WATT-HOUR METER

U-2400 NiCd Battery Charger

RF Attenuators

Multi-user satellite TV reception

New Column: DX Television

I²C Optocard

Digital Audio Enhancer
MULTI-USER SATELLITE TV RECEPTION

This article looks at several options that exist for the distribution of satellite TV signals in small networks. Also, in the face of the formidable increase in the number of programmes that can be received these days, we are hard pressed to have a look at receiver system configurations that allow two or more satellites to be received simultaneously using a single, fixed, dish antenna.

By P. Schrama

MOST of the information that has appeared in satellite TV magazines regarding multi-user networks has been confined to reception of the Astra satellite cluster at 19.2 degrees East. However, as many of you will be keenly aware, there is more on the Clarke belt than the two Astra 'birds'. Consequently, many of you will have wondered what can be done in respect of multi-satellite reception as well. In particular, ex-patriates and foreign workers will often be delighted to be able to watch programmes from the 'home front' in addition to those offered by Astra.

Although an outdoor unit with a motor driven dish allows you to aim at any geo-stationary satellite 'within sight', this is not usually a viable solution if more than one user is connected to the system. In a small apartment, for instance, different users may want to watch different programmes on, you guessed it, different satellites. This is a real problem since not everybody can be allowed to turn the dish as he or she likes — others may hate to see their favourite programme disappearing suddenly from the screen!

In this article we will take reception of the Astra, Eutelsat II-F1 and Eutelsat II-F2 satellites as an example to see what equipment is required to set up a multi-satellite, multi-feed, multi-user receive system based on one, fixed, dish antenna. Further, it is assumed that the receive station is located in Western Europe. However, the principles of multi-feed reception and signal distribution of Ku-band (11-GHz) satellite TV signals are applicable to any other area in the world.

The latest trend: multi-feeds for co-reception

The term 'multi-feed' refers to a mechanical system consisting of a number of LNCs (low-noise converters) and associated feeds, fitted outside the focal area, or off the main focal axis, of the dish, in accordance with the azimuth (horizontal angle) difference between the satellites in question. A multi-feed thus allows programmes to be received that are not on the satellite to which the dish is actually aimed, but only if the 'co-satellite' is not too far away from the 'main' one. In a multi-feed construction, the position of the LNC is always opposite the direction from which the satellite is received (see

Fig. 1. Overview of the components used in the proposed receive system configurations.

![Diagram](image)

Fig. 2. Sat System 900 dish with three LNCs fitted on to the multifeed plate.

<table>
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GP3 14/18V (60mm) or Continental Microwave (40mm)

GP 32

GP 31 AB (2050 MHz)

Feed + OMT + 2 x LNC.

Multiswitch 14/18V

Syntrac II 200 Hz

Coax relay (A/B switch)

Active 4-way splitter

ELEKTOR ELECTRONICS FEBRUARY 1993
Fig. 9. This is because the angle of the reflected beam (to the LNC input) is the same as the angle of incidence. In plain words: the dish and main LNC are aimed at one satellite as before, while a second LNC, fitted outside the focal area, 'squints' to another satellite positioned not too far away from the main one.

Because of some physical properties related to the dish geometry, like reduced illuminance, we have to take into account a reduced dish gain for signals caught outside the focal area. The reduced gain is often compensated by using a dish size that is the next larger relative to the size normally required (e.g., 1 m instead of 80 cm, 1.5 m instead of 1.2 m, etc.). Furthermore, the dish size required for multi-feed co-reception depends on the azimuth difference. The gain reduction of a good dish antenna will be roughly 1 dB per 3° of azimuth difference.

In general, the main axis of a dish antenna fitted with a multi-feed LNC cluster is aimed at the weakest satellite. If, for instance, we wish to receive Astra as well as Eutelsat II-F1 or Eutelsat II-F2, a dish size of 85-90 cm is normally required (in Western Europe). If this dish is aimed at one of the Eutelsats, 'squinting' to Astra obviously does not require a larger dish, since 60 cm is usually adequate for this satellite.

Since both Eutelsats (at 10° and 13° East) are normally equally strong, but weaker than Astra, the dish is conveniently aimed at a point in between the Eutelsats. Using a 90-cm dish, reception of these two satellites will still be adequate, since the dish gain reduction will be limited to about 0.5 dB only, caused by an azimuth difference of 1.5° in each direction.

Larger azimuth differences, for instance, between Astra (19.2° East) and Eutelsat I-F4 (7° East) call for larger dishes, not only to compensate the reduced dish gain, but also because the Eutelsat I-F4 beams a much weaker signal into Western Europe. A dish size of 1.2 m is required anyway to receive Belgrade on the Eutelsat I-F4. This size is still sufficient for the reception of Astra at an azimuth difference of no less than 12°. Fortunately, although 12° is very near the limit of what is possible with a 1.2-m dish, such a multi-feed system allows four satellites to be received simultaneously without investing in a motorized outdoor unit (i.e., a polar mount drive).

Azimuth differences smaller than 3° often cause problems with the size and the mounting of the LNC feeds, which then have to be fitted very close to one another. For instance, the azimuth difference is only 2.3° for co-reception of the Astra (1A and 1B) and Eutelsat I-F5 at 21.5° East. Such a multi-feed is however required for users wishing to view the Zagreb broadcast on the I-F5.

If a normal feed horn is used for the weakest satellite, it is often not possible to fit a funnel-type LNC feed close enough to the other one, since this will have an aperture of up to 60 mm. This problem may be solved by using a Marconi LNC with a dielectrical polyrod feed, which is cone-shaped. This type of LNC may be secured directly to the side of the other feed. Thanks to the much smaller diameter of the polyrod feed, it proved possible to receive Astra at an azimuth difference of only 2.3°. However, the HTV Zagreb programme has meanwhile moved to a transponder on the Eutelsat I-F3 at 16° East. This means that the azimuth difference with Astra is now 3.2°. For azimuth differences larger than or equal to 3°, it is possible to use Sharp LNCs also, since these too have relatively small feeds.

As regards the elevation (vertical) angle, this did not appear to require correction in the case of the above examples, since the differences remain limited to tenths of a degree. Setting the correct elevation angle was found unnecessary since it gave no noticeable improvements.

When a TV receive system for satellite as well as terrestrial broadcasts is to be built from scratch (i.e., a completely new system is installed) it is recommended to use a multiswitch or signal combiner as shown in Fig. 5. A separate power supply for the LNCs is also quite useful in such cases.

**The building blocks**

The system descriptions below are based exclusively on modules that are available on the satellite TV market. Some modules may also be available from other suppliers than the ones mentioned. Compatible models from other manufacturers may also exist.

The outdoor unit is based on a 90-cm...
offset dish supplied by ITS of Bonn, Germany. This dish has a number of constructional characteristics that make it particularly suited to our purpose. This dish has a feed boom on to which a standard, universal, feed can be mounted. However, it is also possible to mount a multi-feed holder. This holder is a simple base plate with holes in it to secure the different feeds. Apart from the small, low-cost, base plate, this dish does not require special (expensive) mechanical add-ons to realize a multi-feed construction. Small elevation angle corrections (if at all necessary — see above) may be achieved simply by bending the base plate. The feed clamps that can be supplied are suitable for 23 to 40 mm dia. and 60 mm (MT1) feeds.

Figure 1 lists the different LNC types, modules and equipment. In the examples, it is assumed that single-polarity, 14-18 V, LNCs are used, such as the MTI-GP3 or Continental Microwave, which have off-set local oscillators and an extended IF range up to 2,050 MHz (GP 31AB). It is also necessary at this point to discuss some of the special features of the Syntrac-II receiver used in the proposed receive system configurations. A series of this type of receiver has an extended input frequency range of 2,000 MHz, which is a necessary feature for some of the proposed receive system configurations. The Syntrac-II also has a function known as 'master programming', which enables all settings relevant to a certain channel to be copied to another receiver. The advantages of this feature are evident if it is known that the Astra programme/channel assignment programmed into the receiver by the manufacturer is confused, if not illogical, and is best changed straight away on receipt of the receiver. Assuming that everyone connected to the small cable network has a Syntrac-II receiver, the 'master programming' feature may save a lot of time, since channel re-arranging is required only once, after which the settings can be copied to all users at the flick of a switch.

A further special feature of the Syntrac-II is its ability to switch between two LNCs without having to forfeit the (partly) required 14/18-V controlled multiswitch or polarization selection. Unusually, the LNC switching function is achieved by a relay operated by a 200-Hz signal, which is superimposed on the 14/18 V LNC supply voltage. The above features make the Syntrac-II a very good choice for use in a multi-feed, multi-user, multi-satellite receive system. Another good receiver offering an extended input frequency range (2,050 MHz) is the BX 100 from Zehnder. Unfortunately, this receiver does not supply the 200-Hz signal to control the coax switch. However it can be used without problems in all example configurations where the 200-Hz signal is not used.

If you are an apartment dweller with a permission to fit a small dish on the balcony railing, and the distance to the receiver is only a few metres, the total equipment layout becomes much simpler, and it may be worthwhile to use a satellite receiver with two IF inputs. An example of such a receiver is the inexpensive, widely available CBM-9200 from Citizen, or the FACE MRD-920, which has a built-in D2MAC decoder.

The total size of the installation can be reduced significantly by clever selection of components and cutting down on the number of control lines that would otherwise be required.

Two satellites — one user (Fig. 3) Figure 3 shows a simple example. Converter A or B is selected via the 200-Hz coax relay. The polarization selection on both LNCs is achieved by the 14/18 V control voltage. This configuration may be used, for instance, for the Astra-Kopernikus pair, or the Astra-Eutelsat II-F1/Eutelsat II-F2 pair.
Fig. 6. Two satellites, two to four users. LNCs used are single output split IF range 2050 MHz types.

Although the interest in the Kopernikus satellite has declined since the start of the Astra 1B in March 1991, many Hi-Fi music lovers will remain keen on receiving the DSR (digital satellite radio) channels packed together on one of the Kopernikus transponders. A DSR programme bundle is, unfortunately, not (yet) available on the Astra. When aimed at the Eutelsat II-F1, the proposed system brings several Turkish, one Arabic and a number of English and French-language programmes on your screen. Among the signals transmitted by the Eutelsat II-F2 are Turkish, Spanish and Italian broadcasts.

Two satellites — two to four users (Figs. 4 and 5)
Whenever more than one user is to be connected to the receive system, there is no way to go round extending it. As illustrated in Figs. 4 and 5, a number of options are available.

For quite some time, the option shown in Fig. 4 was the most widely used. However, the use of 2,050-MHz components (Fig. 5) enables the system size and outlay to be reduced. This is in spite of the fact that the 2,050-MHz compatible GP31AB LNCs are more expensive than GP32's (compare the two parts list). The multswitch shown in Fig. 5 is used to select between two satellites instead of between two LNC polarizations. Syntrac-II receivers are used as in Fig. 4.

Two satellites — two to four users (Figs. 6 and 7)
This configuration is basically similar to the one described above. Here, too, the total cost is reduced considerably by the use of GP31AB converters — the parts list indicates that the use of this LNC economizes on no fewer than four A/B

Fig. 7. As Fig. 6, but using four single-polarization LNCs and two OMTs.

Fig. 8. A modern LNC with electrical 14/18 V H/V polarization switching (Continental Microwave).
Three satellites — one user (Fig. 9) ('single-cable solution')

Many foreign workers, immigrants and other ex-patriates are keen on receiving more than two satellites. As an example, many Turks will be very eager to watch TRT-Int. on Eutelsat II-F1, TeleOn + Star 1 and Show-TV on Eutelsat II-F2, in addition to the Astra programmes. If the coax relay is not energized, Astra is selected via output A. Polarization switching is achieved via the 14/18 V control voltages as usual. These voltages are passed by the coax relay. The II-F2 satellite is selected by applying the 200-Hz switchover signal to the coax relay, which switches to output B, and carries a fixed voltage of 18 V. Unfortunately, the Turkish programmes are transmitted at different polarizations. Since the 14/18 V LNC voltage is required for the selection of two of the three satellites via the multiswitch, we once again use the Type GP31AB LNC, which supplies the signals in the two polarization fields in two output (IF) frequency ranges. When the third satellite is selected, the signal flow is once again via output B of the relay, but this time with a 14 V fixed voltage on the other output of the multiswitch. Although only one polarization is available for the third satellite, this is fortunately sufficient to receive the missing Turkish as well as all other horizontally polarized programmes. If it is required, for some reason or other, to receive both polarizations, the GP31AB LNC may be used.

Three satellites — two to four users (Fig. 11)

While it is possible to choose between a motor driven system and a multi-feed in the case of a single-user receive system, there is no option but to use a fixed dish whenever more than one user is to participate. It will be impossible, however, for every user to install one or more dishes on the roof, if only because of limited space, building licenses, etc. To enable every one to watch his or her favourite programmes, the configuration of Fig. 11 is suggested. Its operation is similar to that of Fig. 9, except that the Astra LNC output signals are taken to an active splitter before they arrive at the coax relays. In this multi-user system, a GP31AB is required for the reception of the two Astra polarizations.

Final thoughts

It should be noted that 2,050-MHz technology is still fairly new, and compatible receivers and LNCS are, unfortunately, still thin on the ground. When designing
a receive system configuration based on 2,050-MHz compatible modules it is, therefore, necessary to be very attentive to the relevant specifications, for which the manufacturers or suppliers may have to be pressed a little.

Not all receivers claimed to be 2,050-MHz compatible actually have a tuner that can reach up that far. In many cases, the gain drops to very low levels above 1,800 MHz or so, which results in noisy signals above this frequency.

When buying a 2,050-MHz LNC, be sure that the conversion gain is the same for both polarizations. Here, too, there are many products that fail miserably. It is therefore good practice to test everything on receipt of the materials, and return anything that does not meet the specifications.
WATT-HOUR METER – PART 1

Design by M. Ohsmann

Since low energy consumption of domestic appliances benefits your bank account as well as the environment (provided that their production is also economic and environment-friendly), the watt-hour meter, which can monitor the energy use of many appliances, will be of interest to many. Even if current and voltage are not sinusoidal, or out of phase with one another, the meter determines the use of energy accurately.

The watt-hour meter can measure current, voltage and energy used by refrigerators, deep-freezers, washing machines, or that provided by a solar panel, and can also monitor the charging of a battery. The available measuring ranges enable the consumption of a 300 mW bicycle lamp as well as that of a 3 kW electric hob plate to be monitored.

Primarily intended as a stand-alone unit, the meter may also be used in conjunction with a computer. This will be discussed further in Part 2.

To measure energy, three other quantities have to be known: voltage, current and time. How it is done with the watt-hour meter is shown in Fig. 1. The voltage and current delivered by a generator to a load are measured relative to time with the aid of a microcontroller, which consists of an analogue-to-digital (A-D) converter with switched inputs and a timer. The current is measured by passing it through a shunt resistor that converts it into a directly proportional voltage. The resistor is shown for clarity only: in practice it is part of the meter.

The range of input voltages is determined by a series resistor that also provides current limiting. Before the voltage and current are applied to the controller, they are amplified and rectified. Their polarity is determined by a comparator whose output is linked to a digital input of the controller. This arrangement makes it appear as if the A-D conversion were upgraded from eight to nine bits.

The energy supplied to a load is determined by measuring the voltage across, and the current to, the load.

Fig. 1. The energy supplied to a load is determined by measuring the voltage across, and the current to, the load.

Fig. 2. Block diagram of the watt-hour meter.

Fig. 3. A front panel foil (here scaled down to 75%) is available: see page 70.
Circuit description

Apart from the two input stages, the circuit is almost a standard application of a controller from the MCS51 family. Note that the circuit does not show the entire meter: the series resistors at the voltage input, the shunt resistors for measuring current, and the display, are absent. The resistors will be discussed later: the display is a ready-made unit about which nothing much needs to be said other than that $P_1$ controls its contrast.

The series resistors form an integral part of inverting amplifier $IC_1$. The meter has two inputs: one for voltages of 100–1000 V and one for 0–100 V. This arrangement saves a (hard to find and expensive) switch rated at 1000 V. Four voltage (peak value!) ranges can be selected with $S_2$, see Fig. 2. The input voltage is adjusted with $P_2$ and $P_3$.

After it has been amplified, the signal is applied to a precision rectifier based on $IC_9$ and $IC_{10}$. The rectified signal is applied to the analogue input, $AN0$, of controller $IC_{15}$. The polarity of the signal, before it is rectified, is determined by comparator $IC_{2a}$, whose output is linked to one of the many I/O gates of $IC_{15}$.

The input stage for current measurements is largely identical to that for the voltage measurement, with the addition

Fig. 4. Circuit diagram of the watt-hour meter.
Fig. 5. Motherboard.
of $R_1/P_1$ and $R_2/P_2$. As for voltages, the meter has two inputs for currents: one for 0-2.55 A and one for 2.55-25.5 A. Again, this saves a (large and expensive) switch that can handle 25 A. Four current (peak value) ranges can be selected with $S_1$—see Fig. 2.

The third rotary switch, $S_4$, enables selecting the quantity to be measured: voltage, $U_p$; current, $I_p$; power (in watts); energy or work done in watt-seconds (W-s); and energy or work done in kilowatt-hours (kWh).

