Active output impedance for ADSL line drivers

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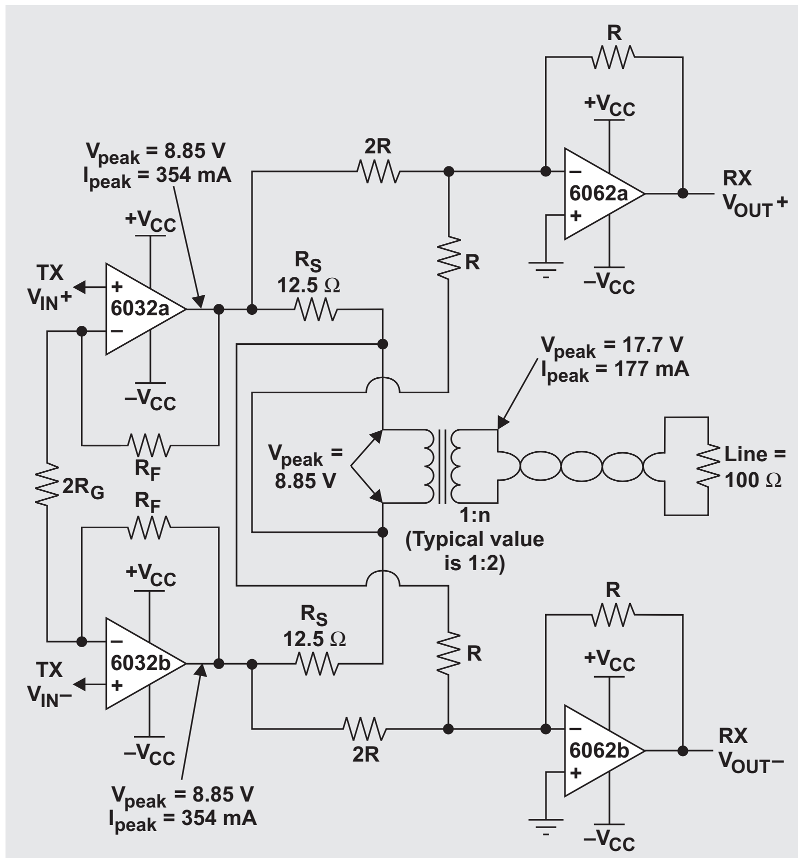
Introduction

The exceptional bidirectional data transmission rates over traditional telephone lines are a major factor for the widespread industry growth of ADSL. The ability to transmit data at over 8 MBps over an existing infrastructure of copper telephone lines with limited costs is exciting. There are several key components within the ADSL system, but this article deals solely with the line driver amplifiers.

Because ADSL is considered to be a full-duplex system, able to transmit and receive at the same time, a receiver must be incorporated into the design. The most common way of accomplishing this is to use a hybrid network. The hybrid's function is to cancel out the transmit signal while still being capable of receiving the signals from the customer-premise equipment (CPE) end (also known as the remote-terminal [RT] end). To accomplish this task, series-matching resistors, R_S, are needed and should be equal to one-half the total reflected transmission line impedance to properly match the line impedances (see Figure 1).

Figure 1. Line driver voltage and current levels to meet ANSI T1.413 requirements



$$R_S = \frac{R_{Line}}{2n^2}, \tag{1}$$

where n is the transformer ratio indicated as 1:n.

The problem with using the series-matching resistor is the associated voltage drop across this resistance. The voltage appearing at the transformer primary side is only one-half the voltage developed at the line driver amplifier output. This is one of the key issues when the power dissipation of an ADSL line driver is considered.

Traditional line driver requirements

ANSI T1.413 specifies that the central office (CO) can nominally transmit at –40 dBm/Hz on a 100-Ω telephone line from approximately 25 kHz to 1.104 MHz. This corresponds to roughly 3.16 V_{RMS} (or +20 dBm) being transmitted on the line. The problem is that ANSI T1.413 also dictates that there shall be a bit-error rate (BER) of 1×10^{-7}. In order to accomplish this feat the ADSL signal must have a peak-to-rms ratio, also known as crest factor (CF), of about 5.6 (15 dB). Taking the crest factor into account, the line voltage must now have a peak voltage of about 17.7 V_{peak} (34.4 V_{PP}). Note that the crest factor can vary from 5.3 to as high as 7, depending on the manufacturer and the system goals involved.

This large voltage requirement is a key reason for using a transformer and two amplifiers configured differentially to drive the line. Differential circuits have several advantages over single-ended configurations. This includes minimizing common-mode signals and interference, improving power-supply rejection, and the obvious advantage of doubling the voltage swing that appears at the transformer leads. Another advantage of the differential configuration is that even-order harmonics are reduced by as much as 10 to 20 dB, resulting in a very low distortion system.

