Simple wideband linear voltage-to-frequency converter

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A pulse generator is described which has a voltage-controlled frequency continuously variable over more than three decades; the control function is linear and independent of supply-voltage changes, and the control sensitivity is stable, independent of transistor characteristics and easily adjustable.

Es wird ein spannungsgeregelter Impulsgenerator beschrieben, dessen Frequenz über mehr als drei Dekaden stetig einstellbar ist. Die Einstellung ist linear und unabhängig von Schwankungen der Versorgungsspannung. Die Einstellempfindlichkeit ist konstant, unabhängig von den Transistorkennlinien und leicht abzugleichen.

Le générateur d'impulsions décrit dans cet article a une fréquence à tension contrôlée à variation continue sur plus de trois décades; la fonction de contrôle est linéaire et indépendante des changements de tension d'alimentation. La sensibilité de contrôle est stable, indépendante des caractéristiques de transistors et d'un réglage aisé.

THE CIRCUIT DIAGRAM of the converter is shown in Fig. 1. The circuit is essentially a version of the conventional astable multivibrator, in which VT_1 and VT_2 are the main oscillating transistors. To ensure linear charging of the timing capacitors, the two charging resistors are replaced by two current sources utilizing transistors VT_3 and VT_4.[2] Transistors VT_5 and VT_6 are emitter followers for rapidly discharging the timing capacitors, hence extending the upper frequency limit. To render the frequency independent of the supply voltage and transistor-parameter variation, the voltage swing on the timing capacitors must be held constant. This is achieved by the use of MRZ_1 in connection with the fast silicon diodes $MR_{2,3,4,5}$; this arrangement also renders the voltage swing independent of temperature, provided that the Zener diode is temperature-compensated.

Circuit function

Let VT_1 be off and VT_2 on; thus VT_1 collector is at a high potential, MR_2 is cut off, and MR_3 and VT_5 are conducting. VT_2 being saturated, its collector is held at a low potential; thus MR_4 is conducting and MR_5 and VT_6 are off. The current source VT_3 provides the base current for VT_2, while the other current source VT_4 charges C_2; therefore the base potential of VT_1 rises linearly. When it reaches the cut-in voltage of VT_1, this conducts, its collector voltage drops and the conventional regenerative action of the astable multivibrator takes place, turning VT_2 off. MR_5 and VT_6 conduct and take up the role of MR_3 and VT_5. C_2 is discharged through the small output resistance of VT_6 and the input resistance of VT_1. Now the current source VT_4 provides base current for VT_1 while VT_3 charges C_1, and so on.

Control function (or frequency sensitivity)

Neglecting the saturation voltage of either VT_1 or VT_2 in comparison with the Zener voltage, it can be shown from the preceding description that the voltage swing at either base of the oscillating transistors is equal to the Zener voltage V_Z. Hence, for a symmetrical circuit,

