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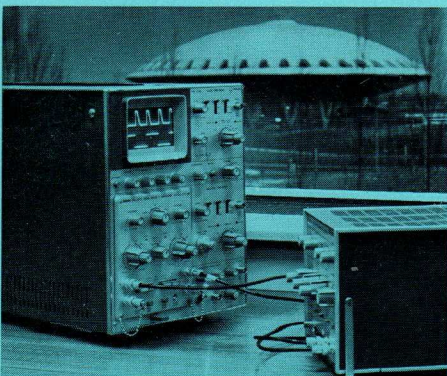
Introduction

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 needs.

The front cover

of this issue shows the PM 3370 150 MHz oscilloscope against the background of the *Evoluon*, the permanent Philips exhibition building in Eindhoven.



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Introduction

The demand for oscilloscopes of greater and greater bandwidth has been increasing steadily of recent years, in particular in connection with the need for measuring very short rise times in circuitry (especially computer circuitry) making use of the very fast semiconductors which are now available.

In this number of Electronic Measuring and Microwave Notes we present a measuring system that goes a long way towards meeting this demand. Apart from its bandwidth of 150 MHz, this system has been designed to offer a large number of measuring facilities via its various plug-in units: this makes it a truly general-purpose instrument. For example, the dual-trace plug-in unit PM 3372 offers a sensitivity of 1 mV/cm over the whole bandwidth of 150 MHz (which corresponds to a rise time of 2.4 ns); a number of different input impedances can be selected for this unit.

Further features such as display switching, the 500 μ V sensitivity at 50 MHz of the PM 3332 and the wideband spectrum analysis provided by the PM 3379 offer a wide range of applications; we are sure that the PM 3370, together with the range of plug-in units described here, will prove to be an invaluable tool for many of your measurement problems.

We have decided not to give authors' names at the head of the various articles, as both the technical design and development work, and the writing itself, have become so much a matter of teamwork that it would be invidious to pick out any single name or names. However, we would like to mention here that the team mainly responsible for making, and writing about the PM 3370 is composed of the following persons —

G. Akkermans, A. v. Beurden, F. Bregman, W. Bussing, P. v. Ensbergen, C. Gijzen, Th. Hooisma, G. Imbens, W. Joosten, J. v. Leusden, E. Polak, P. Poels, M. de Schipper, J. Schuurman, A. Sloots, W. Spapens and J. Zinger.

Summary of the specifications of the PM 3370 oscilloscope

ELECTRICAL

CRT CIRCUIT

Type of CRT:

D13—451 GH/01, rectangular with internal graticule

Effective screen area:

6 cm × 10 cm

Phosphor:

medium long (P31)

Total accelerator voltage:

15 kV

Unblanking:

DC coupled

Graticule:

internal with continuous control of illumination

External Z Modulation

5 V peak-to-peak signal required at 500 Hz to 50 MHz for visible intensity modulation

Beamfinder

push-button for reducing deflection sensitivity for locating the trace on the screen

Y CHANNEL

Final Amplifier

Type: DC amplifier

Delay: by internal delay line 80 ns, effective delay > 20 ns

Rise time:

input I: 1.7 ns (equivalent bandwidth 210 MHz)

input II: 5 ns (equivalent bandwidth 70 MHz)

Plug-in units PM 3332, PM 3334, PM 3342 and PM 3344 automatically use input II

Y Output

Coupling: AC coupled, f_{3dB} low 1 kHz

Volts out: 1 V/cm deflection of the trace; 9 V maximum

Resistance out: 200 Ω

If Y Output terminated with 50 Ω

Volts out: 200 mV/cm deflection of the trace; 1.8 V maximum

Bandwidth: 100 kHz to 50 MHz

Plug-in Unit

See the relevant specifications of the plug-in unit in use

Calibration Unit

Voltage: $R_i = 50\Omega$: 40 mV, 80 mV, 200 mV, 400 mV peak-to-peak
for $R_i = 500\Omega$: 4 V peak-to-peak

Tolerance: $\pm 1\%$

Frequency: 2 kHz square wave voltage

Current: 8 mA peak-to-peak maximum

Tolerance: $\pm 1\%$

Frequency: 2 kHz square wave current

X CHANNEL

Deflection Modes

- Main time-base
- Main time-base intensified
- Delayed time-base
- Alternate between main time-base intensified and delayed time-base
- External

Main Time-base

Sweep speeds:

50 ns/cm to 1 s/cm (1, 2, 5 sequence) in

23 calibrated steps. Continuous control between steps (uncalibrated)

Tolerance $\pm 3\%$ ($\pm 5\%$ in 1 s/cm position)

Magnification:

$10 \times (\pm 2\%)$

with maximum magnification the fastest sweep speed is 5 ns/cm

Mode:

automatic (AUTO)
triggered (TRIGG.)
single shot (SINGLE)

Triggering Source:

internal (INT.)
external (EXT.)
internal at mains frequency (MAINS)

Slope:

+ or —

Coupling:

(LF) 3 Hz to 1 MHz
(HF) 2 kHz to full bandwidth
(DC) 0 to full bandwidth

Sensitivity:

with external triggering;
200 mV (LF and HF coupling)
400 mV (DC coupling)

With internal triggering see the specification of the Y plug-in unit in use

Input Impedance of Trigger Input:

1 M Ω 15 pF

Level:

Internal: continuously adjustable over 6 cm
External: continuously adjustable over 3 V peak-to-peak for medium frequencies (range can be magnified 5 \times)

Main Time-base Intensified

In this position, the portion of the main time-base which coincides with the delayed sweep is intensified. The intensity ratio is adjustable by means of the INTENS. RATIO control.

Delayed Time-base

Sweep speeds:

50 ns/cm to 1 s/cm (1, 2, 5 sequence) in 23 calibrated steps. Continuous control between steps (uncalibrated).

Tolerance $\pm 3\%$ ($\pm 5\%$ in 1 s/cm position)

Magnification:

$10 \times (\pm 2\%)$

With maximum magnification the fastest sweep speed is 5 ns/cm

Mode:

Selected by AFTER DELAY TIME switch;
STARTS immediately after delay time,
EXT. TRIGG. after delay time,
INT. TRIGG. after delay time.

Slope:

Coupling:

Sensitivity:

Input impedance:

Level:

Delay:

adjustable up to 10 seconds. Incremental multiplier linearity, typically 0.2%

Jitter:

<0.005% of the final value of the range

Alternate between Main Time-base Intensified and Delayed Time-base

In the ALTERNATE position, the main time-base intensified and the delayed time-base are displayed alternately. The vertical (TRACE SEPARATION) and intensity ratio are continuously adjustable.

External Deflection

Amplifier:

DC amplifier

Input:

asymmetrical; BNC connector

Impedance:

1 M Ω ; 15 pF

Deflection coefficient:

10 mV/cm and 100 mV/cm switchable continuously adjustable 1:10 (not calibrated)

Bandwidth:

0 to 1 MHz

Undistorted deflection:

10 cm maximum

Output signals

Gate of main time-base and delayed time-base:

0 V to 2 V; $R_i = 1\text{ k}\Omega$; maximum output current is 2 mA

Sweep of main time-base and delayed time-base:

0 V to 8 V; $R_i = 5\text{ k}\Omega$; maximum output current is 1.6 mA

Delayed trigger pulse:

2 V; $R_i = 1\text{ k}\Omega$; maximum current output is 2 mA

Calibration voltage:

As detailed in Calibration Unit

Y Output:

As detailed in Y output

Power supply

Voltage:

adjustable for 110 V, 125 V, 145 V, 200 V, 220 V and 245 V at 46 Hz to 60 Hz. Effect of $\pm 10\%$ variations, negligible.

Power consumption:

180 W maximum

PHYSICAL

Height: 37 cm

Width: 28.5 cm

Depth: 56 cm (overall)

Weight: 25 kg

Plug-in units available

PM 3332 — 50 MHz vertical amplifier (500 $\mu\text{V}/\text{cm}$)

PM 3334 / 60 MHz vertical amplifier (10 mV/cm)

PM 3342 — 50 MHz dual-trace vertical amplifier (10 mV/cm)

PM 3344 — 50 MHz four-trace vertical amplifier (10 mV/cm)

PM 3363 — Test unit

PM 3342 — 50 MHz dual-trace vertical amplifier (1 mV/cm)

PM 3379 — Spectrum analyzer

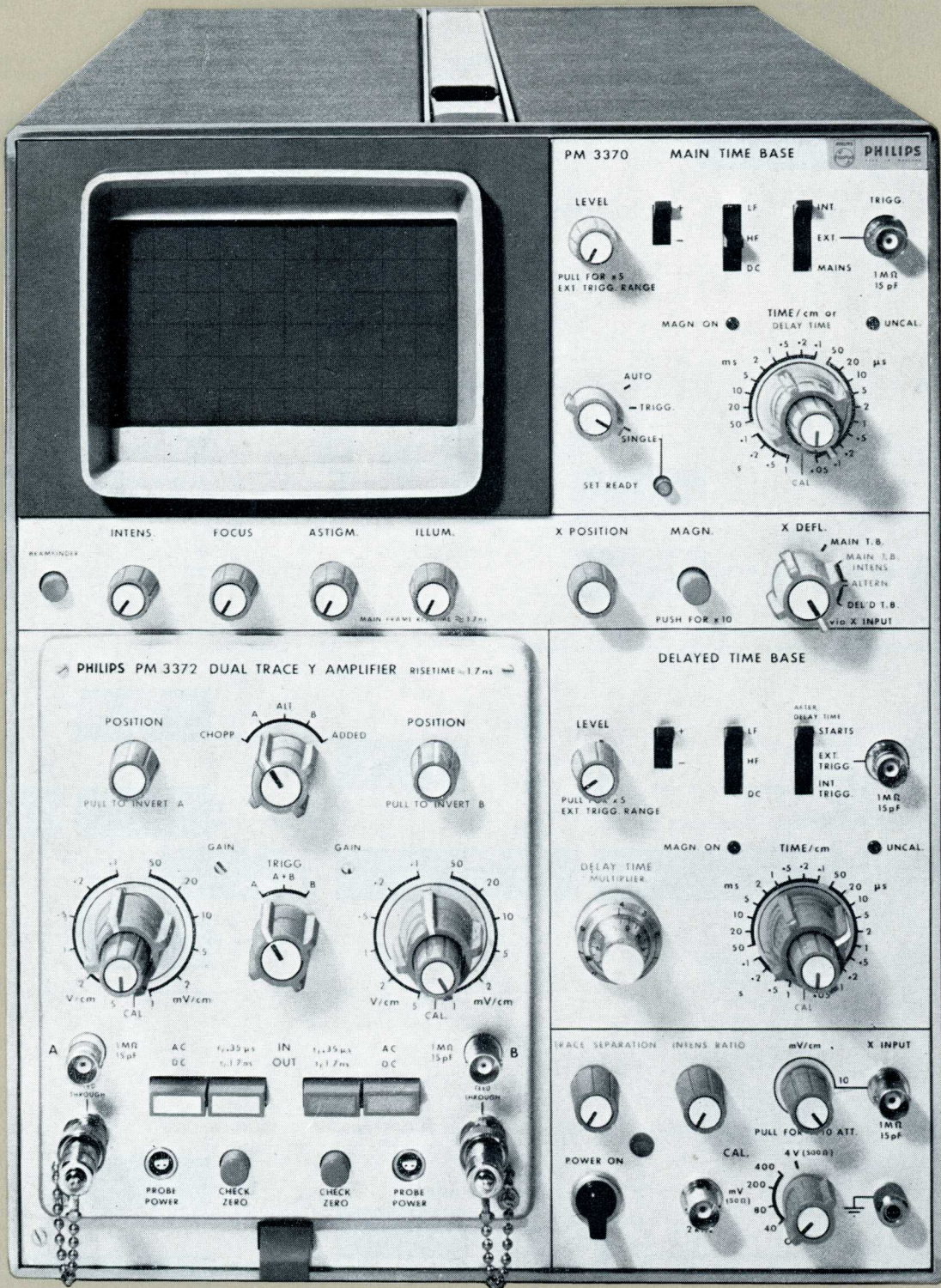
Temperature range

Operating within specification: 0 to 40°C

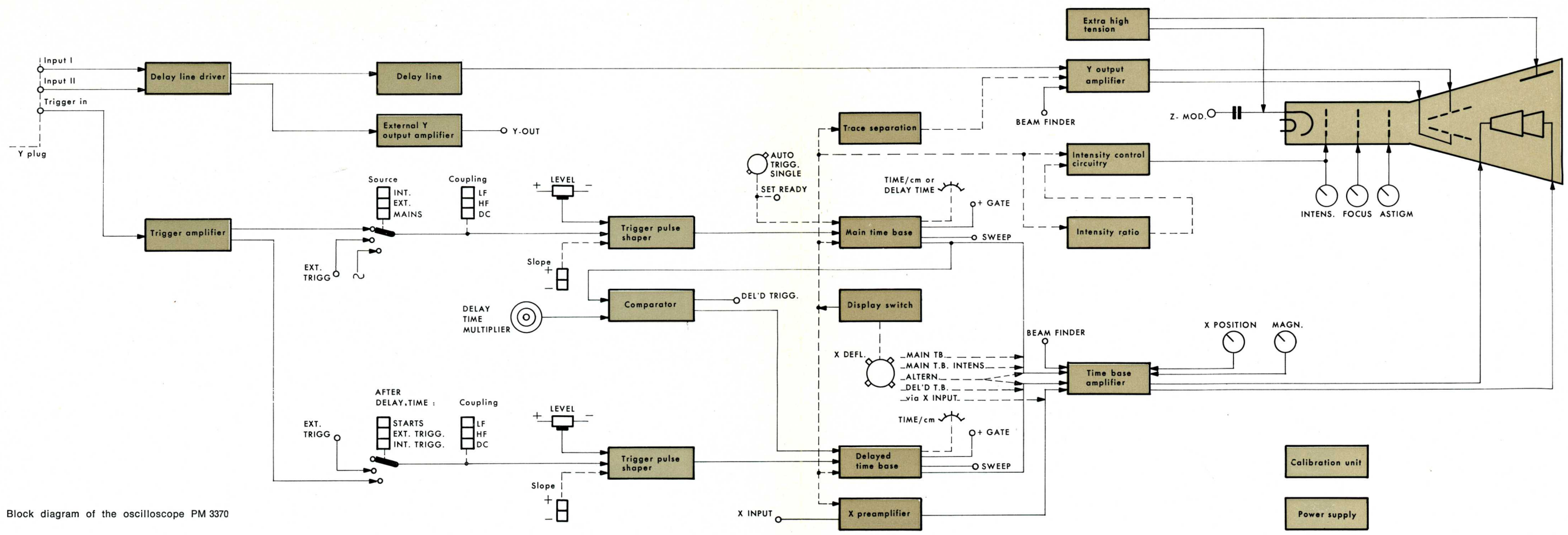
Operating: -10°C to +55°C

Storage: -40°C to +70°C

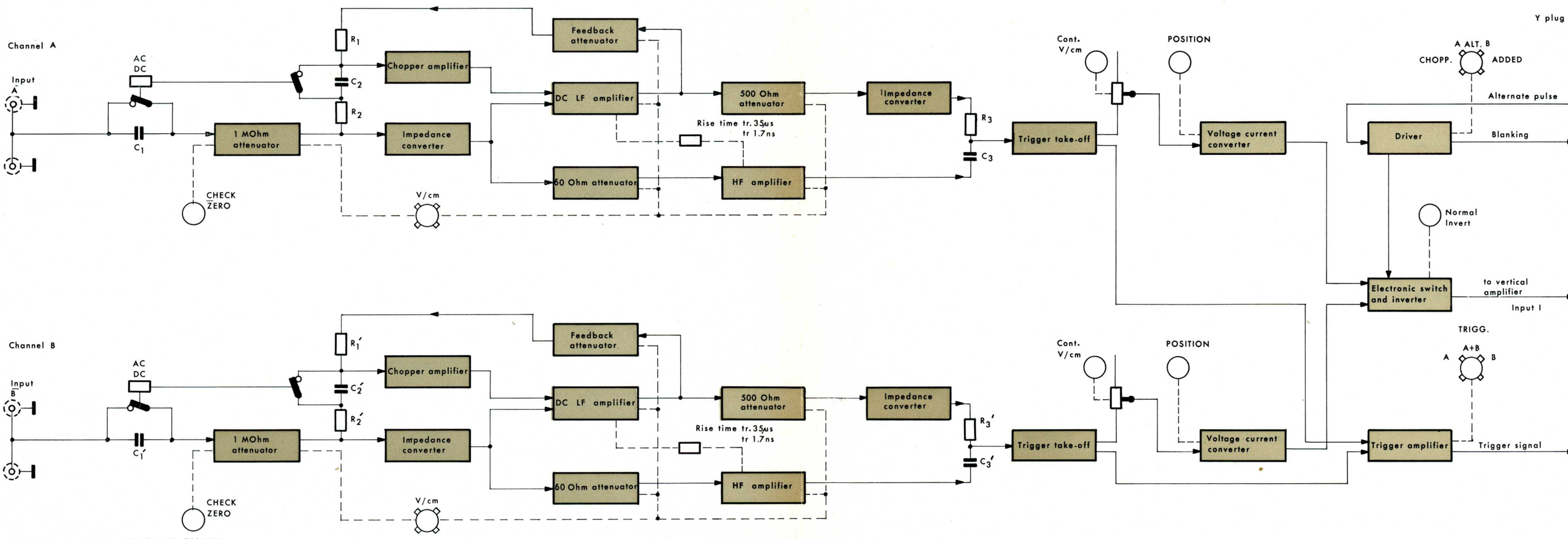
These temperature ranges are in conformity with IEC standard 68



Ergonomic arrangement of the controls. Every item is logically placed relative to its purpose and sequence of use



Block diagram of the oscilloscope PM 3370



Block diagram of the plug-in unit PM 3372

Technical description of the main frame of the PM 3370

General description

The Philips cathode-ray oscilloscope PM 3370 is a general purpose laboratory oscilloscope. A wide range of application is achieved by the utilization of plug-in units, each of which covers one or more fields of applications. For example, the dual-trace plug-in unit PM 3372 provides extreme sensitivity (deflection coefficient = 1 mV/cm) even at the maximum bandwidth of 150 MHz. The instrument takes all PM 33... Y plug-in units except the LF unit PM 3351. When the PM 3333 (the predecessor of the Y plug-in unit PM 3334) is used, the HF-probe input is inoperative. The control panel of the PM 3370 has been designed for ease of operation (see fold-out page).

The controls are functionally arranged in groups. The delayed time base controls have thus been entirely separated from the main time base controls. Important relations between controls in different areas are indicated by the use of identical colours.

The block diagram of the oscilloscope is shown on the fold-out page. The Y-signal derived from a Y plug-in unit is fed symmetrically to the delay-line driver. The latter has two inputs: a current input (I) and a voltage input (II).

The relevant input is automatically selected by the plug-in unit itself. The units (PM 3332 - 3334 - 3342 and 3344)* which were originally designed for the 60 MHz plug-in oscilloscope PM 3330 all make use of voltage input II. The intrinsic rise time for this voltage input II is 5 nsec. When one of the above mentioned units is used, the rise time is automatically set to the appropriate value. The total rise time is then

$$\sqrt{t_{\text{runit}}^2 + 5^2} \text{ ns}$$

The intrinsic rise time for the current input I is 1.7 nsec. When e.g. the dual-trace unit PM 3372 with a rise time of 1.7 nsec is used, the total rise time will be about $\sqrt{1.7^2 + 1.7^2} \approx 2.4$ ns

* These plug-in units have been redesigned and now all have field-effect input circuits.

The delay line consists of two normal 75-ohm coaxial cables, each 18 metres long, giving a delay of about 80 nsec. The cable has a matched load at both ends in order to minimize reflections.

The signal is fed via the Y output stage to the segmented Y deflection plates of the cathode-ray tube. A Y output signal is available at a socket at the rear of the apparatus. The signal is taken off before the delay line. It can be used e.g. for triggering when signals not related in time are displayed using a dual-trace unit in the "ALTERN" mode. The Y output signal is AC coupled ($f_{3\text{dB low}} \approx 1$ kHz) and amounts to 1 V per cm deflection on the screen.

The trigger signals from the Y plug-in unit are fed to the trigger amplifier, which passes them simultaneously to the trigger circuits of the main time base and the delayed time base. The main time base trigger circuit accepts these signals when the source selector is set to "INT". In position "EXT", the time base can be triggered by a signal applied to the EXT. TRIGG. socket. In position "MAINS" a signal derived from the mains transformer is used for triggering.

The frequency range of the trigger amplifiers can be reduced to remove interfering frequency components such as hum or noise from the trigger signal. This is done by means of the coupling circuits. When the corresponding selector switch is set to "LF" the range is 3 Hz — 1 MHz, in position "HF" the range is 2 kHz — full bandwidth, while in position "DC" the frequency range is not reduced (0 — full bandwidth).

In the trigger pulse shaper a DC voltage is added to the trigger pulse to shift the triggering point: the value of this voltage is adjustable by means of potentiometer "LEVEL". When the control knob of this potentiometer is pulled out, the external triggering range is increased by a factor 5. The trigger signal can be reversed in polarity by means of the slope switch



("+ —") to allow triggering by the positive-going or negative-going edge of the trigger signal.

The trigger pulse shaper converts the trigger signal into sharp positive and negative pulses. The negative pulses are used to start the time base generator, while the positive pulses can be used for the automatic free-run circuit. The saw-tooth voltage from the main time base generator is fed to the time-base amplifier in the first and second settings of the "X DEFL" switch.

AUTO mode of the main time base

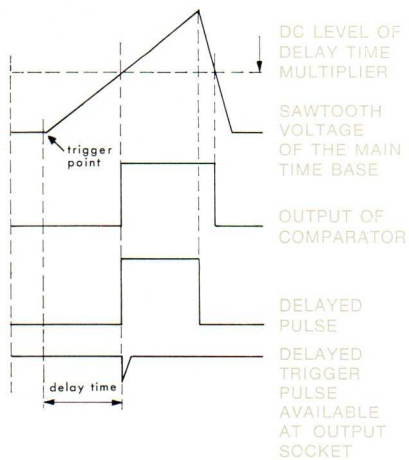
In this mode, the automatic free-run circuit ("AUTO" circuit) is made operative. The function of this circuit is to ensure that the saw-tooth generator is free-running (i.e. produces saw-tooth voltages at regular intervals in the absence of trigger pulses). In this way an image (though not necessarily a stationary one) is always produced on the screen, even though the trigger controls (source, level, slope, coupling) may not be correctly adjusted. This facilitates the search for the correct adjustment of the oscilloscope.

When trigger pulses are received operation automatically reverts to the normal triggered mode, but if trigger pulses are not received for about 0.5 sec., the time base will be free-running again.

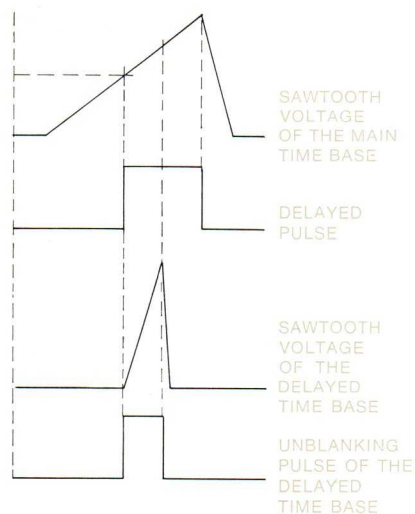
Single-sweep operation of the main time base

When events which occur once only have to be observed (and generally photographed), it is often desirable to ensure that only one saw-tooth is generated, even though several trigger pulses might be produced after the phenomenon of interest. Of course, the single saw-tooth in question must be triggered by a trigger pulse. This can be ensured by setting the mode switch of the main time base to "SINGLE" (for single sweep operation).

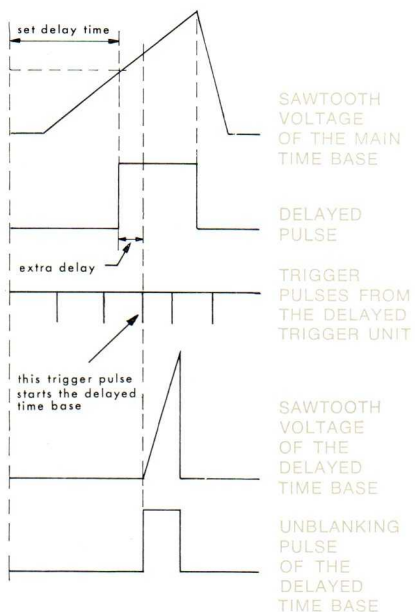
The operation of the system is then as follows:



1. Principle of obtaining the delayed pulse



2. Principle of the delayed time base



3. Principle of the gated operation of the delayed time base

A trigger pulse actuates the main time-base generator, which produces a saw-tooth voltage. Thereafter, the circuit is insensitive to further trigger pulses until it is switched back to its normal triggered mode of operation, or until the "SET READY" push-button is actuated, when the next trigger pulse will again lead to the production of one single saw-tooth.

Principle of the delayed time base

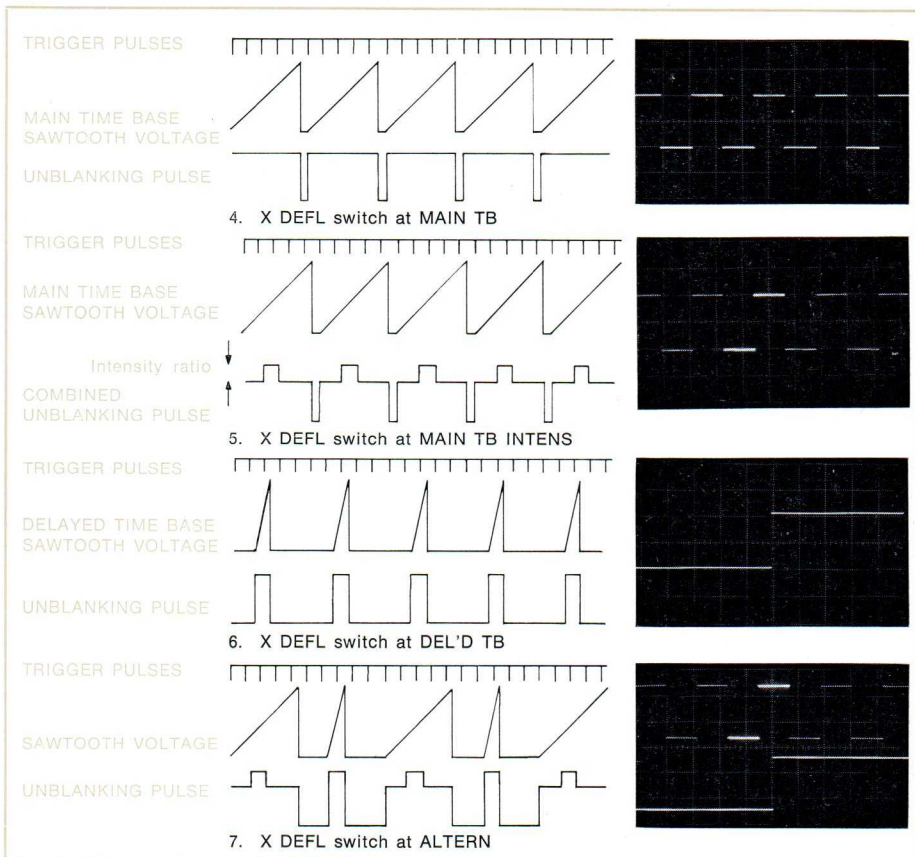
The saw-tooth voltage derived from the main time base generator is compared with an accurately adjustable DC voltage. The output voltage of the comparator is passed to a pulse shaper (reset multivibrator: see page 7) which supplies the delayed pulse. The output voltage of this pulse shaper drops to its original value at the end of the forward sweep of the main saw-tooth voltage. The relevant waveforms

The total delay is now the sum of the set delay time (the product of the values indicated by the knobs DELAY TIME and DELAY TIME MULTIPLIER) and the extra delay indicated in figure 3.

The trigger circuit for the delayed time base generator is identical with that for the main time base generator.

The "X DEFL" switch

When the "X DEFL" switch is set to "MAIN TB", the saw-tooth voltage derived from the main time base generator is fed via the time-base amplifier to the horizontal deflection plates of the CRT. In addition, the gating pulse from the main time base is applied to the control grid (Wehnelt cylinder) of the CRT via the intensity control circuit in order to intensify the trace during the sweep (see fig. 4 and the block diagram).



are shown in figure 1.

The delayed pulse produced as described above is fed to the delayed time base generator. The output voltages from this generator, i.e. the saw-tooth voltage and the unblanking pulse, are fed to the CRT. The waveforms showing the operation of this part of the system are given in fig. 2.

Gated operation of the delayed time base

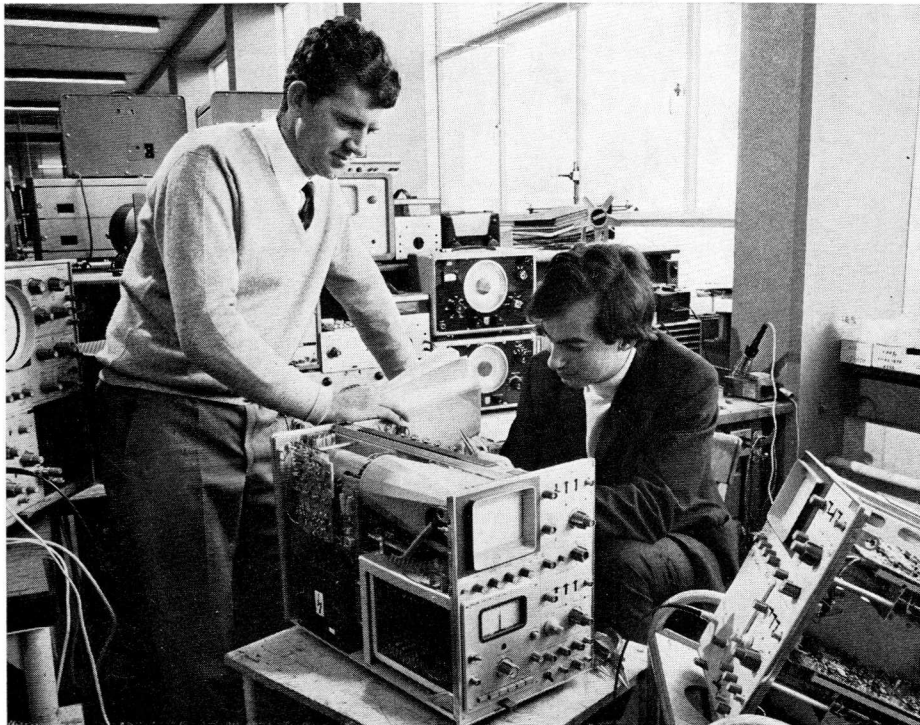
When the AFTER DELAY TIME switch is set to "INT. TRIGG." or "EXT. TRIGG.", the delayed trigger pulse prepares the time base for the normal triggered mode of operation. The next trigger pulse to arrive after the set delay time actuates the delayed time base, which is then locked to this trigger signal. The waveforms showing the mechanism of this gated operation are given in fig. 3.

When the "X DEFL" switch is set to "MAIN TB INTENS", the sawtooth voltage derived from the main time base generator is again fed to the CRT via the time-base amplifier. In this case, the gating pulses from the main time base and the delayed time base are combined and applied to the control grid of the CRT. During the operation of the delayed time base generator the trace undergoes extra intensification. The start of the intensified portion can be shifted with the aid of a ten-turn potentiometer ("DELAY TIME MULTIPLIER"). The brightness of the non-intensified part of the trace can be varied with the aid of the knob "INTENS. RATIO", (see fig. 5 and the block diagram).

When the X DEFL. switch is set to "DEL'D



PM 3370 and PM 3372 in use with a computer-controlled set-up for tests and measurements on silicon wafers for integrated circuits



Final check on the PM 3370

TB", the saw-tooth voltage from the delayed time base generator is fed to the deflection plates of the CRT and the gating pulse from the delayed time base to the control grid of the CRT. As a result, the intensified portion of the display produced by the previous setting is now expanded to fill the entire screen, (see fig. 6 and the block diagram).

When the X DEFL. switch is set to "ALTERN", an electronic switch ensures that the display of fig. 5 and the display of fig. 6 are alternately traced on the screen. The two displays can be separated by varying the voltage applied to the vertical amplifier, and derived from the driving circuit of the electronic switch. This separation is symmetrically variable with the aid of the knob "TRACE SEPARATION", (see fig. 7 and the block diagram).

When the X DEFL switch is set to "via X INPUT" an external voltage applied to the X pre-amplifier is fed via the time-base amplifier to the horizontal deflection plates of the CRT. The deflection coefficient is 10 mV/cm and the bandwidth is from DC to 1 MHz. A 1 to 10 continuous control is incorporated in the X pre-amplifier. When the mV/cm switch is pulled out, the deflection coefficient is stepped up by a factor of ten. The total range is thus from 10 mV/cm to 1 V/cm.

This facility can be used for X-Y applications, where phase relations between the X and Y deflections above frequencies of 50 kHz are not important. This restriction is due to the presence of the delay line in the Y amplifier.

Time-base amplifier

The main purpose of the time-base amplifier is to amplify the saw-tooth voltages. In order to improve linearity, the circuit is designed so that the currents through the output transistors increase during the linear part of the sweep. The gain of the amplifier can be increased 10 times, resulting in a maximum sweep rate of 5 nsec/cm. For a detailed description of this circuit see page 19.

Intensity control circuit

The unblanking pulses from the two time-base generators are fed via the intensity control circuit to the control grid of the CRT. DC coupling of this signal is realized by means of a modulator and a detector including a DC restorer. For a detailed description of this circuit see page 21.

Calibration unit

This unit supplies the voltages for the calibration of the deflection coefficients and for probe adjustments.

Voltages of 400 - 200 - 80 and 40 mV are available with an internal resistance of 50 Ω and a voltage of 4 V with an internal resistance of 500 Ω . The frequency of the square-wave delivered amounts to 2 kHz.

By short-circuiting the output terminal, calibrated currents can be made available for current probes. For a detailed description see page 27.

Power supplies

The power supplies, which are electronically stabilized, are overload- and short-circuit-proof and fitted with automatic reset.

For a detailed description see page 23.

EHT power supply

The EHT voltage is obtained by transforming and rectifying the output voltage of a 40 kHz push-pull oscillator. This power supply delivers:

+ 13000 V with respect to earth for post-deflection acceleration. The rectifying circuit comprises a voltage tripler.

The supply voltage for the cathode of the CRT (—2000 V with respect to earth).

