Chapter 4

Low Noise Amplifiers

Fighting for the Last Few Tenths of s dB in LNA Noise Figure .. Is it Worth it?

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Editor's note: *Dick posted this interesting discussion document on the Internet. We thought it so thought provoking that we just had to publish it in the Newsletter. While it is aimed primarily at the EME operator, there is much for all of us to consider. Our grateful thanks go to Dick for permission to use his document.*

Introduction

Some of this information came from the "Do You Need a LNA Line Driver" presentation by K1FO at the 3/17/01 NEWS (North East Weak Signal) Meeting at Enfield, CT, USA. In many amplifying devices, there is a considerable difference between tuning for the best gain, and the best Noise Figure (NF). Under certain conditions it is possible to tune a particular amplifier for best gain and realize a NF of 0.9dB. That same amplifier may deliver 0.3 dB NF, with slightly less gain (maybe 0.5 dB less), when tuned for best NF. At first you might say, gee, a 0.6dB NF improvement! Is that small improvement worth fighting for? The answer is YES, YES!

In the paragraphs that follow I'll attempt to show you:

(1) Why it is worth fighting for.
(2) Why it is hard to detect the improvement with typical laboratory equipment.
(3) What else in the system must be working right to get the full benefit.
(4) The Antenna Impedance problem.
(5) You'd better have the correct Gain Distribution.

(1) Why is it Worth Fighting For

You're about to have a demonstration that dBs of NF are not linear, when dealing with a cryogenic communication system and a small NF improvement can make a big difference in the Signal to Noise Ratio (SNR). A communication system becomes cryogenic when you aim the antenna into cold space.

Two EME System Examples -- Assume I have an excellent EME antenna that is presently aimed at a high elevation angle and it is aimed at a cold direction in the Universe, where the true Celestial Background is 2.73 Kelvin (the residual of the Big Bang, by Penzias and Wilson, Nobel Laureates, 1968). Under those conditions, that excellent EME antenna might have a total Antenna Noise Temperature (Ta) of 30 Kelvin (they [the experts] call it 30 Kelvin).

System (A) -- If your Low Noise Amplifier (LNA) had a NF of 0.9dB, that equals an electronic Noise Temperature (NT) of 66.78 Kelvin. The formula for NT is:

NT = [ALOG(NF/10) - 1] * 290.

Your total possible system NT (Ts) is now the sum of the Ta and the NT.

Ts = Ta + NT. Ts = 30 + 66.78 = 96.78 Kelvin.

System (B) -- If I were using that LNA when it is tuned at a NF of .3dB, the LNA's NT would equal 20.74 Kelvin. Now the Ts = 30 + 20.74 = 50.74 Kelvin. The only thing that is always linear about a cryogenic communication system is that the total Noise Power is proportional to the Ts in Kelvin. You have to almost ignore the dBs.

When I compare a system with a Ts of 96.78K (the A system) to a system with a Ts of 50.74K (the B system), that ratio is 2.80 dB. In other words, that EME signal will be 2.80dB further out of the noise when I use the LNA with the better tuning (I'm ignoring Moon Noise). That 2.80dB of difference can easily make the difference between a QSO and a missed QSO, when I'm listening to a weak EME signal.

A Better EME Antenna -- If I were able to make further improvements to that EME antenna so that the Antenna Temperature (Ta) was 20 Kelvin (and that is possible), then that ratio would be 86.78 Kelvin versus 40.74 Kelvin, or 3.28dB of system improvement in Signal to Noise Ratio (SNR). All of this is from a 0.6dB NF improvement (0.9dB to 0. dB NF) in the LNA! A ~ 3 dB SNR improvement for a 0.6dB NF change, that's a beautiful (apparent) non-linearity but it really isn't non-linear … it just looks that way (in dBs).

(2) Why is it Hard to Detect With Typical Lab Equipment -- When I make measurements with room temperature laboratory equipment, everything (the pads, signal generators, FM Receiver, etc.) is at approximately 290 Kelvin. In that 290K environment, a change in the LNA's NF from 0. dB to 0.3dB, can create a change in SNR of 0.6dB (at best). Here the dB' are linear. Unless you are using some good laboratory equipment (such as a SINAD Meter or a good Automatic NF Meter) you will probably not detect that 0.6dB SNR improvement. Your ear probably doesn't have enough discrimination to allow you to hear it, when you hit that "Sweet Spot" in NF tuning. As section (3), (4) and (5) will show, you will not be able to realise the benefit of that 0.6dB of NF improvement, unless the rest of the receiver has a NF of nearly 1.0dB or so, has the proper antenna impedance and the proper gain distribution.

(3) What Else in the System Must Be Working Properly to Get the Full benefit -- If I were using a receiver that was Gain Starved or had a second stage NF of 15dB, then the tuning of the LNA takes on an entirely different characteristic.

High Transceiver NF -- Many of the currently used Base Station 2m and 70cm Transceivers have a bare foot NF of 12 to 15dB. This occurs because Japan favours dynamic range over receiver sensitivity. Those Transceivers are front end Gain Starved. If you lived in a dense community where there was a Ham Radio Operator living on each street, you might agree with this approach.

Add an LNA -- Therefore, almost every American SSB operator must add a ~ 20 dB gain LNA in front of his Transceiver, if he desires full sensitivity of his communication system. If he doesn't add that LNA, all his SSB friends will eventually call him an Alligator (he is all mouth), instead of a Rabbit (a guy that is all ears). During a terrestrial contest, everybody can hear that Alligator call CQ, and they answer him, but because of his poor hearing aid (receiver), he only hears the locals and only responds to them.

The rest of the contest operators become frustrated, and learn to ignore him. Unless some local explains this to the "Alligator," he will conclude that there wasn't much activity during the contest. It is well known that the total system's cascaded NF (NFs) is equal to:

NFs = NF1 + (NF2-1)/G1 + (NF3-1)/(G1*G2) + ...,

where:

NF1 = NF of the first stage (as a real, anti-LOGed number).

NF2 = NF of the second stage (as a real number), etc.

G1 = Gain of the first stage (as a real number).

G2 = Gain of the second stage (as a

real number), etc.

Sometimes Best Gain = Best NF - If I were tuning a communication system's LNA, while it is connected to that bare foot Transceiver, I would find that the best system NF would be approximately the LNA tuning with the maximum gain. Here is an example: If you experiment with just the first two terms of that Cascaded NFs formula, and use 15dB for NF2, 0.3dB for NF1 and a gain of 10dB for G1, you will find that the System Noise Figure (NFs) [in dBs] is equal to 6.16dB. If, now, I use 0.9dB for NF1 and 10.5dB for G1, I realise a NFs of 5.98dB. Notice, that I worsened the LNA's NF by 0.6dB, while improving the gain by only 0.5dB, yet the system NF IMPROVED by 0.18 dB.

