Final Report

"All Solid-State Source Technology for Terahertz Applications"

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Methods Assumptions and Procedures

Amplitude stability is defined based on the Electric field or Voltage waveforms. Percent rms is determined by converting all signals into a voltage waveform, calculating the mean, subtracting each measured data point from the mean, squaring the result, averaging all the squares which gives the variance, taking the square root of the variance, dividing by the mean, and multiplying by 100. Fluctuations slower than 10 Hz were measured by detecting the signal and measuring on a multimeter, or downconverting and measuring on an Agilent E4419B Power Meter with an E4413A head. Fluctuations faster than 10 Hz were measured by using a VDI WR1.2ZBD, zero bias detector, with an AD743 low noise operational amplifier. The amplified signal was measured on a Tektronix TDS 744 oscilloscope with AC coupling. The WR1.2ZBD was operated in the linear region, and the DC detected voltage was assumed to represent the voltage amplitude of the RF signal.

Frequency was measured using an HP5343A frequency counter.

Output powers for this report were measured using an Erickson WR10 power meter. The manufacturer estimates the measurement uncertainty to be 5%. The input waveguide is wr10, but the meter also accurately measures higher frequencies. A waveguide taper was used to transition from the waveguide of the device under test to the wr10 waveguide of the power meter. The loss of these tapers varies from 3-33% depending on the frequency. The output power reported here was calibrated based on the taper's insertion loss according to the following table.

Frequency (GHz)	Linear factor
180	1.03
225	1.06
550	1.24
700	1.33

Phase noise was measured by first downconverting and measuring total sideband power on the HP8565E Spectrum Analyzer. It's assumed that this measurement represents a worst case.

I. Introduction

Existing and emerging applications for terahertz technology motivate the need for improved terahertz sources. With the Submillimeter-wave technology gap, terahertz sources lag the performance found in other regions of the spectrum. Many applications such as imaging, chemical agent detection, and spectroscopy would benefit from narrowing the technology gap. This SBIR focuses on improving key aspects of terahertz sources including power, spectral purity, and noise while maintaining good bandwidth, sweep speed, and size for output frequencies from 0.1-1 Terahertz.

For the interim technical report, Virginia Diodes Inc. (VDI) reported power and bandwidth results for two different sources. One source uses all broadband components to achieve the maximum bandwidth, while the other source uses a mixture of high powered and broadband components to achieve the maximum amount of power over 100 GHz of bandwidth. VDI has delivered the high powered version for completion of phase 1. The source operates from 660-760 GHz with greater than 100 μW of power. The source meets or exceeds the phase 1 objectives in all regards except amplitude stability. The amplitude stability of the source exceeds VDI's expectations of what could be accomplished in phase 1 and has been thoroughly analyzed to determine how to meet the specification for phase 2. The source is lightweight, compact, and easy to use. It can be swept across the band rapidly under full computer control.

The source uses a synthesizer to achieve very high spectral purity. The synthesizer was delivered to VDI from Micro Lambda Wireless Inc. very late in the contract, so another similar source operating around 800 GHz was used for many of the measurements. The 800 GHz source, discussed in section B, was developed for another project, but has similar components. Most importantly, the source was made with a compatible receiver that was indispensable for measuring phase noise and searching for spurious signals. Developing a receiver specifically for this contract could not have been done within the time frame, but the performance of the 800 GHz source is most likely very similar to the delivered source. So, measurements that could not have been done with the 660-760 GHz source without a receiver were done with the 800 GHz transceiver. Some of the measurements reported here were made with the 800 GHz source and some were made with the 660-760 GHz delivered source. The delivered source is described in detail including all the necessary information to allow user programming of the source for incorporation into any measurement system. The performance of the source is thoroughly analyzed and reported in the measurement results section.

II. Report

A. <u>660-760 GHz Demonstrated and Delivered Source</u>

A terahertz source was developed and delivered that produces greater than 100 μW tunable from 660-760 GHz. The performance of this source was characterized in terms of phase noise, amplitude stability, spurious frequencies, spectral purity, sweep speed, power, and bandwidth. The performance exceeds the Phase 1 requirements in all areas except amplitude stability. The amplitude stability requirement is the most challenging, and though this source does not meet that requirement it exceeds VDI's expectations of what could be accomplished in Phase 1, and we have identified many of the obstacles that hopefully can be overcome to exceed the specification in Phase 2. The following sections describe the physical layout and operation of this source.

1. Physical Description

The source uses an all solid-state, amplifier multiplier chain (AMC) produced by VDI to multiply the frequency of a YIG-based synthesizer, MLSE-0220, from Micro Lambda Wireless [1]. The synthesizer is phase locked to a low phase noise, 10 MHz, crystal oscillator produced by Wenzel Associates Inc [2]. The crystal oscillator, synthesizer, and AMC are mounted together on an aluminum plate resulting in a compact (0.0062 m^3), lightweight (8.2 lbs), terahertz source as shown in Figure 1. The AMC uses 3 VDI multipliers and a Spacek active doubler to multiply the synthesizer frequency by a factor of 48 [3]. The Spacek active doubler provides about 2 Watts from 27-32 GHz and drives a VDI Q120 which quadruples the frequency to give about 100-200 mW. The Q120 includes two doublers in one block housing similar to the VDI doublers named D60 and D124. Next, a VDI D244 doubles the frequency to drive a WR1.2x3 broadband tripler which is the final stage to bring the output frequency to 660-760 GHz.

