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We thank all our readers wherever they are for their continued support and wish them all a Prosperous and Peaceful New Year!

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MEMBER OF THE AUDIT BUREAU OF CIRCULATIONS
CABLE LEADS THE WAY IN BROADCASTING REVOLUTION

It is forecast in a recent report (*) that by 1994 almost 39 million homes in western Europe may have cable TV.

Subscription rates are set to rise from the 1988 total of nearly £1.25 billion to over £4.5 billion by 1994, while demand for the new installations will push the market for cable TV to £630 million.

At the moment, cable TV has relatively low penetration in western Europe, especially in Germany, France, Italy and the UK. Whereas, in 1988 some 6.5 million homes in the Benelux countries had cable TV only 3.9 million German, 1.1 million French, and 270,000 British homes subscribed.

However, broadcasting is on the brink of revolutionary changes and challenges. The European cable TV industry is ready to meet them with strong confidence in the future. The report adds that there is a strong case for supporting the distribution of satellite TV broadcasts by cable TV networks.

Furthermore, up to now interactive services, such as telephony, data home banking, shopping, and video conferencing have contributed little to the revenues of European cable TV operators. The report forecasts that these services could be worth as much as £15 million by 1994 compared to the current £6 million.

The combined markets of the big four European countries offer the enormous potential of 82 million TV owning households. By 1994, Germany will have overtaken the Benelux, representing the lion's share (almost 23%) of all European cable TV subscribers. The number of Italian cable subscribers is forecast to grow the fastest to from 2.5 million to one million, followed by the UK, where numbers will rise to 3.7 million. In France, the number will rise to 5.3 million from the current 1.1 million, and Scandinavia will also see a healthy increase from 2.52 million to 6.4 million in 1994.

In low density urban areas, cable can be buried for as little as £8,000 a mile, but in city centres that cost can increase tenfold. The market for cable will rise from the current £84 million to £126 million in 1994. However, the major share of the equipment market is that for use in the subscriber's home, and sales in this sector are expected to increase from £1 million to £1.2 billion in 1994.

Among the most exciting developments in cable TV is that of coherent optical networks, which will make important improvements to the present direct detection methods, especially as demand for high-definition television (HDTV) increases. Coherent optical technology is capable of providing very high capacity systems without the noise and distortion associated with photodiodes and preamplifiers in a single detection system.

(*) Frost & Sullivan, The European Market for Cable Television Services and Associated Equipment (E12/90).

600 MILLION RELAYS IN CARS THIS YEAR

With the car, that old faithful among electrical switching devices, the relay, has captured an immense new market. In order to satisfy the increasing customer push for safety, convenience and economy, about 600 million relays will be installed—an average of 12 relays in each of the roughly 50 million motor vehicles produced worldwide.

To obtain approximate values for the stressing of individual components in a car, an average life of ten years is assumed with a total distance covered of 150,000 km in 3,000 hours of operation. With roughly ten uses per day, this results in a total of 50,000 journeys over an average distance of three kilometres at an average speed of 50 km/h. For the relay, a fundamental distinction must be made between systems that are required once for each start (e.g., door locking, battery check, petrol pump, starter), as well as between safety systems for frequent occurrences (airbag, flasher, windshield wiper with its calculated 2.5 million operations) and for frequent occurrences (theft, air bag, short circuit, for which a maximum of 50 operations is allowed).

In the last example of worst case in particular, absolute reliability is demanded, which requires "zero defects" as the quality maxim for the important components. But quality as a guarantee of safety and reliability occupies first place as a matter of principle in the production of automotive relays at Siemens. The necessity to satisfy the highest quality demands, equalling those of spacecraft components, is illustrated by a few examples of the extreme operating conditions in a motor vehicle. The relay is subjected continuously to high vibration and shock stress; must withstand temperatures between -40°C and +125°C; and must be immune to splash water and corrosive liquids. The ability to withstand short-circuits and contact welding, in the case of a locked rotor for instance, are further important requirements.

To meet these requirements, materials are improved continuously, new technologies are employed, production equipment and test methods are optimized and automated continually.

Siemens Ltd. Siemens House, Windmill Road, Sunbury-on-Thames, TW16 7HS.
SAFE WASTE FROM SCRAP FLUORESCENT TUBES AND BATTERIES

Fluorescent tubes are among the most common light sources in the western world. However, their mercury content means that scrap tubes are classified as dangerous waste in many countries. This also applies to button cells that have become widespread in cameras and similar equipment.

The MRT distiller, which is at the heart of the system, is process-controlled and has a high reclaim level.

A new technique, developed by MRT System AB of Sweden, now enables the mercury to be reclaimed and the waste rendered safe. Various kinds of technology were tested before it was discovered that distillation is a promising way of reclaiming the mercury. The company now recycles 4 million of the 12 million fluorescent tubes consumed in Sweden each year and the plant has enough capacity for a further 6 million tubes.


PERSONALIZED ANSWERING AND MESSAGING SERVICE

The 'Meta System' from Millicom Information Services, which combines a telephone answering service with a full message pager, provides a unique service with cost and time benefits over alternative systems.

Epson has recently launched a family of high-quality portable colour LCD television receivers. Top of the range is the 'Vision System', which is a mini TV with 3.3 inch (84 mm) screen and active speakers on a stand. Prices range from £199.99 for the 2.6 in. (66 mm) Pocket TV to £3169.99 for the Vision System (prices include VAT).

NEW PORTABLE COLOUR LCD TV

40 messages, each of up to 296 characters, to be stored. The receiver also automatically records the number of each message for future reference.

Millicom Information Services Ltd, South Bank Business Centre, Ponton Road, LONDON SW8 5BL.

IN-CAR TRAFFIC JAM ALERT

What is claimed to be the world's first real-time traffic information system can rapidly warn drivers in the London area of motorway traffic congestion within 56 km of the city, enabling them to avoid delays and save transport costs by choosing alternative routes.

Designed to provide accurate information on the location, speed, direction and length of any tail-back, Trafficmaster is the first system to provide an in-vehicle 'bird's eye view' of the motorway traffic situation in its area of coverage. In the first phase of operation, it covers the M25 orbital road around London, which is Europe's busiest motorway, and radial motorways within 56 km of the centre of the city. It uses more than 230 infra-red sensors mounted on motorway bridges at about three kilometre intervals to log the speed of traffic passing below and alert drivers with special dashboard-mounted receiver/display units to any problems.

When traffic speed drops below 40 km/h, a sensor signals the fact to a control centre. From there, the data is transmitted via an existing radio paging network to the receiver unit. This displays the information, updated every three minutes until normal speeds are again detected, on a screen in map form. It gives the driver an audible and visual signal when updated details are received, then zooms in to display a close-up map of the area where the traffic build-up is taking place, with a flashing block to show where the jam is. The number of these shows the length of the tail-back.

Fieldtech has announced the arrival of IFR's latest Radio Communications Service Monitor, designated FM/AM 1600S. This top-of-the-line instrument combines a large colour CRT display with microprocessor control to offer the user exceptional clarity of readings and ease of operation for full Radio Communications Monitoring and Servicing, with the inclusion of facilities for data transmissions.

UK AND FRANCE LINK ON ELECTRO-OPTICS

Elevor Electronics research and development in Britain and France is to be pooled as a result of collaboration between major electronics companies in the two countries.

Under an agreement between Thorn-EMI Electronics and Sociéte Anonyme de Télé-

ELEKTOR ELECTRONICS DECEMBER 1990
communications (SAT), their electro-optics divisions will not only collaborate on the research and development of new systems, but also market each other's products.

Current and future products in the areas of thermal imaging and infra-red search and track systems are covered by the agreement. Together the two companies will occupy a leading position in the development, manufacture and marketing of electro-optical systems both within Europe and world-wide.

Thom-EMI Electronics Ltd, 1 Forest Road, FELTHAM TW13 7HE, England.

HYPERBAND CHIP

The introduction of the hyperband range for transmitting new channels in cable networks means that the tuning range of the TV tuner has to be extended. To meet the resulting requirements, Siemens has developed a circuit that integrates the hyper- and UHF-bands. It comprises three tuner sections covering the VHF1 (48-170 MHz), VHF2 (170-470 MHz) and UHF (470-860 MHz) bands.

The TUA 2017 combines on a single chip three combinations of mixer and oscillator for the three bands, an amplifier stage for driving a SAW filter and an amplifier stage for driving a PLL or prescaler.

IEEIE ENCOURAGE LEARNING OPPORTUNITIES

The approach of 1992 and the opening up of European markets bring with them not only a host of new opportunities but also increased responsibilities such as the need to comply with new regulations and standards.

With this in mind, the Institution of Electronics and Electrical Incorporated Engineers (IEEE) is acutely aware of the need to encourage all engineers and technicians to be kept abreast of the latest technical developments.

The Institution's commitment to Continuing Education and Training (CET) is demonstrated by its on-going programme of lectures and symposia, mathematics opening learning programme, its Training Access Point (TAP) and many publications, including the series of technical monographs.

Additionally, the IEEE now offers a programme of one-day or half-day Professional Development Seminars.

IEEE, Savoy Hill House, Savoy Hill, LONDON WC2R OBS, Phone 071 497 9006.

CALLS FOR PAPERS

Papers are invited for the following events.

The Fifth International Conference on HF Radio Systems and Techniques, which will be held at the Edinburgh Conference Centre, Heriot-Watt University, from 22 to 24 July 1991.

The Fourth International Conference on Television Measurements, which will be held in Montreux, Switzerland, from 20 to 22 June 1991. Papers are sought in the fields of cable television, terrestrial television broadcasting and direct broadcasting by satellite.

Conference Services, IEEE, Savoy Place, LONDON WC2R OBL, Telephone 071 240 1871.

The Ninth International Conference of Women Engineers and Scientists (ICWES 9), which will be held at the University of Warwick from 14 to 20 July 1991. The conference will cover a wide range of topics from Acoustics, through Telecommunications and Satellites, to Technology Transfer and Home Banking.

Conference Services Ltd, Congress House, 55 New Cavendish Street, LONDON W1M 7RE, Telephone 071 486 0531.
**ACTIVE MINI SUBWOOFER - PART 2**

by T. Giffard

This second part of the article describes an output amplifier designed for the subwoofer; the fitting of the electronics in the enclosure; and how the subwoofer can be connected to an existing audio system.

**Output amplifier**

Although in principle any output amplifier that can deliver about 50 watts into 8 Ω may be used with the subwoofer, we felt that many readers would want a complete system and so we designed an output amplifier especially for them.

The amplifier is a hybrid circuit consisting of a control section based on an opamp, and a power section that uses discrete transistors. Its circuit diagram is shown in Fig. 8.

The opamp, a Type OP16 from PMI, is a precision type with lpf inputs and a slew rate of 25 V/μs. It has its own power supply of ±15 V, which is derived from the 30-V main supply via Ri5/D1 and Ri6/D6.

The input signal is taken to the non-inverting input of the opamp via Ct. The input impedance is determined almost entirely by Ri (since the opamp has lpf inputs).

The bandwidth of the OP16 is restricted to some extent by a 22 nF capacitor between the output and inverting input, and a 100 Ω resistor between the inverting input and ground. This arrangement may be compared to the compensation capacitor between the outputs of the first differential amplifier in a conventional output stage.

The output of the opamp drives the power section via a current source based on T1. This source ensures a stable setting of the quiescent current through the output transistors. The voltage reference in the source is provided by a high-efficiency LED (D1).

The power section consists of a complementary compound configuration, T3-T6. Normally, a kind of super emitter follower is used in the output to ensure adequate current amplification. In the present design, current amplification alone (a typical characteristic of an emitter follower) is not sufficient, because the signal excursion at the output of the opamp is limited to about ±12 V. Some additional amplification is therefore needed. A compound circuit provides current as well as voltage amplification.

The voltage amplification in the present circuit is determined by the amplification factor of the output transistors and the potential divider, R8-R10, between the output transistors and the drivers. To make sure that the opamp does not provide too high an output voltage, which would limit the output current, the amplification of the compound output circuit has been made ×4 (12 dB).

Notable in this output stage configuration is the location of the emitter resistors of the output transistors, which are connected to the power rails.

Setting of the quiescent current level is accomplished with variable 'zener diode' T2-P1-R4. Transistor T2 is clamped to the heat sink between the output transistors to ensure good thermal coupling. Capacitors C7 and C13 provide a.c. decoupling of the 'zener'.

The feedback loop of the overall amplifier consists of resistors R2 and R3, which set the overall amplification to ×23 (27 dB).

The circuit around T7 and R11 provides a delay of a few seconds between power on and connection between the loudspeaker and the...
The circuit of the power supply is straightforward — see Fig. 9. Apart from the four 10,000 µF capacitors shown here, two more 1000 µF capacitors on the board provide additional decoupling of the power lines.

Construction

The amplifier is best built on the PCB shown in Fig. 10. Apart from the mounting of transistors T2–T6, the construction should not present any problems.

Transistors T2–T6 may be fitted in various ways, depending on the mechanical construction. If use is made of an aluminium L-section, they can be fitted above the board and fastened to the L-section, which in turn is screwed to the heat sink.

It is, however, also feasible to screw the amplifier and filter boards on to an aluminium sheet of suitable size, which then serves as the heat sink. In that case, fit T2–T6 to the sheet first, bend their terminal wires upwards a couple of millimetres above their body and pass these through the relevant holes in the PCB. Make sure that sufficient space is left between the board and sheet to allow solder connections to be made. Also, bear in mind that the transistors must be insulated from the sheet.

For clarity's sake, the latter construction, on a 3 mm thick aluminium sheet, is shown in Fig. 11. The dimensions of the sheet allow it to be fitted in the space in the back of the subwoofer enclosure. For that purpose, glue four triangular wooden supports in the corners of that space to which the built-up sheet is screwed later on.

Fit the boards to the sheet with the aid of 10 mm spacers.
The power supply is fitted as far away from the boards as possible to avoid any possibility of hum.

Note the separate earth connection for the delay circuit (indicated on the PCB by an earth symbol and asterisk) to the central earthing point. **Do not make a direct connection between the two earthing points on the amplifier board.**

Do not yet connect the loudspeaker to the amplifier.

When everything is ready, first set Pt for minimum resistance and then switch on the mains. Next, adjust Pt for a quiescent current through the output amplifier of 100 mA; this is measured with a millivoltmeter across R11 or R12 where the reading should be 22 mV.

Finally, switch off the mains, connect the loudspeaker to the amplifier and close the loudspeaker box.

**Connecting the subwoofer**

There are two ways in which to connect the subwoofer to an existing audio system. If the system has discrete pre- and output-amplifiers, or an external connection between these units when integrated, the best way is to feed the output of the pre-amplifier to the subwoofer via a screened audio cable. If that is not possible, connect the (second pair of) loudspeaker terminals of the system to the banana sockets on the subwoofer.

When the connections between the audio system and its loudspeaker boxes are long, it is possible to extend them from these boxes to the subwoofers, since the latter should in any case be near the loudspeakers for optimum performance.

The low cut-off point of the existing system and the subwoofers may be matched in several ways. When separate pre- and output-amplifiers are used, a simple first-order high-pass filter may be provided by adapting the input capacitor of the power amplifier. If the input impedance, $Z$, of the power amplifier is known, the value of the capacitor for a cut-off frequency, $f$, is given by:

$$C = \frac{1}{2\pi f Z}$$

Another way is adapting the cross-over network in the loudspeaker boxes. This is not so simple, however, because in the low frequency range the resonance peak of the subwoofer will have an effect, so that the filter cannot be terminated into a pure resistance.

