AN RF POWER LINEAR USING IRF MOSFETS

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Here is a 50 watt PEP amplifier, using low priced transistors, which can be used on the lower frequency amateur bands

Introduction

I was interested to read the article by Drew Diamond VK3XU (Notice Notes AR Oct 1988) on the application of the IRF series MOSFET transistor for broadband linear RF amplification. Drew's amplifier used a pair of Motorola IRF510 transistors to generate 5 or 6 watts of peak envelope power from a 13.5V supply and operated within the frequency spectrum of 1.8 to 10 MHz.

I thought it might be of further interest to describe a higher power version of a linear amplifier which I had reason to design, some time age, using two IRF430 transistors. The amplifier was required to deliver 50 or 60 watts of peak envelope power in the MF spectrum. Whilst, at the time, amateur band transmission was not in mind, subsequent tests have shown the circuit can deliver a power off 55 watts at 1.8 MHz, 40 watts at 3.5 MHz and 25 watts at 7 MHz. The IRF430 transistors used were the International Rectifier (IR) type called HEXFET because of their hexagonal source cell structure. (Refer to the appendix at end of the article.)

Secondary Breakdown

An attraction in choosing the MOSFET transistor in preference to the bipolar transistor is the absence of second breakdown. An explanation of this, extracted from the IR handbook, is given in the following paragraph:

One of the outstanding features of IR's power MOSFET is that they do not display the second breakdown phenomenon which is frequently the Achilles heel of bipolar transistors. A simple physical explanation accounts for this superiority. If localized potentially destructive heating occurs within a MOSFET transistor, the carrier mobility in that area decreases. As a result the MOSFET has a positive temperature coefficient and acts in a selfprotective

manner by forcing currents to be uniformly distributed through the silicon die. In contrast and particularly under conditions of high collector-emitter voltage, a bipolar transistor displays "current crowding" in the base region. This causes hot spots. Because of the bipolar's negative temperature coefficient, these hot spots tend to further "hog" the current and cause instantaneous catastrophic destruction of the die.

On the face of it all, the problems of thermal instability, which normally have to be considered in the bi-polar power transistor, could be solved by the use of the MOSFET power transistor. In setting up the amplifier under discussion , this was found to be not quite true. The push pull amplifier was biased to operate in a nominal class B mode with a standing drain current of 150 mA per transistor and a current swing up to 1 amp per transistor at full power. The power supply used was nominally 60 volts. Figure 1 shows the limiting conditions of drain current and source to drain volts for the IRF430 transistor. Pq is the static operating point set and Pm the point of maximum current swing.

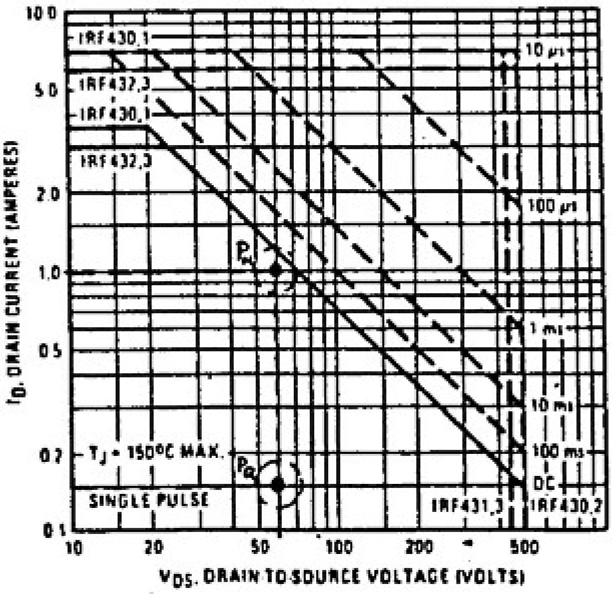


Figure 1 IRF430 - operating condition within maximum safe operating area .

The heat sink was designed so that, at continuous full power, temperature rise at the transistor junctions would be within the limits specified for the transistor. However, under continuous operation at full power, an effect similar to thermal runaway in a bi-polar transistor was experienced.

