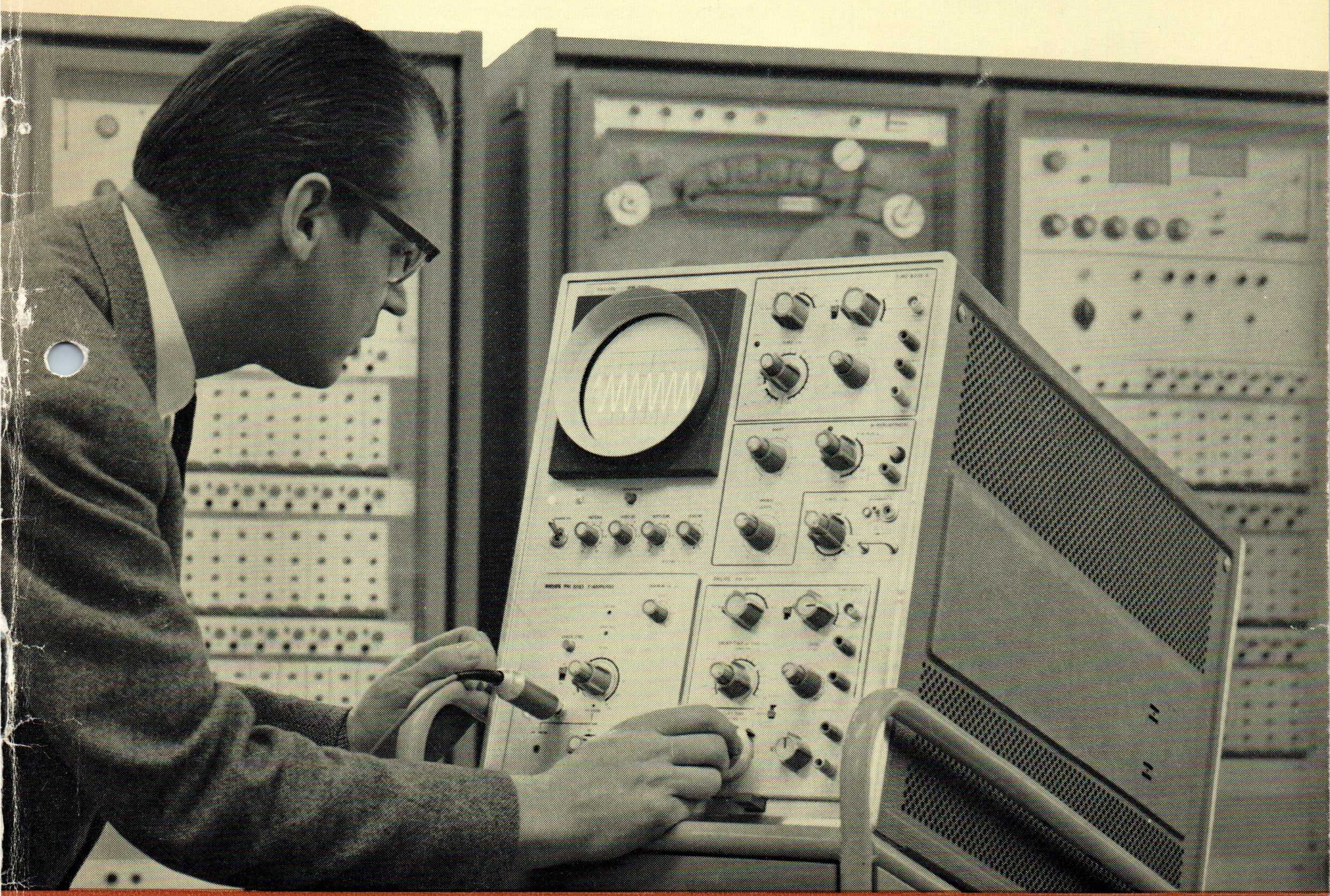
PHILIPS

electronic measuring and microwave notes



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electronic measuring and microwave notes

Introduction The quarterly periodical Electronic Measuring and Microwave Notes, provides information about the application and design of Philips electronic measuring and microwave instruments and also surveys the new instruments which are regularly added to the Philips programme.
The information is intended to assist users in getting the maximum benefit out of instruments which they already possess and to help them in choosing new instruments which will best meet their particular measuring or microwave problems.

In this issue

General information If you are interested in regularly receiving the periodical Electronic Measuring and Microwave Notes and also in more information on the instruments please ask your Philips organisation. If there is no Philips organisation in your country enquiries may be sent to n.v. Philips, EMA Department, Eindhoven, the Netherlands.

Editor T. Sudar, n.v. Philips, EMA Department, Eindhoven.

The front cover of this issue illustrates the 60 MHz plug-in oscilloscope.

60 MHz plug-in oscilloscope with new measuring facilities

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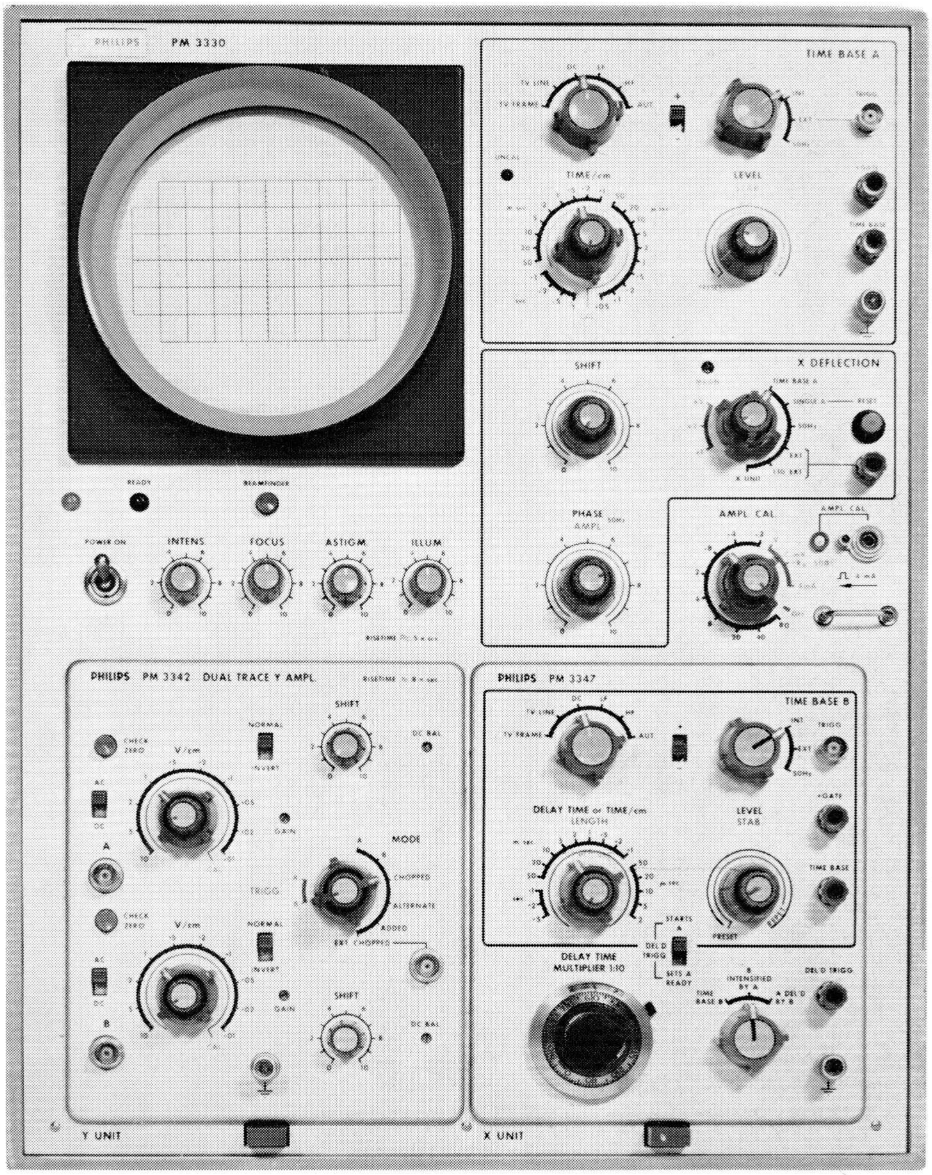

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INTRODUCTION ON THE PM 3330

by J. Poole

Of all types of electronic measuring apparatus the oscilloscope has the most critically regarded specification and probably the one with the most variables. Of these the two things that users ask for most are more bandwidth and more sensitivity. Unfortunately these two, far from going hand in hand, are such that one can only be increased at the expense of the other. This is because an increase of either will increase the noise delivered by the amplifier. Thus a compromise has to be made leaving the user to choose between high sensitivity and wide bandwidth, when he would rather have both and will certainly need one at one time and the other at another. However in the PM 3330 steps have been taken to relieve the user of this difficult decision and these steps go even further, in that they open up new fields of measurement for him.
For example the PM 3332 plug-in unit with DC to 50 MHz bandwidth at 500 μV/cm sensitivity for the full bandwidth.

The PM 3330 is a plug-in 60 MHz oscilloscope. However the key to its wide range of measuring capabilities does not lie in a wide variety of plug-in units, but in the individual capabilities of each of them. This philosophy has been made possible by the development of special circuits for the plug-in amplifiers, which extend their measuring capacities to provide unusually large overlap.

Obviously the inclusion of these special circuits could not be allowed to jeopardize the reliability of the complete instrument. To this end each of the new circuits has been subjected to prolonged testing. As a further safeguard, wherever possibly, none of the components in either the classical or the newly development circuits, are run at more than 70% of their max. rating. Adequate temperature compensation has been incorporated throughout to ensure reliable and accurate operation regardless of changes in the ambient temperature over a wide range.

As an additional safeguard to the user, several of the first production instruments have been subjected to life tests under normal user conditions in the Philips concern; before any have been offered on the market.

It will be appreciated that, through this care for quality, every possible precaution has been taken to ensure that the instrument will not only meet its specification at the time of delivery but will continue to do so for many years to come.

Of course it is not enough when developing an oscilloscope for use in the laboratory simply to provide it with the best possible amplifiers. Thus to ensure that its amplifier capabilities are not wasted, the PM 3330 has been designed to give top class performance in all aspects of its specification. Furthermore the facilities incorporated in the instrument are sufficient to ensure that it is adequately equipped to carry out the most complicated measurements without the user having to resort to the use of external auxiliary units. Moreover the provision of the delaying time base as a plug-in unit; while the main time base is built-in, ensures that these comprehensive facilities are made available to the customer in the most economic way. Thus he needs not buy the delaying time base until it is needed, and when it is needed he only pays for the extra facility as the main time base remains part of the complete system. Furthermore there are 80 watts of power available for user built plug-ins in the 'X' aperture.

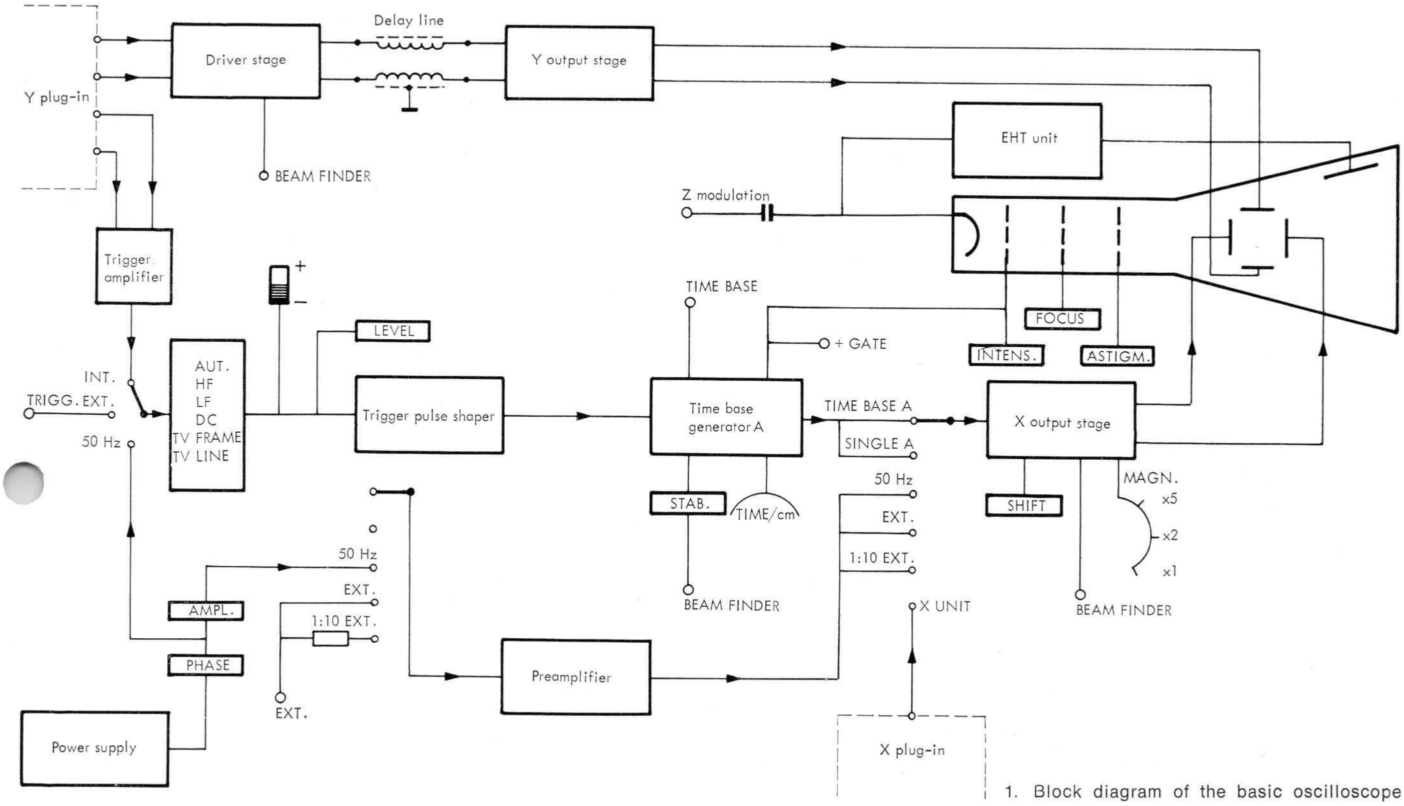
To further enhance the value of this instrument new developments have not been restricted to the plug-in amplifiers but have been spread throughout all the circuitry. These include such items as a complete overload and short circuit protection of the power unit and a specially developed cathode-ray tube which offers a 6 x 10 cm useful scan with a rise time of 5 ns using a conventional amplifier and without resorting to a mesh tube. Thus with 10 kV acceleration voltage a sharp, bright trace is obtained.

Tunnel diode triggering and careful circuit design have been made it possible to provide 60 nanoseconds of visible delay with only 150 nanoseconds of signal delay. In this way it is possible to study the start of pulses used to trigger the time base, including those with a poorly defined leading edge. Furthermore the maximum time base sweep speed of 10 nanoseconds per centimetre, using the 5 times expansion, makes the time base fully compatible with the amplifier rise time.

The vertical amplifier plug-in units offer some particularly interesting and valuable features, such as the special HF probe system, PM 9332, available with the wideband plug-in unit PM 3333. Though this is neither an attenuator nor a cathode follower probe, it offers the best features of both; wide bandwidth, high "tip sensitivity", small size, low capacitive loading and long lead. A second example is the high sensitivity, wide bandwidth, drift free plug-in unit PM 3332. This offers the sensitivity of a low frequency amplifier with true, DC coupled, wide bandwidth without drift and with low noise.

TECHNICAL DESCRIPTION OF THE MAIN FRAME PM 3330

by A. H. J. Sloots



1. Block diagram of the basic oscilloscope

The Philips cathode-ray oscilloscope PM 3330 is a 60 MHz instrument which has a wide range of applications. This is achieved by the utilization of plug-in units, each of which covers one or more fields of application. Plug-in units are available both for the Y and the X channels.

Fig. 1 shows the block diagram of the basic oscilloscope. The Y signal derived from a Y plug-in unit, is symmetrically fed to the „delay line driver". The delay line gives a total delay of about 150 nsec. The cable is terminated on both ends in order to minimize reflections. The signal is fed via the output stage to the Y deflection plates.

The trigger signal is taken-off separately at the Y plug connector and fed to the trigger circuit via a trigger preamplifier. As the trigger signal is thus isolated from the Y signal, triggering will not be affected by the switching frequency when a multi trace unit is used. Triggering can be effected internally, externally or with an internal 50 Hz source. The apparatus offers the following triggering facilities: Automatic, HF, LF, DC, TV frame and TV line.

A tunnel diode trigger pulse shaper is used to convert the trigger signal into pulses, thus ensuring stable triggering.

These trigger pulses start time base generator A. The time base works on the principle of the Miller integrator. An X deflection switch affords selection of the following display modes:

1. Time base A
2. Time base A, single-shot
3. 50 Hz (phase and amplitude adjustable)
4. External 1 : 1
5. External 1 : 10
6. X unit

In modes 3, 4 and 5 the signal passes through a **preamplifier.**

With the selector in the X unit position, the X plug-in unit is brought into use; for instance for a second time base for delayed triggering purposes, or a special X amplifier. The selected signal is then fed via the X output stage, to the X plates.

The X output stage incorporates the x1, x2 and x5 expansion circuit.

The basic oscilloscope is also fitted with a beam finder. By pressing the beam finder button the amplification factors of the Y and X amplifiers are reduced so that the electron beam is not deflected beyond the display area. At the same time the time base A is made free-running. It is then a simple matter to adjust the image to the centre of the screen by operating the shift controls.

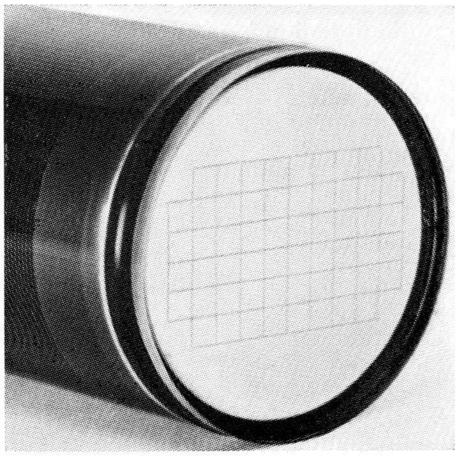
A few specific aspects will now be discussed in further detail.

Cathode-ray tube

The type used is the Philips D13–16GH/1 (see fig. 2). The screen of the CRT incorporates a measuring graticule. This graticule is printed on the inner surface of the glass so that errors caused by parrallex are obviated.

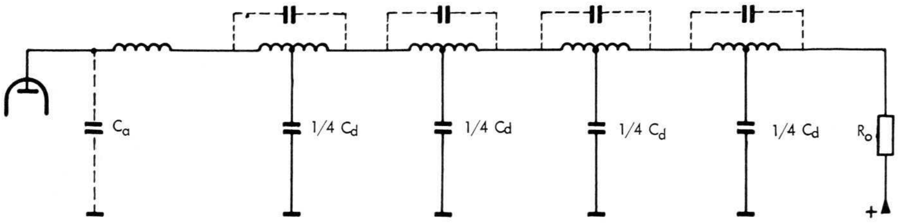
To ensure that the horizontal deflection of the spot coincides exactly with the internal graticule, a magnetic field is produced by a coil mounted inside the mu-metal magnetic screen surrounding the tube. The direction and intensity of the magnetic field are variable.

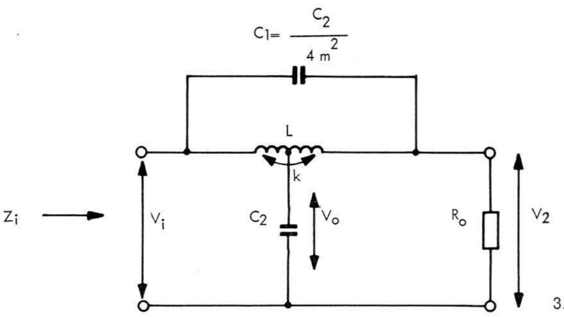
2. Internal graticule and helix of the cathode-ray tube



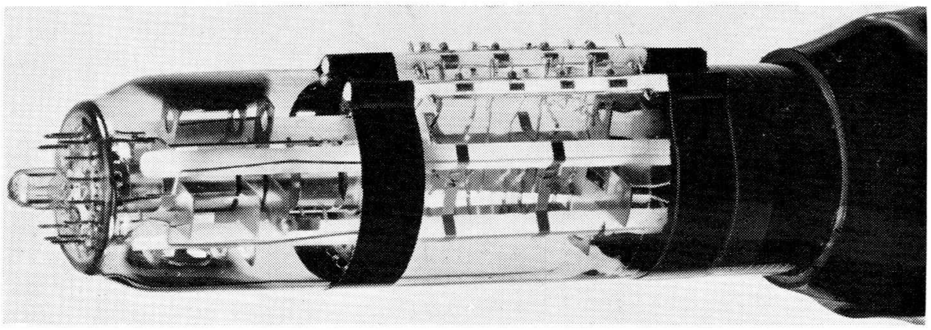
A second coil system is attached to the neck of the CRT. Correct adjustment of the currents through these coils, will ensure that the X and Y deflection are exactly perpendicular and that the useful scan is central for uniform vertical overlap of the graticule. By employing a resistance helix it was possible to select an anode/post acceleration voltage ratio of 1 : 6.
The useful display area of the CRT is 6 x 10 cm and the deflection coefficient in the Y direction is 6 Volt/cm. Unblanking pulses generated by the time base, during the sweep, are DC coupled to the grid of the CRT via a floating grid supply located in the EHT supply.
When a multi trace plug-in unit is used, the cathode is connected to a transistor amplifier. The latter receives blanki[illegible] pulses from the unit. These pulses a[illegible] generated during switch-over when operating in the chopped mode.
The intensity of the beam can be modulated by connecting the modulating signal to the cathode via the built in coupling capacitor.
In order to obtain a large bandwidth, the deflection plates are divided into four pairs, and each pair forms part of an "m"-derived bridged T network (see fig. 3).
The absolute transfer value of V_2/V_i, i.e. the amplitude frequency response of the filter between the output and input, will be unity if the filter is terminated with its characteristic impedance, i.e. $\sqrt{L_k/C_k}$ and

$$C_1 = \frac{C_2}{4m^2}, \text{ where } m = \sqrt{\frac{1+k}{1-k}}$$



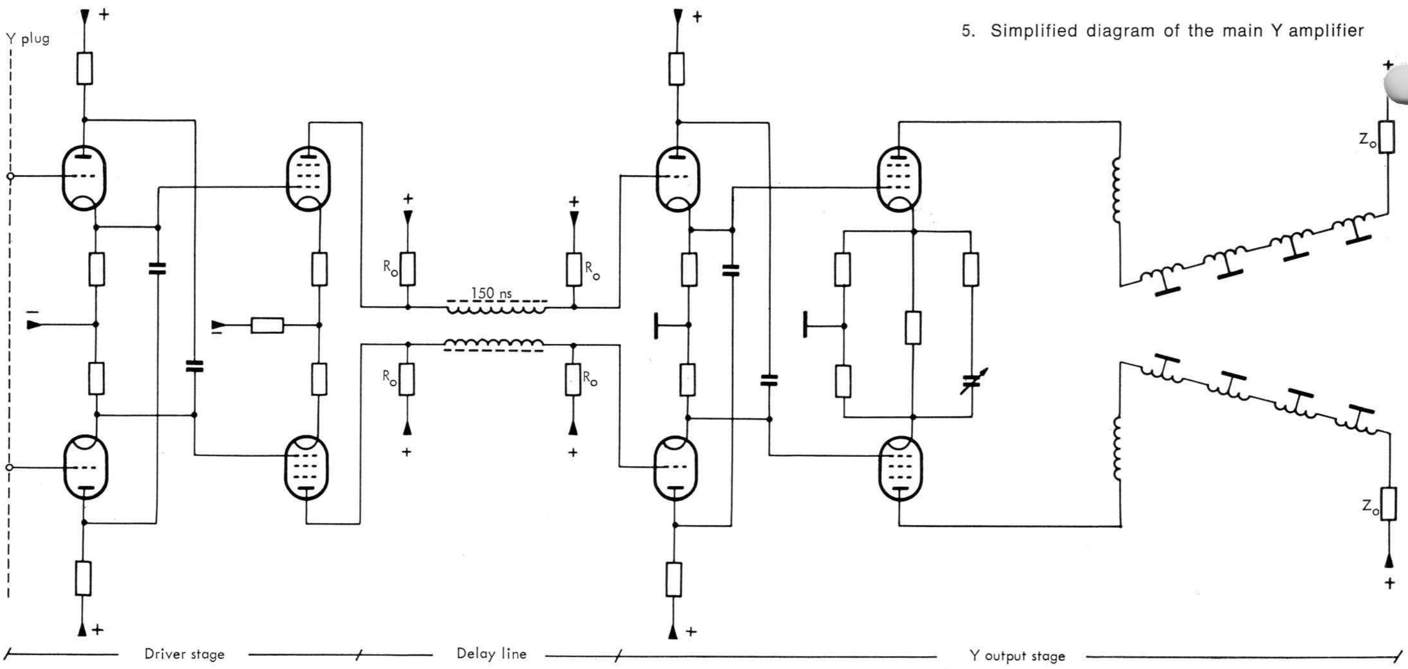


3. Deflection plates included in a filter



4. CRT with filter

5. Simplified diagram of the main Y amplifier



The input impedance Z_i then equals R_o thus Z_i is independent of frequency.
The voltage transfer to the deflection plates V_o/V_i is expressed by the equation:

$$\frac{V_o}{V_i} = \frac{1}{1 - \omega^2 \frac{LC_2}{4\,m^2} + j\omega \frac{\sqrt{L\,C_2}}{2}}$$

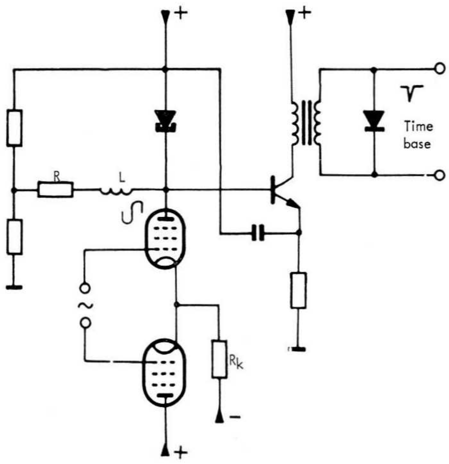
The output voltage V_o, for a unity step function V_i is:

$$V_o = 1 - \frac{2}{\sqrt{4-m^2}}\, e^{-\frac{m^2}{R_oC_2}t} \times \sin\left(\frac{m}{R_oC_2}\, t \sqrt{4-m^2} + \arccos\frac{m}{2}\right)$$

For $m^2 = 3$ the overshoot is 1%. By means of this filter and a 600 Ω anode resistor, as well as extra feedback in the cathode circuit a bandwidth of about 80 MHz has been obtained for the output stage.
The capacitance $C_2/4\,m^2$ is the self capacitance of the coil. Fig. 4 shows this filter fitted to the cathode-ray tube.

