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RF and Microwave Power Amplifier and Transmitter Technologies — Part 3

By Frederick H. Raab, Peter Asbeck, Steve Cripps, Peter B. Kenington, Zoya B. Popovich, Nick Pothecary, John F. Sevic and Nathan O. Sokal

Transmitter architectures is the subject of this installment of our continuing series on power amplifiers, with an emphasis on designs that can meet today's linearity and high efficiency requirements

The building blocks used in transmitters are not only power amplifiers, but a variety of other circuit elements including oscillators, mixers, low-level amplifiers, filters, matching networks, combiners, and circulators. The

arrangement of building blocks is known as the architecture of a transmitter. The classic transmitter architecture is based upon linear PAs and power combiners. More recently, transmitters are being based upon a variety of different architectures including stage bypassing, Kahn, envelope tracking, outphasing, and Doherty. Many of these are actually fairly old techniques that have been recently made practical by the capabilities of DSP.

7a. LINEAR ARCHITECTURE

The conventional architecture for a linear microwave transmitter consists of a baseband or IF modulator, an up-converter, and a power-amplifier chain (Figure 20). The amplifier chain consists of cascaded gain stages with power gains in the range of 6 to 20 dB. If the transmitter must produce an amplitude-modulated or multi-carrier signal, each stage must have adequate linearity. This generally requires class-A amplifiers with substantial power back-off for all of the driver stages. The final amplifier (output stage) is always the most costly in terms of device size and current consumption, hence it is desirable to operate the output stage in class B. In applications requiring very high linearity, it is necessary to use class A in spite of the lower efficiency.

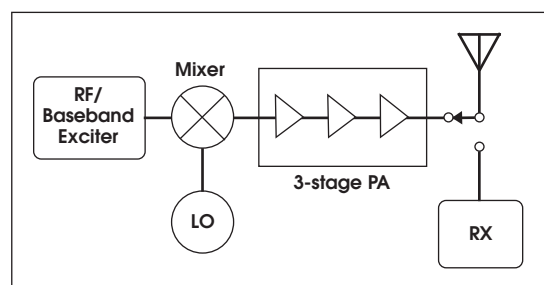


Figure 20 · A conventional transmitter.

The outputs of a driver stage must be matched to the input of the following stage much as the final amplifier is matched to the load. The matching tolerance for maintaining power level can be significantly lower than that for gain [60], hence the 1-dB load-pull contours are more tightly packed for power than for gain.

To obtain even modest bandwidths (e.g., above 5 percent), the use of quadrature balanced stages is advisable (Figure 21). The main benefit of the quadrature balanced configuration is that reflections from the transistors are cancelled by the action of the input and output couplers. An individual device can therefore be deliberately mismatched (e.g., to achieve a power match on the output), yet the quadrature-combined system appears to be well-matched. This configuration also acts as an effective power combiner, so that a given power rating can be achieved using a pair of devices having half of the required power performance. For moderate-bandwidth designs, the lower-power stages are typically designed using a simple single-ended cascade, which in some cases is available as an RFIC. Designs with bandwidths approaching an octave or

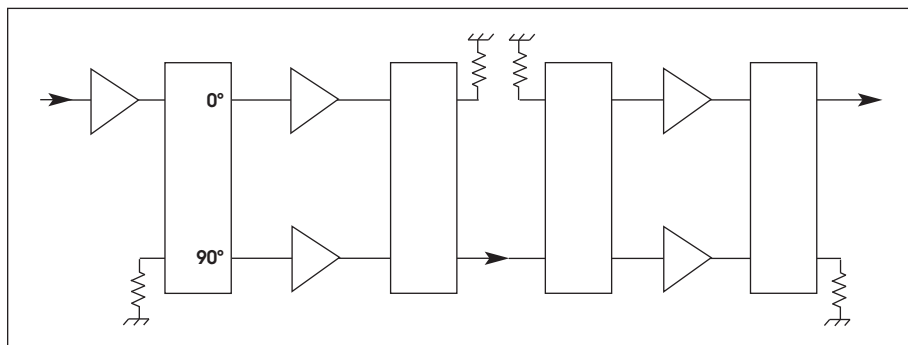


Figure 21 · Amplifier stages with quadrature combiners.

more require the use of quadrature-balanced stages throughout the entire chain.

Simple linear-amplifier chains of this kind have high linearity but only modest efficiency. Single-carrier applications usually operate the final amplifier to about the 1-dB compression point on amplitude modulation peaks. A thus-designed chain in which only the output stage exhibits compression can still deliver an ACPR in the range of about -25 dBc with 50-percent efficiency at PEP.

Two practical problems are frequently encountered in the design of linear PA chains: stability and low gain. Linear, class-A chains are actually more susceptible to oscillation due to their high gain, and single-path chains are especially prone to unstable behavior. Instability can be subdivided into the two distinct categories: Low-frequency oscillation and in-band instability. In-band instability is avoided by designing the individual gain stages to meet the criteria for unconditional stability; i.e., the Rollet k factor [61] must be greater than unity for both in-band and out-of-band frequencies. Meeting this criterion usually requires sacrificing some gain through the use of absorptive elements. Alternatively, the use of quadrature balanced stages provides much greater isolation between individual stages, and the broadband response of the quadrature couplers can eliminate the need to design the transistor

stage itself with $k > 1$. This is another reason for using quadrature coupled stages in the output of the chain.