Because R_S forces the amplifier to swing twice the transformer voltage requirement, the power supplies ($\pm V_{CC}$) must be increased accordingly. This increase in power-supply voltage leads to the primary issue with ADSL line drivers—power dissipation.

Line driver power dissipation

Power dissipation in the line driver amplifier is a dominant factor in CO applications. Let's take an approximation at the power dissipation levels required for the traditional line driver circuit. Let's assume that the amplifier requires at least 2 V of power-supply voltage headroom (i.e., $V_{OUT(max)} = V_{CC} - 2\text{ V}$) and there is about a 10% tolerance on the power supply. Since power dissipation of amplifiers is calculated based on the average current flowing into the amplifiers and the dc voltage, the following line driver amplifier power dissipation approximation can be made:

$$P_{Diss} = \left[2(V_{CC} - V_{OUT_{RMS}})(I_{OUT_{RMS}} \times 0.8^*)\right] + P_{Quiescent}. \quad \textbf{(2)}$$

$$P_{Quiescent} \approx 4 \times V_{CC} \times I_{CC} \times 0.7.^{**} \quad \textbf{(3)}$$

To solve for power dissipation, let

$$V_{CC(min)} \approx V_{OUT(max)} + V_{Headroom} + V_{CC_{Tolerance}}$$
$$= 8.85\text{ V} + 2\text{ V} + 1.1 = 11.94\text{ V}$$

(choose standard voltage 12 V_{DC}).

$$V_{OUT_{RMS}} = 8.85\ V_{peak} \div 5.6 = 1.58\ V_{RMS}.$$

$$I_{OUT_{RMS}} = 354\ mA_{peak} \div 5.6 = 63.2\ mA_{RMS}.$$

$$\text{Let } I_{CC} = 12\ mA_{DC}.$$

$$\therefore P_{Diss} \approx 1.05\text{ W} + 0.40\text{ W} \approx 1.45\text{ W}.$$

As you can see from the calculation, 1.45 W is a lot of power for a single device to dissipate. To compound the problem, there are as many as 72 ADSL lines on a single PCB. This is an enormous amount of heat to try to dissipate while trying to maintain proper silicon die temperatures.

Minimizing power dissipation

Power reduction is easily accomplished by reducing the series-matching resistors (R_S) to a much smaller value. The voltage drop across these resistors is then minimized. The amplifier output voltage is reduced by the same amount that allows the power-supply voltages to be reduced. Because the voltage difference between the power-supply voltage and the rms output voltage is reduced, power dissipation is also reduced. The quiescent power is reduced as well, due to the dropping power-supply voltages. Using the previous example, we can see the amount of power that will be saved by simply utilizing a smaller resistor. Let new R_S equal 13% of the original R_S value.

$$\text{New } V_{OUT(max)} = 1.13 \times \text{old } V_{OUT(max)} \div 2 = 5\text{ V}.$$

$$\text{New } V_{CC} = (5\text{ V} + 2\text{ V}) \times 1.1 = 7.7\text{ V}$$

(choose standard voltage 8 V_{DC}).

$$\text{New } V_{OUT_{RMS}} = 5\text{ V} \div 5.6 = 0.893\text{ V}.$$

$$\text{New } I_{OUT_{RMS}} = \text{old } I_{OUT_{RMS}} = 63.2\text{ mA}.$$

$$\therefore \text{New } P_{Diss} \approx 0.72\text{ W} + 0.22\text{ W} \approx 0.94\text{ W}.$$

This is a savings of 0.51 W, or 35%, per ADSL channel. When there are several channels on a single PCB, this can add up to substantial heat savings. The die temperature is also reduced, allowing for better performance and longer life of the amplifier.

However, this configuration fails to allow for proper line impedance matching. To get the best of both worlds, utilizing small series resistors and matching the line impedance, we need to use an "old" circuit configuration—the active termination circuit (also known as synthesized impedance).

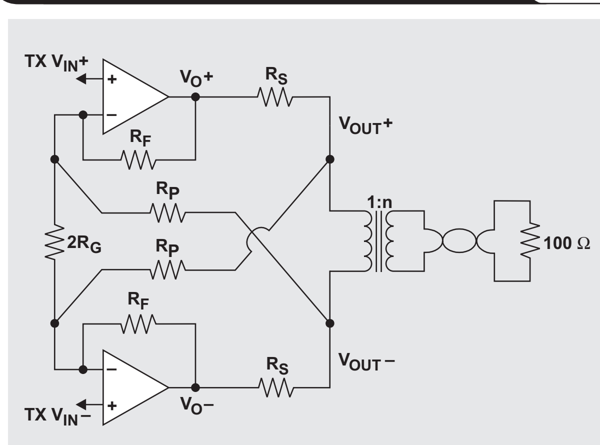
Active termination

Active termination has been around for several years.[1,2] The idea is to use a small ohmic value resistor for R_S. The circuit then utilizes positive feedback to make the impedance of this resistor appear much larger from the line side.

