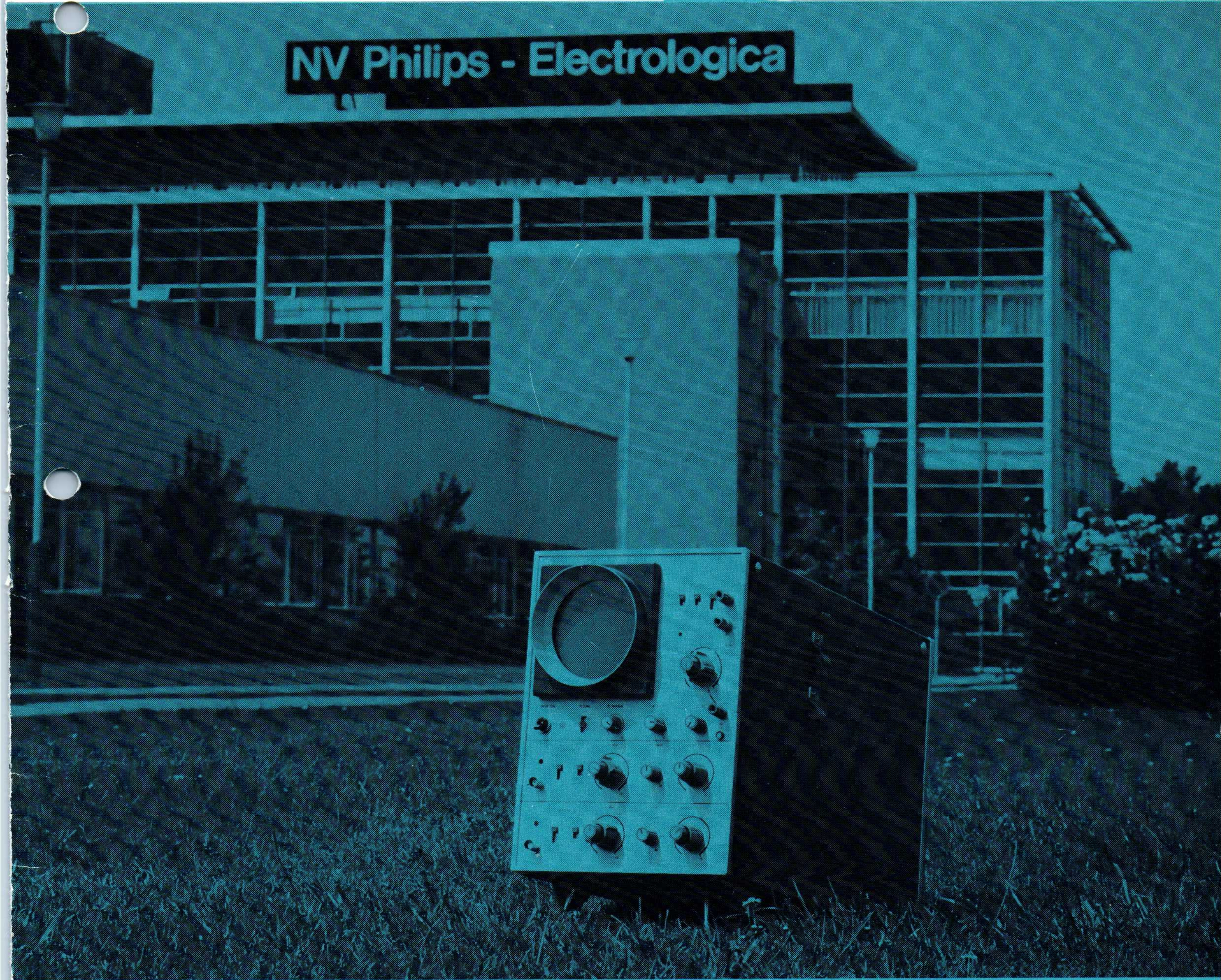
PHILIPS

electronic measuring and microwave notes



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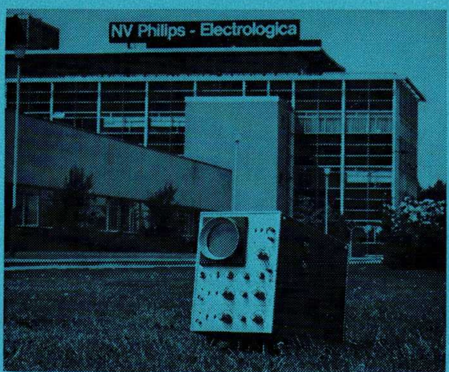
PHILIPS

electronic measuring and microwave notes

Introduction

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



In this issue

General information

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

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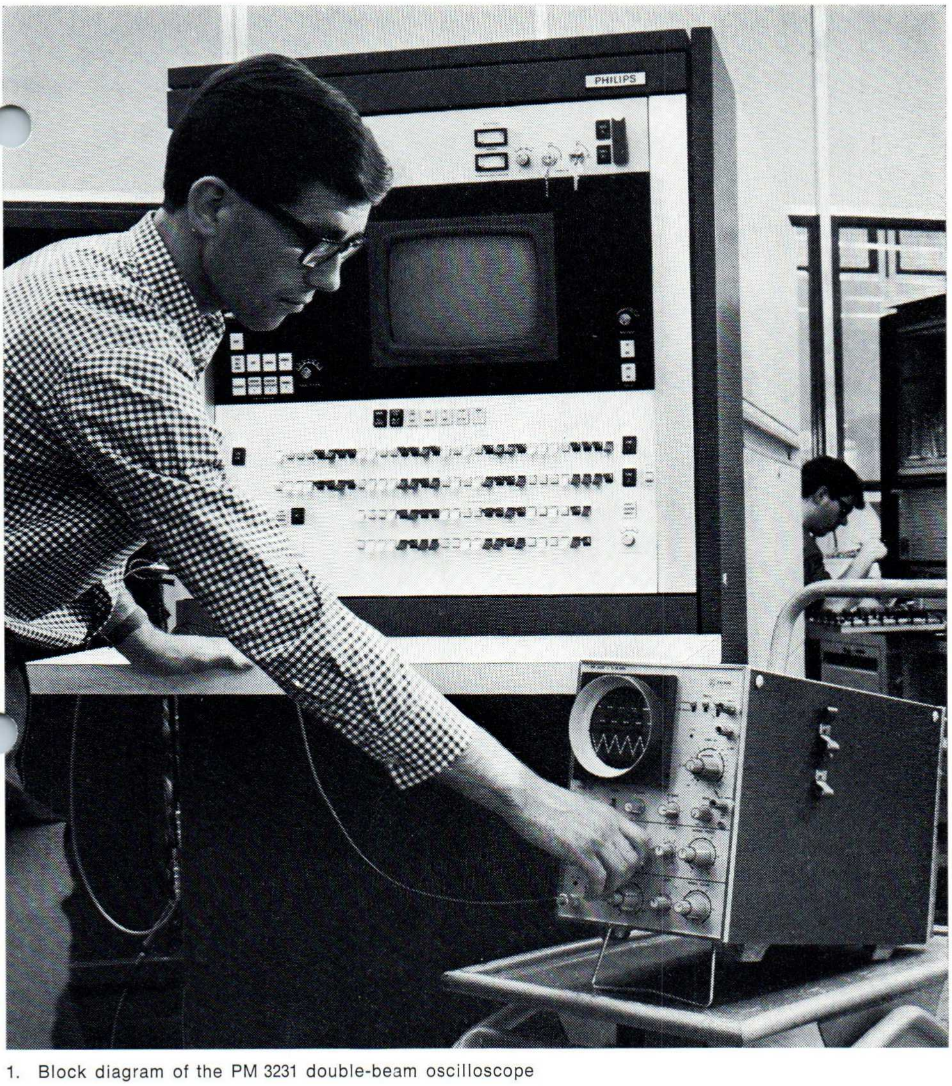
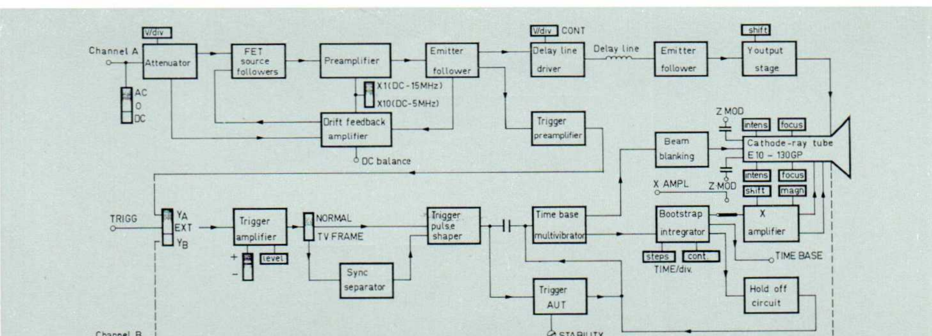
The front cover
of this issue shows new building of the computer factory at Apeldoorn.

A new wide band DC up to 15 MHz double-beam oscilloscope with delay line

by J. Overdorp and J. R. Vogel



1. Block diagram of the PM 3231 double-beam oscilloscope

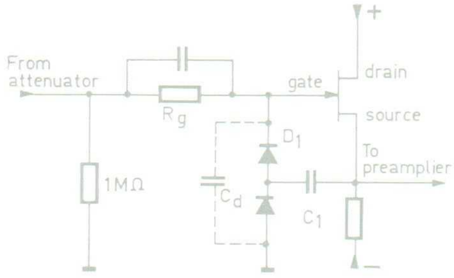


Rapid increase in the application of digital techniques has created a demand for economical and simply operated double-beam oscilloscopes permitting accurate measurement and comparison of the shape, amplitude, rise time, repetition frequency, etc. of pulses. This new double-beam oscilloscope will assist in meeting this demand. Block diagram is shown in fig. 1. The main features of the PM 3231 oscilloscope are:

1. *Two identical vertical channels*
2. *Fully transistorized*
3. *Bandwidth: DC up to 15 MHz at a maximum sensitivity of 10 mV/div*
   DC up to 5 MHz at a maximum sensitivity of 1 mV/div
4. *Field-effect transistor (FET) input both for the two vertical amplifiers and the external trigger input*
5. *Drift-compensation in the vertical amplifiers*
6. *Trigger mode:*
   Triggering with continuous variable level
   Automatic triggering (in absence of a trigger signal the time base starts "free running" and a bright trace is displayed)
7. *Delay-line, providing an effective delay on the screen of at least 50 nanosec*
8. *DC-coupled external trigger input*
9. *Max sweep speed: 200 nanosec/div*
10. *Horizontal magnification up to a maximum of 5x, the highest sweep speed thus becoming 40 nanosec/div*
11. *Fully stabilised power supply*

FET input with overload safeguard

Use is made in the input circuit of the vertical amplifiers of N-channel field-effect transistors circuited as source-followers, see fig. 2. To safeguard these against over loads, the gate is earthed with two series-connected diodes. The diodes become conductive at a negative voltage of −1.4 V. R_g acts as "gate current" limiter in the event of positive overloads. To eliminate the harmful capacitance C_d of the diodes, the signal voltage from the FET source is fedback via C_1 to the anode side of diode D_1, no signal current then passing through diode D_1.



2. The FET input safeguard

In virtue of these safeguards, voltage of up to 500 V_{pp} can be fed to the input without resultant risk to the oscilloscope.

Drift-compensation ("zero-line locking")

A new system of drift-compensation allows the "zero line" of the oscilloscope to be locked. Since the output level of the preamplifier can be further drift-free amplified without special measures thereby being required, the zero-line locking acts only on the preamplifier (see fig. 3).

If, after the normal high-ohmic attenuation of the Y_A signal, the amplifier receives E_i volts, the output voltage E_o amounts to $-A\,E_i$ in the normal state. If drift now gives rise to $\triangle e$ in the preamplifier, the voltage at point X becomes $E_x = \frac{1}{2}(E_i + E_o/A)$. The attenuator R_1R_2 attenuates the signal A times, equal to the amplification of the preamplifier. The drift is corrected by returning the voltage at point X, amplified and in counter-phase, to the preamplifier. The voltage at point II is the $E_{II} = -\frac{1}{2}B\,(E_i + E_o/A)$. At point I $E_I = E_i + \triangle e$. The difference between the voltages E_I and E_{II} is amplified $-A$ times, so that

$$E_o = -A\,[(E_i + \triangle e) - \{-\tfrac{1}{2}B\,(E_i + E_o/A)\}]$$

$$E_o = -A\,E_i - A\triangle e - A\,\tfrac{1}{2}B\,E_i - A\tfrac{1}{2}B\,E_o/A$$

$$E_o + \tfrac{1}{2}A\,BE_o/A = -A\,E_i\,(1 + \tfrac{1}{2}B) - A\triangle e$$

$$E_o = -A\,E_i - \frac{A\triangle e}{1+\frac{1}{2}B} = -A\left(E_i + \frac{\triangle e}{1+\frac{1}{2}B}\right)$$

It follows from this that the drift reduction amounts to $1 + \frac{1}{2}B$. B is approximately 60, so the drift is reduced by a factor of approximately 30.

Triggering

A choice can be made from two methods of triggering, namely, internally from amplifier A or B, or externally. The external trigger input is DC-coupled to the trigger amplifier. The trigger signal passes via the selector switch (Y_A, Y_B or EXT) to the gate of an N-channel field-effect transistor connected as source-follower. This provides a high input impedance, for external triggering namely 1 MΩ; the input capacitance is 10 pF.

Also, by reason of this FET, the feedback of the trigger circuit on the signal to be triggered becomes nil. Choice can also be made between triggering on the positive or on the negative slope of the signal.

The level of the start-point is continuously adjustable on the screen with the aid of the levelling potentiometer. In the automatic triggering position ("AUT"), the start-point is at the zero level of the trigger signal.

In the position external triggering the signal can be levelled over 12 V_{pp}, a facility which experience has shown to be especially important in computer application.

Automatic time base trigger system

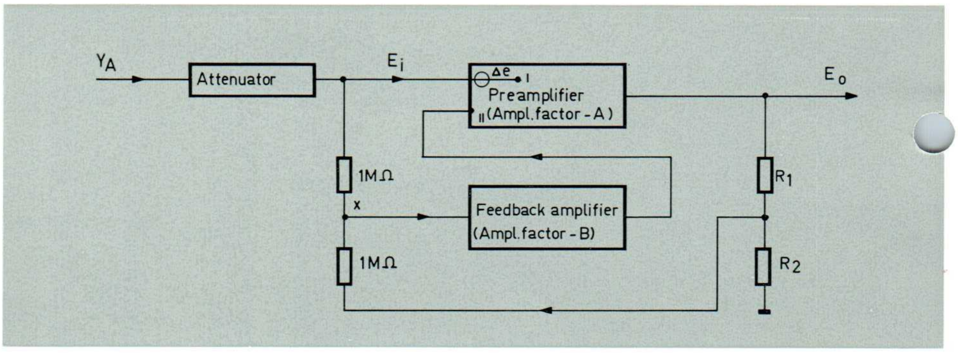
A new, automatic system ensures that the time base trace remains visible on the screen even when no trigger signal is present.

As soon as sufficient signal is present, the time base is switched to the triggered state (see fig. 4).

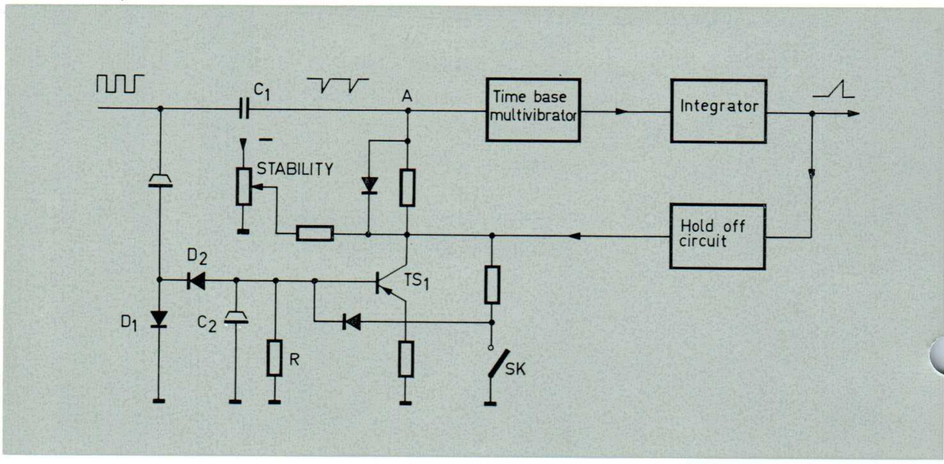
Contact SK is open in the "AUT" position. If no trigger signal is present, the level of point A is determined by the stabilising potentiometer ("STAB"), since in this case TS_1 receives no negative base voltage. This level at A causes the time base to run free. If the trigger multivibrator comes into operation, the base of TS_1 becomes negative through the action of the circuit D_1, D_2, C_2 and R. The level of point A thus becomes less negative and the time base multivibrator will switch over only by reason of the trigger pulses fed via C_1.

The lowest frequency at which the automatic time base trigger system still functions satisfactorily is governed by the time constant RC_2. In the PM 3231, the automatic system operates for sinusoidal voltages of 10 Hz up to 15 MHz.

The automatic time base trigger system is switched off when the levelling potentiometer is at the "level" position. Contact SK is then closed. The level at point A is then such that the time base multivibrator will switch over only by reason of the trigger pulses fed via C_1.