Memory of the meter is contained in a 32-kbyte EPROM, $I_{C3}$, and a 32-kbyte RAM, $I_{C4}$. Both are connected to address lines A14 and A15 in the standard manner. In Fig. 4, A14 and A15 have been interchanged, so that the memory division is more or less the same as that of the 8032/8052 computer board used in last year's assembler course.

The serial input and output of the controller are provided with optoisolators that electrically separate the meter and anything connected to it. This means that a supply voltage must be provided for $I_{C3}$ that is isolated from the rest of the meter. This isolation is provided by $T_{R1}$. Since this transformer is lightly loaded, the secondary voltage may be a few volts higher than stated, but that gives no problems with an RS232 connection.

The isolation provided by transformer $T_{R1}$ is also of real benefit. For instance, without it, the meter could not be connected to the mains supply: one of its primary objectives. Do not confuse the 8 V outputs of the main supply with the 8V supply for $I_{C7}$.

### Construction

Since the meter is often connected to the mains, not only via the two transformers, but also via the measuring inputs, the utmost care must be taken in the construction of the meter.

The mother board is shown Fig. 5 and the transformer board in Fig. 8. The mother board may be fitted to the front panel, but it may also be mounted on a subframe. In any case, the assembly MUST be carried out with man-made fibre bolts, nuts, washers and spacers. Metal ones must NOT be used in any circumstances, since several fixing holes are not sufficiently far away from voltage-carrying parts on the board.

The potentiometers may be mounted on the component side or on the track side of the board, but reset switch $S_1$ always on the track side. Unless this component meets the same insulation requirements as a mains on-off switch, it must be impossible to access it without opening the enclosure.

The rotary switches MUST have man-made fibre spindle.

The display poses a safety problem, because, as usual, this should be mounted as close to the front panel as possible.
This cannot be done, however, in this case since it would not meet insulation requirements. The solution lies in gluing a sheet of perspex to the front panel that is in all directions 6 mm larger than the display.

The display is mounted on the mother board with the aid of 2 bolts, nuts and spacers. After that, make the 14 connections to $K_1$ with circuit wire. A number of resistors must be connected between the input sockets and the mother board as shown in Fig. 6. The mains earth must be connected to the metal parts of the enclosure but NOT to the G(rou)ND of the circuit. A fuse is not needed if the transformers specified in the parts list are used, because these are short-circuit-

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**Fig. 6.** A number of resistors must be soldered between the mother board and the input sockets. The remainder of the wiring is self-evident.

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**Fig. 7.** The input sockets are types that match insulated banana (or similar) plugs. The top diagram refers to the use of European or American mains plugs and sockets; the bottom one to British three-pin types.

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**Fig. 8.** Transformer board.
**SPECIFICATION**

Measuring ranges
- Voltage (true r.m.s.) 25.5 V; 100 V; 255 V; 1000 V (max).
  - $(U_{\text{peak}} = 1000 \text{ V})$
- Current (true r.m.s.) 0.63 A; 2.55 A; 6.3 A; 25.5 A (max).
  - $(I_{\text{peak}} = 25.5 \text{ A})$
- Active power in Watt*
- Energy in Watt-seconds or kilo-Watt-hours*

Frequency range DC–1 kHz
- Sampling frequency 5 kHz
- Time indication (energy 0–10 000 hours (416 days)
  - measuring period)

Alphanumeric display
- 2 lines of 16 characters each

Can be read and controlled via an electrically isolated RS232 interface.

Can be calibrated with a simple digital multimeter.

Freely programmable control computer via integral monitoring program Type EMON52

*Range depends on current and voltage settings

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**proof. Before soldering resistors $R_{59}-R_{59}$, and $R_{60}$ between the board and input sockets, insert them into a length of insulating sleeve. Solder shunt resistors $R_{59}$ and $R_{60}$ direct to the input sockets. It may be necessary (owing to its non-availability in your locality) to make $R_{60}$ from a 1.8 mm thick (2.5 mm² cross-sectional area) piece of wire 145 cm (4 ft 9 in) long. The connections between current input and common sockets and the board must be twisted to prevent them picking up hum. The sockets on the meter and the corresponding test plugs—see Fig. 7—MUST be insulated banana (or similar) types.

**Initial tests**

When everything has been built and checked, set all potentiometers to the centre of their travel and $S_4$ to position cal/V24. Only then, switch on the mains. Leave the inputs open. Adjust $P_1$ for good legibility of the text on the display. It should read ‘EMON52 MODE’ to indicate that the controller works correctly. If that text is not there, check the supply voltages to the various stages of the circuit. If they are all right, check whether the ALE signal is present. If that also is all right, carefully check the boards in detail, paying particular attention to possibly overlooked solder connections or tiny flecks of solder between tracks.

Next month’s instalment will deal with the calibration and operation of the watt-hour meter.
Fig. 5. Motherboard.
PARTS LIST

Resistors:
- R1 = 10 kΩ
- R2, R21 = 1 kΩ
- R3, R13 = 47 kΩ
- R4−R10, R14−R20 = 10 kΩ, 1%
- R11, R12 = 680 Ω
- R22, R23 = 4.7 kΩ
- R25, R26, R27 = 3.32 MΩ, 1%
- R28, R31, R32 = 3.32 kΩ, 1%
- R29 = 10 mΩ, 8 W
- R30 = 100 mΩ, 1 W
- P1, P2, P6 = 10 kΩ preset
- P3, P7 = 25 kΩ preset
- P4, P5 = 5.2 kΩ preset

Capacitors:
- C1, C2 = 22 nF
- C3 = 1 µF, 10 V
- C4, C5 = 12 nF
- C6-C8 = 100 µF, 25 V
- C9 = 1000 µF, 25 V
- C10-C26 = 100 nF

Semiconductors:
- D1-D4 = IN4148
- D5-D7 = IN4001
- B1 = B80C1500
- IC1, IC4 = 7805
- IC2 = LM319
- IC3, IC5 = TLC227
- IC6, IC7 = CNY965
- IC8 = 7805
- IC9 = 7908
- IC10 = 7908
- IC11 = 62256
- IC12 = 6243 (see p. 70)
- IC13 = 74HC00
- IC14 = 74HC573
- IC15 = SAB80C535

Miscellaneous:
- K1, K5 = 2-way terminal block; pitch 5 mm
- K2, K6 = 3-way terminal block; pitch 7.5 mm
- K3 = 14-way terminal block; pitch 5 mm
- K4 = 2-way terminal block; pitch 7.5 mm
- S1 = single-pole, spring-loaded switch with make contact
- S2, S3, S4 = 1-pole, 12-way rotary switch for PCCB mounting with polythene spindle
- T1 = short-circuit-proof mains transformer, secondary 6 V, 1.5 A
- T2 = short-circuit-proof mains transformer, secondary 9 V, 1.5 A
- X1 = 12 MHz crystal
- LCD module Type LTN211F-0 (Philips Components)
- 5 insulated banana (or similar) sockets and plugs
- 9-way female D connector
- 1 mains entry plug
- 1 mains switch
- 1 PCB Type 920148-I
- 1 PCB Type 920148-II

*See page 70.

WATT-HOUR METER – PART I

of \( R_1 / P_3 \) and \( R_2 / P_4 \). As for voltages, the meter has two inputs for currents: one for 0–2.55 A and one for 2.55–25.5 A. Again, this saves a (large and expensive) switch that can handle 25 A. Four current (peak value) ranges can be selected with \( S_3 \) – see Fig. 2.

The third rotary switch, \( S_4 \) enables selecting the quantity to be measured: voltage, \( U_{eq} \); current, \( I_{eq} \); power (in watts); energy or work done in watt-seconds (W-s); and energy or work done in kilowatt-hours (kWh).

Memory of the meter is contained in a 32-kbyte EPROM, \( IC_1 \), and a 32-kbyte RAM, \( IC_2 \). Both are connected to address lines A14 and A15 in the standard manner. In Fig. 4, A14 and A15 have been interchanged, so that the memory division is more or less the same as that of the 8032/8052 computer used in last year’s assembler course.

The serial input and output of the controller are provided with optoisolators that electrically separate the meter and anything connected to it. This means that a supply voltage must be provided for \( IC_3 \) that is isolated from the rest of the meter. This isolation is provided by \( T_1 \). Since this transformer is lightly loaded, the secondary voltage may be a few volts higher than stated, but that gives no problems with an RS232 connection.

The isolation provided by transformer \( T_2 \) is also of real benefit. For instance, without it, the meter could not be connected to the mains supply: one of its primary objectives. Do not confuse the 8 V outputs of the main supply with the 8 V supply for \( IC_2 \).

Construction

Since the meter is often connected to the mains, not only via the two transformers, but also via the measuring inputs, the utmost care must be taken in the construction of the meter.

The mother board is shown Fig. 5 and the transformer board in Fig. 8. The mother board may be fitted to the front panel, but it may also be mounted on a subframe. In any case, the assembly MUST be carried out with man-made fibre bolts, nuts, washers and spacers. Metal ones must NOT be used in any circumstances, since several fixing holes are not sufficiently far away from voltage-carrying parts on the board.

The potentiometers may be mounted on the component side or on the track side of the board, but reset switch \( S_5 \) always on the track side. Unless this component meets the same insulation requirements as a mains on-off switch, it must be impossible to access it without opening the enclosure.

The rotary switches MUST have a man-made fibre spindle.

The display poses a safety problem, because, as usual, this should be mounted as close to the front panel as possible.
**CORRECTIONS**

4-MByte printer buffer insertion card for PCs (April 1993)
The components list on page 56 should be corrected to read:

3 BC557B T1;T2;T3

Contrary to what is stated in the parts list, EPROM 6041 is not supplied with printed circuit board 920009, and has to be ordered separately.

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Watt-hour meter (February 1993)
Contrary to what is stated in the caption below Fig. 3 (p. 14), the front panel foil for this project is not available through the Readers Services.
The following item should be added to the parts list:

1 Metal enclosure LC970 (Telet)

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Video digitizer for PCs (April 1993)
The type number of the relays used in this project should read V23100-V4005-A010, not V23100-A4005-A010.

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ELEKTOR ELECTRONICS MAY 1993
FIGURING IT OUT

PART 2 - CAPACITOR MATHS

By Owen Bishop

This series is intended to help you with the quantitative aspects of electronic design: predicting currents, voltage, waveforms, and other aspects of the behaviour of circuits. Our aim is to provide more than just a collection of rule-of-thumb formulas.

We will explain the underlying electronic theory and, whenever appropriate, render some insights into the mathematics involved.

**Correction**

At the end of last month's instalment it was stated erroneously that this month's issue would contain more applications of KCL and KVL; these will be discussed, however, in our March issue.

The fundamental relationship for capacitors is the equation that defines the unit of capacitance, C, which is the farad:

\[ C = \frac{Q}{U} \]  

([Eq. 14])

The capacitance in farad is the electric charge, \( Q \), expressed in coulombs per volt potential difference, \( U \), between the plates of the capacitor. From this, we obtain the equation for the charge on a capacitor:

\[ Q = CU. \]  

([Eq. 15])

If a capacitor begins by being uncharged, and a constant current \( I \) passes to one plate for a time \( t \), the amount of charge on the capacitor at the end of that period is

\[ Q = It. \]  

([Eq. 16])

This equation defines the coulomb. Combining Eq. 14 and Eq. 15 gives:

\[ U = \frac{It}{C}. \]  

([Eq. 16])

Given the capacitance, the constant current and the length of time for which the current has flowed, we can calculate the potential difference between the plates. For example, if a current of 2 A flows for 0.5 s into a capacitance of 0.01 F (farad), the pd between its plates is:

\[ U = 2 \times 0.5 / 0.01 = 100 \text{ (V)}. \]

**Fig. 15. Capacitor being charged by a constant current.**

In Eq. 16, the term \( It \) is the product of the current and the length of time the current has flowed. In Fig. 15, which plots the current \( I \) in A at each instant of time \( t \) in s (seconds), this product is represented by the shaded area beneath the graph.

This area is the product of the height (current) and width (time) of the line. The amount of charge on the capacitor at any given instant equals the area beneath the line.

There is no difficulty in performing such calculations when the current is constant. There is certainly no need to draw a graph. Complications arise, however, when the current is not constant. Suppose that the current is zero to begin with, but increases steadily at a rate given by the equation:

\[ i = at, \]

where \( a \) is a constant expressing the rate of increase of current in amperes per second, A s\(^{-1}\). Figure 16 shows the situation when \( a = 0.4 \text{ A s}^{-1} \). The graph shows that \( i \) attains a value of 1.2 A after a period of 3 s. Calculating the charge on the capacitor after 3 s is simply a matter of evaluating the area beneath the graph. We use the formula

\[ \text{area} = \frac{1}{2} \text{base} \times \text{height} \]

to obtain the answer which is 1.8 C (coulomb). If the capacitance is 0.01 F as before, the pd between the plates is 1.8/0.01 = 180 V.

In the calculation just completed, we were able to make use of elementary geometry to calculate the area and hence the charge and pd. This method is not practicable if the current is varying in some more complicated way. Suppose the current is varying in an entirely random manner as in Fig. 17. The area has an irregular upper boundary and there is no simple geometrical way of calculating it exactly. The best we can do is to record the current at
very short intervals of time, short enough for the current not to vary significantly during each interval. This may be done by sampling the current at intervals of, say, 0.1 s with the aid of a microcomputer. The result is a series of narrow strips as in Fig. 18. We calculate the area of each strip and sum the areas to obtain the total charge. This method has an inherent error in that the upper boundary of each strip is horizontal, whereas with a continuously varying current it is sloping. The area beneath the curve is not quite the same as the total area of the strips. The error is reduced if we make the time interval smaller. If we could make the strips infinitely narrow, the error would disappear.

Mathematical solution

The technique of sampling the current at very short time intervals has practical limitations, but fortunately the integral calculus comes to our aid. The term integration means 'putting together to make a whole', and the technique allows us to calculate directly the area beneath any curve that can be mathematically defined. It cannot, of course, help with irregular curves such as in Fig. 17, but any curve that can be written as a function is amenable to such treatment.

When we refer to 'a function' in this context, we mean that the way in which the current varies with time can be expressed in an equation. We have already met two such functions. In Fig. 16 the function is

\[ I = 2. \]

Here \( I \) is a constant; its value is independent of time and the variable \( t \) does not enter into the equation.

In Fig. 17, the function is

\[ I = 0.4 t. \]

As already stated, we can solve problems involving such elementary functions by drawing (or imagining) the graph and using simple geometry. But more complex functions such as

\[ I = 2t^2 - 3t + 6 \]

and

\[ I = 2 \sin(4 + 3t) \]

require the integration technique.

Integration

There is no space here to consider the theoretical basis of integration. Instead, we shall state some of the more important results and show how to use them in solving practical problems. The most useful of the results are referred to as the standard integrals.

A few of these are listed in Table 1. Rather than expressing them in terms of variables \( x \) and \( y \), as is usual in maths texts, we have expressed them in terms of \( I \) and \( t \) so that they may be directly applied to problems involving capacitors.

Before we go on to use the technique, there is the matter of the symbols that we use. The idea of integration is that we are summing the areas of an infinitely large number of narrow strips of infinitely small width. The integral expression refers to this in symbolic form

\[ \text{charge } Q = \text{area under curve} = \int_{t_1}^{t_2} I \, dt. \]

The \( \int \) or 'longs' is short for 'summation'. Following this comes a description of what we are to sum: the product of \( I \) (the length of the strip) and \( dt \) (the width of the strip. The 'd' is not a variable: it means 'infinitely small increment', so the term 'd' means 'an infinitely small increment of time'. The \( t_1 \) and \( t_2 \) indicate the period of time over which summation is to be done: the period starting at \( t_1 \) and ending at \( t_2 \). When solving a problem, we replace \( I \) in the expression above by the function of \( t \) which defines how \( I \) varies in time.

We will use integration to solve the problem of Fig. 16 in which the function is \( I = 0.4t \):

\[ Q = \text{area under the curve} = \int 0.4t \, dt. \]

According to the rules of integration, a constant coefficient of \( t \) can be brought in front of the integral sign. This makes sense because any multiplier \( k \) simply makes the strips \( k \) times longer and the area \( k \) times greater. Note that we have not included \( t_1 \) and \( t_2 \) at this stage, as we shall first obtain an expression that applies to any length of time. The integral to be evaluated now is

\[ Q = 0.4 \int dt. \]

Table 1. Some standard integrals expressed in terms of \( I \) and \( t \).
Now look at Table 1. The entry which helps with this example is $t^n$, where $n=1$. Since $n+1=2$, the integral is

$$I/t^2.$$  

Substituting this in the equation

$$Q=0.4(1/t^2)-0.2t^2.$$  

This expression tells us the amount of charge at any given instant, $t$. This is the indefinite integral. In the original problem we wanted to know how much charge accumulated during the given period, defined as $t_1=0$ to $t_2=3$. We find this by calculating the definite integral as follows

$$Q=0.4\int_0^3 t \, dt = 0.4\left\{\frac{t^2}{2}\right\}_0^3 = 0.4(9-2) = 0.4 \times 4.5 = 1.8 \text{ C}.$$  

Here we have written the beginning and end of the time period beside the integral sign. The result of integration is written in brackets with the beginning and end times written after the last bracket. The amount accumulating during the period $t_1$ to $t_2$ is the amount present at $t_2$ minus the amount present at $t_1$, so we can say:

$$Q=0.4\int_0^3 t \, dt = 0.4\left\{\frac{t^2}{2}\right\}_0^3 = 0.4\left(\frac{9}{2}\right) = 1.8 \text{ C}.$$  

This is the same result as we obtained more simply by geometry but serves to demonstrate that the integration method gives the correct result. Before we use integration to solve problems more worthy of its capabilities, there is one point to be considered.

