RF power amplifier design

While many electronics textbooks have a chapter devoted to RF power amplifiers, and also several complete books about this matter exist, I haven't yet found any that explains the whole topic in a way that would enable an electronics hobbyist, or even a practically minded professional circuit designer, to fully understand the matter and achieve designs that work well and for a long time. Most books treat this matter in a totally theoretical way, with excessive use of maths. Practically-minded electronicians most commonly don't understand the sort of mathematical language used in those books, while the mathematicians and theoretically minded engineers who do understand them are usually not the people who would actually design and build an amplifier!

Many people have asked me to write an article about RF power amplifiers, in a way that practical electronicians can understand. Also over many years I have come to acquire a better understanding of these circuits, which often work in ways that are far less trivial than their schematics look. Now, in 2020, confined at home due to the COVID-19 pandemic, time has come to write this article.

Most books about RF power amplifiers are mainly oriented toward the cellphone market, so they place a big emphasis on UHF and microwave amplifiers, often not even mentioning HF. For this reason I will aim my article mainly at HF, extendable to VHF. In fact, as a long-time radio amateur, I have ham radio applications in mind as my primary design objects. So this article is oriented mainly at power amplifiers operating in the range of 1.8 to 54MHz, sometimes to 148MHz, in a power range up to 1500W. I will include driver stage design down to the milliwatt level.

This is a very long web page, almost long enough to be printed in book form, and if you are really interested in the matter, you will need to work through it step by step. It's organized by background color: The main article has my traditional yellow background, while accessory explanations are on blue background. If you already know what I explain in such a Blue Block, you can skip right over it and continue reading the main article.

I will consider the specific quirks of both BJTs and MOSFETs. Since MOSFETs are far more common nowadays, I will draw MOSFETs in all those situations that are applicable to both types of transistors. Throughout this article, when I write "transistor" it means that the matter described is applicable to both types, and I will talk about source, drain and gate as a matter of standards, even when that could be emitter, collector and base. When I write "BJT" and "MOSFET" it means that the matter is specific to one type. Totally obsolete devices, like vacuum tubes, aren't considered. And YES, in 2020 tubes are indeed totally and fully obsolete, in the power and frequency ranges considered in this article! So don't be mad at me, and try not to shed any tears over this fact. Even BJTs are basically obsolete at this time for RF power use, and most are no longer being manufactured, but there is still a lot of equipment around that uses BJTs, so I'm including them.

Classes of amplifiers:

In the beginning of electronics, shortly after Adam invented the vacuum tube and Eve promptly criticized him for doing so, amplifiers were divided into three main classes. Any amplifier in which the active device conducted current all the time, throughout the entire signal waveform, was called "class A". Any amplifier in which the active device conducted current during exactly one half of the driving signal's waveform, usually the positive half, was called "class B". And any amplifier in which the active device conducted current during less than one half the driving waveform was called "class C".

Soon after doing this classification, the Adam of electronics noticed that he had forgotten to consider all those amplifiers in which the active device conducts current during more than half of the signal's waveform, but less than the whole time. Since this mode of operation falls between class A and class B, and given that despite much desperate searching no letter could be found between those two, this operation mode was called "class AB".

Electronicians love to talk in angles. So the whole period of an RF cycle is considered to have 360 degrees. For this reason one can say that transistors operating in class A conduct over 360°, while in class AB they conduct during less than 360° but more than 180°, in class B they conduct exactly during 180°, and in class C they conduct for less than 180° of the signal's period.

Since it is very hard, or even impossible, to bias a transistor precisely to its conduction threshold (which by the way is poorly defined!), so that it conducts **exactly** for 180°, true class B amplifiers strictly don't exist! For that reason some designers use the "class B" designation for any amplifiers that are **close** to true class B, that is, amplifiers that operate in class AB but with a very small idling current, so that they conduct for just slightly more than 180°. I will call such amplifiers class AB, but if you read certain textbooks and papers you need to know that my class AB could be called class B by some authors.

A theoretically perfect amplifier, using perfect components, would achieve a maximum efficiency of 50% in class A and 78.5% in class B, both at the limit of starting to cause distortion by clipping the signal. A class C amplifier has higher efficiency than that. The smaller its conduction angle is, the higher is its efficiency, but the lower is its output power. As the conduction angle approaches zero, the efficiency approaches 100%, but the power output approaches zero, so that's not very useful.

It is often said that class A, AB and B amplifiers are linear, while class C amplifiers are not. But this is incorrect! Firstly, all amplifiers have residual nonlinearities that cannot be fully eliminated. Secondly, an A, AB or B amplifier will turn dramatically nonlinear when overdriven, and sometimes they are deliberately operated in this overdriven, nonlinear mode. And thirdly, and this is a point many people haven't realized, a class C amplifier can be built to be linear! All it takes is devising a biasing scheme that keeps the class C amplifier operating over a **constant** conduction angle, even while the amplitude varies. But this technique is rarely applied, so in practice most class C amplifiers are indeed nonlinear.

Long after Adam had ceased doing electronics, more amplifier classes were added, and these aren't based on conduction angle. And they are usually more efficient than the old ones. In class D amplifiers the transistors are switched on and off, instead of being linearly driven. There are two ways to implement a class D amplifier. In one of them the transistors switch the output **voltage,** which then attains a nearly square waveform, while the output current is defined by the resonant circuit coming after the transistors, and is roughly sinusoidal. In the other kind, the transistors switch the output **current**, resulting in a nearly square current waveform, while the resonant circuit at the output defines the voltage waveform, which becomes nearly sinusoidal. The two types are called, appropriately, "voltage-switching class D", and "current-switching class D" amplifiers.

In class E amplifiers the transistors are also switched on and off, but a resonant output network is used that shapes both the current and voltage waveforms to be non-square, and this largely avoids simultaneous voltage and current on the transistor, further improving efficiency. If a transistor actually could be switched instantaneously, class E wouldn't be necessary, as class D would be perfect - but real transistors can't switch that fast. Class E amplifiers basically work by the tuned output network largely lifting the instantaneous load on the transistor during the non-zero duration switching transitions, which enables getting very high efficiency from relatively slow transistors.

Class F amplifiers also use transistors in switching mode, but use a more complex output network that treats the odd harmonics in one way, and the even ones in another. This results in voltage and current waveforms that optimize both the efficiency and the achievable power output, for a given transistor.

In addition to these 7 basic amplifier classes, there are several more, which have been defined by specific designers and companies, and are often treated as intellectual

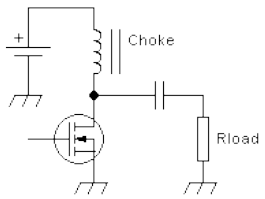
property by them. Two different companies might assign different class letters to basically the same amplifier type.

It should be noted that the definitions of the classic classes A to C, and the new kids D to F, overlap, because the classic ones are defined by conduction angle, and the modern ones by circuit layout and waveforms. Transistors in class D, E and F amplifiers theoretically operate with 50% on, 50% off time, and thus would be considered class B amplifiers by the old definition, driven into saturation. But often in practice their conduction angle is slightly different from 50%, and thus they become either class C or class AB amplifiers by the old definition. This is important to bear in mind, because many transistor datasheets give circuit examples which the manufacturers claim to operate in class AB, but that truly are operating in class D. The power output and efficiency claimed by the manufacturers is very high, typical for class D, but people reading the datasheets and noticing the "class AB" claim will mistakenly believe that the transistor described can deliver that high power and efficiency while operating linearly! This has led many ham experimenters to blowing up expensive MOSFETs by operating them at unrealistically high power levels in linear class AB, trying to achieve the same power that the manufacturer claims for the nonlinear class AB / class D hybrid test circuit.

RF power amplifiers are best designed backwards. that is, you start at the output side, and then continue toward the input side. When designing a multistage amplifer, you start with the final stage, then continue to the driver. Keeping with this logic, let's start looking at **output circuits** for various types of amplifiers.

Single-transistor broadband amplifiers:

This kind of amplifier is usually operated in class A. There is a constant supply current flowing through a choke. The transistor conducts more or less of this current to ground, as commanded by the drive signal. The remainder of the current is forced to go through the load resistance. A coupling capacitor allows the RF current to flow, while blocking the supply voltage.



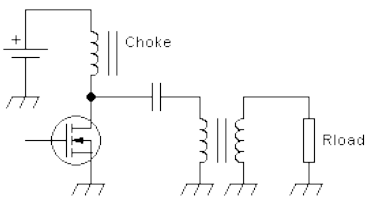
In all circuits on this page, the battery represents a perfect, stable voltage source. In a practical amplifier there wouldn't be a real battery, of course, but a combination of bypass capacitors, typically fed from a regulated power supply, sometimes through an additional choke. And the gate, drawn open here, obviously needs to be connected to a circuit that applies a DC bias and an RF drive voltage.

Since a choke cannot have any DC voltage across it, the supply voltage appears in full at the drain of the transistor. When the transistor is driven harder during the positive half cycle, it conducts more current and thus temporarily pulls its drain voltage down. The additional current flowing in the transistor at that time comes from the capacitor, thus drawing a negative current through the load resistor. Likewise, when the transistor is being driven softer, during the negative half of the driving waveform, it conducts less current than what's coming through the choke, and the surplus must flow through the capacitor and resistor. That's a positive current in the load resistor, and the voltage at the drain becomes higher than the supply voltage.

In normal operation, a typical RF power transistor can pull its drain voltage down to about one tenth of the supply voltage. If the drive signal is symmetric, and the transistor is linear and all goes as it should, then the output signal will also be symmetric, undistorted, and will reach a peak value equal to 90% of the supply voltage. Thus the peak-to-peak output voltage can reach 1.8 times the supply voltage, and since RMS voltage is the peak-to-peak value divided by $2\sqrt{2}$, that's an RMS voltage of about 0.64 times the supply voltage. This is the maximum effective undistorted RMS voltage that can appear on the load. And since power is RMS voltage squared and then divided by load resistance, it's easy to calculate the power available on the load. For example, if the supply provides 12V, and the load resistor is 10Ω, then we get 21.6Vp-p on the load, which is roughly 7.6Vrms, and on 10Ω that makes roughly 5.8W. Note that the exact value depends on the actual saturation voltage of the transistor, which changes according to transistor type and operating conditions. The 10% saturation voltage I assumed here is typical, but individual situations can stray considerably from this value.

The DC current in the choke will be determined by the transistor's biasing. To obtain correct, undistorted operation, it's necessary that the transistor never runs out of drain current, and thus it has to be biased to a DC current larger than the peak signal current. That's class A operation. Since this peak signal current is roughly 0.9 times the supply voltage divided by the load resistance, in the example above it is 1.08A, and you would probably fare well by biasing this transistor to some DC value between 1.2 and 1.5A.

If you need a different output power, and the supply voltage is given, then the only item you can adjust to suit that requirement is the load resistance. But typically there is a requirement to make the amplifier work into a specific load resistance too! So it's most usual that some transformation of the load resistance is required. This could be done by means of a transformer inserted between the amplifier and the load, like this:



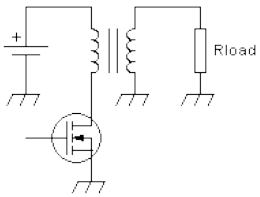
The transformer can be of the conventional type, as shown in this drawing, or it could be a transmission line transformer. In a conventional transformer, the turns ratio is equal to the voltage ratio, inversely proportional to the current ratio, and equal to the square root of the resistance or impedance ratio. For this reason it's important to always make sure you understand whether a transformer said to have a certain ratio, such as 1:4, is meant to have a turns ratio and voltage ratio of 1:4, or an impedance ratio of 1:4. In RF practice it's more common to mean the impedance ratio, while low frequency electronicians almost invariably talk in terms of turns ratio, and some engineers coming from the low frequency world carry this over into the RF world. And since a transformer with a 1:4 turns ratio has a 1:16 impedance ratio, the difference **is indeed** important!

Let's assume that you want an output power of 30W from your 12V-powered amplifier, and that you need to provide this power to a 50Ω load, which is the most usual load resistance in RF power electronics. Multiply the power times the load resistance, take the square root of the result, and you get the RMS voltage. In this case, that's 38.73Vrms. But your transistor's drain still can only swing linearly over a voltage range that gives 7.64Vrms. So you need a transformer having a voltage ratio and turns ratio

of 1:5.07. In practice 1:5 will be close enough. That transformer has a 1:25 impedance ratio, and so your 50Ω load resistance gets transformed down to just 2Ω of load resistance on the transistor!

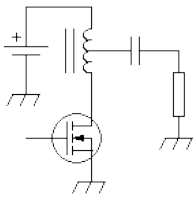
The peak current produced by your 10.8V maximum peak voltage into that 2Ω load is 5.4A, and so you will need to bias your class A amplifier for at least 6, perhaps 7A DC, to get largely undistorted operation. At 12V that's 72 to 84W of continuous input power, for barely 30W peak output power. That's a lousy efficiency, and this is the reason why such class A amplifiers are rarely used at power levels higher than a few watt. But they are nice and simple and produce low distortion, which makes them useful at low power levels.

In fact they get much simpler and more economical by designing the transformer to be able to handle the amplifier's DC current, thus taking over the choke's function, allowing us to eliminate the choke and the capacitor:



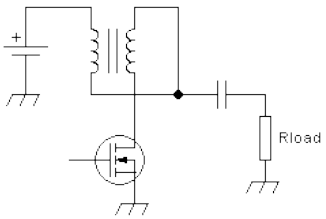
In fact this is the way many such amplifiers are built.

When the turns ratio required is close enough to 1:2 or 2:1, the transformer's size and cost can be shrunken quite effectively by using an autotransformer:



Since the autotransformer provides no galvanic insulation, the coupling capacitor again becomes necessary. But the small autotransformer has better coupling and lower parasitics, and thus gives a better frequency coverage.

This autotransformer is often made with a bifiliar winding, and then it's often drawn in this way:



It's simply the same thing drawn differently. Some authors claim that these bifiliar transformers are transmission line transformers, but that's a matter of debate and definitions. And so it's time for the next Blue Block:

Transformers: Conventional, transmission line, real and fake ones

A conventional transformer consists of two or more windings that share a single magnetic core. Any magnetic field caused by current flowing in one winding will induce a voltage in **all** windings. This allows a transformer to transfer energy from one winding to the others, and by making windings with different turn numbers, it can also convert one combination of voltage and current into a totally different one, keeping the power constant except for some slight losses. It is a very versatile device, that found many applications in electricity and electronics even before Adam invented the vacuum tube. But it has important drawbacks too, called "parasitics". And the big problem for us, when building RF power amplifiers, is that in RF power applications these parasitics are particularly nasty.

All real-world conventional transformers have less than perfect coupling between their windings. This shows up in their behavior as if there were additional inductors connected in series with their windings, these little inductors having their own separate magnetic circuits, being uncoupled to the main transformer. As the frequency gets higher, this "leakage inductance" ends up making a transformer unusable. In most RF power applications, leakage inductance is what determines the upper frequency limit at which a conventional transformer can work well enough. It is possible to reduce the leakage inductance by intimately mingling all windings, for example by using twisted wires to wind all windings together, but this increases another parasitic, which is the capacitance between windings. There is no free lunch here, only a chance to make the best compromise.

The allowable wire length of a winding is limited by the simple reason that current only flows at a speed close to the speed of light. The wire length has to be much smaller than the wavelength, for the transformer to function normally. At HF this can already start to become a problem, while at VHF it's a serious problem. And even in HF amplifiers, transformers are often expected to carry non-sinusoidal signals, which means that there are harmonics present that reach high into the VHF range, and even into UHF. So, winding wire length can become a critical factor even at HF.

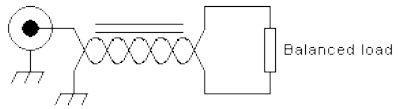
The larger a conventional transformer is, the higher are these parasitics. But in high power applications we need rather large transformers, to handle the high voltages and currents, so this is bad news, and ultimately limits the applicability of conventional transformers in RF power amplifiers.

Transmission line transformers instead work on a completely different principle: One or more transmission lines of a properly chosen impedance, carrying RF energy from

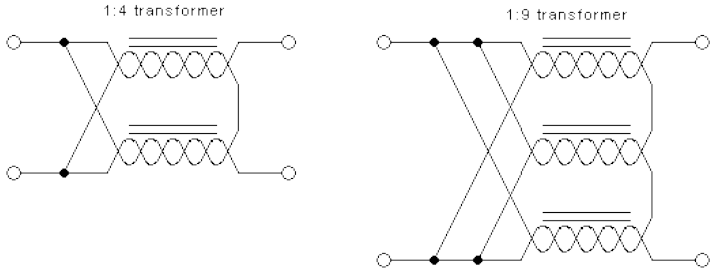
the input side to the output side, and having something on them that makes the output side not "see" the input side, except through the line proper. This "something" is usually a magnetic core on which the transmission line is wound, which makes the line act as a common-mode choke: Any differential current on the two conductors can pass freely, while all common-mode current is largely blocked by the inductance or impedance of the line wound on the core.

The transmission lines can be coaxial cable, parallel line, twisted pairs, etc. Some people draw transmission line transformer sections as coax conductors with a core, or twisted lines with a core, or even two windings on a core. The core is sometimes drawn using the standard schematic symbol, but some designers prefer to physically draw a toroid around the line. There is much creativity in this area of electronics art. I prefer to draw them either as twisted line or as coax line, with a symbolized core, like this:

A single transmission line on a single core can be used as a balun: One end of the line has one conductor grounded, the other conductor connected to an unbalanced signal source. The other end of the line has the same signal voltage between its conductors, at the same signal current, but is no longer ground-referenced. It's a floating, differential signal source. It can be used to drive a balanced load:



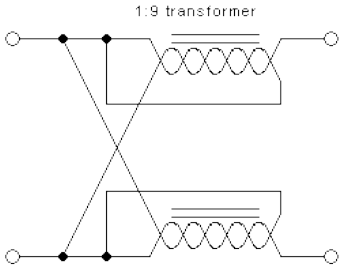
What's more fun is that two such transmission lines, each of them wound on its own core, can be connected with their inputs in parallel and their outputs in series, thus implementing a transformer with a 1:4 impedance ratio. This principle can be extended to three lines and cores for a 1:9 transformer, and so on. These transformers can be freely used with balanced or unbalanced circuits on each side, since they all also provide balun functionality.



The beauty about these transmission line transformers is that they almost don't have a practical upper frequency limit. Even when the transmission lines are long relative to the wavelength, they are all the same length, and so the output signals from all of them are in the same phase, allowing series connection of the line outputs.

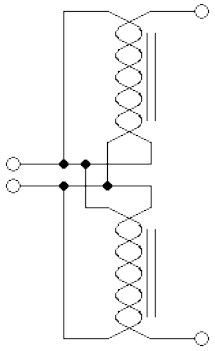
Whenever the transmission lines used to make such transformers have an appreciable length, relative to the wavelength of the highest frequency at which we want them to work, they need to have the same impedance as the signals they are carrying. For example, if the 1:4 transformer shown above is to be used to match a 12.5Ω source to a 50Ω load, then each of the two transmission lines needs to have a 25Ω impedance. The two line inputs in parallel will match to the 12.5Ω source, and the two line outputs in series will match to the 50Ω load. In many practical situations it will be necessary to construct transmission lines of the proper, specific impedances. In other cases it may be necessary to wind such transmission line transformer elements with two or more coax cables and connect them in parallel, to achieve a low impedance. For higher impedances it will be necessary to use parallel transmission lines. And transformers for extremely high or low impedances can be hard or impossible to make.

In some situations, when the maximum operating frequency is low enough so that the line length is very short compared to the wavelength, one can save one transmission line and its core, by connecting the line outputs in series with the signal input:

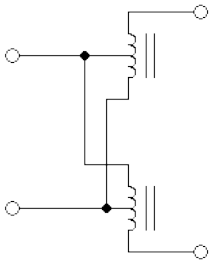


This configuration gets many different names, among them "bootstrapped transmission line transformer", "false transmission line transformer", "non-equal delay transmission line transformer", among various others. The point to keep in mind is that this kind of configuration does have a strict upper frequency limit, given by the requirement that the transmission lines be very short relative to the wavelength. The balun effect is also lost.

This circuit could be redrawn in this way, for improved clarity:



And many people claim that this is nothing else than a combination of two autotransformers, which could also be drawn like this:



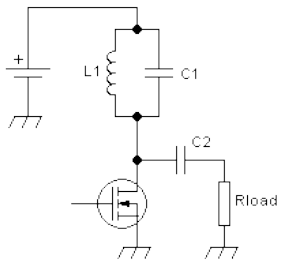
So, it is important to remember one very important point: When you see a combination of transmission lines wound around cores, in which each line's end is connected **either** to the input side **or** the output side, it's a true transmission line transformer that has a very wide frequency range. Instead if you see lines that have the two poles of the **same** end connected to the input **and** output circuit, then it's **not** a true transmission line transformer, but one of those contraptions that suffer from the same wire length limitations as conventional transformers do.

Single-transistor tuned amplifiers:

Instead of using a choke to feed the supply current into the amplifier, and use transformers to adapt the load impedance, it's possible to do both tasks using tuned circuits. The disadvantage of this approach is obvious: The amplifier will only be usable inside a relatively narrow frequency range. But in many cases this is acceptable and even desirable, for example in single-band radios. And there are various advantages in using tuned circuits: They typically don't need ferrite cores, which eliminates some cost and weight, and also does away with all the characteristic problems of ferrite materials, such as losses, DC saturation and nonlinearities.

The narrow-banded nature of tuned circuits makes them suppress harmonics, and this allows operating these amplifiers in classes other than class A. Tuned linear amplifiers can be easily built using a single transistor, with far better efficiency than class A broadband amplifiers. And class C tuned amplifiers, while typically nonlinear, can easily achieve practical efficiencies of roughly 80%.

The basic single-transistor tuned amplifier's output section looks like this:



As you have immediately spotted, it's the same basic circuit as the broadband one, except that the choke has been replaced by a parallel tuned circuit. This tuned circuit is resonant at the operating frequency, or more typically at the center of the desired operating frequency range. It has a Q factor suitably selected to meet several goals: Higher Q leads to smaller bandwidth, better harmonic suppression, smaller inductor, larger capacitor, higher loss and more critical tuning. So the Q is a compromise, but many designers tend to use a loaded Q of roughly 3 in a typical situation. This means that at the operating frequency both L1 and C1 have one third as much reactance as the resistance value of the load. The inductive reactance of a coil is

$Xl = 2 \pi F L$

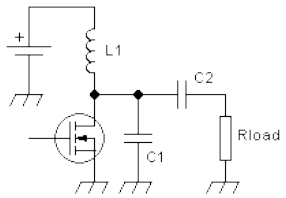
and the capacitive reactance of a capacitor is

$Xc = 1 \div (2 \pi F C)$

where of course F is the frequency in hertz, L the inductance in henry, C the capacitance in farad, Xl the inductive reactance in ohm, Xc the capacitive reactance in ohm, and π is roughly 3.1416.

It's very usual to connect C1 from the drain to ground rather than in parallel to L1. It works exactly the same, because the top end of L1 is grounded at RF through the battery,

or in a practical circuit through the bypass capacitors installed at that location. The circuit then would look like this:



As you can see, C1 ends up directly in parallel with the transistor's output side, which means that the transistor's output capacitance, if significant, needs to be subtracted from the value you use for C1. This is true just the same if C1 is placed in parallel to L1, by the way.

The relationships between supply voltage, drain voltage, load resistance and output power are exactly the same as in the broadband amplifier. But tuned amplifiers are typically biased to a much lower standby current, operating in class AB, or even to no standby current at all, operating in class C. It's interesting to understand what currents will flow in such a case: The resonant circuit L1/C1, if it has a high enough loaded Q, will always force a nearly sinusoidal voltage at the drain, and at the same time can source or sink any instantaneous current needed to achieve this. This is a very different behavior from that of the choke in the broadband amplifier, which delivers a constant current over the RF cycle, regardless of the drain voltage waveform. We can also think of this in the frequency domain: While a choke has a high impedance over the whole RF spectrum of interest, a resonant circuit has a high impedance at the fundamental frequency, but low impedance at its harmonics and all other out-of-band frequencies .

Whenever the transistor is driven on, it will draw a current from the load and the resonant circuit, defined either by the instantaneous gate voltage (while in active range) or by whatever current is available to be drawn (when saturated). As the transistor draws current pulses, over the course of several RF cycles the resonant circuit increases the sine voltage amplitude across it, until reaching equilibrium either with the total average current drawn by the transistor and the current delivered to the load, or until the transistor saturates and the RF voltage gets clipped to the supply voltage. MOSFETs conduct in both polarities when driven on, and this will limit the peak RF voltage to roughly the supply voltage, because the drain can't go significantly negative. BJTs instead do not conduct collector-to-emitter when reverse-polarized. But the base-collector junction does turn on, and this pulls the base voltage down in such a situation, also limiting the collector voltage swing. The clamped RF voltage, applied to the load resistance, defines the load current, and this in turn limits the average current that the transistor can draw. Of course, as the conduction angle of a class C amplifier is reduced, the peak current in the transistor increases, for a given output power.

If a perfect (fully linear) transistor is biased for class B operation, the amplifier will be linear, despite the transistor amplifying only one half of the RF waveform, and staying completely off during the other half. If a real-world transistor is used, which has a somewhat nonlinear transfer curve, acceptable linearity for most purposes can be achieved by biasing the transistor into class AB, with the amount of idling current optimized for lowest distortion.

The very fact that RF power amplifiers can be linear even when the transistor is completely off during half of the time, puzzles many newcomers to RF, and this raises the need for another of my Blue Blocks:

Linearitity at RF and audio - two different worlds

Audio fans are used to just one kind of linearity: The amplifier is supposed to deliver an output signal that is an exact copy of the input signal, except that it has been enlarged. Since audio equipment needs to process a wide wide frequency range, such as 20Hz to 20kHz, a 1:1000 range, and many individual frequencies are usually present simultaneously over this wide range, forming a complex waveform which is often asymmetric, there is really just one way to achieve the required linearity: At every instant the amplifier has to produce an output voltage that is exactly proportional to the input voltage, the relationship being set by a fixed gain factor which stays constant over the whole frequency range.

But in most kinds of RF work, we don't have such wideband signals. A typical ham signal might have a center frequency of 14320kHz, but a total bandwidth of just 3kHz. That's a bandwidth of just 0.02%, so narrow that on a scope it looks like a single frequency. For this reason the waveform of a ham radio signal will always be extremely close to a perfect sine wave, and since we **know** the waveform, there is no need to make the amplifier painstakingly reproduce it. We can allow the amplifier to brutally distort the waveform, and then use a frequency-selective network to eliminate the harmonic frequency components, which fully restores the original sinewave shape. The amplifier only needs to create an output signal whose power varies in exact proportion to the drive power, and that has the same phase modulation, if any, as the input signal. If the amplitude and phase are right, and the sine waveform is reconstructed by frequency-selective networks, then the amplifier will be just as linear for a narrow RF signal as an amplifier that offers perfect waveform reproduction of any input signal.

So, remember: If an amplifier accurately reproduces both the amplitude modulation and the phase modulation of the input signal, it's a linear amplifier for all narrow signals at RF. Waveform linearity is absolutely not required, but typically the sine waveform needs to be reconstructed by filters, to avoid unacceptable harmonic distortion causing interference on higher bands.

Audio guys can't do that, given their complex, wideband signals. On the other hand audio guys get the advantage of transistors that are thousands of times faster than their signals, and that have extremely high gain at their frequencies, so they **can** easily build waveform-linear amplifiers. We RF guys have a much harder time doing so, because the available transistors are slow and have low gain at our frequencies, and even a few cm of circuit board trace cause a phase delay that could be many degrees of our signal's period. So we do whatever is simplest and most cost-effective in achieving our goal, and very often this is building amplifers that have extremely poor waveform linearity but good-enough amplitude and phase linearity, and then we clean up their output waveform using resonant circuits or lowpass filters.

Since in most situations we need to design the amplifier for a load resistance which is different from the drain load needed to produce the desired output power at the available supply voltage, we need a **matching network**. The simplest possible tuned matching network is the L circuit. It's called L because it looks like a lying-down letter L, consisting of either a series coil and a parallel inductor, or the other way around. The first option has some lowpass behavior, the second one is rather a highpass. Since usually we need to suppress harmonics, the lowpass version is by far the most used one. The parallel element is always placed at the side of the L network where the higher impedance is.

The equations for calculating the values for a lowpass L matching network are the following:

$$Xl = \sqrt{(Rlow \times Rhigh - Rlow^2)}$$

$$Xc = Rlow \times Rhigh \div Xl$$

Let's calculate the values for a 12V powered amplifier that delivers 30W into a 50Ω load, like in the broadband example given above. There we found that we needed a transformer with a 1:5 turns ratio, giving us a 1:25 impedance ratio, thus transforming the 50Ω load resistance down to a 2Ω drain load. Let's do the same with a tuned L matching network:

$Xl = \sqrt{ (2 \times 50 - 2^2)} = 9.80$

$Xc = 2 \times 50 \div 9.80 = 10.2$

So we need a coil having 9.80Ω inductive reactance at the design frequency, and a capacitor having 10.2Ω capacitive reactance. If we want to build this amplifer for the amateur 6 meter band, we design it for the center frequency, 52MHz. Applying the reactance equations given higher up in this page, we end up with a 30nH coil and a 300pF capacitor. Note that a 30nH coil might need just two turns of wire, with a small diameter! This is why in amplifiers that work at much higher frequencies than this one, such coils are no longer practical, and circuit board traces (microstriplines) are used instead. And the 300pF capacitor looks easy at first sight, but it's important to make sure that it has negligible stray inductance. A capacitor with wire terminals isn't a good idea here. A parallel combination of two or more SMD capacitors works well.

Such a circuit almost always will need to be tuned after building. It is possible to use trimmer capacitors, while such small coils are most easily trimmed by deforming them, mostly changing the separation of their turns.

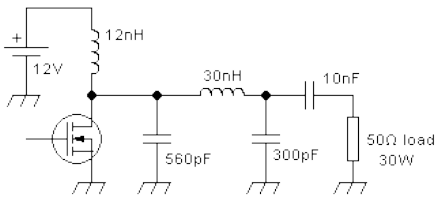
While we are at it, let's quickly calculate the resonant circuit at the drain: For a loaded Q factor of 3, and given the drain load resistance of 2Ω, we need reactances of just 0.67Ω! At 52MHz this translates into an inductance of barely 2nH, and a capacitance of 4.57nF. The drain "inductor" would be just 2 or 3mm of straight wire, while the capacitor would need to be made from several very small SMD parts in parallel to achieve low enough stray inductance.

So these are quite impractical values. I chose this example to illustrate the sort of difficulties often encountered in RF power amplifier design: Very low impedances, requiring impossible components. An amplifier like this, delivering 30W from 12V with a pretty clean sine wave at the drain and a simple L matching network, can often be made in the lower HF range, but hardly at higher HF, let alone VHF. That's one reason why VHF amplifiers typically either work in some other class, such as class E, or else have low efficiency.

We can start making compromises, to get our 52MHz 30W amplifier to operate well enough. The most obvious one is lowering the Q of the resonant circuit, to end up with workable components. The problem is that as we reduce the Q, the waveform at the drain distorts, and we can no longer obtain the power we expect without getting high drain voltage peaks. If we retune the matching network to deliver the desired power while using a very low Q drain tank, the result is sky-high drain voltage peaks, which in the best case reduce the efficiency of the amplifier, and in the worst case destroy the transistor.

Reducing the tank Q to unity, resulting in a 6nH coil and a 1.5nF capacitor, we still get the desired power, at a peak drain voltage of only 27V, instead of the 22V we got with a Q of 3. This is more workable. Reducing it further, to 0.5, using a 12nH coil and a total capacitance of 750pF, which includes the transistor's output capacitance, the drain voltage waveform gets very ugly, but the circuit still works well enough, and our component values start becoming practical. Below that level the drain voltage waveform definitely becomes too peaky, deformed, and the output power falls off dramatically.

So, our 12V, 30W output section for the 50-54MHz band might end up looking like this:



The 560pF is tentative, subject to confirmation of the output capacitance of the transistor chosen. The 10nF capacitor serves just a DC blocking function. Its reactance at the operating frequency is low enough to have no significant effect. I placed it at the output side of the L network, because there the RF current is much lower than at its input side, and the DC voltage across the capacitor stays the same.

When building such an amplifier it's absolutely critical to minimize stray inductances. The 560pF capacitor needs to be soldered directly to the transistor's body, not one centimeter away. It needs to be a chip capacitor, because any wire terminals at all will contribute too much stray inductance. It's best to split it up into two capacitors of half the value, and solder them to the transistor on both sides of the drain terminal. Any deviation from this will make the amplifier behave less like the calculation indicates. And even if you follow these recommendations, you need to keep in mind that this design method is very simplistic, for example by not considering the inductance of the drain connection inside the transistor. One can usually get away with such simplifications at HF, but at 52MHz it's risky, and on even higher bands it definitely no longer works well enough. At VHF and higher one needs to design the matching network for the load impedance the transistor needs at its drain terminal, and not the one it needs inside, at its actual drain on the silicon chip. This load impedance needs to be looked up in the data sheet, which usually gives it for a single or a few operating frequencies and power levels. It often needs to be interpolated, if your application is not a completely standard one for the chosen transistor.

The L matching network is by no means the only choice for such a simple amplifier operating in class AB or class C. In fact, very often there are better choices. That's because the L network doesn't allow you to choose the loaded Q of the network. The Q of an L network is

$Q = \sqrt{ (Rhigh \div Rlow -1)}$

So in this example the loaded Q is 4.9. One of the implications of this is that the network has a -3dB bandwidth of roughly one fifth the center frequency. The usable bandwith is even less, because 3dB power drop at the band edges is not acceptable in a power amplifier. But since the amateur 6 meter band has a width of one thirteenth of the center frequency, the bandwidth of this network with a Q of 4.9 would be sufficient. But there may be other applications in which more bandwidth is needed, making a lower Q desirable. Or there might be a more stringent requirement for harmonic suppression, making a higher Q desirable.

Also it's important to consider the stress placed on the components. A loaded Q of 4.9 also means that the network is storing 4.9 times as much reactive power, as the output power is. So, this little network is storing nearly 150W of reactive power! Even if you make the tiny 30nH coil from really thick wire, it might still have enough RF resistance (which is far higher than its DC resistance, due to skin effect) to heat up so much that it unsolders itself from the circuit! You might need **really** thick wire for it, or perhaps use silver-plated copper sheet. The silver plating helps just a tiny bit, but using sheet instead of round wire helps a lot.

Likewise, the 300pF capacitor is under high stress. It works at the full output voltage of 39V RMS, but also at nearly the full **input** current to the network, specifically at 3.8A RMS. You need a capacitor rated for at least this much ripple current, and that's typically a "transmission type" capacitor, such as a porcelain chip capacitor, or at least a very high quality NP0/C0G chip type. If you use a trimmer, it also needs to be able to handle that much current. A small cheap flimsy little trimmer won't do it, but a big, hefty,

sturdy one will have way too much stray inductance. You need a small, high quality trimmer made for transmitter applications. Which is why experienced designers try to avoid using trimmers in such places.

If you would like to increase the loaded Q, for better harmonic suppression, there are several networks that allow doing it. The most common of these is the LCC network, meaning one coil and two capacitors. It looks exactly like the circuit we already have, and the only change is that instead of using a large value DC blocking capacitor (10nF in my example), we reduce the size of this capacitor to make it an important part of the matching network. To design this network, you pick the value of loaded Q you want, which must be higher than the Q the L network would have, and then you use these equations:

$XL = Q \times Rlow$

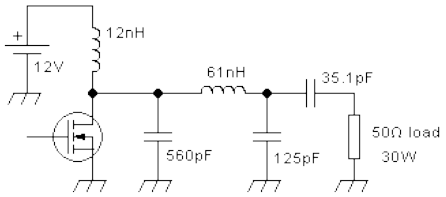
$XC1 = Rlow (1 + Q^2) \div (Q - \sqrt{ (Rlow (1 + Q^2) \div Rhigh - 1)}$

$XC2 = Rhigh \sqrt{ (Rlow (1 + Q^2) \div Rhigh - 1)}$

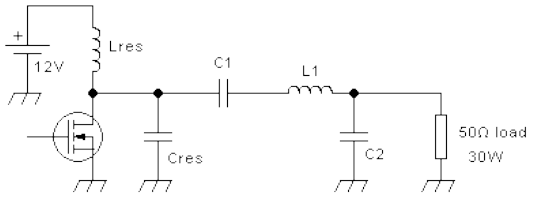
Applying these equations to our example of the 12V powered 30W amplifier for 52MHz, and setting the Q at 10 (which is higher than usual, but OK for this example), the inductor turns out to need a reactance of 20Ω, and thus an inductance of 61nH. The first capacitor needs to have 24.46Ω of reactance, which results in 125pF at this frequency. And the second capacitor, the one that was just a coupling capacitor when using an L network, now needs to have a reactance of 87.2Ω, and thus a capacitance of 35.1pF.

Hint: When using equations like the ones above, be **very** careful about using the correct precedence of operations! While that's just sixth grade maths or so, it's still easy to make a mistake.

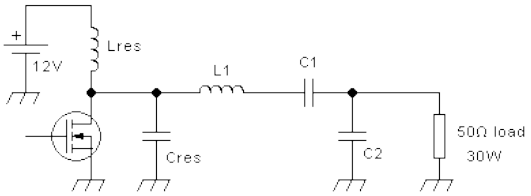
Our amplifier's output section, with the loaded Q changed to 10, ends up looking like this:



Another way to raise the Q is to place a series resonant circuit in series with the inductor of the L network. The two inductances that end up in series can of course be melted into a single inductor of the total inductance value. The circuit then looks like this:



or like this, whichever you like best:



The design equations for this network are the following:

$XC1 = Q \times Rlow$

$XL = \sqrt{ (Rlow \times Rhigh - Rlow^2)} + XC1$

$XC2 = Rlow \times Rhigh \div (XL - XC1)$

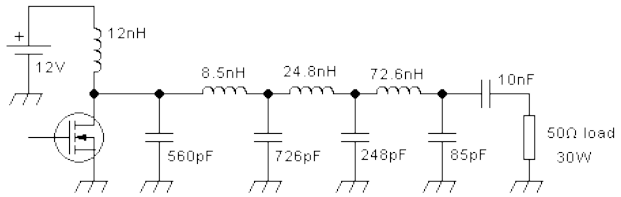
So, for our 12V 30W 52MHz amplifier, designed with a Q of 10, XC1 is 20Ω, XL is 29.8Ω, and XC2 is 10.2Ω. This equates to 153pF, 91.2nH, and 300pF. Lres and Cres of course are resonant with each other at the operating frequency, with a Q high enough to maintain a reasonable sine wave, and low enough to not limit the bandwidth too much and to have practical values, and Cres includes the transistor's output capacitance.

The -3dB bandwidth of these networks with Q = 10 will be only 5.2MHz, so they are really a bit too narrow to cover the entire 6 meter band with near-flat response. Q factors as high as 10 are rarely used in solid state amplifiers. They were more typical in tube amplifiers, where a high Q was necessary to allow the high impedance ratios needed there, and to absorb the sky-high output capacitance of tubes. But tube amp users had to pay for that by tuning up their transmitters to the specific frequency they wanted to transmit on.

In solid state practice it's more common to try achieving **larger** bandwidth, be it to cover a wide band without retuning, such as 88 to 108MHz for a frequency-agile FM broadcast transmitter, or be it to mass-manufacture a design without having to individually tune up each unit. More broadbanded matching networks are less critical in terms of component tolerances and accurate placement on the board.

There really aren't too many options when it comes to making low Q tuned matching networks. The best approach at HF and low VHF is probably cascading several L networks. In this case each L network has a much lower Q than a single L network, and the combination of all of them has a higher Q than the individual L networks composing it, but lower than a single L network that does the entire transformation.

Let's try modifying the amplifier to use this approach, with three L networks cascaded. Each of them will transform over the same resistance ratio. Since the total ratio is 1:25, each L network needs to have a ratio of 1: 2.924, because the cubic root of 25 is 2.924. And each L network will then have a Q of 1.39. So, our first L network matches 2Ω to 5.85Ω, the second one matches 5.85Ω to 17.1Ω, and the third one matches 17.1Ω to 50Ω. Applying the equations for the L network to each stage, the circuit ends up like this:



The -3dB bandwidth of this circuit is nearly 3 times wider than for the simple L network. And the -1dB bandwidth, which is a more useful characteristic for power amplifiers, improves more than threefold, from around 10MHz to 33MHz.

The inductances in such a circuit get so small at VHF that they are usually implemented as a single microstripline on the printed circuit board, with the capacitors soldered to that strip at strategically selected locations, to implement the proper inductance between each capacitor and the next one. Sometimes the width of the microstripline is adjusted for the different inductance values: It starts very broad at the transistor, and then narrows down after each capacitor. At HF instead discrete coils can be used, but it's not very common to use this sort of wideband tuned amplifier at HF, because single HF bands are usually narrow enough to use a simpler matching section, while to cover several bands one needs a true broadband amplifier anyway. Instead a very common application for such a network is a commercial VHF radio covering 136 to 174MHz.

The tuned matching networks that I have described here just barely scratch the surface, but tend to suffice for typical practical amplifiers. There are many other possible configurations, including Pi and T circuits. But not all of those lead to practical component values. And there are many web pages devoted exclusively to impedance matching, and some of them include online impedance matching calculators, so if you are particularly interested in tuned impedance matching, you should search and read some of those pages.

One more comment is due at this point: The data sheets of RF power transistors give the required load **impedances**, not just load resistances, because a typical transistor does need some reactance in the load at most frequencies, to compensate for its output capacitance and for its lead and bonding wire inductance. It's a simple matter to take the series-expressed complex impedance value from the datasheet, design your matching network for the resistive part of that impedance, and then add the required reactance in series with the input of your network, by either reducing or increasing the first inductor's value by the appropriate amount, depending on wether the transistor needs a capacitive or inductive component in the load impedance.

People who are too lazy to do these maths will simply design an approximate matching network, and make two of its elements adjustable. Then they will tune up their amplifiers for best power output and efficiency, by adjusting trimmers or deforming coils. In the case of amplifiers for HF, it's often practical to use coil formers with adjustable slugs in them.

All of these tuned output sections are suitable for use with all traditional classes of operation: A, AB, B and C. Instead the single-transistor broadband configuration is really good for class A only.

Class E

No, don't worry, I do know the alphabet, and I haven't forgotten class D. But class D amplifiers most commonly use two transistors rather than one, and we aren't there yet, so I will treat them later. Class E amplifiers instead are commonly built around a single transistor, and look extremely similar to the tuned amplifiers we were just busy with, and so it makes a lot of sense to treat them now and here.

Class E amplifiers were developed to improve the efficiency beyond that of a class C amplifier. They always operate with the transistor driven into full saturation and full cut-off, either by using a squarewave drive signal, or by strongly overdriving them with a sinewave. This is the first step in achieving high efficiency, because a fully saturated transistor can only cause a small conduction loss, and a transistor that is off cannot cause any significant loss at all. Class D amplifiers do the same. But there is a big problem with class D: Transistors simply cannot switch fast enough to keep the transition time insignificantly small, at radio frequencies. Class E tries to eliminate most of the switching loss by using an output network that shapes both the drain voltage waveform and the drain current waveform, trying to achieve a drain voltage as close to zero as possible while the transistor is switching on, and trying to keep the drain current as close to zero as possible while the transistor is switching off.

The circuit required to do this is almost the same as in the amplifiers we were just looking at. The main difference is that since we will be controlling the drain voltage waveform with our matching network, we don't want any other resonant circuit at the drain. So we replace the resonant inductor that feeds the supply current to the drain by an RF choke, which forces a constant supply current over the RF cycle, allowing strange and even abrupt drain voltage waveforms.

Did I say "RF choke"? Yes. And while discussing class A broadband amplifiers, I talked just about chokes. Time for another Blue Block:

Chokes and RF chokes:

A choke is an inductor of such a high inductance value that it blocks all alternating current that would want to flow, as much as possible. That is, a choke carries essentially a pure DC at all times, as far as imperfections and design compromises allow.

But a choke in a real-world circuit really doesn't need to fully block all frequencies from a femtohertz to a petahertz and beyond. Nor could we make such a choke. In practice, we just need the choke to block the frequencies that we want blocked. Sometimes this could be all frequencies present in a circuit, but at other times we might want a choke to block a certain (usually broad) range of frequencies, but pass another (very different) range of frequencies.

In RF amplifier work we are usually dealing with two such broad frequency ranges. One of them is radio frequencies, usually spanning from the lowest frequency of operation, to the highest important harmonic of the highest frequency of operation. For an HF amplifier operating from 1.8 to 54MHz, the RF range to block might span 1.8

to 500MHz or so.

The other frequency range we must not forget is the envelope frequencies. That is, the frequencies contained in any amplitude modulation of our signals. In an AM signal, this range is directly the audio frequency range our transmitter passes. In SSB the range is wider, theoretically infinite, but from a practical point of view the important components extend to about 10 times the highest audio frequency our transmitter passes. Even a Morse code signal contains a significant modulation bandwidth, which must be cleanly processed by the amplifier to avoid deformed keying and the resulting key clicks.

In RF power amplifiers usually all chokes need to block the RF range. We rarely if ever use chokes that deliberately have to block the modulation frequencies, since such chokes would be big, heavy, expensive, and inefficient. But for chokes in some positions it's irrelevant whether or not they block the modulation frequencies, while for others it's essential that they don't block them! In my opinion it's important to be aware, during amplifier design, whether a choke must freely pass modulation frequencies, or whether it doesn't matter how well it passes them. If you use a choke that noticeably affects the higher modulation frequencies, in a location where the choke needs to pass them, the choke will distort the transmitted signal! Many people have run into distortion trouble they don't understand, due to using a choke that has too much inductance and thus affects the higher modulation frequencies.

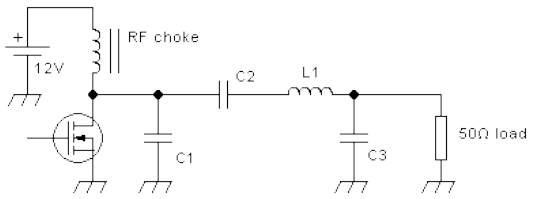
For this reason, throughout this article I'm making an intentional difference between chokes and RF chokes. When I use the term "choke", I mean an inductor that reasonably blocks all RF current, without you nor me needing to care whether or not it also has a noticeable effect on modulation frequencies. Instead when I write "RF choke" I mean an inductor that has enough inductance to block RF currents well enough, while having low enough inductance to pass all modulation frequencies without any noticeable impact on them.

There is another point about chokes that I would like to briefly mention: Many chokes are far from having a pure inductance. Instead they have an impedance, that is, inductance in series with resistance, and both the inductance and the resistance change with frequency. Some chokes are almost purely resistive at the higher frequencies. This behavior is fine, as long as the resistance is high enough to keep the losses acceptable. After all resistance dissipates power, while inductance does not.

It is often even desirable for a choke to be more resistive than inductive, because this reduces the risk for resonances, reduces phase shifts, and thus contributes to amplifier stability. But of course any choke needs to have a low enough resistance to DC, and RF chokes also need to have a low resistance to all modulation frequencies too.

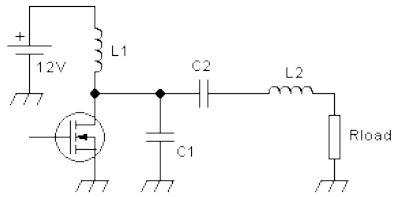
The amount of impedance a choke needs to have, of course depends on the circuit impedances. A typical rule of thumb is making sure that the choke has at least 10 times as much impedance as the circuit it's connected to, at all frequencies in the range it has to block, but specific applications might need even better choking, while others can work well with a significantly lower choking impedance. Typically if the choke's impedance is mainly resistive it needs to be higher than if it's mostly inductive, to keep the losses at an acceptable level. A factor of 100 is more common than 10, for a mostly resistive choke.

A typical, practical, class E amplifier's output section will look like this:



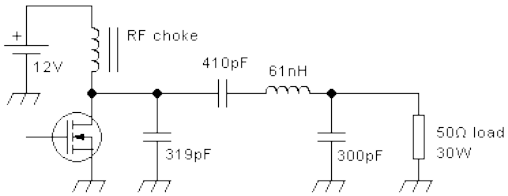
No big surprises there, huh? You might ask what's new at all! The only difference that can be seen, between this class E amplifier and a typical tuned amplifier that can be run in class A, AB, B or C, is the use of an RF choke instead of a small tuned inductor. Even this difference isn't absolute, because class E amplifiers can be designed to work with a small tuned inductor there too, although they work the cleanest way when an RF choke is used, that is, the current supplied to the circuit is forced to be a pretty smooth DC over the entire RF cycle.

The main difference is in the way to calculate the component values. The design equations for a class E amplifier are pretty complex, and that's why I won't copy them here, and rather suggest that you use any of the several online calculators and design programs that exist for class E amplifiers. Most of these calculators, and also the equations given on websites, are applicable to a textbook circuit like this:



You can punch the values you want for voltage, frequency, and power into one of those online calculators, including the value you pick for L1, and get values for C1, C2, L2 and Rload. If you pick a relatively large inductance for L1 the circuit will work better. A very low value for L1 might make you end up with the need for negative capacitances or negative inductances somewhere... And yes, Rload is calculated too, and would be 2Ω or very close to it, if you try our exercise of a 12V 30W 52MHz amplifier. If you need a different load, you have to add a matching section to that circuit. The simplest one is the L, of course, so you calculate an L circuit to match 2Ω to 50Ω and add it, and since that makes you end up with two inductors in series, you can merge the two into a single inductor that has the total inductance value.

Doing this for our 12V 30W 52MHz example, choosing a Q of 5 for the basic class E network, and using an RF choke that has high impedance at 52MHz, makes us end up with these values:



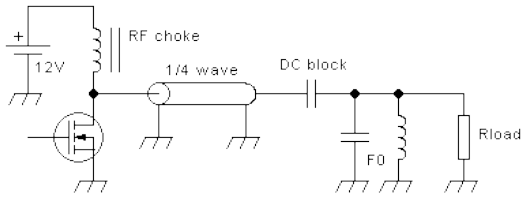
Don't forget that you need to drive this transistor with a squarewave, or with a pretty strong sinewave. You will be rewarded with a high efficiency, that could be as high as 90% if you use a good transistor, but you also have to be aware that the normal voltage peaks at the drain will be at least 3 times the supply voltage, and during transient conditions could be even higher. So you need a transistor with a higher voltage rating in this circuit, than if you use a very similar looking circuit operating in class A, AB, B or C with a sinewave voltage at the drain that doesn't even reach twice the supply voltage.

Many designers have tried to make a class C amplifier, for example for an FM transmitter, and then have empirically tuned and tweaked the matching network, while putting nice strong drive to the transistor, until reaching efficiencies above 80%. In fact they ended up with class E amplifiers, without even knowing it!

Since class E amplifiers operate in saturation, any amplitude modulation of them, even on/off keying for Morse transmission, should be done by modulating the supply voltage with a suitable waveform, while keeping the drive signal going.

Class F

In order to get higher power from a given transistor, without exceeding its drain voltage rating, someone came up with the idea to shape the drain voltage waveform into a square wave, by designing an output network that offers a short circuit to all even harmonics, but an open circuit to all odd harmonics. This was done by placing a parallel resonant circuit in parallel with the load, which offers a short for all harmonics, and then connect it to the transistor via a quarter-wave-long transmission line, like this:



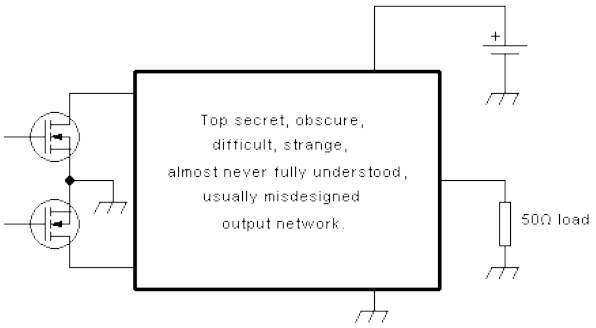
A quarter-wave-long transmission line has the interesting property that on all even harmonics it reflects the same load impedance it sees at its end, while on all odd harmonics, and the fundamental, it transforms the impedance it sees to its image value as seen from the line impedance. So, if we have a load resistance of 50Ω, a resonant circuit tuned to the working frequency, and a quarter-wave-long 10Ω transmission line, then at the working frequency the drain will see a 2Ω load. At all even harmonics it will see the short circuit offered by the resonant circuit. And at all odd harmonics it will see the short circuit of the resonant circuit transformed by the line into an infinite impedance, that is, an open circuit. So the drain voltage will become a square wave, and the drain current too. In theory it all works very nicely, and the efficiency should be very high, along with a higher power output from a given transistor, if the proper supply voltage is chosen. In practice however it's hard to get a transmission line at UHF to perform just right. It would need to be very thin, and then it can't handle high power. On HF instead such a line would be too long to be practical. And anyway, RF transistors don't like square wave voltages on their drains, because that causes large losses due to the need of charging and discharging the transistor's output capacitance at a very high rate, and due to internal parasitic BJTs in MOSFETs turning on at these very fast drain voltage transitions. So I don't see class F amplifiers as being very useful in the context of this article.

