ELECTRONIC BUILDING BLOCKS

PART SEVEN by R. A. DARLEY

FOLLOWING on last month's article on impedance matching devices, we will take a closer look at the Darlington pair and "bootstrap" high impedance amplifier, with special reference to their practical uses. Then we will discuss the theory of some well-known oscillator circuits.

APPLICATIONS OF THE DARLINGTON PAIR

As well as being used in a.f. amplifier circuits, the Darlington pair has a number of other applications. Two such uses are illustrated in Figs. 7.1a and 7.1b.

As was discussed in part 2 of this series (see May issue), when a resistor and capacitor are wired in series and connected across a voltage supply, the potential at the junction of the two components varies in an exponential manner; a time constant is introduced, in which $T = CR$, where T = time in seconds, C = capacitance in farads, and R = resistance in ohms.

By connecting this exponential voltage to the base of a transistor and connecting a relay in series with the emitter or collector, this exponential variation may be used to impart a time delay to the operation of the relay.

If a conventional transistor circuit is used in this application, it is found that the comparatively low input impedance of the transistor stage shunts one or other of the time-constant components and thus reduces the effective value of the component concerned.

By using a Darlington pair circuit as the transistor stage with a high input impedance, this difficulty is largely overcome.

In the circuit shown in Fig. 7.1a, at the moment that S1 is closed and the supply is connected to the circuit, only a very low emitter current is flowing and the relay is not operated; as time passes, the emitter current rises and, after a fixed time delay, the relay operates.

In the circuit of Fig. 7.1b the reverse action takes place; at the moment of switch on, maximum emitter current flows and the relay is operated. As time passes, the emitter current falls in value and, eventually, the relay drops out.

It is important to note that, in both of these circuits, the time delay imparted to the relay depends not only on the values of the time constant components, but also on the characteristics of the relay, the transistor, and the supply potential.

In Fig. 7.1b a switch (S1b) is shown connected in parallel with C and ganged to S1a. Once the circuit has been operated, a voltage is built up across this capacitor. If S1b is not in circuit when the supply voltage is cut off there is no effective discharge path for the capacitor. If the supply voltage is again reconnected it is found that, because the capacitor has retained its charge, the time constant effect is destroyed. The fitting of S1b overcomes this trouble by discharging the capacitor when the supply voltage is cut off.

In the case of the circuit of Fig. 7.1a, the capacitor discharges automatically through the low impedance forward biased emitter-base junctions of the transistors and no additional components are required.

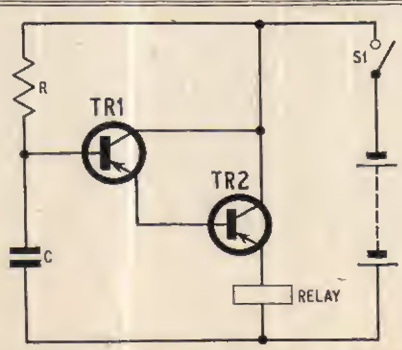

Fig. 7.1a. Darlington pair used to operate a relay

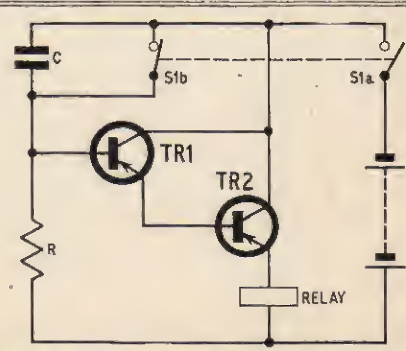

Fig. 7.1b. Darlington pair used to operate a relay, then release it after a time delay