MAIN TIME BASE

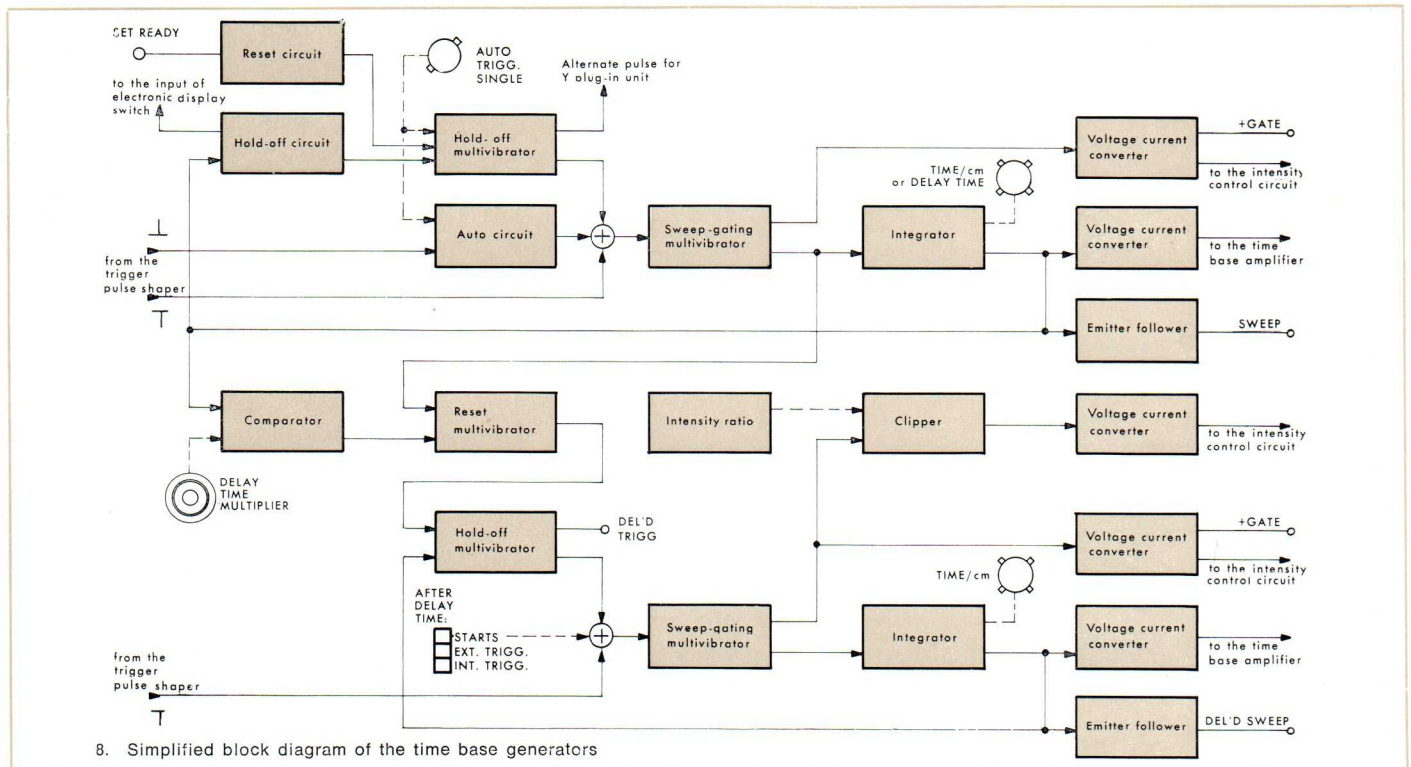
The time base signal is timed by the sweep gating multivibrator (SGM). The input signal of the SGM consists of the sum of three signals: the negative trigger pulse, the positive hold-off pulse and a DC voltage from the AUTO circuit.

If the SGM is triggered by the trigger pulse, a voltage step is applied to the integrator. As a result, a voltage increasing linearly with time is produced at the output of the integrator. This voltage is passed through the hold-off circuit to the hold-off multivibrator (a Schmitt trigger), which switches over at a predetermined level, initiating the hold-off pulse. This pulse resets the SGM to its original state, causing the output voltage of the integrator to return relatively quickly to its original value (flyback). The hold-off multivibrator keeps the SGM insensitive to trigger pulses until the flyback is completed;

position "SINGLE" of the mode switch the input DC level of the hold-off circuit is such that the hold-off multivibrator is switched to the hold-off position. When the push-button "SET READY" is actuated the hold-off multivibrator is switched back, thus making the SGM sensitive to trigger pulses again. A pilot lamp indicates the state of the hold-off multivibrator.

DELAY CIRCUIT

The DELAY TIME MULTIPLIER supplies an adjustable DC voltage to the comparator, to serve as a reference level with which the saw-tooth voltage of the main time base is compared. As soon as the saw-tooth voltage exceeds this level, the comparator will deliver a square-wave signal, which lasts until the flyback. This signal is applied via the reset multivibrator to the hold-off multivibrator of the delayed time base to activate the delayed sweep.



8. Simplified block diagram of the time base generators

This circuit is also overload- and short-circuit-proof. For a detailed description see page 25.

Beam finder

The oscilloscope is also fitted with a beam finder. When the BEAM FINDER button is depressed the amplification factors of the Y and time-base amplifiers are reduced so that the electron beam is not deflected beyond the display area. It is then a simple matter to bring the image to the centre of the screen by operating the position controls.

Over-all operation of the time base generators

The operation of the time-base generators will now be discussed in more detail, with reference to the block diagram of fig. 8. For the description of the individual units involved see pages 13 to 18.

the output voltage of the hold-off circuit is then decreased to such a level that the hold-off multivibrator is reset so that the SGM can be triggered again.

If the mode switch is set to "AUTO", positive trigger pulses from the pulse shaper give the AUTO circuit a positive output voltage. A voltage of this polarity is blocked by a diode and does not affect the normal functioning of the time-base generator, described above, so that normal triggered operation is obtained.

When trigger pulses cease to arrive at the AUTO circuit, it starts lowering its output voltage. At a certain moment the diode starts to conduct and the SGM input voltage will become so low that the saw-tooth generator will start to be free-running.

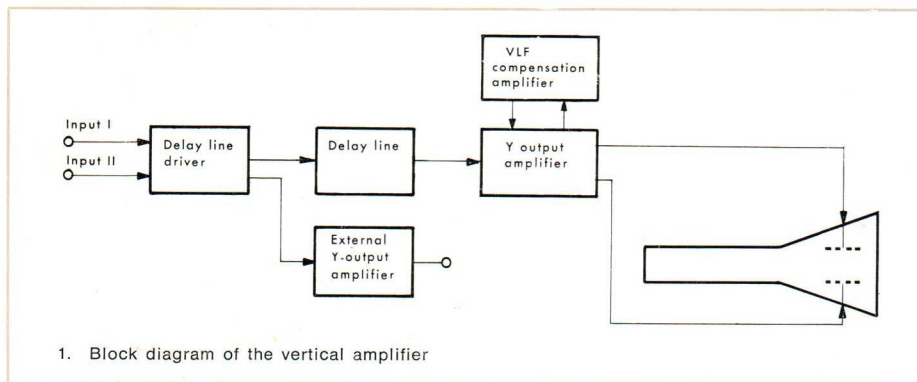
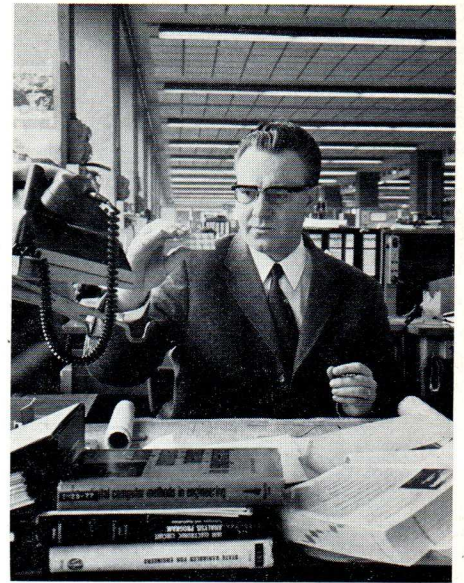
If the mode switch is set to "TRIGG", the output of the AUTO circuit is uncoupled from the SGM and the main time base works in its normal triggered mode. In

DELAYED TIME BASE

The delayed time base is similar to the main time base, but lacks the AUTO circuit and the hold-off circuit.

Its operation is the same as described for the main time base in its single sweep mode, except that the hold-off multivibrator is not operated by the "SET READY" push button but by pulses from the reset multivibrator. Depending on the DC level applied to the input of the SGM, as dictated by the position of the "AFTER DELAY TIME" switch, the delayed time base starts immediately on receipt of a pulse from the reset multivibrator (with the AFTER DELAY TIME switch set to "STARTS"), or is set ready to start on receipt of the next internal trigger pulse (setting "INT") or the next external trigger pulse (setting "EXT").

The vertical amplifier



The vertical amplifier in the main frame is composed of the following parts:

- a) delay line driver
- b) two-stage Y output amplifier
- c) very low frequency compensation amplifier
- d) external Y-output amplifier

The block diagram of the vertical amplifier is shown in fig. 1.

Delay line driver

The delay-line driver has a current input (I) and a voltage input (II). The relevant input is automatically selected by the plug-in itself. The diodes that effect this choice are indicated in the accompanying simplified circuit diagram of the delay-line driver (fig. 2). Diodes D_1 and D_2 are cut off when the voltage input is used. In this case the bases of emitter-followers TS_5 and TS_{10} are driven by the collector-signal from TS_3 and TS_8 via R_1 and R_2 . If now a unit is plugged in which makes use of the current input, collector-resistors R_1 and R_2 are connected via D_1 and D_2 to +24 V so that sequentially the collectors of TS_1 and TS_2 are effectively short-circuited by the conducting diodes. A connection on the plug of those plug-in units which make use of the current input provides the +24 V.

The rise time of the voltage input is restricted by means of simple low-pass RC networks R_6 , R_3 , C_1 , C_2 and R_4 , R_5 , C_3 , C_4 ,

to accommodate the plug-in units which were originally designed for the 60 MHz PM 3330 oscilloscope. These networks also provide correct matching of the DC level. The input sensitivity can be adjusted by means of R_{10} . The DC balance of this part of the amplifier is effected by R_9 . A current-driven input is used, to overcome the difficulties of stray impedances which would have been found voltage transfer. The connections in the plug next to the signal input are earthed, so that this plug can be treated as a short piece of transmission line. Matching of the load-end is obtained by a resistor in series with the input impedance of a grounded base (R_7 , TS_4 and R_8 , TS_9). This configuration guarantees a good signal transfer from broad-banded plug-in.

A bridged-T-filter is inserted in the collector of the grounded base transistor to compensate for the input capacitance of the emitter-follower. (TS_5 and TS_9). The delay line is terminated with a matched impedance at both ends and as a result no reflections are visible on the CRT. A coaxial cable is used for the delay line. HF line losses are partly compensated by C_5 and C_6 .

Two stage Y output amplifier

The amplifier following the delay line (fig. 3) has to satisfy several demands. It must be able to provide large voltage swings to

drive the CRT, wide bandwidth and good overdrive. Careful choice of the transistors took care of the bandwidth and CRT drive. Good overdrive properties were obtained by avoiding saturation at high signal levels. The first stage of the deflection amplifier provides the following facilities:

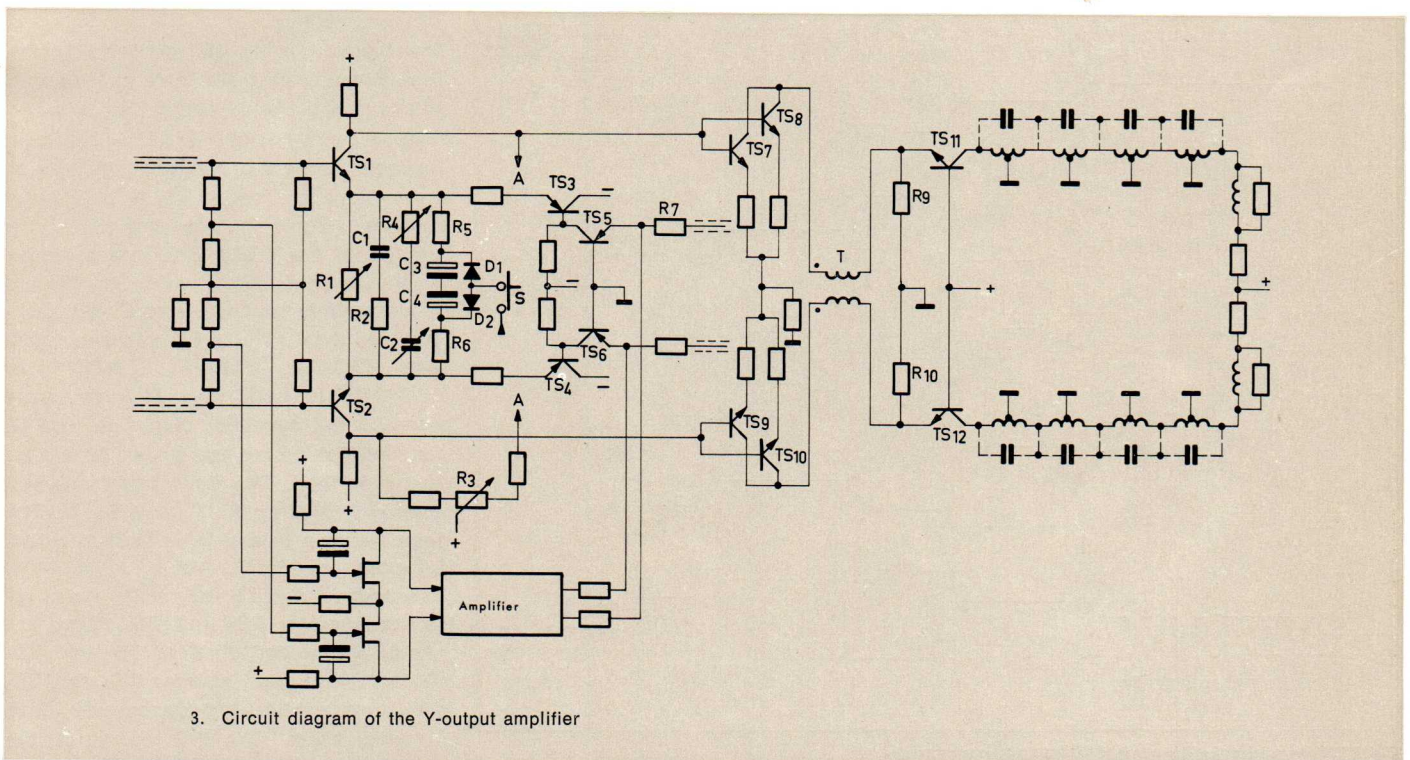
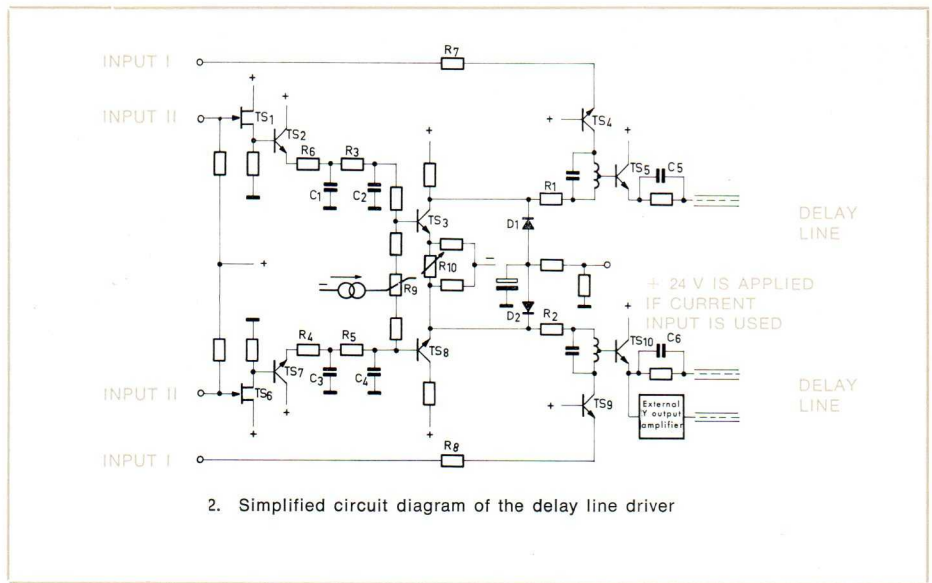
1. Compensation of the skin-effect losses of the delay line R_2 , C_1
2. Gain adjustment R_1
3. LF compensation $R_{5,6}$, $C_{3,4}$
4. DC balance R_3
5. Beamfinding D_1 , D_2 , S
6. Trace separation
7. Compensation HF roll off R_4 , C_2
8. Very low frequency compensation

Trace separation is obtained by applying square-wave signals, derived from the display switching circuit to the emitters of TS_5 and TS_6 via R_7 and R_8 respectively. The ultimate effect of these signals is to displace vertically the traces obtained in the ALT. position of the X-deflection switch. A more detailed description of this feature is given on page 46.

The final stage of the output amplifier is a pair of balanced cascodes. Two transistors in parallel with individual series feedback constitute the lower part of the cascodes. R_9 and R_{10} ensure that TS_{11} and TS_{12} are conducting when the lower transistors are cut of. This also helps to give a good large signal response.

Transformer T improves the HF symmetry of the amplifier. The CRT is equipped with a vertical deflection system, consisting of four pairs of segmented deflection plates. This construction limits transit-time distortion and avoids severe capacitive loading of the Y output amplifier stage. The characteristic impedance of the filter connecting the deflection plates was chosen with a view to the following factors:

- a) signal velocity as close as possible to the electron velocity of the beam.
- b) current-drive capabilities of the final amplifier stage
- c) rise time



200 Ohm was found to give a good compromise between the above mentioned conflicting factors. A bridged-T filter was used as filter cell. The filter is terminated by a network which keeps reflections to a minimum.

VLF correction amplifier

Relatively large changes in dissipation occur when the display is shifted or with large LF signals. The consequent changes in the temperature of the transistor crystals cause a change in the emitter-base potential of the transistor, which acts like an undesired additional drive. To compensate for this effect, a sample is taken from the voltage dividers across the delay-line termination and integrated by a balanced FET Miller integrator. The need for a Miller integrator arose because of the long time constant involved (approx. 30 sec.). The integrated signal is then amplified and

applied to the emitters of TS₅ and TS₆. The resulting effect in the Y output amplifier counteracts the undesired emitter-base voltage changes.

External Y-output amplifier

It is useful to have available at the output a signal which has passed the electronic switch in the plug-in unit (see page 47), for various purposes. For this reason a signal is picked off from the emitter follower TS₁₀ and successively amplified. This signal is available at an output at the rear side of the oscilloscope.

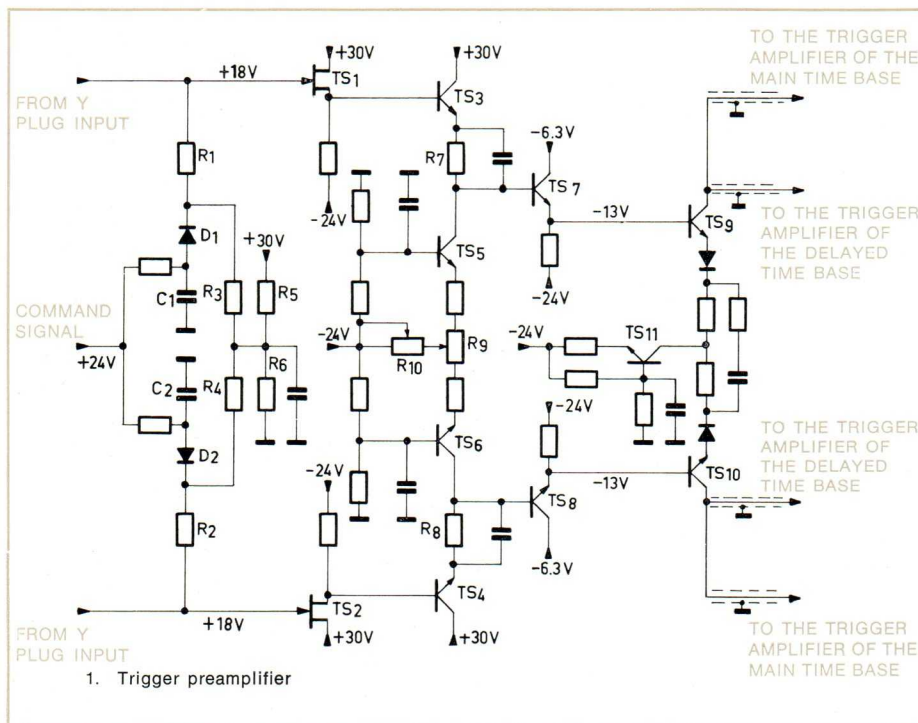
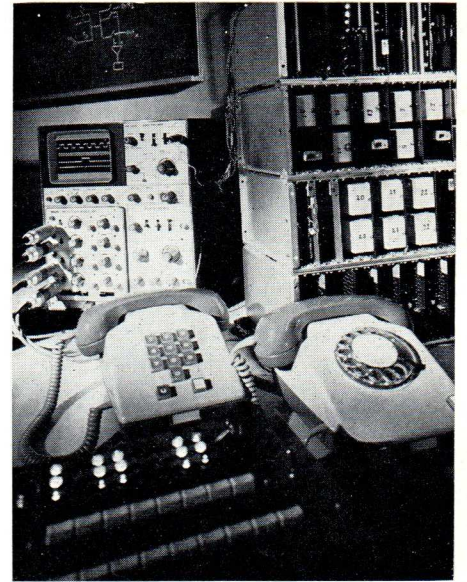
A diode circuit is employed to safeguard the output circuitry of this amplifier against undesired high-voltage inputs.

The output of this amplifier may be used to drive high impedance loads circuitry and can also be used with 50 Ohm loads; in the latter case, the bandwidth is much greater.



Final check on the PM 3372

Trigger circuitry



Trigger amplifier

The trigger signals are provided by the Y plug-in unit at a DC level of +18 V by means of a push-pull circuit (fig. 1).

When e.g. plug-in unit PM 3372 is in use, the input network of the amplifier acts as a low-ohmic termination ($\approx 150 \Omega$). A +24 V command signal is then applied to one terminal of the Y plug and makes diodes D₁ and D₂ conducting. In this way R₁ and R₂ are earthed via capacitors C₁ and C₂.

When plug-in units with voltage output (e.g. PM 3332, PM 3342, PM 3344 or PM 3334) are used, the 24 V command signal is not supplied; diodes D₁ and D₂ are then cut off, so that R₁ and R₂ are no longer earthed. The input resistance will now be formed by R₁, R₃ and R₂, R₄. The point between R₃ and R₄ is held at +18 V by voltage divider R₅—R₆.

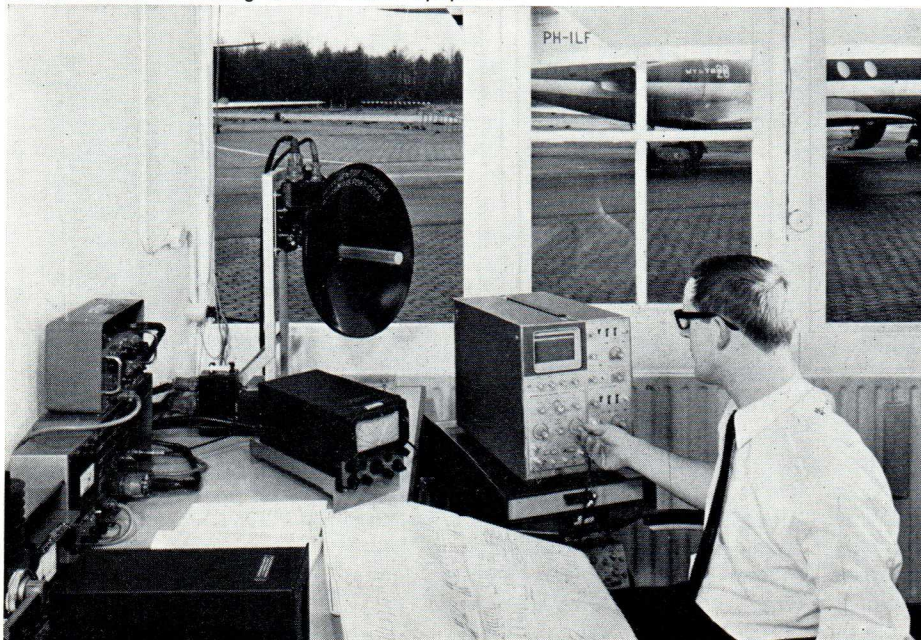
The trigger signals enter the gates of source-follower TS₁ and TS₂. They are passed to emitter-followers TS₃ and TS₄ which have current sources TS₅ and TS₆ acting as emitter impedance with high dynamic value. The DC levels of the trigger signals are shifted without attenuation, by means of R₇ and R₈.

The trigger signals now pass through emitter-followers TS₇ and TS₈, the output DC levels of which can be balanced by adjusting R₉. The voltage level can be adjusted simultaneously by means of R₁₀. The final stage consists of amplifiers TS₉ and TS₁₀, with TS₁₁ as a common current source.

This stage is in a cascode arrangement together with the input stages of the main time-base and delayed time-base trigger amplifiers, (see fig. 2).

Coaxial cables, which transmit the trigger signals as current variations, are used to bridge the physical distance between the various units. These coaxial cables are terminated by their characteristic impedance to minimize reflections. This impedance is formed by R₁₁, R₁₂ and the low-ohmic input impedance of the common-base transistors TS₁₂ and TS₁₃, (fig. 2).

PM 3370 used for testing aircraft radar equipment



The trigger signals are developed as voltage variations across the collector resistors $R_{13/14}$ and are fed to the trigger source switch via cross-coupled emitter-followers $TS_{14/15}$. The output DC levels can be adjusted by means of $R_{15/16}$.

The main time base can be triggered by the trigger signals from the Y plug-in unit if the source switch is set to "INT. TRIGG" (see fig. 3). In position "EXT. TRIGG", triggering is effected by a signal applied to the external input terminal, while in position "MAINS" the triggering signal comes from the mains transformer, via a low-pass filter to suppress high-frequency components.

The delayed time base can be started immediately after completion of the selected delay time, if the AFTER DELAY TIME switch is set to "STARTS"; the input of the trigger amplifier is then earthed.

In position "EXT. TRIGG." the input of the trigger amplifier is connected to the input terminal for an external trigger signal; in position "INT. TRIGG" the trigger signals are obtained from the Y plug-in unit.

After the source switch comes the coupling switch (fig. 3). In position "LF" a low pass-band of 3 Hz — 1 MHz is obtained with R_{17} , C_3 and R_{18} . A high-pass filter with an f_{3dB} low of 2 kHz is inserted in the "HF" position. In position "DC", all signals pass, from DC to 150 MHz.

Having passed the coupling switch, the trigger signals enter a difference amplifier of the trigger amplifier via source-followers TS_{16} and TS_{17} and emitter-followers TS_{18} and TS_{19} (fig. 4).

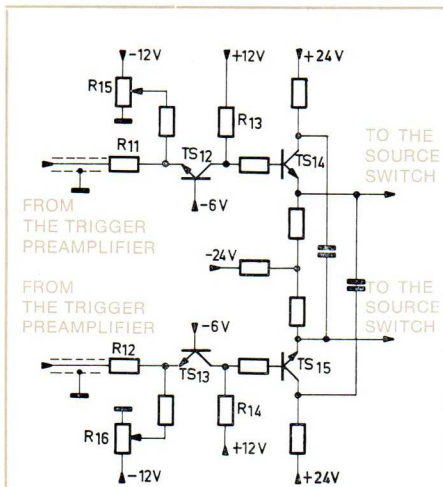
The difference amplifier consists of the transistors TS_{20} and TS_{21} with series feedback. TS_{22} acts as a current source. In conjunction with transistor TS_{23} (with shunt

feed-back), this difference amplifier amplifies the trigger signal and the DC voltage of the level potentiometer R_{19} . The polarity is selected with the aid of the slope switch ("+"—"). Potentiometer R_{20} serves to adjust the middle of the level control range, to compensate mainly for differences in gate-source voltages of the input FET's.

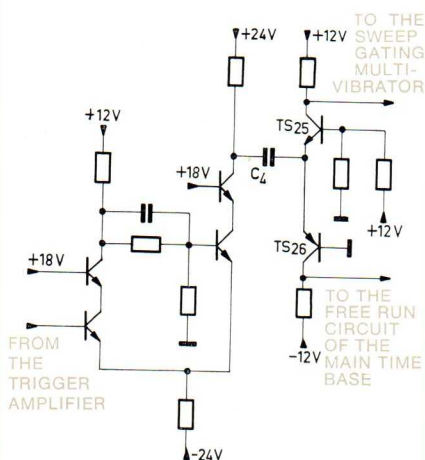
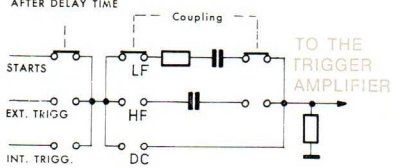
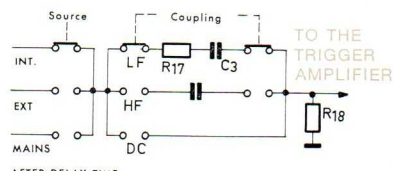
Emitter-follower TS_{24} passes the trigger signals to the trigger-pulse shaper.

Trigger-pulse shaper

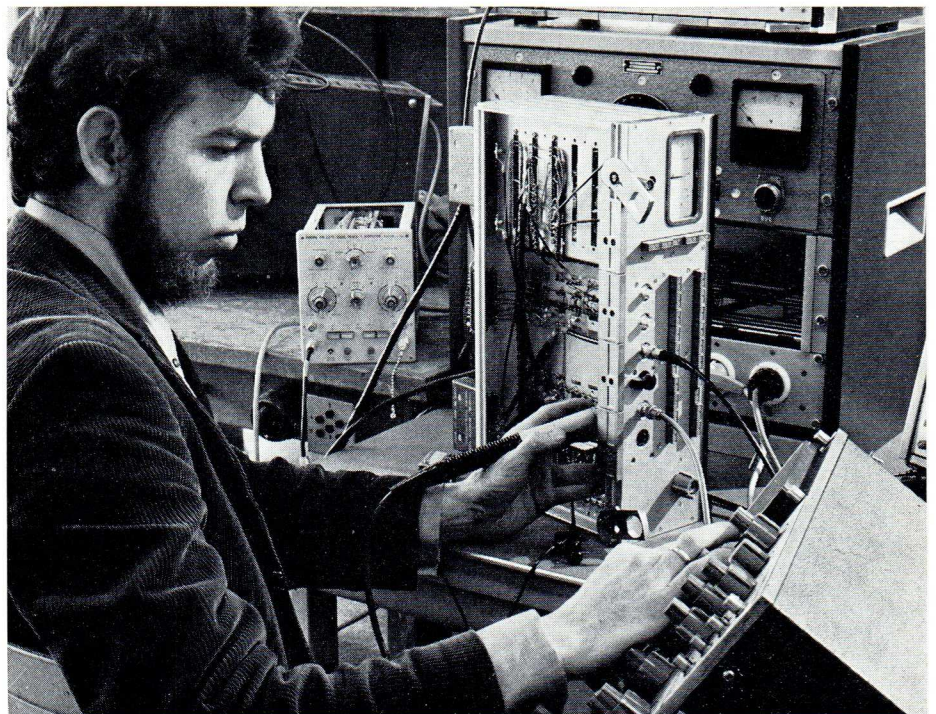
The trigger-pulse shaper contains a Schmitt trigger, which is designed with cascode circuits to obtain fast transients during switching. The output of the Schmitt trigger is converted into short pulses by differentiation by C_4 in conjunction with the low-input-impedance common-base transistors TS_{25} and TS_{26} (fig. 5).



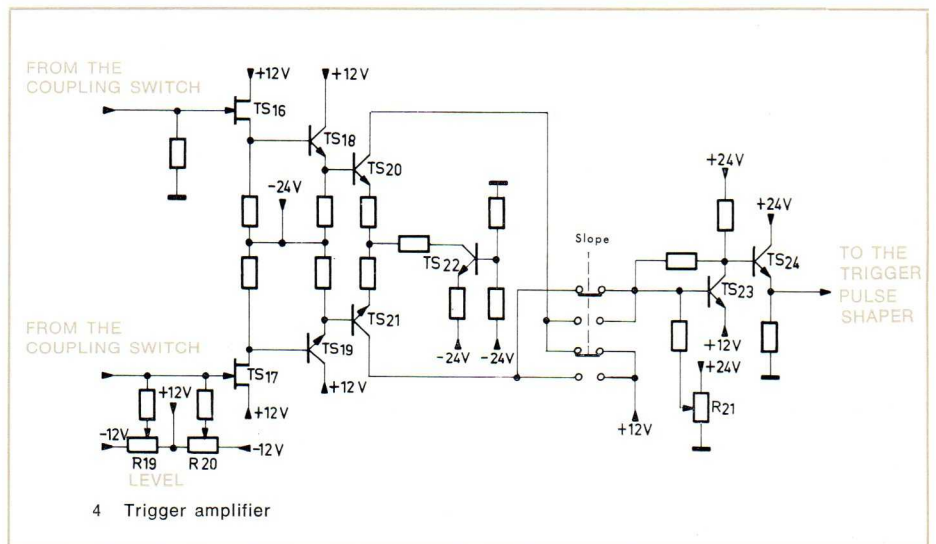
2. Input stage of the trigger amplifiers
3. Source and coupling switches



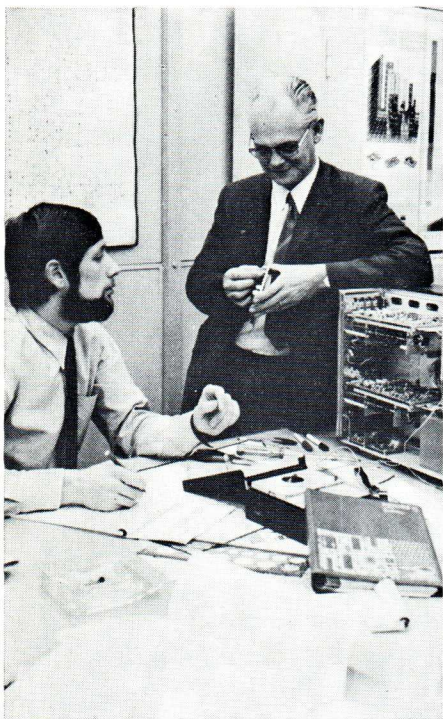
5. Trigger pulse shaper



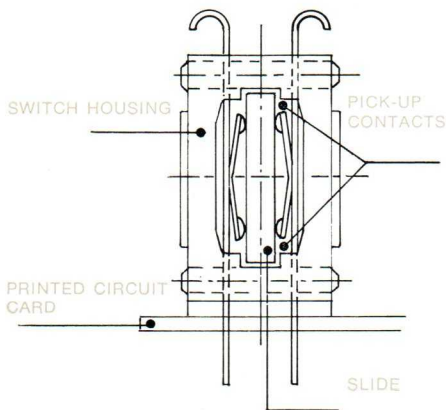
Calibration of test equipment for the PM 3372



4. Trigger amplifier



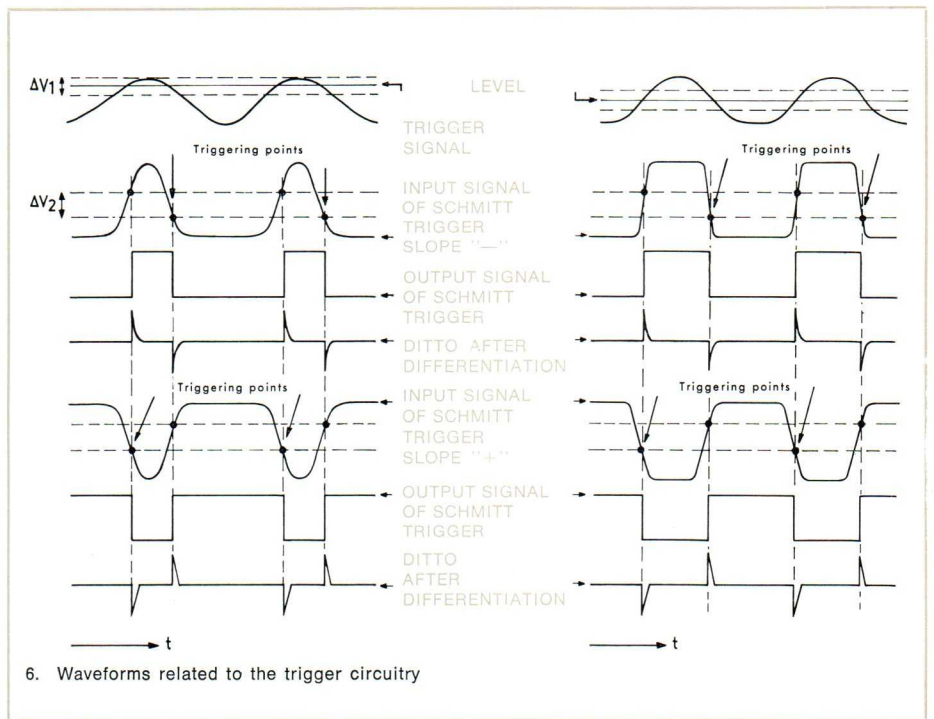
Preparing service manual of the PM 3370.



