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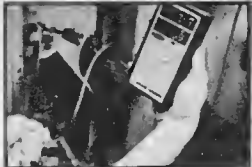
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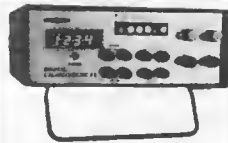
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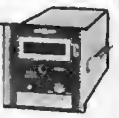
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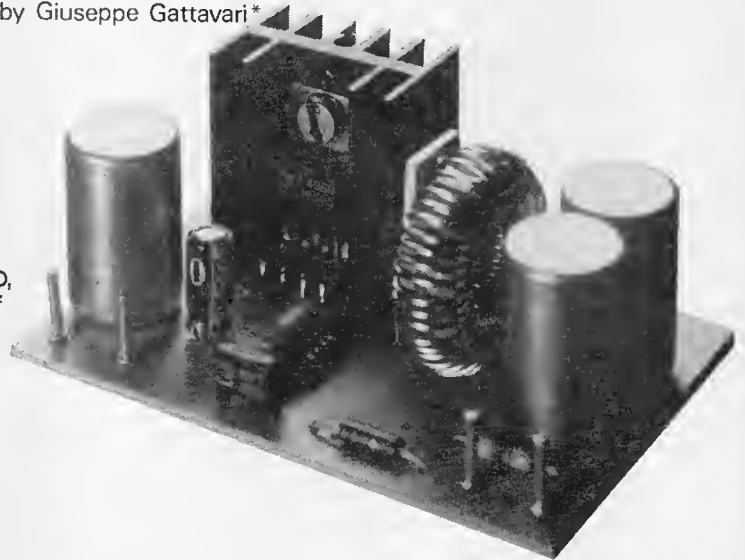
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HIGH CURRENT SWITCHING REGULATOR IC SIMPLIFIES SUPPLY DESIGN

by Giuseppe Gattavari*

Containing a complete 2.5A switching regulator on a single chip, a power IC simplifies the design of many power supply schemes and reduces cost.



While designers are attracted by the high efficiency of switching regulators, they are often deterred by the complexity of circuits based on controllers such as the SG3524 and discrete power transistors. By integrating on a single chip a complete switching regulator capable of delivering 2.5A at 5V

rent Switching Regulator IC offers all the advantages of a switching regulator yet is little more complex to use than a linear regulator. A 2.5A/5V regulator can be built with one L4960 and just eight components (Fig. 1) and higher output voltages are obtained by adding two resistors. Moreover, the device includes current

limiting and thermal protection circuits, eliminating the need to add extra circuitry. A further advantage is that the L4960's source-sink output stage can switch in about 70ns, allowing efficient operation (up to 90%) at switching frequencies of 100kHz. The IC is, in fact, tested dynamically at 100kHz in production. Thanks to

the high frequency operation the output LC filter components can be very small. The IC itself is assembled in the compact Heptawatt 7-lead package—both horizontal and vertical mounting versions are available—and requires only a small heatsink. Considering also the few external components and small LC filter the complete application circuit is extremely compact (see photo). It can even be squeezed into the corner of a system card. A versatile device, the L4960 may be used in a number of different ways. The most obvious is a basic DC-DC converter configuration where a 50/60Hz transformer, rectifier bridge and filter capacitor feed the input of the device with an unstabilized DC voltage.

Multiple outputs

Any number of devices may be combined to produce a multiple output supply, permitting a building block approach to supply design. In multi-chip supplies it is desirable to synchronize the switching frequencies and this can be done by connecting the oscillator pins in common to one RC network (Fig. 2).

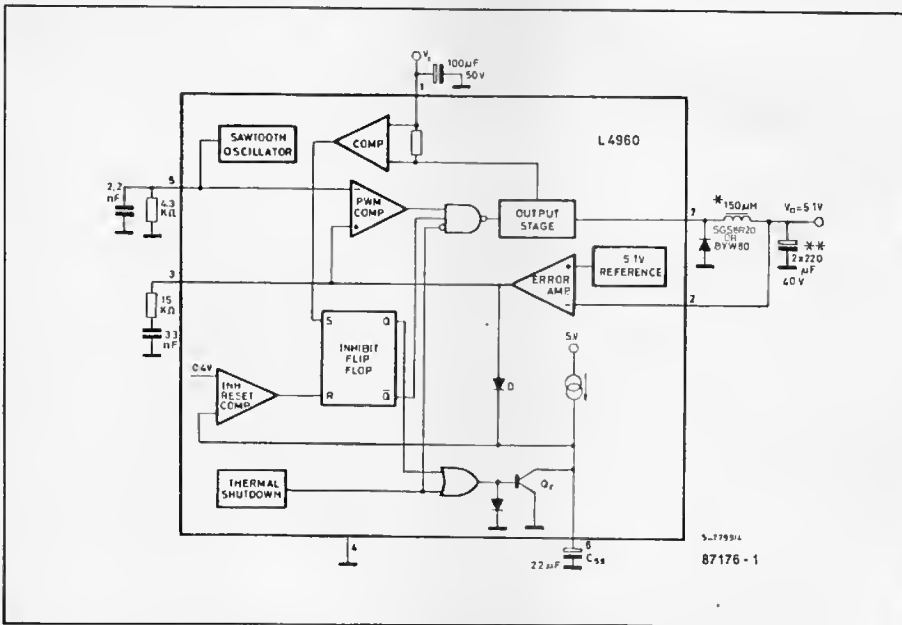


Fig. 1. Containing a complete 2.5A switching regulator with current limiter and protection functions, the L4960 High Current Switching Regulator needs few external components.

current so the diodes must be dimensioned accordingly.

All of these problems are eliminated by using the L4960 as a post regulator as shown in Fig. 7. Note that for all of the auxiliary outputs only one secondary is needed, simplifying the transformer. Cross regulation is no longer a problem and the power dissipated in the stage depends mainly on load current and is almost unaffected by dropout. Moreover, the diode on the secondary can be smaller since the input current of a stepdown switching regulator is always less than the output current.

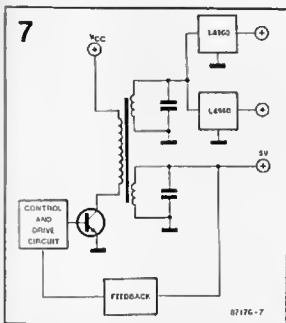


Fig. 7. Using DC-DC converter circuits in place of the linear regulators simplifies the transformer and reduces dissipation.

Finally, short circuit protection is provided for all of the auxiliary outputs by the chip's internal current limiter and thermal protection circuit.

How it works

The SGS L4960 is a monolithic stepdown switching regulator providing output voltages from 5.1V to 40V and delivering up to 2.5A output current.

At the heart of the device is a regulation loop consisting of a sawtooth oscillator, error amplifier, comparator and source-sink output stage. An error signal is produced by comparing the output voltage with a precise 5.1V on-chip reference which is zener-zap trimmed to $\pm 2\%$. This error signal is then compared with the sawtooth signal to generate the fixed frequency pulse-width-modulated pulses which drive the output stage. Gain and frequency stability of the loop are adjusted by an RC network connected to pin 3.

When the loop is closed directly by connecting the supply output to the feedback input (pin 2) an output voltage of 5.1V is produced. Higher output voltages are obtained by inserting a voltage divider in this feedback path.

Output overcurrents at switch-on are prevented by the soft-start function. The error amplifier output is initially clamped by the external capacitor C_{SS} and allowed to rise, linearly, as this capacitor is charged by a constant-current source.

Output overload protection is provided in the form of a current limiter. The load current is sensed by an internal metal resistor connected to a comparator. When the load current exceeds a preset threshold, this comparator sets a flip flop which disables the output stage and discharges the soft-start capacitor.

A second comparator resets the flip flop when the voltage across the soft start capacitor has fallen to 0.4V. The output stage is thus re-enabled and the output voltage rises under control of the soft-start network. If the overload condition is still present, the limiter will trigger again when the threshold current is reached. The average short-circuit current is limited to a safe value by the dead time introduced in the soft-start network.

The thermal overload circuit disables circuit operation when the junction temperature reaches about 150°C and has hysteresis to prevent instability.

* Giuseppe Gattavari is with SGS Microelettronica SpA

ELECTRONMICROSCOPY COMES TO LIFE

by Dr Jitu Shah, H.H. Wills Physics Laboratory, University of Bristol

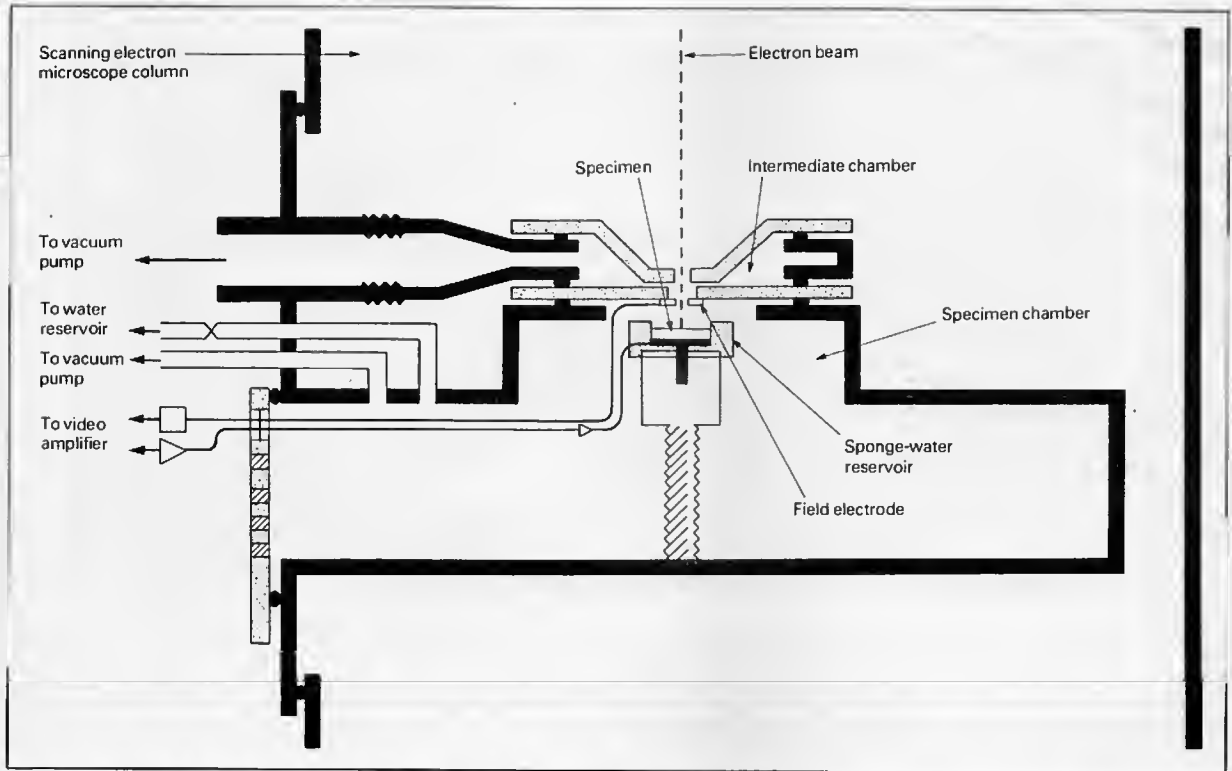
One important limitation of electron microscopes is that the conditions under which the specimen is examined make it impracticable to view living matter. Results from a technique now under development are showing rapid progress in microscopy of biological materials and are paving the way to observing the dynamics of life at high magnifications. Incidentally, it will also provide industry with a powerful tool for inspection and fault finding.

