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SURVEILLANCE RESEARCH LABORATORY

SALISBURY, SOUTH AUSTRALIA

TECHNICAL MEMORANDUM

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HF MOSFET POWER AMPLIFIERS

by

K. Gooley, R. Debnam and P. Hattam

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SUMMARY

Solid state devices are being increasingly used in RF power amplifiers. The work outlined in this paper was performed in order to keep HFRD abreast with the current state of technology, and to develop a potentially useful unit. This report covers the design of RF power amplifiers in a general way, presents the performance test results of two HFRD manufactured amplifier modules (200W and 500W) and compares these with several commercially available units.

DSTO AUSTRALIA

POSTAL ADDRESS: Director, Surveillance Research,
PO Box 1500, Salisbury, South Australia 5108.

CONTENTS

TABLES

FIGURES

1 INTRODUCTION

Since the early 1980's, solid state technology has been used increasingly in high frequency (HF) power amplifiers, as solid state devices capable of handling the necessary power levels have been developed. Solid state devices offer the advantages of smaller size, increased robustness and very short warm-up time. High power bipolar junction transistors (BJT's) were the first to be developed, but with improved technology, field effect transistors (FET's) of comparable power levels have become available.

Little attention has been paid to solid state power amplifiers within HFRD in recent years, and in that time transistor technology has made some vast improvements. FET's each capable of channelling many hundreds of watts have since been developed.

In an effort to bring HFRD up to date with the current state of the field, it was decided to develop a solid state power amplifier.

One of the initial decisions was to use power MOSFET's rather than BJT's. There are a number of reasons for preferring MOSFET's at a given power:

- generally they have a higher power gain, which means a lower power input driver is required.
- they produce lower high order IMD (intermodulation distortion) products than BJT's, unless the bipolar is in a class A configuration, in which case the efficiency of the circuit is less and hence output power is lower.
- FET's are more tolerant to load mismatches.
- they are easier to make broadband due to higher input impedances.
- FET circuits are easier to design and bias since the input and output impedances of the device do not change with frequency or drive level to the extent BJT's do.
- the gain of the circuit can be controlled by varying the bias voltage, making automatic level control simpler.
- BJT's are susceptible to thermal runaway, whereas FET's are not.

This report discusses the design criteria for high power RF amplifiers in a general way, reports specifically on the demonstrator designs and includes a comparison with commercially available units. The report concludes by outlining further work needed.

Apart from keeping the division abreast of the latest technology, the power amplifier development has the potential to be incorporated within future equipment, such as the HFRD field station currently under development; and the remote beacon, still to be developed.

2 AMPLIFIER DESIGN

Power amplifiers have a set of problems which are peculiar to them. The main ones are distortion of the signal, the efficiency of the circuits and heat dissipation. In addition, there are five classes of amplifiers- A, B, AB, C and D - which make the selection of the appropriate amplifier for an application more involved.

2.1 Classes

Each class of amplifier has different output characteristics, efficiencies and uses. The classes of amplifier are defined according to the conduction angle of the active device(s) and generally speaking the smaller the conduction angle the higher the efficiency. The output waveform for a sinusoidal input, for all classes except D, appears in Figure 1.

The class A amplifier produces an amplified image of the input, together with any distortion caused by the transistor, at the output. Its maximum efficiency, at 50%, is the lowest for the five types mentioned here. The low efficiency is the reason this class of amplifier is usually not used in power amplifiers.

The class B amplifier produces an amplified image of the input for half a cycle, and no output for the other half. It's maximum efficiency is 78.5%. The major advantage this class of amplifier has over class A is that for no input signal the power consumption is zero.

The class AB amplifier conducts for more than half a cycle, but less than a full cycle. This gives the class a maximum efficiency of between 50% and 78.5%. Class AB amplifiers are the most popular choice for broadband amplifiers.

The class C amplifier conducts for less than half a cycle and has a theoretical maximum efficiency of 100%. The efficiency is dependent upon the length of time the amplifier conducts each cycle, known as the conduction angle. Efficiency increases as the conduction angle decreases, but power output is also decreased. Maximum efficiency is achieved at a conduction angle equal to zero, where efficiency is 100%, but this means that there is no output power. This class of amplifier is often used in narrowband resonant or tuned circuit amplifiers.

Class D, which is a hard switching method of operation, is of no relevance in discussions about HF power amplifiers, since this class is utilised almost exclusively in audio power amplifiers.

2.2 Efficiency

In conventional small signal amplifiers efficiency is not of great consideration since the total power consumption of the circuit is relatively low, and higher efficiency circuits usually introduce other problems. When dealing with power amplifiers though, efficiency can be an important consideration, depending on the application. For example, if linearity and robust operation in a highly coupled array environment is the key requirement, then an efficiency - linearity tradeoff is inevitably involved.

The efficiency of an amplifier can be improved if an isolating transformer is used between the load and the output. This is because the dc power loss in the load is eliminated. If the loss resistance of the transformer is negligible the ac signal will be symmetric about the dc rail voltage.

As an example, a class A amplifier which is RC coupled has a maximum efficiency of 25%, given by;

$$\eta = 25[(Vce_{max} - Vce_{min})/Vcc]\ \%,$$

whereas, a transformer coupled class A amplifier has efficiency given by;

$$\eta = 50[(Vce_{max} - Vce_{min})/(Vce_{max} + Vce_{min})]\ \%,$$

$$\text{and when } Vce_{min} \ll Vce_{max},\ \eta_{max} \Rightarrow 50\%.$$

As previously mentioned the class A amplifier is the least efficient of the five classes. It also has a much lower maximum output power, as is illustrated in Figure 2a.

The device used for the 200 Watt HFRD demonstrator, a Motorola power MOSFET type MRF151G, has a maximum power rating of 500 W. *If this is used in a class A* configuration the maximum output power possible would be 250W. If the same device were used with two operating push-pull in class B the maximum output power would be 1250 W. These levels are achievable when the configurations are operating at their maximum efficiencies, which, while theoretically possible, is impossible in practice.

As a class AB amplifier has characteristics at some point in between classes A and B, the curves of Figure 2b would need to be modified to accommodate this type of amplifier. The efficiency curve would move from the linear response of the class B down to the curved response of the class A as the quiescent current increased. Similarly, the output power would reduce, and the power dissipation curve would change in a smooth progression from class B to A.

2.3 Heat

Dissipation of heat is a major problem in solid state power amplifiers. The MRF151G has a maximum junction temperature, T_j, of 200°C, a maximum power dissipation, P_d, of 500 W and a junction to case thermal resistance, Θ_{jc}, of 0.35 °C/W. If a device is operated above its rated temperature permanent damage can occur. Using a thermal-electrical analogy, the thermal conduction circuit of an amplifier circuit would be as in Figure 3.

The thermal capacitances of the various stages are an indication of how much thermal energy can be stored. The thermal capacitance of the junction is very small, the copper heat spreader is larger, and the heat sink is larger still. The product of the thermal capacitance and thermal resistance will give the thermal time constant.

ie. $C_j\theta_{jc} = \tau_j$

For the device used τ_j would be of the order of milliseconds, τ_c seconds and τ_h possibly minutes. This indicates that the load line of the device can pass through the area above the maximum dissipation curve for periods very much less than the smallest thermal time constant. This can happen in class B and class AB operation, resulting in the higher possible output powers. However the conservative operation of the HFRD demonstrator amplifier did not allow this at any time.

2.4 Distortion

There are two main reasons that the output signal is or may be a distorted version of the input. Firstly, the amplifying device may be nonlinear, or alternatively, the circuit may be biased such that the active device is driven into cutoff for part of the cycle and consequently the output signal is clipped, as is the case in amplifiers of classes AB, B, C and D. The first cause of distortion cannot be eliminated entirely as it is device dependent, and is present in all classes of amplifiers. Non-linearities in MOSFET devices causing distortion include changes in forward transconductance with drain current and a voltage dependant capacitance between drain and gate. For a single sinusoidal input signal, the distorted output may be considered to be a signal at the same frequency as the input, plus signals at harmonics of the input frequency. These harmonics can be removed by filtering provided the required amplifier is sufficiently narrowband. In broadband amplifiers harmonics may fall within the passband and hence be present at the output.