7. Cross-section of a printed-circuit switch



8. Slide of a printed-circuit switch



6. Waveforms related to the trigger circuitry

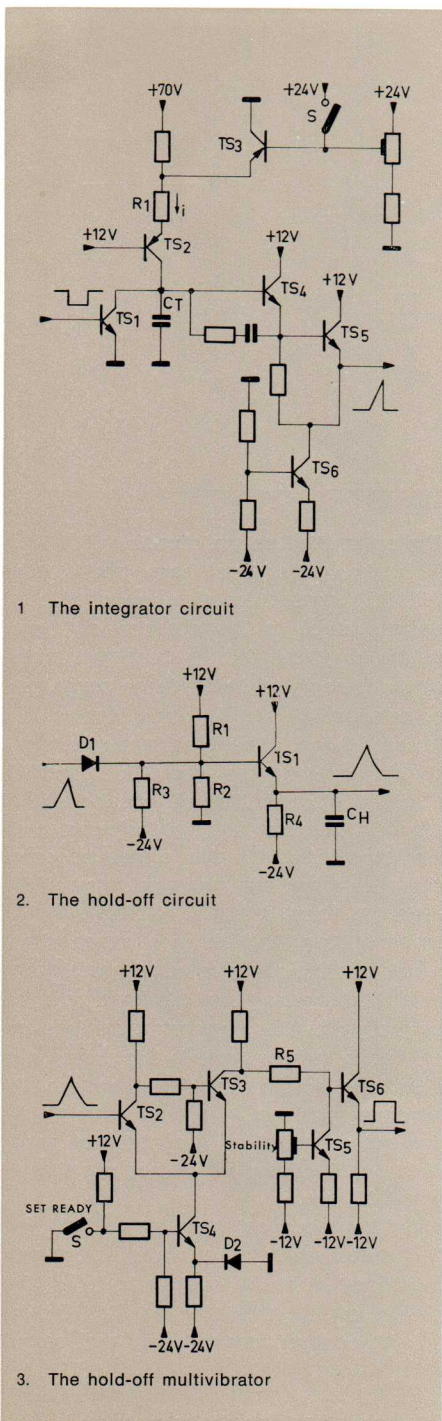
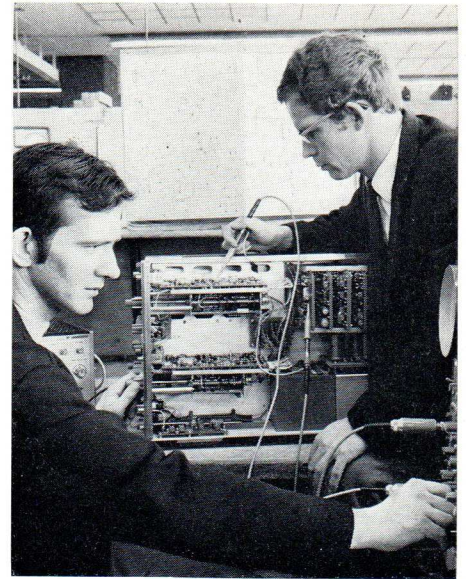
The operator of the whole system is illustrated in fig. 6. ΔV_1 represents the dynamic range of the difference amplifier. Input signals falling within this range give rise to amplifier output signals.

When these signals overlap the hysteresis gap ΔV_2 of the Schmitt trigger, square-wave signals are produced at the output of the latter; after differentiation, these square-wave signals give short positive and negative pulses. The negative pulses are used for the triggering of the time bases and the positive ones for the control of the "AUTO" circuit of the main time base. As may be seen from the figure, when the slope switch is in its "+" position the time-base generator is triggered by the positive-going edge of the original trigger signal and when the slope switch is in its "-" position by the negative-going edge. The precise point on the edge in question at which the time base is triggered can be adjusted with the aid of the "LEVEL" control.

Switch design

The source, coupling and slope switches are specially designed and mounted on printed-circuit cards, so as to obviate long leads in the HF circuit. The switches consist of a slide with two rows of gold-plated contacts on each side, inter-connected in different ways by printed conductors so as to realize the desired switching combinations. The pick-up contacts, which are also gold-plated, press against the contacts on the slide (fig. 7). Fig. 8 shows one half of a slide with the contacts.

Time base circuitry



THE MAIN TIME BASE

The integrator

The integrator contains a capacitor, a charging circuit and a discharging circuit. The desired saw-tooth voltage is obtained by charging the capacitor with a current which is held constant within narrow limits, and then discharging it rapidly, see fig. 1. The constant current is obtained with the aid of a transistor (TS₂) in the common-base configuration, with a large emitter resistance (R₁). When the switch S is closed (CAL setting), the voltage across R₁ is determined by the supply voltages and the base-emitter voltages of TS₂ and TS₃. As a result, the charging current will be known and constant within very narrow limits, and the voltage across C_T will increase linearly with time.

The rate of increase of this voltage depends on the capacitance of C_T (which can be varied step-wise with the aid of the knob "TIME/cm") and the magnitude of the charging current. This current in its turn is determined by the resistance R_T (which is also varied step-wise with the aid of the "TIME/cm" knob) and the voltage across this resistor (which can be varied continuously with the aid of the central knob of the "TIME/cm" controls). Transistor TS₃ acts as an emitter-follower.

The moments at which the forward sweep of the saw-tooth voltage starts and stops are determined by the times at which the switching transistor TS₁ becomes cut off and conducting respectively. This transistor is controlled by the sweep gating multivibrator (SGM). When TS₁ is cut off, it does not form a load for the capacitor, while when TS₁ is conducting the capacitor will be discharged rapidly until TS₁ becomes bottomed (when the capacitor voltage will be constant). In this situation, TS₁ draws only the constant current supplied by TS₂.

The saw-tooth voltage is taken from the capacitor via a double emitter-follower (Darlington pair), TS₄ and TS₅. In order to ensure good linearity, the base current of TS₄ should be kept low and constant. In

order to ensure that the latter condition is fulfilled, TS₄ and TS₅ are fed via TS₆, connected as a current source. The low current setting of TS₄ means that the linearity of the saw-tooth voltage will be poorer at high sweep velocities, as the transition frequency f_T decreases at low currents. In order to compensate for this, an RC filter is shunted across TS₄.

The hold-off circuit

Part of the saw-tooth voltage from the integrator is fed via diode D₁ to the emitter-followers TS₁, see fig. 2. The resistances R₁, R₂ and R₃ determine the voltage level above which D₁ becomes conducting. The leading edge of the signal passed by D₁ to the base of TS₁ will be transmitted undistorted by this transistor to capacitor C_H charging it to the peak value of this signal.

During the flyback of the saw-tooth voltage TS₁ will be cut off as a result of the capacitive load provided by C_H. The voltage now falls relatively slowly to the quiescent level determined by R₁, R₂ and R₃.

The hold-off multivibrator

When the output voltage of the hold-off circuit reaches about +6 V the hold-off multivibrator will switch over, causing TS₂ to become conducting and TS₃ (see fig. 3) to be cut off.

The negative-going edge of the signal switches the multivibrator back to the quiescent position (TS₂ cut off, TS₃ conducting) after the "hold-off time", which is determined by the values of R₄ and C_H of fig. 2.

The hold-off multivibrator receives a constant current via transistor TS₄. This current can be interrupted by means of the push-button switch S of fig. 3 (the "SET READY" button for the "SINGLE" - single-sweep - mode of operation). When this is

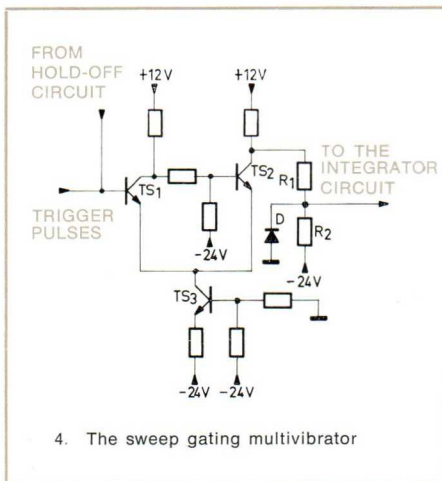
depressed, the base voltage of TS₄ is reduced to such a level that D₂ becomes conducting; this causes TS₄, and hence also TS₂ and TS₃, to be cut off. When the push-button is released, TS₄ will become conducting again, as a result of which the multivibrator will assume its quiescent position.

This possibility of controlling the hold-off multivibrator by hand is made use of in the "SINGLE" mode of the main time base, or with the time base used in the normal way if it is desired to start a new sweep before the old one is completed. The output voltage of TS₃ is fed to the emitter-follower TS₅ via R₅—TS₅. The output voltage level (at the emitter of TS₅) can be controlled by varying the base voltage of TS₅. This adjustment allows the optimum setting of the sweep gating multi-

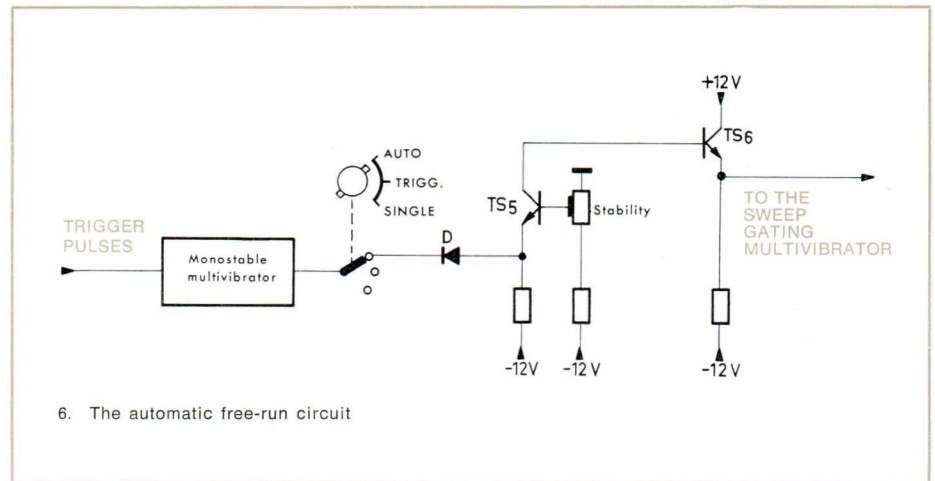
A square-wave voltage is produced at the collector of TS₁. These square-wave pulses serve as the unblanking pulses for the intensity control circuits (see page 21). The current for the SGM is taken from the transistor TS₃, connected as a current source, so that the amplitude of the control pulses is independent of the input voltage of the SGM.

The output voltage of the SGM is fed via the voltage divider R₁-R₂ to the switching transistor of the integrator. During the forward sweep of the saw-tooth voltage, diode D will be conducting. This prevents interference signals from the SGM from reaching the integrator via the switching transistor.

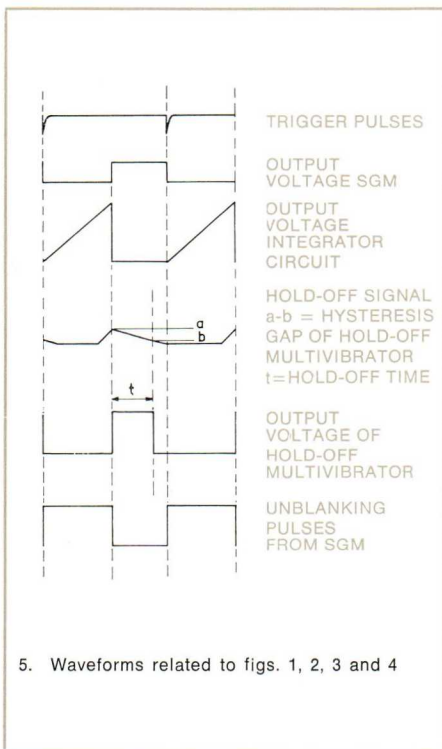
Fig. 5 shows different waveforms related to the main time base circuit.



4. The sweep gating multivibrator



6. The automatic free-run circuit



5. Waveforms related to figs. 1, 2, 3 and 4

vibrator (to which the hold-off voltage is applied) to be obtained.

The sweep gating multivibrator

The SGM is controlled by the output voltage of the hold-off circuit, to which trigger pulses from the trigger pulse shaper can be added. In the quiescent position, TS₁ will conduct and TS₂ will be cut off (see fig. 4). With the main time base set to the "TRIGG" mode, the input voltage of the SGM will be such that TS₁ will remain conducting until a negative-going trigger pulse causes the multivibrator to switch over. TS₁ will then become cut off, and TS₂ conducting. As a result, the switching transistor in the integrator will be cut off, and the saw-tooth voltage will start. When the saw-tooth voltage is so positive that the hold-off multivibrator switches over, the input voltage of the SGM will be so positive that TS₁ will be made conducting and TS₂ cut off. This causes the switching transistor of the integrator circuit to switch over again, so that the saw-tooth voltage falls to the quiescent value. The flyback of the saw-tooth voltage makes the hold-off multivibrator switch over again. The input voltage of the SGM will now be so negative that the next trigger pulse to be received will cause the cycle described above to begin again.

The automatic free-run circuit

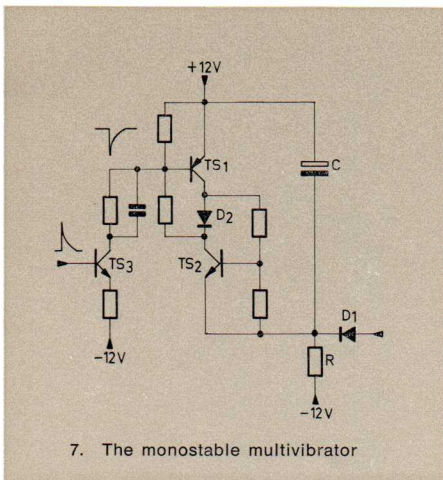
With the main time base set to the "TRIGG" mode, the time base will not operate at all in the absence of trigger pulses or if the further controls are incorrectly adjusted. Under these conditions, no trace will be visible on the screen. In order to facilitate the correct adjustment of the controls, the main time base is provided with a built-in automatic free-run circuit (from now on simply referred to as the "auto circuit"); this circuit becomes operative when the mode selector of the main time base is set to "AUTO". In this setting, the time base is made free-running in the absence of trigger signals, or if the triggering controls have been incorrectly set. If however trigger pulses are applied to the time base in this setting, the circuit is automatically switched to its triggering position. The auto circuit (fig. 6) contains a monostable multivibrator with a recovery time of about 0.5 sec. With the mode selector set to "AUTO", the output voltage of this monostable multivibrator is fed via diode D to the emitter of TS₅ (see also fig. 3). Whether D will conduct depends on the output

TS₁, D₂ and TS₂.

The circuit is dimensioned so that the current through R is not enough to keep TS₁ and TS₂ bottomed under these conditions. A snowball effect will now be produced in the opposite direction, causing TS₁, D₂ and TS₂ to be cut off again. Capacitor C will now be charged again via R. If no more trigger pulses are received, or if the interval between trigger pulses exceeds 0.5 sec, the capacitor will be charged to such a level that D becomes conducting.

The time base should therefore be used in the "TRIGG" mode for triggering frequencies below 2 Hz.

The trigger pulses required for the monostable multivibrator are taken from the differentiator of the trigger unit, are amplified and brought to the desired DC level by TS₃.



7. The monostable multivibrator

voltage of the multivibrator. In the absence of trigger pulses, this output voltage is low and D is conducting. The extra current now flowing through TS₅ reduces the output voltage of TS₆. This change in voltage is sufficient to bring the SGM into the position in which the saw-tooth voltage will start.

When trigger pulses are applied, the output level of the monostable multivibrator will be high. D will now be cut off, and the setting of the SGM will be identical with that for the "TRIGG" mode.

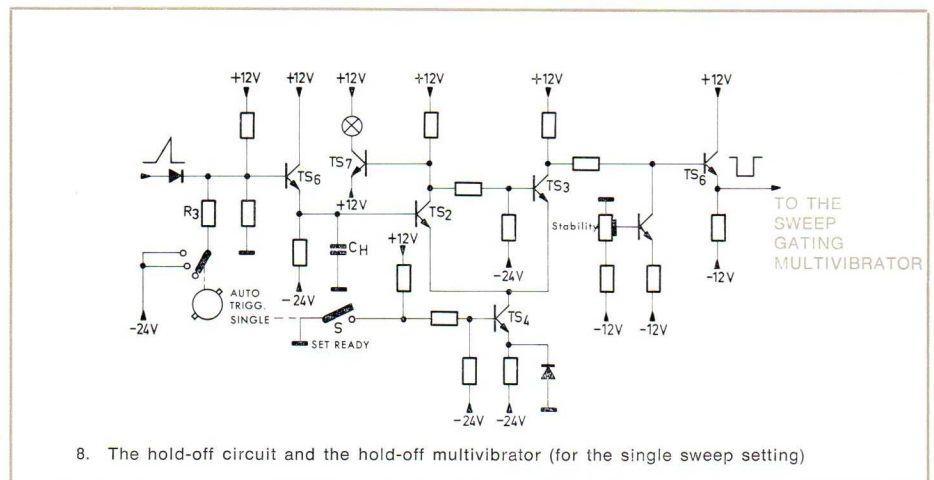
The monostable multivibrator

The monostable multivibrator (fig. 7) is made with the complementary transistors TS₁ and TS₂.

In the absence of trigger pulses, TS₁, D₂ TS₂ and TS₃ will be cut off. Capacitor C will be charged via R to such a level that D₁ will be conducting. A negative-going trigger pulse applied to the base of TS₁ will now cause the latter to become conducting; this will make the base voltage of TS₂ more positive, so that this transistor will start to conduct too. The base voltage of TS₁ will now fall, so that TS₁ becomes more conducting. This gives rise to a snowball effect, causing both transistors to become bottomed in a short space of time. Capacitor C is now discharged via

The saw-tooth generator can be made to start again by switching the hold-off multivibrator over. This is done by cutting off TS₄ — and hence TS₂ and TS₃ — by means of the "SET READY" button (switch S in fig. 8). When S is released, TS₄ (and hence TS₃) becomes conducting again, and the SGM is again in the setting which allows the saw-tooth generator to start (see also the description of the hold-off multivibrator.)

Fig. 9 shows the time relation between the saw-tooth voltage and the collector voltage of TS₂ (fig. 8) for the single-sweep setting. The collector voltage of TS₂ is fed to TS₇, which has a small incandescent lamp in its collector circuit. When TS₂ is cut off, TS₇ will conduct and the lamp will light up. This lamp (mounted in the SET READY button) thus gives the indication that the saw-tooth generator is ready to respond



8. The hold-off circuit and the hold-off multivibrator (for the single sweep setting)

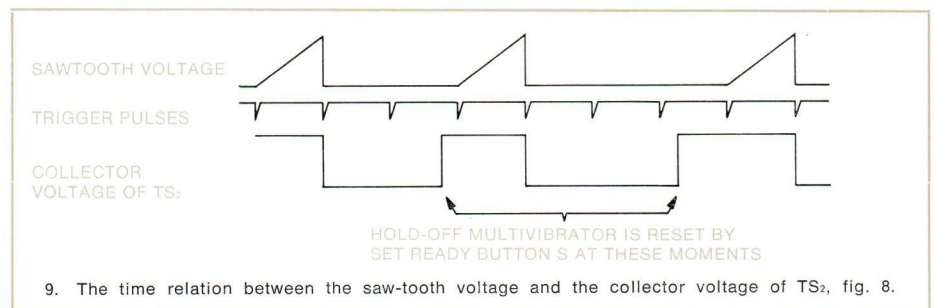
Single-sweep operation

When the time base is free-running or triggered, the hold-off multivibrator switches to and fro thanks to the fact that the output voltage of the hold-off circuit cuts both edges of the hysteresis gap of the hold-off multivibrator (see fig. 5). When the time base is set to the "SINGLE" mode, the lower limit of the base voltage of TS₁ (fig. 8) is raised to such a level that threshold b (fig. 5) of the hold-off multivibrator is not crossed. After a single sweep of the saw-tooth voltage, the multivibrator then remains in the position with TS₂ conducting and TS₃ cut off. This raising of the lower limit of the base voltage of TS₁ in the "SINGLE" mode is obtained by breaking the connection between R₃ and the -24 V supply.

to the next trigger pulse. The lamp remains lit up during the subsequent sweep of the saw-tooth voltage.

Output signals from the time-base generator

The output signals from the time-base generator which have to be fed to other units (saw-tooth voltage, unblanking pulse Y-alternate signal) are converted into currents and passed to their destination via 50-ohm cables with matched termi-



9. The time relation between the saw-tooth voltage and the collector voltage of TS₂, fig. 8.

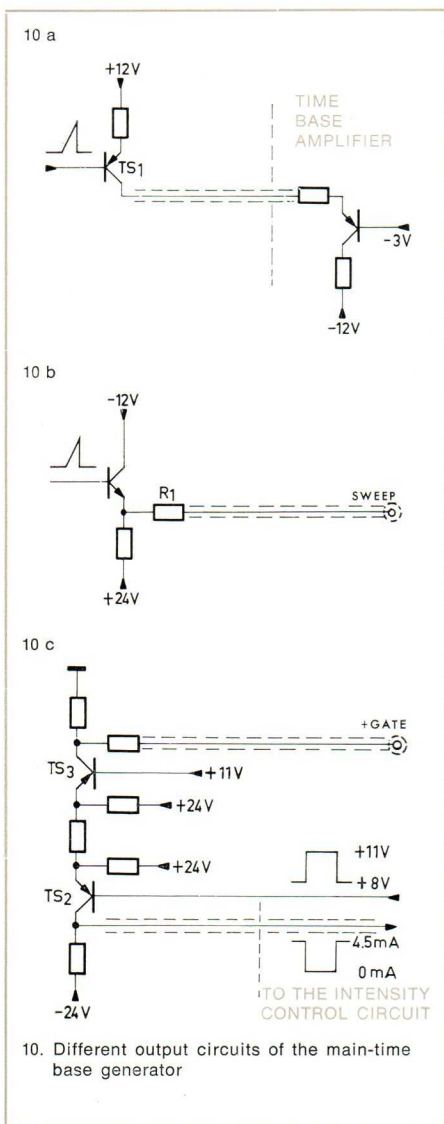
nations (see e.g. fig. 10a). This procedure reduces reflection of these signals to a minimum. Moreover, signal currents can be switched more easily, which is of particular importance for the display switching.

Saw-tooth voltage

The saw-tooth voltage is transformed into a current of about 2 mA in TS₁; this current is applied to a transistor in the common-base configuration in the time-base amplifier (fig. 10a).

The saw-tooth voltage (0 to +8 V) available at the output socket in the back wall of the oscilloscope is taken off via an emitter-follower (fig. 10b).

A series resistance R₁ ensures linear operation with low load resistances (even down to 0 ohm). The short-circuit current is 1.6 mA.



ward sweep of the time base, while current is supplied to the intensity control circuit during the flyback and the waiting time before the next sweep. The voltage available at the "+GATE" socket is taken from the collector of TS₃ via a series resistance. The output voltage is 2 V, positive-going from 0 V. The short-circuit current is about 2 mA.

Y-alternate pulse

For control of the switching between various channels in multi-trace plug-in units, the hold-off multivibrator delivers a square-wave voltage, which is converted into a current with the aid of a transistor in the common-base configuration. This current is passed to its destination in a 50-ohm cable.

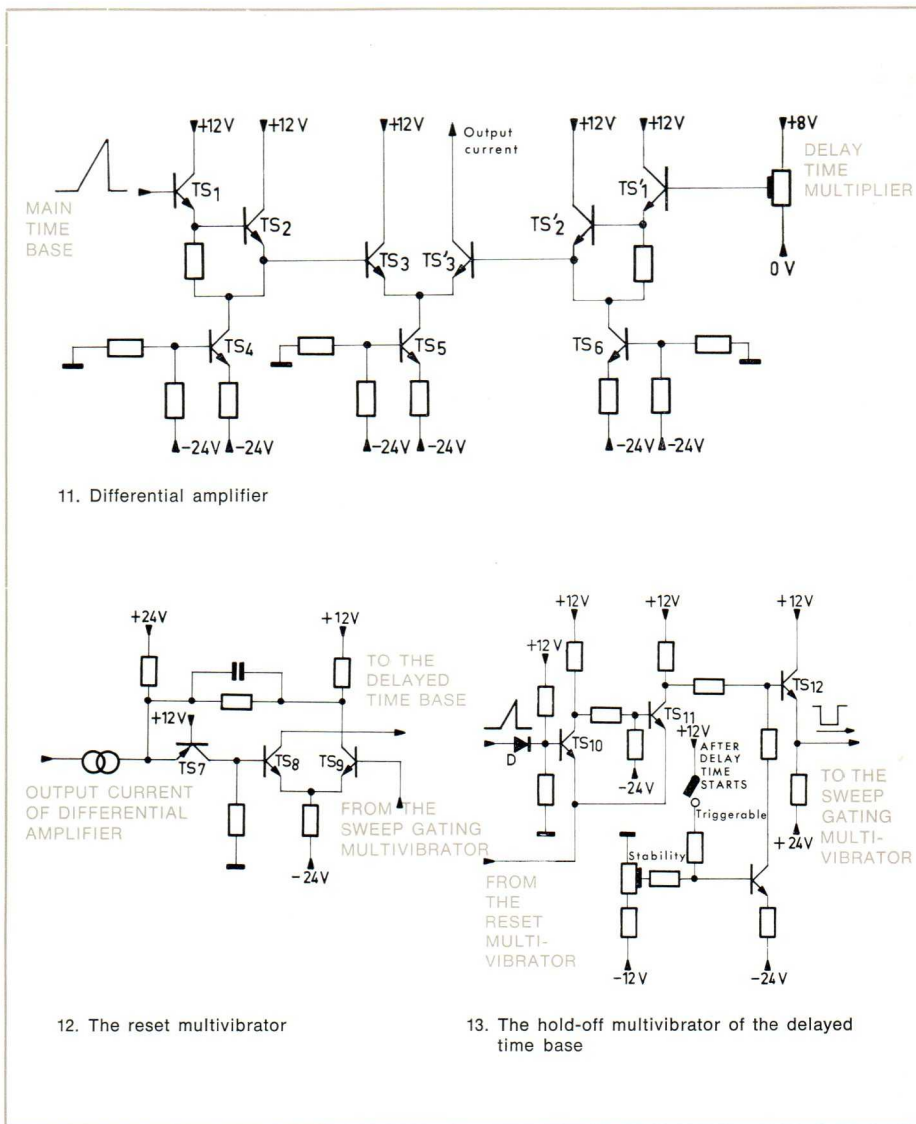
is indicated by a green colour on the control panel.

In this mode of operation, the saw-tooth voltage from the main time base is fed to a differential amplifier and compared with a DC voltage which can be adjusted by means of the DELAY TIME MULTIPLIER. When the saw-tooth voltage exceeds the DC voltage, the differential amplifier delivers an output current which is fed to the delayed time base via a pulse shaper.

Differential amplifier

The differential amplifier is shown in fig. 11. To obtain a very accurate delay time, a ten-turn helical potentiometer of excellent linearity (better than 1‰) has been used. The sensitivity to variations in ambient temperature is kept to a minimum by complete balancing of the amplifier.

TS₁, TS₂ and TS₃ are dual transistors spe-



Unblinking voltage

The unblinking voltage (collector voltage of TS₁, fig. 4) is transformed into a current in the push-pull combination TS₂-TS₃ (fig. 10c). Since the DC base voltage of TS₃ is equal to the maximum value of the unblinking voltage, no current is supplied to the intensity control circuit during the for-

THE MAIN TIME BASE AS A DELAYING TIME BASE

As already mentioned on page 5, the saw-tooth voltage from the main time base is used to initiate the delayed time base after an adjustable delay time.

The delay time is determined by the TIME/cm knob of the main time base and the DELAY TIME MULTIPLIER; this function

cially designed for differential amplifiers. TS₄, TS₅ and TS₆ are current sources used to maintain the average current at a constant value.

Reset multivibrator

The output current of the differential amplifier is fed to the reset multivibrator consisting of TS₇, TS₈ and TS₉ (fig. 12). The base of TS₉ is connected to the output of the sweep gating multivibrator of the main time base.

When the main time base is in its quiescent position, the base of TS₉ is positive with respect to the base of TS₈; TS₉ will then conduct and TS₈ will be cut off.

At the moment the main time base starts, the DC base voltage of TS₉ jumps to a lower value, which is, however, still higher than the DC base voltage of TS₈. When the saw-tooth voltage applied to the differential amplifier exceeds the DC voltage from the DELAY TIME MULTIPLIER, the output current of the differential amplifier causes a positive voltage jump at the base of TS₈. The reset multivibrator will now

of the main saw-tooth voltage, the base voltage of TS₈ will be at its more, positive level, owing to the current from the differential amplifier. In this situation TS₈ will conduct immediately after the start of the main time base, i.e. when the base voltage of TS₉ jumps to its lower level.

This makes it possible to use the DELAY TIME MULTIPLIER over its full range, without the risk that the delayed time base will cease functioning near the zero setting of the DELAY TIME MULTIPLIER.

When the base voltage of TS₉ jumps back to its original value (i.e. when the main time base voltage reaches its peak value) the reset multivibrator switches back to the position with TS₉ conducting and TS₈ cut off.

As TS₈ draws no current, the delayed time base will now return to its quiescent position, if this has not already happened.

ELECTRONIC DISPLAY SWITCH

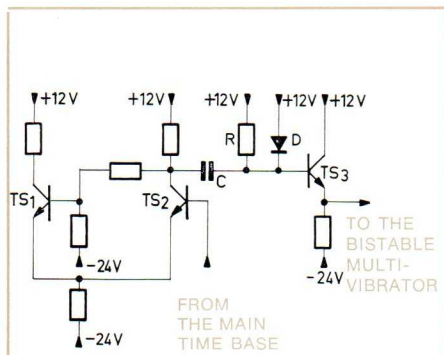
The electronic display switch is driven by a bistable multivibrator; when the X DEFL. switch is set to ALTERN, the horizontal deflection of the CRT receives the two time-base signals alternately.

When the X DEFL switch is set to "MAIN TB" or "MAIN TB INTENS", the multivibrator remains in one of its stable states, while with X DEFL set to "DEL'D TB" the multivibrator remains in its other stable state.

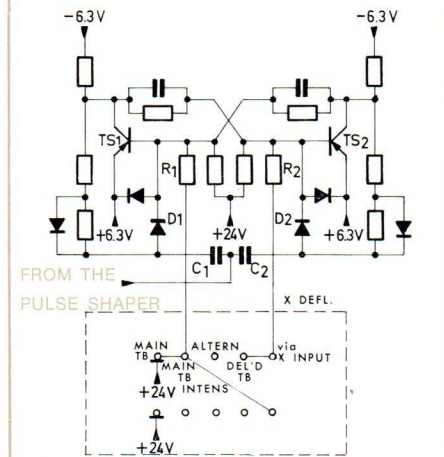
Pulse shaper

The driving pulses for the bistable multivibrator are delivered by a pulse shaper, the Schmitt trigger TS₁, TS₂ of fig. 14. This pulse shaper is driven by a pulse from the main time base.

The square-wave voltage delivered by the pulse shaper is differentiated by C-R. The



14. Pulse shaper



15. Driving multivibrator

switch over to the position in which TS₉ is cut off and TS₈ is conducting. The output voltage of TS₈ is fed to the delayed time base.

Depending on the setting of the delayed time base controls, the delayed time base will either start immediately or be triggerable by the normal trigger sources.

If the DC voltage of the DELAY TIME MULTIPLIER is below the quiescent level

DELAYING TIME BASE

The delayed time base is of the same design as the main time base, except that it lacks the hold-off circuit and the automatic free-run circuit.

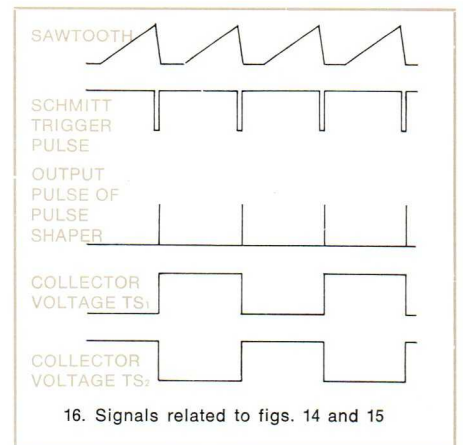
The hold-off multivibrator (fig. 13) of the delayed time base works like that of the main time base when the latter is set for single-sweep operation. The role of the SET READY switch is here played by the reset multivibrator (see the preceding section, on "Main time base as a delaying time base").

When TS₈ (fig. 12) is cut off, TS₁₀ and TS₁₁ will likewise be cut off. The collector voltage of TS₁₁, and hence also the input voltage of the SGM, will now be high. The delayed time base will thus be in its quiescent position.

