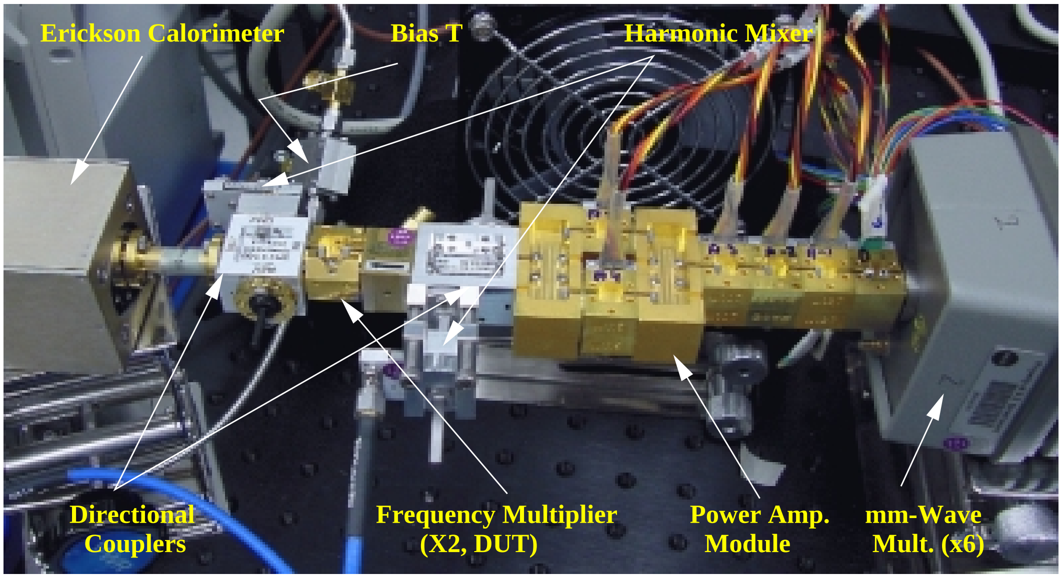


Figure 5. Photo of our measurement set-up. On the right is the X 6 mm-wave multiplier which drives the power amplifier module at 100 GHz range. On the left is the Erickson calorimeter used to measure the output power from the 200 GHz balanced doubler.

used for this experiment. The 100 GHz signal from the through-port of the coupler drives the 200 GHz balanced doubler, whose output power is measured using a calorimeter[15] connected through a WR-5 coupler. The output signal is down-converted to a 2.5 GHz IF by a WR-5 harmonic mixer and a synthesizer, as shown in Figure 4. The frequency of the synthesizer used as the LO signal for the WR-5 harmonic mixer is adjusted to keep the IF output at 2.5 GHz at all times so that the IF calibration is required only at 2.5 GHz. The IF output from the harmonic mixer is measured using a spectrum analyzer as shown. A bias-T is used to bias the doubler and to inject the spurious signal in the bias line. The doubler was first optimally dc biased to generate maximum output power at 200 GHz and then a 2 GHz/5 GHz signal was injected to the bias-T as spurious input to the

