LM109 three-terminal voltage regulator

A look at the design of one of the classic i.c. regulators

by **John L. Linsley Hood**

A description is given of the circuit techniques employed to produce a monolithic three-terminal voltage-regulator i.c., having a very good stability output voltage, and also an internal protection against both output short circuits and thermal overload. The theoretical basis of the energy-band gap voltage reference circuit is examined in an appendix.

In the first article of this series, it was suggested that a closer look at the internal circuitry of some of the common 'building block' integrated circuits could be instructive on two counts – that a better knowledge of the way in which the circuit functioned could be helpful in employing it to its best advantage, and that, for anyone with an inclination to try their hand at circuit design, such circuits were a treasure trove of elegant and innovative design techniques.

However, while there are – literally – thousands of different integrated-circuit designs in common use, there are a lesser number whose availability has had a major impact on the way in which 'linear' circuit designers implement their thoughts, and, of these circuits, next to the ubiquitous operational amplifier, must stand the three-terminal i.c. voltage regulators, of which the LM 109 was one of the earliest, and remains a classic design.

I am restricting my field of view, in this context, to linear or analogue electronics, in that this is still a field which is largely dominated by circuitry assembled from discrete components, and in which the appearance of separate i.c.s. is a departure from previous practice, whereas in the area of digital electronics the presence of anything other than an i.c. – with the possible exception of the odd supply line decoupling capacitor – would be unusual and noteworthy.

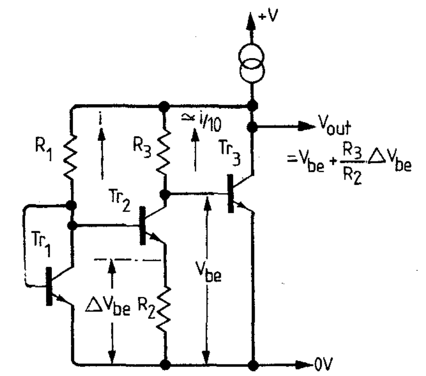


Fig. 2. Temperature-compensated 'band-gap' voltage reference.

In the light of this restriction, it is, perhaps, a little ironic that the i.c. voltage regulator mainly came into existence to meet the need of the t.t.l. logic elements (which were used in vast numbers in the early digital computer circuits, and which were both extravagant in their use of supply current and critical in the magnitude of the supply voltage) for a local 'on-board' voltage regulator circuit which would simplify the provision of an accurate +5 supply in spite of predictable voltage drops along the incoming power supply lines.

The existing 5 volt standard for the supply voltage to t.t.l. digital i.cs. coupled with the relatively heavy current demand imposed by the accumulation of fifty or so on one board, gave rise to some design thinking which has had a lasting effect on the internal circuitry of these voltage regulator units. In particular, the low output-voltage specification made it inconvenient to employ the circuit structure of the typical discrete component voltage regulator of the type shown in Fig. 1, since it is impracticable to make a temperature compensated Zener diode which operates at much less than 6 or 7 volts. Also, Zener diodes have a significant noise component in their operating voltage, which would be difficult to eliminate without the use of substantial values of capacitance, not readily available on a silicon chip.

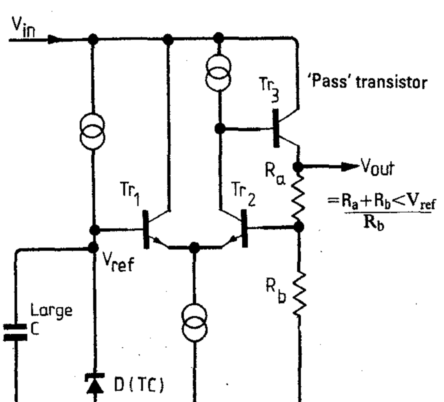


Fig. 1. Simplified arrangement of conventional discrete-component voltage regulator