To make a class F amplifier of a sensible size for HF, instead of the transmission line one needs to use several parallel-tuned circuits (traps), each tuned to one odd harmonic, installed in series in place of the transmission line, and then one needs to add an L network for the impedance transformation. This tends to become too complex to be practical, specially considering that the network will only cover a single band, and the losses in the several tuned circuits tend to eat up any advantages the circuit might otherwise have. So it's not common to see class F amplifiers used for HF.

Broadband push-pull amplifiers:

Most HF amplifiers for radio purposes are broadband ones, not because any broadband signals are to be transmitted, but because operation on several bands is required. And given that single-transistor broadband amplifiers need to run in class A to avoid massive harmonic output, and class A is very inefficient, most broadband linear amplifiers for power levels above a few watt are usually push-pull designs, which provide better efficiency by running in class AB or class D, or a hybrid between those two.

The general form of a push-pull amplifier's output section is like this:



If you try to look up that output network in books about RF power amplifier design, you will hardly ever find anything useful, either because the book doesn't treat HF, or

because the author has no faint idea about the matter. If a book actually describes such a network, the description is usually copied from an older book or magazine article, and the book author copied it with all of the original faults and misunderstandings. If you look into transistor datasheets and application notes, you will find very few broadband push pull circuits, because the application engineers at those companies don't understand them either, get poor results when trying to build any, and the company then prefers to publish some narrow-band design that works better and thus puts their transistors in a better light.

So, let's tackle this matter piece by piece. The method is: Divide and conquer. A very old and well-proven approach.

A push-pull output network needs to perform several functions. Some of them are the same as those a network for a single-transistor amplifier needs to perform:

- Inject the power supply to the transistors.

- Transform the load resistance to whatever value is needed between the drains, for the available supply voltage and the required output power.

- Block the DC from reaching the load.

In addition, a push-pull output network has some additional tasks:

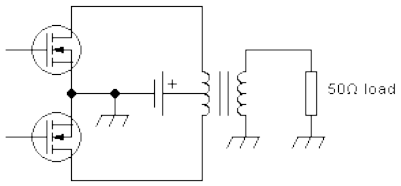
- Act as balun, to transform the single-ended load into a balanced load between the drains.

- Tightly couple the drains together, in phase inversion. This function is required for class AB and voltage-switching class D operation. In class A it's not required, but doesn't hurt if present. For current-switching class-D operation it must be absent.

- Restore a sinewave signal, through bandpass or lowpass filtering. This is required in class D, but optional in class A and AB, depending on whether the amplifier can provide the required harmonic suppression just by operating very linearly. Most often it cannot...

- Properly handle or suppress common-mode signals. This may or may not be necessary, depending on things like the type of feedback employed, if any, the drive method, and the linearity of the transistors. I will treat this point when getting to driving and feedback circuits.

A very simple and convenient way to implement all of these tasks, except filtering, is to employ a conventional transformer having a center-tapped primary:

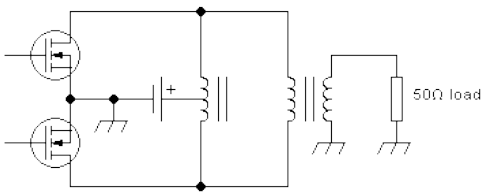


Since this circuit provides drain-to-drain coupling, but no filtering, it's suitable for class A and class AB. If a filter is added, it can be used for voltage-switching class D too. It could then also be used for class C, but that's rarely done. And class B is theoretically also possible, but as explained way above, real-world circuits can never truly work in class B.

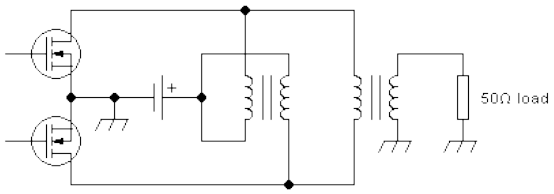
The big problem with this nice and simple circuit is that its practical implementation gets ever harder as the frequency gets higher, the bandwidth gets larger, the power gets higher, and the supply voltage gets lower. This is caused by the shortcomings of real-world transformers, which never have perfect coupling. As the frequency gets higher, the transformer needs to get smaller. But smaller transformers might not offer enough inductance for operation on low frequencies, might not handle the RF voltages without excessive core heating, or might not allow using thick enough wire for the currents that flow. At a given power, a higher supply voltage makes leakage inductance less critical by a square factor, but the core needs to be much larger too, or more turns need to be used, both of which increases the leakage inductance and also the parasitic capacitances.

If we had really good magnetic core materials, life would be much easier. But with the materials available today, it's hard enough to make such a transformer that works well from 1.8 to 30MHz in a 12V 10W amplifier, and I wouldn't know how to reasonably build such a transformer for a 1.8-54MHz, 12V 100W amplifier. Let alone a 1.8-54MHz, 50V, 1.5kW one! Many 100W output stages use a conventional transformer that has a two-hole core and a single-turn primary, but this is **not** the correct type of transformer for this circuit, because it offers no coupling whatsoever between the halves of the primary. I will get to that transformer, and its legitimate uses, just a little further down this page.

One way to make the situation more manageable is to split up the functions of the output network between two magnetic devices. One of them is a thing sometimes called a "bifiliar choke", also called "balanced feed transformer", and several other names, which injects the power supply and couples the drains. The other device is a conventional transformer that performs the load resistance matching, isolation and balun functions. The circuit looks like this:



Some people prefer to draw the feed transformer in some other way that suggests its bifiliar construction, such as this:

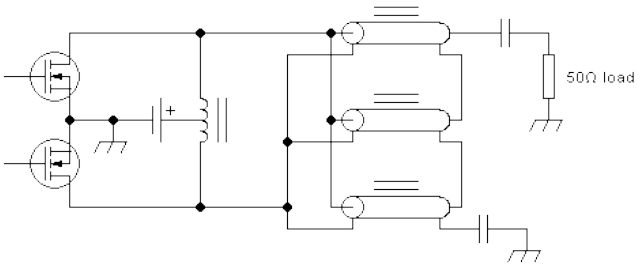


But I feel that this looks like scratching my left ear with my right hand, so I prefer the other drawing style, which also suggests very well that this feed transformer performs some of the same functions as a center-tapped primary winding.

Splitting up the functions between two transformers makes the task easier, but still not really easy. In amplifier configurations requiring drain-drain coupling, it's still hard enough to make a bifiliar feed transformer that couples well from 1.8 to 30MHz, and very hard if you want to get to 54MHz, as soon as the power level of the amplifier is higher than about 10W at 12V, and 100W or so at 50V. It's common to see ferrite toroids wound with 5 to 10 bifiliar turns, but these suffer from so much leakage inductance that they don't properly couple the drains in the higher part of the HF range.

By the way, a twin-hole core with a single-turn primary is perfectly fine as the impedance-transforming, balancing, and DC-isolating device, when combined with an effective bifiliar feed transformer - within its frequency limitations, of course.

Instead of a conventional transformer it's possible to use transmission-line transformers to do the load resistance matching and the balun action. An output section using a TLT with a 1:9 impedance ratio might look like this:

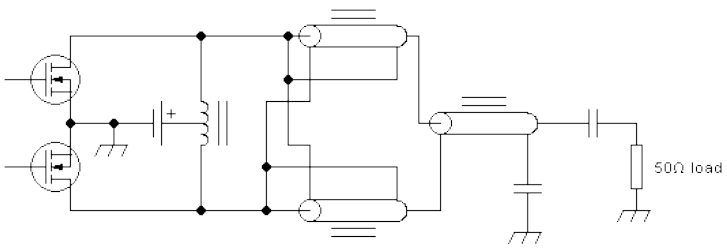


This circuit places a load resistance of 5.56Ω between the drains, and also performs balun action. The true transmission line transformer has a very wide frequency response, so one can pretty much stop worrying about its upper frequency limit. Each of the three transmission lines needs to have a 16.7Ω impedance. The three lines in parallel give 5.56Ω at the input, and the three lines in series give 50Ω at the output. If lines of a wrong impedance are used, the transformer will still work at low frequencies, but will become increasingly bad as the frequency rises, because the lines will start doing some unwanted impedance transformation on their own.

The uppermost line works at the highest end-to-end common-mode voltage, so it also places the largest stress on its core. It should either have the largest core, or the most turns. The lowermost line has the lowest choking requirements, so it can use the fewest turns or the smallest core. Since the lines should all have the same length, to maintain high frequency performance, the best economical compromise is to use the same length of line for each of the three, but use different number or sizes of cores in each.

Since a TLT doesn't provide DC blocking, suitable blocking capacitors have been added. I put them on the output side, where the current is lower.

Sometimes a different configuration is used to achieve the same purpose. The 1:9 transformer is implemented in a bootstrapped configuration, saving one transmission line and its core. But then the transformer doesn't provide balun action (common-mode current blocking), and needs a separate balun (common-mode choke) added! The circuit looks like this:

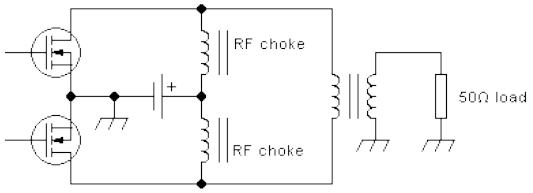


The two lines of the impedance-matching transformer now work at the same end-to-end voltage, and thus are made identical. The balun works at a higher end-to-end voltage and thus needs a larger core or more turns.

Note that "larger core" in this context means a larger ferrite cross-sectional area.

One could ask what point there is in replacing a true transmission line transformer using three lines and three cores, by a bootstrapped version that needs the same number of lines and cores but has a definite and rather low upper frequency limit. The explanation is that the latter version needs only 7/9 as much total ferrite cross section × turns product, as the former one! So, when every penny counts, the latter version is cheaper, and might get the job done, if the required top frequency isn't too high. But often it's impractical to take advantage of this, and designers choose the simpler solution of using only one type of core. Also the former version has the flatter frequency response.

Now that we have a good solution for a very wideband load matching system, it's time to look once more at the remaining trouble spot: The coupling between the drains. Since it's too difficult in practice to make split-primary transformers that produce the required excellent drain-to-drain coupling, and in most cases with mid and high power amplifiers operating to 54MHz it's also very hard or impossible to make a bifiliar feed transformer that does this job, we are often forced by practical reasons to consider amplifier circuits that don't need any drain-to-drain coupling at all. For example, consider this output section:



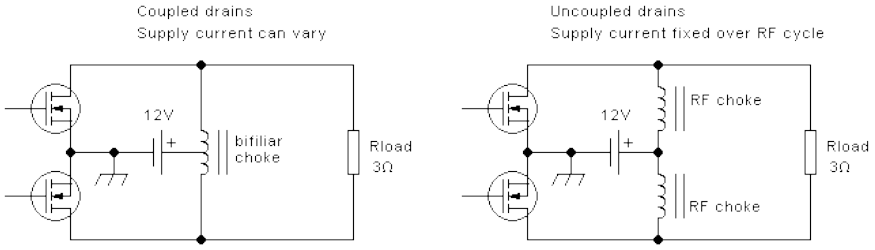
The power supply is now injected through two completely separate RF chokes. They have absolutely no coupling between them, as shown schematically by drawing two separate magnetic cores. Each of these RF chokes will force a constant current over the entire RF cycle. The current can change at modulation rates, though. Each drain can now have its own voltage excursions, unbound and unlimited by the other drain. This is much like having two single-ended broadband amplifiers, with the load connected between their drains, rather than from one drain to ground.

I drew this circuit with a conventional transformer at the output, for simplicity, but the following discussion applies regardless of whether conventional or transmission-line transformers are used for load matching. In either case the load appears differentially between the drains.

Such a configuration works well in class A, and in current-switching class D. It cannot work correctly in class AB, class B, class C, and voltage-switching class D. This is because those classes require the total supply current to vary over the RF cycle, which cannot happen when there are uncoupled and un-bypassed RF chokes in series with it. Instead class A and current-switching class D amplifiers work at a constant supply current over the RF cycle. Actually class A works with constant supply current at all times, so it could even use wideband chokes instead of RF chokes, if that was practical.

At this point it's important to understand the voltage and current waveforms happening in a push-pull amplifier, in each operating class. For this purpose let's consider two output circuits that have the load connected directly between the drains, so it's easiest to imagine and analyze what's happening. Having a transformer-coupled load behaves the same, but needs to consider the transformation ratio in all calculations.

One of these circuits has drain coupling, the other does not:



When operating in class A, using perfectly linear transistors, each of the transistor will conduct a DC (identical for both), plus the full-wave RF current (in opposite phases). In this example, let's assume that we set the DC at 8A per transistor. So, at the zero crossing, each transistor conducts 8A. The bifiliar choke, or feed transformer, conducts 8A in each winding, and since they are in opposed directions, there is no flux in its core. Both drains are at 12V, and the load sees no voltage and no current.

At the peak of the signal, when the amplifier is driven pretty hard but not into saturation, one transistor is being driven to conduct 15A, the other instead is only driven to conduct 1A. Remember that transistors basically behave as controlled current sources. The bifiliar choke cannot suddenly conduct 15A in one winding and 1A in the other, because that would make the net current in it rise from zero to 14A in a quarter of one RF cycle. So, for every ampere the current increases in one winding, it also has to increase the same amount in the other! The bifiliar choke forces this behavior. Also, due to the coupling between the windings in opposed phase, and the fact that the midpoint is held at constant voltage by the power supply, and by bypass capacitors in a real circuit, whenever the voltage at one end rises, the voltage at the other end must drop by the same amount. So this bifiliar choke or feed transformer forces the two drain voltages to maintain symmetry around the supply voltage.

The final result is that the current difference will have to flow through the load resistance, since this is the only available path. The bifiliar choke still conducts 8A per side, and on the side where the transistor is driven softer and conducts only 1A, the excess 7A flow into the resistor. At the other end of the resistor these 7A come out and add to the 8A coming from the bifiliar choke, and the transistor on that side conducts the resulting 15A. The 7A in the 3Ω load resistance cause 21V drop, and the bifiliar choke distributes this drop evenly around the 12V. So, the transistor conducting 15A sees its drain voltage drop to 1.5V, while the other transistor sees its drain voltage rise to 22.5V.

Of course during the opposite half cycle the transistors switch roles.

1.5V is probably as close to saturation as one can get without excessive distortion, so this represents the maximum undistorted power output for this amplifier. 21V peak equals 14.85V RMS, and with the 3Ω load chosen in this example, the power output is 73.5W. The supply current stays constant at 16A all the time, so the input power is 192W, and the efficiency is 38%, typical for a good class A amplifier.

Now consider the same class A operation with the circuit that does not couple the drains. Since the supply current is constant in class A, this circuit happens to work exactly the same! There will be the same currents and voltages as with the drain-coupled circuit, although the two separate chokes will each run with a considerable flux density in their cores, because they conduct a big DC. But there is another difference too: In this circuit there is nothing forcing voltage symmetry around the supply voltage. If there is symmetry, it's caused purely by the transistors being identical, the drive being perfectly symmetric, and everything else also being symmetric. Any asymmetry caused by real-world tolerances will manifest itself in different, asymmetric voltage waveforms at the two drains, while the circuit with a bifiliar choke will largely suppress them.

Now let's consider the nonlinearity of real transistors. Specially when using MOSFETs, the gain is much lower at low currents than at high ones. So, if the transistors are **driven** symmetrically, they will not **conduct** symmetrically. When one is driven to 15A , the other one might still be conducting 3A instead of the 1A it should. So the two of them will conduct a total of 18A instead of the desired 16A. But at the zero crossings each will conduct 8A, for a total of 16A. So the total supply current will want to fluctuate between 16A and 18A, at twice the operating frequency. And this opens a whole new world of complications and trouble!

It's not really terrible when we are using a circuit with good drain-to-drain coupling. In that case simply the amplifier will consume a supply current with this small 2A ripple on it. The drain voltages will stay symmetric around the supply voltage, because the bifiliar choke forces it. At the signal peak, one transistor conducts 15A, the other 3A, the bifiliar choke conducts 9A in each winding, and since the currents in both windings change simultaneously but go in opposite directions, there is no restriction to how fast they can change. On the side where the transistor conducts 3A, the 6A difference gets injected into the load resistor, and on its other end this current adds to the 9A of the bifiliar choke, giving the 15A the transistor there can conduct. The voltage drop on the resistor is only 18V, so one drain is at 3V and the other at 21V. Obviously the amplifier isn't delivering its full power, so we can drive it a little harder to restore the correct output power despite transistor nonlinearity. And the most beautiful thing about class A is

that the lower instantaneous gain of one transistor will be partially compensated by the higher instantaneous gain of the other, so that the overall linearity of this amplifier will be better than the linearity of the individual transistors.

But when you use the circuit with the uncoupled drains, the separate RF chokes will apply whatever brute force is needed to keep a constant current flowing during the RF cycle, in each of them. They absolutely refuse to let the supply current fluctuate between 16A and 18A at an RF rate. And since there is no path to ground in the circuit except through the transistors, this forces the transistor pair to conduct a constant total current over the RF cycle, no matter what happens. The effect of this is that around the peak of the signal the transistors want to conduct more current than the chokes are supplying, and the common-mode drain voltage drops. That is, the transistor that was pulling its drain low, pulls it even lower, and the one that was allowing to let its drain voltage go up, limits this excursion. This process is limited by that fact that the transistor with low drain voltage will saturate, and at that point it no longer conducts all the current it could conduct, limiting the total drain current to the total choke current. But this common-mode drain voltage depression makes the RF chokes slowly pick up a larger current over many cycles, and then we get a real problem around the zero crossings, because at that point of the waveform the transistors want to conduct less current than the chokes are forcing into them, and the drain voltages on both of them will soar up - as high as necessary to **force** them to conduct the current! High enough to make them go into avalanche conduction, and even to destroy them, if necessary.

Now enter class AB. In this class, the difference between the two circuits becomes much worse! All is fine as long as we have perfect drain-to-drain coupling. In such a circuit, class AB works like this: At the zero crossing, both transistors conduct only a small current. Say, 1A each, in this example. At the signal peak, one transistor conducts 14A, the other one is completely off. With the bifiliar choke forcing the currents in both of its windings to be identical, there will be 7A in each winding. On the side where the transistor is off, the whole 7A flow into the load resistor. The outflowing 7A join with the other 7A coming from the bifiliar choke, and the transistor that's on will conduct these 14A to ground. The voltages and output power will be just like in the class A example. But the input power is much lower, because the supply current fluctuates between 2A at the zero crossing and 14A at the signal peak, with a somewhat distorted half-wave sine shape, producing an average current of roughly 9.5A, and thus an input power of 114W, for an efficiency of around 65%, typical for a good class AB amplifier.

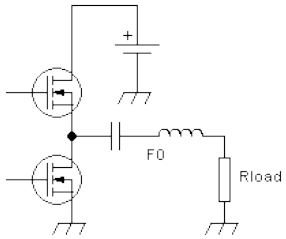
Note that this amplifier has a very large RF ripple in its supply current, requiring really good, high current, low impedance bypass capacitors at the feedpoint, with a very short and wide path to the sources of the transistors. Implementing this is a real problem when constructing a practical amplifier with such a low drain load resistance. By the way, a 3Ω drain-to-drain load is required in 14V 100W amplifiers, and also in 40 to 50V amplifiers operating at the typical ham radio legal limit of 1 to 1.5kW.

If you try to build a class AB amplifier that doesn't have good drain-drain coupling, all hell breaks loose! Over much of each half wave, around the peaks, one transistor is in saturation, the other one showing "strange" waveforms that vary from band to band. And around the zero crossings of the drive signal there are massive drain voltage spikes that are guaranteed to make the transistors break down in avalanche mode, and very likely the transistors will eventually be destroyed. Class AB doesn't work when there is no drain-to-drain coupling! You can make an amplifier work in this configuration, but it will not be in true class AB, but in some sort of hybrid mode between class AB and class D, and the output current waveform will be trapezoidal instead of a sine wave.

Since true class B is impossible in practice, and class C is hardly ever used with broadband push-pull amplifiers, I will skip them. Instead I will finally get to class D! Phew!

Class D amplifiers:

These are the closest thing to switchmode amplifiers. According to textbooks, voltage-switching class D amplifiers have a configuration like this:

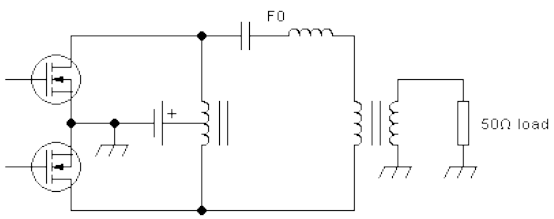


The two transistors are switched on and off by two phase-opposed square waves. The voltage at the mid point switches sharply between ground and the supply voltage. A series resonant circuit at the output offers high impedance to all harmonics, while conducting the fundamental signal to the load. So the current flowing in each transistor is a half sine wave, while the voltage is a square wave.

A transformer or tuned matching network can be inserted between the resonant circuit and the load, to match to the desired load resistance.

This particular configuration is extremely common in switching power supplies operating at frequencies between 20kHz and maybe 1MHz, and even a little beyond. But at HF it's hard to implement, because the upper transistor needs to be driven relative to its source, which isn't grounded, but instead swings up and down all the way between the supply voltage and ground. Driving such a configuration at tens of MHz incurs in some severe trouble with stray capacitances, and also most transistors suitable for such frequencies come in casing styles intended to be operated with a grounded source.

So, an alternative configuration has more chances of being successful at RF:



Here it is again - the good old push-pull circuit. In this incarnation it has coupled drains, and a series-resonant circuit inserted somewhere in the load circuit. If inserted between the transistors and the transformer, like shown here, the transformer only sees the fundamental frequency. If instead the resonant circuit is inserted after the transformer, it's easier to band-switch, but then the transformer will be exposed to the full harmonic frequency voltage, and this requires the transformer to be designed to handle that, both in terms of core loss and stray capacitances.

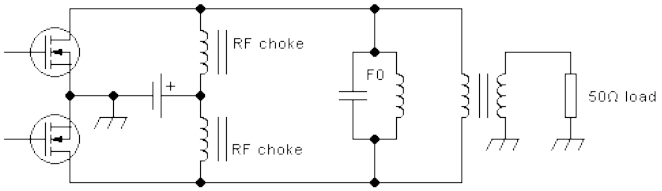
In this configuration the transistors switch their drains between ground and off, and while off, the coupling in the bifiliar feed transformer makes the drain voltage rise to

twice the supply voltage. The drain current conducted by each transistor is a half sine, defined by the resonant circuit. The supply current is a rectified sine.

Instead of a series resonant circuit, it's possible to use a series-input low pass filter. This could allow building a pre-tuned bandswitched amplifier. But the big problem with this configuration is achieving good drain-drain coupling in the bifiliar feed transformer. It has to be good all the way to pretty high harmonics! And that, my dear readers, just can't be achieved in high power amplifiers. Take my word for it, or try it yourself. Let me know if you have success. The higher the power level, the lower is the frequency limit.

And this brings us to current-switching class D amplifiers. These actually **need** separate RF chokes that conduct a constant current over the RF cycle, and they **need** the drains being completely decoupled from each other! So, using current-mode class D turns a very difficult design problem into a very easily implemented design feature!

Such a current-switching class D amplifier's output section might look like this:



Like in the previous case, each transistor switches its drain between ground and off. While at ground, it conducts the total supply current, equal to the sum of both RF choke's currents, and constant throughout the RF cycle. While off, the parallel resonant circuit forces the drain voltage to be a half sine, starting from ground, not from the supply voltage. Assuming perfect transistors that have zero saturation current and instantaneous switching, the voltage across the RF choke on the side where the transistor is on, is the supply voltage. Since that voltage is constant during the half cycle, the peak, average and RMS voltages are all the same. While that transistor is off, the RF choke must restore its DC balance, by forcing an average voltage during the other half cycle that's also equal to the supply voltage. So the half sine voltage at the drain will have an average value of twice the supply voltage, starting from ground, and for a sine wave this is an RMS voltage of 2.22 times the supply voltage, and a peak voltage of 3.14 times the supply voltage! **This is a very important point to be aware of!** While in well-implemented class A, AB, B linear amplifiers, and also in voltage-switching class D amplifiers, the transistors only ever see drain peak voltages of twice the supply voltage under normal operating conditions, in a current-switching class D amplifier they see 3.14 times the supply voltage, with perfect components, and roughly 2.8 to 3 times the supply voltage in a typical real-world implementation. So, a transistor rated for 50V supply voltage, and 125V absolute maximum drain voltage, is fine to be used at 50V in true class AB, but in current-switching class D it should be operated at only 33V to get the same reliability. Operating at 50V in current switching class D it will produce much more power output than in true class AB, but at the cost of much higher peak drain voltage that might well exceed the transistor's absolute maximum rating, and destroy the transistor! I mention this so explicitly because many people trying to build class AB amplifiers end up unknowingly building AB-D hybrids, due to poor drain-drain coupling, and kill transistors like crazy, without understanding why the darn things blow up.

Instead of the parallel resonant circuit, this amplifier can be operated with a parallel input (π configuration) lowpass filter. If the filter is placed after the transformer, the full harmonic current needs to go through the transformer. The voltage instead is close to a sine wave, thanks to the filter, so the core sees no additional stress from harmonic frequency voltage.

True class D amplifiers are usually operated with non-simultaneous switching of the two transistors. To avoid shoot-through and the corresponding high current pulse and loss, voltage-switching class D amplifiers are normally driven in such a way that one transistor switches off a small moment before the other one switches on. So there is a brief "dead time", during which neither transistor conducts. If this dead time is deliberately increased, the output amplitude gets smaller. This is pulse width modulation, and is often used with voltage-switching class D amplifiers at low frequencies, such as in power supplies and audio amplifiers (including modulators). In fact PWM is so common that some people think that voltage-switching class D is necessarily tied to PWM. But this is incorrect. It's perfectly possible to use voltage switching class D without PWM.

Current-switching class D amplifiers instead need to be operated the other way around: Each transistor has to be switched off a little later, not earlier, than the other one is switched on, in order to avoid generating a huge drain voltage spike during crossover, because if there was dead time, the RF chokes would be forcing current into the amplifier, that has nowhere to go. So the drive is arranged in such a way that during crossover both transistors conduct at the same time, and the RF chokes keep the current at the normal value.

In RF amplifiers it's not possible to get absolutely true class D operation, because transistors just don't switch instantly, and the switching time even of UHF transistors is still significant at HF. We can get close to true class D when using UHF transistors on the 160m band, but only close, not fully there. So, in real-life class D RF amplifiers, we never have precise square waves, but instead the current has a trapezoidal waveform with rounded corners. And we rarely get any true clean voltage sine wave, but only something resembling a sine wave, with obvious waveform distortion.

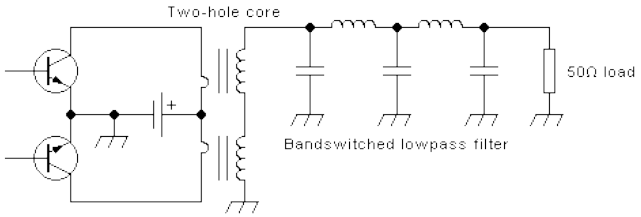
If the filter is placed after the transformer, there could very well be a significant delay between the transistors and the filter, and this causes the reflected harmonic components to be out of phase with the fundamental. As a result it often happens that on some bands we see an approximate square wave on the drains, but on other bands this changes into something closer to a triangle wave, or some other peaky shape. It's not that the harmonic structure has changed. Just the phase relations have changed! But this inconvenient phasing of the harmonic reflections can make the peak drain voltage higher than 3.14 times the supply voltage... and that's bad news for the transistors, and often explains why an amplifier that worked well for years on several bands blows up instantly when tried on a new band. A diplexer filter solves this by not reflecting any harmonics. The drain voltage waveform will then become trapezoidal. But since then the amplifier will produce significant power at harmonic frequencies, and this power will be absorbed in a dummy load placed at the diplexer's highpass output, the efficiency of such an amplifier is poor for a class D amplifier, making the approach rather pointless.

Linear class D amplifiers

If a current-switching class D amplifier is driven with a square wave of variable amplitude, then the two transistors will also conduct a square wave of variable amplitude, instead of saturating. The resulting amplifier output is a squarewave current, or in practical RF amplifiers more like a trapezoidal current, whose amplitude follows the drive signal. A capacitor-input low pass filter will force the voltage at its input to be roughly sinusoidal, and will largely eliminate harmonics from the output. So the configuration of this amplifier's output section is just like that of a current-mode class D amplifier operated in saturation, but it is capable of linear operation simply thanks to the driving scheme used.

And this, folks, is very important, because this is how many, if not most broadband push-pull linear HF amplifiers work! Most people believe them to operate in class AB, and are shocked when they see the waveforms present inside such an amplifier. But really these are linear class D amplifiers, and then the waveforms make sense!
To avoid shocking my dear readers too deeply, I'm willing to consider all those 13V, 100W push-pull amplifiers, that we have used in millions of radios since the 1970's, to be hybrid class AB-D amplifiers. But they are **not** true class AB linear amplifiers as Adam described them, even if widely publicized as such!

Such an amplifier's bare-bones output section typically looks like this:

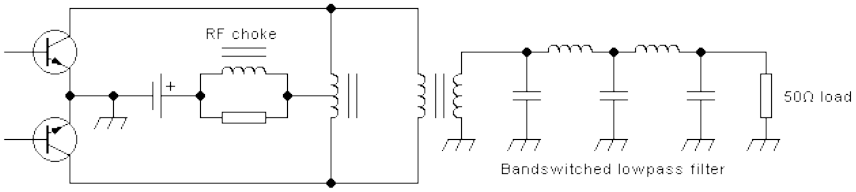


It uses an output transformer made on a two-hole core, with a single-turn primary fed at the midpoint. Since such a transformer has no magnetic coupling whatsoever between the two halves of the primary, I drew it as two separate transformers, connected in series. It behaves exactly like having two separate RF chokes followed by a simple transformer. That is, it provides no coupling between the collectors.

This circuit has been implemented with BJTs for about 40 years, and only later MOSFET versions appeared, which typically have some differences. BJTs allow a very simple trick with biasing and driving, to run them in linear class D. I will come to it in the chapter about biasing.

Most amplifiers in HF communications are used in SSB mode. The peak-to-average ratio in SSB is very high, so that an amplifier designed for 100W PEP will in average be producing only 20W or less output. A practical, well-implemented class AB amplifier might produce 65% efficiency at full power output, but at 20W it will be dramatically worse, and over the entire SSB cycle too. For this reason some linear class D amplifiers use a trick to improve the part-power efficiency. This trick is to **partially** couple the drains together, to achieve an operation mode that lies halfways between class AB and current-switching class D.

This efficiency tuning is most commonly done like this:



The bifiliar feed transformer would couple the drains together, if it had its midpoint at RF ground. But the RF choke lifts it from ground, and the resistor "partially grounds" it through the supply. The resistance value controls the amount of coupling between the collectors.

The resistor can also be connected from the bifiliar feed transformer's midpoint to ground, via a DC-blocking capacitor. But connecting the resistor across the RF choke is usually more convenient.

Back in 1977, Frederick Raab published the mathematical theory behind this optimization in the IEEE Journal of Solid-State Circuits, calling this configuration the "Class BD High Efficiency RF Power Amplifier". And now in 2020, and many million such amplifiers later, still many builders of RF amplifiers don't recognize it even when they are pushed with their noses into it!

Designing those pesky transformers:

In recent years many designers and experimenters have been building high power HF amplifiers using gemini-encapsulated MOSFET pairs such as the BLF188XR, either single or in pairs. To help those experimenters, here is some detail about the design process for that specific case.

The first step is deciding how hard we can load such a transistor. The datasheet states that the RDS(on) is typically 0.08Ω at 25°C. It rises with temperature, so at normal operating temperature it might be roughly 0.13Ω. And this is the value when the gate voltage is driving the MOSFET "fully on". The saturation isn't a knife-sharp process, and thus the practical saturation voltage in linear RF power operation is somewhat higher than the RDS(on) value would suggest. So, if we want to avoid an excessively severe efficiency reduction due to imperfect saturation of the transistor, we need to keep the drain load resistance above 1.3Ω or so.

In a push-pull circuit the drain load resistance on the transistor that is on at a given time is one quarter of the drain-to-drain load resistance. Some people stumble over this, and they believe that it should be half the drain-to-drain load resistance. That *would* be true if both transistors were on and working at the same time, such as in a class A amplifier. But not in AB, nor in class D. And those are the classes we can use with such a large transistor. So it follows that we want a drain-to-drain load resistance of at least 5.2Ω. If it is higher, the efficiency will be better, but the power output for a given supply voltage will be lower.

The impedance transformation ratios we can easily implement are 1:1, 1:4, 1:9, 1:16, perhaps even 1:25, corresponding to turns ratios of 1:1, 1:2, 1:3, 1:4 and 1:5. Since we usually want to match to a 50Ω load, a 1:9 transformer is the only reasonable choice, giving 5.56Ω drain-to-drain load. A 1:4 transformer would produce far lower output power than the transistor is capable of, while a 1:16 transformer would result in a very inefficient amplifier that needs a relatively low supply voltage in order to keep the dissipation manageable. So, 1:9 it is.

Should we use a conventional transformer, or a transmission line transformer? Let's try both, as a design exercise, and also to see which one has what advantages.

The first approach a typical designer tries is a conventional transformer using two ferrite tubes (sleeves, large beads), side by side, with a single-turn primary made from two little pieces of tubing and sheet metal or PCB end plates, and 3 turns of wire wound through the tubes to form the secondary. And the first questions that arise are: How large do the cores need to be, and what ferrite material can be used?

The core needs to fullfill at least two, and sometimes three requirements:

1: The single turn must provide enough inductance for operation on the lowest desired frequency.

2: The losses in the core must remain manageable, at the voltage the transformer operates at, throughout the entire frequency range.

3: If the power supply is applied to the (false) center tap of this primary, the cores need to stay well out of saturation, and inside their linear ranges, when the highest expected DC current flows. This requirement can be dropped if another feed method is used.

The minimum inductance required is a bit subject to negotiation. We might want it to be so high that it has negligible effect on circuit operation. That would be about 10 times the load on the winding. In this case, the single-turn winding operating into 5.56Ω, the inductance would need to have a reactance of at least 55.6Ω at the minimum operating frequency. If we want the amplifier to operate from the 160m band upwards, that minimum frequency is 1.8MHz, and the minimum inductance is then 4.9μH.

If this turns out hard to implement, we can start to negotiate. At half that inductance value, there will be some impact on performance on the lowest band, but not much, so we might just accept 2.5μH as a minimum value. And we can go even lower, if we compensate the transformer for low frequencies! This is done by inserting properly calculated capacitors in series with the primary and secondary, so that they form a T highpass filter with the transformer's inductance. Since the capacitor on the primary side will block DC, this method can only be applied if the supply current is not applied through the transformer. And we need to keep the cutoff frequency of this highpass filter well below 1.8MHz. This method allows us to get by with a primary inductance as low as 0.5μH! This would require a 33nF compensation capacitor in series with the primary, and a 3.6nF capacitor in series with the secondary. Some people see such capacitors in a schematic diagram and scratch their heads about why they are there, given that there is no need to block any DC... Well, they are low-frequency compensation capacitors, not DC-blocking capacitors!

You can use filter design software to design such compensation networks. "Filter Solutions", from Nuhertz Technologies, works very well and is very comprehensive. But much simpler software, online calculators, or even just the plain equations obtained from textbooks or the web, will also do the job.

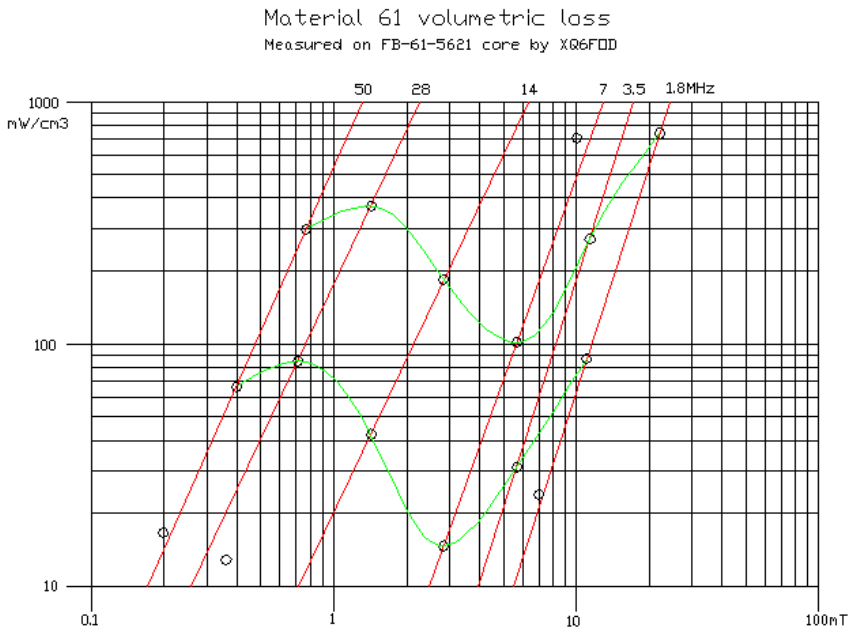
There is a catch, though: If the transformer's inductance is so low that we need to compensate it in this way, then the transformer will operate at increased voltage on the lowest band, and this will increase the core loss!

So, to round up the primary inductance requirement: If we can get 5μH or so, that's plenty. 2.5μH is still OK. Lower values can be used, down to 0.5μH, but require compensation and attention to increased core heating. And if you don't need to operate on 160m, so that your amplifier has a minimum frequency of 3.5MHz, then all these values can be halved.

The volumetric loss in a ferrite core depends in a nonlinear way mainly on the frequency, the volts per turn, and the core's cross-sectional area, and the curves are specific to each particular material. The core's shape also has some influence. So we cannot use equations to calculate the loss, but have to look up tables or graphs. Unfortunately it's very hard to find published loss data for ferrite materials, applicable to the operating conditions in RF power amplifiers. Ferrite manufacturers seem to shy away from publishing such data, for whatever reason. So I measured the loss on several cores, using home lab methods, and published the results in a web page called Ferrite core loss in HF power applications.

Now the question is how much volumetric loss is acceptable. This depends strongly on whether or not there is forced air cooling available for the transformer, and also on many other factors such as the size of the core (due to the Law of Sizes: Larger cores have more volume per surface area than smaller ones), how much heat the windings generate, what's the maximum acceptable temperature for the ferrite and the wire insulation, and so on. And the loss we need to consider is the average loss, so that in an amplifier destined purely for low duty cycle modes such as SSB we can use a much higher design value than in an amplifier intended for 100% duty cycle. So I cannot give you a simple, fixed value for acceptable volumetric loss. Just some very rough guidelines: For large cores like the ones we need in a high power amplifier, 200mW/cm³ is typically OK with forced air cooling in continuous service, and with natural convection cooling in intermittent SSB service. Smaller cores can be driven to higher volumetric losses.

Now notice how steep the curves are! For example, taking the first graph:



If you drive this material to a flux density of 10mT at 1.8MHz, the loss is so low that the core might pass for "cold" when you touch it, even on the worst band, 160m. If you double the flux density to 20mT, the loss becomes roughly 9 times larger, and unacceptable unless perhaps if you use low duty cycle *and* forced air cooling! So, despite the uncertainty we typically have in how much core loss is acceptable, looking at these loss graphs we can get a pretty good idea of what flux density we can use. In case of doubt, choosing a slightly lower flux density will make the loss drop a lot and give us headroom.

Note that with all RF-rated ferrite materials I measured, the highest loss occurs at the lowest frequency, when the core is driven at constant voltage from 1.8 to 54MHz, as is the case with typical ham RF power amplifiers. But some materials show a steady drop in loss toward higher frequencies, while other materials, such as Fair Rite 61, have rather idiosyncratic loss curves. The green curves in my graphs show loss at constant voltage excitation. With any RF ferrite, we need to calculate both the loss and the inductance for the lowest frequency, and the rest of the frequency range will be fine. But this is not true if you try to press low-frequency ferrite material into RF use!

So, looking at this graph, you can see that when using material 61 and designing for a flux density of 10mT at 1.8MHz will produce a low-core-loss transformer, while 15mT might be the sweet spot between core size and losses, and 20mT might be the absolute reasonable limit. If instead you use material 43, you need to design for slightly lower flux density, roughly between 7 and 15mT for the same range of volumetric loss. And material 31 is very close to 43, in this regard, just a tad better. If you use materials that I didn't measure, you need to find loss curves somewhere else. Good luck with that! You will probably have to measure the loss yourself. At least by making a practical test of how much drive your core can take at the lowest frequency, with acceptable heat rise.

Now let's start the maths. The BLF188XR has an absolute maximum drain voltage of 135V. So, whatever class your amplifier operates in, you need to set the supply voltage such that this is never exceeded. And for safety reasons, you should stay somewhat below it. So, 100V peak is fine, and we can design for this voltage, because it doesn't matter if the transformer heats up under conditions that anyway destroy your MOSFET! In true class AB, you get 100V peak with a supply voltage around 53V, and in current-switching class D you get it roughly with a 36V supply.

100V peak is 71V RMS, if we have a sine wave on the drains, which is what we should have both in class AB and current-switching class D, which are the main classes for such an amplifier. The required core cross-sectional area, for 71V RMS across a single turn, at 1.8MHz, for 15mT flux density, is:

$71V \div 4.44 \div 1800000Hz \div 0.015T = 0.00059m^2$

You can look up the equations in my article Transformers and Coils, and apply algebraic transformations to them as needed.

As you can see, everything is in basic units. Afterwards of course we can convert to more practical sub-units, and say that the minimum required cross-sectional area is $5.9cm^2$. This would give us operation around the "sweet spot" with 61 ferrite, and hot-but-perhaps-managable operation with materials 43 or 31. With slightly larger cross-sectional area, materials 43 or 31 would work roughly at their sweet spots.

Amidon 1020-size beads are about the largest cores one can easily obtain at reasonable cost. Not only from Amidon but also from other distributors and made by other companies tha Fair Rite. These beads measure roughly 26mm outer diameter, 13mm inner diameter, and 28mm length. So they have a cross-sectional area of $1.82cm^2$. Two of them side-by-side are definitely not enough to fulfill our requirement, but using 4 of these cores gives us about $7.3cm^2$, which would give low loss with material 61, and acceptable loss with materials 43 or 31.

Just how much loss? The flux density at 71V RMS would be around 12mT at 1.8MHz. With material 61, and according to my measurements on that specific core type and size, this gives a volumetric loss of roughly $100mW/cm^3$ at 1.8MHz and almost the same loss at 54MHz, slightly less at 28MHz, and much less on all other bands. The volume of a 1020-size bead is roughly $11cm^3$, so each bead would lose 1.1W, and the core loss for the entire transformer would be 4.4W on the worst-case bands. With 43 material instead, the loss would be roughly 3 times as high, for 13W or so of total loss. It would certainly need forced-air cooling if used with a high duty cycle!

1020-size beads have rather large holes for such a transformer. Let's try beads with smaller holes, such as the Amidon 5621 size. These are the same length as the 1020 size, but with an outer diameter of 14.3mm and an inner diameter of 6.3mm. So their cross-sectional area is $1.12cm^2$. Six of these would be needed to get enough total cross-sectional area. It would be $6.7cm^2$, about 10% less than with four 1020 beads. This would make the flux density about 10% higher. According to the measurement I did on that exact core, in 61 material, it would result in a loss of roughly $150mW/cm^3$. Given that core volume for each of these beads is $3.6cm^3$, and we have six of them for a total volume of $21.7cm^3$, the loss on 1.8MHz would be arond 3.3W. So, despite having a slightly smaller total cross-sectional area, and thus a higher volumetric loss, the smaller total volume gives us a net improvement in total power loss! But the dissipation surface is also much smaller, and so these cores will get warmer. With 61 material the heating is still low, but with 43 material it might start to become problematic, depending on the duty cycle at which the amplifier runs.

The 5621 cores also have a cost advantage, since six of these are cheaper than four of the larger ones.

The secondary consists of 3 turns of insulated wire looped through the primary. For best coupling the wire should be as thick as fits in the available space. With the 1020-size cores this results in very thick wire, and would have to be implemented using coax cable, using the shield as the conductor, to avoid having a massive, excessively stiff conductor. With the 5621-size beads instead a normal sized wire can be used. Teflon insulation is desirable, because it's highly heat-resistant.

Now let's calculate the inductance. The inductance of a perfect cored inductor is

$$\text{Inductance} = 0.000001257 \times \text{permeability} \times \text{turns}^2 \times \text{cross section area} \div \text{path length}$$

where everything is in base units (inductance in henry, cross section area in meters squared, path length in meters). We can cancel out a lot of zeroes by expressing the equation in microhenries, cross-sectional area in cm^2 and path length in cm, and rewrite the equation for practical work in the quick and dirty way:

$$\mu H = 0.01257 \times \text{permeability} \times \text{turns}^2 \times cm^2 \div cm$$

This equation assumes that all of the flux is in the core. But in reality there is always also some flux in the air around, so the actual inductance is a little higher. When using cores having very low permeability this inaccuracy becomes significant, but not with permeabilty 125.

The mean magnetic path length of a 1020-size core is 5.65cm. So, if we make the transformer using four FB-61-1020 beads, we get that path length, the total cross-sectional area of $7.3cm^2$, and with the permeability of the 61 material being 125, our single-turn primary gives us an inductance of $2\mu H$. This is in a range where at 1.8MHz we will notice some effect from lack of inductance but with some goodwill it should still be usable... If we use low frequency compensation, it's easy to compensate for this inductance and get a flat response to well below 1.8MHz.

To get a flat response on the lowest band without needing compensation, you might consider using material 43. Thanks to its permeability of 850, this material would give us a primary inductance of $13.8\mu H$ - but only at very low frequencies, far below the 1.8MHz point for which we are calculating! You need to be aware of this: The initial permeability stated for a ferrite material is valid at very low frequency, and beyond some higher frequency it starts falling off. For 61 material this fall-off happens above roughly 30MHz, but for material 43 it starts at a few hundred kHz! At 1.8MHz, the permeability is down to roughly 500. So in fact we will only get roughly $8\mu H$ at 1.8MHz - which is still plenty, and 4 times as high as with material 61.

Material 31 has an initial permeability of 1500, and at 1.8MHz it still has about 1200. So with four FB-31-1020 beads we would get close to $20\mu H$, which is not just plenty, but far more than makes sense.

It's good to look at the manufacturer-provided curves for complex permeability, and understand them. But it's also good to realize that these curves are valid only at low flux density, and they change somewhat with the higher flux density used in power amplifiers. So they don't tell the whole story for our purposes.

Let's do the same exercise for the six FB-61-5621 cores: $6.7cm^2$ of total cross-sectional area, 3.2cm mean path length, permeability 125. We get $3.3\mu H$, much better than the $2\mu H$ provided by the fat cores, and enough to use without compensation nor worries at 1.8MHz. Which proves that bigger isn't always better! Shape matters.

Many ferrite beads of different sizes have very roughly the same ratio between outer and inner diameter. The consequence is that the inductance they provide depends on their total length and permeabilty, and is nearly independent from the bead's diameter.

I won't calculate the inductance of six 5621 cores made of higher permeability materials, because it would make no sense to use those lossier materials when the less lossy material 61 can provide enough inductance.

So far we have nailed the core loss, core heating, and the inductance. Now let's see what happens if we feed the amplifier at the false center tap of this single-turn primary. The cores at each side of the transformer will see the DC current consumed by the transistor on that side, which is of course half of the total current. We are applying 71V RMS to a 5.56Ω load, so we have a load current of 12.77A RMS in the load, for a power of about 900W. If the amplifier works in class AB, with just enough idling current to correct cross-over distortion, each transistor conducts close to a half sine of current of twice this value, during its active half cycle. That's 25.54A. During the other half cycle it doesn't conduct. The supply current is then the DC value corresponding to a rectified sine of 25.54A RMS, and that's very close to 23A. This is of course a value that does not include losses in the core and other components, the feedback circuit, etc, so we need to consider a slightly larger supply current, maybe 26A. We should also consider the possibility of higher current due to some load mismatch. So let's take 15A per side as a referential value, and see what happens.

The average flux density in a core is given by

Flux density = 0.000001257 × permeability × turns × current ÷ path length

where again everything is in base units: Tesla, ampere, meter. If you want to get rid of some zeroes, you can express this in millitesla and centimeter:

mT = 0.1257 × permeability × turns × ampere ÷ cm

Note that the cross-sectional area of the core has no effect on DC flux density, since a larger area results in higher inductance, this results in larger voltage while setting up the field, and this larger voltage cancels out with the higher area. See my article about Transformers and Coils about this. There I explain that DC works just like AC in inductors, and why. But for the practical work in RF power amplifiers, let's use the all-inclusive equation above for flux density, to save some work.

Okay... Applying this equation, it turns out that an FB-61-1020 core having 5.65cm mean path length, with 15A flowing through it, will run at an average flux density of about 42mT. Note that this is just the average flux density, though. This core has an inner diameter of only half as much as its outer diameter. The path lengths along the inside and outside surfaces have the same ratio, and for that reason the flux density along its inner surface will be twice that along its outer surface! So the highest DC-caused flux density, along the inner surface of the core, will be roughly 59mT, and along the outer surface it will only be half that much. The ratio between each of them and the mean value is given by √2, of course, since the ratio between inside and outside diameter of this core is 2.

And the actual instantaneous flux density varies at the RF rate, between the DC value plus or minus the RF value. So the peak flux density is the mean DC flux density plus the peak mean RF flux density (12mT at 1.8MHz, our worst case), multiplied by the square root of the diameter ratio. And that turns out to be 76mT, in this case.

76mT is a value at which material 61 is still pretty linear, so we are fine with this design. The FB-61-1020 cores can take the supply current. But we don't have an enormous headroom, given that material 61 starts becoming nonlinear at roughly 120mT.

If we use the smaller FB-61-5621 beads, with their mean path length of just 3.2cm, we get a mean DC flux density of 74mT, plus about 13.2mT RF flux density, making 87.4mT peak flux density at the mean path length, and given the diameter ratio of 2.27, the maximum peak flux density along the inner surface is 132mT. This is a bit high. It would probably still be acceptable, but this transformer would be running pretty much at its absolute limit when also feeding the supply current into the amplifier. It follows that when applying the power supply in this way, it would be better to use the 1020 beads, to be on the safe side.

And what happens if you want to use material 43? Well, using FB-43-1020 beads, with their permeability of 850, the mean DC flux density is 284mT! Add the 12mT coming from RF, multiply by √2 to get the flux density along the inside, and the result is 418mT! But this material becomes nonlinear at about 100mT, and is totally saturated at 300mT! So this isn't workable at all. If you put 15A of DC through an FB-43-1020 bead, the inner zone of the cores will saturate, the permeability drops enormously, and the response gets very nonlinear.

With material 31 the situation is even worse, due to its even higher permeability. The result is: If you want to run DC through a single-turn primary winding, you are stuck to low-permeability ferrite!

Have you lost track?

What we have until here is that with four FB-61-1020 beads we get low loss, ample DC handling, while the inductance is a little on the short side for 1.8MHz operation but could be compensated if no DC needs to be run through the transformer. So either you accept slightly suboptimal performance at 1.8MHz, or you accept the need for a separate feed system - but then you don't need these big cores anyway!

The six FB-61-5621 beads have even lower total loss, but warm up slightly more due to their smaller dissipation surface. They provide enough inductance on 1.8MHz, and could just barely handle the DC if needed, subject to practical testing. This looks like a decent design compromise.

43 and 31 material beads provide ample inductance, but are much more lossy, and absolutely cannot handle the DC in this amplifier.

Of course there are lots of other core sizes, and when you look to other manufacturers than Fair Rite there are several other suitable materials, so you might want to do this evaluation for all cores you can obtain. I would suggest to look specially for materials having permeabilities between 200 and 400.

But let's not jump too happily to closing this chapter, because the most problematic matter is still missing: What will be the highest frequency at which each of these transformers can work well enough! The big problem is that RF transformers have significant leakage inductance, caused by imperfect coupling. Very simply, the primary and secondary conductors do not occupy exactly the same physical space, and so there is some magnetic field that encloses only one of those conductors, and not the other. This creates inductance in each winding that is not coupled to the other winding. This leakage inductance appears in series with the transformer. Both windings have leakage inductance, but for practical purposes it can be considered to be on either side. The leakage inductance of the other winding is transformed by the transformer's ratio to the side we are measuring on.