A desire to see structure, forms and morphology at microscopical scales is inherent to the curiosity of mankind. It was the driving force that led Anthony van Leeuwenhoek, a Dutch clockmaker, to devise a compound light microscope. Nowadays, of course, much larger magnifications can be obtained by electron micro-

scopes. Yet optical microscopy still has a powerful advantage over electron microscopy: it can be performed on living matter without destroying it. It is interesting to note that in 1926, when the US physicist Leo Szilard suggested to the British engineer Dennis Gabor that an electron microscope might be made by assembling electron

lenses, Gabor (later the inventor of the hologram and winner of a Nobel prize in physics) rejected the idea and pointed out that living specimens cannot be placed in a vacuum, which is essential for electron beam optics, and that energy in the focused electron beam would burn and destroy anything placed under it.

In the event, the first such microscope was built by the German scientists Max Knoll and Ernst Ruska in 1931. Advances since then in many branches of materials science, biology and medicine can be attributed to the use of electron microscopy. This has been duly acknowledged by the fact that Ernst Ruska shared a 1986



Schematic diagram of the apparatus used for MEATSEM.

Nobel prize in physics, so it is appropriate now to review how far electron microscopy for biological and other difficult materials has progressed.

With a modern, commercially available transmission electron microscope (the type developed by Ruska) we can visualise features and structures of only a few tens of ångströms. However, such an instrument requires the specimen to be thin, because an image is obtained by passing electrons through it. This means that surface features of a thick, three-dimensional object cannot be examined very well. Additionally, because the specimens have to be very thin, it is extremely laborious to obtain three-dimensional information. These disadvantages are overcome by the scanning electron microscope, a type of instrument first built by the German physicist Manfred von Ardenne in 1938.

Depth of focus

The first commercial scanning electron microscope was made available in 1965 by Cambridge Instruments, a British firm near Cambridge. In this kind of microscope an extremely

small, focused spot of electrons is made to fall on the surface of a specimen and scan across it in a 'raster', just as in an ordinary television tube. The electrons interact with the specimen and release secondary electrons from near the surface. (Some of the primary, incident electrons are absorbed within the specimen while some are bounced back out of the surface; more about this back-scattering later.) Emitted secondary electrons yield information about the surface topography.

In a conventional scanning electron microscope these secondary electrons are collected, point by point, and used to build a picture. Deeper surface features of thick specimens can be imaged in a scanning electron microscope because the depth of focus is much greater than that in an optical microscope, so the technique gives a much more vivid impression of three-dimensionality. It displays morphological and topological features at much higher magnifications than in an optical microscope. Because the instrument is relatively easy to use and can be combined with other analytical techniques, it has become enormously popular. Magnifications available

with modern instruments are only slightly less than those possible in a transmission electron microscope.

When it comes to viewing living matter, scanning electron microscopy has the same serious drawbacks as those pointed out by Dennis Gabor. Therefore we must be able to keep specimens in a microscope in a fully hydrated state, without loss of water. That is to say, they must be kept as near as possible to a living state.

In any electron microscope, a well-focused beam of high-energy electrons, necessary for imaging, has to be produced and kept in a high vacuum. It cannot travel long distances in a high-pressure gaseous environment without being scattered by gas atoms or molecules and losing its energy, and a badly scattered beam cannot render high resolution. This is an obstacle to electron microscopy of biological material in its natural, hydrated state.

For these reasons specimens are, conventionally, deliberately dried out and made stable for viewing by using procedures such as chemical fixation, dehydration and fluid replacement. Additionally, for scanning electron microscopy, dehydrated specimens, which are

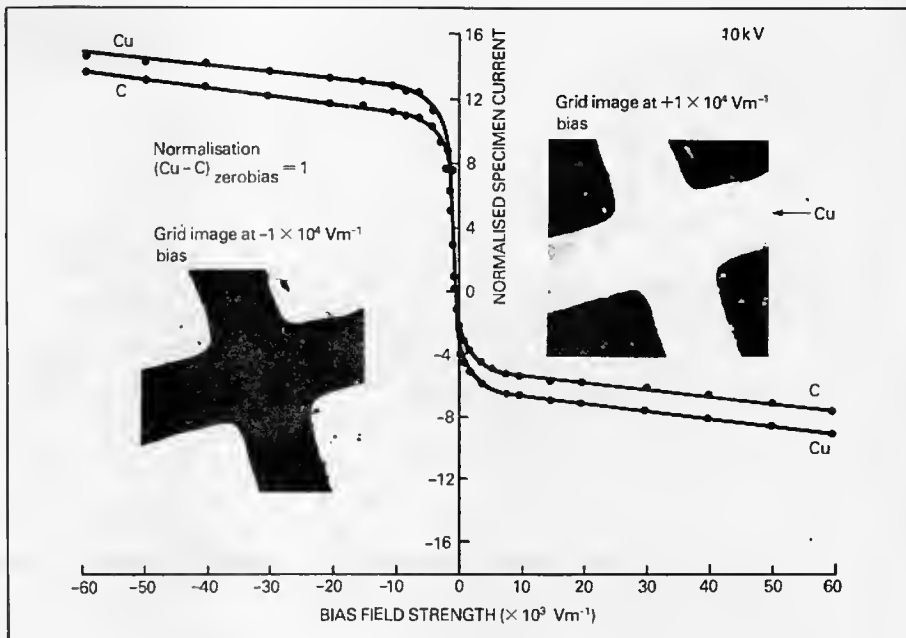
generally poorly conducting, are coated with a thin conducting layer of gold or of an alloy of gold and palladium to avoid a build-up of charge, for detailed features on charged surfaces cannot be imaged well. But these techniques cause a drastic change in interfacial tension forces, which in turn causes delicate biological structures to become distorted and even to collapse. In spite of the development of special techniques for preparing specimens it has not been possible to eliminate damage completely.

Loss of water in a vacuum can be avoided by freezing the specimen and keeping it at a low temperature, at which saturated water pressure is very small. In so-called cryo scanning electron microscopy, specimens are frozen, coated with metal and transferred into the scanning microscope, where they are viewed at a low temperature. However, frozen specimens are not free from damage which takes place through anomalous expansion of water as it changes into ice crystals. Obviously the cryo technique, even if it were made free from specimen damage, could not be used to view living matter.

Two regions

Another approach for solving the problem of loss of water is called moist environment scanning electron microscopy (MEATSEM). This relies on compartmentalisation of a microscope into two regions: the first is a high-vacuum region for electron beam production and electron lens optics; the second is a high-pressure region at room temperature, to surround the specimen and prevent it from losing fluid and gaseous constituents. (If a specimen is kept at saturated vapour pressure of water or 100 per cent humidity it will remain wet, just like clothes hanging on a washing line on a humid day.) Complete compartmentalisation can be achieved by using a window that is transparent to electrons but at the same time tough enough to maintain a high difference of pressure between the two compartments. This approach has the drawback that window materials scatter the electron beam badly. To keep scattering down to feasible levels the window has to be extremely thin, which means it is extremely fragile. Reliable windows with an acceptable loss of resolution are difficult to make.

We have adopted an alternative method of open-window MEATSEM here at Bristol. The focused electron beam passes from the microscope to the specimen chamber via small apertures, as shown in the first diagram. Leakage of gases from the high-pressure region to the microscope column is kept to a minimum by introducing a buffer space, or intermediate chamber, between the specimen region and the electron beam column. The intermediate chamber is bounded by two walls containing concentric limiting aper-



Variation of 'normalised' specimen current with field bias strength for 10-kV accelerated primary electrons. Normalisation is on the basis that the difference between the specimen current from copper and carbon at zero field strength is unity.

tures in the planes normal to the electron beam, and it is pumped continuously so that the pressure gradients can be maintained while the microscope is in use. Water lost through the window is replaced by a continuous injection of water vapour in the vicinity of the specimen. To keep damage due to water loss from the specimen to a minimum, a sponge is introduced to the specimen chamber. The surface area of the sponge is much larger than that of the specimen, so the proportion of the water lost from the specimen to the total loss of water is very small.

Scattering of the electron beam in this arrangement depends largely upon the pressure in the specimen compartment and how far the electron beam travels through the high pressure to reach the

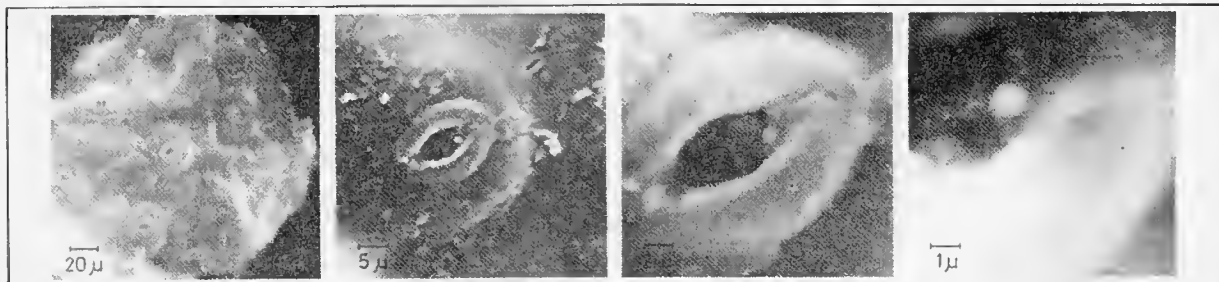
specimen. Pressure in the specimen chamber is kept at the saturated water vapour pressure, at near to room temperature. With careful design of the apparatus the scattering of the primary beam can be kept down to give reasonable resolution. We have used this open-window MEATSEM with some success: it is now possible to insert a fully hydrated specimen into a scanning electron microscope and keep it hydrated for a long time. Nevertheless, preventing desiccation of a specimen in this way poses another problem.