$$IT/2 = CV_Z \quad \ldots\ldots\ldots\ldots\ldots (1)$$

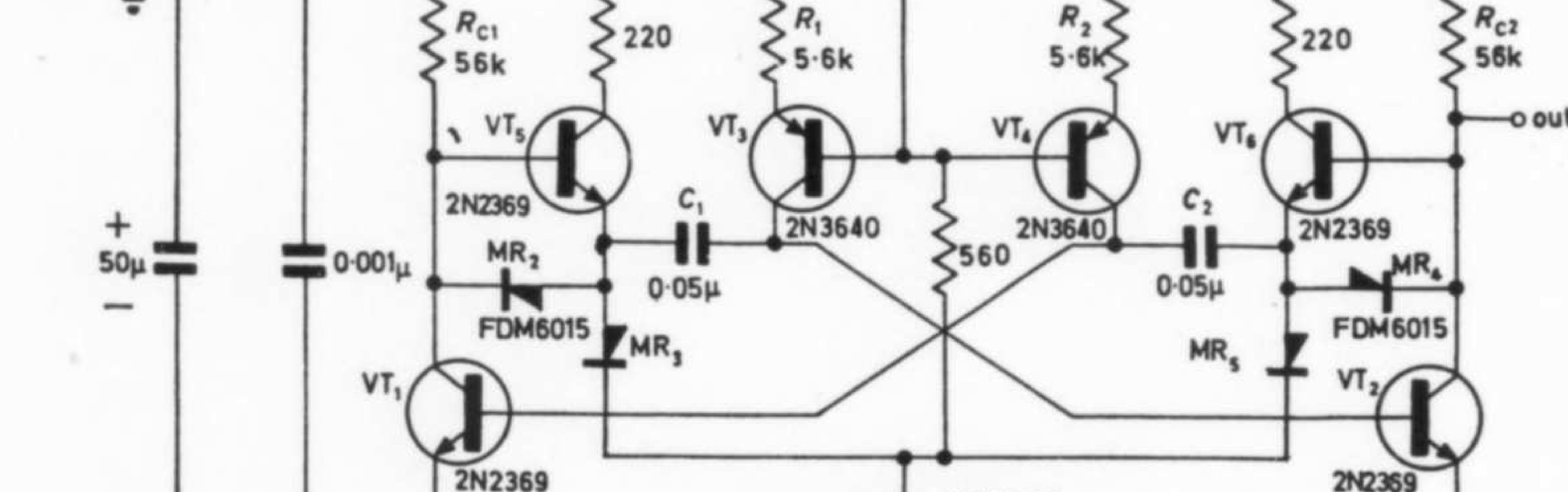

Fig. 1 Circuit diagram of the converter for operation in band 1 (10Hz-10kHz). For band 2 (7kHz-3MHz): $R_1 = R_2 = 1{\cdot}8k\Omega$, $R_{c1} = R_{c2} = 3k\Omega$, $C_1 = C_2 = 470pF$

where I = current provided by each of the two current sources

T = period of oscillation

C = timing capacitance ($C_1 = C_2 = C$)

V_Z = Zener voltage

The current I is given by

$$I = (V_x - V_B)/R \quad \ldots\ldots\ldots\ldots\ldots\ldots \quad (2)$$

where V_x = control voltage

R = sensitivity resistor ($R_1 = R_2 = R$)

V_B = $e-b$ voltage of VT_3 and VT_4*

Combining equations (1) and (2), the frequency of oscillation is given by

$$f = \frac{V_x - V_B}{2CRV_Z} \quad \ldots\ldots\ldots\ldots \quad (3)$$

From which, the frequency sensitivity S (in Hz/V) is given as

$$S = 1/2CRV_Z \quad \ldots\ldots\ldots\ldots\ldots\ldots \quad (4)$$

Therefore S is constant, independent of supply-voltage and transistor-parameter variations, and can be adjusted to the required value by adjusting R (here called the sensitivity resistor).

Design considerations and experimental results

It is appropriate at this point to discuss briefly the limitations imposed on the frequency range, together with the effect of changing various circuit parameters in an attempt to arrive at the optimum design. Experimental results are presented together with suggestions for waveform improvement and temperature stability. A method of direct digital control of frequency is also indicated.

(a) *Frequency range*

The minimum oscillating frequency is determined by the minimum base current required to drive VT_1 or VT_2 into saturation. Provided that the combined switching time of the transistors is small compared with the period of oscillation, the upper limit of the linear control function is imposed by the discharging time of the timing capacitors. If this exceeds the time of half a period, the timing capacitor will not be able to discharge completely, resulting in reduced voltage swing and deviation from linear control. It can be shown (see Appendix) that the frequency range (f.r.) is given by

$$\text{f.r.} = \frac{h_{FE}^2}{x \ln [x/(x-1)]} \quad \ldots\ldots\ldots\ldots \quad (5)$$

where h_{FE} is the transistor d.c. current gain (assumed equal for all transistors) and $X = V_{CC}/V_Z$. Thus choosing transistors of high current gain widens the control range. There is also the possibility of using a value of x which gives an optimum frequency range; however, in choosing V_{CC}, one must consider other limitations, such as the breakdown of VT_1 or VT_2, power consumption, proper operation of the Zener diode etc. It can also be shown that the absolute value of the frequency depends only on the time constant CR_C. Thus it may be seen that, for two reasons, a good choice is a combination of C rather large and R_C rather small. This derives from the fact that, if C is much larger than the storage-capacitive effects of the saturated transistor, the transistor does not impose any limit on the upper frequency range. A smaller value of R_C ensures working less deeply in saturation, thus allowing smaller storage times.

* To simplify matters, assume that the voltage drop on all conducting silicon junctions is equal to V_B (which is approximately 0·7V).

