W Band MMIC Power Amplifiers for the Herschel HIFI Instrument

Robert R. Ferber PhD, Todd C.Gaier PhD, John C. Pearson PhD, Lorene A. Samoska PhD, Mary Wells, April Campbell
Jet Propulsion Laboratory, California Institute of Technology
4800 Oak Grove Drive, Pasadena, Ca 91125
Gerald Swift, Paul Yocom, K.T.Liao
Northrop Grumman Space Technology(NGST)
One Space Park, Redondo Beach, Ca 90278

ABSTRACT

This paper summarizes the development of W Band amplifiers for the Local Oscillator (LO) chains for the Herschel HIFI (Heterodyne Instrument for Far Infrared) Instrument. Key amplifier development issues and their solutions are presented, which have been applied on the way to realizing stable, wide-band amplifiers capable of producing 240 mW or greater RF power output across the 71 to 106 GHz frequency range. Development challenges addressed include: MMIC chip designs which initially had a variety of oscillation or "moding" propensities (mostly out-of-band), signal splitter and combiner development, matching of chip characteristics., power output control and leveling. The chosen design solutions are presented, including device, component and material selection for amplifier operation at cryogenic temperatures. Both room temperature and cryogenic (120 Kelvin) data is presented.

Keywords: GaAs, MMIC, mm-wave, W Band Amplifiers, Local Oscillator, Heterodyne Receiver, Herschel Space Observatory , radiometer

1. INTRODUCTION AND HIFI INSTRUMENT OVERVIEW

The HIFI receivers cover the frequency range of 480 GHz to 1900 GHz in six bands, using low noise superheterodyne receivers having superconducting HEB (Hot Electron Bolometers) or SIS (Superconductor-Insulator-Superconductor) mixers operating at about 2 Kelvin. These receivers, comprising the HIFI Instrument, will be the principal instrument of the Herschel Space Observatory. This European Space Agency observatory spacecraft is to be launched in 2007 to the L-2 Libration Point, where it will perform a 5 year mission to conduct a variety of astrophysics observations of star forming regions in the Milky Way Galaxy, as well as others.

NASA/JPL, with Northrop Grumman Space Technology(NGST,formerly TRW) as the principal contractor partner, is developing the amplifiers for the receiver local oscillator chains. The amplifiers, cooled to an operating temperature of approximately 130K, amplify a 0 dBm "W" Band signal up to + 23.6 dBm, or ~240 mW. The amplifier output is then coupled into a multiplier chain to produce the final LO injection signal in the range of 480 GHz to 1900 GHz.

This paper describes key amplifier development issues and their solutions, which have been applied on the way to realizing stable, wide-band amplifiers capable of producing 240 mW or greater RF power output across the 71 to 106 GHz frequency range. The HIFI power amplifier design embodiment is based on an A-40 silicon-aluminum alloy package with six GaAs(Gallium Arsenide) HEMT(High Electron Mobility Transistors) MMIC(Monolithic Microwave Integrated Circuit) amplifier chips used in each amplifier. Development challenges addressed include: connector and package design to fit available space, MMIC chip designs which initially had a variety of oscillation or "moding" propensities (mostly out-of-band), signal splitter and combiner development and matching across the band, matching of chip characteristics for those chips installed in the parallel power combined arms of the amplifier, power output control and microstrip cavity moding. The chosen solutions to produce the desired results are presented, including device, component and material selection issues for amplifier operation at cryogenic temperatures. Cryogenic (120 Kelvin) and power output leveling performance data is also shown for the amplifier.

2. W BAND MMIC CHIP DEVELOPMENT & PROBLEM SOLUTIONS

In Figure 1 we show a two-stage, 71-84 GHz MMIC power amplifier, of similar type as those in Ref. [1], fabricated by NGST, using a 2 mil thick, 0.1 μm GaAs PHEMT process. The design topology makes use of 4-way and 8-way microstrip power combiners to combine the FETs. The photograph indicates the two stages and the 8-parallel FETs combined in Stage 2. This type of cell is common for combining devices in parallel to achieve more power.