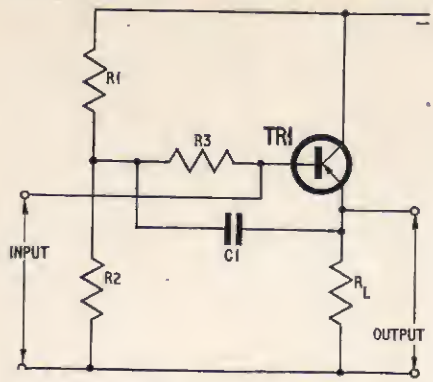

Fig. 7.2a. Basic circuit of a "bootstrap" high impedance amplifier using an emitter follower

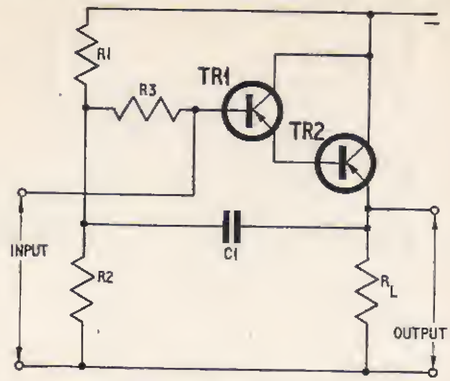

Fig. 7.2b. "Bootstrap" circuit using a Darlington pair

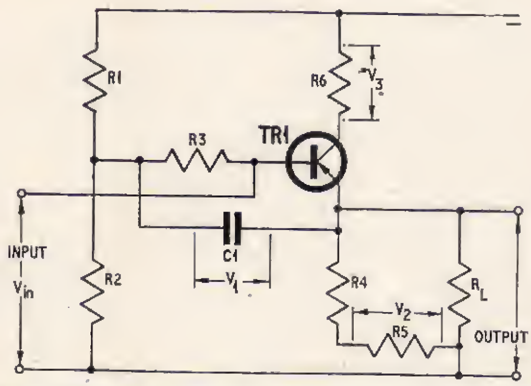

Fig. 7.3 (right). Basic "bootstrap" circuit

Input impedance $Z_{in} > 1{,}000\ M\Omega$.
$V_{in} = V_1 = V_2 = V_3$

THE "BOOTSTRAP" HIGH IMPEDANCE AMPLIFIER

One of the snags with the Darlington pair emitter follower circuits is that the base bias resistors are effectively in parallel with the input to the transistor circuits.

It was also pointed out earlier in this series that, for good stability, the base bias should be provided via a voltage divider network, in which the current flowing through the lower resistor should be at least 10 times greater than that flowing in the transistor base. Quite clearly, the effect of such a base bias network on an emitter follower or Darlington pair circuit is to reduce the circuit's input impedance by a factor of at least 10.

As an alternative to the divider chain method of obtaining base bias, a high value resistor, of such a value as to limit the base current to the required value, can be connected between the base and negative supply line, as shown in Fig. 6.7 last month. This method has the advantage of considerably reducing the shunting effect on the input circuit, but has the disadvantage of giving poor circuit stability.

The circuit known as the "bootstrap" high impedance amplifier overcomes these disadvantages. Fig. 7.2a illustrates the circuit.

TR1 is connected as an emitter follower, with an emitter load R_L. Resistors R1 and R2 form a conventional voltage divider base bias network, but an additional resistor (R3) is connected between the junction of R1–R2 and the transistor base; this is an a.c. isolating resistor. For d.c. purposes, R3 has little effect on the biasing arrangements. A feedback capacitor (C1) is connected between the junctions of R1–R2–R3 and the emitter.

When an input signal is fed to the base of the transistor, an output signal will appear at the emitter, in phase with the input and of almost the same amplitude. If the impedance of C1 is low at the frequency of operation, the signal from the emitter will be fed back to the junction of the voltage divider chain. The a.c. voltages at both ends of R3 are therefore almost equal. If the voltage gain of the emitter follower circuit is exactly 1, the voltages (a.c.) at each end of R3 will be exactly equal. Therefore, no a.c. current is flowing through this resistor; the resistor presents an infinitely high impedance to a.c. It also follows that R1 and R2 can thus have no shunting effect on the input impedance of the circuit.

