InGaP HBT MMIC Development

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Abstract

InGaP HBT is being increasingly adopted as the technology of choice for low voltage PA's, integrated VCO's and broadband DC coupled amplifiers. Devices with high Fmax can be fabricated with moderate emitter geometries and proven, commercial parts are widely available. This article discusses a commercially available InGaP HBT MMIC process and presents the design and measured performance of two specific MMICs.

InGap HBT Technology

Conventional GaAs HBTs use GaAs/AlGaAs to form the heterojunction. They have the benefits of improved noise figure and increased Fmax as compared to conventional Si BJTs of the same geometry. Devices fabricated with a 2µm emitter width can have an Fmax of around 35GHz. GaAs HBT's offer performance advantages over PHEMT and MESFET based processes for certain circuit functions. They are well suited to realising:

- Low phase noise oscillators
- DC coupled amplifiers
- Circuits that require digital control functions on the same die
- Single positive supply, low operating voltage, high efficiency PA's

A more recent development in HBT technology is the Indium Gallium Phosphide (InGaP) HBT, which has a heterojunction of GaAs/InGaP. They offer the same functional advantages as conventional GaAs/AlGaAs HBTs but have a number of additional advantages:

- Higher Fmax (a 2µm device can have an Fmax of over 45GHz)
- More reliable
- Easier to manufacture
- Lower phase noise
- Better linearity
- Improved temperature stability
- Increased current gain

In addition to this, InGaP HBT processes are no longer capitive processes of major Integrated Device Manufacturers (IDM's). They are now readily available at pure-play GaAs foundries such as GCS.

Design Examples

Two Plextek designed InGaP HBT MMICs have recently been manufactured and evaluated:

- A 5GHz VCO for 802.11a/HiperLAN applications
- A DC to 12GHz broadband amplifier

A photograph of one of the VCO's is shown in Figure 1 and a photograph of one of the amplifiers is shown in Figure 2. Both circuits were fabricated on a multi-project mask set and therefore the size of both die has been increased to allow arraying. This is particularly evident with the amplifier where a length of 50Ω transmission line connects the RF output bondpad to the amplifier. A custom mask set for this amplifier would allow the die area to be reduced and over 30,000 die could be manufactured on a single 4” diameter wafer.