A third possibility is to leave everything as it is. Particularly with small loudspeaker boxes where the low cut-off frequency is in any case fairly high—normally 75-100 Hz—it is perfectly all right to just connect the subwoofers into the system.

A fourth solution would be to precede the present output stage by a cross-over network of a type of which we have published several during the past few years. This is a rather exaggerated solution, but it is there if you want.

The location of the subwoofers is not very important, but they should preferably be not too far from the loudspeakers. Critical listeners may like them between the loudspeakers.

The sound level may be set with the potentiometer at the back of the subwoofers.

Finally, the input signals may be inverted with the aid of the phase switch if needed. Some experimentation here may well prove to be interesting.

---

**Fig. 11. Wiring diagram of the output amplifier complete with its power supply.**
Introduction

Throw a stone the size of a golf-ball into a can of water the size of a tea-cup and see what happens: virtually all points on the surface of the water are disturbed simultaneously. This is a rough mechanical analogy to the case of a ‘lumped’ electrical circuit, e.g., a simple resistive potentiometer, comprising two resistors, subjected to a transient input.

Throw the same stone into the middle of a village pond and observe a different effect: all points on the surface of the pond are not affected simultaneously: they are disturbed only as ripples move outward from the point where the stone falls. This is a crude mechanical analogy to the case of a ‘distributed’ electrical circuit, notably a transmission line, with a transient input.

The difference between the two cases arises through the finite time taken for disturbances to be transmitted. A variation of the pond analogy, which has long been used in the study of wave transmission, is the ‘canal’ analogy. In this, we consider what happens when a straight plank is dropped into a canal in a direction perpendicular to its length. Straight ripples, parallel to the length of the plank, move outward from the place where it falls. This analogy is more appropriate in the discussion that follows, because propagation is characterized by movement principally in one dimension.

Transmission lines are very important in digital electronics because of their use in the distribution of fast logic signals, but their operation is sometimes a puzzle to budding engineers (some with a predominantly mechanical engineering background), who have been taught the basic principles of lumped-circuit electronics but who have not studied established Electromagnetic Theory (or been convinced by it, even if they had!).

This introductory article sets out to clarify the understanding of some fundamental aspects of the pulse operation of transmission lines, particularly the popular twisted pair line (t.p.l.). The aim is to concentrate on the basic circuit theory aspects and practically observable waveforms that support the theoretical background.

Line modelling

A section of t.p.l. is shown diagrammatically in Fig. 1. In reel form, this can be purchased commercially (e.g., from RS Components), but for line lengths of a few metres, a t.p.l. may be made up by twisting together, uniformly, two pieces of pvc insulated wire (26 gauge, say) with a pitch of about 5 cm.

If we imagine the t.p.l. as laid along an x-axis perpendicular to the plane of this page, the field patterns that exist when equal-magnitude direct currents flow into the page at ‘a’ and ‘b’ respectively are shown in Fig. 2, where the solid lines indicate the nature of the magnetic field and the dashed lines the configuration of the electric field. These field patterns correspond also to those of the basic propagation mode for line transients discussed throughout this article.

The magnetic flux linking the wires is proportional to the current. The flux per unit current is represented by a series-inductance L per unit length. L is a parameter dependent on conductor geometry and can be estimated by analytical principles well known in field theory but a knowledge of L, by itself, is rarely required by t.p.l. users and, if needed, is best inferred from other readily measurable parameters. The electric field and flux associated with the conductors and the line charge on them are proportional to the p.d. between them, so the t.p.l. has also a per-unit-length capacitance C. As with L, this can be estimated theoretically, if required, but is readily determined practically.

Series losses may be represented by a per-unit-length resistance R and shunt losses resulting from leakage, through wire insulation, by a per-unit-length conductance, G. The t.p.l., although distributed in nature, can nevertheless be considered as made up from as large a number as we wish of tiny lumped sections, each of length dX, connected in series. The idea of using a large number of small discrete lumps to simulate a continuous variable is not unfamiliar in electronics. Thus, a digital time base for an oscilloscope based on a counter and D-A converter produces a horizontal pattern of dots on the
screen. However, for a 10-bit converter, the number of dots exceeds 1000 and on a 10-cm screen the trace appears continuous. The specific configuration of series and shunt components adopted to model an elemental section of line is a matter of sensible choice. All choices must, by definition, be equivalent in electrical choice. We could use a 'T-section', but the 'L-section' shown in Fig. 3(a) is analytically more convenient.

Figure 3(a) is often used for coaxial cable lines in which the outer conductor is 'earthed' but this can be misleading, particularly for a t.p.I., because it may give the false impression that one of the conductors behaves in a different way, electrically, from the other. The alternative model shown in Fig. 3(b) shows $R$ and $\ell$ as equally shared between the two conductors of the t.p.I. and in that respect is conceptually more attractive.

For a t.p.I. a few metres long, series and shunt losses can usually be neglected and the section reduces to the 'ideal' or 'lossless' form ($R = G = 0$); it is tempting to say that this is fortunate for were it not so, the t.p.I. would be of very restricted use.

For this case, the relevant equations lend themselves simply to pictorial interpretation and the essential features of line operation are not obscured by second-order effects.

Line equations

Consider the section shown in Fig. 3(c). The currents, $i_1$, $i_2$, shown flowing in the upper and lower inductance elements must be equal in magnitude to $i$, say.

The reason for this is as follows. If we imagine the line to the right of the points $p$ and $q$ to be contained within the 'black box', the Law of Conservation of Charge requires that $[(i_1 - i_2)] = 0$. This is true only, irrespective of the timescale $t$, if $i_1 = i_2 = i$.

Applying Kirchhoff's Voltage Law for loop voltage drops,

$$ u = (v_1 + i_1 R_0) + (v_2 + i_2 R_0) \neq (v_1 + i_2 R_0), $$

where $v_i$, the inductive voltage drop, is given by

$$ v_i = (L\Delta i)(\Delta i / \Delta t). $$

Substituting for $v_i$ and rearranging:

$$ -iR_0 \Delta i = L(\Delta i / \Delta t) $$

In passing, it may seem contrary to write the p.d. at $(x + \Delta x)$ as $(x + \Delta u)$, with a plus sign for the increment, when physical considerations tell us that it must be less than $u$. However, this is in the tradition of differential calculus. The physics of the problem gives the negative sign in [1].

Kirchhoff's Current Law for Fig. 3(c) gives

$$ \frac{\partial i}{\partial x} = C(\Delta i / \Delta t). \tag{1} $$

In the limit case $\Delta x \rightarrow 0, \Delta t \rightarrow 0$, the approximation sign becomes an equality symbol. We have not proceeded to this limit yet, because the aim is to avoid the distraction of partial differential relationships that arise when a function is dependent on two or more variables. Indeed, combining [1] and [2] to eliminate $\Delta i$ or $\Delta u$ leads to the (partial differential) 'wave equation' for an ideal line, but such a procedure requires us to solve the equation $\Delta t$, at least, quite solutions for it.

An alternative approach is to show that a voltage step at the input to the line travels along it with constant amplitude and uniform velocity. To do this, we must first establish a relationship between $u$ and $i$ and then derive an expression for step velocity that is independent of $y$.

$u / i$ relationship; characteristic resistance, $R_0$

Dividing each side of [1] by the corresponding side of [2] gives:

$$ (\partial u / \partial x) = (L/C)(\partial i / \partial t). \tag{2} $$

or

$$ (\partial u / \partial t)^2 = (L/C). \tag{3} $$

Taking the square root and proceeding to the limit,

$$ (\partial u / \partial t) = \sqrt{(L/C)} = R_0, \text{ say}. $$

We are entitled to express [3] in total differential form because it is valid irrespective of $t$. Equation [3] gives the limit case for small changes. The ratio is given the symbol $R_0$, because $\sqrt{(L/C)}$ has the dimensions of resistance. $R_0$ is known as the 'characteristic resistance'. It is characteristic of the line alone and not dependent on the nature of $u$ or $i$, and is the incremental resistance looking to the right (or left) between terminals $p$ and $q$, or $p'$ and $q'$. The expression 'characteristic impedance' is often used but is un-
necessary for a lossless line. It conjures up thoughts of the frequency variable \( w \) (or \( jw \)) and we are operating here strictly in the time domain.

\( R_p \) is unlike a normal resistor in that it dissipates no power: it is a parameter, dependent on line geometry, that fixes a relationship between the instantaneous changes in \( i \) and \( u \) either of which can be regarded as a stimulus while the other is regarded as a response. In particular, a step change in \( u \) produces a step change in \( i \) and vice versa.

Integrating (3), the instantaneous value of \( u \) is:

\[
 u = iR_p + U_0 \tag{4}
\]

Equation (4) is illustrated in Fig. 4., in which \( U_0 \) is any initial line voltage, shown here arbitrarily as positive, and for changes in \( u \) and \( i \) is accessible only via the series resistor \( R_p \). This accessibility to a line voltage source only via a series resistor \( R_p \) is true also looking to the left at a point on the line, because the line has no built-in directional properties for pulse propagation.

For an initially uncharged line, treated from now on, \( U_0 = 0 \) and the circuit looking to the right between \( p \) and \( q \) reduces to the simpler form of Fig. 4(b).

A step voltage of magnitude \( U \) appearing at one moment between \( p \) and \( q \) appears at a later time between \( p' \) and \( q' \), charging up the line as it progresses with velocity \( v \). There is no loss in amplitude as there are assumed to be no line losses.

**Propagating velocity \( v \)**

The propagation velocity, \( v \), is found as follows. Multiplying each side of (11) by the corresponding side of (12):

\[
 (\partial u / \partial x) / (\partial t) = v = \frac{L}{C} \left( \frac{\partial u}{\partial t} / \partial x \right) \tag{5}
\]

Thus, in the limit,

\[
 v = \frac{dx}{dt} = \sqrt{\frac{L}{C}} \tag{6}
\]

Since \( v \) is independent of \( v \), the velocity is constant along the line. The time, \( t_0 \), is

\[
 t_0 = \frac{1}{v} = \sqrt{\frac{L}{C}} \tag{6}
\]

Let us check (5) another way. We assume that \( v \) is constant and apply the principle of charge conservation. If a step waveform \( U \) travels from \( x = 0 \) to \( x = x' \) in a time \( t = (x/v') \), the charge supplied to the line by the source is \( (xL/v') = (U_0R_p)/(xL/v') \). This must equal the charge accumulated by the line capacitance from \( x = 0 \) to \( x = x' \) and this is \( C \cdot U' \). Thus,

\[
 v = \frac{1}{C} \frac{R_p}{x} \tag{7}
\]

Substituting for \( R_p \), from (3) in (7) gives the same value for \( v \) as in (5).

Note that \( R_p \) and \( L \) are two basic parameters required to be known of a line. From these can be found \( C \) and \( L \), if needed, with the aid of equations (3) and (6).

**Models for step waveform progress**

Progress of a step waveform is so basic that it merits further study.

We consider a mechanical analogy and an electric circuit model.

In a mechanical analogy, we may consider a stationary hopper containing sand over a conveyor belt that is moving to the right. At a chosen moment, the exit pipe from the hopper is opened suddenly. The result is a constantly lengthening, uniform-thickness, trace of sand on the belt. Sand here, of course, analogous to electric charge.

A model for step progress attractive to the engineer more at home with lumped circuit theory involves the concept of a 'sliding source'. Consider the progress of a step voltage wavefront of magnitude \( U \) along an initially uncharged line in the direction of increasing \( x \).

Looking to the right at any point \( x' \), the remainder of the line appears as a resistor \( R_p \) as shown in Fig. 5. Looking backwards, towards \( x = 0 \), the line appears as a voltage source \( U_0 \) accessible via a source resistor \( R_p \) (inside dashed rectangle), both of which appear to slide along the line with velocity \( v \). To produce a step of magnitude \( U \) at \( x = x' \), it is obviously necessary that \( U_0 = 2U \).

This sliding source approach is helpful in calculating what happens at the end of a line of finite length \( l \).

**Line Voltage \( u(x,t) \): step and pulse drive**

To investigate, experimentally, a t.p.i. subjected to a step input, the line can be 'voltage-driven' or 'current-driven' as shown in Fig. 6.

In both cases, the condition of switch \( S \) is assumed to change at \( t = 0 \).

Simple experimental predictions of terminal voltage behaviour based on Fig. 6(a) require a knowledge of \( R_p \) and the certainty of its constancy over the range of the output voltage swing. It is not possible to guarantee constancy using stan-
dard saturated transistor logic circuits (e.g. TTL) to voltage-drive the line.

In the current-drive scheme of Fig. 6(b), the output resistance, \( R_o \), is generally much greater than \( R_f \) and can be ignored by comparison with it. This is the case, in practice, with a switched long-tail pair driver stage. \( u(x, t) \) denotes line voltage as a function of variables \( x \) and \( t \). Of special interest are: \( u(0, t) \), the variation with \( t \) at \( x = 0 \), i.e., the input waveform. \( u_i(t) \); \( u(x, t) \) the waveform at an arbitrary point \( x = x' \); \( u(x, t) \), a plot of line voltage as a function of \( x \) at a specific time \( t = t' \).

In Fig. 7(a), \( i(t) \) is a step of magnitude \( I \) because the line appears initially, at its input terminals, as a pure resistance \( R_o \).

\[ u(x', t) \] in Fig. 7(b) is \( u(0, t) \) delayed by a time interval \( t' = t - t_0 \). The line is uncharged at \( x' \) till the step reaches that point. If the switching action occurs at \( t = t_0 \), the line is charged up to the point \( v(x', t_0) \).

Unlike \( u_i(t) \) and \( u(x, t) \), which can be monitored, \( u(x, t) \) is not a waveform. However, if we choose an appropriate scale on the paper as in Fig. 7(e), we can make the graph appear complementary to that of \( u(x, t) \). This means that the sum of ordinates of the two graphs, at each time point on the horizontal axis, gives a constant value. This scale changing “trick” is useful in deriving \( u(x, t) \) from \( u_i(t) \) for the general case of a line signal that is not a step, as we will show now.

In Fig. 8, \( n \) current sources each of strength \( I/n \) are connected to a t.p.l. via switches \( S_1-S_n \).

\( S \) changes state at \( t = t_0 + \Delta t \), where \( \Delta t = (t_n - t_0)/n \), and pumps a current \( I/n \) into the line. This is followed by successive time intervals \( \Delta t \), by \( S_1-S_n \), respectively, causing additional current steps \( I/n \) to be applied in sequence to the line input.

Equation [4] specifies a linear relationship between \( u \) and \( t \), so the Principle of Superposition is applicable and we can add algebraically the effects of each input taken separately to obtain the overall response.

The resulting waveform for \( u_i(t) \) is a voltage ‘staircase’, which, for \( n = 4 \), is shown in Fig. 9(b). The dashed line joining the edges of the treads intersects the \( t \)-axis at \( t = t_0 \). Suppose now that instead of \( n = 4 \) we let \( n \to \infty \). The staircase edge then becomes the profile of the dashed line in Fig. 9(b) and Fig. 10. We have thus deduced \( u(x, t') \) for a ramp input. The general case for an input of arbitrary shape is worked out similarly by considering steps of unequal magnitude and—if necessary—opposite polarity when the switches in Fig. 5 change state.