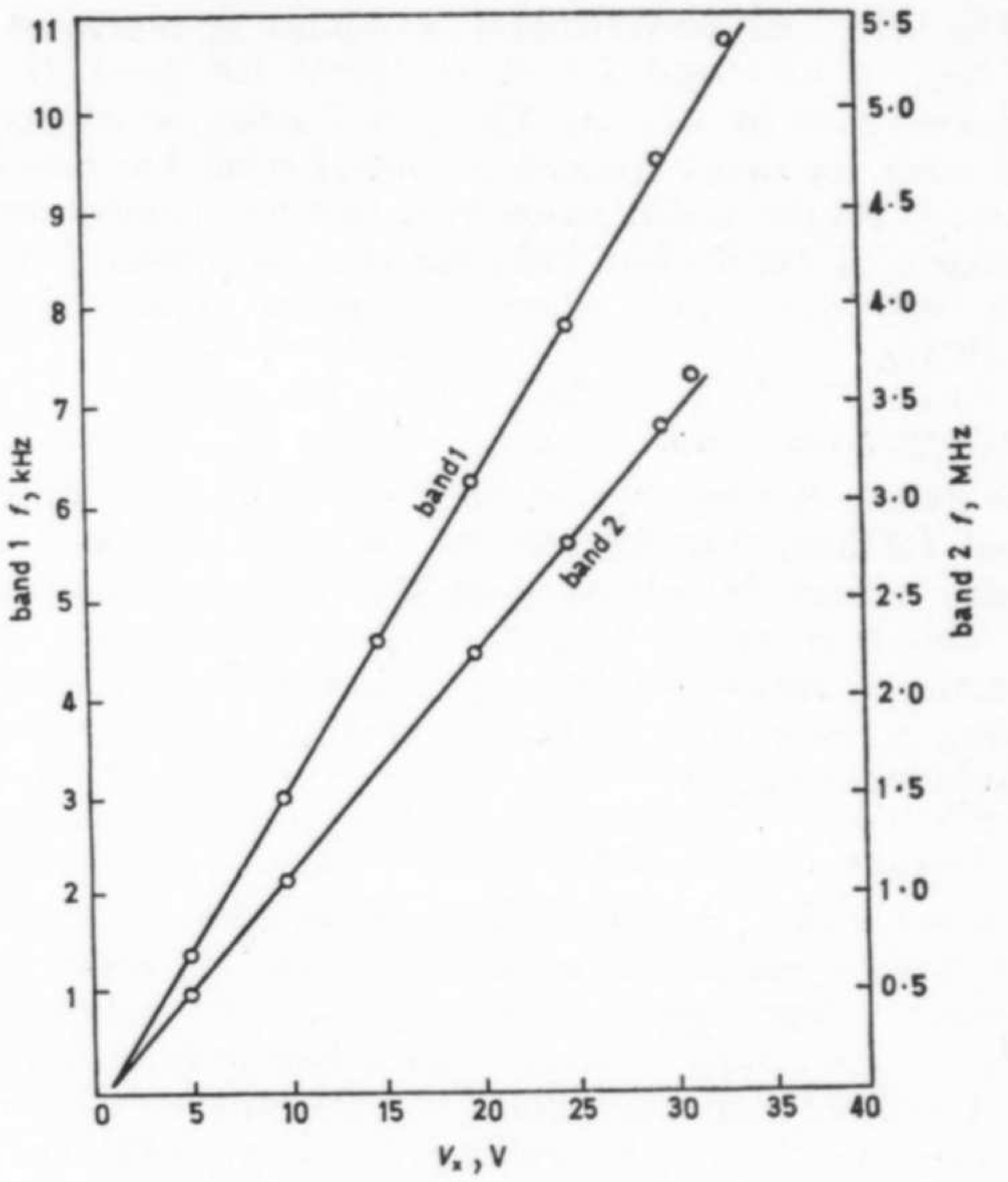

Fig. 2 Measured control characteristics for
(a) band 1 (10Hz-10kHz)
(b) band 2 (7kHz-3·4MHz)

As both the upper and lower limits of the frequency range are set by the limited h_{FE} of both VT_1 and VT_2 and both VT_5 and VT_6 respectively, the frequency range can be extended by replacing any of these individual transistors by a Darlington-connected pair.

(b) *Experimental results*

Fig. 1 is a converter designed to operate in the band 10Hz → 10kHz (1000 : 1). The measured control characteristic in Fig. 2 shows a frequency sensitivity of about 312Hz/V. Calculation from equation (4) would indicate a value of 318Hz/V. The characteristic has an offset of about 0·6V, which is the junction voltage of each of the current sources. This may be minimized by adding a diode, as shown in Fig. 3. As the offset varies with the current I, deviations from linearity result at low input voltages (Fig. 4). By including the circuit shown in Fig. 6, see (d), these deviations are completely eliminated.

The same circuit with similar transistors is transformed to another frequencyband by changing C and R_C and adjusting R. Using the additional values given in Fig. 1, the oscillator covers the band 7kHz-3MHz, with measured characteristics as shown in Fig. 2. Here the upper frequency limit is set by the switching speed of the transistors used.

