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Power Amplifiers and Transmitters for RF and Microwave

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Invited Paper

Abstract—The generation of RF/microwave power is required not only in wireless communications, but also in applications such as jamming, imaging, RF heating, and miniature dc/dc converters. Each application has its own unique requirements for frequency, bandwidth, load, power, efficiency, linearity, and cost. RF power is generated by a wide variety of techniques, implementations, and active devices. Power amplifiers are incorporated into transmitters in a similarly wide variety of architectures, including linear, Kahn, envelope tracking, outphasing, and Doherty. Linearity can be improved through techniques such as feedback, feedforward, and predistortion.

Index Terms—Amplifier, Chireix, class A, class B, class C, class D, class E, class F, Doherty, envelope tracking, feedback, feedforward, HF, Kahn, microwave, outphasing, power, predistortion, transmitter, UHF, VHF.

I. INTRODUCTION

A power amplifier (PA) is a circuit for converting dc-input power into a significant amount of RF/microwave output power. In most cases, a PA is not just a small-signal amplifier driven into saturation. There exists a great variety of different PAs, and most employ techniques beyond simple linear amplification. A transmitter contains one or more PAs, as well as ancillary circuits such as signal generators, frequency converters, modulators, signal processors, linearizers, and power supplies. The classic architecture employs progressively larger PAs to boost a low-level signal to the desired output power. However, a wide variety of different architectures in essence disassemble and then reassemble the signal to permit amplification with higher efficiency and linearity.

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In the early days of wireless communication (1895–mid-1920s), RF power was generated by spark, arc, and alternator techniques. With the advent of the DeForest audion in 1907, the thermoionic vacuum tube offered a means of generating and controlling RF signals, and vacuum-tube PAs were dominant from the late 1920s through the mid-1970s. Discrete solid-state RF-power devices began to appear at the end of the 1960s with the introduction of silicon bipolar transistors such as the 2N6093 [(75-W HF single sideband (SSB))] by RCA. Their dominance in the 1980s brought about the use of lower voltages, higher currents, and relatively low load resistances. The 1990s saw a proliferation of a variety of new solid-state devices including HEMT, pHEMT, HFET, and HBT, using a variety of new materials such as InP, SiC, and GaN. These devices offer amplification to 100 GHz or more and are in many cases grown to order in MMIC form. The combination of digital signal processing (DSP) and microprocessor control allows widespread use of complicated feedback and predistortion techniques to improve efficiency and linearity.

Modern applications are highly varied. Frequencies from VLF through millimeter wave (MMW) are used for communication, navigation, and broadcasting. Output powers vary from 10 mW in short-range unlicensed wireless systems to 1 MW in long-range broadcast transmitters. Almost every conceivable type of modulation is being used in one system or another. PAs and transmitters also find use in systems such as radar, RF heating, plasma generation, laser drivers, magnetic-resonance imaging, and miniature dc/dc converters. No single PA or transmitter technique suits all applications. Many techniques that are now coming into use were devised decades ago, but only recently made possible by advances in signal-processing and control technology.

II. LINEARITY

The need for linearity is one of the principal drivers in the design of modern PAs. Signals such as CW, FM, classical FSK, and GMSK (used in GSM) have constant envelopes (amplitudes) and, therefore, do not require linear amplification. Full-carrier amplitude modulation is best produced by high-level amplitude modulation of the final RF PA. Linear amplification is required when the signal contains both amplitude and phase modulation. Examples include SSB voice,

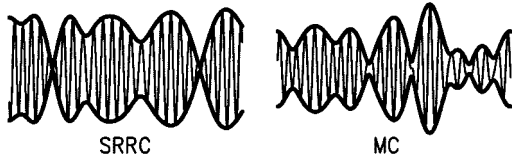


Fig. 1. RF waveforms for SRRC and multicarrier signals.

vestigial-sideband television (both NTSC and HDTV), modern shaped-pulse data modulation (QAM, QPSK, CDMA), and multiple carriers (OFDM).

The requirements for both high data rates and efficient utilization of the increasingly crowded spectrum necessitates the use of shaped data pulses in modern digital signals such as QPSK, QAM, and CDMA. Most systems use raised-cosine shaping, which eliminates intersymbol interference during detection and allows the spectrum to be shaped arbitrarily close to rectangular [1]. This requires the transmission of square-root-raised-cosine (SRRC) data pulses that look much like truncated sinc functions. The resultant modulated carrier (Fig. 1) has simultaneous amplitude and phase modulation with a peak-to-average ratio of 3–6 dB.

Applications such as cellular base-stations, satellite repeaters, and active phased arrays require the simultaneous amplification of multiple signals. The signals can, in general, have different amplitudes, different modulations, and irregular frequency spacing. In a number of applications including HF modems and digital broadcasting, it is more convenient to use a large number of carriers with low data rates than a single carrier with a high data rate. Orthogonal frequency division multiplex (OFDM) [2] employs carriers with the same amplitude and modulation, separated in frequency so that modulation products from one carrier are zero at the frequencies of the other carriers. The resultant composite signal (Fig. 1) has a peak-to-average ratio ξ in the range of 8–13 dB.

Distortion of the amplified signal can be caused by both amplitude nonlinearity (such as a variable gain) or amplitude-to-phase conversion (produced, for example, by a voltage-variable capacitance). The result is splatter into adjacent channels and impairment of detection. Linearity is characterized, measured, and specified by various techniques, depending upon the specific signal and application.

The carrier-to-intermodulation (C/I) ratio, compares the amplitude of the desired output carriers to the intermodulation-distortion (IMD) products [3]. Noise-power ratio (NPR) is the ratio of the notch power to the total signal power when a PA is driven by noise with a spectral notch. Adjacent channel power ratio (ACPR) compares the power in an adjacent channel to that of the signal (Fig. 2). It is currently the most widely used measure of linearity, but defined differently for each application. Error vector magnitude (EVM) is the distance between the desired and actual signal vectors.

III. EFFICIENCY

Efficiency, like linearity, is a critical factor in PA design. Three definitions of efficiency are commonly used. Drain efficiency is defined as the ratio of RF-output power to dc-input

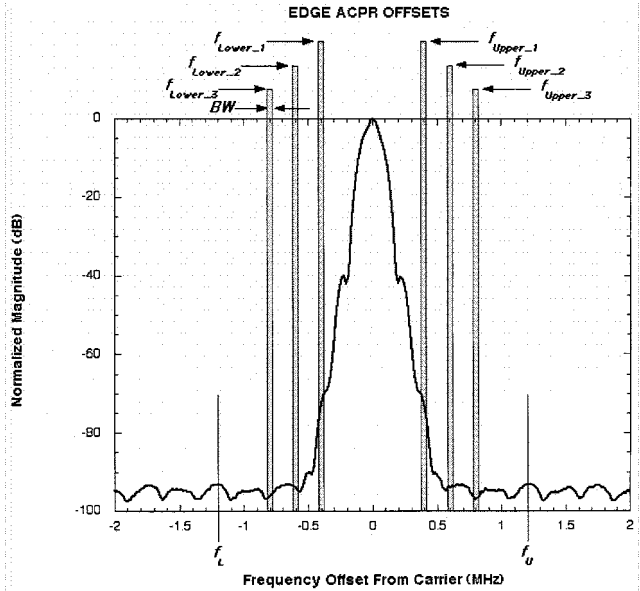


Fig. 2. ACPR offsets and bandwidths.