Thus, the digital input signal of Fig. 11(a), with transition times \( t_i \) and \( t_f \) purposely chosen unequal, produces \( u(x, t) \) in Fig. 11(b). Figure 11(c) may be regarded as a scaled mirror image of Fig. 11(a) displaced along the horizontal axis. An alternative graphical method for obtaining \( u(x, t') \) from \( u(0, t) \) is given in the reference at the end of this article.

**Fig. 12. Current-driven terminated line.**

**Fig. 13. Calculation of terminal voltage at \( t = t_q \).**

**Fig. 14. Generation of \( u_i \) for \( R_f = R_o \): (a) \( R_f > R_o \); (b) \( R_f < R_o \).**

**Fig. 15.** (a) Line voltage for \( 2t_q \geq t \geq 4t_q \); (b) Sliding circuit equivalent-circuit form for (a); (c) Circuit for calculating \( u(2t_q) \).

**Reflections**

It is convenient to imagine a semi-infinite line, stretching from \( t = 0 \) to \( t = \infty \), in an initial discussion of lines because it simplifies the presentation. However, once the progress of a step waveform is understood we can consider what happens at the end of a line of finite length \( l \) when a pulse edge or a pulse of arbitrary shape reaches it.

Consider the scheme shown in Fig. 12, where \( R_f \) is a terminating resistor. A voltage waveform \( u_f \) of amplitude \( U = IR_f \), which we call the forward waveform, starts down the line at \( t = 0 \) when \( S \) opens. It reaches the end of the line in the one-way delay time \( t_q = l/v \).

The terminal voltage \( u_T \) at \( t = t_q \) is calculated from the sliding source equivalent circuit of Fig. 13: \( u_T(t) = 2UR_f/(R_f + R_o) \). Now, \( u_T = u_f \) (the terminal voltage step is equal to the amplitude of the forward voltage waveform on the line) if \( R_f = R_o \).

This is the case of a line ‘matched’ or ‘correctly terminated’ at the receiving end. Then \( R_f \) dissipates energy at the same rate as it is supplied to the line from the source. No energy is reflected, that is, sent back to the source. As far as any effect on the sending end is concerned, the line may just as well be considered semi-infinite despite its actual finite length. There is an analogy here in radar. If the energy in a radar beam is completely absorbed by a target, there is no reflection, that is, the target is “invisible”. As far as the radar receiving equipment is concerned, the target may be regarded as located at a point an infinite distance away. Suppose, however, that \( R_f \neq R_o \). Then, \( u_T \neq u_f \), all the energy associated with \( u_f \) cannot be absorbed by \( R_f \), and a reflected waveform \( u_f \) is pro-

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duced. The amplitude and polarity of $u_{i}$ must be such that the Principle of Superposition is applicable at the termination. Thus,

$$u_{f} + u_{f} = u_{f},$$

or

$$u_{f} = R_{f} - u_{p}.$$  \[9\]

Substituting for $u_{f}$ from [8] and $u_{f} = U$, gives

$$u_{f} = \left[2UR_{f}/(R_{f} + R_{o})\right] - U = \rho_{VT} U$$ \[10\]

where $\rho_{VT}$ is the voltage reflection coefficient at the termination and is defined by

$$\rho_{VT} = (R_{f} - R_{o}) / (R_{f} + R_{o}) \quad \[11\]

Figure 14 shows a geometrical construction giving $u_{f}$ for the cases:

(a) $R_{f} > R_{o}$, and hence $\rho_{VT} > 0$, and

(b) $R_{f} < R_{o}$, and hence $\rho_{VT} < 0$.

For either condition, the reflected voltage wavefront travels back to the source.

A plot of line voltage for $2t_{1} > t_{f}$ is shown in Fig. 15(a) for $\rho_{VT} = 0$. This results from adding $u_{f}$ to the existing line voltage giving a total line voltage $(1 + \rho_{VT})U$ at the position of the wavefront.

The total line voltage is also obtained from the sliding source equivalent circuit which, in this case, comprises a generator $2\rho_{VT}U$ in series with an output resistance $R_{o}$ as shown in Fig. 15(b):

$$u_{f}(2t_{1}) = UR_{o} + 2\rho_{VT}U = U(1 + 2\rho_{VT}) \quad \[12\]

Since there is already a line voltage $U$ and $u_{f} = \rho_{VT}U$, this means a forward, reflected wavefront of amplitude $2\rho_{VT}U$. This also follows from [11] since the voltage reflection coefficient is unity for an ideal current source.

The current-driven line of Fig. 12 with $R_{f} = R_{o}$ is of restricted use: Two cases of reflection of practical interest for a current-driven line with an intentional mismatch at the receiving end are considered next.

With reference to Fig. 16, in which a shunt matching resistor is incorporated at the sending end, the two cases correspond to $R_{f} = 0$ and $R_{f} = R_{o}$.

Consider first the case $R_{f} = 0$. Writing $U$ for $UR_{o}/2$, it follows that $u_{f}(0+) = U$. From [11], $\rho_{VT} = -1$. The equivalent circuit for calculating $u_{f}(2t_{1})$ and $u_{f}(x)$ for $t > 2t_{1}$ is shown in Fig. 17.

In Fig. 18, $u_{f}(t)$ is a pulse of amplitude $U$ and duration $2t_{1}$. The line input current, $i_{i}$, and the energy supplied by the source, $W_{s}$, are shown in Fig. 18(b) and Fig. 18(c) respectively.

An argument based on the Principle of Conservation of Energy leads to an algebraic expression for $i_{f}$. Thus,

$$W_{m} = LR_{f}^{2}/2 \quad \[14\]

Equating $W_{m}$ and $W_{s}$ yields:

$$i_{f} = (U/\rho_{VT}) = (U/2R_{o}) \quad \[15\]

where $\rho_{VT}$ is the voltage reflection coefficient at the termination and is defined by

$$\rho_{VT} = (R_{f} - R_{o}) / (R_{f} + R_{o}) \quad \[16\]

as previously shown in [6]. With reference to Fig. 16, the case $R_{f} = 0$ gives the waveforms for $u_{f}(t)$ and $i_{f}(t)$ in Fig. 19:

**Conclusion**

This article has dealt in detail with some aspects of line pulse operation that are either ignored or skimpily covered in the literature.

**Reference:**


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**PWM CONTROLLER IC**

The Si9120 pulse-width modulation (PWM) controller IC from Siliconix offers a low-cost solution to the provision of a wide input-voltage range for universal-input power supplies. The unique wide-input range of 50-450 V enables the Si9120 to operate directly from rectified 110 V or 220 V AC power lines. All essential controller functions are integrated in the Si9120, including high-voltage start-up circuitry, oscillator, error amplifier, voltage reference, and a non-inverted cross output driver for the external MOSFET. The low supply current of 1 mA allows highly efficient, very reliable operation at high temperatures, and the high frequency (500 kHz) meets the high-performance demands of modern power supplies. Siliconix has manufacturing and sales operations in the USA, United Kingdom, Hong Kong and Taiwan. Other sales offices are located in Germany, France, Italy and Sweden.
This Videotext decoder, designed and marketed as a kit by ELV, allows the decoding and storage of Videotext (or Teletext) pages on an IBM PC or compatible. Among the special features of the PC-VT7000 are fast access to subpages, the possibility of using a video recorder for separate processing of subtitles (particularly useful for the deaf and hard of hearing), and the use of a SCART-compatible TV set for displaying the decoded pages.

Videotext, Teletext, CEEFAX and Oracle are but a few names given by broadcasters to a special information service transmitted during the blanking period of TV signals. The information is brought to the viewer via pages of text and graphics, which can be called up by entering the appropriate number on the remote control of the TV set. Among the subjects in the Videotext service are news items, sports, weather information and TV programme overviews. In most cases, the pages are updated by the broadcaster's editorial staff for the Videotext service.

In the PAL TV system, 625 TV lines are transmitted as two interlaced fields of 312.5 lines each. About 50 of these lines fall inside the vertical retrace (or blanking) period, which is not normally visible on the TV screen. These 50 lines are used to convey test signals and digital information (see also Ref. 1).

Teletext is usually conveyed via lines 11 to 14, and 20 and 21, in the first field, and 324 to 327, and 333 and 334, in the second field. At a field frequency of 50 Hz, the maximum text line rate is about 300 per second, corresponding to about 12 pages per second.

To be able to receive Videotext pages, you need a special decoder. Most modern TV sets, and even some of the latest video recorders, have such a decoder as a built-in unit. Where a decoder is not part of the TV, it may often be purchased and installed as an upgrade.

After entering the requested Videotext page number on the remote control, the decoder starts to search for it. The search process is indicated by the three-digit page counter in the upper left-hand corner of the TV screen. When the page is found, the search process stops, and the relevant information is shown on the screen. Unfortunately, finding a particular page may take quite some time — depending on the reception conditions and the number of pages in the service, wait times of up to 10s are not uncommon. Particularly when frequent use is made of Videotext pages, the long wait time before they are available is a real disadvantage of an otherwise extremely useful information service.
The PC-VT7000 has a number of advantages over a conventional Videotext decoder built into a TV set. To use the unit, you require either a video recorder with a CVBS (chrominance-video-blanking-synchronization, also called composite video) output, or a TV set (with or without a Teletext decoder) with a SCART socket. The CVBS signal taken from this socket is fed to the PC-VT7000. After decoding and processing, the Videotext pages may be displayed either on the TV set (which takes in the video signal via the SCART socket), or on the monitor of the PC. A video recorder may be connected to the second SCART socket on the PC-VT7000 to enable Videotext pages as well as TV pictures with subtitles to be recorded. The latter option is of particular interest to the deaf and the hard of hearing.

A further special feature of the decoder is its ability to produce hard copy of Videotext pages on a printer. By using this option you are in a position to print out, say, the day's programme overview, or the current weather situation (which consists of charts and tables). The final advantage of the PC-VT7000 over a conventional Teletext decoder is that it enables you to have immediate access to subpages. Most conventional Teletext decoders allow you to enter the main page only. To view the subpages that belong with this main page, you have to sit and wait for the decoder to show them one after the other. Normally, a subpage is shown 10 seconds or so before the next appears. There is, however, no way to skip subpages to get to the one you do want to read. Many Videotext users find this irritating and a waste of time.

The PC-VT7000 has a special page memory that solves this problem by offering you immediate access to any subpage.

**Connecting the decoder**

The PC-VT7000 consists of two units: (1) an insertion card for PCs that provides an IC bus interface, and (2) the decoder proper.

The inputs and outputs of the decoder are found on the rear panel of the ELV 7000-series enclosure. These inputs and outputs are used to connect the IC insertion card and the video equipment.

The minimum equipment to run the system is a TV set with a SCART socket, which must be connected to the PC-VT7000 via a SCART cable. One of the pins on the SCART socket of the TV supplies the composite video signal, which is used by the decoder to extract the Videotext information. This information is processed and turned into a video signal that is fed back into the TV set, again via the SCART connection, which thus functions as a bidirectional link.

The second SCART socket on the rear panel of the PC-VT7000 allows a video recorder to be connected. The special use of the VCR for recording TV programmes with a subtitling service has already been mentioned.

The toggle switch on the rear panel is used to select either the TV set or the video recorder as the source of the CVBS input signal for the decoder. When this switch is set to TV, the TV set must be switched on—otherwise, the tuner cannot supply a CVBS signal to the decoder. For the same reason, when the switch is set to the other position, the VCR must be on or in stand-by mode, i.e., its tuner must supply a video signal that contains Videotext information. To enable the PC-VT7000 to store Videotext pages on the PC, or programmes with subtitled on the VCR, the input source switch must be set to TV.

There is one special equipment configuration in which a TV set is not required: when a VCR is used as the CVBS signal source, and a computer screen only to display the Videotext pages.

The IC insertion card is powered by the PC. The cable between this interface card and the decoder also carries the required supply voltage, so that a separate power supply is not required to use the system.

**The hardware: an overview**

The block diagram in Fig. 1 shows the way in which the previously discussed units are interconnected. The heart of the circuit is formed by the Videotext decoder, which communicates with the other units via a two-way multiplexer that forms part of the main decoder. The CVBS signal that contains the Videotext information of the relevant broadcaster is supplied by the tuner in the TV set. As already mentioned, the decoder may also accept the CVBS signal of a video recorder, provided this is not used to play a tape. Unfortunately, owing to their limited bandwidth and recording method, very few videorecorders are capable of reproducing a usable Videotext signal from tape. However, the VCR is perfect for recording and reproducing decoded Videotext pages and programmes with superimposed subtitles.

The Videotext decoder must be connected to store Videotext information (see also Ref. 1). Next, the information is either sent to the PC via the IC bus (see Ref. 2), or fed to the display controller which uses it to build a complete picture that can be displayed on the TV screen.

The system also allows decoded Videotext pages as well as subtitles with the current programme to be recorded (on the VCR) or stored (on the PC). It should be noted that the stored Videotext pages and the subtitles are displayed in black and white, while the VCR recordings are, of course, in colour.

**Control program**

The functions of the PC-VT7000 are controlled from a PC running a special program, loaded from floppy disk or hard disk. This program is called up by typing VT followed by a carriage return. The program automatically prompts the Videotext decoder to search and display page 100 on the PC monitor or the TV screen. The command to do so is issued by the PC insertion card and sent to the decoder via the IC bus. Page 100 is provided as a default value by simple programming, the software can be changed to load any other page on starting the system.

Although the control program supplied with the PC-VT7000 is largely self-explanatory, a help function giving details of all essential actions may be called up at any time by pressing function key FI.

The three-digit number of the requested page is entered on the PC keyboard. A page selection window appears on the PC screen when the first digit is typed. Since the page number invariably consists of three digits, the CR key need not be pressed when the number is complete. The requested page is displayed as soon as it is found in the Videotext datastream. If the page is not active, or can not be found, a message appears after a short while.

As already mentioned, the PC-VT7000 offers a fast way of calling up subpages. After
loading the main page, simply press the 1 or the ↓ key to leaf through the subpages. The action on part of the Videotext decoder is virtually immediate. The other two arrow keys, ← and →, are used to leaf through the main pages. Since the main pages are not stored sequentially, this may take more time than with subpages.

The current Videotext page may be sent to a printer by pressing F2.

Function key F3 allows the currently displayed page to be converted into a data file. After pressing F3 you are prompted to enter comment which helps you identify the page when it is retrieved later. The program automatically assigns the page number to the file as an identifier.

Videotext data files may be retrieved by pressing F4. You are prompted to enter the number of the requested page, which is subsequently loaded from disk and displayed on the PC screen. Note that this function is available even when neither the IC insertion card, the main decoder, nor the TV set are connected.

The list of Videotext data files stored on the computer, along with the associated comment, may be called up by pressing the F5 key.

Finally, the control program may be terminated by pressing the ESC key.

The Videotext decoder

The circuit diagram of the decoder is given in Fig. 4. As already mentioned, this circuit is fitted into a series-7000 type enclosure. The CVBS signal is applied to the decoder either via pin 20 of SCART socket BU1 (TV set), or via pin 20 of BU3 (video recorder). The CVBS signal supplied by the TV set is terminated with resistor R1, and applied to pins 4 and 11 of analogue multiplexer ICs. When the CVBS signal from the VCR is used, this is allied to pins 2 and 15 of the same IC. The required terminating resistance is then formed by the TV set, whose CVBS input is connected to pin 19 of socket BU1. In case a TV set is not connected, switch S1 must be set to the upper position to allow R2 to function as a terminating resistance.