**Initial charge**

The integral tells us how much charge has accumulated on the capacitor. But there is nothing in the equation to take account of any charge that might exist on the capacitor before the current began to flow. If the capacitor had a charge of $2 \text{ C}$ on its plates at $t=0$, this would not affect the rate at which charge accumulates subsequently. At $t=3$, the total charge would be $2+1.8=3.8 \text{ C}$.  

When calculating the definitive integral, we should allow for any possible charge by including it in the equation. In the example above, the modified equation is:

$$Q=0.4\int t \, dt + c,$$

where $c$ is the initial charge, if any. In maths, $c$ is referred to as the constant of integration. In the example, $c=0$, for this was stated when the problem was set out. Being negative implies that the initial charge is of opposite polarity and that the current flowing into the capacitor will be acting to neutralize it and possibly recharge it in the opposite direction.

Unless some statement is made about the initial charge, or we are told the total charge at a given instant in time, we cannot take account of initial charge in the calculation—we only know the amount that accumulates during the given period. Often, the initial charge is known to be zero, so it may be ignored. In other cases, additional information is given, as in this example.

Current is flowing into a capacitor at a rate given by $I=2t+1$. The charge on the capacitor at time $2 \text{ s}$ is 10 C. What is the charge at time $5 \text{ s}$? The total charge is given by:

$$Q=\int (2t+1) \, dt + c.$$  

We integrate each term of the expression individually, using the standard integrals:

$$Q=2\int t \, dt + \int 1 \, dt + c$$

$$=2\left(\frac{t^2}{2}\right) + t + c.$$  

Note that when integrating the constant ‘1’, we consider it to be $1^0$, so that we take $n$ to be zero. The question states that $Q=10$ when $t=2$. Substituting these values:

$$10 = 2\left(\frac{2^2}{2}\right) + 2 + c = 6 + c.$$  

$$c = 4.$$  

The initial charge, or constant of integration, is 4. Now we can evaluate the charge after 5 s:

$$Q=2\left(\frac{5^2}{2}\right) + 5 + 4 = 25 + 9 = 34 \text{ C}.$$  

The constant of integration is ignored when we are calculating definite integrals, as the integrals at time $t_1$ and at time $t_2$ both have added to them. The constants disappear when one is subtracted from the other. Putting it in terms of charge, the amount of charge accumulating during a given period is not affected by whatever charge happened to be on the capacitor already.

If the current function is more complex than those in Table 1, special integrating techniques are required, which will be found in advanced math texts. However, the table covers most of the situations commonly met in electronic circuits. In particular, the sine function is characteristic of audio signals and is, therefore, a feature of many types of circuit, including amplifiers, oscillators and filters. We deal with this in the next section.

**Sinusoidal currents**

It often happens that a capacitor is being charged by a current that is varied according to a sine function. For example:

$$I=2\sin 3t.$$  

How much charge accumulates during 0.1 s? Integrating according to the formula in Table 1:

$$Q = \int_0^{0.1} 2\sin 3t \, dt = \left[-\frac{\cos 3t}{3}\right]_0^{0.1}$$

$$= 2\left[-\frac{\cos 3\times 0.1}{3}\right] - 2\left[-\frac{\cos 0}{3}\right]$$

$$= 2\left[-0.318\right] - 2\left[-0.333\right]$$

$$= 0.03 \text{ C}.$$  

When evaluating the integrals of trigonometric functions, we always work in radians, not degrees, since the standard integrals are derived on that basis. Given the same charging current, what is the charge that accumulates in 1 s? Substituting $t=1$:

$$Q = 2\left[-\frac{\cos 3\times 1}{3}\right] - 2\left[-\frac{\cos 0}{3}\right]$$

$$= 2\left[-0.333\right] + 2\left[0.333\right]$$

$$= 1.326 \text{ C}.$$  

Now find the charge after 2.094 s:

$$Q = 2\left[-\frac{\cos 6.282}{3}\right] - 2\left[-\frac{\cos 0}{3}\right]$$

$$= 2\left[-0.333\right] + 2\left[0.333\right]$$

$$= 0 \text{ C}.$$  

What has happened to the charge? Figure 19 shows that the...
time $t_2=2.094\ s$ is the point at which $3\pi$ equals $2\pi$. This is the end of one complete cycle of the sinusoidal waveform. During the first half of this period (up to $1.047\ s$, which is close to the second example above), the current is charging the capacitor. During the second half-cycle, the current is reversed, removing all the accumulated charge.

Figure 19 also shows that, since the function for the current is a sine function, and the function for the charge is a negative cosine function, the curves are out of phase by $90^\circ$ with the sine curve leading.

In the network of Fig. 20, the charging current is provided by a sinusoidal voltage $U_i$ passing through a resistor:

$$U_i=IR=2Rs\sin3t.$$  
This is the input voltage of the network. The output voltage is that developed across the plates of the capacitor:

$$U_0=\frac{Q}{C}=\frac{-2\cos3t}{3C}.$$  
The input and output voltage are $90^\circ$ out of phase, with $U_i$ leading $U_0$. This network is that of a low-pass filter and the equations above, developed by integration, show why a phase difference exists between input and output signals of such a filter. In a later issue we shall look at the inverse situation in the high-pass filter and examine the inverse mathematical routines of differentiation.

Test yourself

1. A $0.05\ \text{F}$ discharged capacitor is charged by a constant current of $1.5\ \text{A}$ for $0.4\ \text{s}$. The current is then reduced by $1\ \text{A}\ \text{s}^{-1}$ until it is zero. What charge accumulates on the capacitor? What is the potential difference across it?

2. A capacitor is charged by a current specified by the function $I=0.5+3t^2$. What charge accumulates on the capacitor during the period $0.8\ \text{s}$ to $1.0\ \text{s}$?

3. A current $I=-8\sin4t$ is applied to a capacitor which already has a charge of $2.45\ \text{C}$. At what time is the charge on the capacitor first reduced to zero?

Answers will be given in next month's instalment.
**RF ATTENUATORS**

**INCLUDING SOME YOU BUILD YOURSELF**

An attenuator is a circuit that reduces signal level, i.e., has an output signal level that is lower than the input signal level. In this respect, it is the opposite of an amplifier. Precision attenuators are frequently used in electronics to reduce signals a specified amount, usually expressed in decibels (dB), for some purpose such as measurement or stabilization.

By Joseph J. Carr

An attenuator pad will provide the rated attenuation in dB only when the input and output impedances are terminated in a like impedance. For audio systems, it is common to use 600 Ω as the standard system impedance; in radio (and most RF) systems the impedance is 50 Ω; in television systems it is 75 Ω.

When working with attenuators it is the custom to use decibel notation. There is often some confusion on this point because a lot of people simply misunderstand the concept of decibels.

**Decibel notation**

The decibel measurement originated with the telephone industry, and was named after telephone inventor Alexander Graham Bell. The original unit was the bel. The prefix deci means 1/10, so the decibel is one-tenth of a bel. The bel is too large for most common applications, so it is rarely if ever used. Thus, we will concentrate only on the more familiar decibel (dB).

The decibel is nothing more than a means of expressing a ratio between two signal levels, for example the 'output-over-input' ratio of an amplifier. Because the decibel is a ratio, it is also dimensionless — despite the fact that 'dB' looks peculiarly like a dimension. Consider the voltage amplifier as an example of dimensionless gain; its gain is expressed as the output voltage over the input voltage \( U_o / U_i \), so the 'volts' in the numerator are cancelled by the 'volts' in the denominator.

In order to analyze systems using simple addition and subtraction, rather than multiplication and division, a little math trick is used on the ratio. We take the base-10 logarithm of the ratio, and then multiply it by a scaling factor (either 10 or 20). For voltage systems, such as our voltage amplifier, the expression becomes:

\[
\text{dB} = 20 \log_{10} \left( \frac{U_o}{U_i} \right). \tag{1}
\]

Despite the fact that we have mathematically massaged the ratio by converting it to a logarithm, the decibel is still nonetheless nothing more than a means for expressing the ratio between two signals. Thus, a voltage gain of 12 can also be expressed as a gain of 21.6 dB (take your calculator and work out the example).

A similar expression can be used for current amplifiers, where the gain ratio \( I_o / I_i \) replaces the voltage ratio in equation (1).

For power measurements we need a modified expression to account for the fact that power is proportional to the square of the voltage or current:

\[
\text{dB} = 10 \log_{10} \left( \frac{P_o}{P_i} \right) \tag{2}
\]

Decibel notation prescribes a positive sign for a signal gain, and a negative sign for a signal loss. You will see this effect if you calculate either of the two equations above with an output signal level that is less than the input signal level. Thus, if the voltage amplifier gain is 12, the ratio of signals is 12/1, and the dB notation is 21.6 dB. But if the circuit is a 12-dB attenuator, the ratio is 1/12, and the expression works out to be -21.6 dB.

**Adding it all up**

So why bother converting seemingly easy to handle, dimensionless numbers like voltage or power gains to a logarithmic number like decibels? Fair question. The answer is that it makes calculating signal strengths in a larger circuit or system a lot easier. To see this effect, let us consider the multistage system in Fig. 1. Here we have a hypothetical electronic circuit in which there are three amplifier stages and an attenuator pad. The stage
The overall gain of the system (in dB) is

\[ A = A_1 \times \text{Atten} \times A_2 \times A_3 \]
\[ = (20) \times (0.5) \times (15) \times (4) = 600. \]

When converted to dB, the gains are expressed as:

\[ A_1 = 26.02; \text{Atten} = -6.02; A_2 = 23.52; \]
\[ \text{and } A_3 = 12.04. \]

The overall gain of the system (in dB) is the sum (not product) of these numbers:

\[ A_{\text{v(dB)}} = A_1 + \text{Atten} + A_2 + A_3 \]
\[ = (26.02) + (-6.02) + (23.52) + (12.04) \]
\[ = 55.56 \text{ dB}. \]

The system gain calculated earlier was 600, and this number should be the same as above:

\[ A_{\text{v(dB)}} = 20 \log_{10} (600) = 55.56 \text{ dB}, \]
\[ \text{i.e., they are the same.} \]

Converting from decibel notation to straight gain notation is simple: solve either the power or voltage decibel equation for the ratio \( U_i / U_o \) or \( P_i / P_o \), as appropriate. For those who do not want to make the calculation, Table 1 shows common voltage and power gains and losses expressed both ways.

**Special dB scales**

Various people have defined special dB-based scales that meet their own needs. They make a special scale by defining a certain signal level as '0-dB', and referencing all other signal levels to the defined 0-dB point. In the dimensionless dB scale, 0-dB corresponds to a gain of unity (see Table 1). But if we define '0-dB' as a particular signal level, then we obtain one of the special scales. The 'dBm' scale is often used in RF measurements, defines 0-dBm as one milliwatt (1 mW) of RF power dissipated in a 50-\( \Omega \) resistive load.

**Attenuator circuits**

The basic circuit used for RF attenuators is the \( \pi \)-network of Fig. 2. This circuit is unbalanced with respect to ground, so it can be used for coaxial RF systems.

The values of the resistors \( R_a \) and \( R_b \) (two \( R_b \) resistors are needed) determine the input/output impedances and the attenuation ratio. The calculation for the resistors values for very precise attenuators is quite complex, but those readers who want to try it are advised to consult a standard reference work such as *Reference Data For Radio Engineers*. If you use the same 5th edition that I own, check Chapter 10. Other readers might want to use the values shown in Table 2. The resistances are from the table of standard resistor values, so are easily obtained. The attenuation will be close to the values shown in the left-hand column of Table 2, but there will be a small error due to the use of standard resistor values, rather than the 'formula values' (which tend to come out with unlikely

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**Table 1. Common gains/losses expressed in decibels (dB).**

<table>
<thead>
<tr>
<th>Ratio (out/in)</th>
<th>Voltage gain (dB)</th>
<th>Power gain (dB)</th>
</tr>
</thead>
<tbody>
<tr>
<td>1/1000</td>
<td>-60</td>
<td>-30</td>
</tr>
<tr>
<td>1/100</td>
<td>-40</td>
<td>-20</td>
</tr>
<tr>
<td>1/10</td>
<td>-20</td>
<td>-10</td>
</tr>
<tr>
<td>1/2</td>
<td>-6.02</td>
<td>-3.01</td>
</tr>
<tr>
<td>1</td>
<td>0</td>
<td>0</td>
</tr>
<tr>
<td>2</td>
<td>+6.02</td>
<td>+3.01</td>
</tr>
<tr>
<td>5</td>
<td>+14</td>
<td>+7</td>
</tr>
<tr>
<td>10</td>
<td>+20</td>
<td>+10</td>
</tr>
<tr>
<td>100</td>
<td>+40</td>
<td>+20</td>
</tr>
<tr>
<td>1000</td>
<td>+60</td>
<td>+30</td>
</tr>
<tr>
<td>10,000</td>
<td>+80</td>
<td>+40</td>
</tr>
<tr>
<td>100,000</td>
<td>+120</td>
<td>+60</td>
</tr>
</tbody>
</table>

---

**Table 2. pi-attenuator resistor values**

<table>
<thead>
<tr>
<th>Attenuation</th>
<th>( R_a )</th>
<th>( R_b )</th>
</tr>
</thead>
<tbody>
<tr>
<td>1 dB</td>
<td>6.2 ( \Omega )</td>
<td>910 ( \Omega )</td>
</tr>
<tr>
<td>2 dB</td>
<td>12 ( \Omega )</td>
<td>470 ( \Omega )</td>
</tr>
<tr>
<td>3 dB</td>
<td>18 ( \Omega )</td>
<td>300 ( \Omega )</td>
</tr>
<tr>
<td>5 dB</td>
<td>33 ( \Omega )</td>
<td>200 ( \Omega )</td>
</tr>
<tr>
<td>10 dB</td>
<td>75 ( \Omega )</td>
<td>100 ( \Omega )</td>
</tr>
<tr>
<td>20 dB</td>
<td>270 ( \Omega )</td>
<td>68 ( \Omega )</td>
</tr>
</tbody>
</table>

---

**Fig. 3.** a) Circuit for a selectable step-attenuator; b) digital driver circuit to replace switches.
values like '31.234232 Ω'.

When building an RF attenuator, be sure to use the inherently non-inductive carbon composition or metal film resistors. Wirewound resistors tend to be inductors, so will not work well in RF circuits. Also, be sure to use 5% (gold colour code tolerance band), or better, tolerance resistors or the accuracy will suffer accordingly. The 1% resistors are best, but are hard to come by in quantities of less than fifty each.

Another alternative is to purchase ready-made commercial attenuators in standard attenuation values. I have used fixed attenuators made by Mini-Circuits (P.O. Box 350166, Brooklyn, N.Y., 11235-0003) for a number of projects over the years. They also make MMIC amplifier chips, double balanced mixers and other RF and microwave products. The attenuator to select will depend on your application; Mini-Circuits offers both coaxial in-line attenuators and printed circuit mounted attenuators.

Figure 3a shows a multi-stage step attenuator project that you can build. I built this one from five Mini-Circuits printed circuit mounted attenuators offering 1, 2, 3, 6 and 10 dB of attenuation. Note that each section of the attenuator is shielded from the other sections. This is an effort to prevent leakage of signal from one section to another, resulting in a loss of attenuation due to capacitive coupling.

The attenuators are switched in and out of the circuit by cross-connected double pole double throw (DPDT) switches. In one position, the switch shorts the signal path from input to output, while in the other position the attenuator is in series with the path.

You can use either manual DPDT toggle switches, or do like I did and use a small DPDT electromechanical relay as the switch. Parts catalogues can show you a number of relays, many of which are built in dual in-line package (DIP) format for use on IC printed circuit board, that will do the trick. Some of these relays operate from as low as 5 V DC, and as high as 12 V DC, so are compatible with a wide variety of applications. Because they contain diodes across the relay coil, the relays are polarity sensitive, i.e. they must be connected to the DC power supply correctly or damage can result (polarity markings are usually on the package).

One advantage of using relays for the switches in the step attenuator is that you can either switch them with manual SPST switches (as shown in Fig. 3a), or use digital drivers, such as the open-collector TTL inverter shown in Fig. 3b. When the inverter is used, it becomes possible to control the degree of attenuation used from a computer output or other digital circuit.

Figure 4 shows the construction of the multi-stage step-attenuator that I built. It is housed in a die-cast aluminum box (Fig. 4a) that is fitted with a tight cover to eliminate RF leakage (take a look at the box you buy, not all of them are well made and will leak RF if the cover if poorly fitted). The internal shields (Fig. 4b) are made from brass flat stock sheet metal obtained in a local hobby shop (the kind that cater to model plane and train builders). RCA phono jacks are used for input and output, but I recommend BNC or SO-239 UHF coaxial connectors for most readers (match the other instruments or circuits in your system).