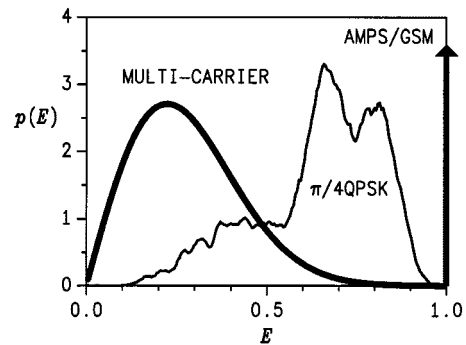


Fig. 3. Envelope PDFs.

power, i.e., $\eta = P_o/P_i$. Power-added efficiency (PAE) incorporates the RF-drive power by subtracting it from the output power, i.e., $(P_o - P_{DR})/P_i$. PAE gives a reasonable indication of PA performance when gain is high; however, it can become negative for low gains. An overall efficiency such as $P_o/(P_i + P_{DR})$ is usable in all situations. This definition can be varied to include driver dc-input power, the power consumed by supporting circuits, and anything else of interest.

The instantaneous efficiency is the efficiency at one specific output level. For most PAs, the instantaneous efficiency is highest at the peak output power (PEP) and decreases as output decreases. When amplifying signals with time-varying amplitudes, a useful measure of performance is the average efficiency, which is defined [4] as the ratio of the average output power to the average dc-input power, i.e., $\eta_{AVG} = P_{oAVG}/P_{iAVG}$.

The probability-density function (PDF) gives the relative amount of time an envelope spends at various amplitudes (Fig. 3). The PDF of an SRRC signal must generally be determined by simulation or measurement. Multiple carriers produce random-phaser sums and, therefore, have Rayleigh-distributed envelopes. The average input and output powers are found by integrating the product of the variable of interest and the PDF of the envelope over the range of the envelope.

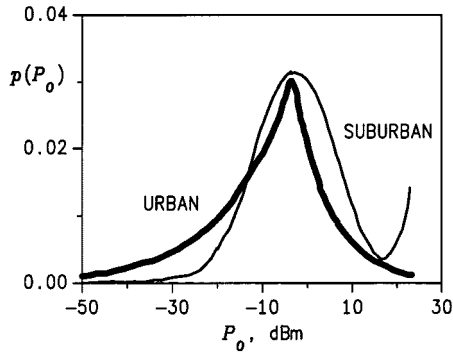


Fig. 4. Power-output PDFs.

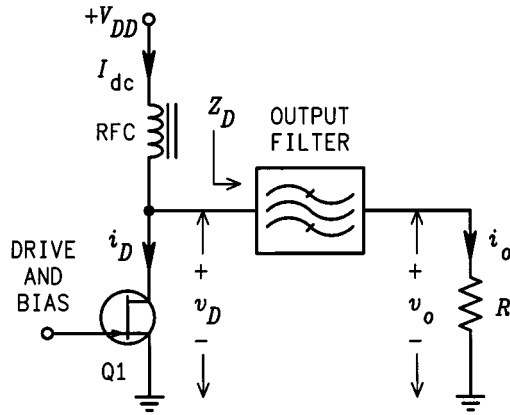


Fig. 5. Single-ended PA.

The need to conserve battery power and to avoid interference to other users operating on the same frequency necessitates the transmission of signals whose peak amplitudes are well below the PEP of the transmitter. Since peak power is needed only in the worst-case links, the “backoff” is typically in the range of 10–20 dB. For a single-carrier mobile transmitter, backoff rather than envelope PDF is dominant in determining the average power consumption and average efficiency. The PDF of the transmitting power (Fig. 4) depends not only upon the distance, but also upon factors such as attenuation by buildings, multipath, and orientation of the mobile antenna [5].

IV. PAs

RF PAs are commonly designated as classes A–F [3]. Classes of operation differ in the method of operation, efficiency, and power-output capability. The “power-output capability” (“transistor utilization factor”) is defined as output power per transistor normalized for peak drain voltage and current of 1 V and 1 A, respectively. The basic single-ended topology (Fig. 5) includes an active device, dc feed, and output filter/matching network. Transformer-coupled and complementary topologies are also used. The drain voltage and current waveforms of selected ideal PAs are shown in Fig. 6.

A. RF-Power Transistors

RF PAs utilize a wide variety of active devices, including bipolar-junction transistors (BJTs), MOSFETs, JFETs (SITs), GaAs MESFETs, HEMTs, pHEMTs, and vacuum tubes [6],

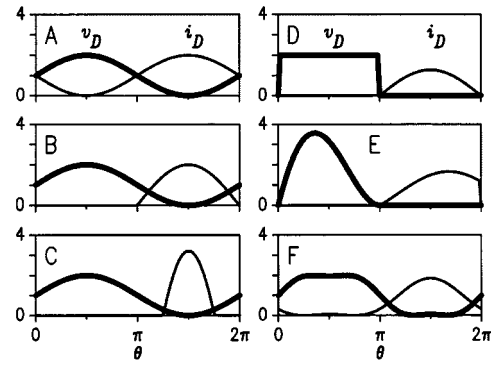


Fig. 6. Waveforms for ideal PAs.

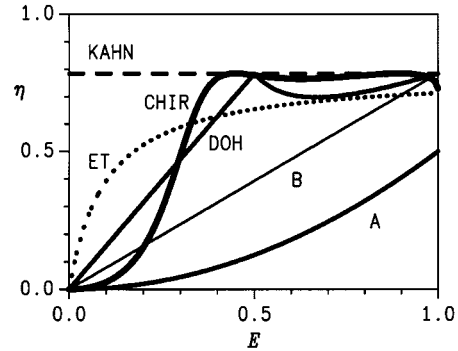


Fig. 7. Efficiency as a function of output (CW).

[7]. The power-output capabilities range from tens of kilowatts for vacuum tubes to hundreds of watts for Si MOSFETs at HF and VHF to hundreds of milliwatts for InP HEMTs at MMW frequencies. Depending upon frequency and power, devices are available in packaged, chip, and MMIC form. Virtually all RF-power transistors are n-p-n or n-channel types because the greater mobility of electrons (versus holes) results in better operation at higher frequencies. While the voltages and currents differ considerably, the basic principles for power amplification are common to all devices.

B. Methods of Amplification

Class A: In class-A amplification, the transistor is in the active region at all times and acts as a current source controlled by the gate drive and bias. The drain-voltage and drain-current waveforms are sinusoids. This results in linear amplification with an output power of $V_{om}^2/2R$, where output voltage V_{om} on load R cannot exceed supply voltage V_{DD} . The dc-power input is constant, hence, the instantaneous efficiency (Fig. 7) is proportional to the power output and reaches 50% at PEP. The average efficiency is inversely proportional to the peak-to-average ratio (e.g., 5% for 10 dB) and backoff (Fig. 8). For amplification of amplitude-modulated signals, the quiescent current can be varied in proportion to the instantaneous signal envelope. The utilization factor is 1/8. Class A offers high linearity, high gain, and operation close to the maximum operating frequency of the transistor.