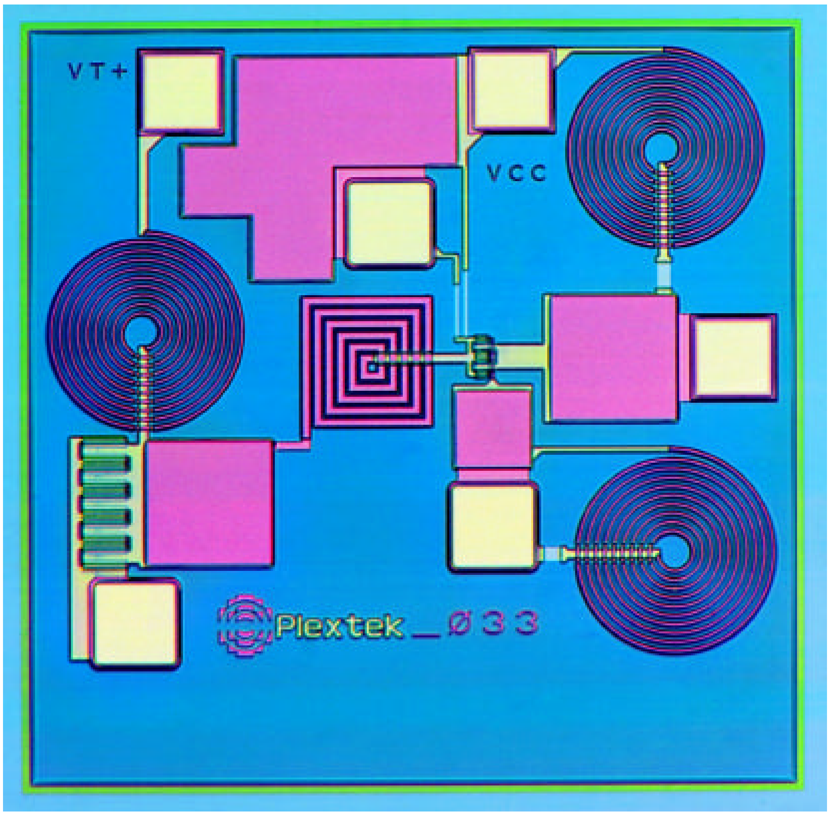

Figure 1: Photograph of the 5GHz VCO

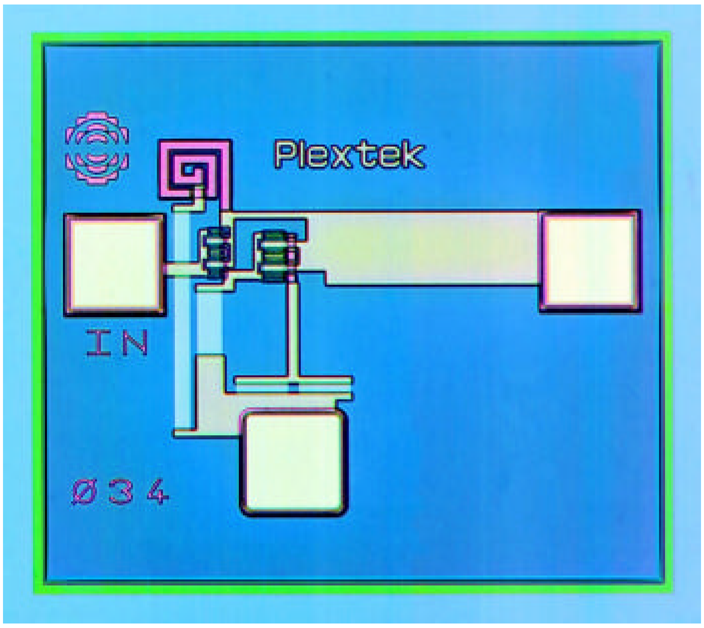

Figure 2: Photograph of the DC to 12GHz amplifier

The VCO design is a Colpitts-based negative resistance oscillator with a series LC resonator. An on-chip varactor is used to allow tuning of the oscillation frequency. The oscillator transistor has four emitter fingers of 2x12µm. The varactor utilises the base-collector PN junction of the HBT and the resonator inductor is a "square" spiral. Lower Q, more compact, circular spiral inductors are used as RF chokes. A schematic of the VCO is shown in Figure 3.