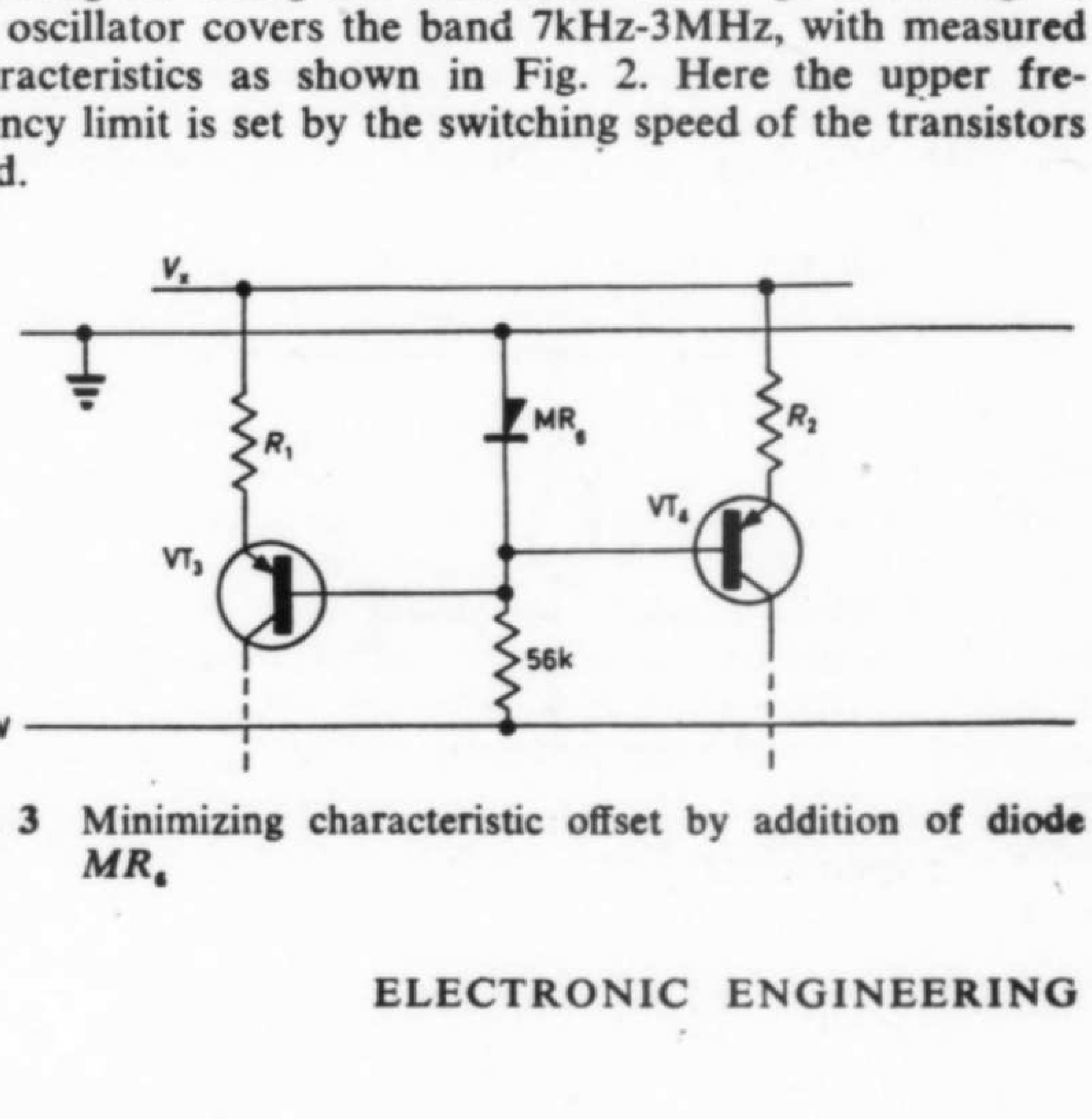

Fig. 3 Minimizing characteristic offset by addition of diode MR_6

The effect of power-supply variation is measured to be 5Hz/V for band I and 100Hz/V for band II. By replacing each of VT_5 and VT_6 by a Darlington-connected pair each, the upper frequency limit of band I is extended to about 50kHz, which results in a 50 000 : 1 control range. Similarly, if Darlington pairs are used to replace VT_1 or VT_2, the lower limit extends to about 0·5Hz (instead of 10Hz).

(c) *Output waveform*

The pulses obtained are of 50% duty cycle and of very small falltime, but the risetime is large. Although one is not necessarily interested in the output waveform of the converter (as it may be a preliminary stage in a system), a reasonable waveform could be obtained by adding a transistor, as shown in Fig. 5, or by more-conventional means.