Since the base voltage of TS₁₁ is higher than that of TS₁₀, the former will become conducting when TS₈ starts to conduct. The collector voltage of TS₁₁ will then fall, and so will the input voltage of the SGM. Depending on the position of the AFTER DELAY TIME switch, the saw-tooth will start either immediately (when the switch is set to STARTS) or on receipt of the next trigger pulse (when the switch is set to TRIGGERABLE).

The top portion of the saw-tooth voltage is fed to the base of TS₁₀ via diode D; this causes the hold-off multivibrator to switch over (TS₁₀ becomes conducting and TS₁₁ cut off). During the flyback of the saw-tooth voltage, diode D is cut off so that the multivibrator remains in the above position and no new saw-tooth voltage is generated. Since the operating time of the delayed time base always falls within that of the main time base, no hold-off circuit is needed here.

When the current from the reset multivibrator is now interrupted, at the end of the flyback of the main time base, both transistors TS₁₀ and TS₁₁ will be cut off. As soon as TS₈ (reset multivibrator) starts to draw current again, TS₁₁ will become conducting, and a new cycle can begin.



16. Signals related to figs. 14 and 15

negative pulse formed in this way is eliminated by the diode D, while the positive pulse is fed to the bistable multivibrator via the emitter-follower TS₃.

Driving multivibrator

With X DEFL set to MAIN TB or MAIN TB INTENS, a positive voltage is applied to R₁, while with X DEFL at DEL'D TB a positive voltage is applied to R₂; in the setting „via X INPUT”, both these resistors receive a positive voltage. (See fig. 15).

As a result, in the settings MAIN TB and MAIN TB INTENS the transistor TS₁ is cut off while TS₂ is conducting. When the X DEFL switch is set to DEL'D TB, TS₂ is cut off while TS₁ is conducting, while in the setting "via X INPUT" both transistors are cut off.

The positive driving pulses are fed to the bases of TS₁ and TS₂ via the capacitors C₁ and C₂ and the diodes D₁ and D₂. The driving pulse makes the conducting transistor cut off, as a result of which the transistor which was originally cut off starts to conduct. The collector voltages of the transistors TS₁ and TS₂ are switched between -6.3V and +6.3V in this way.

Fig. 16 shows the relation between the

main time-base signal and the collector voltages of the bistable multivibrator.

Switching of time bases and unblanking pulses

The principle of the switching of the sawtooth and unblanking currents is shown in fig. 17.

When the cathode voltage of D_1 and D_5 is positive, these diodes are cut off and diodes D_2 and D_6 are conducting. At the same time, the cathode voltage of diodes D_3 and D_7 is negative. This means that D_3 and D_7 will be conducting, while D_4 and D_8 are cut off.

The signals from the main time base are now applied to the time-base amplifier and the intensity control unit.

When the bistable multivibrator switches over, D_2 and D_6 are cut off and D_4 and D_8 are conducting; the time-base amplifier

Y-alternate pulse

As has already been mentioned a square-wave voltage (called the "Y-alternate pulse") is taken from the hold-off multivibrator of the main time base. This is used to switch over from one channel to the other in multi-channel plug-in units when the mode switch of the Y unit is set to ALTERN.

Because the delayed time-base sweep always falls within the sweep time of the main time base and the two time bases have the same repetition frequency, the same Y-alternate pulse is used when the X DEFL switch is set to DEL'D TB.

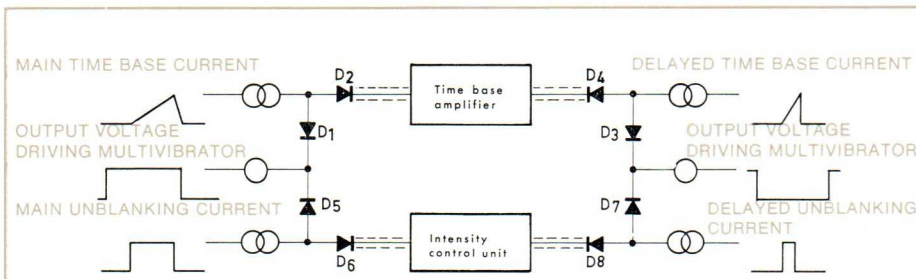
When the time bases are used alternately, the signal on one channel of the Y amplifier must first be displayed with both time bases before the system switches over to the other channel. This means that a Y-alternate pulse with half the frequency of

Trace separation

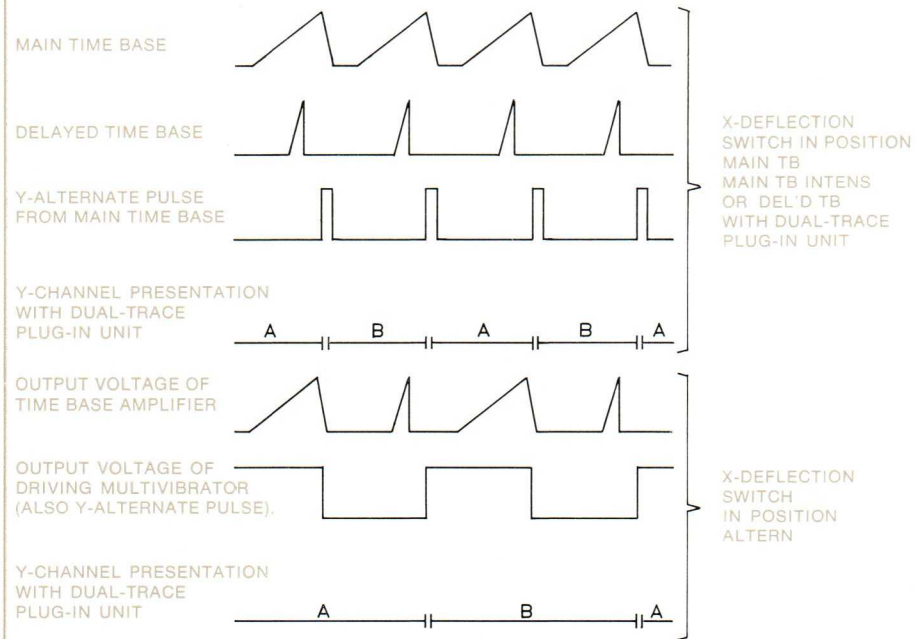
The trace separation circuit serves to ensure that the traces from the two time bases are given a vertical separation on the CRT screen.

Two square-wave voltages in anti-phase, both derived from the bistable multivibrator, are fed via the emitter-followers TS_1 and TS_2 to the clipping network D_1 - D_2 and D_3 - D_4 (fig. 20).

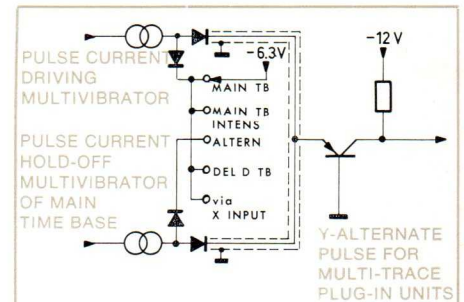
The clipping level depends on the emitter voltage of TS_3 , and can be adjusted with the aid of the TRACE SEPARATION potentiometer R (see fig. 20 and 21). Depending on the setting of R, diodes D_1 and D_3 will thus pass part of the signal to the bases of TS_4 and TS_5 (fig. 20). With the aid of these transistors, the square-wave voltages are transformed into currents and passed to the Y amplifier via 50-ohm cables.



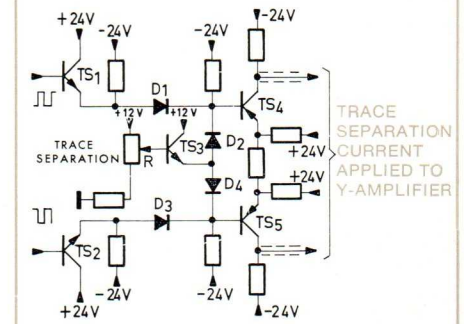
17. Principle of the switching of the sawtooth and unblanking currents



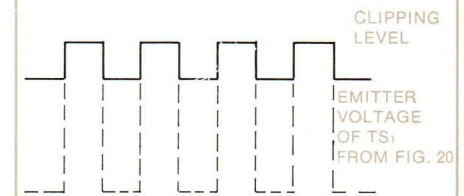
18. The relationship between sawtooth voltages and Y-alternate pulses in the different positions of the X-deflection switch.



19. Current transport of Y alternate pulses



20. Trace separation circuit



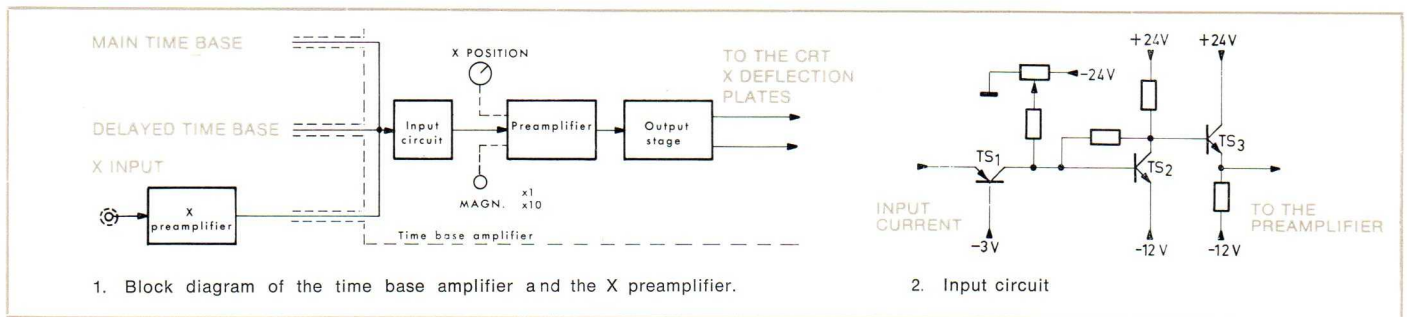
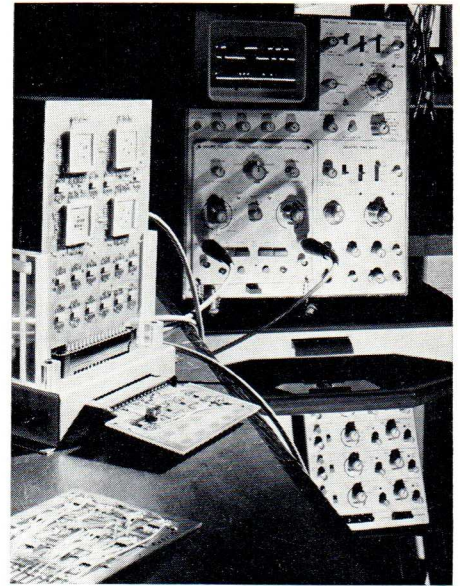
21. Trace separation signal related to fig. 20

and the intensity control unit now receive the signals from the delayed time base. When the X DEFL switch is set to MAIN TB or MAIN TB INTENS, the cathode voltage of diodes D_1 and D_5 remains positive, while in the setting DEL'D TB the cathode voltage of D_3 and D_7 remains positive. With X DEFL set to "via X INPUT", D_2 , D_6 , D_4 and D_8 are all cut off, and the time-base amplifier is driven from the X pre-amplifier.

the original Y-alternate pulse is also required. This pulse is taken from the bistable multivibrator. The time relation between the various time-base voltages, the Y-alternate pulses and the channel presentation of a dual-trace plug-in unit is given in fig. 18. The Y-alternate pulses are converted into currents and passed to the Y amplifier via 50-ohm cables (fig. 19).

The trace-separation control is symmetrical, thanks to the symmetrical output of the bistable multivibrator.

The time base amplifier and X preamplifier



1. Block diagram of the time base amplifier and the X preamplifier.

2. Input circuit



Making the cable form of the PM 3370

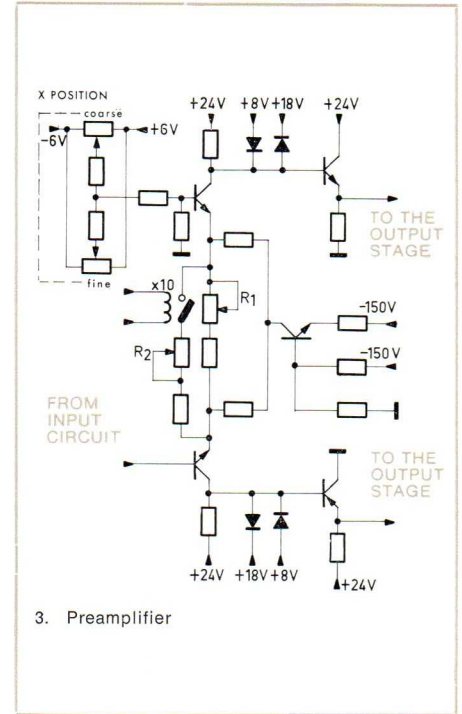
The time-base amplifier has been specially designed for the amplification of saw-tooth voltages. The time-base signals are applied via an input circuit to an amplifier stage whose amplification can be increased by a factor 10.

The DC voltage for the X POSITION control is also applied to this stage. Use of the X preamplifier makes X-Y measurements possible. Its output signal, like those of the two time-base generators,

is fed as a current to the time-base amplifier. The block diagram of the time-base amplifier with X preamplifier is shown in fig. 1.

Input circuit

The input current, from one of the two time-base generators or from the X pre-amplifier, is fed via transistor TS₁ (common-base configuration) to TS₂, which is connected as an amplifier with shunt feedback (fig. 2).



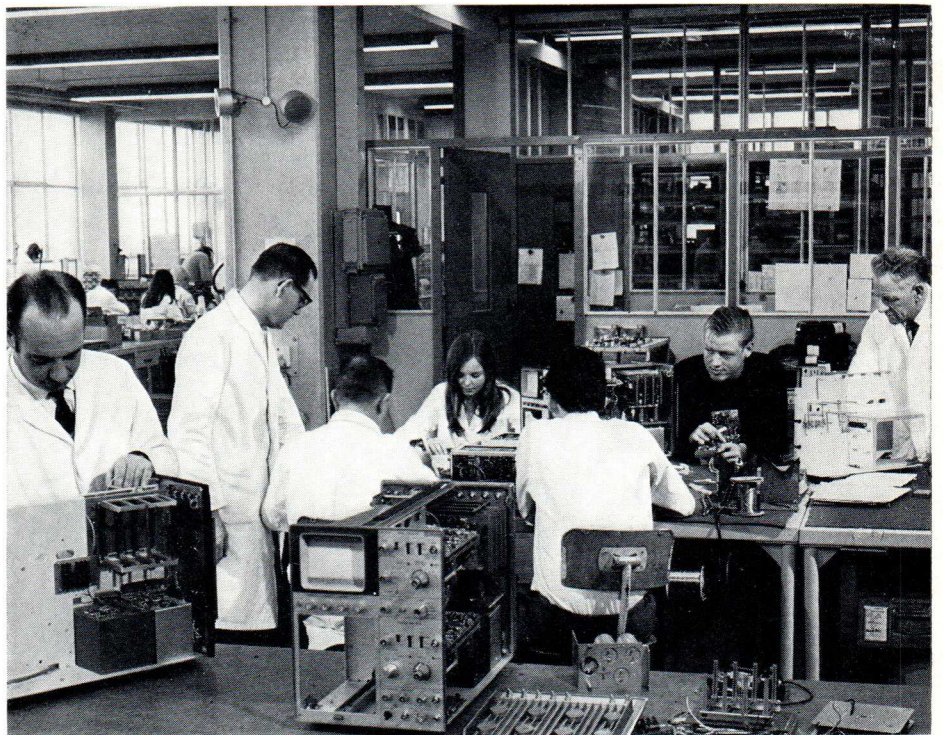
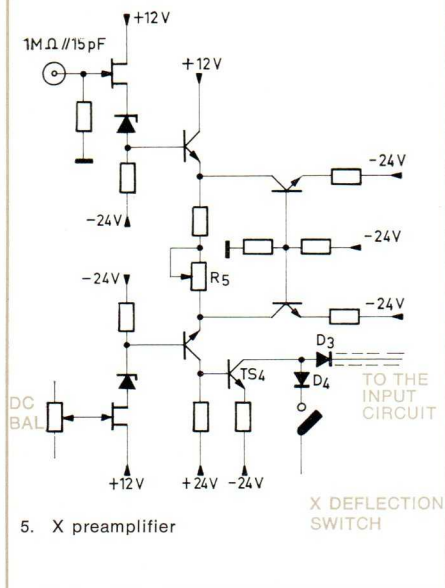
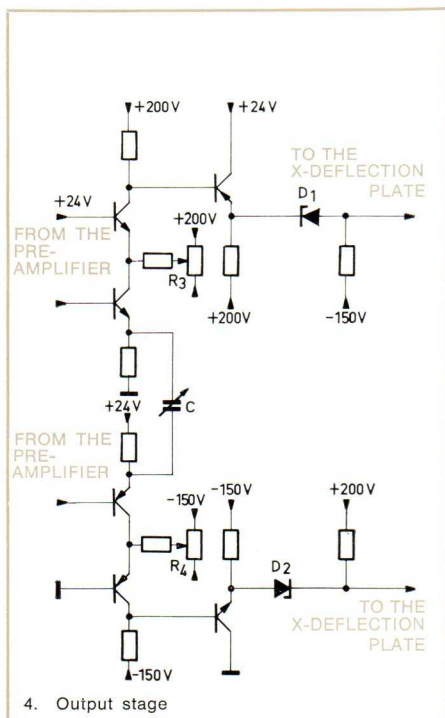
3. Pre-amplifier

The output voltage TS₂ is taken off via emitter-follower TS₃.

Pre-amplifier

This amplifier is built in push-pull (fig. 3). When the MAGN switch is set to "x 1", the amplification is adjustable with the aid of R₁.

With the MAGN switch at "x 10", the feed-



Assembly of prototypes of the PM 3370

back resistance is reduced with the aid of a reed relay. The magnification factor can be adjusted with the aid of R_2 . The DC voltage for the X POSITION control can be varied with the aid of two mechanically coupled potentiometers, one of which provides fine control and the other coarse control, in such a way that the position of the image can be accurately adjusted, even when MAGN is set to "x 10".

Output stage

The output stage is constructed with complementary transistors (fig. 4), and is designed so that all transistors pass an increasing current during the forward sweep of the time base. This makes for good linearity, especially at high sweep rates. Moreover, the transistors can be set to a relatively low quiescent current with this design.

C is used to improve the linearity at high sweep rates.

Because of the use of NPN and PNP transistors, the output voltage levels are different. The Zener diodes D_1 and D_2 are included in the circuit to compensate for this.

The voltages applied to the deflection plates can be accurately adjusted to the desired levels with the aid of the potentiometers R_3 and R_4 .

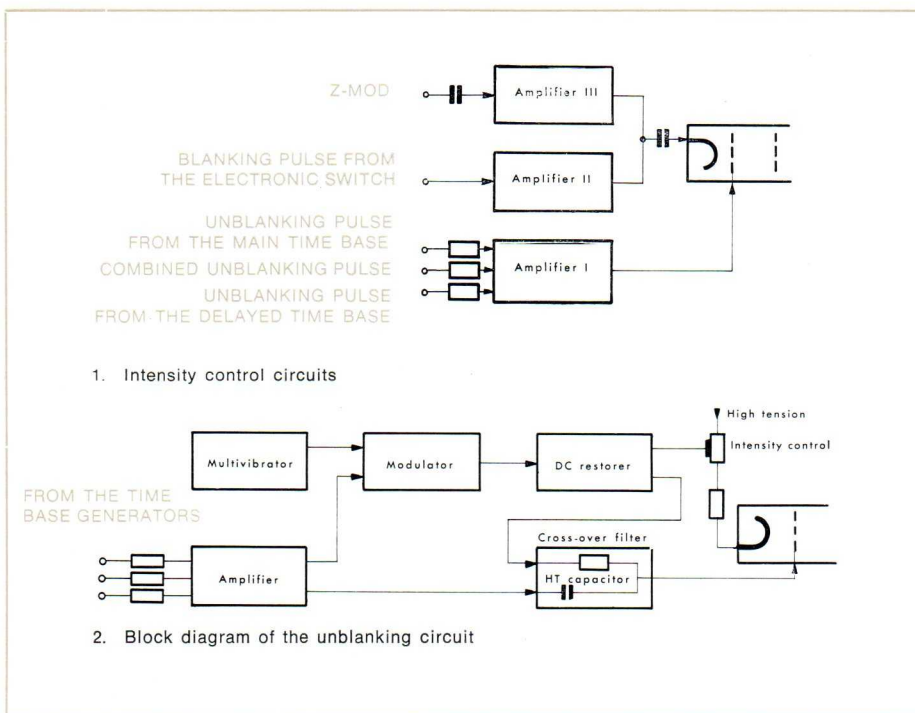
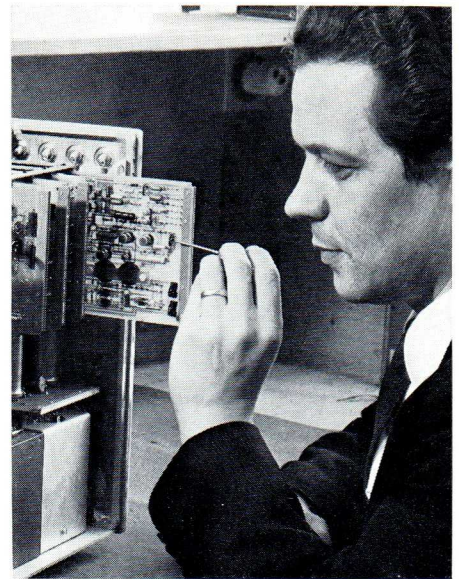
X preamplifier

The X preamplifier is a push-pull stage with FET input ($1 \text{ Mohm} // 15 \text{ pF}$); see fig. 5. The deflection coefficient can be continuously varied from 10 to about 100 mV/cm with the aid of R_5 , which is operated by the "mV/cm" knob.

TS_4 delivers the current for the output signal.

The diode switch D_3 - D_4 , which realizes the connection with the time-base amplifier, is operated by the X DEFL switch.

The intensity control circuits



The intensity control circuits ensure that:

- a trace is visible on the screen of the cathode-ray tube (CRT) during the forward sweep of the time base, and not during the flyback or the interval between successive saw-tooth signals;
- during the switching over of the electronic switch in the vertical amplifier, the edges of the switching pulses are suppressed;
- external Z-modulation signals can be applied to the CRT.

This implies that the various unblanking and blanking pulses must be applied at the right time and with the right amplitude to the control grid and the cathode of the CRT respectively.

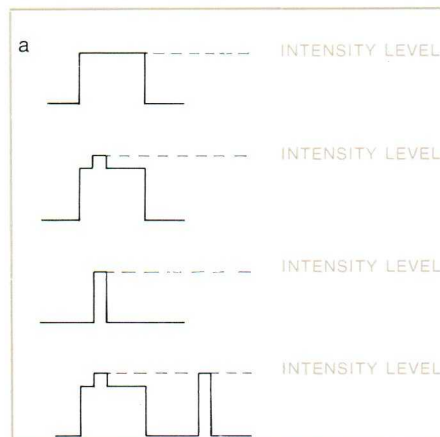
The intensity control circuits consist of three amplifiers, I, II and III (see fig. 1). The following signals can be applied to the various amplifiers, to realize the different functions mentioned above.

The following signals can be applied to the various amplifiers, to realize the different functions mentioned above.

Applied to amplifier I

a) An unblanking pulse from the main time base, when the X-deflection mode selector is set to MAIN TB; see fig. a.

b) A combined unblanking pulse from the main time base and the delayed time base, when the X-deflection mode selector is set to MAIN TB INTENS;



c) An unblanking pulse from the delayed time base, when the X-deflection mode selector is set to DEL'D TB;

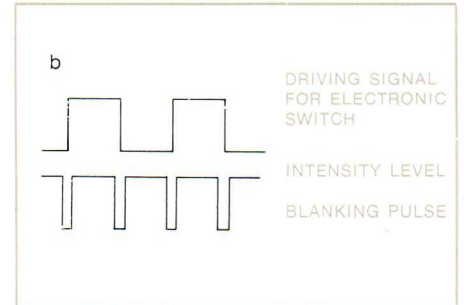
d) The unblanking pulses mentioned under b) and c) alternately, when the X-deflection mode selector is set to ALTERN.

Applied to amplifier II

A blanking pulse from the electronic switch, when a multi-trace Y plug-in unit is used in position CHOPP, see fig. b.

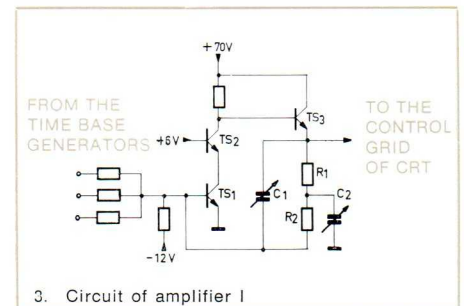
Applied to amplifier III

External Z-modulation signals, as required.



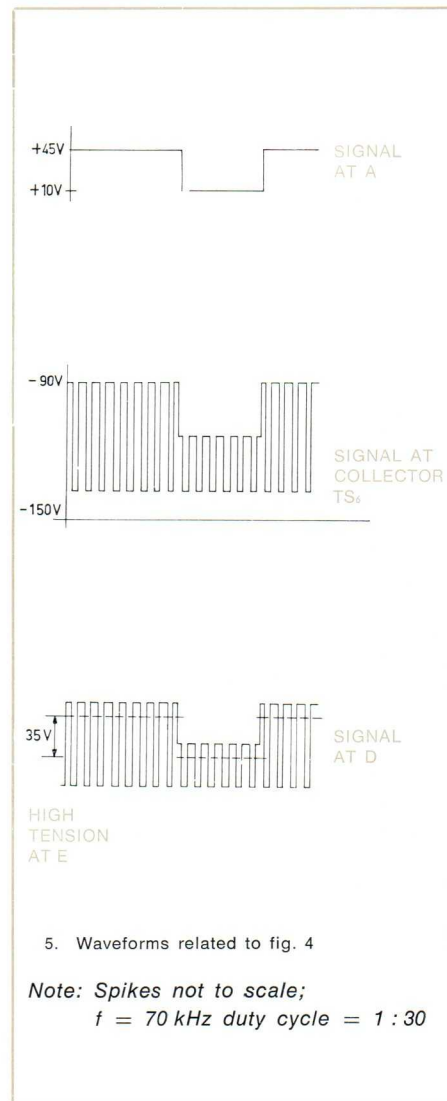
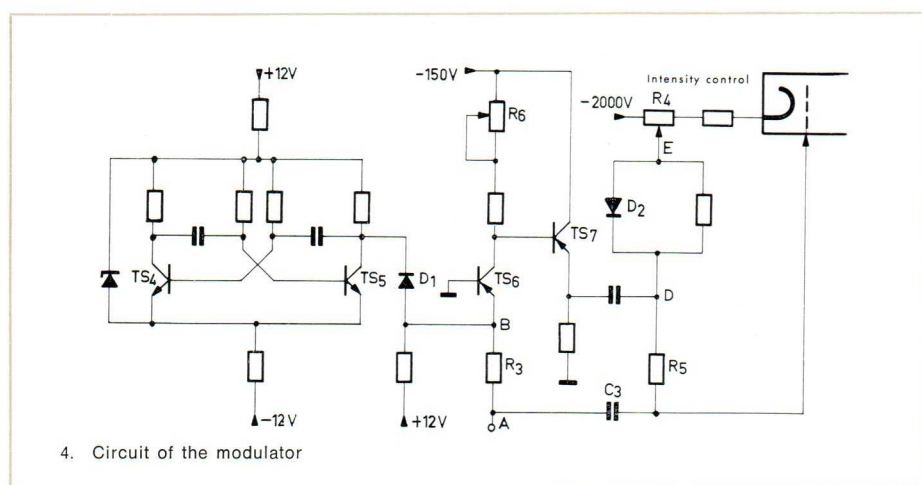
The signals from the two time bases, in the form of current pulses, are applied to the intensity-control circuits via cables provided with matched terminations.

Figure 2 shows the block diagram of the unblanking circuit, and figure 3 the circuit diagram of amplifier I.



The HF components of the various signals are applied directly to the control grid via an HT capacitor. The DC and LF components are applied to the modulator as modulation signals. The switching frequency of the modulator is derived from a multivibrator. The modulated carrier wave produced at the output of the modulator is applied to a DC restorer which is at the same DC level as the control grid. After detection, the carrier wave is eliminated by means of a cross-over filter, which also combines the signal from the DC restorer in the proper manner with the HF components; and the resulting signal is applied to the control grid of the CRT. The over-all effect of the unblanking circuit is thus that the complete unblanking signal is brought to the desired DC level (about -2 kV).

converted into a current with the aid of R_3 , and is fed to point B. The multivibrator $TS_4 - TS_5$ delivers a pulsed signal with a frequency of about 70 kHz and a duty cycle such that the diode D_1 is cut off about 3% of the time. The current signal through D_1 is also applied to point B, in such a way that it is subtracted from the signal applied via R_3 . A signal with the form shown in fig.5 is now applied to the collector of TS_6 . This signal is now fed to a DC restorer via emitter-follower TS_7 . Diode D_2 only conducts during the peaks of the switching pulses; as a result, the signal is clamped to the DC level set at point E by the "INTENS" control potentiometer R_4 . The switching pulses in the signal at D are completely smoothed by the integrating effect of the filter R_5C_3 . The HF components of the unblanking



Amplifier I

This amplifier stage consists of the transistors TS_1 and TS_2 , followed by an emitter-follower TS_3 . Negative voltage feedback is applied across the whole stage, which keeps the input impedance very low. This feedback is realized with a bridged T-filter. The magnitude of the output signal is mainly determined by the input current and the resistances R_1 and R_2 . Optimum step response is obtained with the aid of C_1 and C_2 .

The circuit diagram of the modulator is shown in fig. 4. Point A is directly coupled to the output of amplifier I. The signal is

signal are thus applied to the control grid of the CRT via C_3 , while the other components (including the DC component) are applied to the control grid via R_5 . R_6 ensures a suitable relationship between the amplitudes of the signals at the points D and A. The cathode current of the CRT, and hence the intensity of the trace, can be adjusted by varying the grid voltage of the CRT by means of R_4 .

Amplifiers II and III

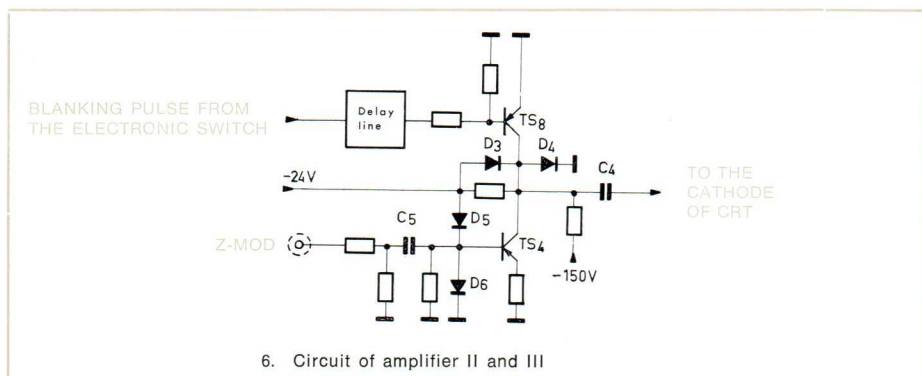
The circuit diagram of amplifier II (for the chopper blanking pulses) and amplifier III (for external Z-modulation signals) is shown in fig. 6.

TS_8 is normally cut off. A negative-going blanking pulse from the electronic switch of the Y plug-in unit is fed to the base of TS_2 via a delay line. This delay line is necessary to compensate for the delay in the vertical amplifier.

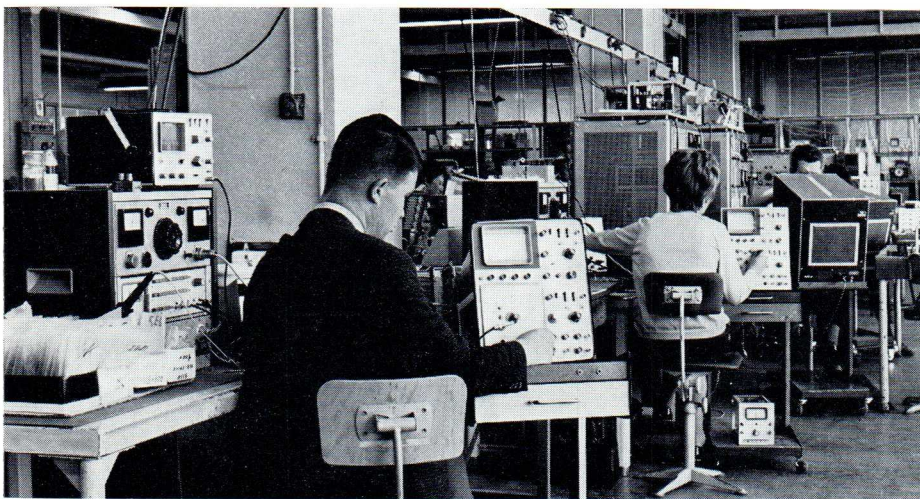
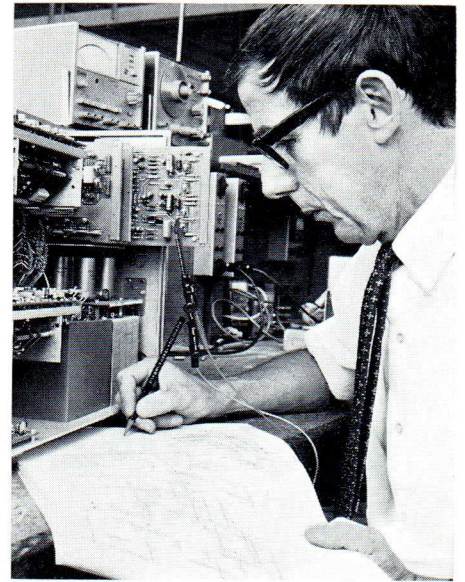
During the existence of this pulse, TS_8 is conducting and a blanking pulse is produced at its collector. This pulse is fed to the cathode of the CRT via the HT capacitor C_4 .

External modulation signals can be applied to the "Z MOD" socket, where they are fed to the base of TS_4 via C_5 . The resulting signals at the collector of TS_4 are fed also to the cathode of the CRT via the high voltage capacitor C_4 .

Over voltage protection for the two transistors is afforded by the four diodes D_3 , D_4 , D_5 and D_6 .



