Polar Transmitters for Wireless Communications

John Groe, Sequoia Communications

ABSTRACT

Future wireless communications will require multimode radio transceivers. This article looks at two possible transmitter architectures: direct upconversion and polar modulation. A brief review of the direct upconversion approach reveals some key limitations. This leads to a detailed study of the polar transmitter focusing on its operation, design challenges, and potential benefits. Its multimode capability and low power consumption are illustrated by a prototype design of a GSM, EDGE, and WCDMA polar transmitter. The efficiency of the polar transmitter is shown as a key enabler of future systems using OFDM modulation with high peak-to-average signals.

INTRODUCTION

Wireless communications are evolving rapidly. The quest for higher data rates is pushing the integration of wideband code-division multiple access (WCDMA) into Global System for Mobile Communications (GSM) and Enhanced Data Rates for GSM Evolution (EDGE) networks. Similarly, the potential of wireless local area networks (WLANs) is spurring the deployment of 802.11a/b/g services. Furthermore, next-generation systems such as fourth generation (4G) and WiMAX promise to add additional capability. It is clear that future wireless devices will need to support an increasing number of modes.

A critical part of the wireless device is the radio transceiver. It consists of a receiver and a transmitter. Wireless receivers typically rely on low-intermediate frequency (IF) and direct conversion architectures that downconvert the received signal directly to baseband frequencies. Wireless transmitters usually employ one of two popular architectures, either direct upconversion or polar modulation. The direct upconversion architecture is widely used in wideband systems, such as WCDMA and WLAN, while polar modulation is almost exclusively employed in narrowband systems like GSM and EDGE.

This article focuses on the development of polar transmitters for multimode applications. It begins by briefly describing the operation and limitations of the direct upconverter architecture. It then describes the operation and design challenges associated with the polar transmitter. Next, it details the performance of a prototype multimode polar transmitter for GSM, EDGE, and WCDMA. The article concludes by summarizing the potential of polar transmitters for other wireless systems.

DIRECT UPCONVERSION TRANSMITTER

The direct upconversion transmitter is based on an I/Q modulator operating at the radio frequency (RF) carrier frequency, as shown in Fig. 1. This convenient technique for generating phase-modulated signals combines a pair of mixers driven by orthogonal local oscillator (LO) signals and produces an output signal centered at the RF carrier frequency. Ideally, the I/Q modulator frequency shifts the spectrum of the input baseband signal to the RF carrier frequency without altering it. In practice, the I/Q modulator is limited by various impairments [1].

Each mixer in the I/Q modulator relies on commutating switches that work best when these devices switch quickly. This reduces noise but produces an output at odd multiples of the RF carrier frequency. (The even multiples are usually cancelled by the circuit structure.) The output spectrum of each mixer is further *cluttered* by mixer distortion and related intermodulation products. These spurious signals are attenuated by bandpass filters in some applications.

The I/Q modulator is also plagued by various circuit and device mismatches. Each mixer produces an output equal to the product of the input signal and RF carrier. Any dc level or offset at the mixer inputs *feeds* a portion of the unmodulated RF carrier to its output, an effect known as carrier leakage. Ideally, the two mixers match exactly with identical inputs and truly orthogonal RF carrier signals. This is difficult at RF, and some phase or amplitude mismatch is expected between the two mixers. Any mismatch degrades orthogonality and causes the inputs to *spill* into each other — a phenomenon known as I/Q leakage. It is possible to minimize carrier and I/Q leakage by using feedback or calibration techniques [2].

The input signals to the I/Q modulator are formed by digital-to-analog (D/A) converters that translate the digitally encoded pulse-shaped data developed in the digital modem. (Although the D/A converters can be located with the modem, the trend is to integrate them with the transmitter.) These converters operate with discrete levels and

