Integrated-Circuit Operational Amplifiers

Edited by

Robert G. Meyer

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Contents

Introduction

Since the introduction of the first monolithic operational amplifier (op amp) in the mid-sixties, these devices have created a revolution in electronics. The utility and low cost of modern integrated-circuit op amps have resulted in their widespread use in all types of electronic systems. Reflecting this activity, many books have been written on the subject of op amp applications, and integrated-circuit manufacturers publish voluminous data books describing many different types of op amps.

The proliferation of op amp types is both a benefit and a problem to the op amp user. The obvious benefit is that the knowledgable user has a wide selection from which to choose an op amp type that closely fits his needs. However the non-specialist in this area often has difficulty in making a logical selection from among the many circuits available. One of the aims of this volume is to illustrate the design techniques and circuit configurations used in the most common types of integrated-circuit op amps. This knowledge allows the op amp user to choose the most appropriate circuit from among those available to satisfy some particular application. For example, if a user understands the difference between the performance of input stages containing super-β transistors and field-effect transistors, he is better able to choose the most appropriate type of op amp for high impedance applications.

A second important aim of this volume is to bring together in one place a selection of the most significant papers published on the subject of monolithic op amp design. This should prove useful to op amp designers as a reference work, and also to analog integrated-circuit designers in general, since the circuit techniques developed for monolithic op amps have wide application to other integrated-circuit types.

This book consists of a selection of journal articles on op amp design, and a series of manufacturers' data sheets describing commercially available circuits. The data sheets have been chosen to illustrate the characteristics of commonly used circuits in various categories, and also to show the practical application of many of the principles described in the journal articles. The book is divided into eight parts covering six different categories of op amps plus parts on general design techniques and aspects of op amp modeling. Each part is preceded by an introduction which gives a perspective of the topic in question and briefly summarizes the salient points of each paper. At the end of each part, additional references are included which should prove useful for further reading in the area. The papers in this book are drawn from several sources, but the largest number by far were originally published in the IEEE Journal of Solid-State Circuits.

Finally, it should be noted that an excellent companion to this volume is the IEEE Press Book entitled "Analog Integrated Circuits" by A. B. Grebene. Some of the basic principles of analog integrated-circuit design are further explored in that volume, and, in addition, some of the principles described in this present book are shown applied to other types of analog circuits.

Part I
Monolithic Op Amp Design Techniques

In this part, several papers have been collected which summarize the basic design techniques used in integrated-circuit operational amplifiers. In the first paper by Widlar, the advantages of active loads are discussed, together with devices such as collector FET's, controlled-gain lateral pnps and super-β transistors. This paper illustrates the evolution of op amp design from discrete circuits, through the μA 709 integrated circuit to the LM 101. The latter is a high performance op amp of similar design to the μA 741, to be described later.

In the second paper, Solomon examines in detail the design tradeoffs involved in op amps of the LM 101 and μA 741 types. Methods of compensation and circuit slew rate are considered, together with the interesting problem of thermal feedback which is unique to monolithic circuits.

In the third paper, Gray and Meyer define some commonly used specifications for op amps, and discuss the relation between offset voltage and common-mode rejection. The frequency response of op amps is considered, and the connection between slew rate and offset voltage is described. Finally, the settling time of an op amp is shown to be dependent on the fine detail of the frequency response of the circuit.

In the last paper in this part, Wooley *et al.* describe a detailed computer analysis of the widely used μA 741 op amp. Typical device data is tabulated, and this is used to simulate many aspects of circuit performance. The influence of the active load on common-mode rejection is shown and the frequency limitations of the circuit are explored.

Since much of the material of this part has centered on the μA 741 type of circuit, a complete data sheet for this device has been included. Finally some additional reference papers are listed together with a selection of books discussing operational amplifiers and analog integrated circuit techniques.

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Design Techniques for Monolithic Operational Amplifiers

ROBERT J. WIDLAR

***Abstract*—The characteristics of recently developed integrated-circuit components will be reviewed and some new devices will be described. Their impact on the design of monolithic operational amplifiers will also be discussed. Emphasis will be placed on realizing particularly good dc characteristics—especially low input current. However, techniques for obtaining higher operating speeds will also be covered.**

Introduction

NEARLY one million monolithic operational amplifiers per month are now being sold. Prices have dropped from $70 in 1965 to less than $2 today, even for relatively small quantities. In fact, the cost is low enough that these devices are used as simple components in applications where operational amplifiers would not even have been considered a couple of years ago.

The 709 [1] was the first monolithic circuit that approached discrete designs in general performance and usefulness, yet could be manufactured with high yields in volume production. It was also the first to be second sourced, and today there are eight major suppliers.