Large RF power devices typically have very high transconductance, and this can produce low-frequency instability unless great care is taken to terminate both the input and output at low frequencies with impedances for unconditional stability. Because of large separation from the RF band, this is usually a simple matter requiring a few resistors and capacitors.

At X band and higher, the power gain of devices in the 10 W and above category can drop well below 10 dB. To maintain linearity, it may be necessary to use a similarly size device as a driver. Such an architecture clearly has a major negative impact upon the cost and efficiency of the whole chain. In the more extreme cases, it may be advantageous to consider a multi-way power combiner, where 4, 8, or an even greater number of smaller devices are combined. Such an approach also has other advantages, such as soft failure, better thermal management, and phase linearity. However, it typically consumes more board space.

7b. POWER COMBINERS

The need frequently arises to combine the outputs of several individual PAs to achieve the desired transmitter output. Whether to use a number of smaller PAs vs. a single larger PA is one of the most basic decisions in selection of an architec-

ture [60]. Even when larger devices are available, smaller devices often offer higher gain, a lower matching Q factor (wider bandwidth), better phase linearity, and lower cost. Heat dissipation is more readily accomplished with a number of small devices, and a soft-failure mode becomes possible. On the other hand, the increase in parts count, assembly time, and physical size are significant disadvantages to the use of multiple, smaller devices.

Direct connection of multiple PAs is generally impractical as the PAs interact, allowing changes in output from one to cause the load impedance seen by the other to vary. A constant load impedance, hence isolation of one PA from the other, is provided by a hybrid combiner. A hybrid combiner causes the difference between the two PA outputs to be routed to and dissipated in a balancing or “dump” resistor. In the event that one PA fails, the other continues to operate normally, with the transmitter output reduced to one fourth of nominal.

The most common power combiner is the quadrature-hybrid combiner. A 90° phase shift is introduced at input of one PA and also at the output of the other. The benefits of quadrature combining include constant input impedance in spite of variations of input impedances of the individual PAs, cancellation of odd harmonics, and cancellation of backward-IMD (IMD resulting from a signal entering the output port). In addition, the effect of load impedance upon the system output is greatly reduced (e.g., to 1.2 dB for a 3:1 SWR). Maintenance of a nearly constant output occurs because the load impedance presented to one PA decreases when that presented to the other PA increases. As a result, however, device ratings increase and efficiency decreases roughly in proportion to the SWR [65]. Because quadrature combiners are inherently two-terminal devices, they are used in a corporate combining architecture

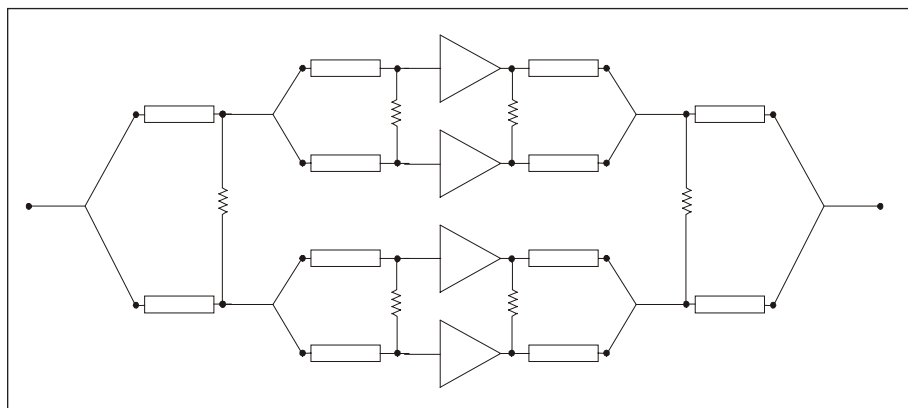


Figure 22 · Multi-section Wilkinson combining architecture.

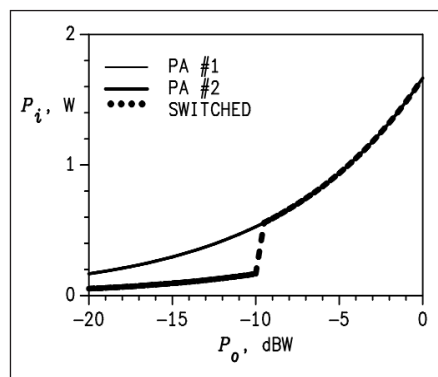


Figure 23 · Power consumption by PAs of different sizes.

(Figure 21). Unfortunately, the physical construction of such couplers poses some problems in a PC-board environment. The very tight coupling between the two quarter-wave transmission lines requires either very fine gaps or a three-dimensional structure. This problem is circumvented by the use of a miniature co-axial cable having a pair of precisely twisted wires to form the coupling section or ready-made, low-cost surface mount 3-dB couplers.