The SCART sockets, BU1 and BU3, are wired in a manner that allows the video recorder and the TV set to be used for recording programmes and playing back tapes just as if the PC-V7000 were not connected, and without having to change any cable or connection.

The CVBS signal with the Videotext information in its vertical blanking period is applied to electronic switch (multiplexer) ICs. Depending on the source selection (Rec/TV) set with S2, either the CVBS signal from the TV set, or that from the tuner inside the VCR, is routed to the parallel-connected IC outputs, pins 3 and 13. This is achieved by S2 determining the logic level at address selection input A of the 4052.

SAA5231 VIP2

The composite video signal arrives at the input of the video processor, ICs (SAA5231) via coupling capacitor C0. The SAA5231 VIP (Video Interface Processor) extracts the Videotext information from the data carried in the previously mentioned TV lines in the vertical blanking interval. The block diagram of the SAA5231, which is manufactured by Philips Components, is given in Fig. 3. Its tasks include:

- separating and regenerating the Videotext information;
- generating a clock signal that is synchronous to the current picture;
- supplying data to the display controller that follows it;
- extracting the synchronization components from the CVBS signal;
- supplying the synchronization components to the display controller that follows it;
- switching to internally generated synchronization when the external synchronization fails;
- supplying the synchronization components at positive and negative polarity;
- locking the internal 13.5-MHz quartz-controlled oscillator to the applied CVBS signal;
- adjusting itself to the level of the applied CVBS signal.

The CVBS signal applied to pin 27 of ICs is fed to an internal adaptive data separation stage with a slicing level of 50% of the CVBS signal amplitude. The slicing level is set to 50% to achieve the highest possible noise immunity. The 8-bit data supplied by the VIP2 consists of 7 databits and 1 parity bit.

As shown in Fig. 3, the CVBS signal is also fed to the input of an adaptive sync separation circuit. The slicing level of this circuit is adjusted automatically as a function of the input amplitude. This is done to compensate low-frequency level variations.

The VIP2 supplies the Videotext data and the associated clock pulses at output pins 15 and 14 respectively, for use by the display controller that follows it.

The output frequency of the 6-MHz VCO (voltage-controlled oscillator) on board the VIP2 is controlled via a phase detector, with the aid of a line-frequency clock signal at pin 28. This is done to ensure that the generated Videotext characters are synchronized to the current picture as required for the subtitled service. The synchronization pulses obtained from the input video signal are also applied to the phase detector. The 6-MHz clock, which is phase-locked to the sync pulses, is coupled out via pin 17 and applied to the relevant input, pin 9, of the display controller, a SAA5243.

Pin 28 of ICs accepts the composite synchronization signal generated in the SAA5243. When the synchronization signal at the CVBS (TV programme) input of the VIP2 falls, this chip automatically switches to the replacement sync signal furnished by
Fig. 4. Circuit diagram of the VideoText decoder. Note that this circuit is controlled via an IC interface connected to PCB header STL1.
the ECCT. The ECCT also generates the sandcastle pulse, which is fed to the VIP2 for use in the Videotext data slicer.

Capacitor C15 and inductor L1 form the external components required to make the 6.938 MHz data clock filter operate. Similarly, quartz crystal Q2 and capacitor C25 enable the 13.175 MHz oscillator to operate.

Pin 1 of the VIP2 supplies the composite-sync signal for the TV set. Resistor R2 sets the polarity of this signal to positive by pulling it to the +12-V supply rail. The sync-locked 6-MHz oscillator operates with external components Q1, C13, C14 and R23. Trimmer C14 allows the synchronization to be adjusted.

**SAA5234 (ECCT) and page memory**

The full identification of the SAA5243, another Philips Components IC, is Enhanced Computer-Controlled Teletext Chip, which is mercifully abbreviated to ECCT. Together with a RAM Type 6264 and the VIP2, the ECCT forms the heart of the present decoder. It should be noted that the VIP2 and the ECCT are also available under the respective type numbers SDA5231 and SDA5243 from 'second source' Siemens.

As shown in the block diagram in Fig. 5, the ECCT contains a character generator, a data acquisition circuit, an 8-bit interface, a clock driver and a memory interface. These standard functions are boosted by the following extras:

- an integrated character generator with 160 alphanumeric and 2x64 graphical characters, each built in a 12x12x10 (HxVxD) matrix;
- user-controlled double-height characters for the upper or lower half of the Videotext page;
- insertion of all characters and colours via commands on the IC control bus;
- the current character position may be identified with a cursor;
- status information above or below the main text line (25);
- automatic switching of the character set to one of six languages by special control bits in the page header;
- simultaneous searching process for up to four pages;
- data capture in all lines of the frame/full-channel mode, offering last page access.

The clock driver in the ECCT communicates with the VIP2 video processor via pins 9 to 12. After checking their validity, the ECCT acquires the data and clock signals received from the VIP2. These data are stored on the external page memory RAM, a 6264, via the memory interface. The data acquisition is organized such that four Videotext pages can be searched for, and stored in RAM, simultaneously. Thus, these four pages are updated at the same time. The page memory is accessed with the aid of signals OE (output enable) and R/W (read/write). Data is carried via pins 22 to 29, and addresses via pins 2, 3 and 30 to 40.

The ECCT supplies the picture information via its three colour output pins, R, G, and B. The character generator has 256 characters, a selection of which is available in each of the six national character sets that can be called up by an appropriate software command.

The blanking signal for use with the RGB components is available at pin 17 of the ECCT. This signal is used during mixed picture operation as, for instance, Videotext annotation. The system has two modes of operation, which are selected by software:

- character insertion (superimpose);
- background suppression.

The Y-signal (luminance or brightness) is provided independently of the selected colour at pin 2 of the ECCT, and is thus only valid for the Teletext characters. A flash function is not provided as standard.

**Output circuits**

The R, G, B, blanking and Y outputs are of the open-drain type, and require external pull-up resistors. Resistors R11-R12 are fitted at the R output, R13-R14 at the G output, and R15-R16 at the B output. The ratios of these resistors determine the signal level at the base of the associated RGB transistor driver stage. Resistor R22 enables the output level of the RGB drivers and that of the Y output to be set to a value that produces optimum contrast of the Videotext characters in relation to the TV picture.

The four outputs are decoupled by diodes D1 to D4. The drivers for the RGB and blanking signals are built around four transistors in common-collector circuits, Ti to T4. Resistors R17 to R24 determine the output impedance and ensure optimum signal matching to the loads formed by the TV inputs. The signals are fed out of the circuit via the SCART socket for the TV set.

**VCR**

As already mentioned, the PCC VT7000 offers the user the possibility of recording Videotext subtitles on a VCR. This works as follows. The Y signal at pin 18 of the ECCT is fed to the base of emitter follower T6 via R31. The composite-sync signal is added to the video via R7. Capacitor C5 provides the necessary d.c. decoupling. The combination of R8 and R9 forms the pull-up resistor at the open-drain Y output, pin 18, of the ECCT.

The buffered VBS (monochrome) signal at the emitter of T5 is fed to the inputs pins 1 and 3, of electronic switches IC8 and IC9. Each second input of these switches, pins 2 and 5, has on it the CVBS signal (the original TV picture). This means that the system can switch between these signals. To make sure that the CVBS signal is at the right level, it is fed, via C14, to a clamping circuit composed of IC4, R3, R5, and C15. Since the positive sync pulse supplied by the VIP2 control the electronic switches, the CVBS input signal is clamped at a potential fixed by R3-R5. This ensures the correct d.c. levels at the second inputs, pins 2 and 5, of the electronic switches.

The control of electronic switches IC8 and IC9 is determined by the blanking signal. The relevant output, pin 15, allows one of three signal configuration to be selected:

- the original composite video signal;
- the Videotext image;
- a mixture of these (superimpose).

When the third configuration is used, the output supplies a signal composed of the CBYS TV signal and the VBS Videotext signal. This mixed signal is fed to a buffer, T5, via coupling capacitor C5. The buffered signal is taken from the emitter of T5, and fed to the video recorder input via pin 19 of the relevant SCART socket.

**Interface to I²C card**

The connection marked STL1 links the Videotext decoder to the IC. This connection carries the supply voltages for the decoder board, and the data.

All functions of the Videotext decoder are controlled via the IC bus interface, pins 19 and 20, of the ECCT. The relevant control signals are conveyed via the IC interface card in the PC. As already mentioned, this card forms part of the project.

Finally, connector STL2 carries a number of control and data signals that may be used for future extensions.

Next month’s second and final instalment of this article will deal with the operation of the IC card, and the construction.

A complete kit of parts for the Videotext decoder is available from the designers’ exclusive worldwide distributors:

**ELV France**

B.P. 40
F-67480 Sterlises-Bains
FRANCE

Telephone: +33 82837213
Facsimile: +33 82838180
As you are probably aware, measuring small resistance values is difficult, if not impossible, with conventional digital and analogue multimeters. While only a few of these instruments have a 1-Ω range with limited practical use, the meter presented here allows very small resistances in the range from 10 mΩ to 5 Ω to be measured reliably.

A. Rigby

That most multimeters have a lowest resistance range of 100 Ω or 1 kΩ is not surprising. The measurement of small resistances poses a number of special problems that do not occur in the kΩ ranges. Take, for instance, the measurement system, which in many cases has to be changed just for the sake of the lowest range. There is, however, a more serious problem in the range up to 10 Ω: the contact resistance of the test lead plugs and the sockets on the instrument, and, of course, the resistance of the test leads themselves. A connection formed by a banana plug and a mating socket, both in new condition, represents a typical resistance smaller than 1 mΩ. This resistance rises to several milliohms as the contact surfaces start to oxidize. Although a few mΩ may not seem much to start worrying about, such values are significant since the instrument discussed here has a resolution of 2 mΩ. The resistance of the test leads is also a factor of some importance. A test lead with a length of 1 m and a cross-sectional area of 1 mm² has a typical resistance of 17 mΩ. For a similar lead with a cross-sectional area of 2.5 mm², this value becomes 7 mΩ. Relating these values to 1 Ω, the error factors are 1.7% and 0.7% respectively. In other words, our measurement starts to become unreliable when these parasitic resistances are not taken into account. Fortunately, there exists a measurement principle that eliminates the effects of these unwanted resistances. This principle is called four-point resistance measurement.

Two terminals, four wires?

Using four wires to connect a resistor with only two terminals to a meter system may seem strange at first. However, since these wires may be divided into two pairs with the same functions, this method allows us to eliminate the effects of parasitic resistances. The principle is illustrated in Fig. 1. The unknown resistor, \( R \), is connected with four wires. The outer two cause a current flow through \( R \). The present meter sends a constant current through \( R \) via terminals 1+ and 1−. The advantage of using a constant-current source is that it is not affected by the parasitic resistance. Hence, we know exactly how much current flows through \( R \). To determine the value of \( R \), all we have to do is...
measure the voltage across it as a result of the constant current. This voltage is fed to the instrument via wires +Rx and −Rx. These wires are connected as close as possible to the resistor body, or to the terminals to which a resistor is to be connected later. In this way, only the voltage drop across the resistor is measured, without the additional voltage across all kinds of parasitic resistances. The system also eliminates the resistance of the test leads, and the contact resistance at the plugs and sockets.

Since the current flow into the voltage meter is negligible with respect to the constant current sent through the resistor under test, it may be concluded at this point that the four-point resistance measurement offers a reliable method of determining the value of small resistors at an accuracy that is not normally achievable with a multimeter.

1 A, and no heat?

Good as the four-point measurement system may be as a basis for the design of a milliohmmeter, there are more aspects to such an instrument that need to be given thought. Among these factors is the heat dissipated by the resistor. To make sure that a low-value resistor produces a voltage drop that is readily measured, it must pass a relatively high current. We can not make the current as high as we wish, however, since the maximum permissible dissipation of the resistor must be taken into account. A 1-Ω resistor with a power rating of 0.25 W, for instance, will not survive the constant current of 1 A supplied by the instrument. The solution to this problem is found in the use of a pulsed constant-current source (see the block diagram in Fig. 2). The resistor under test is fed with an effective current of only 10 mA since the 1-A current source is pulsed at a duty factor of 0.01 (1 ms on, 100 ms off). Even a 0.25-W resistor will not mind such a low effective current. Unfortunately, the use of a pulsed test current has one disadvantage in that resistors with a relatively high reactive component (stray inductance or capacitance) can not be measured reliably.

The test current through the resistor is pulse-shaped because the constant-current source is switched on and off by a pulse generator. The same generator controls a sample-and-hold circuit that stores the measured voltage during the ‘off’ period of the current. This means that the output of the sample-and-hold supplies a constant voltage whose value is in direct proportion to the measured resistance. Depending on the selected range, this voltage is amplified or attenuated before it is fed to a moving-coil meter provided with an ohm scale.

The circuit helps you avoid measurement errors by signalling over-range conditions. This is achieved by monitoring the output current of the current source. When a too large resistor is connected, or when the current wires, +1 and −1, are broken, the current source will no longer be able to supply 1 A, so that the voltage measured across the resistor is no longer a direct measure for the resistance value. However, the meter will still indicate ‘something’ because the measurement circuit and the resistor supply are separate circuits. The fault condition is simple to recognize because the current source then pulls terminal 1+ to ground. A detector circuit that measures the voltage between the input terminal and ground is all that is required to signal over-range conditions. When these occur, the detector causes the error LED to light.

Circuit description

Having explained the principle of operation of the milliohmmeter, we can now look at the way the circuit is realized in practice. Figure 3 shows the circuit diagram of the instrument. The pulse generator is built around opamp IC9. Resistors R1, R2 and R3 cause the opamp to function as a Schmitt-trigger inverter, while components R4, R5, D1 and C1 provide the function of an oscillating pulse generator. The operation of the generator is as follows: when the output of IC9 is high, capacitor C1 is charged via diode D1 and resistor R3, until the voltage across it reaches the upper switching threshold of the Schmitt-trigger. This takes about 1 ms. Next, the output of IC9 goes low, so that C1 is discharged to the lower switching threshold. This takes about 100 μs. The output of the opamp goes high again, and the cycle is repeated. Transistor T1 inverts the output signal of the pulse generator.

The current source in the instrument is built around opamp IC1. This provides a drive signal to transistor T2 that results in a voltage across emitter resistor R5 equal to the voltage at the +input of the opamp. When this voltage is constant, the emitter current is constant too. Since there is a fixed relation between the emitter current and the collector current of T2, it follows that the collector current is also constant. The magnitude of the collector current (which is the test current through the unknown resistor) depends on the value of R5 and the voltage at the +input of IC1. That voltage is supplied by preset Pa, and is stabilized by a precision zener diode, D2. The zener diode is powered by the pulse generator. As a result, the voltage set by Pa at the +input of IC1 will vary between nought and the set peak value. Hence, the test current will also vary between nought and the set peak value of 1 A.

The current sent through the resistor under test can not be drawn direct from voltage regulator IC5 because the peak value (1 A) is about equal to the maximum current the 7810 is capable of supplying. However, since the peak current has a relatively short ‘on’ time, the necessary energy may be obtained from a large electrolytic capacitor, in this case, C5. It will be clear that the voltage across this capacitor is far from constant. This is of little consequence, however, since these variations are compensated by the current source. Resistor R4 between C5 and the voltage regulator keeps the charge current within limits. The relatively long ‘off’ time of the current pulses ensures sufficient time for the capacitor to be charged via this resistor.