0163-6804/07/$20.00 © 2007 IEEE

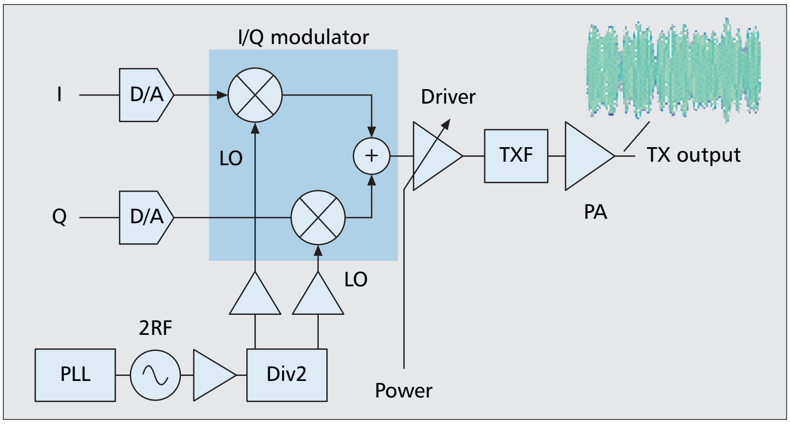
thus introduce quantization noise. Standard Nyquist rate D/A converters produce white noise spread over a bandwidth equal to half the sample rate ($f_s/2$). Oversampled D/A converters using $\Delta\Sigma$ modulation advantageously shape the noise spectrum, pushing most of it toward $f_s/2$. In either case, the output from each D/A converter resembles a sequence of non-return-to-zero pulses. These pulses possess a $\sin(x)/x$ response that affects the signal spectrum and generate replica or alias signals at multiples of f_s. To avoid any passband loss due to the $\sin(x)/x$ response, the data from the digital modem is either appropriately upsampled or precompensated. Any D/A converter nonlinearity — differential or integral — introduces distortion.

The orthogonal RF carrier signals applied to the I/Q modulator are synthesized by a phase-locked loop (PLL) with a voltage controlled oscillator (VCO) operating at twice the carrier frequency and a divide-by-2 circuit. The divide-by-2 circuit produces two signals that are fairly orthogonal provided the VCO's even harmonics are kept low. This is because the even harmonics affect the duty cycle of the VCO signal and consequently the timing between its rising and falling edges. Each of these edges trigger an output from the divide-by-2 circuit, so any timing difference moves the phase offset away from $\pi/2$. Operating the VCO at twice the carrier frequency makes it easy to generate the orthogonal RF carrier signals and also minimizes the potential for injection locking. This phenomenon occurs when the high-power output signal from the PA couples back and affects the frequency stability of the VCO. Even low coupling levels can be a problem in some designs since the PA output signal includes modulation.

The output of the I/Q modulator is the complex transmit signal shown in Fig. 1. Its envelope varies as a result of the pulse-shaping applied to contain its frequency spectrum. The amplifiers following the I/Q modulator must preserve the envelope variation; otherwise, the modulation will spill into nearby frequency channels — an effect known as spectral regrowth. This requires linear driver and power amplifiers that operate inefficiently.

The envelope variation is typically described by its peak-to-average ratio (PAR), which is the ratio of its peak value to its average or root mean square (RMS) value. It is used as a simple guide to estimate the maximum linear power available from the PA. As the signal's PAR value increases, the maximum available linear power from the PA falls. In simple terms, the PA must be *backed off* to operate linearly with higher PAR values. Some typical PAR values are 3.4 dB for EDGE, 3.1 to 6.8 dB for WCDMA, and up to 10 dB or more for OFDM. (All values are for baseband signals. RF signals are 3 dB higher due to the RF carrier.) Alternatively, the transmit signal and its envelope variation can be described by its cubic metric [3]. It measures the signal's likelihood of generating third order distortion and as a result provides a more accurate backoff level. For WCDMA, the cubic metric ranges from 0 to almost 4 dB.

Power amplifiers dissipate more dc power than any other circuit in the transmitter. Their design is difficult and complicated by a variety of factors. At the forefront of these is the desire for high performance and high efficiency. But this becomes more challenging as the PAR and cubic metric of the signal increase. Invariably, the PA is designed to meet performance at peak power levels while minimizing dc power dissipation. The efficiency of a linear PA at its peak output power is approximately 40 percent [4].



■ **Figure 1.** *Direct upconversion transmitter.*

Most wireless systems use some form of power control to minimize interference. The wireless network of base stations generally operates to equalize the power received from different mobiles at each base station. Commands are sent on the downlink control channel to reduce interference and related issues, maximizing system capacity. These power control commands adjust the gain of the transmitter and its output power. The gain is usually adjusted by scaling the D/A converter's reference level or changing the driver's gain, with each approach limited in its range. In practice, the gain distribution in the transmitter depends on the relatively fixed carrier leakage produced by the I/Q modulator and the noise added by the driver and PA.

