PART ONE

THE AUTHOR M. L. MICHAELIS M.A. MAKES FINAL ADJUSTMENTS TO HIS

...beam switching unit

THE conventional oscilloscope allows us to examine *one signal waveform at a time.* If we wish to compare various signal waveforms, then we must feed them one at a time successively into the oscilloscope and remember, draw on paper or photograph each one off the screen for subsequent mutual comparisons. For simple radio and audio equipment, this is neither difficult nor time-consuming and numerous simplifications of procedure are possible. For example, if we are interested in observing the degree of distortion in an audio amplifier, we can feed a good sinewave test signal from a signal generator into the amplifier input. We know what an undistorted sinewave looks like, so that any departures therefrom as the oscilloscope is connected to the outputs of the successive stages of the amplifier chain immediately reveal the faulty section of the equipment. It can speed-up work if we had some means of displaying two signals *simultaneously* on the oscilloscope screen, but this facility remains largely a luxury for simple radio and amplifier servicing and design.

TIMING AND PHASING

Matters are quite different when we turn to more general electronic equipment. We are here not only interested in the correct waveform shapes as a whole in such equipment, but also in the precise timing of each part of any waveform, i.e. in their relative phases and time-leads or time-lags with respect to each other. If the various pulse flanks from different stages are used, say, to set-off responses in electronically controlled machinery, it is immediately obvious that the behaviour of that machinery would be quite erratic if the control flanks in the electronic control circuit waveforms are mixed up or otherwise incorrectly phased. Here we see the first *essential* need to have a means for *simultaneously* displaying *two signal waveforms* on an oscilloscope when designing and servicing machinery control electronics and a host of other logical electronics.

It is clearly not possible to gather information regarding the *phasing* of two waveforms by applying them separately and successively to a normal oscilloscope, since the "synchronisation" circuitry of the oscilloscope timebase deflection always forces the horizontal timebase run to commence at the moment of a predominant flank in the waveform. In other words, the synchronisation arrangement, essential to make the repetitive traces coincide and yield a stationary display, cancels all phase-shifts as far as relative positions on the screen are concerned.

We are obviously no better off when using two oscilloscopes to display one waveform each of a pair of waveforms whose phases we wish to compare. The synchronisation circuit will cause each oscilloscope to display its signal as if there were no phase difference! If we "turn off" the synchronisation action, then the trace will drift about arbitrarily and matters are worse still—we can then not even observe the waveform shape any more, since the successive traces no longer coincide.

TWIN-BEAM CATHODE RAY TUBES

An obvious—but very expensive—way round the problem is to use a special type of cathode ray tube in our oscilloscope, which produces two electron beams, either from two separate electron guns or by some electrostatic means of beam splitting. This double beam can be deflected in the horizontal direction by a common timebase which is synchronised from *one* of the two signals we wish to compare.

This signal, which we will call the *leader*, is applied as vertical deflection to only one of the electron beams, usually the one moving in the upper part of the fluorescent screen. The other signal waveform is simultaneously applied alone to the lower beam. Now it is immediately evident that the two waveforms will appear simultaneously on the screen, in the *correct phase/time relationship*. For example, if the leader is derived from the input to a certain stage in the electronic equipment on test, and the second signal (the *dependant*) from the output of that or a later stage, then the time-delay of signal transfer between the stages in question is accurately portrayed by the horizontal displacement between the salient flanks in the two respective waveforms on the double-beam oscilloscope screen.

Genuine double-beam oscilloscopes of the type described above are manufactured commercially and widely used in professional circles. However, they are very expensive compared to normal single-beam oscilloscopes and seldom found among the offers of oscilloscopes for amateur purchase. This is because there is a cheaper and in many respects more elegant method of achieving virtually the same function with any ordinary single-beam oscilloscope, which need fulfil only a bare minimum of essential prerequisites for the purpose.

BEAM SWITCHING

It is the purpose of this article to present a design for a beam switching unit which may be used in conjunction with most ordinary oscilloscopes to give accurate simultaneous two-signal display, conserving full phase information as well as waveform shapes. This unit may be connected between the signal probes taking the signals off from the test points in the equipment under examination, and the vertical deflection amplifier (Y-deflection amplifier) input of the conventional oscilloscope.

A block diagram of the beam switching unit is given in Fig. 1. It will be seen that the unit has two separate inputs, each with its respective signal pick-up probe and separate pre-amplifier and attenuator (*modules 1 and 2*). The display amplitudes of any two signals of widely different input amplitudes may thus first of all be matched. We remember, one signal is the *leader* (we will call its input channel "Y1" on the beam switching unit) and the other is the dependant ("Y2" channel).

Associated with the Y1 pre-amplifier is a *sync amplifier*. This develops a synchronisation signal from the salient flanks of only the Y1 signal, i.e. from the leader. This signal must be fed to the "external sync" input of the oscilloscope, since the internal synchronisation circuit of the oscilloscope can not work under these conditions.