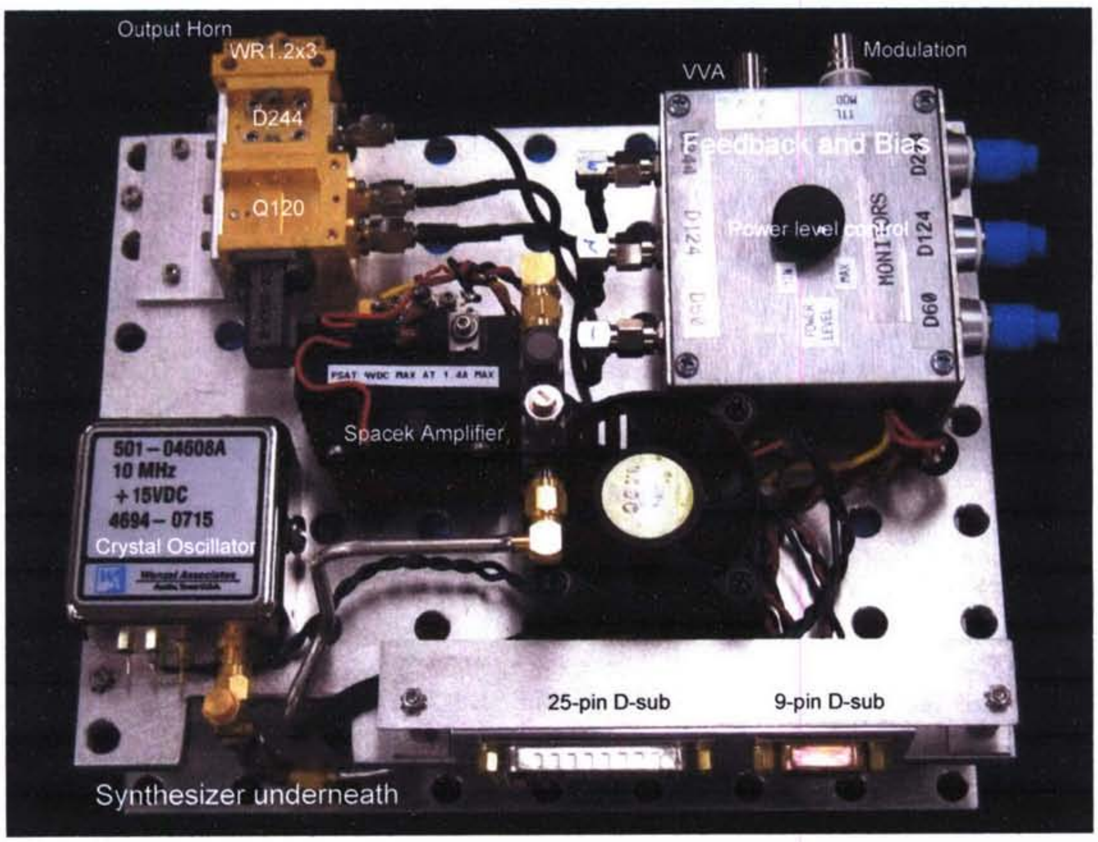

Figure 1: Pictures of the 660-760 GHz source delivered for the contract.

The source receives DC power, multiplier bias, and digital control of the synthesizer through one 9-pin D-sub and one 25-pin D-sub connector conveniently located on the edge of the plate as shown in Figure 1. For added convenience and ease of use, a control box (0.016 m^3, 14 lbs) is included that plugs into a wall power outlet and with a single switch provides all the necessary DC voltages to run the source. Furthermore, a computer can communicate with the control box through a USB connector to control the output frequency. A CD is included which installs a program for fast and easy frequency control. A picture of the entire system including a laptop computer is shown in Figure 2.

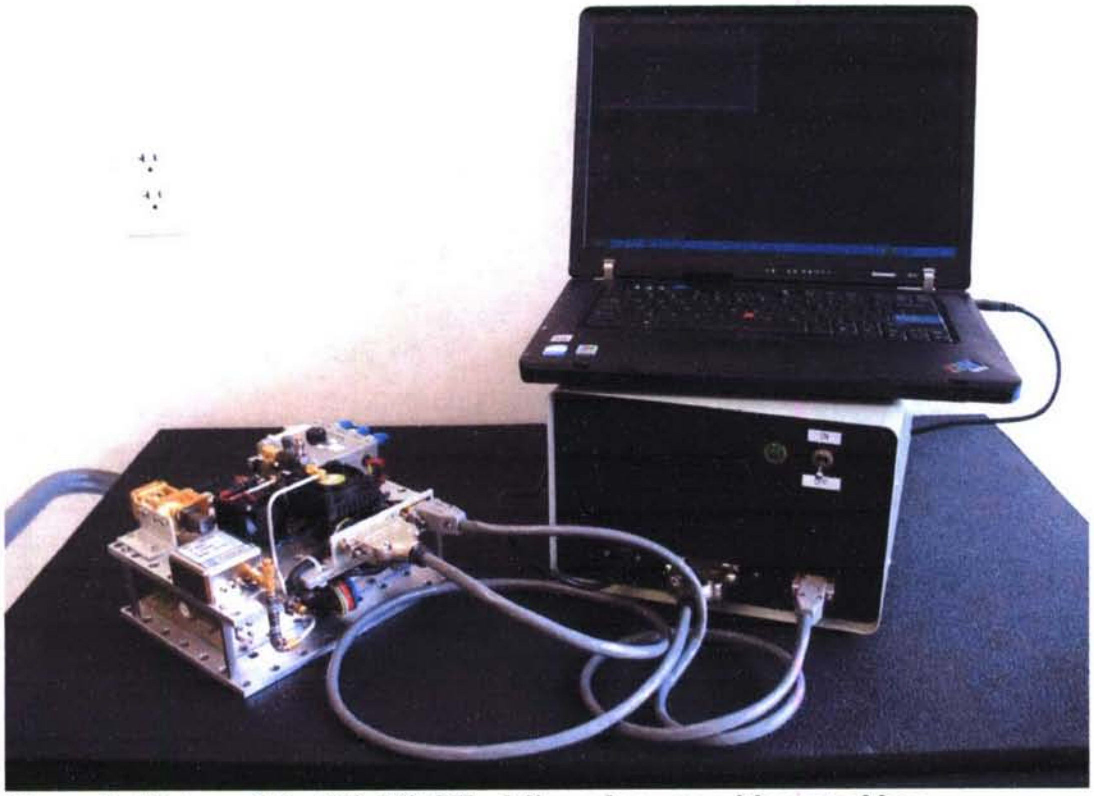

Figure 2: Picture of the 660-760 GHz delivered source with control box.

A feedback control circuit and bias circuit are also housed on the Aluminum plate to provide the optimum multiplier bias and output power control of the Spacek amplifier to achieve the maximum possible output power and bandwidth without damaging the multipliers. This circuit provides a knob which manually controls the source's output power along with two BNC connectors that allow voltage variable attenuation (VVA) and TTL amplitude modulation.

The output power can be precisely controlled by applying a voltage on the VVA port. A 5 V signal on the VVA connector completely attenuates the output power, and 0 V provides no attenuation. The modulation port completely attenuates the output power for all voltages greater than 2.5 V and has no effect for all voltages less than 2.5 V. The VVA port could be used to amplitude modulate the source, but there is an initial overshoot for fast changes that occur before the feedback circuit can adjust. The TTL modulation port allows modulation with no overshoot, but allows only an on and an off state.

The setup procedure is simple and short:

1. Attach the cables between the control box and the source plate.
2. Install the program on a computer.
3. Plug the control box into a wall outlet.
4. Attach the USB cable between the computer and the control box.
5. Allow the computer to automatically install the hardware.
6. Turn on the control switch.
7. Run the program to set the output frequency.

The output power radiates into free space through a diagonal horn antenna shown in Figure 1. There is a directional coupler between the synthesizer and the Spacek amplifier that allows the user to sample the synthesizer output power in order to verify operation. Also, there is an SMA T connector on the 10 MHz crystal oscillator to access the 10 MHz signal for phase locking together multiple instruments with the source such as a frequency counter, spectrum analyzer or a VDI receiver.

The program installation disc will install NI-DAQ 8.5 Runtime 4. This distribution contains all the features in the full NI-DAQmx version except for ADE support for LabVIEW, LabWindows/CVI and Measurement Studio, the DAQ Assistant, the NI-DAQmx documentation, and Real-Time Support. Installation of NI-DAQ 8.5 will result in the removal of all previous versions of NI-DAQmx currently installed. Additionally, NI-DAQ 8.5 only supports LabVIEW version 7.1 and later. All NI-DAQmx support in previous versions of LabVIEW will be removed. The full version of NI-DAQ 8.5, can be download from www.ni.com.

This section described the physical layout, purpose, and use of each component in the source. The next section describes how the source operates along with all information necessary to program the source.