The Wilkinson or in-phase power combiner [62] is often more easily fabricated than a quadrature combiner. In the two-input form (as in each section in Figure 22), the outputs from two quarter-wavelength lines summed into load R_o produce an apparent load impedance of $2R_o$, which is transformed through the lines into at the load impedances R_{PA} seen by the individual PAs. The difference between the two PA outputs is dissipated in a resistor connected across the two inputs. Proper choice of the balancing resistor ($2R_{PA}$) produces a hybrid combiner with good isolation between the two PAs. The Wilkinson concept can be extended to include more than two inputs [63].

Greater bandwidth can be obtained by increasing the number of transforming sections in each signal path. A single-section combiner can have a useful bandwidth of about 20 percent, whereas a two-section version can have a bandwidth close to an octave. In practice, escalating circuit losses generally preclude the use of more than two sections.

All power-combining techniques all suffer from circuit losses as well as mismatch losses. The losses in a simple two-way combiner are typically about 0.5 dB or 10 percent. For a four-way corporate structure, the interconnects typically result in higher losses. Simple open microstrip lines are too lossy for use in combining structures. One technique that offers a good compromise among cost, produceability, and performance, uses suspended stripline. The conductors are etched onto double-sided PC board, interconnected by vias, and then sus-

pended in a machined cavity. Structures of this kind allow high-power 8-way combiners with octave bandwidths and of 0.5 dB.

A wide variety of other approaches to power-combining circuits are possible [62, 64]. Microwave power can also be combined during radiation from multiple antennas through “quasi-optical” techniques [66].

7c. STAGE SWITCHING AND BYPASSING

The power amplifier in a portable transmitter generally operates well below PEP output, as discussed in Section 4 (Part 1). The size of the transistor, quiescent current, and supply voltage are, however, determined by the peak output of the PA. Consequently, a PA with a lower peak output produces low-amplitude signals more efficiently than does a PA with a larger peak output, as illustrated in Figure 23 for class-B PAs with PEP efficiencies of 60 percent. Stage-bypassing and gate-switching techniques [67, 68] reduce power consumption and increase efficiency by switching between large and small amplifiers according to signal level. This process is analogous to selection of supply voltage in a class-G PA, and the average efficiency can be similarly computed [69].

A typical stage-bypassing architecture is shown in Figure 24. For low-power operation, switches SA and SB route the drive signal around the final amplifier.

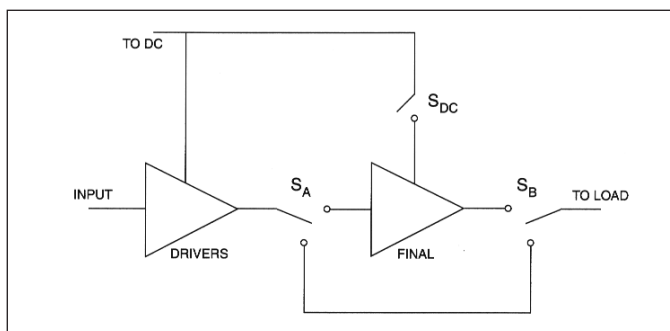


Figure 24 · Stage-bypassing architecture.

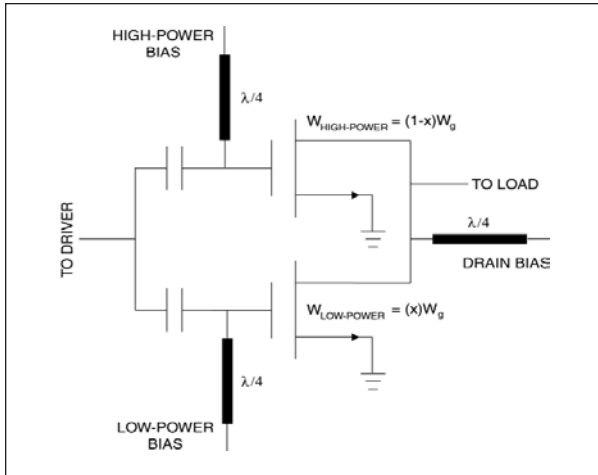


Figure 25 · Adaptive gate switching.

Simultaneously, switch SDC turns-off DC power to the final amplifier. The reduction in power consumption can improve the average efficiency significantly (e.g., from 2.1 to 9.5 percent in [70]). The control signal is based upon the signal envelope and power level (back-off). Avoiding hysteresis effects and distortion due to switching transients are critical issues in implementation.

A PA with adaptive gate switching is shown in Figure 25. The gate width (hence current and power capability) of the upper FET is typically ten to twenty times that of the lower FET. The gate bias for the high-power FET keeps it turned off unless it is needed to support a high-power output. Consequently, the quiescent drain current is reduced to low levels unless actually needed. The advantages of this technique are the absence of loss in the switches required by stage bypassing, and operation of the low-power FET in a more linear region (vs. varying the gate bias of a single large FET). The disadvantage is that the source and load impedances change as the upper FET is switched on and off.

7d. KAHN TECHNIQUE

The Kahn Envelope Elimination and Restoration (EER) technique (Figure 26) combines a highly effi-

cient but nonlinear RF power amplifier (PA) with a highly efficient envelope amplifier to implement a high-efficiency linear RF power amplifier. In its classic form [73], a limiter eliminates the envelope, allowing the constant-amplitude phase modulated carrier to be amplified efficiently by class-C, -D, -E, or -F RF PAs. Amplitude modulation of the final RF PA restores the envelope to the phase-modulated carrier creating an amplified replica of the input signal.