Early versions of this chip and similar chips for the HIFI Program produced oscillations in Ka-Band, which could not be corrected with off-chip bypass capacitors or other stabilization networks. The oscillations occurred between 30-50 GHz upon turn-on of the drain voltage, typically between Vd=0.8 V to Vd=2.4V. The strongest oscillation observed was at 33 GHz. Since the oscillations

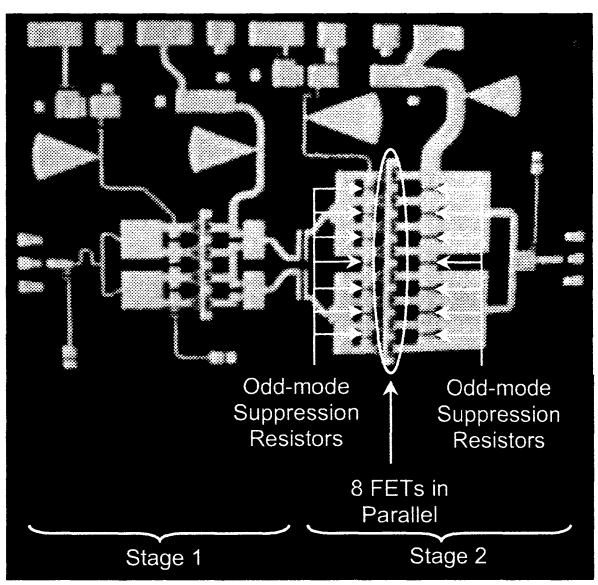

Fig. 1 Two Stage MMIC Power Amplifier Chip

occurred near the operating voltage of 2.5-3V and were expected to be more pronounced when cooled to the 120K operating temperature, we did simulations of the amplifier chips in order to understand and eliminate the oscillations. The analysis we have performed on the MMIC designs into the nature of the oscillations is described below. More details are found in Ref [2].

Two-port small-signal simulations of the amplifier chip in Figure 1 revealed no anomalies between 0-150 GHz, while the in-band stability factor, *K*, was greater than 2 between 60-85 GHz. In addition, the stability condition of each stage was carefully investigated to ensure that there was no overlap between the stability circles of one stage with the source mapping circles of the other stage. We then investigated two other ways to simulate the circuit, to cause the experimentally-verified oscillation at 33 GHz to appear in the simulations.

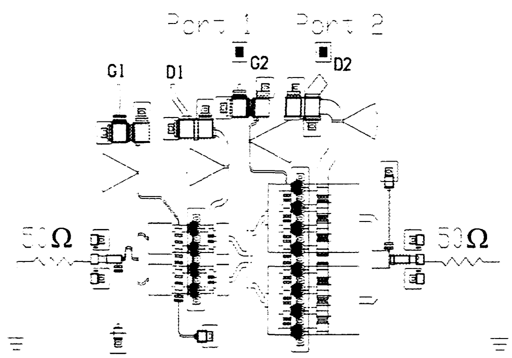

Fig. 2 Configuration for simulated *K*-factor analysis using the second stage bias lines as ports (Ref 3).

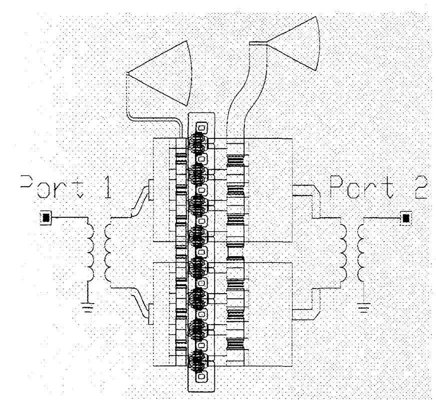

Fig. 3 Configuration for push-pull analysis of the second stage MMIC power amplifier.