A receiver application of the attenuator is shown in Fig. 5. Here we have a switchable attenuator in series with the antenna line to a receiver (the attenuator should be mounted as close as possible to the receiver antenna jack). If a strong signal overloads the receiver (which can cause all manner of messy problems), then cut the attenuator into the circuit while listening to the station (set switch S1 to the 'B' position). When you look for other, less strong stations, remove the attenuator from the circuit by setting switch S1 to the 'A' position. This circuit will also work if you want to limit your tuning to strong stations, eliminating some of the crud underneath the blow-torches of the shortwave bands.

Another application is shown in Fig. 6. Here we see a monolithic microwave integrated circuit (MMIC) device that provides amplification from near DC to 2000 MHz. The particular device shown here is the Mini-Circuits...
MAR-xx series, but there are others available on the market that are based on the same concept. One problem with very wide band amplifiers is that the complex impedances seen by the input and output ports are sometimes capable of creating instability in the amplifier. Impedance variations can be snuffed out by using the brute force method of inserting a 1-dB or 2-dB attenuator in the input and output lines (as shown in Fig. 6), as close as possible to the body of the amplifier IC device. The value given for resistor $R_1$ applies when a supply voltage, $V_s$, of 9 V is used, and the amplifier IC is to work at 7 V. In all other cases, calculate $R_1$ from

$$R_1 = \frac{(V_s - U_{amp})}{0.015} \quad [\Omega]. \quad (3)$$

Attenuators are also used in making comparison measurements in RF circuits. Very often, because we lack well calibrated RF test equipment, we use the S-meter on a receiver as a relative indicator, and then use precision attenuators to adjust signal levels. For example, if you want to measure the gain of an RF preamplifier, you might note the 'with' and 'without' S-meter readings, and then insert the attenuator to see how much the S-meter drops. With this information you know the exact 'dB / S-unit' calibration of the meter.

**Conclusion**

RF attenuators are easy to build, and provide a lot of useful applications to the RF and ham radio experimenter. No good RF workshop should be without a collection of these useful devices.
DIGITAL-AUDIO ENHANCER

Design by T. Giesberts

When a variety of audio equipment (CD drive, D–A converter, CD player, DAT recorder) is connected together, jitter may often be present on the signals between the various units. Investigations have shown that although most people cannot hear jitter, it can be measured. The enhancer resulting from these investigations ensures that the digital signals are free of jitter.

The digital audio signals considered here are of the S/PDIF format as shown in Fig. 1 and 2. The bits to be transmitted are contained in the data signal, which is 'modulated' by the clock, resulting in the biphase mark signal—see Fig. 1. In this signal, a logic 1 is represented by a high-low transition (or a low-high transition) half-way through the bit to be sent and a logic 0 by the absence of such a transition. Furthermore, the logic level changes at the end of each and every bit. In this way the ones and zeros are not represented by levels but by the distances between individual transitions. The advantage of this signal is that it contains not only the data, but also the clock rate at which the data are transmitted. Knowing this rate is essential for the processing of the data after reception.

This signal is still not of the S/PDIF format, for in that each 32-bit word (subframe) is preceded by a preamble. This preamble is a fixed pattern of successive, non-modulated logic levels. The disruption of the biphase modulation is used for synchronization of the decoder. The preamble also shows the decoder what data are contained in the next word and whether the word is the start of a frame (1 frame = 384 words). To ensure a smooth transfer to the modulated data, the preamble may also be inverted, depending on the logic level at which the preceding word ended—see Table 1.

What is jitter?

Jitter occurs when a transition (edge) of a pulse is not exactly on time. It is a kind of frequency modulation that may be compared with the wow and flutter of analogue signals. It may also be noise or an FSK-like modulation on the clock signal. Analysis of such a signal shows that it consists of components as shown in Fig. 3. The top waveform indicates that the clock has two frequency components; the centre waveform is typical of a noise-like variation in the clock signal; the bottom graph shows what the clock should be: a single, stable frequency.

Jitter is produced partly in the equipment that provides the audio signal. Apparently, not all such equipment has a stable clock. Another source of jitter is the analogue transition of the data signal. Analogue? However you look at it, the signal, be it a voltage, a current, or a light-beam, passes through a cable with limited bandwidth and other deleterious properties. The first to be affected by these disruptive characteristics are the transitions (edges) of the signal. The restoring of these transitions after reception may not be wholly correct. In our investigation, the decoding of the clock signal from the biphase-modulated signal in several commercial units was affected by the preamble. This is because the 'not biphase' appearance of the preamble distorts the clock signal that is integrated with the data signal. When that clock is extracted with a straight phase-locked loop—PLL, problems arise because the PLL tends to track the distor-
### Table 1. Cell sequence with the preamble.

<table>
<thead>
<tr>
<th>Preamble</th>
<th>Previous cell 0</th>
<th>Previous cell 1</th>
<th>Word content</th>
</tr>
</thead>
<tbody>
<tr>
<td>B</td>
<td>11101000</td>
<td>00010111</td>
<td>data left channel</td>
</tr>
<tr>
<td>start of frame</td>
<td></td>
<td></td>
<td></td>
</tr>
<tr>
<td>M</td>
<td>11100010</td>
<td>00011101</td>
<td>data left channel</td>
</tr>
<tr>
<td>W</td>
<td>11100100</td>
<td>00011011</td>
<td>data right channel</td>
</tr>
</tbody>
</table>

The perfect clock is a single, stable frequency. The PLL serves to produce a clock that is as nearly free of jitter as possible.

#### Blocking jitter

The enhancer, whose block diagram is given in Fig. 4, is based on bistable ICs. The PLL produces from the input signal a clock signal that is highly stable

![Fig. 3. Depending on the source of the jitter, the clock signal may consist of 2 (or more) discrete frequencies or a band of noise. The perfect clock is a single, stable frequency.](image)

![Fig. 4. Block diagram of the digital audio enhancer.](image)

![Fig. 5. Oscillogram of the five signals highlighted in Fig. 4.](image)
stants, even during the initial synchronization phase of the circuit.

Although the output of the MMV is free of spurious frequency components, it may still contain jitter and, furthermore, its frequency is eight times too low. Both these complications are removed by a PLL. This circuit multiplies the frequency by eight and acts as a filter that reduces the jitter to nought for all practical purposes. The result is a very stable clock (signal 4 in Fig. 4 and 5). The first transition of this pulse is used to clock data bistable IC9. This transition occurs just after the onset of a cell in the input signal. The output of the bistable (signal 5 in Fig. 4 and 5) is thus a signal whose level is derived from the input, but whose transitions are 'new'. Note that the irregular reproduction of the clock on the oscillogram is caused by the digital scope. This instrument was not able to faithfully reproduce the very short (compared with the other signals) periods of the clock.

As you will know, digital audio processing requires three clocks (sampling frequencies): 32 kHz, 44.1 kHz and 48 kHz. This makes the timing of the circuits, which is determined by the PLL, quite complicated.

The PLL contains a crystal-controlled VCO (Voltage Controlled Oscillator). The control voltage for this circuit is divided into three ranges: one for each sampling frequency. From this voltage, a comparator can determine the clock frequency of the input signal.

The output of the comparator is used to set the VCO, a scaler and the MMV. The combination of VCO and scaler makes it possible to obtain three sampling frequencies from two crystal frequencies. Note that designs using one crystal to generate three frequencies, and those using three crystals require more components than the present design.

The sampling frequency, \( f_s \), must be doubled, because each sample must cater for the left-hand channel and for the right-hand channel. Then, it must be multiplied by 64, since each sample, after modulation, consists of 64 cells. The frequency so obtained is that at which the bistable is clocked. The crystal frequency, \( f_c \), is twice or three times that frequency (2 or 3 is the divisor of the scaler between the VCO and the bistable). That is,

\[
\begin{align*}
f_s &= f_c \times 2 \times 64 \times 2, \\
&= f_c \times 2 \times 64 \times 3.
\end{align*}
\]

These calculations yield a crystal frequency of 12.288 MHz for sampling frequencies of 32 kHz and 48 kHz, and of 11.2986 MHz for a sampling frequency of 44.1 kHz.

**Circuit description**

The input of the circuit—see Fig. 6—may be optical via IC1 or coaxial via K1. Jump lead JP must be set accordingly. The (0.5 \( V_p \)) signal at the coaxial input is raised to TTL level by IC9.

The signal is then amplified by IC10, from where it is passed to bistable IC12, whose action has already been discussed. The output of the bistable is applied to a transformer, where the signal level is (re)converted to coax level.

Apart from to the bistable, the output of IC10 is applied to a pulse shaper consisting of XOR gates IC12, IC11. The signal to one input of IC12 is delayed by IC12, IC13, while that to the other input is direct. This means that at every change of level the data at the inputs of IC12 are unequal for a short instant, which results in the output of IC12 briefly going high at each transition of the input signal.

The MMV, which has been discussed already, is composed of bistable IC12 and down counter IC8. The bistable is set via the clock input that serves as trigger input for the MMV. The bistable is reset by the down counter, which in this way thus determines the on time of the MMV.

Since in this arrangement the input signal can influence the bistable only when this has been reset by the counter, the MMV is not retrigerable.

The MMV action is as follows. When the bistable is not set, it enables terminal C3/G4 of IC10 for asynchronous input. The number that is input will be discussed later. When the bistable is set, the counter is enabled via EN 1 and counts down to zero in rhythm with the VCO frequency. At the instant the counter status becomes zero, the CT=0 output also becomes zero, whereupon the bistable is reset.

As already discussed, the VCO frequency is two or three times as high as the frequency that clocks each data cell into bistable IC8. Furthermore, the MMV is quiescent after the seventh and before the eighth cell. Considered in clock pulses for IC10, that is after the 14th or 21st and before the 16th or 24th. The design uses the 15th and the 22nd clock pulse. In other words, IC10 is reset at 15 or 22 via the data inputs. This is effected by two control lines that are each other's inverse (inverter IC13) and are derived from the comparator in the PLL.

The non-inverting output of IC10 is also the output of the MMV, which is linked to phase comparator IC1. This stage compares the clock derived from the input signal with the new clock. The result of this comparison appears as pulses at pin 13, which are applied to the loop filter of the PLL: \( R_c R_f C_f C_p \). This low-pass filter converts the pulses into a direct voltage that is used for fine control of the VCO. Because of the filter, the VCO reacts only to long or large deviations, such as different sampling frequency, but is insensitive to short and small deviations, such as jitter. The second comparator output is amplified by IC2, whose output is used to drive a LED. This diode indicates whether the frequencies at the comparator inputs are equal or not.

Detecting to which input frequency the circuit should lock is effected by splitting the control voltage provided by the loop filter into three ranges: 0.5–1.5 V for 48 kHz; 2–3 V for 44.1 kHz; and 3.5–4.5 V for 32 kHz, and using two comparators, IC8 and IC12, to react to these ranges. The thresholds of the comparators lie at 1.76 V and 3.24 V. The outputs of the comparators are applied to D1–D2, which indicate which sampling frequency has been detected. The LEDs are connected to the comparators in such a manner that only one can light at a time.

As has already been stated, only two VCOs, IC10 and IC10, are used to generate three frequencies. Basically these two oscillators are identical. Their control voltage is derived from the loop filter via IC4, which deducts an offset from this voltage and amplifies the resulting potential \( \times 5 \). In this manner, the ranges of 1 V per sampling frequency are converted to one range of 0–5 V. The offset is derived from potential divider RI1–RI4, via electronic switch IC7. This switch also ensures that the correct VCO is connected to the clock input of counters IC8 and IC12.

Earlier, it was stated that in the timing of the MMV it has to be ensured that the circuit reacts correctly to the sampling frequencies. Comparator IC8a, which has a threshold of 4.55 V, that is, just outside the control range for 32 kHz, prevents erroneous operation if the circuit is set to the 32 kHz mode and a 44.1 kHz signal is input. In that situation, the output of IC8a becomes low, so that the AND gate formed by D5–D6–R3 prevents the output of IC8a becoming high. The divisors of IC8 and IC12 are then, correctly, switched to 2 and 15 respectively.

Since the 1CT=0 output of IC10 is linked to the synchronous preset, G2, of IC10, this counter counts down to zero. Because it has to operate as a +2 or a +3 scaler, numbers 1 and 2 have to be added, for which purpose the clock is used. The output of IC10 is applied to bistable IC12, where it is used for removing the jitter, and to +8 scaler IC10, which feeds the signal back to the phase comparator.

To prevent mutilated data being output when the PLL has not yet been syn-

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**Fig. 6. Circuit diagram of the enhancer. Although the principle of operation is simple, the requirement for these different clock frequencies necessitates the use of quite a few components.**
chronized, the control signal of D4 is used to keep IC3, reset via inverter IC16. Pulse-stretching network D7-R30-C15 prevents the short control pulses of the PLL, which may temporarily be present on this signal, reaching the bistable during the synchronization process.

All that needs to be noted regarding the conventional power supply is that the input voltage of IC3 must be not less than 11 V nor more than 15 V. This means that, since the enhancer draws hardly any current, a transformer with a 9 V secondary may be used (which in these conditions provides about 12–13 V).

Construction

Although construction of the unit on the printed-circuit board shown in Fig. 7 is straightforward, a few points should be noted. Do not mount the crystals too close to the board to avoid any risk of their being short-circuited by the tracks. Preferably, insulate them from the board. If the VCOs do not oscillate readily, use either 22 pF capacitors or 1 kΩ resistors in the C3/R3 positions. It may also be necessary to change the values of C13/C11 and C12/C14, so that the control voltage at pin 1 of IC3, when the circuit is locked, is about 2.5 V.

Transformer Tr1 is a dry device that is wound on a Type C3-3/FT12 toroidal core that has also been used in several recent audio & hi-fi projects. Close-wind 26 turns of 0.5 mm d.t.a. enameled copper wire on to the core and solder the ends to terminals 1 and 2. Then, lay 6 turns of the same type of wire over the earlier turns and solder their ends to terminals 3 and 4. Note that the transformer (and C5, C6 and R9) are not needed if the enhancer is built into the digital-to-analogue converter. In that case, the data signal can then be taken from K9 (at TTL level).
IC4 = 74HC4046
IC5, IC6 = TLC272
IC7 = 74HC4053
IC8, IC9 = 74HC40103
IC10 = 74HC4040
IC11 = TORX173
IC12 = 7805
IC13 = 7808

Miscellaneous:
JP1 = 3-way header with jumper
K1 = audio socket for PCB
K2 = 10-way male box header
K3 = 2-way terminal block, pitch 7.5 mm
X1 = crystal 12.288 MHz
X2 = crystal 11.2896 MHz
Tr2 = mains transformer, 3.3 VA
9 V secondary (see text)

*high-efficiency type

or a wire link may be laid from the C5 to R3 positions on the board.

The enhancer, if not built in, must be connected to the digital-to-analogue converter with as short a coax cable as possible.

Test results

The test signal used was a 10 kHz signal with a sampling frequency of 48 kHz derived from a DAT recorder. This signal was applied via an optical cable and subsequently via a coax cable to the digital-to-analogue converter, first without the enhancer and then with it in both cases. It should be borne in mind that jitter is at least 100 dB below the level of the test signal. The analogue output of the converter was fed to a spectrum analyser after the 10 kHz signal had been removed with the aid of a high-quality notch filter to ensure that only the spurious remnants were analysed. The results are shown in Fig. 8-11: 8 and 9 relate to an optical connection between the DAT recorder and the D-A converter, and 10 and 11 to a coax connection. It will be seen that with the optical connection the enhancer improves the signal, whereas with the coax connection there is little difference between tests with and without the enhancer.
PARTS LIST

Resistors:
R1, R6 = 75 kΩ, 1%
R2, R32 = 100 kΩ
R3, R7 = 4.7 kΩ
R4, R20, R25-R30 = 1 MΩ
R5, R8, R18, R21, R33 = 10 kΩ
R9 = 390 kΩ
R10, R19 = 100 kΩ
R11, R14 = 562 kΩ, 1%
R12, R13 = 1.69 kΩ, 1%
R15, R17 = 1.82 kΩ, 1%
R16 = 1.54 kΩ, 1%
R22-R24, R34 = 1 kΩ
R31 = 1 kΩ (see text)
R35 = 22 MΩ

Capacitors:
C1, C6, C16, C17, C19-C22,
C25-C29, C32, C34, C36-C39
= 47 nF, ceramic
C2 = 39 pF
C3-C5, C23, C24, C31 = 100 nF,
ceramic
C7 = 100 pF
C8 = 1 μF
C9 = 2.7 nF
C10 = 180 pF (see text)
C11 = 220 pF (see text)
C12 = 22 pF (see text)
C13, C14 = 120 pF (see text)
C15 = 220 μF, 10 V
C18 = 22 nF
C30, C33 = 10 μF, 25 V
C35 = 470 μF, 25 V, radial

Inductors:
L1 = 47 μH
T1 = see text

Semiconductors:
D1 = LED, green*
D2 = LED, yellow*
D3, D4 = LED, red*
D5, D6 = BAT85
D7 = 1N4148
D8, D9 = BB212
B1 = B80C1500
T1 = BC557B
IC1 = 74HC04
IC2 = 74HC86
IC3 = 74HC74
IC4 = 74HC4046
IC5, IC6 = TLC272
IC7 = 74HC4053
IC8, IC9 = 74HC40103
IC10 = 74HC4040
IC11 = TORX173
IC12 = 7805
IC13 = 7808

Miscellaneous:
JP1 = 3-way header with jumper
K1 = audio socket for PCB
K2 = 10-way male box header
K3 = 2-way terminal block, pitch 7.5 mm
X1 = crystal 12.288 MHz
X2 = crystal 11.2896 MHz
Tr2 = mains transformer, 3.3 VA
9 V secondary (see text)

*high-efficiency type

Fig. 7. Printed-circuit board for the enhancer: track side.
The battery charger takes its name from the IC it is based on: the well-known Type U2400B processor. This IC was designed primarily for use in battery chargers and is, therefore, equipped with a number of specific features.

