RF Power Amplifiers for Wireless Communications

2nd Edition
Steve C. Cripps

By Chen Gaopeng

Preface

1. The problems of maintaining stability and simultaneously minimizing supply modulation effects can cause as much difficulty for a PA designer as does the design of the RF matching networks.
2. I have always held the view that unless you are able to measure RF voltage and current waveforms you can't say with any certainty what the "mode" of PA operation is, and it was my own measurements using this facility that formed the starting point for some of the less conventional view on harmonic matching to be found in Chapter 4 and 8.

Chapter 1 Introduction

1. Rollett showed that the $0<\left|\Gamma_{S,L}\right|<1$ condition corresponds to a parameter known as the "k-factor" being greater than unity, where $k=\dfrac{1-\left|S_{11}\right|^2-\left|S_{22}\right|^2+\left|S_{11}\cdot S_{22}-S_{12}\cdot S_{21}\right|^2}{2\cdot\left|S_{12}\right|\cdot\left|S_{21}\right|}$. So in practical terms, if $k>1$ the device will never display an input or output reflecting coefficient magnitude which is greater than unity, no matter what passive matching may be placed at its input or output.
2. When the signal gain drops below 10dB, the extra RF drive power required will often cancel out any efficiency advantage that had been carefully designed. The upshot of this is that one is often looking for an optimum situation where k-factor is greater than unity, but not too much greater. Devices with high k-factor also tend to have low gain, and some extra gain can be retrieved by allowing positive feedback around the device, while keeping the k-factor above unity.
3. Something about the k-factor which are worth noting:
   (1) Any device which has a k-factor greater than, but not much greater than, unity displays a more aggressive gain/match characteristic than a theoretical unilateral device. In particular, the final MAG may be considerably higher, in a nearly matched condition, than a simple voltage standing wave ratio (VSWR) mismatch calculation would indicate. For example, such a device displaying a 10dB return loss may show more than the calculated 0.7dB increase in gain when finally matched to -20dB return loss.
   (2) Circuit loss can play havoc with the k-factor, and especially the frequency where it crosses unity. In practice, devices can be safely used some way below the unity k-factor point if the k-factor is based on fully de-embedded s-parameter measurements.
   (3) The circuit environment in which a transistor is placed can modify significantly its

 chen.gaopeng@gmail.com

effective s-parameters, and especially the critical reverse transmission parameter, S_{12}. This is probably the main cause of unexpected, or unsimulatable, stability problems.

(4) K-factor analysis, as presented here in its classical form, is only applicable to a single stage amplifier. In a multistage environment, the condition $0<\left|\Gamma_{S,L}\right|<1$ no longer applies, because the input and/or output planes of an intermediate stage are terminated with active networks. So taking a multistage amplifiers as a single two-port and analyzing its k-factor is a necessary, but by no means sufficient, condition for overall stability. This problem is often bypassed in multistage RF amplifier designs by the use of some form of isolation between stages, although multistage stability analysis and design strategies have been published.

4. Weakly nonlinearities may, for example, be intermodulation distortion at levels lower than, say -30dBc.
   Strongly nonlinear effects refer to the distortion of the signal waveform which is casued by the limiting behavior o of the transistor.
5. Generally speaking, PAs operating at or beyond the 1dB compression point require more careful treatment, since the nonlinearities become “strong” and arise through the cutoff and clipping behavior of the transistor. Beware, especially, of analyses which truncate the power series to include second- and third-order effects. The third-order nonlinearity is, undoubtedly, an important contributor to compression and saturation effects in small signal amplifiers, but in PAs the fifth- and seventh-degree terms become significant as the 1dB compression point is approached and can dominate at higher drive levels.
6. The basic conjugate match theorem only applies in a completely unrestricted case where currents and voltages at the generator terminals are unbounded by physical constraints. The loadline match is a real-world compromise, which is necessary to extract the maximum power from RF transistors and at the same time keep the RF voltage swing within specified limits and/or the available DC supply.
   The reason for the loadline match is to accommodate the maximum permissible current and voltage swings at the transistor output.

Chapter 2 Linear Power Amplifier Design

1. It is widely assumed that the two terms “Class A” and “Linear” are almost synonymous, certainly as far as RF power amplifiers are concerned. In fact, Class A amplifiers are often by no means linear, and highly linear amplifiers are not necessarily, or even frequently, of the Class A type.
2. No matter what criterion (the maximum linear power or 1dB compression power) is used for RF power, the power match gives about the same 2dB improvement. That is to say, the maximum linear power increases with power tuning by about 2dB as well as the 1dB compression power. This is a typical observation, albeit a strictly qualitative one. Across a wide range of devices and technologies, the actual difference may vary over a 0.5 to 3 or 4dB range, but the improvement with power match will be fairly constant over a range of gain compression.

 chen.gaopeng@gmail.com

3. In order to sustain gate-controlled current, it is necessary to reduce the value of the load resistor so that the drain (or collector) voltage is kept above the appropriate knee voltage for the device being used. For BJTs, HBTs, and GaAs FETs, a suitable value for this reduction, denoted by V_k, would be 1 volt. LDMOS devices would require a larger correction, although will also be running at higher supply voltages.