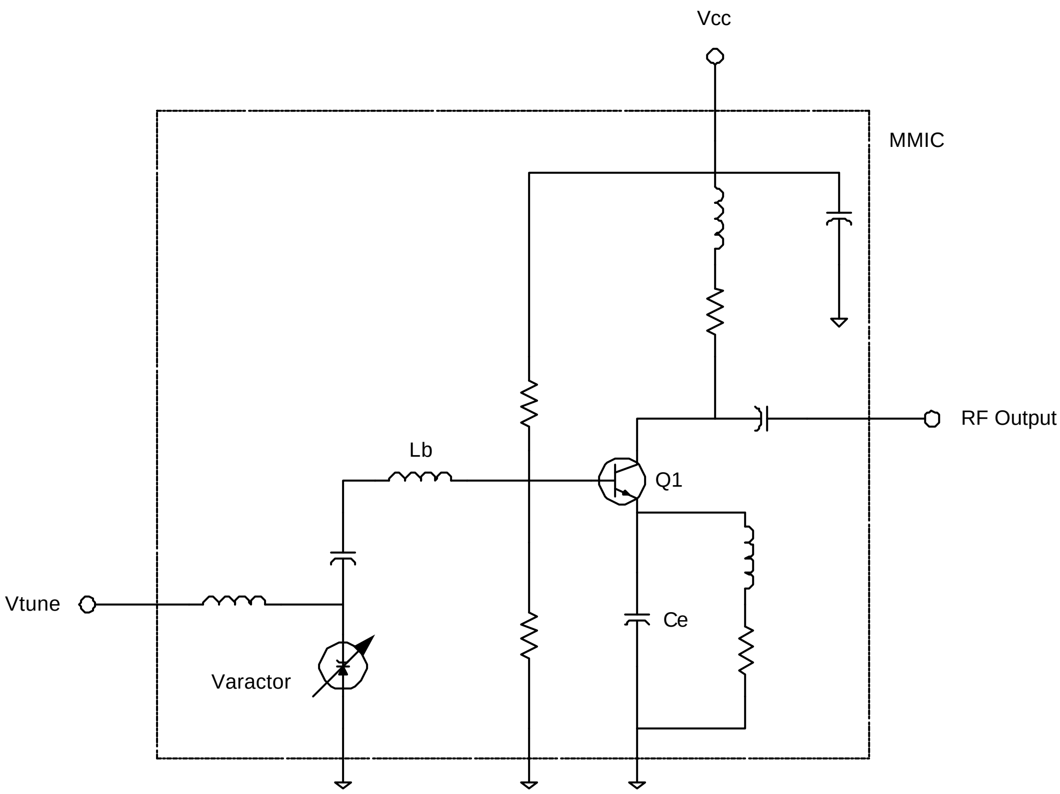

Figure 3: Schematic of the 5GHz VCO

The broadband DC coupled amplifier utilises a Darlington transistor pair. The RF output is collector coupled and a DC supply voltage has to be applied through an external bias tee. A schematic of the amplifier including a simple external bias tee is shown in Figure 4. The DC blocking capacitors at input and output are included to avoid a DC potential being presented to any adjacent stages.

The on chip resistors are selected to both bias the transistors to the required operating point and to optimise the RF performance. The RF performance is optimised to achieve a flat gain response with input and output impedances of 50Ω. Series resistive feedback is used on the output stage (Re_2) together with a small amount of series inductive feedback to improve the high frequency match whilst simultaneously setting the required bias. Shunt resistive feedback around the whole amplifier is also combined with inductance (Lfb). This serves to reduce the effect of the shunt resistive feedback at the top of the band so increasing the amplifier gain.

The amplifier is designed to operate with a DC voltage at the RF output (common to the collectors of both transistors) of 4V and a total supply current (Icc) of 37mA. The DC bias must be connected through an external resistor (R_1) to a DC supply of a higher voltage (Vcc). The value of the external bias resistor R1 is selected by Equation 1.

Equation 1: $$R_1 = \frac{Vcc - 4}{Icc}$$

Reasonable margin is required between Vcc and 4V in order to improve bias stability and reduce variation with temperature. On the MMIC itself, Q_1 acts as a simple constant current source. The collector current for Q_1 is determined by the voltage at its base (Vb_1), its Vbe and the value of the resistor Re_1, as given in Equation 2.

Equation 2: $$IC_1 = \frac{Vb_1 - Vbe_1}{\mathrm{Re}_1}$$

The voltage Vb_1 is in turn set by the DC voltage at the output of the amplifier (Vout) and the potential divider formed by Rfb and Rb_1. Adjusting Vout therefore adjusts the current IC_1. The current through Q_2 is set in a similar manner with IC_1 and Re_1 determining its base voltage (Vb_2). Thus adjusting Vout also adjusts the current IC_2.

The fact that there is a voltage drop across R1 provides a degree of stability to supply variation and helps reduce variation with temperature. For bipolar transistor the temperature coefficient of Vbe ($\Delta Vbe/\Delta T$) is negative. This means an increase in temperature tends to cause an increase in collector current. However, any increase in supply current would also cause an increase the volt drop across R1 so reducing Vout and therefore stabilising the supply current.