EER is based upon the equivalence of any narrowband signal to simultaneous amplitude (envelope) and phase modulations. In a modern implementation, both the envelope and the phase-modulated carrier are generated by a DSP. In contrast to linear amplifiers, a Kahn-technique transmitter operates with high efficiency over a wide dynamic range and therefore produces a high average efficiency for a wide range of signals and power (back-off) levels. Average efficiencies three to five times those of linear amplifiers have been demonstrated (Figure 27) from HF [74] to L band [75].

Transmitters based upon the Kahn technique generally have excellent linearity because linearity depends upon the modulator rather than RF power transistors. The two most important factors affecting the

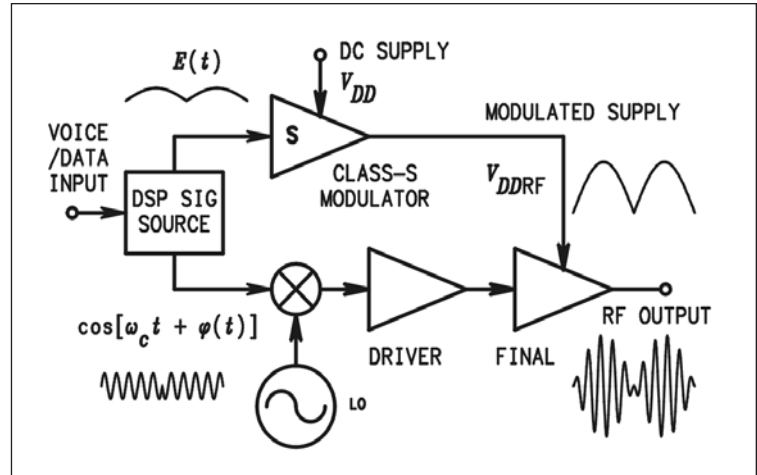


Figure 26 · Kahn-technique transmitter.

linearity are the envelope bandwidth and alignment of the envelope and phase modulations. As a rule of thumb, the envelope bandwidth must be at least twice the RF bandwidth and the misalignment must not exceed one tenth of the inverse of the RF bandwidth [76]. In practice, the drive is not hard-limited as in the classical implementation. Drive power is conserved by allowing the drive to follow the envelope except at low levels. The use of a minimum drive level ensures proper operation of the RF PA at low signal levels where the gain is low [77]. At higher microwave frequencies, the RF power devices exhibit softer saturation characteristics and larger amounts of amplitude-to-phase conversion, necessitating the use of predistortion for good linearity [78].

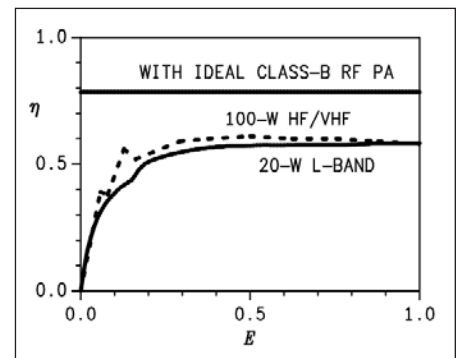


Figure 27 · Efficiency of Kahn-technique transmitters.

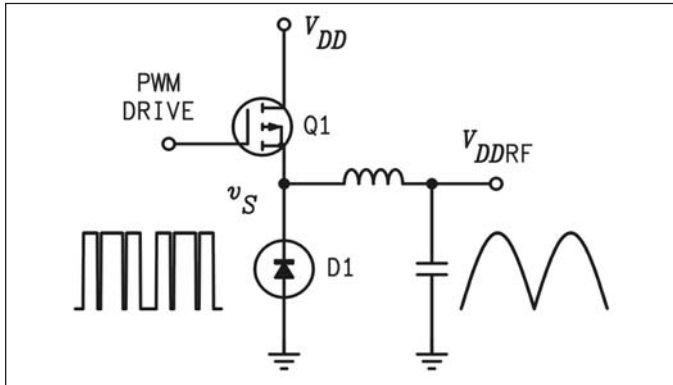


Figure 28 · Class-S modulator.

Class-S Modulator

A class-S modulator (Figure 28) uses a transistor and diode or a pair of transistors act as a two-pole switch to generate a rectangular waveform with a switching frequency several times that of the output signal. The width of pulses is varied in proportion to the instantaneous amplitude of the desired output signal, which is recovered by a low-pass filter. Class S is ideally 100 percent efficient and in practice can have high efficiency over a wide dynamic range. Class-S modulators are typically used as parts of a Kahn-technique transmitter, while class-S amplifiers are becoming popular for the efficient production of audio power in portable equipment. A class-S modulator can be driven by a digital (on/off) signal supplied directly from a DSP, eliminating the need for intermediate conversion to an analog signal.

Selection of the output filter is a compromise between passing the infinite-bandwidth envelope and rejecting FM-like spurious components that are inherent in the PWM process. Typically, the switching frequency must be six to seven times the RF bandwidth. Modulators with switching frequencies of 500 kHz are readily implemented using discrete MOSFETs and off-the-shelf ICs [74], while several MHz can be achieved using MOS ASICs or discrete GaAs devices [75].

