AN-462
Application Note

FET CURRENT REGULATORS — CIRCUITS AND DIODES

Prepared by
Applications Engineering

Two terminal current regulators are having a significant impact on today's electronic technology. This paper provides a brief historical sketch of the development of such devices from vacuum tube days to present. Included are numerous FET current sourcing circuits, along with an extensive treatment of the current regulating diode and its uses as a valuable component in circuit design.

MOTOROLA Semiconductor Products Inc.

FET CURRENT REGULATORS — CIRCUITS AND DIODES

INTRODUCTION

With the present rapid advancement of semiconductor technology, it was inevitable that a two terminal current regulator be developed. The present state of the art in current regulator diodes is merely the beginning of a wide range of current and power handling capabilities. Herein we touch upon the early development of current regulating devices; first the bipolar transistor circuits, then the original diode, and the potted circuits. Finally, field effect transistor circuits and the field effect diode are extensively treated. Included are numerous applications to spark the imagination.

HISTORICAL SKETCH OF CURRENT SOURCES

The constant current device in its original form was simply a high resistance in series with a power supply. A variation of the original sources was the vacuum tube. The most popular tube for this application was the pentode because of its large plate resistance and high voltage capability.

In the early 1950's, with the development of the transistor, current sources were reduced in physical size and improved in power efficiency. The number of terminal connections for the pentode, seven maximum, was reduced to three and even two. Primarily, the reduction in size allowed the encapsulation or potting of circuit assemblies which permitted their use as a unit circuit component.

The constant-current-diode was first introduced at the British Physical Society Exhibition in 1960.[1] As far as this author has been able to find out, it was only a working hypothesis. As a laboratory substitute the reverse leakage of a large germanium junction was used. This was a poor substitute; however, it did allow circuits to be constructed, thereby proving design concepts. Since these diodes were of a lesser quality than desired, they did not become popular.

Field-effect transistors, being analogous to the vacuum tube in many ways, gave rise to a variety of circuits which were quite acceptable as current sources, regulators, and limiters. One of the more important parameters of the junction field-effect transistor, I_{DSS}, is a measure of the drain current flowing with the gate-source shorted. Under this test condition the drain current remains significantly constant over a wide range of drain-source voltages. Coupled with the low output conductance, these devices make good current sources. Consequently, the natural evolution produced a constant current, two-terminal device, simply a junction field-effect transistor with an internal gate to source connection, the "constant-current diode".

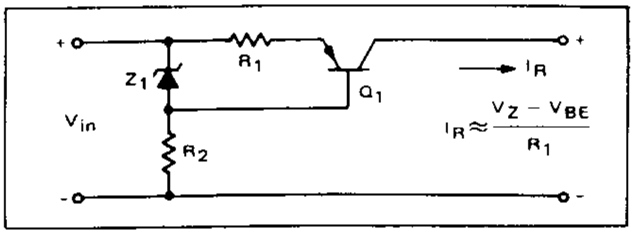


FIGURE 1 — Simple Series Regulator

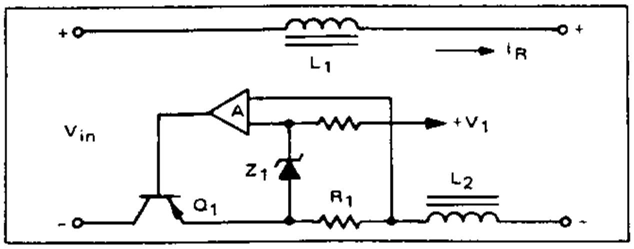


FIGURE 2 — Common-Emitter Series Regulator

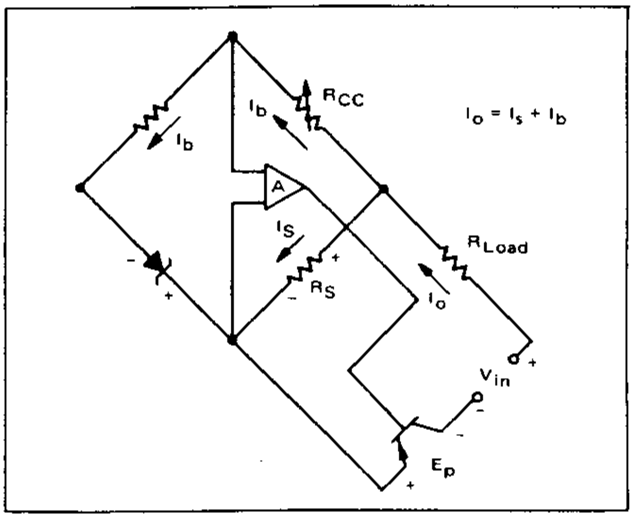


FIGURE 3 — Bridge-Balancing Regulator

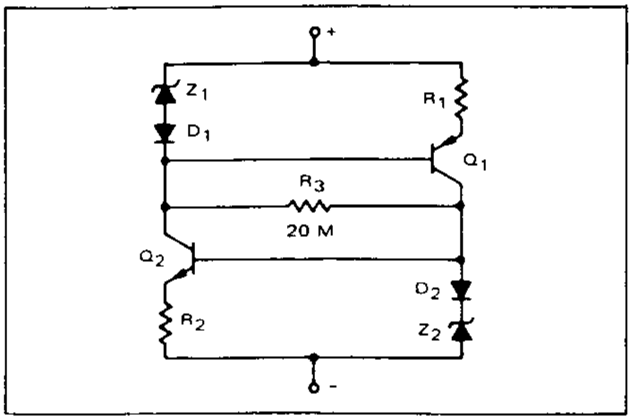


FIGURE 4 — 2-Terminal Constant-Current Device

Although field-effect transistors exhibit degrees of current limiting, these diodes have been especially designed for maximum impedance over the operating range, broad voltage range, and packaged for convenience in installation and space saving.

BIPOLAR TRANSISTOR CURRENT-REGULATING CIRCUITS

The emphasis in voltage regulators is given to the reduction of output impedance. In contrast, the current regulator is constructed to have the highest possible output impedance. Instead of sensing the voltage at the output terminals, the regulated current is passed through a resistor and the voltage developed across the resistor is compared with a voltage reference. Any voltage difference is then amplified and applied to a series regulating transistor which completes the regulating feedback connection by controlling the regulated current flow.

Three representative current-regulator circuits are illustrated in Figures 1, 2 and 3. The first regulator (Figure 1) utilizes R_1 as the current measuring resistor whose voltage is applied to the emitter of Q_1. The voltage from the reference diode, Z_1, is applied to the base of Q_1 so that Q_1 is itself the comparison element. The result is that Q_1 attempts to pass a current which will keep the voltage across R_1 closely equal to the voltage across Z_1.

Performance of this simple circuit is limited because the current measured by R_1 is not exactly the output current, since it also contains the base current of Q_1 and does not contain I_{CBO}. The only regulating gain is that provided by Q_1 itself. Performance improves as R_1 and the voltage across Z_1 are increased, but the output resistance can never surpass the collector-base resistance of Q_1 which is effectively shunting the output.

The output current, I_R, will be regulated over a voltage range extending from a positive limit of about $(V_{IN} - V_{Z1})$ to a lower limit which can even become negative. This assumes, of course, that Q_1 has suitable collector-base voltage and dissipation ratings. In this case the transistor is acting as a common-base amplifier so that the collector-base voltage rating, which is higher than the collector-emitter rating, may be used as long as the total base circuit resistance is held low enough to prevent instability.