What this example demonstrates is that when your receiver system is gain starved, and has a high second stage NF, then the LNA's gain is much more important than it's NF. Even if I were using perfect laboratory instrumentation (such as a perfect NF Meter) while tuning that LNA in that environment, I would end up tuning it for near maximum gain, not best NF.

The conclusion is that to get the maximum benefit of a Super Low Noise LNA, you have to put it into the right environment while tuning it, and using it. Otherwise, you may be "casting pearls upon swine" … you could be wasting your time and your money.

(4) The Antenna Impedance Problem -- It is well known that when you are tuning an LNA for the best NF, you are primarily adjusting the impedance that the front end of the LNA is looking into.

It would be quite wasteful to carefully adjust an LNA stage while it is connected to perfect 50 ohm resistive laboratory equipment, and then connect it to an antenna with a VSWR of 1.41:1. That 1.41:1 antenna could be an impedance that consists of a capacitive (or inductive) reactance of 17.4 ohms in series with a resistance of 50 ohms, or it could be a pure resistance of 70.7 ohms, or 35.4 ohms. In any of these cases, that 1.41 VSWR antenna would drastically change the LNA's NF. Some of those possible impedances would cause more of a NF detriment than others, depending on the particular LNA design.

That uncorrected 1.41:1 VSWR antenna could easily raise your LNA's NF from 0.3dB to 0.9dB and hurt your EME receiver sensitivity by ~ 3dB. However, on transmit, the 1.41:1 would only cost you 0.127dB of transmission loss.

Your choices for correction are either to perfectly impedance match the EME antenna (with a double stub tuner, for instance) to make it look like a 50 ohm resistive load to the LNA, or do the NF tuning of the LNA while it is connected to the antenna -- such as by injecting the NF Meter's Noise Source through a 2 dB Directional Coupler (DC) that is always left in the antenna line.

The DC Line Perturbation -- If, after the NF tuning of the LNA, you made the mistake of removing the Directional Coupler, you would be changing the transmission line length, and that would rotate the antenna impedance to a different place on the Smith Chart. This would disturb the LNA’s NF tuning.

Another solution is to add a carefully chosen length of line to the DC's straight through path, so that the DC plus the extra line is an exact multiple of a half wavelength (electronically). Now you could remove that DC plus extra line, and not effect the antenna's impedance.

(5) You'd Better Have the Correct Gain Distribution -- To realise the system's best possible sensitivity requires that you have enough front end

gain and a low enough second stage NF but this requires a compromise of system NF versus Dynamic range. You usually can't have both all at once. The best possible system NF usually requires a lot of front end gain (sometimes 20 to 30 dB).
But, a system with that much front end gain will saturate 20 or 30 dB sooner from strong local signals -
- that's the problem that the Japanese equipment manufacturer's discovered.
Noise Power Saturation -- Also, bear in mind that even if you live in the "Out Back," and saturation from local operators isn't a problem, there can be another subtle detriment from the use of super high front end gains -
- Noise Power Saturation. It is possible that the later stages of your receiver are being subjected to so much Noise Power, from all the front end gain, that they are beginning to saturate on the instantaneous noise peaks. Even if that saturation is only a fraction of a dB, it can lower the SNR of a weak signal.
It is well known that a limiting stage will suppress a weak signal that's surrounded by noise with, what is called, "Signal-Cross-Noise Terms." In other words, it is possible for a super high gain system to suppress that weak EME signal you're trying to hear, in the later stages of your own receiver.
This phenomenon is quite subtle, and not easy to detect. But, if the gain in your system is shoving the S Meter above S7 from the basic Noise Power, than be wary, it could be happening to you. The only quantitative test procedure I know of to detect this condition is the "Notched Noise Power Fill-In Test," also called the Noise Power Ratio (NPR) Test.
IF Filter BW -- It is also possible that your system is going into and out of Noise Power Saturation, as you change the bandwidth of the IF filter. At first, you would think that the broader IF filter selection would aggravate the problem. However, it is possible that the more narrow filter selection allows less noise power into the final detection stage, and this in turn causes a smaller AGC voltage, which increases the receiver's gain, and causes Noise Power Saturation in an earlier stage.
Sun Noise Problems -- As your EME system becomes more refined, and you experience a larger number of dBs of Sun Noise measurements, it is possible that with the added Sun Noise power, your receiver system could be experiencing Noise Power Saturation. That would give you a pessimistic Sun Noise measurement. One simple method of detecting this problem would be to put a 6dB pad in various places (after the LNA), and repeat the Sun Noise measurement. If you get a better reading, you may have the problem.

The best system for high dynamic range is one that has a gain distribution that's just enough, at each stage, to override the NF of the next stage. The best system NF requires considerably more front end gain than that. Soon we will all pay more attention to the Noise Power Saturation characteristics of our tuneable IF receivers. Then we will simultaneously have the best system NF and high dynamic range.

A state-of-the-art 2.3GHz Pre-amplifier

Grant Hodgson G8UBN

Eagle-eyed readers may have noticed the reference to an ATF-54143 in John G3XDY's notes on the pre-amplifier testing results at the November Adastral Park round table, published in the March 2002 Newsletter. Here are some more details of this pre-amp which has some rather interesting properties :-

The ATF-54143 is the first of a new breed of low noise GaAsFETs. Released by Agilent Semiconductors (formerly HP) in mid-2001, it is less than one year old and offers some remarkable properties. It was designed for the mobile phone base station market, where low noise and good strong signal handling ability have to be achieved at the same time. However, it can be used up to at least 6GHz (although the gain is starting to roll off at this point), and is ideally suited for amateur microwave use in the lower and middle bands.

Agilent call the device an Enhancement-mode Pseudomorphic High Electron Mobility Transistor, or E-PHEMT. HEMTs have been around for some time now, and will be familiar to anybody who has built a microwave low noise amplifier (LNA). 'Pseudomorphic' is a development of the basic HEMT, and refers to the way the Gallium Arsenide is doped during device manufacture; the channel of the FET being made from many thin layers, which form a lattice structure. It sounds very impressive but is only of relevance to the semiconductor physicists involved in the details of the design and fabrication of the device itself.
(Incidentally, PHEMT is pronounced 'pee-hemt - not 'femt, but there doesn't seem to be a universally accepted way to pronounce EPHEMT'!) The really interesting bit is 'Enhancement'. Until now, all low noise and high power GaAsFETs for RF and Microwave use were of the 'depletion' type, which means that the device has to be biased into the operating region by ensuring that the gate is at a more negative potential than the source.

The two most usual ways of achieving this are by grounding the gate at DC and using resistors in the source lead(s), or by grounding the source lead(s) and using a separate negative voltage generator. Enhancement mode FETs require a positive voltage to be applied to the gate, which obviously makes things far easier in terms of circuit design and construction.
The second interesting feature of the ATF-54143 is it's strong signal handling performance.