We have performed two types of stability analysis. We focused our simulations on the second stage of the chip, since experimentally it appeared to be responsible for the 33 GHz oscillation. First, we performed a *K*-factor analysis using the bias lines in each stage of the amplifier chip as ports, as shown in Figure 2, according to Ref. [3]. The RF input and output of the circuit are terminated in 50 Ohms, while ports are inserted at the bias lines of the circuit. Port 1 is applied to the gate bias line while Port 2 is applied to the drain bias line. This method is usually applied to low frequency oscillations < 2 GHz, but it is also applicable to higher frequencies.

The second method we used to detect the instability in simulations was to apply a push-pull analysis to the second stage of the circuit, shown in Figure 3. In this method, the circuit is broken strategically at a point where the power is to be split into two legs, and an ideal transformer is inserted. The transformer serves to drive the two legs of the power splitter 180 degrees out of phase, thereby forcing an odd-mode condition. The transformer is applied at the input and output of the split circuit. Figure 3 shows the setup of the problem in the linear simulator, and the *K*-factor is calculated between the input and output of *only* stage 2.

In order to make the instability more obvious in the simulations, we enhanced the value of transconductance, Gm, to twice its nominal value for the simulations. Although this is a relatively high value to use, such a theoretical exercise is more likely to make potential instabilities appear, and then correcting the circuit for large Gm values will make it more immune to stability problems. For HIFI, we are also planning to operate the circuit at cryogenic temperatures(120-130K), and we had observed an increase in Gm and circuit gain upon cooling.

The two analytical methods described above revealed that the 33 GHz oscillation was an odd-mode oscillation occurring because the damping of the odd-mode resistors separating the power-combined PHEMTs was insufficient to suppress instabilities. In simulation, we changed the value of the resistors from 50 ohms to 10 ohms, and moved the resistors closer to the PHEMTs. The simulations also indicated that a transient oscillation was present at 47 GHz, due to the first stage of the circuit. We made additional corrections to the first stage and also reduced the value of odd-mode suppression resistors to 10 ohms. In general, we have found that keeping the resistors very close to the PHEMTs is critical to making them work to eliminate instabilities.

The results of the simulations were applied to the MMIC circuit designs for HIFI, covering all the possible L.O. bands from 70-113 GHz, and several new circuit layouts were developed. New wafer fabrication runs were done at TRW(HPA3A & 4) and measurements of the chips were performed at JPL using a spectrum analyzer to check for instabilities up to 50 GHz. Whereas the original circuits exhibited 33 GHz and sometimes 47 GHz oscillations, all of the second-generation chips were stable and the oscillations were successfully removed using the methods described above.

3. HIFI AMPLIFIER DEVELOPMENT

Operation at cryogenic temperatures presents unique challenges for this Herschel HIFI application. Size, weight and power consumption also had to be addressed. An aluminum chassis is attractive because of low mass and high thermal conductivity. Yet, owing to its high coefficient of thermal expansion, aluminum becomes a difficult choice for this application where a 200 degree change in temperature will be experienced. Iron based alloys could have been employed to achieve a low coefficient of thermal expansion, but at the expense of high mass and low thermal conductivity. For this application, A40 Al-Si alloy was chosen as a compromise to achieve a moderate and acceptable coefficient of thermal expansion, good thermal conductivity, and low mass.

The five Herschel HIFI LO bands that must be covered span the frequency range of 71 to 113.5 GHz. The most obvious waveguide choice for this frequency range is WR-10, which has a nominal range from 75 to 110 GHz. Although 71 and 113.5 GHz are outside of the traditional WR-10 waveguide band, 71 GHz is still well above the WR-10 waveguide cutoff frequency (59 GHz), while 113.5 Ghz is below the next higher order propagating mode in WR-10 waveguide (118 GHz). Since WR-10 waveguide was chosen as the waveguide medium for all of the amplifier input and output ports, a common amplifier chassis design is used for all Herschel HIFI LO bands.

Figure 4 is a photograph of the Herschel HIFI W Band power amplifier. The WR 10 waveguide input port is visible on the chassis. The output port is on the opposite end.