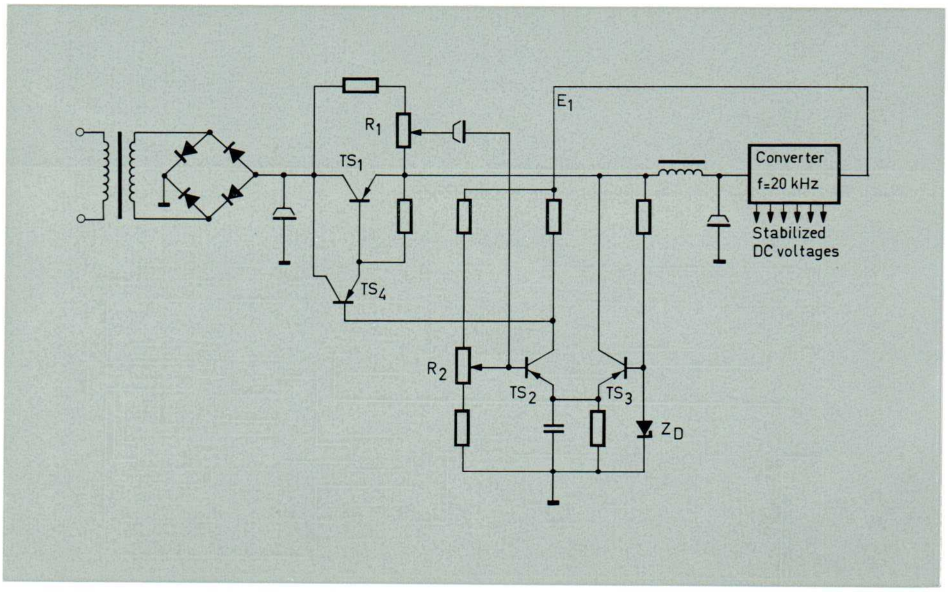


3 Drift compensation

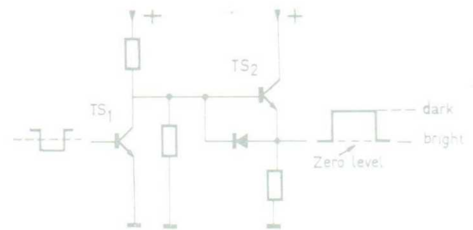


4. Automatic time-base system and time-base

6. Supply unit



5. Beam blanking



Beam blanking

The time base multivibrator controls the opening and closing of transistor TS_1. An emitter follower TS_2 ensures that the screen is rapidly made bright despite the capacitive load of the beam blanking plates, see fig. 5. With this circuit, the period between the starting of the time base and the appearance of the bright image is only 50 nanosec.

Time base accuracy

A bootstrap integrator ensures a very linear time base. The time/div is adjustable in 20 set steps, the tolerance being within plus or minus 5%. Continuous adjustment between the steps is possible.

Stabilized power supply

The power supply unit is so regulated as to equalise mains voltage functuations of plus or minus 15%, see fig. 6. The apparatus can be set by switch for the following mains voltages:

110 V, for voltages between 93 and 127 V, mains frequency 50 to 400 Hz

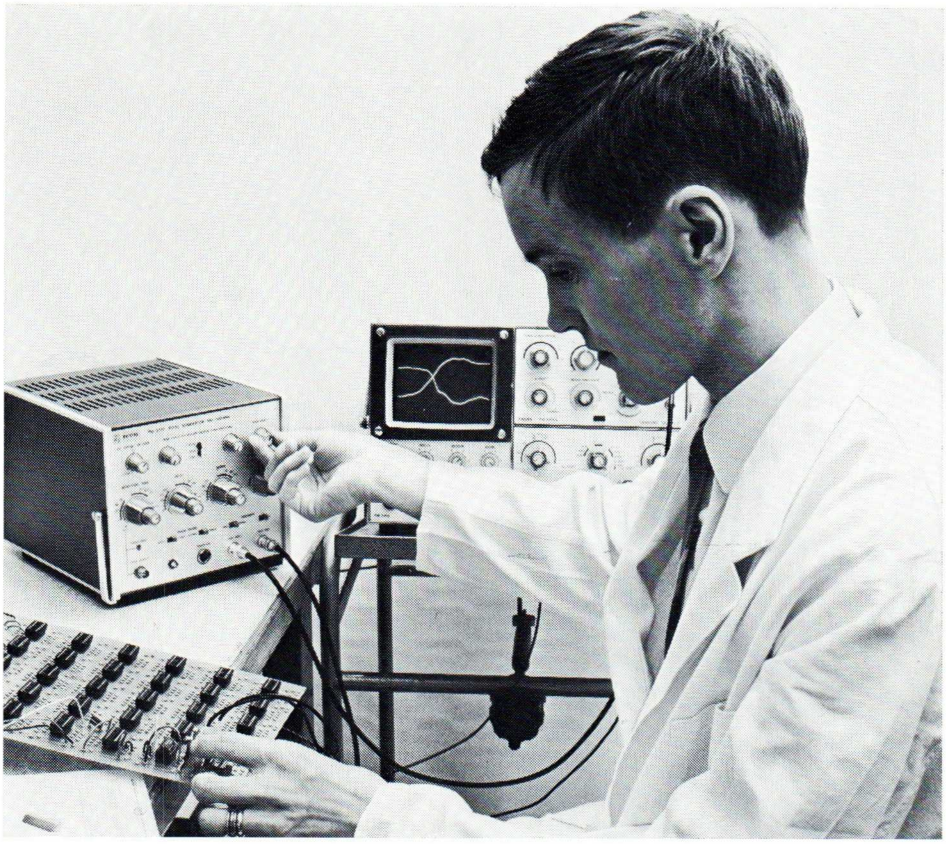
220 V, for voltages between 187 and 253 V, mains frequency 50 to 400 Hz

The output voltage of the converter is stabilized by comparing the secondary voltage E_1 with the reference voltage Z_D in a differential amplifier. This amplifier consisting of the transistors TS_2, TS_3 and TS_4 drives the series transistor TS_1, whilst the voltage E_1 is set to the correct value with R_2. The ripple voltage is set to a minimum with R_1. The converter is an inductive multivibrator which supplies all the voltages, including the high-voltage, for the apparatus. In the event of one of the voltages being short circuited, the converter generator switches off, the supply unit thus being safeguarded against short circuiting.

The E10-130GP cathode-ray tube employed is of the metal-backed phosphor type. This provides a brighter display, in consequence of which pulse shapes of low repetition frequency are still perceptible at high sweep speeds. The screen display may also be photographed with the aid of a camera (PM 9300) and adaptor. Accurate mounting together of the two electron guns ensures a registration error for both deflection systems of less than 0.6 degree in the centre of the screen.

100 MHz pulse generator PM 5770

by L. E. Orrevall



Trends in two important categories of pulse generator can be summarised as an overall improvement in general purpose instruments, referring to basic characteristics such as frequency rise/fall time, pulse delay/width, etc; and, advancing specific areas of performance in special pulse generators. In the latter, exceptional characteristics such as 100 MHz prf (pulse repetition frequency), or sub-nano second rise time are obtained at the expense of limited lower frequency range, maximum pulse-width/delay in the order of a few hundred nano seconds and low duty cycle.

The specific target for the Philips PM 5770 was to achieve a practical combination, between the two categories. This meant extending the overall performance of a "general purpose" generator into areas previously obtainable only with "special purpose" models.

The result is a unique instrument as revealed by the following survey:

Repetition frequency	: up to 100 MHz	but also	down to 1 Hz
Rise/fall time	: down to $<$ 4 ns	,,	up to 100 μs
Delay/width	: down to 5 ns	,,	up to 100 ms
Amplitude	: up to 10 V	,,	down to 30 mV
Polarity	: positive/negative	,,	normal/inverted
Basic features	: external triggering	,,	synchronised gating
	single shot operation	,,	double pulse operation
	synchronising output	,,	DC-offset
Performance and design	: most complete	,,	most compact weight only 7 kg

The above performance was accurately defined by a customer who discribed it as: "All the spec's I have seen somewhere but never realised in one box!"

Lay-out of the instrument

Fig. 1 shows the block diagram of the PM 5770 together with its functional controls and specification ranges. The block diagram is straight forward and complete in the sense that the operator is given the opportunity to vary all the basic pulse parameters.

Functional control arrangement

Fig. 2 shows the high standard achieved by ergonomic arrangement of the controls. Every item is logically placed relative to its purpose and sequence of use.
The upper part shows the details of an individual pulse. The lower part defines the relationship between sequential pulses (drawn in the double pulse mode).
All the parameters shown can be varied individually by means of the front panel controls.

Applications

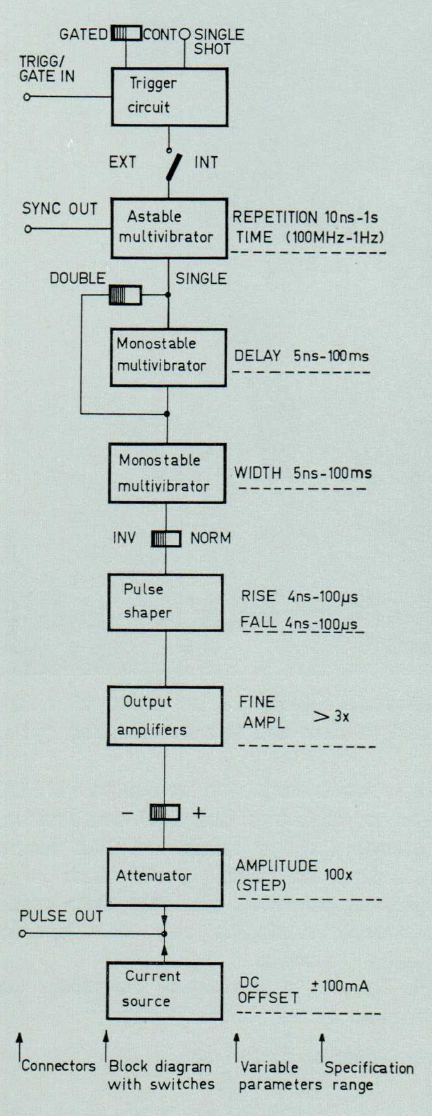
The PM 5770 can be used for all kinds of work with digital circuits. It is well suited for testing integrated circuits (IC) and systems built from IC's. The speed is sufficient for DTL and TTL circuits (and in many cases also E^2CL and ECL).

E^2CL - counter

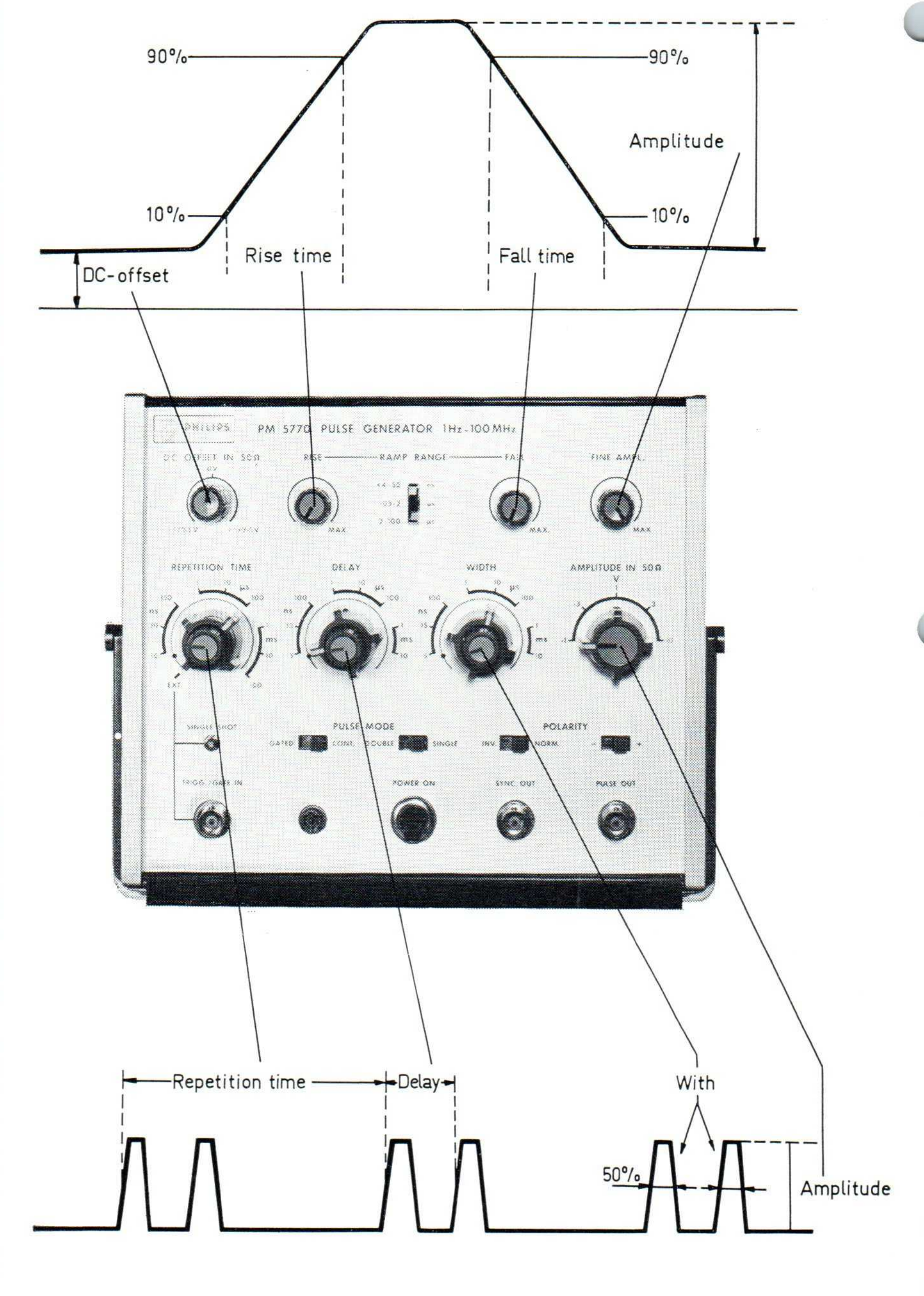
The photo fig. 3 was taken at an exhibition of integrated circuits. It shows a demonstration system made by Mullard, consisting of E^2CL circuits connected as various types of counters. In the situation shown, they were wired to form a divide-by-eight circuit, which performed well up to 40 MHz clock frequency. The PM 5770 was used as a clock generator and delivered a negative pulse with 0.7 V amplitude to the input of the circuit. Both the input and the output waveforms were monitored on a Philips' sampling oscilloscope PM 3419.

Testing TTL-gate

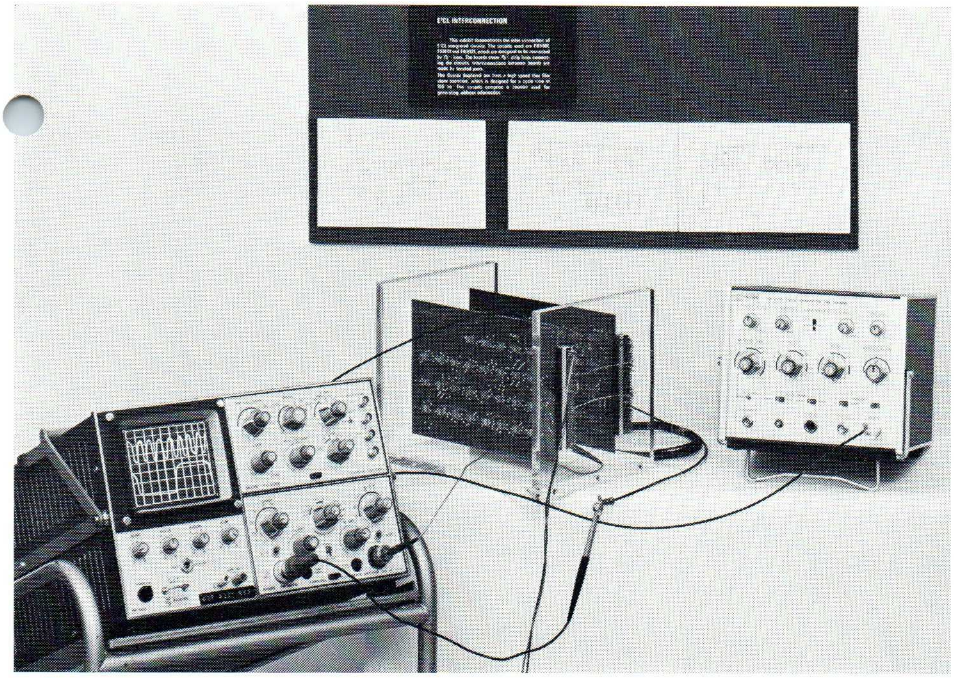
The photo, page 3, shows an investigation of a TTL-gate (Philips FJH III or Texas SN 7420). The measurements were made in accordance with a test circuit that is prescribed by the manufacturer of the integrated circuits. The pulse generator was set to 3 V amplitude, repetition time 1000 ns, pulse width 500 ns. Rise- and



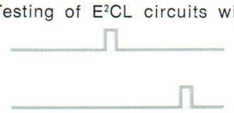
1. Functional block diagram with abreviated specifications of the PM 5770

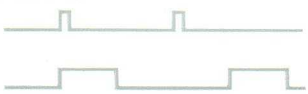


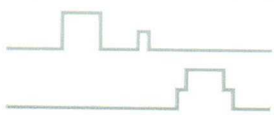
2. Ergonomic arrangement of the controls



3. Testing of E²CL circuits with PM 5770







4. Dual channel test pattern

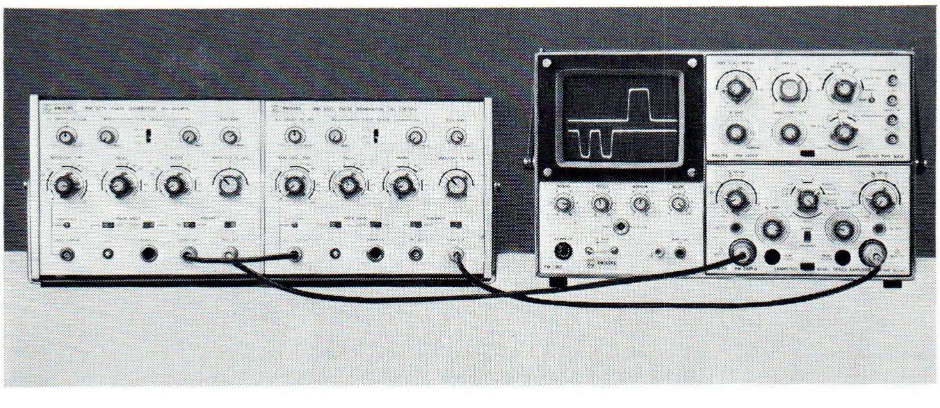
6. Double pulse on two channels

7. Mixed pulses with different amplitudes or complex waveforms

8. Gated pulses, the gate pulse being supplied by one of the generators. Gating, is synchronised

9. Gated double pulses

5. Two PM 5770 pulse generators coupled as dual channel test set



fall time has to be less than 15 ns. The test circuit prescribes that the gate being tested has to be driven by a similar IC gate. The particular item, chosen for the demonstration was a "Dual 4-input positive NAND gate" which contains two identical gates. It was therefore possible to use one gate to drive the other and the only thing which had to be assembled was the load-circuit which simulates a fan-out of 10. The propagation delays could be measured by means of a sampling oscilloscope (PM 3419) for the two transitions, namely from 0 to 1 and from 1 to 0.