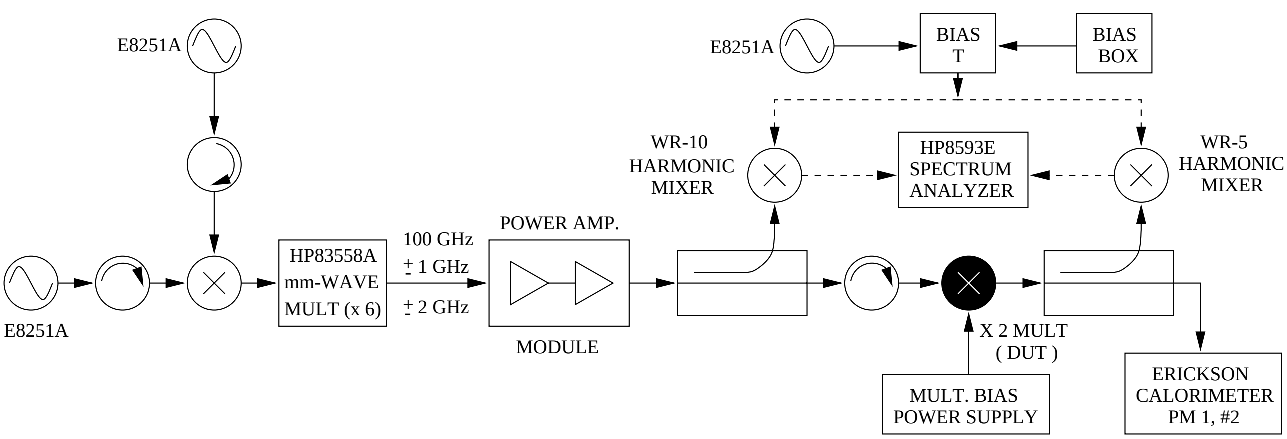


Figure 6. Schematic diagram of our experimental set-up for measuring spurious signal interference through the RF input port of the balanced doubler. The sidebands at ±1 GHz and ±2 GHz at 100 GHz carrier is injected using a mixer and a synthesizer at the input of the power amplifier module. The IF outputs from the WR-10 and WR-5 harmonic mixers are measured using a spectrum analyzer. The dc bias for the frequency doubler is kept fixed at the optimum bias level for this measurement.

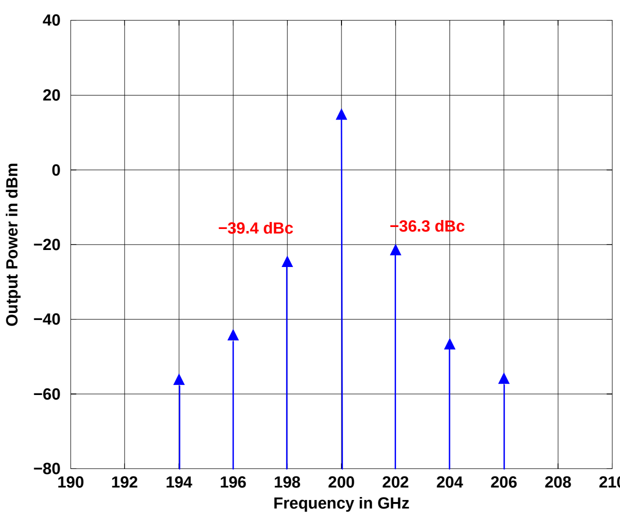
doubler bias line. A photo of the measurement set-up is shown in Figure 5.

Figure 6 shows the schematic of the measurement set-up for spurious signal interference through the RF input port experiment. Here, we introduced sidebands at ±1 GHz and ±2 GHz on the 100 GHz carrier signal using a mixer and a second synthesizer set at 15.67 GHz. The first synthesizer is set at 16.67 GHz, as in the previous measurement. The sideband levels at ±1 GHz and ±2 GHz are controlled through the amplitude level of the second synthesizer. The 100 GHz input signal with sidebands is coupled to a WR-10 harmonic mixer which down-converts it to 1.25 GHz IF signal and is measured using a spectrum analyzer. The output of the balanced doubler is coupled to a WR-5 harmonic mixer which down-converts the output signal to 2.5 GHz IF signal and is measured with the spectrum analyzer. Before we started the experiment, both the input and the output signals at the through-port of the WR-10 and WR-5 couplers are measured with the calorimeter for calibration. The doubler is biased to an optimum fixed dc value to get maximum power at 200 GHz. The LO frequencies for the harmonic mixers are adjusted in such a way that we always have 1.25 GHz IF output for all the sidebands of the WR-10 harmonic mixer output and 2.5 GHz IF output for all the sidebands of the WR-5 harmonic mixer output. As a result, we needed to calibrate the IF outputs only at 1.25 GHz and 2.5 GHz respectively.