Power control places new demands on the PA. Its efficiency is now important over a range of output power levels. Using a self-biasing approach lets the RF input power set the operating current of the PA over a small range of levels. To optimize over a wider output range requires dynamic bias control, dc-dc regulators, and switched PA stages. Nevertheless, the efficiency of linear PAs falls off quickly at moderate and lower power levels [4].

POLAR TRANSMITTERS

POLAR CONCEPT

The direct upconversion transmitter and particularly the I/Q modulator use rectangular or Cartesian modulation to form the complex transmit signal. In contrast, polar modulation techniques use magnitude and phase. This makes it possible to apply the two resulting modulation components (phase and magnitude) differently and more efficiently, as shown in Fig. 2. The phase component PM is applied using the PLL while the amplitude component AM is applied at the PA. Since the amplitude of the phase-modulated signal produced by the PLL remains constant, it can be amplified using very efficient, saturated or compressed amplifiers. This dramatically reduces dc power consumption by the transmitter.

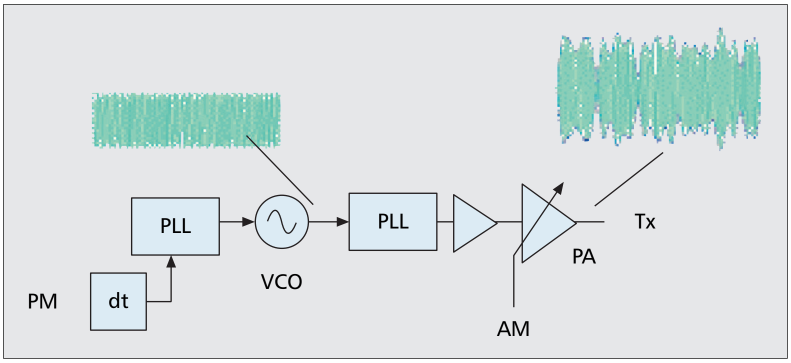


Figure 2. *Concept of the polar transmitter.*

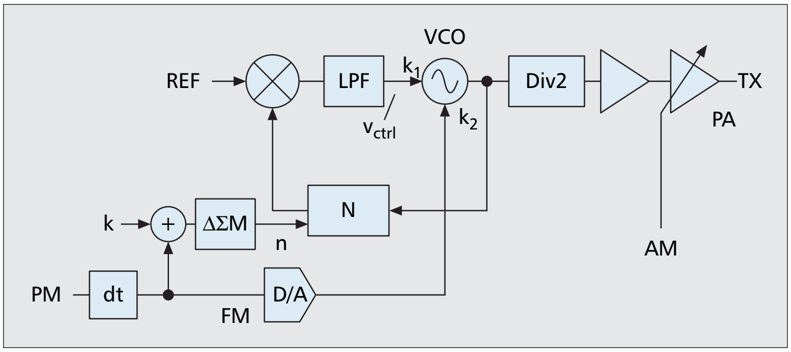


Figure 3. *Fractional-*N *PLL supporting phase modulation.*

Mapping the complex signal trajectory to its phase and amplitude components is a nonlinear process that is complicated as the trajectory approaches the origin. The separate phase and amplitude signals actually *accelerate* near the origin. Furthermore, it is not uncommon for the phase and amplitude signals to abruptly change directions. These effects widen the spectrum of the modulation signals.

PHASE MODULATION

The polar transmitter uses a phase-locked loop to apply phase modulation directly to the synthesized RF carrier and effectively eliminates the I/Q converter with its spurious problems. A detailed view of a fractional-N phase-locked loop supporting phase modulation is shown in Fig. 3. The basic fractional-N PLL consists of a phase/frequency detector, loop filter, VCO, and feedback counter. It operates using feedback to minimize the phase difference between a very accurate crystal reference and the VCO output signal. As such, it produces an output signal at a frequency given by $N \bullet f_{REF}$.

The frequency resolution (or frequency step size) of the PLL is nominally set by f_{REF} since the value of the feedback counter N is naturally an integer. However, it is possible to dramatically decrease the effective frequency step by manipulating the value of N to yield a noninteger average value. For example, if N alternates between 72 and 73, N effectively equals 72.5. Other fractions are created simply by changing the percentage of time the feedback counter operates at each integer value of N. This is the concept of the fractional-N PLL.