Dual channel test arrangement

Many logic applications such as tests on AND gates and shift registers require a dual channel test set, see fig. 4.

Both pulses on channel A and B should be independently adjustable in delay and pulsewidth.

Although the PM 5770 is in principle a compact one channel generator, it can very easily be connected to a second PM 5770 to achieve a dual channel test set, the second PM 5770 is set to the position EXT and triggered by the sync output of the first PM 5770.

The instruments can also be mechanically combined.

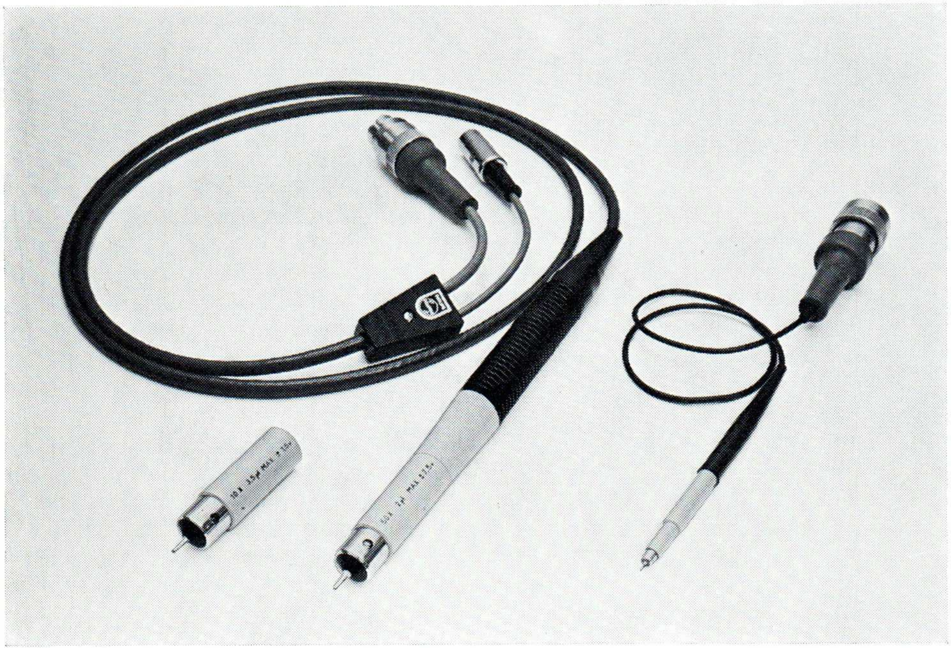
Mounted together, they form a complete 19" wide instrument which can be easily rack mounted if required, see fig. 5.

Using the "gate" and "double" pulse facilities, the above dual channel set is also very useful for many other pulse sequences e.g., see figs. 6, 7, 8 and 9.

The above facilities contribute a degree of general purpose usefulness that makes the PM 5770 welcome in any place where the word "pulse" is employed.

Built-in probes for sampling oscilloscope

by L. E. Orrevall



How to do it yourself:

An earlier article has discussed the use of cathode-follower probes and passive probes, see the photo above. It was also pointed out that it often is advantageous to build the test-circuit in such a way that a 50 Ω cable directly can be connected between the circuit and the sampling oscilloscope. This article will give practical directions for the design of two types of probes, namely resistive and inductive probes.

Restitive probe

This type is used for measuring the voltage at a certain point in a circuit, for instance the collector of a transistor. The principal is then, first to transform the voltage into a current proportional to the voltage by the use of a series resistance. The current is fed into a 50 Ω cable which is terminated at both ends. One of the terminations consists of the input of a sampling oscilloscope. Assume for instance that the collector resistance is 120 Ω. Let us also assume that we shall make a probe 1 : 100. In this case the series resistance shall be 2475 Ω, with 50 Ω terminations at both ends of the cable, making an effective load of 25 Ω. The probe is loading the circuit by less than 5%. The circuit will then look as in fig. 1.

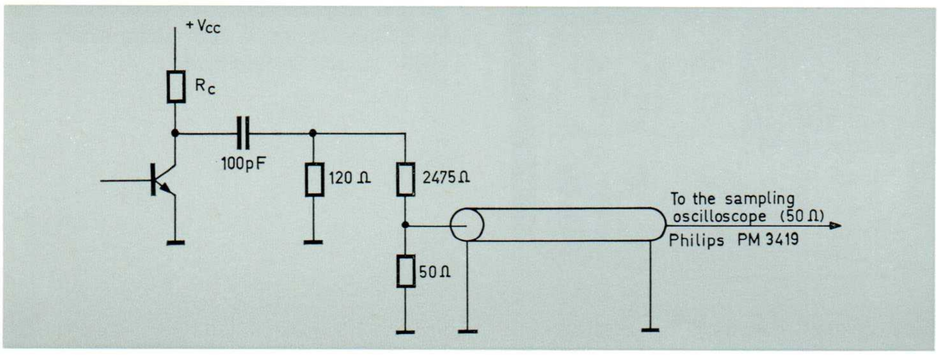
A problem associated with the 2475 Ω resistor is that the shunt capacitance which is approx 0.3 pF would give rise to a serious peak in the 50 Ω load. One way to decrease this is to use two resistors in series. Use 1 kΩ + 1.5 kΩ resistor and select one or both values so that the 1 : 100 division is reasonable correct. In order to further decrease the capacitive feed through it is good to connect a grounded screen with a hole for the first resistor.

The cable should be terminated at both ends. The 50 Ω termination at the beginning of the cable is best made by using two 100 Ω resistors in parallel. The mechanical lay-out can be made as in the photo fig. 2. Resistors are Philips 1/8 W. This type of probe is good enough for 1 ns rise time.

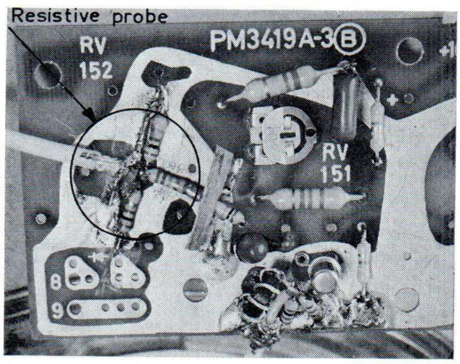
Inductive probe

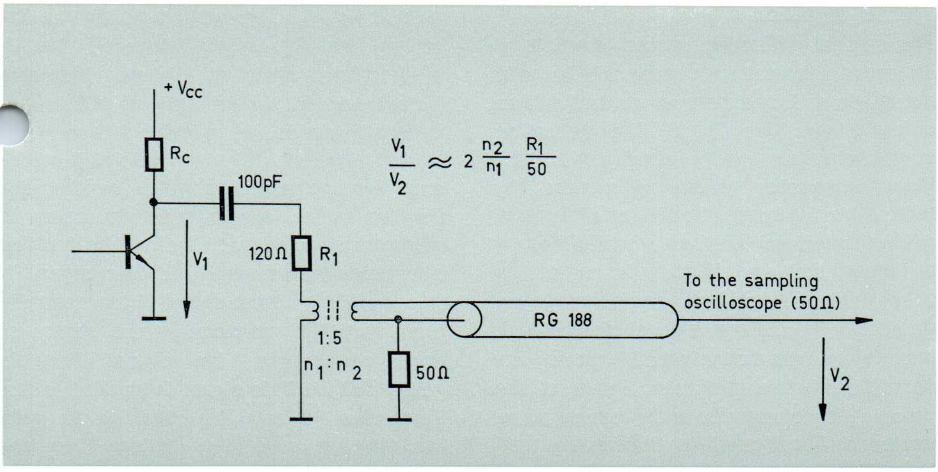
The second possibility to measure the behaviour of the same circuit is by measuring the collector current. As we are only interested in the transient behaviour it is feasible to measure the pulse current by means of a transformer. A good method

1. Fast pulses have to be transported by transmission-lines. 50Ω coaxial cable are simple and inexpensive. Pulses are fed into the cable via a series resistance, the value of which is chosen to give a 1 : 100 voltage division. In this way one can make a passive probe with 100 times attenuation.



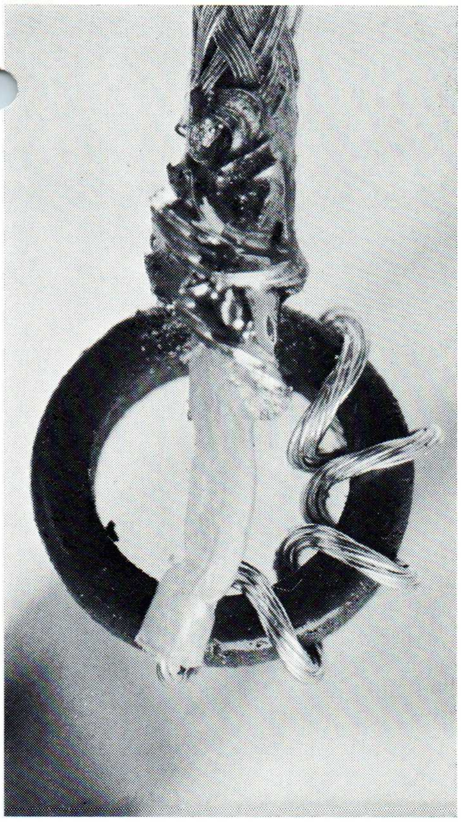
2. The photo shows a practical circuit of the type which is described in fig. 1. The transistor operates in the avalanche mode. The collector voltage is shown in fig. 5





3. Current transformers can be applied for nanosecond phenomena. The figure shows the schematic diagram for an inductive probe and how it can be connected in the same electronic circuit, as in fig. 1

4. The photo is a close-up of the winding of a toroidal transformer



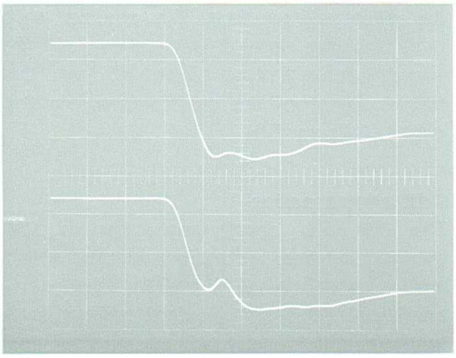
of making such a transformer is to use a ferroxcube toroid (Philips 2 P 65331). The collector current is fed through the hole of the toroid forming a one turn primary winding. The secondary winding can be made in the following way. Use a standard teflon 50 Ω coaxial cable (as RG 188). Strip off the screening and the outer cover but leave approx 5 cm of the inner conductor. This wire is wound 5 turns around the toroid and soldered to the cable screen. Fig. 3 shows how the transformer is added to the same circuit, as was used for the resistive probe.

In this way the transformer adds only 1 Ω in series with the collector resistor. The current conversion ratio of the probe becomes 1 : 10 by the addition of the 50 Ω — termination of the cable in the sending end. Again, the termination is best done with two 100 Ω resistors in parallel (Philips 1/8 W).

Fig. 4 is a photo showing how the winding of the toroid is done in practice.

The last photo, fig. 5, gives a comparison between the waveforms measured with resistive and inductive method. The extra peak from the inductive probe is due to capacitive coupling.

5. The oscillogram shows the waveform at the collector of the avalanche transistor. The upper trace originates from a resistive probe while the lower trace comes from an inductive probe. The oscilloscope is set to 1 ns/cm.



The PM 3200 a new concept in oscilloscopy

by H. Toorens

A full performance oscilloscope which is truly portable

The PM 3200 a new wide application lightweight instrument that offers simplicity of operation, high sensitivity and rugged construction at low cost.

The main characteristics of this instrument are:

Sensitivity: 2 mV/div

Bandwidth: DC up to 10 MHz

Automatic DC balancing and drift control

Trace drift: less than $^1/_4$ div/hour

Sweep rates: 100 nsec/div up to $^1/_2$ sec/div

Fully automatic triggering of the entire frequency range ensures an extremely stable display of any input signal and a bright trace even at the highest sweep speeds

In this article the different circuits and the working principles of this instrument will be described.

Vertical amplifier

The vertical deflection system consists of an input attenuator, a preamplifier with automatic drift and balance control and an output stage. See fig. 1. The deflection coefficients are partly obtained by the input attenuator which attenuates by factors 1 : 1, 1 : 10, 1 : 100 and 1 : 1000, while the factors between are obtained by switching the amplification factor of the preamplifier. The advantages of this system of sensitivity setting are: a simple high impedance attenuator which needs few corrections while the noise figure of the preamplifier is reduced in the less sensitive positions. The Y-amplifier is transistorised to a large extent, and only in the input stage is a valve used. The first stage of the Y-amplifier is a push-pull unit; the second input being used to supply the drift compensation voltage from the compensation circuitry.

The input signals are applied to a cathode follower followed by an emitter follower, thus ensuring a high input impedance. The next stage consists of the transistors TS_2, TS_4 and TS_5, the gain of this stage mainly being determined by the ratio of the emitter feedback resistors (R_s) to the parallel feedback resistor of transistor TS_5 (R_p), thus resulting in a very stable gain. The output stage consists of two balanced transistors TS_7 and TS_8 with a series feedback, followed by two single ended push-pull sections with a shunt feedback, driving the vertical deflection plates of the cathode ray tube. The deflection signal, derived from the collector of TS_5 of the preamplifier, is applied to the base of TS_8, the shift voltage from the vertical shift control being applied to the base of TS_7. The trigger signal is picked-off from the output signal of the output stage via the emitter follower TS_9.

Drift compensation

The PM 3200 oscilloscope has a DC coupled vertical deflection system which has a sensitivity of 2 mV/div. Every DC amplifier shows some output voltage fluctuations which may not be proportional to the input voltage applied. In an oscilloscope amplifier these fluctuations cause zero-line shifting. These fluctuations may be caused by temperature variations or statistical variations in the amplifier parameters. There are several well-known systems to reduce drift figures, a combination of which has been applied in the vertical preamplifier stage of the PM 3200 oscilloscope. To reduce drift error, the whole preamplifier stage is a pushpull circuit. In this description, every voltage variation in the preamplifier, which may cause drift, has been referred to one input of the preamplifier, as V_d. See fig. 2. The input voltage V_i is applied to input. The control voltage is applied to input II. Thus the signal at terminal I is the sum of the input voltage V_i and the drift voltage V_d. Now the gain of the preamplifier is A and the phase shift is 180° while its output voltage is V_o. This output voltage is inserted into the output stage of the Y-channel and also fed into an attenuator (attenuator II), by means of which the voltage is divided by a factor A thus being V_o/A. Every time when the gain of the preamplifier is changed, to obtain another deflection factor, the attenuation factor is changed also, so that the signal behind this attenuator is always V_o/A. This signal is mixed with the input signal V_i by means of the two 1 MΩ resistors. At the centre of the potentiometer the signal then is $^1/_2V_i + {}^1/_2V_o/A$. This signal is applied to the input of a DC amplifier of which the gain is: $-B$. The regulation voltage thus obtained is $V_r = V_{II} = -\,^1/_2\,B\,(V_i + V_o/A)$.

The total signal between the inputs I and II of the preamplifier, then is equal to $V_I - V_{II} = V_i + V_d - (-^1/_2\,B\,(V_i + V_o/A))$

Since the gain of this amplifier is $-A$, the output signal is equal to:

$$V_o = -A\,(V_i + V_d - (-^1/_2\,B\,(V_i + V_o/A))), \text{ or}$$

$$V_o = -A\,V_i - \frac{A}{1 + {}^1/_2B}\,V_d$$

So the factor by which the drift is reduced is: $1 + {}^1/_2B$, resulting in a drift of maximum $^1/_4$ div/hour. Furthermore is it possible, because of the "zero-line locking", to obtain a system of sensitivity setting as described above without jumping of the trace.