The LM101 [2],[1] announced in 1967, was the next major advance in monolithic amplifiers. It has essentially the same specifications as the 709, but it eliminates most of the application problems. The LM101 is not susceptible to latch-up when the common-mode range is exceeded, the inputs and output are protected from overloads, it operates over a wider range of supply voltages, and it is much less prone to oscillations. In addition, frequency compensation can be accomplished with a single 30-pF capacitor—a value that has been included on the silicon chip in the μA741 [3], the RM4101 and the LM107.[2] Hence, with this design it is practical to offer a fully compensated monolithic amplifier.

With the LM101A [4], brought out late in 1968, the input current specifications of discrete amplifiers were finally matched. This device offers input bias currents less than 100 nA and input offset currents less than 20 nA—guaranteed over a -55 to 125°C temperature range. This means that even FET-input amplifiers can be replaced with low-cost monolithic circuits, realizing improved performance in full-temperature-range applications.

The technology is available to make another order of magnitude improvement in input currents, making monolithics competitive with chopper stabilized amplifiers at high temperatures. This leaves only one major area where a significant performance improvement is needed: higher operating speeds. Present general-purpose amplifiers have slew rates of 0.5 V/μs and bandwidths of 1 MHz.[3] Many applications require more than ten times better performance than this.

Impact of New Components

Compared to discrete amplifiers of its time, even the 709 was a weird design, reflecting differences in the relative cost of components in an integrated circuit. This can be seen by comparing a typical discrete component design, shown in Fig. 1, with a 709 schematic in Fig. 2. The apparent complexity of the circuit was increased to minimize total circuit resistance, to make up for the poor gain characteristics of some transistors, and to reduce power consumption to a value that can be handled by integrated-circuit packages. For example, transistors were substituted for large-value resistors, where possible; buffer transistors were added to make the circuit operation insensitive to production variations in transistor current gains; and a class-B output stage was used to reduce quiescent current. In many cases, the integrated circuit gave significantly improved performance. And this was done without paying a price penalty, since the monolithic amplifier is now available at less than one-fifth the price of the discrete version.

Since the 709 was designed in the early days of linear circuits, it did not make use of components that were much different from those employed in digital circuits at that time. The singular exception was the lateral p-n-p [5]. But, the circuit was designed so that it could function properly with p-n-p current gains less than 0.1. Hence, the design was not too adventuresome, as should be the case when working with an emerging technology. The LM101A, on the other hand, makes extensive use of lateral p-n-p transistors, pinch resistors [6], and collector FET's.[4] Judging by the schematic diagram, it is a much more complicated circuit than the 709, as can be seen from Fig. 3. However, schematics are de-

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[1] The LM101 is similar to the LM101A described later in the text.

[2] The RM4101 and the LM107 are compensated versions of the LM101 and the LM101A, respectively.

[3] Slew rates around 10 V/μs and bandwidths of 10 MHz can be realized with certain amplifiers in specialized applications using appropriate compensation networks.

[4] To be described later.

Reprinted from *IEEE J. Solid-State Circuits*, vol. SC-4, pp. 184–191, Aug. 1969.

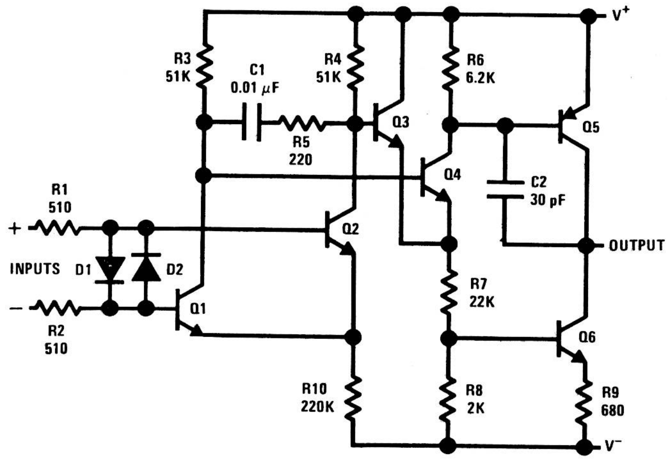

Fig. 1. Discrete component design for a low-cost operational amplifier.