The power supply



Final adjustment of the PM 3370

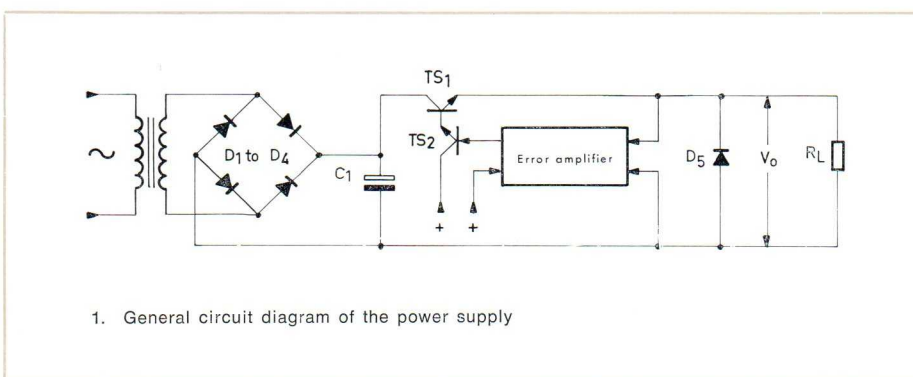
The output voltages of the power supply, which serve the plug-in units as well as the circuits in the main frame of the PM 3370, are electronically stabilized. The power supply is also protected against overload and short-circuits by a simple but effective circuit, with automatic reset. The mains-voltage selector allows various mains voltages to be accommodated. The primary circuit is safeguarded by a slow-blow fuse and a thermal cut-out switch. The latter is encapsulated in the transformer moulding, and breaks the primary circuit if the transformer tempera-

ture exceeds a certain value. Once the transformer has cooled down sufficiently, the cut-out switch cuts in again. To limit the stray magnetic field as much as possible, and hence to restrict undesirable coupling with other circuits, the transformer coils have been wound astatically. The essentials of the power supply are shown in the circuit diagram of fig. 1. The AC voltage delivered by the transformer is rectified by 4 power diodes, D_1 — D_4 , which are arranged in a bridge circuit, and smoothed by the electrolytic capacitor C_1 .

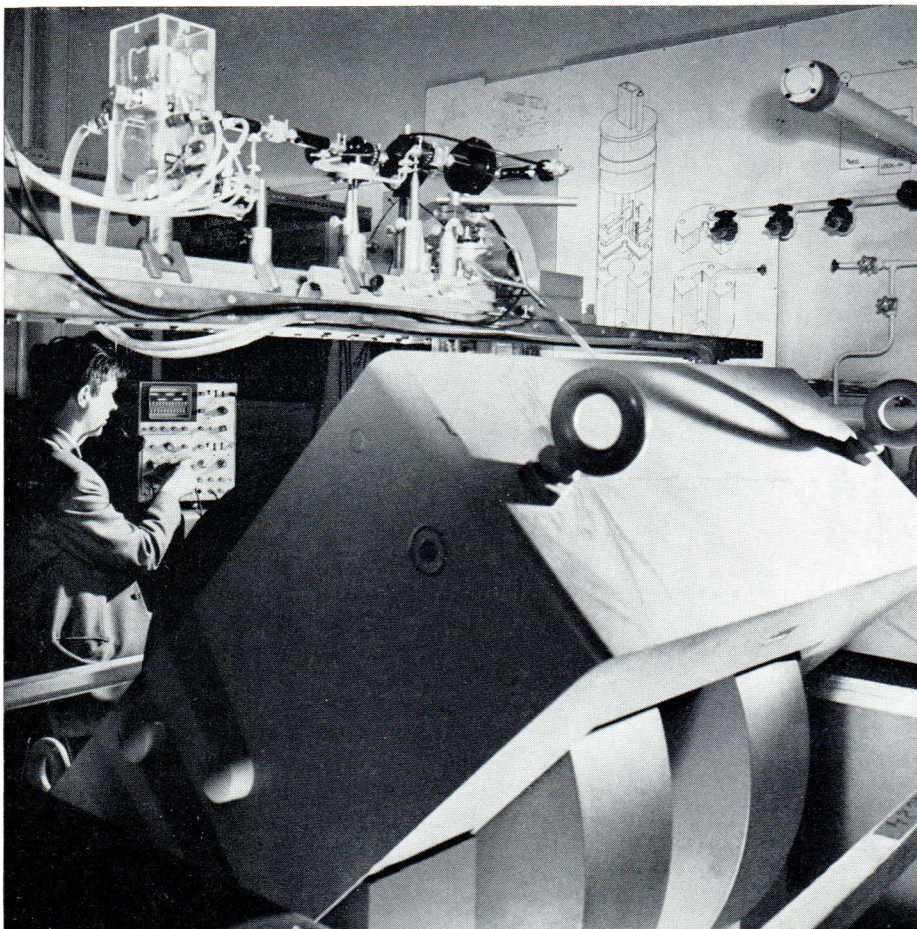
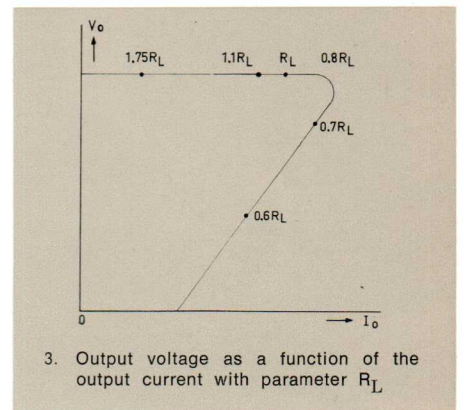
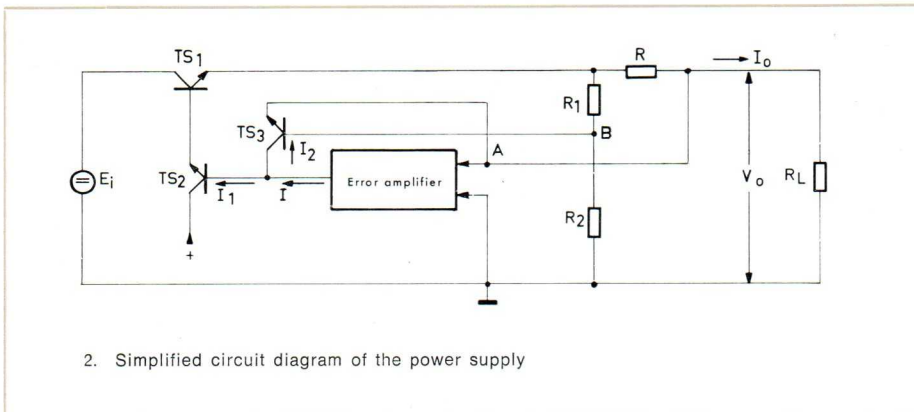
Transistor TS_1 smooths the voltage still further and keeps the output voltage V_o within close limits around the nominal value. This stabilization is realized by feeding part of the output voltage to an error amplifier, where it is compared with a stable reference voltage. The control current delivered by the error amplifier is further amplified by transistor TS_2 , and applied to the base of TS_1 . If the output voltage V_o increases, the control current produced by the error amplifier will be reduced; when this signal is fed back to TS_1 , it causes this transistor to conduct less, thus bringing V_o back to its original value. Similarly, a decrease in the output voltage is opposed by an increase in the control current, which makes TS_1 conduct more.

Diode D_5 ensures that if a negative supply should accidentally be connected to the output terminals, the polarity of these terminals will not be reversed (which could seriously damage the equipment); it does this by effectively short-circuiting both power-supplies involved.

The overload protection is provided mainly by the transistor TS_3 and the resistors R_1 and R_2 shown in the (simplified) circuit diagram of fig. 2. Under normal operating conditions, when the load resistance R_L is equal to the nominal value R_{nom} , point A is slightly negative with respect to point B, so that TS_3 is still just cut off. Under these conditions, the error amplifier can exercise full control over the series transistor TS_1 via the emitter-follower TS_2 . As R_L is reduced, however, the voltage drop across R will increase while the voltage at A remains relatively constant. Consequently, the voltage at B will become steadily more positive with respect to that at A, and at a pre-determined value of R_L the transistor TS_3 will start to conduct.



1. General circuit diagram of the power supply



Measurement of electron spin resonance with the aid of the PM 3370

It is apparent from fig. 2 that $I = I_1 + I_2$. As the load resistance decreases further the resulting base-emitter voltage of TS_3 will increase. Because the relation $I = I_1 + I_2$ has to be satisfied, I_2 now increases at the expense of I_1 . The result is an output voltage/output current characteristic as indicated in fig. 3. It can be shown that the short-circuit output current is approximately equal to

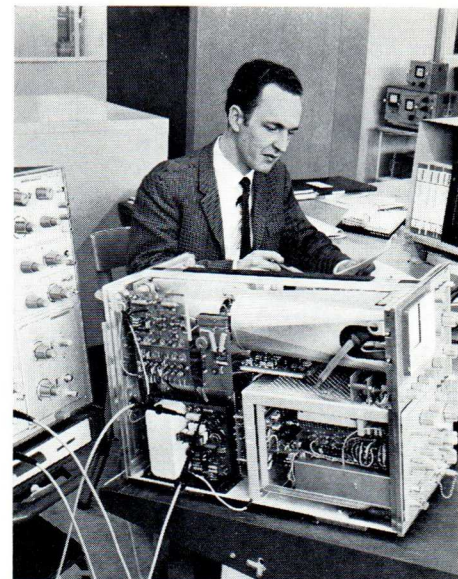
$$\frac{V_{BE}}{R}$$

where V_{BE} is the base-emitter voltage of TS_3 .

When R_L is increased to more than 0.8 times its nominal value, the short-circuit-protection circuit no longer influences the operation of the voltage-stabilization circuit, and the power supply regains its nominal output voltage.

To stabilize sweep rates as much as possible, due attention has been paid to the stability of the -24 V supply. A high loop gain has been achieved by replacing the collector resistor of the error amplifier by a transistor circuit giving a high dynamic resistance. Furthermore, a Zener diode with a low temperature coefficient has been chosen for the voltage reference.

High-voltage supply



The high-voltage supply delivers a positive (13 kV) and a negative (2 kV) voltage. These voltages are derived from a push-pull generator (tuned to 30 kHz) consisting of HT transformer T, transistors TS₁ and TS₂ and inductor L₂ (fig. 1). In this arrangement the transistors act together as one single-pole double-throw switch.

Fig. 2 shows a simplified diagram of this arrangement where L₂ represents the inductor, tank circuit LC the (tuned) HT transformer and S the switching transistors; R_L represents the losses and the load on the HT supply.

From this figure it can be seen that the current I_x through L₂ is split by switch S into I₁ and I₂. The currents I₁ and I₂ are equal, thanks to the symmetry of the circuit; their sum (I₁ + I₂ = I_x), is kept constant by the high self-inductance of L₂. As L is centre-tapped, V_x is equal to half of V₁ or V₂:

$$V_{1m} = V_{2m} = 2 V_{xm} \quad 1$$

Assuming that the DC voltage drop across L₂ is zero, the average value of voltage V_x must be equal to the supply voltage V_b, so that:

$$V_b = \frac{1}{\frac{1}{2}T} \int_0^{\frac{1}{2}T} V_{xm} \sin \frac{2\pi t}{T} dt = \frac{2}{\pi} V_{xm}$$

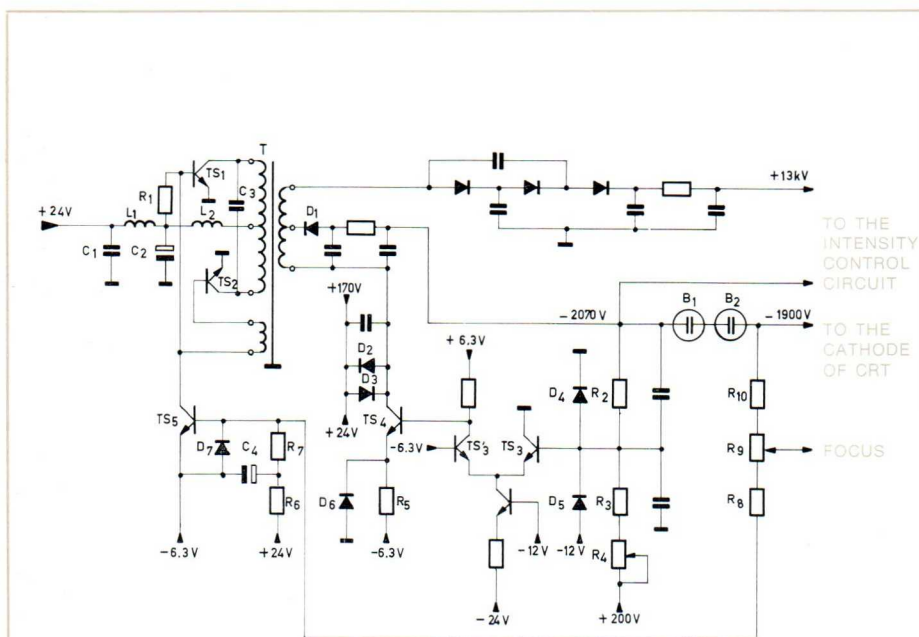
$$\text{From this it follows that: } V_{xm} = \frac{\pi}{2} V_b \quad 2$$

Substituting equation 2 in equation 1, we find:

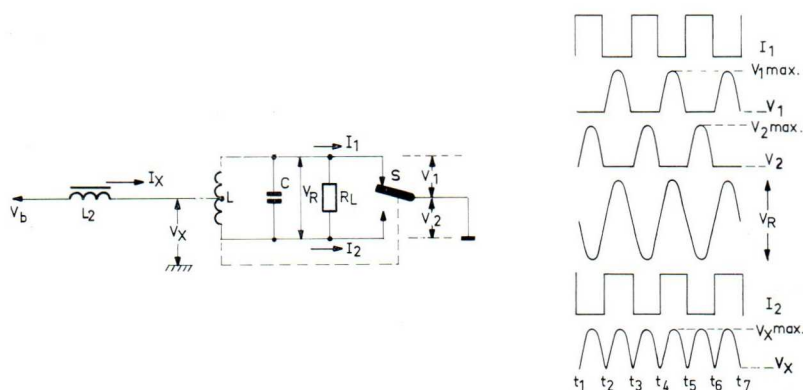
$$V_{1m} = V_{2m} = \pi V_b$$

Fig. 2 shows the phase relations between the various current and voltage waveforms present in the circuit proper. The peak collector voltage (V_{1m} or V_{2m}) occurs when the transistor is cut off.

In the circuit proper the alternating drive required at the bases is delivered by the feedback winding of the HT transformer T.



1. Circuit diagram of the high-voltage supply



2. The phase relations between the various current and voltage waveforms

This winding is given a positive bias by connection to the 24-volt supply across resistor R_1 .

Filter C_1 - L_1 - C_2 prevents interference from reaching the supply line. In case of overload a negative voltage is applied to the feedback winding by transistor TS_5 , thus blocking the oscillator. The overload protection will be discussed in detail below. The secondary winding of T delivers a voltage to a tripler circuit which in its turn delivers a positive supply voltage of about 13 kV with respect to earth. A tapping on the secondary winding delivers a voltage to the half-wave rectifier D_1 . After smoothing, a negative supply of 2070 volt is obtained. Neon stabilizers B_1 and B_2 reduce this to -1900 volt for the cathode voltage of the CRT.

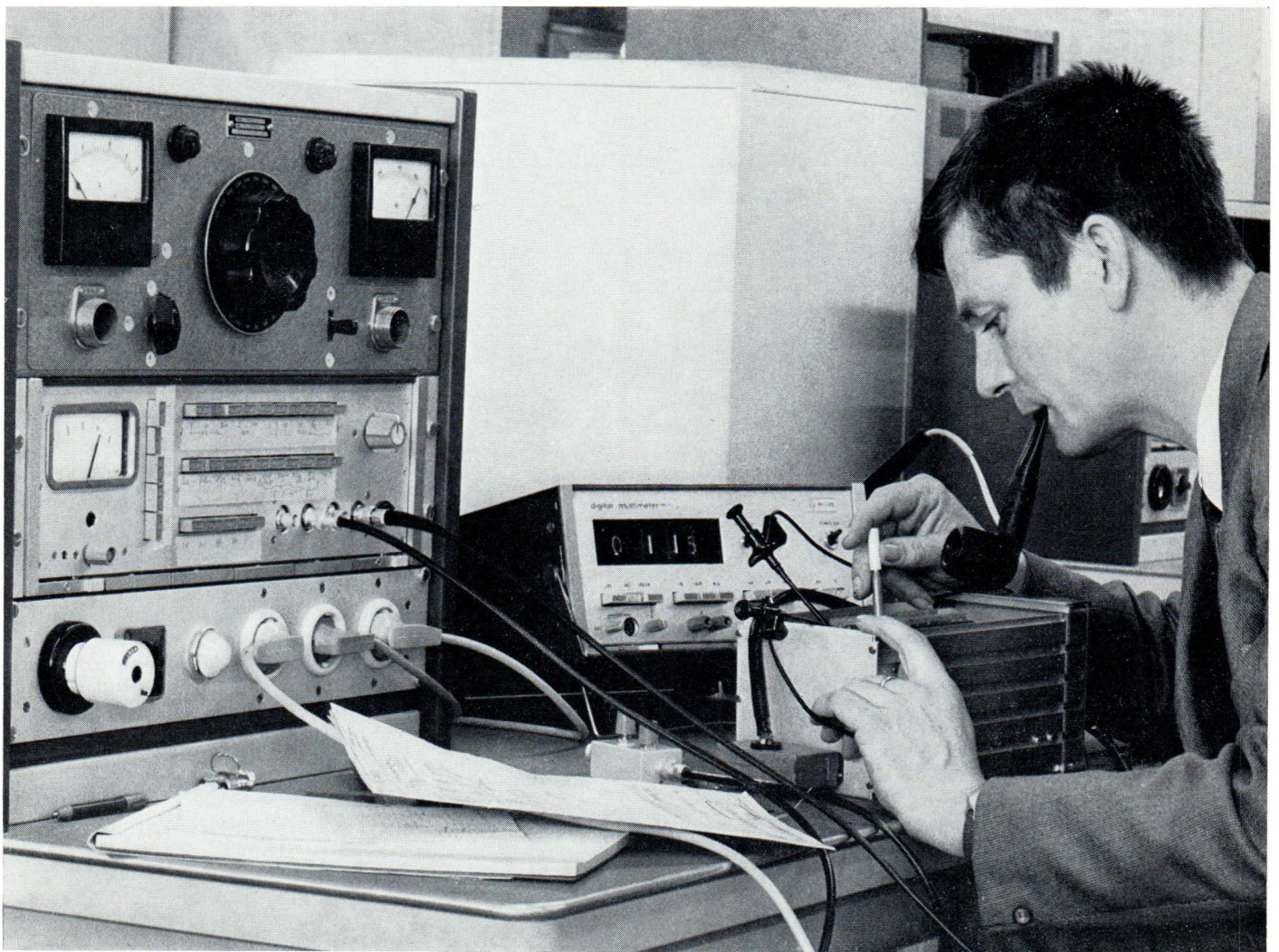
The $-2070V$ supply is stabilized by adding a correcting positive voltage V_c . To this end

the secondary winding of the HT transformer. Depending on the sign of the voltage deviation, transistor TS_4 will become more or less conducting, thus increasing or decreasing the correcting voltage.

The negative high voltage can be adjusted with the aid of R_4 . Diodes D_2 and D_3 limit the voltage excursion, in order to protect TS_4 . Diodes D_4 and D_5 serve the same purpose with respect to TS_3 .

Under normal conditions, diode D_6 conducts, offering a low impedance to the emitter circuit of transistor TS_4 . If the base potential rises too much the diode becomes cut off and the current excursion of TS_4 is limited by resistor R_5 .

The overload protection of the HT generator is controlled by transistor TS_5 . During normal operation TS_5 is cut off, because its base is kept negative with respect to its emitter. This is realized with the aid of



Calibration of test equipment for the PM 3370

the $-2070 V$ supply is related to the stabilized $+200 V$ supply by means of the network R_2 - R_3 - R_4 . The voltage at the point between R_2 and R_3 is applied to the difference amplifier TS_3 , TS_3^1 (dual transistor). The output voltage of this amplifier is passed to transistor TS_4 , which has its collector connected to the lower end of

network $R_6 - R_{10}$, which is dimensioned so that the potential of the point between R_7 and R_8 is so negative that diode D_7 is conducting. Under overload conditions, the $-1900 V$ supply will become less negative. As a result, diode D_7 will be cut off and the base potential of TS_5 can become positive enough to bottom the transistor. In this way the base potential of transistor TS_1 and TS_2 becomes negative and the oscillator ceases operation. The negative

high voltage then drops away, keeping TS_5 bottomed. This way of blocking TS_1 and TS_2 prevents excess heat dissipation in these transistors. C_4 prevents bottoming of TS_5 during the gradual building up of the $-1900 V$ after the apparatus is switched on.

The calibration generator



The calibration voltage for the PM 3370 is provided by an emitter-coupled multivibrator (TS₁ and TS₂; see fig. 1).

The emitter resistors of TS₁ and TS₂ are formed by the transistors TS₃ and TS₄, each ensuring a constant current. The value of this current can be adjusted by means of the voltage divider in the common base circuit of TS₃ and TS₄.

$R_1 = R_2$, which means that the currents through TS₃ and TS₄ are equal. A symmetrical square-wave voltage is thus produced at the output, the frequency of which is independent of the current chosen.

The frequency of the square-wave voltage is in fact determined by the product of the coupling capacitance C and the collector resistance of TS₁,

$$R = \frac{R_3 R_4}{R_3 + R_4}$$

The amplitude of the square-wave output voltage is given by the product of the resistance R₅ and the sum of the currents through TS₃ and TS₄.

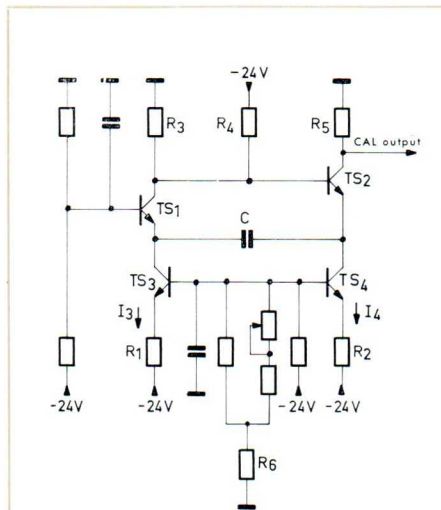
The behaviour of this circuit may be explained as follows (see fig. 2).

Let us suppose that TS₂ is conducting, and TS₁ cut off. The current through TS₂ will then be equal to I₃ + I₄. Capacitor C is charged with a constant current I₃. As a result, V_{E1} decreases linearity with time. When V_{E1} becomes more negative than V_{BE1}, TS₁ becomes conducting and the multivibrator switches over.

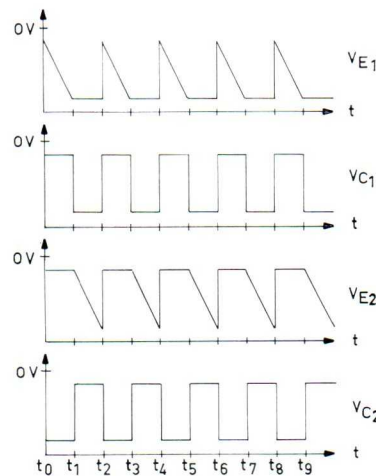
Capacitor C now discharges with the constant current I₄, as a result of which V_{E2} decreases linearity with time until TS₂ starts conducting again.

The emitter voltage V_{E2} now jumps by the same amount as the collector voltages of TS₁, i.e. R(I₃ + I₄). Since V_{E2} alternately decreases and increases by the same amount, we can calculate the duration of one pulse, t₂ - t₁, from:

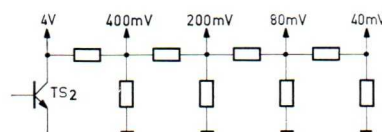
$$t_2 - t_1 = \frac{RC(I_3 + I_4)}{I_4}$$



1. Emitter-coupled multivibrator



2. Waveforms related to fig. 1



3. Step attenuator

Since the two currents are equal ($I_3 = I_4$), the period T_0 becomes: $4RC$ and the frequency:

$$f = \frac{1}{4RC}$$

In order to keep the current through TS₃ and TS₄, and hence the output voltage, constant at different temperatures, an NTC resistor R₆ is included in the common base circuit of TS₃ and TS₄. This compensates for the variation in V_{BE} of TS₃ and TS₄ due to temperature fluctuations.

The output voltages

A step attenuator is connected at the output of the calibration oscillator, allowing a number of different calibration voltages to be selected, viz: 4 V, 400 mV, 200 mV, 80 mV and 40 mV (see fig. 3).

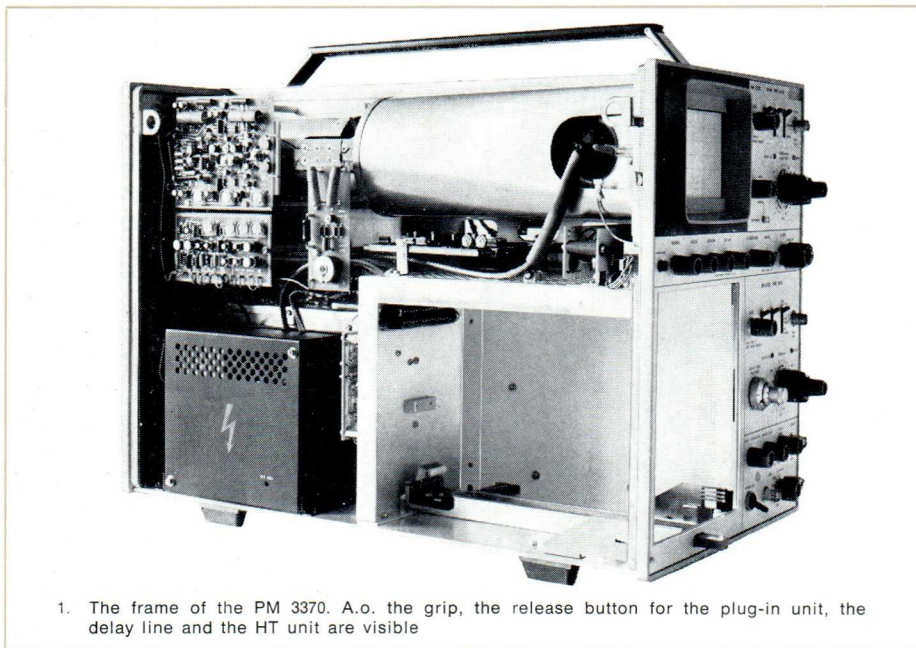
The output impedance is 50 ohm, except in the 4 V setting, when it is 500 ohm.

The above-mentioned values of the calibration voltages apply to the unloaded attenuator.

The impedance of 50 ohm has been chosen in view of the range of applications of the PM 3370, which includes a wide variety of measurements on low-ohmic HF circuits.

The calibration generator can also provide calibration currents as well as voltages. These currents are obtained by earthing the output socket which gives a calibration current of 8, 4, 1.6 or 0.8 mA, corresponding to the attenuator settings 400 mV to 40 mV. In order to ensure an accuracy of 1% for the output voltages and currents, metalfilm resistors with a tolerance of less than 0.25% are used in the attenuator.

Mechanical construction of the PM 3370



1. The frame of the PM 3370. A.o. the grip, the release button for the plug-in unit, the delay line and the HT unit are visible

General constructional details

The apparatus as a whole has been designed for maximum serviceability and strength. The frame consists of front and rear walls of die-cast aluminium, connected by the bottom plate, the middle partition with cage and an aluminium profile, fig. 1.

The front wall has a number of supporting points for the various sub-units involved; the text plate is mounted on its front side. The rear wall serves as a heat sink; it has been found that this design rendered a ventilator unnecessary, fig. 2. Two shock-absorbers are also mounted on the rear wall. The above-mentioned construction makes the frame very rigid, which is needed for the stability of the apparatus as a whole. The various sub-units which are mounted in the frame may all be regarded as electrically independent units; care was taken during the design stage to ensure this. Plug-in connections have been used as far as possible (even for the cable

forms), so that the sub-units can easily be removed from the frame; this greatly facilitates service work. The printed-circuit boards can also simply be pulled out of their connectors, for the same reason.

The side walls and top of the instrument are made in two halves, which can easily be removed as they are equipped with snap latches. No perforations have been made in the top cover, to reduce as far as possible the amount of dust and dirt getting into the apparatus.

The dimensions of the perforations in the side walls are fully in accordance with the safety regulations of the IEC.

The grip by which the oscilloscope can be carried is mounted in the aluminium profile. Slight pressure on this grip releases a pin which pushes the grip out of the profile so that it can be grasped. The space at the bottom left-hand side is for plug-in units; these units are automatically locked in position when they are pushed fully in. Simple pressure on a release button is

enough to unlock them and push them forwards so that they can easily be removed.

Power supply

The power-supply unit has been made in a number of parts, in the interests of serviceability and to get rid of the heat dissipated.

The elco (electrolytic capacitor) unit is made in two parts, one of which is mounted on a printed-circuit board which can simply be removed from the apparatus, while the other part is mounted on a bracket which can be rotated outwards for ease of access, fig. 3. From this elco unit we go to the transformer and a main distribution unit.

The transformer is fixed to the bottom plate and to the middle partition, so that it may be regarded as part of the frame; this arrangement also ensures efficient removal of the heat dissipated in the transformer. The main distribution unit, on which four connector blocks and two smaller distribution units are mounted, supplies the power to the various sub-units. Printed-circuit boards are inserted in the connector blocks; two of these carry part of the power-supply unit.

Main time base

The main time base is built as a rigid sub-unit, which can be removed as a whole from the apparatus, fig. 4. All connections to other units are realized by plugs. All the controls for this unit are mounted on its front plate. The time-base unit is fixed in place with four screws on the front wall of the oscilloscope and 1 screw in the middle partition.

Delayed time base

This has the same construction as the

main time base, fig. 5.

The delay line

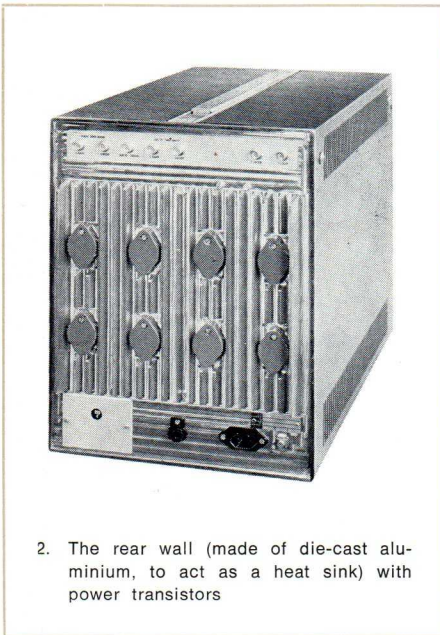
This is formed by a cable wound on two adjacent spools, so as to form a compact unit despite its length.

The HT unit

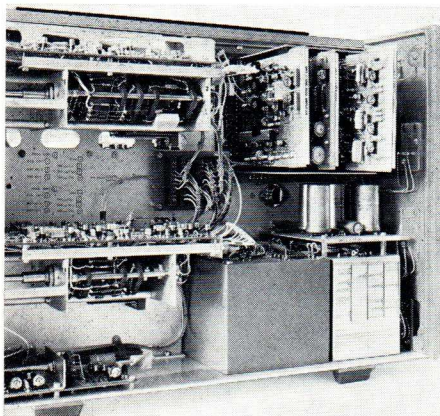
The HT unit is mounted on a printed-circuit board, the high-tension part of which is screened off by a plastic cap. This board is mounted in a tray with a removable cover for easy serviceability. The tray and its cover form a labyrinth seal, to provide good electromagnetic screening of the HT unit as a whole.

Suspension of the CRT

The CRT is given an elastic suspension. The front of the tube is enclosed between

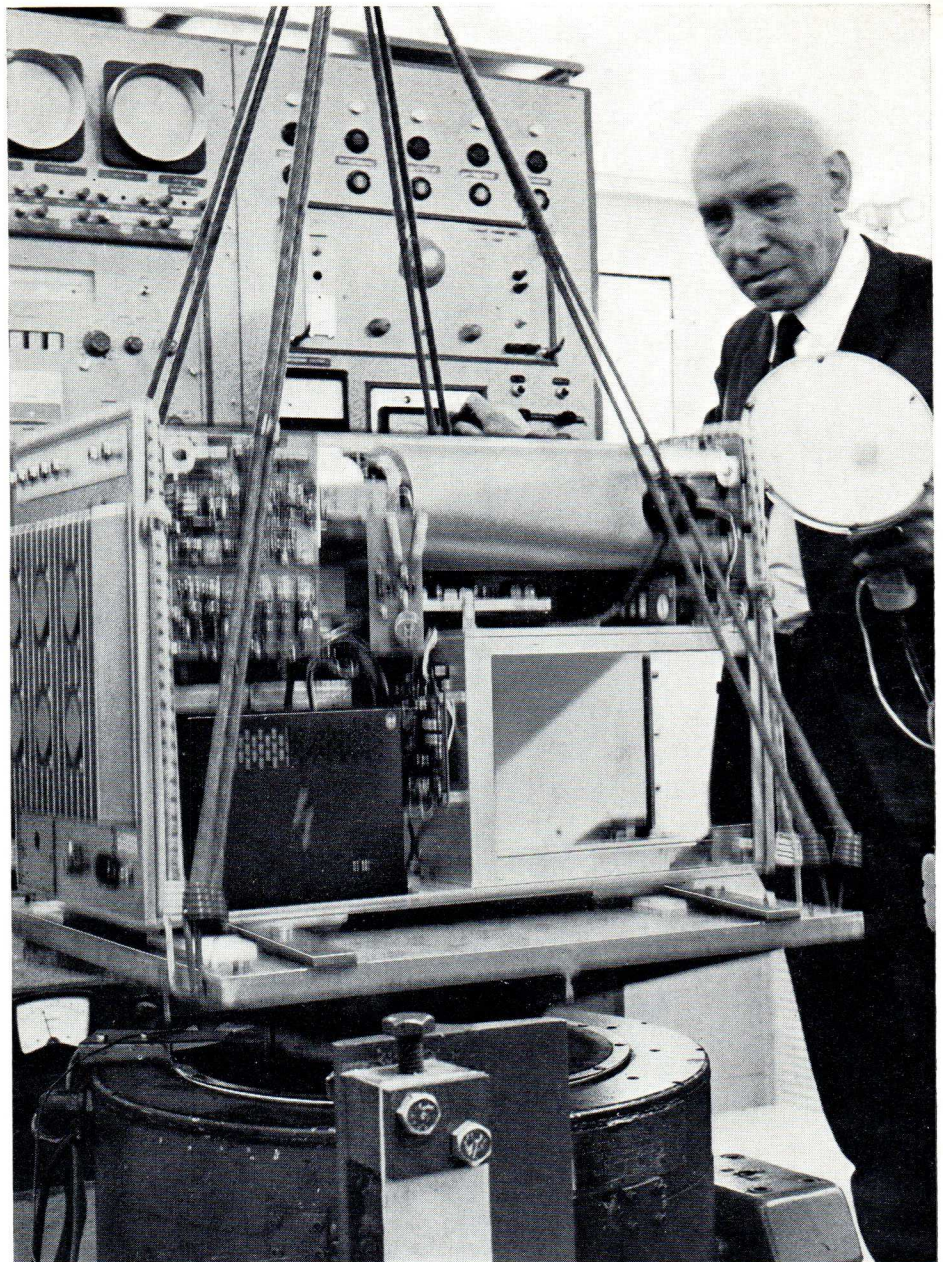


2. The rear wall (made of die-cast aluminium, to act as a heat sink) with power transistors

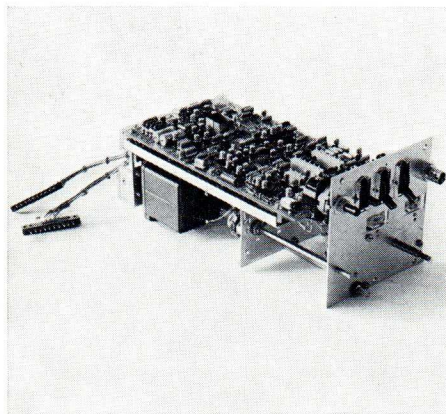


3. The oscilloscope opened, the elco unit is visible

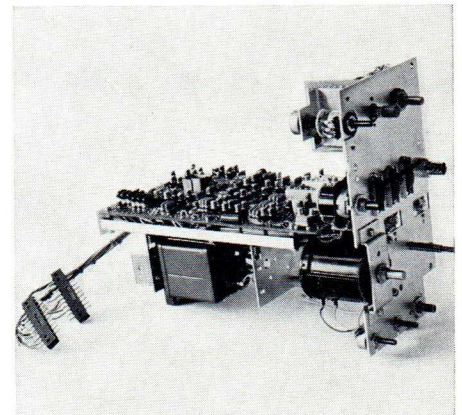
two plastic sheets and two adjustable springs. The plastic sheets have a double function: they provide the illumination of the internal graticule, and they act as shock-absorbers for lateral movement of the tube. The graticule can be adjusted in height by means of two screws of the upper adjustable spring. The rear end of the CRT is clamped in a holder with a



Vibration testing of the PM 3370



4. The main time base



5. The delayed time base

rubber collar. The CRT is screened off by a mumetal tube, the front end of which fits into the front wall of the oscilloscope, while the rear end rests on an insulated mounting on the tube holder. The latter is fixed to the rear wall by two screws. The screen of the CRT can if necessary be adjusted so as to lie flush against the front window by unscrewing the two last-men-

tioned screws and shifting the tube holder appropriately.