Each time the value of N changes, the output phase from the feedback counter jumps. This triggers the phase/frequency detector, which in turn pulses the loop filter accordingly. In fact, the steady-state operation of the fractional-N PLL (where N constantly changes) results in a steady stream of pulses *disturbing* the loop filter. These pulses unavoidably modulate the VCO and generate spurs at multiples of f_{REF}. It is possible to spread out this spur energy by using a ΔΣ modulator to control the value of N. The ΔΣ modulator oversamples the fractional input k to produce a set of values n. By design, the oversampling operation spreads and purposely pushes the quantization noise toward $f_{REF}/2$, where the loop filter can more effectively attenuate it.

The ΔΣ modulator may not completely attenuate spurs. In some cases it may just move the original reference spurs at multiples of f_{REF} to submultiples given by f_{REF}/k. This phenomenon, when present, appears strongest when k is a whole number fraction such as 1/2, 1/4, or 1/8. Fortunately, these fractional spurs can be minimized by careful design of the ΔΣ modulator and either including input dithering or using at least a third order structure.

The phase modulation is applied to the PLL directly at the VCO. In this way, it passes straight to the output. But in order to operate properly, the phase modulation signal must first be differentiated since the control input to the VCO adjusts its output frequency, not its phase. This is easily accomplished using digital techniques and is simply the difference between consecutive PM samples.

The nature of the feedback around the PLL actually restricts any disturbances in the loop such as the modulation applied at the VCO. To counteract this effect, the FM data applied to the VCO is also applied to the feedback counter. This effectively subtracts the frequency modulation applied at the VCO so that the output of the counter represents just the RF carrier frequency. With the modulation applied at two points, this approach is oftentimes referred to as two-point modulation [5].

The differentiation process widens the spectrum of the PM signal. This leads to a high frequency modulation (FM) sample rate to avoid aliasing. In some systems the sample rate is actually set even higher to strategically place the image signals of the FM data outside critical radio bands. As a result, the FM sample rate applied to the VCO is generally higher than f_{REF} (which is approximately the ΔΣ modulator clock rate). Because of this, the FM data applied to the feedback counter must represent multiple samples of the higher-rate FM data applied to the VCO. This is rather straightforward to accomplish since the feedback counter output is discrete. That is, the FM data applied to the feedback counter is simply the average of the FM samples over one period of f_{REF}.

Direct VCO modulation requires near exact control of the VCO's frequency. This is because frequency errors produce phase deviations that accumulate and can potentially grow with time. Fortunately, the PLL's feedback prevents the phase deviations from spiraling out of control, although the PLL's loop bandwidth delays this corrective response. It is interesting to note that within the loop bandwidth of the PLL, the VCO frequency modulation is nearly exact. That's because the VCO output is driven by the PLL's feedback to exactly $N \bullet f_{REF} + FM \bullet f_{REF}$. This is also essentially equal to $K_1 \bullet v_{ctrl} + K_2 \bullet FM$, where K is the VCO's gain and v_{ctrl} is the error signal produced by the phase/frequency detector. Consequently, the error signal v_{ctrl} compensates for any VCO gain errors within the PLL's loop bandwidth.

Outside the loop bandwidth, the feedback and digital correction wane. This makes setting the VCO's gain to its designed value critical and leads to some form of calibration. Since it is impractical to finely adjust the VCO's gain K_2, the calibration instead scales the FM data to achieve a constant frequency step [6]. That is, the parameter α is adjusted to set $K_2(\alpha \bullet FM)$ constant. Figure 4 illustrates the required VCO gain and modulation accuracy for three different loop bandwidth values.

The loop bandwidth of the PLL is a critical design parameter. It not only affects the phase modulator's accuracy, but also the synthesizer's integrated noise, switching time, and stability. In practice, these effects tend to push the loop bandwidth in opposite directions so some compromise results.

The direct FM architecture actually generates the phase-modulated signal at a multiple of the RF carrier frequency. It is then shifted to the RF carrier frequency using a simple divider. This moves the VCO off frequency and reduces injection pulling of the VCO by the high-power PA output.

AMPLITUDE MODULATION

To operate properly, the polar transmitter must also support amplitude modulation (AM). Ideally, the amplitude modulation is applied at the PA as this approach is potentially the most efficient. That is because the sole purpose of the PA is to amplify the transmit signal to a strong enough level to establish a reliable wireless link. As such, it dissipates more dc power than any other circuit in the wireless transceiver. For GSM/EDGE applications, the PA produces up to 2 W of peak RF power.