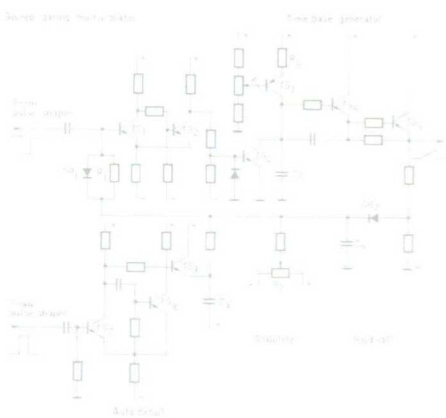
X-deflection

The X-deflection is obtained by means of a balanced cascode amplifier which drives the X-deflection plates of the cathode-ray tube. To this amplifier both the X-deflection signal and the X-shift voltage from the X-shift control are applied. By means of a switch (internal – external X-deflection) a choice can be made between a signal derived from the X-deflection input via an emitter follower, or a signal generated by the sawtooth generator. A block diagram of the total X-deflection channel is given in figure 3.

To obtain an accurate time axis and make time and frequency measurements possible, a voltage, increasing linear proportional to time has to be applied to the X-deflection plates (the i.e. X output amplifier). Such a signal is generated in the sawtooth generator, the working principle of which can be seen in figure 4. The sawtooth generator consists of a capacitor (C_1) which is charged by a constant current source: the collector of TS_3. The charging current is determined and kept constant by means of the series feedback resistor in the emitter lead of TS_3. The different time per division setting are obtained by switching both the load capacitor (C_L) and the load resistor (R_L). The like this generated linear increasing voltage on C_L is then applied to two cascaded emitter followers, from the latter of which the deflection signal is fed into the X output amplifier. When this voltage has reached a certain level, the base of TS_1 becomes more positive than its emitter through diode and resistor R_1 so that the Schmitt trigger (TS_1 and TS_2) will switch. TS_1 then becomes conductive while the TS_2 becomes non-conductive. As a result of this, TS_3 will becomes conductive short circuiting the load capacitor C_L and the output voltage of the sawtooth generator will become zero again.

To get the Schmitt trigger TS_1—TS_2 switched again, the base of TS_1 must become more negative than its emitter. This is not possible, until the voltage on capacitor C_h, which has been charged to the maximum value of the sawtooth signal, by means of diode 1, decreases to a preset value (R_2: stability). During this time (the hold-off time), the capacitor C_h will be completely discharged, so that the next sawtooth will start at the very same value. The stability control R_2 is so adjusted, that the Schmitt trigger TS_1—TS_2 will switch. As a result of this, a new sawtooth will be generated. The sawtooth generator is thus free running. However, to get a stable display of an input signal, every sweep of the sawtooth generator has to start at the same level of the input signal. The sawtooth generator is then triggered.

As may be seen in figure 3 the trigger signal is derived either from the vertical channel, the external input, or the line frequency. After amplification and level setting, this trigger signal is applied to a pulse shaper, i.e. a Schmitt trigger. The output signals of this Schmitt trigger are constant in amplitude and rise time. When this output signal is applied to the base of

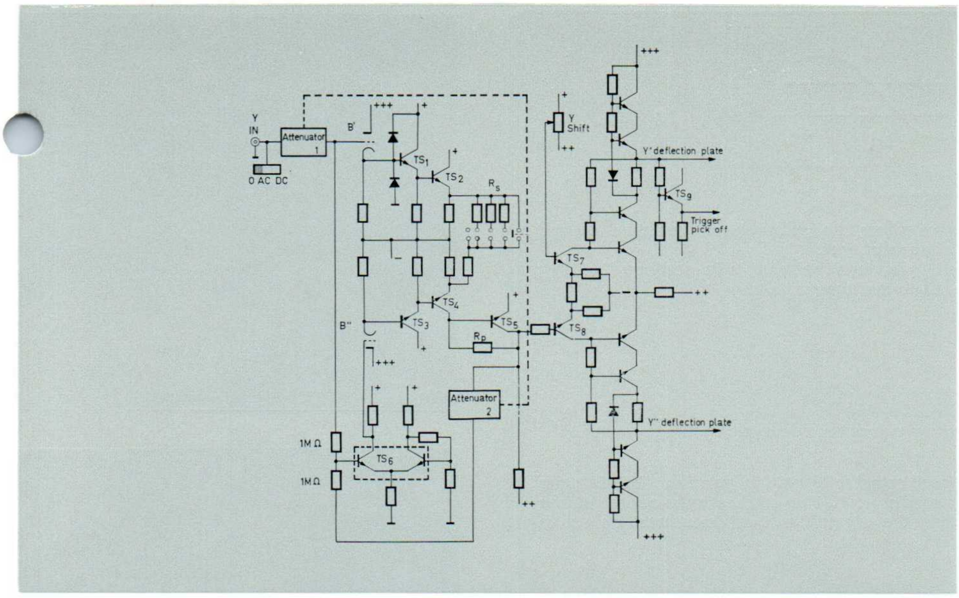


4. Time base generator with hold-off circuit and auto-circuit

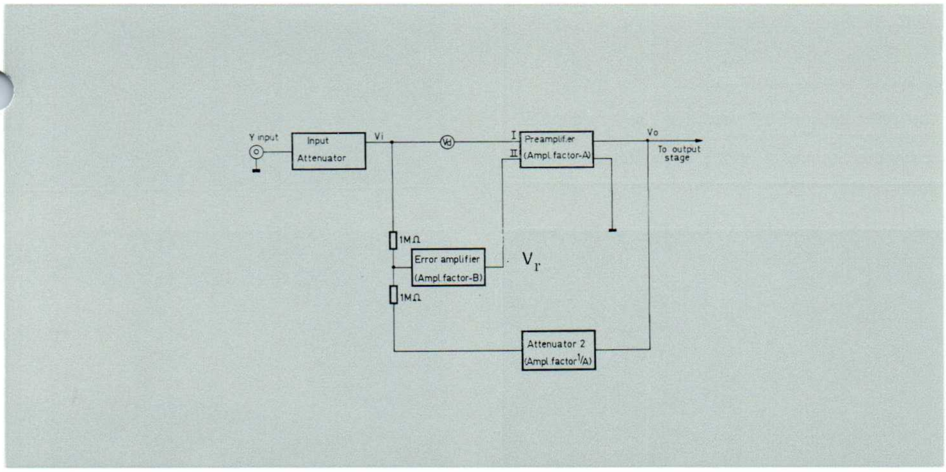
TS_7, after differentiation, the one-shot multivibrator consisting of TS_7 and TS_8 will switch. TS_7 becomes non-conductive, while TS_8 becomes conductive (see figure 4). As a result of this the capacitor C_a in the emitter lead of TS_9 will be charged. As may be seen in figure 4, the tension on the base of TS_1 will raise. This implies that after the hold-off time the sweep gating multivibrator (TS_1 and TS_2) cannot switch again, unless a trigger pulse after differentiation is applied to the base of TS_1.

This trigger pulse is also derived from the trigger pulse shaper. When there are no pulses from the pulse shaper, the time base will not be triggered, and the one shot multivibrator will switch back into its rest position. As a result of this, the capacitor C_a will be discharged down to the level, which has been preset by R_2, so that the time base again becomes free-running.

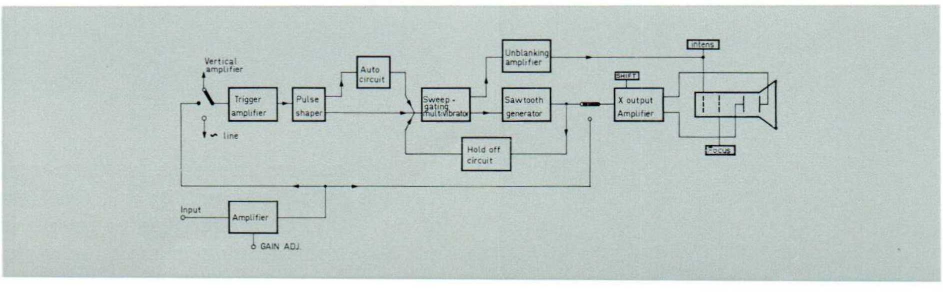
Conclusion: As the auto circuit consists of a one shot multivibrator, even signals with pulse widths of 100 nanoseconds will stop the time base from free-running and trigger the time base generator. It is due to this fact that even signals with duty cycles of 1 : 2 x 10^7 will be within the trigger possibilities, automatically without stability or level controls.

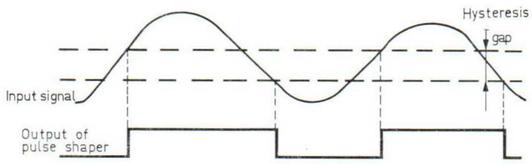


1. Vertical deflection system showing preamplifier, drift compensation circuit, output stage and trigger pick-off

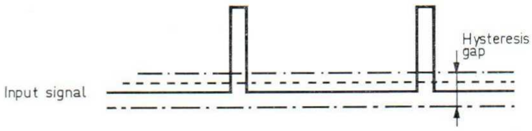


2. Simplified block diagram of drift compensation circuit
3. Block diagram of the X deflection system

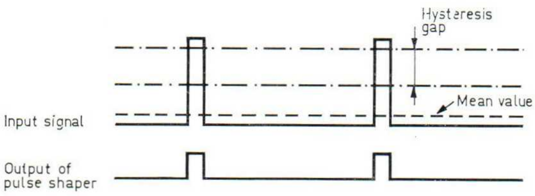




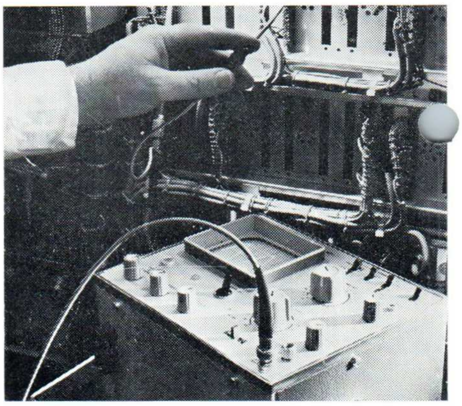
5. Indicates the levels of the input signal at the point where the pulse shaper switches on and off, due to the "hysteresis gap" at the input of the pulse shaper.
The figure shows how triggering occurs when the trigger level selector is in position "MEAN"; the mean value of the input signal is near by or with in the hysteresis gap

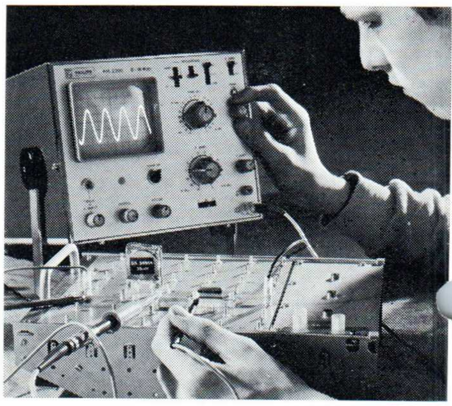


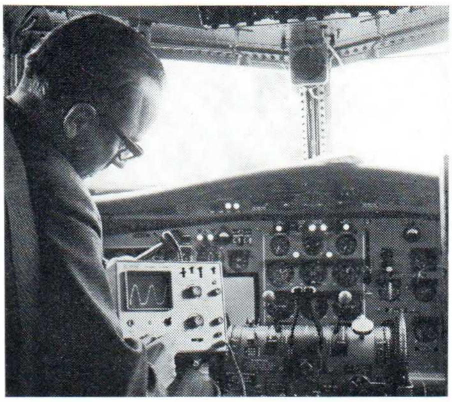
6a. Input signal with large duty cycle while trigger level selector is still in position "MEAN"; the input signal does not pass both sides of the "hysteresis gap". The pulse shaper cannot switch on and off: the time base cannot be triggered. Now the trigger level selector must be set in position "TOP"



6b. Trigger level selector in position "TOP". Due to top detection and DC restoring the "hysteresis gap" is shifted towards the top value of the input signal. The pulse shaper can now be switched; and the time base can be triggered nearby the top value of the input signal







The pictures on this page and the next illustrate fully the versatility and flexibility of the PM 3200, in the field of research and development, for education, industrial maintenance and product line testing.

Level control

The pulse shaper, which is a normal Schmitt trigger, switches on and off at different values of its input level. This is shown in figure 5. The difference is the so-called hysteresis gap. As may be decided from figure 5 the input signal has to pass both sides of the hysteresis gap to obtain triggering. The relationship between the level of the input signal and the level of this hysteresis gap depends on the position of the level control switch:

a. *MEAN.* The trigger circuit is AC coupled. In position Mean and HF reject of the level selector switch, the position of the hysteresis gap of the pulse-shaper is near zero. This results in the pulse-shaper switching on and off due to the AC coupling, near by the mean value of the AC input signal.
When this signal has such a large duty cycle that the mean value is in the hysteresis gap, and only one side of this gap is passed (see fig. 6a), the trigger selector switch must be set in position "TOP".
b. *TOP.* In position TOP of the trigger level selector, the trigger level is automatically derived from the input signal applied. By means of top detection and a DC restorer, the hysteresis gap shifts towards the top value of the input signal. In position "—", the level is set at positive tops. In position "+", the level is set at the negative tops. So changing the polarity means, changing level! (see fig. 6b). It has then the added advantage, that in both positions of the trigger slope selector, the largest possible part of the leading edge is displayed.
c. *HF reject.* In position HF reject, the trigger signal first passes a demodulator and low-pass filter, thus triggering on modulated amplitude signals.

Power supply

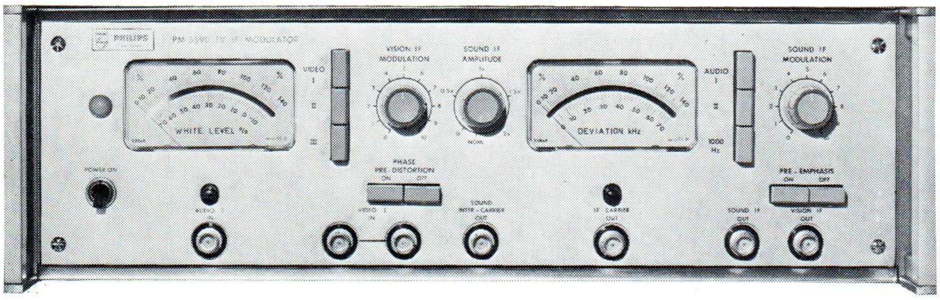
The power supply of the PM 3200 consists of a mains transformer with rectifier, a voltage stabiliser, and a DC-DC converter. All voltages which are used in the PM 3200 are derived from this DC-DC converter. The DC input voltages of this converter are very constant too. This system has two big advantages: All performances of the instrument are independent of power line voltage variations as the input voltage of the stabiliser may be 22 . . . 30 V, also DC powered operation is possible. For this type of operation a battery pack with rechargeable batteries is available.

Conclusion

As may be concluded from the description above, the PM 3200 represents a new class of oscilloscopes. It combines a large bandwidth with a high sensitivity, while the drift figure due to the drift compensation circuit is barely negligible. The new trigger circuitry, the PM 3200 combines simplicity of operation with stable triggering over the entire frequency range. The PM 3200 can be powered from several sources: mains, DC (24 V), or additional battery pack (PM 9391).

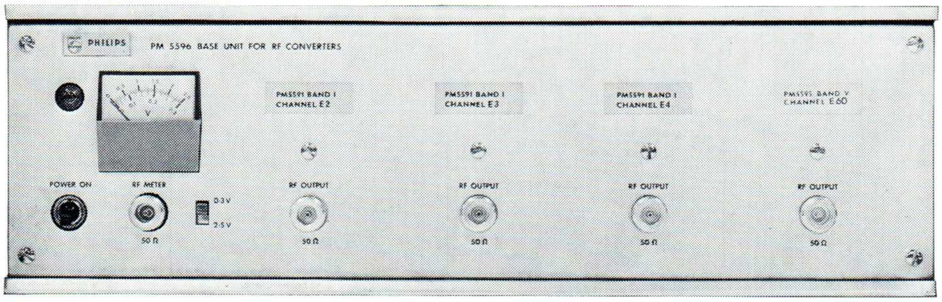
IF distribution of television signals

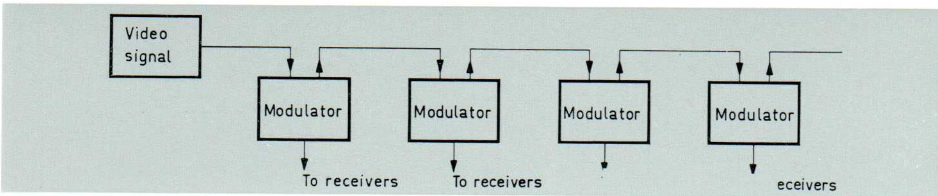
by E. Helmer Nielsen



IF modulator PM 5590

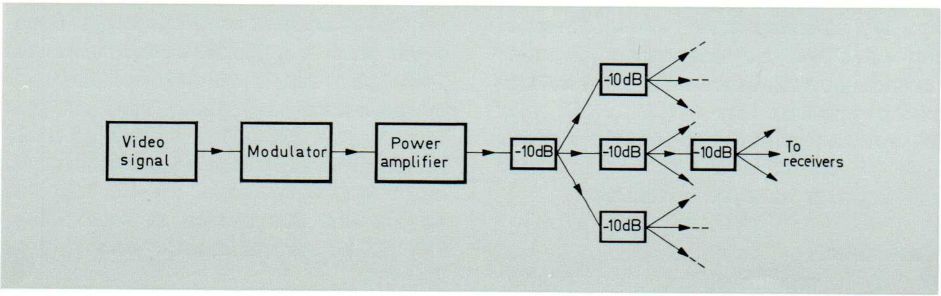
Base unit PM 5596 with four converters





1. Video distribution

2. RF distribution



Distribution of TV signals

It is often necessary to distribute high quality television information over longer distances and to larger numbers of television receivers. This problem is typical for the setmaker industry, where the various signals have to be distributed in the factory and laboratory etc. Also in the more professional fields the same problem arises, for instance in TV broadcasting centres.