A multi-tone sinusoidal input signal will produce not only harmonics, but also product frequencies;

eg. if the frequencies of the two input tones are f_1 and f_2,
product frequencies will be produced at $2f_1 \pm f_2$ etc.

These intermodulation products are likely to fall within the passband of both narrowband and broadband amplifiers.

To reduce second order distortion, amplifiers may be operated in "push-pull" mode. This is a two transistor circuit where each transistor is conducting on opposite half cycles of the input signal, refer to Figure 4.

Push-pull operation has several advantages over single device operation, these are;

- the dc components of the collector/drain currents are opposite through the transformer, which leads to a low net flux due to bias current, resulting in smaller transformers.

- class AB or class B operation with lower distortion.
- higher maximum efficiencies.

There are also disadvantages, namely the distortion is not as low as class A operation, on account of signal levels approaching the clipping point, and class B push-pull amplifiers suffer from "crossover" distortion, refer to Figure 5.

As previously mentioned, the most popular choice for wideband RF power amplifiers is class AB operation. This class is able to produce broadband amplifiers of good efficiency (maximum between 50% and 78.5% with proper design load) and low distortion (in push-pull). These efficiency figures do not take into account power supply efficiency nor the energy used in cooling the amplifier.

3 THE HFRD DESIGN

Two power amplifier modules were manufactured by HFRD, one rated at 200 watts and the other 500 watts. The following discussion refers mainly to the 200 watt design but it is largely relevant to the 500 watt design also. The amplifier modules are illustrated in Figures 10 and 11 respectively.

The 200 watt power amplifier developed by HFRD was based upon an amplifier discussed in Reference 1. The amplifier is a Class AB push-pull design using a dual MOSFET. That is, the two active devices are contained in the same package. This reduces unwanted negative feedback due to non-common source lead inductance and resistance and since the MOSFET dice are well matched by the manufacturer, even order distortion is very low. Negative feedback via resistors between drain and gate stabilises the gain with frequency, reduces distortion and improves the input VSWR of the amplifier. The inductance of the feedback resistors' leads is used to reduce the amount of feedback at the high end of the frequency range. The circuit diagram appears in Figure 6.

The amplifier [1] delivered 300 W in the frequency range 10 - 175 MHz, this differed from our need of a 200 W, 5 - 30 MHz amplifier. Although, superficially, the difference between the two amplifiers may seem minor, the task of extending the frequency range of the amplifier down to a minimum of 5 MHz is by no means simple. The main task in extending the frequency range was redesigning the input and output transformers.

3.1 Transformer

As previously mentioned the use of transformers in amplifiers improves the devices maximum efficiency. Transformers can also be used to match the input and output impedances of the amplifier to the source and load respectively.

The transformers used in this design are transmission line transformers. They differ from conventional transformers in that they do not necessarily provide dc isolation. In addition, they have characteristics which make them far more suitable for broadband matching than alternate solutions.

Unlike conventional transformers, transmission line transformers are not reliant on coupling to produce an output. Across most frequency ranges transmission line effects are responsible for the output. The primary function of the core is to prevent unwanted currents from circulating, except at the low frequency end of the response, where coupling via the core does help to extend the frequency range.

As long as the length of the transmission line is kept below approximately a fifth of the wavelength of the highest frequency required, the ohmic losses of the transformer are minimal (« 1 dB). These small losses mean a close matching of the required transformation ratio across the frequency range. Losses in the core of the transformer also cause problems with conventional transformers. However, transmission line transformers can be wound in such a manner that the currents in the windings are equal and opposite, and cancel the flux generated by the core, thus minimising core losses.

The interwinding capacitances in conventional transformers resonate with the leakage inductance and cause a lowering of the high frequency response. In transmission line transformers these capacitances form part of the characteristic impedance of the line and therefore the bandwidth is not limited.

3.1.1 Core

The choice of core material for a transformer is important, as this can greatly vary the transformers response. This material is usually one of the following:- air, ferrite or powdered iron.

Air cored transformers usually are large and have poor coupling at low frequencies, so were not considered for this design.

The choice is then generally between powdered iron or ferrite cores. Tests [2] have shown that ferrite cores produce better results than powdered iron materials and that toroid shapes with their closed magnetic circuit produce better results than rods. At the frequencies of interest to HFRD ferrites usually have much higher permeabilities, μ, than powdered iron. The higher the μ the less turns required for a given inductance. However, the higher μ ferrites tend to have greater losses at high frequencies. Also to be taken into account is the Curie temperature, T_c, of the material. Above this temperature the permeability of the core falls rapidly, greatly reducing the magnetising inductance of the transformer.

The material first used for the core of the output transformer had a $\mu \approx 850$, and a $T_c \approx 130$ °C. This Curie temperature proved too low and a transistor was destroyed by excessive current as a result of the rapid fall in magnetising inductance. No higher permeability ferrite of adequate Curie temperature could be found so the type chosen was a Philips material type 4C65 with

$\mu \approx 125$ and $T_c \approx 350$ °C. The Curie temperature was more than adequate, but the permeability was low. This meant the output transformer needed to be larger to produce the required impedances. The larger size results in longer transmission line lengths in the transformer windings which can mean higher losses at high frequencies. However, the output transformer design arrived at, has a total primary line length of 100 mm which is 0.01λ at 30 MHz and the transformer loss due to transmission line length is very small.

3.1.2 Transformer Design

The input and output transformers serve two main purposes, namely to transform the active device impedances to the system impedance of 50 Ω and to convert the balanced device impedances to unbalanced. The design of the output transformer will be discussed but the same procedure applies to the input transformer.

The nominal 200 watt amplifier module was designed for a maximum output power of 300 watts leaving a generous margin to give the amplifier more immunity from intermodulation distortion in an array environment.

The active device drain to drain load impedance is therefore calculated for 300 watts output using the following equation:-

$$r_{dd} = \frac{2(V_{DD} - i_{dmax} \times r_{ds})^2}{P_o}$$

where:- r_{dd} = the drain to drain load resistance
V_{DD} = the drain dc supply voltage (50 volts)
$i_{dmax} \times r_{ds}$ = the drain to source voltage at maximum drain current.
The data sheet gives this value as a maximum of 5 volts.
Substituting into the equation gives a value for r_{dd} as 13.5 Ω.

A transformer turns ratio of 1:2 gives an impedance step-up of 1:4. This is close to ideal for presenting a 13.5 Ω load to the active device from a 50 Ω system impedance. The low impedance winding of the transformer is 1 turn in order to use the shortest length of conductor possible. Hence 2 turns are used on the secondary winding. The conductors used are the inner and outer of coaxial cable whose characteristic impedance is the geometric mean between the primary and secondary impedances, i.e. 25 Ω. The outer conductors of the 2 turns are connected in parallel while the inner conductors are connected in series.

3.2 Feedback

Negative feedback is used around the active device to level the power gain versus frequency curve; to improve the amplifier input VSWR; and to reduce distortion due to transistor non-linearities. The data sheet on the MRF151G gives the device open loop power gain as 26.5 dB at 5 MHz and 23.5 dB at 30 MHz. Shunt voltage negative feedback was applied around the active device and the gain was reduced to about 17 dB overall including the transformers. Gain variation was reduced to 2.4 dB. This may not seem to be a large improvement from the data sheet's 3 dB variation but the 2.4 dB includes gain variations in the transformers.

The gate to gate input impedance of the dual MOSFET at 30 MHz is given in a Smith Chart in the data sheet as 4-j5.5 Ω. Extrapolating the curve to 10 MHz involves a high degree of risk of error but we can say that the input impedance at 10 MHz is of the order of 8-j12Ω and even higher at 5 MHz. This large variation in input impedance with frequency would cause serious mismatch problems and therefore unacceptably large variations in input VSWR. Feedback reduced the input VSWR to an acceptable level over the frequency range.

As shown in the circuit diagram of Figure 6 the feedback networks consist of a dc blocking capacitor C_{10} and C_{11}, a resistor of 50 Ω and an inductance which is associated with the leads of the resistor. These are about 20 mm in length and have an inductance of about 20 nH. The effect of this inductance is to reduce the level of feedback at high frequencies reducing the fall off in gain as the frequency is increased. However since the reactance of this inductance is 4 Ω at 30 MHz the effect is not felt until the frequency is substantially higher than the upper limit of interest here.