Class-G Modulator

A class-G modulator (Figure 29) is a combination of linear series-pass (class-B) amplifiers that operate from different supply voltages. Power is conserved by selecting the one with the lowest useable supply voltage [69] so that the voltage drop across the active device is minimized.

Split-Band Modulator

Most of the power in the envelope resides at lower frequencies; typically 80 percent is in the DC component. The bandwidth of a class-S modulator can therefore be extended by combining it with a linear amplifier. While there are a number of approaches, the highest efficiency

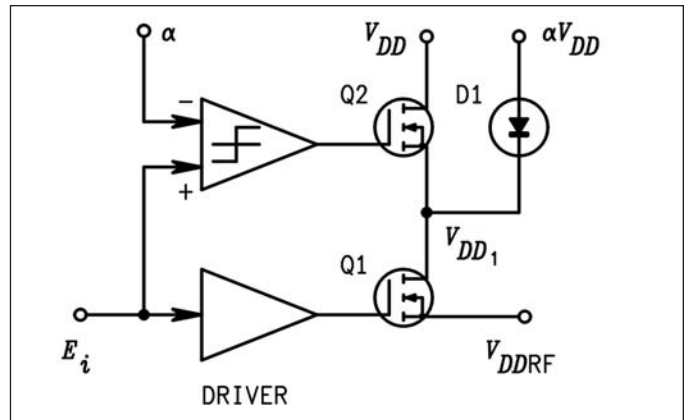


Figure 29 · Class-G modulator.

(typically 90 percent) is achieved by a diplexing combiner. Obtaining a flat frequency response and resistive loads for the two PAs is achieved by splitting the input signals in a DSP that acts as a pair of negative-component filters (Figure 30) [79]. The split-band modulator should make possible Kahn-technique transmitters with RF bandwidths of tens or even hundreds of MHz.

7e. ENVELOPE TRACKING

The envelope-tracking architecture (Figure 31) is similar to that of the Kahn technique. However, the final amplifier operates in a linear mode and the supply voltage is varied dynamically to conserve power [81, 82]. The RF drive contains both amplitude and phase information, and the burden of providing linear amplification lies entirely on the final RF PA. The role of the variable power supply is only to optimize efficiency.

Typically, the envelope is detected and used to control a DC-DC converter. While both buck (step-down) or boost (step-up) converters are used, the latter is more common as it allows operation of the RF PA from a supply voltage higher than the DC-supply voltage. This configuration is

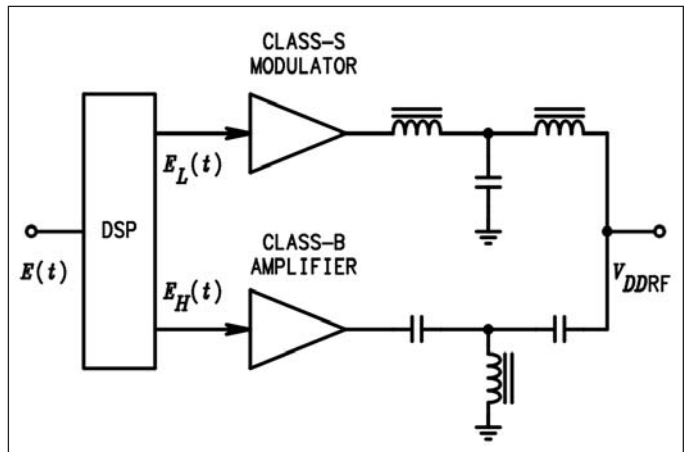


Figure 30 · Split-band modulator.

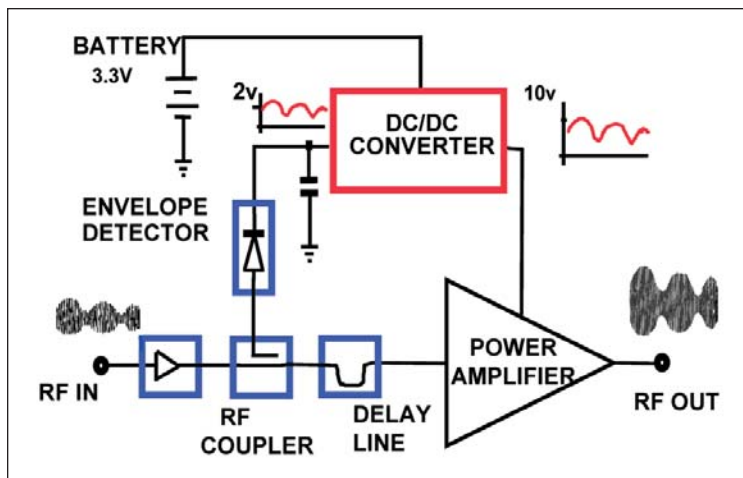


Figure 31 · Envelope-tracking architecture.