2. Operation

The source's output frequency and power can be computer controlled to achieve greater than 100 μW from 660-760 GHz without any mechanical tuning thus resulting in very fast fully optimized frequency sweeps. As described in the physical layout section, a Spacek amplifier and VDI multipliers are used to multiply the frequency output of a synthesizer by a factor of 48. A computer controls the synthesizer frequency and multiplier bias, and a feedback circuit regulates the Spacek amplifier output power to achieve maximum bandwidth, power, and sweep speed without damaging the multipliers. Understanding how the computer controls the synthesizer and multiplier bias is not important for operating the source and setting up frequency sweeps. However, often the source will need to be used within a system with a receiver or power meter, and the system components must be controlled together. In other words, depending on the application it may be necessary to program the source, and the following information in this report may serve as a user manual with all the necessary information.

a. Programming the Source

i. Synthesizer

The computer controls a National Instruments USB-6221 Data Acquisition device (DAQ), and the DAQ controls the Synthesizer and multiplier bias [4]. The synthesizer communicates using 6 digital lines. These 6 lines are connected to digital input and output pins of the DAQ according to Table 1. The synthesizer responds to the ASCII commands shown in Table 3. The DAQ sends an ASCII command in serial binary format, most significant bit first, over the "DATA" line. Each bit is read with each rising edge of a pulse on the "CLOCK" line only when the "ENABLE" line is set to low (0V). Synthesizer responses are received on the falling edge of the clock pulse. Table 2 and Figure 3 are from Micro Lambda's instructions which provide timing, handshaking, and other information.

DAQ Pin Name	25-pin connector pin #	20-pin Synthesizer pin #	Synthesizer Line Name
P0.0	3	9	CLOCK
P0.1	2	10	DATA IN
P0.2	4	11	ENABLE
P0.3	11	12	BUSY
P0.4	13	14	LOCK ALARM
P0.5	15	13	DATA OUT
Digital GND	18	4	LOGIC GND

Table 1: DAQ and synthesizer pin connections.

Select	Input	Active Low. Enables the shifting of data into the internal command buffer. Also serves as a command terminator when it goes HIGH. The status of the BUSY line should be checked before activating this line.
Clock	Input	Data is clocked into the unit on the rising edge (**Positive Edge Clocked**) and Data Out is valid at this time. The maximum clock rate of this line is 50 usec. For best performance the status of the DataOut line should be checked before sending the first Clock. This line should be maintained in a LOW state at the application of Select to prevent confusion.
Data In	Input	Input data pin. Data is sent MSB first. Data must be stable 50 usec. Before the Clock line goes high and 50 usec. After the clock goes low. (**Setup / Hold time**)

BUSY	Output	This line is used to indicate that the unit is busy processing other commands or doing its internal housekeeping. Before sending a Select = TRUE the status of this line should be checked to ensure that it is LOW (**NOT Busy**). Any command initiated by setting Select Low while BUSY is High may result in lost data and uncertain results. NOTE: The unit can be programmed without using this line if sufficient time is allowed between Clocks and between commands. The time required varies between commands. ***This mode is not recommended as there are some events that occupy the microcontroller other than the serial communications.***
DataOut	Output	This line is used to pass internal status information from the synthesizer. Data is guaranteed to be valid on the falling edge of the clock signal. Data is sent out MSB first. In addition, this line is used as a communication 'handshake' line. Once Select has gone LOW the DataOut line will be taken HIGH to indicate that the unit is listening. It will remain HIGH until the first data is sent out which is initiated by the first rising edge of Clock. DataOut will be returned to LOW after Select has been released.

Table 2: Micro Lambda's descriptions of the synthesizer's digital lines. In addition to the lines above there is a unit status line, LOCK Status, which is a static line which only indicates the overall health of the unit – specifically, that all of the internal phase locked loops are locked.

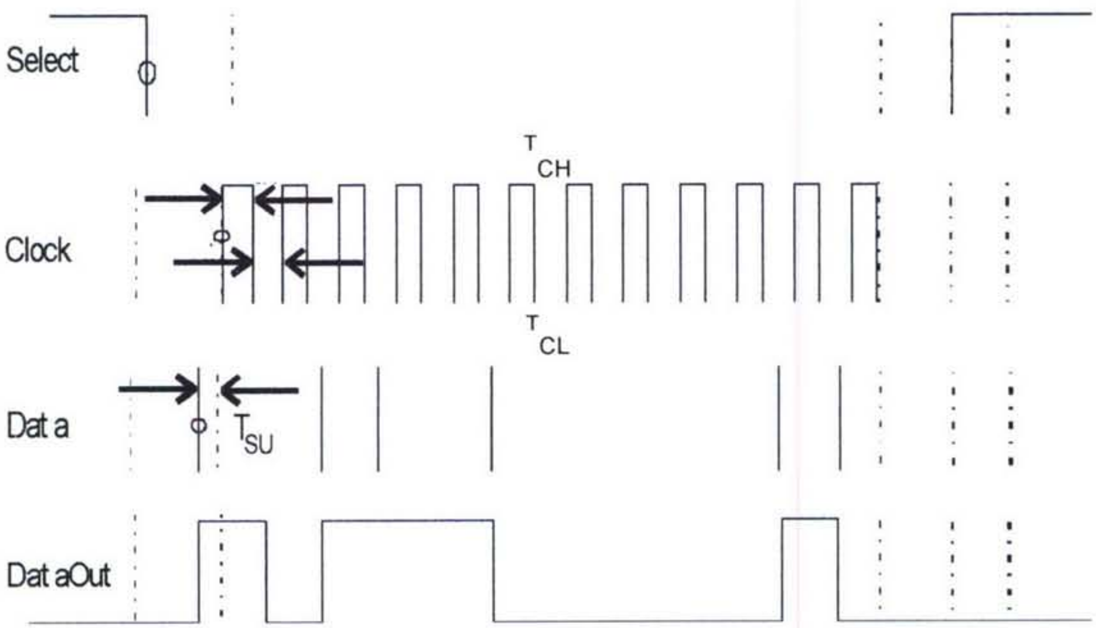

Figure 3: Micro Lambda's timing illustration. Tsc > 2 usec select low before first clock, Tcs > 2 usec clock low before chip select high, Tsu > 2 usec data stable before rising edge of clock, Tch > 2 usec minimum clock high time, Tcl > 2 usec minimum clock low time, Tsb > 10usec (time to wait before sampling 'BUSY' Data and Clock setup / hold time = 10 usec. {hold time is determined from Tch}

Command	Description
?	Reports Status of all internal phase locked loops; three loops are indicated by bits 0, 2, and 3. The other bits are internal variables with of no particular interest to the user.
>	Recalls the synthesizer state from the next sequential saved memory location. If the last location accessed was the 99th, the '>' would recall the 100th location. Using this command the user can get the maximum step rate from the unit. (Used in conjunction with the NR command only.)
AX	Enables the auxiliary RF input port J6. This will allow an external frequency, within the frequency range of the unit (applied to J6) to be passed to the main RF output connector J5. AX0 = Disable, AX1 = Enable. Note: AX1 command disables the internal synthesizer output. This mode is an option.
f	Frequency command (binary). The ASCII 'f' is followed by 34 bits of frequency data. The data is in straight binary format with the LSB representing 1 Hz. (i.e. 3.456789012 GHz would be commanded by f0011001110000010100110101000010100 ('f' followed by CE0A6A14 Hex)
F	Frequency command (ASCII). This accepts the frequency in MHz in straight ASCII format. i.e. 3.456789012 GHz would be commanded by F3456.789012
L	Sets the LOCK ALARM output polarity; L1 sets the unit for LOCK == positive true; L0 sets the unit for LOCK == negative true. (pin 13 of J1; Lock Alarm)
NR	Recalls a synthesizer state from the specified location. (one of 1000 saved using the NS command) 'NR' followed by a hex address of 0x0063 would recall the instrument state stored in the 99th decimal location.