Power dissipation in the feedback resistors is essentially the difference in input power with and without feedback at the low frequency end of the range. As indicated above the gain at 5 MHz is 26.5 dB without feedback. This translates to an input power of 0.45 watts for 200 watts output. The gain with feedback at 5 MHz was measured as 18.2 dB. Input power is then 3 watts. The difference of 2.55 watts is dissipated in the feedback resistors. This is not a significant amount of power compared with the overall dissipation of the amplifier.

4 PERFORMANCE TESTING

4.1 200 Watt Amplifier Module

In order to directly compare the HFRD built 200 watt power amplifier against commercially available equipment a number of performance tests were run on the device. The tests performed were:- gain versus frequency, that is the necessary input power to achieve 200 watts at the output; input voltage standing wave ratio; efficiency; two tone IMD tests; and single tone harmonic distortion tests. The results of tests on the HFRD built power amplifier are tabulated in Tables 1 to 3 and plotted in Figures 7 and 8.

The graphs of gain and efficiency versus frequency appear in Figures 7 and 8 respectively.

Table 1 **Input Power Levels, Input VSWR, Gain, Power Supply Current and Efficiency vs Frequency for 200 W Output.**

Frequency MHz	Input Power forward Watts (dBm)	Input VSWR	Gain dB	dc Current Ampere	Efficiency %
5.0	3.0 (34.8)	1.39	18.2	11.0	36.4
7.5	3.6 (35.5)	1.48	17.5	11.1	36.0
10.0	3.3 (35.2)	1.35	17.8	11.4	35.1
12.5	3.4 (35.3)	1.32	17.7	11.2	35.7
15.0	4.2 (36.2)	1.45	16.8	11.6	34.5
17.5	4.2 (36.2)	1.34	16.8	11.9	33.6
20.0	4.2 (36.2)	1.34	16.8	12.1	33.1
22.5	4.1 (36.1)	1.48	16.9	12.2	32.8
25.0	4.0 (36.0)	1.31	17.0	12.4	32.3
27.5	5.3 (37.2)	1.61	15.8	12.5	32.0
30.0	4.8 (36.8)	1.45	16.2	12.7	31.5

Table 2 **Harmonics Produced.**

Harmonic	Power Amp 100 Watts dB from fund.	Power Amp 200 Watts dB from fund.
Fundamental 15 MHz	0.0	0.0
second	-44.7	-40.5
third	-14.1	-14.9
fourth	-49.4	-46.1
fifth	-52.6	-33.3
sixth	-63.6	-49.5
seventh	-43.3	-39.7
eighth	-68.5	-55.4
ninth	-44.7	-42.5

Table 3 **3rd Order Two Tone Test Results. dB below the level of each tone.**

Freq 1 MHz	Freq 2 MHz	Spectrum Analyser IMD dB below each tone		Amplifier IMD 200 W PEP Output dB below each tone	
		lower	upper	lower	upper
5.0	6.0	-44.1	-44.1	-16.9	-14.7
5.0	5.1	-43.2	-42.8	-16.2	-16.1
5.0	5.01	-42.3	-42.8	-16.2	-16.2
10.0	11.0	-48.9	-48.2	-14.5	-13.6
10.0	10.1	-47.1	-45.8	-13.8	-13.7
10.0	10.01	-46.8	-47.2	-13.6	-13.6
15.0	16.0	-48.3	-47.5	-14.1	-13.2
15.0	15.1	-46.8	-45.0	-12.9	-12.8
15.0	15.01	-46.9	-47.0	-12.6	-12.7
20.0	21.0	-49.1	-47.6	-12.1	-14.1
20.0	20.1	-49.5	-47.4	-12.5	-12.7
20.0	20.01	-47.6	-47.9	-12.7	-12.7
25.0	26.0	-48.9	-47.4	-12.5	-15.7
25.0	25.1	-47.7	-45.7	-13.8	-14.3
25.0	25.01	-46.8	-47.5	*[1]	*

The measurement results listed in Table 3 were taken using the RTS high dynamic range test facility [14]. The two test signals from Fluke synthesisers were combined with filtering and isolation attenuators, and fed to the driver amplifier. The power amplifier output was fed into a high power attenuator before being applied to the HP8568B spectrum analyser. The spectrum analyser IMD columns of Table 3 represent the limit of measurement for the particular test arrangement and show the level of distortion products produced by the instrumentation. These are seen to be very much less than those products produced by the power amplifier.

4.2 500 Watt Amplifier Module

This power amplifier uses two Motorola power FET's type MRF154 in a class AB push pull configuration. Although designed for 1 KW, it is rated at 500 W for reliability and immunity from intermodulation distortion, due to power being fed into the output port in an array environment. The designed operating range is 2 to 50 MHz, however testing so far has been restricted to 5 to 30 MHz at 500 watts into 50 ohms.

[1] The protection circuitry for the device would not allow the measurements marked with the following *.

4.2.1 Set Up

The circuit is as shown in Figure 9. R8 was selected at 8.2 K ohms to give slight reduction in quiescent bias current with temperature rise as measured by the thermistor R9. To set the idle current on the amplifier, first R1 and R2 are adjusted to reduce the gate bias voltage to zero and then R3 is adjusted to give an output voltage from the regulator N1 of 6V (approx twice the fet gate threshold voltage). R1 is then slowly adjusted until the power supply current is 800mA - 1A. R2 is similarly adjusted until the supply current is exactly doubled.

4.2.2 Thermal Considerations.

The measured efficiency of the prototype at 500 W output is 35%, giving a dc input power of approx. 1400 W. This gives a power dissipation per package of (1400-500)/2 = 450 W. The maximum junction temperature of the MRF154 is 200°C derated by a junction to case thermal resistance of 0.13°C/W giving a derating value of approx. 60°C. This allows a maximum safe case temperature of 140°C. A large centrifugal fan combined with a small fan directed onto the top of the amplifier limits the case temperature to approx. 120°C at 500 W output.

Table 4 gives the measured efficiency of the unit at varying power levels operating at 20 MHz.

Table 4 500 Watt Module Efficiency.

0/P POWER (WATTS)	DC I/P (AMPS)	DC I/P (WATTS)	EFFICIENCY[2] (%)
100	11.6	580	17
150	14.5	725	21
200	17.0	850	23.5
250	19.3	965	26
300	21.3	1065	28
350	22.9	1145	30.5
400	24.7	1235	32.5
450	26.4	1320	34
500	28.0	1400	36

4.2.3 RF Performance

Tables 5, 6 and 7 set out the measured performance characteristics of the amplifier over the frequency range of 5 to 30 MHz.

[2]Efficiency figures quoted do not include cooling or DC power supply efficiency.

Table 5 500 Watt Module Gain and Input VSWR at 500 Watts Output.

FREQ. (MHz)	INPUT LEVEL (dBm) (FWD)	(REV)	INPUT VSWR	GAIN (dB)	DC I/P (AMPS)	EFFICIENCY (%)
5	40.3	24.3	1.39	16.7	26.9	37.2
7.5	40.2	27.2	1.59	16.8	26.6	37.6
10	40.6	28.0	1.60	16.4	27.2	36.8
12.5	40.4	27.9	1.62	16.6	28	35.7
15	40.1	29.4	1.81	16.9	28.1	35.6
17.5	40.2	29.0	1.75	16.6	27.9	35.8
20	40.3	29.1	1.75	16.7	28	35.7
22.5	40.6	29.2	1.74	16.4	27.9	35.8
25	40.2	26.8	1.55	16.8	27.5	36.4
27.5	40.3	26.7	1.54	16.7	28.3	35.3
30	39.8	26.5	1.55	17.2	27.2	36.8

Table 6 500 Watt Amplifier Intermodulation Distortion.