The processor controls the charging for a preset time at constant current, provided, of course, that the battery was fully discharged before charging began. To make certain that this is always so, the operation always starts with discharging the battery entirely. To that end, pin 10 of the device—see Fig. 1—is made logic high. An internal stage connected to this pin causes the battery to be discharged until the voltage at pin 6 has dropped to 0.525 V. At that instant, pin 10, and immediately thereafter pin 12, goes low and charging can begin. The low level at pin 12 serves to actuate charging circuit T₂-T₅. After charging at full current has taken place for a predetermined time, trickle charging begins. Trickle charging, which may continue for a long period of time, ensures that the battery capacity does not degrade during the life of the battery. It is effected by the processor driving pin 12 low for 100 ms every 16.8 s, which means that effectively the average charging current is reduced to 6/100 of its peak value.

The initial discharge facility may be disabled by connecting pin 6 to earth. This will be reverted to later.

The charging time is predetermined by an external clock at pin 16 or by an internal clock. The internal timer allows three periods to be preset: 30 min; 1 h; and 12 h. Charging during the 30 min and 1 h periods takes place at full current. During the 12 h period, the charging current is pulsed in a similar manner as in trickle charging but the 100 ms pulses now occur every 1.2 s. Effectively, therefore, the battery is charged at 1/9 of the full current.

The charging current may be altered with P₂. This preset varies the voltage across it, which changes the duty factor of an internal pulse-width modulator (PWM). This modulator, which drives the charging output, is clocked at a rate of 200 Hz. In this manner, the effective value of the charging current is made continuously variable.

To obviate any possible damage to the battery during fast charging, the processor has an internal failure circuit and two protection circuits: one against overheating of the battery and the other against overcharging the battery.

The battery temperature is monitored via an NTC (negative temperature coefficient) resistor, Rₗ, and pin 5. If the level at that pin drops below 0.525 V (equivalent to 40 °C), an error is registered. The NTC sensor must, of course, be in close contact with the battery. Should Rₗ itself, or its connections, become defective, the voltage at pin 5 will rise: when this exceeds 2.95 V, an error is signalled.

If the charging voltage per cell rises above 1.6 V, the potential at pin 4 will exceed 0.525 V, whereupon an error is signalled. The voltage of 0.525 V is obtained via a precise potential divider, R₁₀-R₁₅-R₁₆. The voltage across the

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**Fig. 1. Block diagram of Telefunken's Type U2400B battery charging processor.**

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**U2400B NiCd BATTERY CHARGER**

Design by U. Bangert and W. Ernst

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When a battery is then connected, the charging as long as the error persists. Once the error has been removed, charging is recommenced. If, however, a second error is registered, further action depends on the level at pin 15. If that pin is open, trickle-charging is substituted. In that case, D3 (red) lights. If, however, the reference voltage (3 V, pin 7) is connected to pin 15, as in the present circuit, D3 and D4 flash alternately. When the error is removed, continued flashing indicates that the remainder of the charging can be undertaken.

When the supply voltage is switched on, the U2400B is enabled; if no battery is connected, the red LED will light. When a battery is then connected, the potential at pin 4 will be 200–525 mV, provided that the voltage per cell is not lower than 0.6 V. After about 2 s (during which time the LEDs do not light), the processor will become fully operational. What happens if the cell voltage is below 0.6 V will be discussed later.

During the discharge phase, the red LED flashes; when charging is in progress, the green LED flashes. When the battery is being trickle-charged, the green LED lights continuously.

**Circuit description**

The mains transformer for the power supply is not shown in Fig. 2. It provides a secondary voltage of 18 V at a load current of 3 A. After the secondary voltage has been rectified and filtered, it is lowered to 8 V by regulator IC3, and used to drive control circuits IC1–IC3 and the ammeter. Since the secondary voltage may be as high as 30 V under no-load conditions, IC3 must not exceed 27 V. D3 is inserted in the line to the charging circuit.

The charging circuit, based on T2 and T3, is a simple constant-current source. When the level at pin 12 of IC1 is low, D1 is on, resulting in a constant voltage at the base of T3. Since the potential drop across the base-emitter junctions of Darlington pair T2 and T3 is a steady 1.5 V, the voltage across R56 and R57 is a constant 1.8 V. When S2 is open, the current through the transistors, that is, the charging current, is thus 100 mA. When S2 is closed, the emitter resistors are in parallel, so that the charging current is 2 A. Diode Dp prevents the battery discharging when the charger is switched off. Since the dissipation of T2 may be as high as 35 W (one cell and a charging current of 2 A), it is mounted on a suitable heat sink.

Since the discharge output is also controlled by the PWM, the positive pulses at pin 10 of IC3 charge capaci-
The battery voltage has been preset. Since the battery, Zener diode D₃ is connected to the number of cells contained in the battery. Zener diode D₂ protects pins 4 and 6 of IC₃ from too high a voltage if, for example, S₂ is set to the position for a single cell and a 12-cell battery is connected to the charging terminals. The potential at either of these pins must not exceed 6 V. This switch is a make-before-break type to minimize voltage surges during switching.

Rotary switch S₃ links the U2400B to the external clock, an astable multivibrator (AMV) based on the well-known Type NE555 timer IC. The clock rate is set to 1 Hz, that is, a period of 1 s, with P₁ (this is, by the way, the only calibration required). This switch is a break-before-make type to prevent short-circuiting of IC outputs. The outputs of scaler IC₂ provide six frequencies, corresponding to six different charging times—see Fig. 5.

The voltage set with P₂ controls the PWM and thus the charging current. The voltage range is extended by R₈ and R₉ so that at the two extreme settings of P₂, a narrow 'dead zone' exists. This ensures that the charging current can be set correctly between zero and maximum.

The NTC sensor is connected to the
charger via a switched jack socket, $J_1$. In this way, the resistor may be removed when a sealed battery is being charged. Since the sensor cannot come into close thermal contact with such a battery, temperature monitoring is impossible and it is, therefore, better removed. Resistor $R_5$ then simulates a 'cool' sensor, whereupon charging can commence.

The many 100 nF capacitors and inductor $L_3$ serve to decouple any incoming RF radiation and spurious voltage peaks on the mains supply.

**Construction**

The charger is best built on the PCB shown in Fig. 3. This board has been designed so that the operating controls, $S_7$ to $S_9$ and the LEDs are mounted on one side and the power transistors on the other side. The power transistors must be fixed with the use of insulating washers and plenty of heat conducting paste. A heat sink rated at 1.5 kW$^2$ is then fitted behind the transistors (again with plenty of heat conducting paste)—see Fig. 4. It is also possible to fit the power transistor to the inside rear panel of the enclosure and the heat sink to the outside (again, use plenty of heat conducting paste and insulating washers and bushes). If the heat sink is fitted inside the enclosure, make sure that there is adequate ventilation (drill a number of small holes in the top panel or fit a small extractor fan).

Comence the populating of the board with fitting the wire links. Load resistors must be mounted at least 10 mm ($1/8$ in) above the board’s surface. Resistors $R_{17}$ to $R_{37}$ must be soldered directly to the relevant terminals of $S_7$—see Fig. 2. Only two wires then link the switch to the board.

Preset $P_1$ is adjusted by using either an oscilloscope or a frequency counter between terminal B of $S_7$ and earth.

In the first case, set $P_1$ to obtain a period of 1 s; in the second case, adjust it to obtain a frequency of 1.0 Hz.

If neither of these instruments is available, set $P_1$ to the centre of its travel and $S_7$ to position 30 min (B). Select the number of cells ($S_9$) and the required charging current ($S_8$ and $P_2$). Switch on the mains, whereupon the red LED should light (do not insert the NTC sensor). Connect a battery to the charging terminals, when after two seconds the red LED should begin to flash, indicating that discharging has commenced. Shortly thereafter, charging begins and the green LED starts flashing. Note the time during which the LED continues to flash. Any deviation from 30 min should be corrected with $P_1$. This method may require the calibration to be repeated several times before the length of the period is right.

The ammeter is calibrated simply by measuring the voltage across $R_{37}$, preferably with a standard analogue voltmeter, setting $S_7$ to position 100 mA, and adjusting $P_3$ for maximum current. When the voltage across $R_{37}$ is 1.8 V, adjust $P_3$ for full-scale deflection (FSD) of the meter. This calibration should, of course, be carried out in the charging mode.

**Hints and modifications**

Before connecting a battery to the charger, make sure that the switches are set as required and that the red LED lights...
when the charger is switched on. If it does not, it may help to switch the mains off and then on again. These precautions are necessary in view of the fact that it happens once in a blue moon that undefined levels occur at the IC inputs when the switch is turned.

It is advisable to overcharge the battery by about 20% as a matter of course. That is, charge a 500 mAh battery as if it were a 600 mAh type. Divide this capacity by the desired charging time (in h) to arrive at the charging current to be set. Say, for example, that in the case of the 500 mAh battery the charging time should be 30 min. A quick calculation gives a charging current of 1.2 A. If fast charging is contemplated, make sure that the battery is designed to cope with it.

The charger is not able to register that a battery has been connected to the charging terminals when the voltage per cell has dropped below 0.6 V. Although such a battery can be given a quick (<1 min) boost by connecting it to a mains adaptor with current limiting. It is far better to use the additional circuit shown in Fig. 6 to cope with flat batteries. Spring-loaded switch S1 connects a 10 kΩ resistor, R bel, between pins 4 and 7 of IC3, thereby 'tricking' the circuit into starting charging. These additional components are best connected into the circuit at junction R14-R15 and the +ve terminal of C5. To prevent accidental operation of the switch, it is best to use a flush (set-in) type or to fit it at the rear panel.

It may also prove useful to connect a flush spring-loaded switch, S2, across capacitor C7 with which to end the discharge. When this switch is pressed, charging will commence.

If only small cells or batteries (up to no more than 600 mAh) are to be charged, it may be useful to restrict the peak charging current to 1.2 A. In that case, the rating of the heat sink may be reduced to 2.5 K W⁻¹. R39 to 1.8 Ω, 4 W, and the transformer secondary rating to 1.8 A. The scale of the meter must, of course, also be adapted.

If charging of 10-cell batteries only is foreseen, S1 may be a 10-way switch, R39 and R3 are not required, and the secondary voltage of T1 needs to be only 15 V. If only 6-cell batteries are to be charged, S2 may be a six-way type. R39 to R7 may be omitted, and the secondary transformer voltage needs to be 10–12 V only. In both cases, smaller heat sinks will suffice.

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Fig. 3. Printed-circuit board for the U2400B NiCd battery charger.

PARTS LIST

Resistors:
R1, R13 = 1 kΩ
R2, R7 = 470 kΩ
R3 = 1 MΩ
R4, R9, R30 = 10 kΩ
R5, R6 = 8.2 kΩ
R8 = 10 kΩ, NTC
R10 = 1.5 kΩ
R11 = 33 kΩ
R12, R16 = 100 Ω
R14, R15 = 820 Ω
R17–R27 = 12.7 kΩ
R28, R29 = 150 kΩ
R31, R34 = 47 kΩ
R32, R35 = 3.3 μΩ, 9 W
R33 = 680 Ω
R36 = 1 Ω, 5 W
R37 = 18 kΩ, 0.5 W
P1 = 1 MΩ preset
P2 = potentiometer, 10 kΩ, linear
P3 = 25 kΩ preset

Capacitors:
C1, C4, C6, C9–C12, C15, C17 = 100 nF
C2 = 470 nF
C3 = 15 nF
C5, C7, C8, C16 = 10 μF, 16 V, radial
C13 = 4700 μF, 25 V
C14 = 330 nF

Inductors:
L1 = 47 μH
Tr1 (not shown) = mains transformer, 18 V, 3 A secondary

Semiconductors:
D1, D6 = zener diode 3.3 V, 400 mW
D2 = zener diode 5.6 V, 400 mW
D3 = LED, 3mm, red
D4 = LED, 3mm, green
D5, D12 = 1N4148
D7–D11 = 1N5400
T1 = BUZ11A
T2 = TIP2955
T3 = BC557B
IC1 = NE555
IC2 = 4024
IC3 = U2400B
IC4 = 7808

Miscellaneous:
F1 = fuse, 4 A, with holder for PCB
K1, K2 = 2-way terminal block, pitch 5 mm
M1 = moving coil meter, 100 μA
S1 = 1-pole, 6-position rotary switch
S2 = 1-pole, 12-position rotary switch
S3 = single-pole on-off switch
Heat sink Type SK81 (1.5 K W⁻¹)
PCB Type 920098: see page 70.
Digital Audio/visual system (Multi-purpose Z80 card)

**May and June 1992**
An extensive description of a modification to the memory backup circuit on the Multi-purpose Z80 card is available free of charge through our Technical Queries service.

**FM stereo signal generator**

**May 1993**
Capacitors C17 and C19 should have a value of 33nF, not 3n3F as indicated in the circuit diagram and the parts list of the multiplex generator.

**Workbench PSU**

**May 1993**
The polarity of capacitor C15 is incorrectly indicated on the PCB component overlay (Fig. 5a), and should be reversed. The circuit diagram (Fig. 2) is correct.

Transformer TR2 is incorrectly specified in the circuit diagram (Fig. 2) and in the parts list. The correct rating of the secondary is $2 \times 12V/5A$. Also note that the secondary windings are connected in series to give $24 \text{V}$.

**Audio DAC**

**September 1992**
The polarity of capacitors C25 and C58 is incorrectly indicated on the component overlay of the D-A board (order code 920062-2), and should be reversed.

**CORRECTIONS AND UPDATES**

**U2400B NiCd battery charger**

**February 1993**
The value of resistors R17 through R27 should be 2.7kΩ, not 12.7kΩ as stated in the parts list.

**VHF/UHF receiver**

**May 1993**
In Fig. 4, the connections to ground of the AF amplifier outputs, pins 5 and 8, should be removed. The amplifier outputs are connected to the loudspeaker only. The relevant printed circuit board is all right.
A possible mechanism for the effects of electromagnetic fields on biological cells

By Dr K.A. McLauchlan, F.R.S.

It is widely believed that electromagnetic fields can affect biological systems, including man, although the evidence remains equivocal. Quite apart from the difficulty in establishing a causal relationship, the research field has been held up by a lack of a general mechanism to rationalize why such effects might be expected. Yet, at the same time, it is well established that even very small fields can affect chemical reactions in a way that is completely understood.

Here, we attempt to explain how, to suggest some possible relevances to the biological situation, and to predict what effects might be looked for to confirm the relevance of the mechanism there. This may help to target epidemiological studies on human systems so as to establish the perceived effects.

People are continuously exposed to electromagnetic radiation from their environment, which arises from the normal electric and electronic paraphernalia of modern living, and in various manufacturing processes. It has been believed for some time that it may produce harmful effects, and recommended levels of exposure have been defined in most developed countries. Such evidence as exists comes from both epidemiological studies and from direct biological experiments. The seminal example of the former is the investigation of the incidence of childhood leukaemia in Boulder, Colorado, by Drs Leeper and Wertheimer. They showed that the incidence, apparently, correlates with the proximity of overhead power lines to an alarming degree, although the strengths of the magnetic fields they measured were not substantially greater than that caused by the earth. Subsequent, and continuing, epidemiological studies appeared to confirm an effect, and have led to the re-siting of transmission lines away from sensitive regions of human habitation in some places, and to the detailed planning of their routes in others. Nevertheless, a recent working party of the Institute of Electrical Engineers concluded that epidemiological studies were not likely to confirm the effects, owing to the difficulty in excluding other possible environmental causes.

Many direct experiments on intact biological systems have been described, some with a disproportionate effect on this research field. Thus, Professor Loboff and his colleagues in the United States have published a series of articles which purport to show that very small fields varying at low audio frequencies have direct biological or biochemical effects in many systems. A typical example is the inclusion of $^{46}$Ca in human lymphocytes in which they reported a resonance effect, of greater or less incorporation of the radionuclide, depending on the field strength, at 14.27 Hz. In this, and the other results from Professor Loboff’s group, the resonance frequency appeared to correlate with the cyclotron resonance frequency of the isotope within the field, and this formed the basis of a general theory. Although this is un-physical, it was widely accepted.

More recently, Dr Lednev from Puschino, Russia, reported that the phosphorylation of myosin involving a calcium-binding protein was enhanced when the reaction system was simultaneously exposed to the earth’s field and a field fluctuating at 16 Hz. However, neither do sets of these experiments seem to have been satisfactorily reproduced in other laboratories; nor have the multitude performed on intact specimens. This is despite the expenditure of large funds in a world-wide experimental effort.