Class B: The gate bias in a class-B PA is set at the threshold of conduction so the transistor is active half of the time and the drain current is a half-sinusoid. Since the amplitude of the

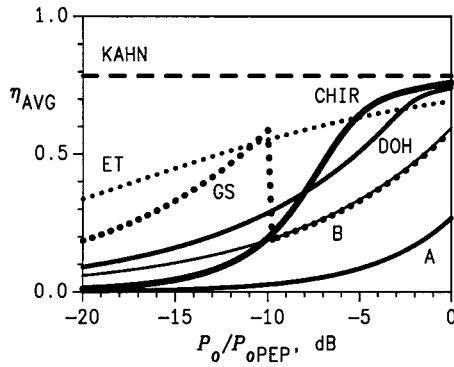


Fig. 8. Efficiency as a function of backoff (SRRC OQPSK).

drain current is proportional to drive amplitude, class B provides linear amplification. The instantaneous efficiency varies linearly with the RF-output voltage and reaches $\pi/4$ (78.5%) at PEP for an ideal PA. For low-level signals, class B is significantly more efficient than class A, and its average efficiency can be several times that of class A at high peak-to-average ratios (e.g., 28% versus 5% for $\xi = 10$ dB). The utilization factor is the same 0.125 of class A. Class B is widely used in broad-band transformer-coupled PAs operating at HF and VHF. It is finding increasing use in microwave PAs, including experimental PAs using complementary devices.

Class C: The gate of a classical (true) class-C PA is biased below threshold so that the transistor is active for less than half of the RF cycle. Linearity is lost, but efficiency can be increased arbitrarily toward 100% by decreasing the conduction angle toward zero. Unfortunately, this causes the output power (utilization factor) to decrease toward zero and the drive power to increase toward infinity. A typical compromise is a conduction angle of 150° and an ideal efficiency of 85%. When driven into saturation, efficiency is stabilized and the output voltage is locked to supply voltage, allowing linear high-level amplitude modulation. Classical class C is widely used in high-power vacuum-tube transmitters, but is generally impractical for solid-state PAs.

Class D: Class-D PAs use two or more transistors as switches to generate square drain-voltage (or current) waveforms. A series-tuned output filter passes only the fundamental-frequency component to the load, resulting in a power outputs of $(8/\pi^2)V_{DD}^2/R$ for the transformer-coupled configuration. Current is drawn only through the transistor that is on, resulting in a 100% efficiency for an ideal PA. The utilization factor ($1/2\pi = 0.159$) is the highest of any PA. If the switching is sufficiently fast, efficiency is not degraded by reactance in the load.

Practical class-D PAs suffer from losses due to saturation, switching speed, and drain capacitance. Finite switching speed causes the transistors to be in their active regions while conducting current. Drain capacitances must be charged and discharged once per RF cycle, resulting in power loss that is proportional to $V_{DD}^{3/2}$ [8] and increases directly with frequency. Class-D PAs with power outputs of 100 W to 1 kW are readily implemented at HF, but are seldom used above lower VHF because of losses associated with the drain capacitance. Recently, however,

experimental class-D PAs have been tested with frequencies of operation as high as 1 GHz [9].

Class E: Class E employs a single transistor operated as a switch [10]. The drain-voltage waveform is the result of the sum of the dc and RF currents charging the drain-shunt capacitance. In optimum class E, the drain voltage drops to zero and has zero slope just as the transistor turns on. The result is an ideal efficiency of 100%, elimination of the losses associated with charging the drain capacitance in class D, reduction of switching losses, and good tolerance of component variation. Optimum class-E operation requires a drain shunt susceptance of $0.1836/R$ and a drain series reactance $1.15R$. It delivers a power output of $0.577V_{DD}^2/R$ for an ideal PA with a utilization factor of 0.098. Variations in load impedance and shunt susceptance cause the PA to deviate from optimum operation, but the degradations in performance are generally no worse than those for classes A and B.

The capability for efficient operation in the presence of significant drain capacitance makes class E useful in a number of applications. High-efficiency HF PAs with power levels to 1 kW can be implemented using low-cost MOSFETs intended for switching rather than RF use [11]. Class E has been used for high-efficiency amplification at frequencies as high as *K*-band [12].

Class F: Class F boosts both efficiency and output by using harmonic resonators in the output network to shape the drain waveforms. The voltage waveform includes one or more odd harmonics and approximates a square wave, while the current includes even harmonics and approximates a half sine wave. Alternately (“inverse class F”), the voltage can approximate a half sine wave and the current a square wave. As the number of harmonics increases, the efficiency of an ideal PA increases from the 50% (class A) toward unity (e.g., 0.707, 0.8165, 0.8656, 0.9045 for two, three, four, and five harmonics, respectively) and the utilization factor increases from $1/8$ toward $1/2\pi$ [13]. The required harmonics arise naturally from nonlinearities and saturation in the transistor. While class F requires a more complex output filter than other PAs, the impedances at the “virtual drain” must be correct at only a few specific frequencies.

A variety of modes of operation in-between classes C, E, and F are possible. The maximum achievable efficiency [13] depends upon the number of harmonics. The utilization factor depends upon the harmonic impedances and is highest for ideal class-F operation.

C. Load-Pull Characterization

RF-power transistors are characterized by breakdown voltages and saturated drain currents. The load impedance for maximum power results in drain voltage and current excursions from near zero to nearly the maximum values. The load impedances corresponding to delivery of a given amount of RF power with a specified maximum drain voltage lie along parallel-resistance lines on the Smith chart. The impedances for a specified maximum current analogously follow a series-resistance line. For an ideal PA, the resultant constant-power contour is football shaped [14].

In a real PA, the “virtual drain” is embedded behind the drain capacitance and bond-wire/package inductance. Transformation of the ideal drain impedance through these elements causes

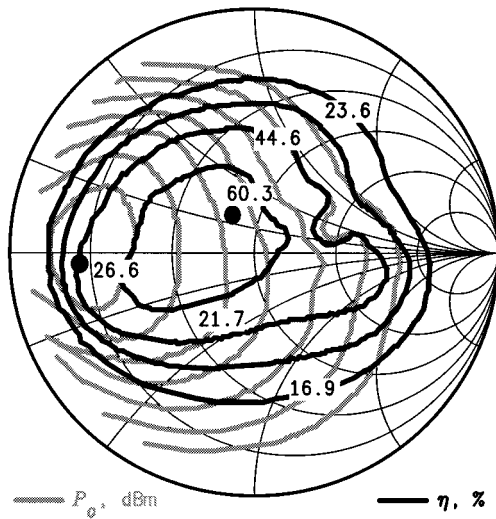


Fig. 9. Example load-pull contours for 0.5-W 836-MHz PA (courtesy Focus Microwave and dBm Engineering).

the constant-power contours to become rotated and distorted. With the addition of second-order effects, the contours become elliptical. As shown in the example of Fig. 9, the power and efficiency contours are not necessarily aligned, nor do maximum power and maximum efficiency necessarily occur for the same load impedance. Sets of such “load-pull” contours are widely used to facilitate design tradeoffs.

Load-pull analyses are generally iterative in nature, as changing one parameter may produce a new set of contours. A variety of different parameters can be plotted during a load-pull analysis, including not only power and efficiency, but also gain, distortion, and stability. Harmonic impedances as well as drive impedances can also be varied. The variable impedance required for load-pull testing can be obtained by mechanical, electrical, or active techniques.

D. Microwave PAs

At microwave frequencies, lumped elements (capacitors, inductors) become unsuitable as tuning components and are used primarily as chokes and bypasses. Matching, tuning, and filtering at microwave frequencies are, therefore, accomplished with distributed (transmission-line) networks. Proper operation of PAs at microwave frequencies is achieved by providing the required drain-load impedance at the fundamental and a number of harmonic frequencies.