The output from the beam switching unit to the normal Y-amplifier of the oscilloscope is a controlled mixture of the leader and dependant signals together with a switching waveform. The oscilloscope is unable to discriminate from this mixture which is the leader, which the dependant and which the switching waveform. As far as it is concerned, all three signals are equivalent and internal synchronisation would try to lock onto any one or all, giving an unsteady and unintelligible display. Thus the oscilloscope must be set to "external sync" and fed with a clear synchronisation signal derived in the beam switching unit from the flanks of the leader signal. Any signal may be taken as leader signal simply by connecting it to the Y1-

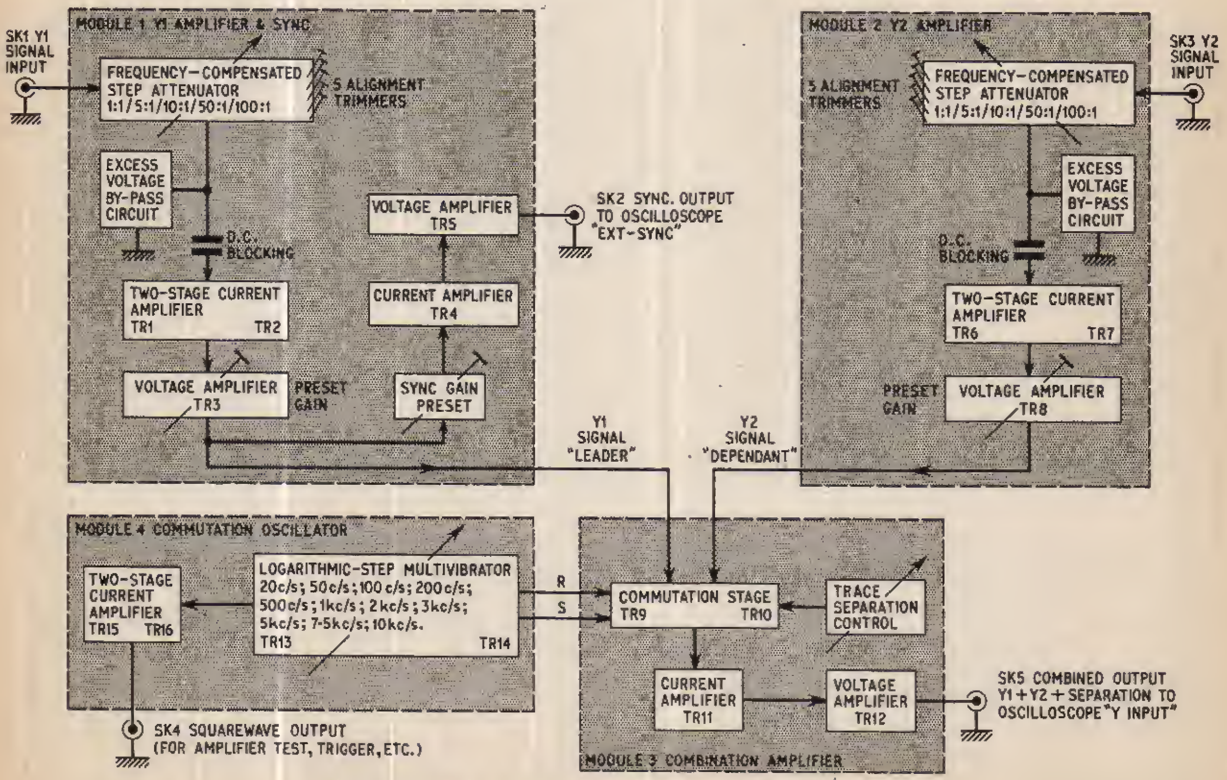


Fig. 1. Oscilloscope beam switching unit. Block diagram

channel input of the switching unit. The signal connected to the Y2-channel input is the dependant.

COMMUTATION

The function of the beam switching unit is to *commutate* the Y1 and Y2 signals alternately through to the output amplifier (*module 3*). For this purpose the unit contains a commutation oscillator (*module 4*), which is a conventional multivibrator generating an accurate symmetrical square wave. The antiphase outputs of the oscillator are fed to gating stages between the outputs of the Y1 and Y2 pre-amplifiers and the common output amplifier (*module 3*). Thus Y1 is connected through to the output amplifier during one half-period of the commutation square wave and Y2 is disconnected because its gate is closed by the antiphase square wave. At the moment the commutation square wave changes over to its other half-period, Y1 gate closes and Y2 gate opens and remains open for the duration of this half-period. Thereafter, Y1 is open and Y2 closed, and so on. As a result, Y1 and Y2 are each chopped through to the output for an average of half the time. Whenever Y1 gate is open, a third signal in the form of a controlled "beam separation voltage" is operative to throw the single electron beam higher up on the screen as a whole.

SWITCHING TRANSIENTS

It is now easy to grasp the resulting overall action. During each interval that Y1 gate is open, the electron beam is higher up on the oscilloscope screen and tracing a segment of the leader signal up there in the correct horizontal (time phase) position. During each interval that Y2 gate is open, the single electron beam is tracing a correctly positioned segment of the dependant signal. In between times, when neither Y1 nor Y2 gates are open, the commutation oscillator is switching over.

These switching transient intervals are very brief and normally negligible compared to the Y1 and Y2 intervals. However, the electron beam is travelling the greatest distances just at these moments, since it is shooting up to the Y1 region, or back down to the Y2 region then. The corresponding switching transient will therefore leave only a very faint trace intensity compared to the chain of Y1 and Y2 segments, since the electron beam is only very briefly at any selected point within the switching flank. Appropriate adjustment of the main brilliance control on the oscilloscope will thus make the switching flanks invisible, leaving only the assembly of Y1 and Y2 segments.

THE COMMUTATION OSCILLATOR

The commutation oscillator in this unit is provided with an eleven-position wafer switch on the front panel, providing eleven spot frequencies for the commutation square wave. These frequencies are staggered logarithmically from 20c/s to 10kc/s. This wide range has been included primarily in the interests of a separate output amplifier for this square wave signal alone, which has been included as a useful extra in our design. This square wave output is available at SK4 (*module 4*).