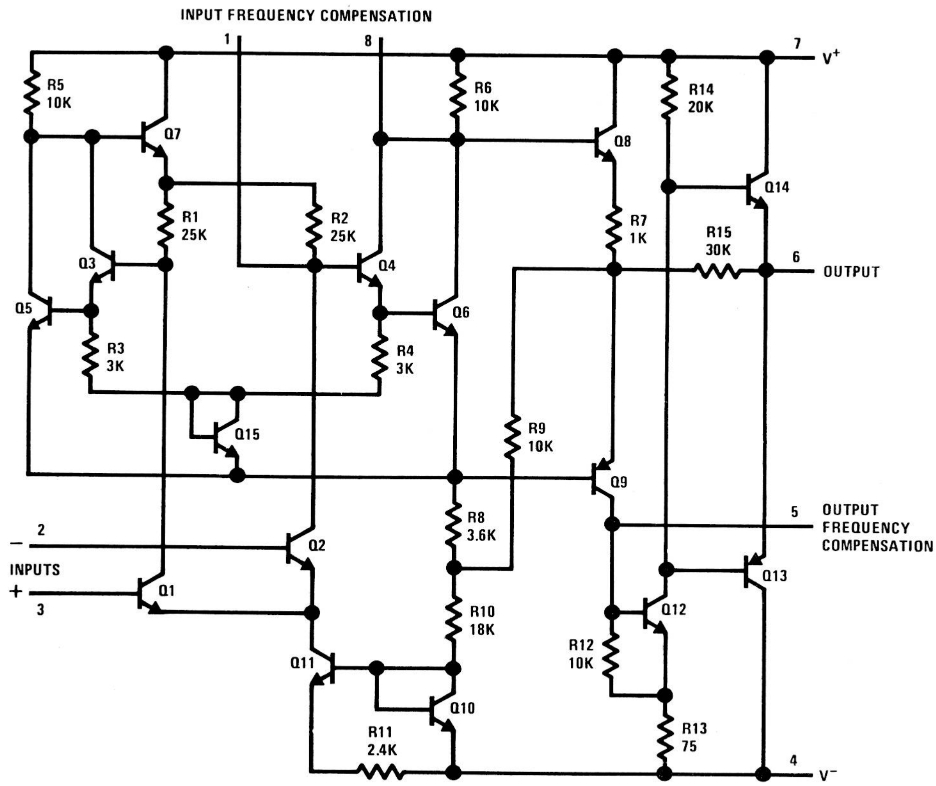

Fig. 2. Schematic diagram of the 709.

ceiving as it is fabricated on a 45-mil-square die, compared to 55 mil square for the 709. Hence, it is clear that new components have made possible much more sophisticated designs without making the circuit more difficult to manufacture.

One of the most significant departures from standard design on the LM101 and LM101A is the extensive use of active collector loads. Referring to Fig. 3, Q_5 and Q_6 serve as the collector loads for the input stage while Q_{17} is the load for the second-stage amplifier, Q_{10}. Active loads have several distinct advantages over the more common resistor loads. First, they permit low-current operation without large resistance values. This is important in reducing input bias currents and power consumption. Second, they do not require that much voltage be dropped for proper operation. This increases common-mode range, increases output swing, and permits the circuit to operate over a wider range of supply voltages. Last, they make possible much higher gain per stage, so fewer stages can be used. This simplifies frequency compensation immensely.

Another component first used in the LM101 design is the collector FET, Q_{18} in Fig. 3. This device, illustrated in Fig. 4(a), has proved invaluable for making low-power high-voltage circuits without large resistor values. The typical characteristics for an FET that is 3 mils wide between isolation cuts, and 20 mils long are shown in Fig. 4(b). Its usage is restricted somewhat by the fact

