Linear Power Control Of GSM Amplifier Power

This technique of controlling PA module output power has advantages in dynamic range and accuracy compared to traditional current-sensing and voltage-sensing power-control methods.

Power-control methods for integrated Global System for Mobile Communications (GSM) power-amplifier (PA) modules are many. New methods include approaches based on sensing current and sensing voltage. But the best performance can be achieved with a linear-in-dB technique that provides an accurate and predictable method of controlling PA output power. In comparing these different approaches, the usual measures include output power and power-added efficiency (PAE). But other areas to consider include the PA's output-power stability as functions of temperature, frequency, load VSWR, and battery power; the dynamic range of the power-ramping function; the ease of calibration; trade-offs between PAE/battery current and output power; and the impact of the power-control circuitry on efficiency.

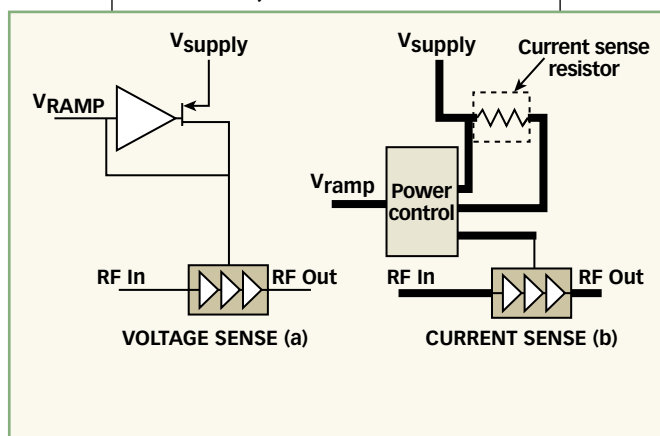The voltage-sensing **(Fig. 1a)** and current-sensing **(Fig. 1b)** methods are often used for PA output-power control. In the voltage-sensing method, a high-speed control loop is incorporated to regulate the collector voltage of the amplifier while the PA stages are held at a constant bias. By regulating the power, the stages are held in saturation across all power levels. As the required output power is decreased from full power down to 0 dBm, the collector voltage is also decreased. The current-sensing method senses the current supplied to the PA through the power supply. A complementary-metal-oxide-semiconductor (CMOS) controller controls the base voltage of the field-effect transistor (FET) with an error voltage generated by applying the ramp voltage, from the DAC, and the current measured to an error amplifier. These are both *indirect closed loop* methods as there is no direct measurement of the PA's output power.

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1. These simple diagrams illustrate the voltage-sensing (a) and current-sensing (b) methods of PA output-power control.

Figure 2 shows a simplified block diagram of a more accurate way to control power by directly detecting the RF power from the amplifier. The module comprises a PA, a silicon power controller, and a number of passive components. The PA output power is controlled by adjusting the bias on the bases of the power transistors. The output power is *regulated* by a classic automatic-gain-control (AGC) loop. This measures the actual output power of the PA by coupling off a small proportion of the output power using a directional coupler. The sensed power level is applied to a logarithmic amplifier (logamp). The logamp measures the power and compares it to a set point, V_{SET}. If there is a difference between the measured power and the corresponding V_{SET}, an error amplifier adjusts the voltage to the bias controller, V_{APC}.

Neither the current-sensing method nor the voltage-sensing technique have any feedback from the output of the PA. Thus, power control is nonlinear in both cases (see Eqs. 1 and 2). The output-power control function in both methods has a nonconstant slope and smaller dynamic range, making ramping to low power levels difficult.