This accomplishes two things: (1) a very small resistance when the line driver amplifier transmits signals to the line, and (2) proper matching impedance between the line and the amplifier. Most of the original designs, however, were single-ended applications instead of the differential configuration used in ADSL systems.

Taking the general idea a step further, we can utilize the fact that the signals from each amplifier are 180° out of phase from each other in the differential system. We use these signals and connect them into the traditional inverting node on the amplifier (minus input) instead of the non-inverting node (plus input) used in the single-ended application. The advantages of this are: (1) The effective impedance of the noninverting inputs is not dictated by the positive feedback resistance and voltage gain; and (2) the active impedance achieves cross-coupling of the signals. Cross-coupling helps minimize differences between the two amplifier output signals, helping to keep the signals fully differential. Figure 2 shows the basic circuit for differential positive feedback.

Figure 2. Basic active impedance circuit



* The ADSL signal is considered to have a Gaussian distribution in the time domain. Because of this, multiplying the amplifier's rms output current by approximately 0.8 yields the average current drawn from the power supply due to the output signal current.

**This multiplication factor accounts for the fact that part of the quiescent current in a Class-AB amplifier gets diverted to the load when there is a signal appearing at the output of the amplifier driving a load. The number chosen is only an approximation and is shown only as a reference. Typical numbers range from 0.4 to 0.9 and are based on numerous circuit parameters internal to the amplifier.

The first question to answer is: How does this circuit configuration increase the effective resistance of R_S when looking from the line? If we assume that the TX inputs are grounded and apply a voltage at $V_{OUT}-$, this creates a voltage at V_O+ equal to $V_{OUT}- \times -RF/RP$. If we also realize that the voltage at $V_{OUT}+$ is equal to $-V_{OUT}-$, then $V_O+ = V_{OUT}+ \times RF/RP$. This makes R_S appear to be a larger impedance, Z, by the following formula:

$$Z(\Omega) = \frac{R_S}{1 - \frac{R_F}{R_P}}. \tag{4}$$

The important thing to consider is that regardless of the forward gain from V_{IN} to V_O, the active impedance value remains constant. The drawback to this arrangement is that the impedance will change at frequencies near the amplifier's bandwidth limit. We must ensure that the amplifier used has a bandwidth high enough not to alter the impedance at the ADSL frequencies from 25 kHz to 1.1 MHz. As a general rule of thumb, the amplifier must have a minimum bandwidth of 10 times the maximum operating frequency, or at least 11 MHz with the amplifier's intended gain.

Active impedance forward gain

Once the return impedance is corrected, we need to turn our attention to the rest of the design parameters. The most fundamental is the forward voltage gain from input to output. For simplicity, we will assume that the amplifier is well within its linear range and ignore bandwidth effects. Equation 5 shows the simplified forward gain from V_{IN} to V_O.

$$A_V = \frac{V_O \pm}{V_{IN} \pm} = \frac{1 + \left(\frac{R_F}{R_G \| R_P}\right)}{1 - \left(\frac{R_F}{R_P}\right)\left(\frac{R_L}{R_L + R_S}\right)} \quad \text{if } R_L \ll R_P, \tag{5}$$

$$\text{where } R_L = \frac{R_{Line}}{2n^2}. \tag{6}$$

In the original circuit (the classic design shown in Figure 1), R_S equaled one-half the total reflected line impedance, which also equaled R_L. We must now choose R_S as a percentage of R_L in the active termination circuit. If we define the variable X as this percentage, where $0 < X \le 1$, then we can start simplifying the preceding equations. Some references use the term "synthesis factor" (SF) to describe the percentage. Synthesis factor is simply 1/X, but the remainder of this article uses the variable X. If we realize that the term

$$\frac{R_L}{R_L + R_S}$$

is held constant, we can make several simplifications. The first sets of assumptions are

$$R_S = R_L X \text{ and} \tag{7}$$

$$\frac{R_L}{R_L + R_S} = \frac{1}{1 + X}. \tag{8}$$

We will also assume that we want the active impedance, Z, equal to the terminating resistance, R_L. Equation 4 is manipulated to achieve

$$R_P = R_F\left(\frac{1}{1 - X}\right) = \frac{R_F}{1 - X}. \tag{9}$$

Equation 9 shows that to properly match the active termination impedance, we need only select an arbitrary value of R_F. Substituting Equations 7 through 9 in Equation 5 leads us to the simplified forward voltage gain of

$$A_V = \frac{R_G[(1 + X)(2 - X)] + R_F(1 + X)}{2R_G X}. \tag{10}$$

If we know the forward gain we want in the system, we can rearrange Equation 10 to solve for the gain resistance, R_G:

$$R_G = \frac{R_F(1 + X)}{2A_V X - [(1 + X)(2 - X)]}. \tag{11}$$

Minimum active impedance forward gain design constraint

Because active impedance utilizes positive feedback, it is possible to create negative impedance instead of positive impedance. Negative impedance makes the series resistance appear to decrease rather than to increase as desired; so we must ensure that there is always positive impedance. We come to our first design constraint of the active termination circuit: There must be a minimum forward gain for the system to work properly.