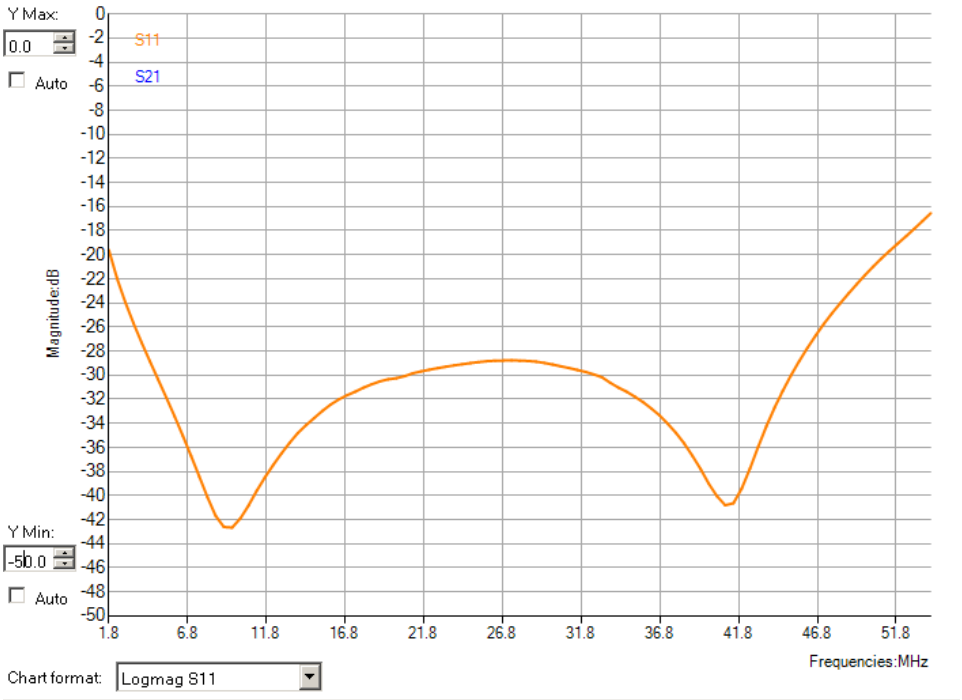
I don't have a good equation to calculate the leakage inductance. Measuring it on such a transformer built with four 1020-size beads, using thin wire for the secondary (about 1mm copper diameter), gives a total leakage inductance of 440nH, as seen from the secondary side, with a meter that uses a low testing frequency. At HF the measured leakage inductance should be slightly lower, due to stray capacitances partially compensating it, but the difference probably won't be much. Note that the leakage inductance is independent from the core permeability. Even the complete lack of a core won't change it!

With that leakage inductance, the transformer would produce decent performance roughly up to 8MHz. Of course this is unacceptable. It is possible to compensate a transformer for high frequency by absorbing the leakage inductance into a lowpass filter of a frequency higher than the highest operating frequency. This is analogous to low-frequency compensation, but the capacitors are added in parallel to the windings, and thus they don't block the DC. The bad news is that with 440nH of leakage inductance on the 50Ω side, optimal compensation flattens the frequency response only to about 20MHz. At 30MHz it's already more than 1dB down, and please let's forget about 54MHz! It would be impossible to operate on the 50MHz band with this transformer, even after compensation.

The coupling factor can be improved by using thicker wire for the secondary, trying to fill out the primary tubes as much as possible. The thickest wire of which I could pass three turns through my thin-walled primary tubes was RG-58 coax cable. Using its braid as my secondary conductor, the leakage inductance dropped to 337nH. Better, but still not brilliant. Using optimal compensation, this version would be entirely usable to 30MHz. The response is down about 0.3dB at 30MHz. But at 54MHz it's still unusable, with an attenuation of more than 5dB! The required compensation capacitors are 67pF across the secondary winding, and 612pF across the primary.

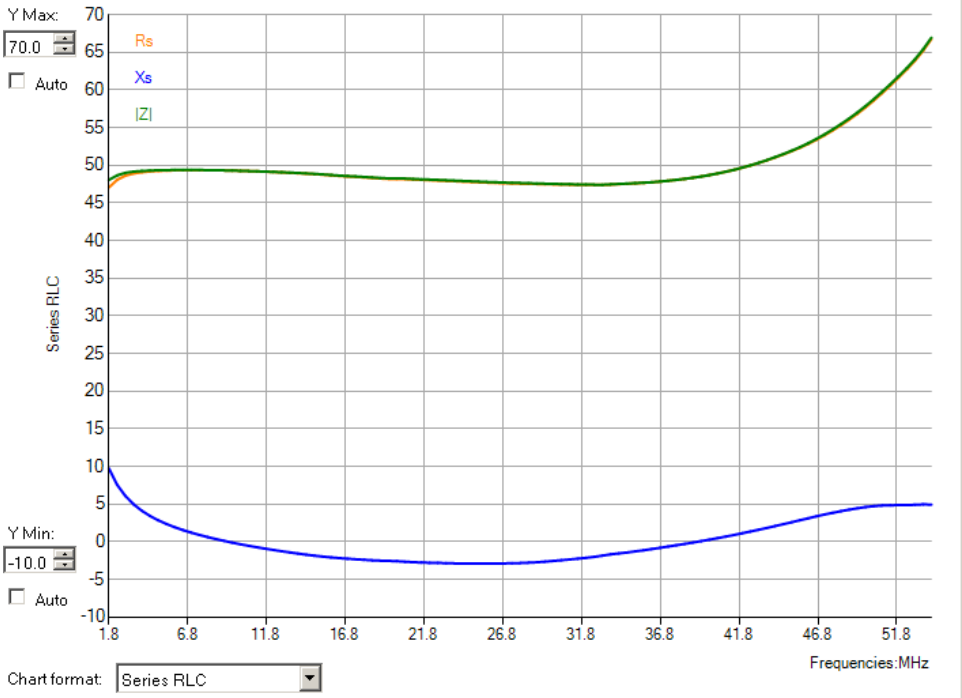
Another way to build such a transformer is to forget the metal tubes, and instead wind three turns of coax cable, then use the intact inner conductor as the secondary, and cut up the outer cover and the braid so as to connect the three loops of braid in parallel, to form the primary. This causes a significant improvement in the coupling factor, because the entire secondary conductor is completely surrounded by primary conductor, centered in it, and each secondary turn doesn't "see" the other secondary turns. I made a transformer in this style, using the same RG-58 coax cable, and obtained 202nH leakage inductance, as seen from the secondary side. Properly compensated, this transformer might work well enough to 54MHz, the response being 0.45dB down at that frequency. The compensation capacitors are 40pF across the secondary and 364pF across the primary.

Since it was easy enough to do, I soldered a 39pF capacitor across the primary, a 330pF and a 33pF capacitors to the secondary, and also 6 resistors of 33Ω each to the secondary. This implements the calculated compensation, and a 5.5Ω load, with hopefully low series inductance. Then I measured the response, using my recently acquired NanoVNA. This is the reflection coefficient I obtained:



It's just like expected! The response at 1.8MHz is just at the acceptable limit, which is usually considered to be a reflection of -20dB.And at 54MHz it's a bit too bad. The reflection crosses -20dB at 50MHz and gets close to -16dB at 54MHz. Really we should try this with coax cable having a lower impedance, thus getting a better coupling factor!

To see what we have, here is the series impedance plot.



The resistance is a tad below 50Ω, simply because my 33Ω resistors are also a tad low, and anyway six 33Ω resistors give 5.5Ω instead of 5.56Ω. Let's split some hairs, right? At the low frequency end we see the effect of the main inductance of this transformer being just at the acceptable limit for 1.8MHz, providing 10Ω of series reactance. This could be corrected by adding series capacitors. From there up the transformer works great, although around 30MHz the impedance transformation is a tiny bit off, which could surely be corrected by fine-tuning the compensation capacitances on both sides. But at the high end the transformation ratio runs away, because we are nearing the cutoff frequency of the compensation network. To fix this, we need to further reduce the leakage inductance.

This could probably be done by using 25Ω coax cable. I don't have any 25Ω coax cable at hand to cut up and try, and I was too lazy to make any. But if I was making an amplifier of this kind that has to operate to 54MHz, I would try this. Note that since the braids of all turns are at the same potential, the outer insulation could be completely removed, to save space and ease the splicing and soldering job. But in that case it would be good to use a coax cable that has a very tight braid, so it won't separate much from the dielectric.

This is my test transformer. Neither tidy nor nice, but good enough to test whether the stuff I'm writing here actually works!

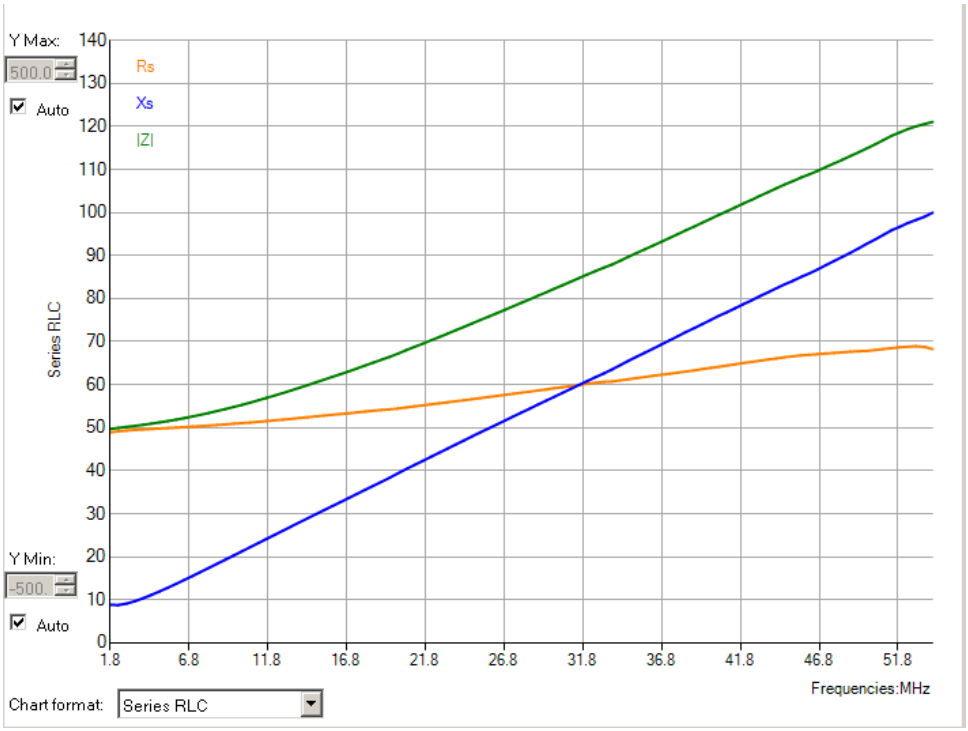
Of course such a transformer should use teflon-insulated coax cable, due to its excellent heat resistance. And if I used it in an actual amplifier, I would bring the coax solder points down to the board and solder them there, so that good nice very wide copper areas can be used to provide low impedance connections. In a practical amplifier the drain connections will always have significant inductance, even if they are just a rectangle of copper 10mm wide and 15mm long. It might be a good idea to try compensating them too, by adding the required additional capacitance to the transformer primary, and using the transistor's output capacitance on the other side. This would require clever dimensioning of the board traces.

Such a coax-wound transformer has input and output at the same end, unless you are willing to do some more cutting and patching. But it's rather easy to accomodate this on a board.

I decided to also try a transformer made with six FB-61-5621 beads, in two rows of three, side by side, with single turn copper pipe primary, and three turns of the same wire used for the first experiment with the four 1020-size beads. A slightly longer piece of the wire was required, due to the longer turn length. The three turns of this wire fit snugly in these cores. The leakage inductance, as measured on the secondary side, turned out to be 443nH, within 1% of what I got with the larger cores! Obviously the advantage of better coupling, due to filling out the primary tubes with the secondary, cancelled out with the disadvantage of the greater turn length.

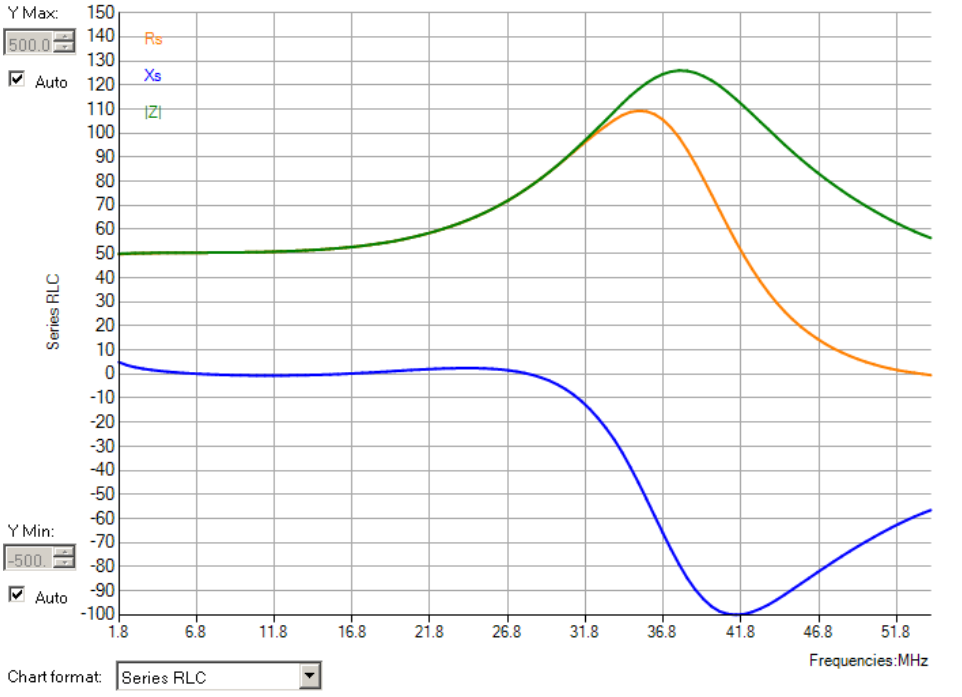
But with this small transformer there is not much to be gained with thicker wire. At most one could use some wire that has a thicker conductor and thinner insulation. And if winding coax cable on these cores, miniature coax cable would have to be used, which can't handle high enough power.

The primary inductance measured exactly 4μH, slightly better than the calculated 3.3μH. This means that indeed the performance of this transformer should be good at 1.8MHz. So I tested this transformer both with and without compensation, to see if it is good for anything. Just as a joke, I tried it first without any compensation, but with a nice, precise 5.56Ω load. This is the series impedance graph, from 1.8 to 54MHz:



At least on the 160m band it's usable...! Maybe on 80 meters too. But beyond that, the leakage inductance kills it. So let's try compensation. A 3-pole lowpass filter with its -3dB point at 36MHz can absorb the leakage inductance of this transformer. It needs an 88pF capacitor across the secondary and a 792pF one across the primary.

I soldered two 390pF capacitors across the secondary, and one 82pF across the primary. This is the result:



With this compensation, the transformer would be fine from 1.5MHz to 22MHz. That's the range in which the reflection stays below -20dB. The reactance remains acceptable all the way to 30MHz, but the transformation ratio deviates too much. I tried to find better compensation capacitances by trial and error, but the ones calculated from theory proved to be the best. So this transformer isn't usable beyond the 15m band, and for that reason most hams won't want it. Case 5621 closed! Posing for the photo at right was the only thing this transformer was good for. Note that this test transformer has primary and secondary connections on the same end, only because I'm getting old and a tad stupid, and such things happen... Of course I would wind it with the connections on opposite ends if it was to be used for real. That would be more practical for board layout.

After this failed attempt, it would be logical to go to the other extreme now, and try a particularly short and fat set of cores, that achives the shortest possible turn length per core cross-sectional area. This would minimize the leakage inductance, but at the same time would increase any trouble coming from unequal flux density distribution in the cores. It would also require a larger total ferrite volumen, increasing weight, cost, and total core loss.

The best core shapes for such a transformer would be RM cores, or pot cores. You would need something like an RM28 core (28mm center leg diameter), made from an RF-capable ferrite. A permeability between 120 and 200 should be optimal. Good luck finding any! We can dream, right?

Since we can't wind a transformer on a core we can't get, our best chance would be extra large beads. Amidon sells some jumbo beads in materials 43 and 31, but unfortunately not in 61. When using such large cores of the lossier materials, we need to use some extra cross-sectional area to keep the flux density low enough to get acceptable loss. Just for the fun of it, I will make a design exercise with two Amidon FB-43-20003 beads. Each of these has an outer diameter of 50.8mm, inner diameter of 25.4mm, and a length of 38.1mm. Yes, American company, inch sizes... Let's round the dimensions to 51, 25 and 38mm.

Two of these beads will have a cross-sectional area of 9.5cm^2, and the turns shape will be close to a square, which is the next best we can get after a circle. The average turn length will be about 20cm. And that's already a tad more than the four 1020-size beads need, so the leakage inductance would be worse! The only way to achieve an advantage might be to stay far away from filling the whole big holes of these cores with wire, and wind tighly with the shortest turn length possible. Three turns of RG-58-size coax cable should end up having a turn length of about 16cm. That would give a very slight advantage over the four 1020 beads.

The average flux density at 1.8MHz would be around 9mT, so the volumetric core loss would be around 150mT/cm^3. But the total core volume is 118cm^3, so we would get close to 18W core loss, at full power output and 1.8MHz! At this point the approach stops looking attractive.

If you can find beads or toroids made of low loss material, in a permeability range of 200 to 400 or so, that can be stacked in such a way as to obtain 6 to 8cm^2 total cross-sectional area, with a round or nearly square winding loop shape, it would be worthwhile to make a design exercise with them. Otherwise just use the four FB-61-1020! Or perhaps look for any core combination that are stacks up a little longer than these. Properly compensated at the high end, they would produce very flat 1.8 to 30MHz response, instead of the roughly 2.3 to 42MHz very flat response of the four FB-61-1020.

Toroids?

Instead of using a single turn primary on a fat core, it is theoretically also possible to use a multiturn primary on a thinner core. The result is usually a longer winding, and thus more leakage inductance, but on the other hand the main inductance is also larger. Just to see what happens, let's consider using a single big toroid and design this transformer on it. I will choose the very well known Amidon FT-240-61. It has a cross-sectional area of 1.6cm^2 and a mean path length of 15cm. Its cross section is square with slightly rounded corners, pretty much the best we can obtain.

At the RF operating voltage it gets in our amplifier, a 4-turn primary will produce a flux density of 13.9mT at 1.8MHz, which is fine. This means that the secondary needs 12 turns. And a bonus is that we can make a real center tap on the primary, since it has an even number of turns, and this allows us to feed the amplifier through this true center tap, allowing us to operate in true class AB. As long as the amplifier is well balanced, the DC in the primary turns will cause opposite fields and thus create no flux in the core.

The problem is that we can't just wind 12 turns of wire on this toroid for the secondary, and then wind 2+2 turns of a wide conductor (braid, metal tape) over it, because the coupling factor would be absolutely terrible! Just for the sake of a laugh, I did it. I got 2.1μH leakage inductance, as seen from the secondary side! It would be possible to reduce this somewhat by tidy construction, but not a lot. So this simplistic approach is a no-go.

It would be possible to do better by twisting three enameled wires closely together, and then make six windings of two turns each. These windings can each sit at one part of the toroid, they don't need to overlap. Then all 36 wire ends need to be freed of their enamel insulation, and interconnected in such a way that one wire from each trio forms the secondary, with all six of them connected in series, while another wire from each trio forms one side of the primary, all six of them connected in parallel, and likewise the last wire from each trio form the other side of the primary. All interconnections need to be made extremely short, to avoid introducing more leakage inductance than what's inevitable.

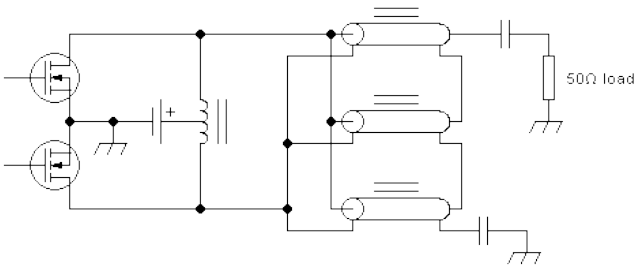
With a core this size, I don't think that the result could compete in terms of leakage inductance with the coax winding on beads described above. And if it was messy to peel, cut and interconnect those coax cables, it's far more messy to scrape away the insulation of 36 very short ends of enameled wire, and then correctly (!) interconnect those 36 wire ends. It would require an amplifier masoquist to do that.

Anyway the primary inductance of this approach is about 2.6μH. Somewhat better than the four 1020 beads, OK for 1.8MHz, without any excess.

So the bottom line is that the best conventional transformer I can come up with, for this amplifier, is three turns of teflon coax cable, probably of 25Ω impedance, wound through four FB-61-1020 beads, using the shields as primary and the inner conductor as secondary. It gives barely enough inductance on 1.8MHz, could easily be compensated to have flat response to a lower frequency, and with 25Ω cable and optimal compensation it would very likely be good enough to 54MHz. It's reasonably easy to make, too. It can handle the DC, if the builder choses to feed the amplifier through the false center tap of this transformer (at the midpoint of all three shields), but of course this is acceptable only for class A and current-switching class D amplifiers, and since nobody in his sane mind would build a class A amplifier around a kilowatt-class transistor, that only leaves class D. For class AB a separate bifiliar feed choke is needed, that couples the drains.

Transmission line transformer calculation:

Let's now see how to design transmission line transformers. First we will look at the 1:9 true, equal-phase transmission line transformer, that doesn't need a separate balun. To refresh your mind, I mean the transformer in this circuit:



The three transmission lines are identical in impedance and length. The impedance must be one third of the 50Ω load, and three times the 5.56Ω they present on the input. This is a direct consequence of the lines being connected in series at the load, and in parallel at the input. The lines do not perform any impedance transformation! They work at their own impedance, front-to-end. The transformation is obtained just by connecting them in parallel at one end and in series at the other end.

So we need transmission lines having an impedance of 16.67Ω, or very close to it. In theory the lines could be coax cable, parallel cable, twisted wires, strips of double-sided flexible PCB, or whatever else you can come up with, that fulfills the impedance requirement, and can handle the voltage and current present in them. Of course, each line works at the full input voltage, one third of the output voltage, and at the full output current, one third of the input current. In this design example that's 71V RMS and 4.3A RMS. The lines must be able to handle this over the full frequency range, without excessive loss. Each line works at one third the output power, so roughly 300W.

There are very few companies that make such special transmission lines, they are hard to find, and absurdly expensive when bought in small quantities. So it's better to make one's own lines. One can buy teflon-insulated wire in various sizes, and use braid harvested from cheap coax cables, to make one's own coax cable in any desired impedance from about 10Ω to 100Ω. Parallel lines are easy to make from about 100Ω to 500Ω or so. Coax lines below 10Ω can be made by making the inner conductor from braid on a passive core, and using very thin insulation, for example kapton tape. With any homemade coax line, after applying the outer conductor one can apply shrink wrap to hold them together. There is even teflon shrink wrap, which works at higher temperature, but is expensive. One could also just wrap them with PVC electrical tape, for testing and non-demanding low-temperature applications. There are online calculators to find out what diameter ratio is required, depending on the desired impedance and on the dielectric material used.

These lines are wound on ferrite cores. For these cores the same requirements and calculations apply as in the case of conventional transformers: One needs to select a suitable core and turns number to provide the required inductance at the lowest frequency, and keep the flux density low enough to keep the core loss at an acceptable level. So we need to know what voltages appear end-to-end on each line, and how much inductance is necessary.

Let's calculate the voltages in a very simple way: Let's consider that at a given instant the drain voltage of the upper transistor is at RF potential "1". Then of course the bifiliar feed transformer forces that the drain of the lower transistor is at RF potential -1. So all three lines have their shields connected to -1, and their center conductors connected to 1, and thus their inputs are in average at potential zero. Each line has 2 between its center conductor and shield.

On the output side, the shield of the lowermost line is grounded through a capacitor. So it is at potential zero. Since it has 2 between its conductors, its output center conductor is at potential 2, and in average the output is at 1. The mid line has its shield connected to 2, so its center conductor is at 4, and in average its output is at 3. The upper line has its shield at 4, so its center conductor is at 6, and in average its output is at 5.

So, the lowermost line has 1 across it from end to end, the middle line has 3, and the upper line has 5.

Since the RMS voltage between drains is 71V, our potential "1" in this case equals 35.5V RMS. It follows that the lower line sees 35.5V end-to-end, the middle line sees 106.5V, and the upper line sees 177.5V. And this means that the three lines need very different minimum values for the ferrite cross-sectional area × turns number product. Since these are minimum requirements, it's the designers choice whether to actually use different cores, different turn numbers, or make all three identical and design them for the requirement of the uppermost line. In the latter case, the cores of the mid and lower lines will work at very low loss. Instead in the former case, it's possible to save a significant amount of ferrite (=cost, weight).

Since identical line lengths are required to keep the output signals properly phase-aligned, a method that is very attractive is using an identical number of turns in each line, but different numbers or sizes of cores.

In the example at hand, if we want to use 61-type ferrite and keep the flux density below 15mT at 1.8MHz, at 177.5V we need a turns × area product of 15cm² for the uppermost line. The middle line would be OK with a turns × area product of 9cm², and the lowermost with just 3cm².

If we choose to wind each transmission line in 3 turns, we need cores totalling 5cm² cross-sectional area for the uppermost line, 3cm² for the middle line, and just 1cm² for the lowermost line. Or we could use identical cores of 1cm², and wind 15 turns, 9 turns and 3 turns respectively. In practice we need to see what cores are available, consider their cross-sectional areas, their winding space, the diameter of the transmission line used, and find out which is the most convenient core and number of turns, in terms of cost, size and loss.

But while we do this, we must also make sure that we get enough inductance. How much inductance we need, depends on how much imbalance we can tolerate. If we want perfect balance in the amplifier, we need infinite inductance.

The input sides of our lines are RF-ground-referenced by the amplifier's supply circuit. And the output side is ground-referenced at the lowermost point, the shield of one line. This confinement makes the lines work at the fixed end-to-end voltages already mentioned, assuming 71V RMS drain-to-drain voltage. So the common-mode currents they inject into the amplifier depend strictly on these voltages and the end-to-end impedance of each line wound on its core. It's good to think of impedance here, rather than reactance, because many ferrites used in this application have a more resistive than reactive behaviour in the upper part of the HF spectrum. But at the lowest frequency, which is usually the worst case and thus the frequency we have to design for, most RF ferrites are mostly inductive, so you are allowed to think in terms of reactance and inductance!

Very simply stated: The uppermost line in this example works at 177.5V RMS end-to-end, and the unwanted common-mode current it injects into the amplifier depends on this voltage, and its choking impedance, by Ohm's Law. The big question is: How much of this unwanted common-mode current is acceptable? There is no obvious, clear answer to this. It's a design trade-off like so many others. We might want to keep this common-mode current to one tenth of the total drain current, as a simple rule of thumb. The smaller we make it, the better balanced the amplifier will be.

Not only the uppermost line injects unwanted common-mode current. The others do it too, and in this transformer configuration, all three common-mode currents are in the same phase, and thus add up. We need to distribute the one-tenth allowance among those three lines, in a way that is practical to implement.

If we intend to use the same number of turns for all three lines, and vary the ferrite cross-sectional area in the ratio 1:3:5, then the inductances will also vary in that ratio, and since the end-to-end voltages have the same ratio, each line will contribute the same amount of common-mode current. So we assign one third of the one tenth allowance to

each line. The total drain current in this example amplifier is roughly 30A, so we will accept 1A of common-mode current in each line. That requires a reactance of 177.5Ω for the uppermost line, which at 1.8MHz is an inductance of 15.7μH. If we design the uppermost line for this value or higher, and then scale the ferrite cross-sectional area for the other lines, all will be fine.

To avoid ending up with enormous ferrite cores, it's a good idea to start with the thinnest transmission line that has the proper impedance and that can carry the required power over the full frequency range. Let's assume that for this application you were able to make a coax line of 16.67Ω impedance, that can carry the required 300W, and has a diameter of 4mm. Then we start trying: The FB61-1020 core could accept a maximum of 7 turns of this line. It gives roughly 0.5μH for one turn. Inductance depends on the square of turns number, so 7 turns gives roughly 25μH. Plenty enough! The ferrite cross-sectional area times the turns number is roughly 12cm^2. But we need at least 15cm^2! Hmmm... We can't wind more turns, because they don't fit. So we need more ferrite, unless we can somehow shrink the coax cable to pass two more turns through this core. That's why thin transmission line is so important!

Assuming that we can't shrink the cable, the next attempt could be using two FB-61-1020 beads side by side. Having twice the ferrite cross-sectional area, 5 turns is enough in terms of flux density. It gives us roughly 12.5μH, and that's a bit low. But 6 turns on these two cores would be fine for the upper line.

The middle transmission line is OK with three fifths the core area × turns product, and inductance. Putting six turns on a single core would give a tad less than that - but we have space for 7 turns! So we use a single core, wind 7 turns on it, and call it a day. The cable length required for 7 turns on a single core is very similar to that required for 6 turns on two cores side-by-side. Of course, for the sake of phase alignment, we should make all three lines identically long. Any slack can be taken up by expanding the turns a little.

And the lowermost line would be OK with fewer turns or a slightly smaller core, but why do it? It's more practical to use the same core size throughout. And since anyway we need to use the same lengths of transmission line, I would just make the lowest line identical to the middle one, 7 turns on a single core. That core will work cold, while the other ones will get warm under the worst operating conditions.

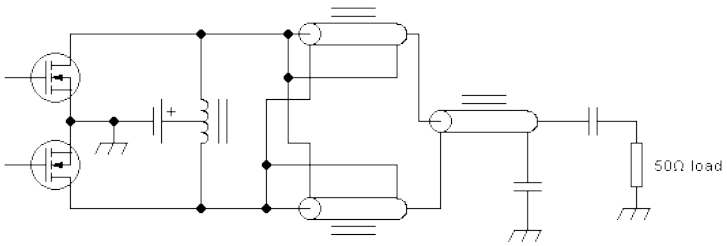
So, we end up with a lot of practical compromises, and using a total of four FB-61-1020 cores for this 1:9 transmission-line transformer. And that's the same amount and type of ferrite we needed for the conventional transformer! Funny, huh?

The lower frequency limit of this transformer is given by excessive common-mode current injection into the amplifer, and we have set it by design to 1.8MHz or slightly below. The upper frequency limit is given mainly by inter-turn capacitance, which tends to short out the choking effect and thus inject excessive common-mode current. In the transformer described this capacitance will be just a few pF, maybe in the low tens of pF, and depends on construction details, such as the thickness and dielectric constant of the outer insulator of our coax cable. We can use the transformer up to a frequency where this capacitance has a reactance of the same 177.5Ω. For example, if we get 10pF total input-to-output capacitance, it limits the operation to a maximum frequency of roughly 90MHz.

If we get too much capacitance for the highest operating frequency desired, we can try reducing it by using a coax cable with thicker outer insulator, made of low dielectric constant material (teflon), also by avoiding packed windings (like in side-by-side cores) and prefering windings airily distributed around a single core, with the largest turn spacing possible. But the parts of the cable that are inside the core will still be very close together, and this can be eased by using a large diameter toroid instead of a bead, and keeping some distance between the first and the last turn. In extreme cases two or more wound cores can be placed in series to form one transmission line choke.

Since the three lines work at known end-to-end voltages, we can in principle wind them on a single core. In that case we must painstakingly follow the 1:3:5 turns ratio, and of course we must get the phasing right, by winding them all in the same sense. But I don't think it's a good idea to do it in this case, because we would need a very large core to fit all those turns. Such a core is more expensive than several small ones - if it exists at all! Also the capacitances between turns become more of a problem. But with transmission-line transformers used at low voltage and low power, core sharing is a definite possibility.

Now let's do a design exercise for the other configuration:



The bootstrapped 1:9 transformer uses two lines of 16.7Ω. Since their outputs are connected in series with the input signal, any phase lag that happens along the lines is detrimental, and so these lines should be as short as possible. Their length are the main factor defining the upper frequency limit.

Like in the previous example, the input side of both lines have an average RF potential of zero, while their output sides have the shields bootstrapped to the inputs, so the upper line has its output side shield at "1", its center conductor at 3, and thus the average output is at 2. The lower line is the same reversed, that is, its shield is bootstrapped to -1, its center conductor is at -3, so its average output is at -2. As a result both lines have identical end-to-end voltage, although in reversed phases. This voltage is "2", and as explained in the example above, this equals 71V in this amplifier.

The common-mode current in these lines appears as inductive loading current on the amplifier. We want to keep it down to a total of one tenth the amplifier current, so we can allow about 1.5A per line at the lowest frequency. That calls for a minimum reactance of about 47Ω at the lowest frequency, which is 4.2μH for 1.8MHz.

Using the methods already explained, and trying our old friend the FB-61-1020, four turns on one core are enough to handle the voltage. They will give us around 8μH, which is also enough. Note that this under-utilizes the 1020-size core. We could try to use a smaller core, maybe two significantly smaller beads side-by-side, which saves a few cents and also makes the design nicer to look at. Or we could keep the 1020 size beads, and have some headroom to use slightly thicker coax cable.

This bootstrapped transformer is ideally suited to winding it on a single core, due to its symmetrical operation of two identical lines under identical stress. If you do so, don't forget that the two lines work with opposed end-to-end voltage, so they have to be wound in opposite senses on the shared core! But this too helps symmetrical construction.

Remember that the best design for this transformer is the one that ends up with the shortest lines. The total electrical length of each line should stay below 8% or so of the wavelength, at the highest frequency that will be used. If that's 54MHz, then the maximum allowable electrical length is around 48cm, and if the cable has a velocity factor of 66%, the maximum allowable physical length is just 32cm. This is probably not yet a problem in this example, but may be close to the limit.

If it does become a problem, there is a simple fix: Instead of making the bootstrapping connections short and direct, they can be made with a third transmission line,

connected exactly like in the previous example. Since this line doesn't need to sustain any end-to-end voltage, it doesn't need to be wound on a core. It's just a delay line. So, cut a piece of your low impedance cable, the same length as the other two, coil it up in the air, and use it for the bootstrap connections. Problem fixed! The line length restriction is now removed, and the upper frequency limit has been extended to whatever the stray capacitances will allow. But at high frequencies, where this line alone has significat end-to-end inductance, the output side of the 1:9 transformer becomes floating, instead of being referenced to the amplifier's grounding, and this can cause imbalance between the two main lines.

The output balun works at 50Ω at both sides, so it's wound with 50Ω coax cable. This cable has to carry the full power, so we can't use miniature coax. RG400 or similar is typically used. Its input side is nominally at average potential zero, and its output at "3", so it sees 106.5V between its ends. The common-mode current it conducts will be injected into the amplifier, so it has to be kept low. Applying the same one-tenth rule of thumb as before, and considering it's just one line injecting this current, 3A would be acceptable in the worst case. So we need at least a reactance of 36Ω or so at the lowest frequency, but more is always better in this case. At 1.8MHz that's a minimum of 3.2μH.

On a single FB-61-1020 we would need six turns, to satisfy the flux density and loss requirement. But six turns of RG400 don't fit. So we need to use two of those cores. 3 turns would give acceptable flux density, along with 4.5μH, which is acceptable. But a fourth turn can be wound, resulting in less core heating and much lower imbalance, so that's what I would suggest.

Funny... Using one FB-61-1020 for each section of the 1:9 transformer, and two for the balun, **again** we ended up with 4 of those cores! Isn't this spooky?

So, it's your free choice whether you use a conventional transformer, a full 1:9 transmission-line transformer, or a bootstrapped 1:9 plus balun. In all three cases you can use the same cores. Only the coax cable needed is different. But there are important differences in the performance: The conventional transformer requires very careful construction and compensation, to achieve just enough performance. Instead both transmission-line versions are easy to get right, and will give a very flat response. That's why most builders of such amplifiers prefer transmission-line transformers. But the transmission-line transformer also has mighty disadvantages, where the conventional transformer wins. One of these is the long delay of transmission-line transformers. Their line, from input to output, might easily have an electrical length of 100 to 150cm! This creates big trouble in class D amplifiers. After all, they **need** the lowpass filter to define the waveform at the drains, either voltage or current! If the lowpass filter is connected through a long line like this, the harmonics reflected by the filter arrive totally out of phase, and lead to horrible and often dangerous waveforms at the drains, which are totally different between one band and another. A current-switching class-D amplifier needs a filter that essentially short-circuits the harmonics. At any frequency where the delay between the drains and the filter is a quarter wave, or an odd multiple of that, the filter-provided short circuit will be transformed by the line into an open circuit, leading to immense drain voltage spikes that kill transistors like flies. And if we have 100cm of electrical line length there, the lowest frequency where that will happen is 75MHz, the third harmonic of our 12m band, and close enough to the 5th harmonic of our 20m band. Longer line lengths make the problem start at even lower frequencies. And even on frequencies where the line is shorter than a quarter wave but still of significant length, it causes impedance transformation, turning the short-circuit reflection from the lowpass filter into a nasty inductive load. Conventional transformers cause less delay, and what delay they do cause, happens because of their leakage inductance, that has to be compensated by absorbing it into a lowpass filter - and the input capacitor of this filter then provides the required near-shorting of harmonics, on the higher bands! This is why conventional transformers are preferable in class D amplifiers. People who build class D amplifiers with transmission-line transfomers often see themselves forced to use a diplexer filter instead of a plain lowpass filter, to absorb the harmonics rather than reflecting them, and have to accept the higher filter complexity and a significant efficiency reduction.

And don't say that this is irrelevant to you because you are building a class AB amplifier, not class D. The bad news is that if you are building a kilowatt-class 50V-powered amplifer, it **will** operate in a mode close to current-switching class D, at least on the higher bands, whether you want it or not! I will come to this a little down this page, when looking at feed transformers...

Another advantage of conventional transformers is that they provide the required DC blocking. This might seem minor, since two little capacitors are cheap and easy to add, but it is important if the amplifier will be modulated through the supply voltage. In that case any coupling capacitor ends up loading the modulator. A typical modulator is a class D amplifier operating with PWM at a carrier frequency around 200kHz, and has a carefully designed lowpass filter at its output to reject the 200kHz switching frequency while passing the modulation frequencies, which might extend to 30 or even 50kHz, for clean SSB operation. This filter would be totally upset if RF coupling capacitors were added to the RF amplifier! Any capacitance in the amplifier's drain circuit needs to be absorbed into the modulator's lowpass filter's design, and there is a limit as to how much capacitance can be absorbed. That makes a conventional transformer highly attractive for such envelope-modulated amplifiers.

Finally a conventional transformer has the advantage of needing less coax cable, of less critical impedances, and in less demanding cases it can be wound with plain wire, and it is more compact.

The conventional transformer and also both versions of the transmission line transformer allow injecting the DC through the transformer, but without coupling the drains. A conventional transformer with a primary that has an even turns number would also couple the drains, but is feasible only if either the size is smaller or the top frequency is lower. So this is reserved mostly to driver stages, and low power transmitters. In high power work, injecting the power supply at the transformer results in uncoupled drains. Good for current-switching class D, but unworkable for true linear class AB.

So, let's look at the feed transformer!

Bifiliar feed chokes, feed transformers, drain coupling transformers, and related animals:

This is the hardest part of an amplifier to get right, if true class AB operation is desired. In this class each transistor conducts essentially a half wave of the RF signal, and is completely off for most of the other half cycle, yet a clean full sine wave voltage and current is expected between the drains, even without the help of a lowpass filter. This is possible only if the two drains are very tightly coupled together, so that the negative half sine, formed by the transistor that is conducting, is transformed to a clean positive half sine by this bifiliar choke, drain coupling transformer, of however you want to call it.

While the output of the true class AB amplifier, taken from drain to drain, is a full sine wave, the feed transformer works with that sinewave voltage, but with half-sine current, a waveform that has its fundamental on twice the operating frequency, and also contains all other even harmonics of the operating frequency. For that reason it is essential that it has a frequency response that includes all strong even harmonics of the highest operating frequency. If clean class AB operation to 54MHz is desired, the feed transformer needs to perform well to at least 500MHz, and it would be better if it reached 1GHz!

For this to happen, its leakage inductance needs to be small compared to the load resistance placed on the conducting transistor, which is one quarter of the drain-to-drain load resistance. And not just the feed transformer's leakage inductance needs to be that low, but also the total stray inductance of the relevant circuit. And that circuit is the entire path from one transistor's drain on its chip, to the drain tab, to the feed transformer, through the feed transformer (leakage inductance), through the bypass capacitors placed at its midpoint, back to the transistor's source connection, and closing the circuit at the transistor chip. The reactance of the total inductance along this path needs to be small compared to the load resistance, at least to the tenth or twelfth harmonic of the highest operating frequency! This is an extremely difficult assignment, and very often it turns out impossible to fulfill.

In the example I'm using here, a BLF188XR amplifier with 1:9 matching, the drain-to-drain load resistance is 5.56Ω, so the load resistance on the transistor that's on is close to 1.4Ω. If we take a factor of ten smaller as a practical value for "small compared to", we need the reactance of the total unwanted inductance along the path, including the feed transformer's leakage inductance, to be no larger than roughly 0.14Ω. At 500MHz that's an inductance of 45pH. Yes, 45 picohenries! In other words, 0.045nH, or

0.000045μH. But 1mm of wire already has roughly 1nH, depending on its thickness. Do you see the problem? It's impossible to solve.

For this reason the only situations in which good performance can be obtained in class AB is when several easing factors concur:

- Low supply voltage, allowing small ferrite cross-sectional area and thus a physically small transformer.
- Low output power, allowing the use of very thin wires, further shrinking the transformer, and also raising the load resistance and thus easing the leakage inductance requirement.
- Reducing the maximum frequency requirement to 30MHz and often even lower.
- Accepting more waveform distortion by requiring the feed transformer to work properly only up to some lower harmonic.
- Using various "patching" methods to apply bandaids where needed, to make the amplifier workable despite inadequate drain-to-drain coupling.

And that's how all attempts at broadband push-pull class AB power amplifiers in the HF range turn out. In the best of all cases they have pretty good drain-to-drain coupling on the lowest bands, allowing good class AB operation there, but on the higher bands they turn increasingly dirty and work only thanks to bandaids, such as adding large capacitance from the drains to ground, or giving up class AB outright and designing for current switching class D instead, or some hybrid class. Designers who don't like an amplifier to change operating class between lower and higher bands often decide to deliberately not have any drain-drain coupling, and thus run their amplifiers in class D on all bands. If that's done well, it produces a very usable linear class D amplifier, but which absolutely depends on lowpass filters, because its output current waveform closely approaches a trapezoid.

Some experimenters strongly dislike having to use a bank of six or more relay-switched lowpass filters, and not realizing that their supposedly class AB amplifier is really operating in class D, they try to force it to produce sine waves by means of very strong negative feedback. This can be done, to some extent, by specific design of the feedback and biasing networks, but instead of true class AB it produces dynamically biased class A operation, with an efficiency around 45% at best. I will come to this later...

I spent quite some time trying to come up with the best possible feed transformer for a 50V high power amplifier. I used the absolutely smallest cores that would provide just acceptable flux density and inductance, and wound them with a single U shaped loop of homemade ultra low impedance coax cable. This cable was made by pulling thick braid over an inert core, then insulating it with just two layers of Kapton tape, then applying another piece of the same braid as the outer conductor. I sized the inert core so that the finished cable just barely fits in the holes of the ferrite core, and only the part outside the core got an outer wrap of Kapton tape to keep the braid from lifting off the dielectric. The terminals of this transformer were made by bending out the four braid ends in the proper arrangement, with small insulating plates between them, made from PCB fiberglass. The idea is to tin these braid ends, trim them nice and clean and very short, then solder them directly to the drain tabs of a BLF188XR, the other ends being soldered directly to a wide row of bypass chip capacitors, installed through the board and soldered down to the copper baseplate (heat spreader) by their other ends. The pigtails of a transmission-line transformer for output matching would be soldered first to the ends of the MOSFET tabs, and then this feed transformer would be installed over those connections, with one or two layers of Kapton tape providing insulation.

At the time of writing this article, I haven't yet built an amplifier using this feed transformer, but other experimenters have, and saw a dramatic improvement of their drain voltage waveforms, and also of the efficiency, even some improvement of IMD levels, over any alternative design. Unfortunately this is mostly true for the lower bands! At the high end of the spectrum even this feed transformer and mounting arrangement has too much leakage and stray inductance.

By the way, these are FB-61-4852 cores, much smaller than the ones I have been mentioning before. They measure 12mm outer diameter, 5mm inner diameter, and 25mm length.

One can see many amplifiers, built by experimenters and even by factories, that instead use a toroidal core wound with 5 to 10 turns of bifiliar enameled wire. Such things really do not couple the drains, because they have sky-high leakage inductance. The behaviour of such a bifiliar-wound toroidal choke is almost the same as that of two separate RF chokes, or feeding the amplifier at the false center tap of a single-turn primary winding: There is no coupling between the drains, and the amplifier can operate in class A or current-switching class D, but not in class AB nor in voltage-switching class D. But still such bifiliar chokes on toroidal cores have three advantages over using two separate feed chokes: They cancel the DC, allowing much smaller cores to be used. They allow the supply current to vary at modulation rates without introducing phase lag, allowing better IMD values than with separate RF chokes. And lastly, they provide a convenient point from which to pick up inductively coupled negative feedback.

Although there are many other possible output circuit configurations, and many more details to mention about them, I have covered the ones that are most commonly used in HF and VHF amplifiers in the power range considered here, and this article anyway is getting much too long. So let's end the discussion of output circuits here. We still have to talk about drive circuits, bias circuits, feedback circuits, and issues like stability. But before getting there, let's insert another Blue Block!

Dynamic class A, and linear class D

Every electronician knowns what class A is: The transistor conducts all the time, amplifying the entire signal waveform. The supply current is constant at all times, and needs to be at least as large as the peak signal current that has to be delivered into the output. The maximum theoretical efficiency in linear operation is 50% at full power, and dramatically lower at partial power.

Every electronician also knows what class B is: The transistor conducts half of the time only, amplifying only the positive half of the signal waveform. Without drive it consumes no current. The supply current varies according to drive. The output signal has a severely distorted waveform, but in theory the amplifier is still perfectly linear in regard to amplitude and phase. The maximum theoretical efficiency in linear operation is 78.5% at full power, and worsens linearly as the power is reduced.

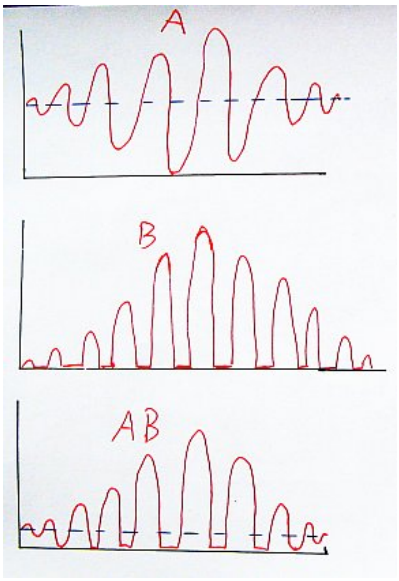
And between class A and class B there is class AB: The transistor amplifies more than half but less then the entire signal. The supply current varies with the signal, but never sinks all the way to zero, resting instead at a low to moderate level in the absence of a drive signal. The efficiency is intermediate too between those of class A and class B.

Is class AB all that exists between A and B? **No!!!** And in fact very few high power linear amplifiers, if any, operate in true class AB!

The reason for this is that classes AB and B require the supply current to vary at the RF rate, and at the low voltages and high currents used with power transistors this poses very difficult problems in the output section. You already read all of this above. Class A instead works with a constant supply current, removing these difficulties. But class A

is very inefficient, particularly in modes that have low average power, like SSB. So designers found alternatives to class AB, that work with constant supply current over the RF cycle, but have better efficiency than class A, even if it doesn't reach that of true class AB.

The trick is simply to operate the transistors in class A, but modulate the bias, so that the supply current varies according to the signal's amplitude (like in class AB), but stays constant over the RF cycle (like in class A). I call this "dynamic class A", because the transistor really operates in class A but with dynamically varying bias. Other people have called it class AAB or class ABA, or various other ways. The supply current when idling is as low as in class AB, but with full drive it's as high as in class A. Consequently the maximum theoretical efficiency is 50% at full power, like in class A, but at partial power it drops in the less terrible way of class AB, instead of that of class A.



Please excuse me for drawing up these waveforms with my old hands, instead of doing it properly in the computer.

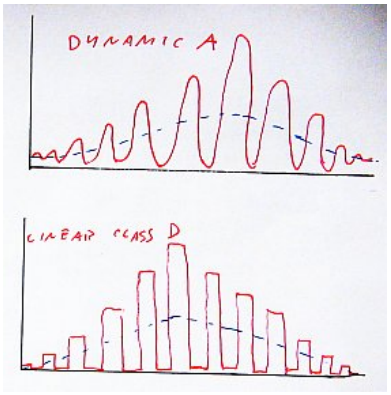
The first graph shows the base current waveform of BJT operating in class A, in red. The dashed blue line is the base bias current. The drive signal alternately adds or subtracts current from it, thus forming the total base current shown.

The second graph shows the base current of a BJT in class B. There is no bias current, only drive current, and the base-emitter junction rectifies this. For such an amplifier to work, there must be a path for the negative half wave of the drive current. Typically this would be the base of the companion transistor, in a push-pull amplifier.

And the third graph shows the base current of a BJT in class AB. There is a small bias current, and the drive signal adds a lot to it, or subtracts all there is. Again a path for the rest of the drive signal needs to be present.

Although it's not relevant to this discussion, I would like to add that class AB is used in practice as a replacement for class B, because class B suffers severe cross over distortion due to a transistor's gain dropping off too much at low currents. In the low current range the two transistors are made to overlap conduction, or a single transistor is made to also amplify some of the opposite half cycle's low current range, in order to at least partially correct this distortion. If transistors were perfect, class B would be linear, and class AB wouldn't be! Due to this distortion of real transistors, my drawn curves with those sharp corners are just wishful thinking, but they capture the idea behind the matter.

Now see what happens in dynamic class A: The bias current starts as low as in class AB, but is modulated so that the transistor never turns off. It always gets just enough bias to remain in class A at any signal amplitude, but just barely. At maximum amplitude the amplifier operates exactly like a class A one.



And the linear class D amplifier is pretty much the same thing, only that a square wave is used for driving. This improves efficiency, but only somewhat, because real world transistors do not switch fast enough to actually produce clean square waves at radio frequencies. Using class linear class D is an attempt at recovering some of the efficiency lost when being forced to move from class AB to an operating class that provides constant supply current over the RF cycle.

Some amplifiers operate with a sine drive at low drive levels, and the drive squares up as they approach full drive. This is easy to implement in BJT drivers, easier in any case than true squarewave driving. Such amplifiers are often called class AB-D amplifiers, or class BD amplifiers.

While I drew my crackly curves for BJT base current, they are also reasonably applicable to MOSFET gate voltage, with some changes: The horizontal zero line of the graph represents the gate threshold voltage of the MOSFET, instead of 0V, and in the case of class AB and B amplifiers the drive voltage waveforms aren't clipped on the negative side, but extend all the way to their normal rounded peaks, even if that means driving the gates negative. Of course that's theory, and in practice they get highly distorted.

Driving a transistor:

While the output side of BJTs and MOSFETs is similiar enough to treat them together, there are fundamental differences between driving a BJT and a MOSFET, so we always need to consider the particularities of each.

A BJT is fundamentally a current-controlled device, while a MOSFET is voltage-controlled. At frequencies low enough to keep parasitic effects low, a BJT's base voltage varies very little with drive, while it takes a relatively large base current. The base input impedance is a very low current-dependent resistance, in parallel with a relatively small capacitance. An RF power MOSFET instead needs from about 1V to several volt of RF drive at the gate, while presenting an essentially capacitive impedance. The gate behaves like a relatively large capacitor, with a small resistance in series.

As we increase the frequency, both for the BJT and the MOSFET the impedance values change. At sufficiently high frequencies the inductances of the gate or base terminal and the internal bonding wires start playing an increasingly important role, eventually becoming more important than the capacitance and the resistance. At such high frequencies the input impedances of BJTs and MOSFETs are much more similar than at low frequencies. For moderate to high frequencies it's best to consult the datasheet to find out the input impedance of a given transistor. In many cases it's given as a Smith plot, in other cases as a table of values for several frequencies.

Instead at low frequencies it's often good enough to consider a MOSFET's gate to behave like a capacitor, whose value is the gate-source capacitance plus the product of the gate-drain capacitance × the voltage gain of the particular amplifier (Miller effect).

A BJT operating at low frequencies can often be considered to have nearly zero RF base voltage for practical purposes, or a very low input resistance, along with a parallel capacitance that is often small enough to ignore. At RF the BJT will require a base drive current that is directly proportional to frequency, for a given RF collector current. Its current gain is equal to its transistion frequency divided by the operating frequency, down to such a frequency where the result equals the specified value for hfe. The current

gain at the operating frequency can be multiplied by the emitter resistance, given by 0.026 divided by the emitter current, to get the base input resistance. For example a small BJT, having an Ft of 500MHz, operating in class A at 30MHz, with a current of 0.05A, would have an emitter resistance of 0.52Ω, a current gain of 16.67, and thus a base input resistance of 8.67Ω in parallel with its base-emitter capacitance plus its Miller capacitance. Larger BJTs, operating at higher current, have an input resistance of a fraction of an ohm, which eventually becomes dominated by ohmic resistance of the connections, and is then dominated by the connection inductances. Since the current varies, the input resistance varies too - which introduces a factor of nonlinearity, making the BJT have a very nonlinear transconductance. The current gain, though, tends to be reasonably constant over a few decades of collector current, making BJTs relatively linear current-controlled devices.

Some people insist that BJTs are voltage-controlled devices, with a pretty clean exponential transfer function. This is true too, of course, but in most practical work it's easier to see the BJT as a pretty linear current-controlled device, requiring a high drive current and a very low, often negligible drive voltage.

It's important to understand that the calculated or datasheet-derived input impedance of a BJT is only the average value over the RF cycle. It varies dramatically over the cycle, due to the emitter resistance and thus the base resistance varying with the instantaneous emitter current. In any operating class other than A, the base-emitter junction stops conducting over a part of the cycle, causing dramatic impedance changes over the RF cycle. When driving the BJT from a resonant network, energy storage (flywheel effect) in this network tends to absorb these variations, but in low Q (broadband) situations the designer needs to be very aware of this variation.

Driving a MOSFET at low frequency is mainly a matter of charging and discharging its input capacitance. The drive current will then also be roughly proportional to frequency, but is 90° out of phase with the driving voltage. MOSFETs are highly non-linear voltage-controlled current sources. The drain current varies fundamentally with the square of the gate voltage change, at least over the lower and mid drain current range. Many amplifier circuits operate completely inside this very non-linear range.