SEE EQ. 1 IN BOX BELOW

SEE EQ. 2 IN BOX BELOW

Figure 3 shows that the output power for the RF power-detection method is linear-in-dB over all GSM power levels. This relationship is stable over temperature, and within each frequency band. The part-to-part transfer functions of the DAC that drives the controller, and the logarithmic detector both vary, however, so it is necessary to calibrate the circuit to obtain precise output power. The straightforward linear-in-dB relationship between the output power and V_{SET}, allows for a single two-point calibration in each frequency band. The procedure simply involves applying a V_{SET} that results in an output-power level close to full specified power (e.g., +33 dBm for GSM). Once achieved, the digital-to-analog converter (DAC) code is noted for
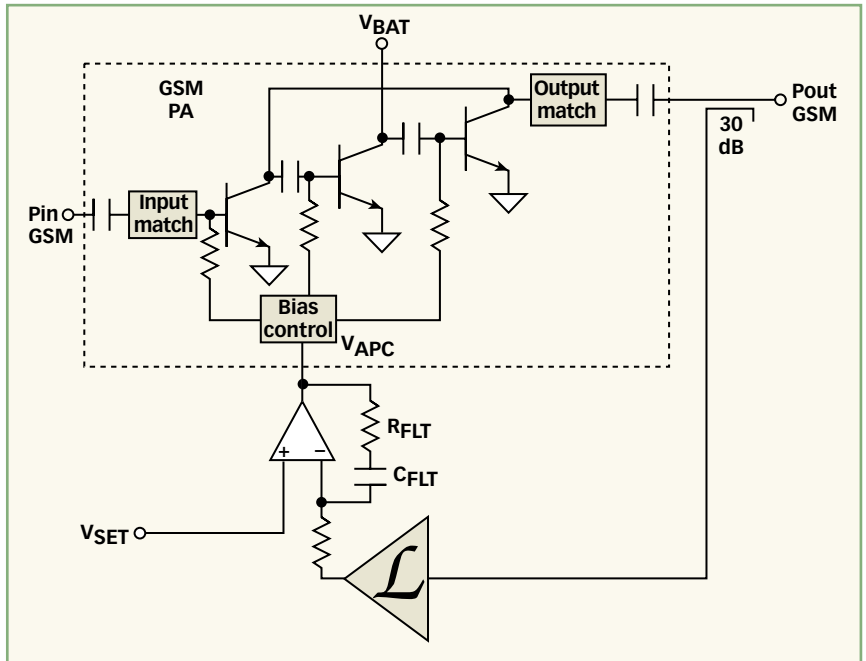
2. **In the RF power-detection method of PA output-power control, the output power versus control voltage (V_{SET}) transfer function is linear-in-dB over a 40-dB dynamic range**.

that power level, which is denoted P_{HIGH}. Then, a V_{SET} is applied that results in an output power close to the minimum output-power level for that standard (e.g., +5 dBm for GSM). Similarly, the DAC code is noted for this power level, which is denoted P_{LOW}. With these four data points, the power output versus V_{SET} can be calculated. This translates into a sim-ple two-point calibration and straight-line approximation.

SEE EQ. 3a IN BOX BELOW

SEE EQ. 3b IN BOX BELOW

Once the slope and the intercept are known, the required code for any trans-

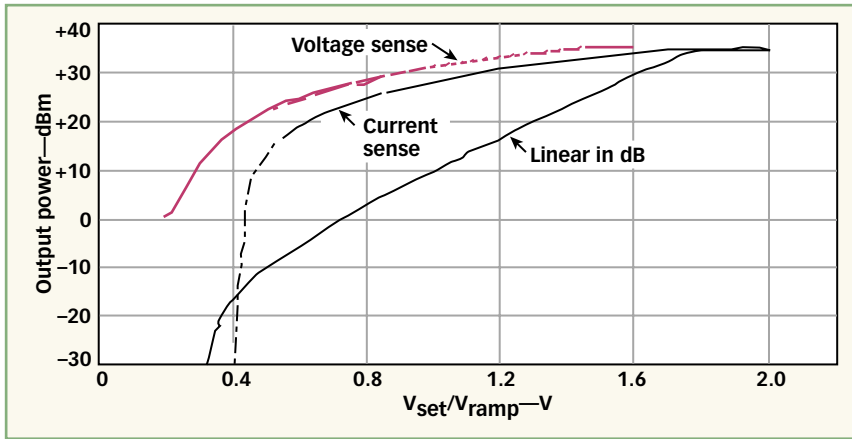$$P_{dBm} = M\log\left(V_{SET} - V_{LOOP\ Threshold}\right) + P_{OFF} \qquad (1)$$