Additional electrons

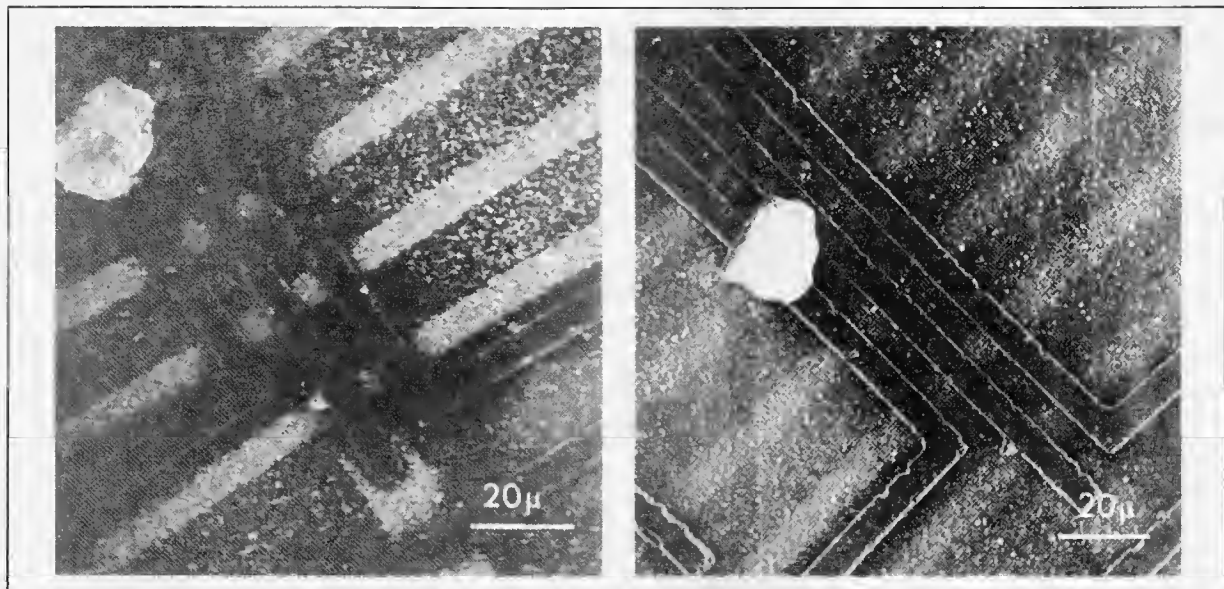
Forming an image under these conditions presents formidable difficulties. The conventional technique of constructing an image by secondary emitted electrons does not work

because secondary electrons, primary electrons and back-scattered electrons ionise water or gas molecules close to the specimen and produce additional electrons. These electrons, which do not carry any information about the specimen surface, have a similar energy range to that of the secondary electrons emitted from the specimen, so they cannot be separated easily from the secondary electrons released from the specimen surface. Without such separation, there is a severe deterioration of the secondary emitted image.

Back-scattered electrons (deflected primary electrons from beneath the specimen) also carry image information. Because they are scattered from a larger volume of specimen, the resolution achievable by their use is not as good as that obtainable by



Stomata, or breathing pores, viewed by open-window MEATSEM at successively higher magnifications, which eventually reveal a tubular structure with a slit within a stoma.



Left: view of part of a circuit on a semiconductor wafer showing charging effects; 'spread' in the darker area is due to charging.
Right: A view of the same circuit after charge neutralisation.

using secondary electrons. Further deterioration in the resolution is also likely because the low-energy of back-scattered electrons means that they cannot be separated easily from spurious electrons.

Interaction of the primary electron beam with the specimen creates a charge which, in turn, generates a minute current in the specimen. The point-to-point variation of this current with scanning of the beam can be made use of for image generation. With a wet specimen the current can be collected without any metal coating. A specimen-current image is also susceptible to deterioration through ambient ionisation. Our research has shown that it is possible to sharpen the image in a way that I shall now describe.

We have incorporated an additional annular electrode in the specimen chamber so that a substantial electric field can be produced at the surface of the specimen. The field has a considerable effect on the specimen current, shown in the second illustration. The images are of copper grid bars on a carbon surface, and the curves represent variations of specimen current with the strength of electric field from copper and carbon. It is thought that the electric field helps to conduct the excess charged carriers, produced by the interaction of primary, back-scattered and secondary electrons,

through the gases, which enhances the contrast in the specimen-current image. Application of the field changes the magnitude of the specimen current rapidly at first and then more slowly as one or other plateau in the curve is reached. The image quality and contrast is re-established with the magnitude of the field applied. It is also possible to invert the image contrast by changing the direction of the imposed field. The curves shown for copper and carbon indicate that the contributions to the current by both the back-scattered electrons and the secondary emitted electrons are recovered to a great extent in the plateau regions. So, once the specimen current reaches a plateau, the contrast, sharpness and resolution of the specimen-current image of an object under high pressure are substantially recovered. The image is comparable in quality to the conventional, secondary emissive image of a similar object in a high vacuum.

A series of pictures in the third illustration shows recovered images of stomata of a fully hydrated leaf, at roughly 16 °C, at various magnifications. Stomata are breathing pores of a leaf, which automatically close and open to regulate exchange of water vapour between the leaf and the air. They are therefore suitable specimens to study by MEATSEM on a fully hydrated leaf.

Fully hydrated internal tissue cells of animals can be imaged, too. The resolution obtained so far is limited by factors not directly related to the principle of the technique, while difficulties stem from the fact that the current from a typical uncoated 'wet' biological specimen is smaller.

Results indicate that we may well see rapid progress in scanning electron microscopy of hydrated biological materials. In turn this will open up means of realistically assessing deterioration from other causes, such as radiation damage due to incident electrons and heat produced by the interaction of electrons with a living specimen. Optimistically, we may expect other advances such as low-voltage scanning electron microscopy combined with MEATSEM, which may well enable us to view live matter without inevitably killing the specimen. This will not only fulfil a long standing dream of mankind but also provide a tool for observing the dynamics of life at high magnifications.

MEATSEM has led to a solution of another long-standing problem in scanning electron microscopy. I have already mentioned that insulating, semiconducting and poorly conducting materials are difficult to image, unless coated, in a scanning electron beam instrument because of charge build-up on the surface of the specimen. It alters the trajec-

tories of primary and secondary (emitted) electrons and the process of emission of secondary electrons; this grossly distorts the image and is also accompanied by loss of details and resolution. Charging can also bring about electric breakdown of the material, which is particularly serious in examining semiconductor chips containing circuits and active electronic devices.

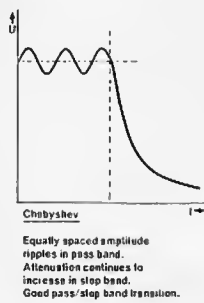
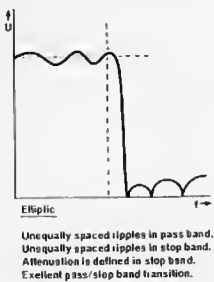
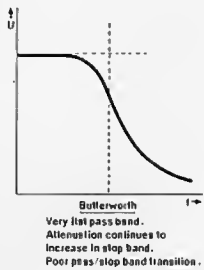
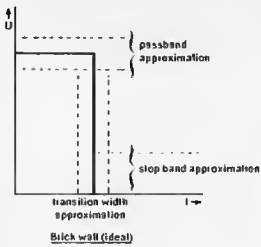
With certain modifications of the open-window MEATSEM, a charge neutralisation mechanism can be employed to reduce or eliminate charge build-up on a surface. The final illustration shows before-and-after images of a circuit on a semiconductor wafer. The improvement was achieved by charge neutralisation. Potentially, the technique is a powerful tool for inspection and fault detection, and promises to have other industrial uses. Cambridge Instruments, who built the first scanning electron microscope, may well be the first company to make the MEATSEM and its associated techniques available commercially.

FILTERS: THEORY & PRACTICE — 3

by A.B. Bradshaw

Television, radar, data transmission, and other techniques developed during the 1940s and 1950s showed up the limitations of the image parameter theory. The higher precision and more exact characteristics required of filters from then on caused the image parameter theory to give way to the modern network theory that uses synthesis techniques and digital computers.

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The underlying principle of the network theory is to synthesize the filter from a knowledge of its **voltage transfer function** (voltage vs frequency characteristic). This is an essentially mathematical approach using the ratio of certain types of polynomials. Known as **approximation theory**, it is a very powerful technique. Using these methods, three of the approximations yield familiar transfer functions. Together with the ideal brick wall response, they are shown in Fig. 36.

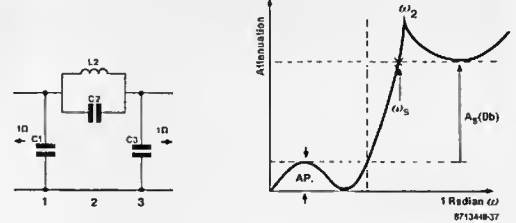
The development of the digital computer greatly aided in the evolution of the polynomial coefficients. This enables filter design tables to be produced from suitable computer programs.

The use of such tables, coupled with a step-by-step design procedure, enables the production of superb filters to the Butterworth, Chebyshev, and elliptic function approximations.

The tables refer to normalized low-pass and high-pass sections. "Normalized" means that the circuit values all refer to a network impedance of 1Ω at an angular frequency, ω , of 1 radian. By using standard multipliers, it is possible to translate the filter impedance to the desired value. Another set of standard multipliers enables the translation of the shape of the response to the required frequency. These two operations are carried out simultaneously and are referred to as **impedance/frequency scaling**.

The basic structures given in the tables can be converted to high pass and a network transformation enables translation into band-pass structures. These new structures are then

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scaled to the desired impedance and frequency to give practical values for the components.

Once a set of filter design tables is available and the designer has familiarized himself with their main properties, it is easy to produce high-quality filter designs without the need to know anything about the coefficients of **real rational functions** or **Hurwitz polynomials**.

Elliptic function tables

The elliptic function tables are chosen as these give the best

results for a given number of components in the design (Ref. 1).

Each table refers to a specific network. Each set of figures is given for a range of stop band attenuations (usually in 5 dB steps for the shorter network), and for a given pass-band ripple in dB. The figures in the columns are the actual component values referring to $\omega = 1$ radian and $Z = 1$.