Having explained the general principle of operation, the individual circuits that together make up the beam switching unit will now be described in detail.

Y1 ATTENUATOR

Refer to Fig. 2. The entire circuitry around S1a and S1b, to the left of the vertical line through D2, constitutes a frequency compensated attenuator network to enable the amplifier to accept high amplitude input signals without overloading.

The signal attenuation is primarily established in the normal manner with a pair of voltage dividing resistors. For example, let us consider the "100" setting of the attenuation ratio selector switch S1. Here the input signal is fed via S1a to R4 and R10 in series, while the output signal is taken off across the small resistor R10 alone, via S1b. Since the value of R10 is a hundred times smaller than that of R4, the signal amplitude passed on to TR1 via S1b is only one hundredth of the input amplitude via S1a. The other steps for ratios of 50, 10 and 5 attenuation ratio function correspondingly, while the final step "1" is in effect a straight-through connection without attenuation.

FREQUENCY COMPENSATION

Such a simple resistive step attenuator will not function above about 5kc/s, because the division ratio due to the parallel stray capacitances, which have then dropped to comparable impedance levels, may be different and arbitrary. This may lead to low or high peaking, whereby a square wave would be distorted to a rounded sawtooth or receive transient noses and a drooping roof, respectively. In either case, this implies phase and amplitude distortion with respect to the signal frequency, and while such an amplifier may be suitable for audio work, it is useless for general oscilloscopy.

The trick to overcome this trouble is to swamp the stray capacitances with comparable or larger intentional capacitors which can be suitably adjusted to make the capacitive division ratio identical to the resistive one. In other words, the product of capacitance (time constant) of every section of the entire network must be the same. The frequency and phase response is then linear, theoretically right up to infinitely high frequencies.

The actual value of the section time constant is not of primary importance, being determined by secondary matters. The greater its value, the easier the adjustment but the greater the damping loading imposed on the circuit from which the input signals are derived. A compromise thus has to be struck, and 100 microseconds is a commonly accepted value. In a 1megohm resistive impedance circuit, this allows 100pF total parallel capacitance, about 60pF of which will be taken up by the self-capacitance of the coaxial cable to the signal probe. The balancing or trimmer capacitors (TC1–TC5) for the attenuator network must thus be adjusted so as to present about 30 to 40pF between S2a slider and chassis in all settings.

THE SIGNAL PROBE

Fig. 8 shows that the probe is simply another "top-section" of a capacitively balanced resistive divider. In the attenuation setting "1", which is a straight-through connection on the module itself, the 10 megohms/10pF of the probe constitute the top section and R14, R15, C12, TC5 and connecting cable represent the bottom section. The probe thus gives a 10:1 attenuation factor in the "1" setting, and correspondingly ten times the attenuation factors of the other settings. It is seen that C12 and TC5 are required only in conjunction with the probe, in order to be able to establish frequency balance therewith in the "1" setting too. They do not interfere when using the input without a probe.

The inputs should under all normal circumstances be used with their probes, since only then is the damping loading and signal falsification of high impedance

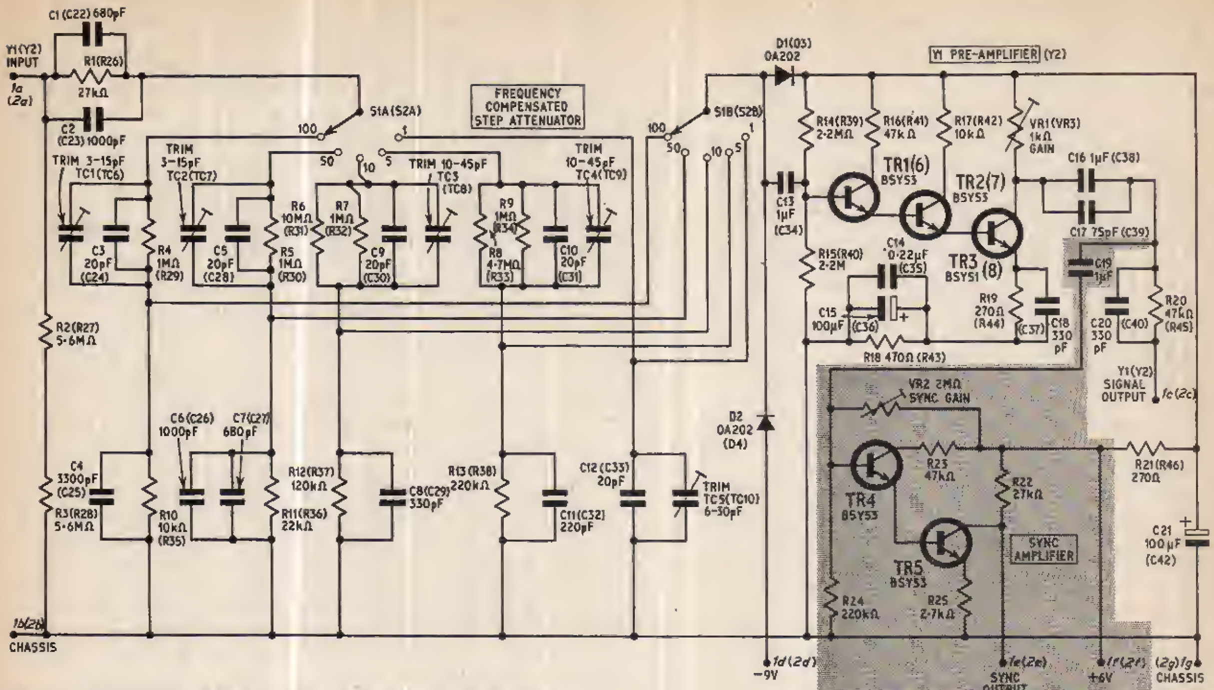


Fig. 2. Circuit of module 1 — Y1 attenuator, pre-amplifier, and sync amplifier. Module 2 is identical to this circuit except for the area shown shaded. Module 2 circuit references are in brackets