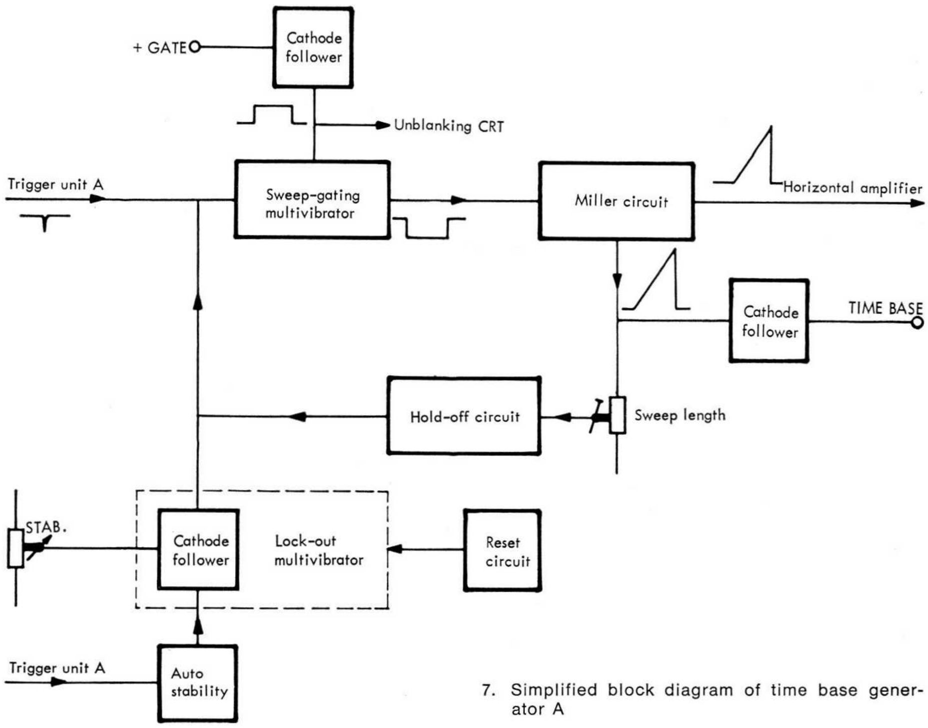
Vertical amplifier
This amplifier is a wideband DC amplifier consisting of two stages (see fig. 5). The simple design ensures faithful response to step functions.
The delay line driver is equipped with high g_m frame grid pentodes E810F.
Apart from the usual feedback circuitry for stabilizing amplification, the cathode circuit also contains several RC networks with different time constants, which serve to compensate losses due to the skin effect of the delay line. As valves with high mutual conductance have a large input capacitance, they must be driven from a voltage source with a very low output impedance, which is obtained by using a cross coupled cathode follower. This is also the reason why the output stage, containing very high g_m tubes (E55L) is preceeded by such a cathode follower. At the same time the delay line is given a better termination, so that reflections are to a large extent suppressed.
Besides the usual feedback circuitry in the cathode circuit of the input stage for stabilizing the amplification, there is a high frequency correction network for step function adjustment.
The delay line consists of two magnetically coupled spiraled cables in the same screening, which has a characteristic impedance of about 270 Ω.

Triggering
In position DC of the trigger mode selector the trigger signal is DC coupled to the trigger circuit whilst AC coupled triggering is provided in the other positions. In position "Automatic" the time base is free-running without a trigger signal. This gives a bright trace even at the shortest sweep times of the time base generator. As soon as a signal is applied, a special circuit automatically resets the time base for stable triggering. To achieve this the trigger signal is rectified, and the DC voltage thus obtained is fed to the time base stability circuit, bringing the time base circuit into triggered operation.
In the positions TV frame and TV line raster pulses and the line pulses respectively, are selected from a complete video signal for triggering.
Fig. 6 shows a simplified diagram of the trigger pulse shaper.
The internal trigger signal is fed symmetrically. The trigger unit is provided with a tunnel diode circuit so that a stable triggering, reliable up to high frequencies is obtained. Operation is explained in appendix, page 9.
As the voltage drop across the tunnel diode amounts to about 0.5 volts, and the time base circuit needs trigger pulses of about 2 volts, an NPN transistor is inserted for amplification. At the collector a 1 : 1 transformer transfers the signal to the time base input circuit. This transformer also differentiates the signal, but the positive-going spikes are practically short circuited by a diode (see fig. 6). The remaining negative spikes are used to trigger the time base.

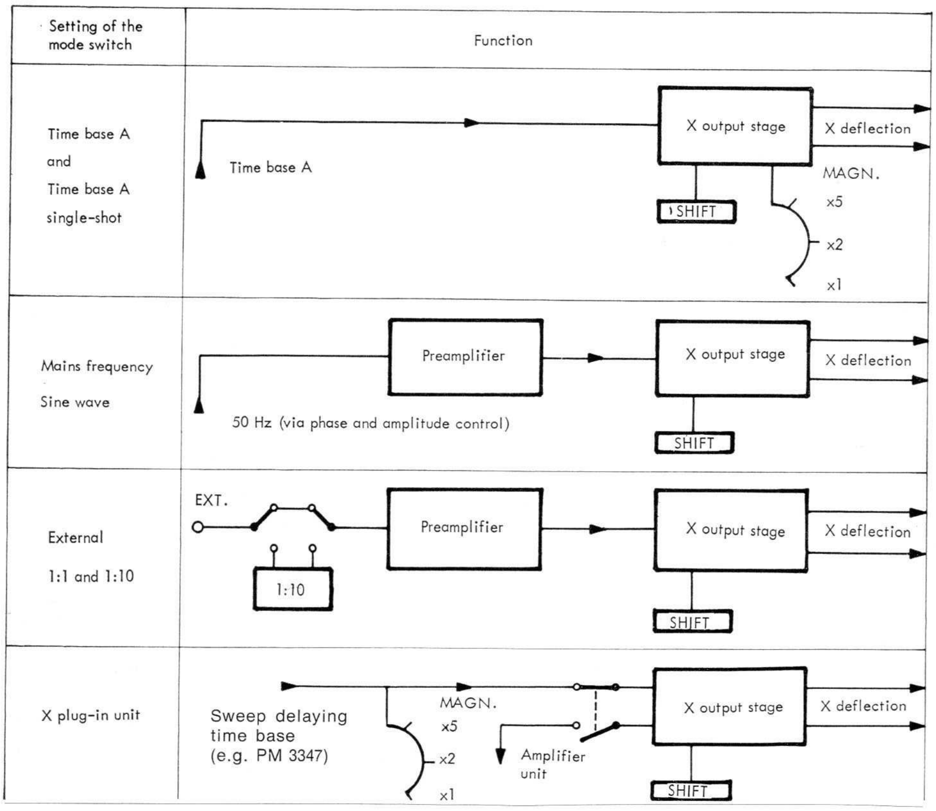


6. Simplified diagram of the trigger unit with tunnel diode

Time base generator A
Fig. 7 shows the block diagram of the time base generator A. Incoming negative trigger pulses from the trigger unit actuate the sweep gating multivibrator which supplies:
– A negative-going pulse which starts the Miller circuit
– A positive-going unblanking pulse to the cathode-ray tube grid to ensure that the forward moving trace is visible on the screen
– A positive gate pulse via a cathode follower, available at socket + GATE
The Miller 'run-up' circuit, driven by the sweep-gating multivibrator, supplies a linear positive-going sawtooth:

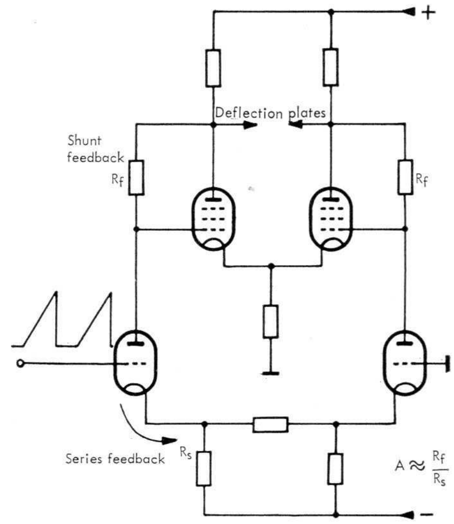


7. Simplified block diagram of time base generator A



8. Survey of functions of the horizontal amplifier circuit

– To the horizontal amplifier for spot deflection
– Via the cathode follower to socket TIME BASE for external use
– Via the SWEEP LENGTH control and hold-off circuit, back to the sweep-gating multivibrator input in order to reset the Miller circuit



9. Paraphase amplifier as used in the X output stage

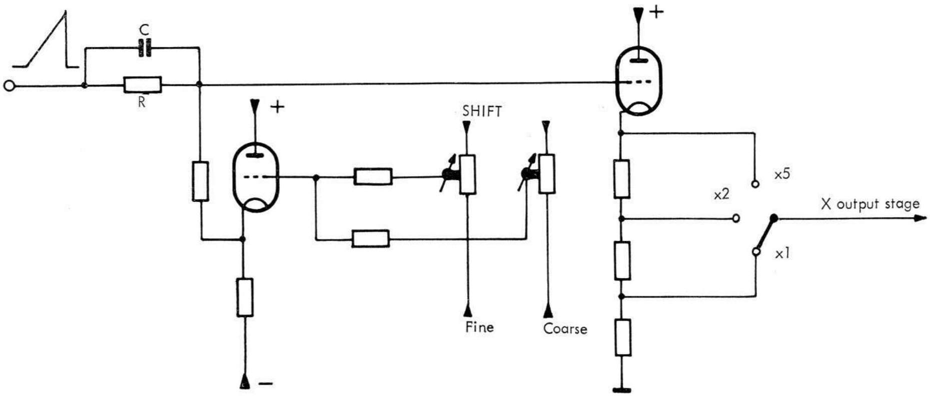
The sweep length control setting determines the instant at which the time base voltages starts to fall to its quiescent level. The hold-off circuit slows down the fall of the voltage from the sweep length control in order to allow time:
– For the time base generator output voltage to attain its quiescent value corresponding to a spot position on the left hand side of the screen
– For any transients in the system, caused by the fast retrace, to die away before the next trigger signal can actuate the sweep-gating multivibrator
The stability circuit makes it possible to set the DC potential at the input of the sweep-gating multivibrator so that stable triggering can be achieved. A "Preset" position for the stability is also provided. If the stability control is at "Repet" the DC potential is such that the time base is free-running. When the trigger mode selector switch is at "Automatic", the rectified trigger signal from trigger unit A is fed to the sweep-gating multivibrator input via the Auto stability circuit. Consequently the time base circuit is set into the normal triggered mode of operation as explained above. The lock-out multivibrator provides the stability circuit with single-sweep facilities:
– In the "free-running" mode; after a sweep of the time base the next sweep can be initiated by depressing the "Reset" button
– In the "triggered" mode; the sweep takes place on the arrival of the first trigger pulse after the "Reset" button has been depressed.

Horizontal amplifier
The horizontal amplifier comprises an X output stage, an expansion circuit and a preamplifier.
These circuits are interconnected by a mode switch having the positions:
Time base A
Time base A, single-shot
Deflection with mains frequency sinewave
External deflection 1 : 1 and 1 : 10
Deflection via an X-plug-in unit
A survey is given in fig. 8.

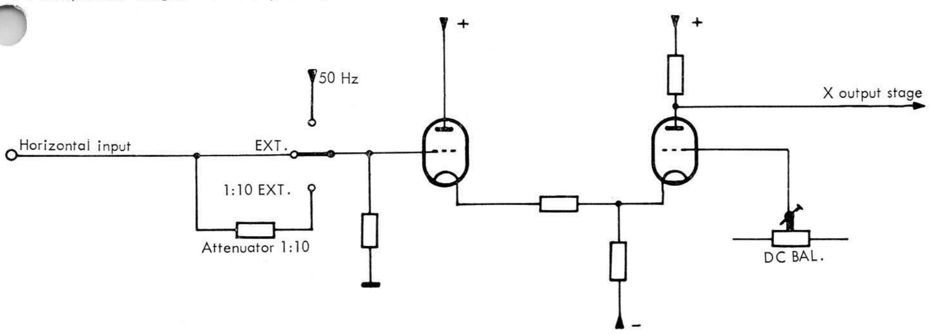
X output stage
In order to obtain stable time coefficients the gain of the X output stage must be very stable. Moreover, the output swing must be very high. Therefore a special output stage has been developed (see fig. 9). The whole circuit is a paraphase amplifier which is driven at one grid whilst the other grid is earthed, so that a symmetrical output waveform results from an asymmetrical input.
The high stability of the amplifier is obtained by using shunt feedback (R_f) on the two pentodes, as well as series feedback (R_s) on the triodes. In this way the gain of the amplifier is almost independent of valve characteristics. The overall gain A is made practically equal to the ratio of two stable resistances ($A \approx R_f/R_s$).

Expansion circuit
This circuit (see fig 10) consists of a cathode follower with a fixed attenuator having 3 positions, x1, x2, x5, in its cathode circuit. The attenuation is maximum at the x1 position, and minimum at the x5 position.
Horizontal positioning of the trace is effected by two front panel controls (shift, coarse and fine) which vary the DC level at the attenuator. Resistor R isolates the shift circuit from the preceeding stage; frequency compensation is carried out by capacitor C. In the position 50 Hz, EXT., and EXT. 1 : 10 of the mode switch, the magnifier switch is rendered inoperative; at the same time the range of the shift controls is reduced by a factor five.

Preamplifier
If it is desired to use an external signal for horizontal deflection, this voltage may be connected to the horizontal input

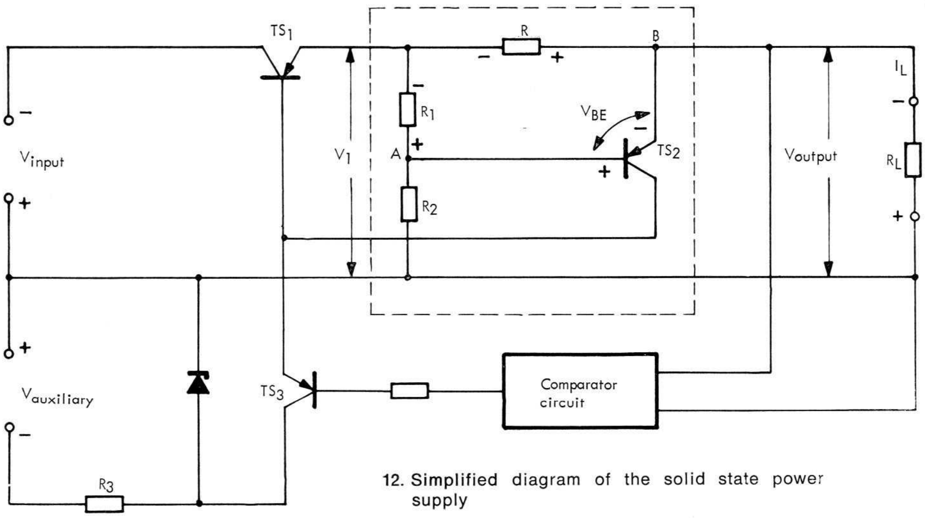


10. Expansion circuit (simplified)
11. Simplified diagram of the preamplifier



terminal. When the mode switch is at EXT. this signal is amplified by a pre-amplifier (fig. 11), and used to drive the X output stage. For large input signals a 1 : 10 attenuator can be interposed between the signal and the preamplifier.
Horizontal deflection with an external signal can also be effected by using an X amplifier plug-in unit. The signal is then applied directly to the X output stage in which case the shift controls and the expansion circuit of the basic oscilloscope are inoperative.

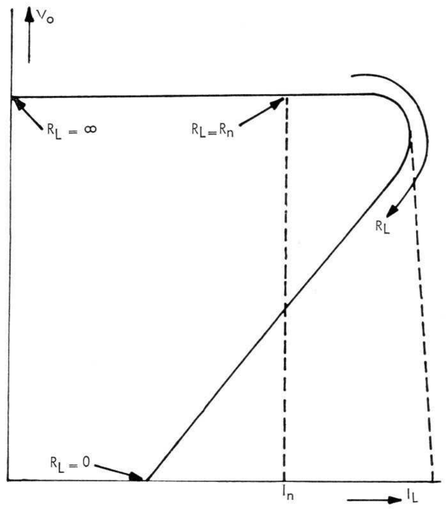
Power supply
The power supplies are constructed with solid-state elements; in the extra high tension supply (10 kV) vacuum-tubes are used. A thermal cut-out prevents the apparatus from overheating. It switches off the mains supply at a temperature of approximately 60°C. The transformer is protected by 2 fuses (slow blow).
The solid-state supply is protected against overload. Furthermore it is short circuit proof and fitted with an automatic reset. Fig. 12 shows the principle of the protective circuit.



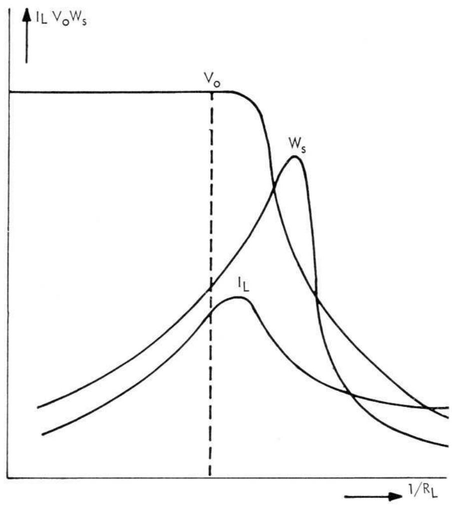
12. Simplified diagram of the solid state power supply

The overload protection circuit consists basically of the transistor TS_2 and resistors R, R_1 and R_2 as shown in the simplified diagram. Under normal working conditions load resistance $R_L = R_{nom.}$, point A is positive with respect to point B, and TS_2 is cut-off. Thus the voltage comparator circuit is able to exercise full control of the series regulator transistor TS_1 via emitter follower TS_3.
However, as R_L is reduced, the voltage drop across R will increase, while the voltage at A will remain relatively constant. Thus as R_L is reduced, the voltage at B will become less negative in relation to that at A, and at a predetermined value of R_L, TS_2 will start to conduct.
When TS_2 conducts, it draws a current I_2 from the emitter of TS_3, which will increase if R_L continues to decrease. As I_2 increases, a point will be reached when TS_3 bottoms and $I_1 + I_2$ is equal to V_{aux}/R_3. Thus any further increase of I_2 will cause a decrease of I_1. At this stage the voltage regulation will no longer operate, and TS_1 will find its own equilibrium as a quasi "constant current" source.
When TS_1 is acting as a quasi "constant current" source, the input to TS_2 must be considered with respect to the loop: TS_2 base, R_1, R and TS_2 emitter.
Thus when considering this loop a further decrease in R_L will cause a decrease in V_0 and V_1. Due to the quasi constant current character of the source the voltage across R will remain constant so that the decrease in voltage across R_1, due to the decrease in V_1, will cause A to become even more negative with respect to B. Thus TS_2 will draw even more current, reducing the current to TS_1 still further. The output of TS_1 will now fall to a new equilibrium point. Now, although the lower current from TS_1 will reduce the voltage across R, this reduction will not be sufficient to restore the voltage difference between A and B to its former value.
In this way further reductions in the value of R_L will cause both I_L and V_0 to fall as shown in fig. 13. The relation between V_0, I_L, the power dissipated in the series transistors W_s, and load conductance $1/R_L$ is shown in fig. 14.

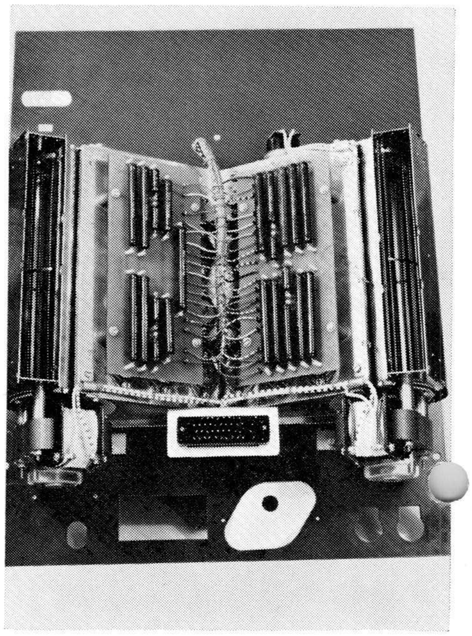
Extra high tension power supply
The extra high tension is obtained by transforming and rectifying the output voltage of a 40 kHz Hartley oscillator (see fig. 15). The high voltage power supply delivers three voltages:
– 8600 V positive with respect to earth for the post-deflection accelerator anode. The rectifying circuit comprises a voltage tripler
– A floating voltage of 1800 V, which is



13. Output voltage as a function of the output current with parameter R_L



14. V_o, I_L and W_s as a function of $1/R_L$



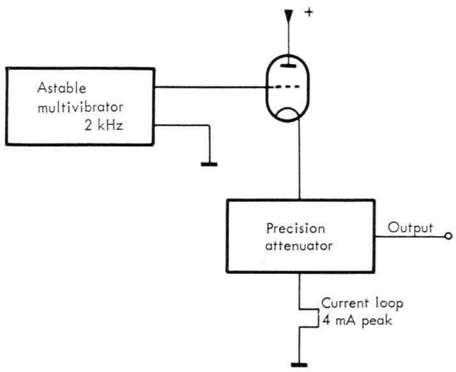
Cooling unit, with power transistors, can be taken out and interconnected with the apparatus by means of a connector block. The transistors on this cooling unit are then interchangeable. Special holder for the transistors renders soldering unnecessary

superimposed on the unblanking pulse from the time base generator to provide DC coupling to the grid of the CRT

– The supply voltage for the cathode of the CRT (−1440 V with respect to earth) A part of this voltage is compared with the stable − 150 V rail. The error signal is amplified and controls the screen grid voltage of the Hartley oscillator tube, thereby controlling its output.

Calibration unit

The calibration unit (see fig. 16) generates a square wave signal with a repetition frequency of about 2 kHz and an accurately known amplitude. This amplitude can be selected in steps between 80 V and 200 μV.

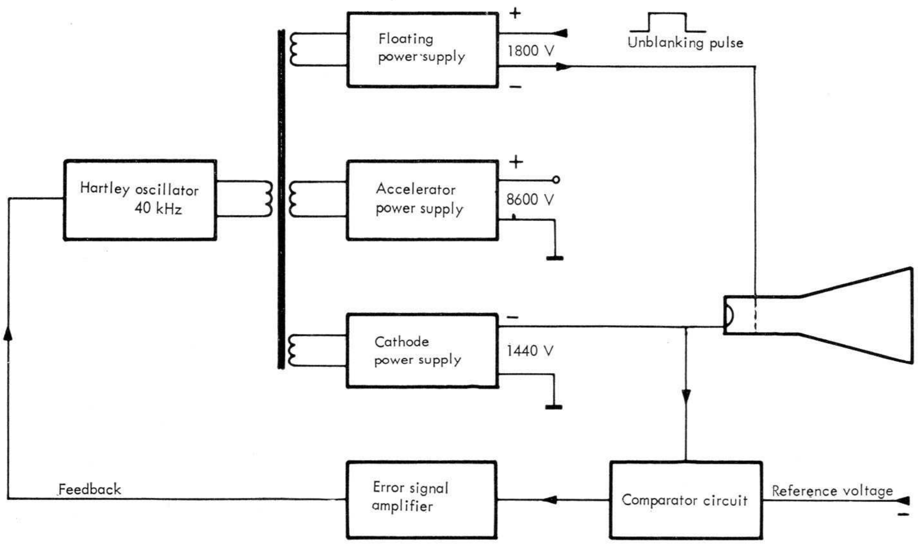


16. Simplified diagram of the calibration unit

An astable multivibrator switches the output voltage between an adjustable voltage of about 80 V and a negative voltage of about −50 V. When the multivibrator output is low, the cathode follower is entirely cut-off. The cathode voltage is then at earth potential.

When the multivibrator output is high, the grid voltage of the cathode follower is determined by the setting of a potentiometer with which the cathode voltage is adjusted to precisely 80 V.

A precision attenuator divides this voltage into steps of 80-40-20-8-4-2-0.8-0.4-0.2 V. If the output switch is in position "V", the amplitude of the output signal is expressed in V. If the output switch is in position "mV", a 1 : 1000 attenuator converts the range to millivolts. In the mV position the output impedance is fixed at 50 Ω. If the output switch is in position "4 mA", no signal is available at the output terminal, but a square wave current with an accurate peak value of 4 mA flows through a loop on the front panel. The fourth position switches the unit off. The Y amplifiers can always be calibrated in all positions of the step attenuator with a 4 cm deflection.



15. Simplified diagram of the EHT supply

Technical data

CATHODE-RAY TUBE
Diameter 13 cm (5 in)
Acceleration voltage 10 kV
Graticule 6 x 10 cm, internally illuminated
Type D13 — 16GH/01
Available phosphors GH, BE, GP

VERTICAL AMPLIFIER
Coupling DC **Signal delay** 150 ns
Rise time 5 ns **Visible delay** 60 ns

CALIBRATION VOLTAGE (CURRENT)
Amplitude 0.2 mV to 80 V; 2, 4, 8 series (4 mA_{p-p})
Tolerance 1%
Frequency 2 kHz square wave

HORIZONTAL INPUT
Bandwidth DC to 1 MHz
Sensitivity 500 mV/cm and 5 V/cm

HORIZONTAL DEFLECTION
Deflection selector Time base A, single-shot, mains supply (adjustable phase and amplitude), external 1 : 1 and 1 : 10 and X unit

TIME BASE GENERATOR
Mode Free-running or triggered
Sweep speed
Calibrated: 50 ns/cm to 1 s/cm
(1, 2, 5 series) max. 10 ns using 5x expansion
Uncalibrated: continuous adjustment between steps
Tolerance 3% (in middle 8 cm)
Expansion 2x and 5x
Expansion tolerance 2%
Time base outputs Sawtooth 90 V_{p-p}, gate 35 V_{p-p} both pos. from zero level

TIME BASE TRIGGERING
Source
Internal (from channel A or B, or A + B when the added mode is used with the dual trace unit)
External, mains supply (with adjustable phase)
Slope Positive or negative
Mode
1. TV frame 4. DC
2. TV line 5. LF over 3 Hz
3. Automatic 10 Hz to 1 MHz 6. HF over 2 kHz
In modes 1, 2 and 3 manual level and stability are inoperative
In modes 5 and 6 high pass filter with RC time 56 ms, 80 μs respectively
Sensitivity
Internal: 3 mm up to 10 MHz, 1 cm up to 30 MHz, 1 cm in aut. mode 10 Hz to 1 MHz, 2 cm p—p video for TV (depending on the unit used)
External: 0.4 V up to 10 MHz, 1 V up to 100 MHz, 1 V positive video for TV position
Input impedance: 1 MΩ shunted by 55 pF (used with DC, LF, HF or Aut.)
Level
Internal: adjustable over 6 cm
External: adjustable over 6 V_{p-p} for midband frequencies

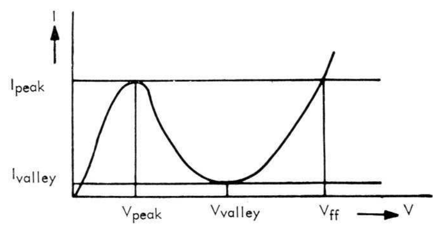
Z MODULATION
Sensitivity 15 V_{p-p} above 300 Hz
Input impedance 25 kΩ shunted by 35 pF
Source External, internal (from multi trace unit chopper) selected by a switch on rear panel

MAINS SUPPLY
Voltages 110, 125, 145, 200, 220 and 245 V
Frequency 40 to 60 Hz (below 50 Hz the voltage must not exceed the nominal value)
Voltage variations 10% (variation can be tolerated with negligible effect)
Max. power consumption 600 VA

DIMENSIONS
H x W x D 46 x 34.5 x 68 cm (18 x 13.5 x 26.5 in)
Weight 42 kg (92 lbs) main frame only

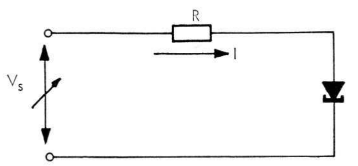
Appendix

Use of a tunnel diode in a trigger circuit

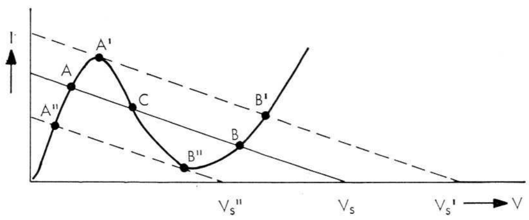


17. Tunnel diode characteristic

increase of the source voltage will cause a very fast voltage transition when the operating point jumps from A′ to B′ (in approx. 1 nsec.). For the operating point to return to A, the source voltage must be decreased to below V_s'' and then restorted to its initial value V_s. The return will be attended by a second very fast transition, corresponding to the jump B″ to A″.
Thus, if an AC voltage with an amplitude exceeding $V_s' - V_s$ and $V_s - V_s''$, respectively is superimposed on the DC bias V_s, the voltage across the tunnel diode will be an alternating voltage with very steep edges. The step voltage can be increased by the provision of an AC impedance formed by an inductance (see fig. 20).