How the tables relate to the network is shown in Fig. 37 and Table 1. This table shows only two lines of the general table with 1 dB pass-band ripples. Table 2 shows two lines of the

Table 1

W_s	A_s	C_1	C_2	L_2	W_2	C_3
2.048	35	1.852	0.214	0.865	2.324	1.852
2.418	40	1.910	0.145	0.905	2.762	1.190

Table 2

W_s	A_s	C_1	C_2	L_2	W_2	C_3
2.921	35	0.958	0.0837	1.057	3.362	0.958
3.542	40	0.988	0.0570	1.081	4.027	0.988

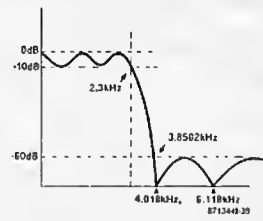
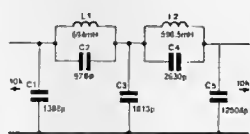
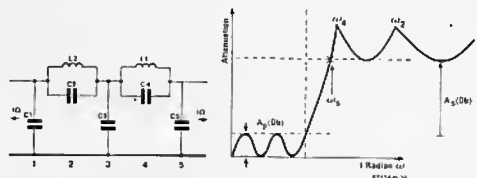


Table 3

Ws	As	C1	C1	L2	W2	C3	C1	L4	W4	C5
1.145	35	1.783	0.174	0.827	1.597	1.978	1.487	0.488	1.174	1.276
1.217	40	1.861	0.372	0.873	1.755	2.142	1.107	0.578	1.250	1.427
1.245	45	1.923	0.293	0.947	1.898	2.296	0.848	0.684	1.313	1.553
1.407	50	1.933	0.223	0.963	2.158	2.392	0.626	0.750	1.459	1.635
1.528	55	1.976	0.178	0.986	2.387	2.519	0.487	0.811	1.591	1.732
1.674	60	2.007	0.141	1.003	2.660	2.620	0.380	0.862	1.747	1.807

general table with 0.1 dB pass-band ripples.

In the worked examples given later in this article, the longer low-pass network of Fig. 38 will be used. For this longer network, figures are given in Table 3 for stop-band attenuations from 35 dB to 60 dB with 1 dB pass-band ripples.

In the tables, ω_s is the frequency at which the specific attenuation begins; A_s is the specified stop-band attenuation;

C_1-C_5 are the values of the capacitors in farad;

L_2 and L_4 are the values of inductors in henry;

ω_2 and ω_4 are the angular frequencies at infinite attenuation.

The values of the capacitors and inductors are very large because the network refers to 1 Ω at 1 radian: they will become more practical during the impedance/frequency scaling. Note that the pass-band edge is the ripple limit and not the -3 dB point.

Low-pass filter for SSB reception

With reference to the last three lines in Table 3, if the pass-band edge is defined at 2.2 kHz (-1 dB), the following frequencies would obtain at the attenuations stated:

- 50 dB: $1.407 \times 2.2 = 3.0954$ kHz
- 55 dB: $1.528 \times 2.2 = 3.3616$ kHz
- 60 dB: $1.674 \times 2.2 = 2.6828$ kHz

If the pass-band edge were

defined at 2.7 kHz (-1 dB), the frequencies at the given attenuations would become:

- 50 dB: $1.407 \times 2.7 = 3.7989$ kHz
- 55 dB: $1.528 \times 2.7 = 4.1256$ kHz
- 60 dB: $1.674 \times 2.7 = 4.5198$ kHz

As regards the filter impedance, generally speaking, the higher this is, the larger the inductances become, and the smaller the capacitances. At relatively high impedances, the capacitors can be matched more easily by using 1% silver mica types. The large inductances may be avoided by electronic simulation, of which more later.

Using the last line of Table 3 and deciding on a filter impedance of 1 k Ω , and a pass-band edge of 2.3 kHz, the two inductances can be calculated with inductance multiplier L' . This multiplier is determined by

$$L' = \frac{(\text{design impedance})(\text{inductance from table})}{2\pi \times (\text{band-edge frequency})}$$

The required inductance values, L'_2 and L'_4 are then calculated as follows:

$$L'_2 = \frac{10^3 \times 1.003}{2\pi \times 2.3 \times 10^3} = 0.069405 \text{ H} = 69.405 \text{ mH}$$

$$L'_4 = \frac{10^3 \times 0.862}{2\pi \times 2.3 \times 10^3} = 0.059648 \text{ H} = 59.648 \text{ mH}$$

The capacitance multiplier, C' , is determined from:

$$C' = \frac{(\text{capacitance from table})}{(\text{design impedance}) \times 2\pi \times (\text{band-edge frequency})}$$

The capacitance values can now be calculated:

$$C'_1 = \frac{2.007}{10^3 \times 2\pi \times 2.3 \times 10^3} = 138.8799 \text{ nF}$$

$$C'_3 = \frac{0.141}{10^3 \times 2\pi \times 2.3 \times 10^3} = 9.75688 \text{ nF}$$

Similarly,

$$C'_2 = 181.2982 \text{ nF}$$

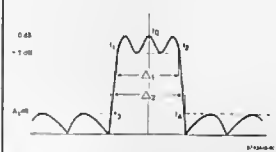
$$C'_4 = 26.295 \text{ nF}$$

$$C'_5 = 125.040 \text{ nF}$$

Since these capacitance values are rather large, it is more practical to decide on an impedance of 10 k Ω , which makes the inductances 10 times larger, and the capacitances 10 times smaller. This will result in a 10 k Ω filter with the same attenuation vs frequency shape, as shown in Fig. 39. This figure shows the complete low-pass filter and its frequency response.

Band-pass filter for CW reception

The desired frequency response curve of the band-pass filter is shown in Fig. 40. Frequencies f_1 and f_2 are the band edges at -1 dB (in this case); f_3 and f_4 are the fre-



quencies about f_0 at which A_s occurs; and f_0 is the centre of the pass band. Note that $f_0^2 = f_1 f_2 = f_3 f_4$ and that $\Delta_2 / \Delta_1 = \omega_s$.

First, a value is chosen for f_0 . Then, from the tables, f_4 for a given A_s is determined, after which f_3 can be calculated.

Subtracting f_3 from f_4 gives Δ_2 . Then, since ω_s is given in the relevant line of the table, Δ_1 can be calculated.

The values of the following are now known: Δ_1 ; Δ_2 ; f_0 ; f_3 ; and f_4 . Now, since $\Delta_1 = f_2 - f_1$,

$$\Delta_1 = f_2 - (f_0^2 / f_2)$$

which is a quadrature in f_2 , so that

$$\Delta_1 f_2 = f_2^2 - f_0^2$$

This is a simple quadratic equation, whose positive root is

$$f_2 = (\Delta_1 + \sqrt{\Delta_1^2 + 4f_0^2}) / 2 \tag{37}$$

$$\Delta_2 = f_4 - f_3 =$$

$$= \omega_s f_0 - f_0 / \omega_s$$

Since $\Delta_2 / \Delta_1 = \omega_s$,

$$\Delta_1 \omega_s = \omega_s f_0 - f_0 / \omega_s$$

from which

$$(f_0 - \Delta_1) \omega_s = f_0 / \omega_s$$

and

$$\omega_s = \sqrt{f_0 / (f_0 - \Delta_1)} \tag{38}$$

This establishes ω_s and the line of the table with which to design the filter.

Taking $f_0 = 850$ Hz (the most popular CW-morse code-tone), and a bandwidth $\Delta_1 = 300$ Hz, gives (from Eq. 38):

$$\omega_s = \sqrt{850 / (850 - 300)} = 1.243$$

The nearest value to this in Table 3 is 1.245, and this would give a stop-band attenuation of 45 dB. Having established the relevant line in Table 3, the tran-

sition frequencies can be calculated. From Eq. 37:

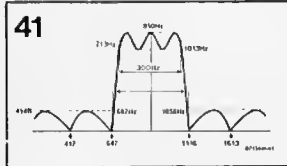
$$f_2 = (300 + \sqrt{300^2 + 4 \times 850^2})/2 = 1013 \text{ Hz}$$

and

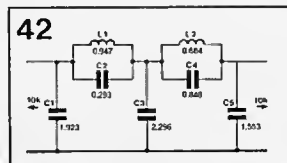
$$f_1 = 1013 - 300 = 713 \text{ Hz}$$

$$f_4 = \omega_s f_0 = 1.245 \times 850 = 1058 \text{ Hz}$$

$$f_3 = f_0 / \omega_s = 850 / 1.245 = 682 \text{ Hz}$$



From these results, the response characteristic can be determined and this is shown in Fig. 41. The actual filter diagram is shown in Fig. 42. If a more rectangular shape or a greater stop-band attenuation at this bandwidth is required, a more complex network needs to be designed.



This may be done by converting the circuit of Fig. 42 to that of Fig. 43. In this, each capacitance and inductance of Fig. 42 is resonated by placing

the opposite reciprocal reactance in series or parallel with it as appropriate.

The two rather untidy middle sections are then transformed to more simple ones by a step-by-step transformation as shown in Fig. 44. For this transformation, let

$$x = L/K$$

$$\alpha = 1 + 1/2x$$

$$\beta = \sqrt{\alpha^2 - 1}$$

$$y = \alpha + \beta$$

$$L_0 = 1/(y + 1)$$

With the aid of these identities and a hand-held calculator, the first mid-section of Fig. 43 is computed:

$$x = L/K = 0.947/(1/0.293) = 0.947 \times 0.293 = 0.277471$$

$$\alpha = 1 + 1/2x = 2.80199$$

$$\beta = \sqrt{\alpha^2 - 1} = 2.61746$$

$$y = \alpha + \beta = 5.41945$$

$$L_0 = 1/(y + 1) = 0.155776$$

The values of the components in Fig. 44 can now be calculated:

$$KL_0 = (1/0.293) \times 0.155776 = 0.531658$$

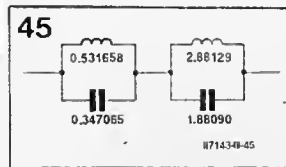
$$yKL_0 = 5.41945 \times (1/0.293) \times 0.155776 = 2.88129$$

$$1/yKL_0 = 0.347065$$

$$1/KL_0 = 1.88090$$

This gives a first mid-section as shown in Fig. 45.