The current regulator in Figure 2 is capable of practically any degree of performance required, provided the regulating amplifier is given the necessary gain and stability. The actual output current is sensed by R_1 whose voltage is then compared with that of the reference, Z_1, which is operated from a separate voltage source. Any error in the two voltages is amplified by the high-gain, direct coupled amplifier and applied to the base of Q_1 which adjusts the current, I_R, to minimize the error. The inductances L_1 and L_2 help to maintain a high output impedance at high frequencies where the regulating gain inevitably decreases and the transistors are unable to maintain a high impedance themselves. The use of inductances here is a direct dual of the use of a shunting capacitor to lower output impedance at high frequencies in a voltage regulator.

The bridge-balancing regulator[2] in Figure 3 is the refinement of the series-sensing type. Here the voltage to be regulated is the drop caused by the load current flowing through R_S, the sensing resistor. The feedback resistor is R_{CC}, a current control. The only current flowing through the current control is the bridge current. When the voltage drop in R_{CC} (due to I_b) equals the drop in R_S (due to I_S), the conditions for bridge balance are met, and a null exists across the input terminals of the error amplifier.

The output current of this circuit can be programmed by varying R_{CC} or changing I_b, either event causing the voltage drop across R_{CC} to change. In order to restore the balance condition, the voltage drop across R_S must change. This happens when the feedback to the "pass" element alters its conduction, so as to permit a change in load current in the proper direction and amount as to restore the balance.

The refinement then, is the capability of varying the output current, without changing the sensing resistance.

A novel variation of Figure 1 is shown in Figure 4. This circuit can readily be used as a component due to its two terminal configuration. The circuit consists of a pnp and an npn current source connected so that each regulates the other's reference. The current through the compensating diodes D_1 and D_2 is the same as the current through the transistor base-emitter junctions, which improves the tracking with temperature. The circuit requires a minimum voltage of $2\,(V_Z + V_D)$ for operation, the maximum is determined by transistor ratings. The value of R_3 is shown as a very high resistance, since its purpose is merely voltage neutralization.

FIELD-EFFECT TRANSISTOR CIRCUITS

Field-effect transistor circuits are very appealing as constant-current sources. Output impedances can range anywhere from the kilohm region to the tens of megohms, depending on the device and configuration used. FETs are available with values of g_{os} ranging from one-half to several hundred micromhos. Saturation voltages, or the minimum operating voltages, are in the region of one volt. The maximum voltage before breakdown occurs presently approaches the one hundred volt level. A nearly zero temperature coefficient is practical if the device is biased properly. Best of all, the FET constant-current source is usually a very simple, easy-to-design circuit. Either a fixed or an adjustable current source can be constructed with nothing more than a FET, a resistor (fixed or variable), and a battery.

Obviously, the simplest constant-current circuit is that of Figure 5, an FET with the gate and source shorted. In this case, the FET operates at I_{DSS}, the zero-bias drain current. The circuit output conductance is equal to the g_{os} of the FET. g_{os} at low frequencies is equal to y_{os}. For higher frequencies, depending on the value of the device output capacitance, series inductance could be added. I_o will change if V_{DS} changes according to the relationship,

$$\Delta I_o = \Delta V_{DS}\, g_{os}. \qquad (1)$$

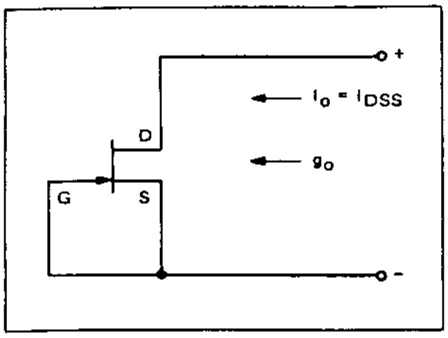


FIGURE 5 – Constant-Current Source Without a Source Resistor

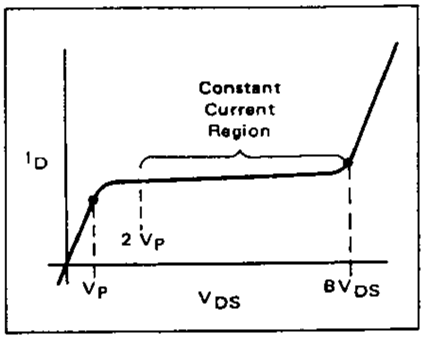


FIGURE 6 – FET Output Characteristic

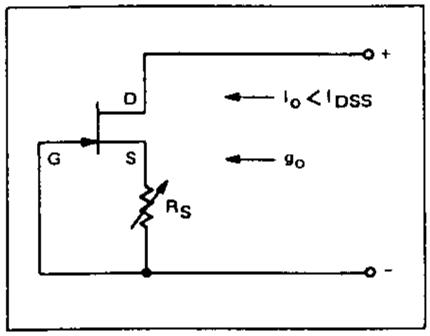


FIGURE 7 – Constant-Current Source With a Source Resistor

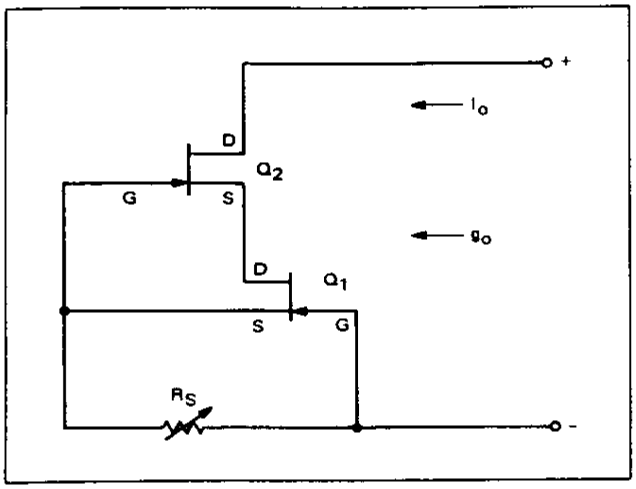


FIGURE 8 – Cascaded FET Contant-Current Source

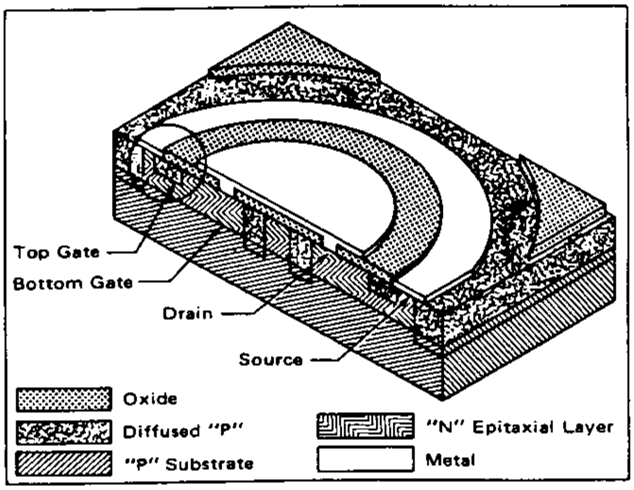


FIGURE 9 – Field Effect Diode

This simple circuit will deliver a relatively constant current from about 2 V_P (the pinch-off or threshold voltage) to BV_{DS} (the breakdown voltage drain to source), as shown in Figure 6.