The test current sent through the unknown resistor via terminal 1+ gives rise to a voltage which is fed to the sample-and-hold circuit via the Rx terminals. The sample-and-hold stores the measured voltage during the

---

**Fig. 2.** Block diagram of the milliohmmeter. The resistor to be measured, Rx, is connected into a four-point network that supplies constant current pulses, and feeds the voltage developed across Rx to a sample-and-hold meter circuit.
'off' time of the test current. In addition, it converts the voltage from floating into one that can be measured with respect to ground. Four CMOS bilateral switches are used to achieve this. When the current source is on, switches IC1a and IC1c are closed, while IC1b and IC1d are open. Capacitor C3 is connected in parallel with RX via resistors R8 and R9, and will be charged until the voltage across it equals that across RX. The resistors and C3 form a low-pass filter to suppress interference. The moment the current source is switched off, switches IC1b and IC1d are closed. This results in C3 being connected to ground via IC1d. The switching can be done without the risk of a short-circuit occurring, because the connection with the floating voltage across RX is broken. Next, the voltage across C3 is fed to C4. This capacitor ensures that the measurement amplifier, IC3, is provided with an input voltage during the time C3 is connected to RX.

Switch S1 selects between an amplification of one and an amplification of 10, for opamp IC3. These amplification factors are used for the ranges 1Ω, 2Ω and 5Ω (×1), and 100 mΩ, 200 mΩ and 500 mΩ (×10). The offset of IC3 is compensated by adjusting P2. The attenuator circuit that follows IC3 consists of a number of switchable potential dividers that drive moving-coil meter M1. The use of 1Ω resistors in the attenuator obviates any adjustments. The attenuator is followed by the moving-coil meter with its series resistors R21-P4.

The over-range detector is formed by comparator IC2b. Resistors R27 and R28 define the switching threshold of this comparator at about 3.3 V. The comparator compares this reference level to the voltage across capacitor C1, which is charged via R26 and can only be discharged when the current source is off. Then, the minimum voltage across C1 is about 0.6 V higher than the collector voltage of T2. When this voltage drops below 2.7 V as a result of a too high resistance between the + and the − terminals, the voltage across C1 drops below the switching threshold of the comparator. Consequently, this toggles, so that LED D1 lights. Calculating the resistance value at which this happens, we find a value of about 7 Ω between these terminals.

**Fig. 3:** Circuit diagram of the milliohmmeter. The instrument is powered by an external mains adapter with a 15 VDC, 100 mA output.
Construction

When the PCB shown in Fig. 4 is to be fitted into the enclosure mentioned in the parts list, the corner near ICI will have to be cut off. Next, fit the parts on to the PCB, starting with the three wire links. Zener diode D2 comes in two different enclosures: a metal type and a plastic type. If you have a metal version, pay attention to the correct polarisation (see Fig. 6). The plastic version presents no problems since its orientation is printed on the component overlay.

As with previous test instruments in this series (see the list at the end of this article), the milliohmeter is powered by a mains adapter. In this case, an adapter with a rating of 15 VDC at about 100 mA is recommended.

The prototype of the milliohmeter is shown in Fig. 7. The completed PCB is fitted vertically at a suitable distance behind the front panel. Use short pieces of solid wire to connect the banana sockets to the relevant points on the PCB. The range selection switch is a type for PCB-mounting that obviates any wiring. The front panel is not fitted as yet.

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**COMPONENTS LIST**

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<tr>
<td></td>
<td>front-panel foil 910004-F</td>
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**Fig. 4.** Single-sided printed-circuit board for the milliohmeter. Note that the range switches are fitted direct onto the PCB.

**Fig. 5.** Completed circuit board, ready for fitting into the enclosure. Note that the left-hand bottom corner of the PCB is cut off diagonally.

**Fig. 6.** The LM336-2.5 precision zener diode comes in two different enclosures.
Adjustment

To adjust the instrument you require two 1% resistors: one of 1 Ω and one of 0.5 Ω (preferred value) or smaller. Where these resistors are not available, two pieces of 0.5-Ω/m resistance wire may also be used with good results. The 1-Ω resistor then has a length of 2 m, and the 0.5-Ω resistor a length of 1 m. In the first case, an error of 1 cm corresponds to a resistance error of 0.5% — in the second case, to a resistance error of 1%. Resistance wire with a different specification may also be used, although the required values of 1 Ω and 0.5 Ω will be a little more difficult to calculate.

The indicated length of the resistance wire applies to where it is connected to the +Rx and -Rx terminals. This means that the wires must be made slightly longer than 2 m or 1 m to allow the ends to be connected to terminals + and -. Having prepared the calibration resistors, put them aside for the moment.

First, null the moving-coil meter mechanically by adjusting the screw on the front. Switch on the instrument, and turn the range switch to select the 100-mΩ range. Connect the +Rs and -Rs terminals, and adjust P2 for maximum meter deflection. Next, re-adjust P2 until the meter just indicates zero. Do not turn P1 any further, since this may cause an unwanted, negative, off-set. Remove the connection between the test terminals. The meter may start to deflect slowly. This is no cause for alarm, however, since it indicates that C1 is charged by the input off-set current. This effect disappears as soon as a resistor is connected to the Rx terminals.

Next, P1 must be adjusted. If you do not have access to an oscilloscope, set the preset to the centre of its travel (this does not affect the accuracy of the instrument). If you do have an oscilloscope, connect the 1-Ω resistor between the terminals of the instrument. Do not connect the resistor to the Rx terminals as yet. Connect the oscilloscope as close as possible to the resistor body, or, when you use resistance wire, at the distance you have previously calculated to produce a resistance of 1 Ω. Adjust P4 until the peak value of the measured voltage is 1 V. This sets a peak current of 1 A. Remove the scope connections, and connect the 1-Ω resistor to the Rx terminals. Switch to the 1-Ω range, and adjust P3 for full-scale deflection of the meter.

Finally, connect the 0.5-Ω resistor, and switch the instrument to the 0.5-Ω range. Adjust P1 until the meter indicates 0.5 Ω.

This concludes the adjustment of the milliohmeter. At this point, you may fit the front panel, and apply the ready-made two-colour self-adhesive foil that gives the instrument a professional look.

Other test instruments in this series are:

- Q meter. Elektor Electronics April 1990.
- 400-W laboratory power supply. Elektor Electronics October 1990 and November 1990.

Fig. 7. Internal view of the instrument.

Fig. 8. Front-panel designed for the milliohmeter. For technical reasons, the meter scale is reproduced in black here, although it is really white. The scale can be cut out of the self-adhesive foil, to replace the one that comes with the moving-coil meter.
After the brief discussion on measuring, errors and tolerances in Part 1, we now turn our attention to practical measurements, more particularly the measurement of voltages.

Measurement of direct voltages

Even measuring a direct voltage is not the straightforward job it is often assumed to be. This may be because of the level of the voltage: very low voltages lie under the noise level and their measurement requires special equipment and techniques, whereas very high voltages require the use of an external prescaler, such as a capacitive voltage divider. But even well away from these extremes there exists the danger that the result will be distorted by the internal resistance of the measuring instrument.

An ideal voltmeter has an infinite internal resistance, but in practice that is, of course, unattainable. Perhaps that is just as well, because a very large resistance produces a high noise voltage and this will affect the measurement. In practice, the internal resistance of the instrument should be appreciably higher than the resistance across which the voltage is being measured, but it should not approach infinity.

Depending on the nature of the measurement, there are two types of voltmeter on the market: those whose internal resistance depends on the selected meter range and those whose internal resistance is constant (normally between 1 MΩ and 10 MΩ).

The first kind includes most low-priced multimeters, a typical example of which is shown in Fig. 8. One of the quality criteria of these Instruments is their characteristic resistance, which is expressed in ohms per volt. The internal resistance, Rm, is calculated by dividing the characteristic resistance (Ω) by the full-scale deflection (V) of the relevant meter range.

The inverse value of the characteristic resistance gives the current that flows through the network at full-scale deflection.

For instance, if the characteristic resistance of the instrument is 100 kΩ/V, its internal resistance over the 1-V range is 100 kΩ, over the 3-V range, 300 kΩ, and over the 5-V range, 500 kΩ.

The input amplifier stage of a typical multimeter shown in Fig. 9 is a grounded-emitter circuit with current feedback. Were the voltage measured across R2, there would be a problem because, if, correctly, the 3-V range is selected, the internal resistance of the instrument is 300 kΩ. Unfortunately, this resistance is in parallel with R2, so that the ratio of divider R1-R2 changes according to the selected range and this will, of course, give rise to incorrect measurements. In such a case, it may, therefore, be better to select a higher range. True, the error will then be larger, but so will the internal resistance and this makes the effect on the divide ratio smaller. None the less, even then precise measurements are not possible.

From the above, it is clear that measurements in high-resistance circuits, such as opamp inputs and base and gate inputs of transistors and FETs respectively, require instruments with a high internal resistance.

It is interesting to calculate how much greater the measurement error is when a 20 kΩ/V instrument is used instead of, say, a 50 kΩ/V one. Once you know the problem, it is quite possible to use a low-priced multimeter for most voltage measurements and guessimate the error. However, in the long run, that is not a satisfactory solution to the problem. Fortunately, manufacturers are aware of this and modern instruments have a much higher internal resistance than their predecessors. This is achieved on the one hand by far more sensitive meters and on the other hand by the use of, for instance, impedance converters (simulating the valve voltmeters of yesteryear).

Instruments with input amplifiers

Instruments with input amplifiers generally have a high input resistance, at least 1 MΩ, and this value is constant, i.e., independent of the metering range. Such instruments are much better suited for use in high-resistance circuits. None the less, even they have their limitations. For instance, in circuits using FETs or electronic valves, an internal resistance of even 1 MΩ can cause errors.

To understand the function of an instrument with input amplifier, consider the circuit in Fig. 10. This shows the layout of an electronic voltmeter, which may actually be
constructed by any electronics enthusiast. Because of the transistor amplifier, the branches of the input divider have a very high resistance. For instance, that for the 1-V range is 500 kΩ, which is equal to the characteristic resistance of the instrument.

A disadvantage of this type of circuit is the temperature dependence of the quiescent collector current. To counter this current, an equal current of opposite polarity, derived from an auxiliary battery, is passed through the instrument. This current is limited by the series network consisting of a 10 kΩ resistor and a potentiometer. Prior to each measurement, the potentiometer must be set to ensure zero reading of the meter.

Figure 11 shows the circuit of a commercial impedance converter for multimeters, which may be used in virtually any kind of multimeter. In use, the instrument must be set to the most sensitive current range. Calibration is effected with R6. Zero setting is accomplished with R8.

Multimeters for industrial use have rather more complex circuits than that in Fig. 11, but they are not necessarily any more exact. Here, as almost everywhere, you get what you pay for: if you want good accuracy, you have to pay a good price.

Figure 12 shows a popular analogue multimeter with integral input amplifier. Its input resistance is 10 MΩ and is independent of the selected metering range.

Digital multimeters generally also have a high input resistance (up to 10 MΩ), but the measurand must additionally be translated by an analogue/digital converter. Their accuracy is therefore, dependent on the accuracy of the converter.

As mentioned before, the last digit of the readout of a digital multimeter is error-prone and should therefore not be taken into account where great precision is required.

**Measurement of alternating voltages**

What has been said about the internal resistance of measuring instruments for direct voltage is equally applicable to those for measuring alternating voltages. There are some additional difficulties as well. For instance, in most instruments, the internal resistance for alternating voltages is lower than that for direct voltages, and is typically 20 kΩ/V and 5 kΩ/V. This is because for alternating voltages to be measured by moving coil meters, it is necessary for them to be rectified. Now, every rectifier diode has a fairly significant threshold voltage and for that reason the full-scale deflection on the lowest metering range cannot be very small.

To reduce the effect of the threshold voltage, it is normal to use diodes in only one section of the bridge rectifier and resistors in the other section—see Fig. 13. This arrangement leads to an additional current through the resistors and this lowers the internal resistance and also the sensitivity. When the instrument has an input amplifier these aspects are of no consequence, because the impedance converter at the input isolates the measurand from the meter section.

There are two other problems in measuring alternating voltages: (1) the r.m.s. value is shown only if the measurand is sinusoidal (and 2) the instrument does not function properly at fair high frequencies.

The first problem is not so bad when a moving coil meter is used, since in that case the arithmetic mean value of the rectified voltage will be calculated and indicated. In digital instruments, the reading is not reliable if the measurand is not sinusoidal.

The second problem is again not too serious in analogue instruments, since the frequency range of them is generally considerably higher than that of digital instruments. Many digital instruments have an upper frequency range as low as 400 Hz, so that even measuring audio signals with these becomes problematic. To add to the problems, there is no indication of the frequency range on many low-priced digital multimeters; that information is normally hidden in the small print in the specification contained in the operating manual. Figure 14 shows the dramatically different frequency ranges of two digital multimeters. One may be used up to 100 kHz whereas the other becomes unreliable above 1 kHz.

(to be continued)
While setting up and connecting audio equipment it is important to have all the units — microphone, loudspeakers and everything in between — 'in phase', that is, interconnected with the right polarity. The low-cost instrument described here is particularly handy for checking out the phase of almost any audio system, whether installed in a living room, in a car, in a studio, or on a stage.

Reversed phase connections in an audio equipment system give strange and unpredictable effects such as the unwanted attenuation or boosting of a particular frequency range, jet-plane effects, whistling noises, or amplifier output power which does not seem to produce any usable sound level. To avoid these problems, use the simple instrument described here. Based on a transmitter and a receiver with a simple good/fault indication, the instrument will check out the system from the input (microphone or line input) right through to the output (loudspeaker or line output).

The transmitter supplies positive or negative needle pulses, which are fed either electrically to an equipment input, via the line-cinch output socket, or acoustically to a microphone, via the built-in loudspeaker. Accordingly, the receiver has an electrical (line) input and an acoustic (microphone) input.

The drawings in Fig. 1 illustrate two ways of using the transmitter and the receiver for phase tests on audio equipment.

Figure 1a shows the set-up used to check the polarity of a microphone, and Fig. 1b that used to ensure a loudspeaker is connected the right way around. The LEDs on the receiver provide a quick indication whether or not the received pulses have the same polar-

**K. Orlowski**

---

**Fig. 1.** Application examples of the phase-check system.
ity as the transmitted pulses. If the receiver indicates the opposite polarity of the transmitter, the chances are pretty high that there is a reversed signal connection somewhere in the system.

The pulse transmitter

The needle pulses are generated by oscillator IC1A (see Fig. 2), which is built from a NAND gate with two Schmitt-trigger inputs. After applying the supply voltage, these inputs take on complementary logic levels, i.e., one is high, the other is low. Consequently, the output of the gate is logic high. Capacitor C2 is charged via resistor R1, until the voltage on it reaches the high threshold voltage of about 5.5 V. Next, the output of the Schmitt-trigger toggles to 0, so that C2 is discharged via D1 and R2, until the low threshold voltage of about 3 V is reached. The NAND gate toggles, and the charging of C2 starts again.

The above process is cyclical and results in a self-oscillating circuit. Since R2 is much smaller than R1, the discharge time of C2 is much shorter than the charge time. As a result, the on-off (mark-space) ratio of the output signal is about 2 ms/1 s, or 0.002. Mind you, 'off' means 'logic high' here since we are dealing with a NAND gate.