The best PA efficiency is achieved with a saturated PA that operates as a switch and toggles between the positive regulated voltage and ground, as shown in Fig. 5. Its design pushes the device voltage toward ground during the output current peaks, minimizing power dissipation in the transistor. As a result, a saturated PA can achieve efficiency levels above 60 percent or higher [7]. This type of PA produces strong harmonics and requires an output filter to select the RF carrier.

In practice, the AM is applied to the saturated PA by adjusting the positive regulated voltage. This is usually accomplished with a dc-dc converter or switching regulator with extremely good efficiency. In practice, the regulator's efficiency depends on the switching frequency, which also affects the supply noise and the AM signal's bandwidth. Using this approach produces nearly ideal amplitude control all the way down to low supply voltages (near the transistor's knee voltage). At these low supply voltages, the transistor acts less like a switch, and distortion results.

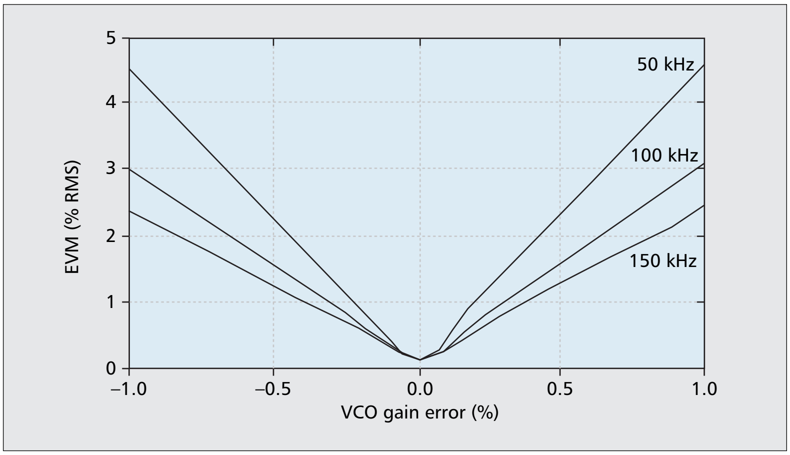


Figure 4. *Sensitivity of modulation accuracy to VCO gain.*

An alternative PA topology using envelope tracking is also shown in Fig. 5; it operates less like a switch and more like an amplifier to reduce distortion. In this PA the dc-dc converter tracks the envelope of the transmit signal to ensure that the amplifier operates linearly. That is, the supply voltage rides above the AM signal. This type of system can adapt if the bandwidth of the dc-dc converter falls short of the AM signal requirements. This approach achieves approximately 50 percent efficiency [8].

Both PA modulation approaches improve efficiency at moderate to high output power levels. Unfortunately, both of these simple approaches also generate too much distortion for most applications. Each requires some form of predistortion or feedback correction. Predistortion techniques compensate the effects of distortion before they occur in the polar transmitter [9]. But this works only when the PA's nonlinear behavior is well known and predictable. If not, then feedback becomes the only way to linearize the system [10].

Amplitude modulation becomes even more

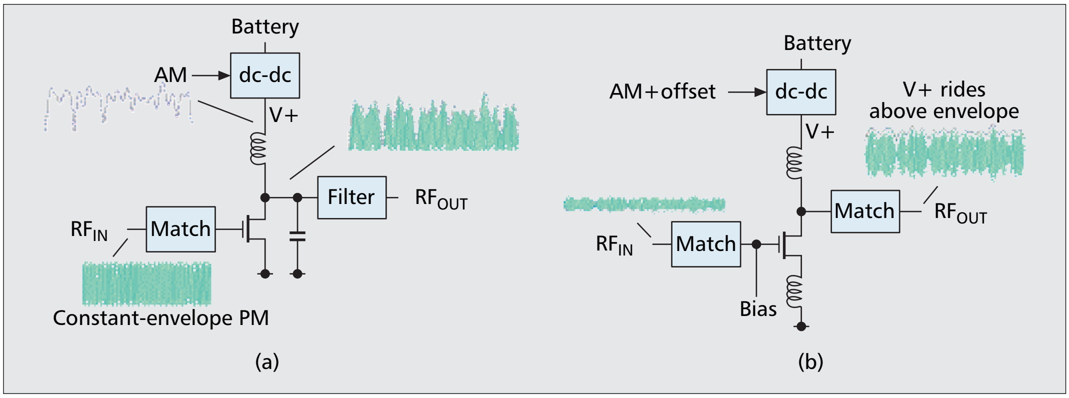


Figure 5. *Amplitude modulation at the PA using a) a saturated PA; b) envelope tracking.*