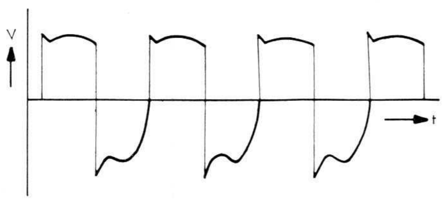


18. Simple tunnel diode circuit



19. Characteristic of the tunnel diode with different load lines

The characteristic of the tunnel diode is shown in fig. 17.
Typical values for V_{peak} and V_{valley} are 60 mV and 350 mV respectivaly at I_{peak} = 10 mA and I_{valley} = 1.3 mA. Another important point is V_{ff} (500 mV) at which the current is again I_{peak}.
Consider a voltage source V_s connected to the tunnel diode via a resistor R (fig. 18). V_s and R can be selected so that the load line crosses the characteristic at 3 points viz. A, C. B (fig. 19), of which only A and B are stable operating points.
Starting from operating point A, with a voltage increase from V_s to V_s', the load line will shift to a new position parallel to the first, until it cuts the characteristic curve at the points A′ and B′. Any further

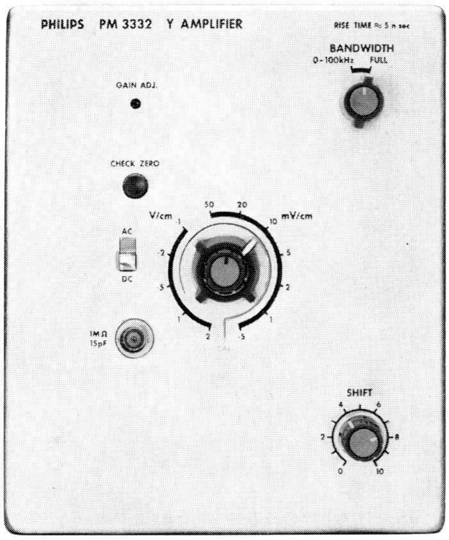


20. The voltage across a tunnel diode with sine-wave input

The L/R time constant must be high compared with the transition time but low compared with the AC period. Then the diode switches from about V_{peak} to V_{ff}.
After amplification and differentiation of this signal the negative going spikes are selected for triggering the time base circuit.

HIGH SENSITIVITY LOW-DRIFT WIDE BAND AMPLIFIER PM 3332

by C. Gijzen



The PM 3332 is a new concept in oscilloscope preamplifiers. It incorporates two innovations which are unique in their use in such amplifiers. These are a chopper feedback amplifier which provides servo stabilization of the DC drift and a split-band amplifier which affords high gain, from DC through to high frequencies, with an extremely good signal to noise ratio. This concept has resulted in a DC plug-in amplifier unit with an overall rise time of 7 ns, 500 μV/cm sensitivity for the full bandwidth of DC to 50 MHz, drift less than 1 cm/week and low noise.

Chopper stabilization

As in most high gain amplifiers drift in the DC amplifier of the PM 3332 is corrected by applying a correction voltage to the first stage of the amplifier. In this case the second input of the differential amplifier is used for the correction voltage as only one input is used for the signal. However, while most high gain amplifiers are fitted with a manual control for this purpose, in the PM 3332 a feedback loop is employed. Thus in fig. 1 input I is used for the signal voltage (V_i) and input II for the drift correction voltage V_c. The total drift voltage occuring in the amplifier can be regarded as a single voltage at the input; V_d in fig. 1. The amplification factors of the DC and chopper (feedback) amplifiers are A and B respectively; both amplifiers giving 180° phase shift. The output voltage of the output attenuator is adjusted to be V_o/A when V_o is output of the DC amplifiers. Thus the inputs to the voltage divider R_1 and R_2 are V_i and V_o/A and as R_1 and R_2 are equal, the input to the feedback amplifier is:
$\frac{1}{2} V_i + \frac{1}{2} V_o/A$ therefore the output of the feedback amplifier (V_c) is given by:

$$V_c = - B\tfrac{1}{2} (V_i + V_o/A)$$

Hence the total input voltage to the DC amplifier V_t will be:

$$V_t = V_i + V_d - (-\tfrac{1}{2}B (V_i + V_o/A))$$

thus when V_t is amplified by − A the output of the DC amplifier (V_o) will be:

$$V_o = - A (V_i + V_d - (-\tfrac{1}{2}B (V_i + V_o/A)))$$

or

$$V_o = -A\, V_i - \frac{A}{1 + \tfrac{1}{2}B} V_d$$

Thus it can be seen that while the input signal V_i benefits from the full amplification A of the DC amplifier the drift voltage is only amplified $A/(1+\tfrac{1}{2}B)$ where B is in the order of several thousands. In this way the drift is greatly reduced. It should be noted that this method of chopper stabilization has no effect on the input impedance as the junction of R_1 and R_2 is a virtual earth point in relation to the input signal.

Split-band amplifier

There are several advantages in splitting the very sensitivity wide band preamplifier into a DC + LF part and a HF part. These advantages are as follows:

LOWER NOISE

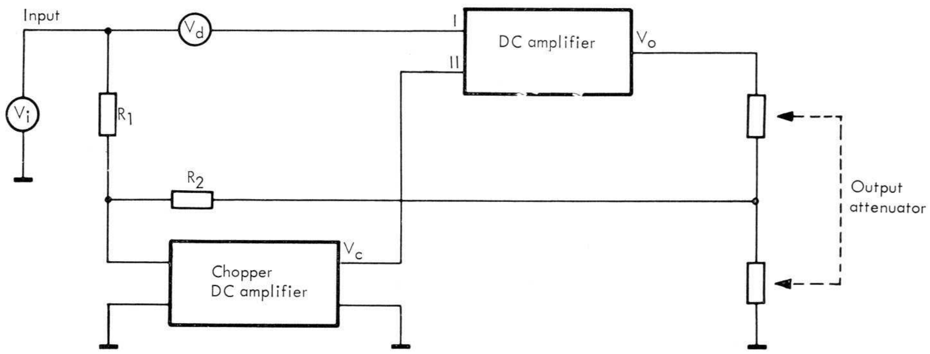
Transistors which have excellent HF properties suffer from a very serious 1/f noise, starting at say 100 kHz, downwards (fig. 2). Although the lowest noise components will be suppressed by the chopper stabilization, discussed in the last paragraph, some noise, due to the limited response speed of the stabilization, will remain. Low frequency transistors, on the other hand, suffer from an increase of noise at the other end of the frequency band. Furthermore these transistors will not reach the rise time requirements. In the split-band amplifier the best properties of HF and LF transistors are used, each in its own frequency range.

BETTER DC BIASING

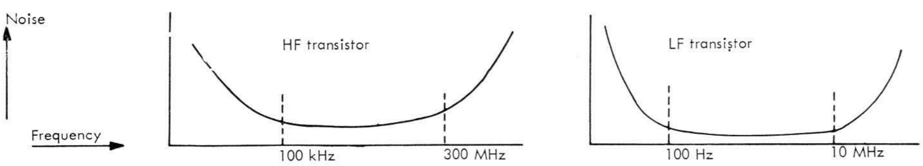
To make use of the excellent low noise properties, that transistor specifications promise, it is necessary to bias an HF transistor to a much higher I_c ($\approx$10 mA) than an LF transistor. These contradictory requirements can only be fulfilled in a split-band amplifier.

NO CONSTANT-DISSIPATION NETWORKS NECESSARY

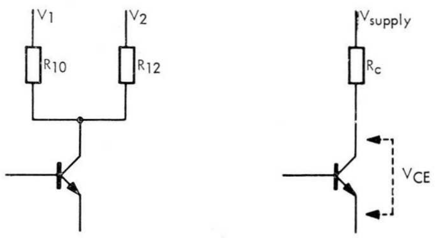
In a wide band amplifier (up to 100 MHz) a rather high I_c and a rather small R_c (few hundred ohms) are generally used. This means that the main part of the supply voltage is across the transistor, and is independent of the instantaneous value of the collector current. Transients of the collector current, with a low repetition rate, give rise to an undesirable fluctuation of collector dissipation; undesirable because these fluctuations cause changes in junction temperature which, in turn vary most of the transistor parameters. The common method of avoiding these variations in collector dissipation, is to insert a by-passed resistance between the collector and the collector load. The value of this resistance must be such, that the voltage across the transistor is half of the supply voltage.



1. Simplified diagram of the chopper stabilization

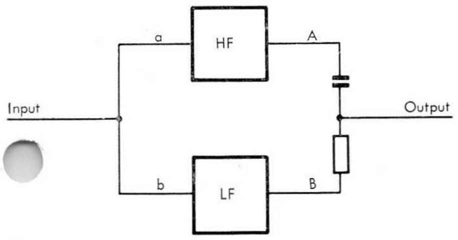


2. Noise frequency curves for HF and LF transistor



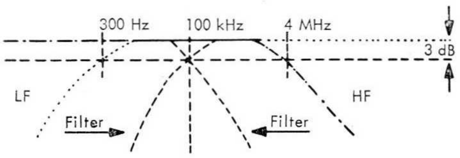
3a. Different supply voltage

3b. Supply voltage is twice the V_{CE}



4. Coupling filter

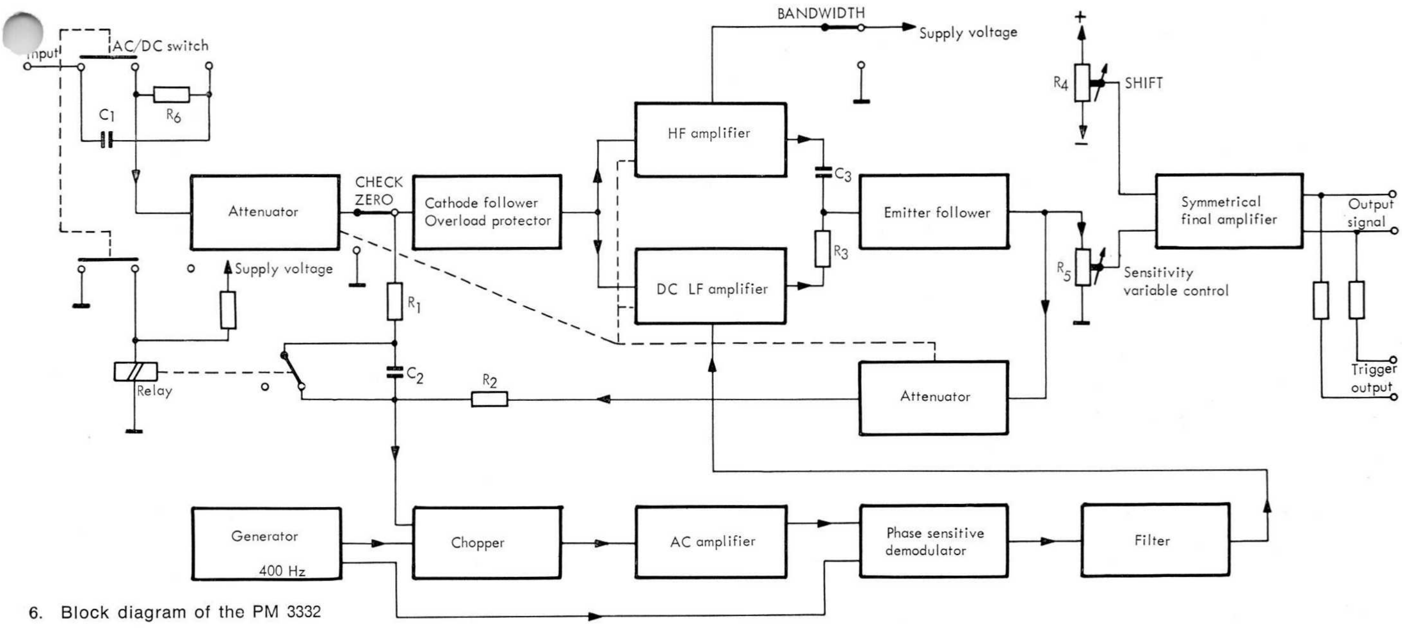
5. Frequency characteristic of the amplifiers and the filter



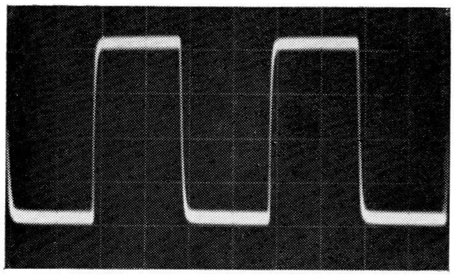
Variations of I_c will now cause proportional variations of V_c (but of opposite sign so that the I_cV_c product i.e. P_c will be nearly constant).
As the HF part of the split-band amplifier does not contribute to the overall characteristic at frequencies below 100 kHz, (−3 dB) this insertion is not required because the time-constant of the thermal response to the collector dissipation is in the order of milliseconds or longer. In the low frequency part of the split-band amplifier, each collector resistance has been chosen such, that half of the supply voltage is across the transistor. In each stage the collector resistor is built up of two resistors in parallel, each fed from a different supply voltage (fig. 3a). The ratio of R_{10} and R_{12} (fig. 3a) can be chosen so that the supply voltage of fig 3b (which must be somewhere between V_1 and V_2 of 3a) is just twice the V_{CE}, whilst R_{10} and R_{12} in parallel will give the desired R_c.

BETTER DECOUPLING OF THE SUPPLY VOLTAGE
As the HF part can be RC coupled, all DC levels can be freely chosen. This means that only one collector supply voltage, which can be adequately decoupled in each stage without causing any deterioration of the low frequency response is all that is necessary. All emitter resistors are grounded in order to suppress interstage coupling. The HF amplifier (from 300 Hz to 100 MHz) and the LF amplifier (from DC to 4 MHz) have a large overlap. This has made the use of a simple RC filter (fig. 4) possible (cross over frequency 100 kHz). Because of this wide overlap the frequency characteristic near the cross over-point is governed only by the filter, whilst the frequency characteristic of each amplifier begins to fall where its contribution to the combination is negligible (fig. 5).

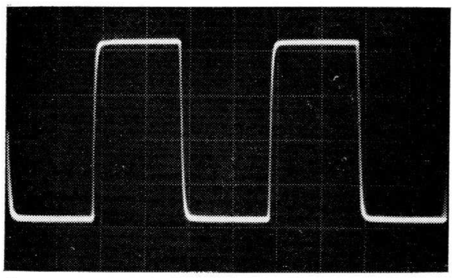
General description of the block diagram
Starting at the input, the AC/DC switch is encountered first. In the AC position it is possible to observe small AC signal "riding" on a max. 400 Volts DC level (fig. 6). R_6 is included to discharge C_1 when the AC/DC switch is changed from AC to DC whilst C_1, is charged. Without this device the charge on C_1 might destroy a delicate test object, even if the unit had been out of use for a long time. Although a quality capacitor was chosen for C_1, the leakage resistance is such, that 400 V DC in the AC position would cause several cms deflection in the 0.5 mV/cm position. However by inserting a capacitor (C_2) in series with R_1 the DC output will be kept at zero by the chopper stabilization circuit, thus eliminating the effect of the leakage in C_1.
From the AC/DC switch the frequency compensated high Z attenuator is encountered next. In the 4 most sensitive positions this attenuator acts only as a feed-through, thus giving no attenuation. Via a cathode follower and overload-protector the signal passes to the split-band amplifier. In order to increase the signal to noise ratio in the less sensitive ranges the amplifier gain is reduced from 0.5 mV/cm to 5 mV/cm by the first 4 steps of the attenuator. This switch is mechanically linked to the input attenuator (viz in the positions 1 to 4). The supply voltage of the HF part of the split-band amplifier can be switched off by the bandwidth switch. This makes operation with a limited (100 kHz = −3 dB) bandwidth and an almost negligible noise, even at full sensitivity possible. After



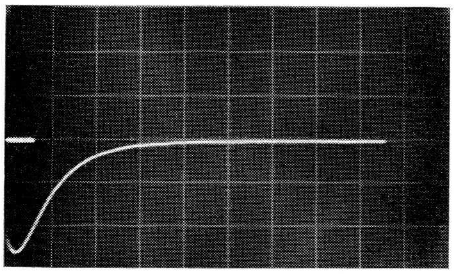
6. Block diagram of the PM 3332



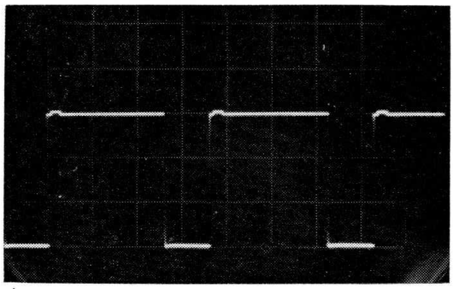
a



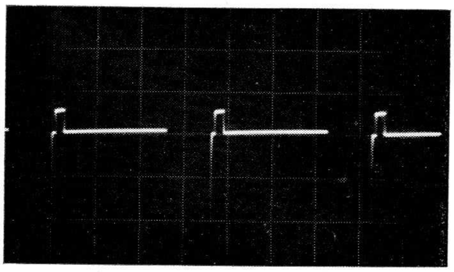
b



c



d



e

the filter, C_3R_3 already discussed, an emitter follower stage is encountered. This protects the filter from being loaded by the low impedance sensitivity control potentiometer. The input voltage of this potentiometer is also fed to the output attenuator, having an attenuation proportional to the overall amplification from the cathode follower input to the emitter follower output. This attenuator is also mechanically linked with the input attenuator. From the output attenuator the voltage which has to be compared with the input voltage, via R_1 and R_2, is obtained. The voltage at the junction of R_1 and R_2 is fed into the chopper which is driven by a 400 Hz generator. The amplified chopper signal is demodulated in a phase sensitive demodulator. The output of which is the correction voltage V_c already discussed. The filter between the demodulator and the DC amplifier eliminates spurious 400 Hz signals in the signal amplifier and includes the phase-correction for the feedback loop which is necessary to prevent oscillations. The symmetrical final amplifier follows R_5 with R_4 supplying the Y shift voltage. This had to be kept outside the stabilization loop, to prevent it being seen by the chopper amplifier, as a drift voltage. The final amplifier has two complimentary NPN/PNP emitter followers having a very low output impedance for both positive and negative transients, affording adequate matching to the basic oscilloscope's Y input and trigger input.

The oscillograms (a, b, c, d, e) illustrate some of the features of the PM 3332 high sensitivity wide band amplifier.
Figures a and b show the degree of noise at 50 MHz bandwidth and at reduced bandwidth respectively.
The sensitivity was set at 500 uV/cm; the signal displayed is 1000 Hz.
Figure c shows the recovery after a short interruption of the line voltage, demonstrating the DC balance stability.
Initially the trace was centered on the screen (as visible at extreme left), then the power switch of the PM 3330 was switched momentanily off and on. The sensitivity during this test was set at 500 uV/cm; the sweep speed 1 cm/sec.
Figures d and e show identical composite signals, obtained from a Philips PM 5720/40 modular pulse system. The sweep speed in both figures is 1 cm/0.1 usec.
However, figure e is a 10-times vertical expansion of that of fig. d.

Technical data

All data for this unit are valid when used in combination with the main frame PM 3330

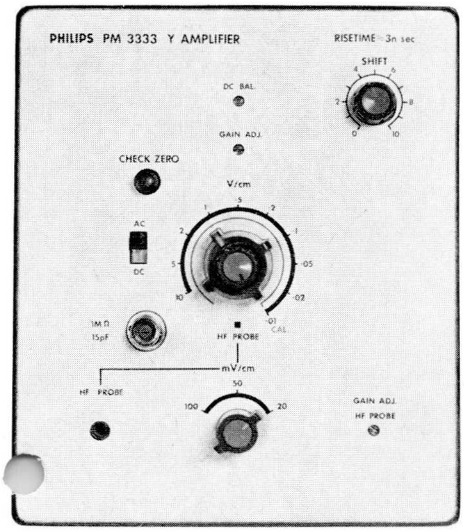
AMPLIFIER
Bandwidth
DC coupled: DC to 50 MHz wide band or
DC to 100 kHz narrow band
AC coupled: 1.6 Hz to 50 MHz wide band or
1.6 Hz to 100 kHz narrow band
Rise time 7 ns
Sensitivity 500 μV/cm
Drift $<$ 1 cm/week, irrespective of switching on and off
Attenuator Calibrated 500 μV/cm to 2 V/cm (1, 2, 5 series) continuous uncalibrated gain between steps
Tolerance 3%
Overshoot $<$ 2% at max. sensitivity
Droop 3% at max. sensitivity
Input impedance 1MΩ shunted by 15 pF (at DC)
Max. input voltage AC coupled 400 V DC
Max. deflection 18 cm mid-band 15 MHz (6 cm window)
Shift range 18 cm

DIMENSIONS
H x W x D 17.5 x 15 x 27.5 cm (7x6x11 in)
Weight 2 kg (4½ lbs)

ACCESSORIES
PM 9331 A/10 PASSIVE PROBE
Coupling DC
Attenuation 10 : 1
Tolerance 3%
Input impedance 10 MΩ shunted by 8 pF
Max. input voltage
DC coupled 1000 V, AC coupled 400 V DC

WIDEBAND VERTICAL AMPLIFIER PM 3333

by A. H. J. Sloots



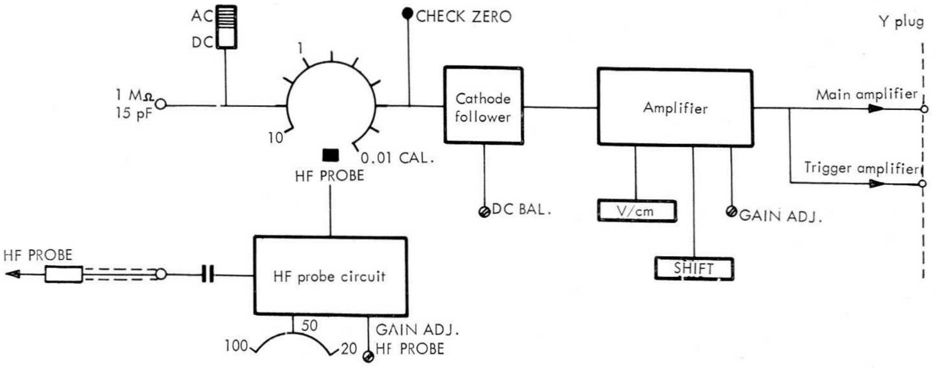
The PM 3333 is a fast rise time vertical amplifier which makes full use of the main frame bandwidth, giving an overall rise time of 6 ns at the full sensitivity of 10 mV/cm. A special high-sensitivity HF PROBE system is used which affords an input impedance of 100 kΩ shunted by 5 pF. The sensitivity at the probe tip is 20 mV/cm. In this way the need for the normally bulky cathode follower probe is eliminated. In addition, a conventional high-impedance probe can be supplied for normal work. Fig. 1 shows the schematic diagram of the unit.

The PM 3333 is provided with two input sockets. The normal input (1 MΩ shunted by 15 pF) is connected via the AC/DC selector to the step attenuator V/cm. The HF probe input has a special circuit, which contains its own step attenuator, calibrated [illegible] mV/cm.

The signal is applied via a cathode follower stage to a transistorized amplifier. This amplifier contains a variable gain control, a SHIFT control and a GAIN-ADJ. control. The output voltage of the amplfier is passed to the vertical amplifier and to the trigger amplifier of the basic oscilloscope via the output plug. A few specific points will now be treated in a little more detail.

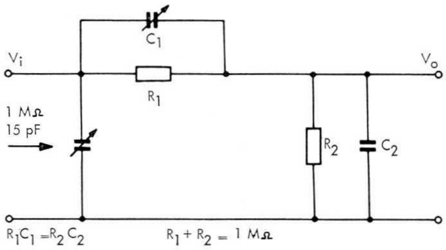
Attenuator

The high ohmic turret attenuator (1 MΩ shunted by 15 pF) consists of 10 RC networks (fig. 2). These networks are switched one after the other and cover an attenuation range from x1 to x1000 in steps of 1–2–5. For high frequencies the attenuation is determined mainly by C_1 and C_2. If the input voltage is attenuated greatly, the value of C_2 is large. The reactance of C_2 is then small, so that the undesired impedances, namely the parasitic self-in-



1. Schematic diagram of the PM 3333

2. RC network



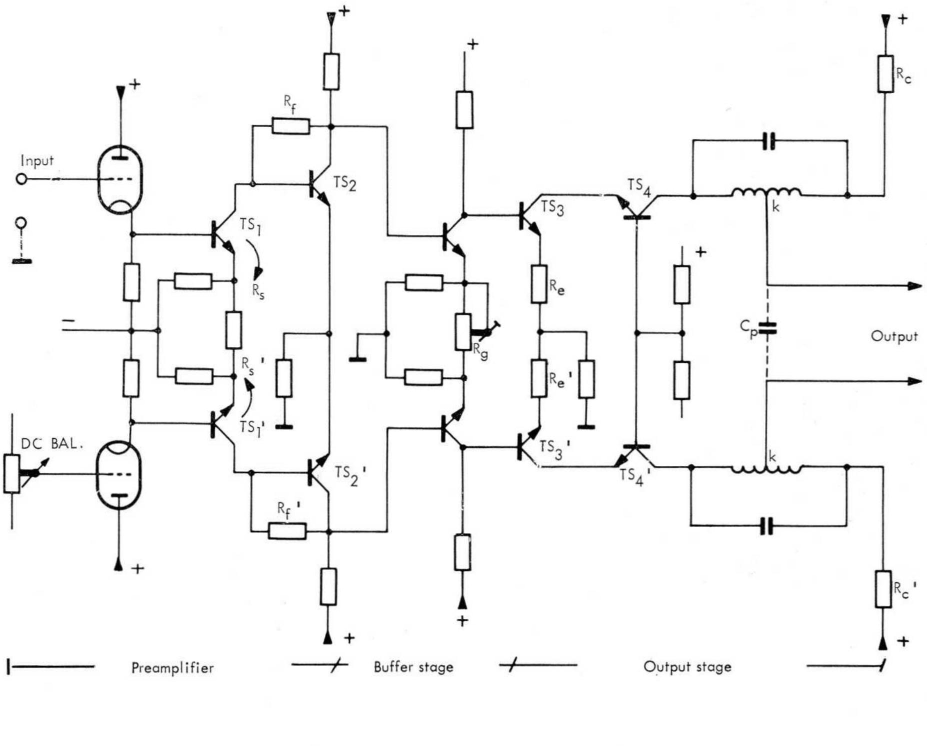
ductance of the capacitors and the earth connections, begin to play a role. In this way the attenuation becomes dependent on the frequency which, in pulse reproduction leads to distortion of the flanks. A great deal of care has therefore been devoted to the attenuator used in the PM 3333, with respect to good high-frequency earthing, and special HF disc capacitors of very low self-inductance have been incorporated.

Transistorized amplifier

This consists of three parts, viz. a preamplifier stage, a buffer stage, and an output stage. A simplified diagram of this unit is shown in fig. 3.