NS	Stores the present state of the synthesizer in the specified location. (one of 1000 available); 'NS' followed by a hex address of 0x0064 would store the present state of the synthesizer in the 100th (decimal) location.
PL	RF Power Level command. It is used to set the RF output level within the usable range of the unit, typically -20.0 to +20.0 dBm. This command accepts the RF power level setting in dBm. The characters are in ASCII format. i.e. +19.5 dBm setting would be commanded by PL+19.5 and -10.1 dBm would be PL-10.1. (Available with the RF Power control option only.)
R	Programs the Reference frequency. Range 5 to 50 MHz; 1 MHz resolution. eg. R25.0 would set the external reference frequency to 25 MHz.
SP	Synthesizer Preset – Clears all nonvolatile memory settings to factory default. Typically used to clear erroneous settings in the unit.
T	Reads internal temperature information. Responds with degrees C in a one byte response (signed char). Range: -40 to +80; Note: Since the T command is only one byte long and the command needs to be processed before the unit can acquire the temperature data, the data returned is always one command behind. Thus if you want the present temperature the T command should be sent twice and the second data used.
VF	Programs the secondary VCO frequency synthesizer (second LO opt., J3) frequency. The characters are sent in ASCII format. i.e. VF1000.0 would set the synthesizer 2nd output PLL for 1000.0 MHz

Table 3: ASCII user commands with Micro Lambda's descriptions.

ii. Biasing

There are two multipliers in the AMC named Q120 and D244 that need a voltage bias. The Q120 has two doublers that need biasing named D60 and D124. The D60 portion of the Q120 receives a fixed voltage bias, but the D124 portion of the Q120 and the D244 receive a variable bias as a function of frequency for optimum performance. The bias voltages are set by placing a voltage on pin 7 of the 25 pin D-sub connector for the D124 and pin 8 for the D244 both relative to the analog ground on pin 9. The DAQ provides the optimum voltage on these pins through two analog outputs. DAQ pin name A0.0 is connected to pin 7 for the D124, and A0.1 is connected to pin 8 for the D244. As the DAQ sets each frequency on the synthesizer, it also sets the D124 and D244 voltage biases interpolated according to the values in Table 4. The voltages on the pins are amplified with maximum and minimum limits, so that the pin voltages are not the same as the multiplier bias voltages. Also, the DAQ cannot output enough voltage to damage the system, so a programmer cannot cause damage with programming errors. The input impedance for both pins 7 and 8 is 15 Kohms.

Frequency(GHz)	D244 DAQ Output Bias (V)	D124 DAQ Output Bias (V)
630	-4.137	-7.715
634.2	-4.182	-5.915
638.4	-4.187	-7.795
642.6	-3.907	-4.395
646.8	-3.217	-3.195
651	-3.312	-1.819
655.2	-3.847	-0.003
659.4	-3.907	0.147
663.6	-3.657	0.097
667.8	-3.032	0.177
672	-3.887	0.177
676.2	-4.157	0.177
680.4	-3.787	-5.817
684.6	-3.372	-6.619
688.8	-3.237	-6.077
693	-3.112	-4.868
697.2	-2.967	-2.62
701.4	-2.997	-0.67
705.6	-3.147	-0.419
709.8	-3.157	-1.005
714	-2.887	-2.715
718.2	-2.797	-0.112
722.4	-2.827	0.177
726.6	-2.557	0.107
730.8	-2.797	0.077
735	-3.312	0.077
739.2	-3.507	0.077
743.4	-3.252	0.077
747.6	-2.827	0.097
751.8	-2.497	0.177
756	-3.489	0.177
760.2	-3.491	0.177
764.4	-3.026	0.177
768.6	-2.505	0.177
770	-2.331	0.177

Table 4: Optimum bias values versus frequency for the D244 which is DAQ pin A0.1 and 25-pin Dsub #8, and the D124 which is DAQ pin A0.0 and 25-pin Dsub #7.

b. Feedback

A feedback circuit maintains safe power levels by limiting the RF generated DC currents in the multipliers. DC currents are generated by the multipliers when exposed to RF power. The currents are typically proportional to power levels and can be used as a rough metric for RF power in the device. Through various tests, VDI has determined safe operating currents. The lifetime of the multiplier is reduced when RF power produces generated DC currents above the safe level. RF power can be controlled by controlling the output power of the Spacek amplifier. Spacek allows access to the transistor drain of their amplifiers through a pin called "Psat", so that the voltage on Psat directly controls the output power. The feedback circuit continually monitors the DC generated currents and decreases the voltage on Psat to reduce the RF power if the currents exceed safe limits. The circuit responds faster than a millisecond, so safe levels are restored before the multipliers reach dangerous temperatures. The Q120 and the D244 are monitored by the feedback circuit, but likely only one will reach the limit at a time. If there is not enough RF power for any of the multipliers to reach a limit then the feedback circuit does nothing. Therefore, the system will not be under feedback for all frequencies. Also, the D244 will limit the power at some frequencies while the Q120 will limit the power at other frequencies. Table 5 shows the feedback status of the delivered source as a function of frequency. The effects of feedback on amplitude stability have been explored and are described in the measurement results section.

Frequency Range (GHz)	Component Under Feedback
630-654	D60
654-658	D244
658-667.8	D60
667.8-670.6	D244
670.6-696	D60
696-700	D244
700-730	D60
730-770	NONE

Table 5: Component under feedback versus frequency. The data was determined with a frequency resolution of 1.4 GHz

B. 797-837 GHz Transceiver

The synthesizer for the delivered source did not arrive until very late in the contract; therefore, an 800 GHz transceiver with similar components to the delivered source was used to complete many of the measurements for this contract such as phase noise measurements and searches for spurious harmonics. Also, a receiver was not available for the delivered source, but was available for the 800 GHz source which likely has similar performance to the delivered source. A picture of the source and receiver is shown in Figure 4 and Figure 5. The source uses a synthesizer from Micro Lambda, MLSN-1012, a Spacek active tripler, A369-3XWB-24, a VDI D70, D142, and D280 frequency doublers, and a VDI WR1.2x3 frequency tripler. The receiver uses a Spacek active doubler, A246-2XWB-31, a VDI D100V3, D200R2, and D400 frequency doublers, and a WR1.2SHM sub-harmonic mixer. The source's output power and the receiver's conversion loss are shown in Figure 6.