(d) *Temperature stability*

As noted earlier, the four silicon diodes $MR_{2,3,4,5}$, together with MRZ_1, connected as shown in Fig. 1, result in a constant voltage swing (equal to the Zener voltage) on either of the timing capacitors. This derives from the fact that the Zener maintains point P at a potential $(-V_{CC}+V_Z)$. In one state, VT_1 is conducting and diode MR_1 clamps

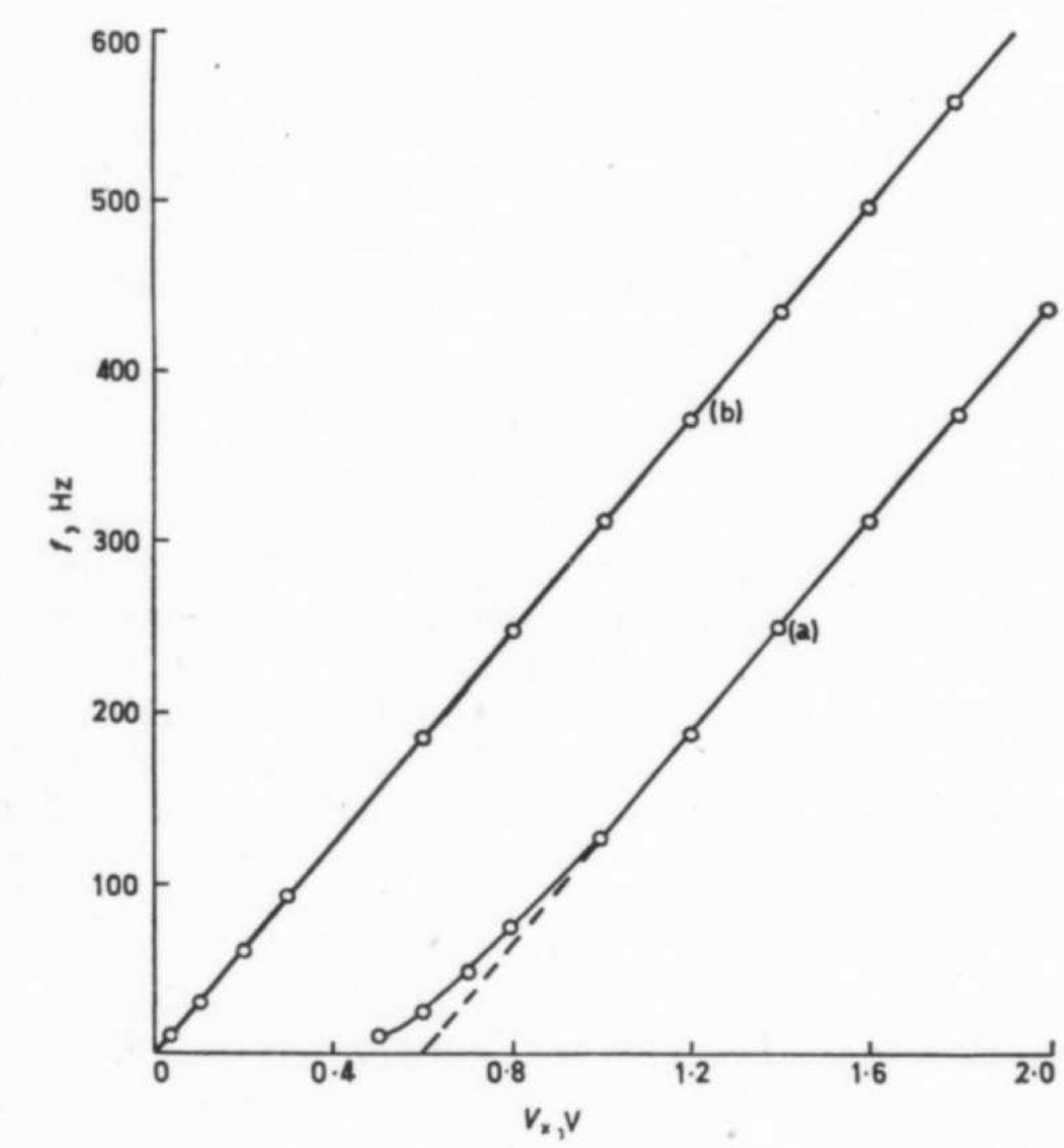

Fig. 4 (a) Control characteristics for band 1 expanded to show deviations from linearity
(b) Including circuit in Fig. 6 eliminates deviations