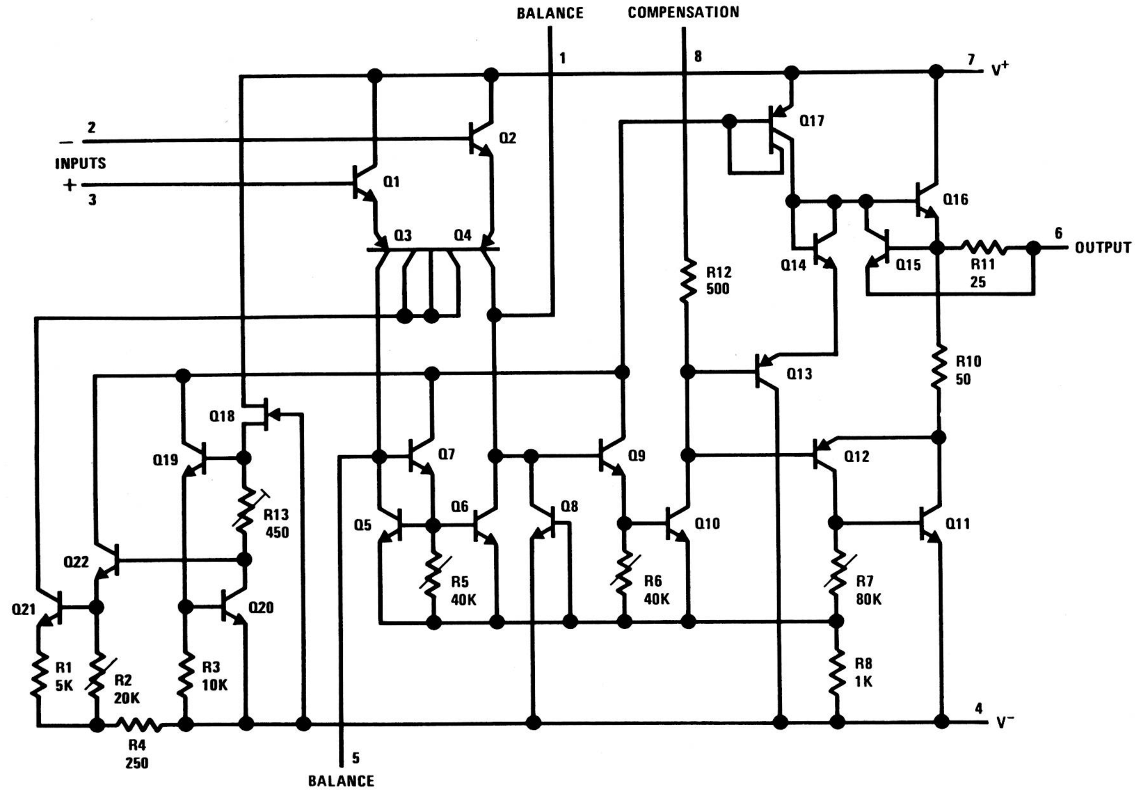

Fig. 3. Schematic diagram of the LM101A.

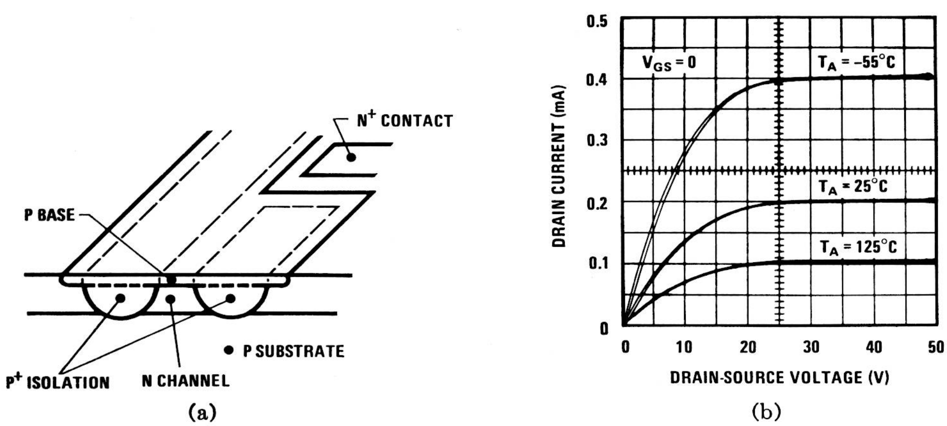

Fig. 4. The collector FET. (a) Sectional view. (b) Characteristics.

that the gate must be committed to the most negative point in the circuit.

The collector FET has proved most effective in circuits where the operating levels are determined by current sources. The FET provides turn-on, or bias, current for the current sources. All the currents can be determined by resistors in the current sources that have relatively small voltages dropped across them, and there are no resistors connected directly across the supplies. This approach not only minimizes chip area as well as current drain but also permits the circuit to operate over a wider range of supply voltages.

Pinched-base resistors [6] can give quite large resistance values in a small area, as they have a nominal sheet resistivity between 10 and 30 kΩ/□. However, it is difficult to maintain resistance tolerances within a factor of 2 or 3 of nominal, because of process variations in production. In addition, the resistors have a large positive temperature coefficient that causes them to vary by a factor of 4 over a −55 to 125°C temperature range. Pinch resistors also have an FET-like characteristic and a breakdown voltage around 6 volts, as shown in Fig. 5. Nonetheless, they have proved valuable for such uses as emitter–base bleed resistors (R_5, R_6,

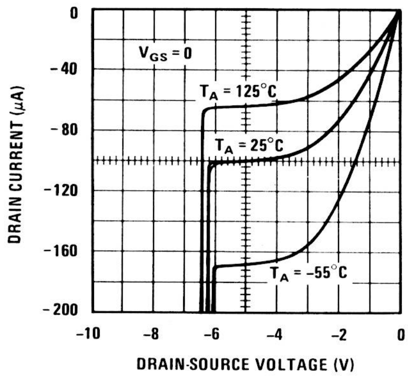

Fig. 5. Characteristics of pinched-base resistors.