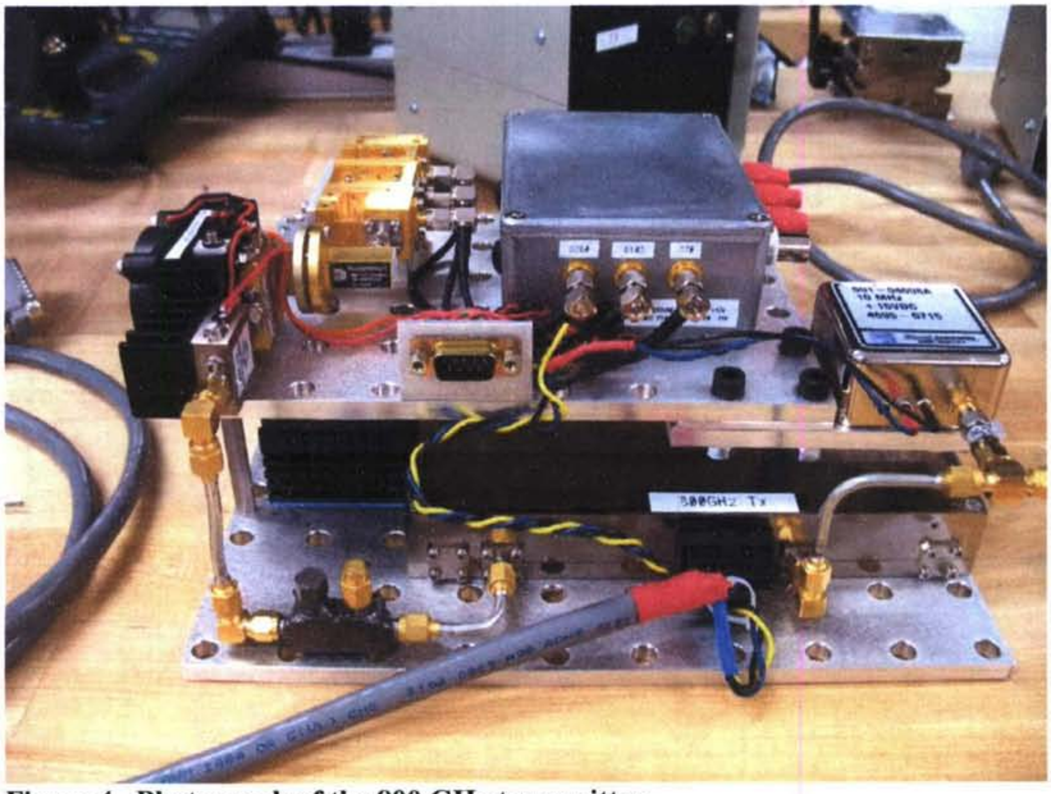

Figure 4. Photograph of the 800 GHz transmitter.

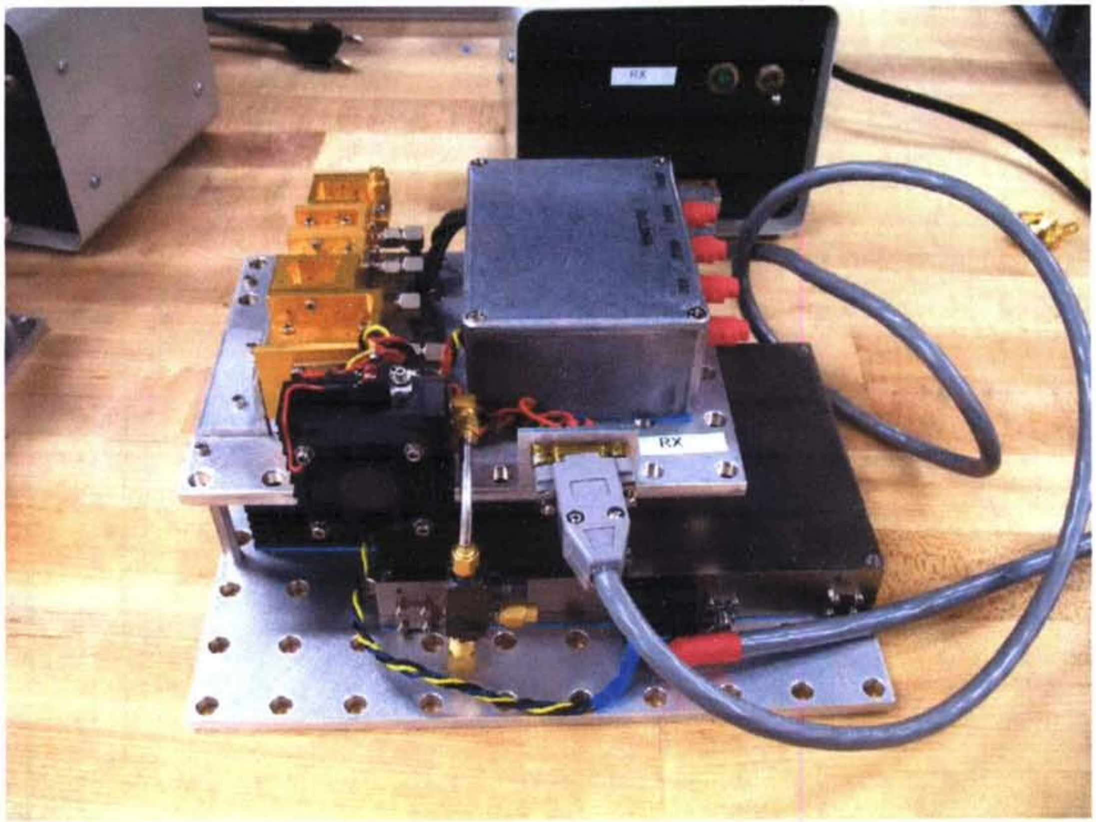

Figure 5: Photograph of 800 GHz receiver.

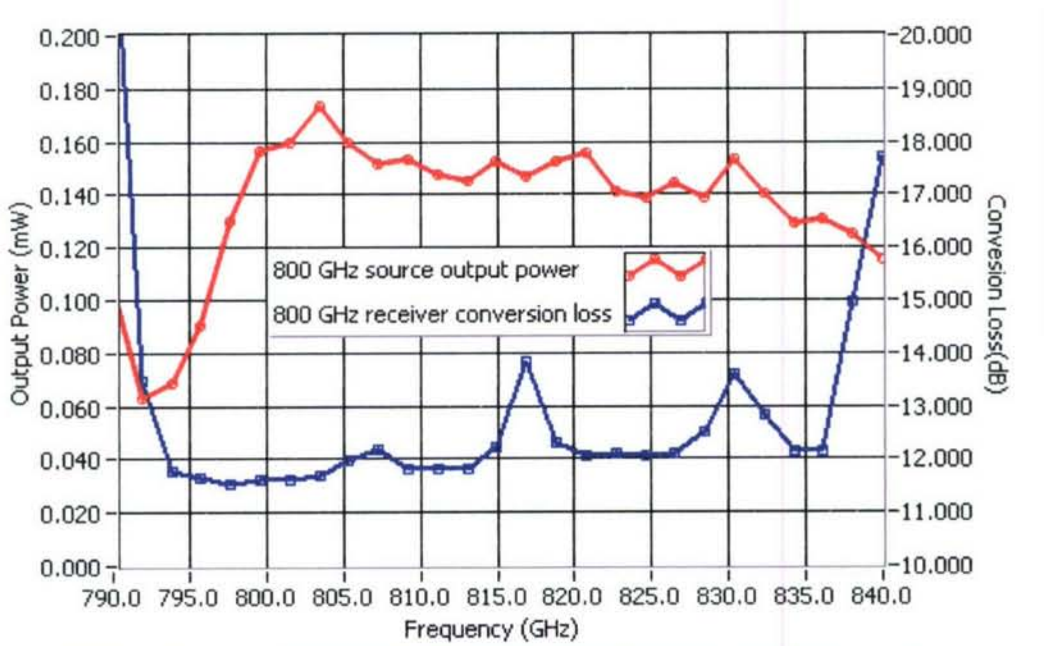

Figure 6: 800 GHz transceiver: source power in red and conversion loss in blue.

C. Measurement Results

1. Tunable Power

The delivered source gives power from 630-760 GHz with greater than 100 μW from 660-760 GHz as shown in Figure 7. The program can sweep the source across the bandwidth using any number of discrete points. The program takes about 25 ms per point, but the synthesizer takes 32 ms to change frequencies, so a 1000 point sweep across the band takes 32 seconds.