Some experiments are still accepted as significant, however, notably the demonstration by Dr Reba Goodman at Columbia that the production of messenger RNA from leukaemic cells is significantly enhanced by application of a 100 Hz field.

Recent reports from the National Radiological Protection Board indicate that it remains unconvinced by the epidemiological studies and sceptical that cellular DNA is damaged directly by exposure to electromagnetic fields.

A limiting feature of demonstrating that biological field effects do exist is that there is no accepted theory for them, which makes specific scientific tests nigh impossible. Furthermore, studies have not been made at the molecular level, where biological chemistry occurs. However, it is known that some chemical reactions—those that involve free radicals—are affected by applied magnetic fields in completely understood ways. What is more, the importance of free radical processes within the body is becoming more and more apparent. It is consequently possible to suggest a mechanism, capable of experimental test, via which fields might have a direct effect in the biological situation. It must be stressed that the link between the chemistry and the biology has not yet been demonstrated. It is to the chemical mechanism that we now turn.

Electrons and their spin in chemistry

The formation of chemical bonds involves the sharing of pairs of electrons between atoms, or common groups of atoms known as radicals (e.g., the methyl radical $\text{CH}_3$). Electrons possess the property of spin angular momentum, and relative to each other the resolved components of this (the ‘spin’) may be either parallel (a ‘triplet’, $T$, state’) or antiparallel (the ‘singlet’, $S$, state’). It has been known for many years that the electrons in the bond have antiparallel spins or, to put it in another way, electrons must have antiparallel spins to form bonds. Normal molecules therefore exist in $S$ states. When a molecule dissociates in a chemical reaction, one electron is taken by each radical and there results a pair of free radicals. These are highly reactive since each contains an unpaired electron. Only recently has it been realized that spin is conserved in this reaction, and that dissociation of a singlet molecule results in the production of a singlet ‘radical pair’; that is, at the time that the pair is formed, the relative electron spins are defined:
This is one of the two crucial features that underlie the effects of magnetic fields on chemical reactions. Since the free radicals are formed very close together in the solution, and they have the correct spin orientation for reaction, we should expect the bond to reform very rapidly, and no free radicals to escape from the initial region in which they are created. Free radical chemistry would exist only if the radicals reacted extremely rapidly with their immediate environment to prevent their recombination. This is, however, against experience, for we know that free radicals are present in many reactions where these competing reactions do not occur. We conclude that a mechanism must exist by which the reactive S state of the radical pair changes rapidly into the un-reactive T one, so that the overall process can be represented as:

\[ \text{molecule} \rightarrow S \rightarrow T. \]

The \( S \cdot T \) conversion is itself reversible, but molecular diffusion in solution normally suffices to separate the radicals before they re-attain the S state and are able to react.

There are, in fact, three T states whose energies become unequal when a magnetic field is applied to the system. They are labelled \( T_\alpha, T_\beta, \) and \( T_\gamma, \) where the subscript represents the resultant spin (each electron has a value of 1/2) in the direction of the field. Since the electron has an electric charge besides its spin, it possesses a magnetic moment. The implication is that two of these triplet spin states of the radical pair are magnetic, whilst the other is not. This is the next crucial feature underlying the normal effects of magnetic fields in chemistry.

**How conversion between the singlet and triplet states occurs**

The mechanism for the conversion between the S and T states of a radical pair consisting of two organic radicals is itself magnetic, and in applied magnetic fields below 10 000 gauss or so in strength is dominated by the hyperfine interactions between the electrons and nearby magnetic nuclei in the radicals. These are normally protons. In very low fields, including the earth’s field, this interaction is capable of changing the S radical pairs into all three types of un-reactive \( T \) state, and as a consequence the overall reaction probability fails (after the conversion has occurred) to one quarter of its original value. In the biological context, this may be a bad thing, for it implies that a greater concentration of possibly damaging radicals escapes into the surroundings from where the radical pair is formed. Hyperfine interactions are, however, very weak and they can only cause interchange between states which have energies that are very close. Now, since the \( T_\beta \) and \( T_\gamma \) states are magnetic, their energies are changed when an external field is applied whereas those of \( S \) and \( T_\alpha \) states are not. Once their energies differ from that of the S state by more than the hyperfine field, the states become inaccessible, and now this can turn into only one un-reactive \( T \) state: \( T_\phi. \) The effect of the field is, therefore, to double the reaction probability, and to lessen the number of potentially dangerous free radicals escaping the initial recombination process. This process is complete in fields of about 100 gauss, and no further field effects occur until very high fields (>10 000 gauss) are reached, when other mechanisms dominate.

The process whereby fields affect reactions is of an entirely kinetic nature and this simple description has included no discussion of the rate of the \( S \cdot T \) mixing process. It is not instantaneous, and the actual magnitudes of the field effects observed consequently depend intimately on the period available for interconversion; this implies that they are increased in viscous media, in micelles, in microemulsions, at surfaces, and so on.

Most environmental fields are very low indeed, and unfortunately it happens that there exists a further mechanism, demonstrable within the laboratory, via which the effect of the field is now harmful, in increasing the free radical concentration, rather than beneficial. Its origin is also in the hyperfine interaction, but it is complex and involves the conservation laws of angular momentum within and outside applied magnetic fields. These have the effect of restricting the mixing of certain hyperfine substates of the radical pairs in zero field but allowing them between all the \( T \) states in slightly higher ones. This effect starts at the lowest field strengths with possible environmental biological relevance. It implies that as the field is increased from zero, the recombination probability of the radical initially decreases before the effect discussed above takes over to increase it. There is consequently a maximum effect which occurs at fields less than 10 gauss for most chemical reactions.

Once again, the actual magnitude of the effect observed in this low-field region depends upon the diffusion conditions within the reaction system, but it is typically a few percent. This may seem small, but it is not necessarily negligible in the biological situation, as is discussed below.

**Experimental results**

A simple system for demonstrating that magnetic fields really do affect chemical reactions involving free radicals is one in which a product is formed by their recombination which fluoresces. The effect of the field is observed directly in the variation of the intensity of the emitted light as the applied field is changed. A suitable reaction involves pyrene (Py), excited to a singlet excited state in a light beam and then reacting with dicyanobenzene (DCB). In detail, this reaction occurs through the formation of a singlet exciplex, which is the entity that fluoresces.

In Fig. 1, the variation in light intensity is shown for the reaction studied following laser excitation of pyrene, and with a radical concentration of about \( 10^{-7} \) M, over a field range from zero to about 400 gauss. It shows a typical sigmoid dependence, with the reaction probability to form
The chemistry and its rationalization are shown in this figure.

Relevance to biology

The chemistry and its rationalization are firmly established, but the jump to biological relevance is at present conjectural. It is known that free radicals are mutagenic, and attack DNA directly. It may be suggested that any increase in the concentration of free radicals in the body is potentially dangerous, and the field, at low strengths, exactly causes this, although at higher values the reverse is true, and the field may be protective. This seems not to have been suggested before. The changes in concentration in this field region are only a few percent, and it might be asked if it is relevant. In the first place, it should be remembered that DNA is a chemical amplifier with a high gain, and, secondly, that the typical time scale of carcinogenic processes is long. In normal healthy people, efficient biochemical pathways exist to remedy the effects of damage to DNA, and the obvious experimental test for the type of mechanism suggested here is to look for field effects in biological systems in which this repair mechanism is already working hard, that is, ones already exposed to carcinogens. This experiment may already have been done, for there are several reports of the synergistic effects of fields with carcinogens. Similarly, epidemiological studies should be targeted on those already exposed to chemical hazards. No studies of this type appear to have been reported. What must be realized is that if this mechanism is the source of field effects on humans, the dose response curves are quite different from those of high energy radiation, for example, and are, in fact, represented by the experimental curves shown in the figures.

It is an intriguing thought that the application of magnetic fields at strengths above about 10 mT may provide increased protection from certain carcinogens to man.

Although this article has dealt with static fields, closely related ideas can be used to rationalize the effects of fluctuating ones, and combinations of the two.
A two-monthly column by Keith Hamer and Garry Smith

Welcome to this new two-monthly column in which we will be covering the fascinating hobby of long-distance TV reception, or DX-TV, as it is universally known. The hobby is in many ways analogous to short-wave listening where pleasure is derived from identifying distant sources of transmissions — the more elusive, the greater the satisfaction. Nowadays TV reception from other countries is taken for granted thanks to satellite technology, but it could come as a surprise to learn that TV signals have been crossing international frontiers for many decades.

DX-TV is well established in Europe and in the USA, and it continues to grow steadily in the UK. Unlike their amateur radio counterparts and short-wave listeners, many DX-TV enthusiasts still tend to be isolated with regard to accurate information for keeping up-to-date with changes in TV broadcasting in other countries. We hope that Elektor Electronics readers will be sufficiently enthused to write in with details of their reception, and help keep other enthusiasts informed by reporting changes in their local TV services. Many services now display a logo in the corner of the picture — details of your local networks would be appreciated, so that we can pass this information on to other enthusiasts via this regular column.

Atmospheric influences

The short-wave or medium-wave DX-er will be aware that these bands become active during the hours of darkness. Sunspot activity also plays a part, especially on the higher short-wave frequencies.

A similar situation arises with TV frequencies, although darkness does not improve the chances of reception. A check through the non-local channels on your TV receiver may reveal weaker relay stations which are always present. There will also be many blank channels. This can create the impression that nothing can ever be received on these channels. In reality this is not so.

Given favourable atmospheric conditions, broadcasts from TV transmitters located hundreds or even thousands of kilometres away can be received on these normally vacant channels. Unfortunately, periods of favourable atmospheric conditions on the TV bands are less frequent than at short-wave frequencies, which means that when conditions are favourable you must make the most of it! A little preliminary knowledge of when and where to tune in is therefore necessary; patience and perseverance are also required, especially when attempting to capture an elusive station.

Under normal reception conditions, the range of a signal emitted from a high-power transmitter is limited to approximately 100 km, although a sea-path can extend this range considerably. Certain atmospheric effects can extend the transmission range even further, albeit only temporarily. There are several types of propagation which can produce signals from distant transmitters. The most spectacular form is sporadic-E ionization, and this is the one which we will describe in this first DX television column.

Sporadic-E ionization

Long-range TV reception is possible due to reflections within the various layers of the earth’s atmosphere, including the E-layer. This particular region is located at approximately 12 km above the surface of the earth. Although it is capable of reflecting short-wave signals, TV signals normally continue through it, and into outer space. However, during the summer months the E-layer becomes highly ionized by the sun, and this can result in signals being reflected, or, more accurately, refracted, back to earth. Reception via sporadic-E can also occur during the winter, but on a much reduced scale.

The unstable nature of the E-layer means that this type of propagation is completely random in terms of direction, distance, duration and signal strength. Reception can last all day or for only a few minutes, but what surprises many newcomers to the hobby is the high field strength of many of the signals encountered, and the simplicity of the aerial required.

Since the signals are returned to earth, a skip distance is involved which is typically 1500 km. Occasionally longer range reception is possible from the Middle East, Africa and North America, but transmitters closer than 300 km are seldom received via this mode of propagation.

Sporadic-E ionization affects TV channels in the VHF bands between approximately 40 and 100 MHz. In Europe, this means that distant signals on the Band I channels E2, E3 and E4 can be received in addition to the channels of Eastern Europe and Italy. The FM radio band can also become highly active.

Tuning in

Unfortunately in the United Kingdom, the budding DX-er will encounter something of a technical obstacle when it comes to sporadic-E reception. This is because TV broadcasting is limited to UHF frequencies only, and as a consequence receivers rarely have a VHF tuner fitted. A few imported models have full VHF and UHF coverage, but these tend to be the exception rather than the rule. In such cases, the VHF tuning scales will be inscribed with channels 2-4 (Band I) and 5-12 (Band III), although this is still no guarantee that a VHF tuner is actually fitted. A frequency up-converter can be used to translate incoming UHF frequencies to a similar spectrum at UHF, but in practice these are only suitable for strong signals.

A special converter system for DX-TV, known as the D-100, has been available for a number of years. Using double conversion techniques, the converter system makes use of a TV receiver which functions as a second IF — the converter simply plugs into the HF aerial socket of a standard receiver. Dial tuning provides VHF and UHF coverage, and the IF vision bandwidth can be progressively reduced to enable weaker signals to be enhanced with the added improvement in selectivity. The unit can be connected to an FM radio to resolve any TV sound carrier even when a reduced vision IF bandwidth is selected. Automatic band-scan is a recently introduced option which allows the TV band to be previewed when reception is imminent, thus freeing the DX-er from constant manual searching.

Aerials for sporadic-E reception

A simple aerial known as a ‘ dipole’ can be used with an overall length of 2.6 metres. Larger aerials featuring a reflector and directors can be used if space permits — these will be more directional and provide increased gain when attempting to receive weaker signals. Band I aerials of this type are normally widely available outside the UK where this band may be used for local broadcasts. In the UK, the DX-er either has to resort to DIY construction, or purchase one from a specialized supplier.

The height of the aerial is not too important because sporadic-E signals arrive at a slight angle. However, a minimum height of 5 metres is recommended.

If a multi-element array is used, some method of rotation is advisable so that the aerial can be positioned for max-
maximum signal. A manually rotatable system is often easy to arrange, although an electric rotator is more convenient because it allows the operator to monitor the screen for best results.

It is advisable to use separate downleads from a multi-aerial installation rather than using a combiner unit at mast-head, feeding a single cable. This is because a combiner introduces a signal loss, which may not be noticeable on local-strength signals, but will certainly be apparent on weak ones. The aim is to preserve as much of the collected signal as possible.

**Be patient**

Regularly check Band 1 channels for signs of unusual activity including pattern from other transmissions in the band such as OIRT FM radio from Eastern Europe, or even cordless telephone carriers from distant countries. Don't forget that, because of the unpredictable nature of sporadic-E propagation, the band may be dead one minute, and choc-a-bloc with signals the next. So keep trying. The D-100 converter with band-scan is invaluable in this respect as it saves having to constantly twiddle with the tuning controls.

**The 1992 sporadic-E season**

Reception proved extremely rewarding last summer with signals encountered almost daily. June was the most productive month, but August was relatively quiet. Broadcasts from the new Baltic states have aroused much interest among DXers, but the lack of updated regional logos on the test card has caused a few problems with identification.

One of the highlights in the UK was a sustained transatlantic opening toward midnight on June 2nd which produced several French and English-language Canadian stations on FCC channels A2, A3 and A4. In the Netherlands and Germany, a number of Arabic countries were resolved during June. These have subsequently been identified from weather maps and news programmes originating in Jordan, Syria, Iran and Egypt.

Andre Gillé in France received the Rumanian test card (TVR Bucureshti) in glorious colour on channel R4 (85.25 MHz) during early June. The picture was remarkably clear and free of interference despite the channel being close to the FM radio band.

Marc Vissers in Belgium logged Tunisia on channel E4 at the end of May, with programmes accompanied by Arabic texts and a logo in the lower right of the screen.

Commenting on a successful season, Marc adds that MRT Pelister (Macedonia), also on E4, and TVM Kisinov (Moldova) on R3 have been identified more times this year than last. Moldova often carries the identification 'tvn' in the lower right hand corner, almost out of screen.

**Service information**

**Czechoslovakia:** a change from SECAM to PAL colour is underway, and test transmissions have already been received using the new standard. The 6.5-MHz sound spacing will be retained for the foreseeable future.

**Portugal:** the current identification caption is 'RTC', and the logo 'C1' appears in the corner of the picture.

**Netherlands:** the identification 'NOZEMA' has been added to the top of the PM5544 test card with the appropriate service below (NEDERLAND 1, etc.).

**Norway:** the transmitter name has been deleted from the test card this summer, and has been replaced by 'TELEVERKET' with a superimposed message relating to a change of an emergency telephone number over its lower half.

**CIS:** following reorganization, the state TV service 'Ostankino l' is broadcast via the former CT-1 network. Identification is easy — a logo consisting of a large enclosed figure 'I' is displayed in the corner of the picture. The news programme is called 'HOBOCTN' (Novosti). The 'RTV' network is shown over the former CT-2 channels. The news programme is called 'BECTN' (Vesti).

This month's service information was kindly supplied by HS Publications (Derby, U.K.), Gösta van der Linden (Rotterdam, Netherlands) and the Benelux DX Club. Please write to the authors at the address below with DX-TV reception reports and news about developments affecting local television services in your particular corner of the world.

Keith Hamer
7, Epping Close,
Derby DE3 4HR
ENGLAND
This month we close the article with a description of the control software developed for the DiAIV system. Also, we help you to test the main unit, and show you how to make a professionally finished infra-red remote control transmitter.

Design by A. Rietjens

The main unit can be tested only if it is linked to an input device, for which there are two options: a PC-XT keyboard and an infra-red (IR) remote control transmitter. If you have the PC keyboard, you can start testing straight away. Not so with the IR remote control, which has to be modified before it can be used. The self-adhesive front panel shown in Fig. 34 fits on an IRC301 remote control from Polcom. This foil is pre-punched to avoid having to cut out the small rectangular holes for the keytops. To enable the foil to be secured on the front panel, you first have to open the unit. Remove the printed circuit board, the rubber feet and the switches. Next, remove the original self-adhesive foil. Before fitting the DiAIV foil, check that this fits properly in the clearance. If necessary, cut off small parts until it fits exactly. Carefully stick the new foil on to the unit, and re-arrange the keytops according to the new layout. Coloured keytops may be used to highlight certain functions (see the front cover photograph of Reference 1). The ‘refurbishing’ work is illustrated in the photograph in Fig. 35.