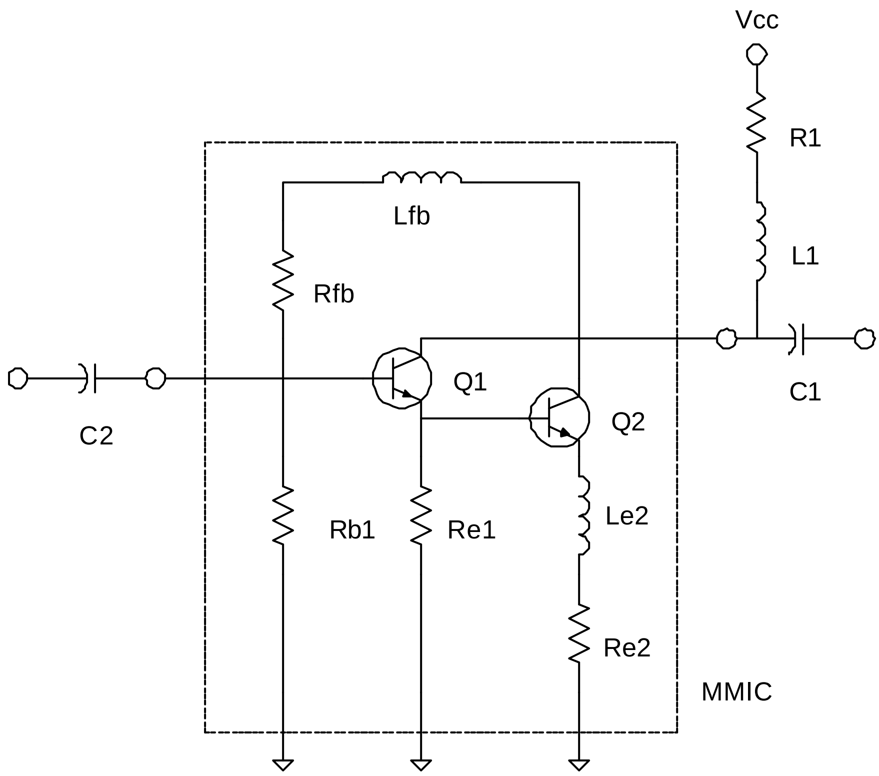

Figure 4: Schematic of the broadband amplifier, no external matching

The operational bandwidth of the amplifier can be extended significantly by the use of external matching. A schematic of the amplifier including external matching is shown in Figure 5. The majority of the components required to provide the bias tee can be adjusted in value to perform the matching function. An additional capacitor (C3) is included to provide an RF ground to the bias inductor L1 and utilise it as a matching component. The matching network is configured to reactively match the top end of the band so increasing the return loss and gain. The DC blocking capacitors are reduced significantly in value and the lower frequency cut-off is therefore increased. This matching structure can be implemented using conventional, low-cost 0402 components and can extend the upper operating frequency by 4GHz. Measured results are presented later in this article.

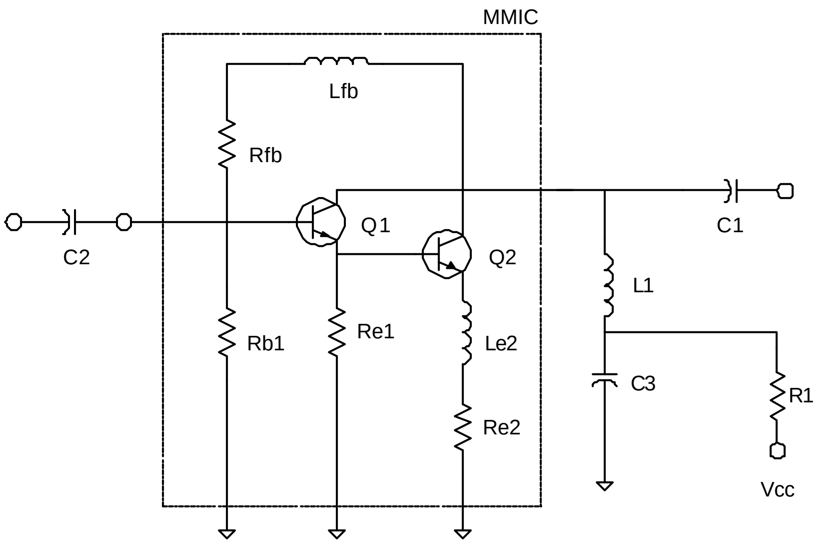

Figure 5: Schematic of the broadband amplifier, with external matching

Measured Performance

The die were assembled onto test PCB's realised on a brass backed PTFE substrate. Measurements of the test PCB's were performed in a test fixture and a Through Reflect Line (TRL) calibration PCB was also fabricated to allow the effects of the test fixture to be calibrated out.

Table 1 provides a summary of the measured performance of the VCO compared to simulated. The oscillation frequency, output power and efficiency are all in close agreement with the simulated performance, whilst the measured phase noise was lower than anticipated. Figure 6 is a plot of the output spectrum of the VCO showing a phase noise of –105dBc/Hz at 100kHz offset.