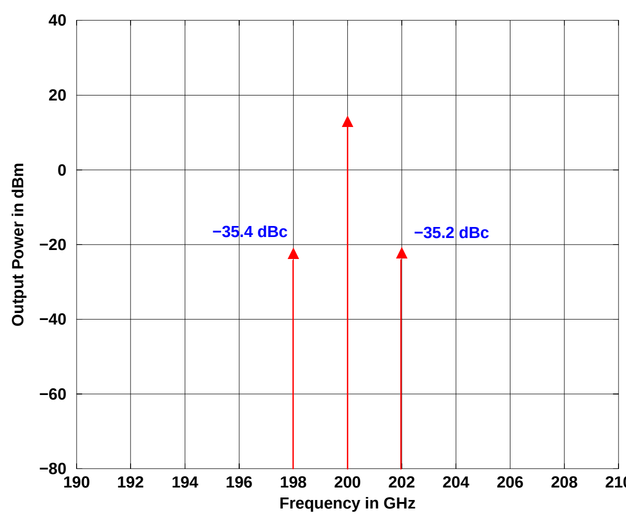
As the overall calibration procedure, we characterized each individual components at all relevant frequencies. The WR-10 coupler was calibrated at 100 GHz ±5 GHz with 1 GHz step size and the WR-5 coupler was calibrated at 200 GHz ±10 GHz with 1 GHz step size. The conversion loss of the WR-10 and WR-5 harmonic mixers was also measured with signal levels close to the signal levels used in the actual measurements.

4. SIMULATION AND MEASUREMENT RESULTS

For the simulations and the measurements of bias line spurious signal interference, we used different amplitudes for the 2 GHz and 5 GHz spurious signals. Figure 7 shows the results for a 185 mV pk-pk signal at 2 GHz spurious signal on the multiplier bias line. Figure 7(a) shows simulated and and Figure 7(b) measured results. The simulations show that the 2 GHz signal on the bias line has generated sidebands at ±2 GHz, ±4 GHz, and ±6 GHz at the 200 GHz output of the doubler. The simulation predicts that the 202 and 198 GHz sidebands are at -36 and -39 dBc respectively. The other sidebands are -60 dBc or lower. The measurement shown on Figure 7(b) is corrected for the conversion losses of the WR-5 coupler and the harmonic mixer. Given that the conversion loss of the harmonic mixer at 200 GHz is about 60 dB, and the noise floor of the spectrum analyzer



(a)



(b)

Figure 7. Simulated and measured results for a 185 mV 2 GHz spurious signal interference through the bias line of our 200 GHz frequency doubler. Figure (a) shows simulated and Figure (b) measured results.

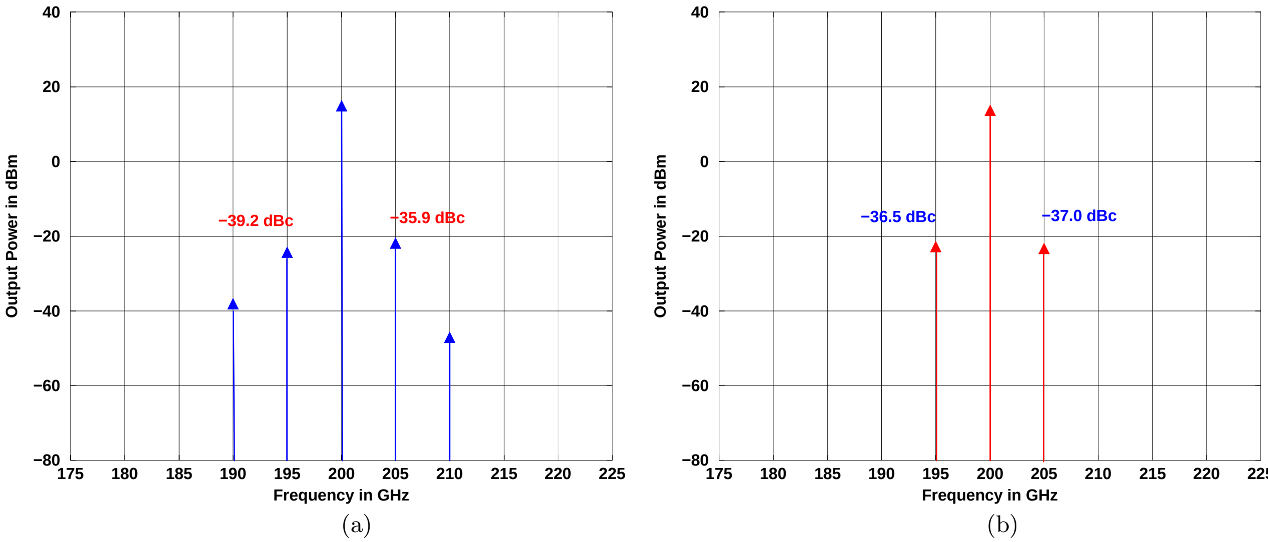


Figure 8. Simulated and measured results for a 200 mV 5 GHz spurious signal interference through the bias line of our 200 GHz frequency doubler. Figure (a) shows simulated and Figure (b) measured results.