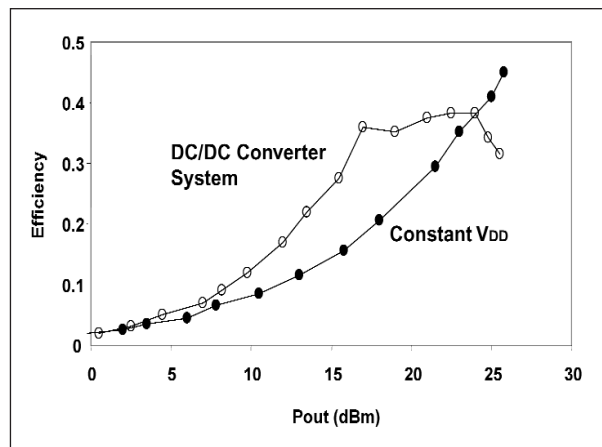


Figure 32 · Efficiency of a GaAs FET envelope-tracking transmitter.

also more amenable to the use of *npn* or *n*-channel transistors for fast switching. The result is a minimum V_{DDRF} corresponding to the DC-supply voltage and tracking of larger envelopes with a fixed “headroom” to ensure linear operation of the RF PA. If the RF PA is operated in class A, its quiescent current can also be varied.

In general, excess power-supply voltage translates to reduced efficiency, rather than output distortion. In principle, perfect tracking of the envelope by the supply voltage preserves the peak efficiency of the RF PA for all output amplitudes, as in the Kahn technique. In practice, efficiency improvement is obtained over a limited range of output power.

A high switching frequency in the DC-DC converter allows both a high modulation bandwidth and the use of smaller inductors and capacitors. The switching devices in the converter can in fact be implemented using the same transistor technology used in the RF PA. Converters with switching frequencies of 10 to 20 MHz have recently been implemented using MOS ASICs [80], GaAs HBTs [83, 84] and RF power MOSFETs [85].

Representative results for an envelope-tracking transmitter based on a GaAs FET power amplifier are shown in Figure 32. The efficiency is lower at high power than that of the conventional amplifier with constant supply voltage due to the inefficiency of the DC-DC converter. However, the efficiency is much higher over a wide range of output power, with the average efficiency approximately 40 percent higher than that of the conventional linear amplifier.

Spurious outputs can be produced by supply-voltage ripple at the switching frequency. The effects of the ripple can be minimized by making the switching frequency sufficiently high or by using an appropriate filter. Variation of the RF PA gain with supply voltage can introduce distortion. Such distortion can, however, be countered by pre-distortion techniques [to be covered in Section 8 (Part 4)].

7f. OUTPHASING

Outphasing was invented during the 1930s as a means of obtaining high-quality AM from vacuum tubes with poor linearity [86] and was used through about 1970 in RCA “Ampliphase” AM-broadcast transmitters. In the 1970s, it came into use at microwave frequencies under the name “LINC” (Linear Amplification using Nonlinear Components) [87].

An outphasing transmitter (Figure 33) produces an amplitude-modulated signal by combining the outputs of two PAs driven with signals of different time-varying phases. Basically, the phase modulation causes the instantaneous vector sum of the two PA outputs to follow the desired signal amplitude (Figure 34). The inverse sine of envelope E phase-modulates the driving signals for the two PAs to produce a transmitter output that is proportional to E . In a modern implementation, a DSP and synthesizer produce the inverse-sine modulations of the driving signals.

Hybrid combining (Figure 33) isolates the PAs from the reactive loads inherent in outphasing, allowing them to see resistive loads at all signal levels. However, both PAs deliver full power all of the time. Consequently, the efficiency of a hybrid-coupled outphasing transmitter varies with the output power (Figure 35), resulting in an average efficiency that is inversely proportional to peak-to-average ratio (as for class A). Recovery of the power from the dump port of the hybrid combiner offers some improvement in the efficiency [88].

The phase of the output current is that of the vector sum of the two PA-output voltages. Direct summation of the out-of-phase signals in a nonhybrid combiner inherently results in reactive load impedances for the power amplifiers [89]. If the reactances are not partially cancelled as in the Chireix technique, the current drawn from the PAs is proportional to the transmitter-output voltage.

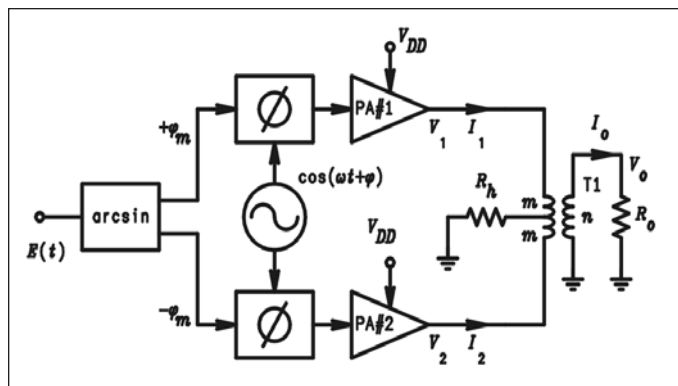


Figure 33 · Hybrid-combined outphasing transmitter.