For these reasons, the voltage against which the voltage output of the circuit is compared is normally derived from a 'band-gap' voltage reference. This takes its name from the energy-band gap of the semiconductor material at 0° Kelvin, V_{go}, which, for silicon, is 1.205V. If, as in the circuit of Fig. 2, two identical transistors, Tr_1 and Tr_2 are operated at substantially different collector currents, so that the current through Tr_1 is, for example, ten times greater than that through Tr_2, there will be a potential developed across R_2 in the emitter circuit of Tr_2 (ΔV_{be}), which is equal to the difference between these two base-emitter voltages. Since this potential has a positive coefficient of voltage against temperature, it provides an elegant means of generating a voltage reference with a near zero temperature coefficient. Referring again to Fig. 2, if the current gain of Tr_2 is sufficiently high, the voltage developed across its load resistor R_3 will be $R_3\Delta V_{be}/R_2$. The total voltage across the circuit, from the positive end of R_3 to the 0V line, will then be $V_{be} + R_3\Delta V_{be}/R_2$, and if these two component voltages, one with a positive temperature coefficient and one with a negative, add up to the 'band-gap' potential of 1.205 volts, the output voltage will have a zero temperature coefficient. This result is analysed in Appendix 1.

Because integrated-circuit voltage stabilizers contain the 'pass' transistor within the package, and are therefore likely to become hot in use, this temperature stability of output voltage is a very important

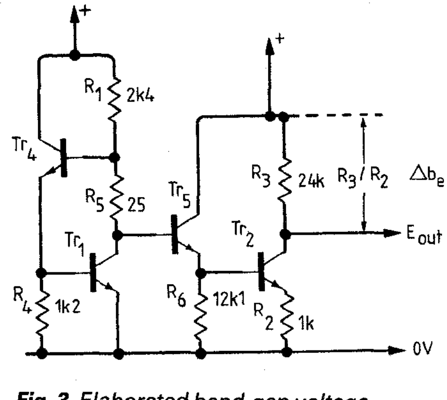


Fig. 3. Elaborated band-gap voltage reference circuit.

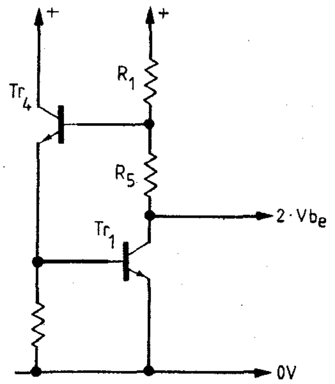


Fig. 4. Rearrangement of circuit of Tr_1 to give lower output impedance.

consideration. Moreover, since the operating junctions are all forward biased, they have a low noise component of output voltage, which eliminates the need for smoothing. Also, the characteristics of forward junctions are much more predictable and controllable in a production environment than highly doped, reverse-breakdown Zener diodes. For these reasons, this type of 'band-gap' stabilizer has been widely adopted as the voltage reference in i.c. voltage regulators, although the transmutation of the circuit layout often makes it difficult to recognize as such.

In the case of the LM 109, the band-gap reference circuit is elaborated into the form shown in Fig. 3, which bears little superficial resemblance to that of Fig. 2. However, it can be disentangled. For example, the original forward-biassed diode-connected transistor, Tr_1 in Fig. 2, has become the two-transistor buffered version of Fig. 4, with an output voltage of twice V_{be}. This is buffered through Tr_5, an emitter follower, to Tr_2 as before. As in the first example, the values of R_1 and R_3 are chosen so that the current through Tr_1 is some ten times greater than that through Tr_3. The small resistor R_5 in the collector circuit of Tr_1 is added to compensate for changes in the transconductance of Tr_1 and make the performance of the circuit less dependent on supply voltage and the absolute values of the resistors fabricated on the chip.