Chapter 3 Conventional High Efficiency Amplifier Modes

1. Straightforward but more detailed analysis than is often presented in elementary textbooks will show that merely reducing the conduction angle of an RF power device is necessary, but frequently not sufficient to obtain a useful improvement in efficiency. In general, it is necessary to increase the drive level substantially from the Class A condition and to provide suitable impedance terminations at harmonics of the signal frequency.
2. Power Utilization Factor: is basically the ratio of RF power delivered by a device in a particular mode under consideration to the power it would deliver as a simple Class A amplifier.
3. Class A,B,AB,C:
   (1) Between Class A and Class B operation, the fundamental RF output power is approximately constant, showing a few tenth dB increase in the mid-AB range over the Class A power output;
   (2) The Class B condition delivers the same power as Class A, but with a DC supply reduced by a factor of $\pi/2$ compared to Class A, giving an ideal efficiency of $\pi/4$;
   (3) The Class C condition shows an ever-increasing efficiency as the conduction angle is reduced to low values; this is, however, accompanied by a substantial reduction in RF output power.
4. Class C's Problem: One of the major problems in utilizing Class C modes is the large negative swing of input voltage, which coincides with drain/collector output voltage peaks. This is precisely the worst condition for reverse breakdown in any kind of transistor, and even small amounts of leakage current flowing at this point of the cycle will have a detrimental effect on the efficiency. For this reason, along with the rapidly decreasing PUF and power gain, true Class C operation has not often been used in solid state amplification at higher RF and microwave frequencies. Some renewed interest is currently apparent, mainly due to the possibility of driving the RF power device using a digital signal.
5. BJTs (and especially the modern GHz derivative, the HBT) can outperform FETs in terms of the tradeoff between linearity and efficiency for variable envelope signals. In any event, it can be stated that gain expansion is an easier form of linearity to correct than gain compression, and there are numerous mechanisms in a practical design which will almost inevitably cause some cancellation and nulling effects in the IM or ACP response of a Class AB BJT PA.

Chapter 4 Class AB PAs at GHz Frequencies

1. Time and again, PA designers find a particular device which exhibits "sweet spots" in Class AB operation, where spectral distortion products drop into nulls over restricted power drive

levels. These effects can usually be attributed to the fortuitous cancellation of the nonlinearity of the classical Class AB current truncation mechanism and the inherent nonlinearity of the device transfer characteristic.

2. The important observation is that the device input voltage is significantly distorted from its original sinusoidal form. If this varactor was the C_{gs} of a device, biased to operate in a deep Class AB mode, the varactor distortion of the voltage remained sinusoidal, as is always assumed in classical analysis. Furthermore, it is clear that the voltage distortion appearing across the C_{gs} capacitor will cause some significant changes in the harmonic components of the resulting transconductive output current.
3. The key point is to recongnize that the input tuning of a PA device can have as much of an impact on power, efficiency, and linearity as the actual quiescent bias point setting and can easily override the intended effects of a classical Class AB design. A strong recommendation is that input harmonic shorts should always be used in order to nullify the nonlinear effects on the device input.
4. In particular, there is the effect of the feedback capacitance, which will have a similar varactor characteristic to the much larger input varactor. There has been a long tradition among PA theorists to assume their active devices can be approximated to be unilateral, but in practice the feedback effects will provide another source for wave shaping on the input. Indeed it can be argued that both input and feedback capacitance have to be considered together in any realistic model.

Chapter 5 Practical Design of Linear RF Power Amplifiers

1. Transmission line elements can, conversely, be used to simulate the effects of lumped element networks. At higher microwave frequencies, it is common practice to design a matching network using a prototype lumped element structures, and then translate it into distributed form. Shun capacitors become Open Circuit Shunt Stubs (OCSS), shunt inductors become Short Circuit Shunt Stubs (SCSS), and series inductors, more restrictively, can be replaced by short lengths of series high impedance line. Series capacitors pose a problem at higher microwave frequencies in that they don not have a simple distributed equivalent. They are generally avoided as matching elements altogether due to parasitic and tolerancing issues. Chip capacitors are used for the larger values required for DC block, but have to be carefully modeled, since they are quite often being used well above their self-resonant frequencies.
2. HBT: The benefits of single supply operation, and the possibility of using many of the circuit techniques employed in analog IC design helped to propel the technology into mainstream use. From a PA viewpoint, the advantage of a single supply was paramount, as was the gradual realization that smaller physical geometries could be used for a given RF power output than comparable GaAs FET or Si MOSFET implementations. What seems to have been largely missed is that the bipolar transistor quite probably makes a better, more linear Class AB PA. Unlike a FET PA design, the gain linearity and efficiency will both be a sensitive function of the input termination.