The explanation of this runaway effect is found in the transfer characteristic curves (refer figure 2). For a junction temperature of 25 degrees C, a standing current of 150 mA is achieved with a forward gate to source bias of 3.5V. At a junction temperature of 125 degrees C and the same bias, standing current rises to nearly 0.5 amp. To maintain the standing current at 125 degrees the same as it is at 25 degrees, bias must be reduced to 3.1V. Clearly, some form of circuit is required which senses the junction temperature and reduces the bias voltage as the temperature rises.

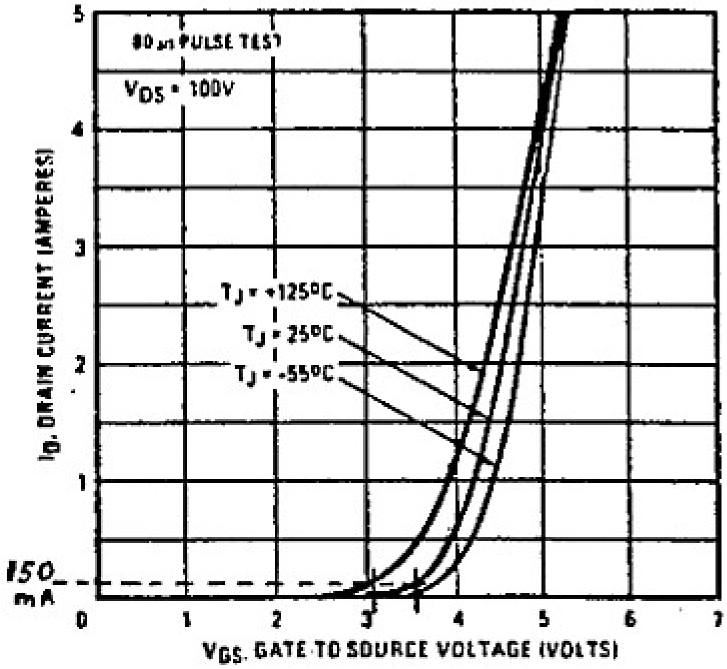
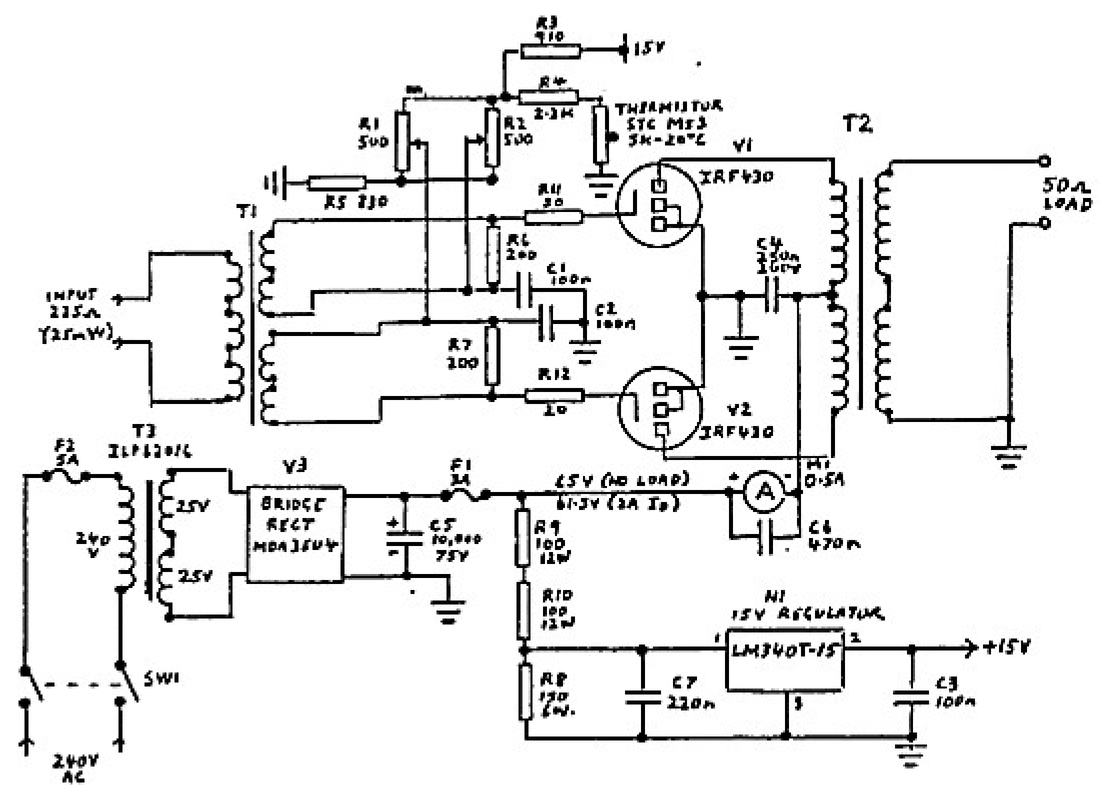