The oscillator output signal is fed to two sub-circuits. One is a small loudspeaker driver based on emitter follower T2. The loudspeaker connections can be swapped by switch contacts S1c and S1d. When an oscilloscope is connected to the loudspeaker, it indicates negative-going needle pulses with the switch set to the centre position, and positive-going pulses with the switch set to the upper position. Likewise, in the other signal branch, the polarity is changed by switching transistor T1 from a common-emitter circuit (S1c at centre position) to a common-collector circuit (S1d at centre position). Coupling capacitor C3 takes the test signal to an attenuator that supplies output levels of 1 Vpp (0 dBV), -20 dBV and -10 dBV.

The receiver

The circuit diagram of the receiver (Fig. 3) shows that two almost identical detectors are used. The test signal is supplied to the two voltage amplifiers T1-T2 and T3-T5 either by the electret microphone, or by the signal source connected to K1. In the latter case, the signal is taken through a high-pass filter, R1-C3, before it arrives at a voltage limiter, D1-D2. The input source, microphone or line, is selected with switch S1. The voltage amplifiers are complementary circuits; T1-T2 amplifies the negative pulses, T4-T5 the positive pulses.

The two monostables in IC1 have different networks at their trigger inputs to enable them to respond to negative pulse edges (IC1A) or positive pulse edges (IC1B). To prevent the trailing edge of a pulse triggering the wrong monostable, IC1A and IC1B disable one another when one of them is activated. The monostables thus allow the circuit to determine whether a pulse starts with a

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Fig. 2. Circuit diagram of the pulse transmitter.

Fig. 3. Circuit diagram of the pulse receiver. The polarity of the measured signal is indicated by two LEDs, D3 and D4.
**COMPONENTS LIST**

### TRANSMITTER

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<td>12-way 1-pole rotary switch for PCB mounting</td>
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**RECEIVER**

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<tr>
<td>6</td>
<td>ABS enclosure, e.g., OKW A9409126</td>
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</tbody>
</table>

**Miscellaneous:**

| 1 | electret microphone | Mic1 |
| 2 | phono socket | K1 |
| 3 | miniature SPDT switch | S1 |
| 4 | miniature SPST switch | S2 |
| 5 | clip for 9V PP3 battery |
| 6 | ABS enclosure, e.g., OKW A9409126 |
| 7 | printed-circuit board 900114-2 |

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Fig. 4a. Single-sided printed circuit board for the pulse transmitter.

Fig. 5. A look inside the completed pulse transmitter.
positive (rising) or a negative (falling) edge. The two LEDs, D3 and D4, indicate the respective polarities. The monostable times are set at about 0.5 s with R4-C6 and R17-C16. This causes the active LED to flicker.

Building and testing

The receiver and the transmitter are best built on the printed-circuit boards shown in Fig. 4. Be sure to fit all polarized components (electrolytic capacitors, ICs, transistors and diodes) the right way around. Also make sure that the two rotary switches on the transmitter PCB are fitted as shown on the overlay (note the '1' mark, and the letters that indicate the poles). On completion of the two units, apply the self-adhesive foils shown in Fig. 6 to the enclosure front panels.

Interconnect the transmitter and the receiver via their line sockets, and check that the LED indication on the receiver is in accordance with the polarity set on the transmitter. When the LEDs remain off, IC1 in the receiver may not have sufficient gain. In that case, adjust P1 and P2 until the receiver does trigger correctly.

Fig. 4b. Single-sided printed-circuit board for the pulse receiver.

Fig. 6. Design of the self-adhesive front panel foil for the transmitter (left) and the receiver (right).
TRANSMISSION LINES

by Roy C. Whitehead, C.Eng., MIEE

Transmission lines may be used both for the direct transmission of information and as circuit elements, sometimes substituted for such components as transformers, capacitors and inductors.

The two main types of metal line are the balanced and the coaxial types as shown in Fig. 1. The familiar pair of wires mounted on porcelain insulators, supported on wooden poles, and the twisted or parallel pairs embedded in solid insulation are shown at (a) and (b) respectively. Where several such pairs are run together, it is customary to employ a physical transposition process, so that mutual interference between pairs encountered along one length of line, is partially balanced out by reversed interference along another length. Such lines are normally operated in the 'balanced' condition, neither conductor being earthed, although sometimes the centre point of an associated transformer or amplifier may be earthed.

A coaxial line, with its central conductor insulated from its outer conductor is shown at (c). These lines are operated in the 'unbalanced' condition, that is, the outer conductor is earthed. The outer conductor does not always provide a very efficient screen at low frequencies, so in some circumstances signals are confined to the spectrum above 50 kHz.

A very important characteristic of any transmission line is its 'characteristic impedance' $Z_0$, which is the ratio $V/I$ for a line of infinite length as shown in Fig. 2. But, of course, there is no such thing as a line of infinite length. However, if a line of finite length be connected with a variable resistor $R_d$ to its remote or distal end, there will be one specific value of $R_d$ that produces a constant ratio $V/I$ for all frequencies and for all lengths of that particular type of line. It is around that particular value of $R_d$, that is, $Z_0$, that complete telecommunication systems are built, just as railway systems are built upon the 'gauge', or spacing, of the rails (which is 4 ft. 8 1/2 in. = 1435 mm in Britain and many other countries).

The two ends of a line are sometimes referred to as the 'proxal' or sending end and the 'distal' or receiving end. Subscripts $p$ and $d$ respectively will be used accordingly.

The characteristic impedance $Z_0$ of a line is governed by the ratio $D/d$ shown in Fig. 1 and the value of the permittivity, $k$, of the dielectric.

Simplified equivalents to balanced and unbalanced lines are shown in Fig. 3. For most practical purposes, the value of $Z_0$ may be taken as $Z_0 = \sqrt{L/C}$, where $L$ is measured with the distal end short-circuited and $C$ with it open-circuited. Details are given in the Appendix.

In Britain, open-wire lines and twisted pairs, singly or in multi-pair cables usually have $Z_0 = 600 \Omega$. Coaxial cables on the other
Hand usually have $Z_o = 75 \Omega$ or, at very high frequencies, $50 \Omega$.

The velocity of propagation in an ideal line that has vacuum insulation and no supports, would equal the velocity, $c$, of an electromagnetic wave in free space, that is $c = 3 \times 10^8$ metres/second. For a line with minimum supports and with air as insulation, the velocity is only slightly less. For a line with solid or gaseous insulation that has a permittivity $\kappa$, the velocity $v = c/\kappa$. The velocity $v/c$ of a line is known as the ‘velocity ratio’. This is usually quoted by the manufacturers: if $\kappa$ varies between 1.2 and 2.8, the value of $v/c$ lies between 0.9 and 0.6.

The type of insulation for concentric cables that is most commonly experienced in laboratories or small installations is polytetrafluoroethylene, normally called PTFE. This has the great merit of being flexible.

For high-power transmitters, where very high voltages are incurred, the insulation may be air, nitrogen under pressure, or helium. No, or very few, intermediate physical spacers may be incurred when transmission lines are installed vertically up masts.

The relationship between attenuation per unit length and frequency is given by the empirical equation: 

$$\text{attenuation} = \alpha + \beta f$$

where $\beta \ll \alpha$. Upto about 16 MHz, the second term may be ignored, but above that frequency the attenuation rises faster.

The increase in attenuation at high frequencies has two causes. The first is that the losses in the insulation rise with frequency. The second is the well-known ‘skin effect’ that takes place in conductors that operate at high frequencies. The higher the frequency, the less deep is the penetration of current into the conductor surface. For this reason, conductors that must carry high levels of current at very high frequencies usually take the form of tubes that have conductivities which are equal to those of solid conductors of similar diameter.

Complete transmission links, say between cities, are usually engineered to produce what is known as ‘zero equivalent’; that is, the combination of attenuation and amplification equals zero decibels. This is to enable communication to be established readily, either directly between two points, or indirectly via other points without change of amplitude of the received signal. To achieve this result, a complete link includes terminal amplifiers to counteract attenuation and ‘equalizers’ to counteract the variations of attenuation over the frequency band. An example of the various parts of a link is shown in Fig. 4, starting with LINE (1), EQUALIZER (2), and so on. The design of an equalizer starts with the design of an attenuator that has an attenuation which is slightly greater than the variation of attenuation of the line over the operating frequency range. A simple attenuator (for an unbalanced line) is shown in Fig. 5 (a). Reactive elements are then added to reduce attenuation at the higher frequencies as shown in Fig. 5 (b). This produces finally an attenuation/frequency characteristic of line plus equalizer that is approximately flat. The equalizer is located at
the receiving end of the line so that it will attenuate not only the lower frequency components of the signal, but also random noise and crosstalk that has been picked up along the line. Finally, a variable-gain amplifier is added to achieve the zero equivalent condition. Along a line there is a limit to the attenuation that can be tolerated between the two terminal amplifiers, otherwise the signal-to-noise ratio of the received signal would be unacceptable. Along a lengthy line, this effect is combated by the introduction, at various stages, of amplifiers that are referred to as 'repeaters'. The power that is required to operate these repeaters is sometimes fed along the signal line as shown in Fig. 6.

The introduction of terminal amplifiers and repeaters implies that such lines can be operated only unidirectionally, so that to enable a conversation to take place, two lines are required. In order that the high-level signal at one end of one line shall not interfere with the low-level signal of an adjacent line that is operating in the reverse direction, two groups of lines are formed physically with screening between them. Each group consists entirely of lines that operate in a given direction as is shown in Fig. 7.

When transmission lines are used for communication, it is usual to operate them between resistive terminations that are equal to the characteristic impedances $Z_0$ of the lines. This produces an attenuation/frequency curve that is smooth as was shown in Fig. 4, curve number 1, enabling a simple equalizer to be designed as was shown in Fig. 5.

It is customary for telecommunication authorities to specify the maximum amplitude of the signals that may be fed into the lines, which is to avoid overloading the amplifiers and also to minimize crosstalk between the various users. Thus, if all users feed into their lines signals of approximately the same magnitude, the overall signal-to-cross-talk ratios will be maximized.

A line that is terminated with a resistance $R_d = Z_0$ will (ignoring attenuation) have a distribution of voltage and current along its length as shown in Fig. 8 (a).

When electrical energy starts to travel down a line, it does so at a rate that is determined by the details of the generator, the velocity ratio of the line and the value of $Z_0$. This is said to constitute a 'travelling wave'. If the termination has a value of $R_d = Z_0$, and energy reaches the termination, a stable condition is established and electrical energy is converted into thermal energy at the same rate at which it was admitted to the line. If, however, $R_d$ does not equal $Z_0$, the termination can no longer dissipate energy at that rate, so information is communicated back to the source by a 'reflective wave' to reduce the rate at which energy is admitted. The final result is a combination of the two waves.

An example of how current and voltage are distributed as travelling waves along a line where $R_d = Z_0$, and there are no reflective waves, is shown in Fig. 8 (a). But if the terminations are an open circuit, and consequently no current can flow in it, the travelling and reflected current waves are in opposition, resulting in no current at the termination. But the voltage travelling and reflected waves are in phase, resulting in a doubling of voltage there as shown in Fig. 8 (b).

The reverse condition applies to a short-circuited line as shown in Fig. 8 (c).

A compromise condition, where $R_d$ is finite but does not equal $Z_0$, is shown in Fig. 8 (d). Because the magnitude of $U$ is rising at the approach to the load, it follows that $R_d$ is greater than $Z_0$.

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**Fig. 7. Signal levels in isolated groups arranged to minimize cross-talk.**

**Fig. 8. Distributions of voltage and current along lines having various terminations.**

Consider again Fig. 8 (d) and note that in comparison with Fig. 8 (a) the values of \( U \) and \( I \) vary cyclically along the line. Because the dielectric and conductor losses are proportional to the squares of \( U \) and \( I \) respectively, the increased losses around the peaks of the waveforms are not compensated completely by the decreased losses around the troughs. The ratios \( \frac{U_{\text{max}}}{U_{\text{min}}} \) and \( \frac{I_{\text{max}}}{I_{\text{min}}} \) are known as the 'standing-wave ratios'. Considering particularly cases of high-power transmitters feeding aerials, the excess amplitudes may cause breakdowns.

However, the small losses of radiated power are usually considered to be significant only in the case of very-high-power transmitters (see References). Standing-wave ratios may be measured using a commercial standing-wave ratio measuring instrument, but a simple test may be carried out with the aid of a neon tube mounted at one end of a stick of insulating material and running this along the line to explore the peaks and troughs.

Having considered the line as a device for the transmission of information, let us now consider it as a circuit element equivalent to, for instance, a capacitor, inductor or transformer.

Consider Fig. 8 (b) and (c). At any point along the axes where \( U \) is finite and \( I \) equals zero, the impedance looking towards the termination must equal infinity. Where \( U \) equals zero and \( I \) is finite, the impedance must equal zero. Where \( U \) is rising, energy is being converted from kinetic into potential form; the nature of the impedance must therefore be capacitive. But where \( I \) is rising, energy is being converted from potential into kinetic form and the nature of the impedance must be inductive. Of special interest are the distributions of voltage and current along quarter- and half-wavelength lines and the extraction of information from Fig. 8 (b) and (c); the distributions for open- and short-circuit conditions are given in Fig. 9.

Extending considerations to a wider variety of effective lengths, some examples with their equivalents are shown in Fig. 10. To a choice exists, it is preferable to use a line with a short-circuit rather than an open-circuit termination, since such a line is easier to fix mechanically. Figure 10 (9), which represents a line with a sliding short-circuiting bar, can be adjusted to present a wide variety of such equivalents.

Although the lines shown are balanced, similar results may be obtained with the use of unbalanced lines. In the case of Fig. 10 (9) using unbalanced lines, however, similar results require the provision of 'trombones', that is, devices that are similar to their musical equivalents in that they provide paths of variable lengths. It must be emphasized, however, that these effects hold good only over very narrow bands of frequency.

A particular use of Fig. 10 (9) is in transmitting stations that house many high-power transmitters which operate on different frequencies. Power that is radiated from one aerial might be picked up by another aerial and this might affect the operation of the second transmitter. This may be avoided by connecting a line of the type shown in Fig. 10 (9) across the output terminals of each transmitter, adjusting the bridge to produce a quarter-wave condition for that transmitter and earthing the short-circuiting bridge. This technique also provides protection against lightning strikes.

The effects of open- and short-circuit terminations have already been dealt with. Now it is necessary to consider the results of employing various other types of termination.

Since the magnitudes of the voltage and current at both ends of a \( \lambda/2 \) line are similar, it follows that if an impedance \( Z_0 \) be connected at one end of such a line, a similar impedance will appear at the other end, that is, the line acts as a 1:1 transformer. This will be so irrespective of the relationship between the values of the load and the characteristic impedance of the lines as is shown in...
The quarter-wave line has been seen to have an inverting effect. Therefore, if the terminating impedance $Z_d$ has a magnitude that equals $\pi Z_0$, the impedance that appears at the other end will have a magnitude $Z_0/\pi$ as is shown in Fig. 11 (b). But not only is the magnitude of the impedance inverted but so also is the sign, e.g., capacitive reactance is transformed into inductive reactance.