Treating the second mid-section of Fig. 43 in the same way, gives



$$x = 0.580032$$

$$\alpha = 1.86202$$

$$\beta = 1.570708$$

$$y = 3.432728$$

$$L_0 = 0.2255947$$

from which

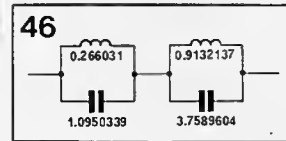
$$KL_0 = 0.266031$$

$$yKL_0 = 0.9132137$$

$$1/yKL_0 = 1.0950339$$

$$1/KL_0 = 3.7589604$$

This gives the second mid-section as shown in Fig. 46.



The complete filter is shown in Fig. 47. For a 10 kΩ network with a centre frequency, $f_0 = 850$ Hz, the inductances and capacitances can be calculated from:

$$L = Z/2\pi f_0 \times (L \text{ in network})$$

and

$$C = (C \text{ in network})/2\pi f_0 Z$$

The following values are obtained:

$$L_1 = 97.369 \text{ mH}$$

$$L_2 = 99.548 \text{ mH}$$

$$L_3 = 539.495 \text{ mH}$$

$$L_4 = 81.550 \text{ mH}$$

$$L_5 = 49.8119 \text{ mH}$$

$$L_6 = 170.991 \text{ mH}$$

$$L_7 = 120.567 \text{ mH}$$

and

$$C_1 = 0.36006 \mu\text{F}$$

$$C_2 = 0.06498 \mu\text{F}$$

$$C_3 = 0.35218 \mu\text{F}$$

$$C_4 = 0.42990 \mu\text{F}$$

$$C_5 = 0.20503 \mu\text{F}$$

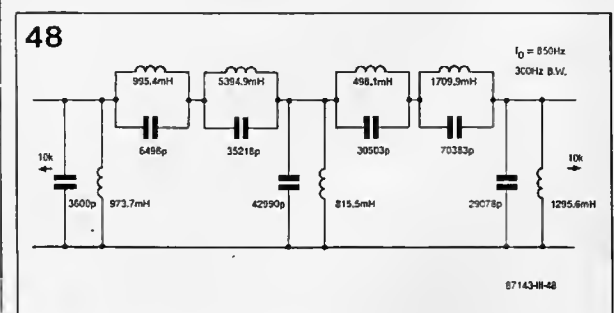
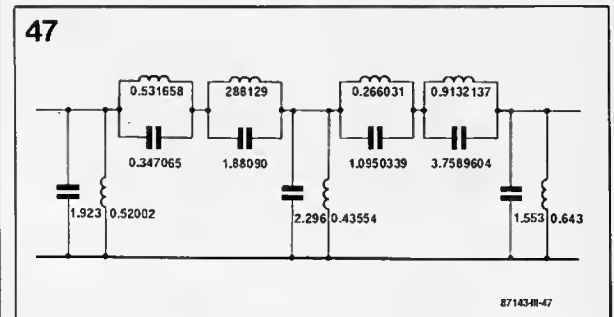
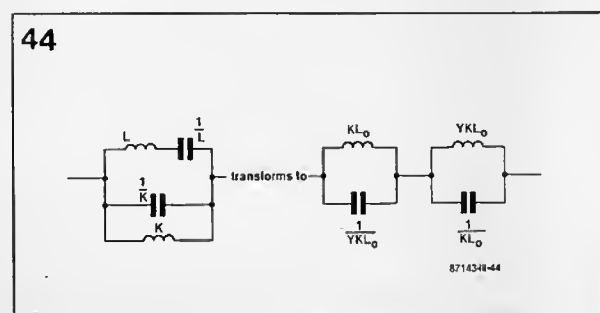
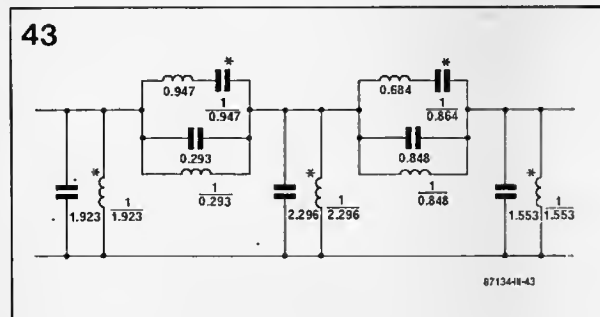
$$C_6 = 0.70383 \mu\text{F}$$

$$C_7 = 0.29078 \mu\text{F}$$

Since these capacitor values are too large for practical purposes, it is propitious to increase the filter impedance to 10 k to make the values of the capacitances 10 times smaller and those of the inductances 10 times larger. This results in the final design shown in Fig. 48.

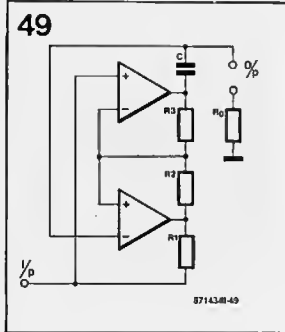
The Positive Impedance Converter (PIC)

As stated earlier, inductances can be electronically simulated and this is done by the use of



positive impedance converters. The theory of these devices is discussed on page 54. When the output port of a PIC is terminated by a resistance, the input port becomes an inductor with a very high Q . In Fig. 49

- $R_1 = 270\Omega$ 1%
- $R_2 = 5k\Omega$ 1%
- $R_3 = 10k\Omega$ 1%
- $C = 10$ nF mica 1%
(2×5000 pF in parallel)
- IC = MC1458 (supply = ± 12 V)



At the input port,

$$L_{in} = KR_0 \quad [39]$$

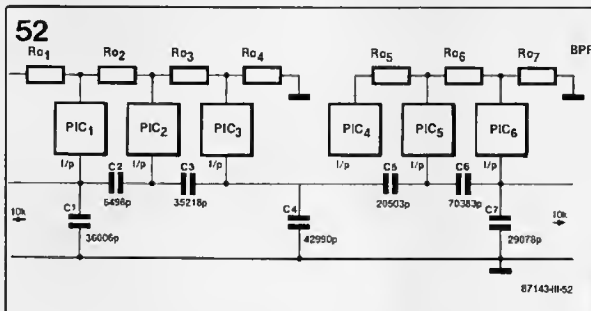
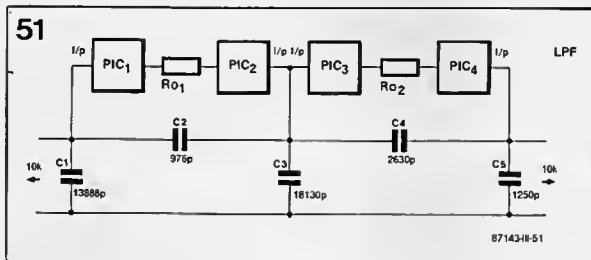
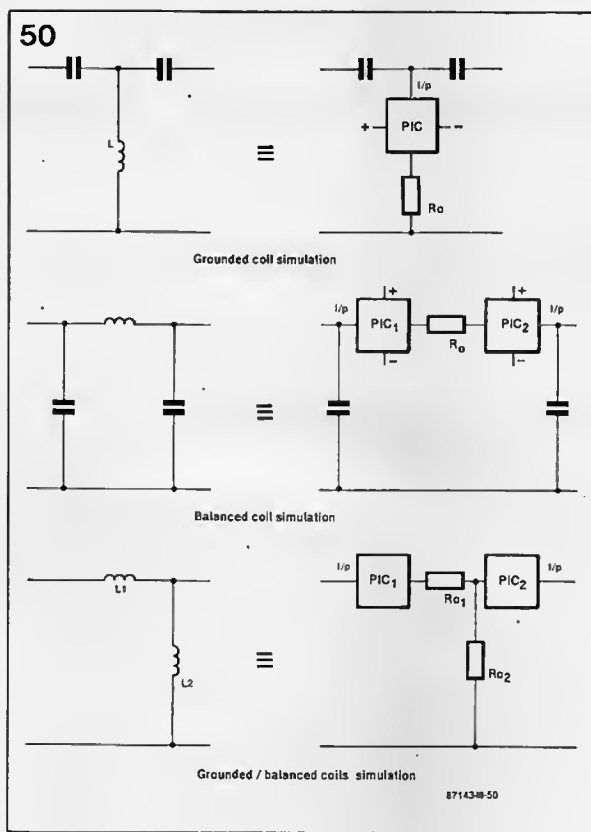
where $K = CR_1R_3/R_2$

With the MC1458, the value of $K = 5.17155 \times 10^{-6}$. The deviation from the ideal K is due to opamp approximations. With this value of K ,

$$L_{in} = 5.17155 \times 10^{-6} R_0$$

Inductances in the filters discussed in this article are simulated by PICs as shown in Fig. 50. In Fig. 50 (c), L_1 is simulated by $PIC_1 + PIC_2R_{01}$, while L_2 is simulated by PIC_2R_{02} . The PIC can be trimmed to have a K value of 5.17155×10^{-6} by adjustment of R_3 in Fig. 49.

When the inductances are replaced by PICs, the low-pass filter of Fig. 39 becomes as shown in Fig. 51, while the band-pass filter of Fig. 48 is



transformed into that of Fig. 52. It is important to match the PICs. The value of each R_0 is calculated with the aid of Eq. 39 ($L_{in} = KR_0$, whence $R_0 = L_{in}/K$). Thus, in Fig. 51:

$$R_{01} = 10^6 L_1 / 5.17155 = \\ = (10^6 / 5.17155) (694 / 10^3) = 134 k$$

$$R_{02} = (10^6 / 5.17155) (596.5 / 10^3) = \\ = 115 k$$

In the same manner, the values of R_{01} to R_{07} in Fig. 52 are found; their values are:

- $R_{01} = 188 k$
- $R_{02} = 192 k$
- $R_{03} = 1.043 M$
- $R_{04} = 157 k$
- $R_{05} = 96 k$
- $R_{06} = 330 k$
- $R_{07} = 233 k$

End

Reference:
On the Design of Filters by Synthesis; IRE Trans: Circuit Theory, Vol. CT-5 1958, pp 284-328, by R. Saal and E. Ulbrich

LOW-NOISE MICROPHONE PREAMPLIFIER

A quality preamplifier that can be configured for low and high impedance, as well as balanced and unbalanced, microphones.