Parameter	**Simulated**	**Measured**
Centre frequency	5GHz	5.072GHz
Output Power	9.7dBm ± 0.5dB	9.5dBm ± 0.5dB
DC Power Consumption	12mA @ +5V	15mA @ +5V
Efficiency	26%	20%
Phase Noise	-100dBc/Hz @ 100KHz offset	-105dBc/Hz @ 100KHz offset

Table 1: Summary of VCO Simulations/Measurement

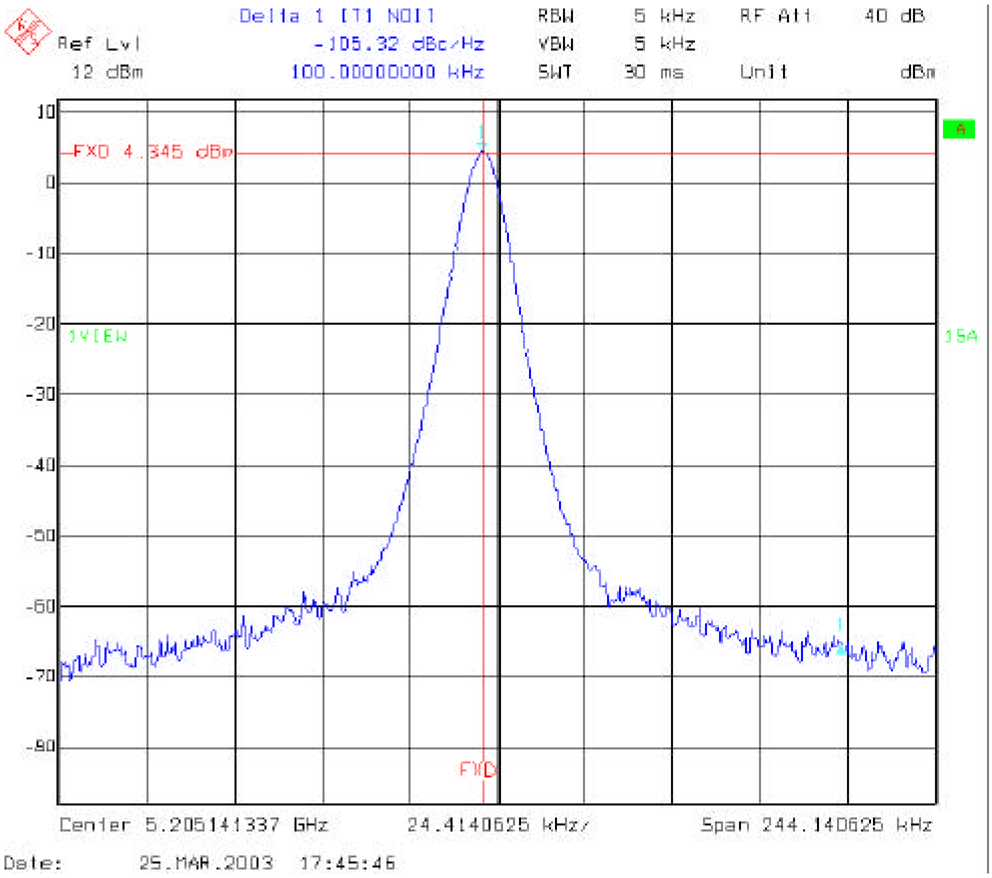

Figure 6: VCO phase noise at Vtune=5V

Figure 7 shows the measured performance of the broadband amplifier including an external bias tee but without external matching. The low frequency gain is 11dB, rolling off gently with frequency to give a 3dB bandwidth of 9GHz. Input and output return losses are both greater than 10dB to 9GHz.