Requirements

The demands to a high quality distribution system are very heavy. The signals are generally used as measuring signals, among others for colour TV, and therefore only negligible deterioration can be tolerated. The primary requirements are small linear as well as non-linear distortion. Also freedom from reflection distortion is important.

The system should be able to supply as well VHF as UHF signals and with equally good quality. Also distribution of broadcasted TV signals may be important.

Video distribution and RF distribution

The traditional distribution system consists of three basic elements:

a. the modulator
b. the distribution network
c. the receiver terminal

Sometimes the sequence of the first two elements are reversed, and it then becomes:

a. the distribution network
b. the modulator
c. the receiver terminal

In case the distribution takes place before the modulation we will call the system a video distribution system. The video signal is distributed to a number of modulators, which will then be placed close to the receiver terminals.

The figure 1 shows in schematic form the set-up used.

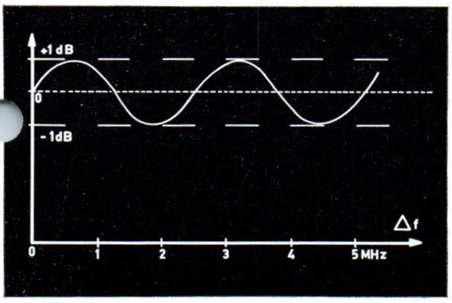
Distribution on video basis is quite flexible because the video signal can be looped-through all modulators. The cable losses are small, and it is therefore possible to use relatively cheap cables. The losses are unfortunately frequency dependent. They increase with the frequency and it is therefore often necessary to use amplifiers with frequency compensation.

The number of modulators used will be large, and it is therefore for practical and economical reasons necessary to use simple types of modulators. Features as vestigial sideband characteristic and phase equalization cannot generally be considered. Distribution of aerial signals is not directly possible.

The other possibility, which was mentioned, use at first modulation and then afterwards distribution of the signal. In this case the signal is distributed as RF signal and then we speak about RF distribution.

The fig. 2 shows in schematic form a typical set-up.

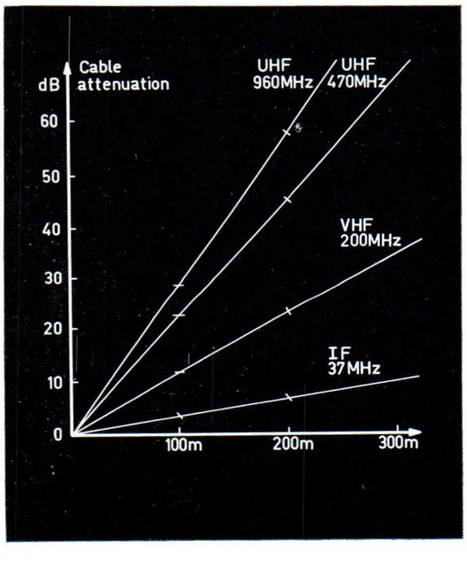
In this case we use only one modulator, and this one can therefore be fitted with all features as for instance VSB (vestigial sideband) filter, phase correction network etc. The distribution network will attenuate the signal noticeably, and therefore the power, which has to be delivered by the modulator, is relatively large. The modulator is often followed by a power amplifier, which then delivers the RF signal at the wanted channel to the network. Instead of the combination modulator (on RF) + RF amplifier one will often choose modulation on IF + RF power converter.
The distribution system consists basically of a number of nodes. In each node the incoming signal is divided between several outgoing cables. The division has usually to take place by means of resistive networks, and the insertion loss in each node



3. Responce due to reflection from approx 50 m cable, VSWR = 1.2

is therefore considerable. The network will often be quite large and complicated, and it is not flexible as regards to changes and extensions. In large systems one has also to be extremely careful with the set-up in order to avoid distortion due to reflections.
Fig. 3 shows an example of the distorted response due to reflections on only 50 m of cable.
At the higher frequencies in question the cable losses play an important role. Fig. 4 shows how the losses especially at UHF very soon increase to intolerable values. Colour TV is often transmitted at UHF, and we see that even 200 m of quite a good cable as RG 213U give an attenuation of 40 – 50 dB at these frequencies.

4. Illustration of the losses by cable RG 213U



This means that distribution of UHF signals is in practice limited to only short distances, say 10 – 20 meters.

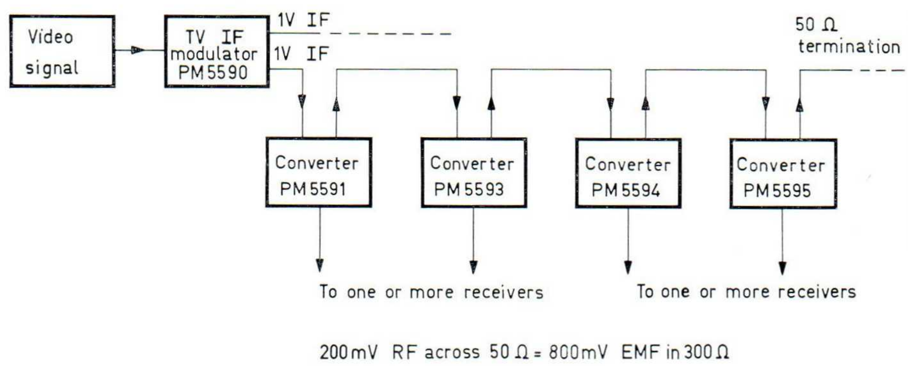
The properties of video and RF distribution
When we compare the properties of the two different distribution systems described above viz. the video distribution and the RF distribution we see that both principles have their advantages and disadvantages. A comparison can best be made if we describe in schematic form the main features for the two systems.

Video distribution
1. The distribution is of the string type. It is flexible and it can be easily extended.
2. The signal is looped-through. There is no or little distribution losses.
3. The cable losses are small.
4. The cable losses are frequency dependent within the video spectrum.
5. There is only little risk of reflections.
6. Large number of modulators necessary. Only simple types of modulators are used in practice.
7. Aerial signals cannot be distributed.

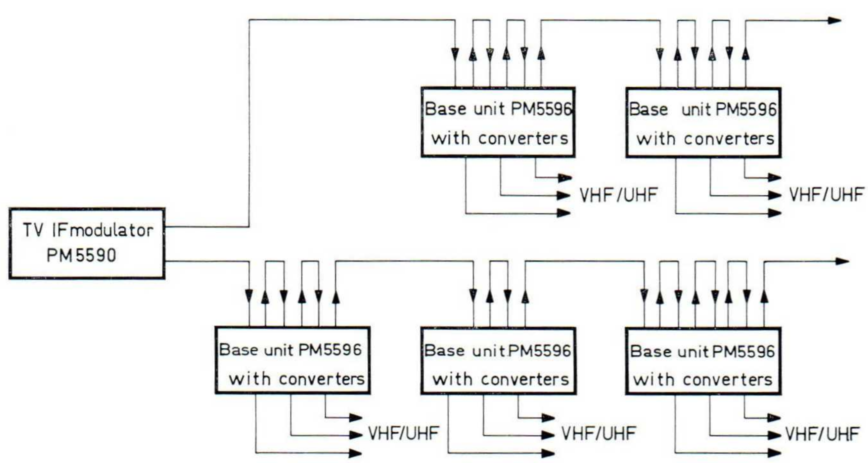
RF distribution
1. The distribution is of the star or nodal type. The system is less flexible.
2. The signal is divided in resistive networks. There are large distribution losses.
3. The cable losses are large.
4. The cable losses are constant within each channel.
5. The cable reflections present big problems.
6. Only one modulator needed. All features as vestigial sideband filter, phase equalizer etc. can be realized.
7. Aerial signals may be distributed.

In the scheme above the points 1, 2, 3 and 5 are in favour of the video distribution and the points 4, 6 and 7 in favour of the RF distribution. The various points are not of equal importance. Point 6 for instance dealing with the properties of the modulator is a very important one.

The IF distribution system
The ideal system should as far as possible combine the advantages of both the above described principles. This seems at the first moment impossible but it is nevertheless realized in a new principle of signal distribution. We will call this system the IF distribution system.
The figs. 5 and 6 show how the system is built-up.
There is only one modulator, which operates on IF. The signal from this modulator



5. The principle of the IF distribution system



6. A practical set-up of the IF distribution system

is distributed in a similar way as a video signal by looping it through a number of RF converters. After each converter a small short secondary RF distribution system can be placed.

If we compare this new system with the two old systems described above we find the following basic properties.

a. The distribution is of the string type just like the video distribution. Therefore the new system has the same flexibility, and it can be easily extended like the video system.
b. The signal is looped-through all converters. The insertion loss is negligible ($<$ 0.02 dB).
c. At the IF of 35 – 40 MHz the cable losses are still small. A length of 300 m RG 213U cable will give an attenuation of 10 dB only. This loss can be compensated for in the RF converters.
d. The frequency is, however, still so high that one can consider the cable losses as being constant over the 5 MHz video band. Therefore frequency dependent cable compensation is not necessary.
e. The system has a very high immunity against cable reflections. This feature is due to the use of a high precision directional coupler at the IF input of each converter. This feature will be described more in detail lateron.
f. Like it was the case with the RF distribution only one modulator is needed. Therefore it is possible to use a modulator of high quality and to incorporate VSB filter, phase equalizers, elaborate clamp circuits, meter indication of modulation etc.
g. Aerial signals can be distributed without demodulation. It is only necessary in a simple converter (TV tuner) to convert the antenna signal to the IF range which is being used in the distribution system.

From the above survey we can see that it is really possible to combine the features and flexibility of the video distribution system with the advantages of the RF system.

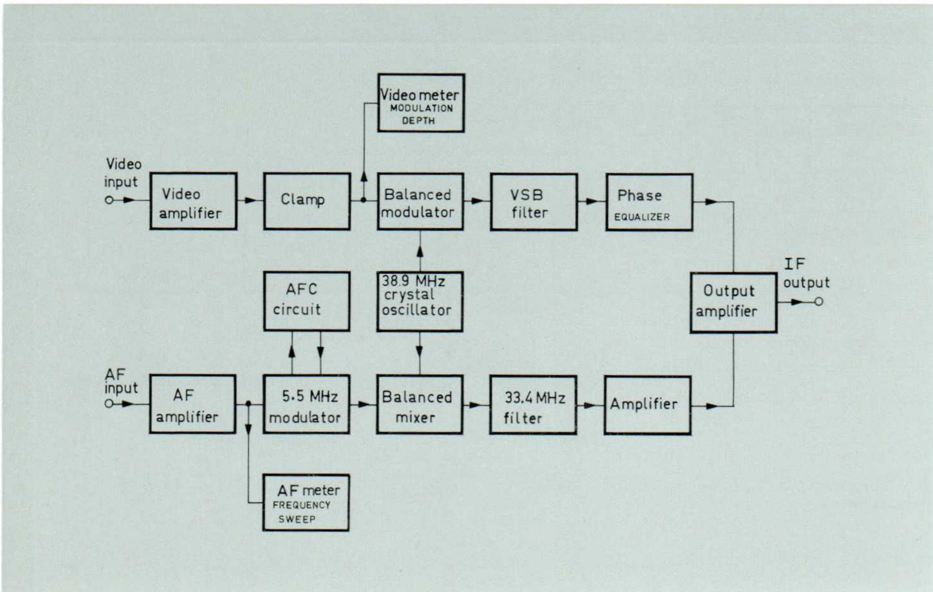
The IF modulator PM 5590

The basic component in the system is the IF modulator (see fig. 7). The type number of this apparatus is PM 5590. This is a combined vision and sound modulator operating at the IF of 38.9/33.4 MHz (Versions for the American, British and OIRT TV system are also available, the latter two with differing IF's at 39.5/33.5 MHz and 38/31.5 MHz respectively). The PM 5590 is fully transistorized and it is housed in a small 19″ cabinet, only 13 cm in height. The video modulator is a balanced diode modulator placed in a temperature controlled oven. The modulator is driven from a balanced amplifier incorporating an elaborate clamp circuit. The video signal is clamped at the back-porch, but without any influence on or from the colour subcarrier signal. To achieve this the signal is split up into a low-frequency and a high-frequency branch, and only in the low-frequency part the DC component is restored; the high-frequency part, however, passes on practically all video information including the colour subcarrier. The video modulation meter measures the peak white level at the input to the balanced modulator. The reading of the peak voltmeter is not influenced by the presence of the colour subcarrier. The modulator is followed by a sideband filter and a phase correction circuit for this filter. Phase correction directly at the modulated IF frequency is far more effective than phase correction of the video signal.

Another phase correction circuit is included for possible phase pre-distortion according to a standard receiver distortion. The phase correction circuits used are of an active type. By isolating the stages in the phase correction network from each other by means of transistors, one obtains the big advantage that each stage can be aligned without mutually influencing the other. It is therefore possible to modify or change the phase correction applied within wide limits.

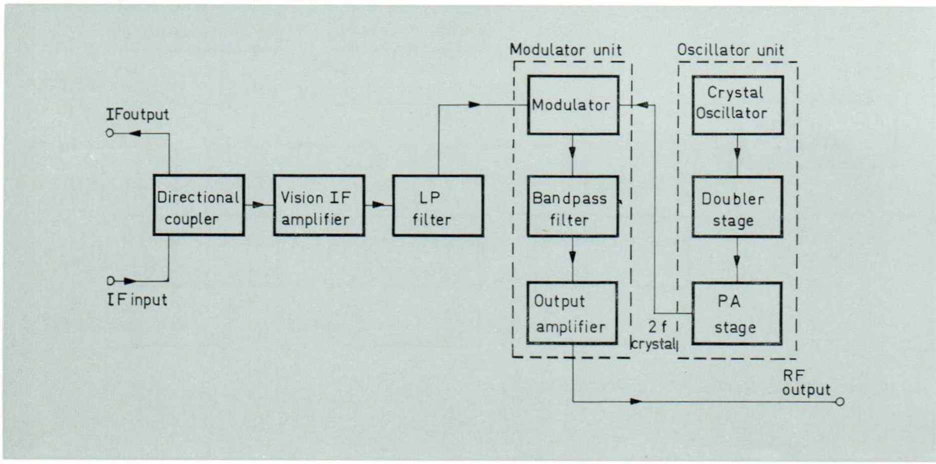
In the sound channel the 5.5 MHz (or 6 or 6.5 MHz depending on system) the sound oscillator is frequency modulated by means of a switching diode. This gives a very stable and linear frequency modulation. In order to reach a high frequency stability of the sound - vision frequency difference, this sound oscillator is controlled in frequency by means of an AFC loop referring to a crystal oscillator. The difference frequency between the two oscillators is measured in a 70 kHz counter type discriminator.

The output sound carrier of 33.4 MHz (33.5 MHz or 31.5 MHz) is obtained by mixing the vision carrier from the crystal oscillator with the FM modulated 5.5 MHz (or 6 or 6.5 MHz) in a balanced modulator. The lower sideband is filtered out and added to the video modulated vision carrier in the output amplifier. The sound carrier level is adjustable within wide limits from the front panel of the apparatus.



7. Block diagram of IF modulator

8. Block diagram of converter units PM 5591



The RF converters PM 5591-93-94-95

The RF converters are small, fully transistorized plug-in units designed for mounting in the 19″ cabinet with the power supply PM 5596 (see fig. 8). The type numbers of the plug-in units are PM 5591, PM 5593, PM 5594 and PM 5595 for the television bands 1, 3, 4 and 5 respectively.

The input circuitry of each converter is a precision directional coupler made in stripline technique. The insertion loss is negligible ($<$ 0.02 dB) and the number of converters which can be fed by one modulator is therefore practically unlimited.

The converters are combined in groups placed in one or more base units PM 5596 and placed in or close to those places where the TV signals are needed. From the same IF signal any combination of VHF-UHF channels can be obtained by choosing the appropriate converters. The gain control in each converter can if wanted be set to compensate for up to 10 dB of the cable losses. This corresponds to a length of about 300 meters of the RG 213U cable. Up to this distance from the modulator the full RF output voltage of 200 mV across 50 Ω can be obtained (≈ 800 mV in 300 Ω). The last converter in the chain has to be terminated by 50 Ω. When more VHF and UHF signals are needed additional converters are simply added at the appropriate place in the chain. The system is therefore extremely flexible; it can easily be changed or extended according to the momentary need for signals.

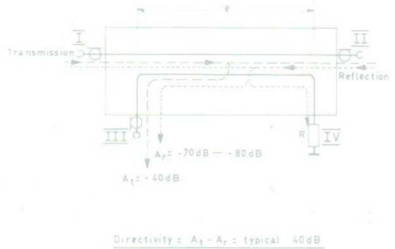
The most important property of this distribution system is its immunity against cable reflections. The requirements to the transmission of measuring signals for colour TV are so strong that only a few percentage of reflection can be tolerated. This corresponds to standing wave ratios being very close to 1 and this will be very difficult to realize in a practical set-up. The solution to these problems is found in the use of a directional coupler in the input circuit of each converter. The directional coupler being used is made in a high precision stripline technique and it has a typical directivity of 40 dB.

The directional coupler consists of two short pieces of transmission lines placed close together in a homogeneous dielectric.

Between these two lines there will be as well an inductive as a capacitive coupling, and a certain amount of the signal on the one line is coupled into the other line.

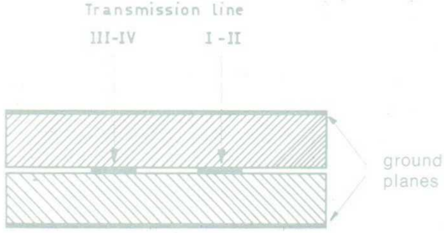
Figure 9 shows in schematic form the circuitry. A signal transmitted on the one line from I to II will because of the combined magnetic and electrical coupling be passed on to the terminal III with a certain attenuation "A_t". In case the signal instead of passing from I to II passes in the direction II to I as indicated as "reflection" in the figure the capacitive coupling will be of the same value as before; the inductive coupling, however, will be reversed in polarity and the two components of the coupling will compensate each other at terminal III. The signal coupled into the second line is in this case transferred to terminal IV, where it is absorbed in the resistor R.