With the addition of a source resistor shown in Figure 7, the previous circuit becomes capable of supplying any current below I_{DSS}.[3] The approximate value of gate-source voltage, V_{GS}, required for a given operating current, I_o, is:

$$V_{GS} = V_P\left[1 - \sqrt{\frac{I_o}{I_{DSS}}}\right]. \qquad (2)$$

The source resistor, R_S, required is then:

$$R_S = \frac{V_{GS}}{I_o}. \qquad (3)$$

Resistor R_S may be variable to provide an adjustable current source. As R_S is increased and I_o decreased, the FET g_{os} decreases. The circuit output conductance decreases more rapidly than the FET g_{os} because of the feedback action produced across R_S. The circuit output conductance is:

$$g_o = \frac{g_{os}}{1 + R_S(g_{os} + g_{fs})} \approx \frac{g_{os}}{1 + R_S\, g_{fs}}, \qquad (4)$$

where g_{fs} is the real part of y_{fs}, the forward transfer admittance.

If two FETs are cascaded as shown in Figure 8, a much lower g_o value for a particular I_o can be obtained. Here I_o is regulated by Q_1 and $V_{DS1} = -V_{GS2}$. The dc value of I_o is controlled by R_S and Q_1. However, Q_1 and Q_2 both affect current stability.

Where $R_S = 0$

$$I_o = \frac{V_{DS2}\, g_{os1}\, g_{os2}}{g_{os1} + g_{fs2}}, \qquad (5)$$

and

$$g_o = \frac{g_{os1}\, g_{os2}}{g_{os1} + g_{os2} + g_{fs2}}, \qquad (6)$$

if $R_S = 0$ and $g_{os1} \approx g_{os2}$,

$$g_o = \frac{g_{os}}{2g_{os} + g_{fs} + R_S(g_{fs2} + g_{os}\,g_{fs} + g_{os2})},$$

$$\approx \frac{(g_{os})^2}{g_{fs}(1 + R_S\, g_{fs})}. \qquad (7)$$

When designing cascaded FET current sources, care must be exercised to ensure that both FETs are operating with adequate drain-source voltage, preferably $V_{DS} > 2\ V_P$, and that Q_2 has a significantly higher I_{DSS} than Q_1.

Notes worth mentioning—all FET circuits described are two terminal devices and can be easily used as circuit elements; all devices shown are N-channel JFETS. For P-channel devices reverse all polarities.

If a zero temperature coefficient (OTC) is required, each transistor must be operated at a specific current, I_{DZ}, the drain current for OTC, which is given by: [4]

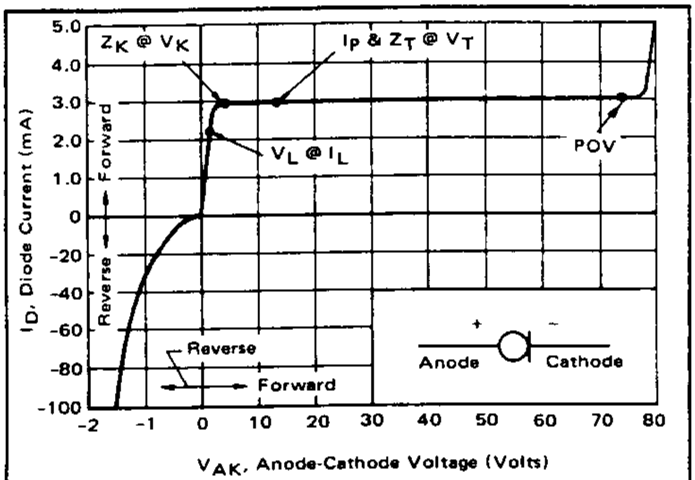
$$I_{DZ} \approx I_{DSS} \left(\frac{0.63}{V_P}\right)^2 . \qquad (8)$$

The gate-source bias voltage required is:

$$V_{GSZ} \approx V_P - 0.63 . \qquad (9)$$

By analyzing equation 8 we find that I_{DZ} increases as I_{DSS} increases. I_{DZ} can be as high as 1 mA for I_{DSS} units of 20 mA.

Operation at $I_D < I_{DZ}$, but near I_{DZ} will yield a positive TC if desired. Conversely, negative TCs will result if $I_D > I_{DZ}$. Negative TCs of 1.2 μA/^{o}C are obtainable at 4 mA currents, with the 2N5361.



SYMBOLS AND DEFINITIONS

I_D — Diode Current.
I_L — Limiting Current: 80% of I_P minimum used to determine Limiting voltage, V_L.
I_P — Pinch-off Current: Regulator current at specified Test Voltage, V_T.
POV — Peak Operating Voltage: Maximum voltage to be applied to device.
θ_I — Current Temperature Coefficient.
V_{AK} — Anode-to-cathode Voltage.
V_K — Knee Impedance Test Voltage: Specified voltage used to establish Knee Impedance, Z_K.
V_L — Limiting Voltage: Measured at I_L. V_L, together with Knee AC Impedance, Z_K, indicates the Knee characteristics of the device.
V_T — Test Voltage: Voltage at which I_P and Z_T are specified.
Z_K — Knee AC Impedance at Test Voltage: To test for Z_K, a 90 Hz signal, v_K, with RMS value equal to 10% of test voltage, V_K, is superimposed on V_K:

$$Z_K = v_K / i_K$$

where i_K is the resultant ac current due to v_K.
To provide the most constant current from the diode, Z_K should be as high as possible; therefore, a minimum value of Z_K is specified.
Z_T — AC Impedance at Test Voltage: Specified as a minimum value. To test for Z_T, a 90 Hz signal with RMS value equal to 10% of Test Voltage, V_T, is superimposed on V_T.

FIGURE 10 — Typical Current Regulator Characteristics

THE FET CURRENT-REGULATING DIODE

The natural evolvement of the circuit in Figure 5 is the current-regulating "diode", (CRD) which is in essence an N-channel JFET with an internal gate-source short. As stated previously, the design of these diodes has been especially optimized for high impedance and current regulating capability.

A semiconductor diode is defined as a two-electrode semiconductor device. Since the FET device met this definition, it was dubbed a diode. Figure 9 shows a typical cross-section of the diode. The circled area shows that the metal distinctively spans the gate-source areas. Typical current regulator characteristics, symbols, and definitions are illustrated in Figure 10.

The equivalent circuit of the diode would simply be either a current generator in series with a parallel combination of Z_T and a capacitance or the same generator shunted by a conductance g_T and the same capacitance. The shunt capacitance associated with Motorola's current regulating diodes is about 6 to 8 picofarads within the useful voltage range of the devices and is relatively constant. Capacitance does tend to increase and peak as the applied voltage nears V_L, but falls to zero as the voltage goes below V_L. Equivalent circuits are shown in Figure 11.