Class F: Typically, a transmission line between the drain and load provides the fundamental-frequency drain impedance of the desired value. A stub that is a quarter-wavelength at the harmonic of interest and open at one end provides a short circuit at the opposite end. The stub is placed along the main transmission line at either a quarter or a half-wavelength from the drain to create either an open or short circuit at the drain [15]. The supply voltage is fed to the drain through a half-wavelength line bypassed on the power-supply end or alternately by a lumped-element choke. When multiple stubs are used, the stub for the highest controlled harmonic is placed near the drain. Stubs for lower harmonics are placed progressively further away and their lengths and impedances are adjusted to allow for interactions.

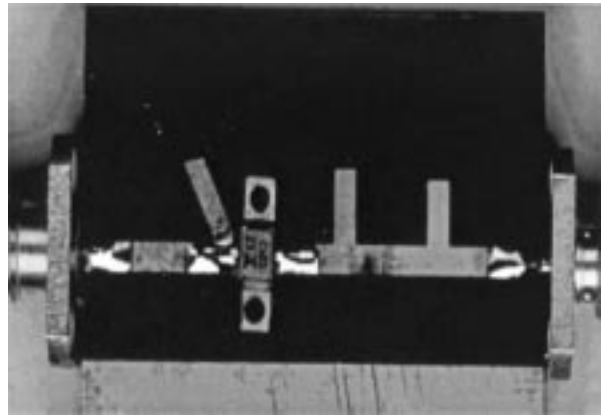


Fig. 10. X-band class-E PA.

Typically, “open” means 3–10 times the fundamental-frequency impedance, and “shorted” means no more 1/10 to 1/3 of the fundamental-frequency impedance [13]. Dielectric resonators can be used in lieu of lumped-element traps.

A wide variety of class-F PAs have been implemented at UHF and microwave frequencies. Generally, only one or two harmonic impedances are controlled. In one X-band PA [16], for example, the output circuit provides a match at the fundamental and a short circuit at the second harmonic. The third-harmonic impedance is high, but not explicitly adjusted to be open. The 3-dB bandwidth of such an output network is about 20%, and the efficiency remains within 10% of its maximum value over a bandwidth of 15%.

Class E: The drain-shunt capacitance and series inductive reactance required for optimum class-E operation result in a drain impedance of $R + j0.725R$ at the fundamental frequency, $-j1.7846R$ at the second harmonic, and proportionately smaller capacitive reactances at higher harmonics. At microwave frequencies, class-E operation is approximated by providing the drain with the fundamental-frequency impedance and preferably one or more of the harmonic impedances [17].

An example of a microwave approximation of class E that provides the correct fundamental and second-harmonic impedances [16], [17] is shown in Fig. 10. The stub immediately to the right of the FET is a quarter-wavelength long at the second harmonic so that the open circuit at its upper end is transformed to a short at its lower end. The line at the drain in combination with drain capacitance and inductance is also a quarter-wavelength to translate the short on its right end to an open at the virtual drain. The remaining lines provide the desired impedance at the fundamental. This circuit uses an FLK052 MESFET to produce 0.68 W at X-band with a drain efficiency of 72% and PAE of 60%.

Methods exist for providing the proper impedances through the fourth harmonic [18]. However, the harmonic impedances are not critical [13], and many variations are, therefore, possible. Since the transistor often has little or no gain at the higher harmonic frequencies, those impedances often have little or no effect upon performance. A single-stub match is often sufficient to provide the desired impedance at the fundamental while simultaneously providing an adequately high impedance at the second harmonic, thus eliminating the need for an extra stub

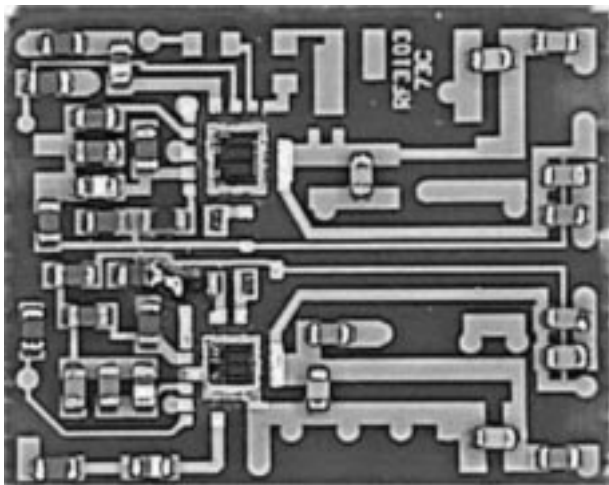


Fig. 11. Internal view of dual-band (GSM/DCS) PA module for cellular-telephone handset (courtesy RF Micro Devices).

and reducing a portion of the losses associated with it. Most microwave class-E amplifiers operate in a suboptimum mode. Demonstrated capabilities range from 16 W with 80% efficiency at UHF (LDMOS) to 100 mW with 60% efficiency at 10 GHz [10], [17], [19].

Comparison: Classes AB and F have essentially the same saturated output power, but class F has about 15% higher efficiency and class E has the highest efficiency [19]. Gain compression occurs at a lower power level for class E than for class F. For a given efficiency, class F produces more power. For the same maximum output power, the third-order IMD products are about 10 dB lower for class F than for class E. Lower power PAs implemented with smaller RF-power devices tend to be more efficient than PAs implemented with larger devices [16].

E. Examples

The PA for a 900-MHz CDMA handset is typically a single GaAs-HBT RFIC that includes a single-ended class-AB PA. Recently developed PA modules also include a silicon control IC that provides the base-bias reference voltage and can be commanded to adjust the output-transistor base bias to optimize efficiency while maintaining acceptably low amplifier distortion over the full ranges of temperature and output power. A typical PA module (Fig. 11) produces 28 dBm (631 mW) at full output with a PAE of 35%–50%.

The thick-film-hybrid PA module shown in Fig. 12 uses four 140-mm LDMOS FETs operating from a 26-V drain supply. The individual PAs have 11-dB power gain and are quadrature-combined to produce a 100-W PEP output at *S*-band. The average output power is 40 W for EDGE and 7 W for CDMA, with an ACPR of -57 dBc for EDGE and -45 dBc for CDMA.

F. MMW PAs

Solid-state PAs for MMW frequencies (30–100 GHz) are predominantly monolithic. Most *K α* -band PAs are based upon pHEMT devices, while most *W*-band PAs are based upon InP HEMTs. Some use is also made of HBTs at the lower MMW frequencies. Class A is used for maximum gain. Typical

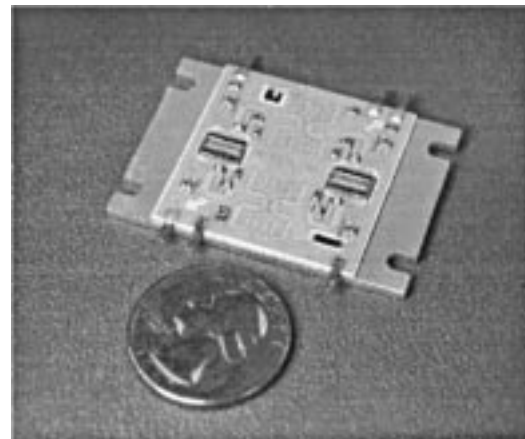


Fig. 12. Thick-film hybrid *S*-band PA amplifier module (courtesy UltraRF).