$$P_{dBm} = 10\log\left[\frac{\left(2V_{RAMP} - V_{SAT}\right)^2}{8R_L \times 10^{-3}}\right] \qquad (2)$$

$$SLOPE\ (dB\,/\,CODE) = \frac{P_{HIGH} - P_{LOW}}{CODE_{HIGH} - CODE_{LOW}} \qquad (3a)$$

$$INTERCEPT\ (CODE) = P_{HIGH} - \left(SLOPE \times CODE_{HIGH}\right) \qquad (3b)$$

$$CODE = \frac{\left(P_{OUT} - INTERCEPT\right)}{SLOPE} \qquad (4)$$

$$10\log\left(1 - \Gamma_L^2\right) - 10\log\frac{1 + \Gamma_L}{1 - \Gamma_L} = 20\log\frac{1}{1 + \Gamma_L} \qquad (7)$$

3. Linear-in-dB power control allows for easy two-point calibration, whereas other control methods require more ramping profiles at lower power levels.

mit power level can be calculated using the formula:

SEE EQ. 4 ON P. 52

The RF power-detection method suffers only ±1 dB error for an output-power range spanning +5 to +34.5 dBm, while the current-sensing and voltage-sensing methods exhibit higher errors for narrower power levels.

The RF power-detection approach uses an internal 30-dB coupler. Because of this low coupling factor, the insertion loss of this coupler is extremely low (approximately 0.05 dB), and has a minimal impact on PA efficiency. This impact can be described by a loss factor of $10^{-0.05/10}$. The PAE of the amplifier module is then the PAE of the amplifier integrated circuit (IC) multiplied by this loss factor. It should be noted that all PAE specifications for the RF power-detection approach include the effect of directional coupler insertion loss.

The FET used in the voltage-sensing approach has a voltage drop of about 180 mV at full power. This will also reduce the efficiency of PA chip inside the module. This loss in efficiency is approximately:

$$Loss\ factor = \frac{(V_{BAT} - 0.18)}{V_{BAT}} \quad (5)$$

This loss factor is similar to the RF power-detection approach at nominal supply voltages. However, it is worse at low supply voltages. Amplifier module PAE specifications for the voltage-sensing approach also include the PAE loss due to the FET.

The log detector controller has a control function that is linear when scaled in dB/V. To achieve the desired raised-cosine RF power profile from the PA, the ramping signal from the *ramp* DAC should also follow a raised-cosine form.

During initialization and completion of the transmit sequence, the PA bias voltage should be held at its minimum level by keeping the external control voltage at some level below 150 mV (this is generally achieved by setting the ramp DAC code to 0). The PA has a clamping mechanism designed to keep the PA off, with high isolation, when the V_{SET} voltage is below 150 mV.
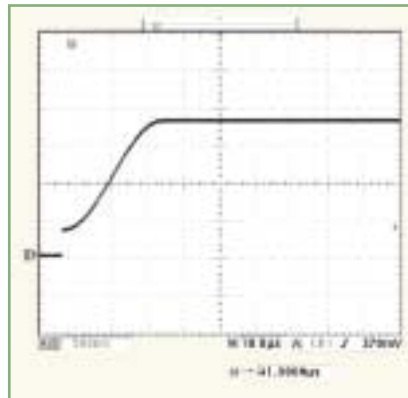

4. The filtered rising edge of ramp DAC output signals (V_{SET} drive signals) used for high-power ramping curves are based on 4× interpolation).

To optimize switching transients a step is applied to the ramp **(Fig. 4)**. The step is used because there is no point in ramping from 0 to 200 mV because the PA is designed to stay off for this voltage range.