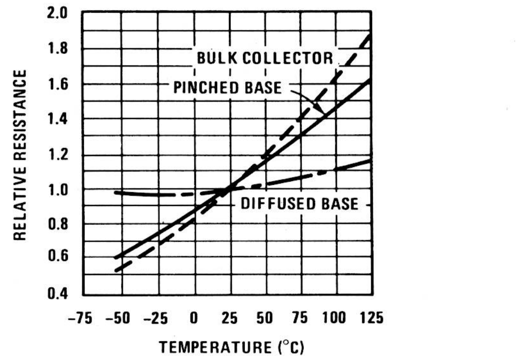

Fig. 6. Temperature characteristics of diffused-base **resistors**, pinched-base resistors, and bulk-collector resistors.

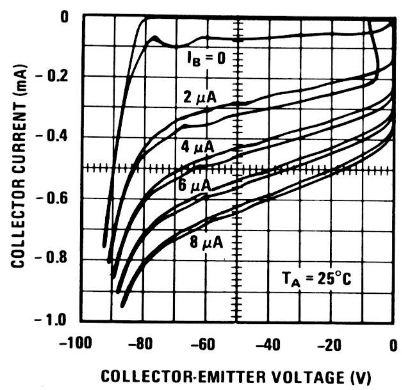

Fig. 7. Collector characteristiccs of a high-gain lateral p-n-p.

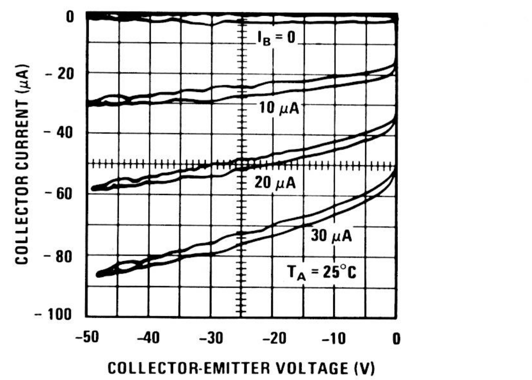

Fig. 8. Collector characteristics of a controlled-gain lateral p-n-p.

and R_7 in Fig. 3), where the voltage and tolerance problems do not materially affect circuit performance. They are also useful in designs that make use of the rough correlation between the resistance value and the current gain of the n-p-n transistors.

Collector resistors are made from the epitaxial collector material of the n-p-n transistors. The resistance tolerance is considerably better than pinch resistors, although it is not quite as good as base resistors. Resistance values from about 100 ohms to over 100 kΩ are practical. The interesting property of collector resistors is that they have temperature characteristics similar to Sensistors[5] and can be used for temperature compensation. One of these devices is used for R_{13} in Fig. 3. The temperature characteristics of collector resistors and pinched-base resistors are compared with standard diffused-base resistors in Fig. 6.

Lateral p-n-p transistors have long been known for their low dc gain. However, devices can now be fabricated with current gains greater than 100. This is done without significantly complicating processing, which is the major advantage of the lateral p-n-p over other complementary structures. High-gain p-n-p's open up new design possibilities. They can even be used for input stages, where their flat gain characteristics over a wide temperature range can be put to good use. The characteristics of a high-gain lateral p-n-p are shown in Fig. 7.

With higher current gain in the basic p-n-p structure, controlled-gain transistors can be made. This is done by breaking the collector into two segments and connecting one segment back to the base. The equivalent current gain is then determined by the relative size of the segments. Fig. 8 shows the collector characteristics of a device that has been designed for a current gain of 2. The current gain is not precisely fixed, varying between 2 and 3 as the collector–emitter voltage is increased to 50 volts. However, this tendency, caused by base-width modulation from the active collector, can be minimized by increasing the distance between the emitter and the collector.

The current gain of an n-p-n transistor depends, for one, on the length of the emitter diffusion cycle. Devices that are diffused for unusually long periods will exhibit increased current gain at the expense of breakdown voltage. Fig. 9 illustrates the characteristics of a transistor that has had the emitter driven in to the point where the device is nearly a collector–emitter short. Current gains in excess of 4000 can be obtained; however, the breakdown voltage is quite low. High-gain

[5] Registered trademark of Texas Instruments Incorporated.