Chapter 6 Overdriven PAs and the Class F Mode

1. Depending on the modulation system, and the specific application for the PA in the system, there are actually many situations in which linearity, especially amplitude linearity, can be traded for efficiency and RF power output. These applications range from single channel, constant envelope systems such as those using FSK and GMSK which can tolerate high level of amplitude distortion, through to intermediate cases such as QPSK and DQPSK systems which can tolerate significant amounts of amplitude distortion, within the constraints of spectral mask regulations.
2. The saturation of device current when heavily driven at the input can result in higher RF power but typically produces little improvement in efficiency other than for low conduction angles. A more interesting scenario is the deliberate use of voltage clipping, which can give significant efficiency improvement. Although voltage clipping inevitably introduces undesirable odd degree distortion effects, it will be shown that there can be a useful tradeoff between efficiency enhancement and linearity which can be utilized in low or intermediate envelope amplitude applications.
3. Input overdriven PA, causing current saturation: first, it shows that significant improvement in power and efficiency only occur simultaneously for the lightest Class AB and Class A cases; once the conduction angle reaches Class B conditions, a modest increase in power is offset by a reduction in efficiency from the nominal 78.5% value. Shorter conduction angles, into the Class C region, show a quite dramatic increase in power from the nominal values, but closer scrutiny shows that this is really a manifestation of the increase in effective conduction angle, the power increase is accompanied by a corresponding decrease in efficiency from the unsaturated case. It has to be concluded that the classical reduced conduction angle amplifier, with shorted harmonics, is not well suited for efficient exploitation of the possibilities offered by overdrive.

Chapter 7 Switching Mode Amplifiers for RF Applications

1. Class E appears to be easier to realize in practice using solid state transistors than a short conduction angle Class C design. Class E mode may just represent a possibility for achieving high enough efficiencies from solid state devices that envelope restoration and outphasing may become mainstream techniques once again.
2. Simulation and verification tests on actual amplifiers seem to support the view that Class E does truly represent an alternative to conventional reduced conduction angle operation, giving higher efficiency without the circuit complexity of more advanced Class F designs. Class F can achieve very high efficiency, but show substantial reduced power output under optimum efficiency operation. Class E has its own disadvantage in terms of peak voltage levels, and the RF PUF has to be traded against this limitation. But the final judgment may be that Class E appears to be easier to realize in practice using solid state transistors than a short conduction angle Class C design.
3. The Class E mode may just represent a possibility for achieving high enough efficiencies from solid state devices that envelope restoration and outphasing may become mainstream techniques once again.

 chen.gaopeng@gmail.com

4. The main action of a Class E amplifier takes place within, and is indeed critically dependent on, the knee region of the I-V characteristic; the true-on is largely implemented by the turn-off is more conventionally implemented by using the transconductive pinchoff. This is a subtle point which partially explains the widespread belief that Class E cannot be made to work at GHz frequencies due to the slow and symmetrical transconductive switching characteristics of RF power transistors.
5. Class E: Theoretical efficiencies of around 90% can be achieved, albeit with an RF output power as much as 3dB lower than that obtainable from the same device in a conventional Class AB configuration. Class E amplifiers are highly nonlinear, but offer possibilities for envelope restoration techniques. On the downside, peak voltages can be as high as three times the DC supply if the power shortfall is to be kept within reasonable limits. There is also the issue of power gain. If a typical Class E amplifier is running at 3-5dB lower power gain, due to the overdrive requirement on the input, than a comparable linear amplifier, the upper frequency limit will be lower for a given technology. Generally speaking, 12-13dB of linear gain would be required from a given device in order to stand a chance of useful Class E operation; however, the same can be said of conventional high efficiency modes with shorter conduction angles.
6. The really critical advantage of the Class E amplifier is efficiency, and with the possibility of reaching into the 90% range, the advantages of low heat generation offer some novel concepts in packaging, modulation, and power combining.

Chapter 8 Switching PA Modes at GHz Frequencies

1. A classical, ideal switching device can be designed into a multitude of circuit configurations which deliver 100% efficiency. This is as much an artifact of the idealized device model, rather than the result of any particular ingenuity in the circuit design. In practice, of course, finite switching speeds and resistive losses will reduce the observed efficiency, and numbers in the literature have ranged anywhere from mid-nineties down to some more lowly numbers in the 60% region.
2. There has been a long tradition of designing RFPA in a “back-to-front” manner, whereby the desired current and voltage waveforms are defined using essentially mathematical representation, and from these waveforms the circuit, and even in some cases the device, characteristics are deduced.
3. Basically, any waveforms or formulas obtained in a PA analysis using the zero-knee assumption will require the DC supply voltage to be increased by a suitable “knee factor” in order to represent a realizable result. This factor may in many cases be about 10%, but it is the variation from one device type to another, along with varying supply voltage requirements in different applications, that justify the ongoing use of the more universal zero-knee assumption.
4. The restriction of the voltage waveform to contain only fundamental and second harmonics is a justifiable assumption in higher frequency PA circuits, where very often the active device will have an output capacitance that will behave as a near-short at higher harmonics.
5. It is appropriate to conclude this section with a brief summary, to confirm the motivations for the detailed waveform function analysis that it has presented. Taking an ideal Class B PA as a

starting point:

(1) The addition of a substantial second harmonic voltage component allows the fundamental component of voltage to be increased by a factor of up to $\sqrt{2}$, giving the potential for power increase of 1.5dB, and an increase in efficiency by the same $\sqrt{2}$ factor.