The signal is passed via a cathode follower to the preamplifier, in which both series and shunt feedback are used. The amplification A ($\approx R_f/R_s$) is stable, being dependent mainly on two resistors. The shunt feedback gives TS_2, the associated stage, a very low input resistance, thus creating a virtual earth point and a low output resistance. Because of the virtual earth, the collector AC voltage of TS_1 is very small, so that the feedback through the collector base capacitance of the first transistor plays a negligible role.

3. Simplified diagram of the amplifier



Because of this and the low output resistance of TS_2 the preamplifier stage has a wide bandwidth. For a detailed treatment of this circuit, reference is made to the articles written by Dr. Cherry and Mr. Hooper, of the Department of Electrical Engineering, University of Melbourne.*

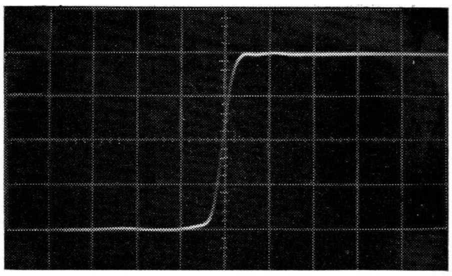
To keep the capacitive load at the output of the preamplifier small, a buffer stage is interposed between the preamplifier and output stages. The amplification i.e. the deflection coefficient of the unit, can be adjusted by means of R_g (gain adjustment).

A cascode amplifier has been chosen for the output stage. The amplification here is also dependent mainly on two resistors, viz, $A \approx R_c/R_e$.

Because of the low input resistance of the second transistor (grounded base configuration of TS_4) the feedback here is also negligible. The output signal of the output amplifier is passed via a bridged T m-derived filter to the Y plug-in connector. The rise time of the unit is about 3 ns ($f_{3dB} \approx$ 120 MHz). The overall rise time, including that of the basic oscilloscope, amounts to about 6 ns (f_{3dB} = 60 MHz). Fig. 4 shows the overall step response.

*1. The design of wide-band transistor feedback amplifiers, Proceedings I.E.E., Vol. 110, no. 2, February 1963.

2. An engineering approach to the design of transistor feedback amplifiers, J. Brit. Inst. Radio Engr., 1963.

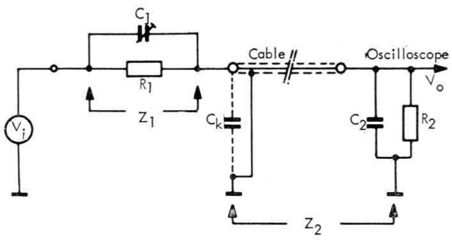


4. Overall step response of the amplifier; rise time of the input signal amounts to about 1ns, time scale 10 ns/cm

PROBES

Passive attenuator probe PM 9331A/10

Fig. 5 shows the principle of the probe.



5. Principle of the passive probe

This probe circuit is based on the principle of voltage transmission:

$$V_o = \frac{Z_2}{Z_1 + Z_2} V_i$$

The output voltage V_o is independent of the frequency, provided that

$R_1C_1 = R_2 (C_2 + C_k)$

Since the cable is not terminated, it can simply be represented by a capacitance C_k. With R_2 = 1 MΩ and C_2 = 15 pF (normal values for input resistance and capacitance of the oscilloscope respectively), C_k = 30 pF (cable length of about 1 m) and R_1 = 9 MΩ, the value of C_1 will be about 5 pF.

The input capacitance of the probe, which is determined by C_1 and the parasitic capacitances at the tip of the probe, now amount to 8 pF. The output voltage V_o will with the given values, be equal to 1/10 V_i.

Since the cable is not terminated with its characteristic impedance, reflection will occur at high frequencies. To damp these reflections, a resistance wire is used for the inner core of the probe cable.

To obtain good pulse transmission it is necessary to attach a compensation network on the oscilloscope end of the cable.

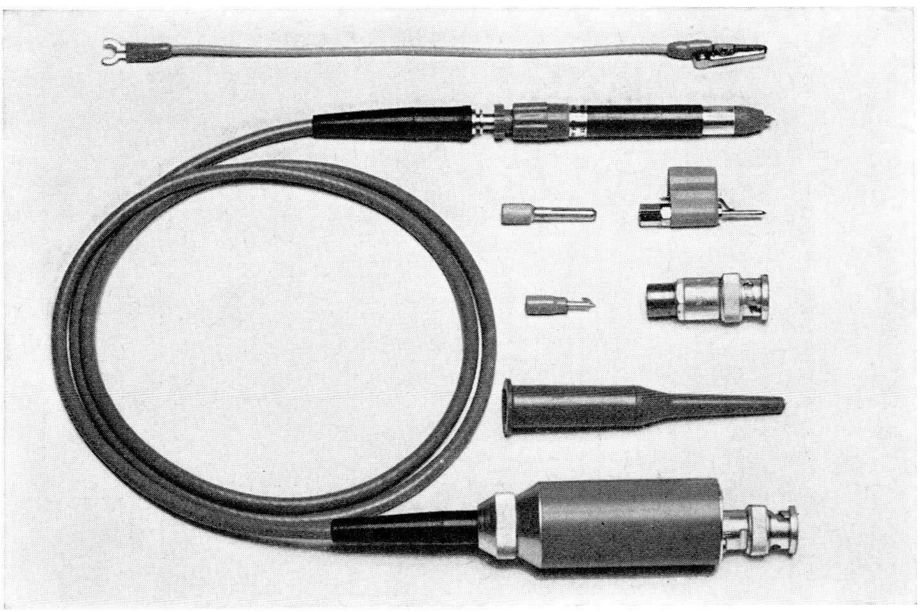
The most significant disadvantages of this system can be summed up as follows:

– Relatively high attenuation
– Relatively high input capacitance
– Difficult frequency compensation for the high frequencies, giving rise to, among other things, a limitation of the length of the probe cable
– Special cable with resistance wire

The advantages of this system can be summarized as follows:

– A very high DC resistance, viz. 10 MΩ
– DC voltage coupling
– A large voltage range in combination with the atenuator in the apparatus

The probe PM 9331A/10, with accessories, is shown in fig. 6.



6. PM 9331/10 Passive probe

HF probe PM 9332

The probe circuit used here is based on the principle of current transmission. It is shown schematically in fig. 7. The measuring cable is terminated with the characteristic impedance Z_o. The current is given by

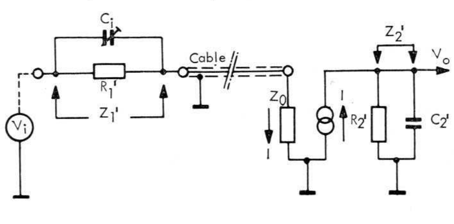
$$I = \frac{V_i}{Z_i + Z_o} \approx \frac{V_i}{Z_i'} \text{ and the voltage by}$$

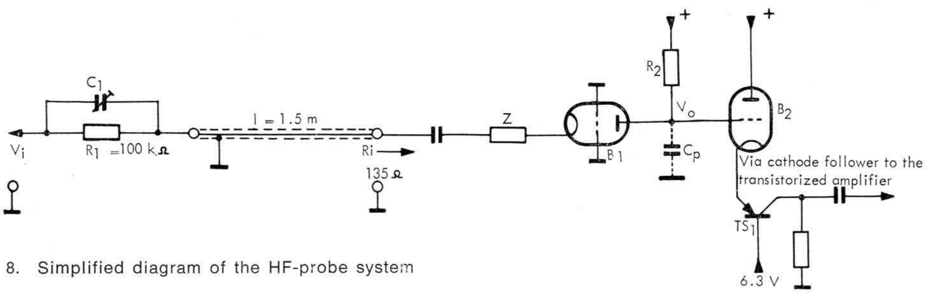
$$V_o = IZ'_2 \approx V_i \frac{Z_2'}{Z_1'} \qquad 1$$

This voltage is independent of the frequency if

$$R_1'C_1' = R_2'C_2' \qquad 2$$

7. Principle of the HF-probe





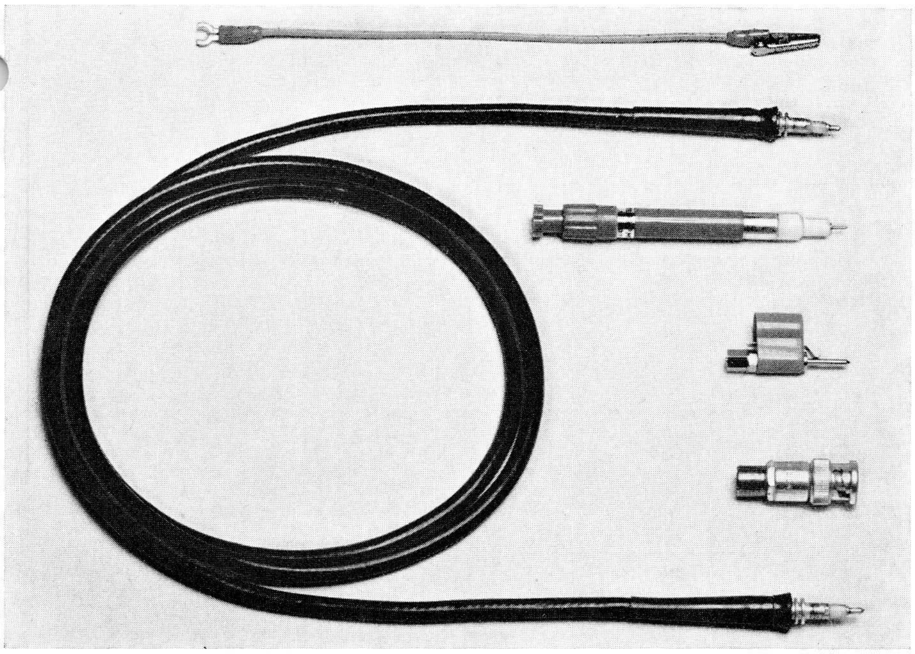
8. Simplified diagram of the HF-probe system

With $R_1' = 100$ kΩ, $R_2' = 33$ kΩ, and $C_2' = 9$ pF, the value of C_1' will be about 3 pF. The input capacitance including the parasitic capacitance measured at the tip of the probe, will now be about 5 pF.

The output voltage V_o will in this case be about $^1/_3$ V_i. Since the cable is terminated with its characteristic impedance, reflections will not occur. Because of this and of the fact that the capacitance of the cable is not significant, as reference to equation 2 will show, a normal HF cable (Z_o = 135 Ω) can be used for connecting the probe. The length of the cable can even be as long as 1.5 m, so that measuring can take place at a considerable distance from the oscilloscope. A simplified diagram of this probe system is shown in fig. 8.

The probe PM 9332 (R_1 and C_1) is connected via a cable with a length of 1.5 m to a triode amplifier tube B_1 in grounded grid configuration taking care of the current transmission. This configuration gives a current amplification of x1 and has a very low input impedance. The cable is terminated by Z and the input impedance of B_1. The output voltage V_o is determined by the quotient of the resistors R_2 (33 kΩ) and R_1 (100 kΩ); the input voltage V_i is thus attenuated threefold. The probe system will be frequency compensated, provided that $R_1 C_1 = R_2 C_p$. Correct adjustment is obtained with the trimmer C_1; the capacitance C_p being a parasitic capacitance.
The output voltage of B_1 is passed to the transistorized amplifier by a circuit containing tube B_2 and transistor TS_1 (fig. 8). The circuit gives small voltage amplification, about x 1.5. The signal passing to the probe is thus attenuated to only half its original value before it reaches the preamplifier. Since the deflection coefficient, measured at the preamplifier, amounts to 10 mV/cm, a voltage of only 20 m $V_{p=p}$, at the tip of the probe, is then sufficient for a vertical deflection of 1 cm. The probe PM 9332, with accessories, is shown in fig. 9.

9. PM 9332 HF-probe



Technical data

All data of this unit are valid when used in combination with the main frame PM 3330

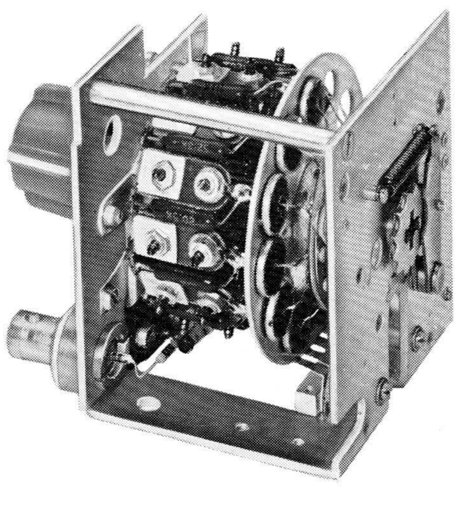
AMPLIFIER
Bandwidth
DC coupled: DC to 60 MHz
AC coupled: 1.6 Hz to 60 MHz
Rise time < 6 ns
Sensitivity 10 mV/cm
Attenuator
Normal input (input I): Calibrated 10 mV/cm to 10 V/cm (1, 2, 5 series), continuous, uncalibrated gain between steps
HF probe (input II): 20, 50 and 100 mV (referred to probe tip), continuous, uncalibrated gain between steps
Tolerance 3 %
Overshoot < 2 %
Input impedance (input I)
1 MΩ shunted by 15 pF
Maximum input voltage
(AC coupled) 400 V DC (Input II can only be used with the HF probe)
Maximum deflection inputs I and II
6 cm over the full bandwidth 18 cm mid-band (20 MHz, window 6 cm)
Shifting range > 12 cm

HF PROBE
(for use with input II only)
Coupling AC
Bandwidth 30 Hz to 50 MHz
Attenuation 1 : 1
Maximum input voltage 400 V DC
Input impedance 100 kΩ shunted by 5 pF

DIMENSIONS
H x W x D 17.5 x 15 x 27.5 cm (7x6x11 in)
Weight 2 kg ($4^1/_2$ lbs)

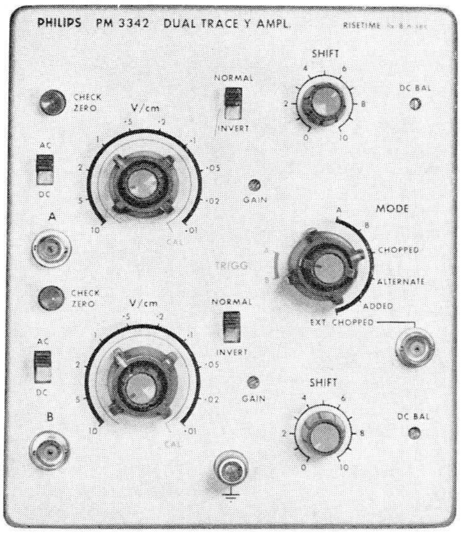
ACCESSORIES
PM 9331A/10 STANDARD PROBE
Coupling DC
Attenuation 10 : 1
Tolerance 3 %
Input impedance 10 MΩ shunted by 8 pF
Maximum input voltage
DC coupled 1000V, AC coupled 400V

Input attenuator as used in high frequency units



DUAL TRACE VERTICAL AMPLIFIER PM 3342

by R. Paans



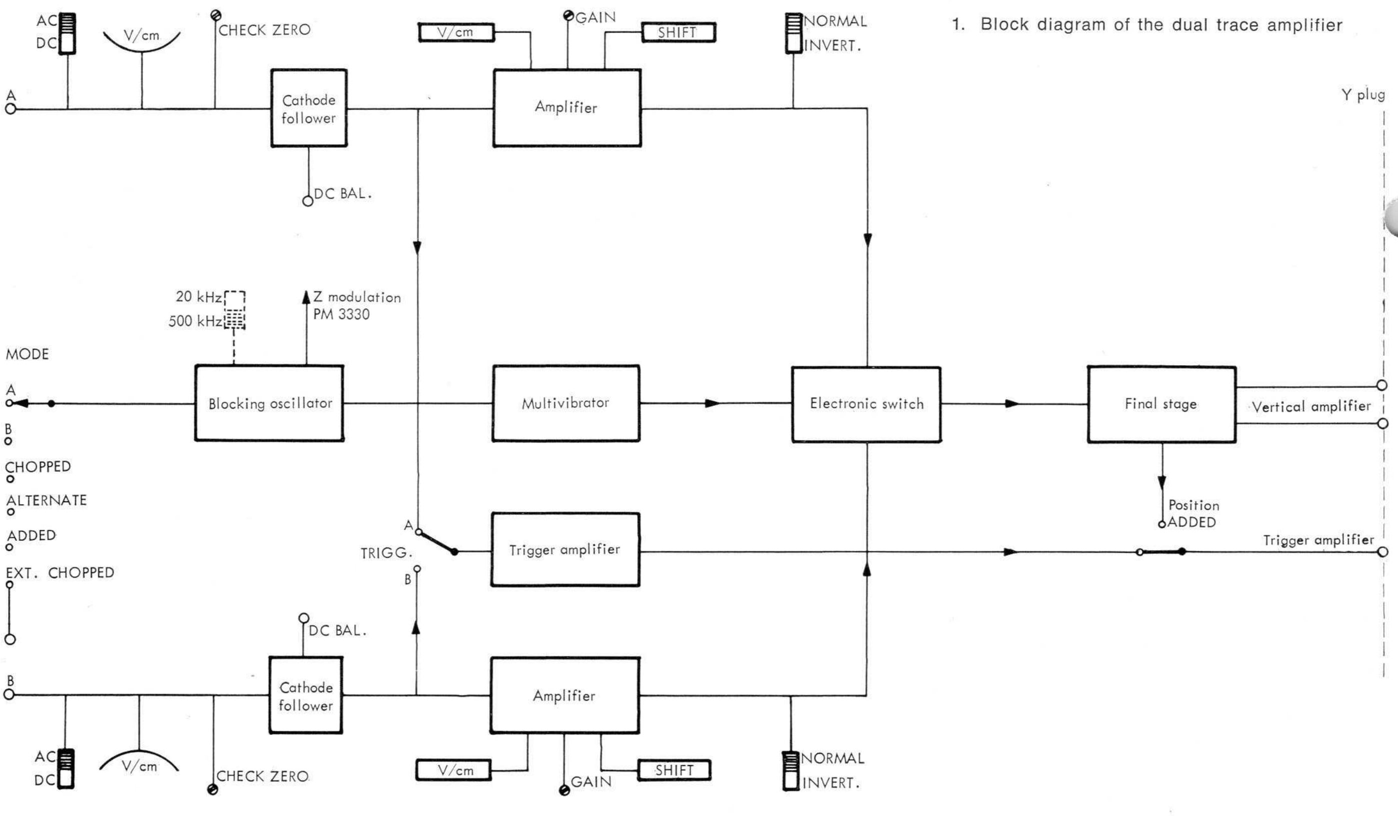
The PM 3342 is a plug-in preamplifier for the basic oscilloscope PM 3330 with two independent input channels (dual trace unit). The signal from either channel can be displayed individually or the two signals can be displayed simultaneously, either on a time-sharing basis (A and B) or algebraically added. Internal triggering is possible in both cases without the switching frequency causing any hindrance.

Description of the block diagram

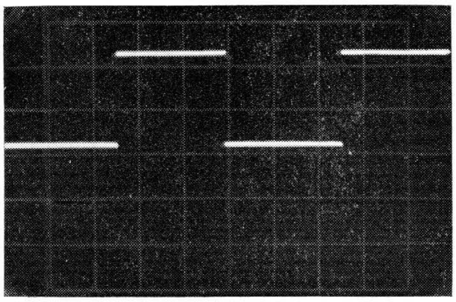
The channels A and B both consist of an input circuit incorporating the AC/DC switch and the step attenuator (V/cm), as well as an amplifier which contains the continuous attenuator (V/cm), the SHIFT and the polarity controls, see fig. 1.
Switching of the channels takes place at the output of these amplifiers, after which the combined signal is applied to an output amplifier. The switching circuit, which consists of a blocking oscillator and a bistable multivibrator (Schmitt trigger), supplies the switching voltage and the blanking pulse for the suppression of the electron beam during switching.
The trigger voltage for the time base generator is derived from one of the two input circuits via a selector switch. This makes internal triggering from both channels possible without the switching frequency causing any hindrance. The trigger voltage is amplified to the required amplitude by a trigger amplifier.
In the position ADDED of the MODE switch, the trigger signal is taken directly from the output amplifier, so that the trigger signal is again derived from the signal to be displayed. If this unit is used as difference amplifier, this method offers the advantage that triggering is not effected by the common mode signal. The operating method of the unit is selected by means of the MODE switch. The switching possibilities are shown in fig. 2.

Electronic switch

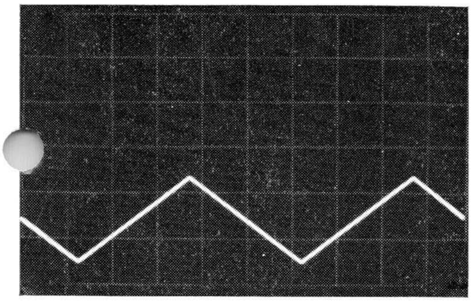
The electronic switch incorporates, per channel, a push-pull cascode amplifier with four switching diodes. The base controlled transistors TS_1 and TS_2 act as a current source (see fig. 3). Due to the low input impedance of the following earthed base transistors TS_3 and TS_4, current control takes place at the junctions C and D. These junctions are, therefore, very suitable for the connection of a diode switch for rapidly making or breaking the connection between the preceeding amplifier and the output amplifier.
In fig. 4 the diodes are represented by switches. Channel A is connected via D_3 and D_4 to the output amplifier; channel B is short-circuites via D_1' and D_2', while the connection between channel B and the output amplifier is, moreover, interrupted by D_3' and D_4'.
If the switch is changed over periodically, channels A and B are alternately connected to the output amplifier. Switching is effected by two square wave voltages which are symmetrical with respect to each other and which are fed to the diodes.



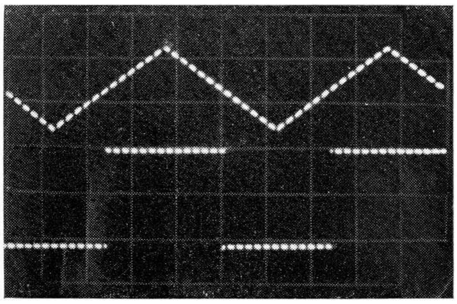
1. Block diagram of the dual trace amplifier



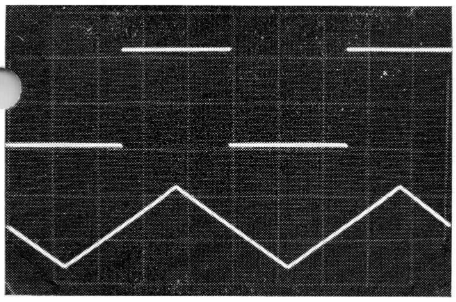
CHANNEL A
Channel A is switched on
Channel B is inoperative



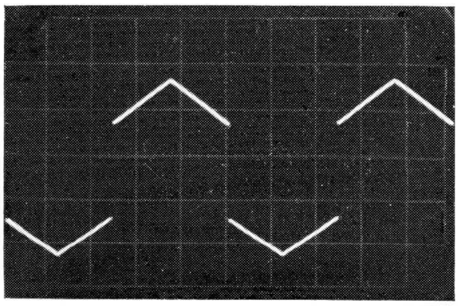
CHANNEL B
Channel B is switched on
Channel A is inoperative



CHOPPED
Both channels are now switched on. The accompanying photograph shows a single-shot exposure from which the switching method is apparent. The frequency at which the channels are switched can be set to 20 kHz or 500 kHz. The frequencies can also be chosen arbitrarily by external triggering of the blocking oscillator

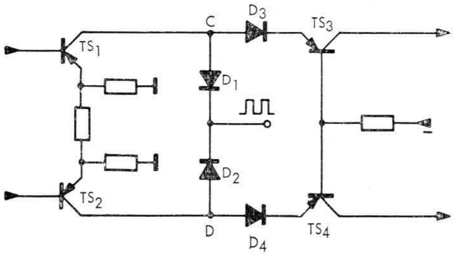


ALTERNATE
The signals at channels A and B are now traced alternately. Switching over takes place during the flyback of the time base



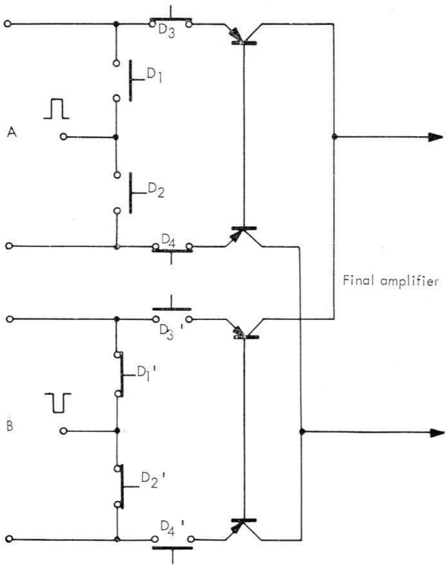
ADDED
The voltages at the output of both amplifiers A and B are added together. Depending on the position of the polarity switches, either the sum or the difference of the input signals is applied to the vertical amplifier of the basic oscilloscope. This makes it possible to use the unit as differential amplifier. Maximum common mode rejection can be adjusted by making the amplification factors of the two channels exactly equal to each other.

2. The switching possibilities of the mode switch

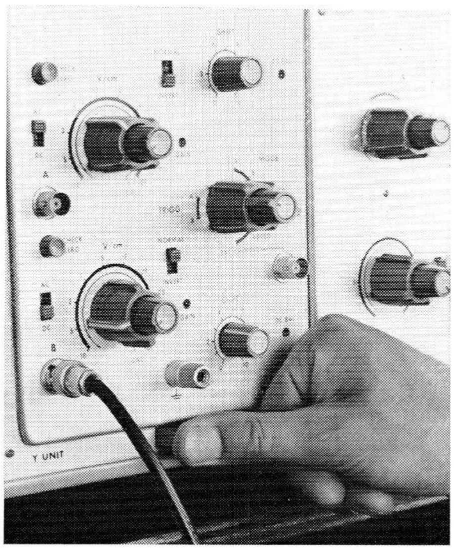


3. Circuit diagram of electronic switch

4. Principle of switching circuit

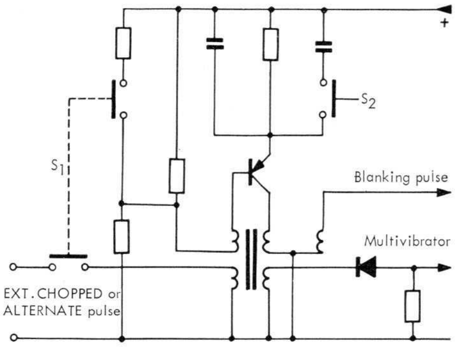


Push a button and the plug-in unit releases and slides forward



Blocking oscillator

The circuit diagram of the blocking oscillator is shown in fig. 5.
This oscillator produces very steep negative pulses. With S_1 opened (chopped position), the blocking oscillator will be free-running. The repetition frequency is determined by the RC-time of the filter incorporated in the emitter of the transistor.



5. Circuit diagram of blocking oscillator

Switch S_2 enables a choice to be made between 40 kHz and 1 MHz, so that interference patterns in the signal to be displayed can be prevented. During switching, a suppression signal is applied to the cathode-ray tube, so that the switching surges are not visible.

In the position ALTERNATE, S_1 is closed. The blocking oscillator is now monostable and produces one pulse per cycle of the applied trigger signal. The trigger signal is generated during flyback of the time base. The channels A and B are therefore alternately switched on and off at each subsequent sweep of the time base.