In practice, of course, the voltage gain is not quite equal to 1, so complete isolation of the base bias components from the input impedance does not take place; sufficient isolation is provided, however, for their effect to be almost ignored. With this circuit, the input impedance is the same as in the emitter follower, ($Z_{in} \simeq \beta\ R_L$) and is in parallel with the leakage resistance.

Using this circuit, quite high input impedances can be obtained with good stability. If a Darlington pair circuit is substituted for the emitter follower, as shown in Fig. 7.2b, impedances in the order of 4 megohms can be obtained using germanium transistors, and as high as 20 megohms using silicon transistors.

ULTRA-HIGH IMPEDANCE AMPLIFIERS

In the circuit just described, the "bootstrap" principal was applied over one resistor only. Circuits are available in which the principal is employed over more than one part of a circuit, however, and as a matter of interest one such circuit will be roughly described.

When discussing the emitter follower circuit, it was mentioned that the input impedance of the circuit could be raised either by increasing the value of R_L or the effective value of β, but if R_L were raised to too high a value difficulties were encountered in meeting the mean emitter current requirements. By using the "bootstrap" principal these difficulties can be overcome.

Fig. 7.3 shows a simplified diagram of a circuit which meets this problem. TR1 is connected as an emitter follower, with emitter load R_L. Base bias is provided by the voltage divider network R1–R2 and isolating resistor R3. The input signal is applied to the base of TR1, and the output of the emitter is fed back to the junction of the potential divider via capacitor C1; resistor R3 is thus "bootstrapped" and, if the voltage (V_1) fed back is of the same phase and amplitude as V_{in}, it presents an infinite impedance to the input signal.

The d.c. emitter current of the transistor flows through resistors R4 and R5 in series; if a signal of the same phase and amplitude (V_2) is applied to the junction of R4–R5, no a.c. current will flow through R4. R4 is thus "bootstrapped" and presents an infinite impedance to a.c.; to d.c., R4 is the nominal d.c. value and no problems are presented in obtaining d.c. values of emitter current. The only path available for a.c. emitter current is through R_L, which is thus the effective emitter load, and may be any value that is required.

By connecting resistor R6 in series with the collector of the transistor and applying a suitable "bootstrap" voltage (V_3) across it, no a.c. voltage drop occurs between the emitter and collector of the transistor. It follows that the transistor leakage resistance is also effectively "bootstrapped" and presents an infinite impedance to a.c.

The input impedance of such a circuit is given as $Z_{in} = \beta \times R_L$. Load resistor R_L can be made any value that is required without imposing any limits on the mean emitter current and causing distortion of the output.

The output impedance of the circuit is given as

$$Z_{out} = r_s/\beta$$

As this circuit is intended for use with inputs having very high values of source resistance the output impedance (Z_{out}) of the circuit will also be very high. It is necessary, if the circuit is eventually to feed a common emitter type of amplifier, to feed it at an impedance Z_{out} into another "bootstrap" amplifier, and thence reduce the output impedance to the required value in a number of successive stages.

With such an arrangement it is possible, using silicon transistors, to obtain input impedances in excess of 1,000 megohms and output impedances of less than 100 ohms.

OSCILLATOR CIRCUITS

Most oscillator circuits can be fitted into one or other of three categories, depending on the type of waveform developed. These categories are (a) sine wave generators, (b) pulse generators, and (c) saw-tooth generators. Each of these three categories can be broken down into a number of sub-divisions, depending on the mode and frequency of operation, and each of these sub-divisions can be further broken down into particular circuits. In the limited space that is available for this series it is not possible to describe all of the circuits.

GENERAL PRINCIPLES OF OSCILLATOR OPERATION

Referring to Fig. 7.4, it can be seen that transistor TR1 is connected as a common emitter amplifier with the primary winding of transformer T1 connected as its collector load. The secondary of T1 is connected in series with the transistor base, but in such a way that any signal fed to the base via the transformer is 180 degrees out of phase with the signal at the collector.