If the design is carefully made and the mechanical and electrical tolerances are kept within narrow limits the compensation can be very accurate. The directional coupler in PM 5591 - 93 - 94 - 95 gives a relative attenuation, called directivity, of typically 40 dB.



9. Directional coupler

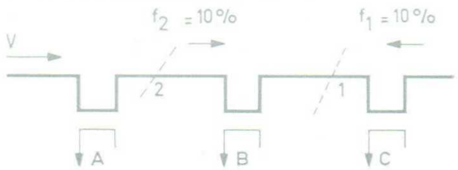
10. Cross section of directional coupler



In order to obtain the necessary accuracy and stability the circuitry is built in stripline technique on a special high quality dielectric. Figure 10 shows a cross section through the coupler. The two transmission lines are made in etching technique on a double sided printed wiring plate. The opposite sides of the two printed wiring boards form two ground planes which perform a screening of the transmission lines. Figure 11 illustraties the effect of this directional coupler. A, B and C are the directional couplers in 3 converters.

The IF signal V is coming from the left and passed each converter in the forward direction. About —40 dB i.e. 10 mV is picked up by the converter, amplified and converted to VHF or UHF. At the point 1 we assume that a certain mismatch gives rise to a strong reflection of 10 %. This reflected signal passes the directional couplers A and B in the reverse direction, and it is therefore passed on to the converter circuits with a further attenuation of typically 40 dB. This means that the reflected signal gives rise to a disturbance not of 10 % but of 10 %: 100 = 0.1 % only. In case another discontinuity exists at point 2, also with a reflection factor of 10 %, this second discontinuity will give rise to a second reflection of the primary reflection coming from point 1. The wave, which is reflected twice now, passes again in the direction from left to right, which is the forward direction of the directional couplers. The disturbance caused, however, is attenuated by two reflections and is therefore as small as 0.1 times 0.1 = 0.01 or 1 %.

11. The effect of directional coupler



We see that even by assuming two large discontinuities and their following reflections we get a converted RF signal which is free from noticeable distortion. Discontinuities as large as assumed here will not occur in a well designed cable network. Therefore the installation has a large safety margin against reflections. The terminating resistor supplied with the apparatus has a reflection coefficient of max. 0.5 % and each converter has a reflection of max. 5 %.

Before the IF signal, which is picked up by the directional coupler, is applied to the mixer itself it is amplified in a wideband IF amplifier. The gain of this amplifier is adjustable to compensate for an eventual decreased IF level on the input cable. The mixer is a balanced diode mixer which is connected as a switching type of ring modulator. This way of operation makes the extremely low distortion possible.

After the mixer stage a passband filter and an output stage deliver the RF signal to the output socket. The oscillator voltage for the mixer is obtained from a crystal-controlled oscillator and a number of frequency multiplier stages.

A frequency change is easily carried out by inserting a new crystal and adjusting two or three multigang condensers for maximum output voltage. In order to facilitate this procedure a small RF meter circuit is built into the cabinet together with the power supply.

Pulse shapers for TV pulses

by E. Helmer Nielsen

In larger TV installations the various apparatus, pattern generators, cameras, colour encoders etc. are all controlled by one common TV pulse generator. This pulse generator delivers the synchronization signal and the blanking signal necessary for making a composite TV signal. Also other auxiliary signals may be distributed, e.g. separate line and frame drive pulses, burst keying pulses etc.

Pulse distortion

The TV pulses are distributed on 75 Ω cables with a nominal pulse amplitude of 4 V_{pp}. The pulses delivered by the pulse generator are quite accurately specified with regard to pulse amplitude and pulse shape but they may often be considerably distorted by the cable distribution network due to reflections, wrong terminations, cable losses etc. In fig. 1 some examples are given showing typical kinds of distortion.

Another disturbance which may often occur is superimposed hum voltage. Fig. 2 shows schematically how this may appear. These disturbances on the TV pulses, must all be eliminated if correct operation of the various apparatus in the installation is to be assured. If we look at fig. 1 we can see that the correct information in all cases shown can be found at the middle of the pulse. The hatched area between the dotted lines at the middle of the pulse gives a good representation of the original pulse information. The circuitry for reshaping the pulses therefore has to be able to select a thin slice at the middle of the pulse and use this for recovering the pulse information. The pulse shaper circuitry has to be designed to separate the middle of the pulse independent of the pulse amplitude and possible reflections. Variations in amplitude of ± 6 dB (2 V_{pp} to 8 V_{pp}) have to be reckoned with. In case the pulse signal also suffers from hum the pulse shaping becomes more difficult. For optimum functioning of the pulse shaper at the middle of the pulses it is necessary to remove the superimposed hum. This presents difficulties because most TV pulses contain information at the same frequency as the hum, i.e. 50 or 60 Hz.

Pulse shaper for line pulses

Line pulses, having a repetition frequency of about 16 kHz, are relatively easy to reshape because they do not contain any low frequency information. Therefore it is simple to separate possible superimposed hum from the pulses.

The basic circuitry, for the pulse shaper is shown in fig. 3. The input pulses are negative-going as shown in fig. 4. The pulses switch the transistor on – off from saturation to cut-off condition. The transition period between these two states has to be placed at the middle of the pulse. By correct design of the circuitry at the base of the transistor this condition can be fulfilled independent of pulse amplitude. Let the pulse amplitude be V volt. Only half of the amplitude has to drive the transistor into saturation, therefore the capacitor C_1 must have an opposing voltage of V/2 volt.

During the negative pulse the transistor is conducting and the capacitor C_1 is discharged by the base current i_b:

$$i_b = \frac{V - V/2}{R_1} = \frac{1}{2}\,\frac{V}{R_2}$$

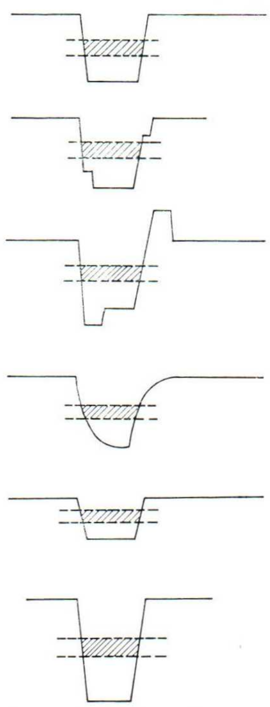
In between the negative pulses the transistor is cut-off and C_1 is charged via $R_1 + R_2$. The charging current i_c is:

$$i_c = \frac{V/2}{R_1 + R_2} = \frac{1}{2}\,\frac{V}{R_1 + R_2}$$

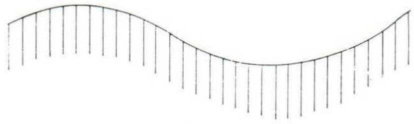
This condition will be static when there is balance between charging and discharging of C_1. This means that $i_b t = i_c (T-t)$

$$\frac{i_b}{i_c} = \frac{R_1 + R_2}{R_1} = \frac{T-t}{t} = \frac{64-7}{7} = \frac{57}{7}$$

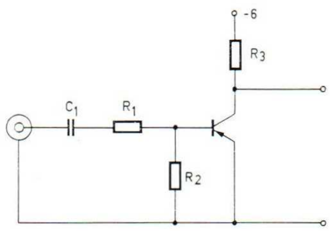
The choice of the value of R_1 is determined partly by the base current necessary and partly by the input impedance. The input impedance has to be sufficiently high for allowing looping-through of the signal. For a maximum reflection of –36 dB the minimum input impedance is 2.3 kΩ. In the bottomed condition the transistor will draw a current of 6 mA using a value of R_3 = 1 kΩ. With a minimum current



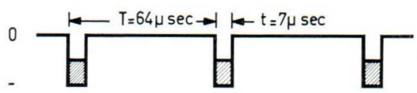
1. Examples of TV pulses distorted by the cable distribution network due to reflections, wrong terminations, cable losses etc.



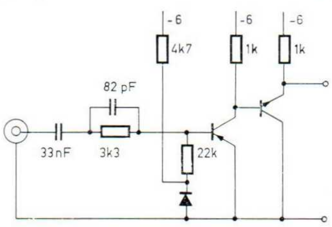
2. This is another disturbance; superimposed hum voltage



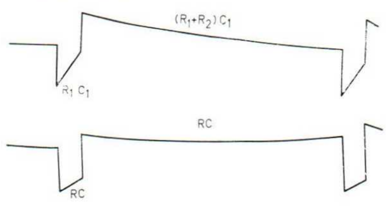
3. Basic circuit arrangement of a pulse shaper for line pulses



4. Input pulses to pulse shaper are negative-going such as illustrated in this figure



5. Complete diagram of the pulse shaper is shown here



6a. Same circuit given in fig. 3 is not advisable to be used for frame pulse because of the difficulty to separate this frame pulse from superimposed hum

6b. If the two time constants were made identical the situation would be more favourable gain of 20 the base current is 0.3 mA, which for a minimum input voltage of 2 V_{pp} corresponds to a value of $R_1 = 3.3\ k\Omega$.

The resistor $R_2 = R_1 \frac{57-7}{7} = 22\ k\Omega$

The capacitor C_1 should be large enough to keep the voltage constant during one line period. A value of 33000 pF will give a time constant of 33000 pF x 22 $k\Omega$ = 700 μsec. This value is sufficiently low for suppressing the 50 Hz hum.

The complete diagram is shown in fig. 5. A speed-up capacitor and a compensating diode have been added. If the output impedance is too high an emitter follower can also be added as shown.

Pulse shaper for frame pulses

The frame pulse is a short pulse having a repetition frequency of 50 Hz (60 Hz). It is difficult to separate this frame pulse from superimposed hum. The circuitry of fig. 3 is not able to do this; Fig. 6a illustrates why. The charging time constant $(R_1 + R_2)\ C_1$ has to be small enough to remove the hum, but then the discharging constant R_1C_1 becomes too small for driving the transistor into saturation during the whole frame pulse.

If the two time constants were made identical the situation would be more favourable. See fig. 6b. It is then possible to suppress the hum considerably without distorting the frame pulse too much. This can be realized if we place an extra C filter in front of our pulse shaper. The RC filter and the pulse shaper have to be separated by means of an emitter follower. The circuitry then looks like fig. 7. The base circuitry of the pulse shaper is designed in the same way as the one for the line pulses. The RC filter at the input is designed for best compromise between hum suppression and pulse distortion.

Pulse shaper for composite sync signal

The composite sync signal is a mixture of line pulses having a repetition frequency of 15625 Hz and frame sync information with 50 Hz repetition frequency.

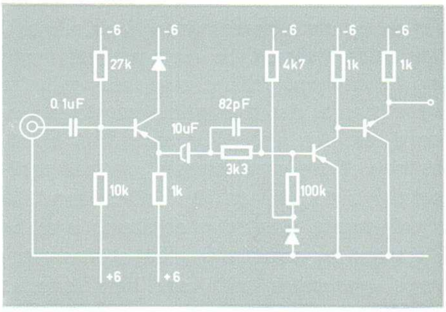
The re-shaping of this signal in case of heavy distortion or disturbances asks for a more elaborate circuitry. Again superimposed hum presents the biggest difficulties. In many TV circuits a so-called keyed clamp circuit is often used in order to restore the DC and LF components of the signal and at the same time remove hum. It is also logical to try to use a clamp circuit in this case for separating the sync signal from hum voltages.

A clamp circuit can schematically be represented by a switch which is closed for instance by the sync pulses. Fig. 8 shows in principle how the function may be. We recognize our basic circuitry from fig. 3. The polarity of the transistor has only been changed from pnp to npn. C_1 is discharged, via R_1, by base current which flows during the period between pulses. During the sync pulses the transistor is cut-off and the switch S closed. With the switch closed C_1 is charged to the voltage V/2 which is assumed to be availabe from a battery B. As the switch is closed during the sync pulses we have, in this way, stabilized the negative peak of all the pulses at V/2 volt.

This means that possible hum is removed and the transistor is switched on-off at the middle of the pulse as desired.

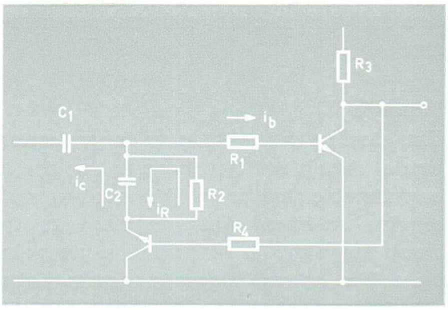
How can we now realize a battery with voltage V/2 and the switch S?

In fig. 9 a simple solution is shown. The battery is replaced by an RC circuit and the switch is a transistor being switched on-off by the output signal of the pulse shaper. The RC circuit R_2C_2 has a large time constant, which means that the voltage is held constant similar to a battery. The voltage adjusts itself automatically until balance between the charging and



7. To realize the waveform shown in fig. 6b an extra RC filter is placed in front of already introduced pulse shaper

9. Simple solution of fig. 8



discharging of C_1 is obtained. The discharging base current i_b flows for each TV line in (64–5) μsec = 59 μsec through R_1. The corresponding discharge of C_1 is i_b x 59 nC. During the sync pulses the transistor switch is closed and the same charge is transferred from C_2 to C_1. The charge on C_2 is recovered through R_2 by means of a current i_R flowing during 64 μsec. Therefore i_b x 59 = i_R x 64 nC. If we further fix the voltage across C_2 at V/2 volt we have

$$i_b = \frac{1}{2}\ \frac{V}{R_1} \quad \text{and} \quad i_R = \frac{1}{2}\ \frac{V}{R_2}$$

The values of R_1 and R_2 can now be determined. We see that $R_2 \approx R_1$.

The current i_c which also has to flow through the transistor will be

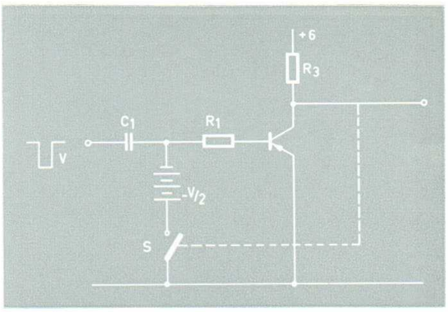
$$i_c = i_b \frac{59}{5} = 12\ i_b$$

This collector current fixes the maximum value of R_4.

i_b maximum = 1.2 mA; i_c maximum = 14 mA

$$R_4 = \frac{6}{14} \beta \approx 10\ k\Omega\ (\beta \geqslant 25)$$

The output pulse is positive and the clamp transistor therefore has to be an npn transistor. At first sight from fig. 9 the transistor seems to be reverse connected. This is not the case, the current I_c does flow in the correct direction from the collector to the emitter. When the negative input pulse has finished, however, the positive step at capacitor C_1 tries to reverse the current in the clamp transistor. Because of the very low reverse current gain in the transistor used (BSY 39) the impedance in the emitter is very high (several $k\Omega$) for the reverse current, and the positive step is, without noticable attenuation, passed on to the pulse shaper, which is then being saturated. Therefore it is very essential for the function of the circuitry to use a clamp transistor with a low reverse current gain.



8. Schematically representation of a clamp circuit

10. Final configuration of the pulse shaper for composite sync signal

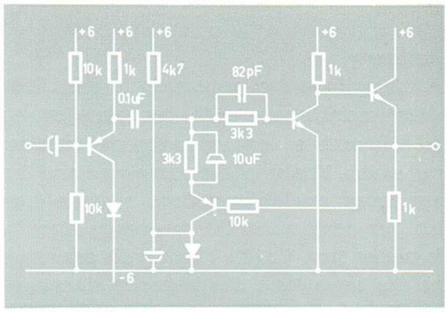


Fig. 10 shows the final diagram. The circuitry is completed with an emitter follower at the output and a compensating diode in series with the clamp transistor. The time constant C_1R_1 is 330 μsec; this is sufficiently long compared to the duration of one TV line, and the value of C_1 is still small enough to allow the clamp circuitry to remove 50 Hz hum completely. A pnp emitter follower at the input is necessary for supplying the relatively large current pulses to C_1 during the charging periods.

Pulse shaper for composite blanking

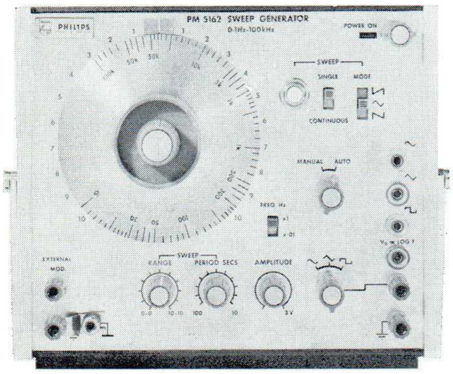
The pulse shaper circuitry for the sync signal can be used, in principle also for the blanking signal. Smaller modifications may be introduced due to the somewhat higher duty cycle of the blanking signal.

Application

The circuits described were developed for use in various TV measuring equipments of the Philips EMA range.

An accurate triangular- wave generator with large frequency sweep *

by G. Klein and H. Hagenbeuk



Summary
A triangular-wave generator is described which combines a wide frequency sweep ($\approx 10^4$) with a very accurate symmetrical waveform. The amplitude response and the symmetry exhibit variations of less than 0.1 per cent, while the frequency stability is better than $1:10^4$. The circuit can serve as the basis for a single-sweep sine-wave oscillator or an FM modulator and voltage (current)-frequency converter.