Freq 1 MHz	Freq 2 MHz	Amplifier IMD at 400 W PEP Output dB below each tone		Amplifier IMD at 1000 W PEP Output dB below each tone	
		lower	upper	lower	upper
5.0	6.0	-15.4	-14.1	-16.2	-14.4
5.0	5.1	-15.2	-14.3	-13.6	-13.2
5.0	5.01	-18.1	-16.8	-12.3	-11.1
10.0	11.0	-14.3	-13.5	-13.7	-13.1
10.0	10.1	-14.4	-13.6	-13.3	-12.6
10.0	10.01	-16.6	-15.7	-12.2	-11.3
15.0	16.0	-13.3	-13.8	-13.0	-13.6
15.0	15.1	-13.5	-13.1	-13.7	-12.8
15.0	15.01	-16.5	-15.3	-11.4	-10.5
20.0	21.0	-13.8	-17.0	-14.4	-14.3
20.0	20.1	-15.2	-14.5	-14.5	-13.0
20.0	20.01	-19.4	-16.2	-10.3	-9.8
25.0	26.0	-18.0	-16.8	-14.1	-13.7
25.0	25.1	-17.6	-16.2	-13.9	-13.3
25.0	25.01	-19.1	-17.3	-11.2	-10.1

Table 7 **500 Watt Amplifier Harmonic Products at 15 MHz.**

Harmonic	Power Amp 200 Watts dB from fund.	Power Amp 500 Watts dB from fund.
Fundamental 15 MHz	0.0	0.0
second	-35.1	-28.1
third	-18.3	-17.2
fourth	-42.2	-47.2
fifth	-44.5	-47.5
sixth	-59.9	-62.3

4.3 Comparison

The three commercial 200 W power amplifiers that were short listed for the recent purchase of an amplifier were the ENI 3200L, the MPE PAS-53-0 and the Kalmus LA200H. All were capable of meeting the requirement of 200 watts power output over a frequency range 5-30 MHz and a listing of comparable specifications appears in Table 8. The device purchased for the prototype transportable field station was the Kalmus.

Table 8 **Amplifier Comparisons**

Model	Frequency Range (MHz)	Gain (dB)	Operating Temperature (°C)	3rd Order Output Intercept (dBm)	2nd Order Output Intercept (dBm)
ENI	0.25 - 120	55 ± 1.5	0 - 40	+62	not spec.
MPE	5 - 50	53	0 - 50	not spec.	+73
Kalmus	0.3 - 100	50 ± 2.5	-10 - 40	+57	+92
HFRD(200 W)	5 - 30	63 ± 1.5	45 max	+55	+93
HFRD(500 W) [3]	5 - 30	67 ± 0.5	45 max	+58	+88

Model	2nd Harmonic	3rd Harmonic	i/p VSWR	o/p VSWR
ENI	not spec.	-18	1.5	2.5
MPE	-20	-12	1.5	∞
Kalmus	-39	-13.5	worst 1.65	∞
HFRD(200 W)	-40	-15	worst 1.61	∞
HFRD(500 W) [3]	-35	-18	worst 1.81	∞

[3]The HFRD 500 watt module figures are given at 200 watts output so as to be comparable with the other amplifiers.

4.4 Gain

The gain of the commercial equipment includes the gain of the driver stage as well as that of the power amplifier. For the HFRD models, the gains of the power amplifiers were 17.2 $\pm$ 1 dB and 16.8 $\pm$ 0.4 dB respectively.

4.5 Intermodulation Distortion Intercepts

The third order output intercept was not quoted in the specifications for two of the models, however the third order intermodulation distortion of the Kalmus amplifier was measured in the laboratory. The levels of the two tone third order products at the output of the HFRD amplifiers are about 12dB below each tone at 200 W peak envelope power (PEP) output. This is a significantly higher level of distortion than expected. The reason for this is not clear and further investigation is required to determine the cause.

The second order intercept points for the 4 amplifiers were inferred from the second harmonic distortion figures. As can be seen from the table, the results for both HFRD amplifiers are comparable with those of the purchased Kalmus unit.

4.6 Harmonics

The push-pull configuration of the HFRD 200 watt amplifier with a closely matched dual active device and well balanced input and output transformers provided very good second order distortion, comparable with that of the purchased amplifier The third harmonic is superior to that of the preferred choice and the MPE, but is marginally worse than the ENI. Further development required to improve the IMD could result in better odd order harmonic distortion figures.

4.7 Voltage Standing Wave Ratio

The input VSWR of a wideband power amplifier is important as it determines how closely the amplifier input matches the 50 Ω impedance required by the driver amplifier. The input VSWR's of the HFRD amplifiers were measured and recorded in Tables 1 and 5 respectively. The output VSWR quoted for the commercial devices is the maximum value of load VSWR that they can tolerate and continue to function. This is due to protection circuitry reducing the amplifier gain and therefore power output when the load VSWR is poor. The HFRD amplifier test arrangement included protection circuitry to remove the input power if the output load VSWR was worse than 2.6.

5 CONCLUSION

The power amplifiers developed thus far by HFRD are only single modules in a complex stand alone system. Still to be developed are protection circuitry for the input, output and dc supply ports, as well as additional temperature protection. Dedicated driver amplifiers are needed in addition to some form of visual feedback to the user, to indicate the operational status of the device. Further work is needed on the transformers to minimise their size and maximise their performance.

Overall the HFRD power amplifiers are good prototype modules which perform better than commercial devices in some areas. They can also be more easily adapted to special requirements than commercial amplifiers of fixed gain and output. Further development is needed to provide the protection circuitry mentioned above for implementing a power amplifier that is robust enough to use in the field; and to reduce the third order intermodulation distortion.

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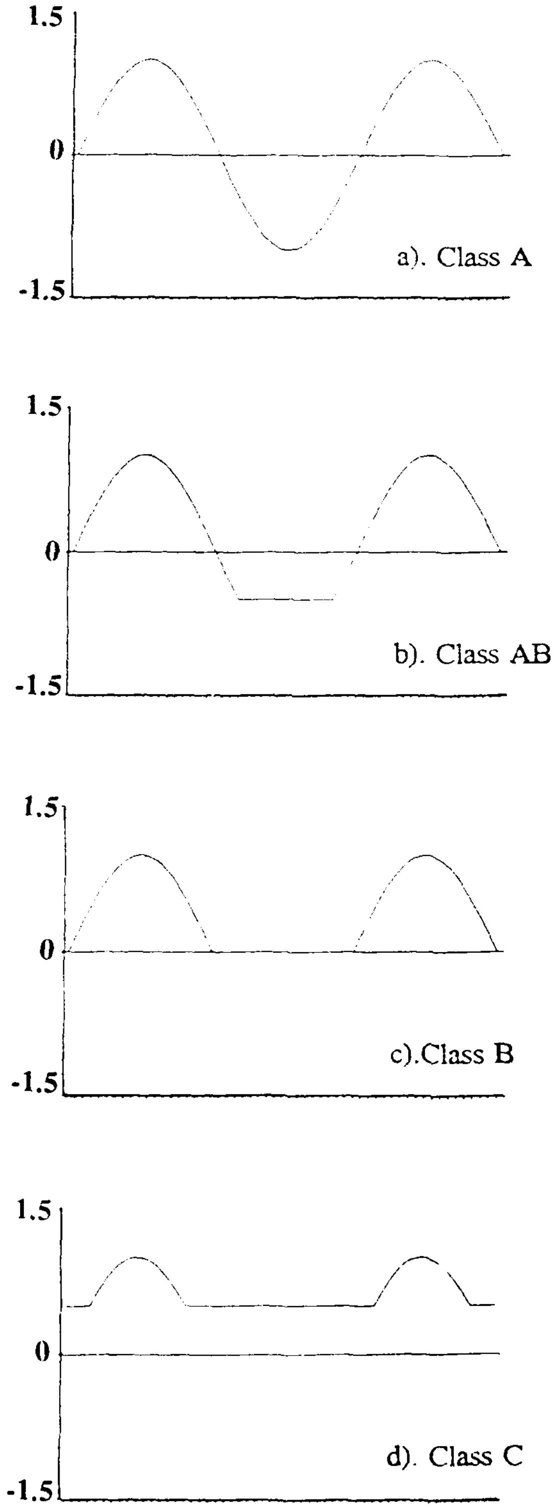
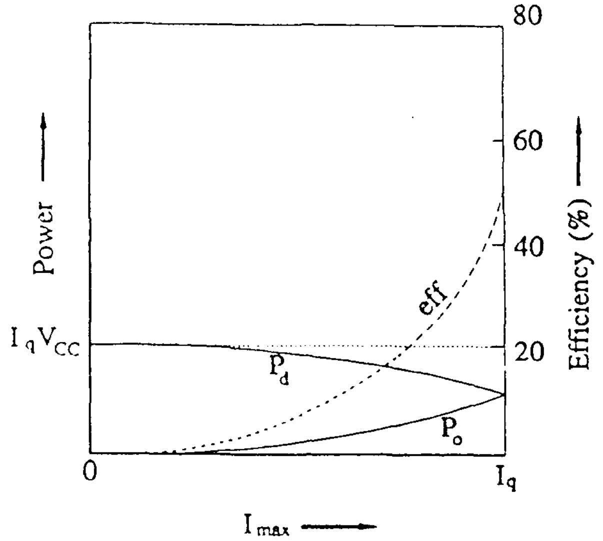
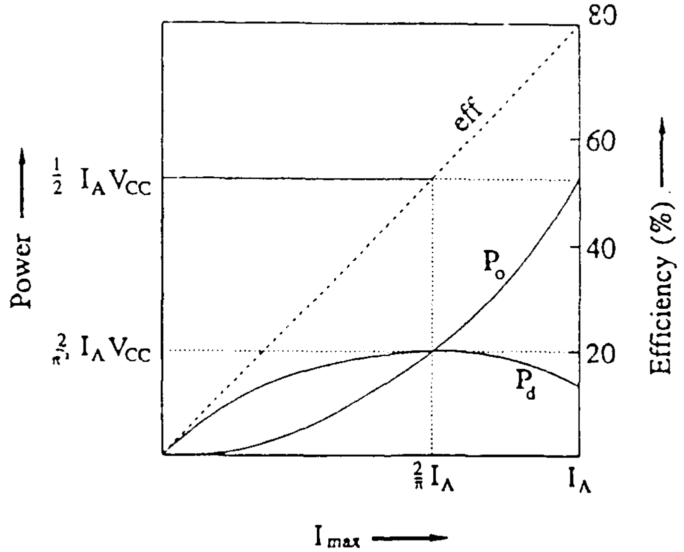