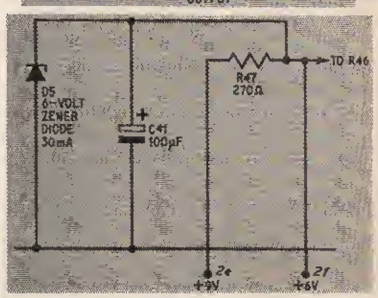


Fig. 3. This section of circuitry applies to module 2 only —Y2 attenuator and pre-amplifier

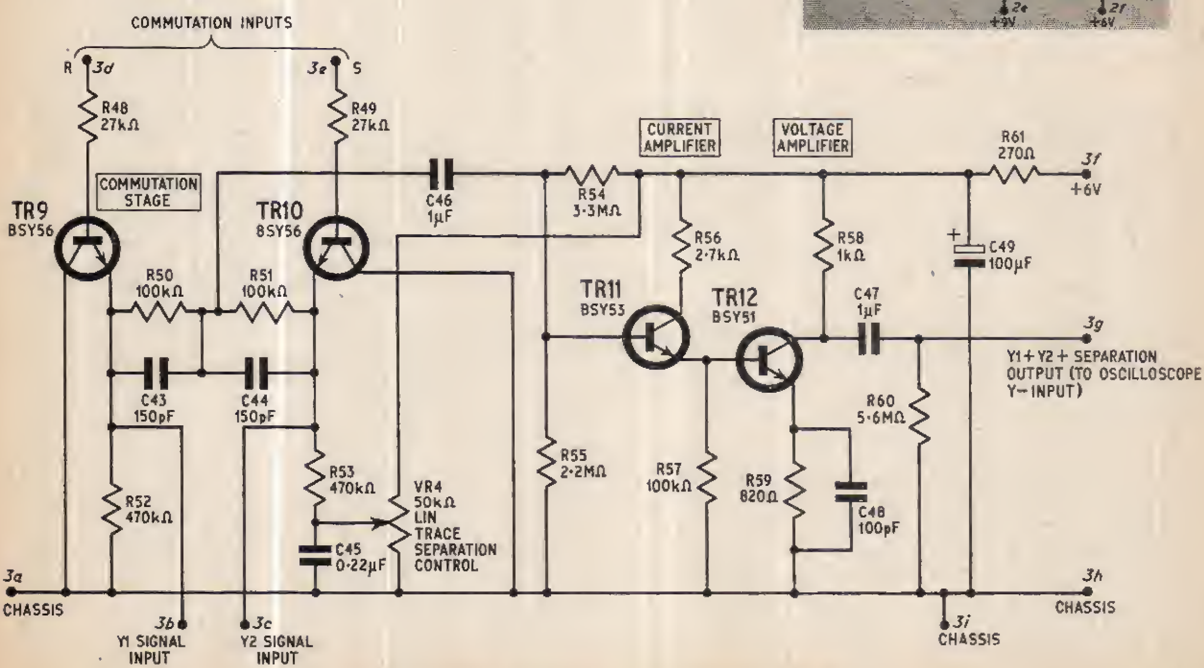


Fig. 4. Circuit of module 3 — combination amplifier

sources sufficiently small. The direct input may be used only at low frequencies and/or on very low impedance sources, when the benefits of the ten fold greater sensitivity are available. However, if the signals are of high amplitude, needing attenuation, the probe should first of all be inserted even for low frequency low impedance sources instead of switching the attenuator to a higher ratio.

With careful adjustment, the entire step attenuator network around S1 (or around S2 in the Y2 module) has level phase and frequency response from d.c. to many Mc/s and the actual cut-off frequencies are determined by the limitations of other parts of the entire circuitry.

SAFETY FUNCTIONS

The remaining components in the Y1 input circuit fulfil safety functions.

R2 and R3 complete a d.c. path for C74 (Fig. 6) in the "1" setting. R1 limits the charging surge current for C74 to about 20mA when the input is suddenly tapped onto a point at the maximum permissible d.c. level of ± 500V. C74 will charge up and block d.c. levels of either polarity up to this magnitude, even if the attenuator is set to a much greater sensitivity to scope small superimposed a.c. waveforms. This is permissible with or without the probe, in any setting of the attenuator. The transient charging surge for C74 appears almost entirely as a voltage pulse across R1, where it is harmless. According to the polarity of the blocked d.c. level, D1 or D2 conducts and limits the transient amplitudes actually reaching the transistors to a harmless level equal to the low supply voltages used to bias these bypass diodes. Both diodes D1 and D2 are normally cut off and thus without effect upon the circuit.