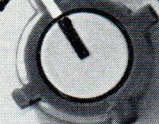
● PHILIPS PM 3372 DUAL TRACE Y AMPLIFIER RISE TIME ≈ 1.7 ns ●

POSITION

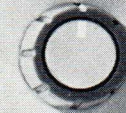


PULL TO INVERT A

CHOPP. ALT. A B ADDED



POSITION

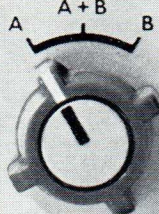


PULL TO INVERT B

GAIN

Gain knob for channel A with scale: .1, .2, .5, 1, 2, 5, 10, 20, 50. CAL. 1. V/cm 5. mV/cm 1.

TRIGG. A A+B B



GAIN

Gain knob for channel B with scale: .1, .2, .5, 1, 2, 5, 10, 20, 50. CAL. 1. V/cm 5. mV/cm 1.

A



1 M Ω
15 pF

AC
DC

t_r .35 μ s
 t_r 1.7 ns

IN
OUT

t_r .35 μ s
 t_r 1.7 ns

AC
DC

1 M Ω
15 pF

B



FEED THROUGH

FEED THROUGH



PROBE
POWER



CHECK
ZERO

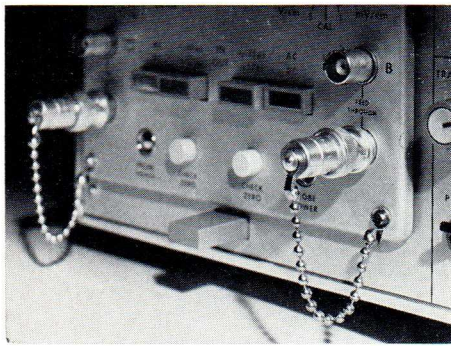


CHECK
ZERO



PROBE
POWER

Dual trace plug-in unit PM 3372



General description

The PM 3372 is a plug-in unit with two independent input channels for the oscilloscope PM 3370.

The signal in each channel can be displayed individually, or the two signals can be displayed simultaneously, either on a time-sharing basis $\pm A$ and $\pm B$ or algebraically added $[(\pm A) + (\pm B)]$.

A trigger signal can be derived from each channel, or from the sum $A + B$. The choice of the trigger source is independent of the position of the display mode switch. Furthermore, this plug-in unit features a high sensitivity combined with high speed, low noise and low drift.

A new approach to input circuitry and a bandwidth reduction switch contribute further to the versatility of this unit.

Operation

All controls of channels A are situated on the left side of the front panel, those of channel B on the right (see photo of control panel).

In the centre we find the display mode selector and the trigger selector for the two channels. As can be seen from the photo, each channel is equipped with two input sockets which are interconnected.

If the $50\ \Omega$ pad is not applied to one of these input sockets, the input impedance is $15\ \text{pF}$ in parallel with $1\ \text{M}\Omega$. The two sockets can also be used in a "feed-

through" mode, preferably in a $50\ \Omega$ coaxial system.

The probe power output adjacent to the input connectors enables optional active devices, such as FET probes, SYNC separators and current probes, to be used.

Block diagram

For block diagram see fold-out page.

An input signal applied to channel A is passed via a coupling capacitor C_1 (bypassed in DC operation by a dry reed switch) to the 3-step $1\ \text{M}\Omega$ decade attenuator, operated by dry reed switches. These reed switches are programmed by a rotary switch (V/cm SELECTOR), which in fact programmes a total of four attenuators and DC LF and HF amplifiers to give 12 different deflection coefficients ranging from $1\ \text{mV/cm}$ to $5\ \text{V/cm}$. For further details, see table on page 32. The CHECK ZERO button, when depressed, overrides the programme of the rotary switch and disconnects the signal from the amplifier, the input of which is earthed.

After the impedance converter (consisting of a source-follower and an emitter-follower), the HF and the LF components of the signal are guided along different paths.

The LF path starts with a DC-LF amplifier with differential input and single-ended output. The signal from the impedance converter is applied to the inverting input. The output of the DC-LF amplifier is passed to the feedback attenuator, the attenuation factor of which is such that its output signal is an inverted replica of the amplifier input signal. The input signal and the inverted amplifier output signal thus cancel out at the junction of the two equal resistors R_1 and R_2 , but the drift voltage caused by the DC-LF amplifier only suffers twofold attenuation. This drift voltage if fed into a very stable DC chopper amplifier and then presented to the non-inverting input of the DC-LF amplifier, giving a considerable reduction in the overall drift (see „Drift correction"). In AC operation, capacitor C_2 is inserted to counteract a DC shift due to leakage current from C_1 ,



when the input signal contains a large DC component.

In order to minimize noise, the amplification factor of both the LF and the HF amplifier (to be discussed later) is decreased in the first four steps of the V/cm selector switch. To keep the voltage between R_1 and R_2 virtually zero for the signal required, the feedback attenuator is stepped down accordingly. The DC-LF amplifier is followed by a $500\ \Omega$ attenuator with attenuation ratios of $1:1$, $1:2$ and $1:5$, to subdivide the decimal steps of the $1\ \text{M}\Omega$ attenuator (except for the first four steps of the V/cm selector, as mentioned above). The HF path is built up in the same way, except that the drift elimination is omitted, because the HF amplifier is AC coupled. The 1—2—5 attenuator is low-ohmic and precedes the amplifier to reduce the dynamic range required. The LF-DC part of the amplified signal is passed through an impedance converter and then recombined with the HF part by means of a filter consisting of R_3 and C_3 .

The recombined signal is applied to the trigger take-off circuit, which supplies the signal current for the trigger amplifier and the voltage for the continuous control.

The input of the trigger amplifier comprises a gate circuit controlled by the three-position trigger selector switch on the front panel. This circuit allows triggering by A, B, or A+B, signals.

The triggering signal selected is amplified and applied to the PM 3370 oscilloscope. The signal fed to the continuous control is passed to the electronic switch via a voltage-to-current converter together with a voltage from the POSITION potentiometer. Control signals for the electronic switch are derived from the electronic switch driver, which is programmed by the display mode switch on the front panel.

The following modes are available: CHOPped, A, ALTerated, B, ADDED Beam suppression (blanking) pulses re-

quired during electronic switching in the CHOPP. mode are also supplied by the driver stage. After passing the electronic switch, the signal is fed to the signal inverter circuit, which is controlled by a pull-push switch incorporated in the POSITION front panel control. The signal is applied to the current input (input I) of the main frame (PM 3370).

Input circuit

The input signal can be applied in three different ways:

1. via socket 1 (or socket 2); $Z_i = 1 \text{ Mohm} // 15 \text{ pF}$
2. via socket 1, with socket 2 terminated by 50 ohm; $Z_i = 50 \text{ ohm}$.
3. in via socket 1 and out via socket 2 ("feed-through").

The input connectors are followed by a coupling capacitor (C_1). When the AC/DC switch is set to DC, this capacitor is bypassed by a reed relay (RE_1). R_1 and R_2 serve to prevent resonances in the circuit formed in this way.

A high-ohmic attenuator consisting of three sections with attenuation factors of 0.8, 0.08 and 0.008 is connected between C_1 and the gate of TS_1 . (See fig. 1). Each section of the attenuator is selected by means of two reed relays. An attenuation of 0.8 is chosen instead of 1 so as to allow R_3 to be given a high value (200 kohm) while C_2 remains low. This ensures that the gate current of TS_1 will be limited if the input voltage is too high. In order to protect TS_1 against the charging current of C_2 , which flows via the drain-gate junction,

a large resistance (R_4) is placed in the drain of TS_1 .

As long as the signal voltages remain within the permissible driving range, D_1 and D_2 ensure that the drain circuit is low-ohmic. D_3 ensures that the maximum permissible emitter-base voltage of TS_2 is not exceeded if highly negative voltages are applied. The above-mentioned measures give adequate protection against over-loading of relatively short duration (a few seconds). If a negative over-voltage is applied to the input for a long time, then RE_2 and RE_4 are opened by a protection circuit comprising R_5 , TS_3 and TS_4 as long as the overloading lasts. This gives foolproof protection of the input right up to $\pm 400 \text{ volt}$. R_6 leads the input signal to a comparison circuit which forms a virtual earth point for the signal (because of the drift feedback). R_7 can supply a minute current (from 0 to 1 nA) for compensation of the leakage current of TS_1 .

TS_1 is connected as a source-follower, and is directly followed by the emitter-follower TS_2 . The emitter circuit contains the 1-2-5 attenuator consisting of R_8 , R_9 , R_{10} and the relays RE_8 , RE_9 and RE_{10} . This attenuator is only used for high frequencies. The LF DC signal is taken directly from the emitter. The 1-2-5 attenuator for the LF DC signal comes after the corresponding amplifier.

Split-band amplifier

General considerations

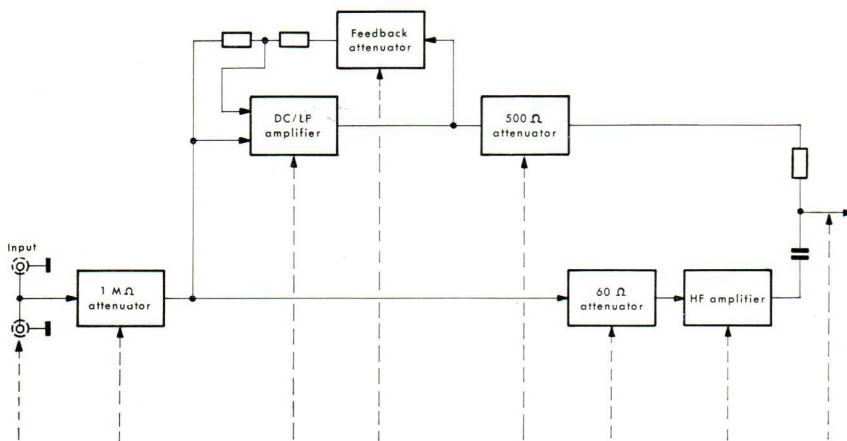
A split-band amplifier provides separate amplification channels for the HF and LF-DC components of the signal, so that each component is amplified under optimum conditions.

In the HF branch, this involves the choice of high-quality T-pack encapsulated HF transistors with high f_T , which operate with high collector currents. In the LF-DC branch, transistors have been chosen with moderate f_T but high h_{fe} and good $1/f$ noise properties; these are used at rather low collector currents. The circuit has also been designed to give a good balance between the collector voltage of the transistors and the supply voltage, to minimize changes in collector dissipation during excitation.

In the first four steps of the V/cm switch, all attenuators remain in the minimum-attenuation position; the deflection coefficients in question (1, 2, 5 and 10 mV/cm) are realized by decreasing the amplification factor of the amplifiers.

The HF branch

The HF amplifier is built up of three identical stages, each consisting of one transistor in the common-emitter configuration and one in the common-collector configuration (fig. 2). Each stage has two different collector loads; when one is in use, the



1	mV/cm	0,8 x	— 70 x	1/70 x	1 x	1 x	— 70 x	56 mV/cm
2	mV/cm	0,8 x	— 35 x	1/35 x	1 x	1 x	— 35 x	56 mV/cm
5	mV/cm	0,8 x	— 14 x	1/14 x	1 x	1 x	— 14 x	56 mV/cm
10	mV/cm	0,8 x	— 7 x	1/7 x	1 x	1 x	— 7 x	56 mV/cm
20	mV/cm	0,8 x	— 7 x	1/7 x	0.5 x	0.5 x	— 7 x	56 mV/cm
50	mV/cm	0,8 x	— 7 x	1/7 x	0.2 x	0.2 x	— 7 x	56 mV/cm
0,1	V/cm	0,08 x	— 7 x	1/7 x	1 x	1 x	— 7 x	56 mV/cm
0,2	V/cm	0,08 x	— 7 x	1/7 x	0.5 x	0.5 x	— 7 x	56 mV/cm
0,5	V/cm	0,08 x	— 7 x	1/7 x	0.2 x	0.2 x	— 7 x	56 mV/cm
1	V/cm	0,008 x	— 7 x	1/7 x	1 x	1. x	— 7 x	56 mV/cm
2	V/cm	0,008 x	— 7 x	1/7 x	0.5 x	0.5 x	— 7 x	56 mV/cm
5	V/cm	0,008 x	— 7 x	1/7 x	0.2 x	0.2 x	— 7 x	56 mV/cm
Deflection coefficient	Ratio between signal in and signal out							Deflection coefficient

other is blocked by a diode, and *vice versa*. This allows a choice to be made between two different amplification factors. In the limited-bandwidth setting, the third stage has no collector supply at any load, and the emitter output is blocked by D.

The LF branch

This amplifier starts with a differential stage; one input receives the signal from the source-follower circuit, and the other that from the drift-compensation circuit, see fig. 3.

The deflection coefficients of 1, 2, 5 and 10 mV/cm are realized by switching the appropriate resistor between the emitters of TS₁ and TS₂.

The second stage of the LF-DC amplifier has an amplification factor which can be varied slightly by means of R₁, to allow

perfect matching between the feedback attenuation and the amplification of this amplifier. The connection between the two stages is earthed via a capacitor if the bandwidth push-button switch (t_r .35 μ s; t_r 1.7 ns) on the front is depressed.

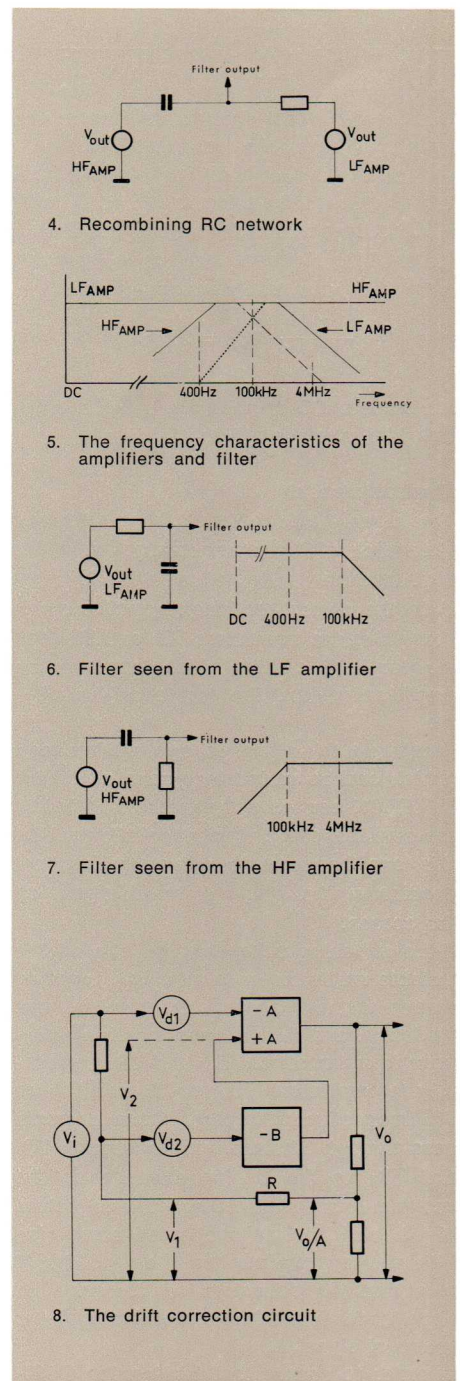
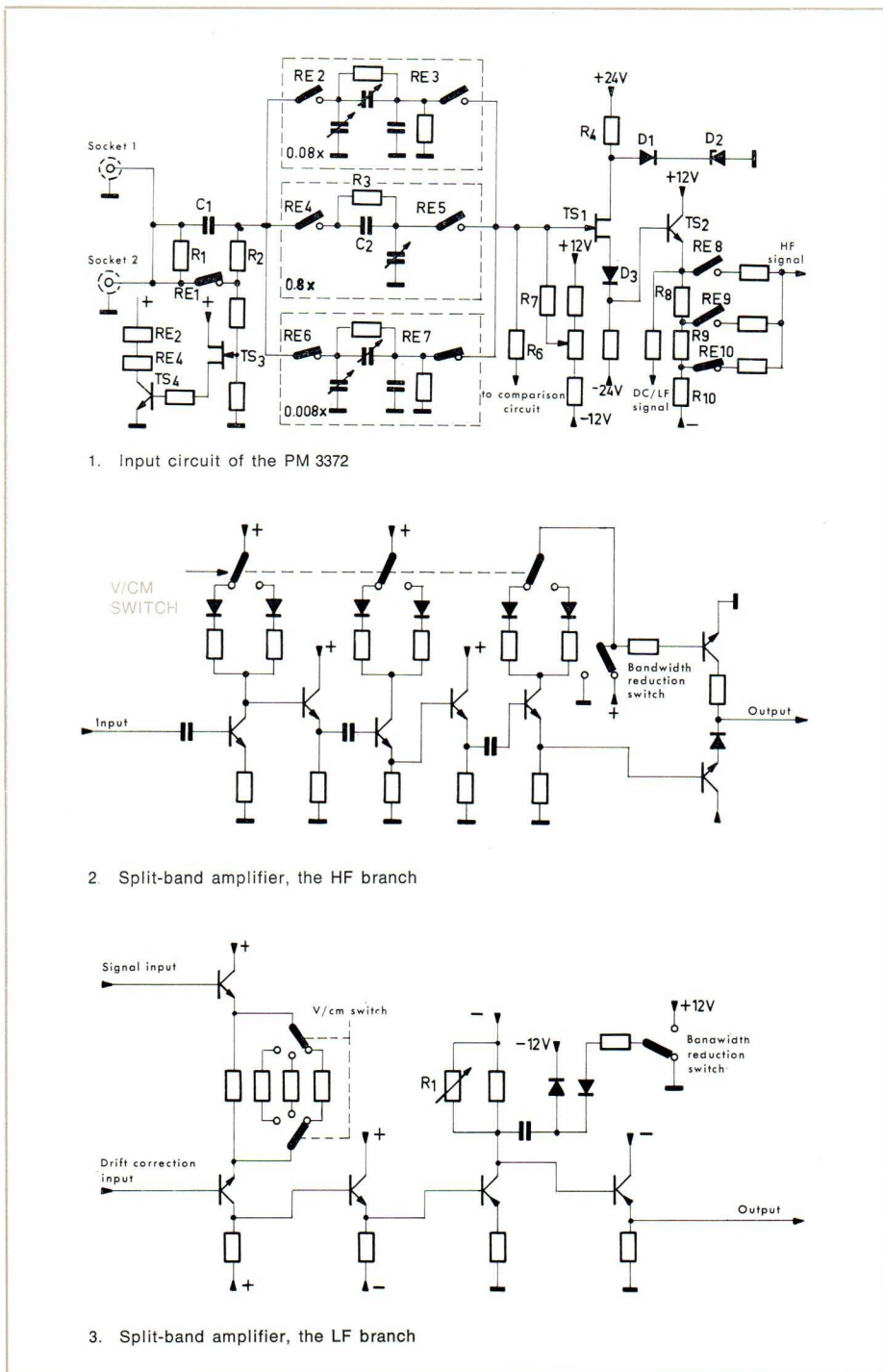
Combining network

The outputs of the HF and LF amplifiers are connected by an RC network (fig. 4). In fig. 5 the frequency characteristics of the amplifiers are represented by the solid lines. The broken lines represent the output voltage of the RC filter, fed from the LF amplifier, supposing the HF amplifier to be shorted. The dotted line gives the filter voltage as seen from the HF amplifier only. It is obvious that in the middle of the passband (i.e. a wide range around 100 kHz), the outputs of the HF and LF

amplifiers are equal, both in magnitude and phase. This means that no current flows in the RC network, and the output voltage of this network is the same as that of either amplifier.

In the frequency domain below 400 Hz, where the output of the HF amplifier decreases with the frequency, the filter output voltage is practically equal to that of the LF amplifier, the contribution of the filter being negligible (fig. 6).

On the other hand, above 4 MHz the filter output is determined by the HF amplifier; in this case, the LF amplifier merely serves as a ground connection for filter resistor R (fig. 7). If the output voltage of one of the amplifiers contains some spurious signals (e.g. 1/f noise from the HF amplifier) besides the desired signal, the 100 kHz filter comes into action.



Spurious signals from the LF amplifier will be filtered according to the characteristic given by the broken line in fig. 5 (higher attenuation at higher frequencies). Spurious signals from the HF amplifier will be filtered the other way round (dotted line in fig. 5).

Drift correction

The drift occurring in the vertical amplifier (gain—A) is corrected with the aid of an indirect DC amplifier (gain—B) consisting of a chopper which converts the DC voltage applied to its input into a square-wave voltage, an AC amplifier and finally a synchronous demodulator which turns the amplified AC voltage back into DC.

The arrangement is given in fig. 8. We regard the drift voltages from the vertical and chopper amplifiers as being due to DC voltage sources at the input (V_{d1} and V_{d2}). We then find:

$$-\frac{V_o}{A} = V_1 + \frac{1}{1 + \frac{B}{2}} V_{d1} + \frac{2}{1 + \frac{2}{B}} V_{d2}$$

The resulting drift voltage referred to the input is given by the last two terms.

We see that the drift of the vertical amplifier is reduced by a factor $(1 + B/2)$, with $B = 1000$ as in the PM 3372 this is thus a factor 500. However, we now have an extra drift due to the chopper amplifier,

$$\text{given by } \frac{2}{1 + \frac{2}{B}} V_{d2} = 2 V_{d2}$$

An indirect DC amplifier is used to keep this drift voltage negligible.

Electronic switch

The signal voltage from the continuous attenuator of the channel-A pre-amplifier is passed via emitter-follower TS_2 (see fig. 9) to the current-feedback amplifier stage containing the transistors TS_3 and TS_4 , the resistor R_2 , R_4 and R_5 (which determine the amplification) and the correction filter $C_1 - R_3$.

Apart from this signal voltage, a DC voltage (which can be varied by means of R_1 , the POSITION control potentiometer) is fed to the amplifier stage via emitter-follower TS_1 : this can be used to adjust the vertical position of the image of the A signal on the screen.

If the switching voltage a from the driving bistable multivibrator is relatively negative, diodes D_2 and D_3 will be cut off, and diodes D_1 and D_4 will pass the signal currents from transistors TS_3 and TS_4 respectively to transistors TS_5 or TS_6 and TS_8 or TS_7 respectively. Whether these currents pass through TS_5 and TS_8 or through TS_6 and TS_7 , to the Y plug depends on the position of switch S_2 . Since the collectors of TS_6 and TS_7 are connected cross-wise with the appropriate terminals of the Y plug and the collectors of TS_5 and TS_8 directly, S_2 determines the polarity of the image of signal A.

In order to ensure that turning the POSI-

TION control to the right always causes the image to shift upwards, no matter what the position of S_2 , things are arranged so that S_2 also determines the polarity of the voltage across potentiometer R_1 .

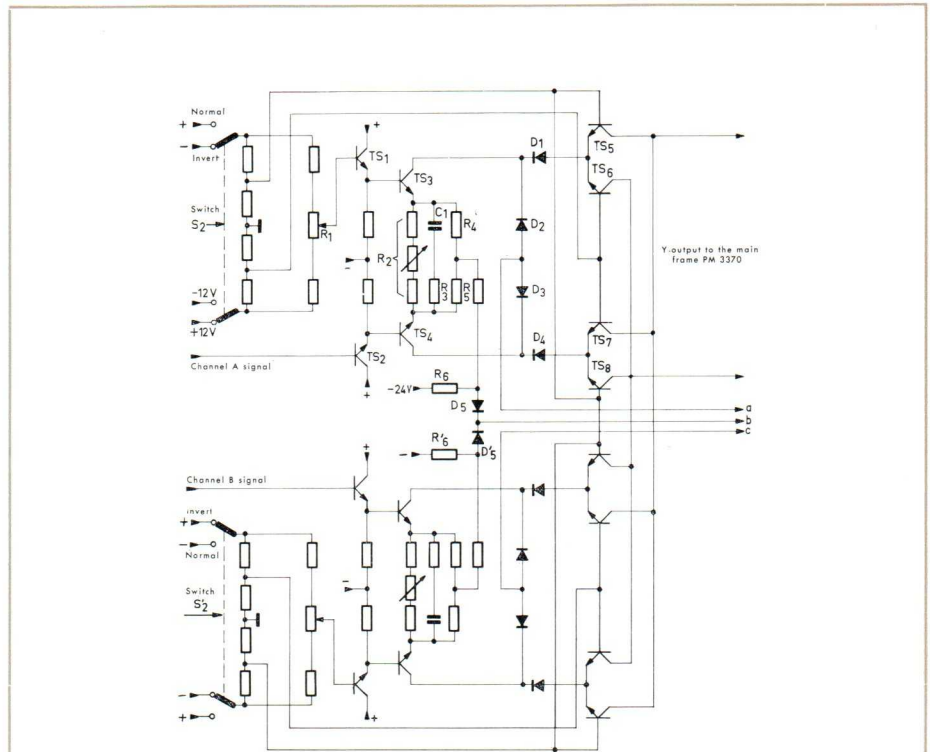
If the switching voltage a is relatively positive, the signal currents from TS_3 and TS_4 will be led off via D_2 and D_3 , and D_1 , D_4 and TS_5 to TS_8 will be cut off.

In general, switching voltage c will be relatively negative under these conditions, so that the signal currents from the B channel will be fed to the output terminals. This B channel is identical with the A channel. However, there is one situation — with the mode switch at ADDED — where switching voltage a is relatively negative as well as switching voltage c . The signal currents from both channels are then added. The setting of the switches S_2 and S_1 then determines whether the image on

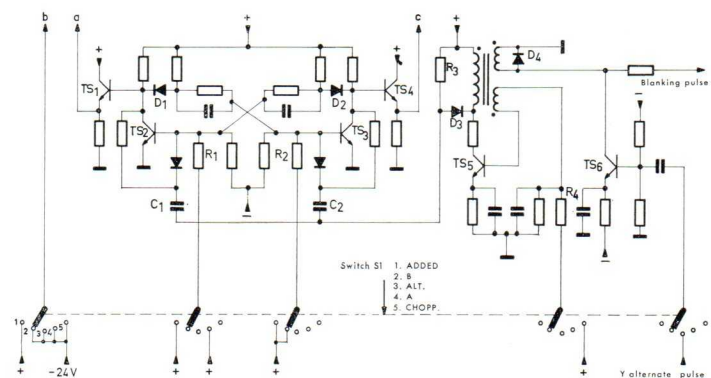
the screen is that of $A+B$, $A-B$, $-A+B$ or $-A-B$.

In order to ensure that the DC components of the output signals have the same value when the signal currents are combined in this way as they do when signal A or B is displayed separately, the total DC current through TS_3 and TS_4 , and through the corresponding transistors in the B channel, is halved. For this purpose, the b signal is made so positive that diode D_5 and D_5' are cut off, as a result of which the resistor R_6 and R_6' are switched into the emitter supply leads. In all other positions of the mode switch, where the signal currents from the two channels are never passed simultaneously, the voltage on the b line is $-24V$, so that R_6 and R_6' are short-circuited.

Fig. 10 shows how the control signals a , b and c are produced. The control circuits



9. Circuit diagram of the electronic switch



10. Circuit of the electronic switch driver

shown in this figure are programmed by the above-mentioned mode switch (S_1). This switch has 5 positions, viz.:

- A *only the A signal is shown on the screen; for this purpose, a is relatively negative, c is relatively positive and b is negative (-24 V)*
- B *only the B signal is displayed; for this purpose, c is relatively negative, a is relatively positive and b is again negative;*
- ADDED *the two signals are added; for this purpose, a and c are both made relatively negative, and b relatively positive;*
- CHOPP (abbreviation for "chopped"): *the A and B signals are displayed in turn, at intervals of $1/2\mu\text{s}$, by using square-wave voltages inverted with respect to one another for a and c. Line b is again relatively negative.*
- ALT (abbreviation for "alternate"): *here again, the two signals are displayed alternately by using square-wave voltages for a and c; but the alternation no longer occurs at a fixed frequency, as with chopping, but always during the flyback of the time base. The control signal here is the Y-alternate pulse from the time-base generator of the PM 3370. The voltage of the b line is again -24 V .*

Control signals **a** and **b** are delivered via emitter-followers TS_1 and TS_4 by a bistable multivibrator with transistors TS_2 and TS_3 . The diodes D_1 and D_2 in the collector circuits ensure that the positive-going edges of the **a** and **c** signals will be steep enough: in the conventional flip-flop circuit without these diodes, the positive-going edges are less steep because the coupling capacitor to the base of the one transistor has to be charged via the collector resistor of the other.

It is important to ensure that both edges of the switching signals **a** and **c** are steep enough (and that they occur at the same time), in order to make the switching time from channel A to B and **vice versa** as short as possible.

During the switching time, no meaningful information can be displayed on the screen; the electron beam is therefore blanked during this period, but this should be kept as short as possible to reduce intensity losses.

The bistable multivibrator is controlled by S_1 via resistors R_1 and R_2 , except when S_1 is set to ALT or CHOPP. With S_1 at A, the transistor TS_2 of the bistable multivibrator is cut off, and TS_3 is conducting; with S_1 at B, the position is reversed; and with S_1 at ADDED we get the situation with both TS_2 and TS_3 conducting (and bottomed). With S_1 at ALT or CHOPP the bistable multivibrator is controlled by the blocking oscillator (transistor TS_5) via the counting circuit with C_1 and C_2 . The bistable multivibrator is switched at the negative-going

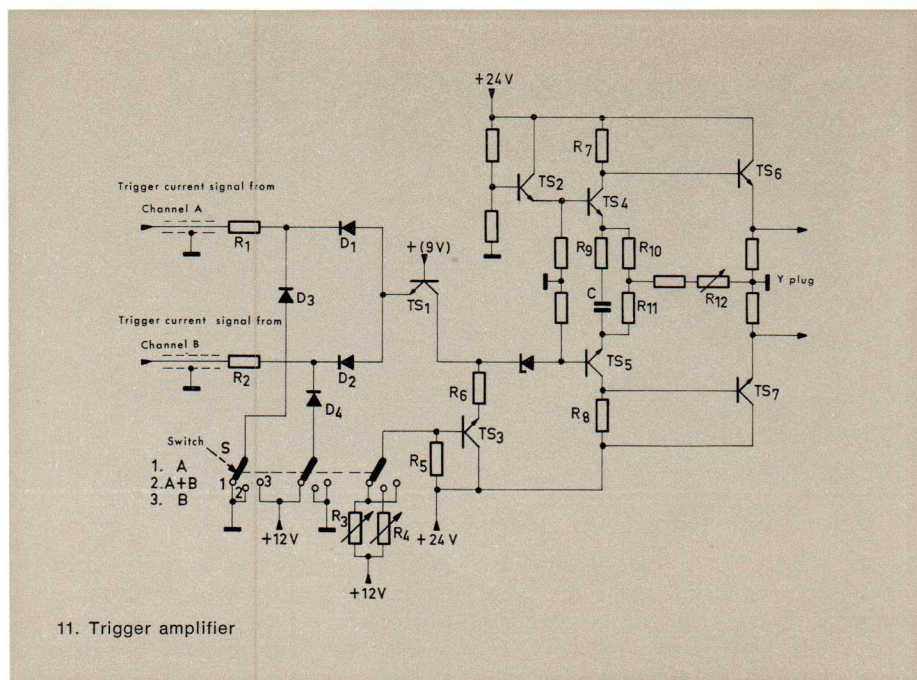
edges of the pulses produced by the blocking oscillator.

With S_1 at CHOPP the blocking oscillator is made free-running by supplying base current to TS_5 via R_4 and S_1 . With S_1 at ALT, on the other hand, TS_5 only becomes conducting (after which it is bottomed by the cumulative effect of the blocking-oscillator circuit) when it is triggered by the Y-alternate pulses amplified by transistor TS_6 .

The blocking-oscillator pulses are also used for the blanking of the electron beam. This is important to prevent blurring of the image due to the edges of the switching pulses on switching from A to B or **vice versa** when S_1 is at CHOPP. Suitable choice of the time constants of the blocking-oscillator circuit ensures that the blanking pulse covers this switching time adequately. Since the output signal from

The desired signal, which is selected by the TRIGGER SOURCE switch S , is fed via the common-base transistor TS_1 to the amplifier with transistors TS_4 to TS_7 . In order to make this possible, two gate circuits are controlled by S . With S at "A", diodes D_1 and D_4 are conducting, and D_2 and D_3 are cut off. The trigger-current signal from channel A is thus passed to TS_1 , while that from channel B is led off to the supply line via D_4 .