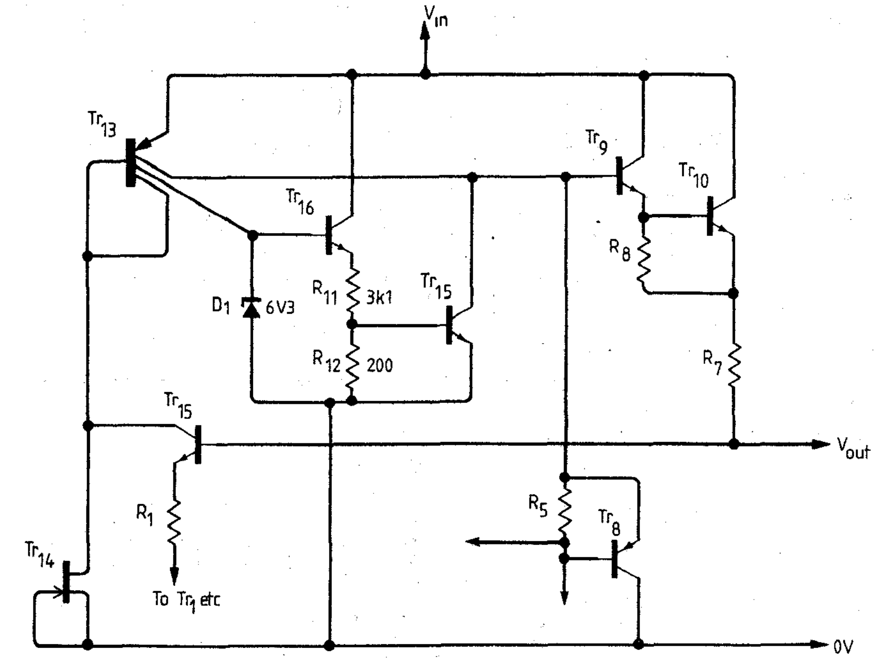


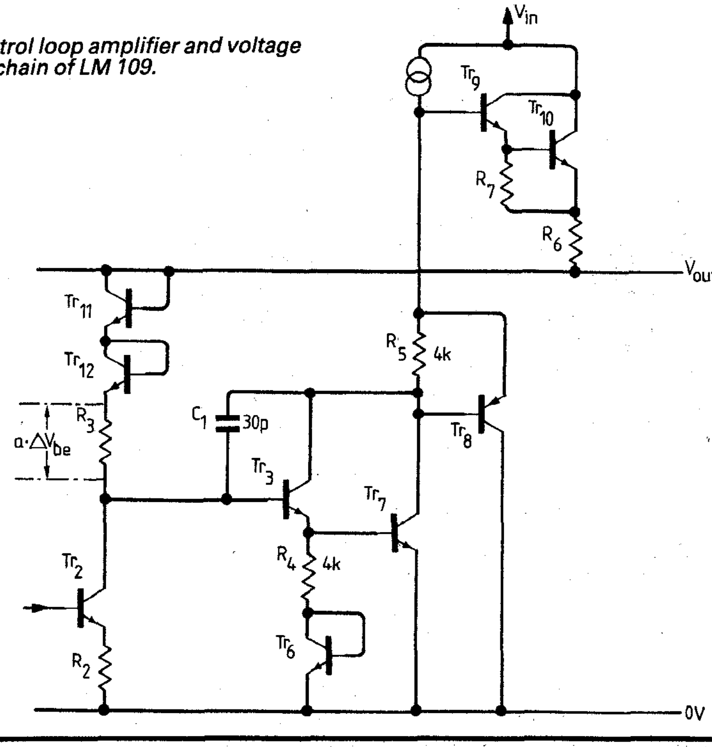
Fig. 6. Thermal-overload shutdown circuitry of LM 109.

In the final circuit of the voltage reference and control loop of the regulator, shown in Fig. 5, the simple band-gap reference system of Fig. 2 has been modified yet again so that the positive temperature coefficient, magnified base-emitter differential potential developed across $R_3(a.\Delta V_{be})$ now appears in series with the four base-emitter junctions (Tr_{11}, Tr_{12}, Tr_3 and Tr_7) forming the reference potential divider chain. Transistors Tr_3, Tr_7 and the 'd.c.-bootstrap' transistor Tr_8 form a very-high-gain inverting amplifier driving the Darlington-connected output pass-transistor pair Tr_9 and Tr_{10}. Transistor Tr_6 compensates for the base-emitter potential differential between Tr_3 and Tr_7, and Tr_8 increases the load impedance seen at the collector of Tr_7 and increases the amplifier stage gain.

Because of the high loop gain of the system, some form of h.f. stabilization is necessary to ensure loop stability on an open circuit load. This is achieved by the 'dominant lag' capacitor, C_1, of 30pF value, connected between collector and base of Tr_3. Although the emitter load of the emitter follower Tr_8 is shown simply as a constant-current source, it does, in reality, employ a circuit artifice much beloved of i.c. designers, the multiple-collector 'current mirror', (Tr_{13}) shown in Fig. 6.