Introduction
Many applications require a generator of symmetrical triangular voltages, the frequency of which can be varied over a wide range by an external voltage or current. The stringency of the requirements made on the accuracy of the amplitude and the symmetry of the triangular waveform depend on the use. For most FM applications, these requirements are not particularly rigid. If, on the other hand, a single-sweep sine wave oscillator is to be designed by combining such a triangular-wave generator with an instantaneous triangle-sine converter, the distortion in the sine-wave voltage will be determined by the symmetry and amplitude stability of the triangle. In order to be most universally applicable a generator was therefore designed to a very stringent specification.

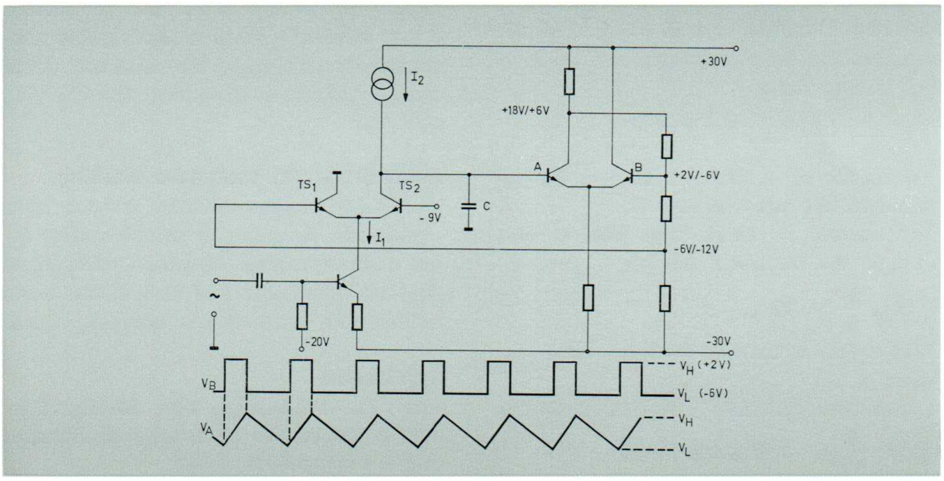
Principle
The starting point for such a generator was a circuit that has been in use for a considerable time as an FM-modulator and already satisfies rigid requirements in many respects. Its principle is given in fig. 1. The Schmitt-trigger is used to make the difference between the base-voltages of TS_1 and TS_2 alternately positive and negative by a few volts so that TS_2 carries either a current I_1 or no current. By this means the capacitor C is alternately charged with a current I_2 and discharged with a current $I_1 - I_2$, provided that $I_1 > I_2$.

The amplitude constancy of the triangular waveform is determined by the difference between the changeover levels V_H and V_L of the Schmitt-trigger (about +2 V and − 6 V in the example) which is, in turn, primarily determined by the resistance values and the supply voltages in the Schmitt-trigger, thus allowing considerable accuracy to be attained. The frequency can be varied over a wide range by changing I_1 but the triangular waveform is not symmetrical because I_2 is maintained constant. For the system to work properly, I_2 would have to vary simultaneously and satisfy the requirement $I_1 = 2\, I_2$, a condition which would be hard to be maintained for large variations of I_1. Fig. 2 gives a possible improvement which has been in use for some years. Transistor of the balanced pair carries the current I_1 in one position of the Schmitt-trigger. Since TS_1 then carries no current, the base voltage of TS_3 will be equal to the positive supply voltage, so that I_2 is then zero. Therefore, in this position, C is discharged by the current I_1. In the other position of the Schmitt-trigger TS_1 carries the current I_1 and the base voltage of TS_3 will be about I_1R volts negative with respect to the positive supply voltage. If the emitter resistance of TS_3 is correctly chosen, I_2 can be made equal to I_1. The symmetry of the triangular wave which can be made perfect at a given value of I_1, is difficult to maintain if I_1 is varied over more than one decade because the base-emitter voltage of TS_3 will not remain equal.

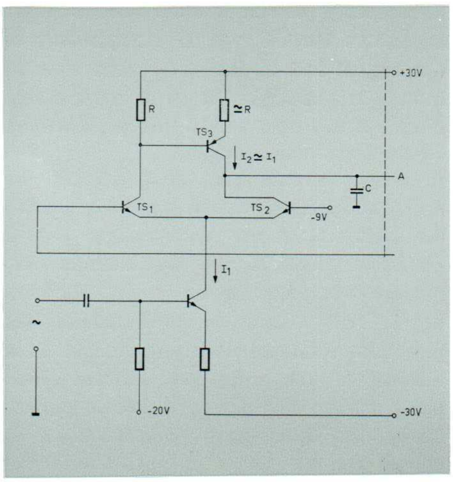
In the proposal put forward here, care is taken to keep the current I_2 accurately equal to I_1, even when the latter varies greatly. Fig. 3 shows how this can be done by making use of a second capacitor C_2. Here, switching transistors TS_1 and TS_2 are symbolically represented by switches S_1 and S_2. These switches, together with S_3 and S_4, are controlled by the Schmitt-trigger as before in such a way that S_1 and S_4 are closed when S_2 and S_3 are open, and vice versa.

The basis principle of the circuit is the use of the voltage on capacitor C_2 to control the current I_2 in such a way that the average voltage C_2 is kept constant.

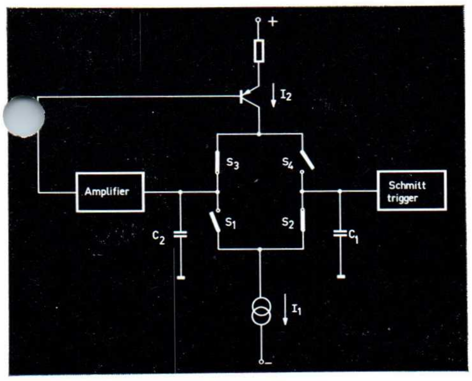
1. Basic circuit for triangular-wave generator



2. Possible regulation of symmetry



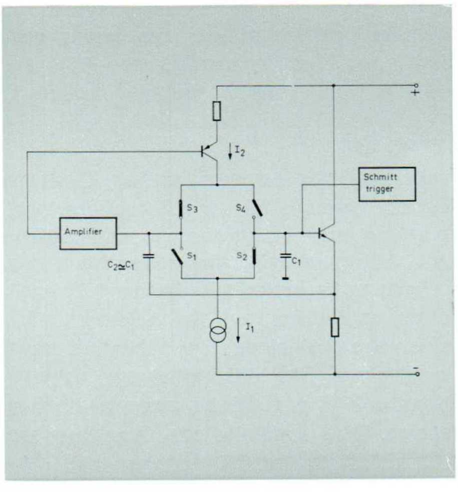
If T_c and T_d are the times during which S_1 and S_4 (S_2, S_3) are closed (open) and open (closed) respectively, the following equalities should apply:
for C_1: $I_1T_d = I_2T_c$ and
for C_2: $I_1T_c = I_2T_d$
Hence:
$I_1 = I_2$ and $T_c = T_d$ as required.
The extent to which these equalities will apply in practice depends on the accuracy with which the currents fed to C_1 and C_2 are equal and this determined by the leakage currents of the switching transistors and the base currents of the input stages of the Schmitt-trigger and the control amplifier. The inequatily between the currents can be kept down to about 10^{-9}A in a transistorized circuit. This means that, if the currents I_1 and I_2 are no smaller than a few microamperes, the currents satisfy the above equations to within about 0.1 per cent. On the other hand, the currents should not be made greater than about 10 mA. This means that where suitable current sources are used, a frequency sweep of 10^4 can be obtained. Furthermore, less than 0.1 per cent asymmetry can be guaranteed. With a reduced frequency sweep, higher minimum values of I_1 and I_2 may be used and the symmetry will be even better.



3. Proposed regulation of symmetry

The principle as given in fig. 3 has only one disadvantage: Since I_2 is controlled by the ripple voltage on C_2 as it charges C_1, the rising ramp of the triangular voltage on C_1 will exhibit a slight deviation from linearity. This effect is reduced by increasing C_2 and/or decreasing the loop gain. This means, however, that the speed with which I_2 follows a change in I_1 is reduced. A simple calculation shows that for this the following relation applies:
$\tau = T/8\delta$ where T = period of the triangular voltage τ = time-constant of the control system of I_2, δ = maximum relative deviation from linearity in the rising flank of the triangular voltage.
Thus, to limit δ to 0.1 per cent, the control would require more than 100 periods.
In the case of an external variation of I_1, the other current may be varied simultaneously by approximately the same amount. The control system would then only have to correct a possible deviation, and this could normally be allowed to take some time.

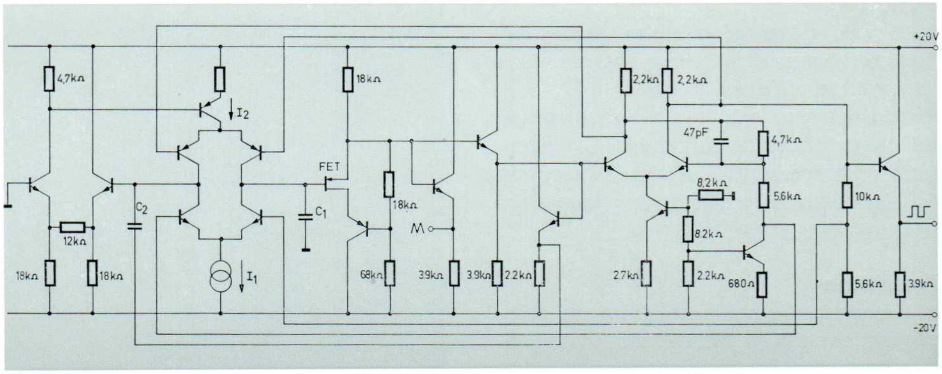
4. Improved regulating circuit



A better solution is given in fig. 4. The ripple in the control voltage can be made zero by making C_2 equal to C_1 and by connecting one side of C_2 to a point that follows the voltage on C_1. In theory, the non-linearity and the control speed are now no longer related. If the adjustment is perfect, the ripple is zero and no distortion occurs. In practice, however, this perfect adjustment cannot be relied upon as deviations in the values of the capacitors will occur. The ripple voltage can, however, easily be compensated to within 1 per cent. Hence, the denominator in equation ($\tau = T/8\delta$) is multiplied by 100 and this is adequate for all practical purposes.

Practical circuit

Fig. 5 gives a simplified diagram of a triangular wave-generator incorporating the principles given above. Where necessary, the transistors were arranged as Darlington pairs. Since only currents are switched, the transistors may be replaced, with advantage, by field-effect transistors. Various current sources may be used for I_1, thus providing, for instance, linear or exponential relationships between the frequency and a voltage which may be controlled externally. I_1 may also be supplied by two or more current sources connected in parallel. Where the switching circuit is designed with Darlington pairs the asymmetry proved to be much smaller than 0.1 % for frequencies below some tens of kilohertz. At higher frequencies an inaccuracy will be introduced by the finite switching time and by parasitic effects of the switching circuit. With the available HF transistors very good symmetry is still possible for frequencies up to some hundreds of kilohertz.



5. Practical circuit

The frequency stability is determined by the constancy of the current source I_1 and the switching level of the Schmitt-trigger. At normal ambient temperatures a frequency stability better than 10^{-4} was easily obtained for linearly variable current sources. Using sources with exponentially varying currents, the stability is generally less by one order of magnitude. The influence of temperature changes can be kept below 0.01 per cent/°C without difficulty. The amplitude is also determined by the constancy of the switching levels of the Schmitt-trigger.
The amplitude constancy with changing frequency is perfect as long as the transit times of the switching circuit are negligible. Due to this effect some increase in amplitude will occur at higher frequencies. It is obvious that the application of high-frequency field effect devices for the switches will give still better results.
The instrument shown in the photograph is the first commercially available generator in which the described circuits have been applied. The main characteristics of this function generator – the Philips type PM 5162 – are:
Frequency range: 0.1 Hz to 100 kHz, divided in two;
Ranges: 0.1 to 10^3 and 10 to 10^5 Hz;
Waveforms: sine, triangular, square;
Output voltage: 3V peak-to-peak into 600Ω.
Sweep mode: internal and external with a maximum sweep ratio of 1 : 10000.

* By kind permission of the editor Mr L. G. Poole this is a reprint of an article recently published in Electronic Communication.

Accurate instantaneous triangle-sine converter

by G. Klein

Introduction

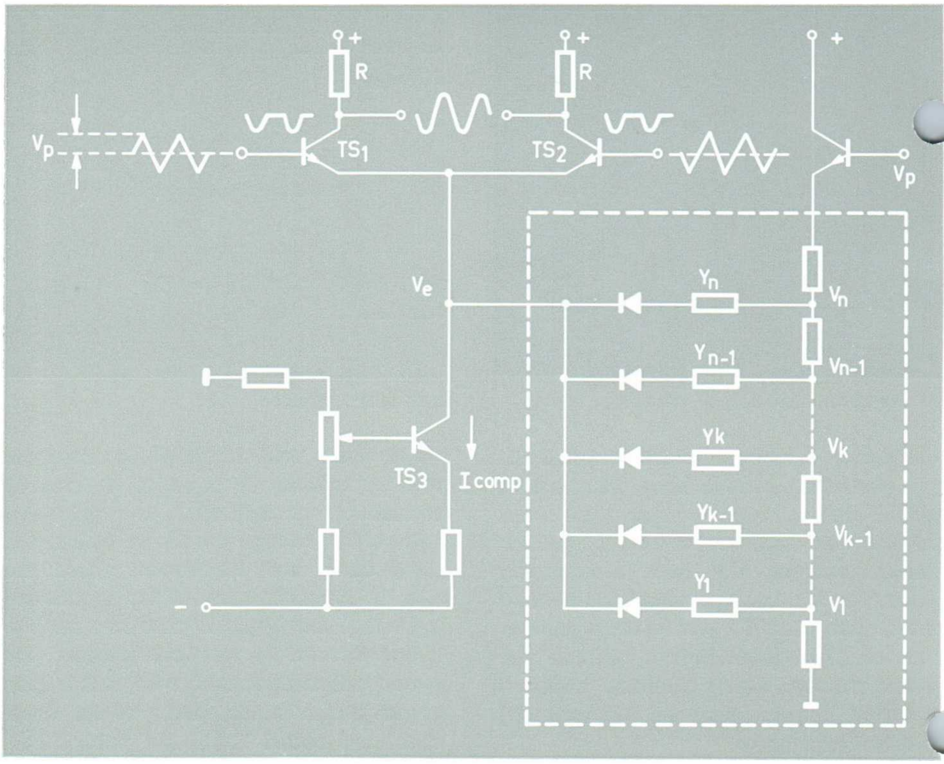
Until now, a few informations on the low frequency measuring instruments are given through the different media.

Here again you will find a reprint of an article published in Electronic Communication, November 1967. A circuit described in this article is applied in our generators PM 5162 and PM 5168.

Describtion

In instrumental electronics, there are many applications for an instantaneous triangle-sine converter. The main reason being that the generation of triangles is much more flexible than that of sine waves. It is, for example, feasible to make a triangular-wave generator with a frequency sweep of more than four decades while maintaining a perfect symmetry and amplitude constancy. If such a circuit is combined with an accurate triangle-sine converter, a sinusoidal signal of good amplitude stability and low distortion is obtained; the frequency of which can be controlled externally over the same wide range of more than four decades.

The known methods for the instantaneous conversion of triangles into sine wave all have some disadvantages. The method employing the inverse-function generator is very accurate but only applicable up to some tens of kiloherz, as it uses a sampling technique. Of the methods which utilize the non-linear characteristics of passive or active elements the one employing a field-effect transistor seems to give the best results: a distortion as low as approximately 1 % can be obtained. For this, however, some critical adjustments are necessary and furthermore the temperature-sensitivity is considerable. The third and most common method makes use of resistor-diode networks with which a well defined non-linear resistance is made. In most published circuits employing this method, separate networks are required for the positive and negative halves of the cycle. This implies that if one wants to keep the distortion low, the tolerances in the corresponding resistor values of the two networks should be made very small. Because of the stringent requirements thus imposed on the components the approximation of the sine wave is then usually done in no more than six steps and this means that the sine wave looks more like a broken than a smooth curve. One of the great disadvantages of this is that such a signal cannot be differentiated properly.



Basic circuit of triangle - sine converter

The circuit described here also uses a resistor-diode network. One feature is that the same network is used for the positive and the negative half cycle so that, even if a distortion of the order of one tenth of a percent is required, large tolerances (5 - 10 %) in the resistor values are allowed as long as a sufficient number of steps are taken in the approximation. This makes the circuit particularly suitable for integration. Other advantages are that no critical adjustments have to be made and that the sine wave has not only a low distortion but is also very smooth.

The principle of the circuit is given in figure. If the height V_p of the two equal, but phase-opposite triangles is made some volts, only one of the transistors TS_1 and TS_2 will be conducting during each half period and will then act as an emitter follower. Supposing for the moment that the common emitter follows the bases perfectly and that the conductance Y in the emitter lead varies proportional to cos x, where x is equal to $\frac{\pi}{2} \cdot \frac{V_e}{V_p}$ and V_e is the voltage of the common emitter, then the current through TS_1 or TS_2 will be proportional to sin x. On the collectors of TS_1 and TS_2 there will thus appear half-sine waves, half a period out of phase with each other, so that between the collectors a full sine wave is obtained. The differential wave is obtained. The differential conductance $Y(x) = Y_0 \cos x$ can be made in the way shown. The resistor-diode branches are connected to the DC voltages $V_1 \ldots V_n$, which are all derived from one DC voltage V_p corresponding to the height of the triangles. The internal resistance of these points is low compared with the value of the branch resistors. At zero-signal all diodes are conducting but as V_e gets more positive the diodes stop conducting one after the other and when V_e is equal to V_p the differential conductance through the resistor-diode network has become zero. By selecting the proper values for the conductances $Y_1 \ldots Y_n$ in the branches the cosine function can thus be obtained. The current source TS_3 is used for compensating the current through the network at zero signal.