Multivibrator

Fig. 6 gives a representation of the bistable multivibrator.
Negative trigger pulses produced by the blocking oscillator are applied via a diode gate to the conducting transistor. As a result, the multivibrator switches over at each pulse and the square waves thus obtained from the two collectors are used to control the diode switch.
In the position channel A (or channel B) of the MODE switch, the multivibrator is so "locked" by a negative auxiliary voltage on the base of the transistor TS_2 (or TS_1) that only channel A (or channel B) is connected to the output amplifier (see fig. 6).
In the position ADDED, the auxiliary voltage is applied to both the base of TS_1 and TS_2, so that neither transistor conducts. The electronic switches of both channels will therefore be closed, so that both channel A and channel B are connected to the output amplifier.

Technical data

All data for this unit are valid when used in combination with the main frame PM 3330

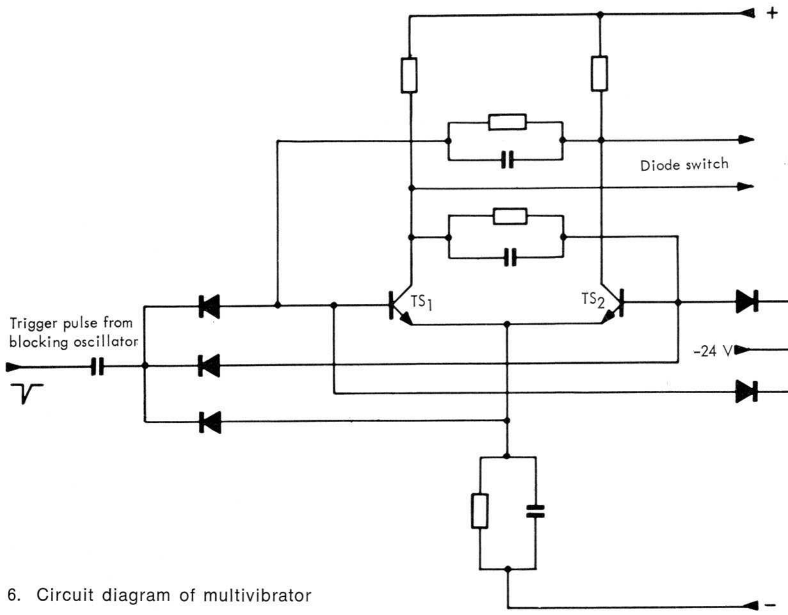
AMPLIFIER
Bandwidth
DC coupled: DC to 35 MHz
AC coupled: 1.6 Hz to 35 MHz
Rise time 10 ns
Sensitivity 10 mV/cm
Mode of operation
1. Channel ± A only
2. Channel ± B only
3. Alternate ± A and ± B
4. Chopped between ± A and ± B: at 20 kHz, 500 kHz or external up to 100 kHz
5 Added ± A ± B

Attenuator
Calibrated 10 mV/cm to 10 V/cm (1, 2, 5 series), continuous uncalibrated gain between steps
Tolerance 3 %
Overshoot < 2 %
Input impedance 1 MΩ shunted by 15 pF
Max. input voltage AC coupled 400 V DC
Max. deflection 6 cm for the full bandwidth, 18 cm mid-band (10 MHz)
Shift range > 12 cm
Common mode rejection 100 : 1 at 1 MHz
Maximum common mode signal
0.1 V (at 10 mV/cm sensitivity)
Trigger signal selection Channel A or B; or A+B when the added mode is used

SIGNAL FOR EXTERNAL CHOPPING
Frequency Less than 100 kHz
Rise time 1 to 100 ns for square waves
Amplitude
1 V to 20 V max. (symmetrical square wave)

DIMENSIONS
H x W x D 17.5 x 15 x 27.5 cm (7 x 6 x 11 in)
Weight 3 kg (7 lbs)

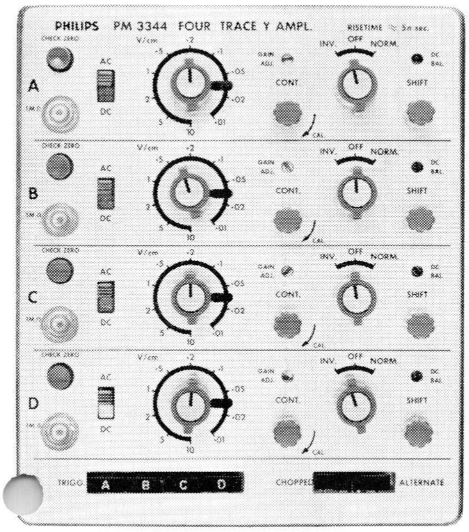
ACCESSORIES
PM 9331A/10 attenuator probe
Coupling DC
Attenuation 10 : 1
Tolerance 3 %
Input impedance 10 MΩ shunted by 8 pF
Maximum input voltage
DC coupled 1000 V, AC coupled 400 V
Cable length 110 cm



6. Circuit diagram of multivibrator

FOUR TRACE VERTICAL AMPLIFIER PM 3344

by N. G. Schultheiss



The PM 3344 is a four trace unit for oscilloscope PM 3330. With this unit four signals can be displayed simultaneously. Moreover there is the possibility of changing the polarity of each channel or simply be switching it off.

There are two methods of switching the various channels:

CHOPPED

The signal of each channel is displayed for 500 ns; any channel out of operation is ignored.

ALTERNATE

The signals of each channel are displayed successively, while any channel out of operation is skipped.

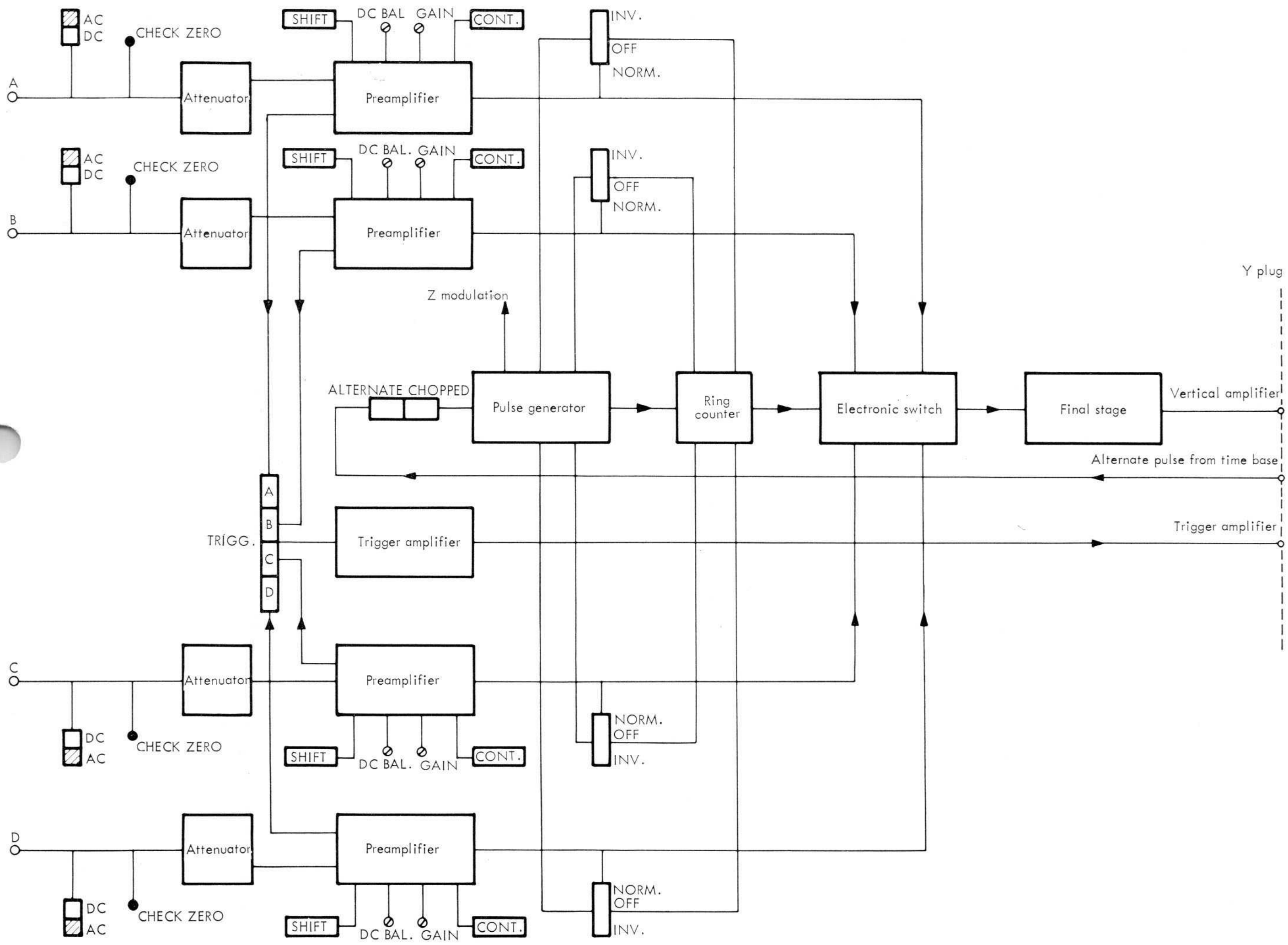
To simplify operation, the unit is fitted with an internal trigger amplifier which transmits the required trigger signal from channel A-B-C or D to the basic apparatus. The block diagram is given in fig. 1.

Attenuator

This is a high-impedance attenuator comprising ten steps of 10, 20, 50 mV/cm up to 10 V/cm. The input capacitance of the attenuator is adjusted so that the measuring probes can readily be interchanged.

Amplifier

To give the input stage a high impedance input circuit one section of a double triode has been connectd as a cathode follower (see fig. 2).



1. Block diagram of the four trace amplifier

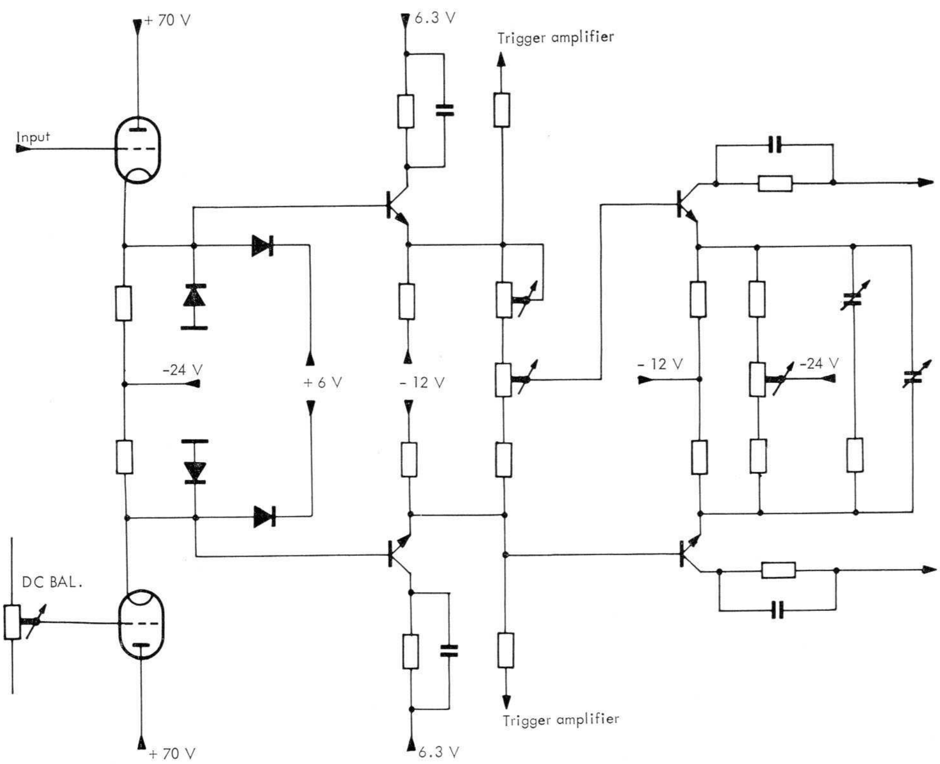
To keep drift in the amplifier to a minimum the other section of the valve is used as a cathode follower in the non-driven part of the amplifier. This part also contains the DC balance.
The above-mentioned circuit is followed by a safety circuit which protects the succeeding transistors against too high an input voltage. The signal is taken-off via 2 emitter followers. These serve for impedance transformation. The output impedance of the triode (1/S), is about 100 Ω and is reduced to a few ohms.
The trigger signal is also taken from the emitters. As the continuous gain control and the gain adjustment are connected between the emitters of this stage, no shift will occur at correct DC balance when the continuous gain control is operated (the emitters have the same potential). Moreover, the trigger voltage is balanced before it is applied to the trigger amplifier.
This emitter follower stage is followed by the preamplifier. The latter consists of 2 transistors, emitter-coupled and base-driven. The shift serves to change the current distribution between the transistors.
Moreover, a compensation network for high frequencies has been connected between the emitters. Owing to the high-ohmic output impedance, the circuit can be regarded as a current generator.
To eliminate low frequency errors RC networks have been inserted in the collector circuit.

Normal/off/invert switch
The normal/off/invert switch is a 3-position rotary switch. In the first and third position the leads of the preamplifier and the electronic switch are interchanged so that the signal is inverted 180°. The centre position of the switch disconnects the relevant channel from the output amplifier by means of the electronic switch.
When 3 channels are switched off, the pulse generator is stopped. No blanking pulses are fed to the cathode-ray tube so that the remaining channel gives the most stable image possible.

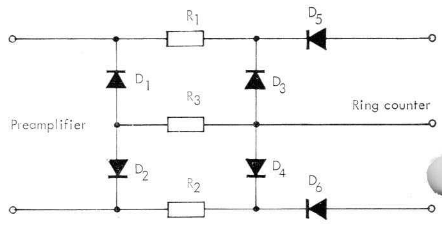
Electronic switch
Each channel of the electronic switch contains a circuit as shown in fig. 3.

If the pulse from the ring counter is negative with respect to the input level of the output stage, the diodes $D_{1-2-3-4}$ will be blocked while D_{5-6} are conductive. This means that the relevant preamplifier is connected to the output stage. If the pulse of the ring counter is positive with respect to the input level of the output stage, the diodes $D_{1-2-3-4}$ will become conductive, D_{5-6} will be blocked. The output stage is now switched off from the relevant preamplifier. The preamplifier is then supplied with current from the ring counter via the conductive diodes $D_{1-2-3-4}$. As a blocked diode always possesses a certain capacitance there could be a chance of the signal being capacitively transmitted. To minimize this chance the circuit is doubled. D_1 and D_2 short-circuit the signal after which any remaining signal is further attenuated by the resistors R_1, R_2 and the conductive D_3 and D_4. R_3 serves to equalize the current through D_{1-2} and D_{3-4}.

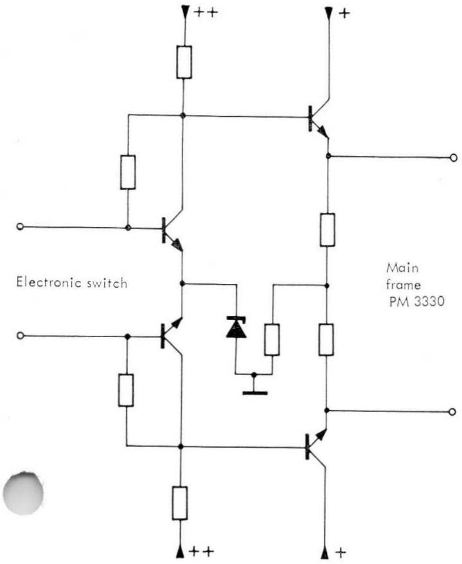
2. Circuit diagram of the preamplifier



3. Circuit diagram of electronic switch



Output stage
The output stage consists of:
– 2 shunt-feedback coupled transistors so that a very low-ohmic input circuit is obtained. The input voltage required for the electronic switch is obtained by setting the emitters of this stage at a fixed voltage by means of a Zener diode.
– 2 output emitter followers feeding the signal to basic apparatus PM 3330, see fig. 4.



4. Circuit diagram of the final stage

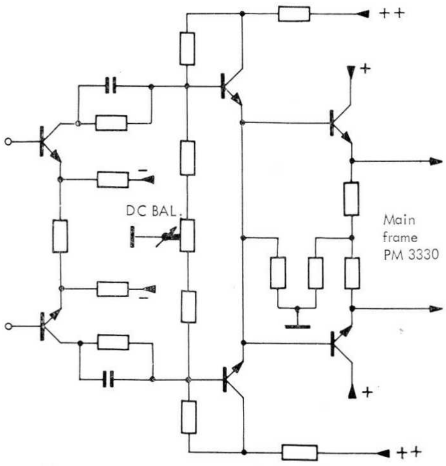
Trigger amplifier

The trigger signal can be picked up from each channel. The choice is made by means of push buttons. The signal is fed to a series and shunt-feedback coupled transistorized trigger amplifier, see fig. 5. Via 2 emitter followers the signal is fed to basic apparatus PM 3330. For perfect LF square wave response, RC networks have been employed in the collector circuit of the first stage.

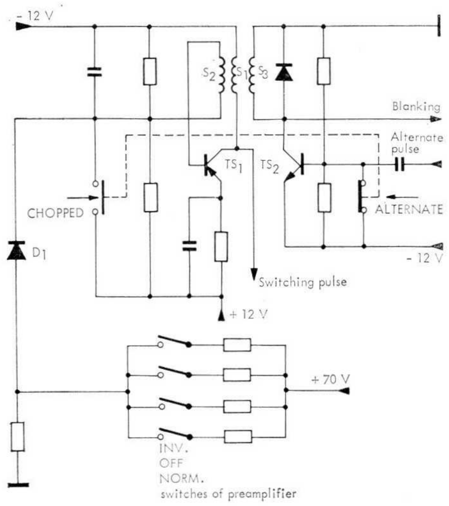
Pulse generator

This generator comprises a blocking oscillator (fig. 6). When the current in PNP transistor TS_1 increases, a negative going voltage will be applied to the base of this transistor by means of the blocking transformer. As a result the collector and also the emitter current increase even more, so that the core material of the transformer rapidly reaches saturation. The negative voltage the transformer supplies to the base disappears so that the transistor is rapidly switched off. The transistor cannot draw current again before the emitter capacitor, which was charged in the meantime, has discharged again. The RC network in the emitter circuit ensures that the repetition frequency of the oscillator is about 2 MHz. When the base rest potential is equal to or positive with respect to the emitter potential, the oscillator has stopped. This happens in two cases:

1. When 3 amplifiers are switched off, by means of a diode, the potential will become



5. Circuit diagram of trigger amplifier



6. Circuit diagram of the pulse generator containing the blocking oscillator and pulse shaper

7. Simplified diagram of ring counter

positive with respect to the emitter, and the blocking oscillator will be switched off so that no further pulses come from the oscillator and no beam suppression takes place.

2. When the alternate pushbutton is depressed, the base quiescent potential will be equal to the emitter potential. At the same time the alternate pulse from the PM 3330 will, after differentiation, be fed to the base of TS_2.

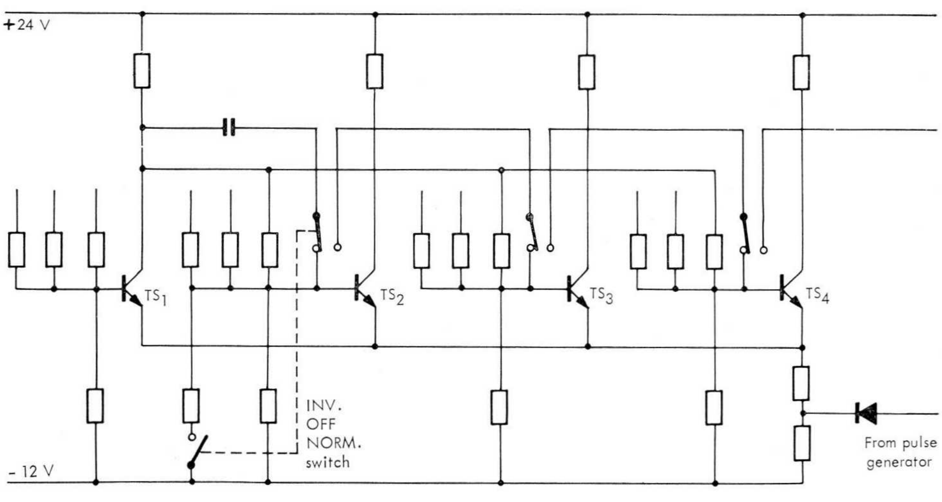
The collector of this transistor is connected to a winding S_3 of the blocking transformer so that an incoming alternate pulse produces a negative pulse on the base of TS_1. Consequently the blocking oscillator will operate once for each arriving alternate pulse.

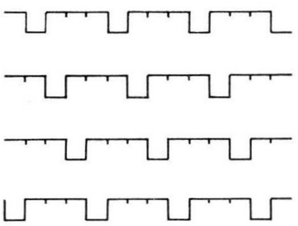
Winding S_3 also serves to supply the blanking pulse. A diode connected in parallel with S_3 ensures that when the blocking oscillator is switched off rapidly the voltage across the transformer does not change polarity; this could cause undesirable restarting of the ring counter.

Ring counter

The ring counter consists of 4 emitter-coupled transistors (fig. 7), so connected that if one draws current, the other three are blocked. This has been achieved by coupling the collector of each transistor to the base of the other three by means of voltage dividers.

The collectors of the transistors are also capacatively coupled to the base of the following transistor via a switch (normal/off/invert). At each incoming positive pulse from the generator the current through the conductive transistor will momentarily fall off to zero so that the collector becomes positive. Via the above-mentioned capa-





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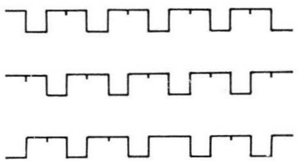

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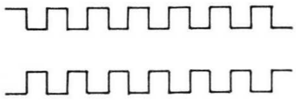

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8. Waveforms on the collectors during the switching
9. Waveforms if only one channel is switched off
10. Waveforms if two channels are switched off

citor this surge is fed to the base of the next transistor which is now in the conductive state.
This process is repeated for each pulse from the generator (in position "chopped" every 500 ns, and in position "alternate" at each fly-back of the time base).
The pulse generator is coupled to the ring counter by means of a diode so that only the peak of the pulse is fed to the ring counter.
The voltage waveforms at the collector are shown in the diagram (fig. 8.) The collectors of the ring counter are connected to the electronic switch. As already mentioned before, the relevant channel can be switched on or off with a negative and positive voltage respectively.
If a channel is not used (normal/off/invert switch in position off), the transistor in the ring counter corresponding to this channel is cut off by connecting the base to the negative supply voltage via a resistor.
Moreover the capacitor connected between the preceding collector and the relevant base is switched over to the base of the next functioning transistor. The collector waveforms are shown in fig. 9. If two channels are switched off, the collector voltage is as shown in fig. 10.
The switching frequency in position "chopped" is always 2 MHz, so that in this position a different channel is displayed each 500 ns.

Technical data

All data for this unit are valid when used in combination with the main frame PM 3330

INPUT
Identical for A, B, C and D.
CHECK ZERO knob for checking zero level and identification

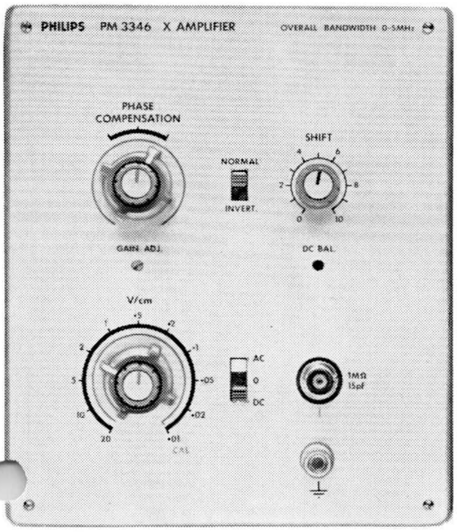
AMPLIFIER
The PM 3344 is a four trace plug-in amplifier
Bandwidth
DC coupled: DC to 50 MHz
AC coupled: 1.6 Hz to 50 MHz
Rise time 7 ns
Sensitivity 10 mV/cm
Mode of operation
Channels A. B. C and D, chopped at 2 MHz or alternate
Each channel can be switched off separately
Blanking during switching
Attenuator
Calibrated 10 mV/cm to 10 V/cm (1, 2, 5 series), continuous attenuation 1 : 3
Tolerance 3 %
Overshoot < 2 %
Maximum input voltage: AC coupled 400 V DC
Magnifier 3x screen height
Polarity reversible for each channel
Triggering
From any of the four channels
Sensitivity
3 mm for frequency up to 10 MHz
Phase shift between channels is not noticeable

DIMENSIONS
H x W x D 17,5 x 15 x 27,5 cm (7 x 6 x 11 in)
Weight 3 kg

ACCESSORIES
PM 9331A/10 passive probes

HORIZONTAL AMPLIFIER PM 3346

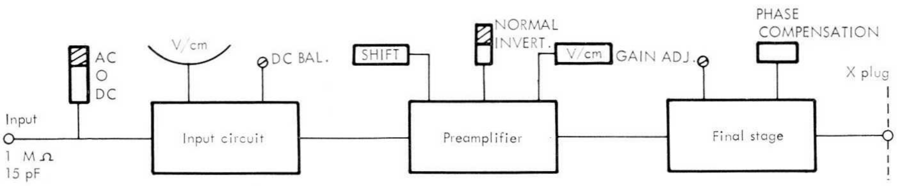
by G. C. W. Imbens



The PM 3346 is a plug-in unit designed for the horizontal channel of the PM 3330. This unit enables the PM 3330 to be used as an X-Y oscilloscope, fig. 1.
The difference in transit time between the horizontal and vertical channel (partially due to the delay line in the vertical amplifier) is compensated for by means of a delay line having adjustable delay, so that the phases of the both channels can be balanced.
The unit can also be used for reproducing signals on a non-linear (logarithmic, exponential) time axis, in which case the sawtooth voltage of the main frame PM 3330 can be used.
The intensity pulse of time base A can be applied to the cathode-ray tube Wehnelt cylinder by means of a slide switch in the unit.

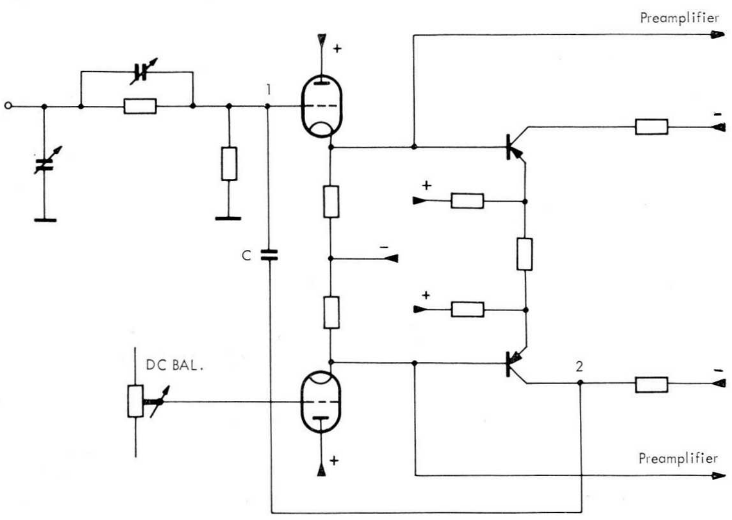
Input circuit
The input signal is fed via an AC, off, DC switch to the input circuit, which incorporates a high-ohmic step attenuator. Despite the use of a simple attenuator an input capacitance of 15 pF is obtained, fig. 2.
This has been achieved by amplifying the output signal of the attenuator in a long-tailed pair circuit and feeding the amplified signal in phase back to the attenuator via a capacitor.
The reduction of the input capacitance is then equal to (A—1), where A is the gain between point 1 and point 2.