Part of the current which is fed to the base and collector of this transistor, connected in parallel, is derived from the common-base connected device (Tr_{15}) whose emitter feeds the Tr_1/Tr_4 compound diode shown in Fig. 4, and described earlier. Since this is a resistor-diode chain connec-

Fig. 5. Control loop amplifier and voltage reference chain of LM 109.



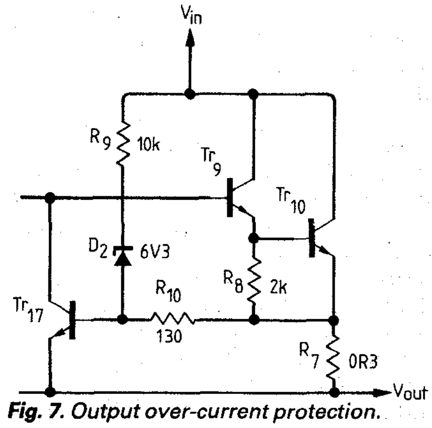


Fig. 7. Output over-current protection.

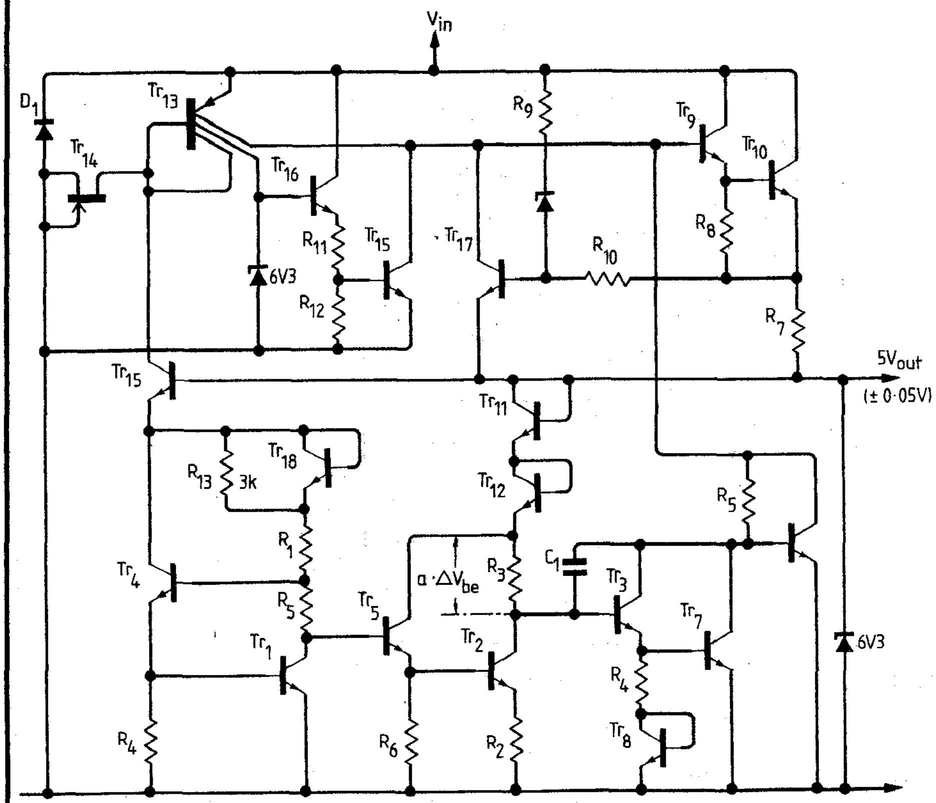


Fig. 8. LM 109 three-terminal voltage regulator.

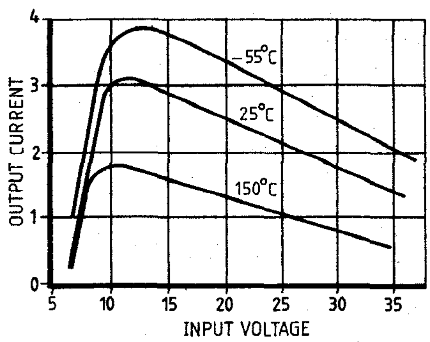


Fig. 9. Output current characteristics.