Both with MOSFETs and BJTs, the manufacturers usually add internal, distributed source or emitter resistance. This serves to evenly distribute the current over the entire chip, but also linearizes the device by introducing negative feedback. Manufacturers select the value of this internal resistance according to the intended application of a transistor: Those intended for linear amplifiers get larger resistance values, improving their linearity, while those intended for nonlinear amplifiers get smaller resistors, which improves the efficiency and gain. Transistors advertised for both linear and high efficiency applications usually have an intermediate amount of resistance built in.

The transfer curve of such a ballasted MOSFET starts with the quadratic curve that's natural for all FETs, and gradually straightens out and then has a pretty linear high-current range. In a MOSFET designed for linear applications, the essentially linear range might start at about 20% of the maximum drain current. BJTs are reasonably linear current-controlled current sources. When seen as voltage-controlled, they become very non-linear, due to the exponential current-to-voltage ratio of a semiconductor junction, but since the base-emitter junction's voltage variation is pretty small over the whole current range, BJTs ballasted for linear service are very linear from a very low collector current onwards, maybe 2% of their maximum current, but they need significant drive voltage, unlike non-ballasted BJTs. In any case, the fact that a BJT's linear current gain range starts at a much lower current than even a highly ballasted FET's transconductance does, means that when operated in class AB a BJT needs far less idling current than a MOSFET, for a given degree of linearity. More about all this comes in the section about linearity.

All ballasting increases the saturation voltage, and thus has an impact on efficiency. It also reduces the power gain of the device, due to the degeneration it introduces. We can see it as built-in negative feedback.

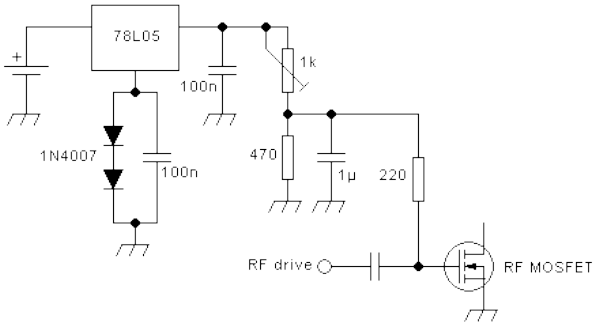
The input circuit of an amplifier stage needs to perform at least two, and often more actions:

- Biasing the transistors into the desired class, often providing temperature compensation of the bias point.
- Matching the input impedance of the transistors to the driving source, and in broadband amplifiers, often provide gain equalization over the frequency range.
- Reducing distortion, and further correcting frequency response, through negative feedback.

The feedback circuit may also influence the bias, and usually it has an important effect on input impedance. So the three functions are intensely interrelated, which makes it hard to discuss them separately. I will do my best, but there will be some mixing of them.

Biasing:

RF power MOSFETs are most commonly biased by applying an adjustable voltage to the gates. Depending on the power level involved, and the MOSFET's characteristics, this voltage might need temperature compensation. The gate does not draw any current from this bias source, so a simple potentiometer connected to a regulated voltage source can be used. It's important to properly design the bias circuit to have a suitable internal impedance all the way from DC to the highest frequency at which the MOSFET has any gain, and this impedance needs to be considered in combination with drive and feedback circuits to either hold the bias voltage constant through the RF cycle (class A, AB, C), or to vary it in the proper way (mainly for dynamic class A).



In this biasing circuit a small 3-terminal regulator provides bias voltage stabilization. Two diodes in series, thermally coupled in the best possible way to the power MOSFET, provide thermal compensation of the regulated voltage. These are slow diodes, so that they aren't prone to rectifying any RF current picked up by their leads. A bypass capacitor across them helps in this. The number of diodes can be varied according to the characteristics of the MOSFET used. It needs to be known that good temperature compensation can never be obtained, very simply because the diodes don't accurately follow the temperature changes of the MOSFETs die! Instead there is both a large attenuation, and a very large time lag in their response. So it is always necessary to design the amplifier in such a way that the MOSFET cannot go into a thermal run-away situation faster than the sensing diodes can react and reduce the bias voltage.

The regulated, roughly thermally compensated voltage is applied to an adjustable voltage divider, that allows setting the desired idling current for the MOSFET. Manufacturing tolerances of the MOSFETs always require the bias voltage to be adjustable. Since the regulated voltage is roughly 6.2V at room temperature, going down at a rate of 4mV/K due to the two diodes, with the divider values shown we can adjust the bias between about 2V and 6.2V at room temperature, with the thermal compensation

varying along with the adjusted voltage. In any practical application, the bias circuit should be tailored so that the potentiometer allows adjusting the MOSFET from fully off (gate voltage below the worst-case gate threshold) to the highest voltage that could be required to set the desired idling current in a particularly high-threshold, low transconductance sample of the selected MOSFET type. This requires a careful look at the MOSFET's datasheet. It is not desirable to make a bias circuit with a much wider adjustment range than necessary, because then it becomes more touchy to set the bias correctly.

Note that the voltage divider in this circuit is configured in a safe way, that is, if the potentiometer fails with its cursor contact opening (a very common failure!), the bias voltage goes down, not up, usually preventing the destruction of the MOSFET.

For further amplifier design, in the areas of driving and feedback, it's essential to understand what the impedance of the bias source is. In this example, over the whole RF range the 1μF capacitor is a good bypass, having negligible impedance, so that the bias source impedance is essentially just a 220Ω resistance. At DC instead this capacitor is an open circuit, and the bias source impedance is the 220Ω in series with the parallel combination of 470Ω and whatever value the potentiometer is adjusted to. So the DC bias source impedance varies between 220 and 790Ω, according to the setting of the potentiometer. And in the transition range between extremely low frequencies and RF, the bypass capacitor has a reactance comparable to the resistances of the circuit, and so the bias source impedance becomes complex. For example at 600Hz, with the potentiometer set midway, the voltage divider has a source resistance of 242Ω, and the capacitor has a reactance of -j265Ω . These two appear in parallel, and that combination is in series with the 220Ω resistor. The resulting bias source impedance is then 352-j121Ω, which is the same as saying 372Ω with a phase angle of -19°. This phase angle needs to be kept in sight, because it will causes a phase shift between currents and voltages at the gate, which detracts from stability. More about this will come further down, in the sections about feedback and stability.

The bias circuit above is just an example. There are many ways to do essentially the same job: Provide a regulated, temperature-compensated, adjustable voltage, with a source impedance that is known from DC to the highest frequencies at which the transistor has any gain, and which is acceptable for designing the rest of the amplifier.

In push-pull amplifiers it is often an advantage to have separate bias voltage adjustment for each MOSFET. But if very well matched MOSFETs are used, this isn't necessary.

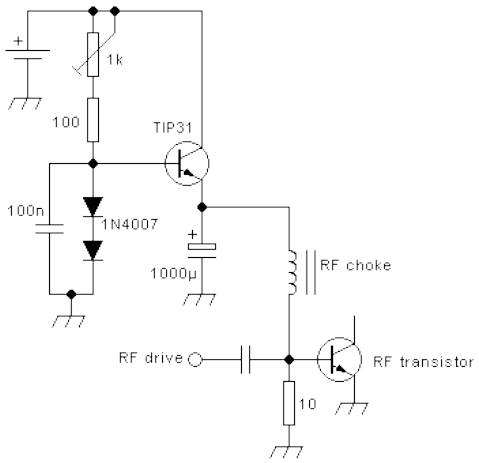
Sometimes active bias stabilization is used, mostly with class A amplifiers. Some sort of current sensor is used in the supply line, or in series with the source, which could be as simple as a low value resistor with a small BJT connected across it, and this is connected to the bias supply in such a way that the operating current is stabilized at a constant value. This eliminates the need for bias adjustment, and also for temperature compensation, at the cost of some more components. The circuit needs to be well designed to avoid instability. It's not very common in practice.

MOSFETs typically require some positive bias voltage even to operate in class C, in order to keep the conduction cycle long enough for acceptable power output, and in some cases also to avoid damage due to excessive negative gate voltage peaks. The latter is true particularly for LDMOSFETs, which typically tolerate far less negative than positive gate voltage. So the bias circuit for MOSFETs in class C is no different from that used in linear classes, except that it must be possible to set the bias voltage low enough under the MOSFET's conduction threshold to obtain the desired conduction angle. Most commonly such bias circuits are configured to allow adjusting the bias voltage between zero and about the highest expected gate threshold voltage.

Power BJTs used in linear service need a bias supply that can be regulated from about 0.5 to 0.8V, and that can deliver a high current. The current required, with some drive schemes, is as high as the maximum collector current divided by the RF current gain of the BJT (transition frequency divided by highest operating frequency). The bias voltage should remain very stable even while the bias current varies over its full range. This means that the bias source impedance must be extremely low, preventing the use of a resistor to inject the bias into the base. Instead an RF choke is required, making the bias source inductive over an extensive frequency range, which often causes stability trouble, mostly at very low frequencies.

The bias voltage absolutely must be compensated for temperature, to avoid thermal run-away of BJTs.

Many circuits have been used, and they are widely published. Designers in the 1970s loved to use the LM723. This IC contains a voltage reference, an operational amplifier, and a medium-current driver transistor, allowing to easily wrap around a current sensing diode, a potentiometer, a power transistor, and a few other parts like bypass capacitors. Of course it's also possible to build a purely discrete circuit, and this could be relatively simple:



The two diodes cause a voltage drop of roughly 1.4V, which varies according to diode temperature, and also according to the current passing through the diodes, which is adjustable. The bias transistor drops this by about 0.7V, depending on its own temperature, and the highly variable bias current drawn by the RF transistor. In a perfect world, we would thermally couple one diode to the RF transistor, and the other to the bias transistor, to get optimal temperature compensation, but in practice perfect thermal coupling is impossible, and so the usual approach is to thermally couple both diodes to the RF transistor, and mount the bias transistor on the same heatsink.

The 10Ω resistor has more than one function. One is providing a certain minimum load to the bias source. This is necessary because when the amplifier is idling, the RF BJT needs only a very small base current. At this small current the base-emitter drop of the bias transistor would drop significantly, producing significantly higher bias voltage. The result is that the bias voltage would be poorly regulated, making the amplifier run at excessive idling current. To fulfill this role, the resistor can be located either where I drew it, or in parallel with the 1000μF capacitor. But for stabiliy reasons its better to place it where I drew it. Otherwise the undamped inductance of the choke might lead to oscillation.

Another role of this resistor has to do with the impedance of the driving source, and operating class. If the driving source has a high output impedance, such as a transistor

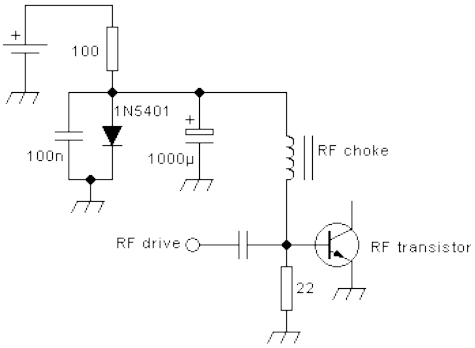
operating without much shunt feedback, it acts as a current source. Without the resistor at the base, it would combine with the rectifying action of the base-emitter junction, and the RF choke's effect of forcing a constant current over the entire RF period, to make the amplifier operate in dynamic class A! This does not happen if the driving source has a low impedance. A suitable resistor allows operating in class AB even when the driving source has a high impedance, and has no effect on this if the driver has a low impedance. Seen in a very simplistic way, it does that by lowering the otherwise high source impedance of a current-sourcing driver.

In addition, of course, the resistor has a swamping effect for the input impedance of the transistor.

The RF choke in such a circuit is a design problem. Its inductance needs to be low enough to be negligible at modulation frequencies, but high enough at RF. This inductance causes serious current/voltage phase shift, so that this choke is very often a big part of the problem when such an amplifer self-oscillates. Often the choke is de-Q-ed by using a very lossy core material, or by adding a resistor in parallel to the choke, which RF-wise has the same effect as the base-to-ground resistor, but does not consume additional current from the power supply. The inductance value of a bias choke typically needs to be pretty low, given the very low base impedance of RF power BJTs. This low inductance might resonate with the transistor's input capacitance, driver capacitance, and stray capacitances, well inside the operating frequency range, causing a high risk of instability. The important thing here is to always stay aware of the total impedance to ground that a signal on the base sees, over a very wide frequency range. The bias circuit is involved, the drive circuit too, also the feedback circuit if present, of course the RF transistor, and the base-to-ground resistor often saves the day by both lowering this impedance and pulling its phase angle closer to zero.

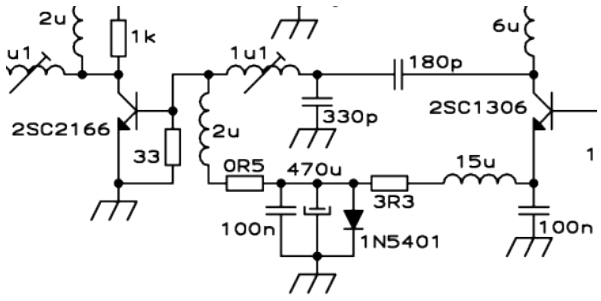
The large electrolytic capacitor provides a low source impedance down to low modulation frequencies. This is important because the bias transistor, operating as emitter follower, has a low source impedance only while actually delivering current. During operation of an amplifier, with its many parasitics and complex modulation, there could be moments in which the choke current gets very low, driving this transistor's emitter resistance way up, and the amplifier might enter self-oscillation at a low frequency if that electrolytic capacitor wasn't present. It's very often not possible to make the value of the base-to-ground resistor low enough to stabilize the amplifier in the absence of a big bias bypass capacitor, because the amplifier would end up having no gain!

In low voltage BJT amplifiers in the power range of about 1 to 20W it's somewhat common to use a further simplified bias circuit:



The disadvantage is that it constantly consumes a rather high current in the bias diode, relative to the amplifier's power, and also that adjustment of the bias level isn't very easy, because it would need a potentiometer with a somewhat highish power dissipation capability. So this circuit is mainly used with highly ballasted transistors, which are a little less critical in their bias regulation requirement. An experimenter can set the proper bias current experimentally by chosing a suitable diode. Larger diodes have a slightly lower voltage drop, at a given current. A transistor connected as superdiode is an option. In any case the current put through the diode must be higher than the maximal average base current needed by the transistor at maximum drive. Very often this results in a resistor lower than 100Ω, and capable of dissipating several watt. It's a simple circuit, but not really good.

When I was a young ham, still a student, and went backpacking with homebuilt radio gear and nickel-cadmium batteries, having an efficient radio was crucial. But it also had to be cheap! So I came up with the idea of using the current standing in the class A driver to bias my class AB final stage. The circuit ended up like this, seen in a cutout of the full schematic:



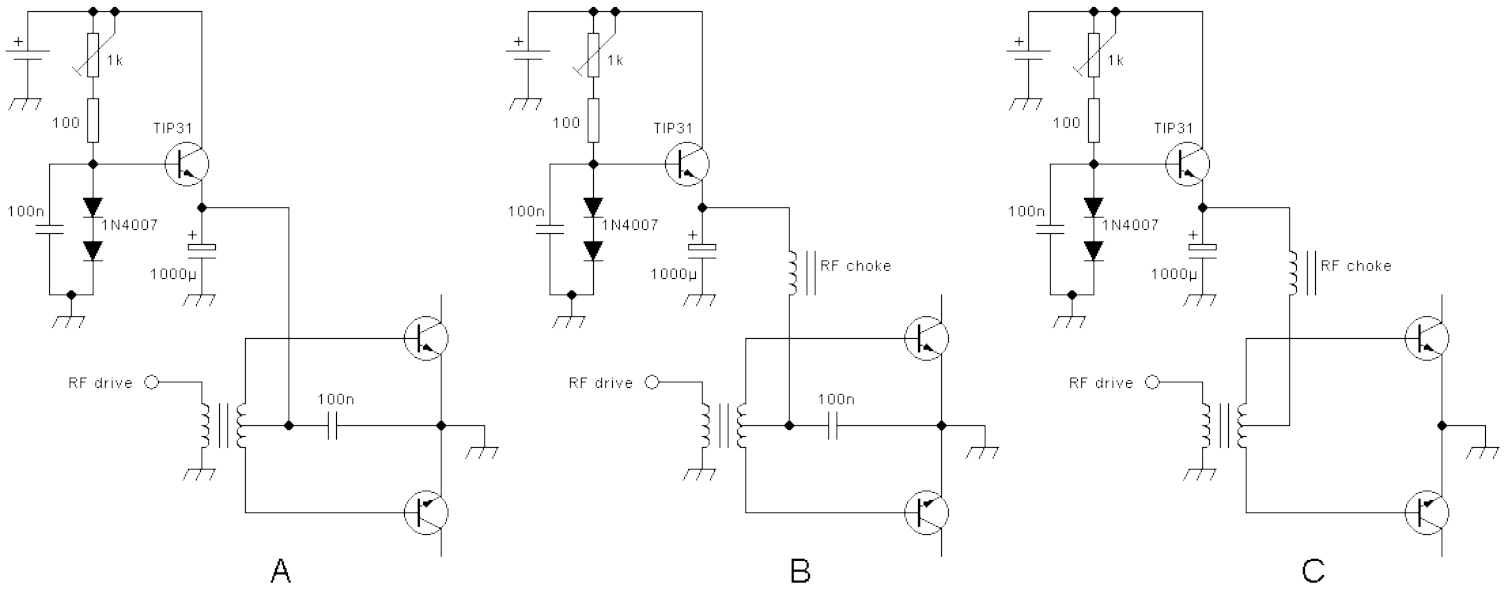
Never mind that signals flow right-to-left in this circuit. Back in those years I was fully omnidextrous... Anyway the emitter current of the 2SC1306 driver was put through the diode, instead of having to feed the diode from the 12V supply via a lossy power resistor. The 3.3Ω resistor is part of the driver's class A bias scheme - more about that later. The 0.5Ω resistor causes a sloping bias curve, which sacrificed a little linearity in exchange for higher efficiency, by moving the power amplifier's operating point from AB to B and even slightly into class C on modulation peaks. An added benefit of this trick is that turning on and off the driver stage will automatically turn on and off the power stage, which is great for very simple TX/RX switching. The driver's current was enough to bias the power amplifier in this radio mainly because the radio ran at just 7-7.3MHz, but used an RF transistor having an Ft of about 100MHz, so the RF current gain was relatively high and the base bias current required correspondingly low. If you want to see this radio in full, it's here.

BJTs can be very conveniently biased into class C operation by grounding the base for DC. If this is done with an RF choke, it typically needs to be de-Q-ed to prevent instability. Often the choke is just a single ferrite bead, made of a very lossy ferrite type, or else a low value resistor is placed in parallel with the RF choke.

It is very important to correctly decide how to inject the bias voltage into a BJT amplifier, in combination with the driving circuit. If it is done in such a way that the bias current can freely vary, or that the drive circuit can contribute current between the base and ground, class AB operation can be obtained. If instead the bias voltage is injected through an RF choke (or two), and the drive circuit isn't ground-referenced (differential drive in push-pull stages), linear class D or dynamic class A operation results. If an

RF choke is used, resistors from base to ground can moderate this, and put the amplifier into hybrid operation between class AB and class D, or even so close to class AB that the remaining dynamic class A effects are absorbed in the idling current band.

Consider the following three circuits of push-pull amplifier biasing circuits, which differ only in the bias injection method:



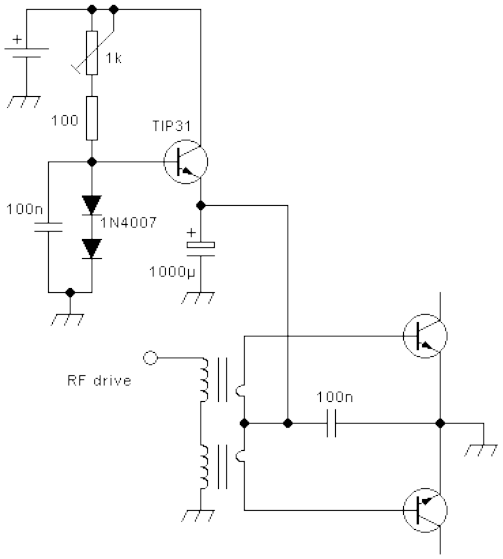
In circuit A the center tap of the drive transformer is firmly held at a fixed voltage of about 0.7V, by the combination of a local RF bypass capacitor, a large electrolytic capacitor, and the bias voltage regulating circuit. No matter what the drive signal does, the center tap of the transformer will stay put at 0.7V (let's disregard for now any small base-emitter drop changes with current). As a result of this, and assuming that the drive transformer has excellent coupling between the two sides of its secondary, the voltage on the bases of the two power transistors will always vary in a symmetric way. If one base voltage goes up a little bit, the other base voltage will go down by the exact same amount. The current drawn by the bases is free to vary. The instantaneous current put into the base of the transistor that is conducting at a given time will depend on the drive current and the transformation ratio, and it will ultimately come from the bias regulator, flow through one half of the secondary, and on into the transistor's base, with bypassing/smoothing provided by the capacitors, so that the TIP31 doesn't need to vary its current at the RF rate (it cannot, since it's too slow!). The other RF transistor's base will conduct no current at all, because it's geting less voltage than its conduction threshold, and so that side of the transformer secondary will carry no current. Of course in the next half cycle this reverses.

This circuit allows to cleanly operate the amplifier in class AB, or class B, or even class A if you want, just by properly setting the potentiometer, and perhaps picking different diodes if necessary.

In circuit B an RF choke has been added. This choke forces a constant current in it over the RF cycle, but allows current changes at the modulation rate. And the bypass capacitor at the transformer's center point defeats the RF-rate current regulation provided by that RF choke, so that circuit B really operates just like circuit A! The RF choke is fundamentally superfluous, and the only reason why some designer might want to use one is if the bias regulator is too far away from the amplifier block, so that there is significant ground RF potential difference between them. Anyway this is an unclean circuit, because any reactance that the RF choke might still have at the modulation frequencies will cause the bias voltage at the power transistors to vary with modulation, and that will cause distortion. Circuit A works cleaner, and is also less problematic in terms of stability.

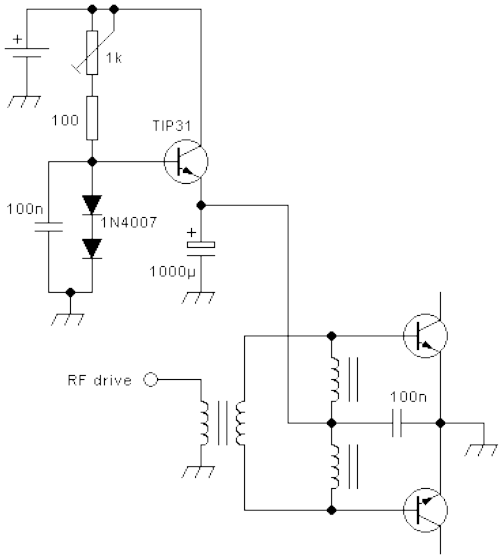
Circuit C operates in a totally different way: Since there is no bypass capacitor at the drive transformer's center point, the constant current forced by the RF choke over the RF cycle also forces the sum of the two base currents to stay constant over the entire RF cycle! The total base current can only vary at the modulation rate, but not at RF rate. So the total collector current will do the same. The drive signal will only act to distribute the total base current between the two transistors, but cannot vary it inside the RF cycle! The result is that this bias circuit cannot produce class AB, B, nor class C operation. It can only be used for class A, including dynamic class A, and for current-switching class D, including linear class D. Whether it operates in class A or class D depends just on the drive signal: A sine wave for dynamic class A, and a square wave for class D.

Now I get to one of the biggest pitfalls in linear push pull amplifier input circuit design. It's the exact same that is common at the output side: Poor or non-existing coupling between the two halves of the transformer's winding!



This circuit shows such an input transformer that has no coupling between the sides. It's what you get when you take a twin-hole ferrite core, and wind a single turn for the low impedance winding. The two magnetic circuits do not couple, and act like two separate transformers. Exactly the same as in output transformers, described roughly five kilometers up this page. An inexperienced designer will very likely think that his amplifier using a single-turn, center-tapped secondary winding drive transformer will behave like circuit A above, but **no**, it behaves like circuit C instead! The designer intended to make a class AB amplifier, and instead he gets a dynamic class A amplifier. Instead of the expected 65 to 70% efficiency he gets only 40 to 45%!

The same happens, of course, if he uses differential drive, without any center tap, and injects the bias through separate chokes:

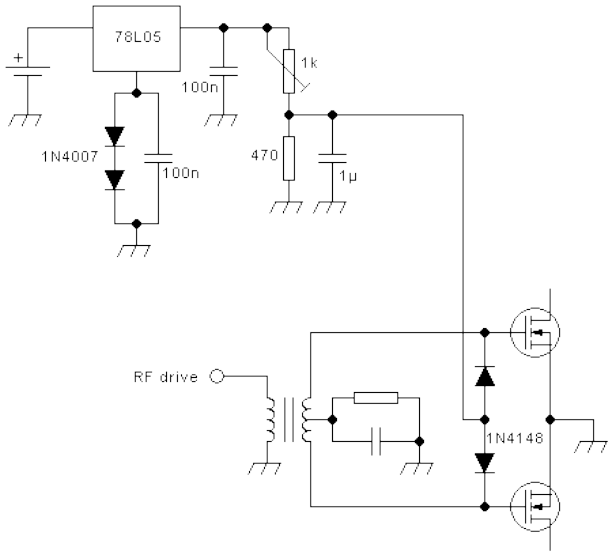


It's just like with the output side of an amplifier: No cross-coupling between sides, cannot work in class AB. The total base current cannot vary over the RF cycle, allowing only class A or class D service.

Just like in the single-ended amplifier, adding resistors from each base to ground reduces the problem, and makes the amplifier operate in an intermediate mode, much like class AB at low drive levels, and more like dynamic class A at higher drive. Many amplifiers exist that operate in this way. They work, but they never have great performance.

Anyway BJT are obsolete for power RF use, so let's return to **MOSFETs**, and see how to bias them for **dynamic class A**! You may think that it's crazy to do so, given that class AB has better performance than dynamic class A, and MOSFETs lend themselves very easily for clean class AB biasing! The problem is on the output side: As described about five and a quarter kilometers up in this web page, it is extremely hard to achieve good cross-coupling between drains in high power, low voltage amplifiers operating in the range of 100W to kilowatts. Hard enough to be practically impossible, if operation up to the end of the HF range, and even beyond into low VHF, is required. This can force a designer to intentionally run his circuit in dynamic class A, and just accept the lower efficiency obtained, or he might square up the drive signal and run the amplifier in class D. To do so, the basic MOSFET biasing circuit needs to be adapted in some way, to modulate the bias voltage according to the drive signal's amplitude.

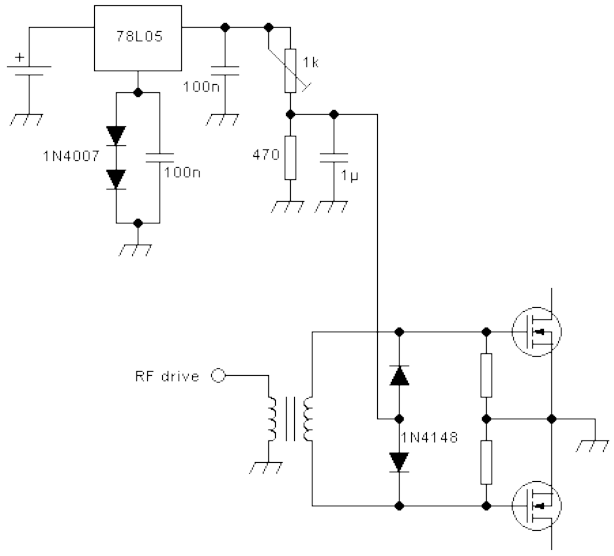
This can be done by common-mode feedback, which I will describe in the section about feedback, or it can be done by rectifying some of the drive signal. BJTs do the latter by themselves, in their base-emitter junction, automatically producing dynamic class A bias if allowed to do so. MOSFETs don't rectify anything at their gates, so the rectifiers have to be added externally. This can be done by injecting the bias voltage through diodes, like this:



In this circuit the gate voltage cannot ever go lower than the voltage at the divider minus the diode drop. So, while there is no drive signal, that's the voltage that will appear. But when there is any drive signal, that's the voltage that will be on each gate during the negative peaks of the drive signal, which means that the actual bias voltage will get a modulation equal to half the peak-to-peak value of the drive signal. The RC pair at the drive transformer's center tap has a cutoff frequency that's much lower than the lowest RF frequency to be amplified, but much higher than the highest modulating frequency. So the centerpoint voltage, which is the effective bias voltage, follows the modulation envelope, but does not follow the RF signal, producing dynamic class A biasing.

Note that since the center tap only needs to conduct low frequencies in this circuit, this works well even if a twin-hole core with a single turn is used.

But we don't even need a center tap! We can equally well place an RC pair directly at each gate. And we don't even need to physically add the capacitor, because the MOSFET gates are very capacitive to start with! So the circuit morphs into this:



The gate to ground resistors need to be dimensioned to give a suitable cut-off frequency with the MOSFETs input capacitance, so they are totally dependent on the MOSFETs used, and both the RF and modulation frequencies. These resistors also have to be included in the calculation of the bias voltage divider. The diode drops too, of course.

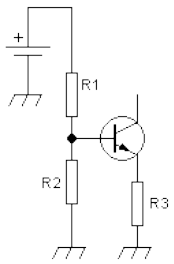
In a practical circuit usually some more resistors are added in various places, to tailor the bias modulation to the optimal compromise between distortion, efficiency, drain overvoltages, and so on. Various capacitors, chokes, beads, etc, might need to be added for stability reasons. If separate bias control for each MOSFET is desired, then one can add capacitors in series with the transformer secondary, to break to DC path, and then feed each diode from a separate voltage divider. And the bias voltage source could of course be totally different from the one I'm showing in this chapter. Since the gate to ground resistors might be relatively low, a bias voltage source with a higher current capability is often required. There are many possibilities, and since this is more a matter of power supply design than RF amplifiers, I will control my urges and force myself to abstain from presenting you two dozen circuit ideas for this...

People casually looking at such bias circuits with diode injection usually don't understand how this works, and why the diodes are present. They just miss too many details to understand the circuit: They don't "see" the capacitance of the gates, they don't know about dynamic class A, and so on. So they often think that the designer's intention, when he added those diodes, was to protect the second MOSFET when the first one fails with a drain-gate blow-through. But this is totally wrong! Designers usually don't care whether or not in the event of a MOSFET failure the other one will fail too, because anyway both will need to be replaced, both for the sake of matching and because the still "good" one has very likely been degraded by the event that killed its companion. The diodes are definitely there to implement a dynamic, modulated bias, not for protection!

Let me warn you that when studying published circuits you may also come across some totally wrongly designed ones. It's very common that someone tries to copy some design he doesn't fully understand, and makes "small" changes which actually end up defeating the very purpose of the circuit he is copying!

To round up the section about biasing, I will quickly mention the biasing of small driver stages. While in higher power class AB and dynamic class A stages it's pretty universal to have potentiometers to set the optimal bias, it would be cumbersome and unnecessary to have bias adjustments for every small stage. But some bias schemes used in small-signal amplifiers don't lend themselves optimally for the typical pre-drivers, which are no longer really small-signal, but are not really power stages either. I mean stages running at an output power level of about 10mW to 1W, which are usually operated in class A.

With BJTs it's very common to use just three resistors to bias them for class A, in a pretty stable and reproducible way:



R1 and R2 form a voltage divider, setting the base of the transistor at some convenient voltage, typically something between 1 and 4V. Then its emitter is about 0.7V lower, giving a pretty well defined voltage across R3, so that the value of R3 determines the idling current in the transistor. The absolute values of R1 and R2 are selected so that a much larger current flows in them, than the expected base current will be. This allows getting a pretty reproducible base voltage regardless of the exact current gain of the actual transistor, and that's important because the current gain of BJTs has a very wide tolerance.

The higher a base voltage is chosen, the more stable will the idling current be when the transistor changes temperature, because the thermal change of its base-emitter drop will be less significant relative to the voltage on R3. But a higher base voltage also means that the maximum possible collector voltage swing gets smaller, reducing the amplifier's maximum power output and efficiency. So this is a tradeoff that the designer must make according to his goals.

In practical circuits part or all of R3 is often used to create negative feedback through emitter degeneration. Part or all of R1 is also often used to create negative feedback, by connecting the upper end of R1 to the collector instead of the supply. More about this will come in the chapter about feedback.

This same circuit can in principle also be used with small MOSFETs, but this is less often done, because the gate-source voltage required for linear operation of a MOSFET is several times larger, and also far less predictable, than the base-emitter voltage of a BJT. So it would be necessary to use roughly 6 to 10V or so at the MOSFET gate, and this makes the approach unsuitable for typical low voltage circuits, typically operating from something between 5 and 15V. But it's a perfectly good method to use when running small MOSFET drivers from 30V or more.

Small MOSFETs running from low voltage can still use this circuit, if a MOSFET with very low and pretty stable gate threshold voltage is used. In other cases small MOSFETs might need a potentiometer for bias setting, just like their big brothers, or an active biasing circuit. Since this is a chore, and anyway small BJTs have far better linearity than small MOSFETs, and are still plentiful and inexpensive, almost everybody uses BJTs in such small predriver stages - or choses the lazy option of using IC RF amplifiers, instead of designing his own discrete circuit!

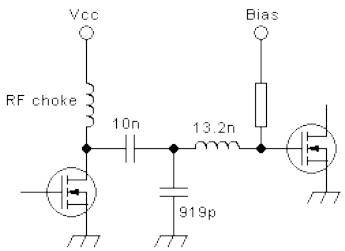
Input impedance matching:

The same methods and circuits that are used for output impedance matching can be applied to the input. That is, single-band amplifiers can use tuned networks like the L network, the LCC to increase the Q, or a series of several L networks to implement the required matching with a lower Q. The same equations given for output matching networks are used, but since the transistor's input impedance is usually lower than the driver's output impedance, the networks are turned over: The parallel element of an L matching network is placed on the input side. In VHF and higher amplifiers, the inductances are often implemented as microstriplines.

As an example let's consider a monoband 52MHz amplifier using a MOSFET that under the expected operating conditions has an input impedance of 0.8-j1.2Ω, glimpsed from the datasheet. This value means that it behaves like a 0.8Ω resistor in series with a 2.55nF capacitor. And the driver for this stage needs to be loaded with 13Ω, free from any reactance. So we need a matching circuit that transforms 0.8Ω into 13Ω at 52MHz, while also absorbing the capacitive reactance of the gate.

Any online L network calculator can quickly spit out that to match 13 to 0.8Ω at 52MHz, a capacitor of 919pF and an inductor of 9.6nH is required. But the series equivalent capacitive reactance at the gate needs to add an equal inductive reactance to the inductor, which brings its value up to 13.2nH. The online calculator can spit out that change too, or it can be calculated the old-fashioned way, with a pocket calculator or even a slide rule, if you are old enough to know what that is.

The matching circuit and its surroundings then looks like this:



The 10nF capacitor is there just to block the DC. The 919pF capacitor would typically be either a trimmer of about 1200pF maximum capacitance, or more likely a parallel combination of a smaller trimmer and one or more fixed capacitors. The inductor could be implemented as a microstripline (a glorified PCB track), which looks clean and is

cheap to make in mass production, or as very tiny coil having 1 or 2 turns of wire. The coil is easy to adjust by bending it, while the stripline can only be adjusted by changing its dimensions, robbing us of one adjustment point. On the other hand, changing the stripline's length can be as simple as soldering the capacitors to it a little closer or farther from the transistor.

The bias injection resistor can double as a load resistor, to dampen down unwanted resonances and thus avoid instability. Of course the top side of that resistor is bypassed to ground, etc... but here we are talking about the matching circuit!

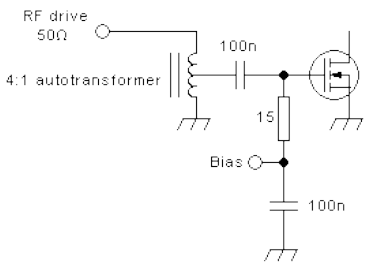
Broadband amplifiers can use conventional transformers or transmission-line transformers. The conventional transformers are often built using small two-hole cores with tube or braid secondaries, and a few turns of wire as primary winding. For low power stages, twisted wire pairs or trios wound on two-hole cores or even toroids are used. As already mentioned in the biasing section, the same problems with lack of side-to-side coupling in push-pull amplifiers that happen in the output side also happen at the input, but since the input transformers are physically smaller, these problems are somewhat easier to avoid, and it's also quite possible to drive a push-pull amplifier in differential mode, with no center tap at all on the transformer.

In most cases of broadband amplifiers it's necessary to compensate for the capacitance in the transistor's input impedance. This is best done by absorbing the capacitance, and also the lead inductance, into a lowpass section of sufficiently high cut-off frequency. In some cases, specially when UHF transistors are used at HF, such compensation isn't necessary, and instead the input capacitance is swamped by connecting a low value resistor from the gate to ground, or better by providing strong negative feedback. The resistor value, or equivalent load resistance offered by the feedback circuit, in parallel with the transistor's internal parallel-expressed resistance, needs to be reasonably low compared to the reactance of the transistor's parallel-expressed input capacitance, at the highest operating frequency.

Let's see examples of this, starting with the swamping technique. Assume that you want to use a MOSFET that has an input capacitance of 100pF, in a 10W amplifier operating from 1.8 to 30MHz. The parallel equivalent resistance in the input impedance will be high at the low end of the frequency range, so that the MOSFET itself needs almost no drive power there, while at the high end of the range the parallel equivalent resistance will be a lot lower. Since the input capacitance is roughly constant, the reactance of this capacitance will also vary dramatically over the frequency range. These variations are what makes gate swamping so attractive! It allows a much more constant input impedance for which to design, and it helps a lot in stabilizing the amplifier, although it also wastes some drive power.

At 30MHz, the 100pF input capacitance has a reactance of 53Ω. At 1.8MHz it will be 884Ω. The resistive part of the input impedance tends to be reasonably close to the reactive one, and it's very often not given in the datasheets for all the frequencies we need, so we have to guess or measure, or just be lazy: If we simply plant a 10Ω swamping resistor across the gate input, then over the whole range the total impedance won't change a lot! If we assume the MOSFET to have roughly about the same input resistance as it has reactance, at any frequency inside this range, then with our 10Ω resistor added we might get a total 9.9Ω with a phase angle of roughly -1° at 1.8MHz, while at 30MHz it will be around 8.3Ω at an angle of -9°. This is close enough to consider the input impedance to be "essentially flat" in practice, over our whole frequency range!

But this input impedance of roughly 9Ω average might not be the most convenient. If we need the amplifier to be driven from a 50Ω source, for example, it would be better to have a more convenient impedance ratio than 5.5:1. So we can tweak our swamping resistor to get what we want: To use a simple 4:1 input transformer, we might use a swamping resistor having about 14Ω. A more available value of 15Ω is close enough, in practice. And we can use this resistor as bias injection resistor too. And then our broadband matching circuit can end up like this:



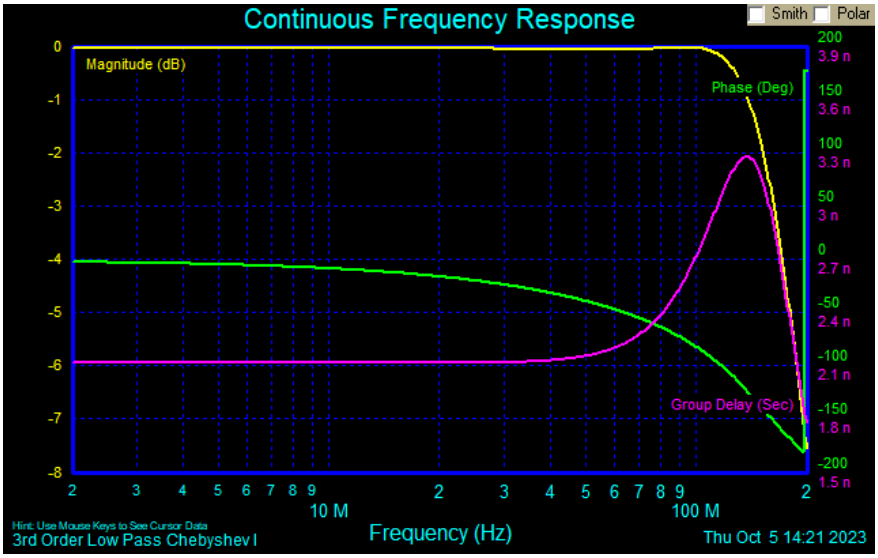
As you can see, it's nice and simple. The autotransformer used here can very easily be wound with a few bifiliar turns of thin wire on a small ferrite toroid or bead. The coupling and bypass capacitors need to be large enough to have a low impedance relative to the roughly 13Ω load present, at the lowest operating frequency. The 100nF shown here is about the minimum suitable value.

A circuit like this does have some droop in gain. At higher frequencies the gate input impedance drops a lot, so that even while swamped by the resistor it does have an effect. Also the gain of transistors, even MOSFETs, drops with increasing frequency. In many applications the droop is acceptable when it is as small as in this case, and it is also possible to further reduce the droop by simply using a lower value of swamping resistor, changing the transformer's ratio accordingly to the new, lower, total input impedance, and accepting the fact that now an even higher drive power is required. That means: Brute-force gain flatness improvement costs gain!

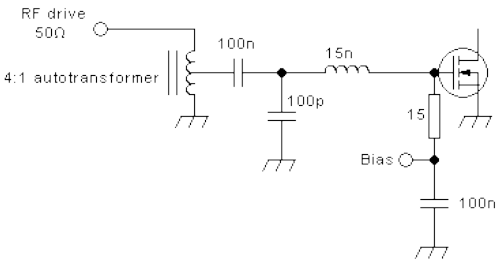
There are some better ways to improve the gain flatness. These become really important when a very wide frequency range is needed, or when the MOSFET used does not have a very high maximum frequency. As already indicated above, one such technique is to absorb the MOSFET's input capacitance into a low pass filter, whose function isn't really any filtering, but simply the seemingly magic disappearance of the problematic high input capacitance of MOSFETs!

For example, let's assume that we want to extend the operating range of our amplifier to 54MHz. Without compensation, and with a15Ω gate swamping resistance, the gain of the stage would be down by a few dB at 54MHz, and the input SWR would begin to be on the poor side. So all we have to do is design a π lowpass filter having a 13Ω impedance, a flat passband, and using 100pF capacitors on each side. If we can do that without this low pass filter cutting off any frequencies we need, then we can use this method!

Low pass filters can be designed the hard way, by lots of manual calculation, or the lazy way, by using software. Take your guess which way I use... Plugging the values into the software, I come up with a low pass filter using a 21.4nH inductor, that has a flat bandwidth of 107MHz, so it's plenty good enough! Its response looks like this:



The magnitude response is essentially flat to well above the 54MHz we need. The phase and delay aren't flat, of course, but these are irrelevant as long as we don't include the filter in a feedback loop, and our signals are all narrow-banded. So we can use this filter in our drive circuit:

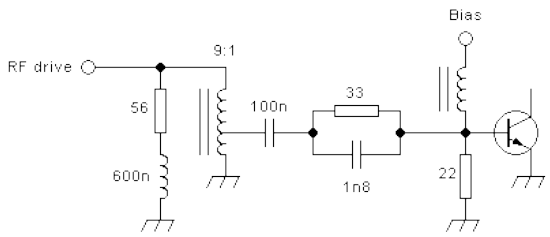


You surely spotted right away that I used a 15nH inductor, instead of 21.4nH. Why? Well, because the MOSFET has some lead inductance! We can absorb that into the low pass filter's inductance, to make it disappear too! The exact value of the inductor will depend on the PCB layout. Often it's easiest to simply use a longish and narrow trace between the MOSFET gate terminal and the 100pF capacitor, calculating it to get the optimal inductance. Then the circuit on paper can be drawn as having no inductor there, but just a capacitor from the gate to ground, and people who don't understand this compensation technique wonder why on earth the designer shunted the MOSFET's disturbing input capacitance with even more capacitance... So, now you know why there are so often capacitors connected directly between gate and source, and often also drain and source! The trick is the transistor's lead inductance! When working in the VHF range, very often even the lead inductance is more than the compensating lowpass filter needs, and so the leads need to be shortened, or the capacitors need to be soldered extremely close to the transistor's body, right where the leads come out.

The larger the lead inductance is, relative to the total lowpass filter inductance, the less well is the swamping/loading resistor placed. This worsens the performance of the circuit. That's why I drew the resistor connected so close to the MOSFETs body! In the example shown here, it still works pretty well, but in VHF and UHF circuits it often makes little or no sense to use any swamping resistance. Anyway at such frequencies the internal resistance of the MOSFETs input is pretty low, and typically such amplifiers are tuned to a single band, so things are all different. Absorbing the gate capacitance into a low pass filter is a technique mostly useful for broadband amplifiers used at HF and into low VHF.

In high power BJT input circuits it was common to use **frequency response compensation** with various arrangements of RC and RL groups, to achieve acceptable gain flatness along with acceptable SWR at the input. This was very necessary, because BJTs have a far more drastic gain droop with frequency than MOSFETs do, and also because in the age of BJTs we often built 1.8-30MHz amplifiers with BJTs having an Ft of barely 100MHz. So the current gain of those transistors was only about 3 at the highest frequency, but over 50 at the lowest frequency! And with power gain being proportional to the square of current gain, that made for a massive gain difference over the frequency range.

A typical BJT frequency compensating drive circuit could have looked like this:



Please don't take the component values given in this schematic as a golden recipe! They need to be optimized for each particular application, depending on the transistor used, power level, frequency range, desired input impedance, tradeoffs between gain flatness and basic gain, and so on. In any case, the idea is that at high frequencies the 600nH inductor has a high reactance, and the 1.8nF capacitor has a low one, so that most of the drive power gets to the transistor, and the autotransformer provides the proper match. At low frequencies instead the 1.8nF capacitor has a high reactance, so there is far less drive current, with most of it going through the 33Ω resistor, the autotransformer sees a relatively high load impedance, and the 600nH inductor has a low reactance, so that the 56Ω resistor can load the drive source to keep the total input

impedance of the amplifier close to 50Ω. The 1.8nF capacitor also introduces a pretty strong capacitive reactance into the circuit's input impedance, while the 600nH inductor introduces a suitable inductive reactance, and if all the component values are properly chosen to work optimally together, then the input impedance will not stray too much from a purely resistive 50Ω.

Such a circuit can be fully calculated, but it's a bit complex. It can also be pre-optimized empirically in a circuit simulator, with the final optimization being done in the good old style, with a soldering iron and a good selection of parts to try. Since these days we rarely use BJTs for RF power amplification, we can probably let this matter rest. It's useful mainly to understand how old, existing equipment works.

Small BJTs are still often used in drivers and pre-drivers below the 1W level. So it's important to understand how to drive these. The input capacitance of such transistors is usually very small, and can be neglected in most cases. But the BJT's input resistance varies dramatically over the frequency range, because it depends on the transistor's current gain, which in turn varies inversely with frequency. The input resistance is the emitter resistance multiplied by the current gain. The emitter resistance is roughly 0.026 divided by the collector current, for silicon transistors, and the current gain is the transistor's Ft divided by the operating frequency. So, if we are building a class A predriver that will operate with an idling current of 30mA, its emitter resistance is

0.026 ÷ 0.03A = 0.87Ω

This is the internal emitter resistance at this current. If the amplifier has any additional external, non-bypassed resistance in series with the transistor's emitter, then this resistance has to be added to the internal one, for all further input resistance calculations.

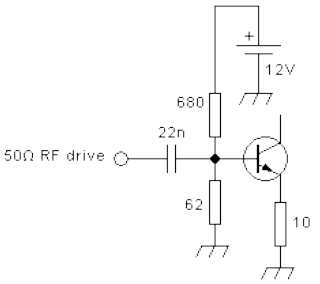
If we are running it at 30MHz at a given time, and the transistor has an Ft of 500MHz, then its current gain is 16.7, and if there is no external emitter resistor, its base input resistance ends up being 14.5Ω. But at 1.8MHz the current gain may be as high as 278, although it will more likely be clipped by the DC current gain (hfe), to a value typically anywhere between 50 and 200. So the base input resistance at 1.8MHz could be anything between about 43 and 170Ω.

Since the input resistance is so variable, we need to swamp it in some way, to get a reasonably stable total resistance to which we can match. This is often done by selecting low value bias resistors, but it's even more common to do it with negative feedback. We will get to this soon!

Often it's a good idea to add a few ohm of external emitter resistance, to drive the transistors input resistance to a value well above 50Ω. Then we can swamp it with a 51 or 56Ω resistor, and get a direct match to a 50Ω source, without any further effort. The cost of this is getting lower gain, but there is also the benefit of better linearity.

The base input capacitance of such small BJTs is just a few pF, so the capacitive reactance is very high compared to the resistance, and we usually don't need to care about it. As a result, we can match the input of our pre-driver stage to the signal source by any of the common techniques, such as autotransformers, plain transformers, transmission-line transformers, or tuned matching networks in the case of single-band amplifiers.

Here is a very simplistic and somewhat stupid little example of a directly matched small amplifier's input circuit:



The bias voltage divider provides 1V at the base, assuming that it isn't loaded by the base current. Assuming 30mA collector current, and an hfe of 100, there are 0.3mA of base current. Relative to the 16mA flowing in that voltage divider, 0.3mA is very little, so we will indeed get very close to 1V at the base. Just a tad less. The emitter will then be at pretty precisely 0.3V, and the 10Ω resistor will indeed set the emitter current to 30mA. So we have done the biasing part for 30mA.

Then comes the RF part: At 30mA we have 0.87Ω internal emitter resistance. Plus the external 10Ω , it's 10.87Ω. Given that this transistor has an Ft of 500MHz, at our highest (worst case) frequency of 30MHz the current gain is 16.7, so the base input resistance is 182Ω. The bias voltage divider has a source resistance of 57Ω. In parallel with the 182Ω, that's 43.3Ω total input resistance. Close enough to 50Ω for direct connection. And at 1.8MHz the transistor's input resistance will be at least 550Ω or so, which in parallel with the divider's resistance is just a little above 50Ω. Close enough for direct connection too! So we don't need any further impedance matching - if we wanted a 50Ω input impedance, of course!

So here you got your first example of a circuit that shares three resistors between biasing and impedance matching functions! Each of those three resistors is used for each of the two functions.

Note that in this circuit I changed the coupling capacitor to 22nF, for no other reason than to remind you that coupling capacitors don't always need to be 100nF! :-) 22nF has a reactance of just 4Ω at 1.8MHz, which is still low enough for good coupling in a 50Ω circuit.

Small MOSFETs are very different animals in this regard. They have a high input capacitance, with very little series resistance. So we basically need to drive a capacitor. This could be 50pF or more, depending on the MOSFET used. This usually cannot be neglected, particularly because there is so little resistance present. So the usual practice with broadband drivers is to swamp it with a resistor, or preferably negative feedback, giving a total input impedance so low that the capacitance remains a reasonably irrelevant part of it even at the highest frequency of operation. But it is also possible to use the low pass filter compensation technique, and this results in much greater gain, since a lot less swamping is needed.

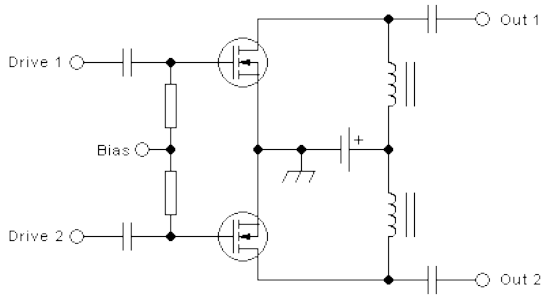
Since MOSFETs are also harder to bias to a stable operating point, and are less linear than BJTs, my advice is to avoid them, and stick to BJTs for small driver stages, as long as such BJTs remain available.

When driving a **push-pull** stage its important to understand what combination of differential-mode and common-mode drive you are implementing. I think it's again time for a Blue Block!

The split personality of push-pull amplifiers

Single-transistor amplifier stages are straight and honest: What goes in at their single drive input, comes out (hopefully amplified and clean) from their single output. But push-pull amplifiers aren't like that. Having two inputs and two outputs, at least internally, they can do all sorts of mischief - and they love to do it! What you put in at only one drive input might come out from only one output, or only the other one, or both in phase, or both in opposite phase, or not at all. What you put into both inputs in the same phase offers the same range of options for where and how it will come out, if at all. Only if you put the same signal into both inputs but in phase opposition, do you have a pretty good chance that it will come out the same way, that is, from both outputs and in phase opposition. Of course it might come out larger from one output than the other, which means that part of it is coming out as it should, while another part is coming out from only one output... You get the idea! You really need to watch them closely, or they will catch you by surprise.

Consider the following, very simple, basic and stupid circuit:



It's a push-pull amplifier that's built basically as two completely separate single-ended amplifiers, that only share the power and bias supplies. As long as we don't interconnect the inputs and outputs in specific ways, this amplifier is capable of being mischievous. Any signal that you apply between the two inputs will come out at both outputs in opposite phase, so that a load placed between the two outputs will see it. But if you place the same signal at both inputs in parallel (that's called a common-mode signal), it will also come out at both inputs in the same phase, which means that a load connected between the outputs won't see it. This is fine, since a good push-pull amplifier is supposed to suppress common-mode signals - but only until you realize that each of the two transistors will still see all of the signal, will amplify it, but will see no load, because the load gets the same signal at both terminals! So both transistors will put out a full-amplitude, clipped, square wave voltage, even if the drive signal is small and a sine wave! In practice this means that if you have such an amplifier, with a load between its outputs, even the smallest common-mode input signal applied together with the main differential drive signal will make it saturate like crazy, and distort the main signal! In other words, your drive signal needs to be perfectly symmetric. And perfection doesn't exist in real-world electronics.