Preamplifier
The preamplifier is a symmetrical series and shunt feedback amplifier (Cherry and Hooper, see also PM3333), fig. 3. The series feedback stage contains a BCY55, so that any drift can be kept at low level. The emitter resistor has been made variable and is used as a continuous gain control. The shift and polarity selection controls are included in the circuit between the output of the series feedback stage and the input of the shunt feedback stage.

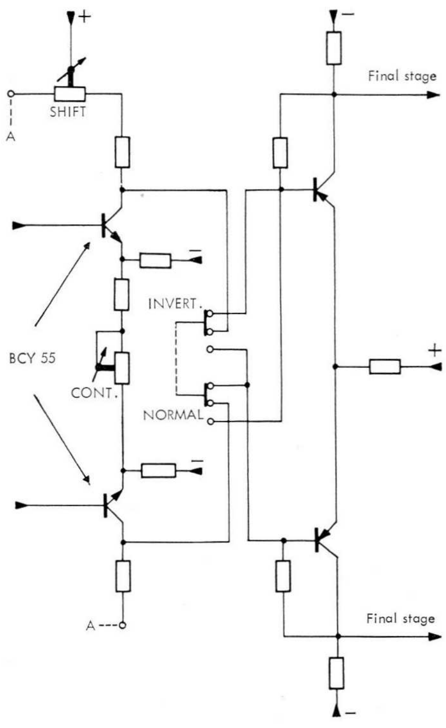


1. Block diagram of the horizontal amplifier

2. Circuit diagram of the input stage



3. Circuit diagram of the preamplifier



Final amplifier

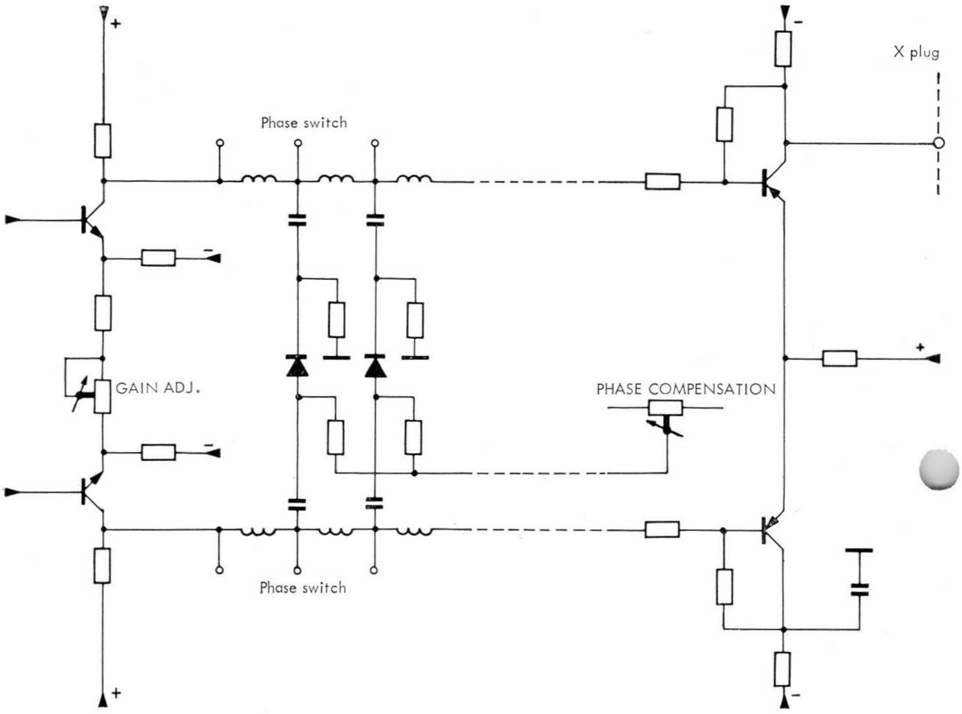
The final amplifier is also a series and shunt feedback amplifier, fig. 4. A potentiometer connected between the two emitters of the series feedback stage provides deflection sensitivity adjustment (GAIN ADJ.). The output stage incorporates a symmetrical delay line consisting of filter sections. The capacitance of the delay line is achieved with variable capacitance diodes (varicaps). The capacitance can be changed by varying the direct voltage across the diodes, so that the delay time of the line can be adjusted continuously. The delay can also be adjusted in steps by filter sections.

Technical data

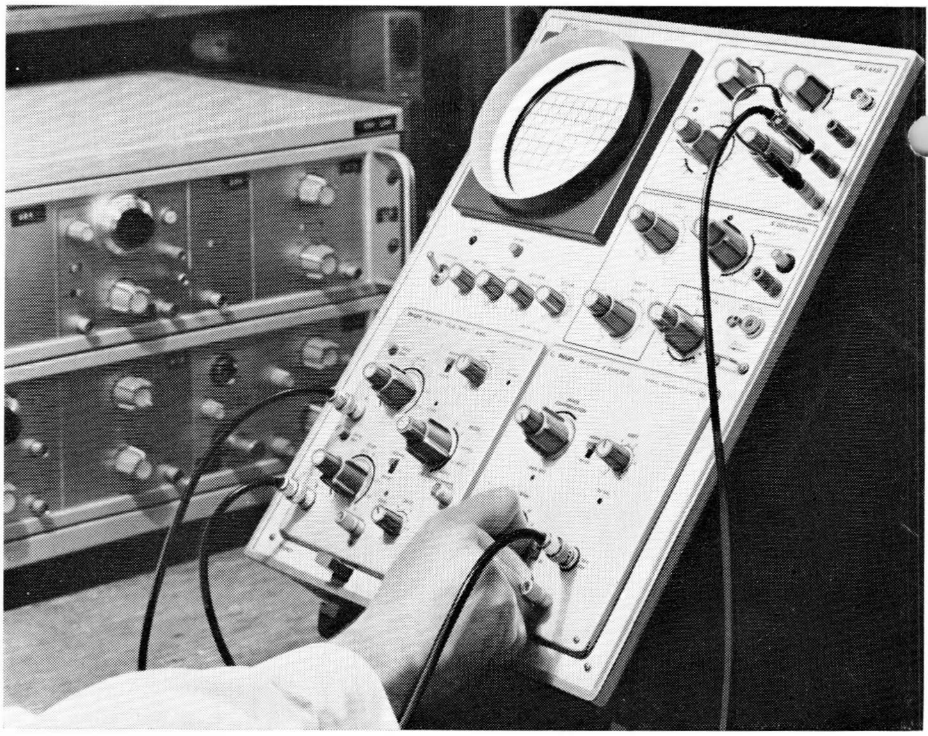
All data for this unit are valid when used in combination with the main frame PM 3330

AMPLIFIER
Bandwidth
DC coupled: DC to 5 MHz
AC coupled 1.6 Hz to 5 MHz
Sensitivity 10 mV/cm
Attenuator
Calibrated 10, 20, 50 mV/cm continuous (uncalibrated 1 : 3)
Tolerance 3 %
Input impedance 1 MΩ shunted by 15 pF
Maximum input voltage AC coupled 400 V DC
Maximum deflection 10 cm for the full bandwidth
Polarity can be either positive or negative

DIMENSIONS
H x W x D 17,5 x 15 x 27,5 cm (7 x 6 x 11 in)
Weight 1,5 kg 3,3 lbs

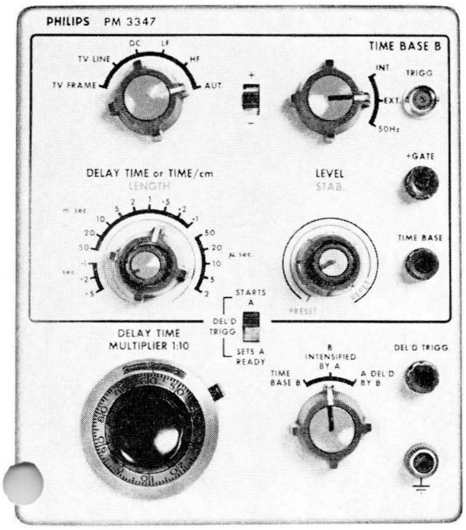


4. Diagram of the final stage



SWEEP DELAYING TIME BASE PM 3347

by A. H. J. Sloots.



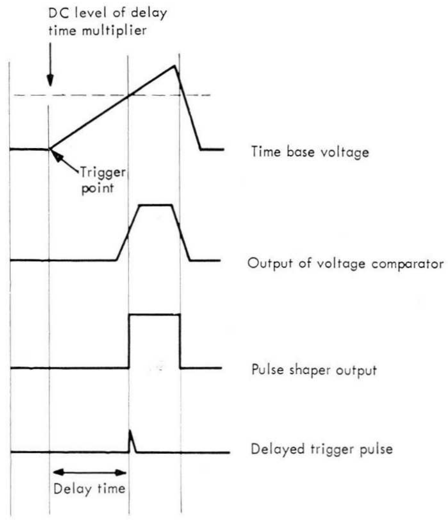
With the unit PM 3347 the oscilloscope is capable of displaying any small part of a signal which is complex with respect to time. Moreover accurate time interval and jitter measurements are possible. This unit can supply both a sawtooth voltage for the horizontal deflection and a trigger pulse for delayed starting time base generator A of the basic oscilloscope. Fig. 1 shows the block diagram of the unit and fig. 2 the relevant waveforms.

The sawtooth voltage derived from time base generator B is compared with an accurately adjustable DC voltage (DELAY TIME MULTIPLIER 1 : 10). The output voltage of this comparator is passed to a pulse shaper, which supplies the delaying pulse. This delaying pulse is fed internally to the time base generator A. The horizontal deflection voltage (sawtooth) is selected with a three-position selector switch affording the following alternatives:

TIME BASE B

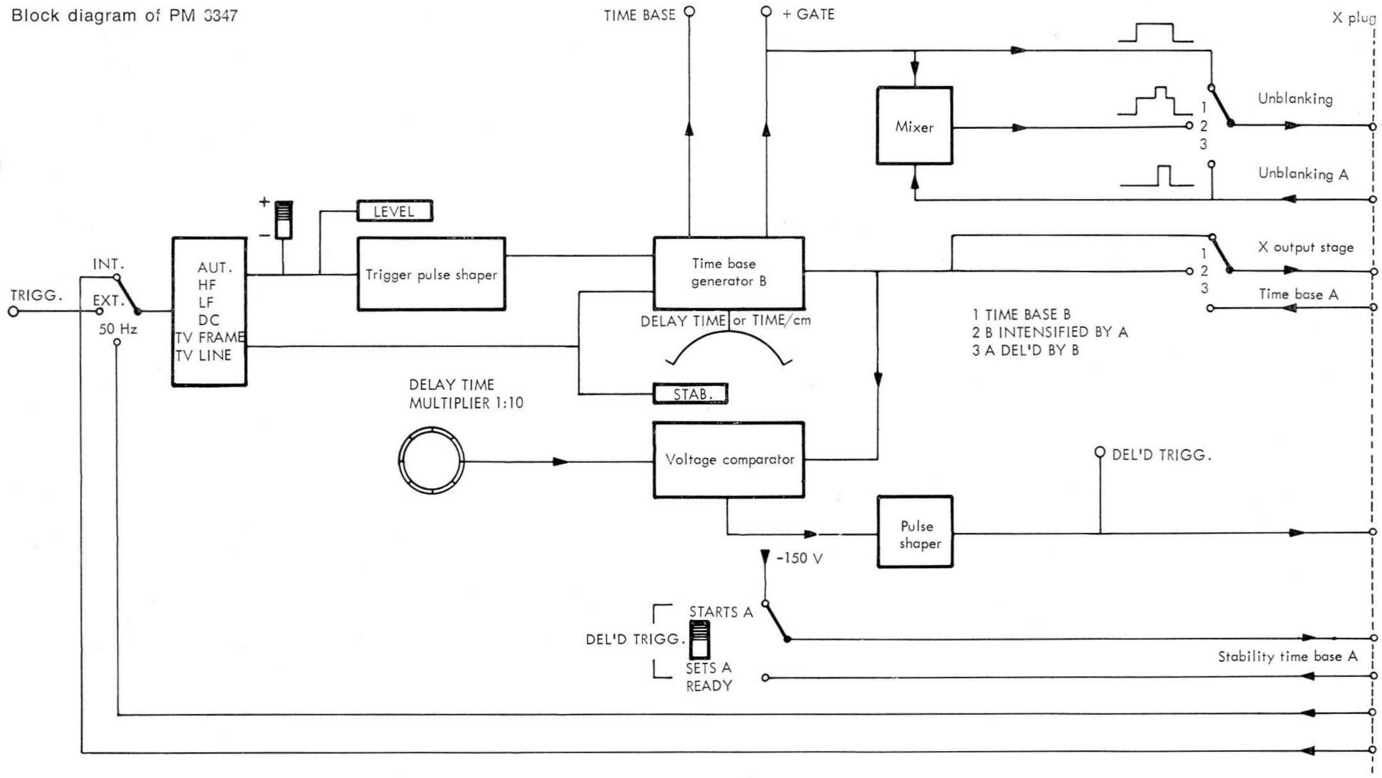
B INTENSIFIED BY A
Time base B, with extra bright modulation of the trace to indicate the interval in which time base A is operating

A DEL'D BY B
Time base A, delayed by B

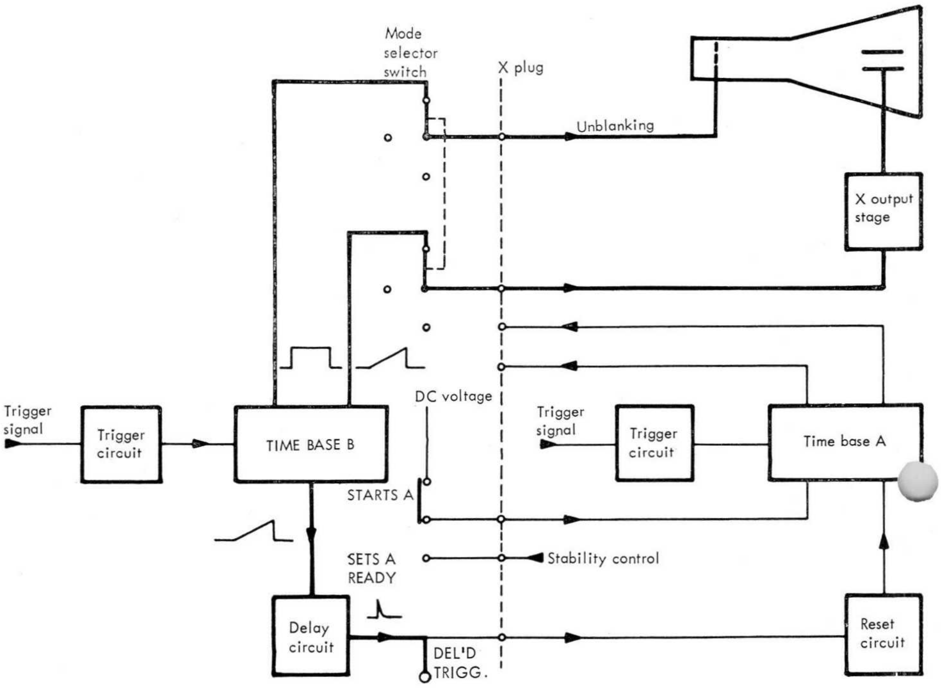


2. Relevant waveforms of fig. 1

1. Block diagram of PM 3347



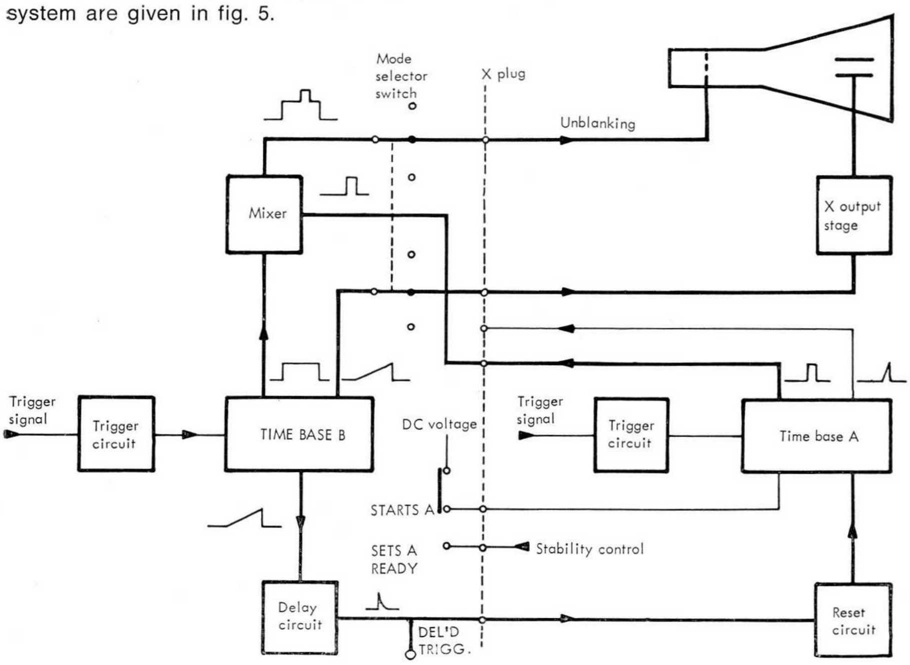
Mode selector switch set to
TIME BASE B
With the X-deflection switch on the main frame PM 3330 set to X UNIT and the mode selector switch of the plug-in unit to TIME BASE B, the complete function of time base A is taken over by time base B. The relevant interconnections are as shown in fig. 3.
The sweep-gating multivibrator in the time base circuit B provides the unblanking pulse for the CRT. The sawtooth output of the time base circuit B drives the X output stage in the basic oscilloscope. A delayed output pulse is available at the output socket.



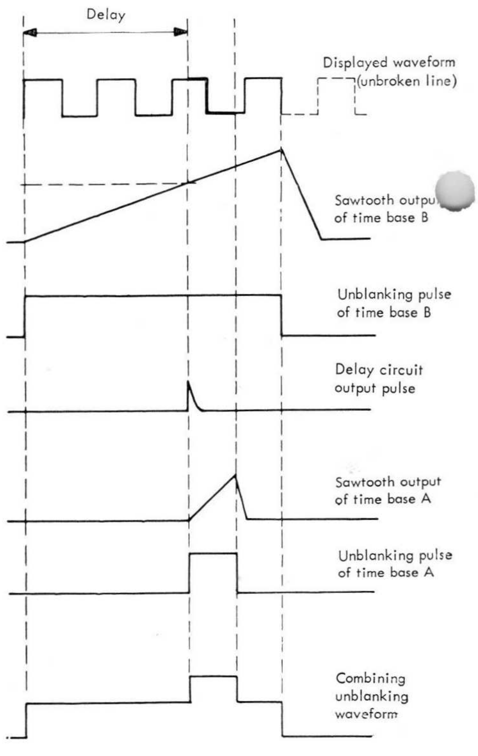
3. Mode selector switch set to TIME BASE B

Mode selector switch set to
B INTENSIFIED BY A
With the X deflection switch on the main frame PM 3330 set to X UNIT and the mode selector switch of the plug-in unit set to B INTENSIFIED BY A the relevant interconnections are shown in fig. 4. The sawtooth output of time base circuit B drives the X output stage and the delay circuit. The output pulse of the delay circuit releases time base A via the reset circuit. Thus time base A generates a sweep and an unblanking pulse after a pre-determined delay time. This unblanking pulse supplements that delivered by time base B. The waveforms obtained in the operation of the system are given in fig. 5.

4. Mode selector switch set to B INTENSIFIED BY A



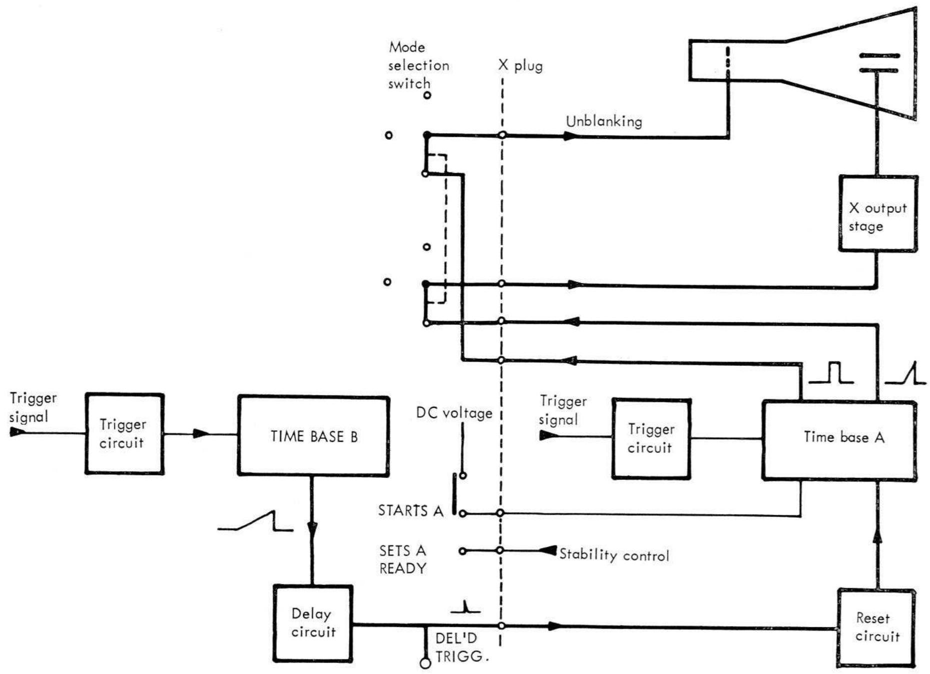
5. Waveforms showing the operation in the B INTENSIFIED BY A mode



Mode selector switch set to
A DEL'D BY B

With the X deflection switch on the main frame PM 3330 set to X UNIT and the mode selector switch of the plug-in unit set to A DEL'D BY B the relevant interconnections are as shown in fig. 6. The X output stage is also driven by time base A. Unblanking being provided by time base A only. The output pulse from the delay circuit releases time base A via the reset circuit.

The portion of the display which was made brighter in the previous position of the mode selector switch is now expanded horizontally to fill the whole screen. The waveforms showing the operation of the system are given in fig. 7.



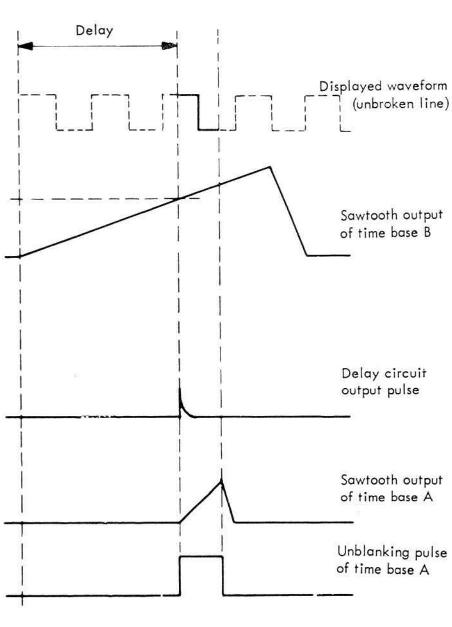
6. Mode selector switch set to A DEL'D BY B

Gated working of time base A
As described above, time base A starts immediately on the arrival of the delayed trigger pulse via the reset circuit. This mode of operation is selected by a small front panel switch when it is set to the position DEL'D TRIGG. STARTS A. In the other position of the switch, DEL'D TRIGG. SETS A READY, the delayed trigger pulse releases time base A in its normal triggered mode of operation. The next trigger pulse from trigger unit A to arrive after the set delay time actuates the time base A, which is then locked to this trigger signal. The waveforms showing the operation of the gated working of A DEL'D BY B are given in fig. 8. The total delay is now the sum of the delay time set on the unit (the product of the values indicated by the knobs DELAY TIME and DELAY TIME MULTIPLIER) and the extra delay as indicated in fig. 8.

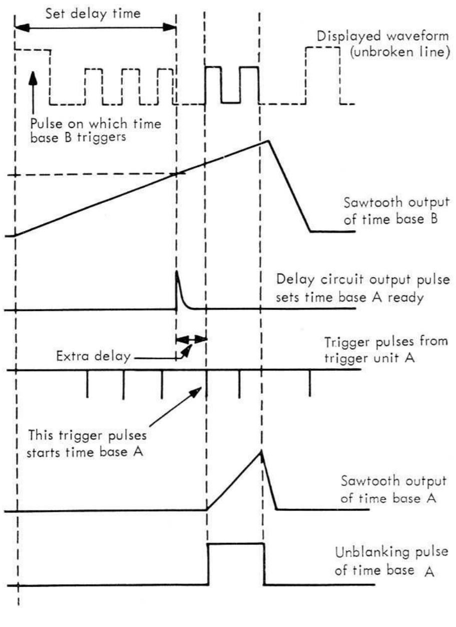
Time base generator B
Time base B, like time base A, is based on a Miller run-up circuit with the following differences:
- The sweep lengh control is no longer a preset adjustment but a front panel control LENGTH. Continuous sweep speed control is not available
- The lock-out multivibrator (single-shot facility) is not duplicated

For a detailed description of the time base circuit the reader is referred to that given for time base A on page 5.

7. Waveforms obtained with the A DEL'D BY B function in operation

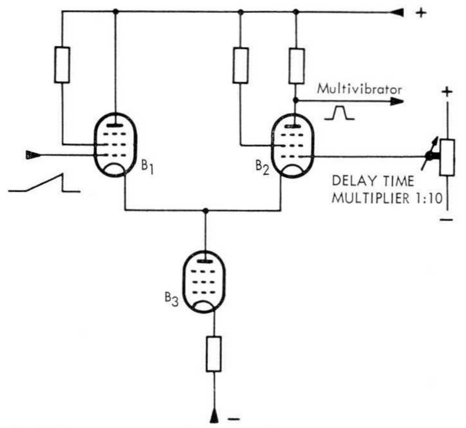


8. Waveforms obtained in operation with the gated working of A DEL'D BY B

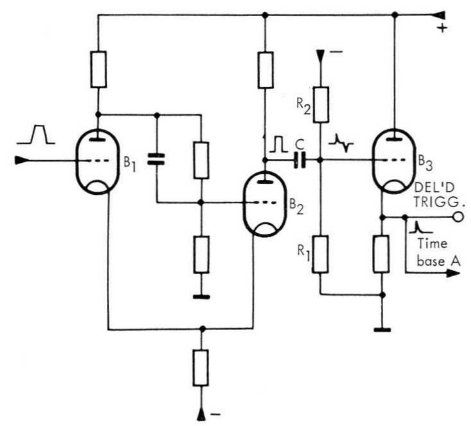


Voltage comparator

The sawtooth voltage from the time base B output is applied to the grid of B_1 (see fig. 9); this valve will start to conduct when its cut-off voltage is exceeded. However, since B_3 is a constant current source supplying the cathodes of B_1 and B_2, an increase in the current through one (B_1) will cause an equal fall in the current through the other (B_2).
Furthermore the use of a constant current source in the cathode makes the current in B_2 independent of the setting of R (DELAY TIME MULTIPLIER 1 : 10), which is used to determine the point at which the take-over of current by B_1 from B_2 occurs. In this way the output from B_2 starts from a constant level and provides a constant amplitude step.