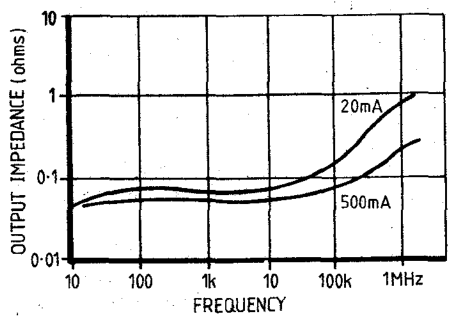


Fig. 10. Output impedance characteristics.

ted across a constant voltage source (the stabilized output voltage of the regulator circuit), it will serve quite well as a constant-current input to the current mirror transistor Tr_{13}. Unfortunately, this would not be self starting on switch-on, so a simple f.e.t. current source, Tr_{14}, is connected in parallel with it.

In the earliest designs of integrated circuit voltage regulator, the pass transistor was kept as an external component to the regulator, in order to minimize the heat dissipated in the regulator chip. With the advent of predictable, well compensated, band-gap voltage reference sources, this problem became less acute, and the way was opened to build onto the chip a temperature-sensitive overload shut-down circuit, so that any combination of inadequate heat sinking, excessive input voltage, or prolonged output current drain which caused the chip to approach its maximum permitted operating temperature, would cause the regulator to shut down. This function is neatly accomplished by the combination of ZD_1, Tr_{15} and Tr_{16} shown also in Fig. 6. This applies a d.c. forward potential of some 0.33 volts to the base of Tr_{15}. When the chip temperature reaches about 180°C, this transistor conducts, and clamps the output Darlington pair to the 0 volt line.

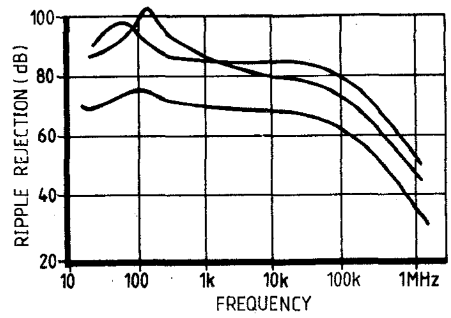
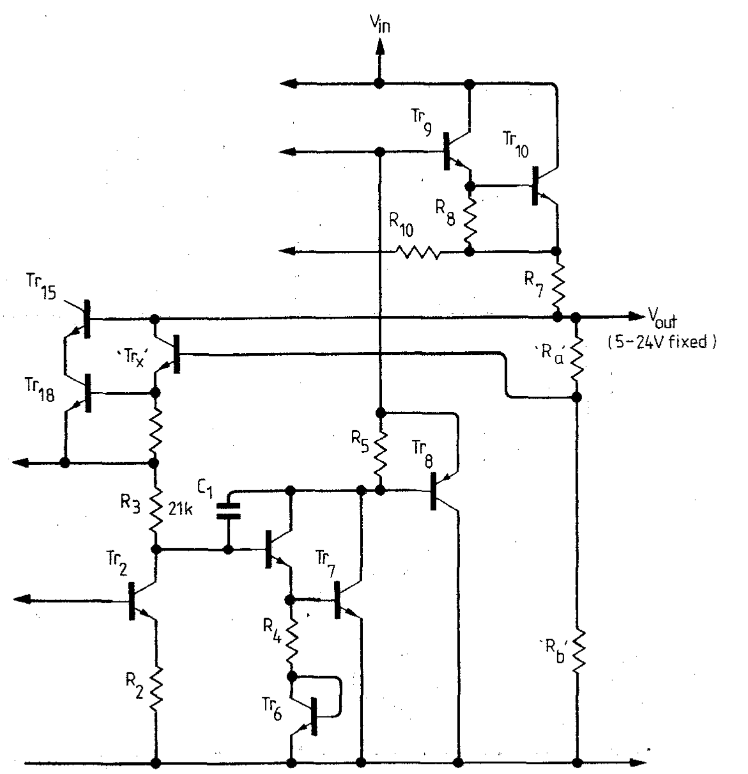


Fig. 11. Ripple rejection of regulator i.c.