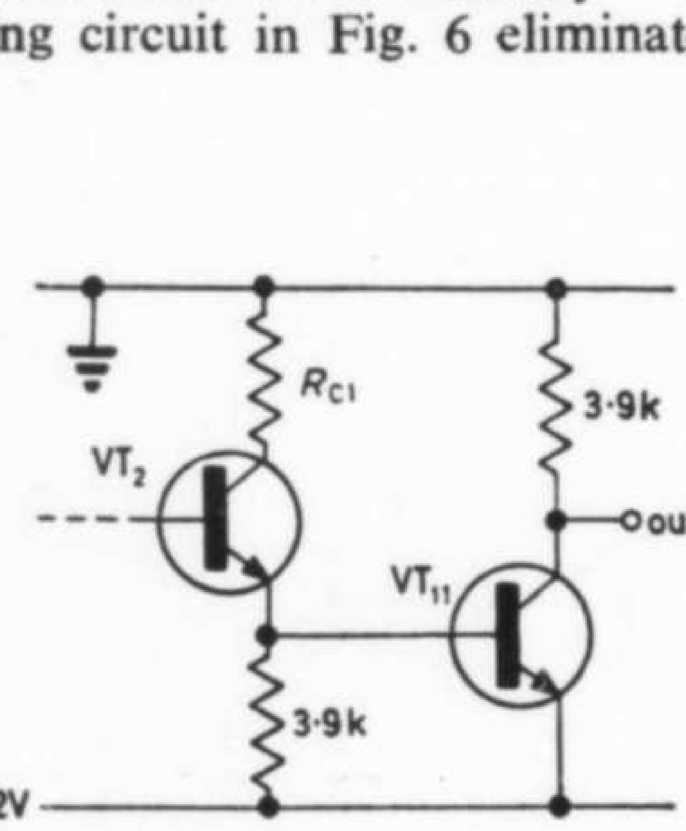

Fig. 5 Improving output waveform by including additional transistor VT_{11}

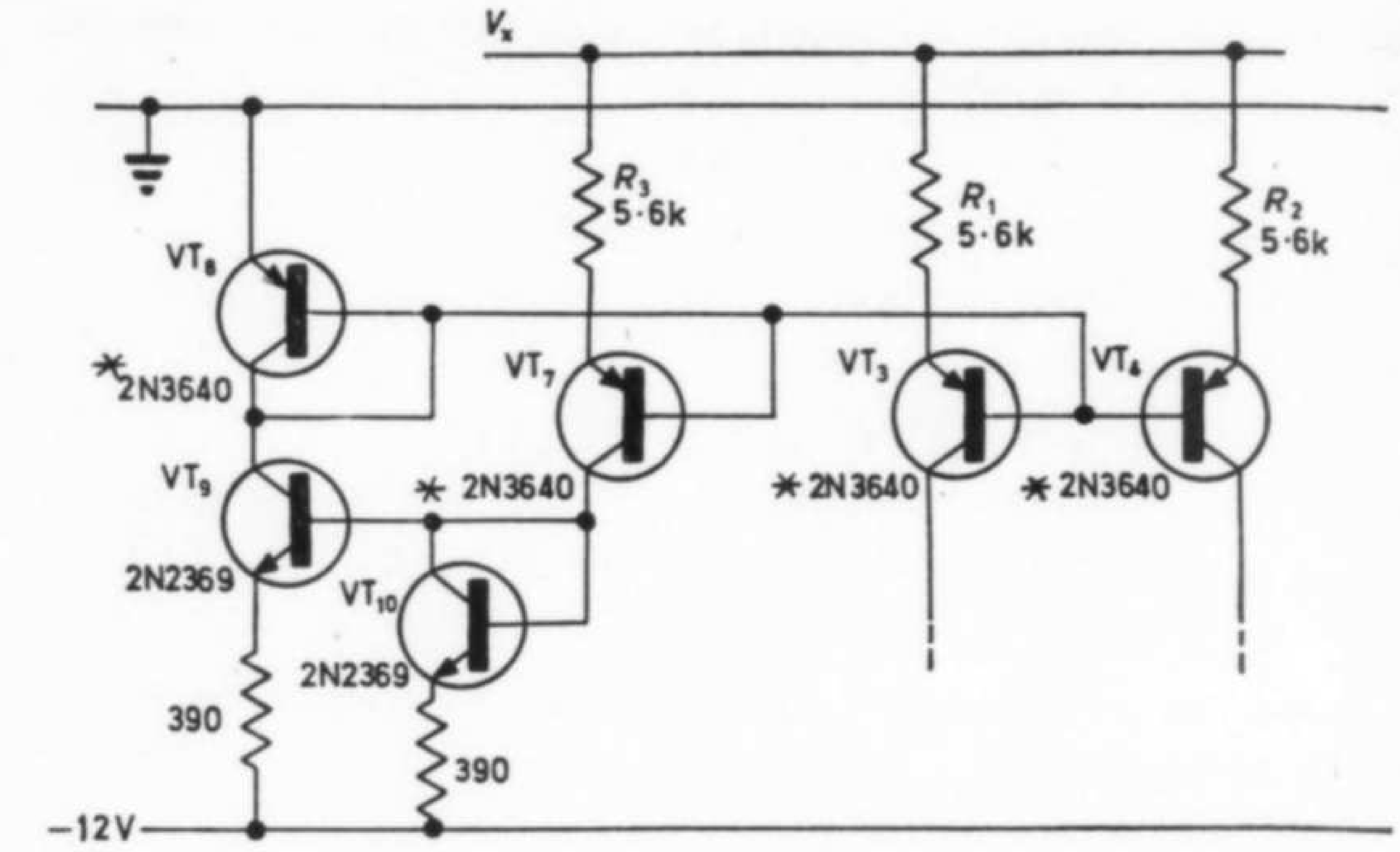

Fig. 6 Addition of four transistors provides complete temperature-independence and eliminates characteristic offset.

* Should have similar V_b and high h_{FE}

the capacitor terminal (marked X) at an approximate potential of $(-V_{CC}+V_B)$ (neglecting $V_{CE(Sat)}$); in the other state, VT_1 and MR_2 are cut off, and point X is now clamped by MR_3 to a potential $(-V_{CC}+V_Z+V_B)$. Thus the voltage swing on the timing capacitor is V_Z. Thus frequency is made independent of supply voltage variations and, provided that the Zener is temperature-compensated, the frequency is also made roughly independent of the temperature variation of either VT_1 or VT_2 or diode parameters.

The only remaining source of temperature instability is the variation in the $e-b$ junction voltage of the current sources VT_3 and VT_4 with temperature. To compensate for this, the arrangement shown in Fig. 6 is suggested. The method consists essentially of holding the emitters of the two current sources almost at ground potential by lowering their bases with a junction drop. For best compensation, this junction must essentially carry an amount of current equal to that of the current sources. Transistor VT_7 is a current source to simulate the original current sources VT_3 and VT_4. By having $R_3 = R_2 = R_1$, VT_7 conducts the same amount of current as either VT_3 or VT_4. Assuming similar transistors and neglecting base currents, it can be shown that the diode-connected transistor VT_8 conducts the same current as that of VT_3 and VT_4. Thus they will