Traditionally, low noise GaAsFETs have had significantly worse performance in terms of being able to handle either in-band or out of band strong signals than bipolar transistors or MOSFETS. The parameter most often used to describe strong signal handling performance is the 3rd order intercept point IP_3, either referred to the input (IIP_3) or the output (OIP_3) [1],[2],[3].

Note that the difference between the IIP_3 and the OIP_3 is simply the gain or loss of the device, so that an amplifier (or transistor) with an IIP_3 of +10 dbm and 6dB gain will have an OIP_3 of +16dBm.

The ATF-54143 has an OIP_3 of up to 37dB at 2.3GHz, with an associated gain of 16dB, giving an IIP_3 of +21dBm which is a very impressive figure indeed. The IP_3 is a function of the bias conditions, and this allows a trade-off to be made between IP_3, gain and noise figure.

On the higher microwave bands, strong signal handling is not usually an issue but there are at least two cases where good strong signal handling could be used to good effect, particularly at 1.3GHz :-

1) The 1.3GHz band has a close proximity to the frequencies used by the huge Civil Aviation Authority radars in some parts of the country. Some 1.3GHz receivers are consequently overloaded.

2) Repeater builders have to go to great lengths to ensure that the repeater output does not de-sensitise the receiver. Traditionally repeaters have required a low loss, high Q cavity duplexer in order to separate the Tx and Rx signals. A receiver front end with a very high IP3 may not be as susceptible to self-desensitisation, allowing the possibility of a lower specification receive filter.

The ATF-54143 could help in both of these situations. The noise figure of the ATF-54143 depends on frequency, but at 1.3GHz the NFmin (minimum noise figure if the rest of the circuit was ideal) is only 0.4dB, and only 0.5dB at 2.3GHz. This is not quite as good as some other devices, such as the ATF-36077, but for the vast majority of cases this will not matter for two reasons :-

i) Noise figure differences of one or two tenths of a dB are only relevant in very special cases for space communications, such as satellite and EME - for terrestrial applications, it makes no real difference to the received signal/noise ratio due to the relatively high level of background noise (approx. 290K).

ii) Use of good RF circuit design techniques, high Q microwave passive components and low-loss PCB materials can reduce the circuit losses in a new LNA design such that the degradation in noise figure due to the these components (i.e. components external to the active device) are very small indeed, typically less than 0.2dB.

'So what's the catch?'

The ATF-54143 is not expensive, currently being only £6 each, which is considerably cheaper than (for example) the Mitsubishi MGF1402. One of the biggest problems is the size of the device - it is obviously in a surface mount package, as with all new devices (apart from those with no package at all - i.e. bare die!), and the package of the ATF-54143 is very small with the leads on a 0.65mm pitch, so some form of optical aid is required when soldering. There is also the usual problem with all HEMTs in that there is a reasonable amount of gain at high frequencies, and this gain increases as the frequency is reduced. This leads to the possibility of instability anywhere from several hundred MHz to over 10GHz, and careful circuit design is required to ensure that the resulting amplifier is stable.

Circuit Description

The positive gate bias can most easily be derived from a simple voltage divider consisting of two resistors. However, the drain current is highly dependent on gate voltage, and the relationship between the drain current and gate voltage (transconductance, Gm) varies from device to device, as do the individual I-V curves. The ATF-54143 data sheet [4] gives details of an active bias circuit, which ensures that the bias conditions (drain voltage and current) are consistent from device to device, and offers a degree of temperature stability. The design presented here is based on the Agilent design, with some subtle modifications.

Referring to the circuit diagram **(figure 1)**, R1 and R2 form a potential divider which keeps the base of TR1 at a constant voltage of approximately

2.7V. The emitter voltage of TR1 is simply the base voltage + 0.65V. This sets the drain voltage of TR2 at approximately 3.4V.

The drain current is set by the resistor R3 to be approximately 30mA. The gate current is almost zero, and can be ignored for the purposes of biasing. The gate voltage is therefore the same as the collector voltage of TR2, and is regulated by TR2 such that conditions for the drain voltage and current are always met. If the drain current of TR2 was to rise for any reason (such as a change in temperature), the voltage at the emitter of TR1 would drop causing the voltage at the collector to drop. This would reduce the gate voltage, causing the drain current to drop, thus maintaining bias stability.

However, the emitter-base voltage of a bipolar transistor decreases as the temperature increases.

Therefore, D1 is included to compensate for this; if the temperature increases, the voltage drop across D1 will reduce, causing the base voltage of TR1 to increase, but the emitter-base voltage of TR1 will also decrease with increased temperature, and consequently the voltage at the emitter of TR1 will be almost constant over a wide range of temperatures. Measurements show that with this arrangement, the drain voltage of TR2 varies by only 100mV and the current varies by only 1.8mA over the temperature range -18°C to +60°C. The temperature performance of microwave circuits is often ignored, but masthead mounted pre-amps can be subject to extremes of temperatures, especially if mounted at the feed point of a dish which is pointed at the sun in order to make G/T measurements. The active bias circuit ensures that consistent bias conditions will be maintained over a wide range of temperatures. Note that this bias circuit could be adapted for use with a negative gate-biased FET to give a similar level of bias stability, but the author is not aware of any amateur designs that actually do this.

The bias conditions of V_{DS}=3.4V and I_D=30mA were chosen to give a low noise figure and reasonably high gain at 2.3GHz, the objective being to reduce the noise figure of the system as a whole. The bias conditions are set only by R1, R2 and R3.