-120 dBm (a combination of the resolution bandwidth, the video bandwidth and the time-base of the spectrum analyzer set this limit), we could measure only one pair of sidebands. The 198 and 202 GHz sidebands were both measured about -35 dBc, about 4 dB and 1 dB higher than the simulation results. When we introduced a 2 GHz 635 mV pk-pk signal (10.7 dB higher than the previous signal), we measured sidebands at -24.4 and -24.2 dBc at 198 and 202 GHz respectively, about 11 dB higher than the previous measurements. This trend continues till the doubler is saturated.

Figure 8 shows the simulation and measurement results for a 5 GHz, 200 mV pk-pk signal on the bias line. It can be seen that for this case, the measured signal is about 3 and 1 dB larger at 195 and 205 GHz sidebands compared to the simulation results. When we introduced a 5 GHz 63 mV pk-pk signal (10 dB lower than the

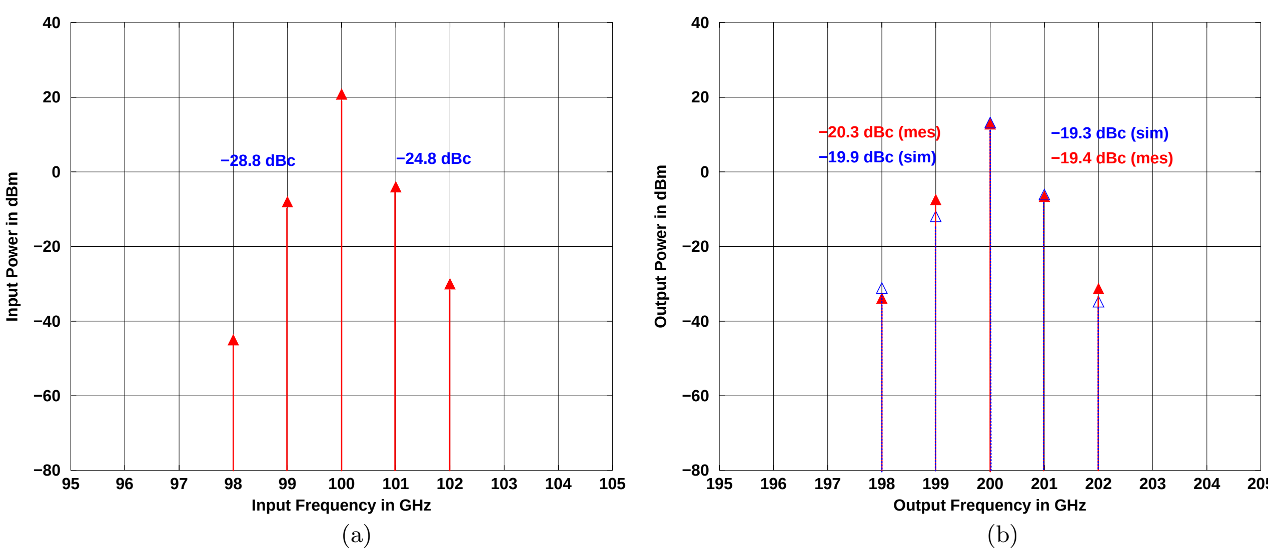


Figure 9. Simulated and measured results for spurious signal interference through the RF input port of our 200 GHz frequency doubler. Figure (a) shows the input signal used to pump the doubler and Figure (b) shows the output at 200 GHz. The open and the close triangles are for simulated and measured results respectively.

previous signal), we measured sidebands at -46.5 and -47.1 dBc at 195 and 205 GHz respectively, about 10 dB lower than the previous measurements.