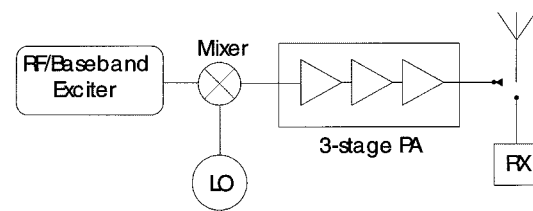


Fig. 13. Linear transmitter architecture.

performance characteristics include 4 W with 30% PAE at *K α* -band, 250 mW with 25% PAE at *Q*-band, and 200 mW with 10% PAE at *W*-band. Devices for operation at MMW are inherently small so large power outputs are obtained by combining the outputs of multiple low PAs in corporate or spatial power combiners.

V. TRANSMITTER ARCHITECTURES

Transmitters use as building blocks not only PAs, 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.

A. 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 (Fig. 13). The amplifier chain consists of cascaded gain stages with power gains in the range of 6–20 dB. If the transmitter must produce an amplitude-modulated or multicarrier signal, each stage must have adequate linearity. This generally requires class-A amplifiers with substantial power backoff 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.

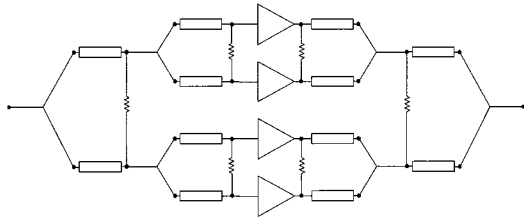


Fig. 14. Corporate architecture with Wilkinson combiners.

B. Power Combiners

Whether to use a number of smaller PAs versus a single larger PA is one of the most basic decisions in selection of an architecture [14]. 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.

In the corporate architecture (Fig. 14), power is split and combined in steps of two. Hybrid combiners isolate the two PAs from each other and allow one to continue operating if the other fails. Quadrature combiners insert a 90° phase shift at the input of one PA and a 90° phase shift at the output of the other. This provides a constant input impedance, 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). The Wilkinson combiner is fabricated using quarter-wavelength lines and can be extended to include more than two inputs or outputs.

C. Stage Bypassing and Gate Switching

Stage-bypassing and gate-switching techniques reduce power consumption and increase efficiency by switching between large and small amplifiers (e.g., the driver) according to peak signal level. This can significantly increase the transmitter efficiency when operating well into backoff, as shown in Fig. 8 ("GS") for ideal class-B PAs. These techniques are particularly effective for mobile handsets that operate over a large dynamic range, and improvement of the average efficiency from 2.1% to 9.5% has been demonstrated [20].

D. Kahn Technique

The Kahn envelope elimination and restoration (EER) technique (Fig. 15) combines a highly efficient, but nonlinear RF PA with a highly efficient envelope amplifier to implement a high-efficiency linear RF PA. In its classic form, a limiter eliminates the envelope, allowing the constant-amplitude phase modulated carrier to be amplified efficiently by class-C, class-D, class-E, or class-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 principle that any narrow-band signal can be produced by simultaneous amplitude (envelope) and phase modulations. In a modern implementation, both the

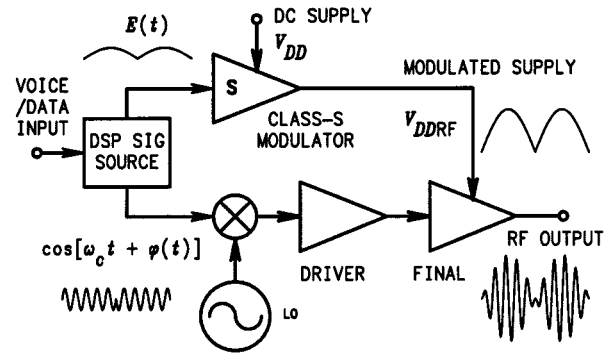


Fig. 15. Kahn-technique transmitter.

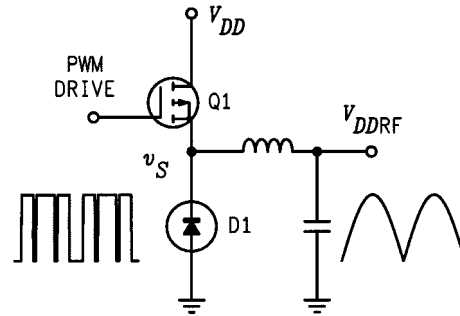


Fig. 16. Class-S modulator.

envelope and 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 (backoff) levels. Average efficiencies three to five times those of linear amplifiers have been demonstrated from HF to L -band [21].

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 linearity are the envelope bandwidth and alignment of the envelope and phase modulations. 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 [22]. In practice, the drive is not hard limited and follows the envelope, except at low levels [23]. 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.

The most widely used high-level modulator is class S (Fig. 16). 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% efficient and, in practice, can have high efficiency over a wide dynamic range. The switching frequency must typically be six times the RF bandwidth. A switching frequencies of 500 kHz is readily achieved with discrete components, and 10 MHz is achievable in IC implementations. Class-G and split-band modulators can be used in wide-band applications.

E. Envelope Tracking

The envelope-tracking architecture is similar to that of the Kahn technique. The supply voltage is varied dynamically to conserve power, but with sufficient excess (“headroom”) to allow the RF PA to operate in a linear mode. The RF drive contains both amplitude and phase information, and the burden of providing linear amplification lies entirely on the final RF PA.

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 also more amenable to the use of n-p-n 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. If the RF PA is operated in class A, its quiescent current can also be varied.

The efficiency is significantly better than that of a linear RF PA operating from a fixed supply voltage, but lower than that of the Kahn technique. The efficiency of a system based upon an ideal converter and class-B RF PA with headroom that is 10% of peak is included in Fig. 7 (“ET”). In practice, power consumption by the converter and other circuits further reduces the efficiency at lower output amplitudes.

A high switching frequency in the dc–dc converter allows both a high modulation bandwidth and the use of smaller inductors and capacitors. Converters with switching frequencies of 10–20 MHz have recently been implemented using MOS ASICs [24], GaAs HBTs [25], and RF-power MOSFETs [26]. The average efficiency for CDMA signals is typically increased from that of a conventional linear amplifier by a factor of 1.5–2.

F. Outphasing

Outphasing was invented by Chireix during the 1930s as a means of obtaining high-quality AM from vacuum tubes with poor linearity 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 (i.e., linear amplification using nonlinear components). An outphasing transmitter (Fig. 17) 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. 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.

Virtually all microwave outphasing systems in use today employ hybrid combiners to isolate the two PAs from each other and to allow 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 (as in a class-A PA), resulting in an average efficiency that is inversely proportional to peak-to-average ratio (as in class A). Recovery of the power from the dump port

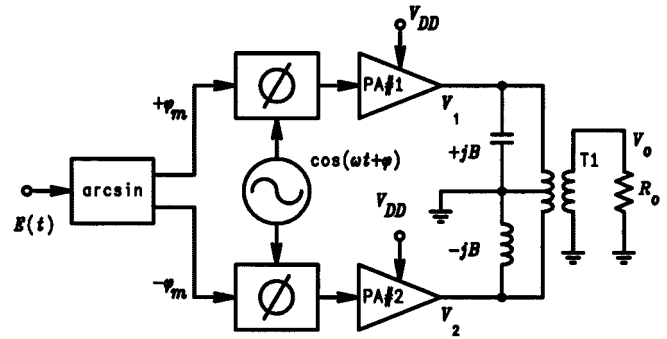


Fig. 17. Chireix-outphasing transmitter.

of the hybrid combiner offers some improvement in the efficiency.