Most dynamic microphones of European make (AKG, Beyer, Sennheiser) have a characteristic impedance of 200 Ω , while those of Far Eastern origin are usually terminated in 500 to 600 Ω . The impedance of electret condenser microphones is type specific, and usually in the range from 600 to 1000 Ω . Most microphones have a sensitivity of 2...3 mV/Pa. The termination impedance, i.e., the input impedance of the preamplifier or the microphone transformer, should be equal to or higher than the impedance of the microphone. The transformer is often precisely matched to the microphone impedance (power matching), while preamplifiers give optimum performance when the termination impedance is slightly higher than the microphone impedance (voltage matching). The use of a transformer with a 1:10 or 1:15 relaxes the requirements for the sensitivity and the signal-to-noise ratio of the preamplifier. However, the circuit may still be susceptible to hum and noise when the transformer is housed in the microphone. Obviously, this calls for high-quality microphone wire. Hum and other interference is minimized when the microphone preamplifier has a signal-to-noise ratio of 70 dB or more. Balanced systems (the microphone, cable, and preamplifier input) are known for their high immunity to noise.

Circuit description

The circuit diagram of the microphone preamplifier is shown in Fig. 1. The amplifier can be configured for use with balanced and unbalanced microphones. Component values for the unbalanced version are shown in brackets.

Components whose value is not shown for a particular version are not required (also consult the parts list).

Balanced version: the input signal is applied to the + inputs of operational amplifiers IC₁ and IC₂. Resistors R₁...R₄ incl. determine the input impedance. When jumper A is fitted, the output signals of the opamps are subtracted in IC₃ for effective suppression of hum and noise. Due to the higher number of active components, the S/N ratio of the balanced version is a few decibels worse than that of the unbalanced circuit.

Unbalanced version: wire link B is fitted, so that IC₃ functions as a non-inverting amplifier. IC₂ and IC₃ are not required. The input impedance is determined by R₃, R₄ and R₁₁. The microphone signal reaches IC₃ via C₁ and the wire link fitted in position R₆. IC₃ and IC₄ raise the input signal to enable driving the line input of an AF power amplifier.

The input impedance of the preamplifier can be switched

between high and low by connecting a switch between S_y and ground, and S_x and ground—the former is not required for the unbalanced versions. If necessary, the input impedance of the preamplifier can be altered to match a particular microphone. The input resistance of the balanced version should consist of 2 equal resistors with a value of half the required termination impedance. In the unbalanced version, R₃ and R₄ may be omitted, so that R₁₁ alone determines the input impedance.

Construction and use

The printed circuit board shown in Fig. 2 can be used for constructing both versions of the preamplifier. Separate parts lists are given to avoid construction errors. Fit jumper A or B as appropriate, and do not forget the 2 wire links. Capacitors of 3 different pitch sizes can be fitted in positions C₁ and C₄. Figure 3 shows prototypes of the microphone preamplifier fitted in diecast enclosures to ensure

robustness and freedom of induced noise.

The symmetry of the preamplifier, and hence the attainable signal-to-noise ratio, depends mainly on the quality and the stability of the components used around IC₁, IC₂ and IC₃. Equally important, however, is the use of a high-quality microphone and a suitable cable. S_v

Low noise microphone preamplifier

Technical specification

Supply voltage:	± 9...15 V
Current consumption:	± 7.5 mA (unbalanced version) ± 15 mA (balanced version)
Signal-to-noise ratio	
NE5534:	-87 dB (unbalanced version) -81 dB (balanced version)
OP-27:	-89 dB (balanced version) -83 dB (balanced version)
Distortion:	≤ 0.003%
Voltage gain:	40 dB
Input resistance:	24 k Ω or 680 Ω (unbalanced version). 45 k Ω or 660 Ω (balanced version).

Measurement conditions:

R_L = 4K7; U_o = 1V_{rms} ± 0 dB; f = 1 kHz.
S/N ratio measured with input short-circuited.

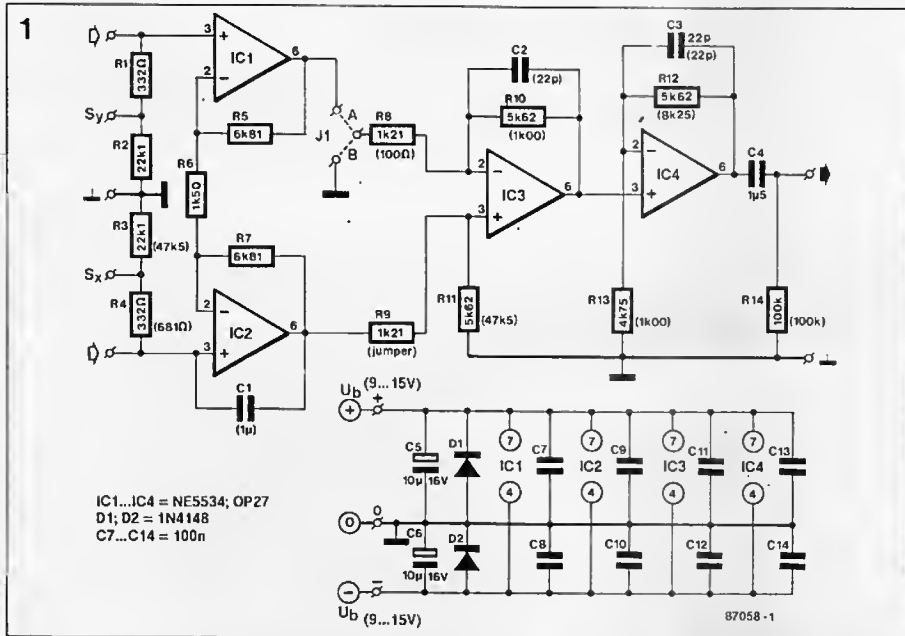


Fig. 1 Circuit diagram of the low noise microphone preamplifier.

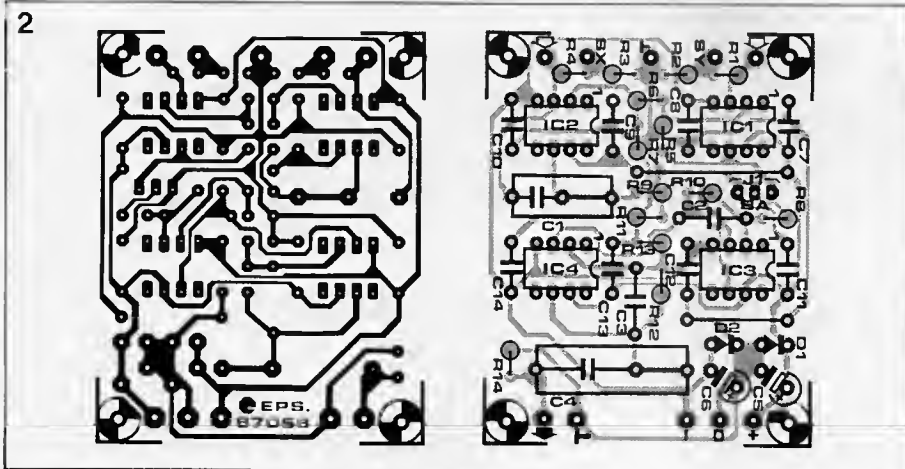


Fig. 2 The printed circuit board for the microphone preamplifier.



Fig. 3 Prototypes of the preamplifier housed in Eddystone enclosures.

Parts list

Balanced version

Resistors ($\pm 1\%$):

- R₁; R₄ = 332RF
- R₂; R₃ = 22K 1F
- R₅; R₇ = 6K81F
- R₆ = 1K5F
- R₈; R₉ = 1K21F
- R₁₀; R₁₁; R₁₂ = 5K62F
- R₁₃ = 4K75F
- R₁₄ = 100KF

Capacitors:

- C₁; C₂ = not required
- C₃ = 22p; styroflex/polystyrene
- C₄ = 1 μ 5; MKT*
- C₅; C₆ = 10 μ ; 16 V; radial
- C₇...C₁₄ incl. = 100n

Semiconductors:

- D₁; D₂ = 1N4148
- IC₁...IC₄ incl. = OP-27 or NE5534

Miscellaneous:

- PCB Type 87058 (available through the Readers Services).

Parts list

Unbalanced version

Resistors ($\pm 1\%$):

- R₁; R₂; R₃; R₅; R₇ = not required
- R₄ = 681RF
- R₆ = 100RF
- R₈ = wire link
- R₁₀; R₁₃ = 1K 0F
- R₁₂ = 8K 25F
- R₁₄ = 100KF

Capacitors:

- C₁ = 1 μ 0; MKT*
- C₂; C₃ = 22p; styroflex/polystyrene
- C₄ = 1 μ 5; MKT*
- C₅; C₆ = 10 μ ; 16 V; radial
- C₇...C₁₀ incl. = not required
- C₁₁...C₁₄ incl. = 100n

Semiconductors:

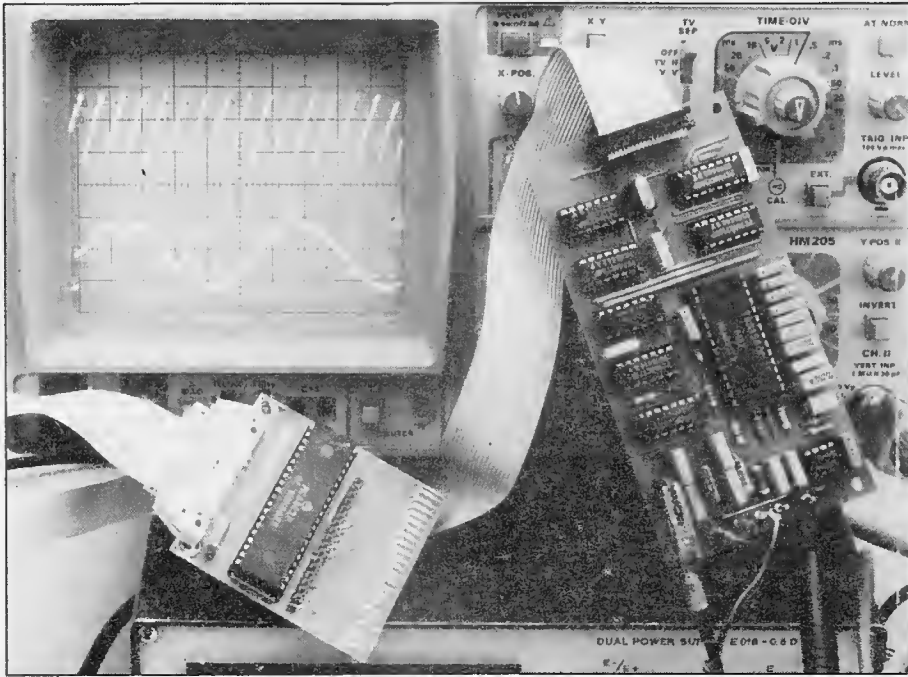
- D₁; D₂ = 1N4148
- IC₁; IC₂ = not required
- IC₃; IC₄ = OP-27 or NE5534

Miscellaneous:

- PCB Type 87058 (available through the Readers Services).