(2) This benefit cannot be used in a practical PA design unless the phase of the second harmonic current component is in phase quadrature with the second harmonic voltage component; a symmetrical Class AB/B/C current waveform has no second harmonic quadrature component and requires further engineering.

(3) A simple 45° differential phaseshift between the current and voltage waveforms of a Class A/AB/C amplifier will neutralize the 1.5dB benefit given by the second harmonic voltage component.

(4) Modification of the current waveform can reduce the differential phaseshift required to enable full use of the second harmonic enhancement factor.

(5) Some plausible waveform candidates have been defined, which enable a useful compromise between enhanced efficiency (i.e., over 80%) and minimum fundamental power reduction.

(6) Optimum waveforms display considerable asymmetry, which will be a common occurrence in practice due to the effect of the IV knee and/or nonlinear effects in the device input circuit.

(7) Asymmetry in the RF current is a feature which is absent in most of the literature on RF power amplifier analysis, Class E being a notable exception.

6. (1) Conventional Class B results are not usually achievable in practice, due to the requirement for a perfect harmonic short at all harmonics.

(2) The second harmonic enhanced results represent closer approximations to what real PA circuits do, using typical realizable matching networks.

6. It would be reasonable to summarize the results and analyses of this chapter as follows:

(1) Class E switching mode analysis is correct and has not been questioned.

(2) The Class E mode can be implemented at GHz frequencies.

(3) Class E designs at GHz frequencies do not have instantaneous switching transient, but it does not necessarily cause significant performance degradation from the ideal classical analysis.

These conclusions make two assumptions, however. The first is that we have a clear, unimpeachable definition of what a Class E amplifier really is; the second is that a reduction of efficiency from 100% down to 85% is “not significant.”

Taking the efficiency issue first; the problem with 85% number is that it represents a value which can be achieved, certainly in theory and quite frequently in practice, using different approaches to the design and mode of operation (e.g., Class C or Class F). So the realization of efficiency performance in this region is certainly not an indication of the mode of operation. As far as mode definitions are concerned, this is usually done suing voltage and current waveforms. Unfortunately, at GHz frequencies, these waveforms are rarely measured

 chen.gaopeng@gmail.com

directly, although can of course be observed on a simulator. The definition of Class A, AB, B, C are universally agreed in terms of voltage sinewaves and truncated current sinewaves. Class D is also universal, other than some confusion caused by use of the term to represent a poulsewidth modulation in audio amplification circles. A precise Class F definition has been the subject of some debate, but only concerning the number of odd harmonics, and their respective amplitudes, that are added to the voltage waveform.

Chapter 9 Nonlinear Effect in RF Power Amplifiers

1. So the input third order intercept point is about 9.64dB higher than the cw, or single tone, 1dB compression point. This will be equivalent to 9.64+3dB=12.64dB difference if the total output power is used to represent the two-carrier case.
2. It is a fundamental assumption in envelope analysis that the modulation frequency is sufficiently slow compared to the RF variations that the conditions at any instant can be considered to be quasi-static, and can be determined from a static, cw measurement at the same input power. In most present day wireless communications systems, this is usually still a reasonable assumption. Essentially, the amplifier has to have sufficient bandwidth that the performance characteristics do not change over the bandwidth of the modulated signal. This is almost certainly true for single channel applications, although multichannel, spread spectrum signals may start to show the first signs of violation on this basic assumption.
3. Clearly, if the mean power level of the variable envelope signal is set to be equal to the conventional, single-carrier 1dB compression power, the peaks of the envelope will incur much greater than 1dB of compression, and will saturate the amplifier for a significant portion of the modulation cycle. On the other hand, for an equal portion of the modulation cycle the signal will be well down in the linear region and incur no distortion at all.
4. In general, AM-PM effect can be traced to the signal-level dependence of several transistor model elements. For FETs the input capacitance and both the depletion and junction resistance of the gate-source diode can be primary culprits. It should be noted that nonlinear resistance, in the presence of linear reactance, can cause AM-PM effects, just as much as nonlinear reactances. For BJTs, the nonlinear base-collector capacitance adds an important additional source of drive-dependent phase shift. All of these effects are detailed, interactive, and highly complex in themselves, and pose great challenges for physical modeling, such as that which is more successfully employed for compression and clipping effects. So the only way that AM-PM effect can be treated in a concise manner is to resort to empirical describing functions fitted to physical measurements.
5. Contributions of AM-AM and AM-PM effects to IM3: Note that under the ongoing assumption of small ϕ, the phase distortion contributions will remain approximately in quadrature with the amplitude ones. So there is no possibility of significant cancellation occurring between the components generated by the two distortion mechanisms. The AM-AM and AM-PM components will always combine to produce IM3 sidebands which have a higher amplitude than either of these individual parts. Experimental measurement of the relative contributions of AM-AM and AM-PM contributions requires the measurement of the relative phase of the IM sidebands, this is not a straightforward task.

 chen.gaopeng@gmail.com

6. Memory effects are an additional source of nonlinear behavior that is typically not accounted for in PA models, and represent a source of error in attempts to simulate the distortion characteristics of any PA. Memory effects can be treated to three main causes:
   - Dynamic thermal effects;
   - Unintentional modulation on supply rails;
   - Semiconductor trapping effects.