Fortunately there are usually so many imperfections present, that they tend to make the circuit more tolerant. One imperfection gives way where another is pushing. For example, the chokes used to feed the drains will be lossy, which means that they act like if there were resistors in parallel to each of them. These resistors form an additional load on the amplifier's outputs, which is independent on each side, and thus can absorb small common-mode output currents and keep the amplifier from saturating on them.

The crucial thing to always keep in mind when working with push-pull amplifiers is that they have two modes of operation: Differential mode (the desired one), and common mode (the parasitic, usually undesired one). You always need to analyze what happens in each of the two modes, to reduce the number of surprises a push-pull amplifier gives you. If you apply two drive signals in opposite phase, as is normal, the two transistors will work in opposed phase, and a load placed between the two outputs will get the output power. Instead if you apply a drive signal in the same phase to each side, both transistors will work happily in the same phase, the differentially connected load will get nothing, but the transistors might actually kill themselves from the inductively generated voltage spikes of the drain feed chokes, which have nowhere to go except straight through the transistors! End even if you apply perfectly differential drive, but the inevitable imperfections in the amplifier turn part of that into a common-mode signal, you can get those damaging common-mode voltages!

Differential and common mode applies not just to drive and output signals, but also to feedback, and to self-oscillation. Even if a push-pull amplifer is driven in a perfectly balanced way, amplifying in the desired differential mode, and loaded in a balanced way, it might self-oscillate in common mode, because the designer forgot to make sure that the common-mode stability is adequate.

Center-tapped push-pull transformers, or balanced (bifiliar) feed or bias chokes, couple the two sides together in opposite phase, preventing significant common-mode voltages from appearing - at least as long as these transformers actually work as intended, which is often not the case...! But designers often forget that electronics is not just a world of voltages, but it's just as much a world of currents! When you suppress common-mode voltages by using such means, usually you are giving totally unrestricted way to common-mode currents! So even if you have excellent side-to-side cross-coupling in a push pull amplifier, you still need to do the common-mode stability analysis, and it's good to learn to "see" voltages and currents in both the differential and common modes.

Feedback:

Negative feedback is a very powerful tool to reduce all kinds of distortion, stabilize the bias and the gain, flatten the frequency response, and make an amplifier's performance largely independent from the specific transistor's characteristics. Some amateur designers are frightened by it and try to avoid it, but in fact all amplifiers have some unavoidable negative feedback! If it's not explicitly implemented in the circuit, it will happen through the transistor's internal output-input capacitance, and through degeneration from internal and external inductance and resistance in the emitter or source connection. Undesired capacitive and inductive feedback in the external circuit is also always present to some extent. Most of these natural negative feedback paths are highly reactive, and thus facilitate self-oscillation, by shifting the phase in a frequency-dependent manner. Intentionally applied feedback instead can be made mostly resistive, and can be made strong enough to swamp the undesired reactive feedback, stabilizing an amplifier that would oscillate without it.

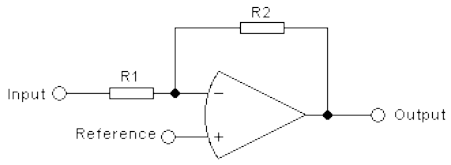
Negative feedback always costs gain. For that reason typically it's not intentionally used in single-band amplifiers operating at a frequency that's close to the highest one the selected transistor is good for, because at such frequencies only little gain is available, and we want it all. But small transistors often have a huge power gain at HF, and LDMOSFETs have a huge power gain at HF even while being high power devices. So we can very well consume some of this gain in negative feedback, to improve the linearity, stability and gain flatness of a broadband linear amplifier.

Transistors are highly nonlinear devices, with MOSFETs being worse than BJTs. If we want to build linear amplifiers using them, correctly applied negative feedback is a

crucial tool to achieve this goal. Amateur designers who avoid it, because they don't understand it, can never achieve the performance of a correctly designed amplifier that uses negative feedback to advantage.

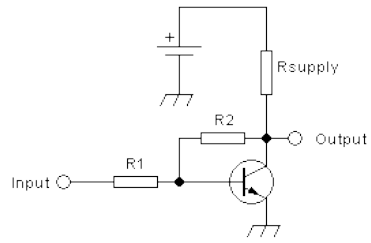
In push-pull stages we need to design the feedback to work in differential and/or common mode, depending on what we want to achive. And in any amplifier we need to make sure that the feedback stays negative over the entire frequency range in which the amplifier has any gain, even far outside our actual operating frequency range, or we will end up with an oscillator instead of an amplifier.

The principle of negative feedback should be known to anyone reading this article, but if you still have a gaping hole in this area of your knowledge, I can quickly fill it in. Negative feedback is very easy to understand, and works in a highly perfect way in operational amplifiers:



The operational amplifier compares the voltage between its two inputs, and amplifiers the difference between them by a very high factor. The two resistors act as a voltage divider between the input and output, and so the operational amplifier will generate whatever output voltage is required to make that voltage divider deliver the exact same voltage to the negative input, as the positive one is getting from the reference voltage source. The output voltage is then the difference between the input voltage and the reference voltage, multiplied by the ratio between R2 and R1. It does not depend significantly on the actual gain of the operational amplifier, and this means that if its gain varies due to frequency, temperature, tolerances, or even if it varies dynamically with voltage level (that is, the operational amplifier causes distortion), the output voltage will still be a very clean, predictable multiple of the difference between the reference and the input voltage, defined by the two resistors. So, as long as the operational amplifier has enough gain, it doesn't need to be good in terms of distortion nor gain flatness, and the complete amplifier will still be excellent, clean and predictable.

Transistors work the same way, only with less gain than an operational amplifier. In exchange for that they can work at much higher power levels, and to much higher frequencies.



The reference in a BJT is the voltage at the emitter, incremented by the base-emitter diode drop of about 0.7V. And since a transistor doesn't have additional pins to connect the power supply, we need to power it at its collector, done here with resistor. Otherwise it's the same circuit as that using the operational amplifier.

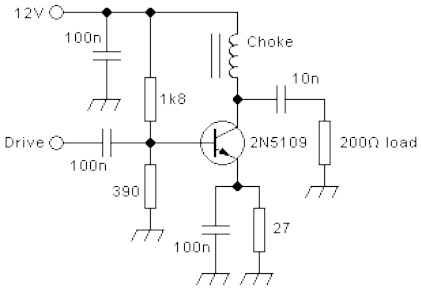
And this circuit indeed works essentially in the same way as the operational amplifier circuit: The voltage gain depends mainly on the ratio between R2 and R1, and the output signal has very little distortion, despite the transistor being a nonlinear device. Also the gain of the amplifier stays essentially constant up to a pretty high frequency, despite the transistor's gain being inversely proportional to frequency over most of the range.

A MOSFET can be used instead of a BJT, but it's less good: The MOSFET has much lower voltage gain in a typical circuit, so the output voltage depends much more on the transistor's characteristics than in case of a BJT amplifier. Also the much higher input capacitance of a MOSFET can cause more limitation of the usable frequency range. But the details depend on the actual characteristics of each particular device. MOSFET technology keeps improving, and each year MOSFETs are taking over some more application area from BJTs.

Negative feedback, implemented in this way, not only reduces the gain, but also reduces the input resistance at the base or gate. It acts just like a swamping resistor in this regard, but while a wamping resistor to ground simply wastes drive power in exchange for a more stable and neutral input impedance, negative feedback wastes that same drive power in exchange for that same more stable and neutral input impedance, plus lower distortion, plus a flatter frequency response! So it's a very much better deal, and every good designer prefers using negative feedback instead of plain resistive input swamping.

I had already mentioned in the input matching chapter that any resistance in series with the emitter of a transistor causes an increase of input resistance. Well, it also causes strong negative feedback, because whenever the collector-emitter path conducts more current, this causes a higher voltage drop on the emitter resistor, reducing the base-emitter voltage and thus reducing the collector-emitter current. This kind of negative feedback is present in every amplifier stage, due to the unavoidable internal emitter or source resistance of transistors, and also due to their lead inductance, which causes the same feedback effect but with 90° phase shifting. Transistor manufacturers intentionally add small amounts of this resistance right inside, where it's associated with the least inductance, when making medium power transistors. In low power transistors the circuit designer can add it externally, and in high current transistors this effect typically becomes larger than desirable, reducing the gain too much, and it has to be reduced by using several very wide emitter or source connections, or even by manufacturing MOSFETs in a laterally diffused fashion with through-the-chip source contacts, so that the entire underside of the silicon chip becomes the source connection, and can be directly soldered to the metal base of the transistor, the heatsink, and the circuit, with extremely short and wide connections.

Let's see what each form of negative feedback does to an amplifier. To begin, I will start with a simple amplifier that has no intentional feedback:



The bias voltage divider would produce a tad over 2V at the base, if there was no base current. This gives around 1.4V at the emitter. With the 27Ω emitter resistor, the transistor takes slightly over 50mA. This in turn creates roughly 0.5mA base current, which loads the divider, which has about 5mA flowing in it. So it gets slightly pulled down, and the transistor will end up biased to slightly less than 50mA.

Its internal emitter resistance is then 0.5Ω. For RF the emitter is grounded via a capacitor, so the transistor's input resistance is simply 0.5Ω multiplied by the RF current gain. The transistor has an Ft of 1200MHz, so at 30MHz its gain is 40, and its input resistance is 20Ω, and this varies in inverse proportion to frequency, except below about 15 or 20MHz, where the current gain gets capped by the (highly part-dependent) hfe.

Let's take 30MHz as a frequency for our calculations. 1mV of drive signal, appearing on the 0.5Ω emitter resistance, create 2mA of collector current. This RF current cannot go through the choke, so it all goes through the load resistance. Since this is 200Ω, there will be 400mV at the output. So this amplifier has a voltage gain of 400 at 30MHz.

Let's see it current-wise: A 1μA drive signal will mostly go into the base, because the bias divider has a pretty high resistance relative to the 20Ω base resistance of the transistor. The transistor has a current gain of 40 at 30MHz, so the collector current is 40μA. All of this can only flow through the load, and so the current gain of this amplifier is 40.

That's a 10:1 ratio between voltage gain and current gain, which makes total sense when you consider that the input resistance is 20Ω and the output resistance is 200Ω.

Many designers like to talk about gain in terms of dB. But dB is a power ratio, not a voltage or current ratio. When input and output resistances are different, one needs to be careful about this! Taking the standard equation to express voltage gain in dB, a gain of 400 is 52dB, but the current gain of 40 is only 32dB. When driving the amplifier with 1mV, the 20Ω input resistance allows 50μA to flow, and that's 50nW drive power. At the same time the output is 400mV, 2mA, and that's 800μW. That's a power gain of 16000, and that equates to a power gain of 42dB, which is the final word on the matter.

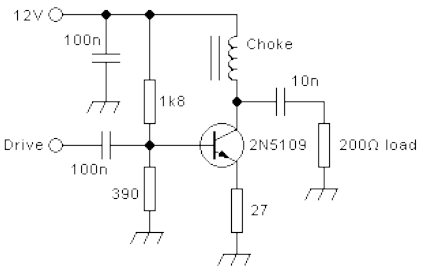
Remember that the gain in dB is 10×log(Pout÷Pin) . Also remember that only when input and output impedances are identical, the gain in dB is 20×log(Vout÷Vin), or 20×log(Iout÷Iin).

How much output power can we get? The transistor can pull its collector voltage down to a little under 2V, so the the possible swing is ±10V around the 12V supply, and that's 7.1V RMS. The transistor is biased to a little less than 50mA. 10V in 200Ω makes 50mA flow, so in fact we will run out of peak current a little bit before we run out of peak voltage capability. Let's take 45mA as the clean peak load current. That's 32mA RMS, and in 200Ω that's 200mW. The input power is given by the 12V supply and the slightly less than 50mA collector current plus 5mA bias divider current, so it's about 620mW. So this amplifier has an efficiency of around 32%, which is totally typical for practical small class A power amplifiers.

Note that in the real world both the gain and the input impedance will be lower, though, because this amplifier is <u>not</u> free from collector-to-base feedback! The transistor used has an internal collector-base capacitance, there is also external capacitance there, and this pushes the input impedance down, and also rotates its phase, consumes some of the output current, and draws additional drive current. The datasheet of the transistor doesn't give the value of this capacitance, so we are on our own there. Also the frequency range over which these calculations hold true will be restricted by parasitic components, such as the inductance in series with the emitter lead and the emitter bypass capacitor.

Anyway the gain of such a feedback-less amplifier is very high, and this high gain usually creates stability problems. Also the entire nonlinearity of the transistor will affect the signal, and both the gain and the input resistance will be strongly frequency-dependent above a certain cutoff frequency, because the current gain of the transistor changes with frequency. So we get an unstable, non-linear, very frequency-dependent, high gain amplifier.

Let's improve that. Let's sacrifice some gain for improvements in the other areas!



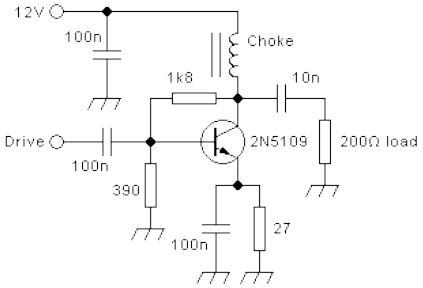
The only difference in this circuit is that I removed the bypass capacitor at the emitter, introducing strong negative feedback through emitter degeneration. The DC biasing has not been changed. But the effective emitter resistance is now the internal 0.5Ω plus the external 27Ω, a whopping difference! At 30MHz, current gain 40, the input resistance of the transistor is now 1100Ω! It gets swamped by the equivalent resistance of the bias voltage divider, 320Ω, to make a total input resistance at 30MHz of 248Ω. That's a whole lot higher than it was before! It will also change a lot less with frequency, due to the swamping.

1mV of input drive , appearing on the 27.5Ω total emitter resistance, makes 36μA of collector signal current, and thus output current, which in 200Ω makes 7.27mV. That's a voltage gain of 7.27. It can also be calculated easier as 200÷27.5.

1μA of drive current will divide into 225nA into the base, and the rest going into the bias resistors. At a transistor current gain of 40, that makes 9mA output current. That's a current gain of 9. And thus a power gain of 66, equating to 18dB. That's certainly a whole lot less gain than before, but it comes with a good gain flatness over a wide frequency range, a much higher and more stable input impedance, lower distortion, and better stability.

An amplifier like this, even while designed for a load resistance of 200Ω, has a very much higher internal output resistance. This means that it behaves largely like a current source. It will tend to force a certain current into the load, no matter what impedance that load has. So this amplifier has no trouble if the output is shorted. It will just put a controlled current into the short circuit, essentially the same current it would put into the load resistance. But if the load is disconnected, it will try to put its current into the almost infinite impedance of the choke! The negative peaks cannot go below a collector voltage of about 1.6V, but there is no real limit set by the circuit for the positive peaks. They can rise as high as required to... guess what? To destroy the transistor!!! So this amplifier cannot handle an open output.

Now let's trade emitter degeneration for shunt feedback:



The capacitor at the emitter is back, turning off almost all of the emitter degeneration (the small amount caused by the transistor's internal emitter resistance remains), and the upper bias resistor has been moved to the collector. This change makes no difference in terms of DC biasing, since the collector is at 12V DC potential, due to the choke being incapable of causing a voltage drop at DC. But now the 1.8kΩ resistor provides a path for RF negative feedback. Let's see what happens:

At 1mV drive, the transistor puts 2mA signal current through its collector, like in the feedback-less case. The total load is now 200Ω in parallel with a little less than 1800Ω. A little less because the base signal voltage is in opposite phase to the collector voltage. Since here the voltage gain is so high, we can neglect this difference and calculate simply with 1800Ω. So the effective load resistance is 180Ω, and the output voltage is 360mV, the voltage gain is 360, not much different from the feedback-less amplifier. But the 360mV make 200μA flow in the resistor, which loads the input, while the transistor's base is only taking 50μA from the input! So the shunt feedback makes the input resistance of the amplifier much lower: At 1mV drive voltage, we now have a total drive current of 250μA, so the input resistance of the amplifier has dropped to just 4Ω!

So we have sacrificed just a little voltage gain, but a considerable amount of current gain, in exchange for better linearity, stability, and flatter frequency response.

The power gain is now 34dB. The fact that we sacrificed less gain is not due to the type of feedback, but to the amount of it. We can of course vary the amount of both types of feedback, by adjusting the resistor values as required.

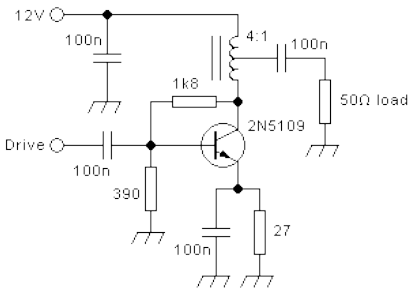
Note what happens if we now disconnect the load: At a given drive voltage, such as 1mV, the collector signal current will then all have to go through the feedback resistor. With the resistor values used in this example, the output voltage would then increase only 10-fold, which is a huge improvement over the amplifier that has no feedback of this sort. Making the shunt feedback a little stronger, we can make this amplifier safe against load disconnections.

Since shunt feedback tends to stabilize the output voltage, no matter how much current the load demands, it makes the amplifier have a low internal output resistance: It behaves more like a voltage source. So it may not survive an output short! If the drive signal is strong enough, it might make the amplifier self-destroy by trying to put more current into the shorted load than the transistor can handle.

So: Shunt feedback protects against open loads, and reduces input and output resistances. Emitter degeneration instead protects against output shorts, and increases input and output resistances. Both of them reduce the gain, and improve linearity. Both of them improve stability, as long as no parasitic effects come into play. In this regard, emitter degeneration can be problematic: Too much of it can degrade stability!

A nice thing in this is that by properly balancing the amount of the two kinds of feedback, we can make an amplifier have a certain desired input resistance, and keep it quite stable over a wide frequency range, while also being relatively tolerant of load impedance changes, have pretty stable gain of our choice over a wide frequency range, have a low distortion, and be free from any self-oscillations. We pay for all this in the maximum achievable gain. The gain is always lower when negative feedback is used. When our transistor has lots of gain that's not a problem at all, but when we need lots of gain and the transistor doesn't have very much we might need to increase the number of amplifier stages, to enable us to get an amplifier with the required frequency response, gain flatness and low distortion.

For this little amplifier I selected a nominal load resistance value of 200Ω, because it was a good fit for the12V supply and the 50mA collector current value from which I arbitrarily started. But when the capability to optimally drive a 50Ω load is desired, it's extremely easy to convert this amplfiier to a 50Ω output: Just replace the choke by a 4:1 autotransformer! This is done by winding the exact same amount of wire, on the same core, but winding it as a bifiliar pair, with the same number of total turns, that is half the number of bifiliar turns:



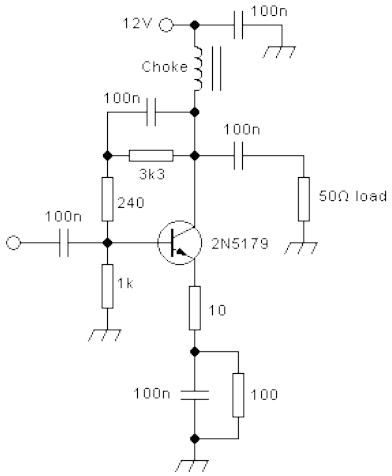
Note that I also increased the value of the coupling capacitor, to account for the lower load resistance. I stuck to the old custom of using round values of capacitance, although of course in an actual design one would check this against the true requirement according to the lowest required operating frequency.

The power supply bypass capacitor becomes more important in this circuit configuration, because it has to carry half of the output's RF current. Ideally the ground terminal of that capacitor should be close to the load's ground point.

With this modification amplifier works just the same, with the same power gain, over the whole frequency range in which the autotransformer has good coupling.

In principle any transformation ratio can be implemented in this way, but 4:1 is the easiest to do.

Now let's consider a much-used, practical, very low power amplifier stage, that has both forms of negative feedback:



This circuit has been published many times in US-American ham literature, almost since the beginning of small RF transistors, with some variations. I'm showing you this circuit to illustrate the use of RRC networks both at the emitter and in the shunt feedback path, which allows separate control over the shunt feedback, the emitter degeneration, the input resistance, the bias current, and the base voltage. It's a good circuit for learning how to tailor such a circuit for specific needs, and that's why I'm mentioning it. Many hams have copied this circuit without fully understanding it. Let's try to do better!

The base bias voltage divider has 3540Ω total in the upper leg, and 1kΩ in the lower. Connecting its upper side to the collector instead of directly to 12V is of course just the same, for DC purposes, because the choke can cause no significant voltage drop at DC. So in the absence of any base current we would get a base voltage of 2.64V, giving about 2V at the emitter. Since the total emitter resistance is 110Ω, plus a small amount inside the transistor that can be ignored relative to 110Ω, the emitter current is around 18mA. The typical hfe of the transistor used is given as 70 in the datasheet, although it has a very wide tolerance. Let's use 70 for the calculation: The base bias current is then 0.26mA. That's about one tenth of the current standing in the base bias divider, so it will pull the voltages and currents a little down, and the transistor will finally end up conducting about 16mA. We can't be ultra-accurate here, because anyway the hfe of a BJT has a very wide tolerance. Bias business done!

With that current, the internal emitter resistance is about 1.6Ω, and for RF purposes this is in series with just the 10Ω of unbypassed external emitter resistance. So we have 11.6Ω total. The typical Ft of this transistor is given as 1.4GHz. So, if properly assembled, with very small, low inductance parts, ideally SMDs, it would be reasonable to expect this circuit to work into the range of a few hundred MHz. At 300MHz the current gain would be 4.7, and the input resistance of the transistor 54Ω. At lower frequencies both of them rise accordingly, down to about 20MHz, where the current gain gets capped by the DC hfe of 70, capping the base input resistance at roughly 800Ω.

Now things start getting more complex: Let's try to calculate the gain, and the feedback effects. Since there are two feedback methods used, which interact, this can get quite entertaining! Let's start assuming that the circuit did not have any shunt feedback. In practice this could be done by connecting the 3.3kΩ resistor and its parallel capacitor to 12V instead of the collector. Any variation of the base voltage will transfer almost unchanged to the emitter, and will make the collector current vary according to the 11.7Ω effective emitter resistance, and Ohm's Law. So, a signal of 1mV amplitude at the input will cause a current signal of 85.5μA at the emitter, and if you want to be pedantic, you might want to subtract the base current from this to get the collector signal amplitude: About 84μA if the frequency is below 20MHz, and decreasing slightly as the frequency rises above that.

This signal current must flow through the load resistor, because it has no other way to go. So, if the load is 50Ω, the current signal will cause a voltage signal of 4.27mV, below 20MHz. That means that the voltage gain of the amplifier, in that configuration, is 4.27. But it's totally dependent on the load resistance! The voltage gain is simply the load resistance divided by the 11.7Ω of RF-effective total emitter resistance, minus a small, frequency-dependent allowance for the base current robbing a little from the collector current.

That was the effect of emitter resistance. Now let's see the effect of the shunt feedback. For this purpose let's imagine the amplifier without any emitter series resistance, not even the transistor's internal one, and consider the shunt feedback as it's shown in the schematic. With the emitter at ground potential for RF purposes, the base also cannot have significant RF voltage on it. Just RF current through it. It's a near-zero impedance point, so the input impedance of the amplifier in such a configuration would be near zero, and the transconductance would be almost infinite. So we need to calculate in terms of current, rather than voltage.

Let's see what happens if we apply a 1μA signal: At frequencies below 20MHz, the current gain is typically 70, so the drain current would be 70μA. Between the collector and the base we have just 240Ω at RF, because the 3.3kΩ resistor is shorted out by the capacitor. So our load on the collector is now 50Ω in parallel with 240Ω, and that's 41.4Ω. The 70μA collector signal would then create a 2.9mV output signal. But the story doesn't nearly end here! 2.9mV on 240Ω puts a current of 12μA into the base node, and this current is in opposite phase to the input current, because the transistor inverts the signal! Higher base current makes collector voltage go down... Of course this can't work!

What then happens is that out 1μA input signal current splits up between a smaller part that flows into the base, and a larger part that flows into the 240Ω resistor. We already saw that for every μA of base current, 12μA show up in the 240Ω resistor, so things are easy: One thirteenth of the drive current flows into the base. So: 1μA drive input, 77nA of that into the base. At a gain of 70, it causes a collector current of about 5.4μA. In a total 41.4Ω load, 5.4μA creates a voltage signal of of 223μV. This voltage applied to the 240Ω resistor causes a current of 923nA, which is the part of the 1μA drive current that did not flow into the base. The 223μV also push a current of 4.46μA into the 50Ω load resistor, which means that the current gain of this amplifier, working without any emitter degeneration but with shunt feedback, would be 4.46.

So you can see that the two feedback systems of this amplifier are pretty well balanced: One limits the gain to 12.6dB, the other to 13dB, assuming equal input and output impedances. When both are combined, by restoring the circuit to how I drew it, but with a 50Ω load connected to the output, the real gain will be the orthogonal composition of the two, which is very close to 3dB below the average of the two individual gains. So this amplifier, running into a 50Ω load, has around 9.8dB gain. It's usually advertised as a 10dB gain amplifier in the literature.

You might ask why anyone should use two methods of feedback combined. As we have seen already, the reason is that they have different effect. Emitter degeneration enhances the natural behaviour of a transistor to act as a controlled current source. An amplifier using only emitter degeneration would have an extremely high output impedance. That means that it will try to force a certain output current, no matter what load the user connects to it. So the output power will depend dramatically on the load. The amplifier would be tolerant to a short circuit at the load, but if left open it would instantly saturate, and might even self-destruct. Instead shunt feedback acts the other way around: Like in the typical operational amplifier circuit, it tends to make the amplifier create a specific output voltage, no matter what the load is. It creates a very low output impedance. It's perfectly tolerant of a load disconnection, but in the event of a shorted load, it would try to put a high current into it, possibly self-destroying... But if both feedback methods are combined, the amplifier is much more tolerant of load variations, the power gain remains much more stable in the event of load variations, and so it's a better all-round, all-purpose amplifier. Instead in applications where the load is known and fixed, amplifiers with just one feedback method are perfectly OK.

We are not nearly through with this little amplifier! The next step is calculating the input impedance of the total amplifier. This can be done without too much difficulty, now that we know that the amplifier has a gain of 9.8dB, a voltage gain of about 3.1, when loaded with 50Ω. Let's calculate for frequencies below 20MHz, because there the current gain of this transistor is stable.

1mV drive signal at the input will then cause 3.1mV at the output. The two signals are in opposite phase. That means that the RF voltage across the 240Ω resistor is 4.1mV. This loads the collector with 17.1μA, and also takes those 17.1μA from the drive input. The 50Ω load, which sees 3.1mV, loads the collector with another 62μA. The total collector signal current is then 79.1μA. At a current gain of 70 for this transistor, in the low frequency range stated, that needs 1.13μA of base current. And then there is the 1kΩ resistor to ground, which at 1mV RF voltage takes an RF current of 1μA. The currents through the base, the 240Ω resistor, and the 1kΩ resistor, add up to 19.23μA. 1mV divided by 19.23μA is 52Ω!

At higher frequencies the transistor's current gain drops, making the base signal current increase, and among many effects this ends up making the input resistance value go slightly down. And that's why this little amplifier is advertised as having a 50Ω input impedance! It's indeed very close to that. A tad above in the low frequency range, and dropping slightly below 50Ω as the frequency goes into the VHF range. If we want to be pedantic, the transistor's total input capacitance of about 4pF, plus any circuit stray capacitance, appears in parallel with the 50Ω input resistance. But this starts becoming important only above roughly 100MHz. So this is a good HF plus low VHF amplifier stage, if properly constructed.

Now it's time to see what happens at low frequencies. The designer of this circuit was so lazy that he used five identical capacitors. This isn't necessarily a good choice! Let's start with the input and output coupling capacitors. Both work into 50Ω loads, so it does make sense to make both capacitors the same value. A 100nF capacitor has a 50Ω reactance at 31.8kHz. The combined -3dB cutoff of the two will then be at 45kHz, which looks like a pretty good choice if coverage from LF to VHF is desired. But at this point I would like to tell you one of the trade secrets of electronics: Building a circuit for a wider frequency range than needed is not a good idea, because it is an unnecessary invitation to trouble! It's better to tailor the frequency response of any circuit to just the amount needed by the application. It can easily be done by adjusting the capacitors accordingly.

The shunting capacitor in the emitter circuit is relevant in comparison to the 100Ω in parallel to it. At very low frequencies, where that capacitor has a high reactance, the effective emitter series resistance will be 111.7Ω, dramatically reducing the gain of the amplifier, but never all the way to zero. The -3dB frequency is half as high as that of the input and output capacitors, because it works against 100Ω instead of 50Ω. So the choice of using the same capacitance here as for the coupling capacitors is fine, because it will keep the emitter degeneration working as expected down to a frequency somewhat lower than the cutoff set by the coupling capacitors.

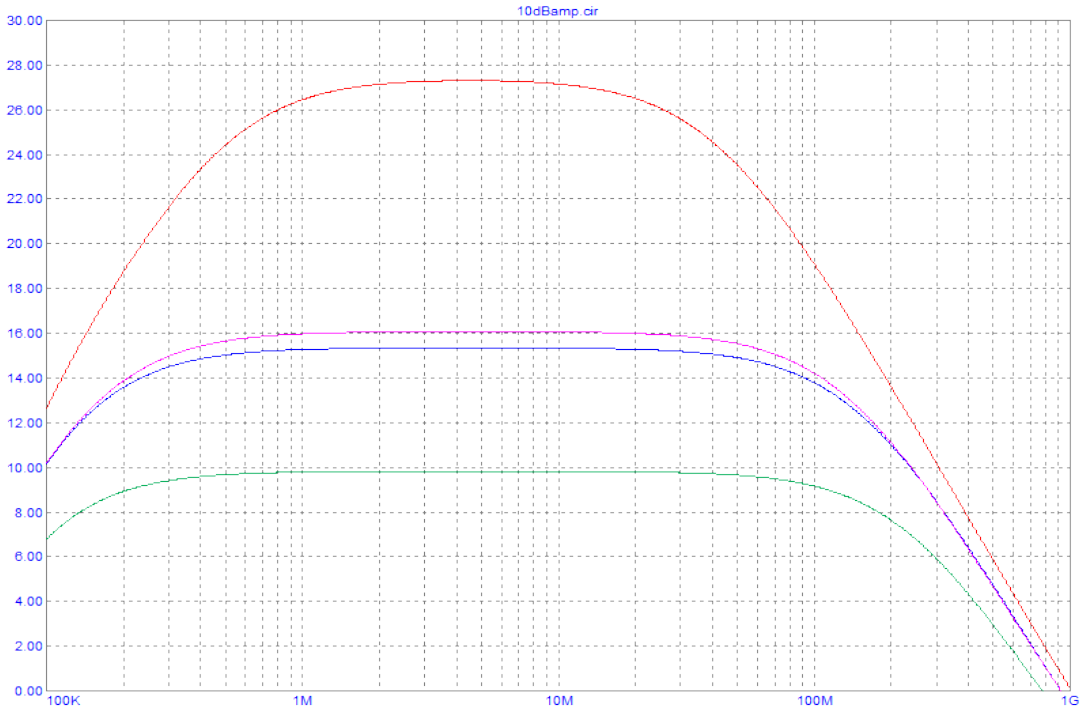
The capacitor shunting the 3.3kΩ resistor is a different matter. At frequencies low enough to make it ineffective, the gain of this amplifier will rise, since the negative feedback will be weaker. But the gain is still capped by the emitter degeneration, and anyway the 3dB cutoff frequency of 100nF with 3.3kΩ is only 482Hz! This is way lower than necessary. The only harm this does is slowing down the stabilization of operating conditions after turn-on. If that's a problem, this capacitor can be downsized, down to 10nF or even somewhat less.

Of course, to set a different low frequency limit for the amplifier's response, all the capacitors can be changed in the same proportion. It's entirely possible to use electrolytic capacitors here, properly oriented, to make this amplifier work throughout the audio range and all the way up to VHF. Of course the choke needs to be suitably too for the low frequencies, and that's a problem...

The bypass capacitor's value, and its effects, depend more on what's happening in the power supply bus than in the amplifier, since the choke should be blocking essential all signal feedthrough. Given that chokes aren't perfect, it's good to have a bypass capacitor that's highly effective particularly in those frequency ranges where the choke isn't.

This circuit was not designed as a power amplifier! This becomes apparent when we analyze the maximum output power than we can get from it: The collector voltage is free to go down to less then 3V, and since the circuit is choke-fed, it can go above the 12V supply as much as it goes below it. So this amplifier could in principle deliver a ±9V swing to the output. But... can it really? Nope! Not even anywhere close to that! Because 9V in a 50Ω load would need 180mA, plus some additional current for the 240Ω feedback resistor, but the amplifier is biased for just 16mA! So the current to which this amplifer is biased allows only about ±13mA into the 50Ω load, which means a voltage swing of only ±650mV, which is a mighty poor use of the available ±9V swing. And that's the peak current and voltage. The corresponding output power is just about 4mW. The power drawn from the supply instead, 12V×18.6mA (don't forget that the bias divider also draws current!), is 223mW, and so this circuit has a maximum efficiency of not even 2%! Instead it has excellent linearity, provided by the strong dual feedback and helped by the small voltage swing. So this is a broadband, low gain, low distortion, robust, small-signal amplifier, with a well-controlled 50Ω input impedance.

How good is such an amplifier, and how does it compare to one without negative feedback? To prepare some nice graphs of it, I used a simulator to calculate the gain from 100kHz to 1GHz, using a 50Ω signal source:



The red curve is the amplifier's gain without any feedback. The violet one is with just emitter degeneration. The blue one is with just shunt feedback. And the green one is with both feedbacks working, as shown in the schematic. You can clearly see that feedback reduces the gain a lot, but also flattens the gain curve, extending the frequency range over which the gain is essentially constant.

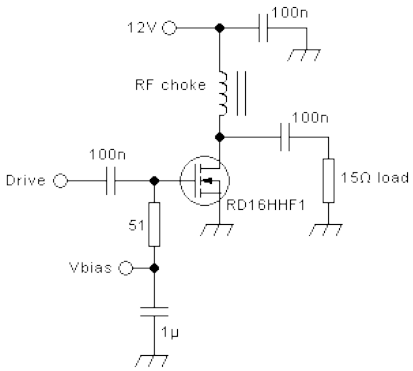
Note that the gains when using only one type of feedback do not agree with the ones I calculated above. That's because my calculation was based on a fixed voltage or current signal at the input, while in the simulation I used a signal source with a 50Ω internal source resistance. The very different amplifier input impedances that result from using each feedback system result in the difference between the simulator's curves and my calculated values. The situation with both feedbacks enabled instead shows excellent agreement between my calculation and the simulator, because it results in an input impedance of almost exactly 50Ω, matching the signal source resistance, so that both calculation methods agree.

What you cannot see here is how much the linearity is improved when using feedback. I tried to get distortion information from the simulator, by making a Fourier analysis of the output signal, but the simulator told me that there was absolutely no distortion ever! Of course... duh! The transistor model this simulator uses is a simplified, linear one! That's the problem with circuit simulators: They idealize the world, and while some of their output is useful, some other is totally misleading. Simulators always require knowing beforehand, at least roughly, what to expect, and then doing a sanity check on anything they tell. Spice-based simulators are basically linear simulators, and simulating any nonlinearity is hard for them.

Likewise the gain curves shown here do not include the effect of all the imperfections of components and interconnections. How well the gain of a real circuit will match the simulator output, depends on how perfectly (free from strays of all sorts) the circuit is built, in addition to how good the semiconductor models in the simulator are.

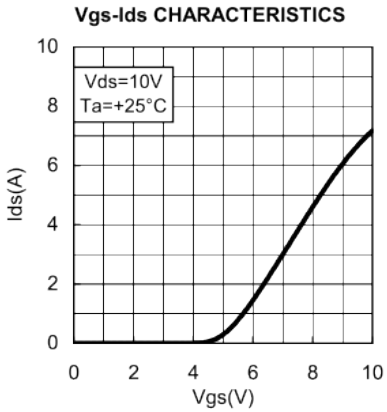
In **MOSFET** amplifiers it's rather uncommon to intentionally use source degeneration as a feedback method, due to the high input capacitance and low voltage gain of these devices. So the feedback is usually implemented as shunt feedback, but there is alwas also some source degeneration present, coming from internal MOSFET source resistance. Since MOSFETs used in an amplifier tend to have a gate-to-drain voltage gain that's quite comparable to the base-collector voltage gain exhibited by a BJT operating with a typical amount of external emitter degeneration, we can analyze and design MOSFET amplifier stages with shunt feedback pretty much in the same way as we analyze a BJT stage having both forms of feedback. Although the transconductance of a given MOSFET is variable, depending on the drain current, and thus causing very nonlinear amplification, we can usually come close enough by taking the transconductance the given MOSFET has at the average current it will be conducting.

Let's first consider a simple 3 watt MOSFET HF driver stage that has no external feedback:



Let's assume that the bias voltage is adjusted so that the MOSFET conducts an idling current of 1A, allowing it to operate in class A. To develop 3W into its 15Ω load, we need an RMS output voltage of 6.7V, and an RMS output current of 0.45A. This equals a peak output voltage of 9.5V, and a peak output current of 0.63A. Due to both the RDSon of the MOSFET, and the distortion caused at higher frequencies by the very strong and nonlinear increase of its drain-gate and drain-source capacitances, we can't make it pull its drain all the way to ground, and the ±9.5V swing around the 12V supply voltage is about the best we gan get, so the supply voltage is well utilized. Instead it

might seem that biasing the transistor to 1A when only needing 0.63A peak load current is overkill and wasteful, but look at the published transfer curve of this MOSFET to see why it is necessary:



As you can see, the reasonably linear part of the curve starts at roughly 0.7A or so. In this amplifier the part of the curve being used is the one that falls between 0.37 and 1.63A, and this already includes some of the very nonlinear zone! So when adjusting the bias of this amplifier, one needs to make a compromise decision between linearity and efficiency. I chose 1A for this example. If instead we bias this amplifier to only 0.7A or so, its distortion would shoot through the roof, despite still operating in class A!

Do you remember the difference I'm making between chokes and RF chokes? Well... In these MOSFET amplifiers, despite being class A, which implies that in principle the supply current is always constant, allowing the use of any choke without caring about whether or not it blocks modulation frequencies, we need RF chokes, that pass modulation frequencies freely. This is simply due to the strong nonlinearity of MOSFETs. During high amplitude operation the average drain current over the RF cycle might well differ a little from that at low or zero amplitude, and the choke needs to accept that current modulation, or else the drain supply voltage will get modulated by the envelope signal, causing additional distortion!

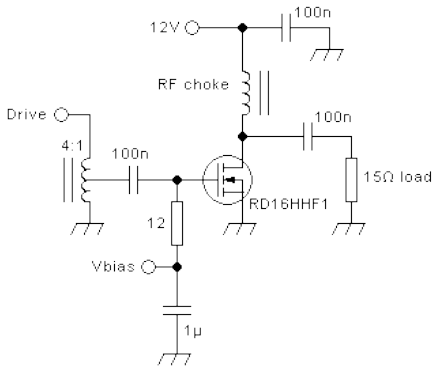
For 0.37A drain current we need roughly 5.2V on the gate, and for 1.63A we need about 6.1V. So our p-p gate voltage swing is 0.9V, for a p-p drain current swing of 1.26A, and that gives us a transconductance of 1.4 siemens.

The input impedance of this amplifier is basically the 51Ω gate swamping resistance, which I also used to inject the bias voltage, in parallel with the MOSFET's input capacitance. This capacitance is the combination of the roughly constant gate-source capacitance of approximately 47pF, and the Miller capacitance, which in turn is the product of the drain-gate capacitance (very roughly 3pF, varying a lot with instantaneous drain voltage) multiplied by the voltage gain. And this voltage gain is the output voltage swing (19V) divided by input voltage swing (0.9V), so it's 21.1. So the Miller capacitance is roughly 63pF, which brings the total input capacitance of the MOSFET to roughly 110pF. The result is that the input impedance of this amplifier is 51Ω ‖ 110pF.

Note that this combination of input resistance and capacitance produces a pole at 28.4MHz, so the input matching of the amplifier will be poor at the high end of the HF range. This wouldn't be really acceptable in an amplifier supposed to be roughly flat to 30MHz. A lower swamping resistance would be required, to flatten the response to a higher frequency.

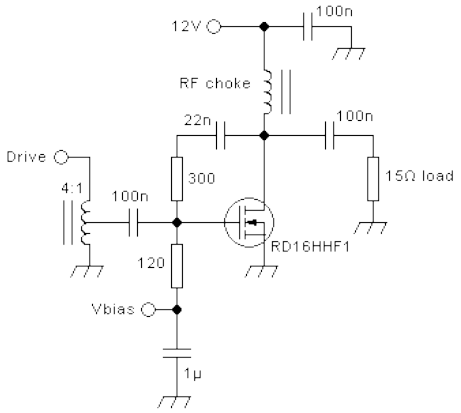
The drive power required is given by the input resistance and the RMS input voltage. Since the p-p input voltage is 0.9V, the RMS value is 0.32V, and across a 51Ω load that's only 2mW. Note that the parallel input capacitance doesn't require any additional drive power, but just additional drive current, which is 90° out of phase with the drive voltage! Of course a typical driving source would have trouble with this additional out-of-phase current. Anyway, since the amplifier produces 3W output, the gain is a whopping 32dB! But the frequency response is poor, and the linearity too.

Let's fix the input first. By quadrupling the load on the gate, we push the problems 4 times up in frequency, which should be good enough. If we want to keep a ~50Ω input, we then also need to add a matching transformer:



Now the input impedance is 48Ω ‖ 27pF, giving us an input pole frequency of roughly 114MHz, which is fine for HF use. But it came at a price! Not only the added autotransformer, but also we now need twice the drive voltage, four times the drive power, and this means that the gain of the modified amplifier is 6dB lower. That's still a gain of 26dB, which is pretty good, and might actually be too much in some applications.

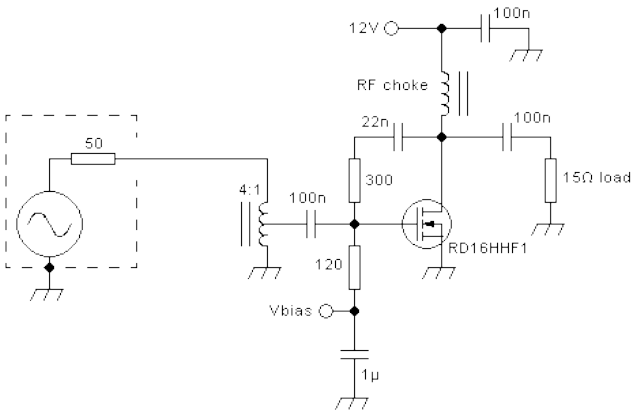
But now lets finally introduce intentional shunt feedback, in parallel to the capacitive shunt feedback that's always present due to the drain-gate capacitance of the MOSFET. Let's arbitrarily replace about 90% of the gate swamping by shunt feedback:



As you can see, I increased the value of the swamping resistor tenfold, and added shunt feedback. Since the voltage gain of the amplifier (given by MOSFET transconductance and load resistance) is 21.1, and the output voltage is in phase opposition to the drive voltage, the total RF voltage between drain and gate is 22.1 times the drive voltage. So the value of the feedback resistor needs to be 22.1 times higher than the part of the swamping resistor it replaces. Since that one is about 13Ω (because 13Ω in parallel with 120Ω is roughly 12Ω), we need close to 300Ω of shunt feedback resistance. A DC blocking capacitor of course needs to go in series, and its value needs to be large enough to keep the feedback functional down to a sufficiently low frequency.

Now the drive current splits about 10% into the swamping/bias resistor, and 90% into the feedback resistor. What did we gain by doing this? Well... errr... blush... nothing, so far! In fact we have added two parts, we are now wasting some of our output power into the feedback resistor, but the drain signal current still depends strictly on the drive voltage, which depends on the driver! So does the distortion, frequency response... really so far there is no improvement!

Of course this changes when we factor in the internal resistance of our signal source. Let's do that explicitly:

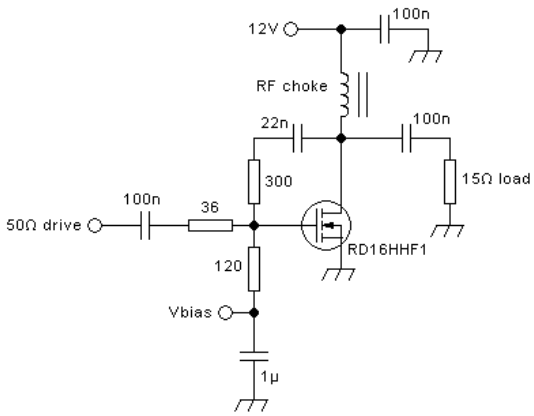


Assuming that the 4:1 autotransformer works as it should, and that the coupling capacitor is transparent at RF, the 50Ω internal resistance of the signal source, transformed down to 12.5Ω by the transformer, forms a voltage divider with the impedance at the MOSFET gate, which is the parallel combination of the 120Ω swamping and bias injection resistor, with the 300Ω feedback resistor divided by the voltage gain plus one, the MOSFET's gate-source capacitance, and the drain-gate capacitance multiplied by the voltage gain plus one.

Now we do get a benefit from having mostly replaced pure dumb gate swamping by feedback: Lower distortion! Because any distortion products appearing at the drain will feed back into the gate, and will actually be able to alter the gate voltage, which is no longer firmly tied to a stiff voltage source in the form of a zero impedance driver. This causes an opposing drain current change, attenuating that distortion. Also any gain reduction at high frequencies caused by runtime effects in the transistor, or by source lead inductance, will also be moderated, resulting in some extension of the flat gain range.

It's interesting to understand what source impedance the gate itself sees, over the frequency range. In the normal operating frequency range, where the transformer works fine and the MOSFET too, this is the 50Ω drive source resistance, transformed down to 12.5Ω by the transformer, in parallel with the 120Ω bias injection/swamping resistor, in parallel with the roughly 13.5Ω of effective resistance provided by the feedback circuit. That makes a total of roughly 6Ω.

But at very low frequencies this condition falls apart: The transformer no longer works well, making it look like having a parallel low value inductor to ground. At the same time the coupling capacitor no longer has near-zero impedance. This forms a pretty series-resonant circuit from the gate to ground, which interacts with the internal MOSFET capacitances, so that we get a series and a parallel resonant frequency! It's very easy to unwillingly build an oscillator in this way! Negative feedback should help prevent oscillation, but our negative feedback path also becomes capacitive at those very low frequencies, as the 22nF capacitor's reactance starts to dominate over the 300Ω resistor. The parallel resonance at the gate still gets loaded and thus de-Q-ed by the swamping resistor, which is connected to the bias source, which hopefully has a low impedance down to DC. But that doesn't help with the series resonance, which can still make the circuit oscillate! We typically need to load and de-Q that one too, by adding a small resistor in series with either the gate or the coupling capacitor. More about this sort of trouble comes in the chapter about amplifier stability! For now I will just suggest that by adding this series resistance our amplifier not only gets more stable, but also gets more independent of the signal source's internal impedance. In an extreme case we might want to put in a series resistance so large that it provides matching to the driving source, without needing the transformer:

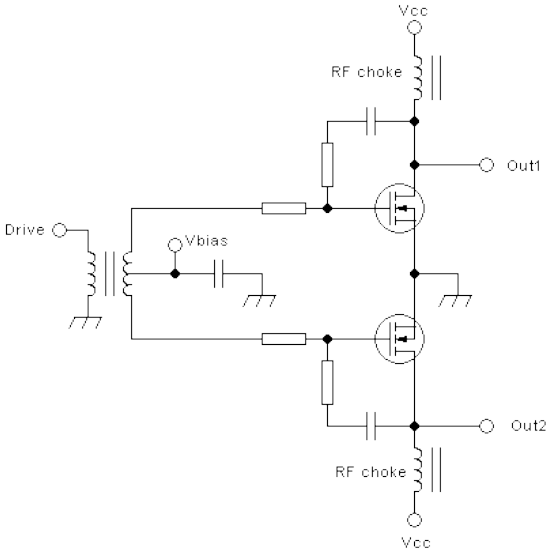


The 50Ω input resistance is now given by the 36Ω resistor in series with the effective 12Ω of the swamp+feedback combination. The cost of this is that we need to again double the drive voltage, quadruple the drive power, and thus the replacement of the transformer by a resistor costs us another 6dB of gain. But it provides excellent stability, combined with a good gain flatness, and reasonably low distortion, and the mentioned moderate degree of independence from the driving source's internal impedance.

The voltage gain of this amplifier would be about 5.5. The current gain is about 18, given that the load resistance is much lower than the input resistance. That's a power gain of 100, which is 20dB. Not bad!

In the most recent Blue Block, way above, I told you about the two modes of **push-pull amplifiers**, and I hinted that they apply to feedback too. It's time to look deeper into this.

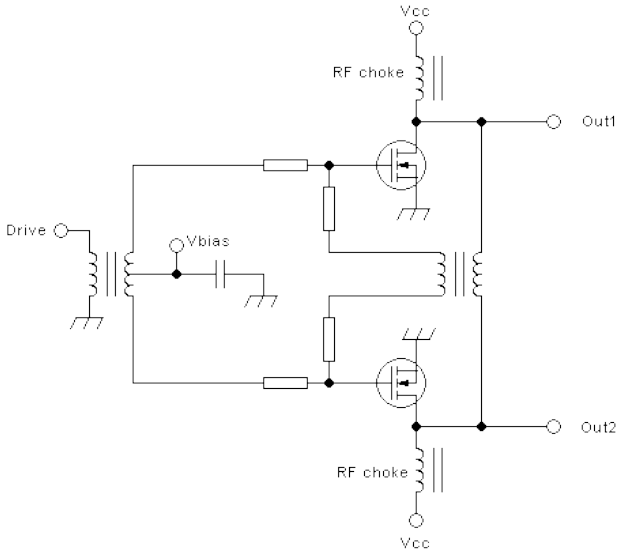
This circuit shows the input and feedback configuration of a push-pull amplifier built as two separate amplifiers in opposite phase:



As you can see, the two sides are totally independent, each with its own shunt feedback and series resistor. The only element joining them (apart from ground and Vcc, of course) is the drive transformer, whose secondary's center point is grounded for RF and modulation frequencies (the latter thanks to a low impedance bias supply). The drive transformer then forces equal and opposite signal voltages driving the two gain sections. As long as the resistor values are chosen such that they force a voltage gain that is much lower than the voltage gain resulting from the MOSFETs' transconductance and the load impedance value, each of the MOSFETs will produce a well-controlled output voltage, equal in magnitude and of opposed phase, to be delivered to any sort of output transformer.

In this way this circuit implements full feedback: There is both differential-mode and common-mode feedback. This can be easily seen by playing the "what if" game: What if the instant output voltage difference is less then it should? In other words, if the output differential voltage is too low? Well, the feedback circuit will make both gate voltage change in opposite directions, such that the error gets largely corrected. And what if both outputs go too high, together? That is, a common-mode output voltage appears? Well, the feedback circuit will then pull both gates a little higher, largely correcting that common-mode deviation.

Now let's modify that circuit, adding a feedback transformer:



This is a pure differentia-mode feedback. Any difference between the two output voltages creates an output from the feedback transformer, which is applied differentially between the gates. Instead any common-mode voltage on the drains is simply not seen by the transformer, causing no feedback.

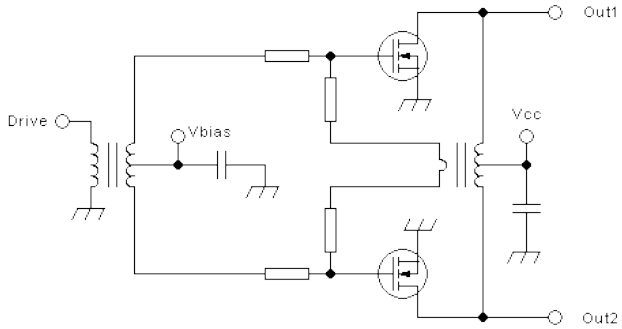
Usually the secondary winding of the feedback transformer has far fewer turns than the primary, often just a single turn. This allows reducing the value of the feedback resistors, and specially, reducing the power loss in them. This is an important advantage in amplifiers powered from relatively high voltages, which tend to require high power RF resistors when using direct drain-gate feedback. Anyway this reduction of power wasting is a poor reason to choose differential-only feedback, because there is no reason why it should be impossible to use two separate feedback transformers, thus implementing full differential and common-mode feedback, combined with low power loss in the resistors. But the circuit is more complex, with one additional transformer and several additional capacitors, so it's rarely done.

Differential feedback is fully capable of controlling the amplifier's differential gain, reducing the distortion of the differential output voltage, extending the flat gain bandwidth, but it does nothing to keep each drain voltage centered around Vcc! At DC, and up to the frequency where the RF chokes start becoming effective, the RF chokes will force that centering, but not at RF. So it's very possible, and even quite usual, that an amplifier configured like this will show totally different, asymmetric, extremely distorted drain voltage waveforms - but the differential voltage, from drain to drain, will still be reasonably clean! The problem is that as one transistor saturates, while the other compensates for that, inevitably there will be some additional distortion in the output, and often one transistor will heat up more than the other one, possibly leading to thermal run-away.