Figure 2 Over a temperature range of 25 degrees C to 125 degrees C, a change in gate to source bias volts of 3.5 to 3.1 volts is needed.

One might be tempted to draw another conclusion that because of the insulated gate, with extremely high input resistance, the MOSFET can be driven from a virtual voltage source supplying negligible power. However, the power MOSFET has very high input capacitance; in the case of the IRF430, around 700 to 900 pF. For a broadband amplifier, this capacitance must he loaded down with shunt resistance so that the impedance presented to the driving amplifier is reasonably constant over the required frequency range. A small amount of drive power is therefore consumed in this resistance load. The other alternative is to have a tuning system at the input which is used to resonate the circuit at whatever specific frequency is in use.

The Circuit



NOTES:
1. V1 & V2 mounted on 8 inch mullard 55D heat sink & insulated with beryllium washers.
2. STC M53 thermistor glued to heat sink.
3. T1 - 18 turns, 7 filar wound 35 SWG on 9 mm toroidal core u = 128.
4. T2 - 18 turns, quadfilar wound 22 SWG on 29 mm toroidal ferrite core u = 800.
5. Set Rl, R2, for static Id of 150 mA in each transistor.

Figure 3 Linear Amplifier Circuit Diagram.

Detail of the circuit is shown in figure 3. Gate to source biasing is developed from a 15V supply through a resistive network. The individual static current in each transistor is set to 150 mA by potentiometers R1 and R2. A thermistor, cemented to the transistor heat sink as close as possible to the transistors, senses the temperature and reduces the bias voltage to correct for the rise in drain current as the temperature rises. (Referring back to Drew's article, it can be seen that he used a diode, fixed to the heat sink of his amplifier, for the same purpose.)

Each transistor gate is loaded down with a 200 ohm resistor selected to mask the input capacitive reactance within the MF range. (These same resistors were left installed for the amateur frequency tests.) At lower frequencies, drive power is essentially that required to develop the required voltage swing across these resistors. At amateur band frequencies, input impedance is essentially the input capacitive reactance of the MOSFET transistors and lower values of load resistance could have been considered. R11 and R12 are parasitic suppressors which should be mounted directly on the transistor gate pins. These were included as a precaution rather than because of any problem experienced.

Based on continuous tone modulation and 60 watts PEP output in the MF region, power dissipation is around 30 watts per transistor. (This is a nominal value as both maximum power output and efficiency fall as the frequency rises. We discuss this further in a later paragraph). Thermal resistance of the transistor junction to its case is 1.67 degree C per watt. Allowing a further thermal resistance of 0.2 degree C per watt across the transistor insulating washer, 30 watts dissipation gives a temperature differential between the junction and the heat sink of 56 degrees. Defining maximum ambient emperature as 40 degrees C and given the maximum allowable junction temperature which is 150 degrees C, the maximum allowable temperature rise in the heat sink is (150 - 56 - 40) = 54 degrees. For two transistors dissipating 30 watts, the heat sink must therefore have a thermal resistance of 0.9 degrees C per watt.

If tone tests were limited to short transmission periods and if the amplifier were only intended for SSB speech, average dissipation per transistor could be considered to be around 10 to 15 watts. Calculating on this dissipation, the thermal resistance of the heat sink could be around 2.5 to 4.5 degrees per watt. A 150 mm length of Mullard 35D heat sink would do nicely.