Figure 1 Amplifier Classes

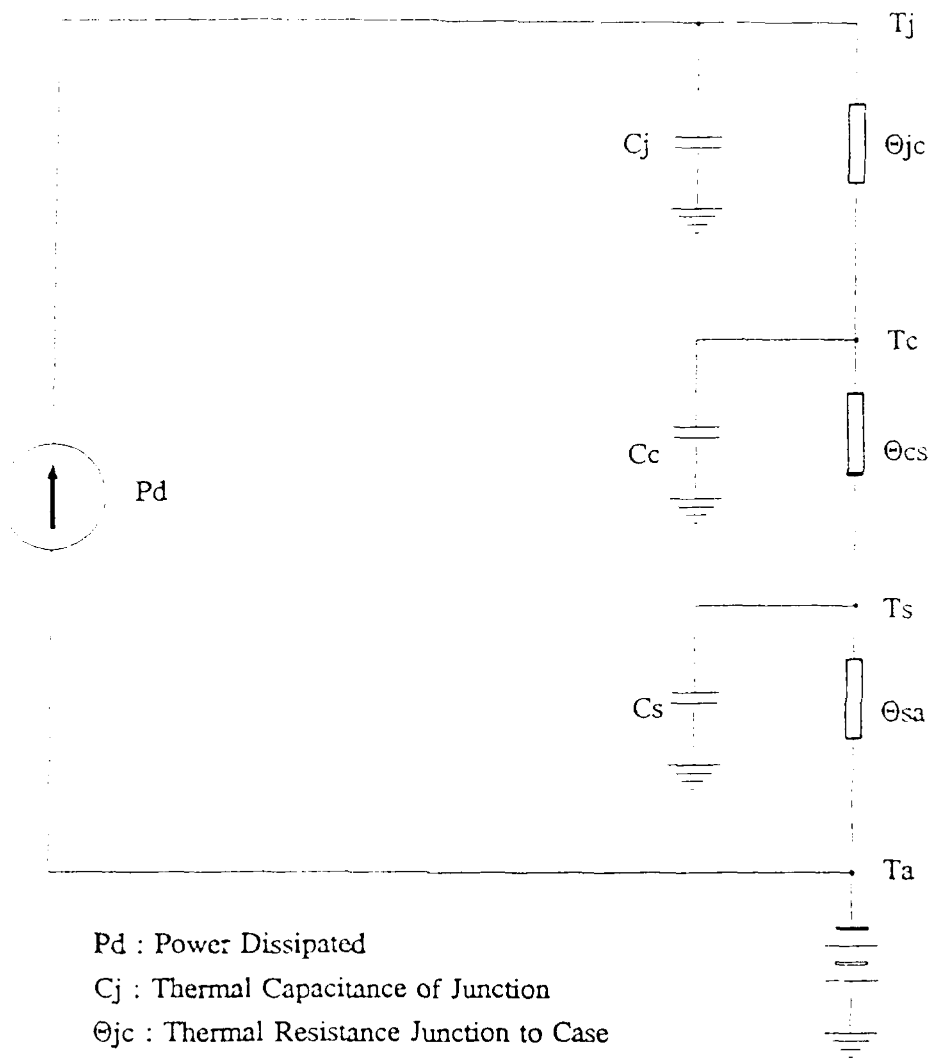


a). Class A



b). Class B

Figure 2 **Power Output, Power Dissipation and Efficiency of Class A and Class B Amplifiers**



Pd : Power Dissipated
Cj : Thermal Capacitance of Junction
Θjc : Thermal Resistance Junction to Case
Tj : Temperature of Junction
Cc : Thermal Capacitance of Case
Θcs : Thermal Resistance Case to Heat Sink
Tc : Temperature of Case
Cs : Thermal Capacitance of Heat Sink
Θsa : Thermal Resistance Heat Sink to Ambient
Ts : Temperature of Heat Sink
Ta : Ambient Temperature