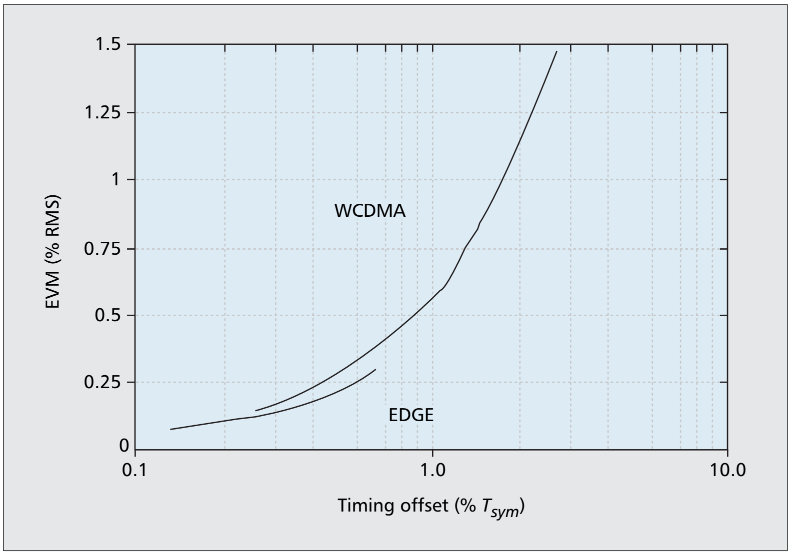


Figure 6. *Timing sensitivity.*

difficult with power control as it extends the AM range. That is because the AM must *slide* with the output power level to be as efficient as possible. Unfortunately, it is impossible for the PA's output to vary enough for some applications, such as WCDMA, so AM control needs to be extended to the driver amplifier. This is especially difficult with the driver and PA amplifiers located on different integrated circuits.

COMBINING PHASE AND AMPLITUDE

An ideal polar modulator combines the amplitude and phase components to describe the exact trajectory of the complex transmit signal. To accomplish this requires near exact alignment of these signals — a task made difficult because these signals are applied at separate points. Any timing offset causes the error vector magnitude (EVM) and emissions to grow. In practice, the required accuracy depends on the data rate (symbol rate or chip rate in spread spectrum systems) of the baseband transmit signal, as illustrated in Fig. 6. Note that a one percent timing offset corresponds to 3.6° phase shift. This is only 2.5 ns in a high-data-rate system like WCDMA.

Any filtering by the PLL or the AM network introduces group delay that must be removed or compensated. Fortunately, this is strictly an analog problem affecting only the AM and FM signals. The RF delay from the VCO to the PA can be ignored since it is necessarily small.

A MULTI-MODE POLAR TRANSMITTER

The most widely used application of polar modulation is GSM. This system employs constant envelope modulation with Gaussian minimum shift keying (GMSK). As a result, the trajectory of the complex signal lies on a unit circle, and the modulation can be described entirely by its phase component.

The EDGE system triples the GSM data rate using differentially encoded 3π/8 8-phase shift keying (PSK) modulation. The system introduces AM so that the transmit signal occupies the same 270 kHz bandwidth as GSM. This and other similarities provide ample motivation to extend the GSM polar transmitter to EDGE.

WCDMA — the evolution path for GSM and EDGE — provides high-speed data by bundling multiple data channels and using spread-spectrum hybrid PSK (HPSK) modulation. The multiple channels create a set of superimposed quaternary PSK (QPSK) patterns with different gains resulting from different spreading factors. A root raised cosine filter limits symbol smearing and restricts the transmit signal bandwidth to 3.84 MHz. The complex signal trajectory for each of these three modulation formats is shown in Fig. 7.

GSM, EDGE, and WCDMA present different challenges to the design of the transmitter. GSM and EDGE systems need to operate with excellent phase linearity, low phase noise, and high efficiency, while WCDMA systems must function accurately over very wide bandwidths and very wide amplitude ranges. Figure 8 illustrates the expansion of the WCDMA transmit signal into its separate frequency and amplitude components. The FM spectrum appears fairly flat.