all have equal junction drops. In this way, the emitters of VT_3 and VT_4 are held at an almost zero potential. To ensure this equality of junction drops at all environmental temperatures (as it is now guaranteed at all currents), the four transistors $VT_{3,4,7,8}$ must be included on the same heat sink. If the circuit were integrated, this problem would be solved directly. By virtue of this additional circuit, almost complete temperature stability of the pulse generator can be achieved. It must be noted that this arrangement eliminates the offset of the control characteristic and ensures accurate frequency at the lowest control voltages (Fig. 4).

(e) *Fine tuning and digital control*

If fine frequency adjustment is to be provided, another two sensitivity resistors of higher values (for example, 10 times the values of R_1 and R_2) are joined to the emitters of VT_3 and VT_4 with their other free terminals connected to another control-voltage source. This indicates the possibility of direct digital control of the frequency by supplying several pairs of binary weighted sensitivity resistors with each pair connected to the appropriate bit of the digital control signal.

Conclusions

In comparison with previous designs[1], the converter described is simple and economic. A brief discussion of the various design parameters has been given to help in arriving at an optimum design for any special requirements. Its applicability to direct digital control has also been indicated.

References

1 JONES, R. J. P.: A wide-frequency-range voltage-controlled multivibrator, *Electronic Engng.*, 39, 14 (1967)
2 BIDDLECOMB, R. W.: Latest multivibrator improvement, *Electronics*, 65 (April 26th, 1963)