Two practical examples of $\lambda/4$ lines used as transformers are shown in Fig. 12. If the line and aerial impedances be $Z_L = 600 \Omega$ and $Z_a = 75 \Omega$, the matching impedance will be $Z_m = \sqrt{(600 \times 75)} = 212 \Omega$.

One of the equations in the Appendix gives for a balanced line:

$$D/d = \frac{1}{2} \text{antilog} \left( \frac{Z_0 \lambda}{276} \right) = \frac{1}{2} \text{antilog} \left( \frac{212}{276} \right) = \frac{1}{2} \text{antilog} 0.768 = 2.93$$

That is, the spacing/diameter ratio will be 2.93 to provide the transformation. The line conductors may then be squeezed closer together just under three times their diameter as shown in Fig. 12 (a). This, however, might bring them dangerously close together. The alternative would be to maintain the same spacing for the transformer as for the line and to construct the transformer from tubes that have diameters of one third of the 600 $\Omega$ line spacing as shown in Fig. 12 (b).

It is interesting to consider whether the presence of standing waves could present a problem on power supply transmission lines operating at 50 Hz. The worst possible case would be where the line represented the $\lambda/4$ condition:

$$\lambda/4 = \frac{V}{Af} \text{ [metres]}$$

and allowing for a velocity ratio of 0.8:

$$\lambda/4 = \frac{3 \times 10^5 \times 0.8}{(4 \times 50)} \text{ [metres]} = 1200 \text{ km}.$$ which is the distance from, say, London to Madrid, Venice or Oslo. Therefore, even with 60 Hz mains supplies, there is little or no trouble likely to be experienced even with very long lines.

**Appendix**

Consider the line formed by a drum of cable or two identical lines that are looped at the distal end, represented by inductors and capacitors as shown in Fig. 13. The two variable resistors are ganged and always have equal values of resistance. The generator voltage is kept fixed and its frequency is varied. $R$ is then varied to produce a constant value of $U$. Then, $Z_0 = R$.

The kinetic energy stored in the inductors will be $\frac{1}{2} L I^2 = \frac{1}{2} L V^2/R^2$, which is dependent on the value of $R$.

Equal energy will be stored in capacitances and inductances when the value of $R$
is such that
\[ V^2 = \frac{1}{2}CV^2 = \frac{1}{2}L I^2/R^2, \]

\[ C = L/R^2; \]

and \( R = \sqrt{L/C} \).

This value of \( R \) is termed the ‘characteristic impedance’ and is given the symbol \( Z_0 \).

A fuller equation, taking into account the series resistance \( R \) of the conductors and the shunt leakance \( G \) (in siemens) of the insulation, is

\[ R = j \omega L + j \omega C + \frac{G}{\omega} \]

where \( j = \sqrt{-1} \) and \( \omega = 2\pi f \).

However, as \( R \ll \omega L \) except at very low frequencies, and \( G \ll \omega C \) except at very high frequencies, the simpler equation is normally accepted as adequate for practical purposes.

It is also possible to determine the value of \( Z_0 \) from the physical construction of the line by the use of one of the following two equations.

For a twin-wire line:

\[ Z_0 = \frac{138 \times \frac{2}{\pi} \log 2D}{d} \]

or

\[ D = \frac{1}{2} \frac{\log \left( \frac{Z_0 \sqrt{k}}{276} \right)} {d} \]

where \( D \) is the distance between the centres of the conductors, \( d \) is the wire diameter, and \( k \) is the permittivity of the insulation (= unity for air).

For an equipotential line:

\[ Z_0 = 138 / k \times \log (DA/d) \]

or

\[ D = \frac{1}{2} \frac{\log \left( \frac{Z_0 \sqrt{k}}{138} \right)} {d} \]

Fig. 14. Balanced lines with air dielectrics: divide \( Z_0 \) by \( \sqrt{k} \) for other dielectrics.

Fig. 15. Coaxial lines with air dielectrics: divide \( Z_0 \) by \( \sqrt{k} \) for other dielectrics.

### HIGH DYNAMIC RANGE S-MMIC AMPLIFIERS FOR 50 OR 75Ω SYSTEMS

Designated MSA-1110, the MMIC (monolithic microwave integrated circuit) is a general-purpose, cascadable gain block intended for use in broadband IF and RF amplifier design, and provides high performance in both 50Ω and 75Ω microstrip systems. This makes the device ideal both for communications and test equipment for the 50Ω transmission line environment, and for television and other equipment using 75Ω impedance standards. The MSA-1110 is offered in a 100 mil-diameter gold-ceramic surface-mount package suitable for industrial purposes.

Avantek's MODAMPTM MSA series of silicon bipolar amplifiers is fabricated using a 10 GHz fr. 25 GHz \( f_{max} \) silicon bipolar MMIC process based on nitride self-alignment, ion implantation and gold metallization to achieve excellent uniformity, performance and reliability. The availability of this wide-ranging family of general-purpose MMIC die gives the designer of hybrid circuit assemblies a choice of an easy-to-apply, stable and reliable gain block for almost any application ranging from intermediate frequency through microwave. MMIC models in this series are: MSA-0100, MSA-0200, MSA-0300, MSA-0400, MSA-0600, MSA-0700 and MSA-0800. At 1 GHz, this family offers gains as high as 22.5 dB, noise figures as low as 3.2 dB, and power outputs of up to +12.5 dB.

For more information, including datasheets, contact any authorized Avantek sales office, representative or distributor, or contact Avantek, Inc. • 3175 Bowers Ave. • Santa Clara • CA 95054-3292. Telephone: (408) 970-2659.

References:


Amateur Radio Technique by P. Hawker, RSGB.

HF Antennas by L.A. Moxon, RSGB.


VHF/UHF Manual, RSGB.
CHOPPER-STABILIZED OPERATIONAL AMPLIFIERS

Chopper-stabilized opamps are in many cases the only feasible alternative when we want to amplify very small direct voltages. In this article we will explore why chopper opamps have such excellent d.c. characteristics. A theoretical background to the operation of interesting new devices is given, followed by a discussion of some inherent problems (and, of course, proposed solutions). The article is closed off with an overview of the most popular chopper opamps currently available.

by J. Ruffell, with contributions from B. Marshall (Texas Instruments) and G.J. van Os (Acal Automation)

FOR a long time to come, instrumentation amplifiers will be required to operate at the highest possible accuracy. This expectation is based on the trend towards ever higher resolution of DACs (digital-to-analogue converters) and ADCs (analogue-to-digital converters). It will be clear that high resolution in a measurement is not achieved just by the use of converters with a high resolution. After all, it makes little sense to perform a measurement at an accuracy of 18 bits when the analogue amplifier used has a maximum resolution of, say, 16 bits. In practice, the accuracy of the hardware for analogue signal conditioning must be doubled for every additional bit to be measured.

Analogue signals are preferably conditioned and/or amplified by a.c.-coupled circuits, mainly because these can be built by relatively simple means and at low cost. There are, however, many applications where the wanted signal is applied in the form of a direct voltage or a direct current. Devices used in such applications include thermocouples, photodiodes and, on a larger scale, the digital multimeter, which is an example of a data acquisition system. Since these devices and circuits can only be d.c.-coupled, the designer is faced with offset voltages and drift of the linear amplifier he intends to use. The origins of input offset voltages and their stability is discussed in an earlier article on new opamps, see Ref. 1.

Although conventional operational amplifiers such as the OP07 and the OP77 are good choices for d.c. signal conditioning, there are devices whose extremely low drift and offset voltages make them far better suited to the application. The type of operational amplifier we have in mind is generally referred to as a chopper opamp, or, more accurately, a chopper-stabilized opamp.

Chopping: the classic approach

During the valve era, the terms chopper amplifier and indirect d.c. amplifier were familiar to almost anybody in the field of electronics. At that time, chopping was taken very literally. A kind of electronic guillotine was used to convert the low-frequency alternating voltage (or the direct voltage) to be amplified, into a signal with a higher frequency. Next, this higher-frequency signal was raised in an a.c.-coupled amplifier, and subsequently restored to its original frequency by a synchronous detector. In practice, the chopping element used to be a relay or, a little later, a bipolar transistor or a FET.

Figures 1a and 1b show the basic schematic of a classic chopper amplifier and the associated waveforms. The input voltage, $V_i$, is converted to a pulsating waveform, $V_o$, by switch $S_1$. The d.c. component is removed before $V_o$ is amplified by a.c.-coupled amplifier $A_1$. It will be clear that the original waveform (with a higher amplitude) must be recovered from $V_o$. The recovering, or demodulation, of $V_o$ is effected by switch $S_2$. This electronically operated switch connects the right-hand side of capacitor $C$ to ground on every second half cycle of the oscillator signal. The waveform of $U_0$ indicates that the switching results in a shift of the direct voltage level. Finally, an integrating filter recovers the amplified voltage, $U_0$, from $V_o$.

Although this type of amplifier allows good drift specifications to be achieved, it suffers from a number of inherent shortcomings. The chopper, for instance, often introduces glitches at the output. Also, the amplifier lacks a differential output, while its bandwidth is limited to a few hundred hertz.

Integrated

Modern chopper opamps no longer work as described above. These days, the signal to be amplified is no longer chopped to pieces and then rebuilt. Instead, use is made of a control loop which compensates the input offset-voltage of a normal differential amplifier. As a result, these new circuits look quite similar to the standard opamps you have grown accustomed to in many circuits in this magazine.

Chopper opamps, like standard opamps, have a differential input circuit. Because of this likeness, and because their principle of operation is based on the old chopper model, the new devices are generally called chopp-
Automatic off-set compensation

The off-set compensation control applied with chopper opamps is in many ways similar to a technique used to compensate the input off-set voltage, $U_{os}$, of a standard opamp. This technique entails off-set compensation by fitting a voltage source that supplies $-U_{os}$ in series with the non-inverting input of the opamp (see Figs. 3 and 4). Automatic input off-set voltage compensation thus requires a circuit capable of measuring $U_{os}$ and supplying an accurate 'negative copy' $-U_{os}$ at the non-inverting input.

You may start wondering at this point how $U_{os}$ can be measured when the opamp is already part of an existing circuit. Assuming that a simple electronic circuit is used, it can be shown that the input off-set voltage is best measured between the input terminals of the opamp in question. Figure 5 shows how this is done in an inverting amplifier set up around the ideal opamp model. Equation 1 describes the voltage between the non-inverting and the inverting input of the opamp. True, the equation looks fairly complex. However, assuming for the moment that $U_{in}$ does not contain an alternating voltage component, you will easily discover that the expression in equation 1 is virtually equal to $-U_{os}$. This is because the open-loop gain, $A_{ol}$, is high (say, 100,000), so that $E$ (see equation 2) approaches 1. The upshot is that equation 1 can be simplified to give equation 3. The output voltage is approximated as described by equation 4.

The schematic in Fig. 6 shows a circuit designed on the basis of the above discussion. An auxiliary amplifier is used to measure and compensate the input off-set voltage of the main opamp. Equation 5, which describes the output voltage, indicates that the effect of the input off-set voltage is reduced by a factor of 1-$E$. Assuming an open-loop gain of 100,000, and $R_1 = R_2$, the reduction amounts to no less than 50,000 times. Compared to the off-set error of about $2U_{os}$ in the output signal of the circuit in Fig. 5, a specification of the order of 1/25,000$U_{os}$ is quite impressive for the circuit in Fig. 6. Thus, equation 6 may be applied with confidence for d.c. applications.

It should be noted that the off-set of the opamp can only be compensated successfully if the auxiliary amplifier is sufficiently compensated. This is why we have shown the auxiliary amplifier as an ideal device, i.e., an opamp without input off-set. It will be clear that such a device does not exist. And yet, the circuit can be extended in a way that does allow automatic off-set compensation to be achieved. Basically, the auxiliary am-
Operational amplifier model with input off-set voltage $U_{os}$.

The input off-set voltage may be compensated by placing a voltage source $-U_{os}$ in series with the non-inverting input.

Off-set compensation thus consists of two successive phases. During the first phase, the electronic switch $S_1$ is set to position A. This causes the inputs of the auxiliary opamp to be short-circuited, so that the output voltage of this amplifier is virtually equal to its own input off-set voltage, $U_{os1}$. Just before $S_1$ switches to position B, a sample-and-hold circuit, S&H-1, connects $U_{os1}$ in series with the inverting input of the auxiliary amplifier. This results in compensation of the off-set error of this amplifier at the start of the second phase. During the second phase, $S_1$ connects the positive input of the auxiliary amplifier to the positive input of the main opamp. This, in fact, creates the circuit in Fig. 6. The sample-and-hold circuit still compensates the off-set of the auxiliary amplifier, whose output is at a potential of practically $-U_{os2}$. To retain this voltage, a second sample-and-hold, S&H-2, is introduced. As shown in Fig. 7, this causes $-U_{os2}$ to be connected in series with the non-inverting input of the main opamp. At least in theory, the result is as may be expected: the input off-set voltage is automatically compensated.

The off-set compensation of the two amplifiers may be optimized by repeating the two phases. Depending on the repeat rate, input off-set drift as a result of temperature changes or supply voltage fluctuations may be eliminated. Preventing these factors from affecting the stability of the instrumentation amplifier.

Main amps and null amps

The above information will, no doubt, enable you to take a well-prepared look at the block schematic diagram of a chopper-stabilized opamp. The functional diagram used by most manufacturers is shown in Fig. 8. The term main amp refers to the main operational amplifier, while the term null amp is meant to identify the auxiliary amplifier. The switches and the oscillator should not surprise you by now. The two sample-and-hold circuits are not so easily discovered, because they appear in the form of two capacitors, $C_a$ and $C_b$. The only new blocks are a clamping circuit and a circuit to suppress intermodulation. These two sub-circuits are of vital importance to a good chopper opamp, and their function will therefore be reverted to a little further on in this article.

During the first phase, also called the clock phase, the null amp compensates itself. Switch $S_1$ is closed, and short-circuits the amplifier inputs. The output voltage is stored in external capacitor $C_a$ via switch $S_2$. Since there is no input signal, the voltage on $C_a$ is equal to the input off-set voltage of the null amp. Furthermore, the capacitor voltage is fed back to an additional inverting input, so that the off-set error of the null amp is eliminated. During the second period of the clock signal, switch $S_2$ is closed, and $S_1$ is open. The null amp then measures the input off-set voltage of the main opamp, and stores it in capacitor $C_b$. At the same time, the measured voltage is applied to the non-inverting input of the main amp, so that the input off-set voltage is compensated. Thus, the system compensates $U_{os}$ of both amplifiers at the rate of the clock or chopper-frequency, $f_c$.

It will be noted that the chopping operation is effected only by the main opamp. The glitches mentioned at the close of the section on the classic chopping amplifier are virtually absent with chopper opamps because the amplified signal is always passed via the continuously operating main opamp.

Recovery time

The decision to use chopper opamps in a practical circuit instead of standard opamps may lead to some surprising problems. First, chopper opamps typically require a much longer time to recover from an overdrive condition, which may occur, for instance.
when the output circuit is driven into saturation. Saturation occurs readily and is perfectly normal in, for instance, a comparator circuit.

After an overdrive condition, the main amp no longer works as a linear amplifier. As a result, the voltage difference between the inverting and the non-inverting input is large relative to $V_{os}$. The auxiliary opamp responds to this condition by charging the two capacitors, $C_A$ and $C_B$, to the maximum level, i.e., the supply voltage. Inevitably, the main opamp requires some time to remove these capacitor charges when the overdrive condition is passed. In the datasheets, the discharge time is referred to as the overload recovery time. For a conventional opamp, this time is about 10 μs. A chopper opamp, however, may need up to 4 s to recover.