When ramping to lower power levels, the same initial offset voltage should be applied before ramping begins. The raised-cosine portion of the ramp should be scaled to set the desired power level. The ramp-down profile can be a simple mirror image of the ramp-up signal (i.e., the same codes can be used). Alternatively, the ramp DAC signal can be a simple raised-cosine signal that falls all the way to 0 V. This is not true for the voltage-sensing and current-sensing control methods. These methods require more than one ramp profile, especially at low power levels, in order to achieve good ramping and switching transients.

A filter (C_{FLT} and R_{FLT}) must be used to stabilize the loop and ensure optimum conformance to the time mask and switching transient specifications (as shown in Fig. 2). The choice of C_{FLT} and R_{FLT} will depend to a large degree on the gain-control dynamics of the PA. The optimum values for the control loop have been determined to be 220 pF for the capacitor and 3 kΩ for the series resistor for GSM and 4.3 kΩ and 150 pF for the resistor and capacitor, respectively, in DCS/PCS systems. This gives the loop sufficient speed to follow the required ramping profiles, while still meeting the switching transient requirements at all power levels. Depending on the board layout and choice of transmit components, these values may have to be adjusted slightly. Generally, meeting these requirements is most difficult at full power.

The specific-absorption rate (SAR) is an indication of the amount of radiation that is absorbed into the body (usually the head) when using a cellular telephone. Due to strict SAR regulations, it is important to accurately control the output power of the PA. A large-impedance mismatch, e.g. 10:1, can occur at the antenna. However, due to losses in the switchplexer placed
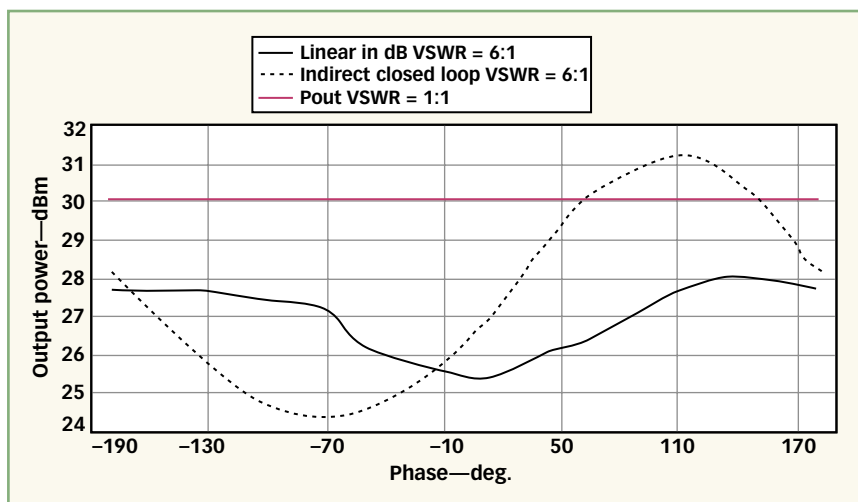
5. For a VSWR of 6:1, the output power of the indirect closed-loop approach varies by more than ±3 dB with phase. The high-directivity directional coupler used in the RF power-detection method ensures a worst-case variation of ±1.3 dB.

between the antenna and the PA, the VSWR at the PA will be reduced, but still can be as high as 6:1. **Figure 5** shows the variation in output power due to a VSWR of 6:1 for all phase angles. The baseline output power level (i.e., a VSWR of 1:1) is +30 dBm. As expected, the baseline output power level drops due to the increased VSWR. As the phase changes, the output power of the indirect closed-loop approach varies by well over ±3 dB. In contrast, the 14-dB coupler directivity of the RF power-detection approach reduces this variation to ±1.3 dB. This translates to a 14-dB reduction in the reflected power presented to the PA.