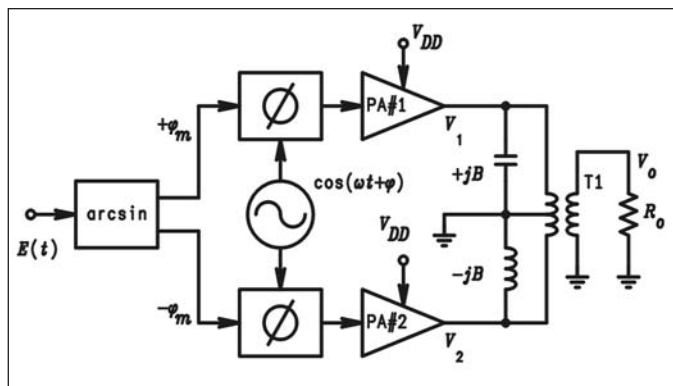


Figure 36 · Chireix-outphasing transmitter.

This results in an efficiency characteristic similar to that of a class-B PA.

The Chireix technique [86] uses shunt reactances on the inputs to the combiner (Figure 36) to tune-out the drain reactances at a particular amplitude, which in turn maximizes the efficiency in the vicinity of that amplitude. The efficiency at high and low amplitudes may be degraded. In the classic Chireix implementation, the shunt reactances maximize the efficiency at the level of the unmodulated carrier in an AM signal and produce good efficiency over the upper 6 dB of the output range. With judicious choice of the shunt susceptances, the average efficiency can be maximized for any given signal [89, 90]. For example, a normalized susceptance of 0.11 peaks the instantaneous efficiency at a somewhat lower amplitude, resulting in an average efficiency of 52.1 percent for an ideal class-B PA and a 10-dB Rayleigh-envelope signal (vs. 28 percent for lin-

ear amplification).

Virtually all microwave outphasing systems in use today are of the hybrid-coupled type. Use of the Chireix technique at microwave frequencies is difficult because microwave PAs do not behave as ideal voltage sources. Simulations suggest that direct (nonhybrid) combining increases both efficiency and distortion [91]. Since outphasing offers a wide bandwidth and the distortion can be mitigated by techniques such as predistortion, directly coupled and Chireix techniques should be fruitful areas for future investigation.

7g. DOHERTY TECHNIQUE

Development of the Doherty technique in 1936 [92] was motivated by the observation that signals with significant amplitude modulation resulted in low average efficiency. The classical Doherty architecture (Figure 37) combines two PAs of equal capacity through quarter-wave-

length lines or networks. The “carrier” (main) PA is biased in class B while the “peaking” (auxiliary) PA is biased in class C. Only the carrier PA is active when the signal amplitude is half or less of the PEP amplitude. Both PAs contribute output power when the signal amplitude is larger than half of the PEP amplitude

Operation of the Doherty system can be understood by dividing it into low-power, medium-power (load-modulation), and peak-power regions [96]. The current and voltage relationships are shown in Figure 38 for ideal transistors and lossless matching networks. In the low-power region, the instantaneous amplitude of the input signal is insufficient to overcome the class-C (negative) bias of the peaking PA, thus the peaking PA remains cut-off and appears as an open-circuit. With the example load impedances shown in Figure 37, the carrier PA sees a 100 ohm load and operates as an ordinary class-B amplifier. The drain voltage increases linearly with output until reaching supply voltage V_{DD} . The instantaneous efficiency at this point (–6 dB from PEP) is therefore the 78.5 percent of the ideal class-B PA.

As the signal amplitude increases into the medium-power region, the carrier PA saturates and the peaking PA becomes active. The additional current I_2 sent to the load by the peaking PA causes the apparent load impedance at V_L to increase above

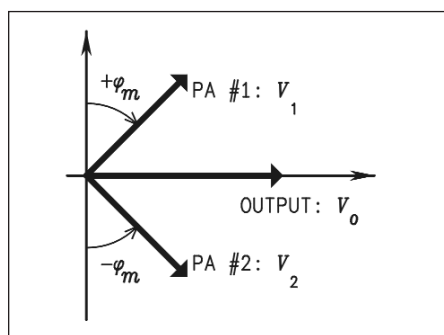


Figure 34 · Signal vectors in outphasing.

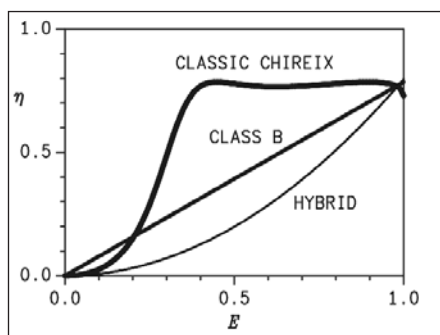


Figure 35 · Efficiency of outphasing transmitters with ideal class-B PAs.

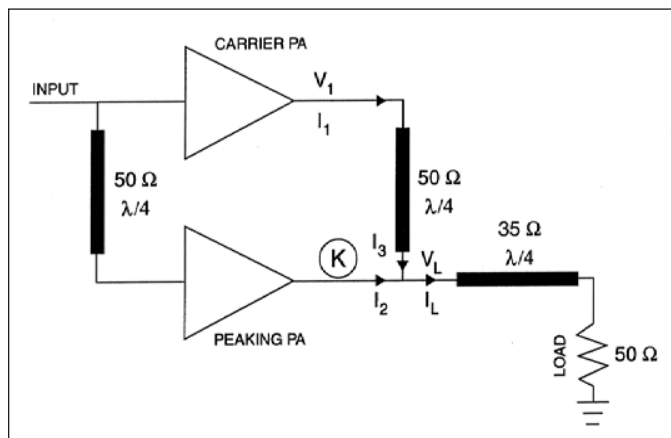


Figure 37 · Doherty transmitter.

the 25 ohms of the low-power region. Transformation through the quarter-wavelength line results in a decrease in the load presented to the carrier PA. The carrier PA remains in saturation and acts as a voltage source. It

operates at peak efficiency, but delivers an increasing amount of power. At PEP output, both PAs see 50-ohm loads and each delivers half of the system output power. The PEP efficiency is that of the class-B PAs.