A prototype of a polar transmitter for GSM, EDGE, and WCDMA has been developed. It uses two-point frequency modulation with AM at the driver amplifier to satisfy the different requirements of the three systems. It also makes use of some innovative signal processing to contain the frequency spectrum of the AM and FM signals. The prototype includes all bias circuits and integrates the output matching network to provide 6 dBm peak output power. (The design includes an on-chip back termination resistor to ease output matching even though it reduces driver efficiency.) It also provides over 80 dB of amplitude range and power control.

The highlights of the polar transmitter are listed in Table 1. It automatically adjusts for each mode to deliver outstanding performance. The EVM results show that the synthesized

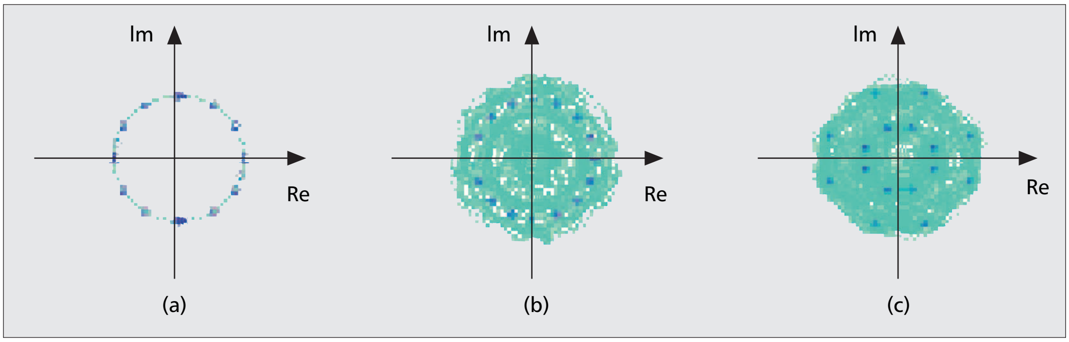


Figure 7. *Complex signal trajectory for a) GSM; b) EDGE; c) WCDMA.*

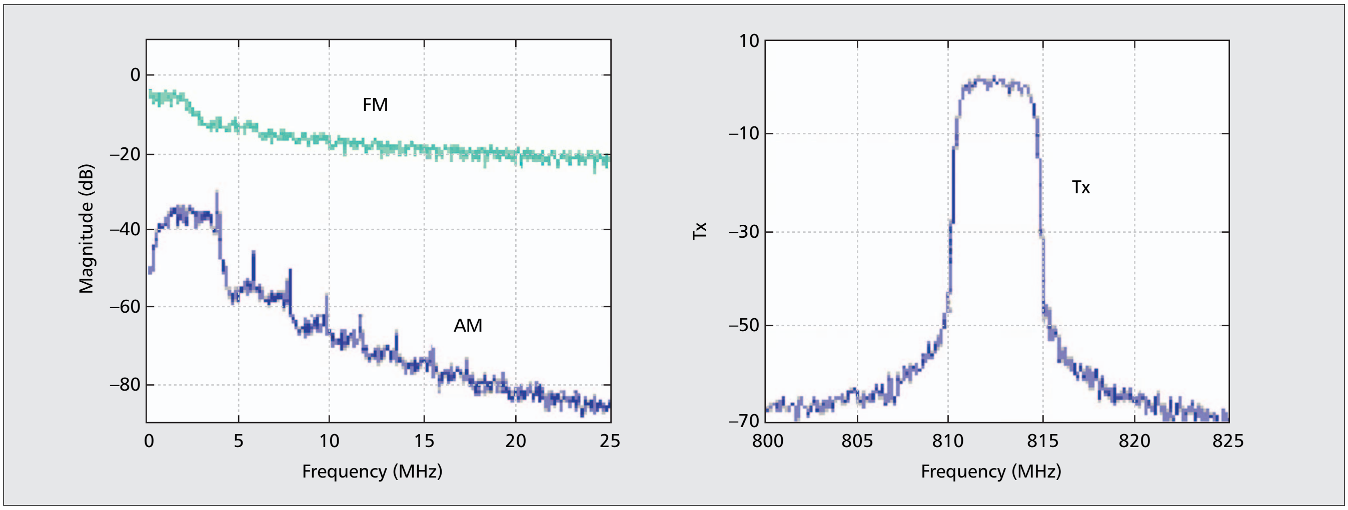


Figure 8. *Expansion of modulation signals that combine to form the WCDMA transmit signal.*

transmit signal closely follows the baseband complex signal trajectory. The emission levels meet system targets, and indicate adequate AM and FM signal bandwidth. The current consumption for the EDGE and WCDMA modes rivals the efficiency of the GSM polar approach.