   The second of these is probably the most common cause of asymmetrical IM sidebands, and can be potentially cured by more attentive design of the biasing networks to the PA stage.
   PA memory effects are very trouble some for the process of predistortion.
7. The most important standard modulation scheme is the GSM (GMSK) system used throughout Europe, the 2.5G EDGE derivative which is finding more widespread use than the GSM replacement for which it was originated, and the various generations of CDMA implementation, including the 3GPP WCDMA system. Orthogonal Frequency Division Multiplex, or OFDM, is a more recent addition and is now standardized in various 802.11 and 802.16 variants aimed at Wireless Local Area Networks (WLAN) and high speed data access (WiFi, WiMax).
8. In performing the most critical function of predicting the ACP (Adjacent Channel Power) response of a PA using a stipulated modulation format as the excitation, the key problem is how to model the PA characteristics. Most commercial packages still use the envelop simulation approach for two-carrier, and multi-carrier signals. These packages also, for the most part, make the quasi-static assumption, that the PA can be entirely characterized by its static AM-AM and AM-PM response, usually cast into a power series format. The power series will have maybe around nine complex coefficients for a ninth degree model. This has been shown to have sufficient complexity to model most PA characteristics.
9. QPSK: Quadrature Phase Shift Keying, is a modulation system which uses the phase of the RF carrier to define a number of states, each of which is assigned a binary symbol. A Binary Phase Shift Keying (BPSK) system has just two phase states, denoted by a 180° phase shift, and in this simplest case each state would be a binary 1 or binary 0, respectively. QPSK systems, by definition, use four phase states, so each state can denote a 2-bit symbol. 8PSK and 16PSK systems are also widely used in the various mobile communications systems.
10. EDGE: It is somewhat analogous to the $\pi/4$ DQPSK system, in that it is a pair of 8QPSK systems. Zero crossing are eliminate by inserting a $\pi/8$ phase rotation at each symbol transition. An important distinguishing feature is the use of a Gaussian filter whose characteristics are carefully prescribed so that an EDGE spectrum cannot be distinguished from that of a GSM signal.
11. CDMA: “direct sequence” CDMA multiplexing system: Each data stream is multiplied by a pseudo-random “chip” sequence. The chip rate will typically be many times higher than the original data bit rate, and can thus be considered to “spread” the data stream over a much larger bandwidth than its original form. At any pint in the process, the original data stream can be retrieved by a suitable inverse process, provided the pseudo-random “key” sequence is known. The chipping process is, in effect, a form of digital modulation and is usually encoded

using one flavor or another of QPSK. The key point in the CDMA process is that the various chipped data stream can now be literally added together, with full confidence that the digital keys will be able to unlock the individual components at any later stage in the process.

12. It should be noted that in most cellular systems, the ACP requirements for mobiles are considerably more forgiving than the basestation specifications, due to the nonuse of the mobile transmit channel in adjacent cell.
13. WCDMS: In its basic form, “wideband” CDMA (WCDMA) follows much the same principles and implementation as the IS95 CDMA system, the main change being the use of a higher chip rate, now mainly standardized at 3.84Ms/sec.
14. The difficulty of specifying and testing PAs for 3GPP applications has been recognized in the form of test standards, which define configurations that representative of typical expected operation. Unfortunately, early experience with pilot 3G installations has caused the concept of “typical” operation to change, to the ongoing frustration of PA manufactures.
15. From the PA designer’s viewpoint, the multi-carrier basis is troublesome, since it inevitably causes high PAR, significantly higher than other single carrier QPSK systems so far encountered.
16. Any signal whose peak envelope power exceeds the power level at which gain compression or AM-PM occurs will cause distortion, which in the frequency domain will appear as inter-modulation or spectral regrowth. If the mean power level of the signal is backed off to the point where about 1dB compression occurs at the peak envelope power, the spectral regrowth will drop to a level which is frequently acceptable for single channel applications, but probably unacceptable for multi-channel applications. In the latter case, external linearization will enable a compromise in the amount of additional power back-off required in order to satisfy the regulatory spectral requirements.