The indirect closed-loop approach operates open loop, and the output power will vary directly with the VSWR, yielding an error of:

$$Error = 10 \log \frac{1 + \Gamma_L}{1 - \Gamma_L} \quad (6)$$

The indirect closed-loop approach will have greater error in its transmitted output by a factor of:

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The error created by the RF power-detection approach will always result in a lower transmitted power, and thus will
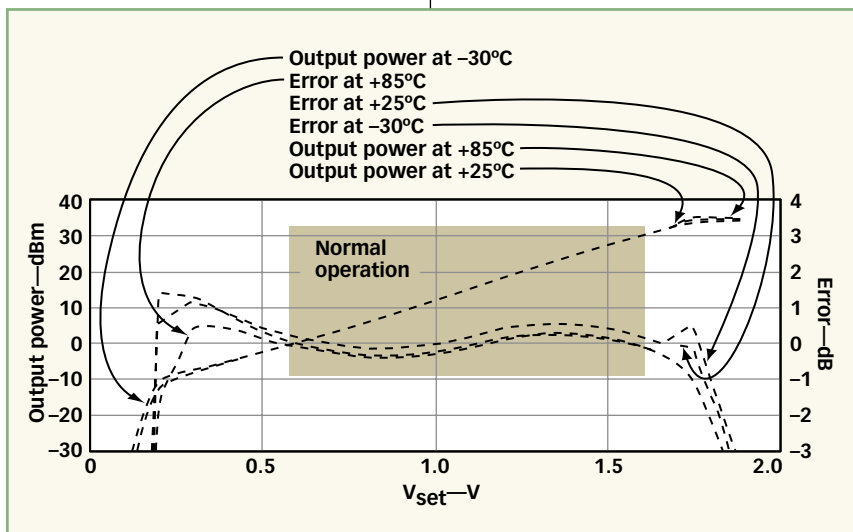
6. These curves show output power versus V_{SET} for PCS (static) at +3.5 VDC.

never exceed maximum safety limits.

The accuracy of the coupler and the detector in the RF power-detection method allows for precise power control. This error of control is within ±1 dB. The required peak power out at the antenna for GSM 900 is +33 dBm. The specification allows a ±2-dB margin. The RF power-detection approach allows cellular telephones to be operated at a lower power level of +32 dBm and still meet specification, in the process saving battery life.

> *The accuracy of the coupler and the detector in the RF power-detection method allows for precise power control.*

When the battery voltage decreases, the PA output power generally decreases. In the RF power-detection method, however, the PA output power will not decrease until the battery voltage drops below +2.9 V. This is because the closed loop senses the decreasing output power and drives the bias circuit (V_{APC}) harder, thereby keeping the output power constant.

In the indirect closed-loop approach at high battery voltages, the output power is well regulated. However, when the battery voltage to the PA decreases, the PA has trouble delivering the requisite output power. In addition, to avoid excessive switching transients at high power levels, V_{RAMP} must be limited according to the equation:

$$V_{RAMP} \leq \frac{3}{8} \times V_{BATT} + 0.18 \qquad (8)$$

As battery voltage decreases, the output power of the indirect closed-loop approach at high power also decreases. Assuming that a V_{RAMP} of 1.6 V is required to achieve full power, this limits the minimum battery voltage

to +3.25 VDC. As the battery voltage drops below this level, V_{RAMP} must also be reduced. In the case of the RF power-detection approach, the battery voltage can drop to +3 VDC before V_{SET} must be adjusted to prevent excessive switching transients. In practice, this adjustment will not be necessary as most cellular telephones are turned off at around +3 VDC.

As temperature varies, so does PA output power. The detector in the RF power-detection method detects this power change and adjusts the biasing to the PA to correct it. This method has good performance over temperatures from –30 to +85°C. **Figure 6** shows that the error is within ±1 dB and shows good linearity over all power levels within the normal output-power operating range. The indirect closed-loop approach has higher variation at lower temperatures, and requires more than one ramp profile. This increases production test time and thus final cost.

> *AS battery voltage decreases, the output power of the indirect closed-loop approach at high power also decreases.*

The linear-in-dB method described in this article is used in the models ADL5551 and ADL5552 6 × 8-mm GSM quad-band X-PA™ PA modules from Analog Devices. The inventions used in these advanced products are protected with patents and other intellectual property rights, including United States Patents Nos. 4,990,803, 4,929,909, 4,604,532, 5,572,166, 6,144,244, 6,172,549, 6,525,601, 6,489,849, and corresponding patents in other countries. **MRF**