SUMMARY

The polar transmitter provides the potential for very efficient multimode wireless transmitters. It provides a single architecture for different systems that eliminates RF mixers with their associated spurious and leakage problems. It also dramatically boosts efficiency, extending battery life and leading to higher output power capability.

Extending the polar transmitter architecture to OFDM, 4G, and WiMAX systems is daunting. The wider spectrums and tighter accuracy requirements will place greater demands on phase and amplitude modulation systems. But these systems will also generate signals with much higher PAR values, severely taxing linear power amplifiers and making them virtually useless. Polar modulation uniquely solves this problem.

The multimode polar transmitter prototype illustrates the potential of this architecture and its application to a wideband system (WCDMA). Extending this architecture presents a tremendous opportunity that in some ways enables future systems.

Parameter	GSM	EDGE	WCDMA
RMS error	0.8°	1.7%	3.8%
Emissions			
400 kHz[1]	−65 dBc	−63 dBc	
1800 kHz[2]	−81 dBc	−72 dBc	
2.5 MHz[1]			−61 dBc
8.5 MHz[3]			−64 dBc
20 MHz	−164 dBc/Hz	−160 dBc/Hz	
Carrier leakage	−66 dBc	−63 dBc	−53 dBc
I/Q leakage	−60 dBc	−57 dBc	−53 dBc
Current	69.5 mA	74.3 mA	71.7 mA

[1] Measured in 30 kHz bandwidth

[2] Measured in 100 kHz bandwidth

[3] Measured in 1 MHz bandwidth

Table 1. *Performance of multi-mode polar transmitter.*

REFERENCES

[1] J. Groe and L. Larson, *CDMA Mobile Radio Design*, Artech House, 2000.

[2] G. Brenna *et al.*, "A 2-GHz Carrier Leakage Calibrated Direct-Conversion WCDMA Transmitter in 0.13-mm CMOS," *IEEE J. Solid-State Circuits*, Aug. 2004, pp. 1253–62.

[3] "Comparison of PAR and Cubic Metric for Power De-Rating," TSG RAN WG1 #37, May 2004.

[4] G. Zhang *et al.*, "WCDMA PCS Handset Front End Module," *IEEE MTT-S Microwave Symp.*, June 2006, pp. 304–07.

[5] C. Durdodt *et al.*, "A Low-IF Rx Two-Point DS-Modulation Tx CMOS Single-Chip Bluetooth Solution," *IEEE Trans. Microwave Theory and Techniques*, Sept. 2001, pp. 1531–37.

[6] J. Groe, "Highly Linear Phase Modulation," US patent 10/420,952.

[7] J Gering *et al.*, "A Quad-band InGaP HBT Power Amplifier Module with 60% EGSM Power Added Efficiency and 35:1 Open Loop Ruggedness," *2003 IEEE Topical Wksp. Power Amplifiers for Wireless Commun.*, Sept. 2003.

[8] D. Kimball *et al.*, "High-Efficiency Envelope-Tracking WCDMA Base-Station Amplifier using GaN HFETs," *IEEE Trans. Microwave Theory and Techniques*, Nov., 2006, pp. 3848–56.

[9] E. McCune, "High-Efficiency, Multi-Mode, Multi-Band Terminal Power Amplifiers," *IEEE Microwave*, Mar. 2005, pp. 44–55.

[10] T. Sowlati *et al.*, "Quad-band GSM/GPRS/EDGE Polar Loop Transmitter," *IEEE J. Solid-State Circuits*, Dec. 2004, pp. 2179–89.

BIOGRAPHY

JOHN GROE [SM] (jgroe@sequoiacommunications.com) received a B.S.E.E. degree from California State University at Long Beach in 1984 and an M.S.E.E. degree from the University of Southern California in 1990. In 1980 he joined TRW where he designed PLLs, data converters, and RF integrated circuits. From 1990 to 2000 he designed RF systems for various wireless communication devices at PCSI and Nokia. In 2000 he founded Sequoia Communications, where he is currently the chief technology officer. He is a co-author of *CDMA Mobile Radio Design* (Artech House, 2000). He has been granted 24 U.S. patents.