While this will prevent the regulator chip from being damaged by thermal overload, it would be of little value against a condition of instantaneous excess output current, due to an output short-circuit. This function is provided by a series resistor in the output circuit, R_7, in combination with a transistor Tr_{17}, arranged as shown in Fig. 7. When the output current drawn through R_7 exceeds about 2 amperes, Tr_{17} conducts, and causes the regulator to operate in a constant-current mode. Further, if the voltage difference between input and output exceeds some 7 volts, the Zener diode ZD_2 will conduct, and progressively reduce the permitted maximum output current.

The circuit of the complete regulator is shown in Fig. 8. The only two features not already covered are the output Zener diode, ZD_3, also 6.3 volts for convenience

Fig. 12. Modification to LM 109 circuit to permit range of fixed voltage regulators, as in 78xx series.



in fabrication, included to prevent the regulator from being embarrassed by voltage transients fed into the regulator through its output line, and the on-chip diode, D_1, included to prevent damage through a reverse polarity input supply voltage.

Although the number and variety of three-terminal fixed, and variable, voltage regulator i.cs has multiplied since the introduction of the National Semiconductors LM 109 in the early 1970s, the general circuit techniques and performance characteristics of these newer types have a lot in common with their forerunners. Because of the relatively low cost and convenient packaging of many of these i.cs, it is now a practical proposition to include these devices in analogue circuitry wherever supply line isolation is desirable, and where, in the past, large value smoothing or decoupling capacitors would have been obligatory, permitting not only a reduction in cost and bulk of the equipment, but also, in many cases, an improvement in circuit performance.

The claimed output current, output impedance and ripple rejection characteristics of this i.c. are shown in Figs. 9, 10 and 11.

Appendix 1

The voltage which will appear across a forward biased base-emitter junction in a transistor can be defined by the equation

$$V_{be}=V_{go}\left(1-\frac{T}{T_o}\right)+V_{beo}\cdot\frac{T}{T_o}+\frac{nkT}{q}\log_e\frac{T_o}{T}+\frac{kT}{q}\log_e\frac{I_c}{I_{co}} \quad (1)$$

where V_{go} is the energy-band-gap voltage of the semiconductor material at 0°K, k is Boltzmann's constant, q is the charge on the electron, n is the operational constant determined by the method of construction of the device (about 1.5 for a double diffused n-p-n structure), V_{beo} is the extrapolated base-emitter voltage at zero forward junction current and T_o, and I_c is the collector current.

The difference in the forward base-emitter voltage between two identical transistors operated at different collector currents could be derived from two such equations, by subtraction. Since it is only the last term in the equation (1), above, which is dependent on collector current, all the other terms cancel leaving only the equation

$$\Delta V_{be}=\frac{kT}{q}\log_e\frac{J_1}{J_2} \quad (2)$$

where J_1 and J_2 are the current densities through the base-emitter junctions of Tr_1 and Tr_2. If the transistors are of identical area, the terms J_1 and J_2 can be replaced by I_{c1} and I_{c2}. Restating the equation, therefore,

$$\Delta V_{be}=\frac{kT}{q}(\log_e I_{c1}-\log_e I_{c2}) \quad (3)$$

This has a positive temperature coefficient so long as the collector current of Tr_1 is greater than that of Tr_2, which is the only practical possibility in the operation of Fig. 2.

Because kT/q is only of the order of 0.025, the last two terms of equation (1) can be neglected in the practical determination of the total voltage produced by the negative temperature coefficient term V_{be} and the positive temperature coefficient term ΔV_{be} which will be

$$V_{out}=V_{go}\left(1-\frac{T}{T_o}\right)+V_{beo}\frac{T}{T_o}+a\frac{kT}{q}\log_e\frac{I_{c1}}{I_{c2}} \quad (4)$$

in which 'a' is a circuit magnification factor.