One can easily calculate what accuracies are required for the resistors and diodes in the branches and for the voltages $V_1 \ldots V_n$. From these calculations it follows that in order to keep the deviation from a pure sine wave at every point smaller than a few tenths of a percent the approximation should be made in about 20 steps: 15 x 5°, 3 x 3°, 2 x 2° and 2 x 1°. In the resistor values tolerances of 10% are then allowed while the deviations in the voltages $V_1 \ldots V_n$ should be less than 0.02 V_p. This means that even without taking special measures the basic circuit should give good results and that the hatched part of figure lends itself excellently to integration: only a moderate accuracy is required as long as enough components are used.

The integrated network consists of 43 thin-film nickel-chromium resistors and 21 monolithic diodes. The optimum values of the resistors were determined experimentally, taking into account the loading of the voltage divider and the switching properties of the diodes. In order to make Y_0 equal to $(1\ k\Omega)^{-1}$ the resistors in the branches were given values between 12 and 100 kΩ. For the resistors in the voltage divider a value of 11 Ω per degree was taken, thus making the total emitter resistance 1 kΩ. The actual size was 2 x 2 cm.

A practical circuit, basically the same as that of figure, gave a distortion which varied from 0.2% for low frequencies to 0.8% for some hundreds of kiloherz. Furthermore the distortion can be made completely insensitive to changes in the height of the triangles. The influence of changes in the ambient temperature was very small indeed: a change from 20°C to 50°C increased the distortion by no more than 0.1%. Apart from the frequency-limitation there is indeed only one shortcoming in the simple circuit: the take-over of the current by TS_1 and TS_2 at zero signal.

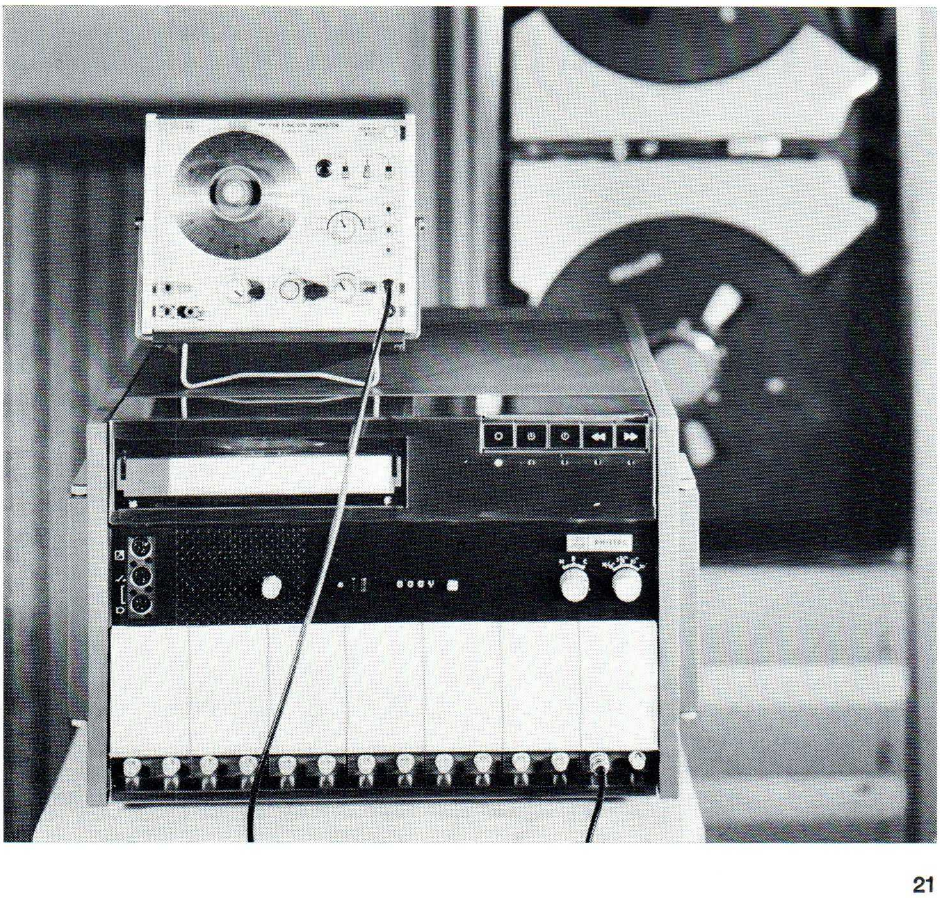
Depending on whether I_{comp} is too large or to small a jump or a flat part occurs. The best result is obtained by making I_{comp} so much larger than the current through the network at zero signal that the slope of the take-over is equal to Y_0. However, as the crossover in a difference amplifier is not completely linear, there still remains a small deviation. A possible improviment is obtained by applying feedback in such a way that the common emitter is forced to follow the positive-going triangle more accurately. In this way a completely smooth curve with a distortion well below 0.1 % was obtained.

The integration of the circuit brings the total parasitic capacitance C_{par} of the common emitter down to some picofarads. Above the frequency for which the admittance of this capacitance becomes, say, 1% of the total conductance Y_0, the distortion will increase rapidly. Thus the maximum frequency for which the circuit can be used without any special measures is given by

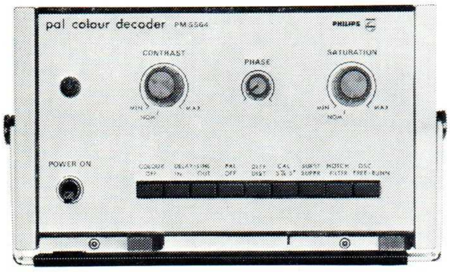
$$\omega_{max}\ C_{par} \approx 10^{-2}\ Y_0$$

With the chosen values this given an upper frequency of approximately 1 MHz.

A detailed description of the circuit will appear in the British journal "Electronic Engineering".



PM 5564 PAL colour decoder



The actual programme of professional television measuring equipment consists of instruments supplying video signals for monochrome- and colour television as well as instruments for modulation of the video signals on the various RF channel frequencies.

Several times the need was felt to extend our programme with a device to check or monitor the generated PAL coded video signals either visually on a colour monitor or via vector- or line display on an oscilloscope.

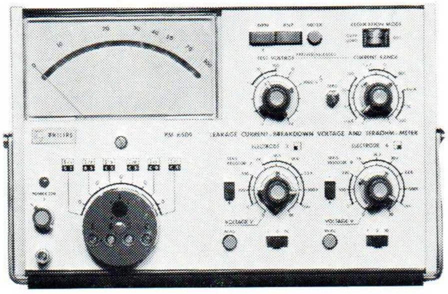
The instrument meeting these requirements is the new PAL colour decoder PM 5564, which has the following features:

- Decoding with or without delay line
- Output signals:
  2 set of RGB outputs
  1 set of colour difference outputs R-Y, B-Y and G-Y and the luminance output Y
  1 set of special outputs for measuring differential phase and gain
- Calibration facility (5° and 5%) for determining differential distortion

Application field

The PAL decoder is a valuable instrument for a number of applications as found in:
CTV research labs – checking properties of PAL system
CTV development labs – design of colour TV circuitry
CTV production – checking of distributed signals
Measuring vans – checking transmission characteristics
CTV studios and O.B. vans – monitoring of pictures

PM 6509 fully protects semiconductor under test



A self-contained semiconductor test instrument which fully protects devices while measuring two of the most common parameters, leakage current and breakdown voltage, has been introduced by Philips of Eindhoven. Called the PM 6509, it employs a constant current/constant voltage mode of operation which ensures complete protection for the semiconductor under test. In addition, it can carry out measurements under "in circuit" conditions.

The PM 6509 can be used with two, three or four electrode devices, including PNP and NPN transistors. It can measure leakage currents of from 3 pA to 100 mA at test voltages between 100 mV and 1000 V, and breakdown voltages up to 1000 V. Measurements can also be made of resistance in ranges up to 100 TΩ and, when a special adapter is used, of semiconductor amplification factors at low currents.

For both r.&d. and production applications

The instrument can be used equally well in r. & d. laboratories and in test and production departments. In the former it finds ready application in optimizing worst-case designs and selecting devices for particular applications and functions. In the latter it can be employed to check semiconductor reliability and determine tolerance spreads. For batch testing, the test voltage can be switched on and off through an external contact.

With three or four electrode devices, leakage ccurrent and breakdown voltage can be measured with the third and/or fourth terminal open circuit, with one or both these terminals connected to the reference electrode via a resistor, or with one or both connected to a voltage. Two additional voltage supplies are provided for this last purpose, and their outputs can be adjusted between –10 and + 10V, with or without a series resistor in the line as required.

Semiconductor fully protected under all test conditions

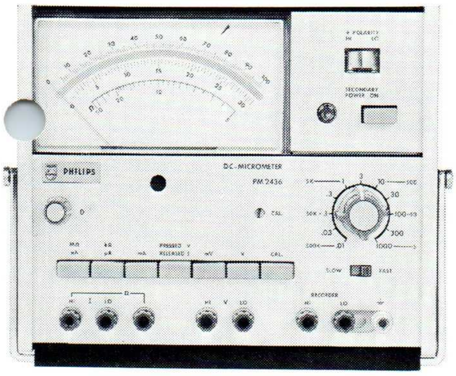
In measuring leakage current and breakdown voltage, the applied voltage and current are limited so that the semiconductor is always fully protected. For leakage current measurements, for example, the maximum current can be limited to a value between 10 and 110% of the full scale reading (lowest value 10 pA), on any of the instrument's ten ranges. A selector switch permits the test voltage to be applied between any two electrodes, and therefore permits I_{cs}, I_{ce}, I_{cb}, I_{eb}, I_{es} or I_{bs} (where s = screen) to be measured. For each switch position, an indication is provided of how the third and fourth electrodes should be connected.

For breakdown voltage measurements, the PM 6509's design allows the maximum current setting to be independent of the test voltage. This permits the current to be set to limit at a value slightly higher than the leakage current before breakdown and the voltage can then be adjusted until the breakdown value, is reached. Under these conditions the semiconductor is fully protected at all times.

Front panel design makes operation simple

The PM 6509's front panel has been designed to make operation of the instrument as simple as possible. As well as the controls for setting the leakage current and breakdown voltage ranges, there is a 'regulation mode' indicator which shows whether the instrument is operating in the constant current or constant voltage mode. Press-button switches are provided so that the voltages applied to the third and fourth electrodes can be checked at all times, and recorder output is available to drive such an instrument if a permanent record of measurements is needed.

DC microvolt & picoampmeter



A new DC microvolt and picoampmeter with high sensitivity and input facilities which make it suitable for practically all DC measurements.

Philips introduce an all solid state DC microvolt and picoampmeter with automatic polarity indication and separate floating inputs for voltages and currents which enable measuring leads to be connected to both inputs at the same time.
Suppression of hum signals is > 100 dB, accuracy better than 1.5 %. Designated the PM 2436, this new instrument lends itself for accurately measuring quantities which can be converted into DC values, in addition to direct DC measurements.
It may be used as a high-gain preamplifier for low DC signal levels (maximum gain 500.000 in 10 μV position) by linking up the recorder output. Other measuring facilities are made available by the following accessories: high tension probe GM 6071; VHF probe PM 9200; and T-piece for VHF probe PM 9250.
An interesting feature of the instrument is the inverted guard system for the power supply unit. This completely isolates the mains earth from the instrument eliminating any errors which may arise from unwanted earth currents.
Operating in the DC voltage range 1 μV - 1.000 V and the DC current range 1 pA - 1 A, this new DC microvolt and picoampmeter offers versatility matched by a high degree of sensitivity.

Philips mobile instrumentation show on European tour

The Electronic Measuring Instruments Division of Philips uses a bus, that has been especially designed as a mobile exhibition. Unlike normal tradeshows, which unvariably give a complete product survey, this exhibition focusses exclusively on *new equipment.*
Equipment like the compact LF generator PM 5125, 150 MHz counter timer PM 6630, or the pulse generator PM 5770.
The advantages of such a bus are obvious:
maximum information in a minimum amount of time, the show comes to you, just a walk away from your office. Timely knowledge on the newest products available can be a great asset when making up your instrument requirements for the next budget!

The bus left Holland at the end of August and will cover the whole of Western Europe from Madrid up to Stockholm. Several weeks are being spent in each country and a great many towns will be visited before it returns to the Netherlands.
The display will be kept up-to-date and alive with new products throughout the tour.
We shall be happy to welcome you, the readers of this journal, whenever we are in your area.
If you require further details of our visit to your country, please write or ring the national Philips organisation, the address is on the back cover.
We are sure you will find the visit worth while.

Some other Philips publications

Philips Technical Review: A monthly publication dealing with technical problems relating to the products, processes, and investigations of Philips Industries (Subscription: Dfl 24 per annum)

Philips Research Reports: A bimonthly publication containing papers on the research work carried out in various Philips Laboratories (Subscription: Dfl 30 per annum)

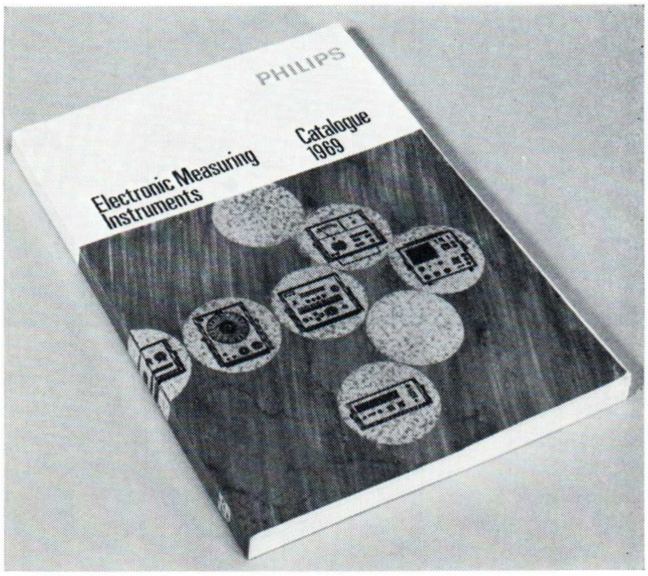
Electronic Applications: A quarterly publication containing articles on electronics (Subscription: Dfl 15 per annum)

Philips Telecommunication Review: A quarterly publication containing articles on all subjects in the field of telecommunications (Subscription: Dfl 16 per annum)

All above mentioned publications can be ordered from n.v. Centrex, Eindhoven, postbus 76, the Netherlands.

Philips in Science and Industry: A quarterly publication containing articles on new developments and applications of products and systems for industry and research. Available free of charge, via local Philips organizations only.

New literature



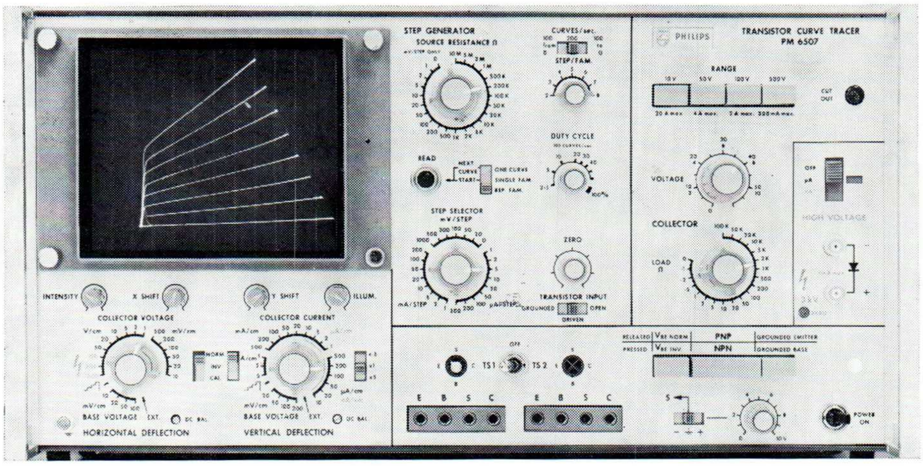
New catalogue with full details of Philips electronic measuring and microwave instruments edition 1969, is available now. This 150-page-book contains extensive data on the range of electronic voltmeters, oscilloscopes, recorders, LF measuring system, television test equipment, pulse and square-wave generators, transistor measuring equipment, bridges, sound measuring equipment and microwave measuring instruments. It is illustrated with a multitude of photographs and is available in English, French and German. Ask for your free copy with the Philips organization in your country (see adress on back cover of this magazine).

She joined on Monday
Now she's testing 220 transistors an hour
-It's still Monday morning

That's how simple it is with Philips new PM 6507 Transistor Curve Tracer. Programme it and it's a semi-automatic "go/no-go" batch tester of FETs, MOSTs or any other four terminal semiconductor. For accurate measurements of individual parameters, it traces single curves or families of up to eight. It shows the behaviour of transistors in the breakdown region, without damaging the devices under test — and of course you can photograph the results. The use of **pulse techniques** enables you to test high power devices without employing elaborate cooling systems.
You'll find the PM 6507 in semiconductor and equipment plants, development laboratories, universities and technical colleges.

Look at these features:

Traces dynamic characteristics of FETs, MOSTs.
10 nA sensitivity and 3 kV supply for diode testing.
Collector current range up to 5A/cm.
Polarity reversible in both display axes.
Overall accuracy ± 3%.
Built-in "fourth connection" supply.
10 x 12 cm flat screen display.
Variable duty cycle to test with minimum dissipation.



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