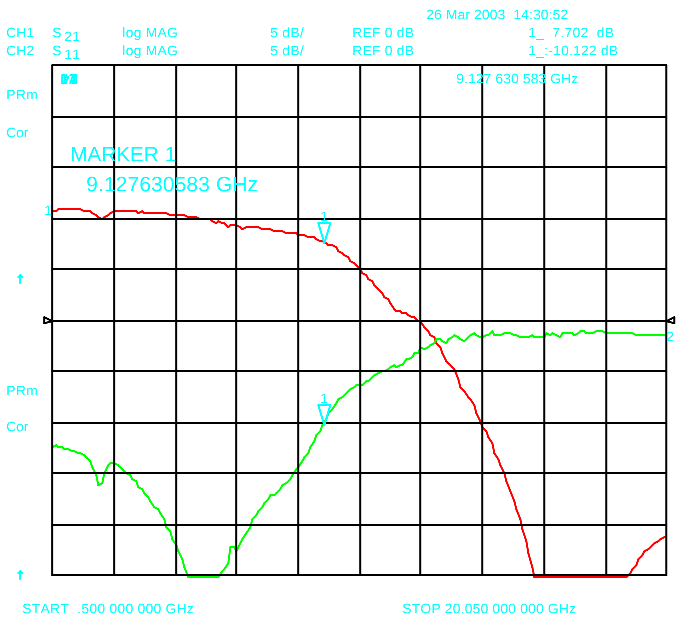

Figure 7: Measured S21/S11 of amplifier without external matching

The measured power compression performance of the amplifier agreed very well with simulated, as indicated in the performance summary of Table 2. This gives confidence in the large signal model data for the transistors. Noise figure is 2.5dB at 2GHz, which is low for this configuration of amplifier.

Parameter	Simulation	Measured
Nominal Gain	~11dB	~ 11dB
3dB bandwidth	DC to 12GHz	DC to 9GHz
DC Power Consumption	37mA @ +7.5V (Using external bias tee)	42mA @ +7.5V (Using external bias tee)
P_{-1dB} output referred	+10.3 dBm @ 2GHz +7.0 dBm @ 10GHz	+9.9dBm @ 2GHz +6.9dBm @ 10GHz
Noise Figure	3.5dB @ 2GHz	2.5dB @ 2GHz
Return Losses	> 10dB, DC to 12GHz	> 10dB, DC to 9GHz

Table 2: Summary of measured and simulated performance of broadband amplifier without external matching

The addition of the external matching circuitry, described above and depicted in Figure 5, allows the operational bandwidth of the amplifier to be extended to 13GHz. It uses just one additional component, beyond those required for the output bias tee, but extends the upper operating frequency by around 4GHz. Figure 8 shows the measured performance of the amplifier with external matching components. The gain is just under 9dB at 4GHz falling to just below 7dB by 12GHz with an output return loss of greater than 13dB. It should be noted that the matching components were conventional 0402 SMT parts, no specialist microwave components were required.

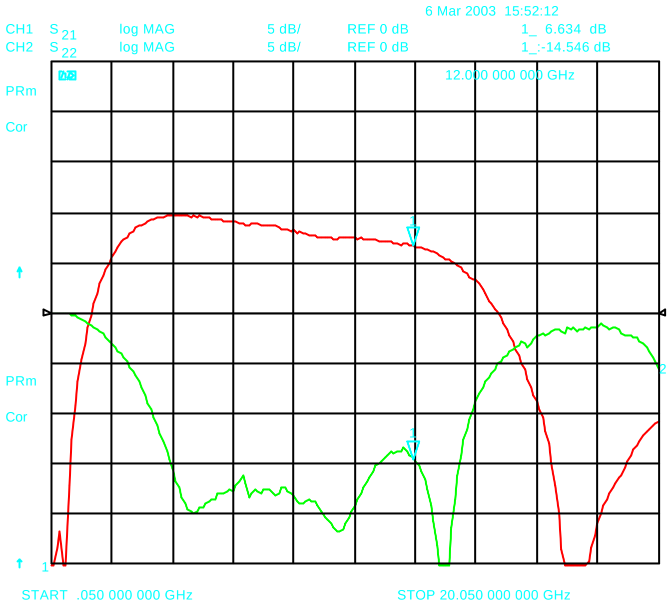

Figure 8: Measured S21/S11 of amplifier with external matching

A comparison of the measured versus simulated power transfer characteristics at 4GHz is shown in Figure 9 . The 1dB gain compressed output power level is 11.5dBm and excellent agreement between measured and modelled performance is demonstrated. A summary of the measured and simulated performance of the broadband amplifier with external matching is given in Table 3.