Figure 3 Thermal Electrical Analogy

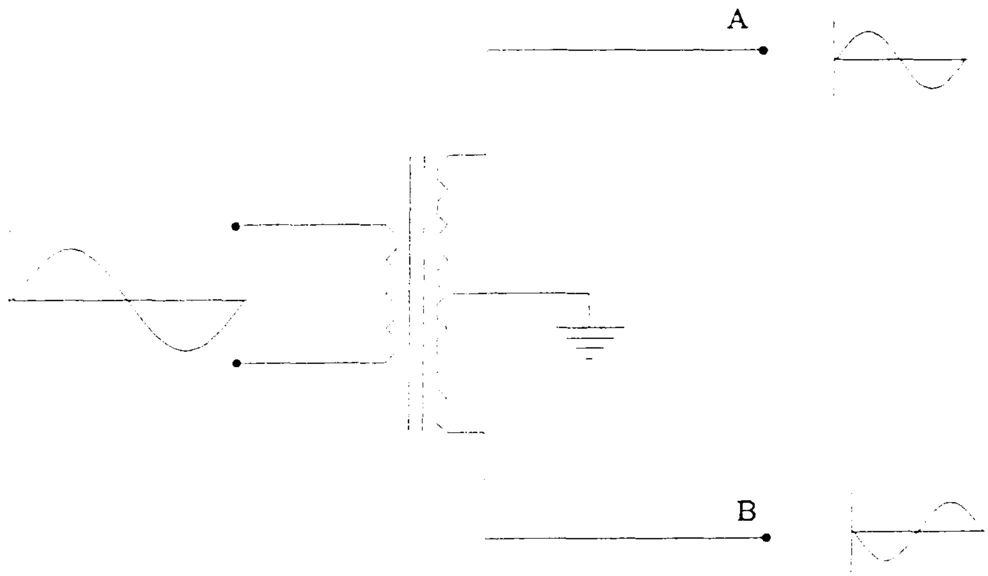


Figure 4 **Push-Pull Operation**

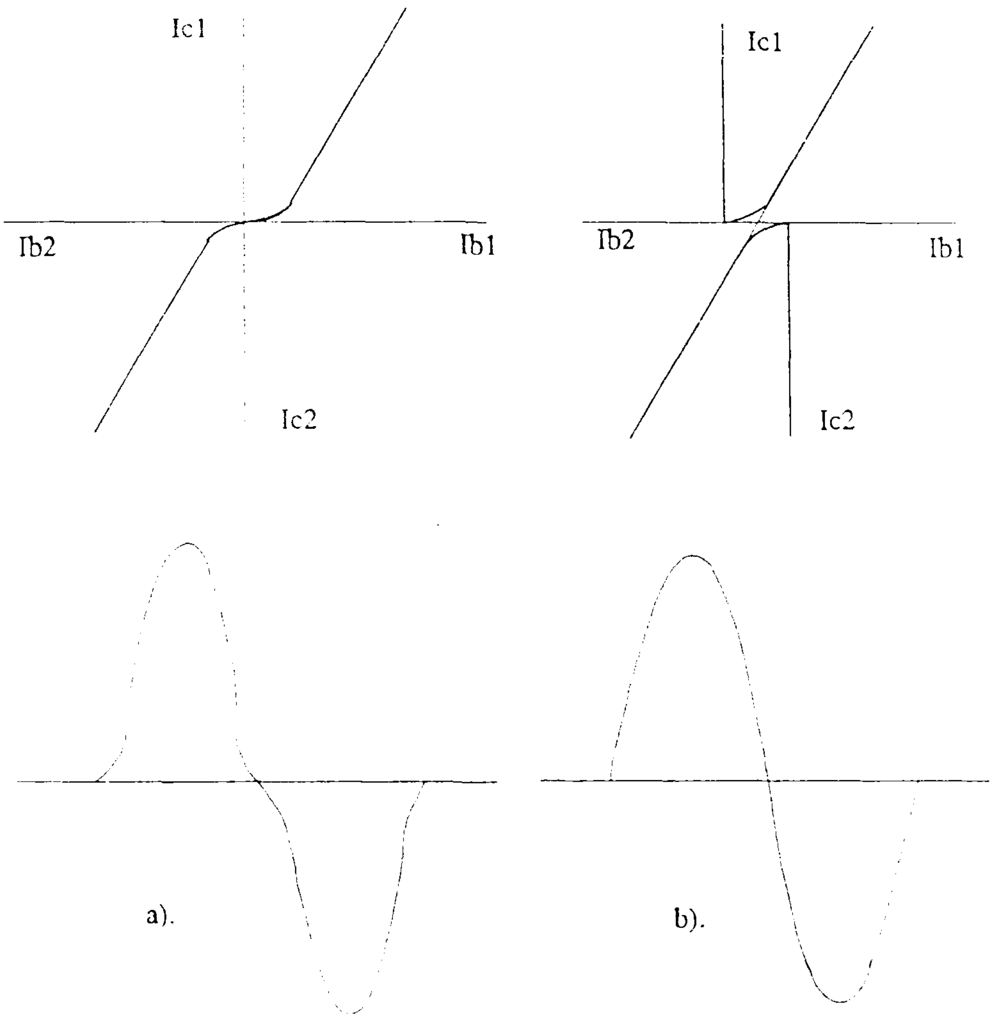


Figure 5 **(a) Class B Crossover Distortion. (b) Class AB Without Distortion**

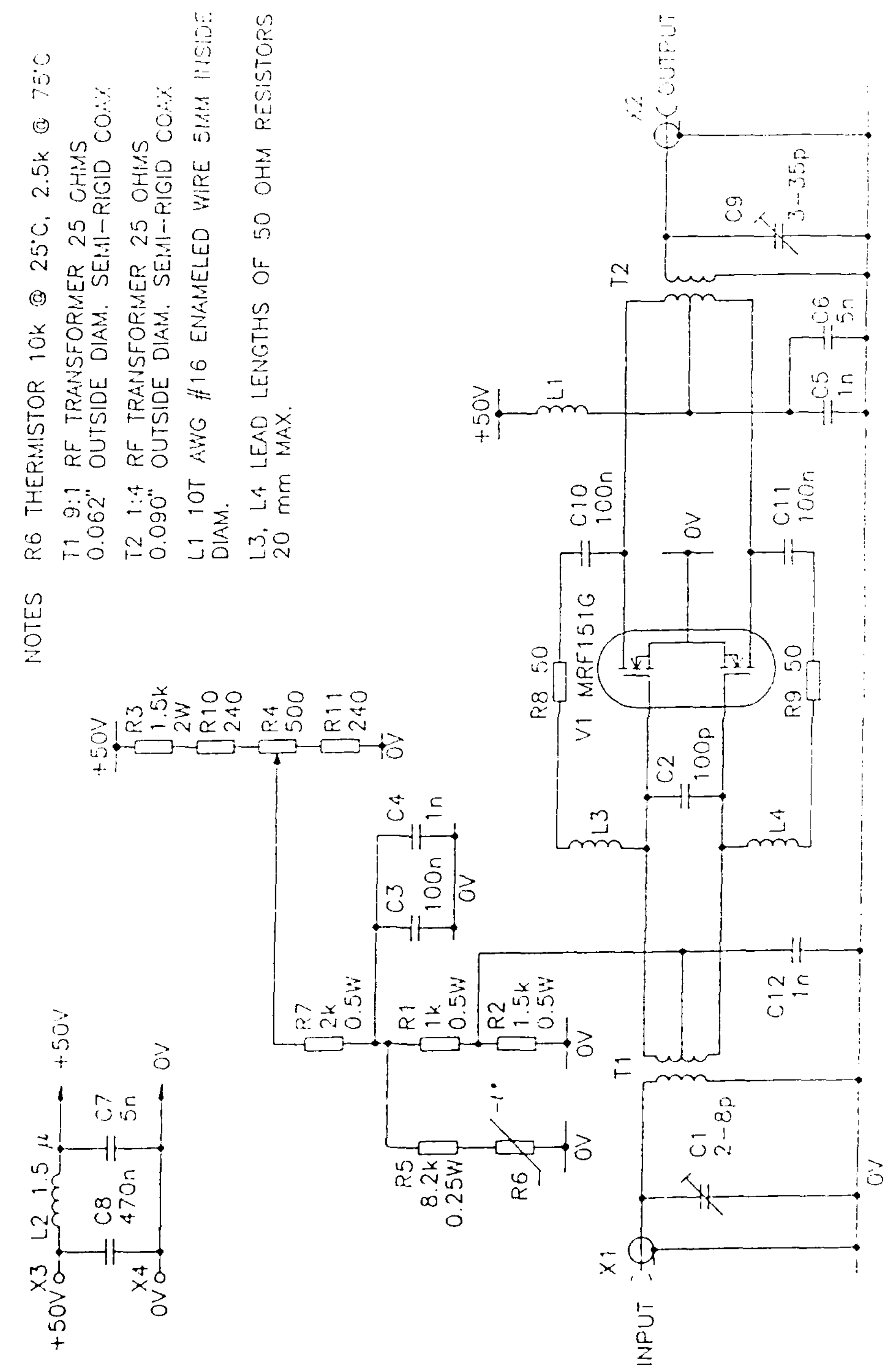


Figure 6 **200 Watt Amplifier**

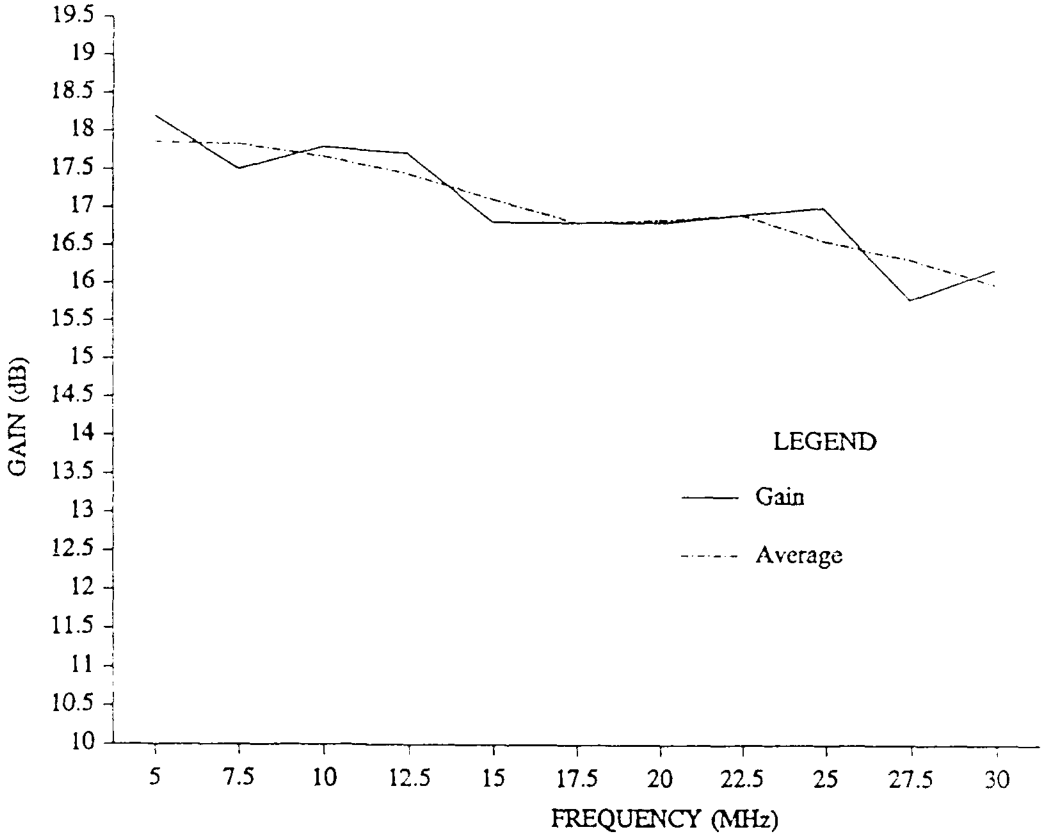


Figure 7 **Gain vs Frequency**

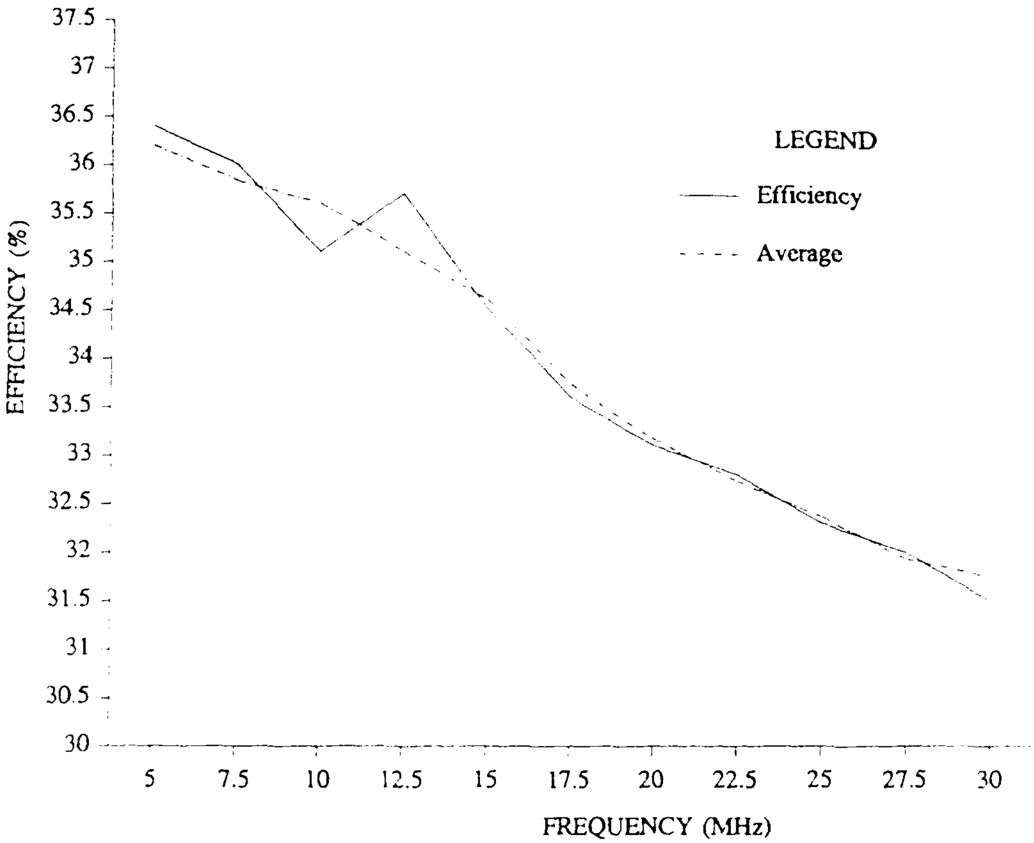