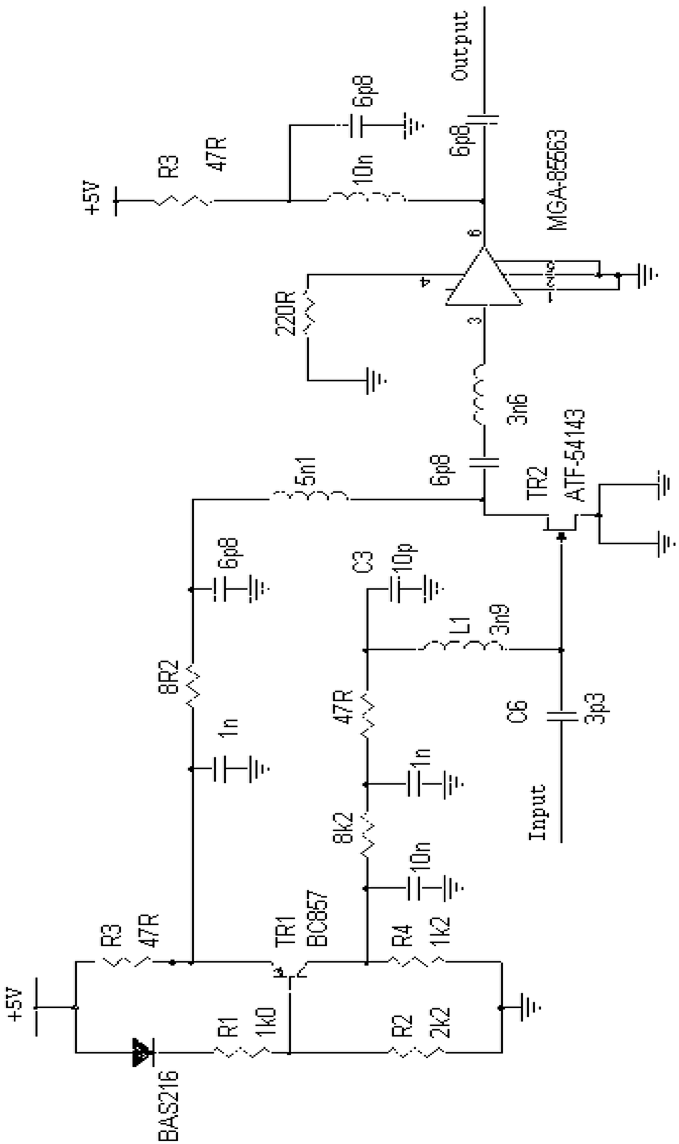
Source Degeneration

Traditionally, HEMTs and PHEMTs have required the absolute minimum inductance from the source lead(s) to RF ground for best performance. This is why FETs usually have two source leads; the inductance to ground is reduced which increases the gain. The ATF-54143 has been designed to allow some source inductance to be used in order to raise the input impedance to be closer to 50 ohms. The input impedance is of little importance in the majority of amateur radio applications, the exception once again being in repeaters where the performance of the duplexer is a function of the load presented to the ports.

For the majority of amateur applications, the input SWR of a low noise amplifier is of secondary importance to the noise figure and gain. However, the PCB (see below) has been designed to allow some source degeneration if required.

This is achieved by etching two parallel tracks for each of the source leads. One track connects to the device source lead, the other is connected to ground with a row of vias. A small (0402 size) zero ohm link is used to connect the source lead to ground; the position of the link setting the source inductance. For 2.3GHz (and above) the inductance is set to minimum, although the resulting impedance is not zero – the size of the zero ohm link acting like a short transmission line with a nonzero inductance. This causes a reduction in gain of approximately

Figure 1: 2.3GHz preamp: circuit diagram



2dB, although the noise figure is unaffected. More details of source degeneration can be found in reference [5].

Input match

As with all discrete low noise amplifiers, the optimum noise figure does not occur when the device is matched to 50 ohms. This causes all manner of confusion to some, but suffice to say that the best (lowest) noise figure is obtained when a certain mismatch occurs. The input impedance is transformed to this optimum impedance with C6 and L1; C3 ensures that the junction of L1 and R6 is at RF ground. In order to achieve the best noise figure possible, the losses associated with these input matching components must be kept to a minimum, which means that high Q components must be used. Traditionally, inductors at these frequencies would either be printed on the PCB or formed from wire of the correct size. The Q of printed inductors is limited by the loss tangent of the PCB material,
and it is difficult to change the inductance of a track once the PCB has been manufactured. The use of surface mount wound inductors allows for very high Qs, and also allows for a single PCB to be used at different frequencies. The cost of wound inductors have recently been reduced to partly due to intense competition between the manufacturers, which is good news for the microwave constructor.

There are no variable capacitors or inductors that need to be adjusted for best performance – this is a true 'no-tune' design.

Recent advances in ceramics technology (developed for the mobile phone industry) have led to the development of surface mounted inductors in a range of different values and sizes, some of which have a very high Q at low, and even middle, microwave frequencies. Once again, the bad news is the size of these parts – the highest Q inductors are in the 0603 or 0402 size. Whilst these are 'industry standard' sizes, they do pose some problems for the amateur at home – 0603 components measure approximately 1.6 x 0.8mm; 0402 components are considerably smaller! Hand soldering is possible with the right equipment, but the techniques are very different to those required for more conventional, larger components.

The inductor chosen for the input match in this design is a Coilcraft 0603CS-030 which has an inductance of 3.9nH +/-5%. The actual inductance varies with frequency, but the graph of L vs. f is very flat and the actual inductance at 2.4GHz is very close to 3.9nH. The Q of this inductor at 2.4GHz is approximately 100, and this high Q value is a major factor in keeping the noise figure of the amplifier down.

Second Stage

Some applications will benefit from a second stage of amplification. This is achieved with an Agilent MGA-85563 low noise Monolithic Microwave Integrated Circuit (MMIC). The MMIC is fairly easy to use compared to a discrete FET, requiring only one inductor to match the input and an RF choke and some decoupling capacitors on the output. An external resistor sets the bias current, and that's it!

The bias current of the second stage has been set to approximately 15mA. The second stage has a noise figure of approximately 1.6dB and a gain of 18dB. The IIP_3 of the MGA-85563 is approximately –7dBm, which is considerably worse than the OIP_3 of the ATF-54143. This means that the overall system IP_3 is limited by the second stage, not by the first stage, and the resultant IIP_3 is approximately –20dBm. However, as mentioned previously, the strong signal handling performance at 2.3GHz or 2.4GHz is not really an issue, and this level of performance will be found to be more than good enough.

'Universal' Printed Circuit Board
The PCB used for this project has been manufactured professionally with 0.4mm plated through holes and a tin-lead finish. This is obviously not the cheapest solution, but it is felt that this was only option given the small size of some of the components used. Professionally etched PCBs offer a high degree of repeatability which is essential for consistent performance, and completely eliminate any uncertainty due to etching errors or mistakes.

The PCB dielectric is a material called AR320, made by Arlon Inc. in the USA. This is a hybrid material of PTFE and fibreglass, offering some properties of both. The losses are much lower than with conventional FR4, but the cost of the material is much less than even the most common PTFE only material, Rogers Duroid 5880. Also, the hybrid material is much stronger than Duroid. The dielectric constant is (perhaps not surprisingly) about half way between the two, at 3.2 (hence the name AR320) and the thickness is 0.8mm. The PCB uses surface mounted components exclusively.

This allows the same PCB to be used for different bands; all that is required is for use at different frequencies is to change the value of the matching components, and possibly the bias resistors. It is hoped that the same PCB can be used for a number of different pre-amps from 432MHz or below to 3.4GHz and maybe even 5.7GHz. It is fully appreciated that 0.8mm Arlon AR320 is an unusual choice of dielectric material. This material was chosen for it's suitability to do the job, and also because the author has stock of this material, it being used on other projects. It is certainly not as readily available as FR4 or Duroid, and so 'home-brewing' the PCB will not be easy, if it is possible at all. (There may be an alternative PCB material such as Rogers 4003, although this has not been tried). However, this LNA is presented to show the levels of performance that can be achieved with a state-of-the-art design, not necessarily something that can be etched in the kitchen sink. If any constructors wish to have a go at etching their own PCBs then please get in touch with the author who will supply artworks. It may be possible to modify the PCB artworks for different dielectric materials, but this has not been tried and there are many pitfalls for the unwary.