Considering the machining tolerances of the multiplier blocks and the fabrication tolerances of the devices, simulation and measurement results show very good agreement. We also simulated and measured bias line spurious signal interference for various other amplitude levels at 2 and 5 GHz. We plotted here only those results for which spurious sideband levels at the output were about -35 dBc.

Figure 9 shows the results of the simulations and measurements for spurious signal interference through the RF input port of the frequency multiplier. Figure 9(a) shows the input signal at 100 GHz with sidebands at ± 1 GHz and at ± 2 GHz, which is used to pump the balanced doubler. Figure 9(b) shows the simulated and measured results at the doubler output. Open triangles show simulated results and the close triangles measured results. When we introduced 10 dB lower sidebands at the input of the multiplier than the previous case (-38.8 and -34.8 dBc at 99 and 101 GHz respectively, and keeping the 100 GHz signal level the same), the output sidebands measured were at -30.2 and -29.2 dBc at 199 and 201 GHz respectively, which are about 10 dB lower than the previous measurements.

Here again we tried different sideband levels at the input to pump the doubler, but we plotted only those results for which the first sidebands at the output are about -20 dBc. It can be seen that although the input had sidebands at ± 1 GHz and at ± 2 GHz at 100 GHz, the doubler output produced sidebands ± 1 GHz and at ± 2 GHz at 200 GHz, because of the different mixing and inter-modulation products generated through the nonlinear doubler.

5. CONCLUSIONS

When a frequency multiplier experiences spurious signal interference, either through the bias line or through the RF input port of the multiplier, the signal is subjected to an input and output filter with symmetric diodes (because the multipliers are balanced) performing mixing of the signal with the input pump power and with the spurious signal itself. The conversion of the sidebands in a balanced varactor diode multiplier is highly dependent on the circuit design. Because of the inherent symmetry of the balanced doubler, the odd harmonic products are suppressed, depending on the degree of balance in the circuit.

We observed from the simulations and the measurements that when the spurious signal levels on the bias line of the multiplier are increased or decreased by about 10 dB, the sidebands at the output of the 200 GHz doubler increases or decreases by the same amount, till the doubler reaches saturation. We also noticed that a signal around 200 mV (pk-pk) in the 1–5 GHz range produces highest level sidebands in the -35 dBc level. We should remember that as these sidebands propagate through the multiplier chain, they will produce more sidebands, but will be at a lower level than the closest sidebands. However, other multipliers in the chain will produce their own sidebands due to spurious signal interference through their bias lines. To reduce these sideband levels, one needs to look in the bias filtering of the multiplier very closely. Incorporating bandstop filters in the bias line of the frequency multiplier in the IF frequency band will certainly help alleviate this problem.

As for the spurious signal interference through the RF input port, the sideband levels change linearly with the input sideband level changes; when we decreased the input sideband levels by 10 dB, the output sideband levels also decreased by 10 dB. The spurious signal interference through the RF input port shows up slightly differently at the frequency multiplier output than the spurious signal interference through the bias line of the multiplier. We observed that sideband signals at ± 1 GHz of the 100 GHz input signal at -25 dBc level produces sidebands at ± 1 GHz of the 200 GHz output signal at -20 dBc, a 5 dB increase in the sideband levels from the input to the output. However, we should keep in mind that these signal levels are very much dependent on the exact circuits used for the frequency multiplier. To reduce the spurious signal interference through the RF input port of the multiplier, one has to pay attention to the filters used on the bias line of the millimeter wave power amplifier modules. Overall, the simulations and the measurements showed that the multiplier did not degrade the spurious signal interference of the LO as the spurious signals propagated through it.

In future, we are planning to run further tests and simulations by cascading more multipliers to the chain. We also want to study the spurious signal interference through the gate bias of the power amplifiers and its effects on the frequency multiplier and the LO chain. Since the power amplifier modules are operated in saturation, the spurious signal interference through the drain bias of the power amplifier will not affect its spurious performance. We are also planning to run simulations and carry out tests with spurious signal frequencies higher than 5 GHz on the bias lines.

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