The reverse situation occurs with S at "B", when diodes D_2 and D_3 are conducting. In the middle position "A+B", diodes D_1 and D_2 conduct, while D_3 and D_4 are cut off. Both trigger signals are now applied to TS_1 , and their sum is amplified. This setting has the advantage that only one of the signals need be present, if it does not matter which of the signals does the triggering.



11. Trigger amplifier

the electronic switch is further delayed by the vertical delay lines of the PM 3370 before being displayed, the blanking pulse also passes through a delay line in the PM 3370 before being fed to the cathode-ray tube together with the other blanking signals.

The positive back swings produced after the negative blocking-oscillator pulses in the collector circuits of TS_5 should be got rid of. This is done by means of diode D_4 for the blanking pulse, and diode D_3 and resistor R_3 for the control of the flip-flop. If these backswings were not eliminated, those associated with the blanking pulse would give rise to an undesirable increase in intensity for some time after switching over, while those associated with the flip-flop control pulse could give rise to double triggering.

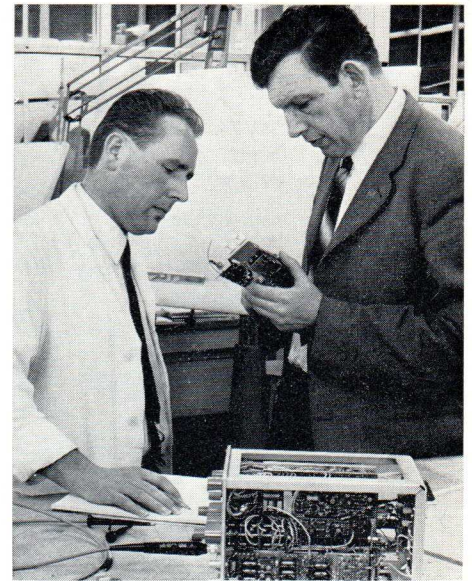
Trigger amplifier

Each of the pre-amplifiers of channels A and B delivers a trigger-current signal, see fig. 11.

The resistors R_1 and R_2 serve as part of the termination of the connecting cables. Since both trigger signals are of course superimposed on a DC component, provision must be made to prevent offset of the amplifier when S is set to "A+B". This is done by making S increase the base voltage of TS_3 in this setting, so that the extra voltage drop across R_6 is compensated for. In positions A and B, R_3 is used to equalize the output DC levels of the trigger amplifier, and in position A+B R_4 is used for this purpose.

This amplifier consists of a current-feedback stage with the transistors TS_4 and TS_5 , the amplification of which is mainly determined by R_7 , R_8 , R_{10} and R_{11} , and two emitter-followers TS_6 and TS_7 . C and R_9 provide a correction for the amplification at high frequencies, and R_{12} serves to adjust the DC voltage level at the two outputs to the proper value.

Mechanical construction of the PM 3372

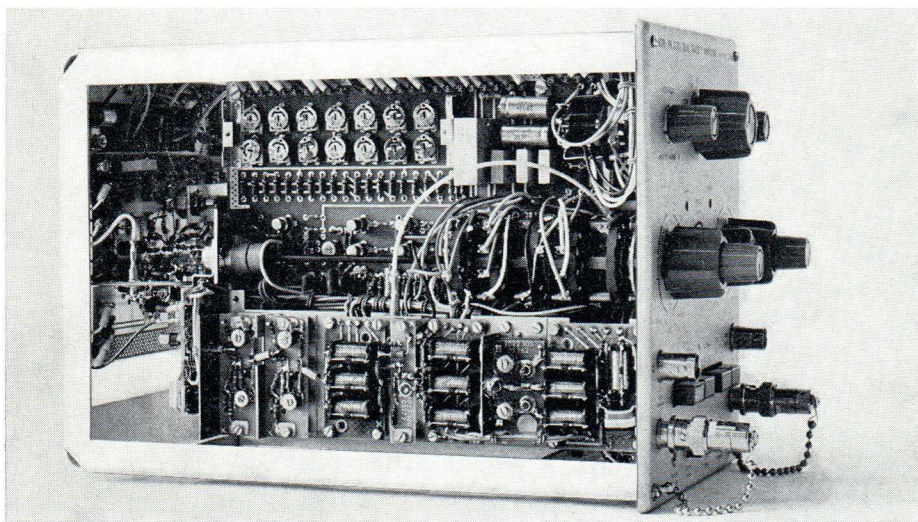


driver is mounted centrally in the apparatus, in the form of a printed-circuit board which can be simply inserted into its connector block. This sub-unit is completely screened off by a tray with a lid, the latter also serving as an earth connection between the front and rear of the apparatus. Under this tray is a metal plate which is formed so as to protect the printed-circuit board against falling out and against lateral forces.

The asymmetrical arrangement of the connector block ensures that the printed-circuit board cannot be inserted the wrong way round.

The design of this plug-in unit was based on the idea that the circuitry should as far as possible be divided up into a number of functional units.

All the above-mentioned considerations concerning the production and serviceability of the apparatus have not been allowed to detract from the ease of operation, which is ensured by the ergonomic design of the controls.



PM 3372 is designed to give not only optimum electronic functioning but also maximum serviceability and mechanical strength. The basic chassis is formed by a frame punched and bent out of aluminium sheet, and an aluminium front wall. At the rear of the frame is a multi-pin plug, and an earthing pin; the latter also ensures that this unit cannot be inserted into the PM 3330 main frame by mistake.

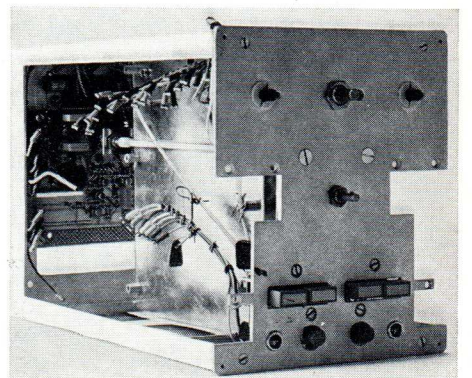
The front wall is provided with a number of supporting points for the text plate, for various brackets and for the different sub-units involved. The holes cut in the front

wall allow the unit of channel A and that of channel B to be taken out of the apparatus once the text plate has been removed. Both units consist of a number of smaller ones mounted on a main bracket. In order to increase the serviceability of the apparatus, a number of components which have to be connected to the main bracket are first mounted on a smaller support, which is in its turn screwed on to the main bracket. Both units (A and B) are screened off with a metal cap.

The sub-unit comprising the drift-correction amplifiers and the electronic switch

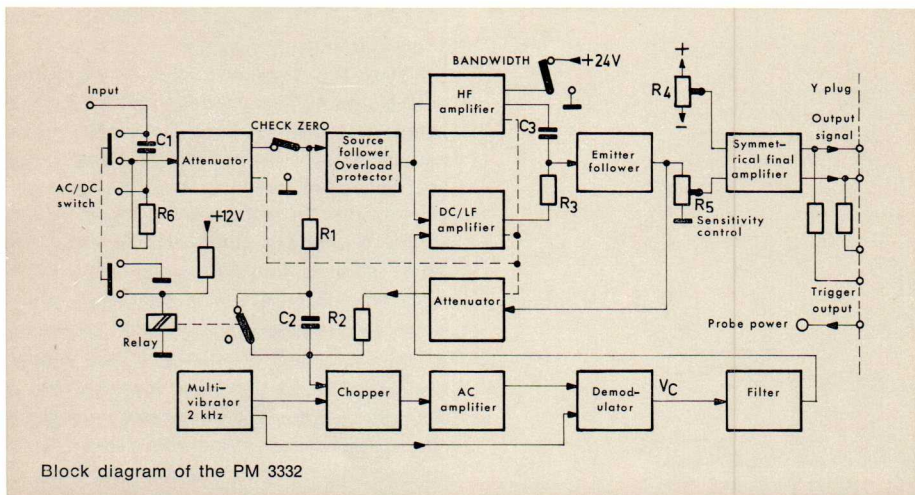
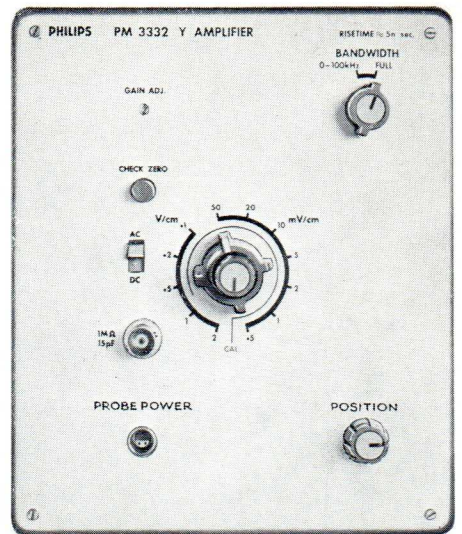


The materials planning and ordering department an essential link in the chain



Electronic measuring and microwave notes

High sensitivity low-drift amplifier PM 3332*



Block diagram of the PM 3332

The PM 3332 is a low-noise single-trace vertical amplifier with chopper stabilization and split-band amplification. The overall rise time at full sensitivity (0.5 mV/cm) is 7 ns and the drift less than 1 cm/week. For details of the chopper stabilization and split-band amplifier, see the description of the PM 3372 unit in this issue.

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 signals "riding" on a DC level of at most 400 V. R_6 is included to discharge C_1 when the AC/DC switch is changed from AC to DC. Without this, 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 high-quality capacitor was chosen for C_1 , the leakage resistance is such, that 400 V DC in the AC position would cause a deflection of several cm 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 . After the AC/DC switch, the frequency compensated high-ohmic attenuator is encountered. In the 4 most sensitive positions this

attenuator gives no attenuation. Via a source follower with overload protection the signal passes 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 (in 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 possible, even at full sensitivity. After the filter, C_3R_3 , an emitter-follower stage is encountered. This protects the filter from being loaded by the low-impedance sensitivity-control potentiometer R_5 . The input voltage of this potentiometer is also fed to the output attenuator, having an attenuation proportional to the overall amplification from the source follower input to the emitter follower output. This attenuator is also mechanically linked with the input attenuator. The voltage which has to be compared with the input voltage, via R_1 and R_2 , is obtained from the output attenuator. The voltage at the junction of R_1 and R_2 is fed into the chopper which is driven by a 2 kHz multivibrator. The filter

between the demodulator and the DC amplifier eliminates spurious 2 kHz signals in the signal amplifier and provides the phase correction for the feedback loop which is necessary to prevent oscillation. The symmetrical final amplifier follows R_5 with R_4 which give the Y position control. 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 with 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 drift occurring in the amplifier is corrected with the aid of an indirect DC amplifier, consisting of a chopper (which converts the DC voltage applied to its input into a square wave voltage), an AC amplifier and finally a demodulator which turns the amplified AC voltage basis into the correction voltage (V_C).

* This description is valid for the /04 and subsequent versions.

Technical data

All data are valid for this unit when used in combination with the main frame PM 3330 or PM 3370

AMPLIFIER

Bandwidth/Rise time

DC coupled: DC to 50 MHz, 7 ns or DC to 100 kHz 3.5 ns

AC coupled: 1.6 Hz to 50 MHz or 1.6 Hz to 100 kHz

Sensitivity 500 μ V/cm

Drift < 1 cm/week, irrespective of switching on and off

Attenuator Calibrated 500 μ 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 1 M Ω shunted by 15 pF

Max. input voltage AC coupled 400 V DC

Max. deflection 18 cm mid-band up to 15 MHz (6 cm window)

Shift range 18 cm

Probe power: 3-pole connector for supply of accessories such as the miniature probe PM 5352 or the FET probe PM 9353

DIMENSIONS

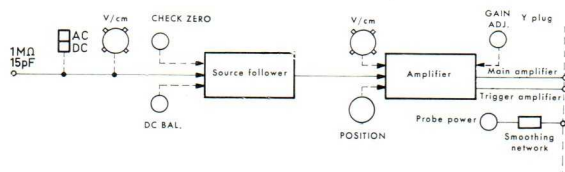
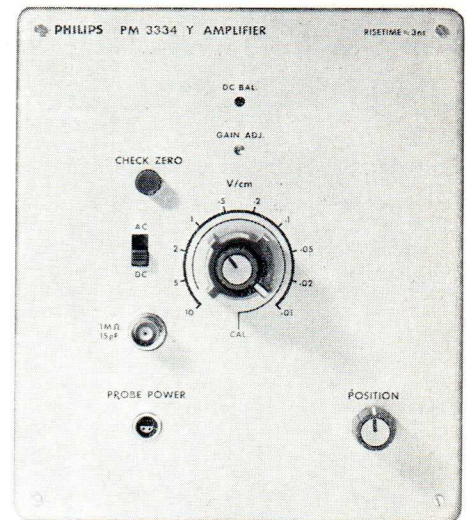
H x W x D 17.5 x 15 x 27.5 cm (7 x 6 x 11 in)

Weight 2 kg (4.5 lbs)

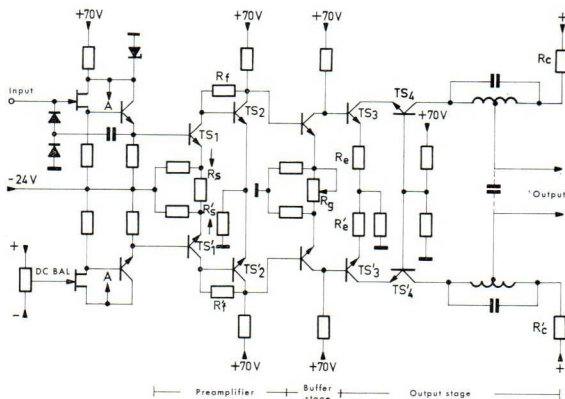
ACCESSORIES

PM 9150, PM 9352, PM 9353

Single trace vertical amplifier PM 3334



1. Block diagram of the PM 3334



2. Simplified diagram of the amplifier

The PM 3334 is a single trace vertical amplifier giving an overall rise time of 6 ns at a sensitivity of 10 mV/cm. Fig. 1 shows the block diagram of the unit.

The input (1 MΩ shunted by 15 pF) is connected via the AC/DC selector to the step attenuator V/cm.

The signal is applied via a source follower stage to the amplifier. This amplifier contains a variable gain control, a POSITION control and a GAIN-ADJ. control. The output voltage of the amplifier is passed to the vertical amplifier and to the trigger amplifier of the main frame via the output plug. The probe power plug is connected via a smoothing network (+ and - 24 Volt). 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.

These networks are switched one after the other and cover an attenuation range from x1 to x1000 in steps of 1—2—5.

Amplifier

This consists of three parts, viz. a pre-amplifier stage, a buffer stage, and an output stage. A simplified diagram of this unit is shown in fig. 2.

The signal is passed via a source and emitter follower to the pre-amplifier, in which both series and shunt feedback are used. The amplification $A (R_f/R_s)$ is stable, being dependent mainly on two resistors. The shunt feedback gives TS₂, 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 TS₁ is very small, so that the feedback through the collector base capacitance of the first transistor plays a negligible role. Because of this and the low output re-

sistance of TS₂ the pre-amplifier stage has a wide bandwidth.

To keep the capacitive load at the output of the pre-amplifier small, a buffer stage is interposed between the pre-amplifier and output stage. The amplification and hence the deflection coefficient of the unit can be adjusted by means of R_g (gain adjustment). A cascode amplifier (TS₃—TS₄) has been chosen for the output stage. The amplification here is also dependent mainly on two resistors, viz. $A R_c/R_e$. Because of the low input resistance of the second transistor (grounded-base configuration of TS₄) the feedback here is also negligible. The output signal of the output amplifier is passed via 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 main frame, amounts to about 6 ns ($f_{3dB} \approx 60$ MHz).

Technical data

Amplifier

Bandwidth:

DC coupled: DC to 60 MHz

AC coupled: 1.6 Hz to 60 MHz

AC coupled via 10: 1 probe: 0.16 Hz to 160 MHz

Risetime: < 6 ns

Sensitivity: 10 mV/cm

Attenuator

Input AC coupled via 10: 1 probe: 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

Maximum input voltage:

AC coupled 500 V DC

Maximum deflection:

6 cm over the full bandwidth

18 cm mid-band (20 MHz) window 6 cm

Shift range: 8 cm

Dimensions and weight

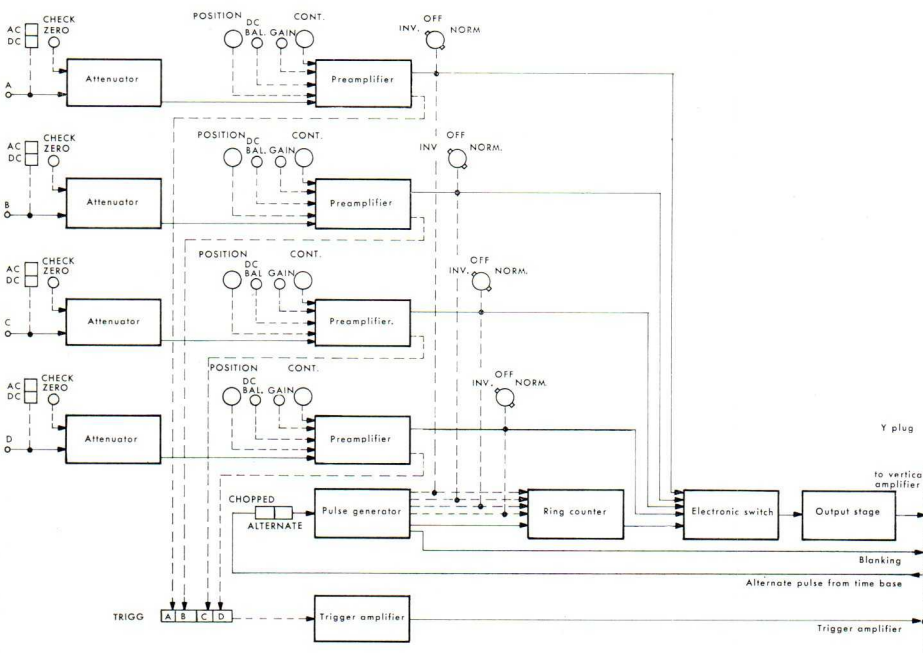
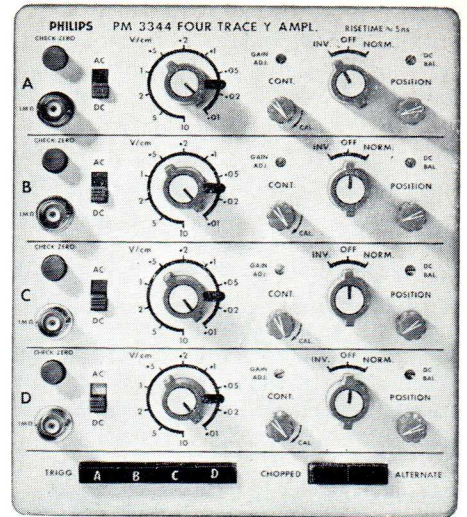
Height: 17.5 cm

Width: 15 cm

Depth: 27.5 cm

Weight: 2 kg

Four trace vertical amplifier PM 3344



1. Block diagram of the four trace amplifier

The PM 3344 is a four-trace unit for oscilloscopes PM 3370 and 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 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 com-

prising ten steps of 10, 20, 50 mV/cm up to 10 V/cm.

Amplifier

The input stage of the amplifier, see fig. 2, consists of a field-effect transistor (FET) connected as a source-follower, followed by a second transistor connected as an emitter-follower.

This circuit gives a high-ohmic input impedance and a low-ohmic output impedance. The input stage is made symmetrical, to limit drift. The signal from the attenuator is applied to one of the inputs, while the other receives an adjustable DC voltage which is used to balance the amplifier. The driven FET is protected by current and voltage limiting. The two diodes connected to the gate of the driven FET and the Zener diode in the drain circuit limit the voltage, while the series resistor in the gate circuit limits the current.

In order to reduce the influence of the diode capacitance at the input of the

amplifier, the signal is fed back from the low-ohmic output to the point between the two safety diodes.

The input stage is followed by two emitter-followers. The continuous amplification regulation and the adjustment of the deflection coefficient occur in this circuit. If the DC balance is correctly adjusted, there will be no voltage difference between the two emitters, so that no zero shift occurs when the continuous amplification regulation is used.

The trigger signal is taken from the same emitter-follower, and is thus fed to the trigger amplifier at a constant voltage level.

The emitter-follower stage is followed by the preamplifier. The latter consists of 2 transistors, emitter-coupled and base-driven. The position control 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 due to thermal phenomena, RC networks have been inserted in the collector circuits, see fig. 2.

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° phase shift). The centre position of the switch disconnects the relevant channel from the output amplifier by means of the electronic switch.

Electronic switch

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

If the level of the relevant output of 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 level 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 and D_{5-6} will be blocked. The output stage is now disconnected from the relevant preamplifier. As a blocked diode always possesses a certain capacitance the signal is capacitively transmitted. To minimize this 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} .

Output stage

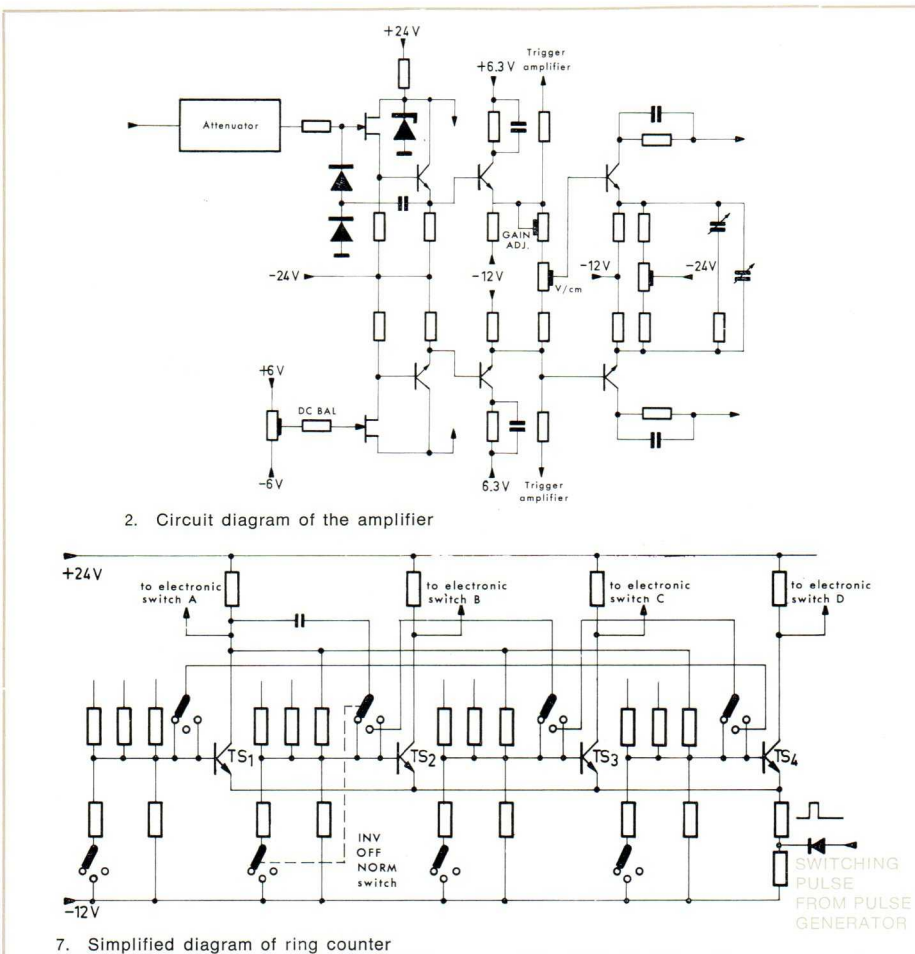
The output stage consists of:

- 2 shunt-feedback coupled transistors giving a very low-ohmic input circuit

oscillator is about 2 MHz. When the DC potential at the base of TS_1 is equal to or positive with respect to the emitter potential, the oscillator will stop. This happens in two cases:

1. When 3 amplifiers are switched off D_1 conducts and the base potential of TS_1 becomes positive with respect to the emitter. The blocking oscillator now stops so that no further pulses come from the oscillator.
2. When the alternate push-button is depressed, the quiescent base potential will be equal to the emitter potential. At the same time the alternate pulse from the oscilloscope 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



- 2 output emitter followers feeding the signal to the Y plug, see fig. 4.

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 trigger amplifier, see fig. 5. The signal is fed to the Y plug via two emitter-followers.

Pulse generator

This generator contains a blocking oscillator (fig. 6).

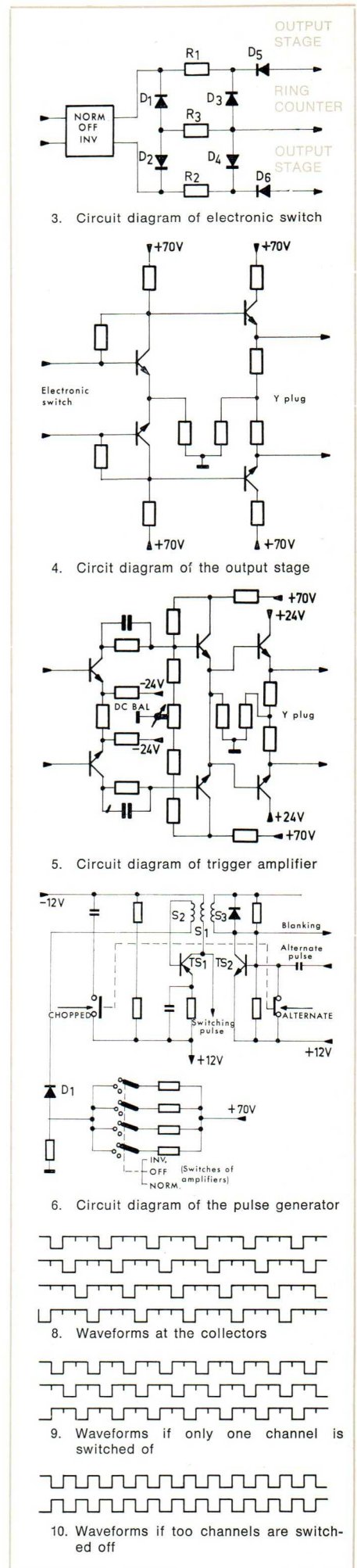
The RC network in the emitter circuit ensures that the repetition frequency of the

TS_1 . Consequently the blocking oscillator will operate once for each alternate pulse arriving.

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 resistors.

The collectors of the transistors are also capacitively coupled to the base of the transistor selected by the 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. This surge is fed via the above mentioned capacitor to the base of the selected transistor which then becomes conductive.

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 voltage waveforms at the collector are shown in fig. 8. The collectors of the ring counter are connected to the electronic switch. As mentioned above, 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 each channel is displayed 500 ns.

Technical data

All data for this unit are valid when it is used in combination with the main frame PM 3370 or PM 3330.

INPUT

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

AMPLIFIER

Bandwidth

DC coupled: DC to 50 MHz (f3dB)
AC coupled: 1.6 Hz to 50 MHz (f3dB)

Rise time 7 ns; unit alone 5 ns

Deflection coefficient 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
Phase shift between channels negligible
Blanking during switching

Attenuator

Calibrated 10 mV/cm to 10 V/cm (1, 2, 5 series)

Tolerance 3%

Continuous attenuation

Overshoot < 2%

Maximum input voltage AC coupled 400 V DC
Dynamic range 3x screen height up to 10 MHz
Polarity reversible for each channel

Triggering

From any of the four channels

Required trace height

3 mm for frequency up to 10 MHz, 2 cm up to 50 MHz

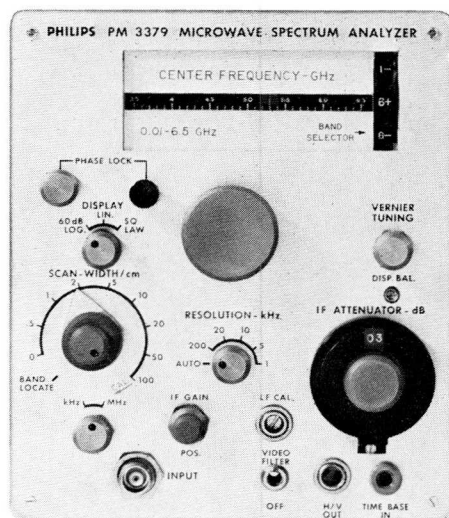
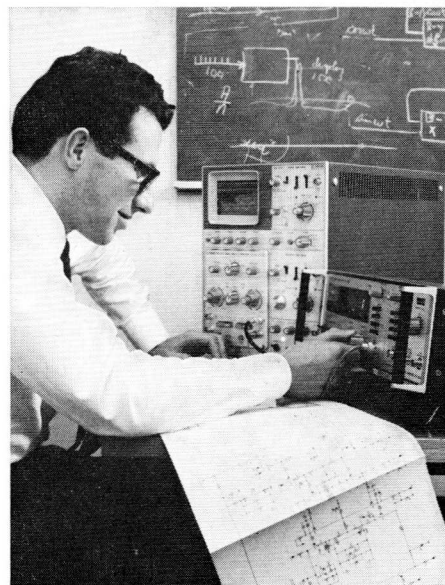
DIMENSIONS

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

Spectrum analyzer PM 3379



Measurements of electrical quantities as a function of frequency are being used more and more, to complement the more usual measurements as a function of time; together, these two types of measurements allow more information to be gained about electrical signals. These measurements require a spectrum analyzer unit, which is why such a unit is included in the plug-in range available with the PM 3370 main frame.

The PM 3379 spectrum analyzer, used in combination with the basic PM 3370 oscilloscope, analyses the frequency components of a signal and displays the relation between amplitude and frequency. This unit covers the frequency range from 10 MHz to 6.5 GHz. Detailed specifications are given below.

A more complete article about the PM 3379 spectrum analyzer will be included in a coming number of these Notes.

Technical data

Central frequency range 10 MHz to 6.5 GHz

Tuning dial accuracy

± 5% of dial reading or 15 MHz, whichever is greater (may be calibrated to within ± 5% at any frequency below 300 MHz by means of "LF cal" front panel adjustment)

Frequency identification

In the "Band locate" mode, the band switch is rotated until the trace is centered on the screen.

The central frequency may then be read off directly. Mixing mode is simultaneously indicated on the band switch.

Sensitivity

Frequency range	Band	Min. sensitivity*	Flatness
0—0.5 GHz	1—	—92 dBm	± 3.0
1.0—1.5 GHz	1+	—92 dBm	± 2.0
0.5—1.5 GHz	2—	—87 dBm	± 2.0
1.5—2.5 GHz	2+	—87 dBm	± 3.0
2.5—3.5 GHz	3+	—83 dBm	± 3.5
3.5—4.5 GHz	4+	—78 dBm	± 4.0
4.5—5.5 GHz	5+	—73 dBm	± 5.0
5.5—6.5 GHz	6+	—65 dBm	± 5.5

* Signal + noise = 2 x noise at 1 kHz resolution

Stabilized tuning

Phase lock automatically activated at dispersions of 100 kHz/cm or less. Front panel control provides lock adjustment

Dispersion/Scan width

17 calibrated positions selectable with front-panel switches

1	0 kHz/cm	10	500 kHz/cm
2	0,5 kHz/cm	11	1 MHz/cm
3	1 kHz/cm	12	2 MHz/cm
4	2 kHz/cm	13	5 MHz/cm
5	5 kHz/cm	14	10 MHz/cm
6	10 kHz/cm	15	20 MHz/cm
7	20 kHz/cm	16	50 MHz/cm
8	50 kHz/cm	17	100 MHz/cm
9	100 kHz/cm		

Continuous control provides adjustment between the steps

Dispersion accuracy ± 10%

Resolution/IF bandwidth

Automatically programmed with dispersion (5 mV/cm scan speed)

Uncoupled: 1, 5, 10, 20 and 200 kHz

Selectable with front panel switch

Vertical displays

Linear, 60 dB logarithmic and 13 dB square law

Selectable with front panel switch

Image separation

1 GHz (500 MHz first IF)

Display dynamic range 60 dB

Incidental FM

Less than 1000 Hz for fundamental mixing

Input impedance 50 ohm TNC connector

Input signal level —20 dBm (maximum)

IF attenuator

59 dB range in 1 dB steps; ± 0.1 dB/dB

IF gain

40 dB (nominal)

Continuously variable with front panel control

Scan rate

Automatic: 1/sec up to 60/sec

Video filter 1 millisecond

Horizontal and vertical outputs

0 to 1 volt from 1 K maximum DC coupled at ground

Two-circuit phone jack on front panel

Power requirements

All power and voltages from oscilloscope

Probes

time through the cable is no longer negligible, undesired reflections must be expected. This makes the pulse response much worse. In order to damp out these reflections, the inner conductor of the cable is given an ohmic resistance (about 200 Ω /metre).

In addition the cable is terminated for the highest frequencies and a correction network is added to compensate for the pulse aberration due to losses in the cable.

Correction network

The termination for the highest frequencies is mainly determined by $R_4 + R_5$ of the cable fig 2. C_4 only blocks the DC and LF signals. As C_4 adds to the input capacitance of the oscilloscope at lower frequencies it is kept as small as possible (8 pF). The HF attenuation can be made equal to the LF attenuation by means of R_5 .

L_1 serves as a M derived filter section, terminated with R_4 , to minimize the influence of the oscilloscope's input capacitance. R_6 is made variable to correct the sagging of the pulse top due to the ohmic losses in the cable. R_7 and C_5 improve the step response. C_3 serves for the LF compensation already mentioned; this adjustable capacitor is accessible from the outside.

Input impedance

The input impedance varies with frequency. At DC and very low frequencies the input impedance is 10 M Ω in parallel with 10 pF. At very high frequencies the input impedance is equal to the characteristic impedance of the cable (\approx 200 Ω) in parallel with 4 pF (stray capacitance). At other frequencies the input impedance is always between these extremes. Fig. 3 gives this impedance (split into X_D and R_D) as a function of the frequency.