Performance
The noise figure of the ATF-54134 increases with frequency, and much work has been done to reduce the noise figure at 2.4GHz to an absolute minimum, for use with the AO40 S-band downlink at 2401MHz. The resultant noise figure of the two stage amplifier is 0.6dB (Te=48K) and the gain is approximately 28dB. The ATF-54143 can be used as a single stage amplifier, in which case the noise figure is slightly less, and the gain is 13dB. The output of the first stage (and the input of the second stage) is matched to 50 ohms, making it easy to use the first stage on it's own if required. The simulated IIP_3 of the two stage amplifier is -20Bm. This is yet to be measured.

Further Developments
A 1.3GHz version of the pre-amp is currently being developed. The noise figure is expected to be approximately 0.4dB, although the biasing will be changed in order to improve the IP_3 performance which may lead to a slight increase in noise figure. The second stage will be re-designed in order to improve the IP_3. There are

other reasons why a high IP_3 is of advantage at 1.3GHz; full details will appear in a future edition of the newsletter.

Conclusion

The above design demonstrates that the amateur microwave community can benefit from the massive amount of research and development that is being carried out by the major component manufacturers. New devices are being introduced literally on a daily basis, and some of these devices can be put to good use by radio amateurs.
Other designs for amateur radio use of the ATF-54143 have been published [6], but the author is not aware of any other designs for the 2.3GHz band using this device.

The author does not claim any originality for this design; all the individual ideas and circuit blocks are used elsewhere in commercial receiver designs. However, these ideas are now put together and demonstrated in a practical way for the amateur microwave experimenter.

Apart from the very small size, the ATF-54143 has no real drawbacks. Now we have a device with very low noise figure, positive gate bias, superb strong signal handling and easy availability. Active bias circuitry ensures that bias adjustment potentiometers are not required, and the use of high Q lumped components gives a low noise figure and no need for tuning. So not only can we have our cake, we can eat it and have second portions as well!

References

1.Agilent Test & Measurement 'Spectrum Analysis Basics', Application note 150
2. ftp.agilent.com/pub/semiconductor/morpheus/docs/Measuring_IP3.pdf
3. R A Write, 'Spectrum and Network Measurements', Prentice Hall Inc., 1993
4. ATF-54143 data sheet http://literature.agilent.com/litweb/pdf/5988-6275EN.pdf
5. D VanStone 'Unique inductive feedback LNA design', RF Design, March 2002
6. Dubus Q1/2002

6cm Wideband Amplifier Using a Surplus Sat TV LNB

Mike Parkin, G0JMI

The wideband amplifier described here is based on the two-stage GASFET amplifier taken from an Amstrad "Blue Cap" Low Noise Block (LNB) satellite TV down-converter.

Performance

Measured gain for 1mW drive at:
2320.1MHz: 17dB approx.
3400.1MHz: 17dB approx.
5760.1MHz: 15dB approx.
Typical output power out:
2320.1MHz: 60mW
5760.1MHz: 50mW
Measured using HP Power Meter 430c into 50 ohms load.

Summary of Construction

The GASFET amplifier was built using two of the three amplifier stages from an Amstrad "Blue Cap" LNB TV receiver that link the input to the mixer.
The LNB module was obtained from a Rally for about £1. The pcb is removed from the LNB's box with access to it being gained by drilling out the four outside rivets and then removing the screws from the internal screening.

The screw holding the +5 Volts d.c. regulator needs to be removed and the output wire cut before the pcb can be removed from the box.

Using a sharp knife or scissors cut out the two-stage GASFET circuit as shown in **Figure 1**.

Remove the copper track runs for the integral tuned circuits and track runs as shown in Figure 1.

A pcb drill fitted with a small grinding tool was used for this purpose, however a needle file can be also used with care.

Note that the input capacitor C1 uses the modified first inductor where it is soldered into place.

Remove sufficient of the inductor to accommodate the surface mount capacitor.

Note that the output capacitor C3 uses the modified circuit track from the Drain of GASFET F2.

Remove sufficient of the track to accommodate the surface mount capacitor.

Capacitors C1 and C3 are desoldered from the remaining LNB board, use the "white" capacitors used to couple the MMIC amplifier chips together or to decouple them.
For 5670MHz use (6cm) leave the stripline capacitor in place. For 9cm and 13cm remove enough of this capacitor to accommodate another "white" surface mount capacitor recovered from the remaining pcb (C2).

Keep the GASFET +Ve supply and bias circuits intact, but remove supply lines.

Drill 1mm holes close to the +Ve supply circuits (marked R and C in Figure 1) and counter-sink them with a 4mm drill on the earth-plane side of the board to provide sufficient insulation for the +5 Volts supply lines that are fed to the supply circuits from under the board (P1 and P2).

Drill a 2mm hole as shown in Figure 1 for the -1.5 Volts bias supply.

Once this process has been completed, the box for the amplifier can be made up from double sided copper clad board as shown in **Figure 2.** Solder the LNB two-stage amplifier to the long side of the box (solder both the earth-strip on the upper side of the board and the earth-plane lower side).

Suitable holes are drilled in the shorter sides as shown in Figure 2 to accommodate BNC sockets that are attached using 8BA (or equivalent) short length nuts and bolts. Note that the input and out BNC sockets are not symmetrical because of the LNB board layout. Add the BNC sockets and then solder the end walls together to form three sides of the box (both inside where possible and the outside as shown in Figure 2). Solder the BNC socket inner connectors to points X1 and X2 as shown in Figure 1.

Once the three sides of the box are soldered together, solder the earth-strip of the LNB to the "output" wall and solder the lower side's earth-plane to the these side walls where possible on the inside.

Remove the 7805 +5 Volts Regulator from the remaining LNB. Drill a hole in the longer side wall to accommodate the Regulator under the LNB two-stage pcb (i.e. the earth-plane side) and secure with a 4 BA nut and bolt. Drill suitably sized holes to solder feed-through connectors for the +Ve Supply (12v) and the –1.5 Volts bias supply (e.g.: a battery). Solder the input rail wire of the Regulator to the feed-through, the earth rail to the side of the box and the output rail to the +5 Volts supply rails run to the RC supply circuits to the GaAsFETs F1 and F2.

The Regular's input and output rails are decoupled to earth using 1000pF disc-ceramic capacitors. Solder a suitable length of wire to the feed-through connection point on the outside of the box.

Next add the GASFET bias rail. This is thin strip of pcb with the two 22k Ohms pre-set variable resistors soldered to it. The "variable" rail from each variable resistor being carefully soldered to the RC junction of the surface mounted bias circuit on the LNB. Solder the remaining variable resistor rail to the wall of the box. Solder the –1.5 Volts d.c. supply wire (insulated) to the pcb strip and pass this through the 2 mm access hole mentioned earlier.