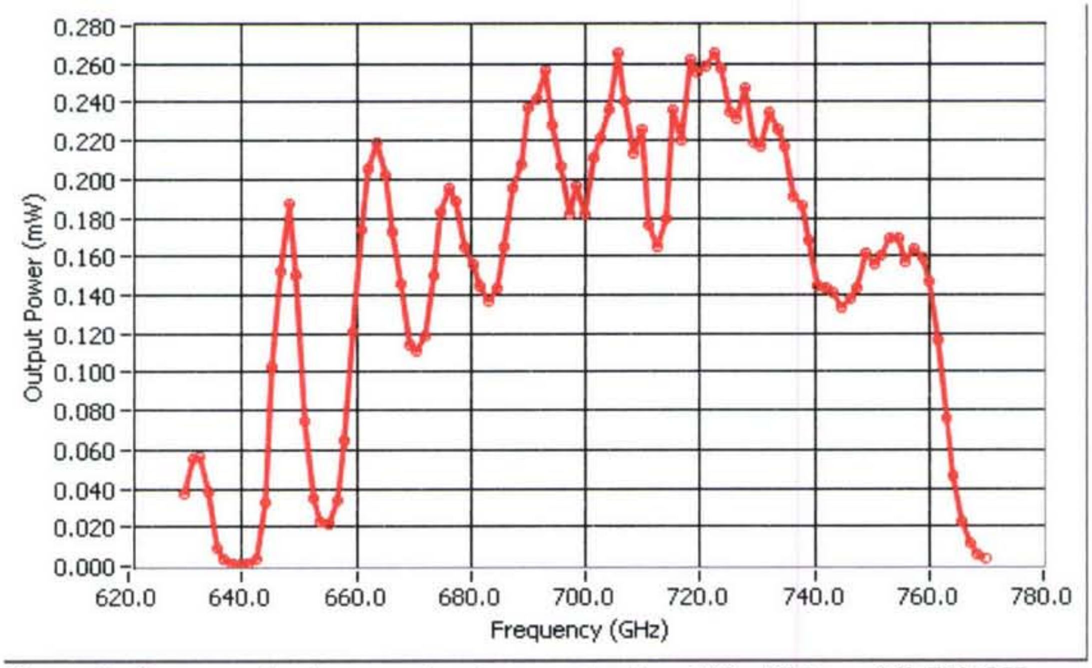

Figure 7: Source output power showing greater than 100 μW from 660-760 GHz

2. Spectral Purity

The synthesizer has a resolution of 1 Hz. The frequency multiplication factor is 48, therefore the output frequency resolution is 48 Hz. Sometimes, sources have power at unwanted frequencies called spurs. Spurs can appear for some input frequencies and disappear at others. A frequency selective power measurement from 660-760 GHz is necessary to measure spurs. Typical spectrum analyzers with extenders barely reach above 300 GHz, and even with state of the art microwave equipment, exhaustive spur searches can be very time consuming.

Perhaps the most effective way of searching for spurs is to use a 660-760 GHz receiver to downconvert the source's output within the range of a 50 GHz spectrum analyzer. Building a Terahertz receiver for this was not within the time frame of this

Phase 1, however a transceiver at 800 GHz with very similar components but less bandwidth was under development for another customer, and this system was used to make measurements for this contract. The 800 GHz receiver downconverted the source's output power to be measured on an HP8565E Spectrum Analyzer. The Local Oscillator of the receiver was fixed while the RF was swept across the band. There were no spurs found in the source within the noise floor of the Spectrum analyzer which was about 20-40 dB below the IF signal.

3. Phase Noise

Phase noise was measured using the 800 GHz transceiver with a spectrum analyzer. The 800 GHz output was downconverted by mixing with a local oscillator and then measured on a spectrum analyzer. Theory predicts that the output phase noise should be greater than the crystal oscillator by 20*Log(N) where N is the harmonic number. Figure 8 shows the comparison between measured and predicted phase noise for the 800 GHz system for different RF and LO frequencies. The spectrum analyzer cannot distinguish between amplitude noise and phase noise or between transmitter, receiver or analyzer noise. So, the plots in Figure 8 show the worst case scenario.

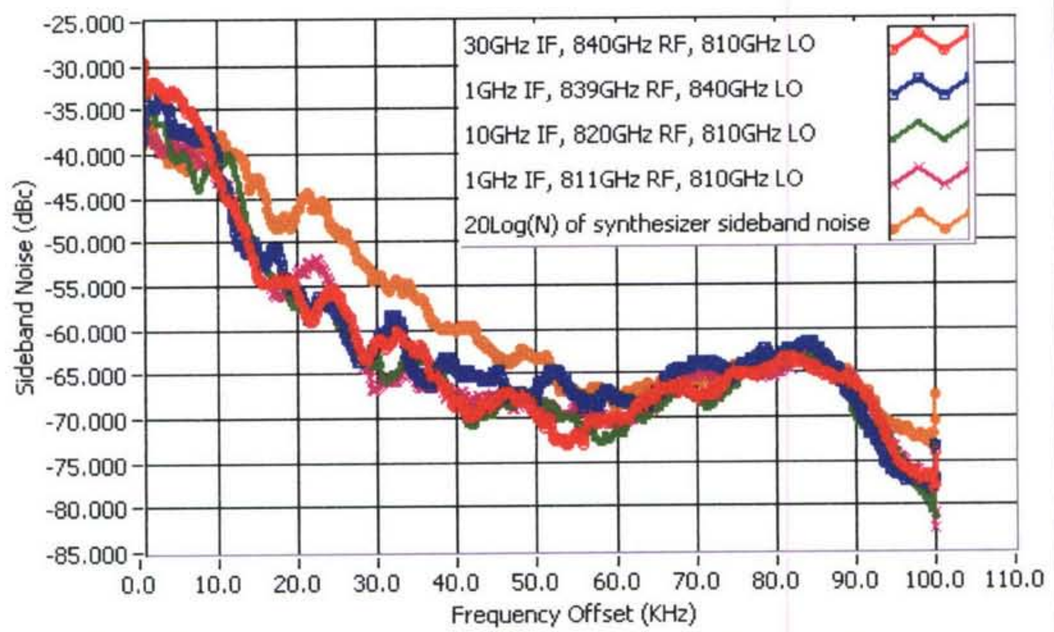

Figure 8: 800 GHz source phase noise for different RF and LO frequencies, compared to 20*Log(N) of the synthesizer phase noise.

4. Amplitude Stability

Amplitude stability is the most challenging specification for this contract. Various noise sources such as thermal noise, phase noise, and ground loop noise cause small fluctuations in the output power. These fluctuations can occur very rapidly or very slowly. The rapid fluctuations represent noise contributions whose frequencies are large compared to the sweep rate. The noise can be filtered or averaged by measuring over many cycles of the noise at each output frequency. The amount of time to average at each point can be determined by the amount of desired noise reduction and the noise power spectral density. Long averaging times would be necessary to achieve high stability in the presence of low frequency noise power.