The 1N5283 through 1N5314 is a family of current-regulating diodes that cover a 220 microampere to 4.7 milliampere range in 32 different currents. Performance of several selected diodes is shown in Figure 12. These 10% tolerance units operate over a -55^{o}C to +200^{o}C range and have a moderate temperature coefficient (unless a OTC device is selected) that must be taken into account in precision circuitry which must operate over a significant temperature range. Maximum diode dissipation is a healthy 600 milliwatts derated at 4.8 mW/^{o}C above 75^{o}C T_L, the lead temperature, as explained in Figure 13. The mimimum operating voltage, V_L ranges

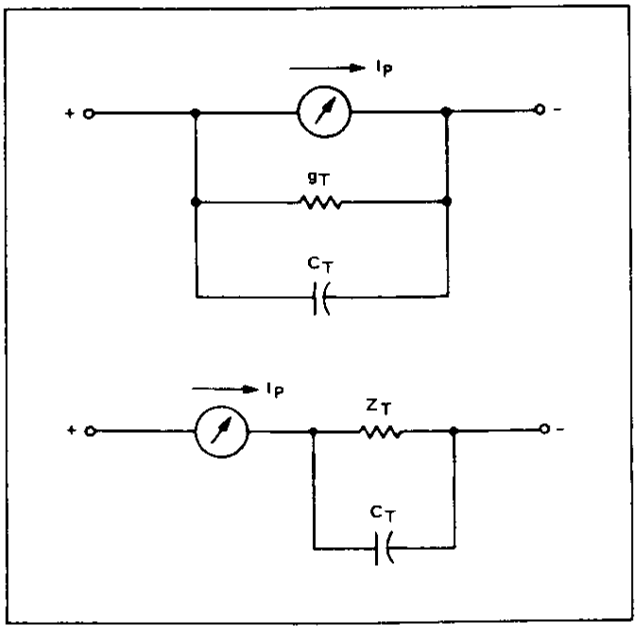


FIGURE 11 — Current-Regulator Diode Equivalent Circuits

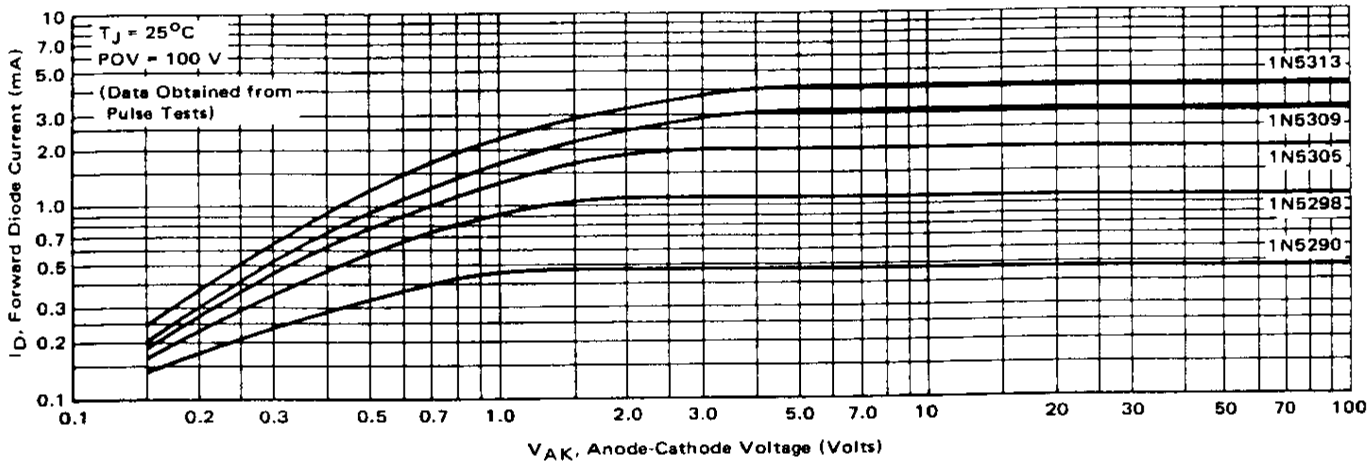
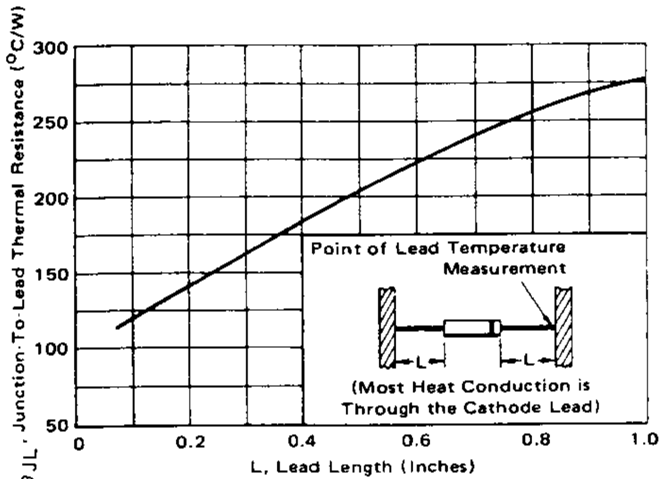


FIGURE 12 — Typical Forward Characteristics



(a) Typical Thermal Resistance

(b) Application Note

As the current available from the diode is temperature dependent, it is necessary to determine junction temperature, T_J, under specific operating conditions to calculate the value of the diode current. The following procedure is recommended:

Lead Temperature, T_L, shall be determined from:

$$T_L = R_{\theta LA} P_D + T_A$$

where $R_{\theta LA}$ is lead-to-ambient thermal resistance, T_A is the ambient temperature

and P_D is power dissipation.

$R_{\theta LA}$ is generally 30-40°C/W for the various clips and tie points in common use, and for printed circuit-board wiring.

Junction Temperature, T_J, shall be calculated from:

$$T_J = T_L + R_{\theta JL} P_D$$

where $R_{\theta JL}$ is taken from Figure 13(a).

For circuit design limits of V_{AK}, limits of P_D may be estimated and extremes of T_J may be computed. Using data sheet information, changes in current may be found. To improve current regulation, keep V_{AK} low to reduce P_D and keep the leads short, especially the cathode lead, to reduce $R_{\theta JL}$.

FIGURE 13 — Thermal Resistance and Power Dissipation Calculations

from 1 to 3 volts, while the forward breakdown voltage is specified at 100 volts for the 1N series, giving a 97 to 99 volt range of constant current operation. Further detailed information is found on the device data sheets.

VARIANCES OF FET DIODE CONFIGURATIONS

Various diode configurations can be implemented to extend their maximum operating voltage and current ranges.

An extension of the dynamic voltage range is achieved by placing the devices in series as shown in Figure 14(a). Here it is necessary to introduce voltage-balancing resistances. The resistive values should be high since they shunt the output resistance. This technique is synonymous to the current-balancing required when paralleling zener diodes.

The current range of the diodes can be extended by the simple expedient of paralleling these devices as Figure 14(b) indicates. No special precautions are required. The resultant current is merely the summation of the individual currents.

There are instances where a bipolar device, one which regulates or limits in both directions, is of use. In this case, the diodes can be connected in series-opposing fashion as shown in Figure 14(c). During the cycle when one is limiting, the other is a forward-biased junction, producing a diode voltage drop.

GENERAL APPLICATIONS

(a) Low voltage references, those below normal zener voltages of about 2.5 volts, or voltage references with good characteristics below about 6 volts can be easily attained using the current-regulator diode in series with a resistor. These diodes serve well as a precision millivolt reference source, where required. The diode simply drives a known resistor, producing an output reference voltage the value of which is determined by Ohm's law as indicated in Figure 15.

(b) As a zener diode source element the current-regulating diode has several prime advantages. Figure 16(a) and (b) illustrates the unmodified and modified zener regulator circuits, respectively.