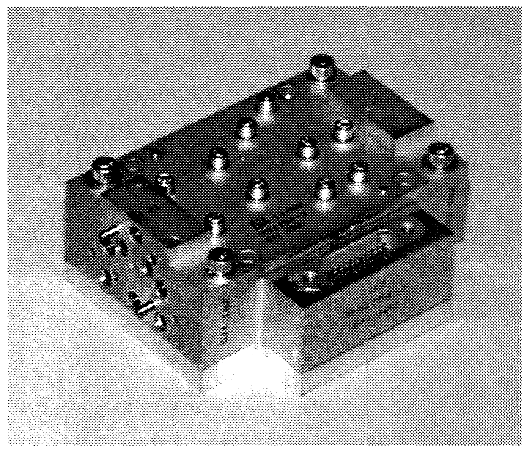

Fig. 4 Amplifier Photograph

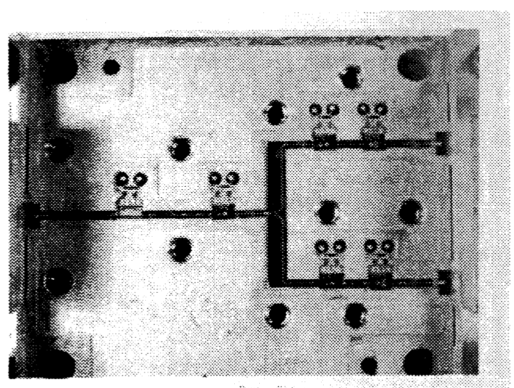

Fig. 5 Amplifier RF Cavity

Figure 5 shows the RF circuitry channelized inside the amplifier housing. Waveguide to microstrip transition is accomplished on the input port and on the output ports prior to power combining in the waveguide Magic "T" hybrid. The dc bias circuitry resides below the RF cavity and is brought up to the amplifier devices with glass bead feedthroughs, as can be seen in Figure 5. The six amplifier MMIC chips are clearly visible, along with the Wilkinson power splitter. The substrate material used for the microstrip elements is 125 micron quartz.

The TRW 0.1 micron GaAs process was chosen to achieve the required MMIC amplifier performance to well in excess of 100 GHz. The average Gm(transconductance) for the process is typically 695 mS/mm with Ft and Fmax greater than 120 GHz and 200 GHz, respectively, Ref. [4]. The Herschel HIFI LO bandwidth specification imposed a limit on the device size that could be employed to achieve power at such frequencies. A single power device might have been able to achieve close to the required 240 milliwatts from the amplifier, but this would have required that the amplifier be operated at drain voltages that might compromise long term reliability. For this reason, two power devices are operated in parallel, at reduced drain voltages to achieve the necessary 240 milliwatts output to drive the L.O. multiplier chains.

Several possible power combining approaches were examined for this application. These included Lange type hybrids, magic tee hybrids, branchline couplers, and Wilkinson and short slot hybrids. Figure 7 shows the various approaches considered. Figure 8 presents the results of the amplifier splitter/combiner trade studies.

For low loss power combining at the output of the amplifier, a waveguide magic tee hybrid was chosen. The magic tee hybrid provides close to full band performance. For this application, two tee-buttons were designed such that all five Herschel LO bands were addressed by use of only two magic tee designs.

A Wilkinson style power splitter was chosen because it is best suited for a low loss implementation in microstrip that is compatible with interconnecting substrates within the amplifier chassis. The five Herschel LO bands were covered with three Wilkinson designs.

Different combiner options that were considered.

A uses a Wilkinson power splitter on the input and output
B uses 90 degree phase shifter lines to improve return loss
C uses a magic tee for the output
D uses 90 degree lines to improve return loss
E uses magic tees for the input and output
F uses 90 degree lines to improve return loss
G and H consider alternate options with Lange couplers, and Short Slot Hybrids