What's even worse, this amplifier will exhibit extremely high and uncontrolled common-mode voltage gain. Any signal leaking into the gates, such as capacitive coupling between the feedback transformer windings, will appear in common-mode at both gates, causing common-mode drain current. Also any common-mode signal created by the MOSFETs themselves when driven differentially, due to their nonlinearity, will do this. Since the drains aren't loaded for common mode, due to the load being connected differentially, a very large common-mode drain voltage signal results. It might easily exceed the drain voltage limit, making the transistors go into avalanche conduction, which can easily destroy them. And last but not least, such a high, undesirable, unneeded common-mode gain could very easily cause the amplifier to oscillate in common mode.

All this said, I should again mention that neither this amplifier, nor any amplifier, is completely free from direct drain-gate feedback. The reason is the drain-gate capacitance of the MOSFETs, which is quite important at the higher frequencies, even becoming the dominating factor. But at lower frequencies this internal feedback path gets less effective. Also, being a capacitive path, the feedback it provides is 90° out of phase, doing half of the job required to make an amplifier oscillate! If the other half comes from phase shifts in all other circuit parts, we get a low-frequency oscillator instead of an amplifier.

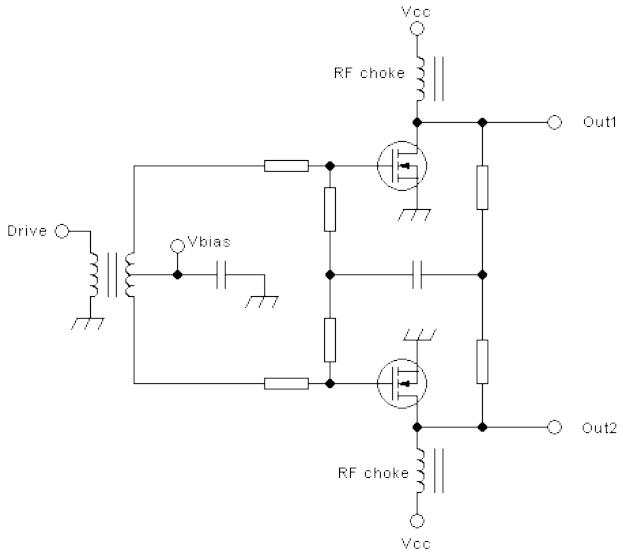
Even so, this transformer-based differential feedback circuit is very often used in commercially made radios! Some of that may even be due to designers not fully understanding what they are doing. But their are indeed ways to get such an amplifier under control even in common-mode. Remember the chapter about feed configurations for push-pull amplifiers: By using a bifiliar feed choke, balanced feed transformer, or whatever you want to call it, the drains are forced to carry mirror images of any signal present, and that shorts out any common-mode voltage at the drains! When there is no possibility of common-mode drain voltage, there is also no need for any common-mode voltage feedback. Also a dedicated feedback transformer becomes unnecessary, because a simple, single turn on the feed choke can be used in its place. This is very commonly done, and looks like this:



Very simple, nice, and cost-effective. That's why manufacturers love to use it! Also this circuit allows true class AB and class B operation, unlike the amplifier using separate chokes, which is only good for class A, dynamic class A, and current-switching class D. There is just one problem, and it's a big one: The extreme difficulty in making such a coupled feed choke that actually maintains good coupling to the highest frequencies required, and in physically building the drain-feed-bypass circuit with low enough stray inductance.

In theory it would also be possible to use a single transformer for the whole drain feeding, load matching and feedback business. But in this case the difficulties are even greater. At power levels of about 100W and greater, with the typical low supply voltages used, the leakage inductance becomes impossible to manage even in the mid HF range, let alone its high end. I have never seen an amplifier in this power class using such a do-it-all transformer.

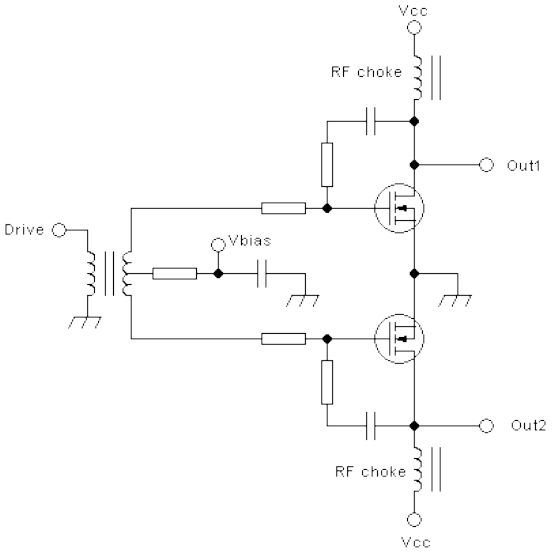
If a designer ever wanted to implement only common-mode feedback, without any differential-mode feedback, it can certainly be done, such as in this example:



The two resistors connected to the drains react purely to the common-mode drain voltage, and this is applied in the same phase to both gates. The common-mode gain can be controlled by the resistance values, while in differential mode this amplifier operates in open loop.

The main problem with this circuit is that no matter how hard I think, I can't come up with a good use for it!

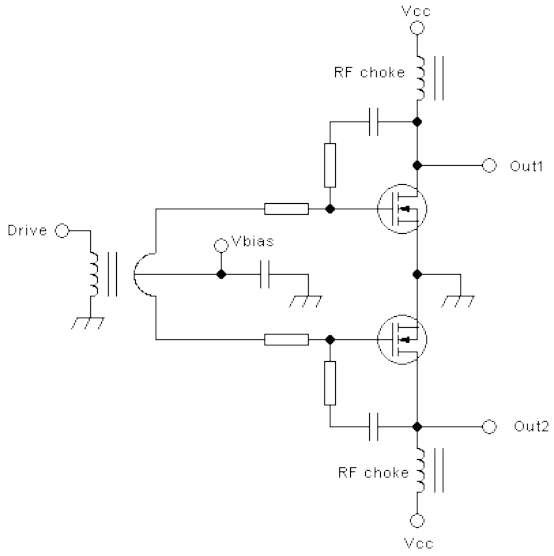
Let's instead make a small modification that allows implementing a stronger common-mode feedback than differential-mode feedback. Such a configuration is useful if we want to forcefully keep the drain voltages from developing any significant common-mode signals, while having a weaker differential-mode feedback in order to still get enough gain. I will base this circuit on the one that has two completely separate sides, with just one small modification:



The modification is, obviously, adding a resistor in series with the drive transformer center tap. The differential feedback still depends just on the ratio between the shunt and drive series resistors (including the transformed impedance of the driver), while the common-mode feedback depends on the ratio between the shunt resistors and the series combination of the gate series resistors and the center tap resistor.

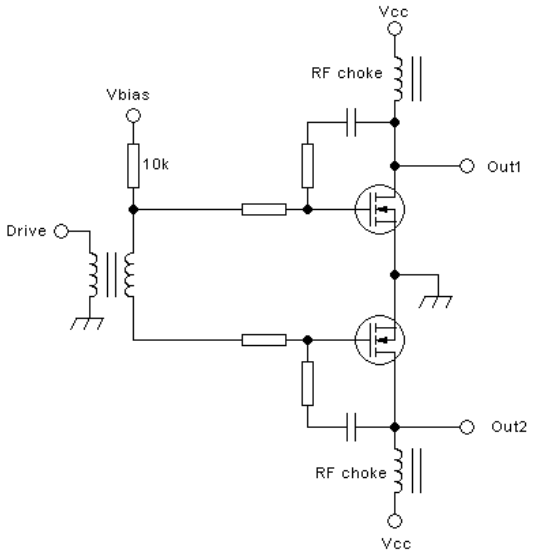
If this new resistor is replaced by an RF choke, the common-mode feedback becomes extremely strong, forcing essentially complete absence of common-mode signals at the drain. But the reactance of the choke is a stability hazard in the frequency range where the choke starts becoming ineffective.

Note that if a drive transformer with a single secondary turn is used, as is often the case in large amplifiers, the center tap is essentially uncoupled, so it behaves like a series choke! So this amplifier has very low common-mode gain:



Such transformers are deceiving, because many people people assume that their center tap actually works at RF, which it really doesn't!

The same effect can be achieved by using no center tap at all, and injecting the bias voltage through a high value resistor:



This circuit might look crazy, because the bias is injected into just one side. Of course this really doesn't matter, because the drive transformer has essentially zero DC resistance, and the bias resistance I chose is so high that the RF imbalance it causes is totally negligible, at least as long as the driver used doesn't have near-infinite impedance. But if you build a circuit like this, and you want perfect symmetry, just for the aesthetic value of it, then I gracefully allow you to use two bias resistors, one to each side...

The advantage of this configuration over the one that has a resistor at the center tap is, of course, that no center tap is required, simplifying drive transformer construction. This may be particularly important when using a transmission-line transformer for driving.

An amplifier like this works with purely differential drive, optimal differential feedback, and very strong common-mode feedback. It forces a good balance between sides, despite having no direct drain-to-drain coupling. But one needs to use great caution when designing such an amplifier, because there is a trap built in: The common-mode gate voltage can be pulled and pushed very heavily by the feedback, possibly destroying the MOSFETs by exceeding the negative gate voltage spec! Getting too high a positive gate voltage is impossible, because when the gate voltage goes pretty high, the MOSFET conducts so strongly that it pulls down its drain, no matter what, and thus limits the positive gate voltage excursion. The problem is just with negative gate voltage. If for any reason both drain voltages are pulled down fast, the gates follow that, the MOSFETs turn off, and since they can't turn "offer" than off, they cannot keep the gate voltage from going strongly negative. In amplifiers operating from 12V, and using VDMOSFETs with their ±20V or even ±30V gate voltage ratings, this doesn't pose any risk, but with 50V-powered amplifiers using LDMOSFETs having a gate voltage rating of perhaps +13/-6V, damage can easily happen. So it would be wise to either not use a circuit with such strong common-mode feedback in 50V LDMOSFET amplifiers, or add gate voltage clamps that will prevent excessive negative gate voltage spikes.

Increasing the common-mode feedback strength is very useful for controlling drain voltage overshoot. You may remember what I explained about push -pull amplifier feed arrangements in the section about output circuits, way up this page, and in the Blue Block comparing Dynamic Class A to Linear Class D. The main point is: When there is any significant uncoupled inductance in the supply path of each drain, this inductance tries to keep the supply current constant throughout the RF cycle, but a class AB amplifier needs to vary that current from almost zero to maximum, twice per RF cycle. So the two requirements clash, and the result is extremely high inductive voltage spikes on the drains, which at best only create distortion and lower efficiency, and at worst kill the transistors. A feedback scheme that implements very strong common-mode feedback, like the ones we were just looking at, can largely control this problem. Whenever the MOSFETs try to suck more current than the drain chokes are carrying, the common-mode drain voltage will sag, and this will reduce both gate voltages, limiting the current drawn by whichever MOSFET is conducting at that time. Likewise, when the drive signal is crossing zero, so that the MOSFETs would have just the bias voltage on their gates and conduct very little current, but the drain chokes are still forcing in

the same high current as before, the common-mode drain voltage would soar. In even clearer words, both drains would see a very dangerous high voltage spike, at the same time. With strong common-mode feedback, this won't happen, because both gates will be pulled up, enabling both MOSFETs to conduct the current standing in the chokes. So, by using strong common-mode feedback, we turn a class AB amplifier with dangerous high voltage spikes on the drains into a dynamic class A amplifier, that is largely free from these spikes! The price we pay is that we get a full-power efficiency more like that of a class A amplifier, than a class AB one. But at least the idling current follows the signal amplitude, and during modulation pauses the dissipation is as low as in a class AB amplifier.

In this way feedback can have a strong effect on bias, despite there being no DC feedback path, and not even a low-frequency feedback path!

We can also employ feedback to implement **frequency response compensation**. With BJT amplifiers this can be quite important, because many BJTs used for amplifiers operating in the HF frequency range have an Ft of just 100MHz or little more, so their current gain drops with increasing frequency at a steady rate through most of our intended operating frequency range. The gain would be very much lower at the high end of the range than at the low end. Proper compensation for this can be included in the drive path, or in the feedback path, or even in both of them. Since this involves inserting reactive components, such as inductors in series with the shunt feedback resistors, great care has to be used to make sure the amplifier remains stable. The general idea is to reduce the strength of the negative feedback as the frequency rises, to compensate for falling gain in the transistors. This also tends to keep the input impedance more constant, since the base or gate impedance of transistors drops with increasing frequency.

MOSFETs used at HF have a much flatter frequency response, because their Ft is so high that usually we don't need to worry about it when operating at HF. But their input impedance drops even more strongly with frequency than that of BJTs, and this is one of the limiting factors to a MOSFET's practical operating frequency range in a broadband amplifier. Not so in a tuned amplifier, where the capacitances can mostly be tuned out. That's why a given MOSFET might work well in a tuned 200MHz amplifier, while being suitable only for up to 50MHz or so in a broadband amplifier.

Given the high gain of LDMOSFETs, usually we can get enough gain flatness by simply using plain simple feedback, strong enough to lower the gain so much that our intentional feedback controls the gain throughout the frequency range, rather than the internal drain-gate capacitance. And we also either swamp the input capacitance of the MOSFETs through strong feedback, or we can help the situation by absorbing it into a lowpass filter section, as explained in the chapter about driving. Again, this involves reactive elements, and care must be used to avoid instability.

And since I have been writing so much about stability and oscillation, it's time to begin a proper chapter about it!

Amplifier stability

Let's begin with Murphy's Law of Amplifier Stability: *Any amplifier that can oscillate, will oscillate*. There are several extensions and additions to this law. For example: *An amplifier that is conditionally stable will start oscillating at the worst possible time*. That might be precisely when you demonstrate your amplifier during an international symposium about RF design, or while an EMC inspector has his spectrum analyzer connected to it. Amplifiers that go into satellites will start oscillating precisely when they reach orbit, never earlier. There is also a combinational extension to this law: *When you connect a stable driver to a stable final stage, the combination will oscillate*. There is even a corollary that applies to individual components: *Whenever an amplifier gets damaged from oscillation, the most expensive and hardest to replace component is the one that will fail*. A sub-corollary to this is: *The likelihood that an RF power transistor will be destroyed during oscillation is directly proportional to the cost of that transistor.*

So we clearly need to take all required measures to make sure that our amplifiers just cannot oscillate, no matter how hard they try. There should be absolutely no condition, not even the least likely one, that could result in oscillation. But this, my dear readers, is much easier to write or say, than to achieve!

Let's review the theory of oscillators: It simply says that any device that has feedback will oscillate, if there is any frequency at which the feedback is exactly in phase, and there is enough gain to overcome the losses in the entire feedback loop. So you need just three things to get oscillation: Gain, feedback, and a suitable phase of the feedback.

Any amplifier has gain, or it wouldn't be an amplifier. So we can't attack oscillations by eliminating gain, at least inside the required operating frequency range. One out, two to go.

You might think that it would be wisest to avoid building feedback into an amplifier, to prevent oscillation. Unfortunately this isn't possible, because there is always some feedback. Every transistor has shunt feedback through the drain-gate or collector-base capacitance, also there is always some feedback through the impedance of the emitter or source connection, and there is additional feedback through capacitive and inductive coupling between the input and output circuitry, and by feedthrough on the power supply line into the bias supply, and into the main supply of the other stages in an amplifier chain, and, and, and... So, feedback is unavoidable, and instead of going to extremes in fighting it (called "neutralization"), it's very often easier and better to overwhelm the inevitable undesired feedback with strong, intentional, correctly designed feedback. One more out, just one to go.

And that's the phase of the feedback. In any amplifier, or chain of amplifier stages, we must make absolutely sure that under no condition there is any frequency at which the phase of the feedback is the correct one for oscillation, while at the same time there is any gain in the loop. This is the golden rule of stability!

The phase required for oscillation, of course, is such that a signal applied to a point of the loop gets amplified, fed back, delayed, inverted, inverted again, shifted forward, shifted back, whatever, and ends up returning to that same spot in the exact same phase as it was injected. It doesn't matter whether it's phase shifted by 0°, 360°, 720°, or even more full periods. Any of those is good for oscillation.

When we design an amplifer, in principle we apply negative feedback, that is, the feedback is 180° shifted from the original signal. That's a simple phase inversion. At least that's our first goal. Inevitably there will be some additional phase shifts, forward or backward, that make the feedback phase angle deviate from the optimum 180°. Our second goal is to keep the combination of all these shifts under control, in such a way that there is no frequency where they add up to an additional 180° shift, which would put the feedback signal in phase with the drive, and at which there is still gain.

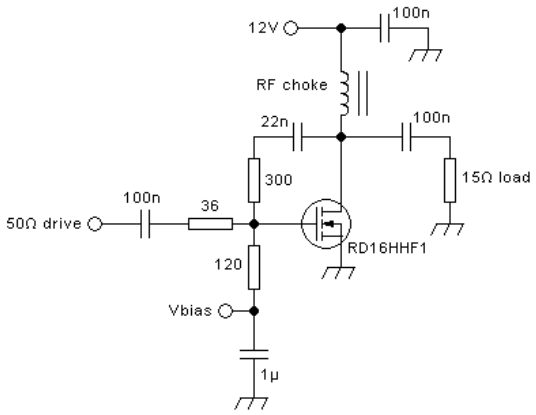
A feedback loop will self-oscillate only when the total phase shift is exactly right. The oscillation will automatically happen on the frequency where this perfect phase shift happens. That's why oscillators can be tuned by varying a capacitance or inductance, after all. If there is no frequency where the correct phase shift happens, there will be no self-oscillation, but when the phase shift is close to the required one, the amplifier will *ring*. This means that when excited near that frequency, or with a pulse that contains all frequencies, this will kick-start an oscillation at the resonant frequency, with an amplitude that decreases over time until fading into the noise. The closer the phase of the feedback is to the critical one that causes self-oscillation, the stronger and longer will be this ringing. Likewise, the higher the loop gain is, the more ringing will occur.

How much ringing is acceptable, depends on the application. A rule of thumb that's often used is that the designer should keep at least ±45° of phase margin, ideally ±60°, and at least 6dB of gain margin, ideally 10dB. Translated into a slightly more human-compatible lingo, it means that the total phase delay, including phase inversion, along the complete feedback loop, must be anywhere between 45° and 315° at all frequencies where the loop gain is at least unity, and at any frequency where the total phase delay is 0°, the loop gain must be ≤ -6dB. And if you can get the delay to fall between 60° and 300° at all frequencies where the loop gain unity or higher, and the gain to be ≤ -10dB at any frequency where the phase is at 0°, you can call it an excellent result. These numbers don't need to be super precise. When comparing signal amplitudes on a

scope screen, it's plenty accurate enough if we consider 6dB to be a ~2:1 voltage ratio, and 10dB a ~3:1 ratio.

Of course it's almost impossibly difficult to calculate the phase shift and gain for every frequency, in an amplifier that has many unknown values, such as varying and poorly specified internal capacitances of transistors, unknown wiring inductances, and ferrite-cored chokes that vary their inductance and loss in a non-linear way according to the current flowing through them. I often do just some educated guesses to determine whether there is an obvious problem in my design, or if it seems to be on the safe side. Then I try it. This is really tickling Mr. Murphy and asking for trouble, but with some care it works, and is much more fun than filling 50 sheets of paper with extremely complex equations - and making one maths mistake that messes up the whole calculation and gives a totally wrong result!

Enough said about the theory. Let's see how it's done in practice, by analyzing the stability of a real amplifier. For simplicity, let's use one of the examples I used in the previous section:



Let's assume that the RF choke has a value of 22μH, that the 12V and bias sources have zero internal impedance, and that the drive signal source has a 50Ω internal impedance, free from reactance.

To start, imagine what happens if there was a signal of 1mV on the gate, at 2 MHz. We calculated way above in this page that this transistor, operating at an idling current of 1A, has a transconductance of roughly 1.4S. Thus the 1mV gate signal will cause a 1.4mA drain current signal. At this low frequency, the delay in the MOSFET is negligible, so we can consider the drain current to be precisely in phase with the gate voltage. But the current is sunk rather than sourced, so it's actually 180° out of phase. Since this phase inversion isn't associated to time delay, it makes the same sense to consider it positive or negative. For the fun of it, let's consider it to be -180°.

To calculate the resulting drain voltage signal, we need to calculate the total impedance seen by the drain. This is the result of the 15Ω load in series with the 100nF coupling capacitor, in parallel with the 22μH of the choke, in parallel with the series connection of the 22nF and the 300Ω, with the added complication that this latter branch does not return to the RF ground, but the the 1mV gate signal point, and at this moment we don't even know the phase difference between the drain and gate voltages!

This is too complex to be practical, so let's make the approximation of considering 1mV to be negligible. This is quite daring, but let's try to get away with it. So we have 3 branches in parallel, two of them composed by resistance and reactance, and to this we must add the two internal branches of the MOSFET: The drain-source capacitance, and the drain-gate capacitance, where the latter again requires either considering or neglecting the 1mV gate signal! To put the cherry on the pie, these two capacitances vary like crazy during the signal period, and also their averages over the period vary with amplitude... But hey, we are doing small-signal analysis here, so let's consider them constant!

You see, the matter is complex, and we need to make practical simplifications to ever get anywhere.

Resolving this network of impedances at the drain, which can be done by hand and calculator, or using software, or even using online calculators, I get 14.3Ω at a phase angle of -0.3°. Which means that at this frequency the reactances in the circuit have a very small effect, and largely compensate each other. The resulting impedance is almost exactly just the 15Ω load in parallel with the 300Ω feedback resistor.

This makes things nice and easy. The 1.4mA drain signal current, when applied to this impedance, creates a drain voltage signal of 20mV. The phase of this voltage signal is -180.3°. Great.

Now let's see what the shunt feedback does to the gate voltage, to complete the loop: 300Ω in series with 22nF is an impedance of 300.02Ω at a phase angle of -0.7°. A negligible difference from a pure 300Ω resistor. Let's add the drain-gate capacitance of the MOSFET, about 2.2pF, in parallel. The result is a tad less than 300Ω, at -1.2°.

Due to the almost exact phase opposition between gate and drain voltages, there is 21mV across the 300Ω, which injects a current of 70μA into the gate note, at a phase angle of -179.1°. That's because dividing a voltage by an impedance that has a negative phase shift results in a current with a positive phase shift, so the phase is -180.3° +1.2°.

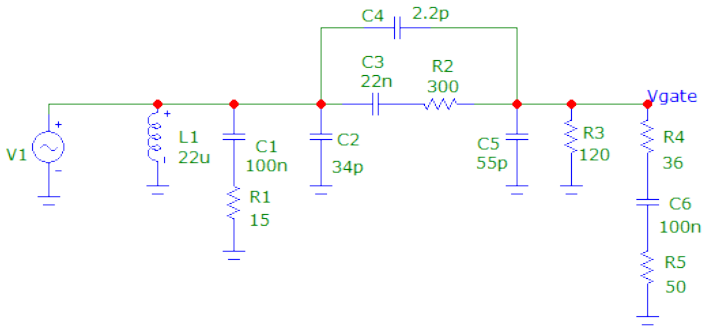
This current sees 3 parallel paths to ground. One is the gate-source capacitance of the MOSFET, 55pF. Then there is the 120Ω bias resistor, which is RF-grounded at the other end, and the series combination of the 36Ω input resistor, the 100nF coupling capacitor, and the 50Ω internal resistance of the drive source. Resolving this network, I get almost exactly 50Ω, at -2.3°.

70μA in 50Ω makes 3.5mV, and the phase current of -179.1° in an impedance of -2.3° makes a voltage phase of -181.4°.

So, the loop gain of this amplifier at 2MHz, under the conditions shown and assumed, is 3.5 (a little more than 10dB), with a phase shifted only -1.4° from perfect negative feedback. This is very good, and there is not the faintest chance of any oscillation at 2MHz.

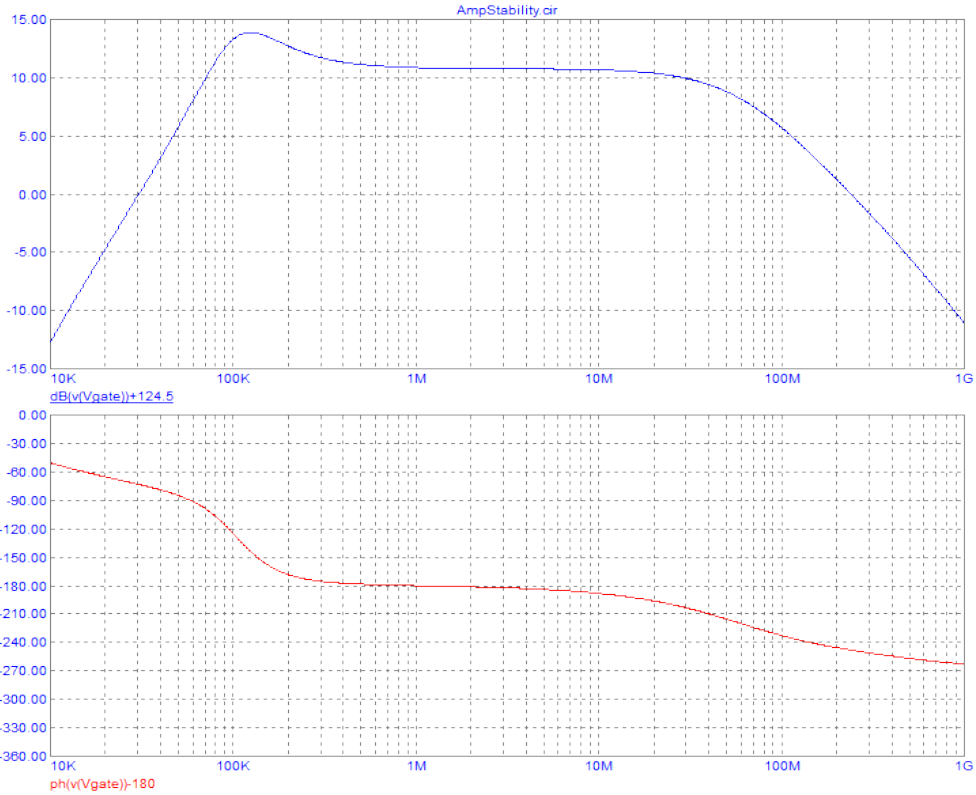
One could do a stability analysis by putting all these calculations into a little program, or even a spreadsheet, and having the computer calculate the loop gain and phase for a very wide range of frequencies, from 1Hz to 1GHz or so, and make sure that the conditions for oscillation don't happen at any frequency, and that the phase and/or gain margins are OK. Or if you are fluent using a circuit simulator, this is a great opportunity to use it, but you need to use it carefully, because transistor models in simulators are often very poor, and lead to wrong results. I prefer simulating just the passive circuitry in my simulations, and add myself what I assume the transistor should be doing.

I put a passive model of the feedback circuit into Micro-CAP:



The signal source is configured to create 1.4kV, with a 1MΩ internal resistance, to create a pretty well defined 1.4mA current injection into the circuit. This simulates the drain current signal created by 1mV on the gate. L1 is the supply choke, C1 the output coupling capacitor, R1 is the load, C2 is the MOSFET's drain-source capacitance, C4 its drain-gate capacitance, C5 its gate-source capacitance. C3 and R2 are the shunt feedback parts, R3 the bias resistor, R4 and C6 the input parts of the amplifier, and R5 is the official signal source's internal resistance.

Then I made a sweep from 10kHz to 1GHz, to create a Bode plot (gain and phase) of the feedback loop. I adjusted the gain scale to correct for my crazy current source and the transistor's gain, and I added the 180° phase shift occuring in the transistor. This is the result:



In the desired operating frequency range, from 1.8 to 30MHz, the feedback is of pretty constant strength, and its phase drops from the nominal -180° to about -200°, so it's excellent negative feedback through that whole range.

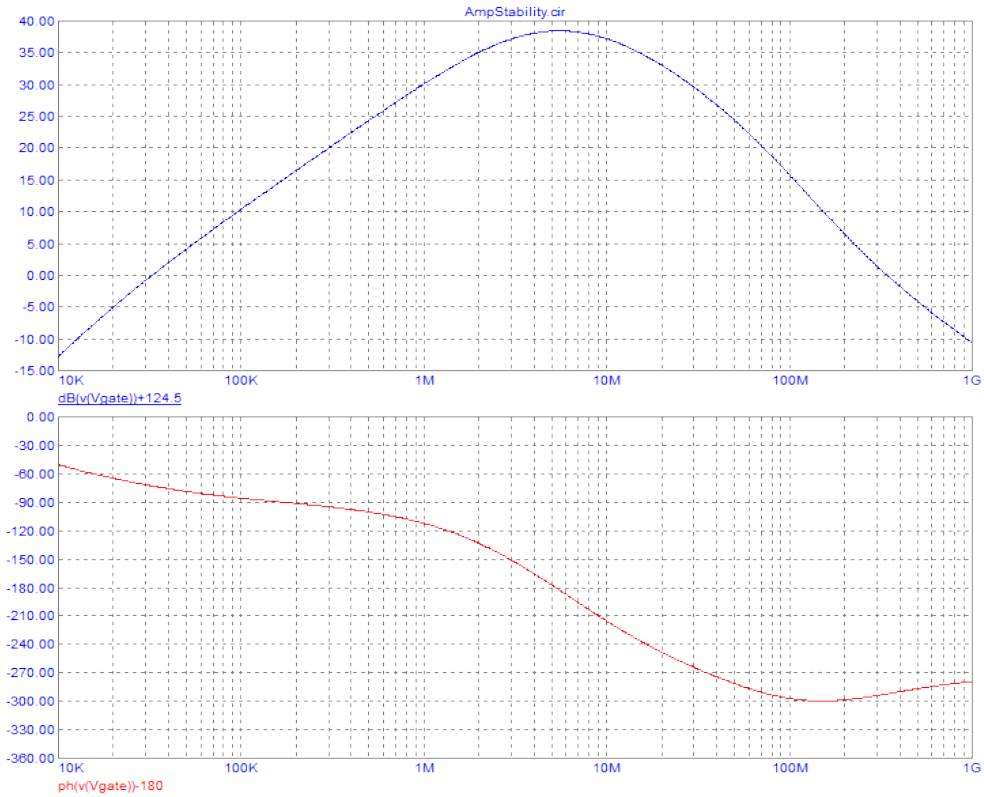
In the low frequency area we see a gain peak, then a fall-off, along with the corresponding phase shift. Unity loop gain (0dB) is reached at 30kHz, and the phase margin is about 72°, which is excellent. The frequency at which the feedback loop reaches zero phase is way out of the graph, and the loop gain at that frequency is way below unity, probably -20dB or even more down, so the gain margin is very ample. There is no risk of low frequency oscillation.

At the high frequency end this Bode plot is not accurate, because it does not include the effects of the transistor's internal delay and gain drop. This data isn't given in its datasheet, so I have to go by feeling and experience, or else built a test setup and measure this MOSFET's behavior. Anyway, without including the VHF/UHF effects contributed by the MOSFET, this amplifier has an excellent stability profile, with a unity loop gain crossing at roughly 230MHz, and a phase margin of about 110°. When the roll-off of the transistor is added, this will get worse, but hopefully not nearly enough to make the amplifier oscillate at VHF.

So there is a good chance that this amplifier will be unproblematic.

But I was analyzing it under optimal conditions! A clean resistive load, and a signal source with a constant, clean internal resistance. In practice this is very often not so, and depending on what load the amplifier has to drive, and what impedance the signal source has, the picture could be quite different! For a real application, we need to simulate the loop gain and phase under the actual operating conditions, with the true load and true signal source connected.

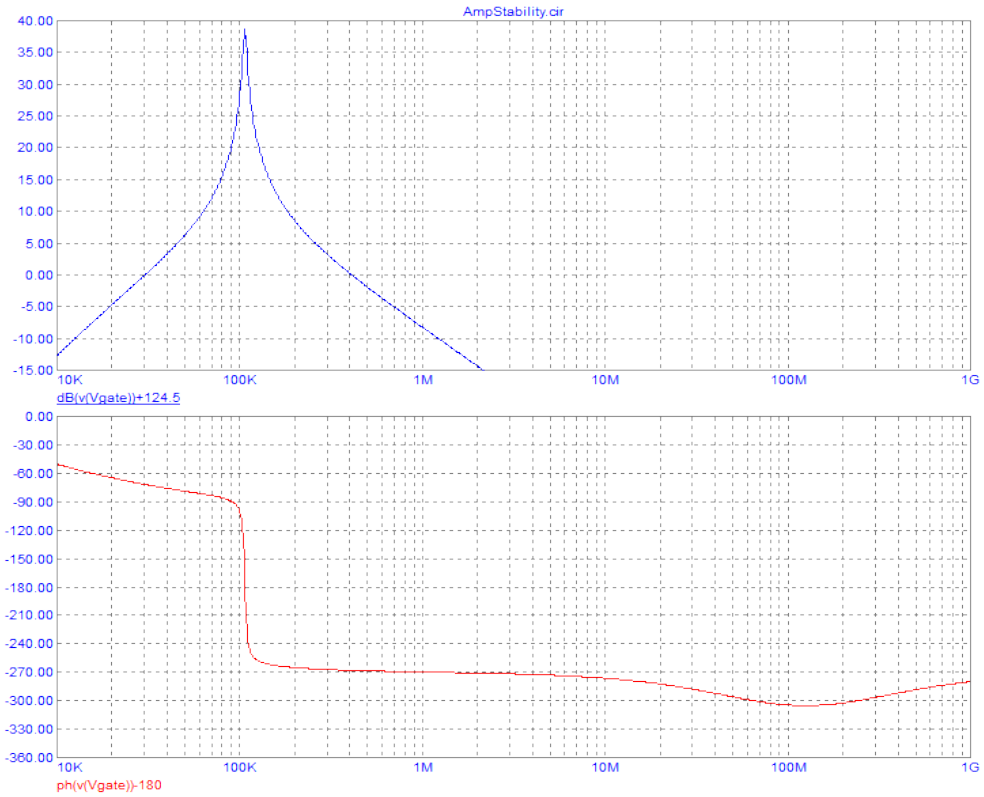
Let's use the simulator to play the "what if" game. First test: Remove the load, let the amplifier run unloaded. The Bode plot changes to this:



Note the change of scale in the gain graph! There is a very high loop gain around 6MHz, but it's no problem, because the phase stays close to -180°. At the low-frequency end there is no change, because there the very low impedance of the supply choke dominates over the load, or the lack of it. But watch the range around 100MHz: The phase margin is only 60° with a loop gain of 10 to 15dB, which is no problem by itself, but after adding some additional phase delay in the MOSFET, things could get tight here! There is some risk that the phase might drop all the way to -360° before the loop gain has dropped to 0dB, and in that case the amplifier will do what amplifiers love more than anything: Oscillate!

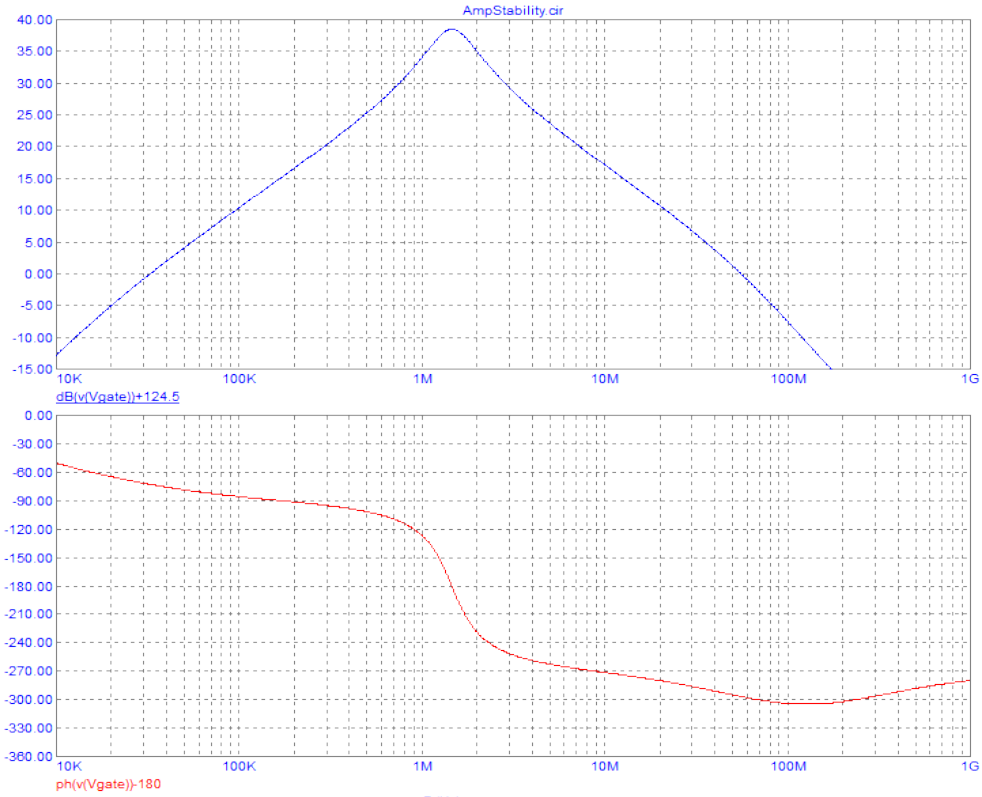
So, in case that the application of this amplifier might include an open-load situation, it would be wise to find out the transistor's time delay and gain curve, and include it in the simulation. Another possibility is just building the amplifier and testing it without a load, at various supply voltages, bias settings, temperatures, and see if it does any mischief. If it oscillates, the frequency would be somewhat lower than 100MHz. If it doesn't oscillate, it would be good to inject a small signal, sweep it over the suspect range, and see how much the amplifier rings.

And what happens if we short the load? Just see:



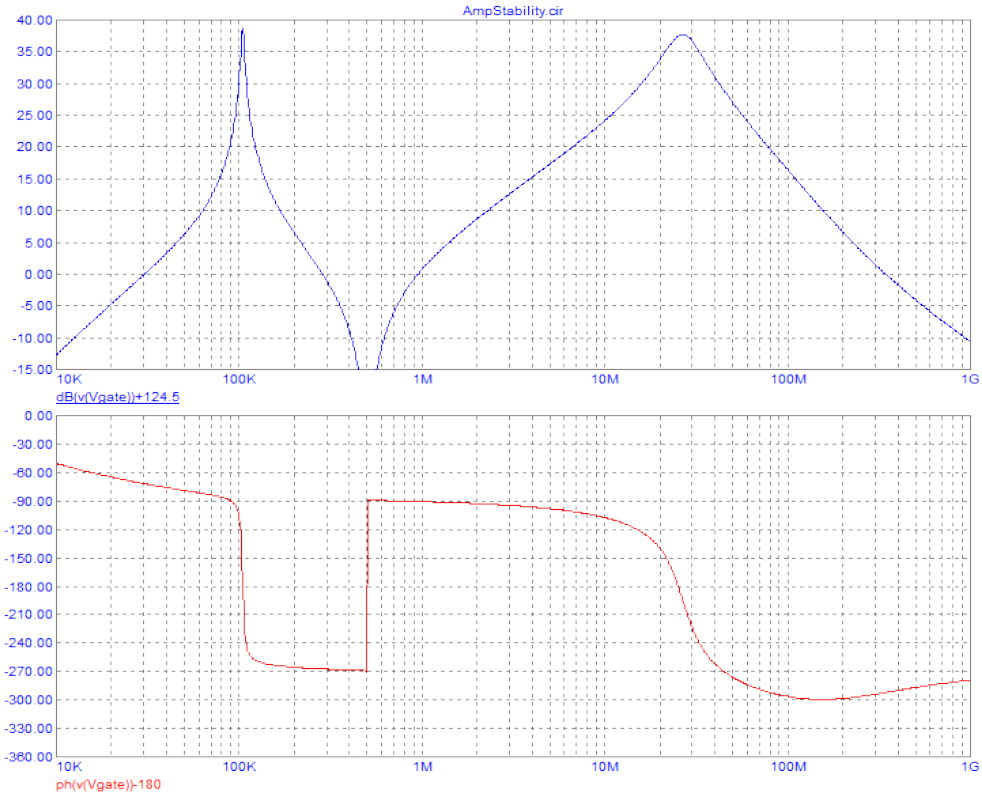
We get a nasty, sharp resonance near 100kHz, because the feed choke and the output coupling capacitor form a parallel resonant circuit at the drain. The phase changes through almost 180° in a very small frequency span. But still there is no problem, because the phase margin on the low end is essentially unchanged at about 72°, and on the high side it's slightly over 90°, at only 400kHz, where the MOSFET adds almost no delay. So there is no risk of oscillation when shorting the load.

And what happens if the load is purely capacitive? Well, that depends on the capacitance value. Let's try 500pF, and see what happens:



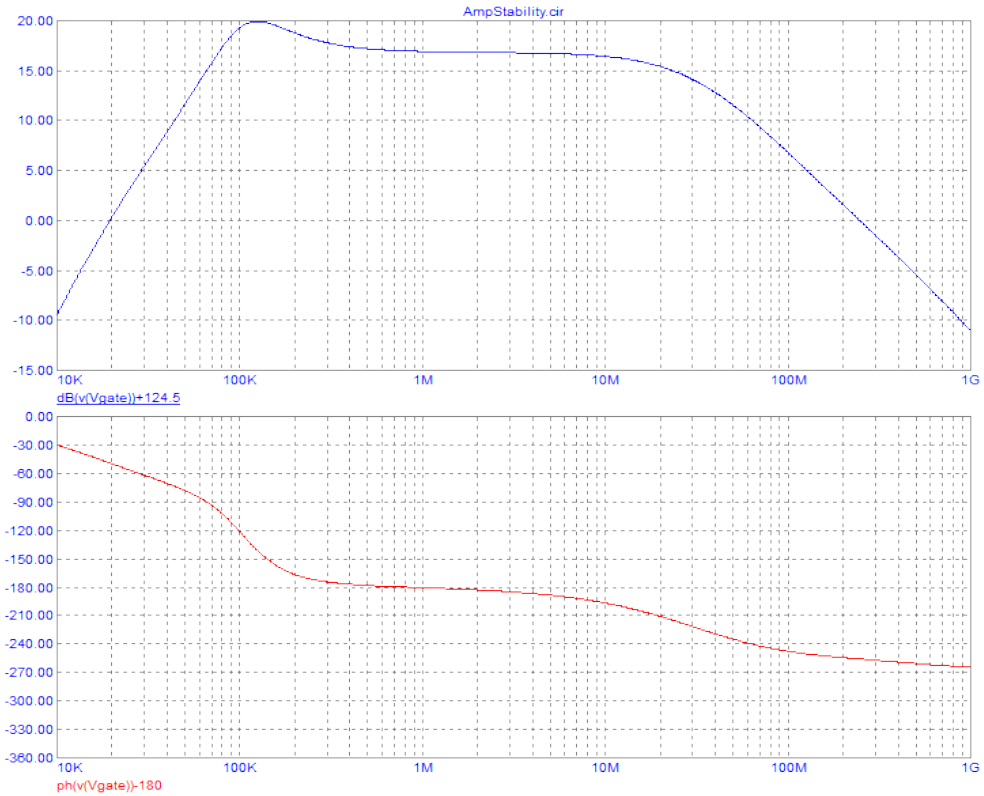
The resonance moves up, and broadens. It's simply an intermediate case between the shorted load, and the open load. We need to begin watching what happens when adding the transistor's delay.

And what if the load is inductive? Let's try with 1μH:



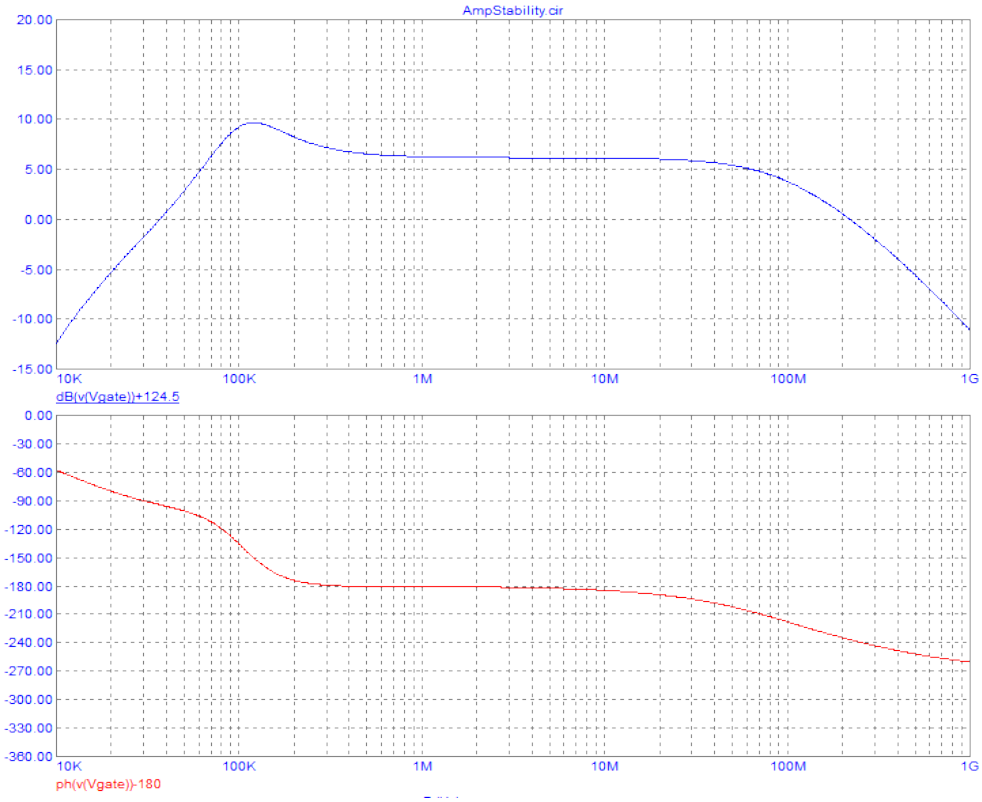
The plot turns pretty wild! The inductive load forms a series resonant circuit with the coupling capacitor, at 500kHz, while at 100kHz this inductance turns quite insignificant, acting as a short and allowing the coupling capacitor to form a the same parallel circuit with the feed choke as it did before. The important point, for stability purposes, is that the phase margin is still about 72° at the gain zero-crossing frequency of 30kHz, so there will be no low frequency oscillations, and that around 100MHz we still get some risk of oscillation, if the transistor adds significant delay there, which I don't know.

Then I returned to the normal resistive load, and instead changed the driving source to have infinite internal impedance. This is what you would approximately get by using a driver stage that has no shunt feedback:



Note the change of scale again. And note that due to the lower loading of the gate circuit, the whole loop gain curve is higher, resulting in the low frequency 0dB crossover shifting down to 20kHz, with a phase margin of only 50°. This is still OK, but approaching the acceptable limit. On the high frequency side instead there isn't much of a problem, with a 0dB gain transition at about 240MHz, and a phase margin of about 100°, which gives room for quite some delay in the transistor, without anything bad happening.

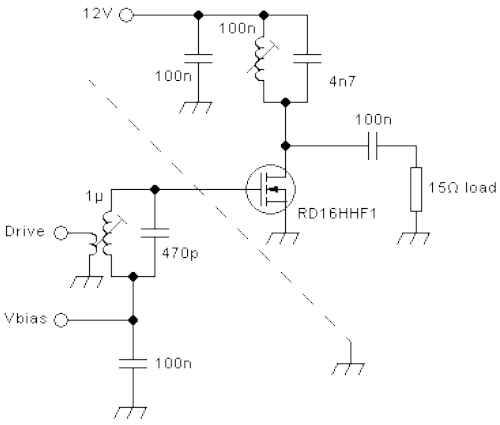
And then let's try a driving source with zero internal impedance. An RF operational amplifier would have that behavior, and also a transistor driver using very strong shunt feedback. An emitter follower stage can also approximate that behavior. What we get is:



It's a very benign curve, with phase margins better than 90° on the low side, and about 120° on the high side. This shows that using strong shunt feedback in each stage of an amplifier chain can be a very good idea! The 36Ω input series resistor, grounded through the zero driver impedance, gives us this advantage. A low gate swamping resistor to ground would do the same for a high-impedance driver. Of course, it would reduce the amplifier's gain too.

Now I will show you examples of **what really makes amplifiers oscillate**! There are so many pitfalls that I can only mention a few, to give you an idea. First example: Let's suppose that there is this new ham, who is learning to build radios. He bought an RD16HHF1 transistor, and wants to build an amplifier with it. He doesn't want the complexities of band-switched low-pass filters, and in fact he doesn't fancy multiband operation at all, in addition he lives in a country where getting ferrite cores is hard, so he decided to make a tuned amplifier, just for the 40m band.

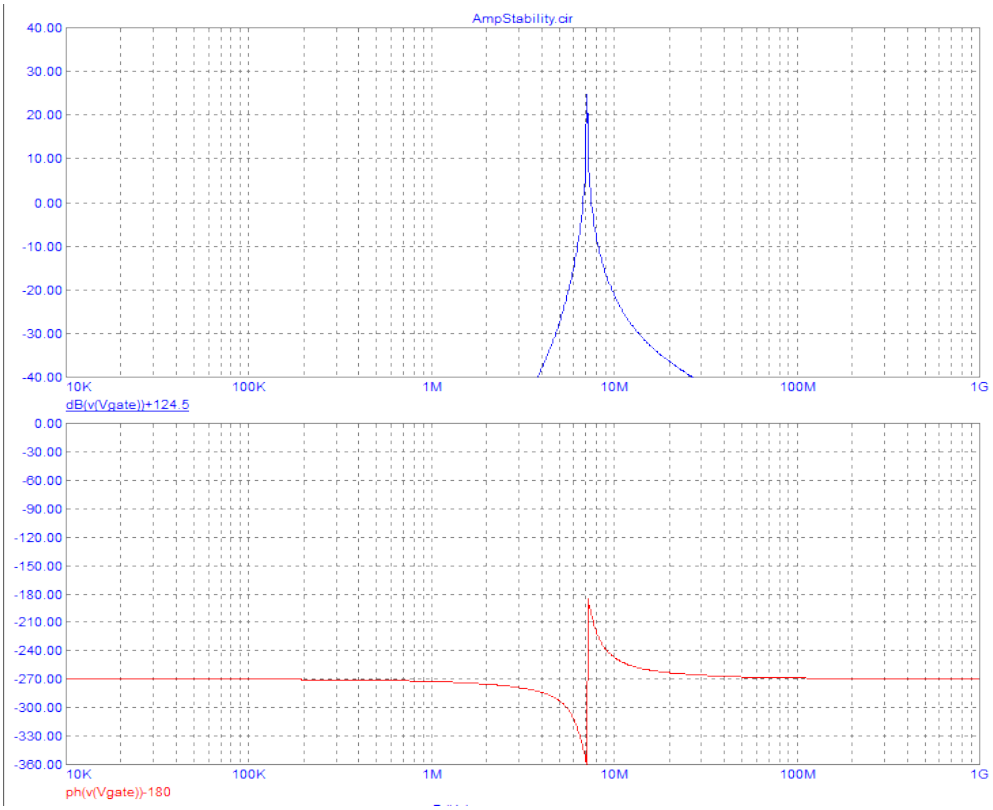
Since he knows that tuned amplifiers are prone to oscillating, he decides to be very careful to avoid feedback paths. He installs very good power supply decoupling, bias supply decoupling, and he avoids using shunt feedback, for fear that the negative feedback will end up turning into positive feedback at some frequency. He even is so careful as to add a shield between the input and output circuitry, to prevent any coupling between the tuned circuits. The result of his effort is this:



He builds the circuit nicely into a shielded box, and as soon as he connects the power supply and bias, he gets a strong, solid carrier at the output, right in the 40m band - before even connecting a drive source!

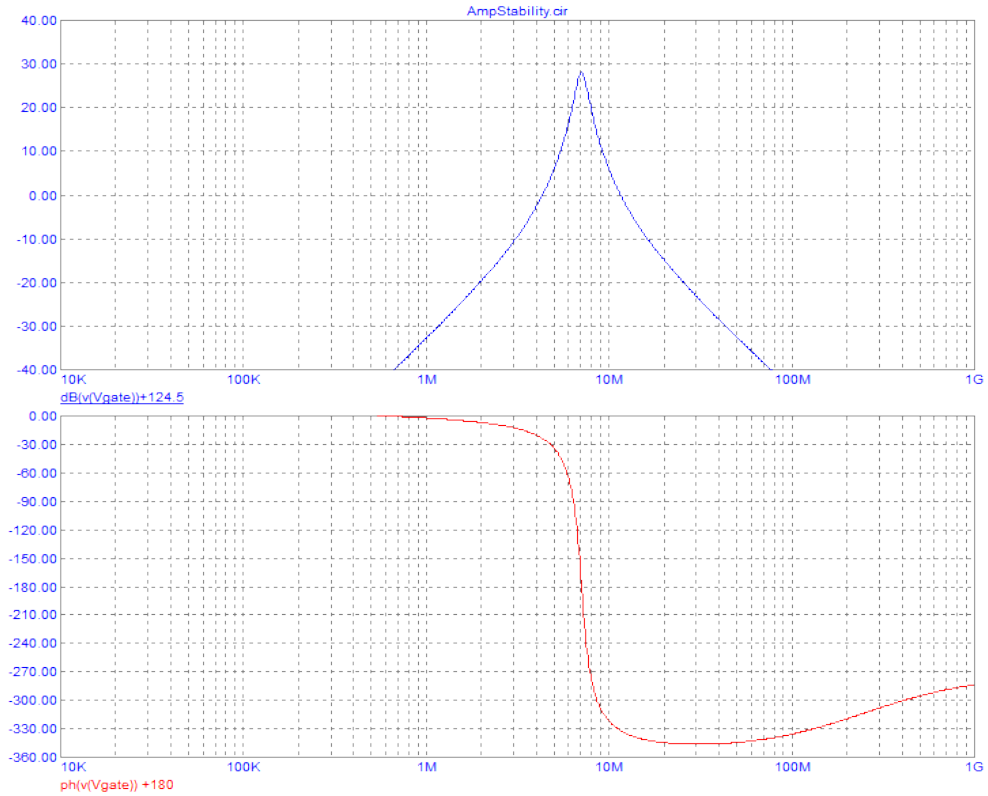
He has built an oscillator.