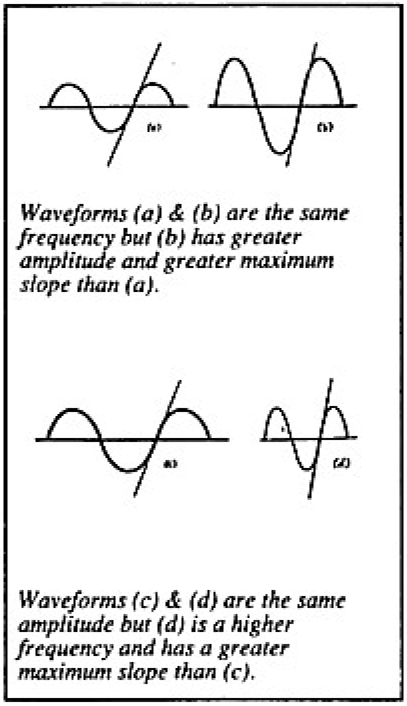
The heat sink actually used was 200 mm of Mullard 55D type material. Conservatively selected on the basis of continuous power in the amplifier, it is somewhat of an overkill for sideband speech transmission with lower average power and for this application, a smaller heat sink could be used, as suggested in the previous paragraph. Beryllium insulating washers were fitted because of their low thermal resistance. (There is not much point in having a large heat sink if a high temperature gradient is allowed to build up across the insulating washer.)

The output coupling transformer, made up by quadfilar winding on a 29 mm ferrite toroidal core, reflects a drain to drain load resistance of 50 ohms at its primary windings from the 50 ohm output circuit. Maximum output power (Pm) is calculated from the following:

$$Pm = \frac{2(Vp)^2}{Rdd}$$

where Vp = peak voltage swing
and Rdd = drain to drain load resistance.

At low frequencies, the value of Vp approaches the voltage of the supply rail giving an output power calculation of over 100 watts. As frequency is increased, a limiting factor on the value of Vp called slew rate comes into effect. Slew rate is the maximum instantaneous change in output voltage that the amplifier can deliver in a given time. If we consider an output signal waveform, the maximum rate of change in voltage of the waveform, or the maximum slope of the waveform, increases with both frequency and amplitude of the waveform (refer figure 4). For a given frequency, maximum possible output voltage is achieved when the slope equals the amplifier slew rate. When frequency is increased, the maximum voltage swing is reduced and so also is the maximum power output. This explains why we can only get 55 watts at 1.8 MHz and an even lower value of 25 watts at 7 MHz.



Waveforms (a) & (b) are the same frequency but (b) has greater amplitude and greater maximum slope than (a).

Waveforms (c) & (d) are the same amplitude but (d) is a higher frequency and has a greater maximum slope than (c).

Figure 4 Maximum slope of waveform increases with both amplitude & frequency.

The problem of limiting slew rate can be dealt with by using a higher frequency type of power transistor, but it will cost more. The IRF430 can be purchased for about 4 or 5 dollars, considerably less than a 'hot' RF type. This gets back to the basis of Drew's article in making an amplifier suitable for the lower frequency amateur bands with moderately priced transistors. Actually, the IRF510 transistors used by Drew are a little faster than the IRF430 but of course rated for lower power.

Power Supply

The power rail selected was 60 volts, supplied via an ILP 225 VA toroidal power transformer, bridge rectifier and simple capacitance filter. Regulation was considered adequate at around 6%. For SSB speech transmission, the transformer rating is a considerable overkill and a more economical approach might he to rewind the secondary of an old valve receiver power transformer. A secondary winding of 50 volts at around 1.5 amp rating should be satisfactory.

A further 15V rail was derived from the 60v supply via a voltage regulator. This rail, used elsewhere as a general system supply, is also used as a source to bias the IRF430 transistors.