Because we want to match the line properly, we must first arbitrarily choose R_F. Using Equation 9 dictates a specific fixed value for R_P. This leads to R_G solely dictating the forward voltage gain for any given value of X. The minimum forward voltage gain allowed is when R_G is not even in the system, resulting in

$$A_{V(min)} = \frac{2 + X - X^2}{2X} = \frac{(1 + X)(2 - X)}{2X}. \tag{12}$$

Luckily, for most ADSL systems, the gain of the amplifiers is typically greater than 10 V/V. Meeting the minimum gain requirement is usually not an obstacle as long as the value of X is greater than about 10%. As long as the minimum forward gain is met, the low-power active termination system will work properly.

Line impedance changes

Up until now, we have assumed that the line was a fixed value (usually 100 Ω). But in reality, we know that the line impedance is highly complex. Typically the line impedance can range from as low as 50 Ω up to as high as 300 Ω over the ADSL frequency spectrum. Since the positive feedback is obtained between R_S and the reflected line impedance (R_L), it stands to reason that the forward voltage gain will be affected.

To quantify the exact change in forward voltage gain, the variable Y is introduced. Let the variable Y equal the percentage change in the reflected line impedance (R_L). This leads to the new forward voltage gain:

$$A_V = \frac{R_G[(2-X)(1+X+Y)]+R_F(1+X+Y)}{R_G X(2+Y)}. \quad \textbf{(13)}$$

Figure 3 illustrates the percentage change in forward gain with varying values of X. The forward gain with a 100-Ω line impedance will be used as the base line for comparison.

It is interesting to note that the change in percentage gain is independent of the transformer ratio, n; feedback resistance, R_F; gain resistance, R_G; and initial amplifier gain, A_V.

The minimum forward gain will also vary with the line impedance. The minimum forward gain becomes

$$A_{V(min)} = \frac{(2-X)(1+X+Y)}{X(2+Y)}. \quad \textbf{(14)}$$

Figure 4 illustrates the minimum forward gain with varying line impedance.

When an active termination system is designed, it does not matter what initial design line impedance is used. As long as the minimum gain criterion is met, the system should not create negative impedances.

Line impedance changes and the amplifier output voltage

In a real system it is quite common for forward voltage gain to change ±20%, which must be accounted for. If not, the input signal can be amplified too high and clipping could easily occur. Excess distortion, data transfer rate, line reach, and even power dissipation could become worse if the line impedance is not handled properly within the active impedance circuit design.

Examining the circuit of Figure 2 and using Equation 7 will help us calculate how the line impedance changes the amplifier's output voltage. We will assume that R_S is designed for a 100-Ω system and is held constant. We will also assume that the power on the line was done with a 100-Ω line impedance and is +20 dBm. This corresponds to a line voltage of 3.162 V_{RMS}. The formula used to find the corresponding amplifier voltages is

$$V_{O_{RMS}} = \frac{V_{Line_{RMS}}(R_{Line}+2n^2R_S)}{2nR_{Line}}. \quad \textbf{(15)}$$

The important number is the peak output voltage of the amplifier ($V_{peak} = V_{RMS} \times CF$) because a given supply voltage determines how much voltage swing can occur. Failure to plan for varying line impedances can cause

Figure 3. Forward gain change with varying line impedance

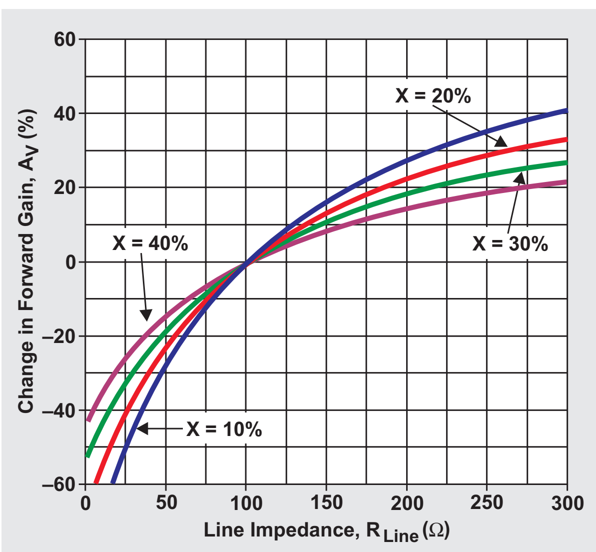
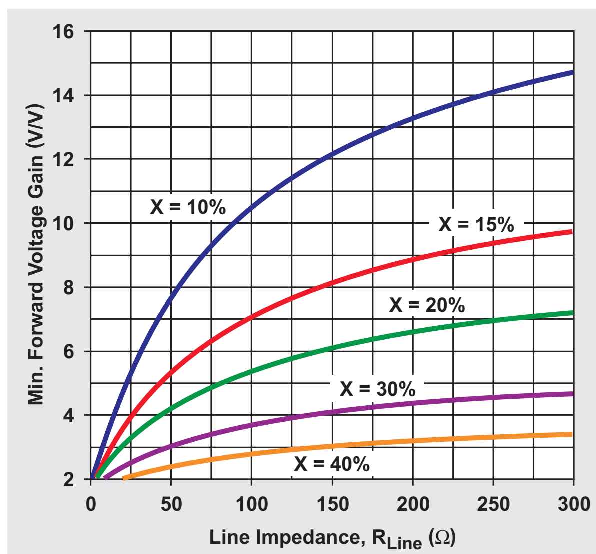