How? Well, there is feedback, of course, through the drain-gate capacitance of the MOSFET. That's a nasty feedback, because it's 90° phase-shifted, being capacitive. And the two resonant circuits provide all the additional phase shift needed for oscillation. The Bode analysis shows it clearly:



You can see a huge resonance, which is to be expected in a tuned amplifier with undamped input circuit. The phase of the feedback is -270° over most of the frequency range, but touches -360° at the lower side of resonance. At that frequency the loop gain is a little above 0dB, so we have perfect conditions for oscillation.

Let's try to fix this amplifier. First, I will add shunt feedback, with a 300Ω resistor. The resonance broadens, and the phase response changes dramatically:

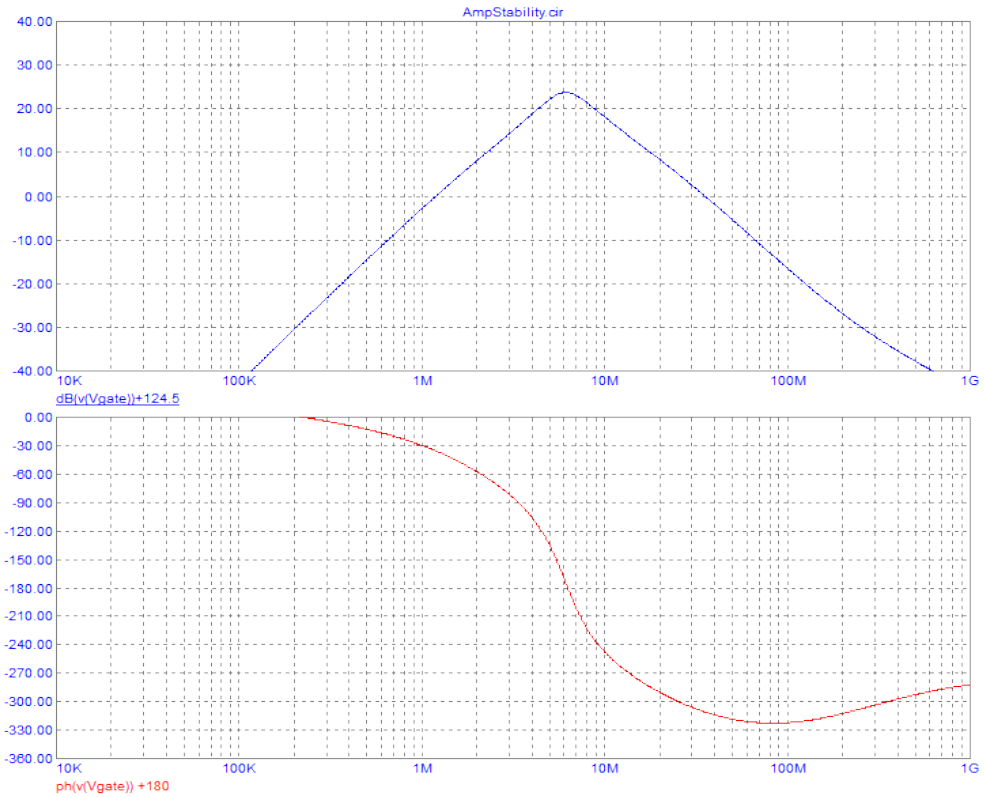


Why this drastic difference in the phase plot? Well... Two resonant circuits can cause twice as much phase shift as a single one. In the previous example we had no resistors between the two resonant circuits, so at the exact resonant frequency they were strongly coupled through the drain-gate capacitance, acting as a single resonant circuit. But now, with resistive feedback overwhelming the capacitive one, we actually separated the two resonant circuits, getting twice as much phase shift! I had to change the -180° addition for the transistor's phase inversion to +180°, to bring the curve back into the graph's range, because it had descended out of range!

Despite the larger phase shift range, the situation is now much better. At the frequency of maximum gain, the feedback is nicely centered on -180°. The phase margin at both the lower and the upper gain crossover frequencies is about 25°, not really good, but at least the amplifier now wouldn't oscillate. It would just ring like crazy.

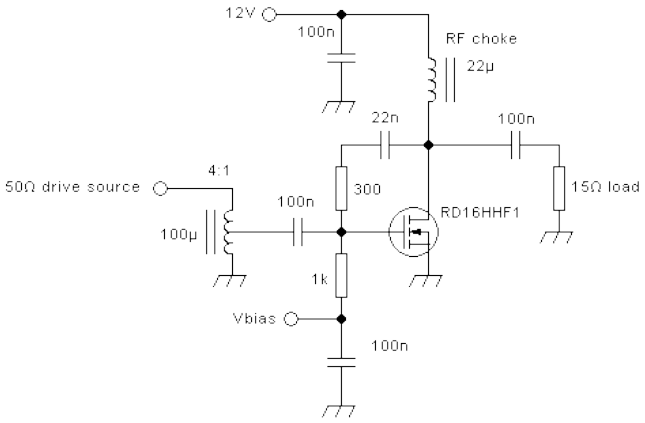
Adding a drive source with an internal 50Ω helps a little, but not enough. And anyway an amplifier should be inconditionally stable even if the input is left open.

Then I drastically de-Q-ed the circuit, by changing the output tank to 1μH and 450pF, and the input tank to 3μH and 150pF. The drive source has 50Ω, the input matching link is 30% of the turns of the main coil. The result is:



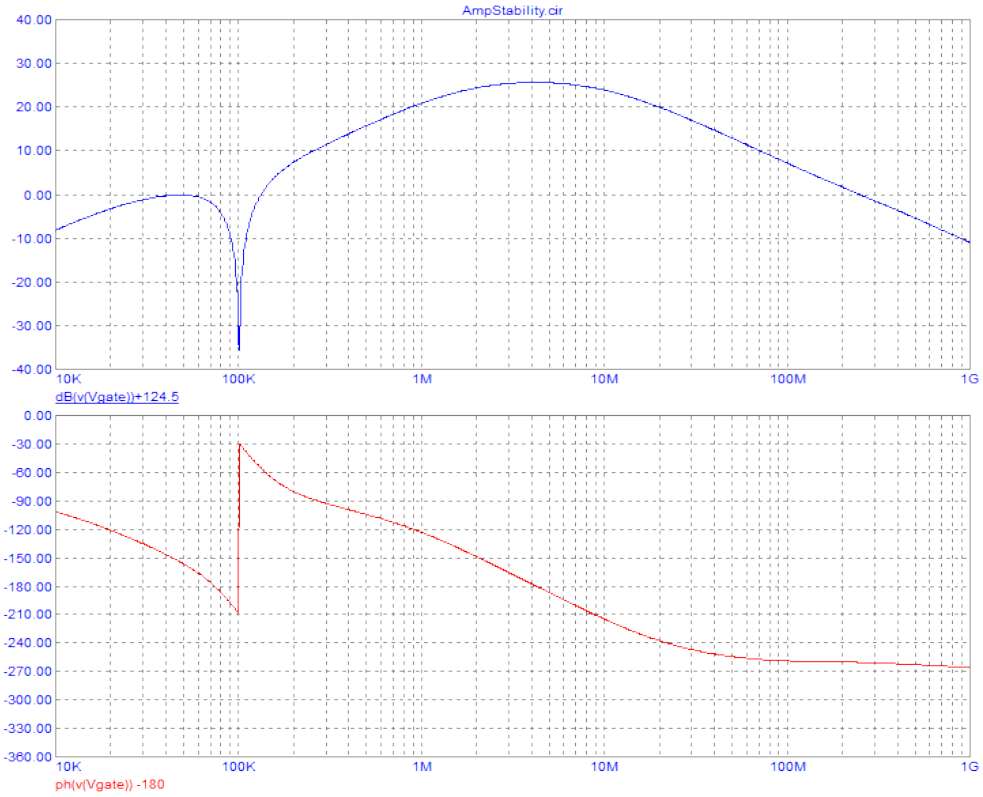
So the resonance is much broader, the phase shift happens over a much wider frequency range, the phase margin improved to 35° on the low side and 47° on the high side. That's better, but still a little in the ringing range. The conclusion is that tuned-input, tuned-output amplifiers are always a little touchy, and it's a good idea to make the loaded Q of any tuned circuits only as high as strictly necessary. This also goes to the benefit of efficiency.

Many times people run into trouble caused by resonant circuits they don't even see. Consider this variation of out pet amplifier:



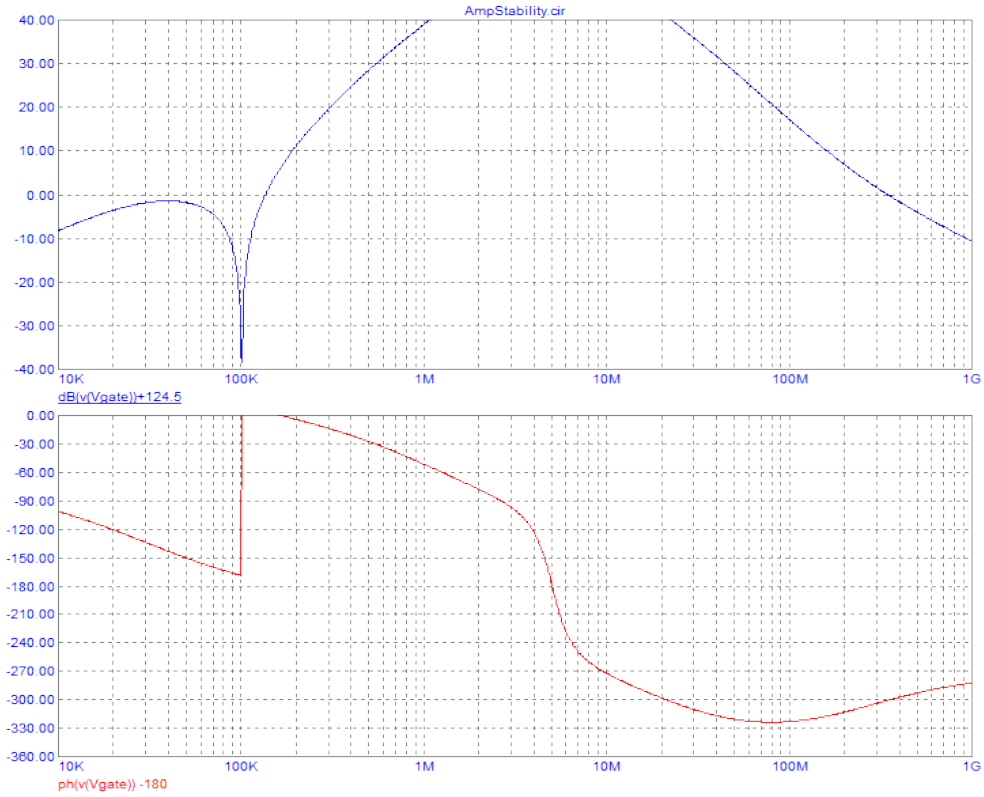
It's a plain broadband amplifier with a 4:1 input transformer, simple choke-capacitor output, a good shunt feedback to achieve good linearity, stability, and flat frequency response, and even a bias injection resistor doubling as gate swamping resistor at those frequencies where the 100nF input coupling capacitor, and the 22nF feedback capacitor, become ineffective. It should work fine, right?

Well, here is its feedback Bode plot, as usual without including any delay caused by the transistor:



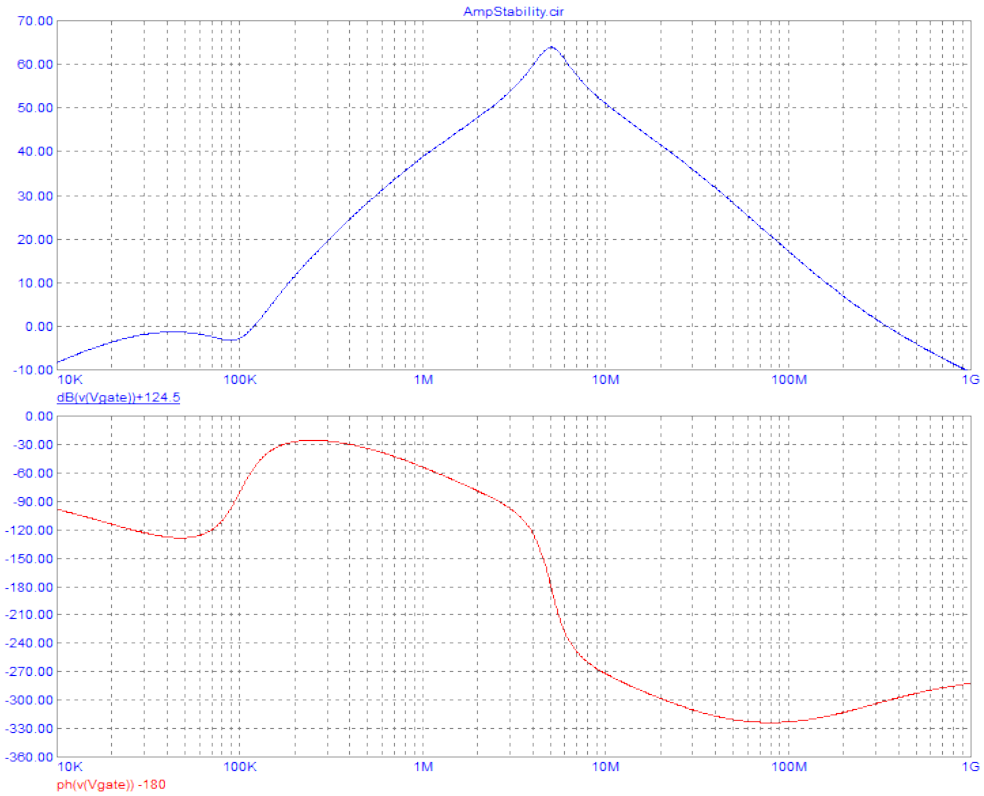
It's pretty much as expected, except for the huge series resonance of the input transformer with the 100nF coupling capacitor! That one causes a sudden 180° phase shift, but it happens at a frequency where the loop gain is already low enough, so it's OK. The phase margin is roughly 60°, which is actually fine. Note that below the resonant frequency the gain rises just to 0dB again, but in that range the phase is fine. So this circuit won't oscillate.

But what happens if the load is disconnected? Hmmm... Let's see... No-load conditions are always risky, because then the transistor works into the choke's inductance, adding a large phase shift to the feedback signal.



Yep, we are in trouble! With the load removed, the phase of the feedback crosses zero at a frequency where the loop gain is about 8dB. Our amplifier will oscillate if the load is disconnected.

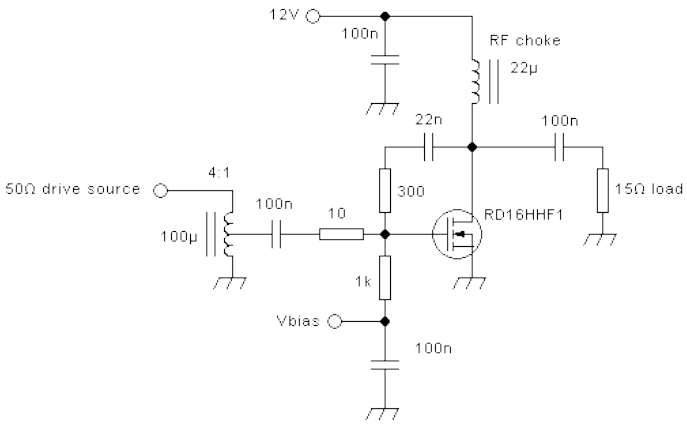
What can we do to fix that? Well, we can use some damping of that series resonance, by adding a resistor in series with that input coupling capacitor. I chose 10Ω, just to try. The result is:



Well, much better! The series resonance at 100kHz is now quite weak, and the phase shift it causes is far less dramatic. The phase margin is still not great, but good enough to avoid oscillation. Since we don't have any drive in that very low frequency range, a tendency to ring down there isn't too bad. But we definitely should always avoid any risk of ringing inside the operating frequency range... And now look at the area around 80MHz! There too the phase margin is a bit tight, while the loop gain is still around 20dB. We need to watch what delay causes our MOSFET, to make sure it doesn't cause trouble there, but that's unlikely. We might end up with a poor phase margin at 30MHz, and ringing, but remember, I'm analyzing the amplifier operating without a load! It doesn't matter much if it works somewhat dirty, when no load is connected. The important point is that it doesn't break into potentially destructive oscillation.

By the way, the strong resonance at 5MHz is caused by the feed choke resonating with the transistor's output capacitance. Normally this resonance is very well damped by the load, but when the load is gone, we see it.

So, our fixed circuit, tolerant of load disconnection, is this:



I didn't check the input impedance matching now. Of course it's way off... A good practical approach is to design a circuit including such a series resistance in the input, calculating it for proper impedance matching, and then simulate the loop to see if it will be stable.

Do you like bad news? Stupid question, sure! But whether or not you like them, I need to tell you some: The above stability analisis is extremely oversimplified, because I considered only a single feedback path, the one inside the amplifier cell proper. A real circuit will have additional feedback paths: Through the power supply, through imperfect ground, through capacitive and inductive coupling between components and connections, and through the source or emitter lead inductance (very important)!

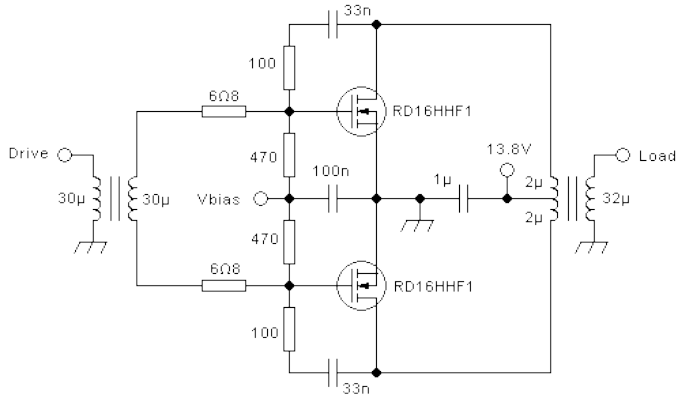
And in a transmitter we rarely use a single amplifier stage just by itself. The mixer or DAC used to generate the transmit signal on the actual transmitting frequency will typically deliver only about 1mW. But an HF transmitter typically needs to produce anything between perhaps 10W and over 1kW output power. So we need between 40 and over 60dB of gain, and that can't be done in a single stage. So we get absolutely beautiful chances to create a big, complex oscillator, due to the many feedback paths available! Feedback inside each stage, over two stages, over three stages, over all stages, even outside the amplifier block or whole transmitter, by coupling between input and output cables, and so on. Detecting, tracking down, understanding, and curing all those instabilities provides never-ending entertainment, joy and fun to any builder of RF

power amplifiers.

Yes, I'm being sarcastic.

Oh, I almost forgot: In any **push-pull** stage of course you need to understand, calculate or simulate the stability both for differential mode and common mode. Even if the common mode isn't used for amplification, it is present, and needs to be stable.

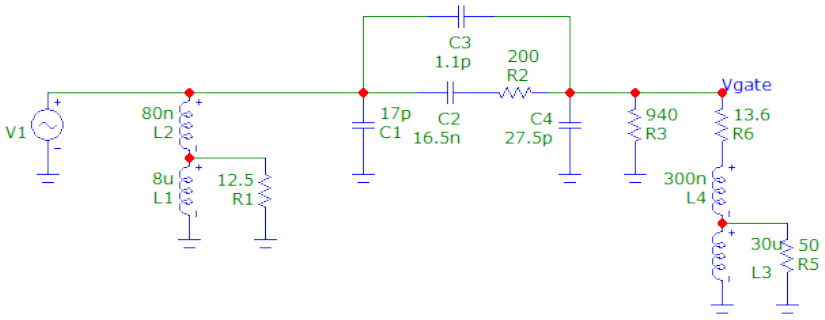
Let's consider this 20W class AB amplifer:



First let's calculate its basics: The output transformer has a primary of 1 turn on each side, and a secondary of 4 turns. I will assume that the leakage inductance of the transformers is 1% of each winding, which is pretty optimistic. When the load is 50Ω, the transistor that is conducting at any given instant sees a drain load of 3.1Ω in parallel with 2μH, the feedback path, and its own output capacitance. On a good day and with some luck, it might be able to pull down its drain to 2.5V or so, given that the RDSon of these MOSFETs isn't brilliant, so we have roughly 11.5V peak amplitude, or 8.1V RMS at each drain, and 32.5V RMS at the load, making a tad over 20W. Each transistor needs to conduct a peak current of about 3.7A. The transfer curve up this page shows that if we bias the amplifier to an idling current of 0.5A in each transistor, we need to pull the gate about 2.5V peak above the bias voltage. Since these 2.5V give us an 11.5V drain excursion, the voltage gain from gate to drain is 4.6. The 100Ω shunt feedback resistor, which sees 14V peaks, will conduct 0.14A peak, so at 2.5V peak on the gate it loads the gate node with 17.9Ω. The 470Ω bias/swamp resistor lowers this to 17.2Ω. The transistor's input capacitance has an effect at the high end of the HF range, but not a big one. The 6.8Ω resistor adds to the 17.2Ω to produce 24Ω load resistance on each side of the drive trafo's secondary. The two sides in opposition then have a drive input impedance of 48Ω, plenty close enough to 50Ω to use a little bifiliar-wound 1:1 balun as input transformer. 2.5V peak is required on each gate, and due to the 6.8Ω resistors, forming a divider with the 17.2Ω gate node resistance, 3.5V peak is needed at each side of the drive transformer secondary, making 7V peak total. That's very close to 5V RMS, and the same is needed on the primary side, thus this amplifier requires a drive of 0.5W to produce an output of 20W. So its power gain is 16dB. The low frequency limit is given by the mutual inductance of the transformers, the high frequency limit is a combination of the transformers, the transistors and the wiring, with the leakage inductance of the transformers being the main limiting factor. The feedback path has a low frequency cutoff of 48kHz. It has to be low enough to keep the feedback from phase-shifting at frequencies that the transformers will pass well.

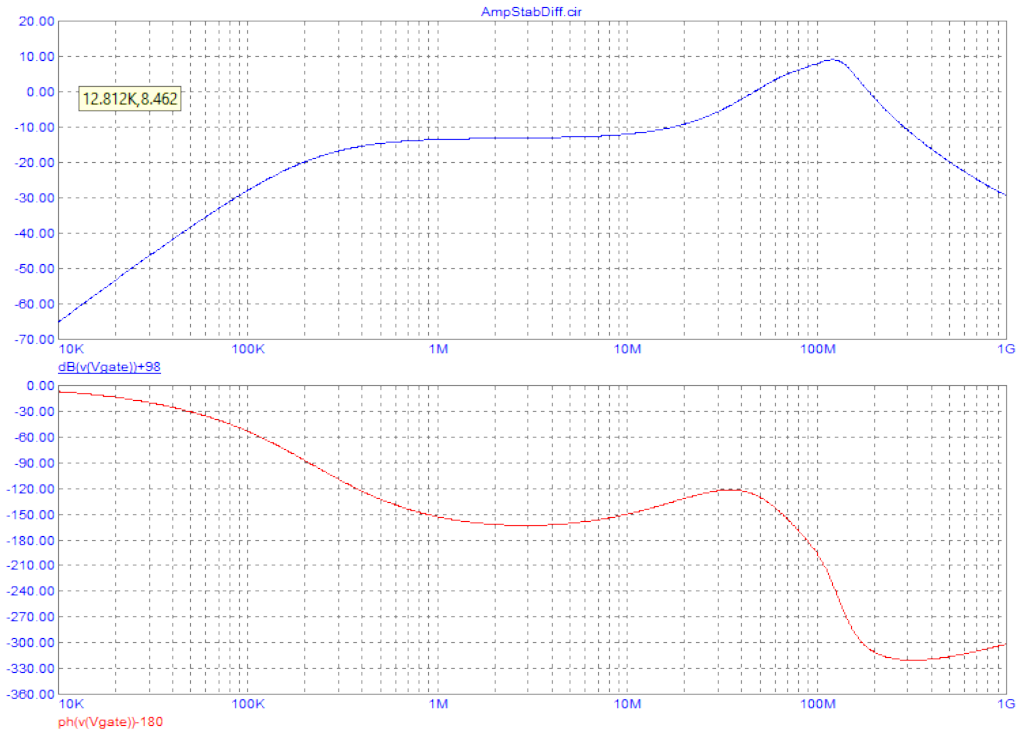
When oscillations happen, they start with microscopic amplitude, so we can do small-signal analysis for them. But they can start at any drive level, and at any point of the driving waveform, so we need to do this small-signal analysis not only at idling conditions, but also when a transistor is conducting a high current. Since its transconductance varies with drain current, stability conditions also vary. And since the capacitances vary with voltage, that requires even more stability analysis... Life isn't easy!

Let's make a first analysis in the differential mode under zero-signal conditions, assuming an idling current of 0.5A per side. Both transistors are on, and on the transfer curve diagram we can see that the transconductance is about 0.8S. So, by assuming a test signal injection of 1mV from gate to gate, one transistor draws 0.4mA more, the other one 0.4mA less drain current. For all analysis purposes they are in series, and so we can draw the following equivalent diagram for feedback loop analysis in a circuit simulator:



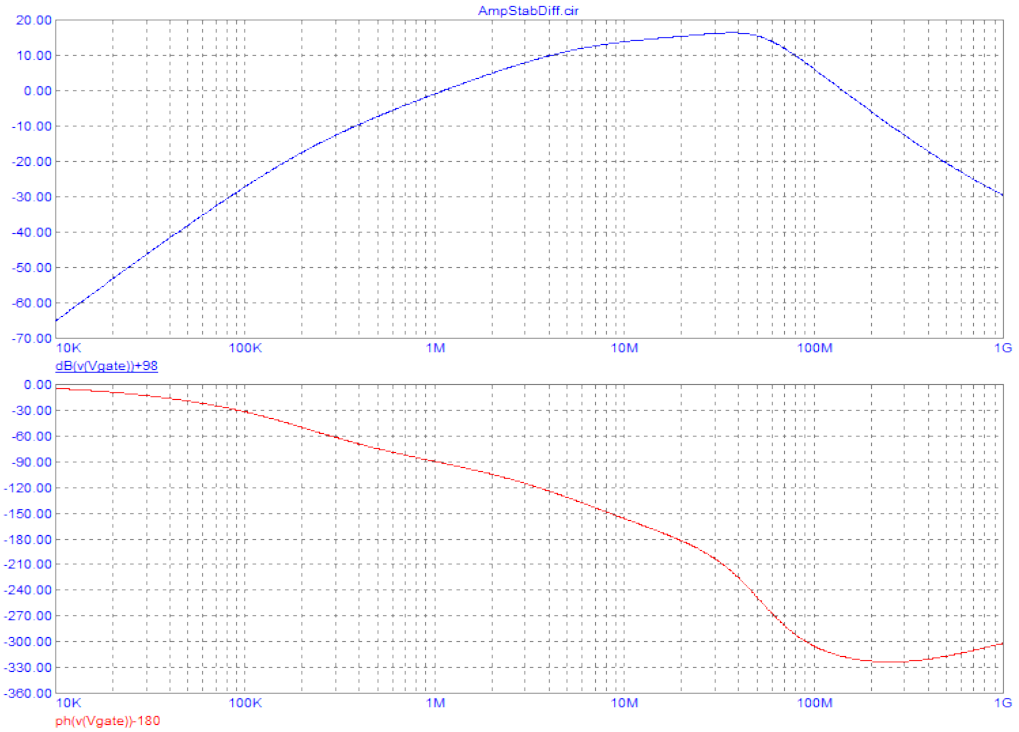
As before, V1 is configured as a current source, injecting the calculated drain current for 1mV between gates. L1 is the mutual inductance of the output transformer, seen on its entire primary. Since each side of the primary has 2μH, both in series and coupled have 8μH. L2 is the leakage inductance of the output transformer. R1 is the output load, transformed down, as it appears between drains. C1, C3 and C4 are the transistor capacitances, but expressed in pairs of two in series. Likewise C2 and R2 is the shunt feedback, both in series. R3 is the two bias resistors in series, R6 is the two gate series resistors in series, L4 is the drive trafo leakage inductance, L3 its mutual inductance, and R5 is the internal resistance of the driving source.

Under these optimal conditions, I get this Bode plot:



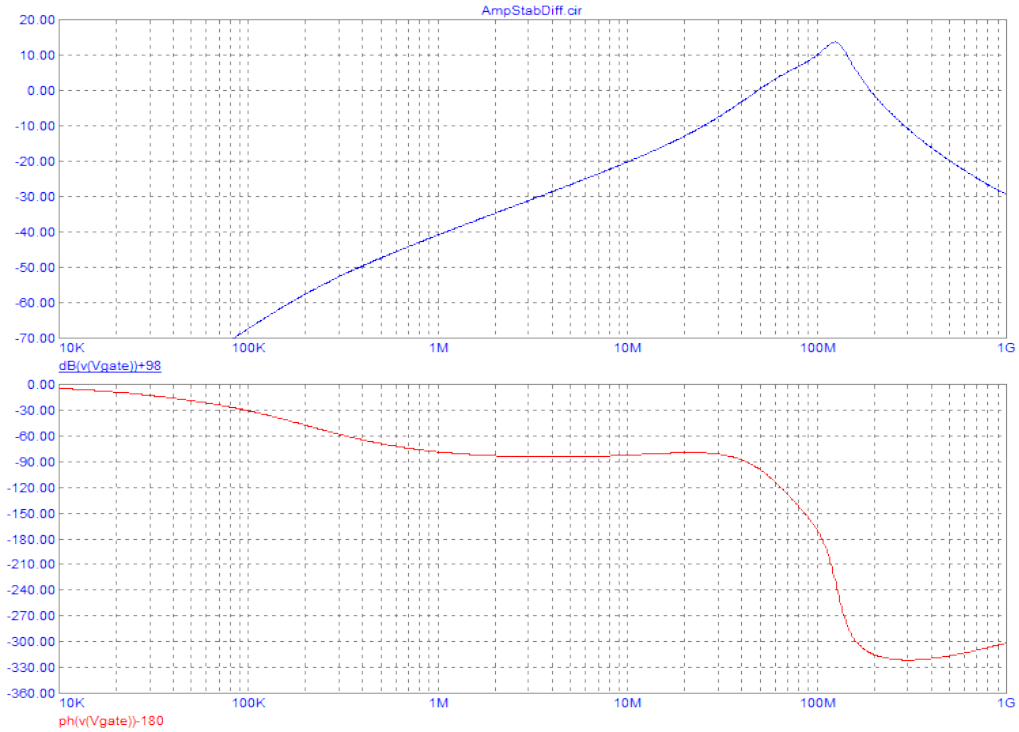
The loop gain is very low. That's caused by the low MOSFET transconductance at the bias current level, the low load resistance, and a considerable dividing factor in the feedback network. The whole low frequency side and the operating frequency range have no risk of oscillating, thanks to the low loop gain. At the high side, the phase margin is about 50°, good enough. Above 100MHz there is a peak in the loop gain, caused by resonance between leakage inductances and transistor capacitances. The fast phase shift in that frequency range is caused by this too, of course.

Let's disconnect the load:



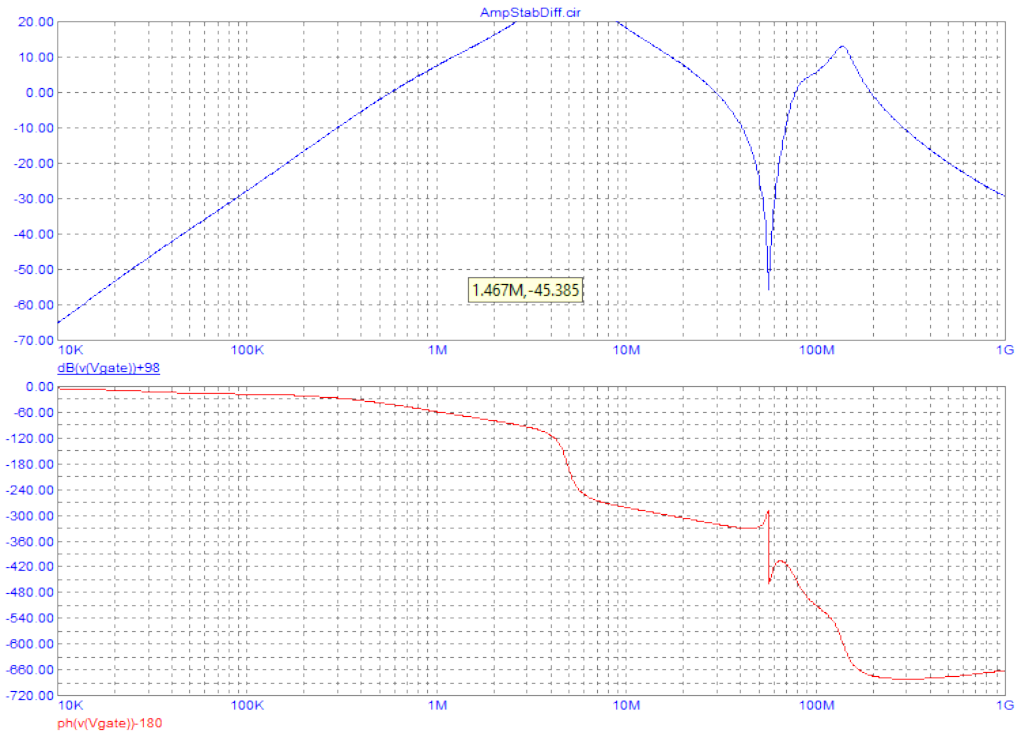
Now there is much less attenuation in the drain circuit, so the loop gain is higher to a lower frequency. Still the phase margin at the low gain crossover is excellent, at 90°, while the one at the high frequency gain crossover is only 40°, a bit small for comfort. If the transistors add enough delay, we might see oscillation near 100MHz if the load is disconnected.

With a shorted output, we get this:



Again no problem at low frequencies, and a somewhat meager phase margin a little below 200MHz, but it should be OK.

Removing the drive source is fine, too. But then I tried to connect devilishly chosen capacitive loads of 100pF to the output, and 26pF to the input. See what happens:



Note that the phase scale had to be changed, to span two full revolutions! There is a gain crossover at 30MHz with a phase margin of only 30°, too small for comfort, and some pretty wild phase behavior that could lead to further trouble when including the MOSFET delay. This behavior comes from those capacitive terminations resonating with the transformers. It probably wouldn't be impossible to make this amplifier oscillate, by simply connecting short pieces of unterminated coax cable to the input and the output, which can happen rather easily in practice. So, if you build such an amplifier, keep it decently loaded!

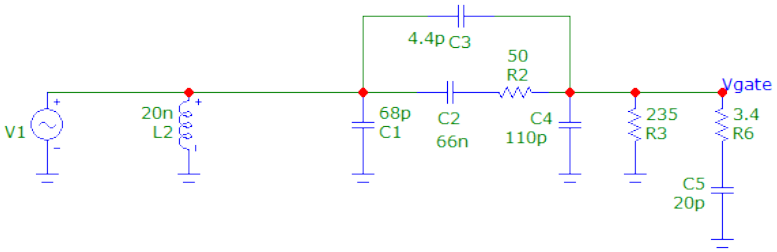
At least the -360° phase crossing happens at a frequency where the loop gain is way down below 0dB.

When the amplifier is being driven at **full amplitude**, we have to look what happens at the instants when the drain current is at about two thirds of the maximum value, which is where the MOSFETs transconductance is largest. While at constant drain voltage the transconductance reaches a peak at about 4A, it is lower when at high current the drain voltage gets very low. That's why amplifiers typically are most unstable near that two-thirds amplitude point of the RF waveform. With a fast oscilloscope one can sometimes observe little wiggles starting roughly there, which are VHF oscillations mounted atop that part of the main HF waveform.

At that time only one of the MOSFETs is conducting. Since I'm simulating just the resistive and reactive elements of the circuit, not the MOSFET transfer curve itself, the circuit used for simulation doesn't change. Just the amplitude of my current source increases, because of the higher transconductance. It is very close to 1.4S, so for 1mV between gates, we have just 0.5mV on the gate of the active transistor, and thus it's sinking 0.7mA on its side of the output transformer. But if we keep the simulation schematic unchanged, we have to halve that value, because the center-tapped primary acts as an autotransformer/balun to put the voltage caused by that current on the other side too, while the transformer acts as a 1:16 impedance transformer, so that 3.1Ω load resistance appear across the transistor that is on. We could change the diagram to reflect that, and then have the current source inject 0.7mA, but the result will be the same. So let's keep it simple.

In this case, it turns out that the resulting 0.35mA of effective current injection into our model is even lower than the 0.4mA that happened when I analized the amplifier at zero signal, with 0.5A idling current in each transistor. This is rather unusual in practical amplifiers, and happened because I chose 0.5A idling current per side simply to start the current excursion where the transfer curve begins its nearly straight part - but this is really not the correct thing to do! For lowest distortion the idling current should be selected so that the transconductance of each transistor at the bias point is one half that at high current, and for this MOSFET that would be at roughly 350mA or so per side. In that case the stability analysis would give the same result at idle and at high drive. In reality many people set the idling current of a class AB amplifier even lower, looking for a compromise between linearity and efficiency, and in that case the total transconductance of both transistors at idling is lower than that of a single one at high drive, and then what I wrote above is true: The amplifier has more risk to oscillate at strong drive than when idling! Phew! My face has been saved...

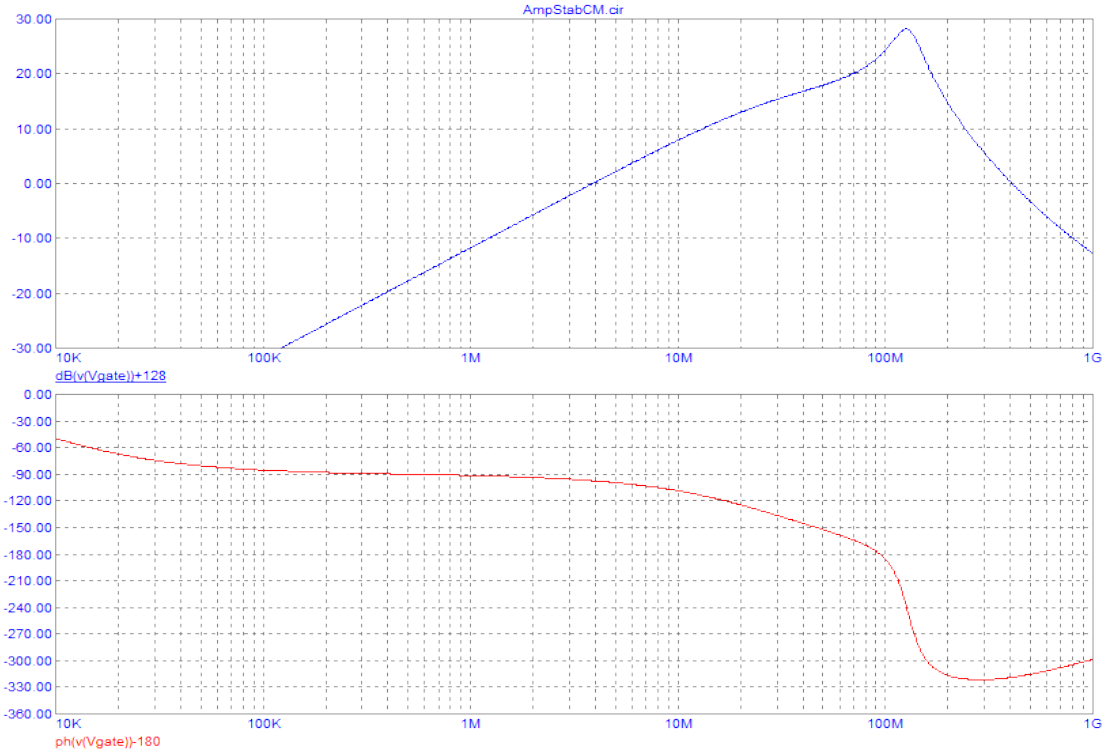
Let's now look at the common-mode stability. The equivalent circuit for feedback analisis changes a lot! Because the transistors, working in equal phase, don't see the load, which is applied only differentially through the transformer. Also they don't see the mutual inductance of the output transformer, because it's in opposite phase between sides, and thus cancels. But they still see the leakage inductance between primary halves, 20nH. The two transistors act in parallel, so the transconductance doubles, all the reactances and resistances are halved, and so on. The input transformer simply disappears, since it's balanced too, except that its interwinding capacitance might possibly become important, so I will include it in the equivalent schematic. So we end up with this:



Always keep in mind that such equivalent circuits are only coarse approximations to the reality, which contains many unknowns. The intention is to try to include all important bits and pieces, and very often this attempt is unsuccessful...

I assumed 20pF interwinding capacitance for the input transformer. Whether that comes close to the truth, depends on how it's made.

The Bode plot looks like this:

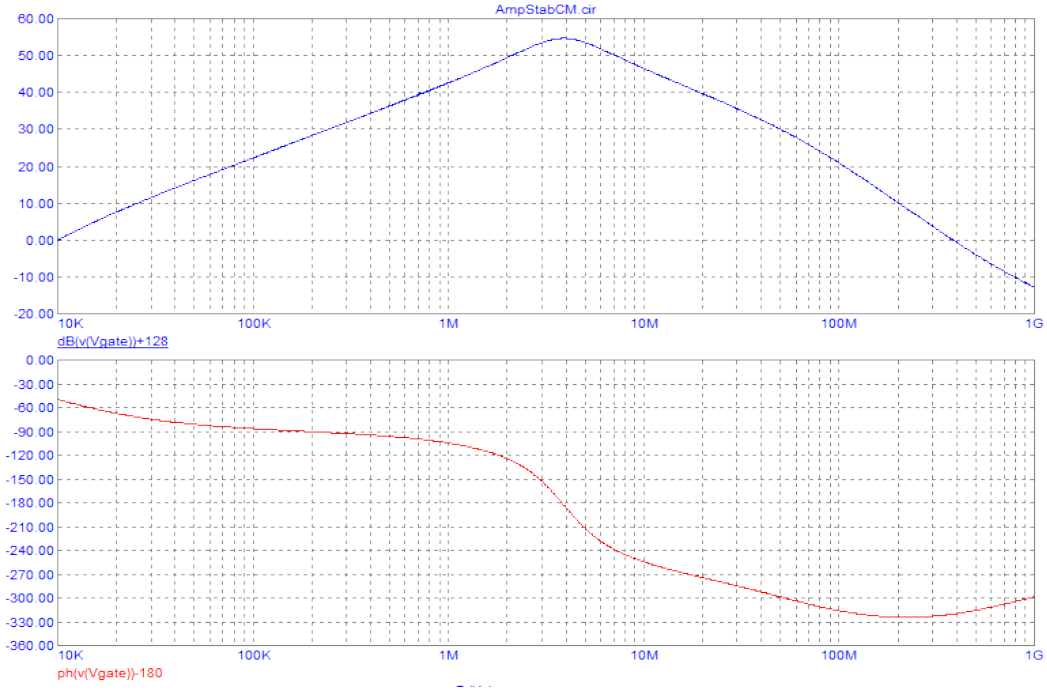


The resonance of the output transformer's primary leakage inductance with the transistors' output capacitance is obvious. The high side gain transition occurs at 40° phase margin, but at a slightly lower frequency there is only 35° phase margin, a bit tight, but it should do. In the low frequency range the feedback has a phase close to -90° instead of the desirable -180°, due to the fierce effect of the output leakage inductance almost shorting the drain current, but still the phase margin is excellent at the lower gain crossover frequency.

Disconnecting or changing the load or drive won't have any effect on common-mode loop gain, because the common-mode configuration just doesn't see those.

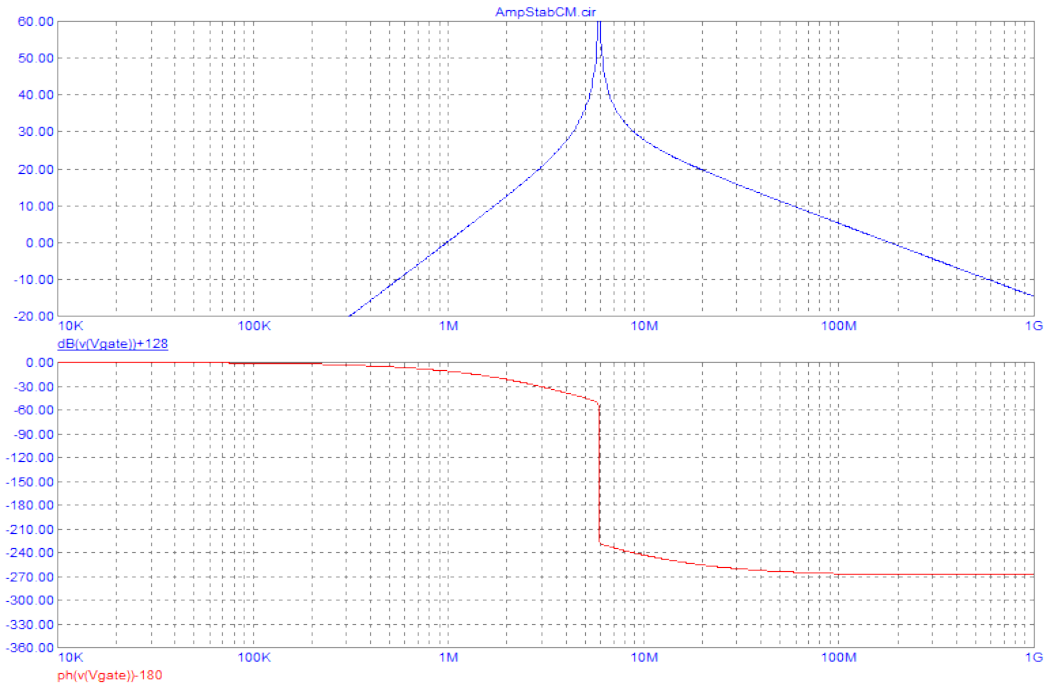
So it's all reasonably good, and this amplifer has a fair chance of being sufficiently stable in all ways, for practical use!

Next part of the "what if" game: What happens to the common-mode stability if we change the feed configuration to two separate RF chokes? The equivalent simulation circuit, is then the same, except that we get the full inductance of those chokes causing a rather high common-mode drain signal voltage, from the drain current injected there. The result is that the gain curve shifts up a lot, but stability isn't compromised, because there is still about the same phase margin at the gain zero crossings, due to the flatness of the phase response:



The point to watch is still the VHF range, where the transistors might cause trouble due to their signal delay.

But now let's do something evil. Some hams trying to build high power amplifiers have done it, and have suffered the consequences! That evil thing is removing the shunt feedback from such an amplifier that uses separate RF chokes. This is what happens:



The phase margin at the lower gain crossover is only a few degrees! Almost certainly there will be enough strays in the circuit to move the phase just a little further up, and we will end up with a 1MHz power oscillator. And even in case it's so well built that there aren't enough strays to give it that little phase boost, it will badly ring when operating at 1.8MHz, with a phase margin of only 20° and a loop gain of 10dB!

Such oscillation looks like both drains going down and up together, and since the drain current can't go anywhere else than into the relatively high inductance RF chokes, enormous drain voltage spikes show up in each cycle of the oscillation, making the transistors go full and heavy into avalanche conduction. Heavy enough to kill them in an instant! There are hams who have killed a dozen or more high power LDMOSFETs, costing over $200 each, due to this mistake! In short: Common-mode oscillation due to lack of both common-mode feedback and common-mode load.

Amplifiers that have coupled drains, through a well-built bifiliar feed transformer, or a center-tapped output transformer whose center tap actually works, are instead tolerant to the lack of common-mode feedback. That's because only the leakage inductance of the output transformer or feed transformer appears on the drains in common mode, drastically reducing the common-mode loop gain, and also preventing the appearance of those high inductive drain voltage spikes. Of course at least differential-mode feedback should still be used with them, to keep the gain, distortion, and frequency response flatness under control, and the leakage inductance between drains should be kept really, very, extremely low, making that point the priority #1 of the whole design!

There is also the nasty phenomenom of **internal MOSFET instability**. Some MOSFETs will self-oscillate just by themselves, when they get the right conditions. A MOSFET, after all, is a complex thing. In addition to the MOSFET proper, it contains a parasitic diode, a parasitic BJT, several resistances (which depend on temperature), several capacitances (which are voltage-variable), several inductances, and in addition an RF power MOSFET is composed of hundreds or thousands of tiny separate little MOSFET cells! Be very careful when using a MOSFET outside of the operating parameters recommended by the manufacturer. Crazily what tends to make these MOSFETs oscillate is powering them from a lower voltage than the nominal one, or driving them at a frequency below the lowest frequency recommended by the manufacturer! For example, you might find a VHF/UHF power MOSFET rated to operate from 1 to 500MHz, and another one rated for 100 to 500MHz. If you use the latter in an HF amplifier, it will certainly amplify - but it might oscillate internally. And there might be no way to stop this oscillation from the outside!

The best thing to do is to only use MOSFETs rated for a frequency range that completely includes the intended one, and then design a circuit with adequate phase margin. And when building amplifiers that use a modulated supply voltage, make sure that you arent' getting this kind of oscillation at the times of low supply voltage. Typically such oscillation is in the UHF range, and appears only during a part of the main RF cycle. You need a good, fast oscilloscope and test probe to see it. Sometimes such instability can be cured by using more resistance in series with the gates, but in other cases nothing works to fix it, and you have to choose a different MOSFET.

Here are some **suggestions for building stable amplifiers**:

- Design the amplifier for just the gain you need. The more gain there is, the more likely it will be unstable. The best way to reduce excessive gain is through negative feedback, rather than using attenuators.

- Keep the frequency response of the amplifier limited to just the range you need. Coverage of any additional frequency ranges is pointless, and increases the chances for instability.

- In a multistage amplifier, let one stage, preferably the middle one, do the strict bandwidth limitation, and keep the response of the other stages wider. This avoids the situation where several stages contribute phase shift on the same frequency, causing the overall phase response to rotate through the oscillation boundary before the gain is down enough. This technique is called dominant-pole compensation in some other areas of electronics.

- Resistors are my friends. Print this short sentence, frame it, and hang it on the wall beside your workbench. If you want, add Never trust inductors nor capacitors. The reason for these two rules is that resistors don't cause phase shift, while inductors and capacitors do. And phase shift makes the difference between negative and positive feedback, and thus between a stable amplifier, and, well, an oscillator. Every capacitor and every inductor can cause up to 90° of phase shift. Keep their number small. Swamp them with resistors, both in parallel and in series, wherever necessary.
Even a small amount of resistive damping will transform a 90° phase shift into a more moderate one, and very often this makes the difference between a stable circuit and one that isn't. If you can inject the bias through a resistor rather than an RF choke, use the resistor. Make sure that MOSFET gates always connect only to resistive paths. Specially avoid connecting any inductive element directly to a MOSFET gate! A capacitor in series with the inductance doesn't count. It has to be a resistor! And don't forget that chokes and transformers are also inductors, nor that transformers have invisible leakage inductors built in!

- Always keep in mind that all bypass capacitors, coupling capacitors, RF chokes, transformers, even if they have a reactance high or low enough to be negligible over the entire operating frequency range, will absolutely have significant, phase-shift-causing reactances above or below that range. Analyze their behavior essentially from DC to daylight, not just over the intended operating frequency range.

- When physically building an amplifier, keep the strays small. That means, make low impedance interconnections extremely short and wide, to reduce stray inductance. Make high impedance connections very thin, and keep them clear of other stuff, to reduce stray capacitance. Analyze which connections are most critical in this regard, and design the layout to absolutely optimize them. A typical connection to keep extremely short and wide is the emitter or source connection of RF power transistors that operate at high current. That's why such RF power transistors have two to six emitter or source tabs, or use the mounting flange as emitter or source connection.

- Build amplifiers on a continuous ground plane. When using transistors that use the mounting surface as source or emitter connection, solder them to this ground plane. If you have to solder them to a heat spreader, solder the underside of the board (used as ground plane) to the heat spreader too, all around the transistor, to make a continuous ground plane.

- Don't ever do the mistake that even some supposedly professional designers love to do: Using several capacitors of different values in parallel, for wide-range bypassing! It's nonsense, and it doesn't work, unless correctly done with a full understanding. It's nonsense because even pretty high value capacitors are physically small enough to have low impedance up into VHF and beyond, and it doesn't work because when connecting two different capacitors in parallel, the larger one starts acting as an inductor at a frequency where the smaller one is still capacitive, and you get a parallel resonant circuit that causes extremely poor bypassing at the resonant frequency. The only case when such capacitor paralleling works is when at least one of them is very lossy, causing enough damping of the resonance they form. This could be the case when paralleling a poor quality aluminium electrolytic capacitor, that has a high ESR, with a small SMD ceramic chip capacitor.