Summation of the out-of-phase signals in a nonhybrid combiner inherently results in variable reactive PA-load impedances. If the combiner is untuned, the current drawn from the PAs is proportional to the transmitter-output voltage, resulting in an efficiency characteristic that varies with signal amplitude, as in a similar class-B PA. The Chireix technique uses shunt reactances on the inputs to the combiner to tune out the drain reactances at a particular amplitude, which, in turn, maximizes the efficiency in the vicinity of that amplitude. In the classic implementation, the efficiency is maximized at the level of the unmodulated AM carrier and remains high over the upper 6 dB of the output range (Fig. 7) and for about 8 dB into backoff (Fig. 8). With judicious choice of the shunt susceptances, the average efficiency can be maximized for any given signal [27]. For example, the average efficiency for a multicarrier signal with a 10-dB peak-to-average ratio can be boosted from the 28% of class B to 52.1%. Simulations suggest that nonhybrid combining of microwave PAs increases both efficiency and distortion [28].

G. Doherty Technique

The classical Doherty architecture (Fig. 18) combines two PAs of equal capacity through quarter-wavelength 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 [29]. In the low-power region, the peaking PA remains cut off and appears as an open circuit. The carrier PA, therefore, sees a 100- Ω load and operates as an ordinary class-B amplifier. The instantaneous efficiency increases linearly with output, reaching the 78.5% of ideal class B at saturation of the carrier PA at –6 dB from transmitter PEP.

As the signal amplitude increases into the medium-power region, 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 the 25 Ω of the low-power region. Transformation through the quarter-wavelength line results in a decrease in the load presented to the carrier PA. The

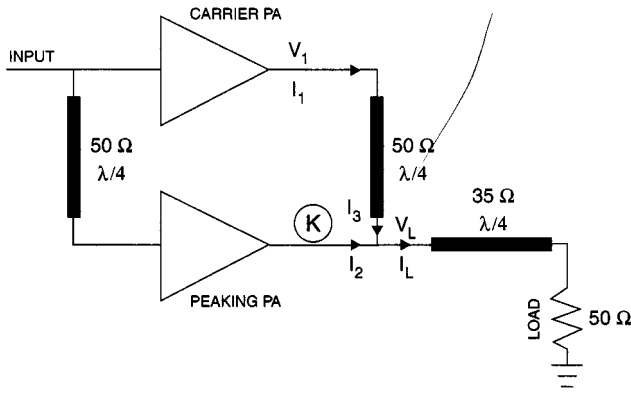


Fig. 18. Doherty transmitter.

carrier PA remains in saturation and acts as a voltage source. It operates at peak efficiency and delivers an increasing amount of power. At PEP output, both PAs see 50-Ω loads and each delivers half of the system output power. The PEP efficiency is ideally the 78.5% of class-B PAs.

The classical power division 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, a transition at 36 percent of PEP voltage results in a 60% 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 [30].

Much recent effort has focused on accommodating nonideal effects (e.g., nonlinearity, loss, phase shift) into a Doherty architecture [31]. In a modern implementation, DSP can be used to control the drive and bias to the two PAs, resulting in more precise control and higher linearity. The power consumed by the quiescent current of the peaking amplifier is also a concern. Nonetheless, S-band Doherty LDMOS transmitters exhibit an average efficiency nearly twice that of a quadrature-combined PA with the same ACPR. It is also possible to use three or more stages to keep the instantaneous efficiency relatively high over a larger dynamic range [32]. The average efficiency of a three-stage Doherty with ideal class-B PAs is 70% for a Rayleigh-envelope signal with 10-dB peak-to-average ratio [29].

VI. LINEARIZATION

Linearization techniques are used both to improve linearity and to allow more efficient, but less linear methods of operation. The three principal types of linearization are feedback, feedforward, and predistortion.

A. Feedback

Feedback linearizes the transmitter by forcing the output to follow the input. It can be applied either directly to the RF amplifier (RF feedback) or indirectly to the modulation (envelope, phase, or I and Q components).

In RF feedback, a portion of the RF-output signal from the amplifier is fed back to and subtracted from the RF-input signal without detection or down-conversion. The delays involved must be small to ensure stability, and the loss of gain at RF is a more significant design issue. The use of RF feedback in discrete circuits is usually restricted to HF and lower VHF frequencies, but it can be applied within MMIC devices well into the microwave region [33].

Envelope feedback reduces distortion associated with amplitude nonlinearity. It can be applied to either a complete transmitter or a single PA [33]. The RF input signal is sampled by a coupler and the envelope of the input sample is detected. The resulting envelope is then fed to one input of a differential amplifier, which subtracts it from a similarly obtained sample of the RF output. The difference signal, representing the error between the input and output envelopes, is used to drive a modulator in the main RF path. This modulator modifies the envelope of the RF signal, which drives the RF PA. The envelope of the resulting output signal is, therefore, linearized to a degree determined by the loop gain of the feedback process. For a VHF BJT amplifier in which amplitude nonlinearity is dominant, two-tone IMD is typically reduced by 10 dB.

The polar loop overcomes the fundamental inability of envelope feedback to correct for AM-PM distortion by adding a phase-locked loop to the envelope feedback system. Envelope detection and phase comparison generally take place at the IF. For a narrow-band VHF PA, the improvement in two-tone IMD is typically around 30 dB. The envelope bandwidth must be at least twice the RF bandwidth, but the phase bandwidth must be at least ten times the RF bandwidth.

The Cartesian-feedback technique overcomes the problems associated with the wide bandwidth of the signal phase by applying modulation feedback in I and Q (Cartesian) components. Since the I and Q components are the natural outputs of a modern DSP, the Cartesian loop is widely used in mobile-radio systems. Two identical feedback processes operate independently on the I and Q channels (Fig. 19). The inputs are applied to differential integrators (in the case of a first-order loop) and the resulting difference (error) signals are quadrature-up-converted to drive the PA. A sample of the output from the PA is attenuated and down-converted in quadrature and synchronously with the up-conversion process. The resulting quadrature feedback signals then form the second inputs to the input differential integrators, completing the two feedback loops. The phase shifter shown in the up-converter local-oscillator path is used to align the phases of the up- and down-conversion processes. The use of Cartesian feedback with a class-C PA amplifying an IS-136 (DAMPS) signal improves the first ACPR by 35 dB and the allows the signal to be produced with an efficiency of 60% [33].