Figure 4. Minimum forward gain change with varying line impedance



some serious problems. Figure 5 illustrates this issue with X = 20% (SF = 5) and a crest factor of 5.3.

Obviously, as the crest factor increases, the peak output voltage will also increase. Additionally, when R_S increases, the amplifier output voltage will also increase. The obvious question is: Why not use the smallest resistance possible? There are several reasons for this that the remainder of this article explains in detail.

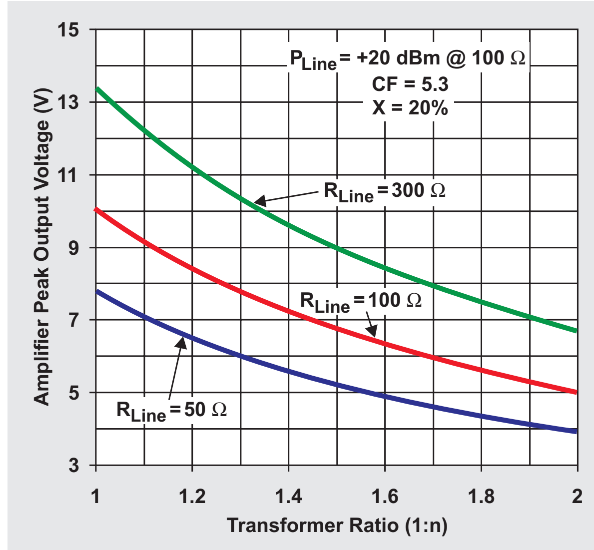
Lab tests

Setup

The first test examined how the resistor values affect the system. Because the THS6032, like most ADSL line drivers, is a current feedback (CFB) amplifier, the feedback resistance (R_F) dictates the bandwidth and the stability of the amplifier. Keeping a high bandwidth increases the amplifier's excess open-loop gain in the ADSL frequency band and reduces distortion. At the same time, however, the amplifier bandwidth may be high enough to interact with the transformer's resonance frequency, which can cause possible instabilities in the overall system. This is especially true when active impedance circuits are used, as R_S can become very small, resulting in very little isolation between the amplifier and the transformer. When you consider Equations 13 to 15 along with the transformer's impedance at resonance, it is apparent that the system can potentially become unstable. Using a simple RC snubber across the transformer can be a simple solution for instability concerns.

To circumvent this potential issue, two new amplifiers from Texas Instruments, the THS6132 and the THS6182, incorporate special internal circuitry. These new amplifiers yield extremely low distortion at the ADSL frequencies yet have a bandwidth of only 10 to 20 MHz—depending on the system design. For all other line drivers, the trade-off of bandwidth and stability needs to be managed. As a side benefit of reducing the feedback resistor, the overall output noise of the line driver system can be significantly reduced.

Figure 5. Amplifier peak output voltage with X = 20%



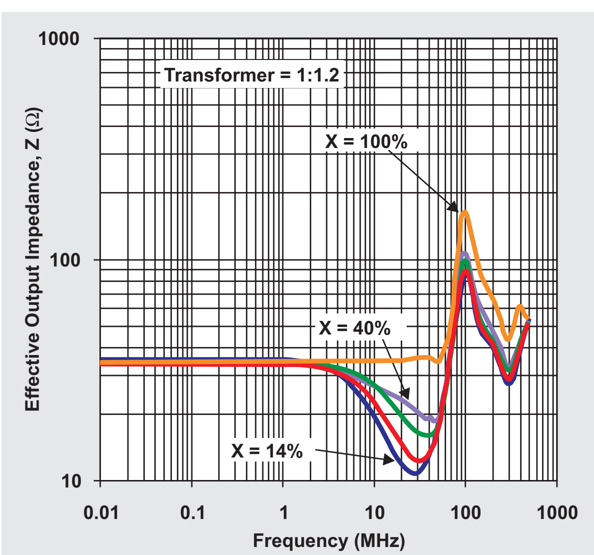
For the THS6032 testing, a feedback resistor value of 1150 Ω was chosen. The rest of the system component values were then easily calculated with the previous equations. The only other variable was that the gain of each amplifier was set to approximately +12 V/V. This allowed testing of the X = 10% system where the appropriate minimum gain requirement was about 10.5. As R_S was increased, the gain also had to be increased to account for the additional voltage drop from the added series resistance.