Before closing the remote control again, we have to change its transmit address. A suggested address is 63. This can be achieved by leaving all jumpers on the RC-5 receiver module open, disconnecting pin 3 of the IR transmitter IC, and connecting pin 6 to pin 9 (see the example in the article covering the RC-5 IR receiver. Reference 2). The advantage of using address 63 is that the operation of the DiAIV can not be disturbed by other remote controls, such as those used for most audio/video equipment. Having re-assembled the remote control, you are ready to embark on testing the main unit.

Main unit testing

The main unit can only be tested if the Z80 card is fitted with the 27512 system EPROM and the two GALs that are supplied through the Readers Services as order code 6181. If the main unit has been constructed using the components listed in last month’s parts list, the test and further construction will not cause problems. We base our test on that of the multi-purpose Z80 card. The differences are listed below — the points correspond to the test procedure given in the article on the Z80 card.

1. No remarks.
2. ICs and ICs are now replaced by DiAIV decoders 1 and 2. Since we are to install 64 KByte of RAM, jumpers JP1 to JP9 have to be fitted at the RAM side. Also, since a 27512 EPROM is used, the memory configuration must be set to ‘type 3’ (con0 and con1 both at ‘1’).
3. IC1 is now the DiAIV EPROM (Type 27512). Since the MAX690 is not yet...
fitted at this stage. It may be necessary to manually reset the Z80 card (press S1).

4. Fit extra IC2 (43256). IC6 (Z80B-PIO), and connect the mini-keyboard described in Reference 3. On starting the system, the display indicates a RAM size of 64 kByte. Two beeps will sound, and the ‘halt’ LED will remain off. Since the DIAV control program and the Z80 BIOS are both contained in the EPROM, the software will enter the main program rather than the test routines. However, we first wish to test the card’s functions, and this can be achieved by pressing a key on one of the input devices just after the first beep. The system then enters the test routines described in the article on the Z80 card. You are now ready to check the operation of the XT keyboard, the infra-red remote control and the mini-keyboard (depending on what has been connected). If the mini-keyboard is not connected, random key codes may appear that upset the normal operation of the system.

5. No remarks.

6. Since the AD7569 is not used, this test is dropped.

7. Connector K9 is omitted because the main unit has no separate printer output. The dissolve unit may be connected straight to K9. The text ‘Test Dissolve Unit Y/N’ appears instead of ‘Test printer Y/N’. Selecting ‘Y’ (yes), causes all dissolve units to be driven one after another, while the display shows a certain pattern.

First, the lamp intensity changes from off to fully on. Next, the system issues nine forward and nine backward slide carrier commands. This is indicated by the displays counting up from 1 to 10, and down again. This test does not take account of any projectors actually connected. If there are, they are best switched off. If the IR remote control is used, check that the cursor keys can be used to toggle the default option: the indication ‘Y/N’ then changes into ‘Y/N’. Next, if any key is pressed, the test will be executed. Pressing ‘ESC’ or ‘N’ ends the test routines, whereupon the main program is started.

8. IC6 is already fitted, and reads the mini keyboard. The EXT1 and EXT2 connector may be used to hook up the timecode interface (or two of these using both connectors), as described in the previous instalment. This link is tested automatically by the DIAV software. In manual mode, ‘m’ is used to switch to monitor mode, and ‘c’ to set the speed as well as the inputs and outputs used (see the software description).

Having tested the basic functions of the main card as described above, we can concentrate on the DIAV control software proper.

**Main unit control software**

After switching on, or after running the built-in system tests, a copyright message appears on the screen, and you are prompted to press a key. Next, you are taken into the main menu as shown in Fig. 37. The indicated functions may be selected either by pressing the cursor keys, followed by a ‘return’ (‘GO’), or by typing the first letter of the option (this menu structure is, incidentally, supported by the Z80-BIOS, which may be of interest to those of you who use the Z80 card for applications other than the DIAV system). Below are brief outlines of all functions and features may be found in the DIAV user manual, which may be printed from the disk supplied with the software package.

The remote control key legend shown in Fig. 36 provides a rough indication of the available functions. The keys are arranged in functional groups wherever that was possible — the emphasis is on the most frequently used functions.

For instance, a separate key field has been reserved for the selection of effects, carrier movement and speed. Another field has been reserved for the main menu. Options of the main menu and submenus that are not indicated may be selected with the aid of the
cursor keys.

Below are short descriptions of a number of functions offered by the system. The idea is to give you an impression of how the DIAV is operated, rather than to render a complete functional description.

**Manual**
The system status menu appears after selecting the 'manual' option. This menu may be used at any time to see what is happening. An example state is shown in Fig. 38. An indication is provided about the current slide in the projector, and whether or not this slide is projected. In the top right-hand corner you can read the time sent to the
timecode interface, and, next to it, the setting of this interface. The effect selected is 'superimpose' at a duration of 2,500 ms (adjustable in steps of 100 ms using the '+' and '-' keys).

Evidently, one of the most important functions in manual mode is being able to control a number of projectors in the simplest possible manner, in combination with the infra-red remote control. The key sequence 'I', 'D', 'N' and 'GO' ('enter' or 'space') on the PC-XT keyboard, for instance, causes projector 1 to do a 'dissolve', while all other projectors that were switched on do a fade, followed by a forward carrier change. For the next dissolve action, all you have to do is press the 'GO' key, whereupon projector 1 is dimmed and changes, and the selected projector lights.

Since it is not easy to keep track of the projector selection, the following function has been added to the manual mode. After the system start, no projectors are selected. After each projector selection follows a comparison with the highest and lowest projector number entered so far. In case the new number (projector selection) is higher or lower than the one stored, the new value is stored, i.e., it overrides the previous one. In this way, the main unit 'remembers' the highest and lowest projector numbers. The last selection is stored, too. This data is used when the cursor-up and cursor-down keys are pressed. On pressing the '1', '4' and 'D' keys, for instance, the main unit stores the value '4' as the highest (and last) projector number, and the value '1' as the lowest number. The effect chosen in this way is 'dissolve'. If 'cursor-up' is pressed next, the next projector, '1', is selected to do the dissolve. If you press 'cursor-down', the previous projector, '4', will do a reverse slide change, after which a fade-over is done. Slide changing will only be done if this is possible, so that the carrier is not pushed out of the projector! In this way, you can easily do fade-overs with several projectors, without being concerned all the time with knowing the next projector number.

If the timecode interface is connected, and the C (code) command is used to select output B, all selected functions will be recorded on to tape in the order as entered. This allows a slide series to be programmed in real time. A better and more comfortable way of doing this, however, is to enter a series via the editor function, and have it played back in automatic mode.

**Automatic mode**

An automatic series (of which an example is shown in Fig. 39) consists of the same commands as those used in manual mode. Such a series can be entered into the system via a PC, via the editor function, or via the serial interface. Cues (effects) are separated by a space, or the 'end of line' command, which equals 'GO' in manual mode. The maximum number of characters per line is 80. To be able to keep the command lines apart, it is best to enter a short description of its function at the end. Everything typed after a semicolon (:) is treated as comment. It is up to you to decide whether only one cue is used, or more cues. By contrast, most professional systems allow only one cue to be programmed per line. The advantage of being able to put two cues in a program line is that a slide series can be entered very rapidly.
In tape mode, all commands over) to enliven the presentation. If in those locations where we want to synchronize with the sound track. In example (3), such a loop is only used to define the start point of the series. The complete series is started on detection of a pulse from the recorder. The waiting takes place in the loop. If we want to synchronize more accurately, for instance, once every four slides, the wait period is inside the brackets, so that every cycle can be synchronized individually (in example 4), the last 'W1000' is, of course, omitted). The series can be perfected further by replacing all 'W1000' commands by 'W' commands. However, in practice it then becomes difficult to get all pulses in the right places, while erasing one of them is not all that easy either. In such case, it is more convenient to write out the above example in the form of five separate lines, where synchronization takes place at one point in every line, and the absolute wait times are adapted for every individual slide. In example (5) we have added a command that 'closes down' the last projector, so as to create a good 'end' to the show.

Defining exact start points is much easier if the timecode interface is used. First, the timecode signal is recorded on a non-used track. During automatic mode, the time read back from the tape is taken as the system time. In the program, we enter 'XX:XX.XX' instead of the 'W' command (example 6). This enables the desired time to be filled in automatically on pressing a random key (7). In this way, the start point can be fixed by the timecode synchronization.

The timecode interface gives you a very accurate control over the whole timing of the slide presentation. Fine tuning is also much simpler than with pulse synchronization: after setting the times intuitively, you may correct them manually in very small steps (10 ms), as illustrated by example (8). So far, the examples have been relatively simple. We do, however, want to give a more extensive example that illustrates the use of only three slides (loaded on projectors 1, 3 and 4). Example (9) starts from a quenched projector, and fades in (i) a slide (3) slowly (x), after which a fade-over is done (II1 OL3/) into a second slide (1), whereupon a fade-out (0) is done. At the same time, a third slide (4) flashes (f) at a changing rate (w300, w500 and w1000). The images used for the fade-over could be those of a pop group, while the flashing slide provides the light effects. The user manual contained on the diskette supplied with the DIAV software package gives short examples to illustrate the effect of each of the user options.

**Debug and synchro**

Programming a slide presentation is, of course, never as simple as described above. In nearly all cases, errors will be made which require simple and effective debugging tools. For these, the DIAV control software offers you the 'debug' and 'synchro' functions. In debug mode, you can execute either individual line or each individual cue, in both directions! The projectors simply 'follow' the cursor, which enables you to spot the faulty command very quickly, and correct it using the editor mode. If you want to look for a specific position without seeing all cues along the program faithfully executed, use the 'synchro' function rather than 'debug'. Synchro allows you to step through the program without projector action. Once you have arrived at the desired position (and, if necessary, found the corresponding tape position), you can change to 'debug' or automatic mode to run or correct the series from there on. For this it is necessary that the projector carriers are automatically moved to the desired positions, and the relevant lamp intensities are set up. This allows the series to be continued from that point onward.

**SDL (Special Data Language) and tape mode**

After a slide show has been put together with the aid of timecode syn-

---

Fig. 40. In tape mode, all commands read from the tape recorder are displayed on the LCD.
chronization, it is desirable to link the control program to the sound tape. Apart from the series itself, the current projector parameters (lamp intensity, carrier position) have to be recorded also. Being able to read this information from tape at any tape position allows us to adapt the projector states as required. This is achieved by the SDL code, which, apart from the commands that make up the series, also contains the current projector parameters such as the effect chosen, fade-over time and carrier position. Writing to tape is effected in parallel with the music, so that the series is securely linked to the location where 'it is to happen'. If the music tape is started somewhere in the middle of a series, the DIAV main unit is still capable of moving the projector carriers to the desired slide position, and start again from there. In practice, this will not be done too often, but such a test clearly demonstrates the error correcting capabilities of the SDL code. If, during the show, errors occur owing to tape drop outs, you can be sure that the system is capable of correcting these, and making the show continue practically without a hitch.

Furthermore, the SDL code itself is protected against read errors. All codes receive an error correction block before they are sent to the tape. This is achieved by 'Hamming' encoding, a technique widely used in telecommunications. This encoding system affixes a number of extra bits to each byte to be transmitted: three for each nibble. That is why an SDL byte is actually transmitted as two bytes. The structure of the Hamming codes enables a byte of which one bit goes wrong to be reconstructed on the basis of the remaining information. In short, this results in an extremely reliable decoding process.

The SDL code is sent to tape via the 'B' output on the timecode interface. In manual or automatic mode. During projection in automatic mode, the timecode interface is automatically given the proper settings, and the SDL code is always transmitted. In manual mode, the SDL codes are only transmitted if output 'B' has been selected on the timecode interface, with the aid of the C (code) command.

Reading and recording SDL codes takes place in 'tape' mode. Commands read from tape are sent to the main unit display in chronological order (see Fig. 40).

**Remaining functions and features**

The functions described above form the core of the software. In addition to these, there are functions to control the RAM disk (read, write, erase and directory), and to change a miscellany of basic settings (with battery backup), including the 'one-button/two-button' projector type selection. A full description of these features may be found in the user manual contained on the system diskette.

**DIAV software for IBM PCs and compatibles**

Since it is often more convenient to type in a lengthy program on a larger screen, the DIAV system has been endowed with a program download feature that works in conjunction with IBM PCs and compatibles. The download utility allows you to control the main unit (via the PC's COM1: or COM2: serial port) as if this had its own, local, keyboard. In addition to this function, the download utility also allows programs to be sent to, and read back from, the main unit. Additionally, the software offers the possibility to make a back-up copy of the entire RAM-disk. This may be done to save complete programs on to hard disk or floppy disk, from which they can be retrieved and downloaded to the DIAV system at any time. Making back-up copies of programs on disk is good practice because the memory back-up battery in the main unit will not last forever! It should be noted that the DIAV system disk does not contain software to drive the dissolve unit or the timecode interface from the PC.

In contrast with the main unit local control software, the PC DIAV down-loader does not contain a text editor. However, any word processor capable of saving and loading straight ASCII files may be used for this purpose, even the most rudimentary of these. EDLIN (found in your DOS directory). To make sure that the DIAV programs are easily found back, they are best stored in a separate subdirectory.

**Conclusion**

This last article installment has aimed at giving you a first impression of the many features of the DIAV system, without attempting to be complete in any way. As already mentioned, all system functions are described and illustrated at length in the 40-odd page user manual which you can print from the system diskette.

After the first faltering steps in operating and programming the DIAV system, you will soon get the hang of it, and agree with us that things look a lot less complex than initially feared. Every module, however, takes some getting used to. For example, the comfort of using the timecode interface will not be appreciated until you have grown accustomed to using this unit.

**References:**

I²C OPTO/RELAY CARD
PLUS A NEW I²C DEVICE DRIVER (V 1.1)

The popularity of our series on I²C devices is still growing, and that is why we venture to propose a real-world interface for the I²C bus. The eight channels on the interface described here are electrically isolated by optocouplers or relays, and operate under the control of I²C bus commands. If you need more than eight channels, it is possible to put up to seven more cards on to the bus.

Design by J. Ruiters

Since its introduction, the software we proposed to control I²C devices via a PC has been improved in a number of ways. Before discussing the relay/opto card, we will have a look at the improvements.

A time-out check has been added to the Wait Byte routine contained in the device driver I2CDRV2.SYS (I2CDRV2.ASM). This addition produces an error code if the specified wait time (default: 1 s) is exceeded when waiting for a byte to arrive via the I²C bus. This effectively prevents the PC being lost in an endless wait loop when there are errors in an I²C communication.

Furthermore, one bug has been fixed, and the communication between the PC and the PCD8584 (I²C bus controller) has been improved. The latter modification was made when we found that the CS pin has to be inactive sufficiently long in between two cycles. The datasheet mentions the following in this respect: “If the I²C controller runs at a clock speed of 8 MHz or 12 MHz, the minimum number of clock cycles between any two actions on the data bus is six. At lower clock speeds, this may be reduced to three cycles”. Mind you, the datasheet mentions the controller’s clock speed, not that of the PC. In practice, the modern generation of PCs, for example 486-core 33-MHz machines, are far too fast for the I²C controller. To solve the problem, a subroutine doing a number of NOPs (no operation) is called after each input or output instruction. The number of NOPs required depends on the clock speed of your PC, and may be defined by trial and error (as indicated by the datasheet, there is a certain spread). The number of NOPs inserted by the driver is conveniently defined via the CONFIG.SYS file, which may look like this:

DEVICE = I2CDRV2.SYS B:300 C:0 D:00FF

where the ‘D’ parameter indicates the number of NOPs. The default value is 00FFH, which works fine on a 33-MHz 486 PC. The optimum value (i.e., the smallest possible value) may be established by experiment only. The smallest delay you can enter is 0001. Finally, the driver now includes routines that read the ‘D’ parameter, and puts it on the screen. These routines go by the name read_nop_num and print_nop_num.

The PASCAL unit for the I²C interface has also been modified slightly. The new unit, I2C2.PAS, enables error reports to be handled from within the main program. The new routine is contained in the G_IO subdirectory on the floppy disk, together with G_IO.PAS, a demo for the opto/relay card.

The new software may be obtained as order code 1821, and replaces the old version, 1671. For price and ordering information, you are referred to the Readers Services on page 70. Please note that the older version can not be exchanged for the new one.

The opto/relay card

The circuit diagram of the opto/relay card is drawn in Fig. 1. The heart of the circuit is formed by IC1, a PCF8574. This IC may be familiar from the A-D/D-A card in the I²C module series (Ref. 1). The PCF8574 is an 8-bit parallel I/O port located at base address 4xH on the I²C bus. The value of ’x’ is determined by jumpers JP1, JP2 and JP3 which are connected to address lines A0, A1 and A2. The least significant bit of the address byte is used to indicate write or read actions to or from the IC. In binary form the address looks like this:

0 1 0 0 A2 A1 A0 R/W

where W\ stands for the inverted write signal. The jumpers thus allow you to select between eight addresses, which is also the maximum number of opto/relay cards that can be put on to the I²C bus. However, any other cards that have a PCF8574, such as the A-D/D-A, are to be included when tolling up!