See Figures 1 and 2 for details.

Once the +/- Supply rails and the Regulator are in place, add the floor of the box that is also made from double-sided copper clad pcb.

Add the remaining wall of the box and solder wire tabs across the corners to ensure good connectivity.

Add a suitable wire to allow the –1.5 Volts to be connected.

Add two suitable length earth wire leads for the supply and bias.

Testing

Connect the amplifier's output (X2 BNC) to a suitable 13, 9 or 6cm receiver and connect the –1.5 Volts bias supply before the +Ve supply (12v was used).

The F1 and F2 Drain volts are set to about +3volts using a Digital Volt Meter to check the setting by adjusting the Bias variable resistors. (Check that F1 and F2 Drain volts can be varied from about +1.8 to +4 volts).

Connect a suitable aerial to the input (X1 BNC). Tune the receiver to a suitable signal source (e.g. beacon).

Adjust each 22k Ohms variable resistor for best maximum signal.

For use as a low power transmit amplifier, supply about 1mW of drive at 13, 9 or 6cm. Then adjust the 22k Ohms variable resistors for best output (one circuit gave about 80mW at 6cm.

In Use

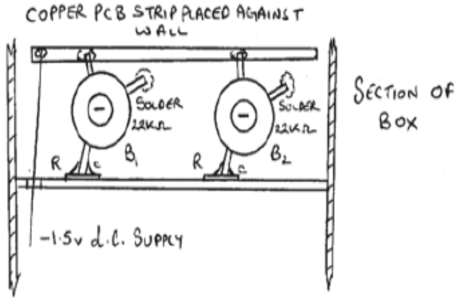
The design has been used from 13cm to 6cm on both the Rx and Tx systems for my transceivers.

Many QSOs have been made, including across the Channel, using these amplifiers as Rx and Tx amplifiers on these bands. For 6cm, two such modules are cascaded to give sufficient gain for the receiver.

A four-stage unit has been made up from two LNBs and tuned to act a multiplier from 1152MHz to 10368MHz, using two tuning pots made 13mm plumbing copper closed cylinders and copper 4 BA bolts to link stages.

FIGURE1: LNB BOARD CONVERTED FOR USE.

GASFET BIASING ARRANGEMENT



LNB MODIFICATIONS

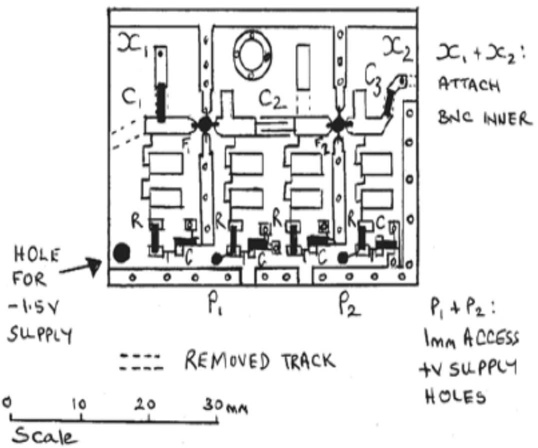
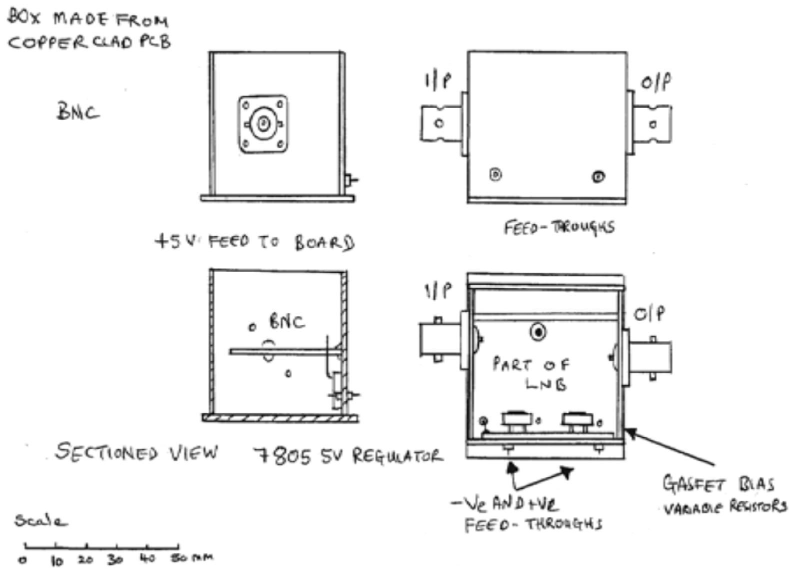
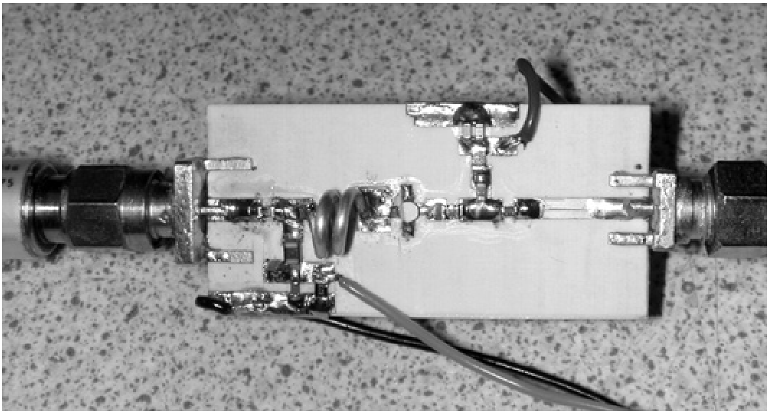


FIGURE 2: BOX CONSTRUCTION



23cm HEMT Tropo Preamp

Chris Bartram, GW4DGU



This preamp has been designed to be an excellent tropo preamp and a good second stage for EME. In practice, a pair of the amplifiers cascaded would probably be just about good enough for EME, but some redesign could probably shave 10K or more off of the noise temperature, at the expense of simplicity and possibly unconditional stability.

Please don't expect this to be a detailed constructional article. It's not. I'm trying to informally describe aspects of a project I've completed for my own entertainment and to possibly provide a small amount of inspiration for others. Experienced microwave equipment builders should find enough information here to duplicate the design but I hardly have enough time to earn a living, look after a small farm, and to play radio, let alone properly support a constructional project! If somebody wants to take the design further, and make PCBs available, I'd be happy to help, though.

I set-out to design a preamp covering the whole 1240 - 1315MHz range which would be easy to make with good, but not necessarily spectacular, noise figure, adequate gain, unconditional stability, good linearity and output return loss. The bandwidth of the preamp is very large. In gain terms, although not in terms of noise figure, it's still usable at 432! As it has good intermodulation performance, in many locations the amplifier could be attached directly to a 1296MHz antenna and not suffer from intermodulation and out of band issues.