Chapter 10 Efficiency Enhancement Techniques

1. The power amplification of amplitude modulated RF signals has two inherent problems. The first is that the envelope, and hence the modulating signal, will be distorted to some degree, if the power amplifying device is used at its full rated RF power level. The second is that conventional power amplifier designs only give maximum efficiency at a single power level, which is dependent on the circuit design but usually near the maximum rated power for the device. As the device power is backed off this point, the efficiency drops sharply and the heat dissipation can increase, even though the RF output power is also decreasing. The overall effect is therefore to measure a mean efficiency which is much lower than the efficiency at the maximum power, or PEP, level.
2. Three classical efficiency enhancement techniques: the Doherty Amplifier, the Outphasing Amplifier initially proposed by Chireix, and the Envelope Elimination and Restoration (EER) technique demonstrated by Kahn in the early days of single sideband (SSB) transmission.
3. The key issue, when dealing with a signal which has a time-varying envelope amplitude, is that the full power capability of the PA is only needed at the peaks of the signal envelope. When the envelope dips down to much lower amplitudes, the capability of the PA is wasted, to a greater or lesser extent. Reduced conduction angle modes, even in their original basic form, do alleviate this problem in that the current peaks adapt to the amplitude of the

chen.gaopeng@gmail.com

incoming signal envelope. Due to the rectified nature of a Class B, or deep Class AB current waveform, the DC supply will also automatically reduce under low envelope excitation. This will not happen in a Class A amplifier, and even a lighter Class AB mode will be particularly susceptible to low efficiency during envelope minima.

4. In conclusion, the Doherty technique appears to offer some useful possibilities for solving some of the problems which arise in both mobile and basestation amplifiers in modern wireless communications systems, and it has gradually been receiving more attention, as witnessed by many research papers in the last few years. But actual commercial products still seem slow to emerge and from this it is has to be concluded than some of the implementation details remain troublesome, for a technique that at the outset looks tailor-made for such applications.
5. So the key element in an outphasing PA is the generating of two constant amplitude input signals which have the amplitude modulation information encoded as a differential phase shift. These signals can then be amplified by a nonlinear device and the original AM is then recovered in a summing operation. Furthermore, this procedure operates in an orthogonal manner to any phase modulation on the original signal, so signals having both AM and PM can be fully reconstituted at the output. So ideally the final distortion levels at the PA output depend on the integrity of the AM to PM conversion process in the modulator, and not on the PA themselves, which operate at constant RF signal amplifiers.
6. Notwithstanding the challenge of implementing a suitable power converter, it should be further noted that the EER technique is potentially immune to RF bandwidth and matching issues, which become severe limitations of the Doherty and Outphasing techniques for bandwidth greater than 10%.
7. Linearity problems, as always, can in principle be handled using DSP techniques, however this introduces some serious overheads to the cost and complexity of the system. In particular, the need to calibrate each individual RFPA device characteristics is a detraction from the ER and Polar techniques.
8. Envelope Tracking: Recalling the analysis of a conventional Class B amplifier, we have already seen that if the RF load resistor is decreased in inverse proportion to increasing drive voltage amplitude, the efficiency remains constant at its maximum value while the power increase as the square root of the drive power; this is an important mode of dynamic behavior in the Doherty amplifier. An alternative, and considerably simpler, scenario is that the load resistor remains fixed and the supply voltage is increased in proportion to the increasing drive voltage. In this case, maximum efficiency is maintained (due to full rail-to-rail voltage swing) and the output power increases linearly with input driver power. This process can be continued down to a selected point on the lower side, and up to a point where the RF swing reaches breakdown level. In this manner, maximum efficiency can be maintained over a wide linear power range.
9. The interaction of any efficiency enhancement techniques with the system linearity always seems to be a negative one, to the extent that there seems to be no single approach which does not carry with it the excess baggage of linearization, usually in the form of digital predistortion. It is indeed ironic that the oldest efficiency enhancement technique in the book, that of reduced conduction angle PA operation, still remains very much the default approach in the industry.

chen.gaopeng@gmail.com

Chapter 11 Power Amplifier Bias Circuit Design

1. The oscillatory habits of RF power transistors are an infamous aspect of RF power amplifier design. This much-dreaded skeleton jumps out of its closet at the most inconvenient times, frequently causing destruction and panic in its wake. Its effects are most prevalent at lower frequencies in the MHz to VHF range, where the terminating impedances of an RF power device are mainly defined by the bias insertion networks. The design of bias networks for any RF power amplifier therefore plays an important part in establishing stable operation.
2. The very term "DC", as applied to the supply requirements of a typical Class AB RFPA becomes inappropriate. Any RFPA stage which is operating in a Class AB mode will draw an alternating current though its supply circuitry, the variation following the peaks and dips of the amplitude-modulated signal envelope. Ideally, the on-board biasing circuitry of any RFPA should smooth out these variations in supply current such that under all operation conditions the user, or system interfaces, is not required to supply any alternating components. Unfortunately, physical limitations on capacitor technology can make this a near-impossible task for higher power RFPAs, and as a result problems of supply modulation can arise from the manner in which an RFPA is connected to the primary power source, as well as the on-board bias network itself.
3. The stability criteria, or assumption, is that any negative resistance component at any frequency over the validity range of the analysis constitutes an oscillation hazard when the decoupling capacitor at the port in question is replaced. Of course, oscillation will in fact be restricted to the frequencies at which the reflection back from the decoupling capacitor has the appropriate phasing. In many practical cases such a frequency may not exist, and the device will operate without oscillation despite the existence of negative resistance over certain frequency ranges. The safest design approach, however, is to seek to eliminate the negative resistance as much as possible, notwithstanding the probability that in many cases we would get away with it.
4. In a formal stability analysis, the existence of a negative resistance component is merely a necessary and not a sufficient condition for oscillation. The input termination has to present a load having a specific magnitude and phase for oscillation to occur, and the analysis can be extended to consider this. But in general the goal should be to eliminate the negative resistance region as much as possible, especially at higher frequencies where the phase of the load will rotate much more quickly as a function of frequency.
5. The overall conclusions from this analysis are quite surprising and are worth summarizing:
   (1) Almost any RF power transistor (e.g., device with $g_m > 1S$) will show a negative resistance at the input bias port, over a wide range of baseband and VHF frequencies, if the output bias is fed through an inductance anywhere between 10nH and 1mH.
   (2) This is the primary culprit for low frequency oscillation in RFPAs.
   (3) The same, precisely, applies to the output impedance in the reverse situation.
   (4) A series resistor can be used to cancel the negative resistance on the input port.
   (5) This in turn greatly reduces the magnitude of the output negative resistance.
6. Class AB amplifiers draw a current from the main output bias supply which varies from a low