When a transistor is connected in the common emitter mode, the signal appearing at the collector is 180 degrees out of phase with that at the base. In the circuit of Fig. 7.4, however, an additional 180 degree phase shift has been imparted to the signal by the transformer, so that the signal appearing at the secondary of T1 is in phase with the base.

If it is assumed that the gain of the circuit between the transistor base and the transformer secondary is greater than 1, the action of the circuit will be as follows.

When S1 is closed and power is supplied to the circuit, a base current will flow via the base bias network R1 and R2, which will cause a collector current to flow also. Although no signal is directly connected to the base, a small a.c. signal will appear at the collector, due to random electron movement (noise) in the transistor; this signal will be transmitted back to the base via T1. This base signal is then amplified by the transistor and appears at a much greater amplitude at the collector, where it is again fed back to the base and amplified even more. This action is almost instantaneous and is cumulative, the signal eventually becoming so large that the transistor is driven into saturation and can no longer amplify further.

The transformer is an a.c. device, and will only transmit changes in potential; thus, as the transistor runs into saturation and amplification ceases, signals also cease to appear at the transformer secondary and, therefore, at the base of TR1. With no base signal applied, the transistor drops out of saturation and reverts to its original condition, whence the complete cycle of events again repeats itself.

Such a circuit acts as a self-energising switch, rapidly switching itself on and off, the output waveform being a series of pulses. The frequency of operation and rise and decay times of such a circuit are determined by the characteristics of the components used, i.e. by the transit time and junction capacitance of the transistor, inductance and stray capacitance of the transformer windings, and so on. A simple circuit of this type can, for practical purposes, be considered as "uncontrolled", in that no special circuitry is employed to control the frequency of operation or the waveform.