Living in the primary service areas of two major TV broadcast transmitters, I can use the amplifier connected directly to a 1296MHz Yagi antenna without noticing intermodulation products but I may be lucky! In more stringent situations, a low-loss bandpass (or highpass) filter could be connected to the input. From a systems viewpoint, it makes good sense to separate the functions, rather than trying to design a narrowband preamp. If the filter and antenna have good return-loss at 1296MHz, the amplifier will still see something close to 50Ω, so the noise match will be unaffected and the noise figure of the filter-amplifier combination will be the sum of the preamplifier Noise Figure and filter loss. I'm working on the design of a simple-to-make, but very low loss bandpass filter, which I'll be putting ahead of this amplifier as a precaution. There are a number of options for a post-amplifier filter. My 'final' solution will probably be to use a second low loss input filter as an interstage filter, as my planned new transverter will have an excellent narrowband response from dielectric resonator filters.

Filters and antennas present impedances which can be far removed from 50Ω outside their passbands. For an amplifier to be stable with any combination of passive input and output load over the range of frequencies where the active device has gain is a highly desirable, but often difficult requirement to meet. It's even more difficult to prove in the real world. This preamplifier has been designed to be stable using the usual stability measures, and a detailed model been simulated to beyond 15GHz. It shows none of the usual signs of instability in simulation and I've so far failed to see any on the bench. That doesn't mean that there isn't a frequency somewhere in the spectrum where a combination of passive source and load impedances couldn't provoke instability. It just means that I've not yet found it!

Performance:

My prototype achieved the following performance at 1296MHz:

Noise figure: 0.55dB (T ≅ 40K)
Gain: 12.1dB
Input 3rd order intercept: -4dBm
Input 1dB gain compression: -11dBm
Output return loss: 20dB

The measured data agrees well with my simulations and also meets Bartram's First Law of LNA design: in the absence of linearisation circuitry, the input third order intercept of any low noise device, biased for low-noise operation, will be of the order of 0dBm. The plots were obtained from my VNA and written to a file via the IEE488 bus and then plotted using Open Office Calc running under Linux. The return loss graph looks a little noisy, as I accidentally read the data at 1dB resolution and didn't notice until I came to plot it.

My simulations also suggest that the noise figure remains good across the1.3GHz amateur band and is still OK at 1420MHz. I'd expect that to be the case, in practice, but I haven't yet measured the amplifier at other frequencies. Although apparently relatively simple, noise figure is a difficult parameter to measure accurately, particularly with modern test equipment! I am sceptical of many claimed noise figures. My measurement was made using a professional semiconductor noise source and my spectrum analyser, preceded by a low noise broadband amplifier. I guesstimate that the accuracy of my measurement is within about 0.3dB. A more accurate - and probably appropriate - method for amateur measurement of modern LNAs would be to return to the basic physics and use a low thermal mass 50 ohm termination dipped alternatively in melting ice and boiling water. However, that doesn't quite have the cachet of a £30k item of test equipment and it requires a modicum of understanding rather than the ability to read a display uncritically...!

Background

I have a number of Fujitsu FHX05 HEMTs in my component drawers following a successful EBay bid(!). OK, there are better devices, but not much better, and at £1 each.....?! Using Fujitsu's published device models, it was clear, following a lot of analysis, (using the Eagleware Genesys software

I use in my work) that it wouldn't be possible to guarantee unconditional stability out of band using a 'source feedback' topology. As an aside, although the source feedback topology has been around for twenty years or more, and is highly trendy - probably because it's possible to obtain a reasonable input match without degrading the noise figure - I've never found it possible to make a completely stable amplifier when using it. It's nearly always possible to find a region of instability, often at tens of GHz! I know of at least one other UK 1.3GHz preamp project (using Agilent PHEMTs with source feedback) which foundered for exactly that reason. For an LNA, good input match isn't actually necessary. Good input matching won't provide any more sensitivity (it's a matter of getting the right 'mismatch') but instabilities can completely wreck a potentially good NF.

The design that emerged after a number of simulations employs a conventional mismatched input circuit realised using lumped inductors for the input impedance transformation and gate bias feed and a shunt capacitance formed by a short length of microstripline. This forms a shunt-L, series-L, shunt-C network, which allows gate bias to be introduced at a relatively insensitive circuit node.

The HEMT source connections are grounded via 1mm diameter pins, cut and filed flush to the top (component side) surface of the PCB. This is critical for stability. The other component grounds are made by wrapping a piece of copper foil around the edge of the PCB at the appropriate places.

The output circuit is broadband. In order to control potential instabilities, caused effectively by the output resistance of the FET going negative at some frequencies, a small series resistor has to be inserted in the drain circuit. This causes a small degradation in the obtainable noise temperature but not as much as the device hooting away merrily at 30GHz! There is also a gain penalty, but gain is cheap, nowadays! The drain load is an inductor-resistor series network. This gives a good broadband output match.

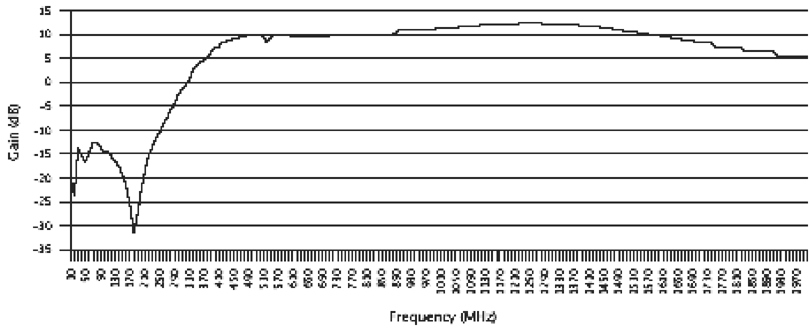
My prototype PCB was cut by hand from 0.75mm (0.030 inch) thick Rogers R4350 pcb material. I used just a scalpel, and a steel rule, working under a X10 binocular microscope. Using this technique I've made prototype LNAs to past 10GHz, and power amplifier boards to several GHz. R4350 is a low-cost, low-loss, PCB material intended for large quantity RF/ microwave applications. It can apparently be processed by using standard production techniques employed for FR4 material. It's probably not part of the manufacturers design remit, but R4350 is also particularly easy to work by hand! I have generated a set of Gerber files for the PCB layout and put them on the <www.blaenffos.org> website. These may be used for non-commercial applications.

The drawing here has been through a number of format translations and is adequate as an illustration. It isn't to scale and a number of 'funnies' have crept in. I'd recommend that even if you intend to adopt the scalpel approach to PCB production, you download a Gerber viewer, such as PREVUE and use that to print scale drawings of the PCB.