At some point, the output power fluctuations occur so slowly or non-cyclically that averaging or filtering is not effective or results in excessively long time constants. These slow fluctuations are often described as "drift". In this report, amplitude stability is divided into two categories: noise and drift, where noise represents fluctuations that occur faster than 10 Hertz and drift represents fluctuations that occur slower than 10 Hertz. The methods for measuring and solutions for minimizing these two types of fluctuations are different. Ten Hertz is the dividing frequency because it is the low frequency limit for the technique used to measure the noise described in the following section. The feedback circuit used to maintain safe operating power levels is effective for reducing both noise and drift, so the following sections describe the differences in noise and drift with and without feedback.

a. Noise

The voltage amplitude stability in regards to noise fluctuations for this source is less than 0.045% rms with feedback. The noise was measured using a VDI WR1.2ZBD which is a detector that operates over the wr1.2 waveguide band (600-900 GHz) without needing a bias voltage. The detector has a diagonal feed horn that was placed very close to the source's output horn. The detected signal was amplified with an ultra low noise bi-fet operational amplifier, AD743 (gain=100), and then measured on a Tektronix TDS 744 oscilloscope. The performance of the source is so good that the noise is difficult to measure in the presence of the noise floor of the oscilloscope and amplifier. A large detected signal is necessary to be above the amplifier noise floor, therefore the source and detector horns were placed very close together and the detector was operated in the linear region where the output voltage is representative of the square root of the power.

The noise on top of the detected DC signal could not be measured accurately without first using the AC coupling mode on the oscilloscope to decouple the noise from the DC signal. The lower frequency limit for AC coupling with the oscilloscope is 10 Hz. Therefore, the noise measurements only include noise with frequencies greater than 10 Hz. The oscilloscope was set for 500 ms per division with 50,000 points. The total rms power for this sweep with feedback was 1.34 mV for a DC level of 2.9 V. The noise floor with a short circuit in place of the ZBD detector is 318 uV. The oscilloscope and AD743 noise contribution was removed by subtracting the square of the rms voltages (1.34^2-0.318^2) and then taking the square root.

Since the detector is operated in the linear region, the detected signal is proportional to the Electric Field or Voltage level of the oscillator. Therefore, the final % rms amplitude stability is the ratio of the result from the equation above divided by the DC level. The same process without feedback resulted in 0.16% stability, but most of this noise is from the fans. After stopping the fans, the rms voltage dropped to 0.036% without feedback and 0.028% with feedback showing that feedback reduces the fan noise but is not entirely effective. The fans provide a lightweight means of controlling the temperature, but they can easily be removed and replaced with the best heat sinking method for the application.

The FFT of the noise time waveforms gives noise amplitude versus frequency. The area under the FFT function gives the amplitude stability. The high frequency noise can be filtered or averaged to improve the amplitude stability by reducing the total area under the curve. Therefore amplitude stability can easily be calculated in terms of an equivalent noise bandwidth which can then be equated to sweep time. For example, to achieve 0.01% rms amplitude stability resulting from noise power above 10 Hz, the detected signal must be filtered with a time constant of ~4 ms resulting in a sweep delay of 4 s for 1000 points.

b. Drift

Drift can not be eliminated by averaging, because there can be a non-cyclical change in the power level slower than the averaging time period. Again, amplitude stabilization can be defined based on different time periods, such as over several seconds, over many minutes, or over many sweeps. Each time period of drift may have a different cause and a different solution. Perhaps the most significant cause of drift is due to temperature changes. But, these temperature changes can be in the Spacek amplifier, in the VDI multipliers, and in the feedback circuit, and each may need a different solution.

Feedback loops and temperature stabilization can decrease drift. The multiplier current feedback circuit was shown to decrease drift. Figure 9 shows a comparison between the output power with feedback and the output power without feedback as measured on the 800 GHz system described earlier. The source output was downconverted with an LO of 810 GHz and then measured on an Agilent E4419B Power Meter with an E4413A head. The system was operating at an arbitrary frequency and then adjusted to a new frequency. The graph shows the output power as a function of time beginning after the synthesizer frequency was changed. The non-feedback system shows a sharp change in output power after the frequency was adjusted until the temperature stabilizes, whereas the circuit with feedback stays relatively constant. Most of the power change results from the Spacek amplifier. The feedback circuit stabilizes the multiplier currents which are proportional to power after the Spacek amplifier. Therefore, the feedback circuit continually adjusts the amplifier's power to offset the temperature changes.

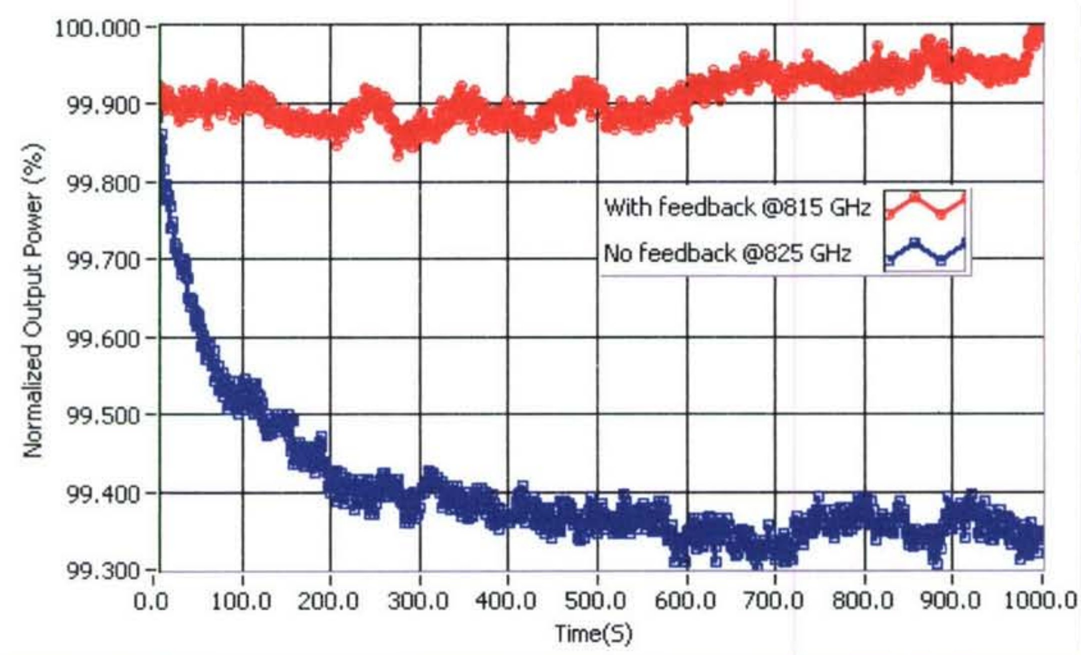

Figure 9: Drift for 800 GHz system with and without feedback.

Each component in the chain has a unique frequency response. Therefore, the power consumed in the devices and thus the steady state temperatures can vary as a function of frequency. If two frequencies have the same steady state temperature, then the output power will be stable when switching back and forth. However, when switching to another frequency with a different steady state temperature, the output power can vary until the temperature stabilizes. This effect is most pronounced without feedback, because all the components in the chain are involved. With feedback, the power changes of components before the device under feedback are compensated by the feedback circuit. Temperature stabilization that brings the same steady state temperature regardless of frequency can improve amplitude stability. Fans mounted on the source help to maintain temperatures, but they introduce noise as discussed in the previous section. Depending on the overall measurement system and technique, noise could be more troublesome than drift in which case the fans can be replaced with stationary heat sinking materials.