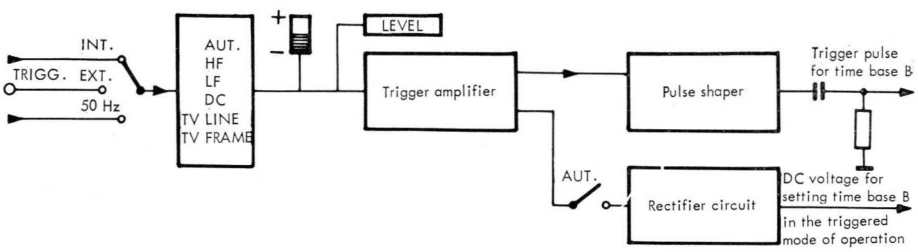


9. Voltage comparator circuit



10. Pulse shaper and its associated circuits

11. Block diagram of the trigger circuit



Pulse shaper

The output of the voltage comparator is applied to a pulse shaper (Schmitt trigger circuit) consisting of valves B_1 and B_2 (see fig. 10). In the quiescent state B_1 is cut-off and B_2 is conducting. When the voltage, applied to the grid of B_1, rises above the switching level of the Schmitt trigger the circuit switches over. Thus the anode voltage of B_2 rises abruptly.
When the input voltage drops below the switching level (during the flyback interval) the anode voltage of B_2 falls back to the quiescent value. Thus a square wave voltage is produced at the anode of B_2. This square wave is differentiated by capacitor C and resistors R_1 and R_2, so that sharp positive and negative peaks appear at the grid of the cathode follower B_3. When there is no input signal, B_3 is biased beyond cut-off so that only the positive peaks appear at the output. These pulses are also available at the socket DEL'D TRIGG.

Trigger circuit

Compared with the trigger unit of the basic oscilloscope, this trigger circuit is of simpler design in that the pulse shaper consists of a Schmitt trigger circuit (see fig. 11). The selection facilites for source, slope and mode are identical to those of the main time base.
The amplified signal of the trigger amplifier is fed to a Schmitt trigger circuit. The latter switches when the applied signal reaches the predetermined level. Thus a square wave orginates at its output, which is differentiated by an RC network. The resulting negative peaks start time base B. With the selector switch set to AUT., in the absence of a trigger signal, the time base is free-running. Thus in the absence of a signal a bright line is still obtained, even with the fastest sweep speeds of the time base.
When a signal is applied in the AUT. mode, the output voltage of the trigger amplifier is rectified, and the DC voltage obtained biases the time base into triggered operation.

Technical Data

All data for this unit are valid when used in combination with the main frame PM 3330

DISPLAY MODE
1. Time base B
2. B intensified by A
3. A delayed by B

TIME BASE B (as a main time base)
Mode Free-running or triggered
Sweep speed 2 μs/cm to 0.5 s/cm (1, 2, 5 series)
Tolerance 3%
Expansion 2 or 5 times via main frame PM 3330
Length 4 to 10 cm
Time base outputs
Sawtooth + 90 V_{p-p} from zero level;
gate + 35 V_{p-p} from zero level

Trigger source Internal, external or mains supply (adjustable in phase)
Trigger slope Positive or negative
Trigger mode

1. TV frame
2. TV line
3. Automatic 10 Hz to 1 MHz
4. DC
5. LF over 3 Hz
6. HF over 2 kHz

In modes 1, 2 and 3 manual level and stability are inoperative
In modes 5 and 6 high pass filter with RC time 56 ms, 80 μs respectively

Trigger sensitivity
Internal: 3 mm up to 1 MHz, 1 cm for automatic setting, 2 cm p—p video for TV
External: 0.5 V_{p-p} up to 1 MHz, 1 V_{p-p} in aut. position, 1 V_{p-p} positive video for TV

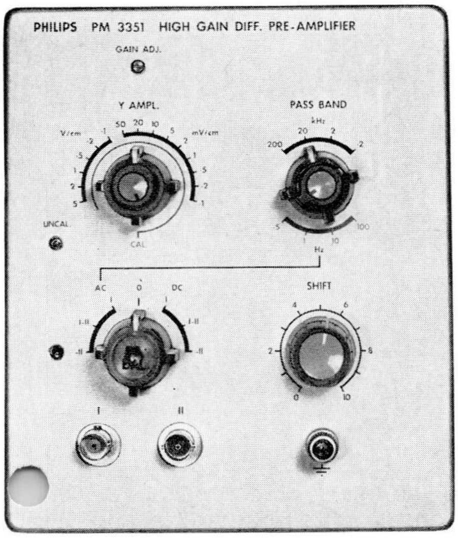
Trigger level
Internal: adjustable over 6 cm
External: adjustable over 6 V
Trigger input impedance: 1 MΩ shunted by 55 pF (used with DC, LF, HF or AUT.)

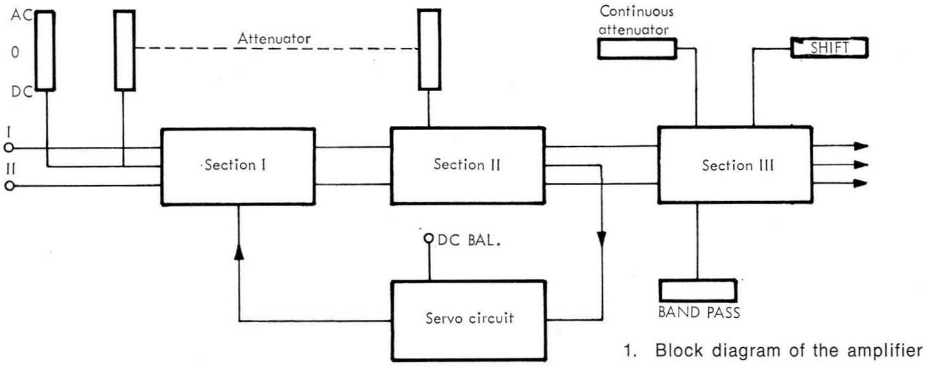
A DELAYED BY B (as a delaying time base)
Delay range Calibrated 2 μs to 5 s (1, 2, 5 series)
Tolenrance 3% + 200 ns
Intermediate adjustment Calibrated ten turn h pot
Tolerance 0.2% of the full scale time
Jitter < 1 : 20.000 of the full scale time
Delayed trigger pulse output 3 V_p

DIMENSIONS
H x W x D 17.5 x 15 x 27.5 cm (7x6x11 in)
Weight 2 kg (4½ lbs)

HIGH GAIN DIFFERENTIAL PREAMPLIFIER PM 3351

by G. Akkermans





1. Block diagram of the amplifier

The PM 3351 is a sensitive LF vertical plug-in unit, 100 μV (200 kHz), which is used in conjunction with the basic oscilloscope PM 3330. The PM 3351 consists of three amplifier sections (see fig. 1).

Section I is formed by two balanced cascode amplifiers and a cathode follower. This section is resiliently mounted for minimizing microphony.

Section II is a multi-stage amplifier with feedback. The gain of this amplifier can be step-controlled. By means of network attenuators at the input 15 different deflection coefficients can be selected in steps of 1, 2 and 5. This section also incorporates a control circuit for the semi-automatic control adjustment of the amplifier DC balance.

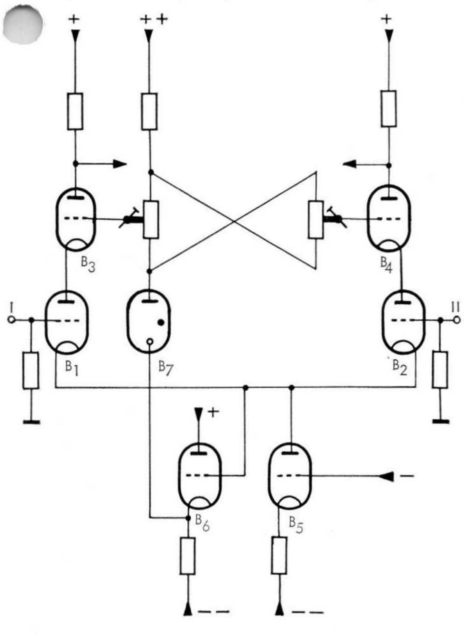
Section III is a multi-stage amplifier incorporated the continuous attenuator, the shift control and the bandwidth limiting switch.

A diagram of the cascode amplifiers in push-pull operation in section I is shown in fig. 2. The common-mode rejection factor of the amplifier, which depends on the equality of the gains of the two balanced halves, is high, because a valve is used instead of the cathode resistor. The gain is accurately adjusted by means of a DC voltage applied to the grids of B_3 and B_4. To ensure that the gains do not vary, if the input valves B_1 and B_2 are driven with voltages of equal phase, these grids are connected to the common cathode connection of the input valves via neon tube B_7 and cathode follower B_6. A voltage variation at the common cathode connection is now fed to the anodes of B_1 and B_2 via the above-mentioned coupling and the control grids of B_3 and B_4, so that the anode voltages of these valves, and consequently the respective gains, will remain constant. At higher frequencies the rejection factor decreases because the cathode impedance decreases owing to parasitic capacitances, and the gains of the two halves begin to differ.

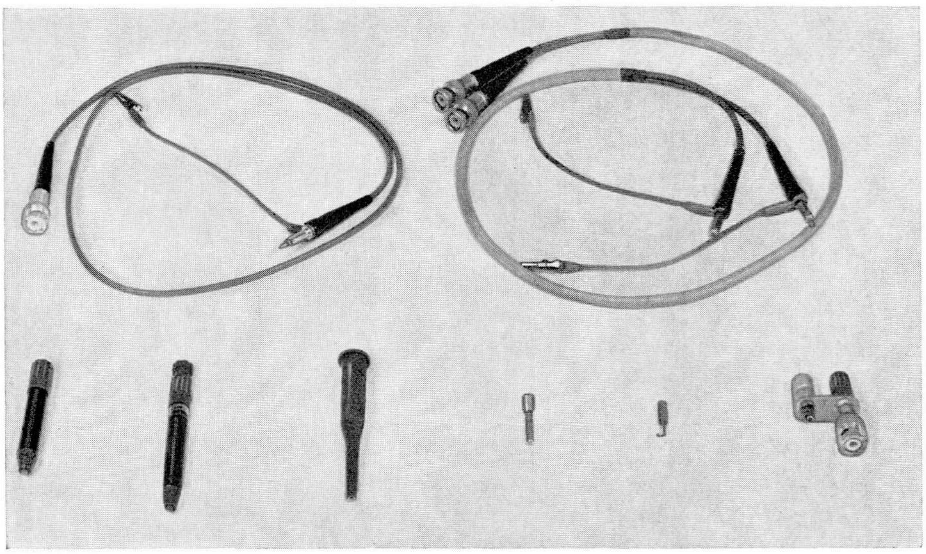
As the rejection factor also depends on the input circuit of the amplifier, a few special measures have been taken. The attenuations of the symmetrical network attenuators have been adjusted to equal values. Moreover the input capacitors are matched within 0.5% and the input RC-times are accurately adjusted to the same value so that the high rejection requirements are also satisfied in the AC positions of the AC/DC switch.

The signals from section I are fed to the second section via the cathode followers. The gain of this section is determined by a negative feedback circuit, the feedback factor of which can be adjusted in steps by means of the attenuator switch. In the first seven positions of the attenuator switch the gain of the complete amplifier depends on section II, so that in these positions the rejection factor is not affected by the input circuit.

2. Diagram of the push-pull cascode amplifier



3. Passive probe set, type PM 9328



In the AC position of the AC/DC switch large capacitors are included in the feed back circuit. The amplification of very low frequency signals, and also the drift, is now very low. The result is that the drift in the AC positions, caused primaraly by section I, is approximately a factor of one hundred smaller than in the DC positions of the AC/DC switch.
With the AC/DC switch in position 0 the DC balance of the amplifier can be semi-automatically adjusted by means of push-button DC balance. In this position the grids of the input valves are connected to earth. If, a voltage difference occurs at the output of section II due to unbalance, this difference is fed to a control circuit which governs a motor via a polar relay. This motor is coupled with a potentiometer in section I thus resting balance. With the AC/DC switch in position AC the pass-band switch can switch-in high-pass filters between sections II and III for −3 dB frequencies of 0.5, 1, 10 and 100 Hz.
Because of the high-pass frequencies input of section III has been given a high impedance. For this purpose emitter followers with a rather high emitter resistance, and transistors with a high h_{fe} are used. The next stage is a push-pull amplifier containing the continuous attenuator and the gain adjustment. By means of the pass-band switch various capacitors can be connected between the collectors of this amplifier stage so that in conjunction with the collector resistors and divider resistors, they form low-pass filters for −3 dB frequencies of 200, 20, 2 and 0.2 kHz.
Via resistive dividers, which include the shift potentiometer, the signal is fed to the output stage where the shift unbalance can be compensated by means of an adjustment potentiometer. Via plugs, the signal is fed from the collectors to the vertical amplifier and trigger amplifier of the basic oscilloscope PM 3330.

Technical data

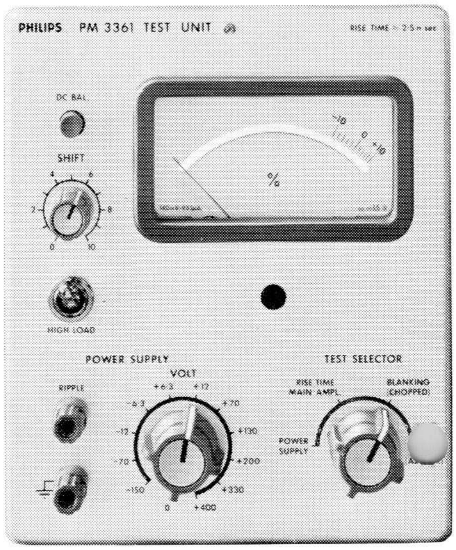
All data for this unit are valid when used in combination with te main frame PM 3330

AMPLIFIER
Bandwidth
DC coupled: DC to 200 kHz
AC coupled: 0.5 Hz to 200 kHz
Filters
Low pass: 0.2, 2, 20 and 200 kHz (3 dB)
High pass: 100, 1 and 0.5 Hz (3 dB)
Sensitivity 100 μV/cm
DC drift
DC coupled < 500 μV/h after first hour
AC coupled < 5 μV/h after first hour
Attenuator
Calibrated 100 μV/cm to 5 V/cm (1, 2, 5 serics) continuous uncalibrated gain between steps
Tolerance 3 %
Common mode rejection
at 0.1 to 10 mV/cm: 50,000 : 1 for 50 Hz
at 0.1 to 10 mV/cm: 1.000 : 1 for 200 kHz
at 20 mV/cm to 5 V/cm: 2,000 : 1 for 50 Hz
Maximum common mode signal
10 V at 0.1 to 10 mV/cm, 20 V at 20 mV/cm
50 V at 50 mV/cm, etc. up to 500 V max.
Input impedance 1 MΩ shunted by 50 pF
Maximum input voltage AC coupled 500 V DC
Maximum vertical deflection
18 cm undistorted (6 cm window) mid-band (100 kHz)
Shift range > 12 cm
Input selector
DC/AC channel I only
DC/AC channel I and II
DC/AC channel II only
O amplifier input disconnected from the input socket and connected to earth
DC balance Using servo system

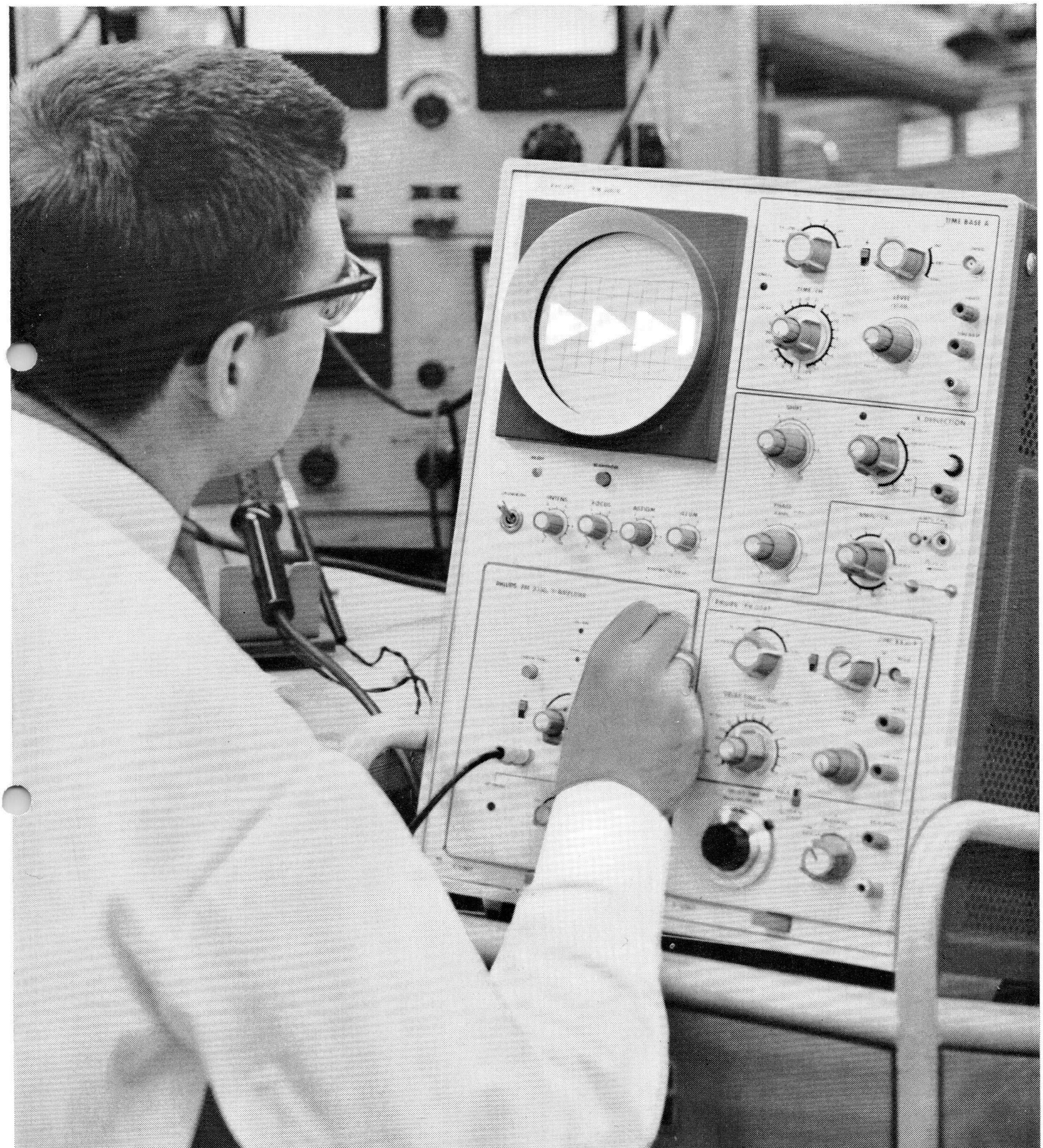
DIMENSIONS
H x W x D 17.5 x 15 x 27.5 cm (7 x 6 x 11 in)
Weight 4 kg (9 lbs)

ACCESSORIES
PM 9329 passive probe set
The PM 9329 consists of 2x PM 9324A/10, 2x M 773 204, a "siamesed" twin probe and 2x PM9051
PM 9329A/10 (blue)
Coupling DC
Attenuation 10 : 1
Tolerance 3 %
Input impedance 10 MΩ shunted by 12 pF
Maximum input voltage
DC coupled 1000 V,
AC coupled 500 V DC
M 773 204 head only (black)
Coupling DC
Attenuator 1 : 1
The "siamesed" twin probe cable uses low noise co-axial cable so that maximum use can be made of the high sensitivity and the effect of stray hum pick-up is reduced to a minimum.
Cable length 110 cm
PM 9051 BNC to 4 mm plug adaptor

PM 3361 Plug-in test unit

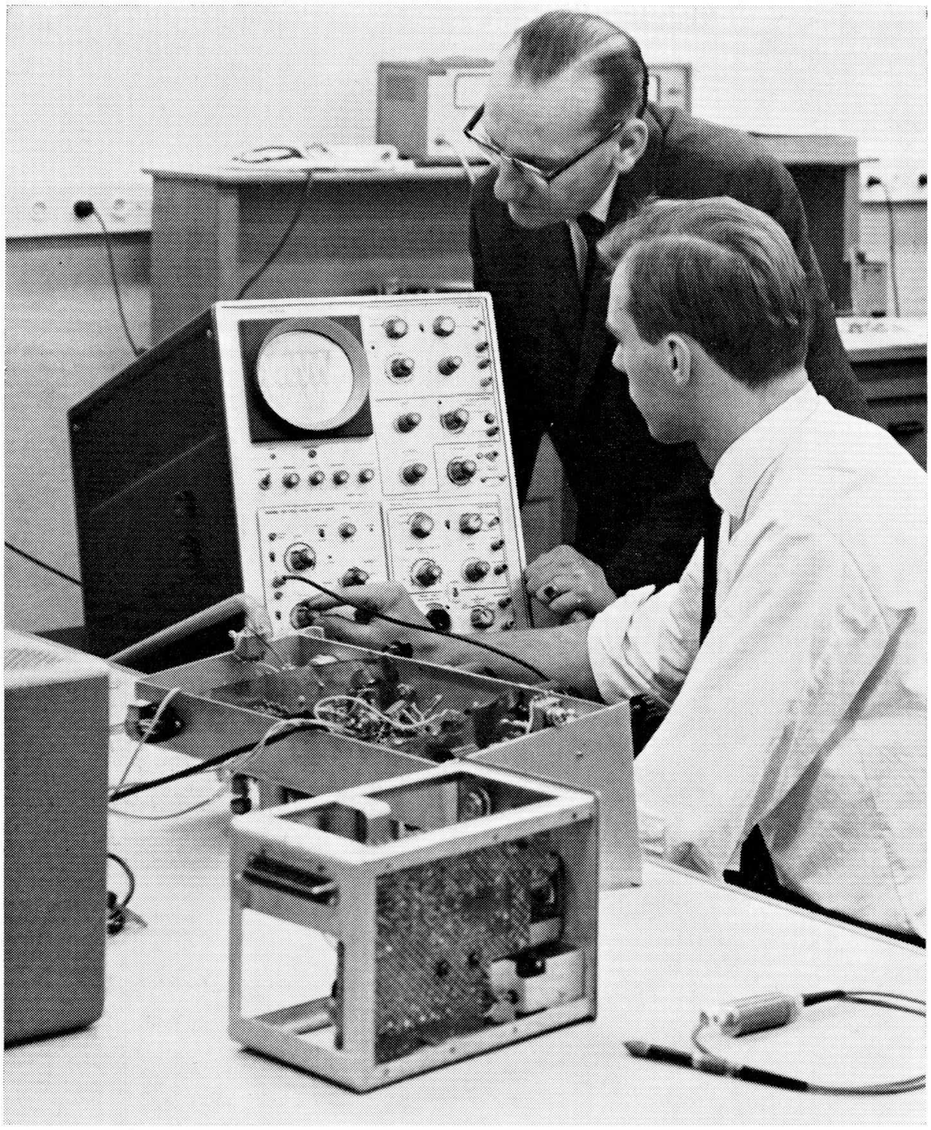


This plug-in facilitates testing of the main frame, output amplifier, power unit and blanking circuit. It provides:
I. A symmetrical square wave for testing the rise time of the vert. output amplifier.
Amplitude: 1.2 V_{p-p} = 4 cm deflection
Risetime: 2.5 ns
Frequency: 100 kHz and 500 kHz
Shift: Via panel control
Zero level: A press button on the front panel connects the output amplifier input to the zero level reference voltage
II. An alternate display with two traces 4 cm apart for adjusting the output amplifier sensitivity and checking the performance of time base switching pulses as used by multitrace amplifiers. This display is obtained by driving the square wave oscillator from the time base, thus all data other than frequency is identical to I above.
III. A built-in voltmeter with a selector switch for input, and range for each supply rail, facilitates rapid testing of the power unit. The output voltage can be checked for normal load and maximum permitted load. At the same time ripple voltage can be measured by connecting an oscilloscope.
IV. Blanking pulses are available to check that the main frame blanking circuit, used in the chopped mode of the dual trace unit, is functioning satisfactorily.

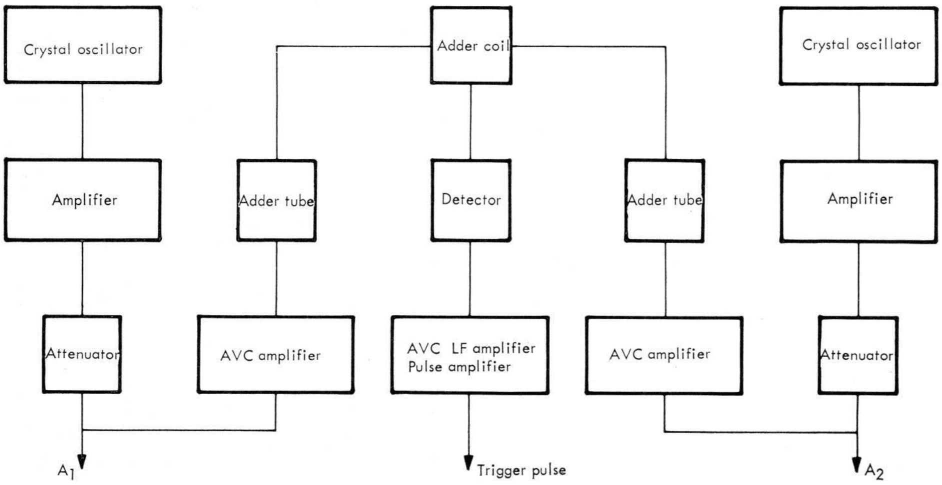

TIME BASE A
X DEFLECTION

DIFFERENTIAL FREQUENCY PHASE WOBBULATOR

by T. W. G. M. v.d. Staay
n.v. Volt, AT Laboratory, Eindhoven, the Netherlands



1. Block diagram of the phase wobbulator



A colour TV set employs several phase-discriminators and demodulators which serve both for demodulating the coded colour signal and for synchronizing the oscillator. It is possible to measure these phase-sensitive circuits and to align them in a completed receiver.

If, however, the above mentioned circuits are separately developed, tested and manufactured, it is necessary to have available signal sources which can be used for alignment. For this purpose two signals are required whose phase, with respect to each other, can be varied. This phase variation can be carried out in various ways. For instance, phase angle φ can be adjusted with a variable delay line. This method is quite suitable for static measurements. It has, however, also been found necessary to measure these phase sensitive circuits dynamically. For this purpose a phase wobbulator has been developed which employs the following system, fig. 1. Two crystal oscillators are adjusted to the same nominal frequency. By changing the parallel capacitance of the oscillators the frequency of both oscillator I and II can be detuned by 0, 10, 500 or 850 Hz. Each of the signals from the oscillators is amplified in an final stage which employs a tuned circuit in the anode. This has been done in order to keep distortion of the signals at a minimum (distortion $< 0.25\%$ at V_0 max $= 1.5\ V_{rms}$).

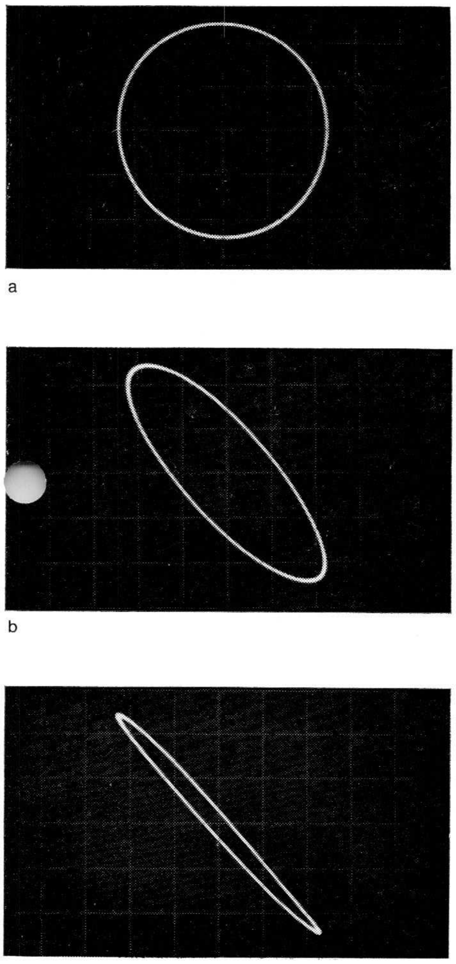
At the same time the impedance is transformed to 75 Ω, and the signal is fed to a coaxial attenuator. In order to adjust the two oscillator stages and the output stages use is made of an oscilloscope PM 3330 and dual trace unit PM 3342. The photograph to the left shows the amplitude of the two signals and the phase difference between them.