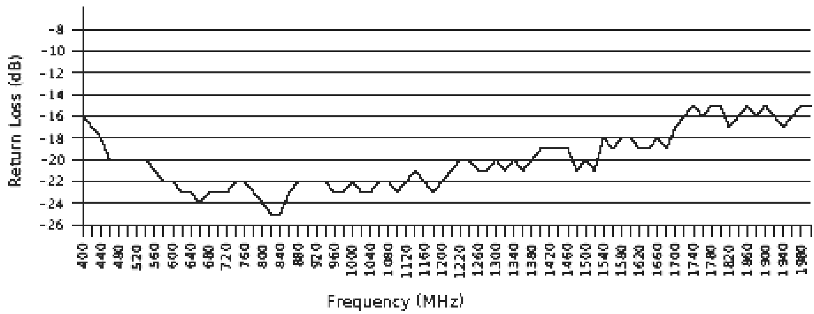
Components

The most critical components are in the input circuit. The 6p8 capacitor really needs to be a low loss part. I used an 0603 AVX 'Accu-P' capacitor. Porcelain capacitors could be used, and even, at a pinch, a standard 0603 COG cap although that would have an effect on the obtainable NF. The 18nH shunt inductor is reasonably critical. A low-Q device could degrade the noise figure by some tenths of a dB. My

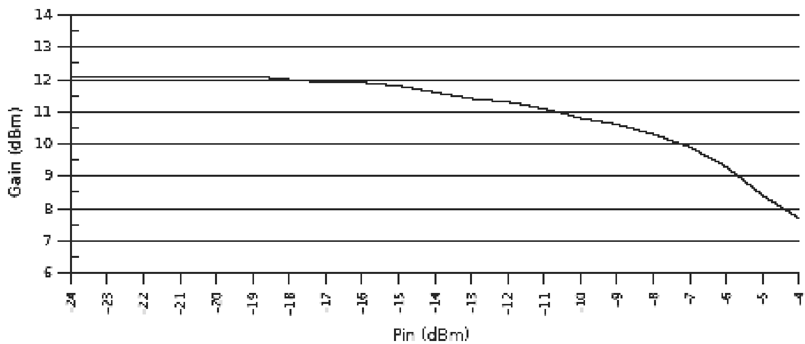
1.3GHz Preamp – Gain



1.3GHz Preamp -Output return loss



1.3GHz preamp - Power Sweep near compression



choice was a Coilcraft 0805CS series wound inductor. Perhaps the most critical part is the hand wound inductor. This is wound with a constant 2.5mm pitch on a 4mm mandrel e.g. a 4mm drill shank, such that the spacing of the centre of the start and finish of the winding is 5mm. There are no leads. Simply trim the coil so that it is a two turn helix. I used silver plated copper wire because I had some! Enamelled copper wire would also be suitable. In practice, unless high conductivity silver plating is used - and it's protected from corrosion, there's little advantage over copper. Don't even think of using tinned copper wire.

It may be necessary to slightly squeeze or stretch the inductor to optimise the noise figure, depending on exactly how the amplifier is built. If you feel this is necessary, thoroughly ground the gate of the HEMT before you do anything and remove the solder at one end of the coil. Make your modification, resolder and then remove the gate grounding. Otherwise you'll either lift a pcb track or damage the HEMT, or both! I know!

All other passive components were standard 0603 parts. The capacitors should have COG dielectric. In my case, the 5n6 inductor was a Coilcraft 0603CS wound part, but monolithic inductors would also be suitable. The FHX05 is the middle device in a series. It looks from the data sheet as though the devices are selected for noise figure in the 11GHz satellite television band.

The FHX04 has better guaranteed NF at that frequency, and the FHX 06, slightly worse. I suspect that there would be very little difference between them at 1.3GHz.

Enclosure

Although it might just be sensible to use a milled enclosure for a preamplifier built on a flexible substrate like ptfe/glass, that's really overkill. I tend to solder preamp pcbs into a lidless brass shim 'case'. This is more for physical protection than for RF screening, as I don't treat screening as a universal prophylactic! In this case, the bare PCB is small, the substrate material is relatively rigid and, providing the board isn't maltreated by being flexed, the amplifier will work entirely adequately just hung in the wiring! Flexing will break surface-mount components very easily and, even with a microscope, it's sometimes difficult to detect this visually.

Although there are, of course, many situations where good screening is mandatory, it can bring its own problems. This is particularly true of amplifiers using devices with bandwidths of tens of GHz.

There are hazards in packaging microwave amplifiers: putting-on lids leads to many potential problems as it is frighteningly easy to excite cavity resonances. Identifying and killing these can be as much work as designing the amplifier in the first place! HEMT precautions It's very easy to damage HEMTs and not realise it. They are extremely susceptible to static damage during assembly and from supply line transients in operation. This doesn't usually show as a significant change in the dc parameters, just as a change (for the worse!) in noise figure. Be VERY careful and take extreme precautions to avoid static damage when soldering the device into circuit.

Power supplies

The preamp requires a drain supply of +2.8V, 10mA. The gate bias required for this will be of the order of -0.6V. As HEMTs, along with most LNA devices are susceptible to supply-line transients, I now operate my preamps from isolated power supplies. In this I use a cheap packaged transformer-coupled inverter which converts 8 - 36V dc into ±5V, this largely isolates the preamp from supply line transients, and lets me power the preamp from my 28V antenna relay supply, as I energise the antenna c/o relays on receive. Following the inverter, I filter and clamp the ±5V lines and use these to power active bias networks. I'll write-up this 'preamp power supply for the paranoid' in the near future, along with details of the sequencing circuitry I use.

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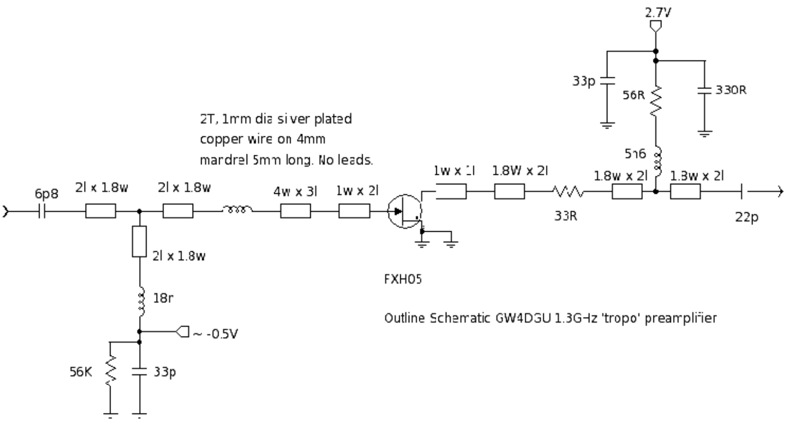
He is accessible by email at:
gw4dgu@blaenffos.org



Outline Schematic GW4DGU 1.3GHz 'tropo' preamplifier

PCB