Note that it is NOT permissible in this circuit to obtain d.c. blocking transient bypass with the help of a pair of low voltage Zener diodes connected back-to-back between S1b slider and chassis, because all low voltage Zener diodes which the author was able to trace in makers' lists have self-capacitances of 100 to 1000pF, which are far too great for present purposes. It is thus necessary to use biased ordinary diodes in the positions shown. Make sure that low-capacitance r.f. types, e.g. television video detector diodes, are employed if adopting substitutes to the specified ones. The self capacitance at some 6V reverse bias should not exceed a few pF in the makers' ratings.

All capacitors to the left of D2 in Fig. 2 (and to the left of D4 in Fig. 3) must be 500V working rating, whereas all components to the right of these diodes including C12 (C34) may be low voltage types, since no high voltage transients pass to the right of the bypass diodes.

Note that the described bypass arrangement will also give full protection if excessive input signal amplitudes are applied, i.e. if the attenuator is set to the wrong step position in relation to the a.c. signal amplitude. The output signal reaching the oscilloscope will then approach a square wave for any input waveform, i.e. distortion will be tremendous, but no damage is suffered on a.c. input signals up to 500V peak-to-peak amplitude, whatever the attenuator setting. Nevertheless, do not prolong the application of excessive signal amplitudes in an incorrect attenuator setting, since R1 could otherwise gradually overheat if the frequency is high.

Also note that R1 is shunted with C1 and C2, since it represents another section of the attenuator network and must therefore be shunted capacitively to the same time constant.

Y1 SIGNAL PRE-AMPLIFIER

TR1 and TR2 constitute a two-stage current amplifier (emitter follower cascade) whose function is to step-down the impedance level. TR3 is a voltage amplifier stage with a gain of about 3–4 maximum, adjustable with VR1. This compensates the slight voltage loss in the chain TR1 to TR3 emitter, as well as the subsequent division via R20 and the chopper gate circuits on the combination amplifier module. The intention is to make the overall gain exactly unity from Y1 or Y2 input terminal to combination output terminal, in the respective "1" settings of the attenuators. The beam switching unit as a whole then involves zero insertion loss when connected to the oscilloscope.

As is already evident with C20 across R20, the output feed to the combination amplifier involves considerable capacitive loading. Two measures have been adopted to make this tolerable without undue restriction of bandwidth. Firstly, VR1 is of very low value, permissible thanks to the impedance step-down of TR1 and TR2. Secondly, emitter compensation has been applied by shunting R19 with C18 to yield an emitter circuit time constant roughly equal to the collector circuit time constant of VR1 with the feed capacitances. R19 at the same time provides a.c. negative feedback, and together with R18 heavy d.c. negative feedback over all three stages to stabilise the operating points.

Y1 SYNC AMPLIFIER

The sync amplifier consists of TR4 and TR5. The circuit configuration is similar to the final two stages TR2 and TR3 of the Y1 pre-amplifier, but without capacitive compensation since wide bandwidth is not necessary. R25 provides strong a.c. and d.c. negative feedback so stabilising the gain and operating point set with VR2. This preset control should be adjusted for maximum possible undistorted output swing, as will be described later.

"Y2 AMPLIFIER" MODULE

Module 2 (see Fig. 3) is in every way identical to the corresponding sections on *module 1*—but the sync amplifier section is absent.

The Y1 and Y2 step attenuator switches may of course be set to very different positions, for matching two signals of widely different amplitudes for simultaneous display. Assuming that this is tolerable on other considerations (see above), one channel may be operated without a probe and the other with a probe. In the extreme case, this permits matching of signals differing in amplitude by a factor of 1000:1 (60 dB) for approximately equal height display on the dual trace.

The graded steps of the attenuators are normally adequate, so that there is no need for a continuous fine control whose inclusion would have added complications of frequency balancing.

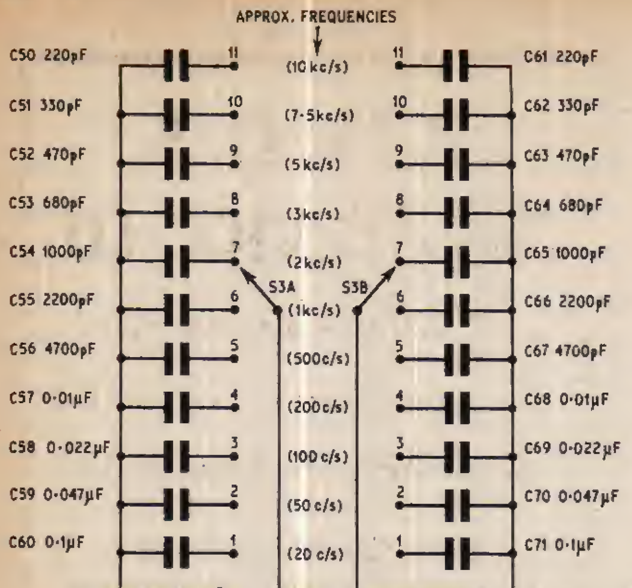
COMMUTATION GATES

The circuit diagram for the combination amplifier is given in Fig. 4.

R20 (*module 1*), R50, R51 (*module 3*), and R45 (*module 2*) constitute a balanced resistive bleeder strung between the output collector circuits of the channel pre-amplifiers. The combination signal to the actual combined signal amplifier TR11, TR12 is taken from the centre of this balanced bleeder, i.e. from the junction of R50, R51.

The two transistors TR9 and TR10 are the chopping gates which effect the required signal commutation by chopping the chassis connection on the balanced bleeder to alternating sides of the pick-off point for the combined signal amplifier. The bases of the gate transistors are fed via R48, R49 respectively with anti-phase square waves from a multivibrator as commutation oscillator. These square waves commutate between −3V and +6V at the input ends of R48 and R49 which go directly to the respective collectors of the commutation multivibrator.