Figure 8 **Efficiency vs Frequency**

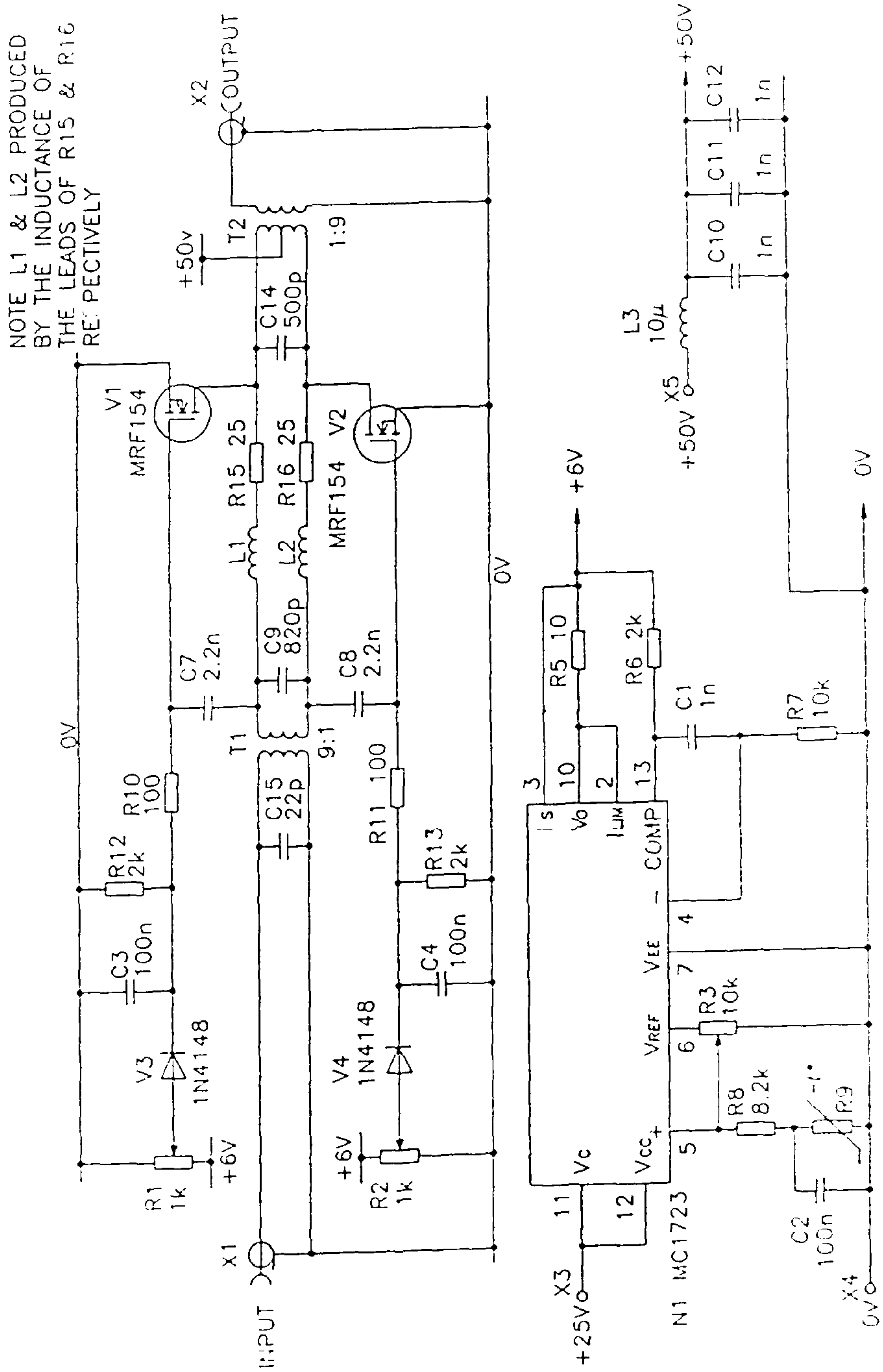


Figure 9 **500 Watt MOSFET Amplifier (5 - 30 MHz)**

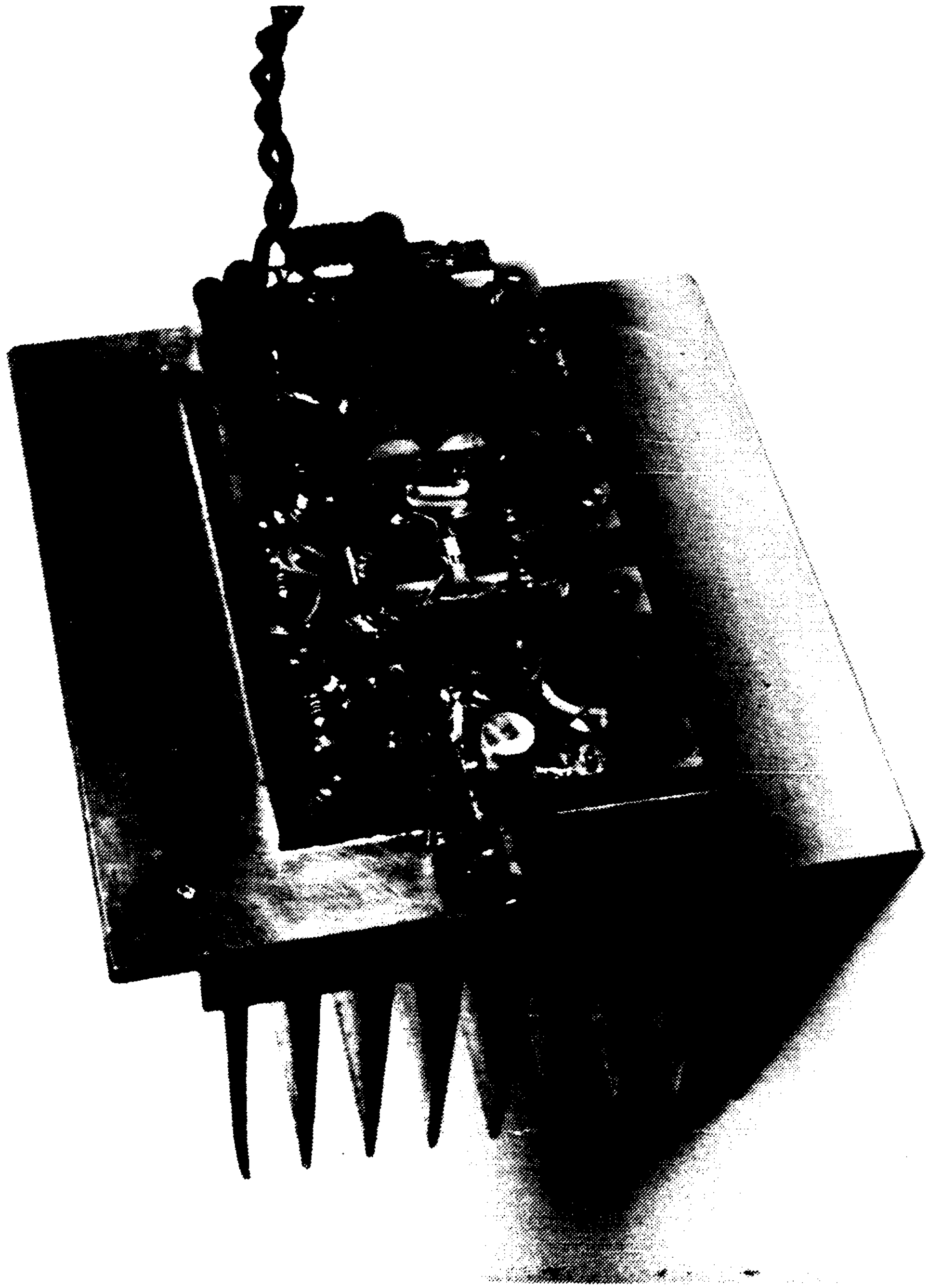


Figure 10 200 Watt Amplifier Module

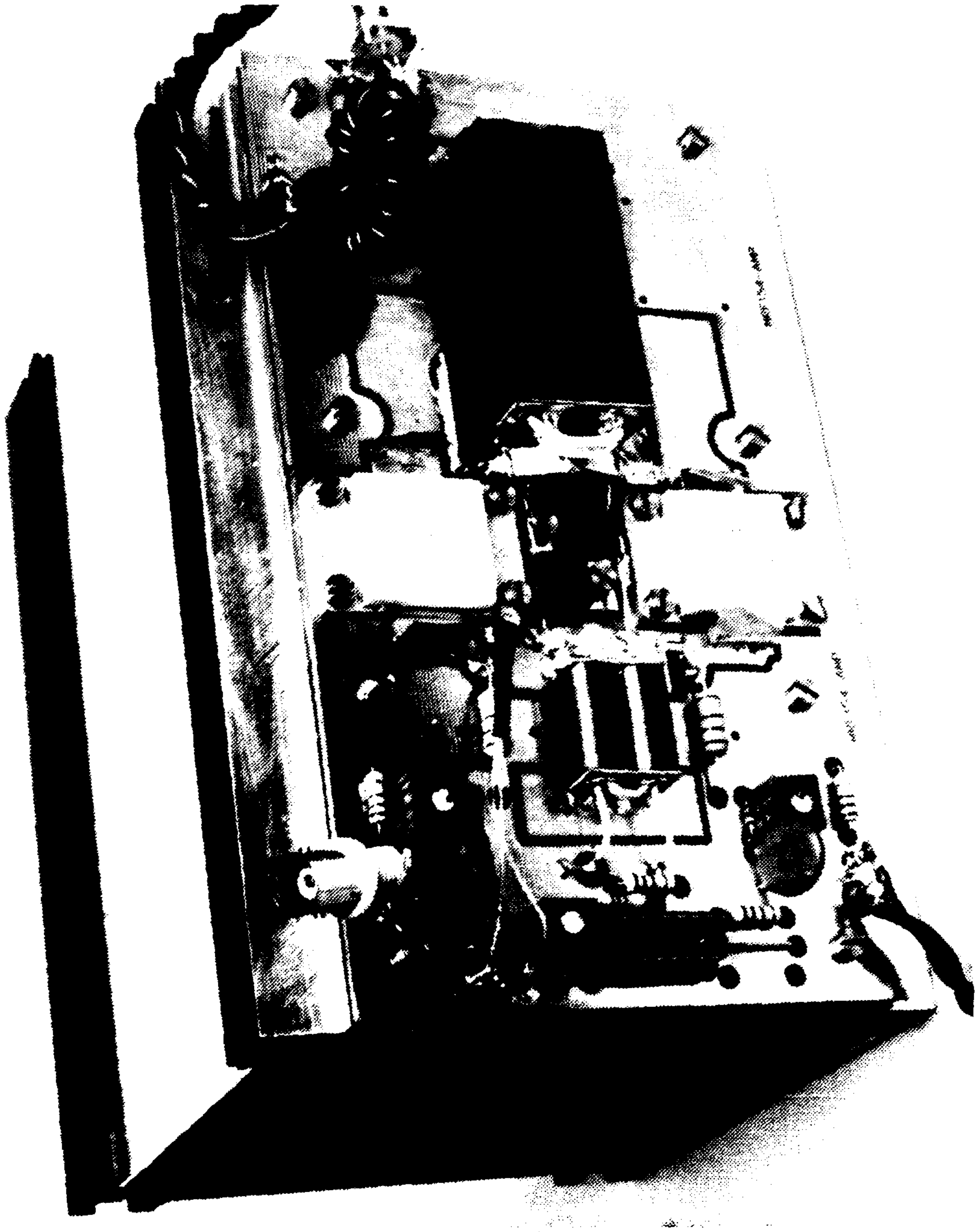


Figure 11 500 Watt Amplifier Module

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16. Abstract

Solid state devices are being increasingly used in RF power amplifiers. The work outlined in this paper was performed in order to keep HFRD abreast with the current state of technology, and to develop a potentially useful unit. This report covers the design of RF amplifiers in a general way, presents the performance test results of two HFRD manufactured amplifier modules (200 W and 500 W) and compares these with several commercially available units.

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