The active impedance test

Figure 6 shows the impedance looking back into R_S from the transformer primary. It clearly shows the amplifier's closed-loop bandwidth effects. Eventually the amplifier's own output impedance takes over regardless of the termination system used. At this point the impedance is out of the designer's control. Since the ADSL spectrum is well controlled, the system will meet its designated functionality as a low-power line driver.

One area of concern with using active impedance is that lightning surge tests could overwhelm the amplifiers' internal circuitry and cause failures due to a decreased real resistance between the amplifier and the transformer. The larger the resistance, the better the chance that no damage will occur within the amplifier. If the active impedance configuration is utilized, then R_S should be a "respectable" value and not something trivial (for example, 1 Ω). Most systems should strive for a value of 20 to 30% of R_L (SF = 3 to 5). This allows for respectable power savings and reasonable isolation from surges on the line.

Figure 6. Amplifier effective output impedance viewed from transformer primary winding



Power dissipation and distortion

The line impedance used in the testing was a 100-Ω resistor. Variable line impedance issues are not of concern but should constrain the final system design. As a result, the power dissipation numbers shown should be considered optimal for a particular test setup. When a varying line impedance is thrown into the mix, the power-supply voltages will need to be adjusted accordingly and the power dissipation will increase.

The other factor hampering the power dissipation is that the THS6032 requires 4-V headroom from the power supplies. This is due to the Class-G architecture requiring multiple series transistors in the output stage. If a very low headroom amplifier were used (such as the THS6132 or THS6182), the power-supply voltage could be reduced by at least ±2 V, decreasing power even more. As we are concerned with power savings in general, these results can be used to draw some general conclusions about the use of active termination in an ADSL application.

Keep in mind that when you compare power numbers from amplifier to amplifier, the entire system configuration needs to be divulged. This includes things such as crest factor; accounting for varying line impedances; accounting for power-supply tolerances; and, of course, the synthesis factor. Because of the numerous options available, doing a true apples-to-apples comparison is often very difficult when you just look at manufacturers' data sheets.

As a reference for the active termination testing, a THS6032 was tested with the traditional configuration shown in Figure 1. To really see the effects of the Class-G circuitry in action, refer to Figure 7, which shows how changing the V_{CC-L} supply voltages alters the power dissipation. For reference, it also shows the power consumed in each set of supplies. In Class-AB mode, power dissipation is about 1.8 W; but in Class-G mode, the best power achieved is approximately 1.35 W with V_{CC-L} at ±6 V. The multitone power ratios (MTPRs) were –70 dBc for Class-AB operation and –68 dBc for Class-G operation.

Figure 8 shows how the crest factor affects power dissipation with a 1:1.2 transformer and X = 20% (R_S = 6.94 Ω). The power-supply voltage was chosen to give an additional ±0.5-V headroom for a design margin. In the lab, we could set the supplies ±1 V lower before clipping started to occur; but this is not considered good practice, as power-supply tolerances and component tolerances could come into play. The power dissipation numbers shown are thus considered to be realistic and within the safe operating area of the system.

When compared to the traditional circuit design, the active termination circuit saved a huge 47% in power dissipation. This was true for both Class-AB operation and Class-G operation. For the active termination data, the use of Class-G operation saved an additional 20 to 25% power dissipation compared to the Class-AB operation. As expected, when the crest factor increased, the power dissipation also increased by as much as 25%. This was mainly due to the increase in power-supply voltage required to handle the larger peak voltages.