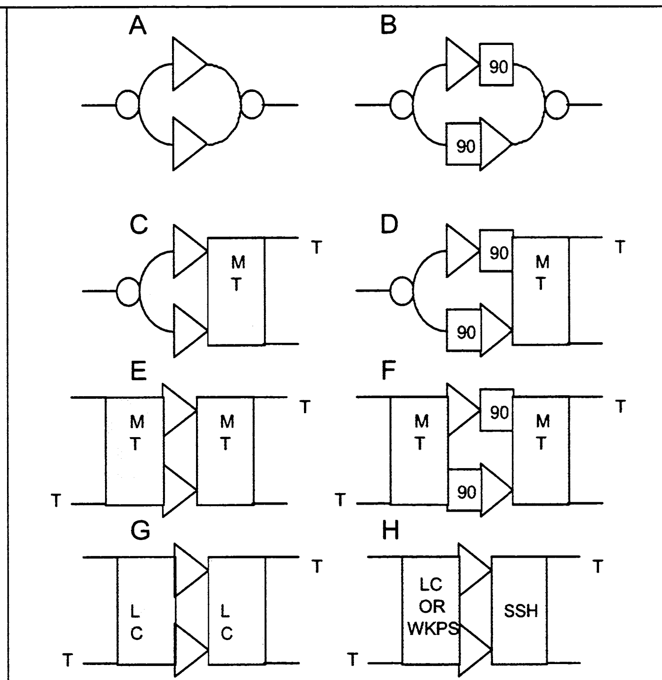

Fig. 7 Splitter/Combiner Options

Circuit	Advantages	Disadvantages
A	1) Low Cost 2) Allows single chip output	1) Moderate loss (1 dB?) 2) Output Return Loss not as good as "B"
B	1) Low Cost 2) Allows single chip output 3) Better Return Loss than "A"	1) More substrates needed 2) Better return loss than "A"
C	1) Magic tee provides good isolation 2) Magic tee is lower loss than Wilkenson	1) poorer output RL than "D" 2) Mechanical complexity
D	1) Magic tee provides good isolation 2) Magic tee is lower loss than Wilkenson	1) Mechanical complexity 2) Better return loss than "C"
E	1) Ideal for waveguide modules 2) Magic tee provides good isolation	1) Mechanical complexity
F	3) Ideal for waveguide modules 4) Magic tee provides good isolation	1) Mechanical complexity 2) Better return loss than "E"
G	1) Classic broadband balanced design 2) Good Output Return Loss	1) Lange Coupler Loss (2 dB?) 2)
H	1) Broadband 90 degree Hybrid 2) SSH works well with split block	3) Mechanical complexity 4) Less well characterized

Fig. 8 Splitter/Combiner Topology Design Trades

In-phase power combining and splitting techniques have the disadvantage that the resulting input and output return losses are for the most part, the same as the power-added gain of the devices that are being combined. Approach D of Fig. 7, using 90 degree phase delay for the combined devices, partially overcomes this disadvantage.

Consider the input Wilkinson power splitter when a 90 degree phase delay offset is included in one of the arms. The power is equally split, but the two signals arrive at the amplifier device offset by 90 degrees (at the design frequency). Reflections from the amplifier devices are also offset 90 degrees. On the return path, the same 90 degree phase delay offset is encountered and the reflected signals are 180 degrees out of phase when they arrive back at the power combiner. The reflected signals are then absorbed in the odd-mode isolation resistor, and ideally do not propagate any further. In this manner, excellent input and output return losses are achieved for the combined devices. An output isolator may not be required in this case.

Because the phase shift introduced by adding physical length in the combined arms is actually a time delay, the actual phase shift will vary as a function of frequency. As a result, it was necessary to employ three different lengths of substrate to achieve sufficiently close to 90 degrees across the five Herschel HIFI LO operating bands.