Distinct advantages of the zener-CRD combination are that the maximum permissible V_{in} is determined by the maximum current-regulating diode voltage (100 volts), rather than maximum zener dissipation, and that variations in V_{in} have virtually no effect on V_{out}. In either circuit

$$\Delta V_{out} = \Delta V_{in} \frac{Z_{ZT}}{R_S + Z_{ZT}} \qquad (10)$$

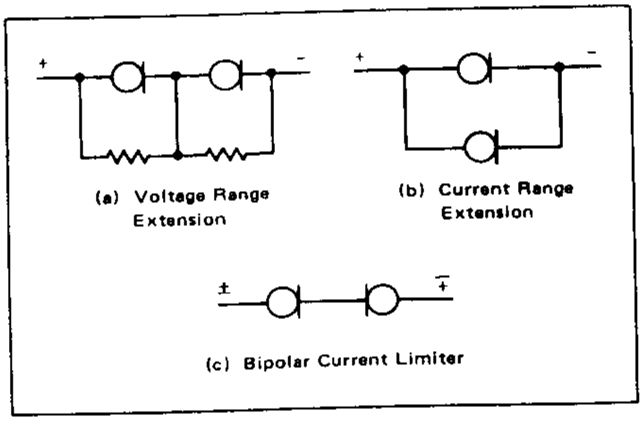


FIGURE 14 – CRD Variations

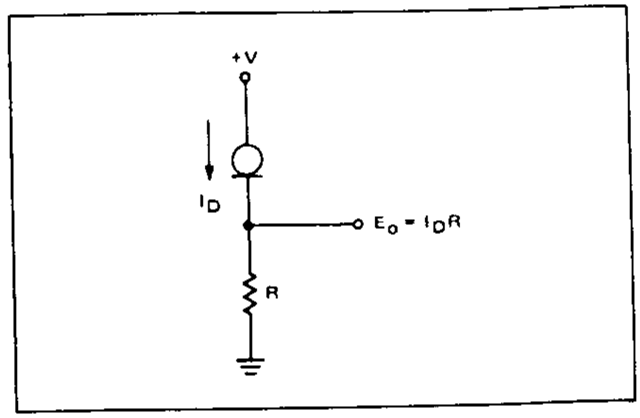


FIGURE 15 – Low Voltage Reference

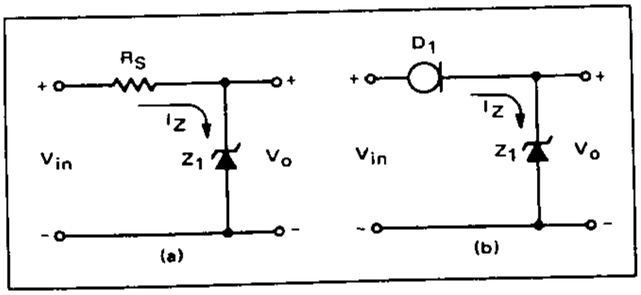


FIGURE 16 – (a) Standard and (b) Modified Zener Voltage Regulator

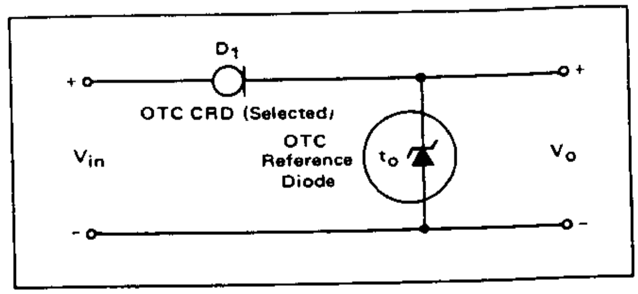


FIGURE 17 – A Most Absolute OTC Voltage Reference

where R_S is the series resistance, either discrete or the dynamic impedance (Z_T) of the CRD, and Z_{ZT} is the zener dynamic impedance.

Since Z_{ZT} is relatively low, it is obvious that as R_S increases in value the regulation, V_{out}/V_{in}, is improved. In practical circuits R_S is normally in the hundreds or thousands of ohms, while the CRD can present megohms. This yields improvements of three to four orders of magnitude.

As V_{in} varies I_Z also varies according to

$$\Delta I_Z = \frac{\Delta V_{in}}{R_S + Z_{ZT}}. \qquad (11)$$

From the relationship given it is obvious that as V_{in} increases, the current through, and consequently, the power dissipated in the zener increases. V_{in} is thus limited by the power rating of the zener. Using a CRD alleviates this problem entirely.

A most absolute voltage reference can be implemented using a zener-CRD combination as Figure 17 suggests. The CRD exhibits a zero temperature coefficient characteristic at currents around 0.5 mA. A Motorola family of OTC reference diodes also operates on zener currents of 0.5 mA. A combination of the two with OTC characteristics, would yield a constant reference voltage between the input voltage limits of 2 V_L + V_Z to POV + V_Z (practically, about 8.4 to 106.4 volts with a 6.4 V zener), and the temperature range of 0°C to 100°C. A change in V_{out} of about 10 mV could be expected over the V_{in} range. A TC of 0.001%/°C could be expected over the temperature range. The devices suggested are the 1N4569A reference diode and a selected 1N5290 current-regulator diode.

When voltages lower than system power supplies are required and a zener is used to provide same, another advantage of the CRD is one of tremendous decoupling of either ripple or noise on the supply lines. Due to the high ratio of the dynamic impedance of the CRD as compared to the zener, an attenuation of about 100 dB can be realized at frequencies up to several hundred kilohertz.

(c) A standard form of dc coupling, as shown in Figure 18(a), is much improved by the substitution of a zener diode for resistor R_1, as illustrated in Figure 18(b). This modification substantially reduces the loss of gain introduced by the coupling circuit. A similar increase in gain is achieved by the substitution of a CRD for R_2, as Figure 18(c) indicates. Furthermore, as might be expected, a coupling circuit with very desirable characteristics results from a combination of these methods. Figure 18(d) shows an amplifier stage incorporating this approach.

(d) The common mode rejection ratio of emitter-coupled differential amplifiers is directly proportional to the common-emitter impedance. For this reason current sourcing in the emitter is commonly used.[5] Figure 19(a) illustrates the use of a resistor voltage combination, while Figure 19(b) is the biased-transistor version. An improvement over both of these methods is attained by the use of a CRD as shown in Figure 19(c).

When high input impedance is required, the differential amplifier incorporates the Darlington input configuration