B. Feedforward

The very wide bandwidths (10–100 MHz) required in multi-carrier applications can render feedback and DSP impractical. In such cases, the feedforward technique can be used to reduce distortion by 20–40 dB. In its basic form (Fig. 20), a feedforward amplifier consists of two amplifiers (the main and error amplifiers), directional couplers, delay lines, and loop control

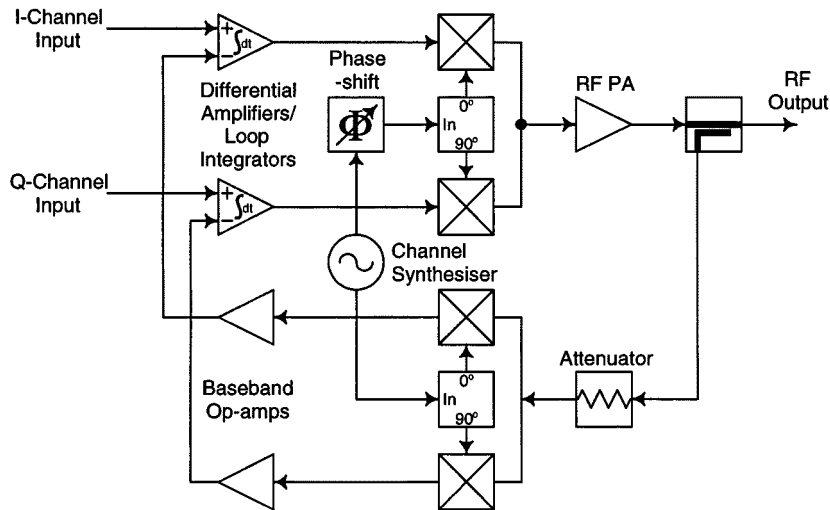


Fig. 19. Cartesian-loop transmitter.

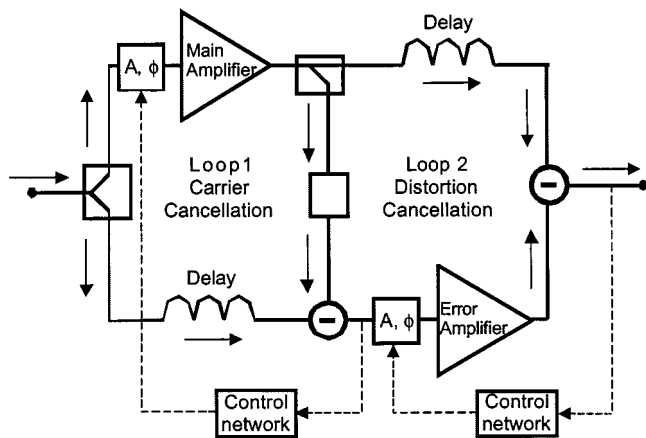


Fig. 20. Feedforward transmitter.

networks [34]. The directional couplers are used for power splitting/combining, and the delay lines ensure operation over a wide bandwidth. Loop-control networks, which consist of amplitude- and phase-shifting networks, maintain signal and distortion cancellation within the various feedforward loops.

The input signal is first split into two paths, with one path going to the high-power main amplifier, while the other signal path goes to a delay element. The output signal from the main amplifier contains both the desired signal and distortion. This signal is sampled and scaled using attenuators before being combined with the delayed portion of the input signal, which is regarded as distortion free. The resulting “error signal” ideally contains only the distortion components in the output of the main amplifier. The error signal is then amplified by the low-power high-linearity error amplifier, and then combined with a delayed version of the main amplifier output. This second combination ideally cancels the distortion components in the main-amplifier output while leaving the desired signal unaltered.

Successful isolation of an error signal and the removal of distortion components depend upon precise signal cancellation over a band of frequencies. For a 30-dB cancellation depth, the amplitudes must be matched within 0.22 dB and the phases

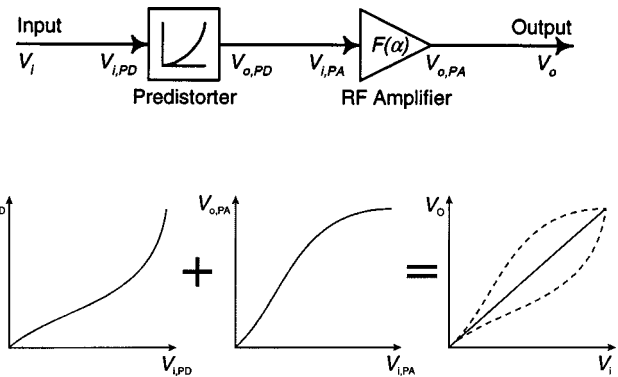


Fig. 21. Predistortion concept.

within 1.2° [34]. For manufactured equipment, realistic values of distortion cancellation are around 25–30 dB. The limiting factor is nearly always the bandwidth over which a given accuracy can be obtained.

The outputs of the main and error amplifiers are typically combined in a directional coupler that both isolates the PAs from each other and provides resistive input impedances. For a typical 10-dB coupling ratio, 90% of the power from the main PA reaches the output. For the same coupling ratio, only 10% of the power from the error amplifier reaches the load, thus the error amplifier must produce ten times the power of the distortion in the main amplifier. The peak-to-average ratio of the error signal is often much higher than that of the desired signal, making amplification of the error signal inherently much less efficient than that of the main signal. As a result, the power consumed by the error amplifier can be a significant fraction (e.g., one-third) of that of the main amplifier. In addition, it may be necessary to operate one or both amplifiers well into backoff to improve linearity. The overall average efficiency of a feedforward transmitter may, therefore, be only 10%–15% for typical multicarrier signals.

Since feedforward is inherently an open-loop process, changes in device characteristics over time, temperature, voltage, and signal level degrade the amplitude and phase matching and, therefore, increase distortion in the transmitter

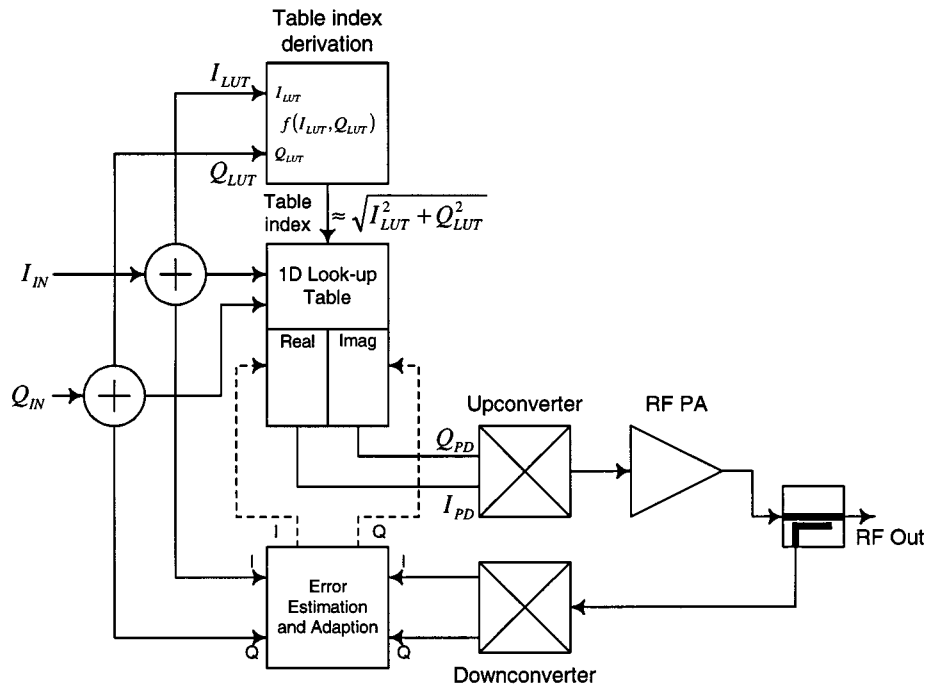


Fig. 22. Constant-gain predistorter.

output. An automatic control scheme continuously adjusts the gain and phase to achieve the best signal cancellation and output linearity. The first step is to use FFT techniques, direct power measurement, or pilot signals to determine how well the loop is balanced. Both digital and analog techniques can be used for loop control and adjustment.