Whichever gate transistor of the pair TR9, TR10 happens to be connected to the −3V level during any commutation half-period is cut off and may thus be considered as if it were not present. The other transistor is at the same time fed with +6V at the input end of its base resistor, so that it is cut on hard and presents an effective short circuit to chassis between emitter and collector. In other words, it places a chassis connection on to the balanced signal bleeder at the point of connection of its emitter thereto. For the other half period of the commutation square wave, the gate transistors change over their roles.



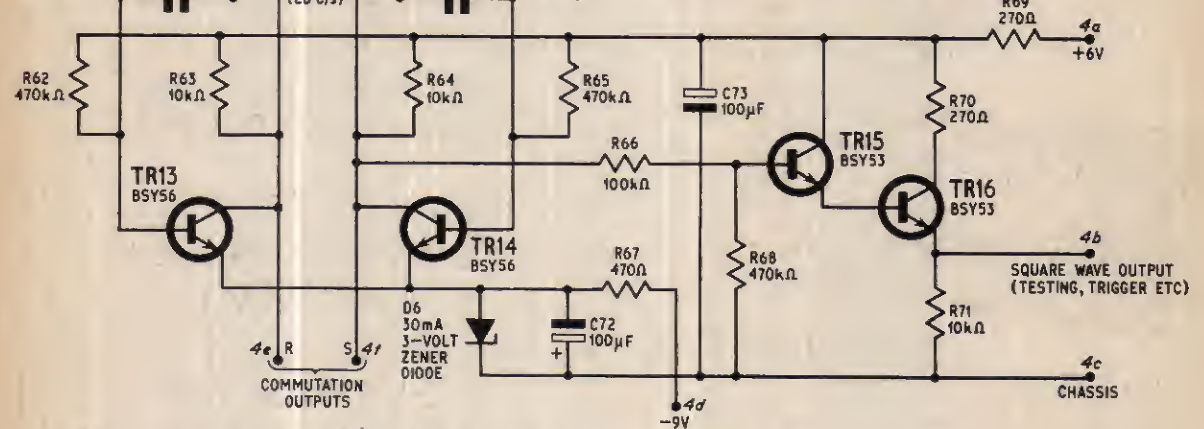


Fig. 5. Circuit of module 4—commutation oscillator

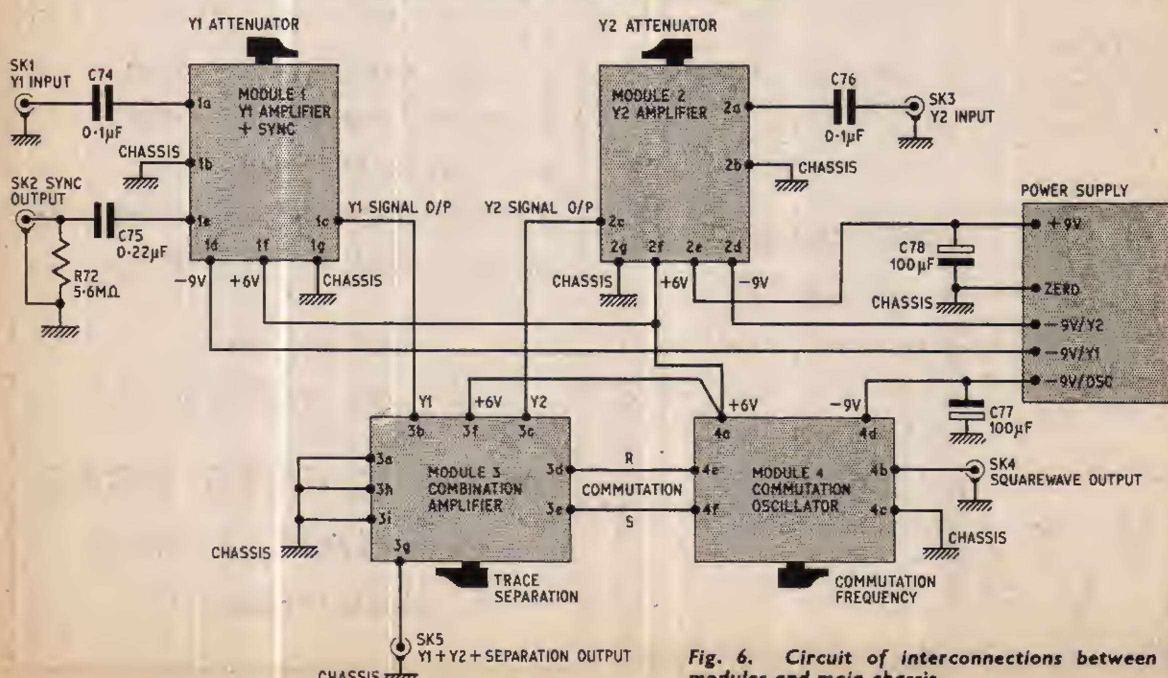


Fig. 6. Circuit of interconnections between the modules and main chassis

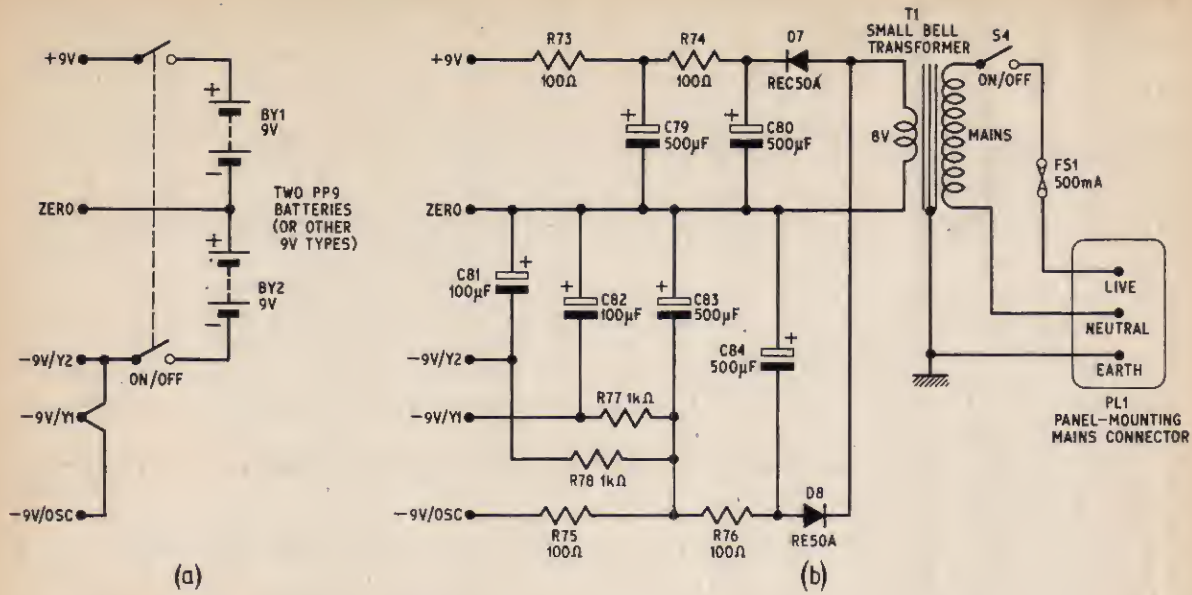


Fig. 7. Power supply circuits: (a) arrangement for battery operation;(b) mains operated power unit

TRACE SEPARATION CONTROL

R53, R51, R50 and cut-on TR9 represent a d.c. voltage bleeder to chassis for the voltage at VR4 slider whenever TR10 is cut-off. When TR10 is cut-on, it shorts-out this d.c. voltage along with the Y2 signal. There is thus a positive square wave component (controlled by VR4) at the junction of R50, R51 whenever and only when the Y2 signal is being fed-through. In other words, Y2 is given a positive d.c. component and Y1 is left as pure a.c. After phase inversion in TR12, Y2 has a *negative* chopped d.c. component and Y1 is pure a.c. Both signals are also again "erect", i.e. of the same polarity as at the inputs to their respective modules, since TR3, TR8 had already caused one phase inversion.

A correctly designed oscilloscope gives upward beam deflection for positive and downward beam deflection for negative signal inputs. Thus the beam switching unit maintains this convention and always makes the Y2 (dependant) trace appear lower down on the screen than the Y1 (leader) trace when trace separation voltage is inserted.

COMBINED SIGNAL AMPLIFIER