chen.gaopeng@gmail.com

value, maybe 10% of the peak value, as the amplitude modulation on the input signal swings the envelope amplitude over its full dynamic range. Any impedance which is placed in the bias supply path will cause voltage modulation to appear at the output terminal of the device; this will in turn modulate the gain and phase of the amplifier and cause additional distortion products. The designer therefore has to ensure that this series impedance is sufficiently low that the resulting voltage modulation has an acceptably low amplitude.

7. Just as it is possible to predistort the input signal to a PA in order to compensate for the PA distortion, it would seem equally possible to predistort the supply to a PA in order to remove the impedance effects of known, fixed, passive elements in the bias supply chian.
8. SCSS for matching purpose: it has been noted that the device output capacitance has the effect of transforming the device loadline impedance in the wrong direction, increasing the impedance transformation requirement from the output matching network. The use of a Short Circuit Shunt Stubs to resonate the output capacitance, removes this undesirable feature and creates a convenient, and optimally positioned, biasing point. Such an element could also be used as part of a broader-band matching network.
9. Secret of Bias Tee: The secret, such that it remains, is usually the use of a large bias inductance which is cleverly designed to eliminate resonances over a very wide band of frequencies, perhaps 10MHz to 20GHz.

Chapter 12 Load-Pull Techniques

Chapter 13 Power Amplifier Architecture

1. The contribution of the drive stages to the overall efficiency of the assembly is reduced by the factor of the final PA stage gain, and it is quite justifiable to design all the driver stages as simple Class A, or light AB, amplifiers. This has great benefits in terms of linearity and AM-PM issues, and avoids the demanding harmonic termination requirements of higher efficiency modes. The key problem in designing a PA driver chain is to limit the distortion, or gain compression, of the drivers so that most of the distortion takes place in the PA stage itself.
2. There is a tendency, especially in the lower frequency PA regime, to regard the push-pull configuration as mandatory, and to ignore the quadrature balanced approach. Conversely, the microwave ECM amplifier community use quadrature balanced modules almost exclusively, and claim little or no benefit from push-pull operation. The wireless communication bands occupy an interesting middle-ground position, and both techniques should be examined and well understood before making any decision on the architecture of a new PA assembly.
3. Balanced Amplifier: Two identical amplifiers are fed from an input power splitter which produces two signal in phase quadrature, the outputs being recombined using a similar device connected in reverse. The principal advantage of this configuration is that any mismatch reflections from the amplifiers pass back through the couplers and appear in antiphase and therefore cancel at the RF input (or output) port. Reflected energy is diverted to the terminated coupler ports.

chen.gaopeng@gmail.com

4. Balanced amplifiers are much more stable than their single-ended equivalents. So much so, in fact, that in the balanced world some of the dire predictions of k-factor analysis can be quietly set aside.
5. Insertion Loss of Wilkinson Combiner: Radiation losses, and simple copper losses due to the use of thin dielectric board material (and hence thin conductor strips) at higher frequencies, can cause the losses of such structure, especially double section designs, to creep up to unacceptable levels for RFPA work. The use of a good low loss dielectric board material does not, unfortunately, guarantee low loss performance in these applications.
6. Waveguides Power Combiner: In particular, the properties of waveguides can be sued to make very effective multiway power combiners that can have almost negligible losses, but these designs would appear to be impractical at lower frequencies due to the physical size of the waveguide. The loss performance may nevertheless become sufficiently attractive as the final power requirement reaches toward kilowatt levels.
7. Driver chains become more challenging when overall linearity is important. The key issue is to control the generation of nonlinear products in the driver stages, while maintaining efficiency and power supply specifications. At first, this may seem a simple enough task of selecting the periphery of each driver and predriver, taking due account of the gain in front of each respective stage. Assuming the output PA stage is designed to run at a given level of backoff in order to meet a given specification, it would seem obvious that the drivers all have to run at the same level of relative backoff, but with some added margin in order to ensure that most of the nonlinear effects come from the output stage; 3dB extra backoff is a typical rule-of-thumb often used.
8. It is clear that the AM-PM phase shift can be either positive or negative, depending on the specific device characteristics, the input and output tuning, and the bias conditions. In a given situation, it is quite possible to find a fortuitous cancellation between the AM-PM characteristics of the driver and the PA stage. This cancellation only becomes possible if the driver stage itself is driven into compression, and is biased more towards cufoff. Well backed-off Class A drivers may be more linear in themselves, but do not provide the opportunity for such helpful cancellation scenarios. Such fortuitous cancellations are pleasing to observe, but have some dangers associated with their long-term reproducibility.
9. An interstage matching network has to include both output and input bias insertion networks for the driver and PA devices, respectively, including a DC block.