Technical data

Attenuation : 10x
 Rise time : (of probe only) 1,2 nsec
 Aberrations : $< \pm 3\%$ (ringing or roll off)
 Voltage rating : 500V DC + AC peak

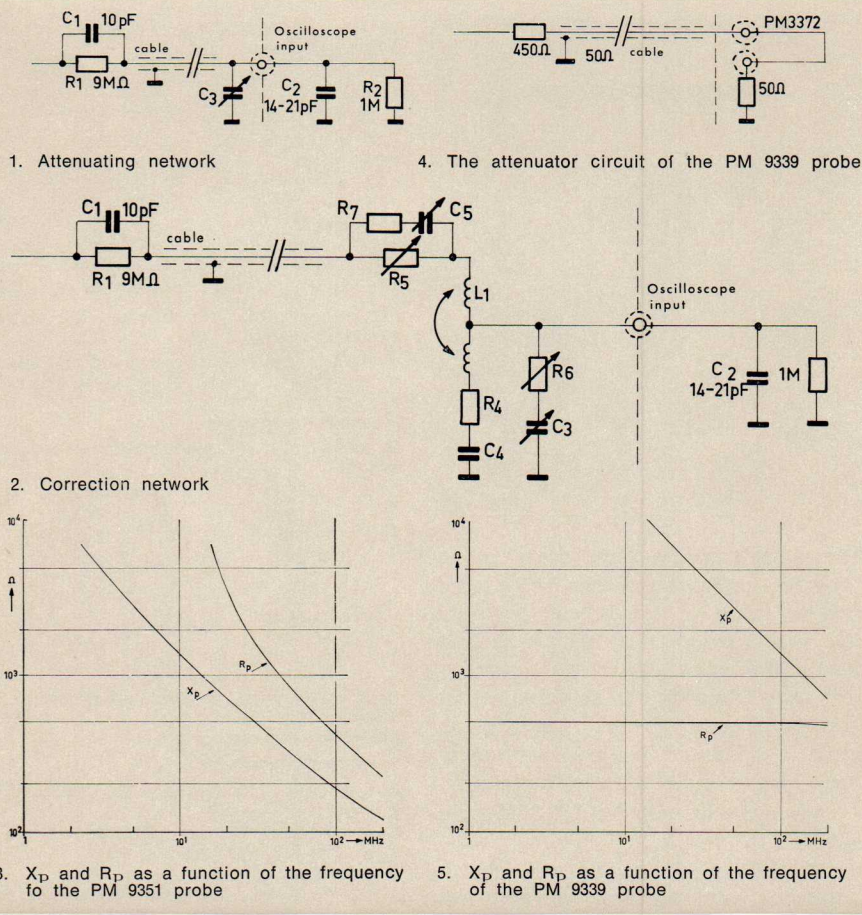
The PM 9339 probe

As may be seen from the description of the PM 9351 the construction of a high-impedance probe involves quite a few compromises, due to the way the cable behaves between probe tip and oscilloscope input.

At low frequencies a high-impedance cable acts as a mere capacitance, which is compensated for in the probe tip.

At high frequencies, it acts as a lossy cable, terminated at the oscilloscope side.

Although this solution yields a probe that is most suited for general use, the PM 9339 follows another principle that has its own advantages, mainly at the top of the frequency range; it makes use of the 50 Ω input impedance available with the PM 3372 plug-in amplifier, to terminate a 50 Ω cable.



The PM 9351 probe

The PM 9351 is a passive attenuating probe, for use with oscilloscopes with a bandwidth of up to 150 MHz and an input capacitance of between 14 and 21 pF in parallel with 1 M Ω .

Attenuating network

This network consists of a 9 M Ω resistor (R_1), which in combination with the 1 M Ω input resistance of the oscilloscope (R_2) gives an attenuation of 10 times for DC and LF signals. fig. 1. For AC signals at higher frequencies the attenuation is governed by the ratio of C_1 and the capacitance in parallel with the 1 M Ω input resistance of the oscilloscope.

This capacitance consists of the input capacitance of the oscilloscope itself, which must be between 14 and 21 pF, the capacitance of the cable and a trimmer capacitor C_3 . To make the attenuation of this circuit independent of frequency the following demand must be fulfilled:

$$R_1 : R_2 = \text{total capacitance in parallel with } R_2 : C_1$$

C_1 has been chosen to meet this demand with a cable 1.5 m long and an oscilloscope input capacitance of 21 pF, while C_3 is in its minimum position. With an input capacitance < 21 pF, C_3 must be increased to compensate for this.

The cable

With HF signals, where the propagation

A probe tip with a $450\ \Omega$ resistor is mounted in front of this cable, thus giving a 10 : 1 attenuation ratio (see fig. 4).

The input impedance of this probe is $500\ \Omega$ in parallel with only $1.5\ \text{pF}$. Although this resistance is fairly low compared with the $10\ \text{M}\Omega$ of the PM 9351, it remains virtually constant over the whole frequency range whilst its parallel capacitance, which is far more important at the upper frequency limit, is only a fraction of that of a high-impedance probe.

Other advantages of this approach are that the cable can be given any desired length, because this length hardly influences the LF properties of the attenuator network and that no correction network whatsoever is needed.

This results in very small pulse aberrations.

Fig. 5 shows the input impedance of the PM 9339 as a function of frequency.

When we use the PM 9339 with the $1\ \text{M}\Omega$ input mode of the PM 3372, we have a probe without attenuation, thus giving a 1 : 1 voltage ratio; this may be useful for DC and LF measurements.

Technical data

- Attenuation : 10x
- Rise time : (probe only) < 0,5 nsec
- Aberrations : $< \pm 1\frac{1}{2}\%$ (ringing)
- Voltage rating : 15V DC + AC peak or 15 V_{ACeff}

OTHER NEW PROBES

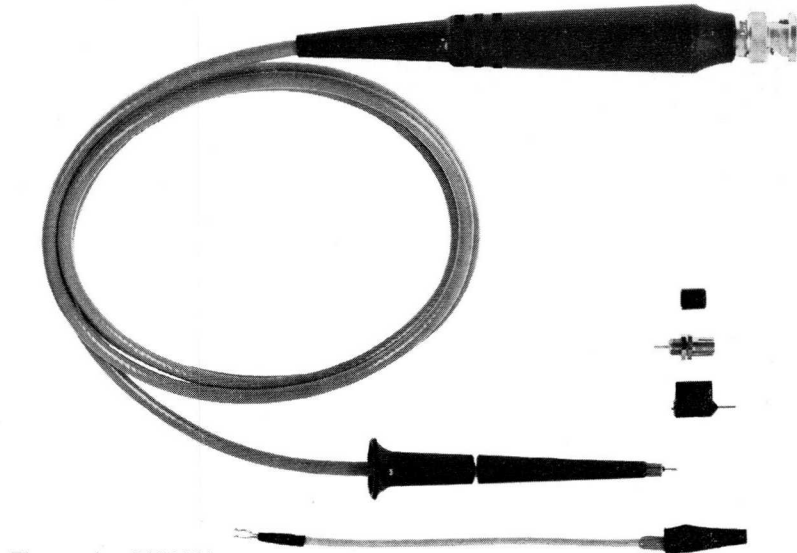
PM 9352

This is a very small probe specially suited for measurements on miniature circuits. A vast range of accessories makes it adaptable for many special applications. The probe power outlet of the PM 3372 plug-in amplifier is used for this probe. A detailed description of this probe will be given in the next issue.

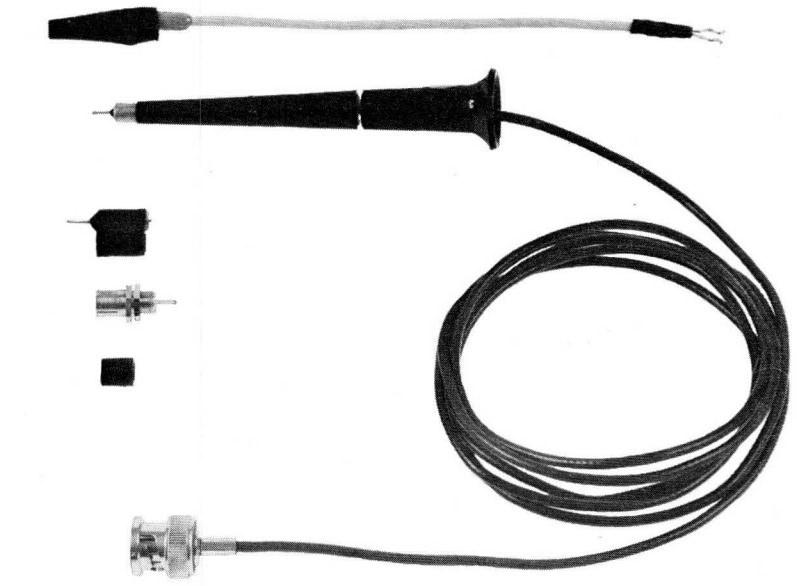
PM 9347 TV probe

This probe combines the functions of the PM 9350 10 : 1 attenuator probe and a TV sync separator. It also makes use of the probe power outlet of the oscilloscope, for the sync separator: a more detailed article in our next issue will deal with this completely new design.

The probe PM 9352

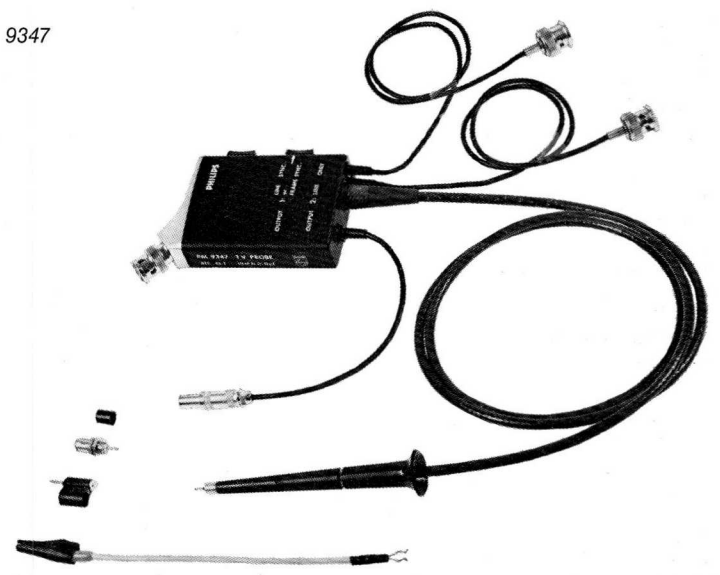


The probe PM 9351



The probe PM 9339

The probe PM 9347



Applications of the PM 3370 oscilloscope



An oscilloscope designed for a wide range of applications is, inevitably, a complex instrument. As modern instruments offer more and more facilities, so the task of operating them increases in complexity. The PM 3370, although of logical design and built for ease of operation, may present some problems to the occasional operator confronted with such a large number of controls. However, an understanding of the techniques for making basic measurements together with a recognition of the less obvious triggering problems and their solution, should help to dispel any apprehension on the part of the user.

Accurate Time Measurements using the Delayed Time base

The delayed sweep is invaluable for making accurate time measurements. The advantages of using the delayed time base method can best be understood if we consider the factors that contribute towards measurement errors when using an oscilloscope.

These are:

- the inaccuracy of the time base,
- the reading error.

The inaccuracy of the time base is generally specified for the middle 8 divisions of the deflection (due to slight non-linearities at the extremes of the scan) a tolerance of approximately 3% being common for professional equipment. This is often regarded as the measuring error of the oscilloscope, the reading error generally influence on the accuracy of the measurement. The factors that influence the reading error are:

- line thickness,
- the angle between the signal and the horizontal graticule line at the point where the reading is made,
- parallax.

Line thickness is dependent on the focusing and astigmatism of the CRT-spot and upon its intensity. For optimum line thickness, the intensity should not be too high.

44

The angle between the signal and the horizontal graticule line at the point of reading should be as large as possible to obtain a clearly defined crossover point. It is therefore advisable to work with the biggest possible display and, if possible, to take the reading at the steepest part of the trace.

With the PM 3370, the reading error due to parallax is eliminated as the CRT has an internal graticule.

The value of the reading error which results from taking measurements may be as much as 0.05 of a centimetre for each measuring point. As all time measurements are in fact the result of two measurements, the total error can be 0.1 of a centimetre. This is an additional factor contributing towards the total inaccuracy, and for measurements over small distances it adds a considerable error to the reading. For example, over two centimetres

$$\frac{0.1}{2} \times 100\% = 5\%$$

By using the delayed sweep method of measurement it is possible to considerably reduce the influence of the reading error.

The following methods illustrate the various uses of the delayed time base. Note that the DELAY TIME MULTIPLIER potentiometer introduces a slight inaccuracy due to its non-linearity which is 0.2% typically.

COMPARISON OF METHODS

The attendant advantages of measuring time differences by means of the delayed time base can be shown by considering a specific example and applying the various methods of measurement at our disposal. Example 1.

Assume a signal as illustrated in fig. 1 where it is required to measure the time difference between the points t_1 , t_2 , t_3 , t_4 .

Method A

Measuring the time difference by means of the main time base.

Assume that the TIME/cm setting of the

main time base is 10 $\mu\text{s}/\text{cm}$ and that the horizontal distance t_1 to t_4 is 8.2 cm. The measured time difference between them is therefore $8.2 \times 10 \mu\text{s} = 82 \mu\text{s}$. The measured distance t_2-t_3 is 0.2 cm which represents 2 μs . The measured distance t_3-t_4 is 1.4 cm which represents 14 μs . Therefore, the reading errors for the respective times are:

$$t_1-t_4 = 2 \times \frac{0.05}{8.2} \times 100\% = \pm 1.2\%$$

$$t_2-t_3 = 2 \times \frac{0.05}{0.2} \times 100\% = \pm 50\%$$

$$t_3-t_4 = 2 \times \frac{0.05}{1.4} \times 100\% = \pm 7\%$$

As the error of the time base may be as much as 3%, this must be taken into account when calculating the total error.* It will be observed that the accuracy is rapidly degraded as the distance on the time axis decreases. For distances less than 1 cm, the $\times 10$ MAGN. can be used. In the example for t_2-t_3 , the distance is expanded to 1.7 cm, which gives 1.7 μs . The reading error is

$$2 \times \frac{0.05}{1.7} \times 100\% = 6\%$$

Therefore the max. total error could be $3\% + 2\% + 6\% = 11\%$

The 2% error factor is due to the use of the $\times 10$ magnifier.

By using this method, the accuracy is improved by a factor of 5.

Method B

Measuring the time difference by means of the delayed time base.

With the X DEFL switch set to MAIN TB INTENS, the DELAY TIME MULTIPLIER potentiometer may be adjusted in conjunction with the TIME/cm switch of the delayed time base to give an intensified display of that portion of the trace which embraces t_2-t_3 .

* The figure for the timebase error can be checked by displaying an accurately known frequency in the used positions of the time base controls and a better figure than 3% can be obtained.

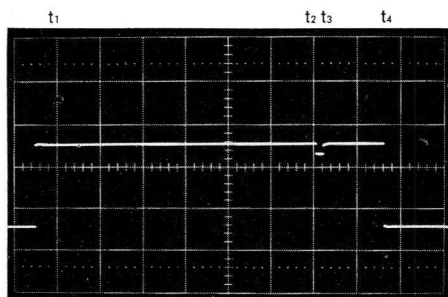
By switching the X DEFL switch to DEL'D TB an expanded image, as shown in fig. 2, is obtained. Assume that the TIME/cm setting of the delayed time base is $0.2\mu\text{s}/\text{cm}$. The measured distance between t_2 and t_3 is now 8 cm and t_2-t_3 is $1.6\mu\text{s}$. The read-off error expressed as a percentage is now:

$$2 \times \frac{0.05}{8} \times 100\% = 1.2\%$$

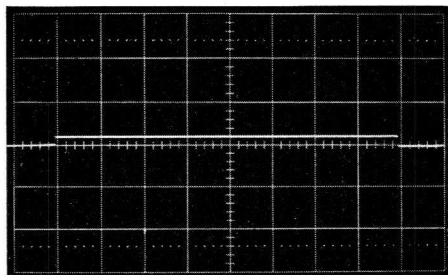
The total measuring error is thus $3 + 1.2 = 4.2\%$. When t_3-t_4 is measured by the same method, the largest distance on the time axis will be a setting of $2\mu\text{s}/\text{cm}$ for the delay time base. The distance is 6.8 cm, therefore, the time is $13.6\mu\text{s}$, and this reading error is 1.5% giving a total error of $1.5\% + 3\% = 4.5\%$.

Method C

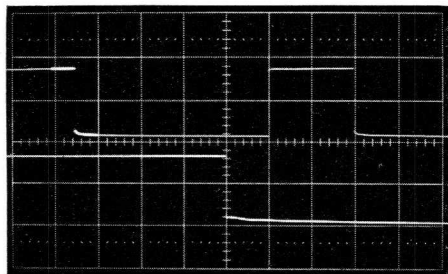
Measuring the time difference by means



1. Example of signal to be examined



2. Time measurements using the delayed time base



4. Time ratio measurements

of the delayed time base and the DELAY TIME MULTIPLIER.

Obtain an expanded display as in the previous method, then rotate the TIME/cm switch of the delayed time base to $100\text{ ns}/\text{cm}$ so that the trace is expanded by a factor 100. Locate t_2 on the screen and align it with the vertical centre-line of the graticule by means of the DELAY TIME MULTIPLIER dial. Note the dial reading. Then rotate the dial until t_3 coincides with the vertical centre-line and note the new dial reading.

If these readings are, for example, 7.08 and 6.92, then the distance between the two measuring points has been expanded to 16 cm and the time difference is

$$16 \times 100\text{ ns} = 1.6\mu\text{s}.$$

The total measuring error using this method, is the sum of the maximum main time-base error and the error due to the non-linearity of the DELAY TIME MULTIPLIER potentiometer (0.2%) together with the read-off error.

The read-off error is now the sum of the relative reading error and the interpolation error of the delay time multiplier.

The relative reading error is

$$2 \times \frac{0.05}{16} \times 100\% = 0.63\%$$

The interpolation error of the delay time multiplier is

$$2 \times \frac{0.005}{16} \times 100\% = 6.3\%$$

Thus the maximum total error

$$3\% + 0.2\% + 0.63\% + 6.3\% \approx 10\%.$$

Measuring t_3-t_4 in the same manner will give dial readings of 7.08 and 8.445 so the

Interval	Method A		Method B	Method C
	$\times 1\text{ MAGN}$	$\times 10\text{ MAGN}$		
t_1-t_4	$82\mu\text{s} \pm 4.2\%$	—	—	$82.1\mu\text{s} \pm 3.3\%$
t_2-t_3	$2\mu\text{s} \pm 53\%$	$1.7\mu\text{s} \pm 11\%$	$1.6\mu\text{s} \pm 4.2\%$	$1.6\mu\text{s} \pm 10\%$
t_3-t_4	$14\mu\text{s} \pm 10\%$	—	$13.6\mu\text{s} \pm 4.5\%$	$13.65\mu\text{s} \pm 4\%$

3. Table of results

difference is 1.365. Thus the distance between the two measuring points has now been expanded to 136.5 cm, the time difference being $136.5 \times 100\text{ ns} = 13.65\mu\text{s}$. The reading error on the display is

$$\frac{0.1}{136.5} \times 100\% = 0.07\%$$

the dial read-off error is

$$\frac{0.01}{1.365} \times 100\% = 0.7\%$$

therefore, the total error

$$3\% + 0.2\% + 0.07\% + 0.7\% \approx 4\%.$$

Measuring t_1-t_4 with method C will give a dial reading difference of 8.21. The time is therefore $82.1\mu\text{s}$ and the expanded distance is 821 cm. The reading error of the display is

$$\frac{0.1}{821} \times 100\% = 0.012\%$$

which can be neglected. The dial reading error is

$$\frac{0.01}{8.21} \times 100\% = 0.12\%$$

$$3\% + 0.2\% + 0.12\% \approx 3.3\%.$$

Conclusions

It will be seen from the table of results fig. 3 that Method B is best when very small distances on the main time-base axis are to be measured. However, when these distances are greater than about 1 cm, the read-off error of the DELAY TIME MULTIPLIER becomes negligible and Method C is to be preferred.

Time ratio measurements

The delay time multiplier offers the facility for the accurate determination of time ratios. Any significant errors incurred are due only to the incremental non-linearity of the ten-turn potentiometer, and to the reading error which can be kept small by using the delayed time base.

This is illustrated by reference to the following example fig 4. The upper trace shows the pulse displayed by the main time base at the $20\mu\text{s}/\text{cm}$ setting. The lower trace represents the intensified portion of the upper trace as displayed in the ALTERN. position of the X DEFL. switch.

The purpose of the measurement is to find the mark/space ratio of the signal. For suitable expansion of the leading and trailing edges, the delayed time-base TIME/switch was set to $1\mu\text{s}/\text{cm}$.

The negative-going leading edge was dialled to the centre vertical graticule line, as illustrated, and the dial setting of the DELAY TIME MULTIPLIER noted; this was 1.770. Then the next positive-going edge was dialled to the vertical centre-line, the setting being 6.265. Similarly, the final negative-going edge was dialled and a reading of 8.21 obtained.

Thus the negative-going portion of the trace gave a measured reading of: $6.265 - 1.77 = 4.495$.

The positive-going portion of the trace gave a measured reading of:

$$8.21 - 6.265 = 1.945.$$

Therefore, the mark/space ratio is:

$$\frac{1.945}{4.495} = 0.433.$$

The accuracy of the positive-going portion is determined by the reading error and the incremental non-linearity of the dial potentiometer.

As the main time base is expanded by a factor of 20, the pulse length is $20 \times 1.945 = 38.9$.

Therefore, reading error

$$2 \times \frac{0.05}{38.9} \times 100\% = 0.26\%$$

dial reading error

$$2 \times \frac{0.005}{1.945} \times 100\% \approx 0.5\%$$

potentiometer non-linearity = 0.2% .

Total accuracy for the positive-going portion is therefore:

$$0.26\% + 0.5\% + 0.2\% = 0.96\%.$$

For the negative-going portion: reading error

$$2 \times \frac{0.05}{89.9} \times 100\% = 0.11\%$$

dial reading error

$$2 \times \frac{0.005}{4.495} \times 100\% = 0.23\%$$

potentiometer non-linearity = 0.2% .

Total accuracy for the negative-going portion is therefore:

$$0.11\% + 0.23\% + 0.2\% = 0.54\%.$$

The total ratio error is therefore:

$$0.96\% + 0.54\% = 1.4\%.$$

The resultant measurement gives a ratio of $0.433 \pm 1.4\%$; i.e. within 0.427 and 0.439.

Display Switching

The PM 3370 is equipped with display switching, a new feature for the Philips range of oscilloscope. This now offers the instrument user a simultaneous display of the signal on the two time scales provided by the main time base and by the delayed time base.

Detailed examination of a certain portion of the main time base display is enabled by expanding the time interval of interest by means of the delayed time base. Expansion is achieved by selecting a correspondingly faster sweep for the delayed time-base. TIME/cm control, and positioning the time interval by the DELAY TIME MULTIPLIER potentiometer.

The part of the signal under detailed observation by the delayed time base appears as an intensified portion of the main time base display. This not only facilitates

An example of this is shown in fig. 6 where a sine wave of variable amplitude is fed into an overdriven amplifier which functions as a squarer. The amplitude of the square wave output, unlike the rise time, is independent of the input voltage. To examine the relationship between rise time and input voltage, the following procedure may be employed.

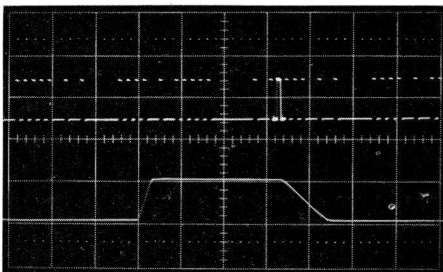
- Using a multi-trace plug-in in the ALTERN. mode display the output voltage as the upper trace and the input voltage as the lower trace, as shown in fig. 6.
- Set the X DEFL. switch to ALTERN. and adjust the delayed time base controls for a suitable display of the leading edge of the square wave (fig. 7).
- Adjust the Y deflection and POSITION controls of the plug-in unit and the TRACE SEPARATION control of the

As can be seen, the double triggering is apparent by the closed loops that are present at various points on the trace.

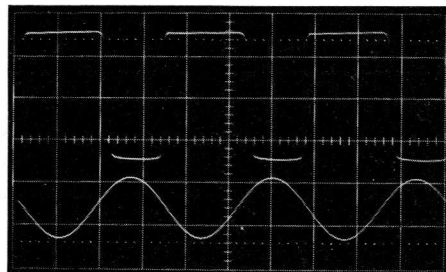
Assume that the sawtooth starts on the first pulse of the composite waveform. During the flyback period the input is made insensitive to further trigger pulses by means of the hold-off pulse.

When this pulse terminates, the next trigger pulse will initiate the sawtooth again. However, if this trigger pulse is derived from the second pulse of a multi-pulse waveform then a situation is obtained as illustrated in fig. 9.

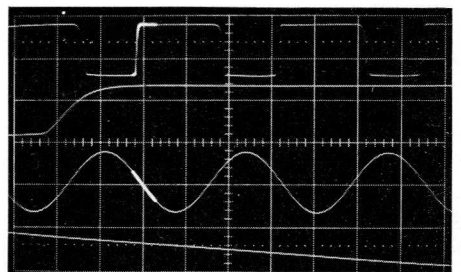
Now because both time bases are of the same length the next flyback period will end after a second pulse in the waveform and the new sawtooth will again be triggered on the first-pulse. So the sawtooth waveform will be alternately triggered by the first and second pulses resulting



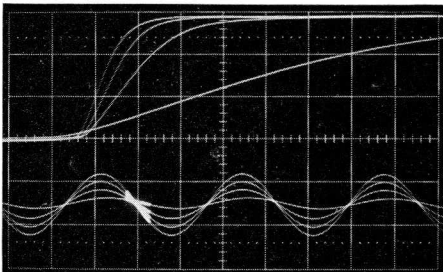
5. Showing easy location of the expanded portion



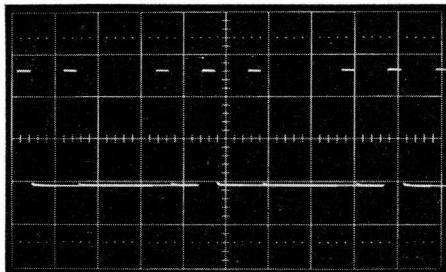
6. Output signal (upper trace) of a squarer, driven by a sinusoidal input signal (lower trace)



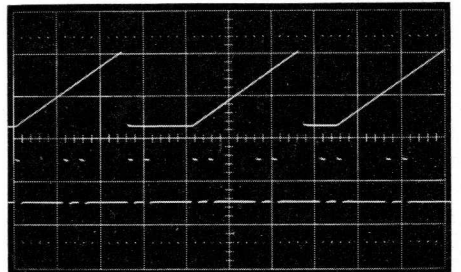
7. Showing the leading edge (second trace) of the squarer wave



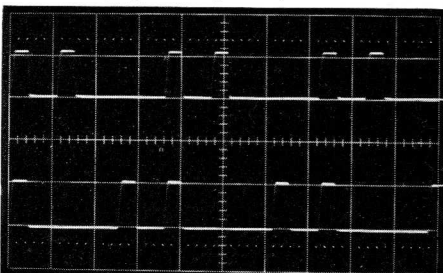
8. Multi exposure photograph showing the rise time varying with input amplitude



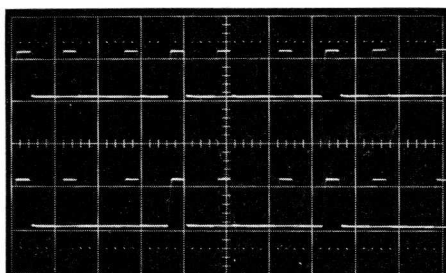
9. Effect of double triggering



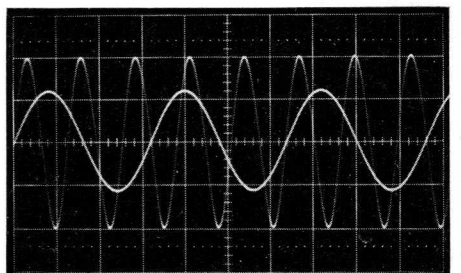
10. Double triggering on pulse train



11. Double triggering in ALTERN mode



12. Double triggering in CHOPP mode



13. Using the Y output signal

the location of the required detail during dialling but also serves as a visual indication of which portion of the overall trace is being examined. (See fig. 5). Thus the operator can immediately correlate the detail with the overall signal, which may be extremely complex, without the necessity of switching between MAIN TB and DEL'D TB as required with most other instruments.

This feature is extremely useful when the aspects of the detail observed are changes caused by the variation of some other signal displayed simultaneously by the main time base.

main instrument to remove the extreme traces of fig 7 for clarity.

By employing a multi-exposure photographic technique, the relationship between input voltage and output rise time can be effectively displayed as illustrated in fig. 8.

Double Triggering

Double triggering is an effect, inherent in all triggering systems, that can be produced when the signal point at which the circuit triggers occurs more than once in a single repetition time T as shown in fig. 9. This phenomenon is exhibited particularly in the display of pulse trains.

repetition time T , and which is related to the displayed pulse (e.g. via the EXT. trigger input), then the obvious remedy is to vary the sweep speed by means of the continuous time base control. If a calibrated display is required, then the alternative is to vary the time-base switch control in steps to eliminate the double triggering. However, if this produces an unsatisfactory time base speed, the display in an oscillogram as illustrated in fig. 10. If it is not possible to trigger on another pulse that occurs only once during the

can be rectified by switching the TIME/cm potentiometer to the UNCAL position and calibrated measurements can be made by using the delayed time-base.

In some instances, it is not obvious from the display that double triggering is occurring and, in these circumstances, time errors will inevitably result. If, for example, we apply the same signal as in fig. 9 to both inputs of a dual-trace unit, then with the electronic switch in the ALTERN. position there will be a display as in fig. 11. This is because the two channels are each triggered by different pulses. Consequently, when the sawtooth is triggered by the first pulse, for example, channel A is displayed and for the next sawtooth which is triggered by a second pulse, channel B is displayed. Since we know that the two waveforms are derived from the same source, then the displayed time displacement error is immediately apparent.

However, when we are observing two signals whose time relationship is unknown (and requires to be measured), this time error due to double triggering may invalidate the measurement. If the trigger signal is such that it offers any possibility of double triggering, e.g., when the trigger signal crosses the trigger level twice or more during time T, the ALTERN. display position should not be used.

In the CHOPP. mode of the multi-trace unit, double triggering will appear as "closed loops", as illustrated in fig. 12.

Thus switching to the chopped mode when double triggering is suspected provides a useful check. The remedies are the same as those already outlined.

Y Output signal

Many advantages accrue from the fact that in Philips multi-trace oscilloscope the internal triggering signal can be derived from either channel and it is always taken off before the electronic switch. The main advantages are as follows:

- the trigger signal is merely a part of the input signal and is therefore devoid of interfering components such as are present in the chopped mode of operation which could result in display instability;
- in the alternate mode, the possibility of time relationship errors between the displayed waveform is avoided since the trigger signal is not a combination of two or more signals;
- the trigger is unaffected by the continuous VOLTS/cm control, the POSITION control and the polarity selector switch of the Y plug-in unit;
- a simple method is provided for triggering from either channel without the necessity of changing input connections or applying one of the signals to the external trigger input.

In this way, the system used on the PM 3370 and its plug-in units safeguards the operator against unmeaningful displays whilst providing him with the facilities to obtain a stable display very easily.

However, there are some instances where a stable display may be required from two signals the frequencies of which are unrelated. The Y output signal, available at a socket outlet on the rear of the instrument, derived from the Y plug-in unit after the electronic switch, may be used for this purpose.

An example of this occurs when it is required to adjust a frequency to be a certain ratio of an accurately known frequency.

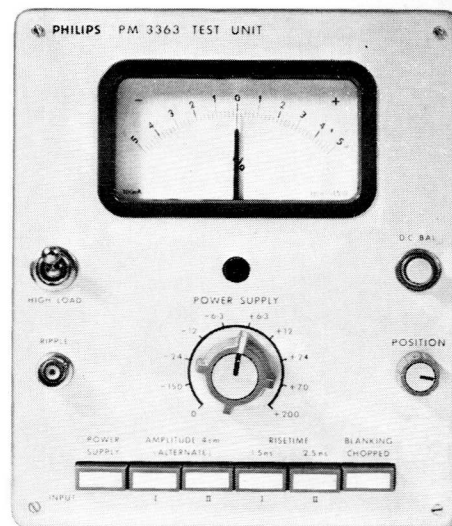
Assume that the known frequency is exactly 2 MHz and that the other frequency has to be adjusted to 5 MHz. The dual-trace plug-in unit used with the PM 3370 must be operated in the ALTERN. mode. The main time base is set to display about 3 complete cycles of the 2 MHz signal. The input signals will be present at the Y output socket alternately for a complete sweep time. The Y output socket must be connected to the EXT. TRIGG. input socket of the main time base. The DC level of each signal can be varied by means of the relevant POSITION control. Set the coupling selector to DC or HF to obtain main time base triggering, and adjust the LEVEL control so that the sweep starts for both signals on the horizontal centre-line of the graticule.

The frequency can now be adjusted to the desired value by making the end of the fifth cycle coincide with the end of the second cycle of the known frequency, as shown in fig. 13.

This method is extremely accurate since measurements are independent of time base inaccuracies.

In order to make the read-off error as small possible, the coincidence area on the main time base can be displayed enlarged by the delayed time base using the ALTERN. mode of the display switch.

The intensified portion of the main time base can be set to the point of interest by means of the DELAY TIME MULTIPLIER dial and the delayed display used for accurate adjustment.



Plug in test unit PM 3363

The PM 3363 plug-in unit enables quick checking and adjusting of the PM 3370 vertical amplifier and power unit.

The test unit provides:

1. A symmetrical square wave, current or voltage, to test the risetimes of both inputs of the vertical amplifier.

Amplitude: 4 cm deflection

Risetime: 1.5 ns for input I, 2.5 ns for input II

Frequency: 2 or 100 kHz switchable

Position: via front panel control

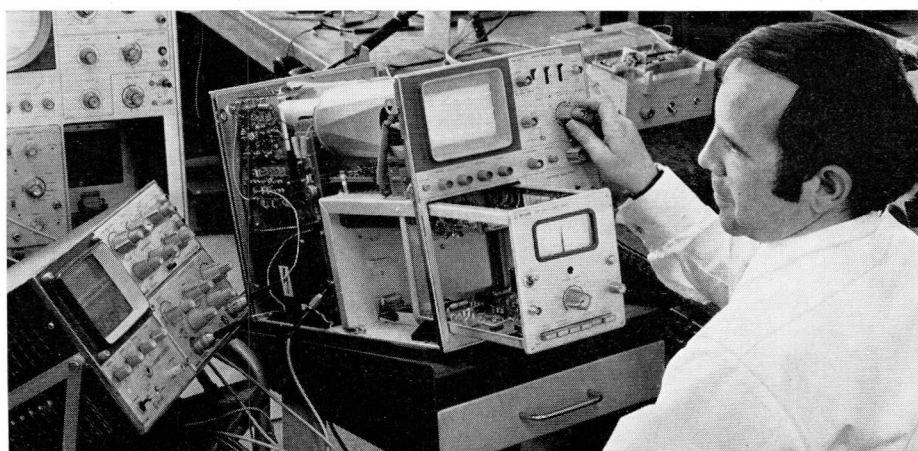
Zero level: The DC balance button on the front panel connects the output amplifier II to the zero level reference voltage.

2. An alternate display with two traces 4 cm apart. This allows adjustment of the output amplifier sensitivity and checking the performance of time-base switching pulses as used by multi-trace amplifiers.

3. A built-in voltmeter gives rapid testing of the power supply.

Ripple voltage can be measured by connecting an external oscilloscope.

4. Blanking pulses are available to check satisfactory functioning of the main frame blanking circuit in the chopped mode.



During the development of the test unit

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