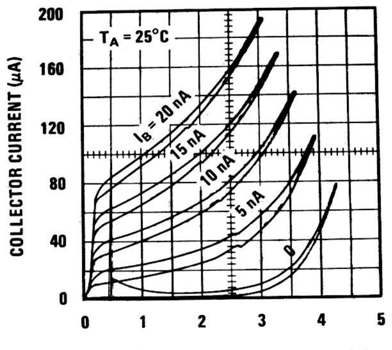

Fig. 9. Curve tracer display of high-gain transistors.

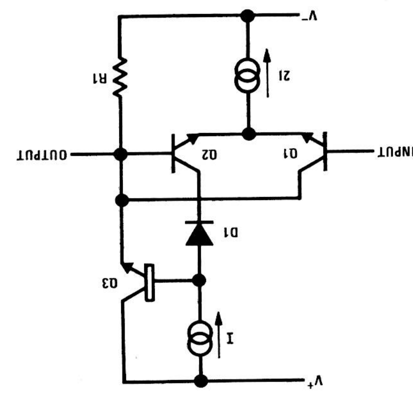

Fig. 10. Voltage follower using both high-gain and high-voltage transistors.

transistors, like these, can be built on the same chip with standard n-p-n transistors by using two separate emitter diffusions. With this technology, it is possible to design circuits that take advantage of the high current gain, yet can be operated at high voltages.

Reducing Input Currents

The low-voltage super-gain transistors can provide another order of magnitude improvement in the input current specifications of monolithic operational amplifiers like the LM101A. A circuit using them is the voltage-follower design shown in Fig. 10. High-gain primary transistors are used in the input stage to get very low input bias current. D_1 is included to operate Q_2 at near-zero collector–base voltage, and the collector of Q_1 is bootstrapped to the output for the same reason. Hence, low-voltage transistors can be used on the input. The only transistor that sees any voltage is Q_3, which buffers the output. Its current gain requirements are not stringent, so a moderate-gain secondary transistor can be used. In this particular design, the high-gain and high-voltage transistors are combined to take advantage of the best characteristics of both, without complicating the circuitry.

Because the input transistors are operated at zero collector–base voltage, high-temperature leakage currents do not show up on the input. Field-effect transistors, which, in the past, have been an obvious choice for the input stage of low-input-current operational amplifiers, suffer from leakage problems because there is no way to operate them with no voltage across the gate junction [7]. In applications covering a -55 to 125°C temperature range, super-gain transistors, which can give worst-case bias currents of 3 nA and worst-case offset currents of 400 pA, have a distinct advantage over FET's. With existing technology, they can equal FET's over a -25 to 85°C temperature range, and it is not difficult to foresee their superiority over a 0 to 70°C temperature range. Furthermore, matched pairs of super-gain transistors exhibit typical offset voltages of 0.5 to 1.0 mV with temperature drifts about 2 μV/°C, compared with 40 mV and 50 μV/°C for FET's. Certainly, discrete FET's can be compensated or selected for better offset or drift, but at a substantial increase in cost. MOS transistors, which do not have leakage problems, are no alternate solution [7] because they exhibit gross instabilities in offset voltage.

Standard bipolar transistors in a Darlington connection have been tried in a number of IC designs to get very low input currents. First-order calculations indicate that a Darlington should be competitive with super-gain transistors on input-current specifications. However, differential amplifiers, using Darlington-connected transistors, have problems that may not be immediately obvious.

The offset voltage depends not only on the inherent emitter–base voltage match of the transistors but also on the percentage match of current gains. A 10 percent mismatch in current gains give a 2.5-mV offset. Within a given process, bias currents drop faster than offset currents as the transistor current gains are raised. Hence, the better the transistors, the worse the offset voltage.

In addition to being a major contributor to offset voltage, this dependence on current-gain matching causes other problems. In a simple differential amplifier, the offset voltage drift can be correlated with offset voltage [8], [9] because of the predictable nature of emitter–base voltage. This is not so with Darlingtons, as bias current matching is not predictable over temperature. At high temperatures, this effect is aggravated further by leakage currents, so it is not possible to predict performance over a wide temperature range based on room-temperature tests.