Differentiating this with respect to temperature yields

$$\frac{\delta V_{out}}{\delta T}=-\frac{V_{go}}{T_o}+\frac{V_{beo}}{T_o}+a\frac{kT_o}{q}\log_e\frac{I_{c1}}{I_{c2}} \quad (5)$$

for zero change in output voltage as a function of temperature, the right hand side of this equation should equal zero, from which

$$V_{beo}+a\frac{kT_o}{q}\log_e\frac{I_{c1}}{I_{c2}}=V_{go}=1.205\text{ volts}, \quad (6)$$

so that for some predetermined ratio of collector currents in Tr_1 and Tr_2, it should be possible to find a circuit magnification factor ('a'), usually in the range 15-30 if $I_{c1} = 10I_{c2}$, which will make the sum of these two potentials add up to the energy-band-gap potential of 1.205 volts, at which the condition for a zero temperature coefficient will be met.

Appendix 2

Although, as mentioned above, the LM 109 was specifically designed to meet the problems inherent in the need for good temperature stability in a regulator circuit having a low output voltage, which is to say at voltage levels below those for which temperature-compensated Zener diodes would be available, the success of this design was such that it was extended to cover higher-output fixed voltage units, such as the 78** and 78M** series, covering the voltage range 5-24 volts, by the simple expedient of replacing the upper two diodes in the voltage reference chain(Tr_{11} and Tr_{12} in Fig. 5), by a conventional voltage divider, as in Fig. 1 (R_a and R_b), and an emitter follower output, (Tr_x).

This small modification is shown in Fig. 12. Clearly, by an appropriate choice of R_a, R_b and the temperature compensating resistor R_3, this will allow the manufacturer to choose the required output voltage at will. Inevitably, there is a small penalty to be paid in a slightly lower precision of voltage control as the ratio of output voltage to reference voltage is increased. However, the other excellent performance characteristics of the circuit remain unimpaired.

Literature Received

A wallchart from Ferranti gives brief specifications of their products in **data conversion**. It includes fixed and trimmable voltage references, monolithic digital voltmeters, and A-to-D and D-to-A converters. A glossary of the terms used is included. Copies of the wallchart may be obtained from: The Publicity Department, Ferranti Electronics Ltd, Fields New Road, Chadderton, Oldham, Lancs OL9 8NP.

WW401

Chart recorders that can cope with Z-fold paper as well as rolls are described in a leaflet published by Allen Datagraph and available from Techmation Ltd, 58 Edgware Way, Middlesex HA8 8JP. The recorders, from the 2100 series are available in one- or two-pen models with 13 calibrated spans and ten chart speeds.

WW402

The first five parts of a new Standard, BS 6160: **Methods of measurement for radio equipment used in mobile services,** has been published by BSI. The first part is concerned with general definitions and standard conditions of measurement, the others relate to specific equipment: A3 or F3 transmitters, A3 or F3 receivers, s.s.b. transmitters (A3A, A3H or A3J) and s.s.b. receivers for the same bands. The Standard is identical with IEC Publication 498-1 to 5. British Standards Institution, 101 Pentonville Road, London N1 9ND.

WW403

The KEF Constructor series of leaflets (CSI to 5) give designs for the construction of five **loudspeaker** enclosures using KEF drive units. Details are given of the construction and the circuit diagram for suitable crossover networks are provided. The designs cover a range of sizes from a miniature bookshelf loudspeaker to a large three-way system. KEF Electronics Ltd, Tovil, Maidstone, Kent ME15 6QP.

WW404

D.c.-d.c. converters in the Avel C101 range provide 12 or 24V outputs from supply voltages between 12 and 220V d.c. They are described in a leaflet available from Avel-Lindberg, South Ockendon, Essex RM15 5TD.

WW405

113 pages are needed by Harting to describe their ranges of **multi-pole electrical connestors.** The plugs and sockets range from low current/low voltage applications up to 35A and 750V requirements. Included are a variety of accessories an a users' guide. The manual is available free from Harting Electronik Ltd, Airport Estate, Biggin Hill, Kent.

WW406

C & K have produced so many new **switches** that they have issued a new product supplement to their catalogue. It includes solid state push buttons, d.i.p. slide and toggle, rotary and rocker switches. Copies are available from Roxburgh Switches Ltd, 22 Winchelsea Road, Rye, E. Sussex TN31 7BR.

WW407