The PM 3330 is also used as an X-Y oscilloscope for measuring frequency (PM 3346 is then used as the horizontal amplifier). The frequencies of the two crystal generators are compared over a period of eight hours, giving the result of a phase shift of 80° between the two signals (see fig. 2). This stability is obtained by using a crystal oven (Overnaire product) set to a temperature of 70° ± 0.1°C.

Used as a wobbulator

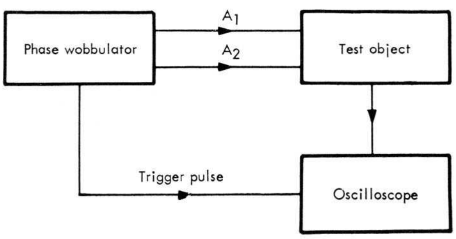
When used as a phase wobbulator one of the oscillator is tuned to a frequency f_1 (signal A_1) while the other is tuned to a frequency $f_1 + \Delta f$ (signal A_2). The "phase difference" between the two oscillators increases, per sec, by Δf times the full cycle from 0 to 360°.

The signals A_1 and A_2 are now fed to the phase-sensitive circuit which is to measured (see diagram fig. 3).

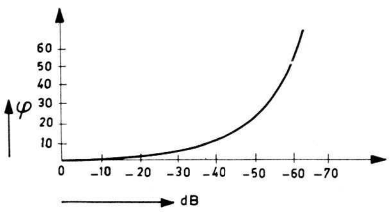


2. Stability test
a. start b. after 4 hrs c. after 8 hrs

3. Block diagram for phase sensitive circuit to be measured



The output voltage of this circuit depends on phase angle φ. If it is required to measure the output voltage, represented on the vertical axis of the oscilloscope, as a function of phase angle φ of the input voltage, the horizontal axis of the oscilloscope must be calibrated as a function of φ. The scale of φ is linear so that the



4. Phase difference φ as a function of attenuation

time base of the oscilloscope can be used for horizontal deflection. If the sweep-time of the time base is adjusted to equal $1/\triangle f$, the width of the screen corresponds to a sweep of 360°.
For oscilloscope PM 3330, where the length of the horizontal axis is 10 cm, 1 cm corresponds with $\triangle\varphi = 36°$ or, with an expansion of 5x, $\triangle f = 36/5 = 7\ {}^1/_5{}°/\text{cm}$.
For easy readability it is now only necessary to make the beginning of the horizontal axis correspond with $\varphi = 180°$. For this purpose a trigger pulse for starting the time base is obtained from the signals A_1 and A_2, at the moment that the phase difference $\varphi = 180°$, as described below.

Trigger pulse

The trigger pulse must be provided when the two output signals are in anti-phase. As the attenuator provides a phase shift as a function of attenuation (see fig. 4), it is necessary to compare the signals after the attenuator.
The output voltages A_1 and A_2 are added (see fig. 1) by means of the two adder valves which have a common anode impedance.
At the input of the detector one obtains a signal whose amplitude depends on the frequency difference $\triangle f$. Depending on $\cos \triangle\omega t = \pm 1$ the signal is either maximum or zero. It is given in the next explanation including fig. 5.

$Y_1 = A \sin 2\pi f_1 t \quad Y_2 = A \sin 2\pi (f_1 + \triangle f)t$
$X_1 = A \cos 2\pi f_1 t \quad X_2 = A \cos 2\pi (f_1 + \triangle f)t$

$$P = A \sqrt{(\sin 2\pi f_1 t + \sin 2\pi (f_1 + \triangle f)t)^2 + (\cos 2\pi f_1 t + \cos 2\pi (f_1 + \triangle f)t)^2}$$
$$= A \sqrt{2 + 2 \cos 2\pi \triangle ft}$$

5. Vector diagram for two signals

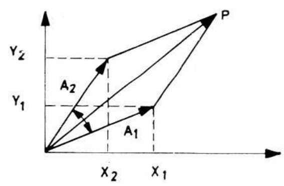
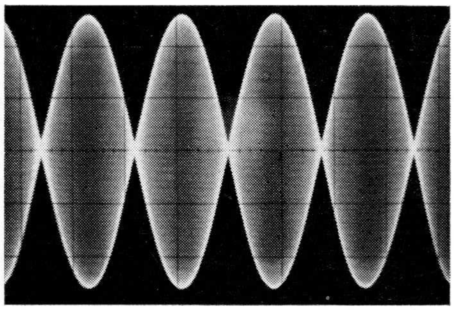


Fig. 6 shows the signal before and after detection (6a) and (6b) respectively. The signal is then fed through an amplifier which is biased so that only the peak of the signal is amplified.
Fig. 7 shows the pulse, obtained at the pulse output, which can be used to trigger the oscilloscope.
In order to obtain trigger pulses of constant amplitude and width, AVC amplifiers have been connected between the two signal outputs and the adder circuit, and also between the adder circuits and the trigger pulse output (fig. 1). Of the total 65 dB variation of the output signals A_1 and A_2, 30 dB is eliminated in the first two AVC amplifiers while the remaining part is compensated in the AVC amplifier following the detector.

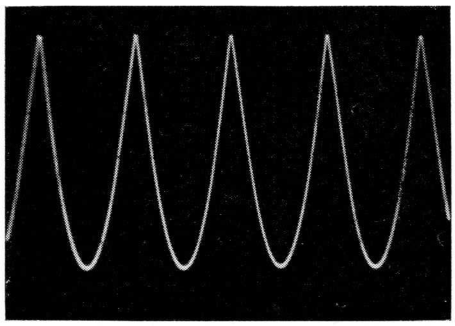
6. Signal before (a) and after (b) detection

7. Pulse used to trigger oscilloscope

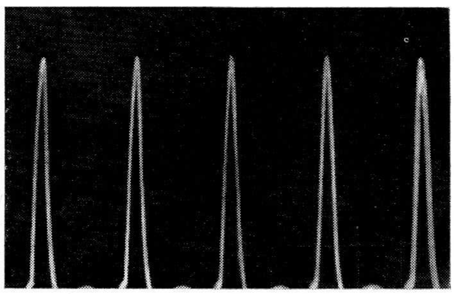
6a



6b



7



PAIRSPECTROMETRY IN NUCLEAR PHYSICS

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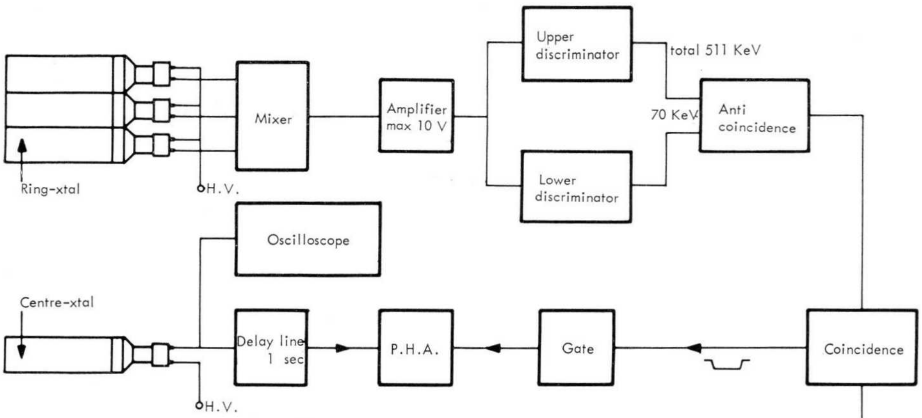
Harshaw-v.d. Hoorn n.v. de Meern, the Netherlands

A special gamma radiation spectrometer is described, which has several advantages over the conventional single crystal spectrometer. These advantages are:

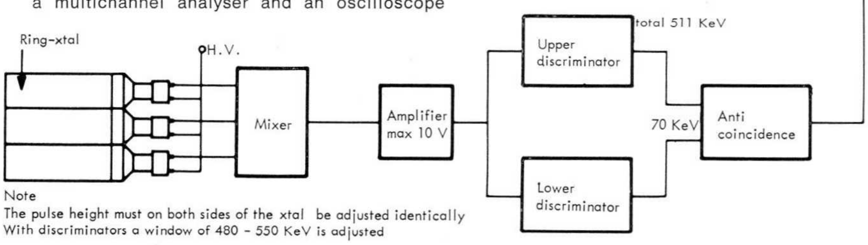
- Almost complete surpression of Compton and escape events
- Improved resolution

Before going into a more detailed description of the gamma radiation detector shown above, it is believed to be useful to mention some basic phenomena relative to the manner in which gamma radiation interacts with matter in general.
The three most important interactions are the following:

- Photoelectric absorption
- Compton scattering
- Pairproduction



1. Adjustment of a pairspectrometer by means of a multichannel analyser and an oscilloscope



Note
The pulse height must on both sides of the xtal be adjusted identically
With discriminators a window of 480 - 550 KeV is adjusted

Photoelectric absorption

In this process, a gamma quantum is absorbed in an atomic shell resulting in the liberation of a so-called "photoelectron". The kinetic energy imparted to this electron is equal to $E\gamma - E_k$, where E_k is the amount of energy needed to liberate the electron. As a result of this process, the atom is left in an excited state and consequently returns to the ground state through emission of an X-ray quantum or Auger-electrons, which takes place simultaneously with the liberation of the electron.
In favourable circumstances the X-ray quantum will be absorbed in the same medium, thus liberating a second electron and so on. The photo-electrons and Auger-electrons will dissipate practically the total energy $E\gamma$ in the absorbing medium.

Compton scattering

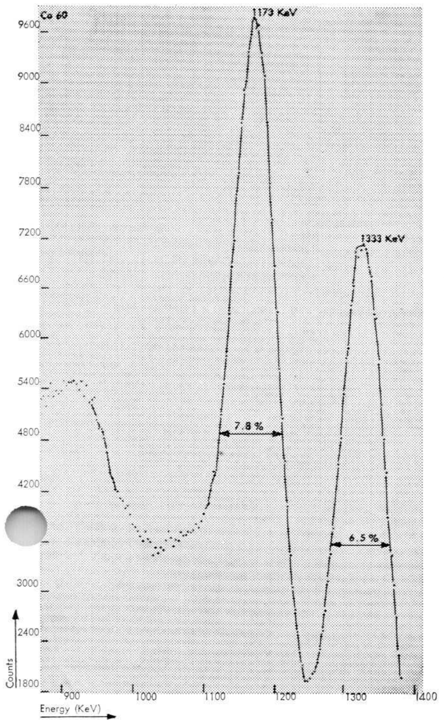
The Compton scattering of a gamma quantum results in the transfer of only a portion of its energy to an electron, whereas the degraded gamma quantum $E'\gamma$ continues in a different direction.
The relationship between $E'\gamma$, $E\gamma$ and the scattering angle is given by the socalled Compton formula:

$$E'\gamma = \frac{mc^2\,E\gamma}{mc^2 + E\gamma\,(1 - \cos v)}$$

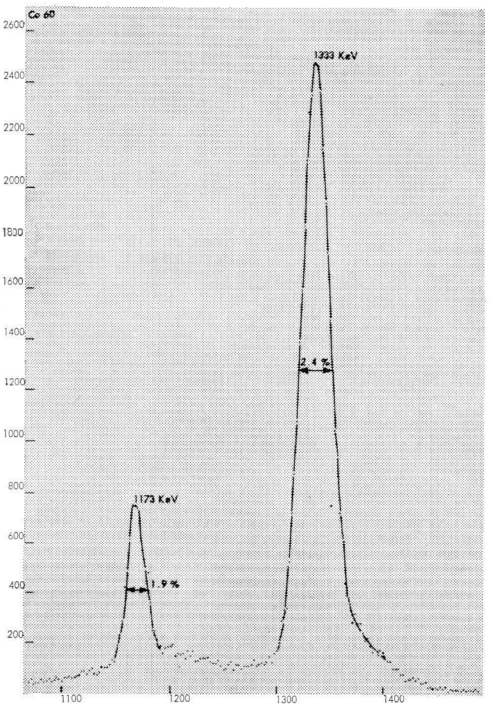
In this formula mc^2 is equal to the rest energy of the electron = 0.51 MeV.

Pairproduction

In the process of pairproduction, energy is directly converted into matter in such a way that an electron positron pair is created. This can only happen in the vicinity of a nucleus. The minimum amount of energy necessary for the creation of the two particles amounts to $2mc^2$ and the residual energy is imparted to both particles as kinetic energy. When the velocity

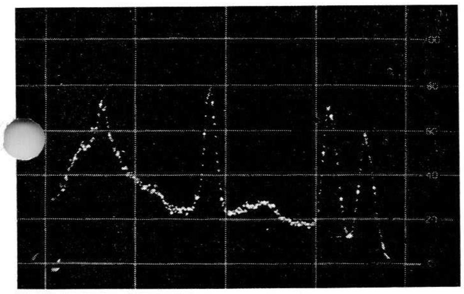


2. Co^{60} spectrum recorded with a conventional single crystal spectrometer consisting of a 1³/₄" diameter x 2" thick NaI (Tl) crystal coupled to a 2" photomultiplier tube



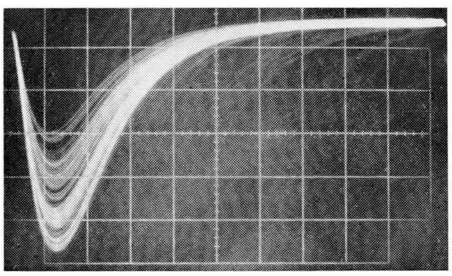
3. Co^{60} spectrum recorded with the pairspectrometer described in this article

is sufficiently reduced by loss of energy to the absorbing medium, a positron will interact with an electron resulting in total annihilation of both particles with production of two oppositely directed quanta of energy $mc^2 = 0.51$ MeV each.



4. Spectrum of C_s 137 (661 KeV) and $C_o{}^{60}$ (1.17 MeV + 1.33 MeV) on a spectrum analyser

5. Input pulses for the analyser corresponding to the spectrum in fig. 4 on a PM 3330 oscilloscope (The low energy pulses are eliminated)



A scintillation pairspectrometer

We are especially interested in the pairproduction process in connection with the detector mentioned earlier in this text. The detector consists of two parts, namely a centre crystal, coupled to a suitable photomultiplier tube, surrounded by an annulus of NaI (Tl), which in turn is viewed by several P.M. tubes. The crystal in the annular detector is divided in an even number of segments, which are optically separated and viewed by one P.M. tube each. When a gamma quantum with an energy exceeding 1.02 MeV enters the centre crystal, the possibility exists that pairproduction occurs according to the above described principle. As explained above, two diametrically opposed gamma quanta of 0.51 MeV each will ultimately be created and as long as the centre detector is not too big, these quanta will leave the crystal without causing interactions. There is a great probability however (depending on the diameter of the annular crystal) that they will be absorbed in two separate segments of the surrounding annulus. If this is the case, the arrangement will give three pulses, namely one from the centre detector and two from two different segments. These three pulses are fed into a coincidence unit which determines whether or not the three pulses do indeed arrive simultaneously.
In order to reduce the possibility of so-called "random coincidences", which may be caused by cosmic rays and other types of background radiation, pulses coming from the annulus segments are fed into the coincidence unit only if they correspond to an energy of 0.51 MeV. This gives the experimenter the certainty that he is indeed using the escape quanta for his coincidence requirement. This is achieved by means of a discriminator type of circuitry which allows pulses within a known finite energy interval to pass to the coincidence unit.
Only if pulses from two annulus' sections and the centre crystal arrive simultaneously at the input stage of the coincidence unit will the pulse from the centre detector be permitted to enter the spectrum analyser. In the event of only one or two pulses no signal is admitted to the analyser. The result of this method is that only interactions through pairproduction are registered which is extremely important since a number of other phenomena, such as background radiation and Compton scattered photons, will not be measured. In normal gamma ray spectroscopy, the latter phenomena are quite capable of obscuring a considerable number of details in a spectrum.
It should be noted however, that the pulse height of the pulses coming from the centre detector are no longer directly proportional to the energy of the incident radiation, since two quanta of 0.51 MeV have left the detector. Consequently, all of the values obtained in the spectrum will be 1.02 MeV lower than the incident energy $E\gamma$. This however, gives us the opportunity to use a pairspectrometer as a detector with considerably improved resolution. This possibility is perhaps best explained in an example:
Suppose Co^{60} gammas are detected with a pairspectrometer and that the second peak (1.33 MeV) is being studied. First the spectrum is recorded with the central detector alone. A normal resolution for this detector at 1.33 MeV will be about 6% which gives us a peak width at half maximum of about 80 KeV.
In using the pairspectrometer, only those events will be recorded that have left $E\gamma - 2 \times 0.51$ MeV in the crystal. In deducting 1.02 MeV from the 1.33 MeV mentioned earlier, we find that 0.31 MeV is absorbed in the centre crystal. At this energy level the resolution of the centre detector will be about 11% or 34 KeV. It can easily be seen that the peak width is reduced by a factor of more than 2 and that consequently the resolution has been improved from 6% to about 2.5%. Figures 2 and 3 show the difference between a Cobalt spectrum recorded with a single crystal arrangement and a complete pairspectrometer. The difference in resolution is quite evident.

Announcing a new sampling oscilloscope

by L. E. Orrevall
Philips Industrielektronik AB, Solna Sweden

Are you a little shy of a sampling oscilloscope? Maybe because you don't quite know how to use trigger hold-off, DC-offset or samples/cm. These are really no reasons to keep you from using Philips' sampling oscilloscope. It is as easy to operate as a conventional oscilloscope!

Why sampling?
Because it is a smart way to get around the problem of direct amplification of high frequency signals. A conventional oscilloscope has two fundamental limitations: firstly, the gain bandwidths product of existing amplifiers, secondly, the transit time of the electrons in the deflection system of the cathode-ray tube itself.

It is possible to use special oscilloscope tubes which may work up to 1000 MHz but they have no amplifiers and the sensitivity is therefore poor (approximately 10 V/cm). Generally wide band oscilloscopes have no better sensitivity than approximately 10 mV/cm. This is to be compared to a calibrated sensitivity of 1 mV/cm at an apparent bandwidth of 1000 MHz for Philips' sampling oscilloscope PM3419.

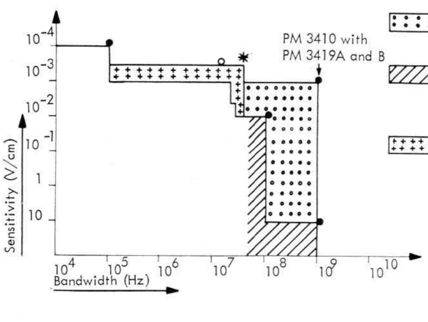
When to use a sampling oscilloscope?
As stated above a sampling oscilloscope has a much better sensitivity than a conventional oscilloscope with comparable bandwidth. This points, of course, to an important, exclusive area of use, which is dotted in the graph.

However, it is some times not recognized, that a sampling oscilloscope is often more convenient to use than a fast, conventional oscilloscope at moderate frequencies too. Let us assume, for example, that one wants to measure in pulse circuits, with rise and fall times of 5 ns. This corresponds to 70 MHz bandwidth and therefore an 80 or 100 MHz oscilloscope might be suggested. Observed rise times would in such a case be 6 or 7 ns, that is, longer than the real rise time, which has to be calculated from $T_{obs}^2 = T_{real}^2 + T_{scope}^2$. To repeat such calculations often is a nuisance. A further complication is that the amplitude of a needle pulse also appears to be lower than the real value. With a sampling oscilloscope there is no need to correct observed values as the oscilloscope limitations are far beyond the actual need.

Another point to consider is that parasitic high frequency oscillations and ringing might be so heavily attenuated that they remain undetected by a 50–100 MHz oscilloscope.

How to work with sampling!
Philips' sampling oscilloscope is simple to operate. There are only two controls, which are special for a sampling oscilloscope; smoothing and the special X deflection. Everything else is just as it is in a normal, conventional oscilloscope.

The smoothing control is a switch, by which the random noise can be decreased by a factor 2, and that is not difficult to understand. With the X deflection one can choose between different numbers of samples/cm. 100 samples/cm is the normal choice. 1000 is used when one wants to look at fine details. 10 samples/cm is used with waveforms, which have a low repetition rate, as otherwise the picture would flicker too much. When using the control for sampling density, the user must know that fine details and sharp rise times may be lost or distorted at low sampling densities and that the smoothing also has some influence on these matters. But in fact, after the operator has learned the exact nature of this interdependence from the oscilloscope manual, he can work as he used to do with his earlier, conventional oscilloscopes.



← *Application area in which there is no alternative to the use of sampling techniques*

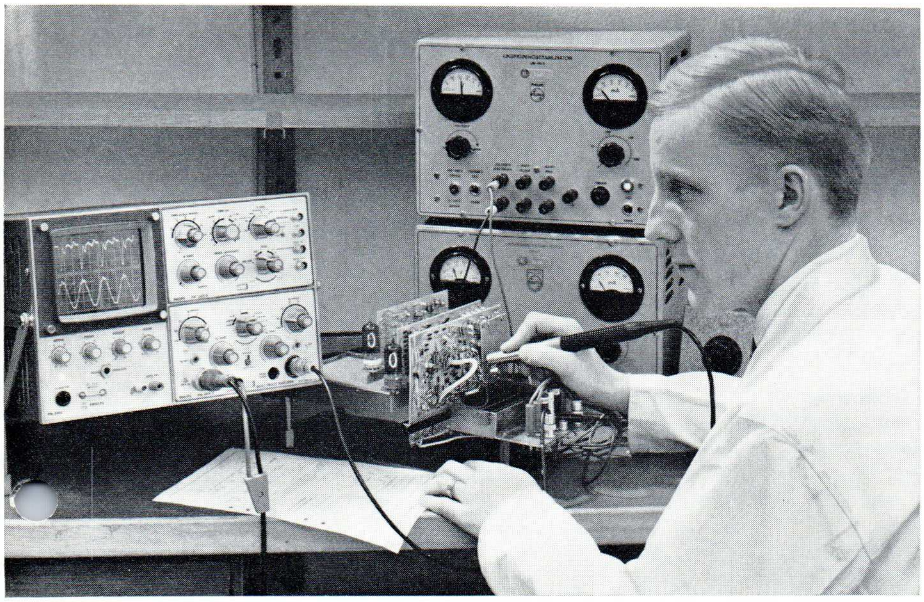
← *Application area in which the use of sampling techniques is often superior to the use of conventional techniques*

← *Application area which to date is only accessible with conventional oscilloscopes to users of the Philips PM3410 + PM3417 + 18° and PM3330 + PM3332**

Full details on the sampling oscilloscope will be given in the next issue

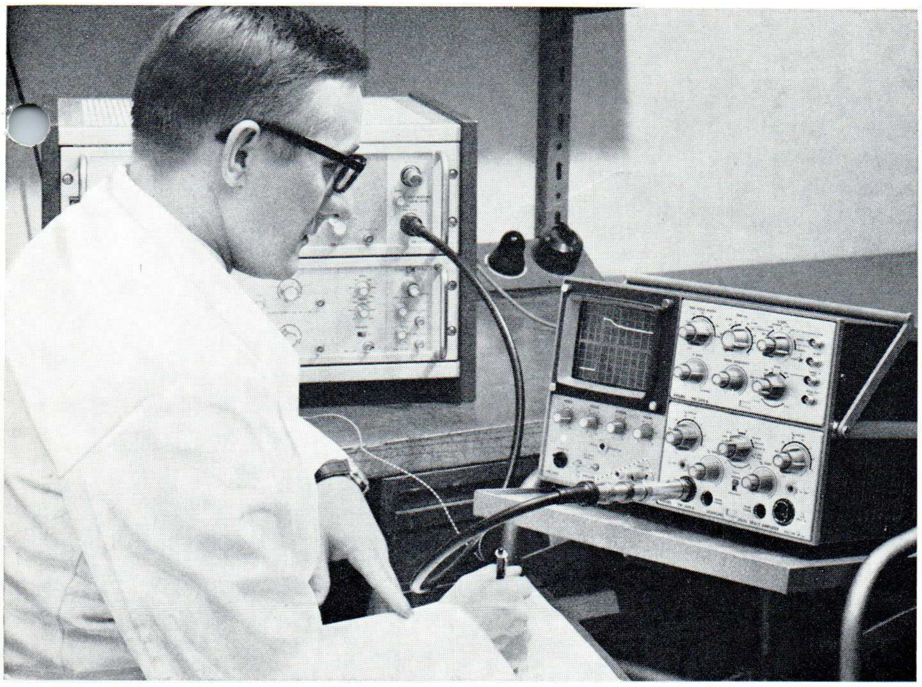
New literature

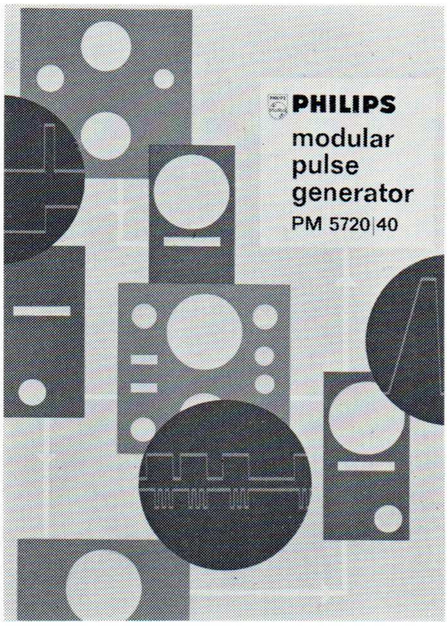
New brochure with extensive details of the Philips PM 5720-40 Modular Pulse Generator



The application picture above shows how the Philips sampling oscilloscope is applied to development work with fast digital circuits. In this case it is necessary to test a circuit, which is connected on a printed circuit board. Frequent use is then made of probes (in this case a cathode follower probe) because the points normally cannot be loaded with the 50Ω input of the oscilloscope. The use of a dual trace unit is very convenient, because one can watch the waveforms at two different points in the circuit and measure time delays between them.

The photo below shows a PM 3419 sampling oscilloscope used for the measurement of the transient behaviour of fast diodes. The diode is placed in a coaxial mounting piece, which is directly connected to the 50Ω-input of the oscilloscope. The driving signal can for instance originate from an output unit of the Philips modular pulse generator PM 5720–40.





The need for flexibility in pulse generation is increasing. Users of pulse generating equipment require more choice of repetition rate, pulse delay, pulse width, and pulse shape, as well as the ability to generate pulse bursts and pulse codes. Meeting these needs effectively and economically requires careful consideration of cost, versatility and reliability.

A new 32-page brochure gives a full description of a system optimising all these factors — the Philips PM 5720-40 Modular Pulse Generator. Unlike the typical self-contained universal pulse generator, this system consists of small, functional, interchangeable units that can be interconnected with flexible cables to produce a wide variety of pulsed signals.

In the first section of the brochure the basic principles are explained.

The second section gives extensive application data. Eleven examples of possible combinations of units are discussed, ranging from a simple set-up requiring only four units which are contained in a single 19 inch rackmounted panel or shelf, up to double- and triple shelf systems for the generation of double pulses, triple pulses, burst pulses, pulses on different channels and other complex pulse patterns.

In the last section full specifications of the individual units are given, including the recently introduced output unit type PM 5729. This unit provides an output voltage of 50 V with a rise time of less than 0.8 nanoseconds and a fall time less than 5 nanoseconds.

The brochure (reference number 80.091) is available free of charge from the Philips organization in your country.

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