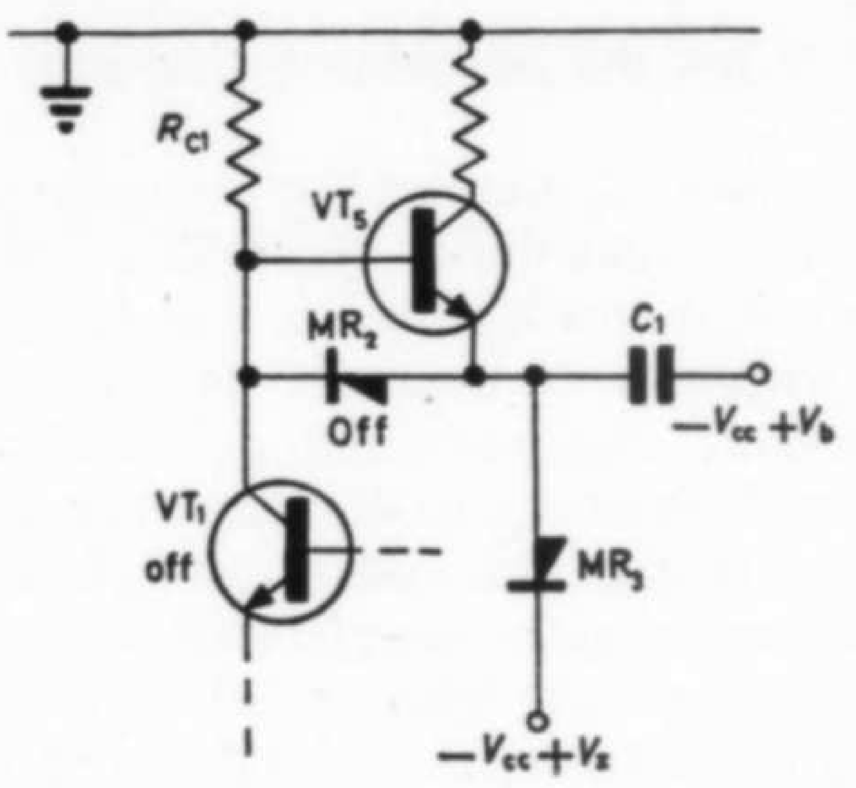

Fig. 7 Calculation of recharging time of C_1 through output resistance of VT_5

Appendix

Dependence of the linear-controlled frequency range on supply voltage

From equation (1),

$$f = \frac{I}{2CV_Z} \qquad \text{(1a)}$$

The minimum frequency is determined by the minimum current I required to drive both VT_1 and VT_2 into saturation; hence,

$$f_{min} = \frac{I_{C(sat)}}{2Ch_{FE}V_Z}$$

and neglecting $V_{CE(sat)}$

$$f_{min} = \frac{V_{CC}}{2Ch_{FE}R_CV_Z} \qquad \text{(6)}$$

The maximum frequency is determined by the time required to recharge the timing capacitors (recovery time). This is calculated by reference to Fig. 7, which shows the emitter follower VT_5 used to recharge C_1. One terminal of C_1 is clamped at $(-V_{CC}+V_B)$ while the other terminal starts from $(-V_{CC}+V_b)$ and rises until it is clamped at $(-V_{CC}+V_Z)$. This time T_r is given by

$$T_r = \frac{CR_C}{h_{FE}} \Big/ \ln\frac{V_{CC}-2V_B}{V_{CC}-V_Z-V_B}$$

Equating T_r to $T_{min}/2$, the maximum frequency is given by

$$f_{max} = \frac{h_{FE}}{2CR_C} \Big/ \ln\frac{V_{CC}-2V_B}{V_{CC}-V_Z-V_B} \qquad \text{(7)}$$

Combining equations (6) and (7), assuming $V_B \ll V_{CC}$ and denoting V_{CC}/V_Z by x, the frequency range is given by

$$\text{f.r.} = \frac{f_{max}}{f_{min}}$$

Therefore

$$\text{f.r.} = \frac{h_{FE}^2}{x \ln[x/(x-1)]} \qquad \text{(5)}$$

A value for x could be found to give the optimum frequency range by differentiation of f.r. with respect to x.

F.E.T. isolation unit for a nerve stimulator

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A pair of f.e.t.s followed by two silicon transistors form an effective, inexpensive and easily constructed stimulus-isolation unit. The unit is suitable for use with a mains-driven stimulator for class studies of nerve *in vitro*.

Ein Paar Feldeffekttransistoren mit zwei nachgeschalteten Silizium-transistoren bilden eine wirksame, leicht aufzubauende Trennschaltung für Erregerimpulse. Die Schaltung ist in Verbindung mit einem netzgespeisten Erregergerät geeignet für Klassenuntersuchungen von Nerven *in vitro*.

Une paire de transistors à effet de champ, suivie de deux transistors au silicium, constitue un ensemble d'isolement d'excitation efficace et de construction aisée. Cet ensemble peut être utilisé en liaison avec un simulateur entraîné par un courant secteur pour l'étude des nerfs sous verre.

IN EXPERIMENTAL WORK, nerves are usually stimulated electrically using a squarewave pulse. Such stimulation not only initiates an action potential in the nerve fibres but transmits the stimulating potential along the bundle of nerve fibres to the recording electrodes and indicating oscilloscope. As the potential of the squarewave is usually 500mV or more, compared with some 5mV or less of action potential detected by recording electrodes outside the nerve, the stimulation artifact can be sufficient to unbalance the amplifier of the oscilloscope so that the action potential immediately following is masked. Most of this artifact can be eliminated when working with nerve *in vitro*, by placing an earthing electrode on the nerve between the stimulating and the recording electrodes, as is well known[1,2] but when a mains-driven stimulator is used sufficient artifact remains after the squarewave to distort the action potential as recorded on the oscilloscope, Fig. 1. This persistent artifact is due to resistive and capa-