Reducing the collector current of double diffused silicon transistors to very low currents (below 1 μA), as is done in a Darlington, does not improve input currents as much as might be expected. Lowering the collector current by a factor of 10 reduces the bias current by about a factor of 7 and the offset current by a factor of 3. These numbers are typical; the results obtained near the edges of a production distribution are significantly worse. In addition, the variation of input currents with

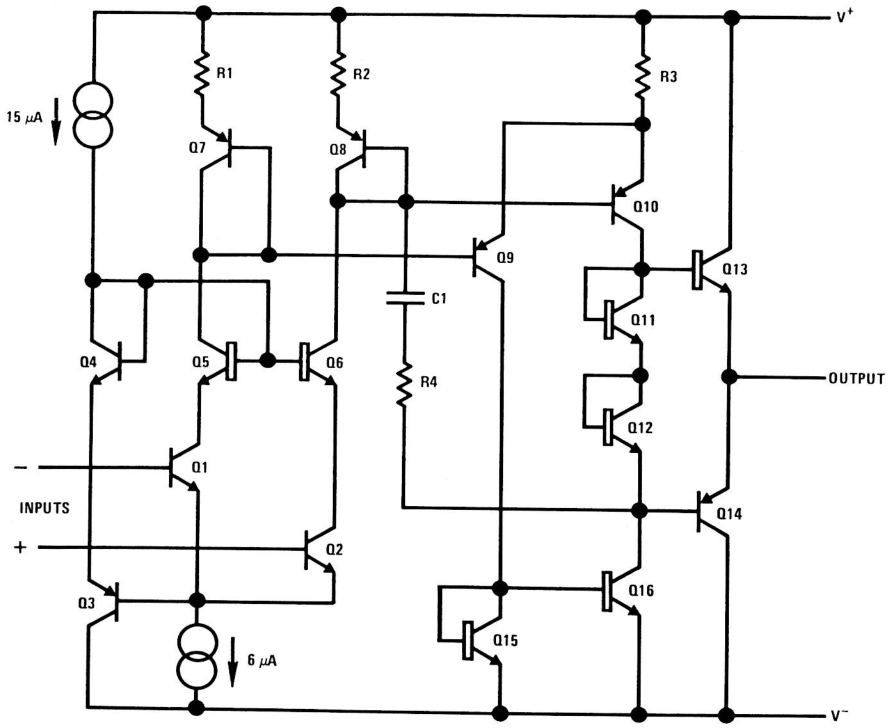

Fig. 11. General-purpose operational amplifier using bootstrapped high-gain transistors in the input stage.

temperature goes as the square of the current gain. Since the gain of integrated-circuit transistors falls off by a factor of 2 to 5 going from 25°C down to −55°C, the input currents obtained at the minimum operating temperature are considerably higher than at room temperature.

Other limitations of Darlingtons are that they have higher noise, lower common-mode rejection, and reduced common-mode slew rate. Further, they have one-half the transconductance of a simple differential stage; this doubles the effect of dc offset terms in their output circuitry.

To summarize, Darlingtons can give typical input currents competitive with super-gain transistors. However, if the full range of production variables is taken into account, along with −55 to 125°C operation, the performance is degraded considerably (or the yields reduced) both with respect to offset voltage and input current.[6] This is substantiated by the fact that a large number of IC operational amplifiers using Darlingtons have been marketed, but none have become industry standards. Nonetheless, the Darlington connection might become useful to compound super-gain transistors with standard transistors. This approach could give input currents less than 50 pA over a 0 to 70°C temperature range, which equals the best FET's. The offset voltage would not be as good as super-gain transistors alone, but it would be substantially better than monolithic FET's. It should be emphasized that this approach is not likely to work at temperatures much above 70°C.

Present data indicate that super-gain transistors are indeed the most effective and economical solution to the problem of making high performance integrated operational amplifiers. This approach cannot be considered to be speculation since an integrated voltage follower, the LM102 [10], which uses super-gain transistors to provide input bias currents less than 20 nA over a −55 to 125°C temperature range has been in volume production since early 1968.

Incorporating super-gain transistors in the input stage of a general-purpose operational amplifier is not as easy as in a voltage follower. Because the circuit must operate over a wide range of common-mode voltages, there is no simple way to operate the input transistors at zero collector–base voltage. However, this is certainly not impossible; and one circuit for doing it is shown in Fig. 11. The design uses more active components than a standard design, but this poses no problems in a monolithic circuit.

The input pair, Q_1 and Q_2, is operated in a cascode connection with Q_5 and Q_6, which stand off the common-mode voltage. The bases of Q_5 and Q_6 are bootstrapped to the common-mode voltage seen by the input transistors by Q_3 and Q_4. Hence, the input transistors are always operated with near-zero collector–base voltage.

[6] Modified Darlingtons, where the operating current of the input transistors is made large by comparison to the base current of the output transistors, do not suffer from problems caused by current-gain matching.