4. AMPLIFIER PERFORMANCE

Room temperature amplifier performance example

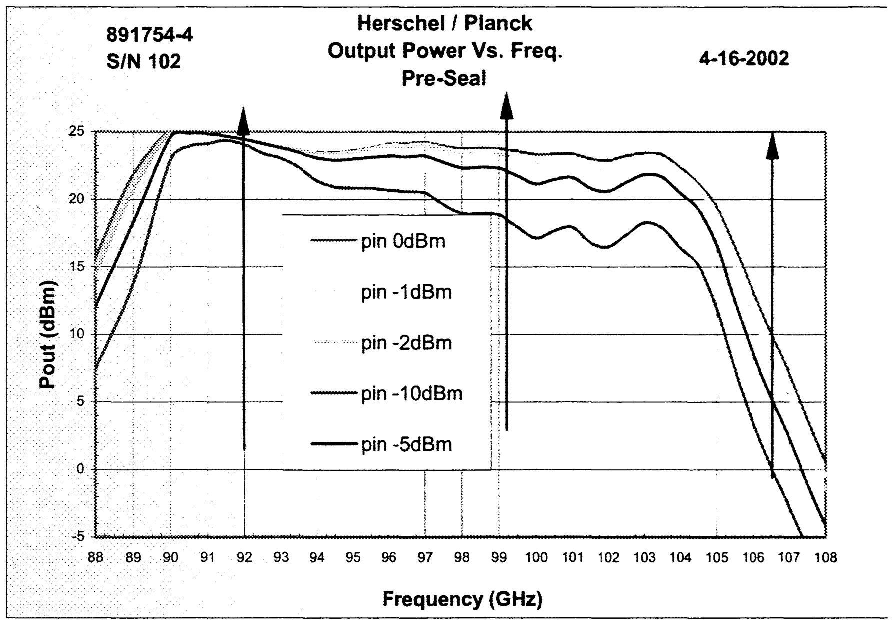

Fig. 9 Engineering Model(EM) Amplifier 102 Room Temperature Performance

Figure 9 shows the room temperature performance of Engineering Model(EM) amplifier serial no. 102. Greater than 23.6 dBm was achieved across the 92 to 106 GHz band, except for the high frequency edge of the band. Revised PA MMIC designs will correct this room temperature performance deficiency at the upper band edge.

Cryogenic performance example

Figure 10 shows the performance of EM amplifier serial # 102 at 120 Kelvin. A total of 9 EM amplifiers were built and all show repeatable and stable performance at room temperature and when cooled to temperatures as low as 60K.

5. POWER OUTPUT MODULATION AND CONTROL

A major challenge to the HIFI instrument is the adjustment of the local oscillator power so the mixers in the focal plane unit are optimally pumped. Additionally it is very useful as a diagnostic to be able to under or over pump SIS mixers. The result is a local oscillator subsystem requirement to be able to provide +3dB above the optimal and –10dB below the optimal LO power in something close to 0.5dB steps. Unfortunately no one knows what pump level the mixers really need or what the final in-flight or even in-lab coupling will be. As a result, power output control flexibility over at least 15 dB is desirable in the local oscillator subsystem, as is extra power margin capability.

The power amplifier control unit(power supply) has the ability to adjust the amplifier drain voltage over 0V to +3.4 V in 2^{10} steps. The amplifier control module also has gate control with –4V to +2V adjustment in a nearly arbitrary 2^{16} steps. The output amplifiers are operated in saturation, so controlling the drain voltage changes the output voltage swing. With two additional wafer iterations and many MMIC design modifications, it has now been confirmed that the amplifier MMICs do not oscillate on chip under any bias condition that has been imposed. As a result, it was possible starting with the later EM amplifiers, to experiment using the drain bias to control the output power.

Figure 10 shows the 120K power vs output drain bias from less than 1V to more than 2.5V for S/N 102. All the other parameters (gates, input drains, RF level) were kept constant. From Figure 10, it can be shown that the output power increases slowly with turn on, then rapidly between 1.3 and 2.3V, and then saturates between 2.3 and 2.95V. We have found that the best power output, when at 120K, occurs at 3.1 (+/-0.1) V, however, long term operation at this high drain voltage may create a reliability liability, especially at high levels of RF input power.