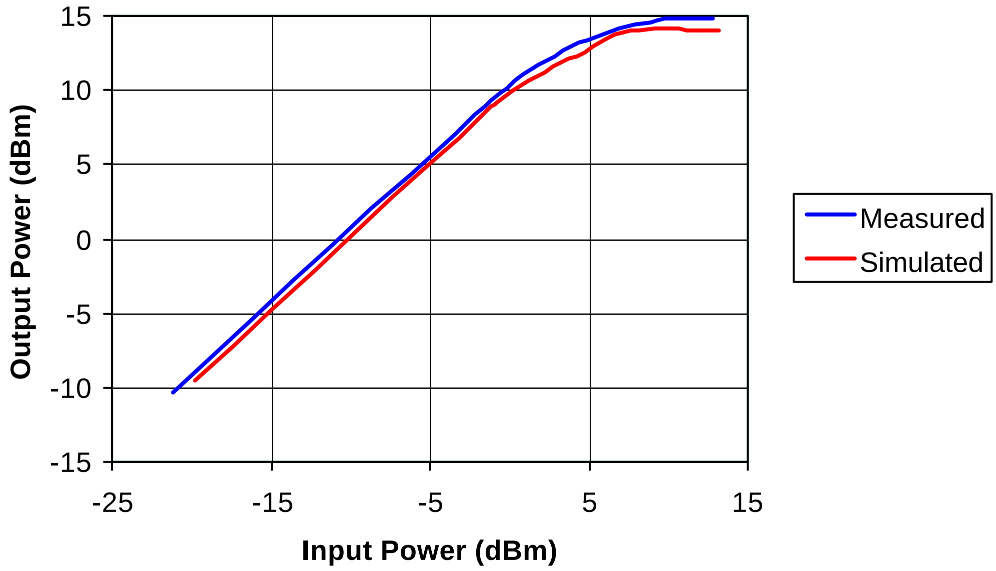

Figure 9: Power versus simulated compression characteristics at 4GHz

Parameter	Simulation	Measured
Nominal Gain	~ 10.5dB	~ 10dB
Operational bandwidth	4.5 to 14GHz	4 to 13GHz
DC Power Consumption	41mA @ +7V	42mA @ +7.5V
P_{-1dB} output referred	+11.9 dBm @ 4GHz +6.6 dBm @ 12GHz	+11.6 dBm @ 4GHz +6.2 dBm @ 12GHz
Output return loss	> 13dB, 4.5 to 14GHz	> 13dB, 4 to 13GHz
Input return loss	> 10dB, 4.5 to 14GHz	> 10dB, 4 to 13GHz

Table 3: Summary of measured and simulated performance of broadband amplifier with external matching

Measurements of the third order intermodulation point (IP3) were also made. Figure 10 shows the output referred P-1dB and IP3 points versus frequency. Both decrease with frequency and a difference of around 12dB is observed between the two.

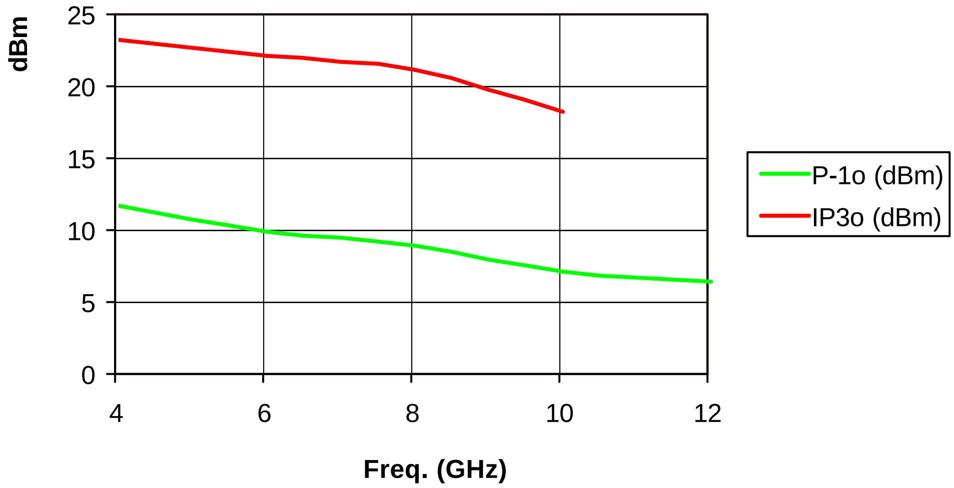

Figure 10: Power Compression & IP3 Performance vs. Frequency for Amplifier with External Match

Conclusions

InGap HBT MMIC technology can allow the realisation of low-cost, high performance components. The technology is well suited to the realisation of DC coupled amplifiers, low phase noise oscillators and high efficiency, low supply voltage PA's. Two example MMICs have been designed, manufactured and evaluated. Excellent agreement between simulated and measured large signal performance is demonstrated. All design work was carried by Plextek (a UK based design consultancy) and MMIC fabrication was performed by GCS (a US based pure-play foundry).