Chapter 14 Power Amplifier Linearization Techniques

1. Techniques which improve efficiency are of paramount importance in mobile systems, where battery lifetime and thermal management are critical. There are other PA applications where efficiency becomes an important, but secondary consideration, in comparison to linearity. Such applications would typically be single or multichannel basestation transmitters in ground or satellite communications system.
2. Feedback techniques, most notably those which use input and output detection in order to form a corrective drive signal to the PA, have been largely sidelined by the increasing bandwidths of modern communications signals. Even a single carrier WCDMA signal has a bandwidth in excess of 1MHz, which poses great problems in a feedback linearization system

chen.gaopeng@gmail.com

due to the virtually irreducible delays in detecting, differencing, and amplifying a corrective signal.

3. Predistortion is the generic term given to techniques which seek to linearize a PA by making suitable modifications to the amplitude and phase of the input signal. Feedforward, a technique which has thus far dominated multicarrier PA linearization, applies a corrective signal at the PA output.
4. It becomes clear that AM-PM distortion effects can contribute significantly to IM products in the precompression zone, and in order to reduce the IM level down to -60dBc the linearizer has to reduce gain compression down to 0.01dB and AM-PM to well under 1°. The AM-PM curve is probably less familiar, and very significant. It shows that correction of AM-PM down to one or two degrees, over the whole dynamic signal range, will always be necessary to achieve more than about 10dB reduction in spectral distortion products.
5. Subsequent generations of DPD configurations have gradually got on top of the memory effects problem, through three main developments:
   - Reduction of memory effects through more careful PA design;
   - Addition of an adaption loop in the DPD system, enabling continuous monitoring of the linearization integrity, and LUT refreshing;
   - Use of DPD algorithms which can correct shorter-term memory effects through the inclusion of some of the recent “history” of the signal.
6. Physical modeling of memory effects in PAs could be straightforward enough if, as discussed in Chapter 9, they can primarily be attributed to dynamic thermal, and bias supply modulation, effects.
7. Analog predistortion usually take the form of attenuators with an expansive insertion loss characteristic. It is worth noting that the analog version would be unlikely to fare any better than a memoryless DPD scheme in its ability to handle PA memory effects.
8. To the considerable chagrin of the DPD camp, a feedforward loop performs its error corrections in real time, literally at lightening speed, and as such is essentially immune to PA memory effects.
9. It is clear that the feedforward will always cancel the IM distortion, regardless of the sampling coefficient α setting, provided that the EPA is not generating any comparable levels of IM itself, and also assuming that the loop is kept perfectly gain and phase tracked. For any particular signal environment, it is usually possible to find a setting which minimizes the error signal entering the EPA; in practice this is mainly an issue of arranging for cancellation of the amplified signal component. Provided this is done slowly enough, so that it does not interfere directly with the intended ongoing action of the FFW loop, significant improvement in linearization performance, for a given EPA power, can be achieved.
10. Clearly, if the α parameter is varied in envelope time, to keep track of the signal envelope, such that no error signal ever gets to the error amplifier, then the FFW loop will not generate any corrective signal and no linearization action will tack place.
11. Direct application of classical feedback theory to microwave amplifiers is somewhat irrelevant. It is worth noting however, that the RF world is a little careless about assigning the term “feedback” to various configurations which include some amount of lossy matching to improve bandwidth at the expense of gain. Microwave circuit designers use the term “feedback” to categorize amplifier designs that use a moderately high value resistor connected

chen.gaopeng@gmail.com

from the drain to gate, or collector to base. This resistor acts mainly as a damping element, which improves stability and decrease the Q-factor for broader bandwidth applications. But these amplifiers do not usually show any significant linearity benefit from using the feedback element.

12. The Polar Loop is a complete modulator, rather than a linearizer, but it include some linearization in the modulation process. The Polar Loop has nevertheless come back into favor in the handset PA zone, especially for challenging modulation systems such as EDGE, which make excessive demands on the efficiency-linearity tradeoff in conventional Class AB approaches. Claims and counterclaims seem to be made concerning the feasibility of making the system work for a 300kHz bandwidth signal in its original closed loop form.
13. Despite their universal impact in electronic engineering, feedback techniques are of limited use for the linearization of RFPAs. An exception is the widespread use of indirect feedback techniques, especially the Cartesian Loop, in single carrier transmitter applications which have modulation bandwidths less than 100kHz.
14. PA linearization can essentially be performed in three ways: modification of the input signal, modification of the output signal, or dynamic modification of the amplifier characteristics.

chen.gaopeng@gmail.com