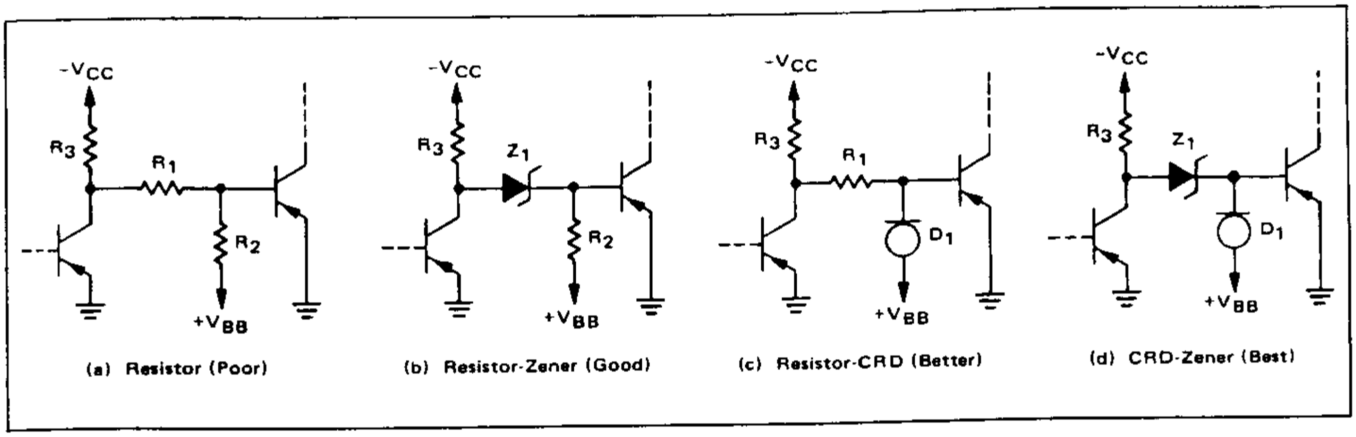


FIGURE 18 – DC Coupling Methods

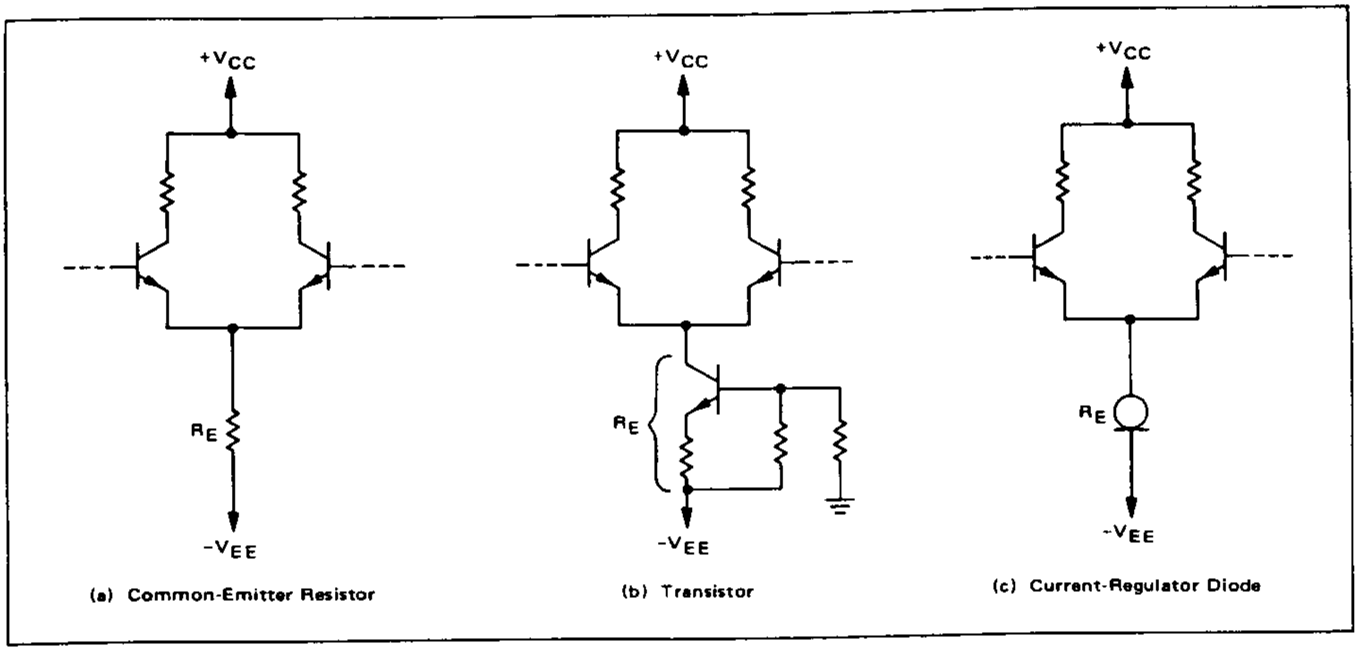


FIGURE 19 – Current Sources in Differential Amplifiers

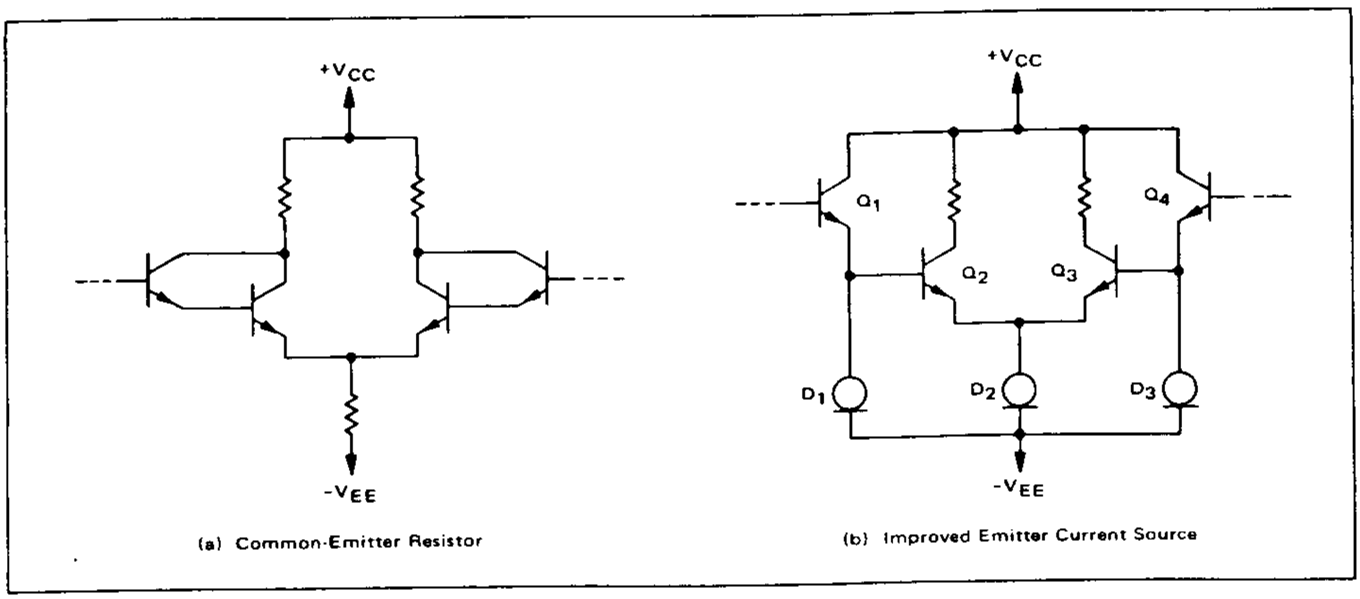


FIGURE 20 – Common-Emitter Current Sourcing for Darlington-Input Differential Amplifier

of Figure 20(a). The modified amplifier of Figure 20(b) has 3 current sources. Current source D_2 provides the high common mode rejection, while D_1 and D_3 provide fixed currents in Q_1 and Q_4. Using this method, the characteristics of transistors Q_1 and Q_4 are not a function of the betas of the input transistors Q_2 and Q_3, as would be the case with a standard Darlington circuit.[6]

(e) The emitter resistor of emitter-followers can often be replaced by a CRD, resulting in a significant increase in input impedance, a gain closer to unity, and less obvious is a lower transistor dissipation when supplying a heavy external load. Figure 21 illustrates an example where an output of 5 volts peak is required in a 600 ohm load, with ± 10 volt supplies. When the input peak is +5 V the load current is 8.33 mA. The drop across R_1 is now 5 V and in order that Q_1 is not cut off, R_1 must pass more than 8.33 mA. Thus R_1 should be less than 600 ohms. Under quiescent conditions, the emitter of Q_1 is near zero and so quiescent current is 16.7 mA and Q_1 is dissipating 167 mW. When R_1 is replaced by an 8.33 mA CRD, the quiescent dissipation will be 83 mW. Thus power consumption is halved, and since the transistor load is 600 ohms instead of 300 ohms, the voltage gain is nearer unity.