Instead paralleling identical bypass capacitors, to provide a wider bypass path, or for current sharing, is fine. But in most cases it's best to use a single capacitor that is electrically large enough, and physically small enough, to provide good bypassing over the whole frequency range needed. Ceramic multilayer chip capacitors are pretty good for this.

- When building several stages, arrange them linearly, rather than in a circle or a compact blob. The output should be far away from the input. This applies to connections and all RF parts, but even more so to the magnetics.

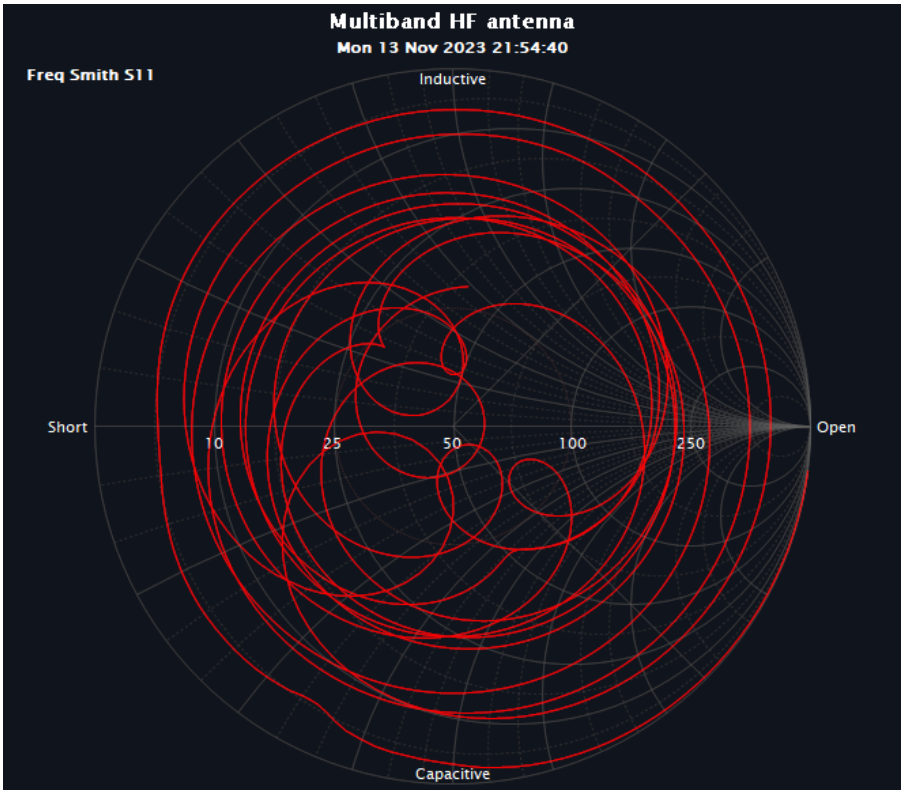
- Use proper shielding, at the proper places.

- When using the same power supply for several stages, be very careful with the decoupling. It should be effective over a wider frequency range, both down and up, than the range over which the amplifier has any gain.

- Watch for saturation of any magnetic cores used in power supply chokes, and in other DC-carrying inductors and transformers. Many people make mistakes in this area. The result is adequate decoupling when there is no signal, degrading into insufficient decoupling when the signal is strong and the DC is high, resulting in oscillation that starts only above a certain drive level. Very often such oscillation keeps going after the drive is removed, because the oscillation makes the amplifier consume enough current to keep the culprit in saturation.

- When building the final stage of a transmitter, it's critically important to understand and duly consider the effect of load impedance on the phase and amplitude of the

feedback. That's because the impedance of an antenna, at the end of a feedline, is totally unknown, usually quite complex, and totally different between one antenna and another. You could get literally any load impedance on any frequency. So you need to analyze your amplifier considering a drain load impedance that varies over the whole range from totally capacitive to totally inductive, and from zero to infinite magnitude. Now don't tell me that you are very careful with your antennas, and they all have close to 1:1 SWR! That's only the case on the few narrow bands for which they are built - but the amplifier sees them pretty much from VLF to somewhat above the cutoff frequency of your lowpass filter! And the impedance in that range can be crazy. Here is a Smith chart of my HF antenna's impedance, from 50kHz to 30MHz. That's 4 dipoles, well separated, two of them having small capacitive loads, and fed through a roughly 30m long 50Ω coax cable, giving low SWR on 80, 40, 20, 17, 15, 12, and 10 meters:



You see? In the bands of interest, the impedance loops close to the 50Ω center point of the graph, but outside those bands it goes in many circles between open, almost shorted, very capacitive and very inductive. That's what the final stage of a transmitter must handle, without becoming unstable! The lowpass filter will change and complicate this impedance further, the output transformer will modify it too, and so on.

I have to make a **disclaimer**: I didn't double-check all my calculations and simulations about stability. So there might be errors in them, even big errors. If you find any, let me know. My intention was to show you a method of doing stability analysis, rather than to give you practical, proven circuits with accurate calculations. And there are other, better methods to do it, such as using the best simulation software you can find, with exceptionally accurate non-linear transistor models that work even with large signals and at RF, and then simulate the entire circuit, including the transistors and all the many strays.

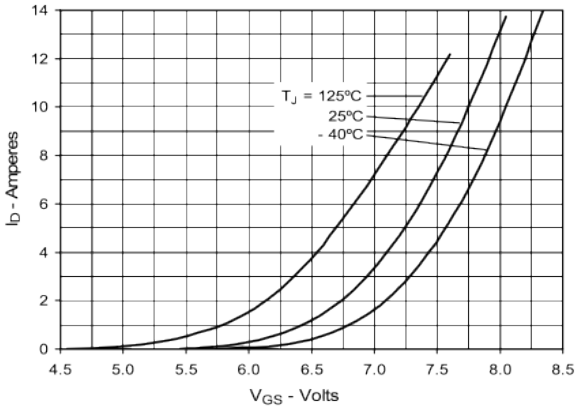
It's also possible to build a circuit and test it, but with expensive RF power transistors that's often a quite costly way, when you burn out a dozen of them before you discover why they oscillate! And while an amplifier that is definitely unstable will allow you to notice, in one way or another, an amplifier that is just on the brink of getting unstable might seem to perform well, only to start oscillating when conditons change a little. Load impedance, temperature, supply voltage, bias setting, drive level, whatever.

If you weren't already totally seasick from the many Bode plots, I assume that the plot of my antenna's impedance definitely gave you the rest! So let's end the chapter about stability. There is more to do.

Linearity

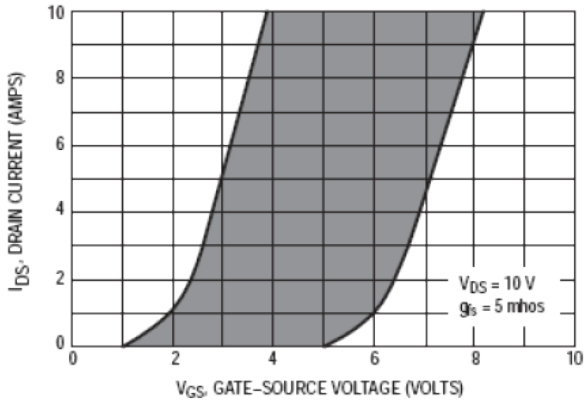
A sad thing in the life of anyone building linear amplifiers is that transistors are fundamentally nonlinear devices. They introduce several kinds of distortion, most of which can be broadly grouped into amplitude distortion and phase distortion. Let's start with amplitude distortion.

The transfer curve of a MOSFET theoretically follows a square law. This can be seen very clearly in the data sheets of transistors that haven't been optimized for linearity, but for efficient switching. This is a typical example:



The parabolic shape of the curve is very obvious. But note that the high temperature curve tends to straighten at high currents, more than the curves for lower temperatures. This is because the internal resistances of a MOSFET increase when hot, and that causes the always- present source degeneration to become stronger. Very generally, MOSFET transfer curves start with a parabolic shape at low current, which bends into a relatively straight curve at high currents, and does this more so at higher temperatures.

A manufacturer wishing to make a more linear MOSFET can do so by deliberately increasing the internal source resistance. In this case the parabolic section is much smaller, confined to the low current zone, while from there up the MOSFET's transfer is relatively linear. Such as in this example:



This graph shows a range, and the actual curve of the MOSFET will fall somewhere inside, depending on temperature and tolerances. But you can clearly see the response being exponential in the low current zone, and pretty linear at high currents.

To get reasonably linear operation from a MOSFET like this in class AB, you should bias it to a current that gives about half the transconductance (curve steepness) as in the linear zone. In this example, this would be about 1.2A. Due to their large nonlinear zone, MOSFETs require relatively large idling current in class AB, if decent linearity is desired.

Biasing a transistor to the optimal point will still not result in perfect linearity in the low current zone, but it's the best we can do. If we want to build an amplifier that does not have this sort of small-signal distortion, we need to build a class A amplifier, biasing the transistor well into the linear range of the curve. That is, if we need a linear drain current swing of ±3A, we need to bias this transistor to at least 5A idling current! This would make a class A amplifier that has pretty good amplitude linearity, at the cost of dismally poor efficiency.

At very high current the transfer curve of a MOSFET the transconductance will again get smaller, due to saturation effects. But in linear amplifiers we hardly ever get into that range, because usually the maximum power dissipation of the device limits us to lower currents.

A MOSFETs drain current also depends to some extent on drain voltage. This can be seen very well in a graph that shows drain current versus drain voltage, for several gate voltages:

Fig. 1. Output Characteristics @ $T_J = 25^oC$

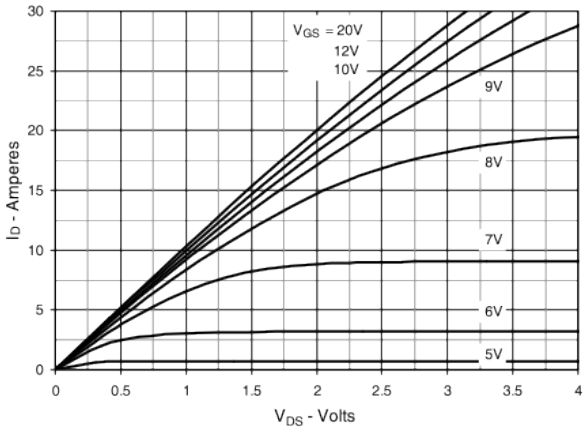
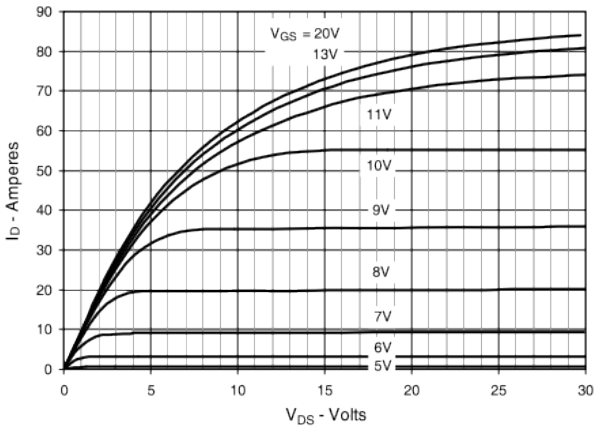


Fig. 2. Extended Output Characteristics @ $T_J = 25^oC$



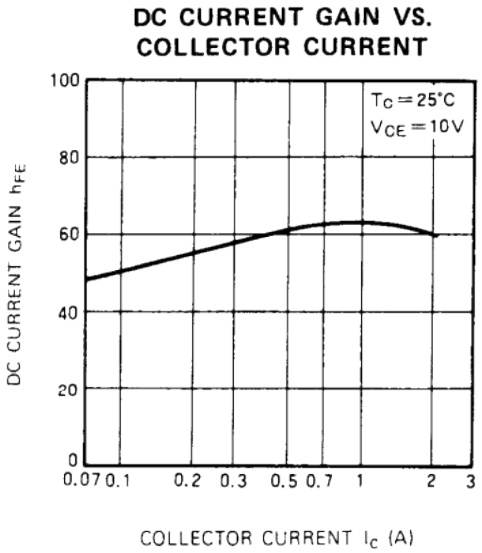
Here are two graphs showing the same thing, but the one at left is a detail of the low drain voltage range, while the one at right shows the curves for the same MOSFET over a higher drain voltage range. In the right-side graph we can see that in the middle of the current range, and at suffciently high drain voltages, the current depends only very slightly on drain voltage. That is, the MOSFET is a very good current source in that range, controlled by gate voltage. It has a very high internal impedance, as all good current sources do. Instead at lower drain voltages the curves bend. Operating a MOSFET in that zone causes distortion. For example, if we used this MOSFET for a power amplifier, operating at a peak drain current of 50A, we can see by interpolating between curves that the transfer would start showing saturation when the the drain voltage is down to about 11V. But this same MOSFET can conduct the 50A with only 6.5V on the drain, if we apply enough gate voltage! The gate voltage needs to be much higher, though. So the ultimate waveform clipping of this amplifier would happen at 6.5V on the drain, but driving it to go below 11V would cause increasing distortion. In other words, we shouldn't drive it too close to saturation, if linear operation is desired.

Now this data is for a MOSFET that has a 250V maximum drain voltage rating. Such a device would be used in an RF amplifier with a supply voltage of about 80V. Avoiding distortion from flat-topping (good old term!) requires us to sacrifice 11V of those 80V, limiting the drain voltage swing to ±69V around the 80V supply. This limits both the available clean output power, and the efficiency. If we didn't care about linearity, we could run this MOSFET at ±73.5V, getting a distorted sine wave of higher power and at higher efficiency, or we could overdrive it like crazy to get a ±73.5V near-square-wave, getting even more power and efficiency. This is fine with FM transmitters, and with supply-modulated linear transmitters for AM or SSB, but not for linear amplifiers.

In a graph showing such a family of curves, the basic linearity from gate voltage to drain current can be seen by the separation between the flat horizontal regions of each curve: They should be equally spaced, for a reasonably linearized MOSFET. In curve pair above is for a switching transistor with their typical square-law curve, which can be seen by the spacing between the lower curves roughly doubling from one interval to the next. Above about 60A this MOSFET enters the pretty linear zone.

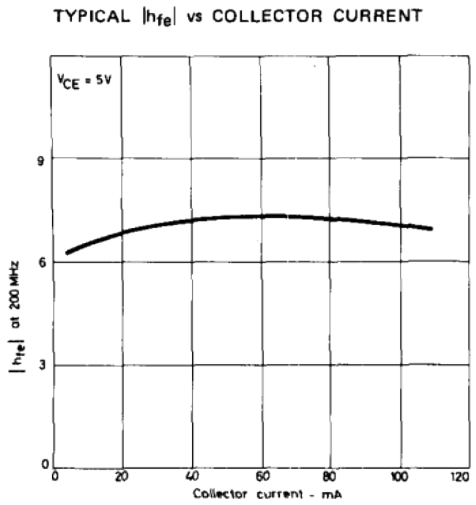
BJTs have a strongly exponential transfer curve, if you express it the same way as for MOSFETs: Base voltage versus collector current. But since the base voltage required to drive the collector over its whole useful range is very small, it's typically more convenient to consider the BJT to be a current-controlled current source. So we need to talk about its current gain, hfe.

The hfe for DC, valid up to a certain pole frequency, is usually given in the datasheets, but has an extremely wide tolerance. The hfe at higher RF instead is given by the transistor's transition frequency, Ft, also given in the datasheets, divided by the operating frequency. Both values of hfe change with collector current, but not nearly as much as the transconductance of a MOSFET. Here is an example of the hfe at DC, for a typical BJT intended for VHF amplifiers of a few watt output power:



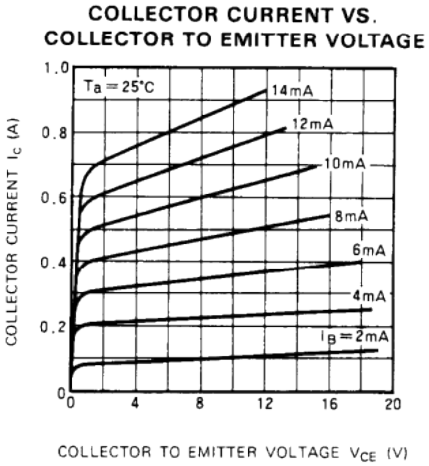
Only about 25% variation of the hfe, from a very small current right up to the transistor's maximum rated current! This allows biasing BJT in class AB to a very low idling current, and still get good linearity.

The RF hfe varies even less, seen here for an even smaller VHF BJT:



Even while the gain variation is much smaller than in a MOSFET, the same basic pattern can be observed: Highest in the middle of the current range, lower at both ends. But note how low the current gain is. This transistor has a rated Ft of 1400MHz, so at 200MHz its current gain averages about 7.

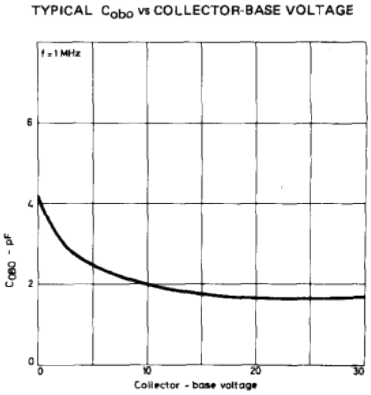
The family of curves that shows collector current versus collector voltage, for different base currents, typically looks like this:



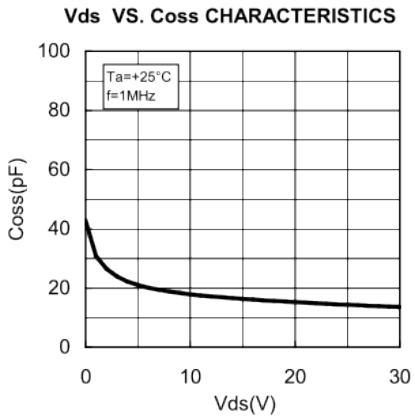
Comparing this graph to the MOSFET's, you can see that the curves are quite evenly spaced, meaning good basic input-output linearity, and also that the saturation level is much lower than for MOSFETs of the same voltage rating, while also the knee, where the curves stop being straight (linear), is lower. This allows driving a BJT's collector closer to ground than a MOSFET's drain, allowing better amplifier efficiency. This has to be taken with a grain of salt, though, because the actual values depend on the exact transistor chosen, and there are now mighty good MOSFETs available. Just when comparing transistors of the same technology level, typically BJTs are both more linear and more efficient, while MOSFETs have more power gain and much flatter frequency response, giving them a progressive advantage towards higher frequencies.

The curves for this BJT are not as horizontal as they are for the MOSFET example I chose. This means that this BJT has a lower internal output resistance, being a less good current source than the MOSFET example. Since in RF power amplifiers we always have some shunt feedback, intentional or not, that lowers the output impedance anyway, this difference is irrelevant for RF amplifier use.

Phase distortion is caused by transistors through several phenomena. One is that the internal capacitances vary strongly with the instantaneous drain or collector voltage. Varying capacitances in an RF circuit will cause phase shifts. Here is an example of output capacitance for a typical small BJT:



And here is an example for a typical small MOSFET, intended for roughly the same frequency range, output power and supply voltage as the BJT:



As you can see, the variation range is about the same for both, roughly 3:1, but the MOSFET's capacitance is about 10 times larger than the BJTs! For this reason, MOSFETs generate much stronger phase distortion than BJTs, and with MOSFETs it's much more important to avoid driving them into onsetting saturation, when good linearity is desired. The reason why MOSFETs have so much more capacitance is that they work at a lower current density, so a larger silicon chip is required to make a MOSFET for a given current. The good side of this is that the MOSFET will then have a higher maximum dissipation rating.

Another phenomenom that causes phase distortion is that the delay a transistor adds to the signal tends to change with current, because the Ft changes.

Memory distortion is of a more special kind. The term refers to all those distortions that depend on previous states of the amplifier. For example, thermal effects depend on the instantaneous temperature of the silicon chip, which is higher when the the amplitude of the signal is 70% *after* a modulation peak, then when it is 70% *before* that modulation peak. At both times the driving signal is identical, but the transistor is still operating at a different condition, and might amplify a little less or a little more. This distorts the signal.

Another way to produce memory distortion is to do what many designers love to do: Plant an electrolytic capacitor across the supply line at the amplifier. Typically the amplifier will be powered by a very well regulated power supply, through a long wire. That wire causes voltage drop. While the drive signal is getting stronger approaching a modulation peak, the supply current increases, the voltage drop increases too, but the capacitor will deliver some current, reducing that drop. When the signal gets weaker after that modulation peak, the current drawn is the same, but the capacitor is at a lower charge state, is starting to recharge, and its voltage is lower than it was before the modulation peak. So the transistors have slightly larger capacitances, a little bit less gain, and we get distortion. It would be better to either eliminate that electrolytic capacitor, or use an extremely large one, or use very short and fat supply wires, or use a power supply that supports remote sensing, and wire it up to regulate the voltage at the amplifier, rather than at the supply.

Memory distortion effects tend to be rather small, anyway. They are usually masked by basic amplitude and phase distortion. But when these distortions are corrected for, memory distortion becomes measurable.

While the transistors are the main causes of distortion, some other components can contribute some. Watch out for transformers and any chokes or inductors that have magnetic cores and carry DC, because it's rather easy to make a design mistake that results in saturation of the core from DC. RF current will almost never be strong enough to cause saturation, because the core will get extremely hot from losses at an RF flux density very much below the saturation level. But when a core gets somewhat into saturation due to DC, which happens easily, the small RF current amplitude riding on top of this DC varies the saturation degree of the core at the RF rate, causing severe distortion. This is in addition to the other troubles caused by saturation, such as a drastic drop in inductance.

Many kinds of capacitors vary their value according to voltage. This isn't a big problem with bypass and coupling capacitors, because they are usually chosen with adequate capacitances to cause very little RF voltage drop across them, but any capacitors used in filters, resonant circuits, for compensation or such must be of a type that is stable in the presence of variable voltage. These are C0G ceramic capacitors, called NP0 in olden times, porcelain capacitors, mica capacitors, PTFE (teflon), the now-little-used polystyrene capacitors, and of course air dielectric and vacuum capacitors. Some more probably exist. Just make sure that you don't use any class 2 or worse ceramic capacitors in such places of the circuit, such as X5R, X7R, Z5U, Y5V, X7S.

Also be careful with any diodes used in the RF area of the amplifier, because these too have a voltage-variable capacitance when reverse-biased, and a current-variable resistance when forward-biased.

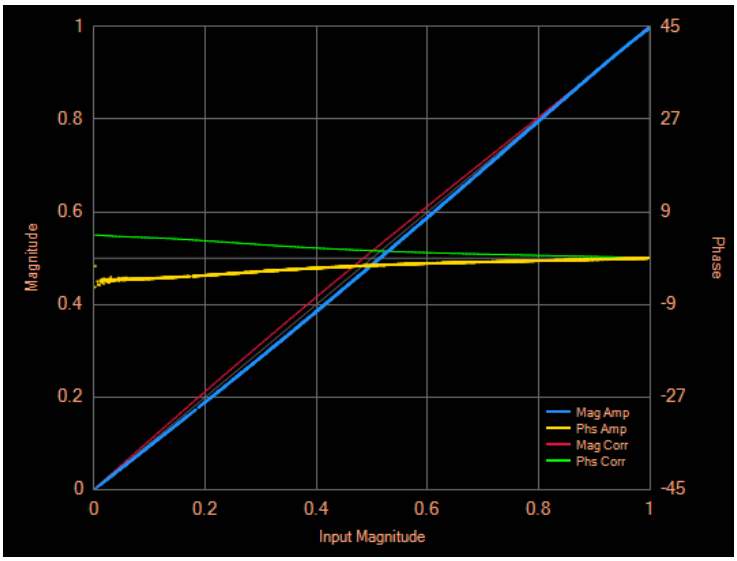
Resistors are our friends... Most of them behave pretty well. Just stay away from carbon composition resistors. Modern film resistors are much better. Old textbooks often recommend carbon composition resistors for RF work, but that was simply because film resistors were not yet available, and the only alternative would have been terribly inductive wirewound resistors!

Negative feedback reduces all distortions. But it's no magic bullet. It can only reduce them a little, according to how much gain we sacrifice for that purpose. A a typical RF power amplifier will have a few dB less distortion with negative feedback than without. People coming from the audio world, using operational amplifiers having an open loop gain of 120dB, to make an audio amplifier having 20dB gain, and using up the excess 100dB gain in negative feedback to correct distortion, can achieve a linearity that we RF guys can only dream about. But even if what we can do with negative feedback might not seem like much, it's definitely an improvement, so we should always make good use of negative feedback.

Yes, **it is possible to correct distortion** in a more complete way in RF amplifier systems, and achieve an extreme degree of linearity. But we can't do the trick of the audio guys, and use negative feedback spanning several stages in a single loop. Given the phase delays in each of our stages, such an approach is guaranteed to produce an oscillator instead of an amplifier. What we need to do instead is more complex: Take a sample of the amplifier's output signal, continuously compare it in amplitude and phase to the drive signal, detect and "learn" the distortion curve of each, updating it continuously, and then pre-distort the driving signal in amplitude and phase, to make the nonlinear amplifier produce a clean output!

This can be done with analog circuitry, but it's complex and touchy. It's better done with digital signal processing, and lends itself very well for inclusion in software-defined

radios. I use PowerSDR software to play with this. Here is a graph of distortions and corrections of a well-designed two-stage amplifier, using a single BJT driving a Gemini LDMOSFET in push-pull:

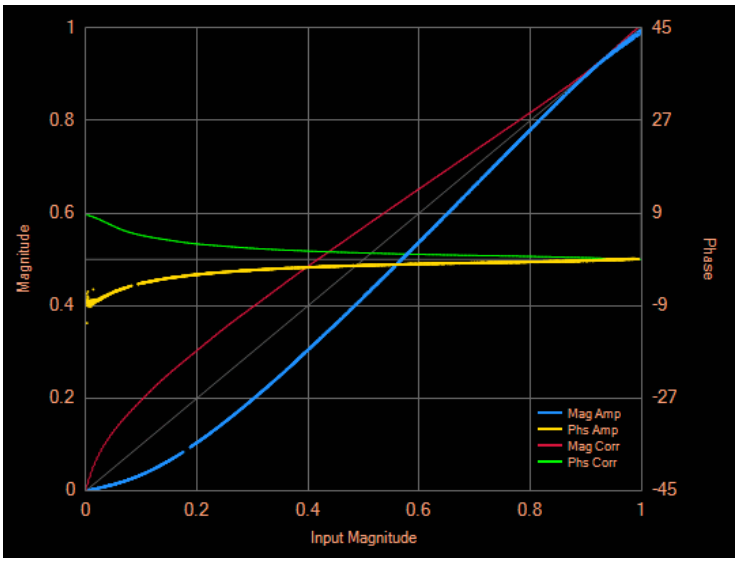


In blue is the amplifier's amplitude transfer curve, with the amplifer optimally biased into class AB, and driven just to the onset of compression. You can see that the beginning of the curve is very straight (linear crossover region), then bends a little up to become steeper (higher gain in the mid-current area), then again bends very slightly down (the very beginning of saturation).

In yellow is the phase response. It's normalized to the phase at full power. You can see that between very small signals and full power, there is about 5° of phase difference. This measurement was made at 7.1MHz, a pretty low frequency. The phase distortion gets larger at higher frequencies, and smaller at lower ones.

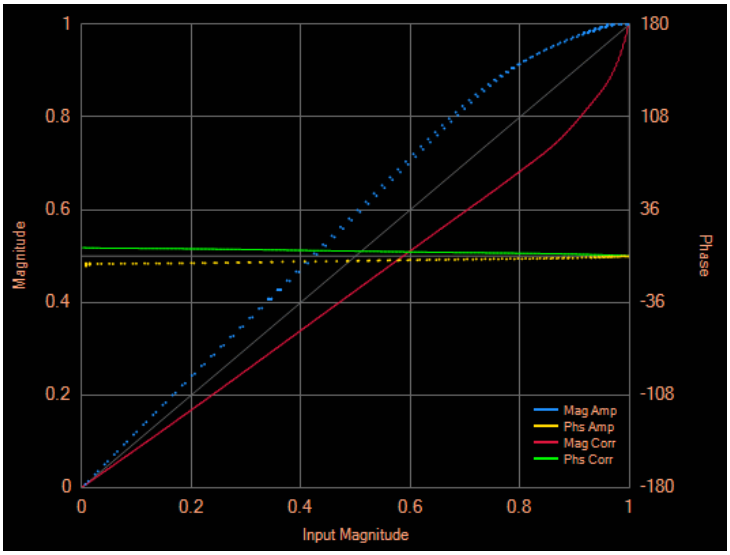
The red and green curves are simply the correction functions needed for predistorting the drive signal, to clean up these little residual distortions of my amplifier.

The next curve shows the same amplifier, operated with much too low bias, giving just 10mA of idling current:



You can see the obvious, severe cross-over distortion. Basically the amplitude response of the amplifier follows the MOSFET's gate-drain transfer curve, slightly straightened by negative feedback. The phase distortion is also higher, probably because at very low current the MOSFET gets slower, causing more phase lag. That's why now the phase at low power is 9° behind the phase at full power.

And the third graph shows the same amplifier, correctly biased, but severely overdriven:



Never mind the dotted curves. I only made a brief test with a stable two-tone signal, collecting few data points. What you see is a pretty linear amplitude response up to about 70% of full drive, then the amplifier goes progressively into saturation. At full overdrive it's almost fully saturated, the output power barely increases anymore with additional drive. There is also more phase distortion (note the change of phase scale done by the software), with the phase distortion making a slight bend in the saturation zone. That's the internal MOSFET capacitances shooting through the roof as the negative peak drain voltage gets very close to ground.

The beauty of adaptive predistortion is that I can run the amplifier with moderately low bias, and overdrive it a little, getting higher output power and much better efficiency, while at the same time producing an exceptionally clean signal.

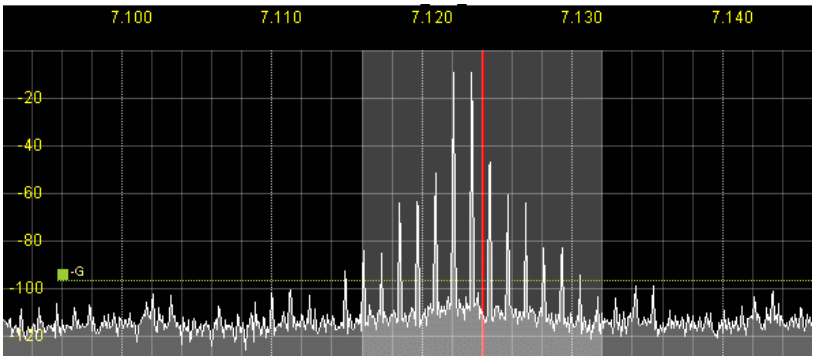
This version of the software cannot correct for memory distortion. Doing so would require additional real-time computing. But thanks to a stable supply voltage right at the amplifier, and very low temperature variations due to this LDMOSFET having a huge dissipation rating relative to the actual dissipated power, memory effects in this amplifier are very low.

Such a distortion correction system samples the output signal only over a specific bandwidth. In the case of PowerSDR, this is a little less than 48kHz. Any distortion products falling outside that bandwidth aren't "seen" by the software, and thus cannot be corrected. When the software predistorts the drive signal to correct the signal inside the bandwidth it sees, the distortion products outside that bandwidth actually increase! For this reason it's important not to overdo the overdriving and the bias reduction, to keep these far-away distortion products small.

Of course whenever building or adjusting a linear amplifier, it's essential to do proper **distortion measurement**. If you check very old radio handbooks, you will find several pages showing you photos of oscilloscope screens displaying various patterns, usually created by feeding a dual tone signal into the transmitter and watching the RF output. The book author will try to teach you to visually recognize a signal having various forms of distortion, such as flat-topping (saturation), crossover distortion (incorrect bias setting), and others. Indeed this was a good method for those times, particularly because vacuum tubes don't cause the strong phase distortion of transistors, so the main kind of distortion back in those years was amplitude distortion. Still, measuring the intermodulation products on a spectrum analyzer is a much better method. It will detect all usual forms of distortion, and will accurately measure how much they impact the signal quality.

In ancient times, in the past millenium, spectrum analyzers were excruciatingly expensive, and were to be found only in labs belonging to rich companies. But nowadays a lot of very inexpensive spectrum analyzers have become available. Their performance is typically not at the level of a professional instrument, but plenty good enough to measure the signal purity of transmitters. Tiny handheld ones exist, also PC-based ones, and many digital scopes also have spectrum analysis functions. If you buy any of these inexpensive things for amplifier testing, make sure before you buy that it has enough frequency resolution. To be really useful, it would be good if the resolution was no worse than about 100Hz.

Here is an example of a spectral measurement done with PowerSDR, of my little 60W amplifier, without predistortion, well clear of its saturation, and with optimal bias setting:



Never mind the red line, which indicates the suppressed carrier frequency of my LSB signal, nor the gray bandwidth shown around it.

The two tones transmitted should appear nominally 6dB below the zero line, because each contains one quarter of the peak envelope power, but I drove the amplifier only to half power, so they are at -9dB each. The next shorter lines are the two 3rd-order intermodulation products. They are about 40dB below each tone, or 46dB below PEP. The next two are the 5th-order products, 53dB or so below each tone. The 7th-order products are almost of the same strength, but the 9th-order ones are much weaker, more than 70dB below each tone, making them totally irrelevant.

Such a result is considered excellent signal quality, for communications purposes. Most commercial radios are much worse than this. But even this amplifier is very poor when compared to audio amplifiers. Expressed in audio terms, the above spectrum would equate to roughly 1.7% total distortion. A golden-eared HiFi fan would break all olympic records while running away from such an amplifier! Audio guys shoot for distortion below 0.01% or so, and at least 90dB dynamic range. In communications use, even 3% distortion and 40dB dynamic range are considered to be quite good.

When varying the power, such a spectrum will not move up and down as a block. Instead individual IMD product pairs will tend to rise more or less, in various patterns. Typically a little into saturation the 5th-order products get as strong or even stronger than the 3rd-order ones, and at even higher drive they fall back again. So it's never enough when a manufacturer claims "3rd-order IMD is down by 40dB". He should at least also tell how strong the 5th-order IMD was at that power setting! Clever people love to play tricks here, turning the power control for the lowest possible 3rd-order IMD, even if at that setting the 5th-order one is much higher! Specmanship at its best...

Time for a Blue Block, just in case...

Intermodulation

When any signal flows through any nonlinear device, distortion is created. When several signals flow through that device, much more distortion is created. Every signal (where one "signal" is a single-frequency) mixes with each of the others, and with itself.

When signals mix, they create an infinite series of new frequencies: All combinations of sums and differences between them. These are called intermodulation products. How many signals or copies of the same signals interact to create a new one, defines the order of the intermodulation product.

A single signal going through a nonlinear device can only mix with itself. For example a single 7100kHz signal will add to itself, producing 14200kHz, this being a 2nd-order signal. This one will again add to the original signal, creating 21300kHz, a 3rd-order signal. The higher the order of a product, the lower is its amplitude. These intermodulation products of a single signal with itself are called the harmonics of that signal, because they are harmonically related: 1:2, 1:3, 1:4, and so on. Interestingly there are cultural differences in how they are counted: In English and Spanish we call the 2nd-order product the 2nd harmonic, the 3rd-order product the 3rd harmonic, and so on, while Germans call the 2nd-order product the 1st harmonic, the 3rd order product the 2nd harmonic, etc. Go figure... I found this out the hard way. Don't ask.

The difference frequency between a single signal and itself is zero Hz, that is, DC. This is the IMD view of why a diode creates DC from AC, when acting as detector or rectifier. It's a way to see a simple process in a complex way...

Other difference signals between harmonics fall on the same frequencies as other products, so there is no point in counting them separately.

When two different signals mix, matters get more complex. For example, when mixing 7100kHz with 7101kHz, we get:

1st order: 7100 and 7101
2nd order: 1, 14200, 14201, 14202
3rd order: 7099, 7102, 21300, 21301, 21302, 21303
4th order: 2, 14199, 14203, 28400, 28401, 28402, 28403, 28404

And so on. The higher the order, the more products there are, and the weaker is each.

Note that even-order intermodulation products of nearby frequencies all fall far away from the original frequencies, either at audio or into harmonic bands, while odd-order intermodulation products fall both into harmonic bands and into the operating band, very near to the original frequencies.

Audio products are easily suppressed by making an amplifier that doesn't respond to audio. This is done using small enough coupling capacitors, RF transformers with a not too low cutoff frequency, and using RF chokes instead of wideband chokes. The products that fall into harmonic bands are also easily filtered out by using a lowpass filter after the last amplifier stage. A problem happens just with those odd-order products that fall into the operating band. They cannot be filtered out by any practical means, so if we want an amplifier that doesn't cause severe splatter into neighboring channels, we need to make it linear enough to suppress these products as much as required by the application.

Thermal design

There is just one way of killing RF power transistors that is even more common than massive common-mode overvoltage spikes on the drains or collectors: Simple, plain and old overheating! And in fact, when transistors fail from overvoltage, this is typically also a case of overheating. Just instead of overheating the whole silicon die, typically just a small spot of it is overheated, through localized avalanching.

Hams and other amateurs are particularly famous for overheating their transistors. Don't get me wrong, I don't mean to imply anything negative in the word "amateur"! After all, it means "lover". An electronics amateur is someone who loves to do electronics, even if he may not be a professional in the field. Such a person enjoys my deepest respect, since I also started that way, at age 12, and I have very fond memories of doing things with great enthusiasm, but without the faintest idea of how to do them.

There are many rules of thumb which are simply not true, such as "If you can keep your finger on it for 5 seconds without feeling pain, it's not too hot." And a very common mistake made by amateurs is underestimating the huge temperature gradient between a heatsink and the silicon chip inside a transistor package. Many times when I told someone that he simply cooked his transistor to death, I got the surprised statement "That can't be the case, because the heatsink was barely lukewarm when the amplifier failed." My standard reply to this is: "Well, what exactly burned out? The heatsink or the transistor?"

My point is that one can easily end up with such a high thermal resistance between the silicon die and the heatsink fins, that the heatsink stays stone cold while the die dies from heat stroke.

The problem starts with the sad fact that most amplifiers dissipate a lot of power. A simple old rule of thumb (another one!) says that an amplifier will dissipate about as much power in heat, as it puts into the antenna. Unfortunately this rule falls very short of the truth, particularly for broadband linear amplifiers used in SSB! Most such amplifiers end up working in dynamic class A. They reach a peak efficiency, at full power, of slightly under 50%, but their average efficiency over the whole envelope waveform might be as poor as 20%. So, four times as much power goes into the heatsink as into the antenna! Such RF amplifiers are really room heaters with an auxiliary RF output! Since hams, ship operators, and many other users of the HF bands only operate sporadically, and transmit for very short times, efficiency isn't important to most. But it's

essential to make a proper thermal design of such amplifiers, to keep the transistors alive over a long time.

The essential points here are:

- Calculating the worst-case power that the transistor needs to dissipate,
- Deciding how hot we can allow the silicon to run,
- Finding out what our highest environmental temperature will be,
- Calculating the required thermal resistance from the transistor chip to the ambient, and
- Actually implementing the required low thermal resistance in our amplifier!

The dissipated power varies a lot over the RF cycle, and also over the envelope cycle. It also varies with load conditions. We don't need to consider the variations over the RF cycle, because thermal inertia of the silicon chip proper is plenty large enough to average the temperature of the chip over a whole RF cycle, and even several cycles. But we do need to consider the dissipation fluctuations over the envelope waveform, and the easiest way to do that is to simply consider what will happen at each particular drive level, when transmitting a plain carrier. The highest dissipation will typically not happen at full power output, but at some lower level. The exact power output that causes maximum power dissipation varies with the bias level, amplifier type, and so on. If we make sure that the silicon chip stays cool enough at the worst possible drive level with a plain carrier, it will also be fine over the whole envelope of an SSB signal.

Datasheets always state an absolute maximum allowed junction temperature. Depending on transistor characteristics, such as the metal used to contact the silicon, this absolute maximum temperature may be anything from 150 to 225°C, for silicon transistors. But this does not mean that we can operate a transistor for a long time at the rated temperature! Instead its life span will be short, if operated that hot. We need to keep it cooler if we want it to last. How much cooler, depends on how long a life we want it to have, also on current density in the silicon, and other things. Manufacturers often provide information relating chip temperature to mean time before failure, for specific operating conditions, specially current. These can be used if available. Otherwise, as a rule of thumb for typical intermittent use by hams and similar users, running a transistor about 30°C cooler than the absolute maximum rated temperature usually results in an acceptable life span. But if high reliability is desired, the temperature should be kept even lower than that. Keeping a silicon chip below 100°C results in unlimited life span, on a human time scale.

Ambient temperature depends on where the equipment will operate. When using it in a fresh climate, or in air-conditioned rooms, and the heatsink is located outside the radio in free unconfined air, 25°C is a good value to take. But in most cases we have no assurance that a radio will always be used under such benign conditions. Depending on the situation, we might need to assume a maximum ambient temperature of 35°C or even more. And if the heatsink is inside the radio, or enclosed in any other way, we need to carefully evaluate how high the air temperature in that space might get.

Once we have the maximum actual dissipated power, the maximum junction temperature that will give a good life span, and the maximum ambient temperature, it's trivial to calculate the total junction-to-ambient thermal resistance needed.

Achieving it is a totally different matter, and is the hardest part of the process. Transistors normally come with their internal junction-to-case thermal resistance clearly specified, but from there on we are on our own. The total thermal path of a typical transistor mounted on a heatsink is a series connection of these:

- Junction to case
- Case to heatsink
- Inside heatsink
- Heatsink to air

Junction-to-case thermal resistance is known, but the others usually aren't. The case-to-heatsink thermal resistance is dramatically dependent on the mounting method: Soldering the transistor to a smooth heatsink surface is good. Bolting it down with a very thin layer of good heat-conductive paste is bad. With a less good paste, or a thicker layer, it's even worse. Dry mounting, without any paste or other filler, is horrible. And any sort of insulated mounting, be it with mica, ceramic, kapton, silpad, is worse than horrible. Silpad mounting is about as bad as mica installed with heat conducting paste on both sides. In all these cases, clamping is often better than bolting, and smooth flat surfaces are better than rough, irregular ones.

I don't mean that insulated mounting, thermal grease, bolts and clamps, can't be used. They do have their place, because in a great many situations their performance is good enough. But when you need the best thermal performance, there is no alternative to soldering the transistor to the heatsink.

The thermal resistance inside the heatsink is the worst-understood item of all these. The transistor might have a mounting surface of only 1 to 4cm^2. But the heatsink is large. The heat has to travel from the transistor mounting spot to every last corner of the last fin. Some paths are short, like a few millimeters from the mounting spot straight through the base plate to the closest spot where air touches. And some paths are long, like 20cm from the transistor mounting spot to the corner of the farthest fin of a moderately large heatsink. In between there are infinite other path lengths, and they have varying cross sections. Calculating this precisely requires relatively advanced maths, so many electronicians will instead prefer to use simplified calculations, approximating the final value by dividing the heatsink into just a few sections for calculation, or use online thermal calculators to get the job done. The all-important point here is to never forget the thermal resistance of the spot under the transistor, and in the lateral conduction inside the heatsink's base plate! Too many people forget to consider them, and their transistors overheat.

When using a commercially made heatsink, the heatsink-to-air thermal resistance is usually given by the manufacturer. But there are two big catches, and a small one. The first big one is that this value is valid only for very specific conditions, such as free natural convection with a specific temperature difference, or perhaps it's rated for a given airflow rate, using a specific fan. When using natural convection, the heatsink-to-air thermal resistance varies dramatically with heatsink temperature, because a hotter heatsink causes much stronger convection. So a very nonlinear ratio results between dissipated power and heatsink temperature. With a small power it already heats up a lot, but with higher power the additional heating is progressively smaller. It also varies a lot with the orientation of the heatsink. Vertical fins, with free access from below and above, is best. When using a fan, or some other form of forced convection, the heatsink-to-air thermal resistance becomes much lower, and almost independent from the dissipated power and orientation.

The second big catch is that when a manufacturer specifies the thermal resistance of a typical heatsink consisting of a flat base plate with fins on one side, he assumes that the heat source will be evenly distributed over the entire flat surface of the base plate! But this is very rarely the case with RF power amplifiers, and so the actual thermal resistance from the transistor's mounting spot to the air is higher, and often very much higher.

Finally, the small catch is that this rating is only valid at sea level. As soon as you use the radio in a higher location, the air is thinner, so the heatsink-to-air thermal resistance gets worse. You either need to recalculate it for the altitude at which you will operate the equipment, or for the highest altitude at which someone may want to operate it. And that's why most equipment has a maximum operating altitude specification!

When designing an amplifier, you need to be specially aware of **thermal bottlenecks**, where a lot of heat flow concentrates. It's most important to keep thermal resistance low at those places. The heat is all generated in a tiny silicon chip, and silicon is a relatively poor thermal conductor, so there we have the most important bottleneck. Then the heat spreads out a little in the transistor's metal base, and that's the second bottleneck. Both of them are added up and specified by the manufacturer as the junction-to-case thermal resistance. We can't change the insides of an existing transistor, but we can select a transistor that is adequate for our needs. And if none exists, we can combine two or more transistors, which results in a lower total internal thermal resistance, since they are thermally in parallel.

The next bottleneck is the case-to-heatsink interface. This one is under our complete control, and we need to make its thermal resistance low enough. This is done by avoiding insulated mounting when so required for low enough thermal resistance, using good thermal paste if needed, and soldering the transistor to the surface when even better performance is needed. The ratio between the thermal resistance of an insulated mounting, and taht of a soldered one, can easily be as much as 100:1! And the other important thing is using a material under the transistor that has good enough thermal conductivity. Most heatsinks are made from some aluminium alloy, that has worse thermal conductivity than pure aluminium, and much worse than copper. That's why in demanding situations a copper heat spreader is used between the transistor and the heatsink. Its main benefit is reducing the thermal resistance in that thermal bottleneck right under the transistor, but an added benefit is improving the lateral thermal conduction along the heatsink's base plate. Its downsides are lengthening the thermal path from the transistor to the air, and adding an additional interface layer, between the spreader and the heatsink, which is usually only bolted and greased, but can be made large enough to have low thermal resistance anyway.

Since the thermal calculations are rather complex, I use online thermal calculators to do them. I prefer not giving links here, because such web pages come and go. Just search for some, and try them. It takes some time understanding them and learning to use them properly.

Water cooling is attractive when a large amount of heat has to be removed. Since water has an extremely large thermal capacity, only a small flow rate is needed, and this small amount of water needs only narrow channels, so it can be passed extremely close to the transistor, by using a properly designed cooling block. A small copper block with many very fine water channels in it, fed with pressurized water, is most effective. The flow will be highly turbulent, producing low copper-to-water thermal resistance, and the fine structure allows having a large transfer surface very close to the transistor, minimizing the thermal path length through copper.

Many people instead use a large copper block with fat water pipes or channels in it. This tends to still be better than air cooling, but not nearly as good as a micromachined small cooling block. Those large blocks are best for distributed heat sources, like some large modules often used in industrial power electronics.

The water can either be cooled back down in a radiator, with or without a fan, or for intermittent, low duty cycle use, it can simply be kept in a large enough container, mostly just storing the heat while the amplifer is used, and releasing it slowly and continuously through the walls. For normal ham operation at the legal limit power, a 20 liter cannister under the desk is usually sufficient. With such a system the only noise of the cooling system is the pump, and if an immersion pump is used it can be inaudible.

Protection

Brave people don't include protection circuitry in their amplifiers. They try to always use the correct antenna, make sure that no connectors are loose, badly soldered, and that no bird ever sits on the antenna while transmitting. Sometimes they are lucky for a long time, but eventually Murphy gets them.

People aware of Murphy's Laws add several levels of protections, and then some. But Murphy still gets them, by making some protection misbehave in a way that kills the amplifier while shutting it down, or whatever.

When you build an amplifier, only you can decide if you want protections, and which ones. In case you are of the brave kind, skip this section!

The problem is that almost all higher power amplifiers can be destroyed by incorrect loading of some sort. Not all bad load conditions will destroy a given amplifier, but some may. Some amplifier might survive a shorted load but burn out when left open, or the other way around. Some amplifier might burn out with any excessively inductive load but survive all capacitive ones, or the other way around. And many broadband amplifiers will be susceptible to dying on one condition on one band, and a totally different condition on another band! That has to do with phase shifts inside the circuitry, which are totally different according to frequency.

Other conditions that can kill amplifiers are excessive temperature, overvoltage at the supply, reverse polarity, overdrive, driving at a frequency outside its range, leaving the input open, nearby lightning, not to mention water ingress, condensation, snails, spiders and other assorted critters making a home inside, etc.

Many people use SWR sensing circuits, that reduce the drive power or even shut if off when the SWR exceeds a certain level. Such a protection can indeed save an amplifier from many mishaps, but it needs to be fast enough. Many aren't. Transistors can take huge overloads of some kinds, but only for a very brief time, microseconds to milliseconds. To save a transistor, the protection must react within this time.

An SWR sensor placed between the low pass filter and the antenna output will not protect against accidentally switching in the wrong filter, or transmitting on a frequency above the cutoff of the selected filter, or a relay failure, which are common. If instead the sensor is placed right after the amplifier block, before the low pass filters, it will protect against that, but the protection might trigger on the normal reflection of harmonics from the filter. Depending on the amplifier type (producing more or less harmonic output), a decision needs to be made. Some amplifiers even use two SWR sensors, one before and one after the filters, the first one being less sensitive than the second one.

Any condition that causes higher dissipation than the maximum one assumed in the thermal design, can make a transistor burn out. To effectively protect against this, a protection circuit needs to continuously calculate the actual dissipated power. This can be done by sensing the supply voltage, supply current, forward power and reflected power, and then making the calculation:

Dissipated power = voltage × current + reflected power - forward power

It's easiest to make these calculations in a microcontroller. It needs to be fast enough. How fast it needs to be, requires a transient thermal resistance calculation, using data provided in the transistor's datasheet. If a microcontroller turns out to be too slow, analog multipliers and op amps can be used instead. They are fast, but analog multipliers tend to be somewhat expensive.

Such a circuit will protect against many accidents. Particularly it can protect against lowpass filter switching errors and relay failures, without requiring a sensor between the amplifier and the filters. It can also avoid many cases of destruction from self-oscillation. But it does not provide sufficient protection from conditions that cause abnormally high drain voltage peaks, because those make a transistor go into avalanche conduction, and the power dissipation capability of a transistor during avalanche conduction is much lower than during normal conditions, because the avalanching tends to happen in a very small area of the silicon chip, concentrating all the heat there.

So it's a very good idea to add direct drain overvoltage protection. It's a circuit that takes a sample of the drain voltage, on both sides for a push-pull amplifier, detects the peak voltage, and acts very quickly if this gets dangerously close to the absolute maximum rated voltage of the transistor.

Another protection that is good to have is fast-acting overcurrent protection in the power supply.

A builder has to decide what to do when a protection circuit detects a dangerous condition. Reducing the drive power is effective in some cases, but not in others, such as those brought on by self-oscillation. Completely cutting off the drive is not much better. Shutting down the power supply is very effective, but tends to have some delay due to charged electrolytic capacitors. If the delay is too long, an additional MOSFET switch has to be added after those capacitors, instead of shutting down the power supply through its control circuit. Reducing or shutting off the bias isn't a very good protection method for LDMOSFETs, because they are not very tolerant of negative gate voltage. If the bias is turned off but drive is kept on, gate damage might result, killing the LDMOSFETs, particularly if ALC is used, that reacts to the sudden drop in output power by increasing the drive! In any case, if protection is implemented via bias shutdown, the bias voltage needs to go down slow enough to avoid causing a drain voltage spike by induction in any chokes present in the supply path. Such a spike can make a bad situation worse, by momentarily adding to the avalanche conduction!

Complete shutdown of the amplifier in the event of an abnormal condition is better than drive reduction. Of course it results in an interrupted transmission, but it may save the amplifier! The protection circuitry should clearly display the reason of the shutdown, until the operator resets it, because having to make the protection shut down the amplifier over and over, to find out the reason, is extremely risky.

Perfect, foolproof protection doesn't exist. Add that to your list of rules. Fools are foolish enough to kill any amplifier. But one can drastically reduce the risk for normal people killing amplifiers, by using good protection circuits.

Don't believe false advertising!!! A video has been circulating for years, in which a company pretends to demonstrate that one of their high power LDMOSFETs is indestructible. The video shows a test setup with power supply, drive source, power meter, load, and the operator disconnects the load, shorts the amplifier's output, drawing very pretty sparks, while the amplifier survives unharmed. But the video totally fails to mention that the amplifier is being driven by a pulsed signal, at a low duty cycle! The RF power meter is a peak-reading one, so at first glance it looks like a continuous RF signal is used. Not so! By paying attention to the power supply displays, it's clear that the amplifier is operating at a low duty cycle.

Also the datasheet of that transistor, and many others, claim that it will survive very high or infinite SWR at all phase angles. Again, that's only true under the operating conditions of the test circuit shown, with excellent cooling, and with the low duty cycle specified in the fine print. In a typical broadband circuit, in SSB, let alone RTTY, and realistic/practical dissipation conditions, these transistors will definitely not survive that high SWR at all phase angles!

You have reached

The End

Phew!

This article took barely four years to complete.

Back to homo ludens electronicus.