TR11 and TR12 constitute the combined signal amplifier which handles the commutated mixture of Y1, Y2 and separation signals. Its output feeds the normal Y-amplifier input of the oscilloscope. The values of C46 and C47 determine the bass cut-off frequency and thus the lowest usable commutation frequency. The values of 1μF each as shown in Fig. 4 represent the highest convenient ones which are possible with modern sub-miniature printed circuit capacitors without resorting to electrolytics.

Due to the high impedance at TR11 base, leaky electrolytics would lead to trouble, but good ones may be tried. However, the values of 1μF shown for C46 and C47 already give a cut-off frequency around 1c/s which is better than most a.c. oscilloscopes.

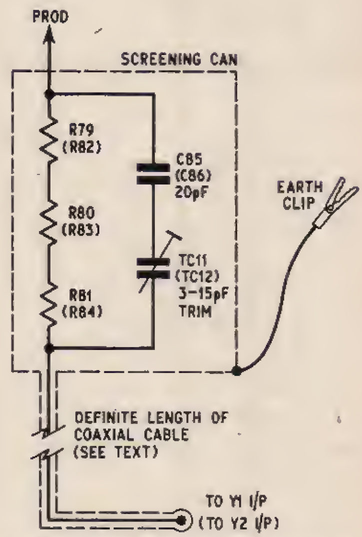


Fig. 8. Circuit of input signal probe

TR11 is a current amplifier (emitter follower impedance step-down stage) and TR12 is a conventional voltage amplifier stage with a gain very slightly greater than unity, determined by the ratio of the emitter and collector resistors. This compensates the slight voltage loss in TR11. TR12 also provides the essential final phase reversal, to compensate the first phase reversal due to the voltage amplifier stages TR3 and TR8 in the respective pre-amplifier modules.

Almost any oscilloscope may be used immediately, without any special matching measures. However, the design basis was that a cable (coaxial) of about 60pF self capacitance is connected to feed an oscilloscope Y-amplifier with about 30pF input capacitance and any resistive input impedance component greater than 100 kilohm (usually 1 megohm).

This places a total capacitive shunt of around 90pF across R58, giving a cut-off frequency around 1Mc/s without compensation. The cut-off time constant is about 90 microseconds and has been duplicated by placing C48 across R59 in the emitter circuit. This gives additional boost around the h.f. cut-off, so that the overall bandwidth of the entire circuit including all modules is at least 1Mc/s. Some experiments may be worthwhile using other values for C48 if the sum of the connecting cable and oscilloscope input capacitances differs from about 90pF.