The amplitude stability resulting from drift with feedback for the plots in Figure 9 is about 0.006% rms over the first 100 seconds and 0.015% over 15 minutes. This was calculated by first taking the square root of the power to find the voltage, and then calculating the % rms variation around the mean. The power drifts 0.6% without feedback until the temperature settles, and then has 0.009% rms over the last 100 seconds. The drift over 15 minutes with feedback is most likely due to temperature changes in VDI's components on the output of the multiplier used for the feedback circuit. The multiplier current is a good indication of output power of a multiplier for a given temperature. The feedback circuit maintains a constant multiplier current, but temperature fluctuations change the relationship between multiplier current and output power. Also, at a given frequency, the D60 may be under feedback, but there are three more multipliers not under feedback after the D60 whose efficiency can change with temperature. Figure 10 shows the detected power as the temperature on the final output tripler was changed while under feedback. A DN515-2840 temperature chip from ThermOptics was mounted on the outside of the tripler block [5]. The chip tries to maintain constant temperature through a feedback loop. At first the chip was off and the output power was relatively flat, but then the chip was turned on after 100 s, and the output power dropped significantly until the steady state temperature was reached (2° C change). The heat from the chip was adequately isolated from the feedback circuit which shows that the changes are entirely due to temperature dependencies in VDI's components.

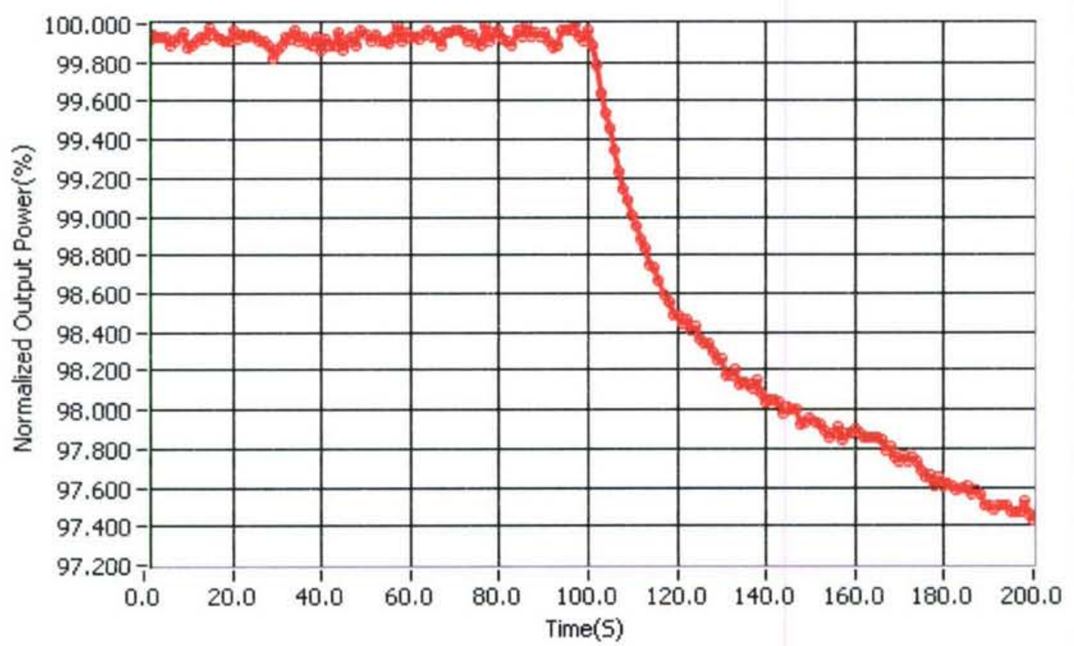

Figure 10: Normalized output power of 800 GHz source showing change in power after a heater changed the temperature by 2° C beginning at 100 s.

The drift due to noise and temperature fluctuations within the feedback circuit may also contribute to the overall output power drift. Measurements of the D142 current under feedback had 0.003% rms stability over the first 100 s, but 0.02% over 15 minutes. The D142 current is not directly representative of output power. For the non-feedback case in Figure 9, the D142 current changed by 9% while the output power only changed by 0.6%. If the same reduction applies for the feedback case then the 0.02% fluctuation in D142 current is most likely negligible.

The feedback circuit improves amplitude stability, but there is still more that can be done. First, noise and drift can be decreased in the feedback circuit. Next, temperature stabilization and feedback techniques can be used to stabilize the temperature without the use of fans. Finally, there needs to be a way always to operate under feedback. In the present system, if the multiplier currents drop below their limits, feedback has no effect. This occurs for low input powers and typically near the upper band edges where the optimum reverse voltages are large and the amplifier power is decreasing. Also, the feedback needs to occur as close to the output as possible. The last stage tripler, VDI WR1.2x3, can not be used with feedback, because there is no access to the bias currents. Perhaps a detector coupled onto the output with computer controlled frequency dependant limits could provide the best feedback stabilization for Phase II sources. If some or all of these changes are implemented for Phase II, 0.01% amplitude stability should be achievable over all time periods.

III. Conclusion and Future Work

The first half of this Phase I contract showed two chains that met the bandwidth and power requirements of the solicitation. The source with more power but less bandwidth was thoroughly analyzed, reported here, and delivered for completion of Phase 1. These two basic designs illustrate the power vs. bandwidth trade-off. In Phase 2, VDI will explore this trade-off further and attempt to find the optimal solution to achieve the overall program goals with the maximum power across the band and the lowest number of components required. VDI will also develop methods to alleviate this trade-off, including increased amplifier power, improved multiplier designs and optimized feedback control methods. A design plan for Phase 2 will be presented in the Phase 2 proposal.

There are many different methods for using a source in a transceiver system such as doing several continuous sweeps across the band versus stepping among discrete frequencies while averaging at each point. The chosen method can have important ramifications for optimizing the design of the source. For example, the synthesizer has extraordinary frequency resolution and control and can be set quickly, but can not be linearly swept in frequency. A free running YIG could be a better driver depending on the application. The synthesizer delivered with the source for Phase 1 runs from 2-20 GHz, so it should have enough bandwidth to be the fundamental drive source for all the frequency extension modules that would be delivered for Phase 2. But, any microwave source could be substituted in order to meet specific needs. Also, the feedback circuit may have some unexpected consequences when the source is used in certain applications. Since there are regions of the band where feedback is not engaged, the user can experiment with and without feedback to determine which is best, so that the Phase 2 effort can be directed towards insuring constant feedback, no feedback, or feedback where most convenient. In summary, with a healthy dialog, VDI's phase 2 efforts can be prioritized to deliver a source that will be most useful for DARPA's needs.

IV. Summary

A 660-760 GHz source was delivered for completion of the Phase 1 contract that has greater than 100 μW across the band. The source can be controlled by a computer through a USB connection with a program included on an installation CD. All the programming information was included in this report to allow full programming control for integration into a system. The source together with the control box is less than 0.022 m^3 and 22.2 lbs. The source frequency can be changed in 32 ms with a frequency resolution of 48 Hz. The amplitude stability varies from 0.028% to 0.16% rms depending on whether feedback is engaged, if fans are used to cool the devices, and the time period of interest. Without feedback initial power drifts of 0.6% have been seen until temperatures reach steady state. Feedback has the greatest stabilizing effect on slow fluctuations when engaged near the output. Stability most likely can be improved by careful temperature control. A table was included which tells when and how feedback is engaged versus frequency to allow the user to determine which operating mode is optimal in order to help determine the highest priorities for Phase 2 sources.

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