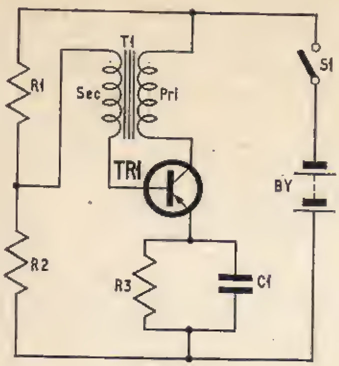

Fig. 7.4. Uncontrolled self-energising pulse oscillator

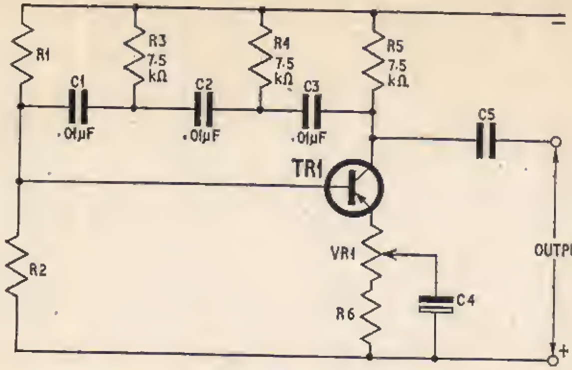

Fig. 7.5. Phase shift oscillator tuned to 1,000c/s

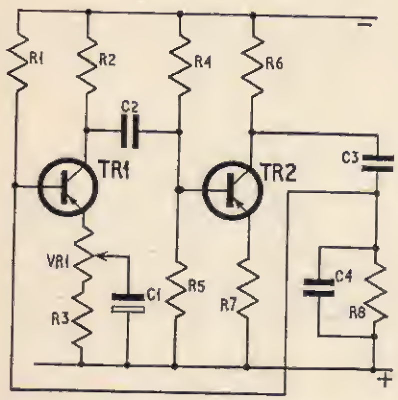

Fig. 7.6a. Simple Wien bridge audio oscillator

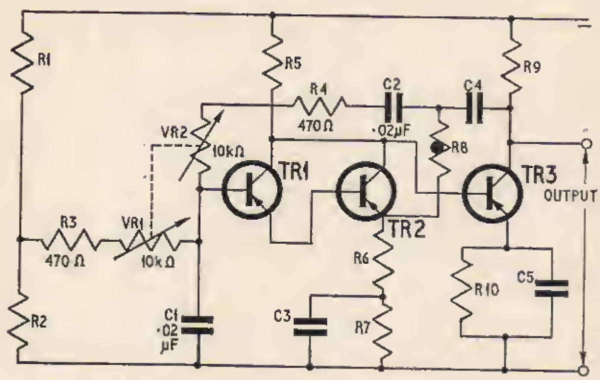

Fig. 7.6b. More advanced type of Wien bridge oscillator; frequency is variable between 1,000 and 10,000c/s

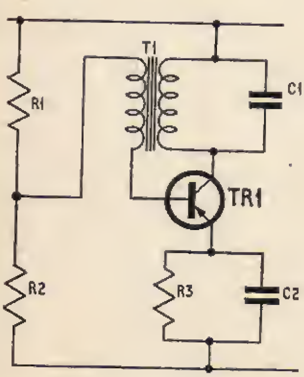

Fig. 7.7. Simple r.f. oscillator with a tuned collector circuit and inductive feedback to the base

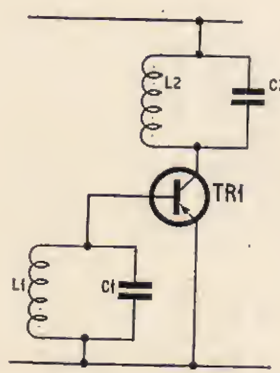

Fig. 7.8a. Basic tuned collector, tuned base oscillator

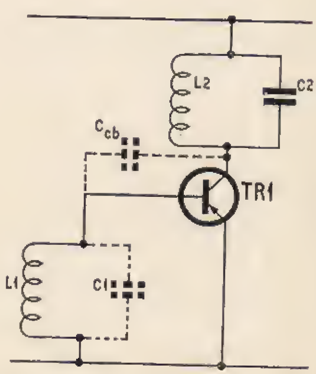

Fig. 7.8b. Extension of the circuit shown in Fig. 7.8a using the Miller effect capacitance for tuning L1

SINE WAVE OSCILLATORS

If a tuned circuit or filter network is incorporated in either the feedback network or the transistor loading circuits, the rise and decay times and the frequency and waveform, will be controlled, a sine waveform being obtained.

Generally speaking, low frequency oscillators can be recognised by the use of RC networks, while r.f. oscillators can be recognised by the preferred use of LC networks. Inductances are avoided in low frequency work because they tend to be rather large. Crystal oscillators are generally operated at high frequencies.

PHASE SHIFT OSCILLATOR

Fig. 7.5 illustrates the circuit known as the "phase shift oscillator". Transistor TR1 is connected in a conventional common emitter mode with resistors R1 and R2 providing the base bias network. R6 and VR1 make up the emitter load with C4 as the decoupling capacitor. VR1 enables the circuit gain to be adjusted by means of negative feedback due to the undecoupled part of the R6–VR1 chain.

The CR networks, C1–R3, C2–R4, and C3–R5, act as phase shifting devices or filters in the feedback loop between the collector and base, a 180 degree phase shift being obtained.

If an a.c. signal is passed through a pure capacitance, a phase shift of 90 degrees is imparted to the signal; if a resistance is connected in series or in parallel with the capacitance, the resulting phase shift will be less than 90 degrees, the precise amount of phase shift depending on the values of C and R and the frequency.

If three such phase shift networks are connected in cascade, a total phase shift of 180 degrees can be obtained if the component values and frequency of operation are suitably chosen. Note that, with any given set of CR values, the overall phase shift will be exactly 180 degrees at only one frequency; thus, the network not only gives the 180 degrees phase shift required for operation, but it also controls the frequency and the waveform.