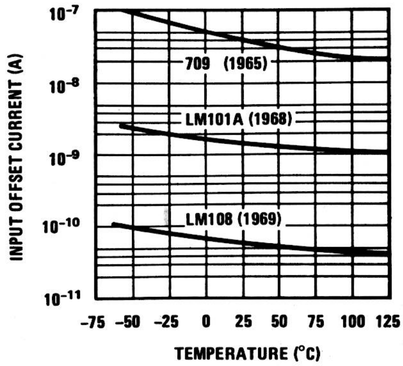

Fig. 12. Illustrating improvements in input current specifications.

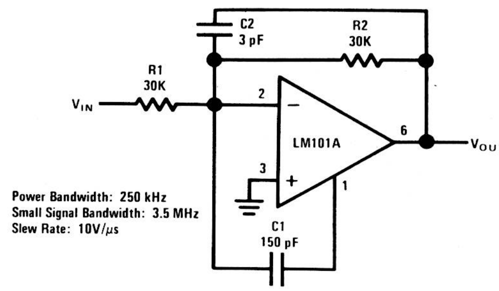

Fig. 13. Fast compensation for the LM101A in summing amplifier applications.

Fig. 12 shows the results of using this approach. The input currents are more than ten times better than the LM101A which, itself, is pretty close to the limits of what can be done with conventional transistors—FET or bipolar.

Fast Amplifiers

The ideal operational amplifier has been described as one with zero offset voltage, zero input current, infinite bandwidth, zero standby current drain, etc. No one can dispute that zero offset, no input current, infinite gain, and the like are desirable characteristics. However, a little practical experience quickly demonstrates that infinite bandwidth is not an unmixed blessing. High-speed amplifiers are definitely ore difficult to use. Capacitive loading, stray capacitances, improper supply bypassing, and poor physical layouts can all cause oscillation problems. Furthermore, fast amplifiers will, in general, have considerably higher power consumption; and it is harder to get effective protection of the input and output without adversely affecting the stability.

Most applications require an operational amplifier with excellent dc characteristics but moderate high-frequency performance, so that a reasonable amount of slop can be tolerated in the physical layout before oscillation problems are encountered. However, there are a substantial number of applications where fast operation is definitely needed. It should be possible to make better high-frequency amplifiers with monolithic technology than with discretes. The basic reason is that stray and wiring capacitances can be virtually eliminated; an integrated circuit transistor has a collector–base capacitance of less than 0.1 pF. This value cannot be approached with discretes, especially if one considers the package and wiring capacitances. All that is needed for monolithic circuits to excel is a suitable circuit design.

There are two major parameters to consider in the design of a fast operational amplifier: small-signal bandwidth and slew rate. The major problem encountered in trying to improve small-signal frequency response has been the poor frequency characteristics of the lateral p-n-p [11], which has been used for level shifting. The response above 2 MHz, especially the excess phase shift, gets so bad that it cannot be used in a feedback amplifier at higher frequencies. A definite connection between the small-signal bandwidth, the unity gain slew rate, and the transconductance of the input transistors can be demonstrated [12] for amplifiers using 6 dB/octave rolloff.[7] With a simple, differential input stage using bipolar transistors, the p-n-p bandwidth forces the slew rate to be less than 1 V/μs. Higher slew rates can be obtained by reducing the input stage transconductance with emitter degeneration resistors, although this will reduce gain, decrease common-mode rejection and increase the offset voltage somewhat. It is possible to obtain any desired slew rate with degeneration, but slew rates much above 5 V/μs are not of much practical value. With faster slew rates, high-frequency gain error takes over as the biggest problem; and there is not much improvement in the end result. For example, with equal bandwidths and a 2-volt output step, an amplifier with a 50-V/μs slew would take about as long to settle within 1 percent of final value as an amplifier with a 5-V/μs slew. Hence, slew rates much above 5 V/μs will not be too useful without a proportionate increase in bandwidth.

Feedforward compensation [13] may be used to improve the high-frequency performance of standard amplifiers, in certain configurations. An example of this is shown in Fig. 13. An LM101A can be compensated by bypassing the lateral p-n-p to get slew rates of 10 V/μs, a power bandwidth of 250 kHz, and a small-signal bandwidth as high as 10 MHz. Further, high-frequency gain error is reduced (for example, by a factor of 50 at 100 kHz compared to standard compensation) so it is possible to take advantage of the faster slew. The feedforward compensation, which is no more complicated than standard compensation, provides more than an order of magnitude better performance in high-frequency applications. However, feedforward only works here when the device is used as a summing amplifier where the compensation capacitor does not have to be charged and discharged to swing the output.

[7] This relation does not apply if compensation is applied to the input terminals, but input compensation is not satisfactory in the majority of applications.