* MKT capacitors are Siemens style, blue cased, Type B32529. Available from ElectroValue • 28 St Judes Road • Englefield Green • Egham • Surrey TW20 0HB. Telephone: (0784) 336603. Manchester branch: (061) 432) 4945.

14-BIT DIGITAL TO ANALOGUE CONVERTER



A high resolution D-A converter that enables computer control of test, measuring, and electrophonic equipment.

Most D-A converters (DACs) that find applications in home-made peripheral equipment are 8-bit types. The converter chip used in this project, however, is a 14-bit type originally designed for use in compact disc players. The programmable number of steps of this chip is 2^{14} , or 16,384. The present converter board has a parallel input for ready connection to most types of microcomputer. Applications of the board include a computer-controlled function generator (see the above photograph), a power supply controller, or a high quality driver for synthesizers and effects equipment. Many other applications are within reach provided the appropriate software is available—and that is where the programmer's own ingenuity and creativity come in.

The Type TDA1540 14-bit DAC

The standards, and therefore the manufacturing technology, for ensuring the accuracy of a 14-bit DAC are quite different from those applied to, say, 8-bit converters. In essence, the conventional DAC based on the use

of a R-2R ladder network (Fig. 1) requires no trimming if up to 10 bits are converted; the stability of the resistors is simply increased to meet the required standard. For DACs with more than 10 bits it becomes necessary to calibrate the resistor ladder by means of a process called *laser trimming*, but

this also has its practical limitations. Applied to 14 and 16-bit DACs, the calibration of the chip would be upset when this is fitted in its plastic or ceramic enclosure.

A different method of dividing the current in the network is referred to as *dynamic element matching*—see Fig. 2a. A 14-bit version of this circuit can attain the required accuracy without the need for trimming, stabilization, or other adjustments. The current supplied by the reference source is divided in 4 equal currents by 4 matched transistors. It is evident that the inevitable production tolerance on the transistors causes small deviations, ΔI , from the ideal current distribution:

$$(I + \Delta I_1) + (I + \Delta I_2) + (I + \Delta I_3) + (I + \Delta I_4) = 4I$$

Table 1

TDA1540 14-bit digital to analogue converter

Supply voltages:	$\pm 5\text{ V}; -17\text{ V}$
S/N (full scale sine wave)	85 dB
at analogue output:	
Non-linearity ($T_a = -20$ to $+70\text{ }^\circ\text{C}$):	$\frac{1}{2}$ LSB
Maximum clock frequency:	12 MHz
Full scale temperature coefficient at analogue output:	$\pm 30 \cdot 10^{-8}\text{ K}^{-1}$
Total power dissipation:	350 mW

1

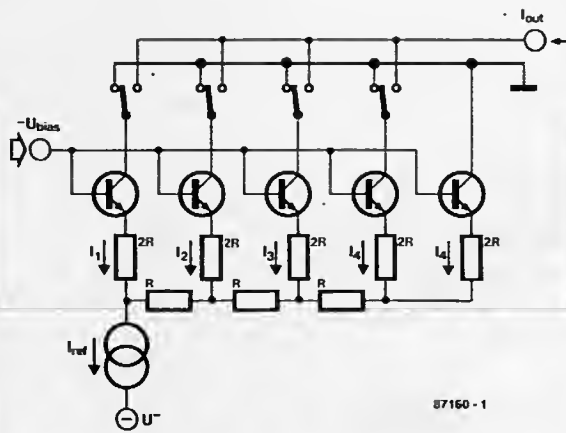


Fig. 1 The accuracy of an R-2R ladder network is insufficient for a 14-bit DAC.

whence

$$\Delta I_1 + \Delta I_2 + \Delta I_3 + \Delta I_4 = 0.$$

The 4 currents ($I + \Delta I_n$) are alternately fed to the 3 outputs. This effectively time-averages the error currents ΔI , so that outputs 1, 2 and 3 carry currents whose average values are related as $I_1:I_2:I_3 = 1:1:2$ (see Fig. 2b). Ripple currents are suppressed with the aid of an R-C filter. The ripple frequency is 50 kHz, i.e., one fourth of the oscillator frequency. The R-C filters also ensure that the average current is used in the D-A conversion when samples with a frequency of more than 50 kHz are applied.

Figure 3 shows the internal organization of the TDA1540 14-bit DAC from Signetics. The current divider for the most significant bits, the reference and the amplifier form a current mirror. The reference current source also supplies the current for the most significant bit, eliminating the need for additional filtering. A different current divider is used for the least significant bits. To avoid the need for an increase in the negative supply voltage for the chip, the current is not divided by connecting matched transistors in parallel, but by designing and manufacturing these in accordance with the required distribution.

For optimum operation the output voltage should be kept virtually nought, since this ensures the effectiveness of the

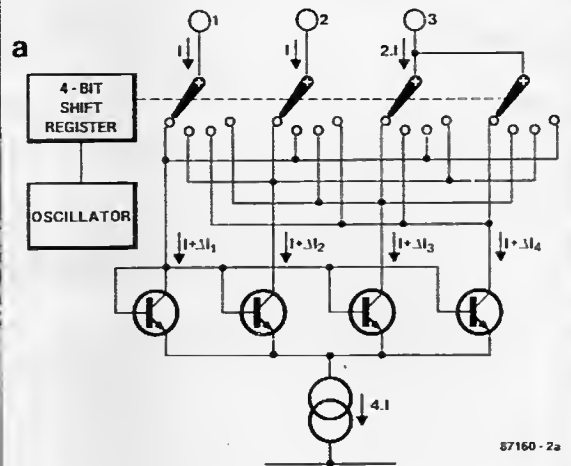
R-C filters while avoiding spurious products caused by the switch circuitry. This is important because samples seem to come off the disc at a rate of 176,400 per second when the CD is a type with quadruple oversampling. The high sampling frequency lowers the overall resolution to some extent because the currents are averaged over a longer period. A compact disc is essentially a serial storage device, and the TDA1540 is, therefore, designed for serial processing of digital data. The most significant bit is first fed into the shift register. Data is stored in a latch to eliminate spurious analogue output currents. The principal technical specifications of the TDA1540 are shown in Table 1.

From parallel to analogue information

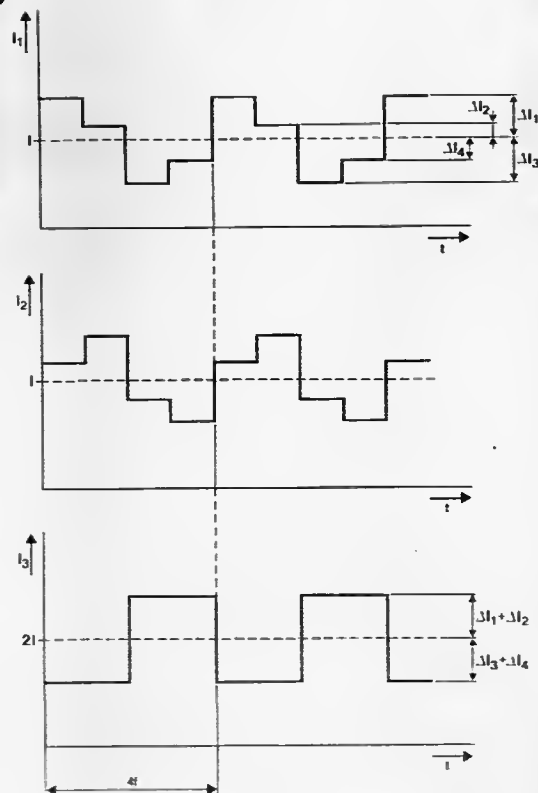
The quickest and simplest way of outputting computer data to a peripheral device is via a parallel port. As already stated, the TDA1540 handles serial data, but the standard serial (RS232) port on a computer is generally too slow for direct interfacing. A parallel-serial converter is therefore required to enable straightforward driving of the DAC board with the aid of 2 parallel output ports.

The block diagram of the 14 bit DAC board is shown in Fig. 4. The 14 bits are applied as 2 bytes, which are converted into serial format by a shift register.

2



b



87160-2b

Fig. 2 An accurate current ratio is obtained by time-averaging deviations ΔI_n in a distributor circuit.

The most significant bit (MSB) of each byte is not used. Lines PA6 and PA7 are used for starting the shift register, and also indicate which byte (MSB/LSB) is to be loaded. The control circuitry on the DAC board activates the RDY line to signal readiness for reception of the next 2 bytes when the shift operation is com-

plete, and the D-A conversion has been started.

A transimpedance amplifier at the output of the DAC converts the output current into a corresponding output voltage.

Circuit description

With reference to the circuit

3

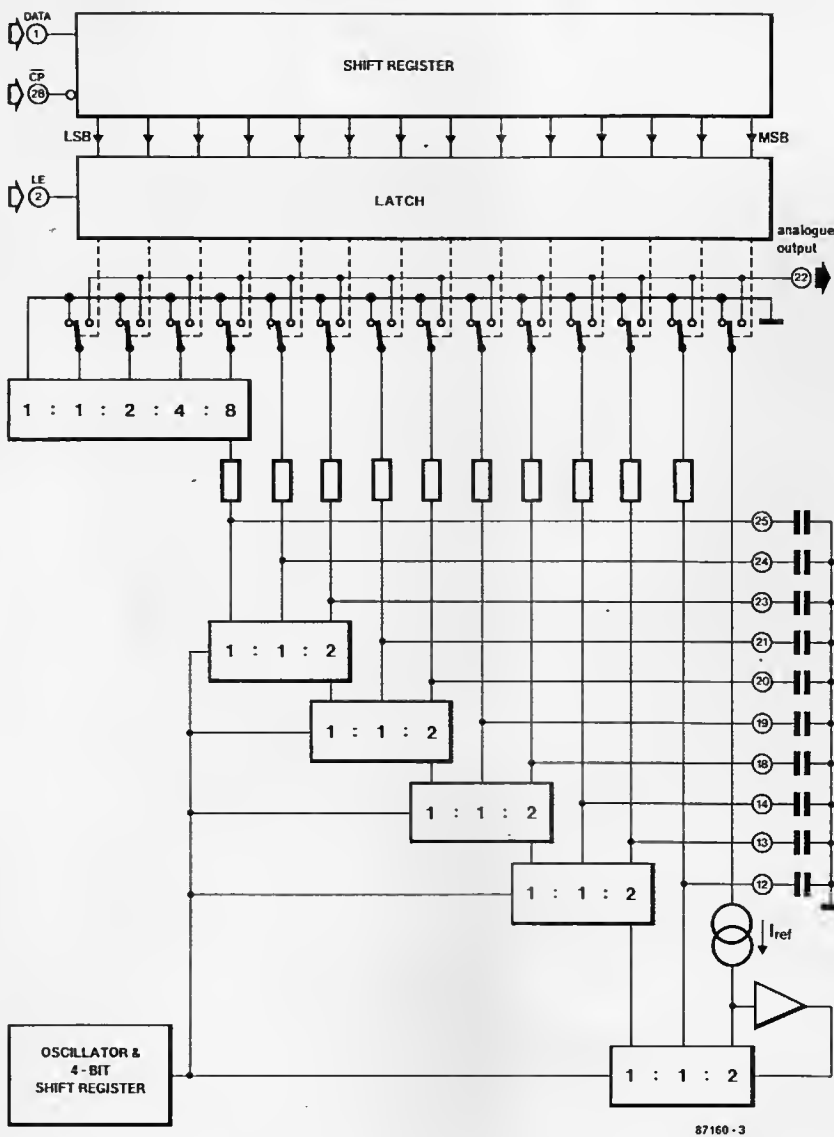


Fig. 3 Block diagram of the TDA1540. The filter capacitors are fitted externally.