It is common practice to make all the resistors and all the capacitors of such a network of equal value. In such a case the frequency at which the phase shift is 180 degrees is given by the formula:

$$f = \frac{1}{2\pi CR\sqrt{6}} = \frac{1}{15{\cdot}4 \times CR}$$

The attenuation factor of the circuit at the frequency at which the phase shift is 180 degrees is 29; thus, in an idealised circuit, a transistor gain of only 29 is required and the circuit will oscillate at a frequency given by the formula above. In practice, it is not quite as simple as this and the following points should be taken into account:

(a) The range of values that can be selected for the phase shift resistors is limited; the selected values are a compromise between a value which would appreciably load the transistor input circuit (which has a low impedance) and a value that would be appreciably loaded by the transistor output circuit (high impedance). Values in the order of 2·7 kilohms to 27 kilohms are generally recommended.

(b) The frequency of operation of the circuit is modified from that given in the formula by the input and output impedances of the circuit and by additional phase shifts in the circuit. The frequency of operation is usually higher than that calculated. Remember that to increase the frequency of operation the CR values of the phase shift network are decreased. In the circuit shown, where 7·5 kilohm resistors are used, the calculated frequency of operation is about 850 c/s, but in practice the circuit will oscillate at about 1,000 c/s.

(c) Although the theoretical attenuation of the phase shift network is only 29 at the frequency of operation, it will be found in practice that additional losses occur due to the transistors input and output impedances; generally speaking, the transistor gain is required to be at least 60 in order to overcome these losses. This gain can be obtained either by using a single high gain transistor or by using two low gain transistors connected as a super alpha pair in place of TR1.

Practical circuits of this kind have three main disadvantages: (a) the frequency of operation is subject to considerable change (up to 10 per cent) with variation of the gain control (VR1), which should be set to give minimum distortion of the signal; (b) the circuit is not generally suitable for use as a variable frequency oscillator, as a three-gang tuning control is required; (c) output amplitude and frequency are both subject to change with different loads attached to the output circuit, and a buffer stage is thus generally required between the oscillator and any load.

WIEN BRIDGE OSCILLATOR

A rather different approach to the filter network is employed in the circuit of Fig. 7.6a, which is a basic version of the circuit known as the Wien bridge oscillator. In this case the filter network comprises four components R6 and C3, and R8 and C4. This circuit is very similar to one side of the Wien Bridge described in the July issue.

If a signal is applied across the complete network, it will be found that a definite phase shift occurs across the lower arm at all frequencies except one; at this frequency the phase shift is zero and the actual frequency at which this condition occurs is given by:

$$f = \frac{1}{2\pi\sqrt{(R_6C_3R_8C_4)}}$$

It is common practice to make $R = R_6 = R_8$, and $C = C_3 = C_4$, in which case the above formula simplifies to:

$$f = \frac{1}{2\pi CR}$$

At the frequency of zero phase shift, the attenuation of the network is 3.

Returning to the circuit of Fig. 7.6a, transistors TR1 and TR2 are each connected as common emitter amplifiers. There is a phase shift of 180 degrees between the base and collector of each transistor, giving a total phase shift between TR1 base and TR2 collector of 360 degrees. The output of TR2 collector is effectively applied across the Wien network, and the signal at the junctions of the upper and lower arms is fed back to TR1 base. There is a total phase shift of 360 degrees at one frequency only, and, providing that the gain is correct, the circuit will oscillate at this frequency.

To give a sine wave output which is pure, the transistors are required to provide a gain of 3, to compensate for losses in the Wien network. If the transistor circuits were simply designed to give this degree of gain, they would exercise some degree of control over the frequency (through phase shift effects) and waveform (through non-linearity of frequency response). These snags are overcome by designing each transistor circuit to give a very high gain, which is then reduced to the required low value by the application of heavy negative feedback, i.e. the emitter resistors are left totally or partially undecoupled.