Fig. 8. Typical block diagram of a chopper-stabilized operational amplifier.

Fig. 9. This clamp circuit reduces the overload recovery time of the ICL7650.

Fig. 10. The clamp circuit is actuated by connecting the clamp input to the inverting input of the opamp. Figure 10a shows a comparator with very low off-set, and Fig. 10b an inverting direct voltage amplifier.
The clamp circuit provided in the latest chopper opamps serves to reduce the recovery time. The ICL7650, manufactured by Maxim and Teledyne, for instance, has a recovery time of only 300 ms. The clamp circuit used in this chip is shown in Fig. 9. The circuit is actuated by connecting the clamp terminal to the inverting input of the amplifier. Figure 10 shows two circuits that make use of this option.

The clamp circuit is really quite simple, and consists of a mere switch that closes automatically when the output voltage is too close to the supply voltage. When that happens, the switch shunts the externally connected feedback resistor, so that the amplification is reduced. The clamp thus effectively prevents the amplifier being driven into saturation. The very latest chopper opamps have an additional circuit that limits the voltage across the sample-and-hold capacitors. The result is an even shorter recovery time—Texas Instruments' TLC2652, for instance, has a recovery time of only 40 ms.

A further problem with chopper opamps may not be noticed until you are dealing with alternating voltages. Unfortunately, an alternating input voltage may cause unwanted sum and difference frequencies because it is mixed with the clock signal. The cause of this annoying effect, called intermodulation, can be traced back to the fact that the voltage between the inverting and the non-inverting inputs of the opamp corresponds closely to the off-set voltage. It should be noted, however, that this is valid for direct voltages only, when the main opamp has a very high open loop gain, and equation 1 may be replaced by equation 3. As soon as an alternating voltage is applied to the opamp, the open-loop gain drops rapidly, as shown by the graph in Fig. 11.

Equation 1 allows us to deduce that the limited value of $A_{ol}$ is $\left(\frac{1}{1 + \frac{R_2}{R_1}}\right)$ also includes a part of the input signal:

$$\left(\frac{1}{1 + \frac{R_2}{R_1}}\right) \cdot m$$

Furthermore, this part increases with frequency since variable $E$ deviates more and
Open-loop gain becomes smaller (see equation 2). Hence, this alternating voltage component appears also at the output of the auxiliary amplifier and at the input of the chopper amplifier. This component is generated as a result of the sampling operation, which causes sum and difference frequencies. To prevent these frequencies rising to a level high enough to be resolved by the input of the chopper amplifier, this component is treated as offset during the nulling of the OP177B at the end of the list. The 741 and the OP177B are not considered here for reference.

<table>
<thead>
<tr>
<th>TYPE</th>
<th>( U_{op} ) (( \mu V )) max.</th>
<th>( dU_{op}/dT ) (nV/K) typ.</th>
<th>INPUT BIAS (pA) typ.</th>
<th>NOISE ( \mu V_{pp} ) typ.</th>
<th>SUPPLY CURRENT (mA) typ.</th>
<th>SUPPLY VOLTAGE (V) max.</th>
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<td>2.0</td>
<td>44</td>
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</table>

Notes: 1. \( 0 - 10 \) Hz 2. \( 1 - 100 \) Hz 3. \( V_+ \) to \( V_- \)

Table 1. Overview of the most popular chopper opamps, and their main technical characteristics. The 741 and the OP177B are not included here for reference.

Practical notes

Chopper-stabilized opamps usually have the same pinning as standard types. This allows them to be used as upgrades in existing circuits, replacing opamps with worse d.c. specifications. The only components that need to be added are the two external capacitors, \( C_A \) and \( C_B \). This is not required, however, with some amplifiers. The LTC1049 and LTC1050 from Linear Technology, for instance, have on-chip capacitors. Unfortunately, production techniques limit the maximum capacitance of such integrated capacitors to about 450 pF, which gives these opamps a low performance in regard to noise. The usual values of the external capacitors lie between 0.1 \( \mu F \) and 1.0 \( \mu F \). In all cases, high-grade capacitors are required to bring out the specific qualities of a chopper opamp. Film capacitors like polystyrene and polypyrrole types are well worth using.

Unfortunately, the use of high-grade capacitors is not a guarantee that a d.c. amplifier is obtained with a small off-set and a low drift. There is another factor, which has not been mentioned so far: thermovoltages. Thermovoltages occur where two different materials are in contact. As indicated by the name of the phenomenon, the voltage is temperature-dependent. In practice, a thermovoltage readily amounts to a few microvolts per Kelvin. The average drift of a good chopper-stabilized opamp is of the order of \( 10 \) nV/K. However, this value is not usually achievable in a practical amplifier without paying attention to thermoelectric effects in and around the circuit. Components which form connections without soldering, such as switches, relays and connectors, must not be used in the input circuit. Where parts are soldered, it is best to use components with a low thermoelectric specification, such as a cadmium alloy. Errors brought about by thermoelectric effects may also be kept to a minimum by arranging a symmetrical circuit at the opamp inputs. The most sensitive part of the amplifier is thermally balanced by using the same components in the two branches (even if they are really superfluous for the function of the circuit, see Fig. 13), and by forcing an equal number of solder joints. Furthermore, temperature differences as a result of, say, ventilation or power dissipation, must be kept as small as possible.

Guard

An additional advantage of chopper-stabilized opamps is the extremely low input currents. The TLC2652, for instance, has an average input bias current of 4 \( pA \) at an ambient temperature of 25 °C. In practice, however, little use is made of this characteristic because the external leakage currents are much higher. Nonetheless, these leakage currents are easily kept in check. The necessary measures may already be taken during the printed-circuit board design phase. For instance, the solder spots near the inverting and the non-inverting inputs of the opamp can be surrounded by a screening copper area, called a guard. The principle is illustrated in Fig. 14. It is desirable that the guard be held at about the same potential as the inputs of the opamp. Thus, the guard is connected to ground in an inverting circuit, and connected to the input of the opamp in a non-inverting circuit. It will be clear that guards must be provided at both sides of the PCB. Finally, the PCB is cleaned with alcohol before fitting the components.

The differences

From the above discussion you will have gathered that there are many types of chopper opamps available. A selection of the most popular types, along with their main specifications, may be found in Table 1. The good old 741 opamp, which is a chopper, is also included for your amusement. The OP177B at the end of the list represents the latest in bipolar technology, and is a competitive alternative to chopper opamps, according to the manufacturer, PMI.

Finally, a word of warning to those of you who want to start immediately replacing standard opamps by chopper types: as yet, these devices are quite expensive (expect to pay around £10 per amplifier) and difficult to obtain as one-offs.

Reference:
DROITWICH TIMEBASE

Roughly two years ago, the carrier frequency of the 400-kW long-wave Droitwich transmitter was changed from 200 kHz to 198 KHz. This was done by the BBC to comply with the internationally agreed 9-kHz spacing for broadcast stations in the medium- and long-wave bands. The frequency change of 2 kHz was largely unnoticed by thousands of listeners of the Radio Four and BBC World Service programmes. Not so, however, by the many users of frequency standards and timebase circuits which derived their stability from the 200 kHz carrier. All these circuits became useless overnight since the new frequency, 198 kHz, can not be divided to give multiples of 10 Hz. Fortunately, there is a way to get your timebase ticking again. An update for our own Droitwich receiver, an immensely popular project which goes back as far as 1977, is described here.

K.F. Ruwisch

Although the difference of 2 kHz is hardly noticed on the tuning scale of the vintage radio in the introductory photograph, the output frequencies supplied by an unmodified Droitwich timebase are useless for most, if not all, digital circuits. This is because they are no longer exact multiples of 10 Hz.

The problem is obvious: we can no longer use our receiver plus timebase because the Droitwich transmitter is at 198 kHz instead of 200 kHz. All is not lost, however. The good news is that the stability of the Droitwich carrier is still just as good as before the change from 200 kHz to 198 kHz. So, the solution to the problem is also obvious: to enable us to use our timebase circuits, we must convert the 198-kHz output signal of our Droitwich receiver to 200 kHz.
Up by 2 kHz

The block diagram of the circuit we have in mind is shown in Fig. 1. Assuming that the unit is provided with the 198-kHz digital output pulses from a Droitwich receiver, it supplies a rock-steady 200-kHz output signal. No changes are required to the existing Droitwich receiver.

At the input of the upgrade circuit we find a special divider around a 4040, wired for a divisor of 99. Its output signal has a frequency of 2 kHz and is used as a reference for a phase-locked loop (PLL) circuit based on the well-known 4046. The voltage-controlled oscillator in the PLL is set to operate at 200 kHz. Its output signal is divided by 100 by a 4518 dual decade counter to give 2 kHz.

To understand how the output frequency of the circuit is kept stable, let us assume that the VCO drifts from the nominal frequency of 200 kHz. This drift, however small, causes a frequency difference between the 2-kHz reference signal (derived from Droitwich) and the 2-kHz signal supplied by the 4518. The frequency difference causes the phase comparator in the 4046 to supply an error voltage. In this way, the VCO is automatically retuned to minimize the frequency difference. The upshot is that the VCO output frequency of 200 kHz is 'locked' to the carrier received from Droitwich.

It could be argued that the PLL is not required because the 2-kHz signal from the 4040-based divider may be fed, at a suitable point, into an existing divider cascade to give the previously mentioned decade timebase frequencies. We feel that the up-conversion to 200-kHz is required, however, to make sure that the formerly available frequencies of 100 kHz and 10 kHz are retained without any change to the existing divider cascade.

In other words, all the functions of the Droitwich timebase you were just about to throw away are restored simply by installing the proposed up-converter.

Circuit description

The circuit diagram of the upgrade is shown in Fig. 2. The diodes at the Q0, Q1, Q5 and Q6 outputs of the 74HC4040 form an AND gate at the reset input of the chip, and define a divisor of 99. The 2-kHz output signal of the HC4040 is fed to the 4046 PLL, whose internal organization is shown in Fig. 3. The VCO frequency is defined by external parts 142 and 143. Here, phase comparator 2 is used. Network R2-C2 forms the PLL loop filter at the control input of the VCO. A LED indicates that the PLL is locked to the Droitwich signal.

The 200-kHz VCO signal is divided by 100 in a 4518 dual BCD counter. The 2-kHz output signal at pin 14 of this IC is fed to the Cin (phase comparator in) input of the 4046.

The 200-kHz output signal of the upgrade circuit is digitally compatible with a swing of 5 Vpp, and can be fed to any existing divider cascade based on TTL ICs or CMOS ICs, operating at a supply voltage of 5 V.

Construction

Construction of the upgrade circuit is straightforward on the small PCB shown in Fig. 4. The input of the board is connected to output A of the Droitwich receiver (see Ref. 1). The output of the board is connected to the existing 200-kHz output socket of your frequency standard, and to the input of any divider cascade you may have built into the

Fig. 1. Block diagram of the frequency converter. The circuit comprises a divider and a phase-locked loop.

Fig. 2. Circuit diagram of the timebase upgrade.
 Assuming that you use the Droitwich receiver described in Ref. 1, carefully adjust the aerial, and then preset Pt, until the lock LED on the receiver board lights. Use the earpiece, and check that you are tuned to Droitwich by listening to the programme. Next, check that output A of the receiver supplies digital pulses to the upgrade board. If the pulse train is steady, the lock LED on the upgrade board lights, and the output should supply a stable 200-kHz signal.

Reference:

Fig. 3. Block diagram of the 4046 phase-locked loop used in the upgrade (Illustration courtesy RCA/Harris Semiconductor).

Fig. 4. Single-sided printed-circuit board for the timebase upgrade.

Fig. 5. Completed prototype of the Droitwich receiver described in the June 1977 issue of Elektor Electronics.

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TSIEN (UK) LTD

by Bernard Hubbard

In just under eighteen months, two young men have created a company and a software system that has made a significant impact in CAD.

The company is called Tsien (UK) Ltd and the software is BoardMaker I & II. BoardMaker I was launched in July 1989 and Tsien have sold 1200 copies of it so far, which lends considerable credence to their claim that BoardMaker I is easy to use, speedy and offers remarkably more benefits to CAD users than any other system priced at under £200. Some features, notably automatic ground planning with obstacle avoidance are found only in far more expensive systems.

BoardMaker 2 was introduced at last April's CAD-CAM Show and since then the customer base has already grown to around 3000, including the free conversion from BoardMaker I. Even Tsien's rivals showed great interest in BoardMaker 2 at the show. According to John Ellis, Tsien's Sales and Marketing Manager, that is not surprising, because "BoardMaker's price-performance ratio is second to none".

With BoardMaker, which is used on IBM PC XT, AT or 100% compatible computers, you are encouraged to explore the menus and experiment with the options available.

According to John Ellis: "You will find Tsien's sensible user interface a straightforward link to well-thought-out editing features that make the design of printed circuit boards easy. It is equally well suited to conventional single-sided boards and multi-layer surface-mount designs.

BoardMaker V2.23, the latest version, has net-list capability, being able to import nets in Tango, Racal Redac and Konnect formats, with more to follow. Customers without schematic capture can create a rats-nest with BoardMaker and the net-list export facilities then allow them to easily check this against the schematic. As with all other net-list-based PCB CAD-CAM packages, the benefits are quick routing (reference to schematics being minimized) total accuracy and global net-based rule checking to confirm automatically your layout.

One of the attractions of BoardMaker is that it is designed to run quickly on an ordinary PC with the obvious benefits that it is speedy to use on a 286 or 386 system. A Eurocard with a dozen assorted chips will redraw at any zoom level in around a second, minimizing the need for high-resolution displays because there is no penalty for movement around the layout. Speed of operation is coupled with a WYSIWYG Display, mouse/menu or key operation, laser, matrix printed or pen-plot outputs for prototype artworks, photoplot and CNC outputs for production, giving the basis of a highly user-friendly professional package.

In addition to the capabilities of a typical net-based PCB layout, BoardMaker has other features designed to ease the process and make the package useful in a wider range of environments. The next net facility allows the user to scan quickly and demonstrate any unrouted nets or portions of nets and the automatic component designation speeds placement and block copying. The ability to understand individual design rules for each net maintains quality and safety on mixed-voltage boards.

The company have also recently introduced a new autorouter. Called BoardRouter, this is an integrated gridless autorouter module that overcomes the limitations normally associated with autorouting. You specify the track width, via size and design rules for individual nets, BoardRouter then routes the board based on these settings in the same way you would route it yourself manually.

This ability allows you to autoroute mixed technology designs (CMOS, Analogue, digital, power switching, etc) in one pass, while respecting all design rules.

BoardRouter will automatically place 1, 2 or even 3 tracks between pins. You can freely pre-route any tracks manually using BoardMaker prior to autorouting. Whilsts autorouting, you can pan and zoom to inspect the routes placed, interrupt it, manually modify the layout and resume autorouting.

BoardMaker V2.23 and BoardRouter are priced at £295.00 each, which includes 3 months free software updates and full telephone technical support. The pair together can be bought for only £495.00 direct from Tsien (UK) Ltd. Cambridge Research Laboratories, 181A Huntingdon Road, CAMBRIDGE CB3 0DJ. Telephone (0223) 277777, Fax (0223) 277747.
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