Observe that the parallel operation of two CRD's is used to obtain the 8.33 mA.

(f) In amplifier circuits where the collector voltage is defined by an external feedback loop, the collector load resistor can be replaced by a CRD to give a greatly increased voltage gain. Using T equivalent parameters, the voltage gain, Av, of a grounded-emitter amplifier approaches the value given by:

$$Av = \frac{r_c}{r_e}. \tag{12}$$

In order to realize this gain a large collector load is required. This can be attained by replacing the collector resistor with a CRD, resulting in actual voltage gains from 700 to 1000 with currently available small-signal transistors. Voltage gain will be reduced if the amplifier is substantially loaded.

Similar results can be obtained using the CRD as drain elements in FET circuits.

PARTICULAR APPLICATIONS

(a) Simple Sawtooth Generators[7]. The two circuits of Figure 22 make use of the Motorola four-layer and field effect diodes to provide exceedingly simple, fixed-frequency sawtooth generators with linear output waveforms. Shown are circuits for both positive and negative going ramps. The principal design equation for these circuits is:

$$T = \frac{CV_{BR}}{I_P}, \tag{13}$$

where

T = period of one cycle
I_P = pinch-off current of the current limiting diode
C = timing capacitor in μF
V_{BR} = breakover voltage of the four layer diode

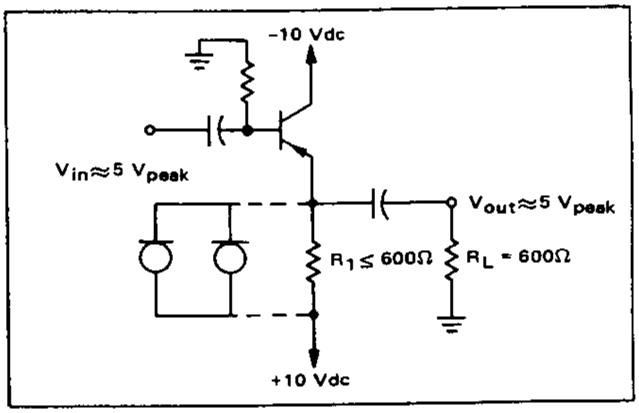


FIGURE 21 — Improved Emitter-Follower

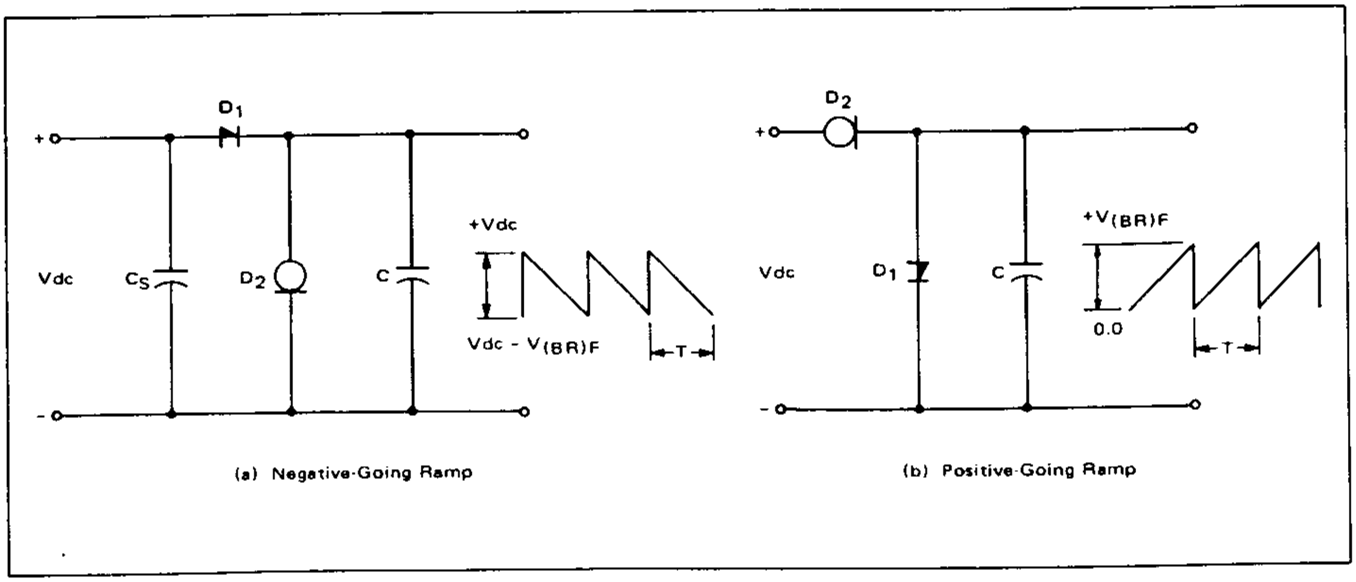


FIGURE 22 — Sawtooth Generator Circuits

(b) Triangular-Wave Generator. The use of the CRD in series–opposing fashion makes possible the generation of a high-quality triangular wave from a sine or square wave source as Figure 23 illustrates. Square wave drive results in a better waveform at the zero crossings. Here the output frequency is identical to the input frequency. The peak to peak amplitude is given by:

$$V_{O\ p\text{-}p} = \frac{It}{C} \tag{14}$$

where $V_{O\ p\text{-}p}$ is in volts, C is in microfarads, I is in milliamperes and t is in milliseconds.

and $V_{O\ p\text{-}p} < V_{in\ p\text{-}p}$.

(c) Square Wave Generator or an Improved Clipper. A popular circuit and the improved version is shown in Figure 24(a) and (b) respectively. Note the vast improvement in the output waveform as illustrated. The output frequency is naturally the same as the input frequency. The peak value of the output waveform is:

$$V_{O\ pk} \approx (0.7 + V_Z) \text{ volts} \tag{15}$$

where V_Z is the zener voltage in volts.

The improved circuit also has the additional advantages of increased efficiency and reduced power dissipation in the zener.

(d) Stairstep Generator. This circuit operates on the same principle as the triangular wave generator. A single CRD is used and the ratio of I to C is much greater. The circuit would appear as shown in Figure 25. The height of each step is defined by Equation 14. The time between steps is governed by the period of the input pulse. In this application $V_{CC} - V_{CE(sat)}$ of Q_1 must exceed the level of the highest step by the least 2 V_L of the CRD.