Figure 7. Traditional circuit design power dissipation results

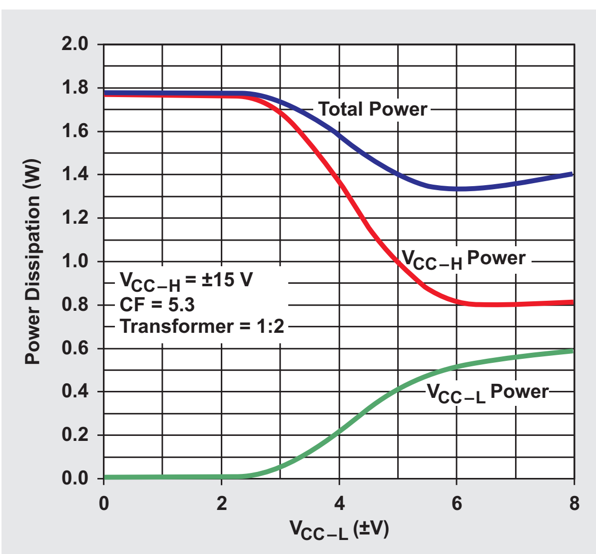


Figure 8. Power dissipation with 1:1.2 transformer and different crest factors

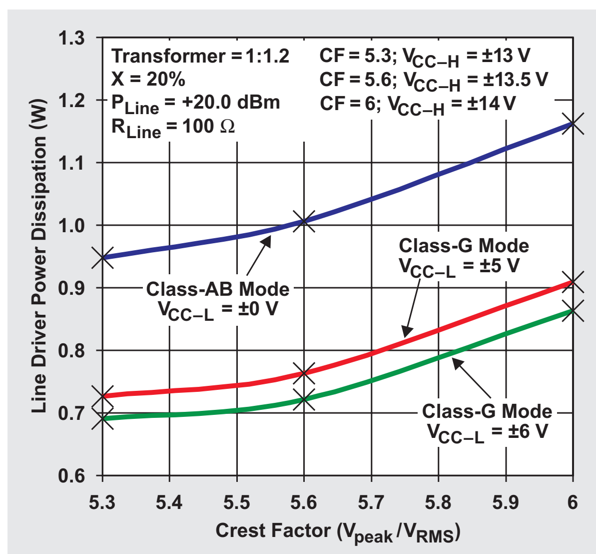


Figure 9 shows how changing R_S affects the power dissipation. A common crest factor of 5.3 was used to illustrate the change in the system.

If the power-supply voltages had been held constant and no clipping had occurred, the power dissipation would have decreased with an increase in R_S; but the testing was done to show the best possible performance with a given set of constraints. The power-supply voltages thus were increased as R_S was increased to compensate for the increase in output voltage required from the amplifier. The power-supply voltages ranged from ±12.5 V (X = 14%) to ±14 V (X = 40%).

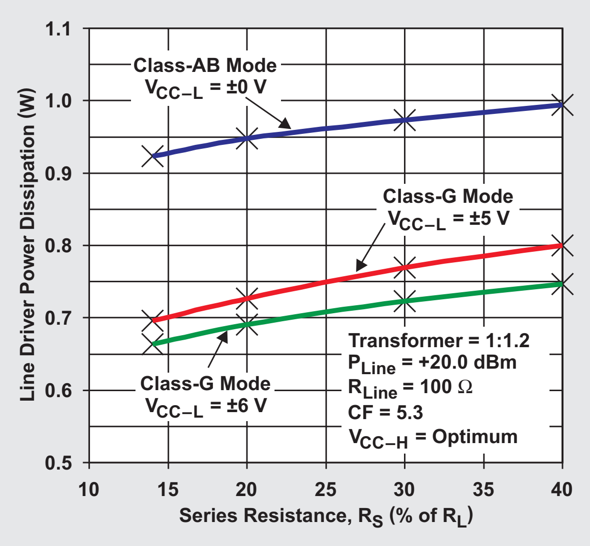
The last thing to check was the effect of MTPR distortion on the system.

Figure 10 shows us that as R_S increases, the MTPR distortion decreases. The designer has to choose between lower distortion and lower power dissipation. As stated earlier, a series resistance of 20 to 30% of R_L should give good results for both requirements.

Power dissipation and MTPR with multiple transformer ratios

The purpose of the next series of tests was to find out if there is a general relationship between the transformer ratio and the power dissipation. For each transformer ratio tested, the corresponding resistor values and power-supply voltages were accordingly changed. Figure 11 shows how changing R_S affects power dissipation with varying transformer ratios.

Figure 9. Power dissipation with 1:1.2 transformer and varying R_S



Regardless of the power-supply voltages and the mode of operation, as R_S increases, the power dissipation increases. This is generally dominated by the amplifier's overhead