Although the circuit diagram would appear to show that the databits are connected to a relay and an optocoupler, they are, in fact, connected to a relay or an optocoupler. The selection between optocoupler and relay is made by a jumper on the printed circuit board. The two can not be used at the
same time, which would be very strange anyway from a point of view of electronic design.

None the less, the opto/relay circuits are designed such that the same PCB layout is used to build either an optocoupler input or a relay output. It should be noted that the bits defined as an input must be initialized by writing logic 1s to them.

The relays used are polarized types. The cores of the relay coils contain small permanent magnets whose force is just too small to pull in the contacts. By virtue of the magnetic field built up by the permanent magnet, the coil has to overcome only a small force to pull in the contacts and keep them closed. The advantage is clear: the coil current required to energize the relay is considerably reduced. However, this is only true if the coil is supplied with the correct polarity. If not, the purpose of the permanent magnet is defeated, since the coil and the core then operate against each other.

The power supply on the card contains a jumper (JP4) that enables the relays to be powered by a separate supply. This may be necessary in some cases to reduce the total load on the I²C bus 5-V supply line. If the jumper

**MAIN SPECIFICATIONS**

- **Relay outputs**: 2 A; 150 Vdc; max. 60 VA
- **Port outputs**: 5 V; 25 mA sink; 400 µA source.
- **Opto inputs**: 5-15 V (4-14 mA) with polarity reversal protection. Other ranges possible by changing R1-R8.
- **Port inputs**: TTL level
- **Base address**: 4xH (max. eight PCF8574 modules on bus)

---

**Fig. 1.** The circuit diagram of the 8-way I²C input/output card is simplicity itself. Note that only one LSI IC is used.
You do not need to resort to a war. The artwork of the single-sided construction of the present I is used, since the 5-V supply in the PCB can be seen, it is impossible to connect may then omit components ICs. C8, C11 and D17 (supply reversal protection).

**Construction**

The construction of the present I²C extension card is entirely straightforward. The artwork of the single-sided printed circuit board for the I²C opto/relay card is given in Fig. 2. As you can see, it is impossible to connect an optocoupler and a relay to one and the same bit.

The parts list indicates components required for an 'input' configuration (optocoupler) with one asterisk (*), and components required for an 'output' configuration (relay) with two asterisks (**).

The specification and the PCB layout make it quite clear that the circuit is not suitable for connecting to the mains.

**Testing 1, 2, 3...**

The program G_IO.PAS may be used to test the hardware, provided it is adapted to match the input/output configuration of the card. It is also eminent for the design of your own applications. Those of you who do not have a Pascal compiler may use the executable file G_IO.EXE. Note, however, that G_IO.EXE assumes that there are relays in positions Re1-Re4, and optocouplers in positions ICs-ICs. This program runs a switching sequence on the relays, and subsequently puts the current input levels on the screen.

G_IO.EXE also assumes that the opto/relay card is located at address 40H (JP1-JP3 in position 'A').

**Reference:**

### Components List

- **Components for input configuration (optocoupler):**
  - * = Components for output configuration (relay).

#### Resistors:
- 8 $1k\Omega^*$
- 2 $330\Omega$
- R1-R8
- R9; R10

#### Capacitors:
- 8 $820pF^*$
- 1 100µF 16V radial C9
- 2 100nF C10; C13
- 1 1µF 16V radial C11
- 2 10µF 16V radial C12; C14

#### Semiconductors:
- 8 1N4148**
  - D1; D3; D5; D7
  - D9; D11; D13; D15
- 8 1N4148*
  - D2; D4; D6; D8; D10; D12; D14; D16
- 1 1N4007 D17
- 8 CNY17-1** IC1; IC8
- 7805 IC9
- 1 PCF8574 ** IC10

#### Miscellaneous:
- 8 3-way PCB terminal block; pin grid 5mm K1-K8
- 2 6-way PCB-mount mini-DIN socket K9; K10
- 1 16-way angled box header K11
- 8 V23040-A0001-B201 relay (5V; 320Ω) Re1-Re8
- 1 Printed circuit board; order code 930004 (see page 70)
- 1 Software on disk; order code 1821 (see page 70)
- 1 Siemens product. Suggested retailer -
1.2 GHz MULTIFUNCTION FREQUENCY METER

PART 3: THE PC LINK

Design by B.C. Zschock

This month's installment explains how the frequency meter can communicate with a PC via an RS232 serial link. Obviously, to be able to write and download user control programs to the instrument, you require full descriptions of the various functions available. These descriptions are given here, but before we start listing them it must be mentioned that PC control of the counter is optional, and offers no extra functions over and above manual control (described last month). So, do not worry if you do not have a PC — this installment is simply not for you. The PC link is, however, a jewel for anyone involved in automated test, measurement and data logging systems, since it allows measurements to be set up and timed, and measurement data to be collected, in the absence of an operator.

The control of the instrument via the PC is basically the same as manual control. Last month we already mentioned that manual control (i.e., pressing the keys on the instrument) builds a string of commands that is stored internally and executed after the start command. Well, the PC link does the same: commands are entered into the instrument in the form of codes, which you may have spotted already in the large menu overview presented in part 2 of this article (hexadecimal codes in the upper left hand corners of most of the boxes). The PC link uses the same command storage locations in the instrument as manual control, while the same options and parameters are available. The measurement results stored in the instrument may be read by the PC for processing. In the following description, the storage locations addressed by the PC are called registers.

Hardware

Two points should be noted in relation to the RS232 link between the counter and the PC. First, the PC should have an active RS232 port, because the RS232 line voltages are used to power the level converter in the counter. Second, bear in mind that the counter’s ground is connected to the RS232 interface ground, which means that it is also connected to the mains earth via the PC. Consequently, measurements carried out with the counter are not potential-free, which requires extra care in some cases. An electrically isolated RS232 interface that can be used on the counter is planned for a future issue of Elektor Electronics. Such an interface enables a potential-free measurement system to be set up.

PC link basics

Start by connecting the counter’s serial port to the RS232 port on the PC. Next, initialize the PC port so that it runs at 2,400 baud, 8 data bits, no parity. This may be done under DOS using the MODE command, or by a communication program. Make sure that the RTS line is properly initialized. Although this is done automatically by the DOS MODE command, it may have to be arranged otherwise on other computers.

The next phase is simple: switch the counter to its start (default) state. This is achieved by switching on the instrument, or pressing the BREAK key.

The communication between the PC and the instrument is based on the full ASCII code set. The characters 0H to FFH are used for the commands, and the remainder of the set (20H and higher) to identify the functions (command parameters). ASCII values up to FFH have three-letter codes which are remnants from the telegraphy age. These letter codes are used in the descriptions below, instead of the hexadecimal values. An overview of ASCII values and corresponding telegraph codes is given in Table 1.

So far, so good. Now, the PC waits for data. If it receives a DLE code (10H), this must be taken to mean that connection is not possible. In that case, the counter has to be switched off and on again, or force it to transmit a SYN (16H) code by pressing the BREAK key. Reception of a SYN code at the PC side means ‘connection possible’. The ensuing wait state is marked as ‘NC’ (for ‘not connected’) in the following diagrams. Manual control of the counter is only possible in NC mode.

On reception of a SYN code, the PC can request the connection to be set up by issuing an ENQ code. In response, the counter establishes the link, and supplies an ACK (acknowledge) code. From then on, the counter is in connect mode (C), which puts the PC in control. In connect mode, a measurement can only be started. All setting-up has to be done beforehand. Thus, before a measurement can be started, the PC must send a series (string) of commands and parameters that selects, for example, the counter function, the input, the gate time, etc. This can only be done in command entry (CE) mode, which is selected by sending an STX code to the counter. In this mode, each received code is written into the counter’s command memory, until the command is complete. An ETX code switches the counter back from CE mode to C mode. The command string built by the PC is fetched from the memory and executed on receipt of a DC2 code.

Sending an EOT code enables the PC to take the counter out of C mode, and back to the start (default) state.

---

Table 1. ASCII command set

<table>
<thead>
<tr>
<th>Decimal</th>
<th>Hexadecimal</th>
<th>Code</th>
</tr>
</thead>
<tbody>
<tr>
<td>0</td>
<td>00</td>
<td>NUL</td>
</tr>
<tr>
<td>1</td>
<td>01</td>
<td>SOH</td>
</tr>
<tr>
<td>2</td>
<td>02</td>
<td>STX</td>
</tr>
<tr>
<td>3</td>
<td>03</td>
<td>ETX</td>
</tr>
<tr>
<td>4</td>
<td>04</td>
<td>EOT</td>
</tr>
<tr>
<td>5</td>
<td>05</td>
<td>ENQ</td>
</tr>
<tr>
<td>6</td>
<td>06</td>
<td>ACK</td>
</tr>
<tr>
<td>7</td>
<td>07</td>
<td>BEL</td>
</tr>
<tr>
<td>8</td>
<td>08</td>
<td>BS</td>
</tr>
<tr>
<td>9</td>
<td>09</td>
<td>HT</td>
</tr>
<tr>
<td>10</td>
<td>0A</td>
<td>LF</td>
</tr>
<tr>
<td>11</td>
<td>0B</td>
<td>VT</td>
</tr>
<tr>
<td>12</td>
<td>0C</td>
<td>FF</td>
</tr>
<tr>
<td>13</td>
<td>0D</td>
<td>CR</td>
</tr>
<tr>
<td>14</td>
<td>0E</td>
<td>SO</td>
</tr>
<tr>
<td>15</td>
<td>0F</td>
<td>SI</td>
</tr>
<tr>
<td>16</td>
<td>10</td>
<td>DLE</td>
</tr>
<tr>
<td>17</td>
<td>11</td>
<td>DC1</td>
</tr>
<tr>
<td>18</td>
<td>12</td>
<td>DC2</td>
</tr>
<tr>
<td>19</td>
<td>13</td>
<td>DC3</td>
</tr>
<tr>
<td>20</td>
<td>14</td>
<td>DC4</td>
</tr>
<tr>
<td>21</td>
<td>15</td>
<td>NAK</td>
</tr>
<tr>
<td>22</td>
<td>16</td>
<td>SYN</td>
</tr>
<tr>
<td>23</td>
<td>17</td>
<td>ETB</td>
</tr>
<tr>
<td>24</td>
<td>18</td>
<td>CAN</td>
</tr>
<tr>
<td>25</td>
<td>19</td>
<td>EM</td>
</tr>
<tr>
<td>26</td>
<td>1A</td>
<td>SUB</td>
</tr>
<tr>
<td>27</td>
<td>1B</td>
<td>ESC</td>
</tr>
<tr>
<td>28</td>
<td>1C</td>
<td>FS</td>
</tr>
<tr>
<td>29</td>
<td>1D</td>
<td>GS</td>
</tr>
<tr>
<td>30</td>
<td>1E</td>
<td>RS</td>
</tr>
<tr>
<td>31</td>
<td>1F</td>
<td>US</td>
</tr>
</tbody>
</table>
**Table 2. Counter control function overview.**

<table>
<thead>
<tr>
<th>Line structure:</th>
<th>Command</th>
<th>Hex code</th>
<th>Description</th>
</tr>
</thead>
<tbody>
<tr>
<td><strong>General control functions</strong></td>
<td>ENQ</td>
<td>005</td>
<td>Call to set up connection</td>
</tr>
<tr>
<td></td>
<td>ACK</td>
<td>006</td>
<td>Command character received o.k. and accepted</td>
</tr>
<tr>
<td></td>
<td>NAK</td>
<td>015</td>
<td>After ENQ: call can not be accepted. else: faulty character received</td>
</tr>
<tr>
<td></td>
<td>EOT</td>
<td>004</td>
<td>End of communication</td>
</tr>
<tr>
<td></td>
<td>SOH</td>
<td>00E</td>
<td>Next character determines the number of transparent characters to follow (usually number of bytes)</td>
</tr>
<tr>
<td></td>
<td>RS</td>
<td>01E</td>
<td>Reset</td>
</tr>
</tbody>
</table>

| **Control functions from PC to counter** | STX     | 002      | Start of command communication |
|                                           | CAN     | 018      | One function back in command |
|                                           | HT      | 009      | One function forward in command |
|                                           | VT      | 00B      | Pause duration follows |
|                                           | ESC     | 01B      | Reserved |
|                                           | ETX     | 003      | End of command communication |
|                                           | DC1     | 011      | Register follows |
|                                           | DC2     | 012      | Execute command in command memory |
|                                           | DC3     | 013      | Return register set |
|                                           | DC4     | 014      | Return command from command memory |
|                                           | US      | 01F      | Break (interrupt measurement) |

| **Control functions from counter to PC** | SYN     | 016      | Ready for communication; set up link |
|                                          | SO      | 00E      | Ready for command reception |
|                                          | FS      | 01C      | Receive character, wait for answer |
|                                          | GS      | 01D      | Command memory full |
|                                          | DLE     | 010      | Stop communication; break link |

**State diagrams**

Figures 9 and 10 illustrate the communication flow between the PC and the counter with the aid of so-called state diagrams. To explain what happens, we take the state diagram in Fig. 9 as an example. All diagrams have a dashed vertical line between the computer side (left) and the counter side (right). The horizontal lines mark intermediate states. An arrow pointing to the PC indicates a character being sent by the counter to the PC. Likewise, an arrow pointing to the counter indicates a character sent by the PC to the counter. The relevant character is found to the left or the right of the arrow. After a character has been sent, a new intermediate or final state arises. Occasionally, an arrow remains at the counter side, while it may also go back to an earlier state. In both cases, a state change occurs that is not caused by the 'other side'. An action can occur (Fig. 10c) during such states.

Two parallel lines indicate an intermediate state with a pause, of which the duration may be programmed by control function VT.

Returning to our example in Fig. 9, this can be read as follows: In the default state (not connected, NC), the counter sends a SYN code to the PC. After a short pause, the PC responds with an ENQ code. This prompts the counter to change to connect mode (C), which is confirmed by sending an ACK code to the PC.

If the start or end state in a diagram is marked by 'C/CE' (connect/command entry), the diagram is valid in connect mode as well as command entry mode. Further, the following conventions apply: a register number (1,…,n) is always conveyed as an ASCII number; the number of bytes <Nmb> as a binary value; the register identification (R, I) as an ASCII character; the pulse duration <#PD> and the register bytes <Byte> as binary values.

Figures 10a to 10p show state diagrams for each possible configuration. Figures 10a to 10e illustrate the command strings used to change between the three modes: not connected/default (NC), connect (C), and command entry (CE). In the default mode, the only valid code is an ENQ received from the PC. All other codes received cause the counter to return a DLE, and refuse to perform the mode change (Fig. 10b).

Whenever the counter transmits more than two characters, for instance, when conveying register contents, or the command string, a 'serial data input overflow' may occur in the PC. To prevent this, the counter has a built-in wait loop, which inserts a short pause between transmitted characters. The pause duration ( pacing) can be set to a value between 1 (short) and 256 (long) by the PC (Fig. 10f). After a reset on the counter, or after it is switched on, the pacing is 256.

With some types of measurement, for instance, frequency and revolution counter, the instrument expects a number of parameters, such as the gate time, before the start command. Such a parameter must be written into the relevant register by the PC (Fig. 10g). The register designations are listed in Table 2. Register loading is started by a DC1 code, which can be used to convey only one register (contents), marked by 'R' or 'I' and the associated number (1,…,n). The following is important: the counter transmits <Nmb> (for 'number') in binary code to tell the PC how many characters are expected. The PC must start the message with the lowest-value byte. If, for instance, the counter requests five bytes, the PC must send exactly this number — not more, not less. This
means that the PC may have to transmit nulls. In the case of the above example, assuming that three bytes are sufficient to form a particular value, the PC must still add two null characters to make a total of five bytes.

In connect mode, the DC2 character causes the counter to execute the command string received from the PC (Fig. 10b). Depending on the type of measurement, for instance, frequency with a gate time of 10 s, the execution of the string may take some time. The counter acknowledges receipt of the DC2 code by returning an FS code to the PC. An ACK code follows when the commands have been executed.

Continued next month.

Fig. 10. State diagrams showing how the PC and the counter communicate.
February 1992
The PCD8584 used in this project is no longer manufactured by Philips Semiconductors, and replaced by the PCF8584. This is a fully compatible IC and only improved as regards the 4-wire long-distance mode, which did not work correctly on the PCD8584.

Real-time clock for 80C32 computer
June 1993
Contrary to what is implied by the description of the parallel connection of the SmartWatch IC pins with the EPROM pins, pin 1 of the SmartWatch should be connected separately to +5V, for instance, to EPROM pin 28, via a short wire. This is necessary because pin 1 on the SmartWatch is 'reset', while on the EPROM it is address line A14, which may be made high by 'high' addressing or glitches, causing the clock to be reset.

VHF-low converter
June 1993
The parts list should be corrected to read:
1 2µH2   L3
1 0µH1   L5
The circuit diagram is correct.
The sub-1µH chokes used in this project are available from, among others, Cricklewood Electronics.

1.2 GHz multifunction frequency meter
December 1992
The recommended LCD module Type LTN211-F10 is no longer manufactured by Philips Components, and may be replaced by the compatible types LM016L from Hitachi, or the LM16A21 from Sharp.