AVAILABLE OUTPUT AMPLITUDE

TR12 circuit in the combination amplifier has been designed to give a maximum undistorted output amplitude (peak-to-peak) of at least half a volt. It is thus suitable for feeding the Y-amplifier of an oscilloscope set to a sensitivity of 0·5V for full-screen deflection. This corresponds to some 100mV/cm for the usable trace height of a scope using a 3in c.r.t. or 50mV/cm for one using a 5in c.r.t. These are normal and common figures for the maximum sensitivity settings of the majority of such oscilloscopes as used for amateur, servicing and educational purposes.

The beam switching unit should thus be used with the oscilloscope Y-amplifier set to maximum gain, and signal attenuation undertaken with the help of the input probes and respective step attenuators on the beam switching unit.

Note that Y1 and Y2 are not present simultaneously, but only alternately, so that the full 0·5V peak-to-peak swing is available for both signals. However, trace separation is in fact added to the Y2 signal, whose swing must be kept correspondingly smaller when trace separation is inserted. The Y1 signal is unaffected. The maximum available trace separation at the output from TR12 is about 0·3V, or about 60 per cent of the screen height under optimum conditions. The Y2 signal must then be kept down to 0·2V amplitude to avoid overloading. More space is not available on the screen anyway when the traces are separated that far.

COMMUTATION OSCILLATOR

Fig. 5 shows the circuit of the commutation oscillator (*module 4*).

TR13 and TR14 constitute a conventional symmetrical multivibrator operating between −3V and +6V stabilised supply lines. The switch-over is very rapid, so that the waveform at each collector is a good square wave commutating between −3V and +6V. The negative lower level, instead of the otherwise customary choice of chassis potential as lower level, is here necessary to make sure that the gate transistors in the combination amplifier are cut off hard under all circumstances when the oscillator collector feeding them is at the lower commutation level.

SWITCHED CAPACITORS

The use of constant resistors and switched capacitors assures the same rise and fall time ratios for the output square waves at all frequencies, so that direct comparisons of amplifier performance at all frequencies are possible. Were variable potentiometers to be used as the base resistors with fixed capacitors, the higher frequencies would yield very rounded and useless "square" waveforms.

BASE-TO-EMITTER VOLTAGE RATING

A final point of some importance in the multivibrator circuit concerns the moment just after a switch-over relaxation. The transistor which has thereby just been cut off is momentarily driven negative at the base to an extent such that the reverse voltage between base and emitter is equal to the full supply voltage (3 + 6 = 9V in our case). Very few transistors are rated for such high base-to-emitter reverse voltages, the common limiting ratings being 5 or 7V for silicon transistors, sometimes even lower. The author has tested the circuit with a large variety of silicon transistors in the 5 and 7V range, and even some with 3V limiting rating. All performed quite satisfactorily for long periods. Upon reporting these findings to a manufacturer, the author was told that all base to emitter limiting ratings are well below voltage breakdown, which normally does not take place until values two to three times as great are reached. They are more concerned with other characteristics of the particular transistor. Silicon transistors may be used in relaxation circuits with virtually perfect reliability under any conditions where the peak base to emitter transients do not exceed twice, better still 1·5 times the static limiting rating. Types with a V_{EBO} rating of 5V, preferably 7V, are therefore quite satisfactory for the present circuit.

GATE DRIVE

The commutation oscillator drives the gate transistors TR9, TR10 on the combination amplifier module, via the two 27 kilohm resistors R48, R49. These resistors are essential to prevent the waveform collapsing at the oscillator collectors when the gate transistors open.

The greater the values of R48, R49, the less the collapse of the positive part of the oscillator waveform at the oscillator collectors and the better the output of the square wave test amplifier. The maximum permissible value of R48 and R49 is dictated by the need to still turn on the gates hard enough. For this purpose, one may treat each gate transistor as an emitter follower which divides this base feed resistor by its current gain and presents the resulting low resistance value as bottom-end section of a resistive bleeder for the signal feed it is supposed to be shorting right out.

This leads to a finite cross-talk factor. With the value of 27 kilohm for the base feed resistors and a current gain of 120 for the specified transistors under the given conditions, this cross-talk resistance is about 200 ohms. This is forming a bleeder with the 47 kilohm resistor from the pre-amplifier output which it is supposed to be shorting out. The division ratio is thus about 100, multiplied by a further factor of 2 due to the resistors R50, R51. The cross-talk is thus about 0·5 per cent. This is of the same order of magnitude as the spot diameter on the c.r.t. and thus not seriously noticeable. It may be slightly greater near the h.f. cut-off frequency if unbalanced capacitive effects then arise. To minimise this, it is advisable to choose a matched pair for the gate transistors, though not essential. The use of gate transistors with lower current gain will lead to intolerable increase of cross-talk between Y1 and Y2 with the given circuit values.

POWER SUPPLIES

The circuit of the power supply section appears in Fig. 7b. A bell transformer T1 provides an output of approximately 8V from the mains. This output is applied to two separate rectifier and filter circuits; one of these providing a 9V positive output, the other a 9V negative output.

If preferred, the beam switching unit can be driven from a pair of 9V batteries as indicated in Fig. 7a.

Part Two next month will include the complete components list, detailed constructional drawings, and setting up and alignment instructions.