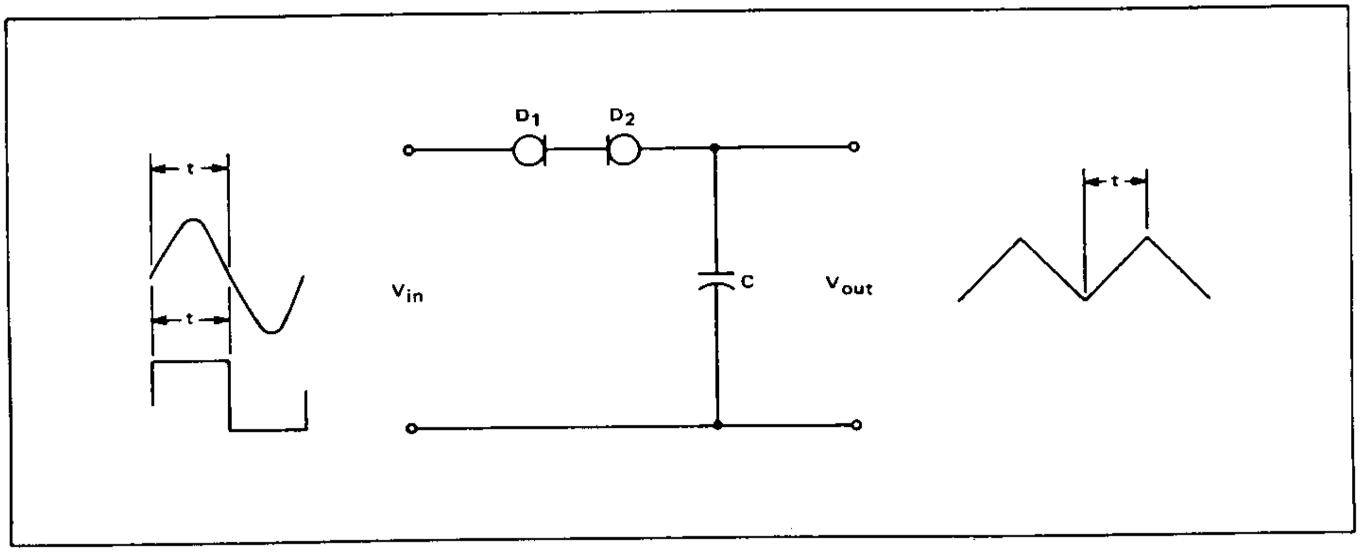


FIGURE 23 – Triangular-Wave Generator or Integrator

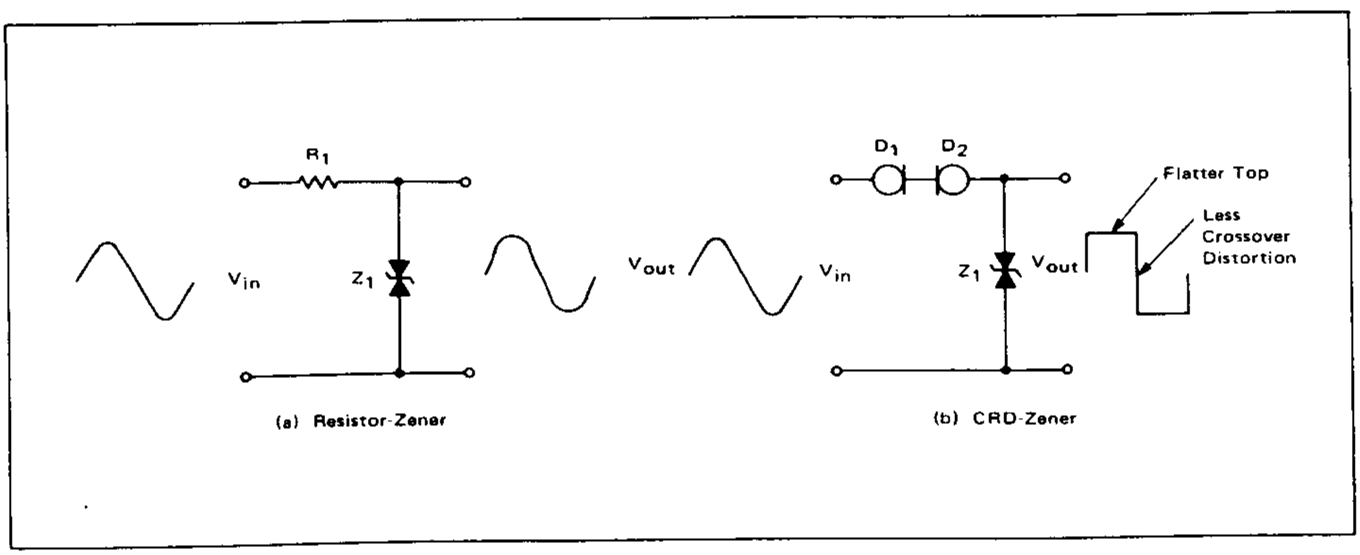


FIGURE 24 – Square-Wave Generator or Clipper

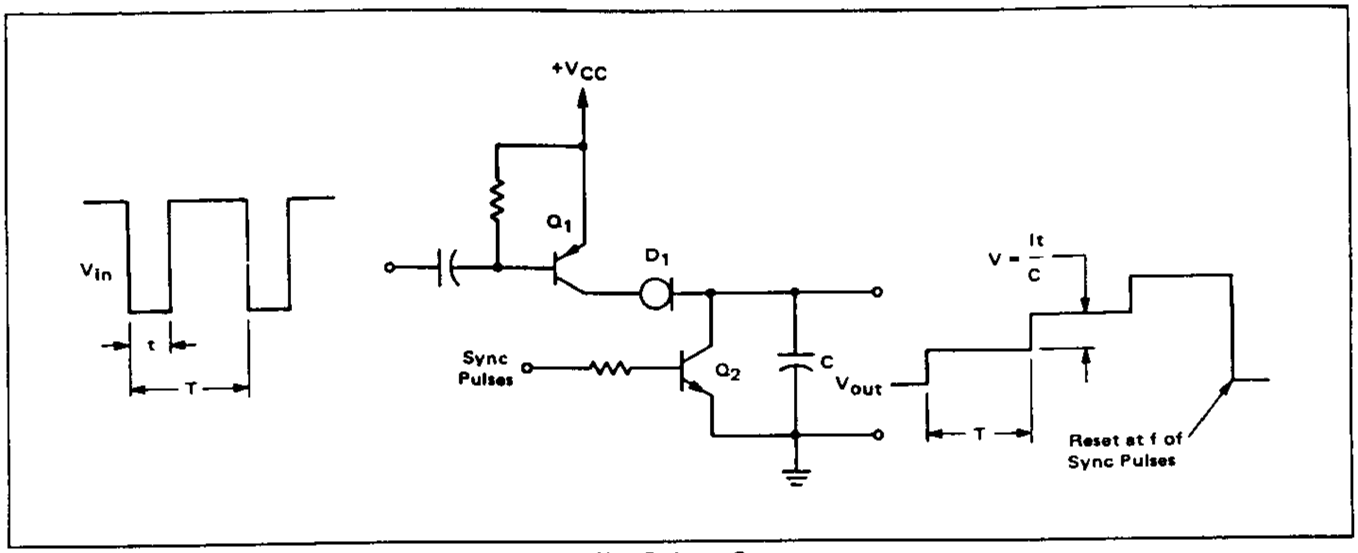


FIGURE 25 – Stairstep Generator

CONCLUSION

In the course of this paper, the current-regulating diode has been shown, by theory and practical application, to be a most versatile component in the design of a multitude of basic circuits. Their application to the general fields of electronics is only limited by the imagination of the user.

If the advantages gained by using the CRD outweigh the disadvantage of added cost, the CRD is highly recommended. An inexpensive plastic FET can also serve the purpose, but with poorer overall characteristics.

REFERENCES

1. "Applications of the Current-Regulating Diode", Hemingway "Electronics" October 20, 1961.
2. "Power Supply Handbook", Paul Berman Kepco Inc., Flushing, N. Y.
3. "FET Applications Handbook", Eimbinder TAB Books.
4. "Field Effect Transistor Applications", William Gosling, Wiley & Sons.
5. "Differential Amplifiers", Middlebrook, J. Wiley & Sons.
6. Hemingway, Op. Cit.

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