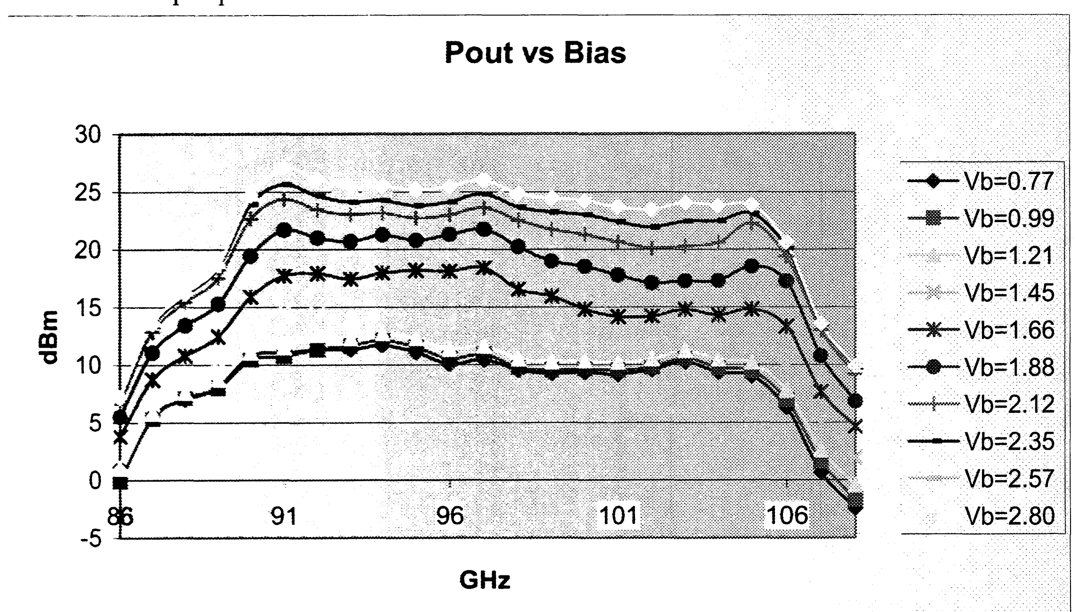

Fig. 10 S/N 102 power vs output drain voltage at 120K with +0dBm input

In order for the output power adjustment to function as desired, the power change with bias step needs to be evaluated. The bias supply can step in 3.3 mV steps, which correspond to approximately 60 steps between curves in Figure 12. The fastest change in output power is near 1.7 Volts, with S/N102 at 97 GHz between 1.45 and 1.65V. being the limiting case. This was 14.15 to 18.40 dBm or 4.25dB, which translates into a 0.072dB maximum step (if exponential). This is one order of magnitude finer than the 0.5dB adjustability needed by the HIFI system. A more typical case is the 79 GHz points between 2.5 and 2.7V, where the power changes from 24.04 dBm to 24.62dBm. This is 0.58dB or 0.010dB/step(if exponential). If the step is evaluated in milliwatts of output power, these cases range from 0.6 to 1.0mW change per drain voltage step.

The data shows that the power amplifiers can be used to adjust the RF power output by more than 13dB by controlling one bias line(drain 2). The local oscillator control unit is currently designed with sufficient resolution to step the output power in steps of 0.08dB or less per step. The variability of the power amplifier dissipation (assuming we need 1V to 3V to adjust the power) would be on the order of 2 Watts.

6. SUMMARY AND CONCLUSIONS

The NASA/JPL Herschel HIFI Project, in partnership with TRW, has developed GaAs "W" Band power amplifier MMICs, which are stable at all frequencies(no in-band or out-of-band oscillations or other instabilities. These MMIC chip designs have been developed in three design types or categories for use in the power amplifiers developed for the Herschel HIFI instrument receiver L.O. chains.

Using these MMIC devices, multi-chip amplifiers have been developed for the HIFI Instrument L.O. chains. The power amplifiers utilize six MMICs to produce up to 240 mW(or more when operating at cryogenic temperatures) "W" Band RF output to drive the LO multiplier chains. The Herschel HIFI amplifiers include single MMIC amplifiers, 5 chip DM amplifiers, and the 6 chip Engineering Model amplifiers described in this paper. The 25 flight and flight spare amplifier deliveries are expected to be made to SRON(Space Research Organization, Netherlands) and ESA(European Space Agency) during 2003.

7.ACKNOWLEDGEMENT

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