Figure 10. MTPR with 1:1.2 transformer and varying R_S

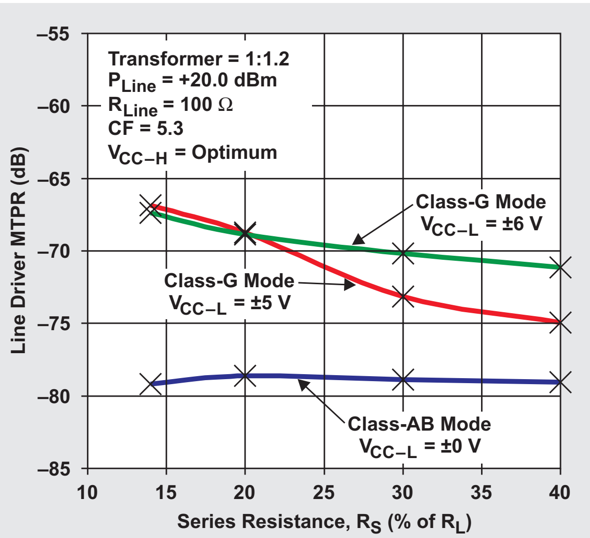
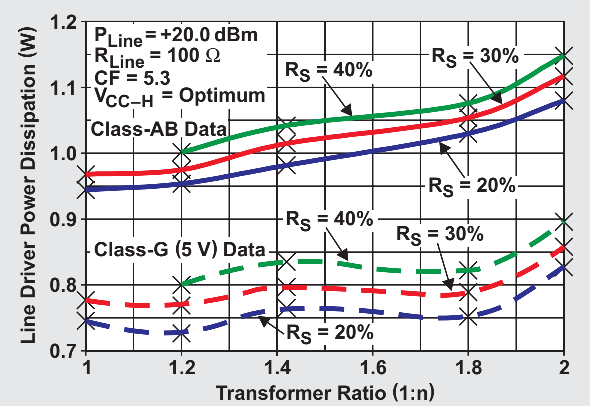


Figure 11. Power dissipation with varying transformer ratios



voltage requirements and quiescent current. We now come to the final test—determining the effects of varying transformer ratios on MTPR distortion. Figure 12 shows the effects of R_S on MTPR distortion with a changing transformer ratio and the same setup that was used before.

The data tells us that increasing the physical value of R_S lowers MTPR distortion. This is because distortion in operational amplifiers generally gets better with an increase in load resistance. In the case of the ADSL configuration, increasing R_S also helps isolate the complex loading that the transformer places on the amplifier. Comparing the 1:2 transformer data with the traditional circuit design shows that MTPR performance degrades by 4 to 5 dB as the transformer ratio increases.

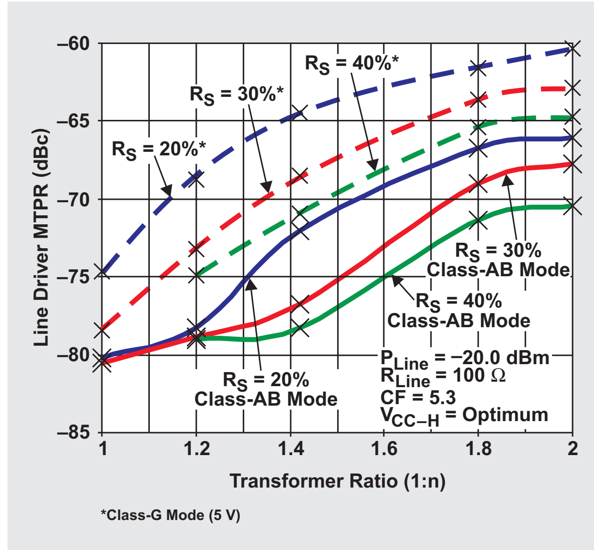
Conclusion

Reduced power dissipation is the main goal for using active termination in ADSL systems. Using a 1:1.2 transformer saved 47% of power regardless of the mode in which the THS6032 was used. This translates to a savings of up to 0.85 W with Class-AB operation and 0.63 W with optimal Class-G operation. In light of the distortion and power savings, choosing a value for X of 0.2 to 0.3 (SF = 3 to 5) shows about the best overall performance.

Using TI's newest amplifiers, THS6132 (Class-G) or THS6182 (Class-AB), can save substantially even more power. Initial testing with the THS6132 in Class-G operation shows a total power consumption of as low as 0.53 W, which is a power dissipation of roughly 0.43 W over the THS6032. However, keep in mind the design constraints of the active termination system. The line impedance variations, the minimum power-supply voltages, and the system crest factor all contribute to the power consumption of the line driver.

With any electrical circuit, there are trade-offs to using one configuration over another. The active impedance circuit is no exception. The trade-off to achieving lower line driver power dissipation is that the receiver circuitry will require more voltage gain to overcome the voltage reduction appearing across R_S. This can play a significant role in the noise performance of the system. One way to help alleviate this problem is to use a smaller transformer ratio; but the power-supply voltages will have to be increased, which can increase power dissipation. The added benefits of an increased series resistance can help in many other areas of the system, including distortion and surge isolation. Ultimately, the goal of saving power can still be met while satisfying all requirements of the ADSL line driver system.

Figure 12. MTPR distortion with varying transformer ratios



Additional information will be available in an application note to be released by January 2003, at www-s.ti.com/sc/techlit/sloa100

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2. Donald Whitney Jr., "Design Ideas - Circuit Adapts Differential Input to Drive Coax," *Electronic Design News* (May 8, 1997), pp. 132-34.

Related Web sites

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