C. Predistortion

The basic concept of a predistortion system (Fig. 21) involves the insertion of a nonlinear element prior to the RF PA such that the combined transfer characteristic of both is linear. Predistortion can be accomplished at either RF or baseband.

An RF predistorter typically creates the expansive predistortion characteristic by subtracting a compressive transfer function (such as that of a diode) from a linear transfer function. Improvements in the ACPR by 10 dB are typical. As with feedforward, the operating bandwidth is limited by the gain and phase flatness of the predistorter itself and of the RF PA. In addition, memory effects in the PA and the predistorter limit the degree of cancellation. Better performance can be achieved with more complex forms of RF predistortion such as Adaptive Parametric Linearization (APL), which is capable of multiorder correction [33]. Most RF-predistortion techniques are capable of broad-band operation with practical operational bandwidths similar to, or greater than, those of feedforward.

D. Digital Predistortion

Digital predistortion techniques exploit the considerable processing power now available from DSP devices, which allows them both to form and to update the required predistortion characteristic. They can operate with analog-baseband, digital-baseband, analog-IF, digital-IF, or analog-RF input signals. Digital-baseband and digital-IF processing are most common. The two

most common types of digital predistorter are termed “mapping predistorters” and “constant-gain predistorters.”

A constant-gain predistorter (Fig. 22) requires only a single-dimensional lookup table, indexed by the signal envelope to generate the expansive predistortion characteristic. It is simple to implement and requires only a modest amount of memory for a given level of performance and adaption time. A mapping predistorter utilizes two lookup tables, each of which is a function of the I and Q components of the input. This type of predistorter is capable of excellent performance. However, it requires a significant storage and/or processing overhead for the lookup tables and their updating mechanism, and has a low speed of convergence.

An example of linearization of a PA with two 3G W-CDMA signals by a digital baseband-input predistorter is shown in Fig. 23. The linearized amplifier meets the required spectral mask with a comfortable margin at all frequency offsets. The noise floor is set by the degree of clipping employed on the waveform, which limits the ACPR improvement obtained. It clearly demonstrates, however, that digital predistortion can be used in broad-band, as well as narrow-band applications. Fig. 24 shows an example of a commercial 3G transmitter with digital predistortion.

VII. EMERGING TECHNIQUES

The ever-increasing demands for more bandwidth, coupled with requirements for both high linearity and high efficiency create ever-increasing challenges in the design of PAs and transmitters. These problems are especially acute in base-station and satellite transmitters, where multiple carriers must be amplified simultaneously, resulting in peak-to-average ratios of 10–13 dB and bandwidths of 30–100 MHz. A number of emerging techniques may prove useful in these applications in the near future.

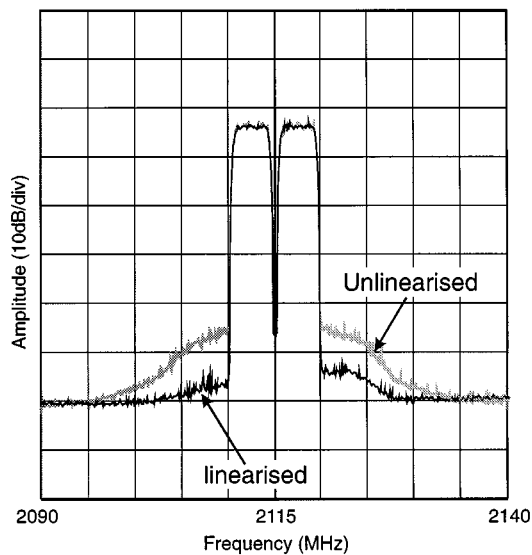


Fig. 23. Linearization of 3G W-CDMA PA signal by digital baseband-input predistorter (courtesy WSI).

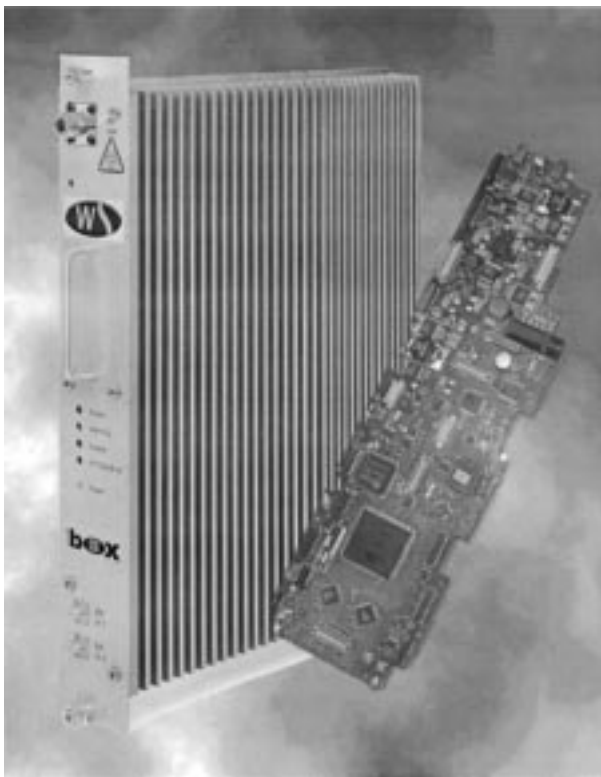


Fig. 24. Multicarrier S-band transmitter with digital predistorter (courtesy WSI).

RF pulsewidth modulation (RF PWM) varies the duty ratio of a class-D RF PA to produce an output signal with a time-varying envelope. The amplitude of the output is proportional to the inverse sine of the pulsewidth. The spurious products associated with PWM are located in the vicinity of the harmonics of the carrier and, therefore, do not limit the modulation bandwidth.

Delta-sigma modulation also directly modulates the carrier produced by a class-D RF PA [35]. The PA is driven at a fixed clock rate (hence, fixed pulsewidth) that is generally higher than

the carrier frequency. The polarity is toggled by a quantizer so that the average carrier amplitude is the desired value. The quantizer forces most of the quantizing noise to fall outside of the signal band where it can be removed by a narrow-band output filter.

Carrier PWM drives the RF PA with bursts of the RF carrier frequency. The width of the bursts is varied in proportion to the instantaneous output amplitude. A narrow-band output filter passes the desired average carrier amplitude and rejects the sidebands associated with the burst frequency. While the process is analogous to that in a class-S modulator, the burst frequency can be 100 MHz or more to accommodate wide-band signals.

Electronic tuning [36] allows frequency agility, matching of unknown and variable loads, and amplitude modulation. Components for electronic tuning include p-i-n diode switches, MEMS switches, MEMS capacitors, semiconductor capacitors, ceramic capacitors (e.g., BST), and bias-controlled inductors. "Load modulation" uses an electronically tuned output filter to vary load impedance and thereby the instantaneous amplitude of the output signal. The modulation bandwidth can be quite wide, as it is limited only by the bias feeds to the tuning components. With judicious choice of the impedance locus used in the modulation process, the efficiency remains high over most of the dynamic range, resulting in average efficiencies three times those of a class-B PA.

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