Harmonics

Odd order harmonic level at the output of the amplifier is quite high and as is usual when using a broadband amplifier to feed an antenna, an efficient harmonic filter for each band is required. The filter should have at least 50 dB of rejection at the third harmonic frequency. Design of suitable low pass filters is given in amateur radio handbooks. Design information has also been given in previous issues of AR, in Drew's article (ref 3) and in one of my own (ref 4). If toroidal cores are used to construct the inductors, iron dust cores and not ferrite types are recommended. The ferrite cored inductors have been found to change their value of inductance when high power is pumped through them, resulting in de-tuning of the filter. The iron dust cored inductors are more stable in this respect. As an alternative, air cored coils can be made with quite high Q and are quite satisfactory. They also cannot saturate so that they are very stable and are certainly cheaper to construct, even if they are a little larger.

Operation

With no signal input, the transistors are individually set for a static drain current of 150 mA. To do this, the alternate drain circuit is opened whilst the relevant bias potentiometer (R1 or R2) is adjusted.

To cheek maximum power output at a given frequency, the amplifier is connected to a 50 ohm dummy load and the AC output voltage across the load is monitored with a CRO. The input can be fed from a standard RF signal generator as around 25 mW is all that is needed to drive the amplifier to full power. The input level is increased until the output waveform shows limiting and the peak to peak voltage (Vpp) is then recorded. Maximum power output (Pm) is given by:

$$Pm = \frac{(Vpp)^2}{400}$$

Total drain current should not be allowed to exceed 2.2 A and if the heat sink is not designed for continuous signal operation, the current should not be sustained for more than a brief period.

As discussed previously, the higher the operating frequency, the lower is the peak to peak voltage swing (and hence drain current swing) achievable without serious waveform distortion. Amplifier efficiency is decreased with a rise in frequency, but as maximum drain current is also decreased there is no problem of increased power dissipation and subsequent rise in heat sink temperature at the higher frequencies.

Summary

The reasonably priced IRF MOSFET transistor can be used to provide moderately high power amplification on the lower frequency amateur bands. Maximum power output decreases as frequency is increased but quite reasonable performance can be achieved at 1.8 and 3.5 MHz using an amplifier such as the one described. The amplifier is still usable at 7 MHz but at reduced power.

References

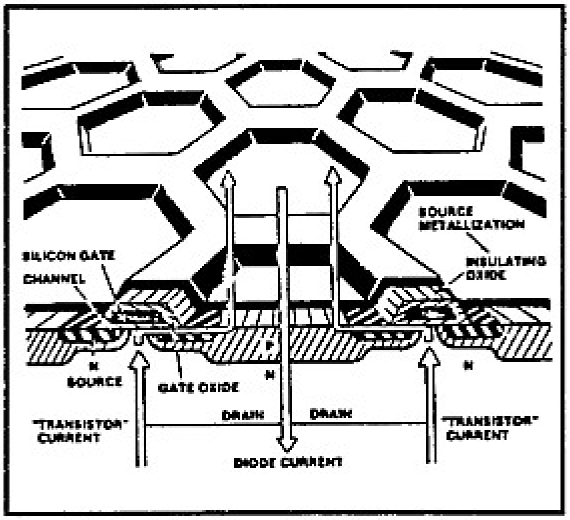
1 International Rectifier HEXFET Databook - Power MOSFET Application & Product Data.

2 Motorola Handbook - Power MOSFET Transistor Data.

3 Drew Diamond VK3XU - MOSFET Power Amplifier for 1.8 to 10. 1 MHz Novice Notes, Amateur Radio, Oct 1988.

4 Lloyd Butler - Tank Circuits & Output Coupling - Amateur Radio May 1988, plus corrections July 1988.

APPENDIX - The HEXFET Structure (from Reference 1)



The HEXFET surface is characterized by a multiplicity of closed hexagonal source cells (over 500,000 per square inch) from which the name HEXFET is derived. In cross section, the HEXFET is based on a double-diffused (DMOS) structure. A channel is formed by double diffusion at the periphery of each hexagonal source cell. An insulating gate oxide layer covers the channel. A silicon gate then overlays both the insulating oxide and channel. The silicon gate in turn is insulated from the source by an additional oxide layer. All of the hexagonal source cells are then parallel connected by a continuous sheet of metalization which forms the source terminal.

Transistor action occurs by penetration of an electric field into the channel area which modulates the conductivity between drain and source. Conventional current flow is from the drain substrate, across the channel surface, and vertically out the source terminal.