The simple circuit shown in Fig. 7.6a may be used in a number of applications, but for general use it has three drawbacks:

(i) the input impedance of TR1 shunts the lower half of the Wien network, thereby limiting its value;

(ii) resistor R8 also acts as the lower half of the TR1 base bias network, and the circuit is thus not suitable for use as a variable frequency oscillator;

(iii) the gain of the circuit is not fully stabilised against changes of temperature and frequency.

A better circuit is shown in Fig. 7.6b. The single low input impedance transistor is replaced by two transistors, TR1 and TR2, connected as a super alpha pair, and having a high input impedance. A variable resistor VR2 is connected in the upper arm of the Wien network and VR1 in the lower arm as shown. Variations in the setting of VR2 have virtually no effect on the d.c. bias conditions of the circuit, but VR2 does enable the effective values of the Wien arm to be varied. VR1 and VR2 are ganged together. The circuit will oscillate at frequencies between approximately 1,000 and 10,000 cycles per second.

An additional feedback circuit is inserted between the output of TR3 and the emitter of TR2. The thermistor R8 is used as the feedback link; if the output of TR3 increases for any reason, the current taken by R8 will also increase, and in so doing will cause its resistance to fall, which in turn will result in a larger portion of the output of TR3 being applied to the emitter of TR2, thereby reducing the circuit gain. Effective amplitude stabilisation is thus obtained.

The three-section phase shift and the Wien network l.f. oscillators described are some of the most widely used types in general use; many other types or variations do exist. For example, the control network may be a "twin-T" or "parallel-T" type; a phase shift network using 4, 5, or 6 sections in place of the more usual 3 may be used.

R.F. OSCILLATORS

In the case of RC oscillators it is generally essential that the overall circuit gain be 1, any greater value may give distortion. In the case of r.f. oscillators, however, this requirement is not as important, since the LC tuned circuits that are used give a "flywheel" effect to the circuit waveform.

The requirements of these oscillators are simple: the overall gain must be greater than 1; the output must be fed back to the input in phase; and the circuit must be frequency selective.

Figs. 7.7 and 7.8 illustrate two of the many circuits that may be used as r.f. oscillators, very little explanation being required in most cases. Fig. 7.7 shows a circuit with a transformer used between collector and base to give 180 degree phase change, the transformer primary being tuned by C1 to make the circuit frequency selective.

In some cases, there is a tuned circuit in the collector and another tuned circuit connected to the base, but with separate coils used instead of a transformer. In such cases, the circuit relies on the Miller feedback effect to cause oscillation. Between the base and collector junctions of a transistor there is some small capacitance; this capacitance provides a feedback path between collector and base.

Thus, a transistor circuit with a tuned collector and a tuned base, as shown in Fig. 7.8a, can be re-drawn as in Fig. 7.8b, where C_{cb} is the effective Miller capacitance. In this circuit, C1 and L1 do not form a major tuned circuit; hence C1 is shown dotted. It is C_{cb} and L1 which form the tuned circuit, of the series type.

Across a series LC circuit there is a phase shift of 180 degrees in voltage. In this circuit, at the frequency of oscillation, the series circuit is detuned so that the reactance of L1 is smaller than that of C_{cb}, so that the voltage across L1 is smaller than, and in opposite phase to, that across the series combination, i.e. the series circuit acts as an inductive reactance. C1 enables this inductive reactance to be tuned to the required frequency. Due to the resistive losses in the LC circuit, the phase shift is slightly less than 180 degrees; the collector tuned circuit (L2 and C2) is therefore slightly detuned to compensate for this. It is interesting to note that, at the frequency of oscillation, neither the apparent collector tuned circuit, the series tuned circuit, nor the apparent base tuned circuit are actually tuned to the frequency of oscillation.

Next month: Basic circuits for cathode ray oscilloscopes