The resulting instantaneous-efficiency curve is shown in Figure

39. The classical power division ($\alpha = 0.5$) approximately maximizes the average efficiency for full-carrier AM signals, as well as modern single-carrier digital signals. The use of other power-division ratios allows the lower

efficiency peak to be shifted leftward so that the average efficiency is increased for signals with higher peak-to-average ratios. For example, $\alpha = 0.36$ results in a 60 percent average efficiency for a Rayleigh-envelope signal with a 10-dB peak-to-average ratio, which is a factor of 2.1 improvement over class B. Doherty transmitters with unequal power division can be implemented by using different PEP load impedances and different supply voltages in the two PAs [97].

Much recent effort has focused on accommodating non-ideal effects (e.g., nonlinearity, loss, phase shift) into a Doherty architecture [93, 94, 95]. The power consumed by the quiescent current of the peaking amplifier is also a concern. The measured ACPR characteristics of an S-band Doherty transmitter are compared to those of quadrature-combined class-B PAs in Figure 40. The signal is IS-95 forward link with pilot channel, paging channel, and sync-channel. The PAs are based upon 50-W LDMOS transistors. Back-off is varied to trade-off linearity against output. For the specified ACPR of -45 dBc, the average PAE is nearly twice that of the quadrature-combined PAs.

In a modern implementation, DSP can be used to control the drive and bias to the two PAs, for precise control and higher linearity. It is also possible to use three or more stages to keep the instantaneous efficiency relatively high over a larger dynamic range [96, 98]. For ideal class-B PAs, the average efficiency of a three-stage Doherty can be as high as 70 percent for a Rayleigh-envelope signal with 10-dB peak-to-average ratio.

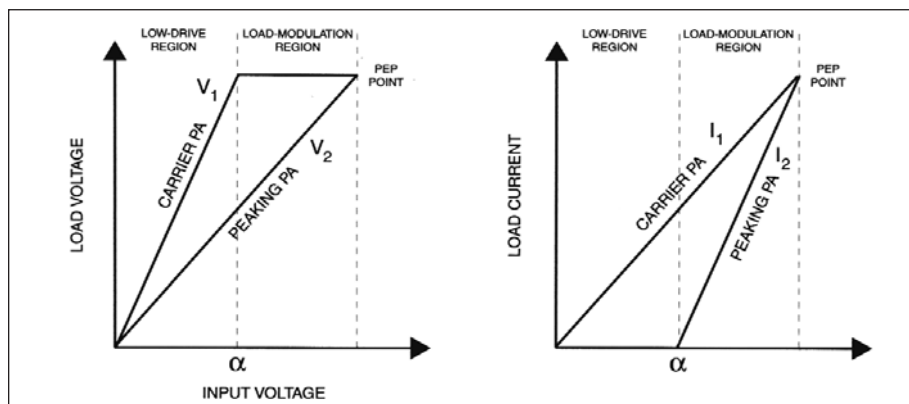


Figure 38 · Ideal voltage and current relationships in Doherty transmitter.

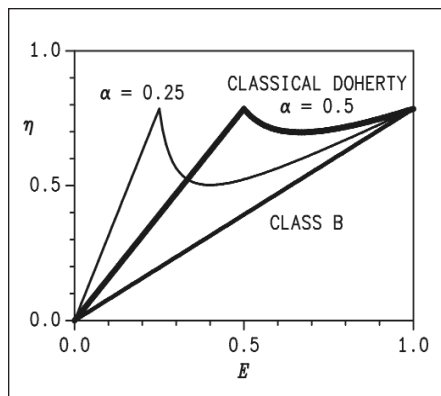


Figure 39 Instantaneous efficiency of the Doherty system with ideal class-B PAs.

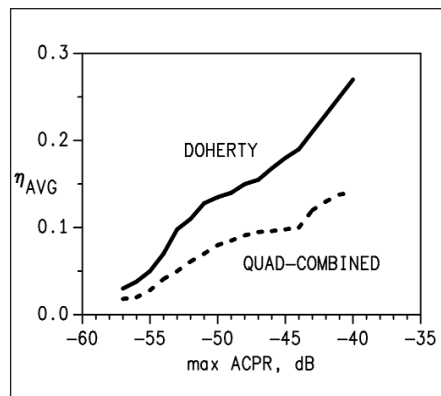


Figure 40 · Measured ACPR performance of an S-band Doherty transmitter.

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Acronyms Used in Part 3

EER	Envelope Elimination and Restoration
AM	Amplitude Modulation
LINC	Linear Amplification with Nonlinear Components

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