4

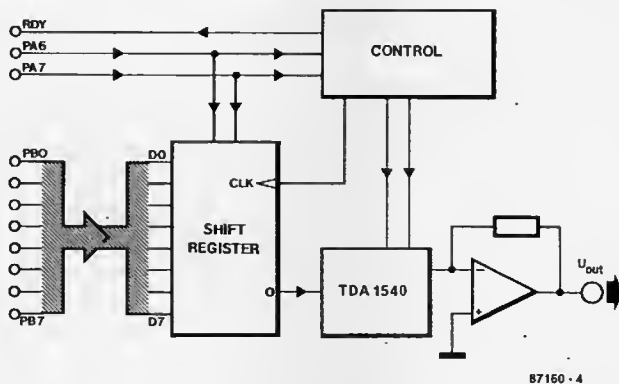


Fig. 4 Block diagram of the complete 14 bit DAC.

diagram of Fig. 5, the DAC board is connected to the computer ports via connector K1. A PIA (peripheral interface adapter) for the Commodore C64 computer was described in Reference (1), while MSX users are referred to Reference (2). The operation of the circuit is explained with reference to the timing diagram of Fig. 6, and the flowchart for generating a sawtooth voltage shown in Fig. 7. The numbers in between brackets refer to the pulses at particular points in the circuit diagram. The timing diagram is valid for a single FOR/NEXT loop. At the onset, data is loaded into the shift registers, IC9 and IC10. The least significant byte on port lines PBx is loaded by making PA6 logic low, PA7 logic high. This must happen simultaneously to prevent N3 prematurely supplying a start pulse. Now the circuit can load the second (MS) byte, whose bit 6 and 7 are kept logic low to ensure the correct operation of the converter. PA7 is subsequently made logic high to enable the circuit to start shifting data into the D-A converter. Bistables FF1 and FF2 enable the clock signal (4) to be fed to the shift registers. Counter IC4 counts the number of output bits from the shift register. When the 15th and 16th bit have been shifted out, and the 14th bit appears on output QH, N10 is enabled by N7 to pass clock pulses (7) to the DAC, IC7. After a total of 16 clock pulses, all bits are available in the DAC, point (6) goes high, and N12 supplies an output pulse (8) that causes the 14 bits to be latched in the DAC, and FF1 and FF2 to be reset. The ready signal (RDY) goes high, and the computer can output the next 2 bytes. The pulse diagram shows that the circuit is not activated until the bits are loaded. This arrangement effectively prevents the production of spurious signals. Operational amplifier IC8 translates the output current from the DAC into a corresponding voltage. A virtual ground potential is created to ensure a low offset at the output of the TDA1540. The maximum output voltage of the opamp is $4R_2 V$, where R_2 is given in k Ω . The minimum load impedance is 600 Ω . Separate +15 V and -17 V sup-

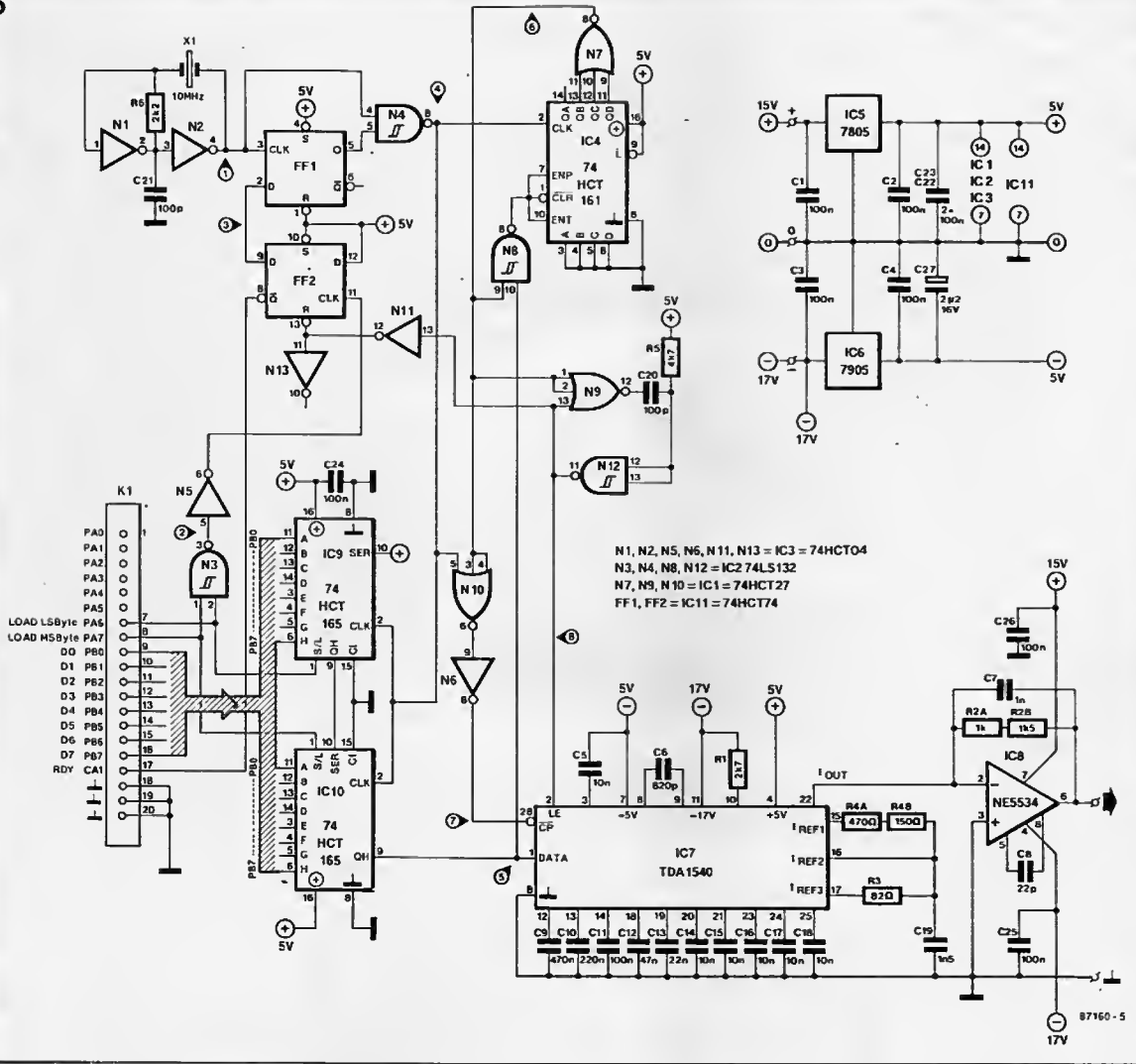


Fig. 5 Circuit diagram of the 14 bit DAC.

plies are required to feed the DAC board. Standard designs with a Type 7815 regulator (+15 V), and an LM337 (-17 V) are adequate for this purpose. Resistors R₃ and R₄ are high stability (1%) types. The values stated for R₂, R₃ and R₄ ensure optimum performance of the converter. R₂ and R₄ may be made from series connected resistors as shown in the circuit diagram and the parts list.

Construction

The ready-made printed circuit board for building the converter is shown in Fig. 8. Be sure to fit all wire links shown on the overlay; the 2 long links in between IC₂ and IC₃ should be made in insulated wire. The voltage regulators are fitted direct onto the board, and may

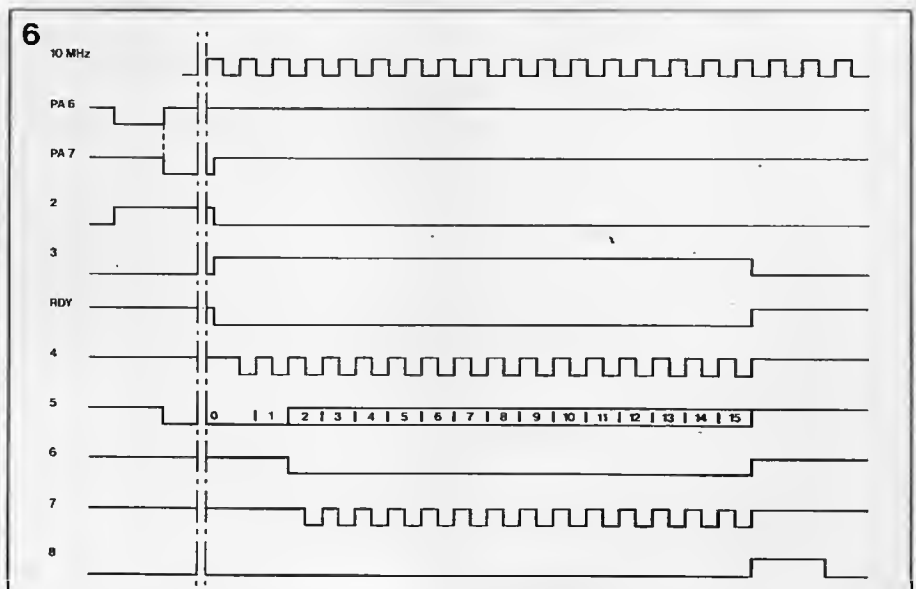


Fig. 6 Timing diagram for a single sampling cycle.

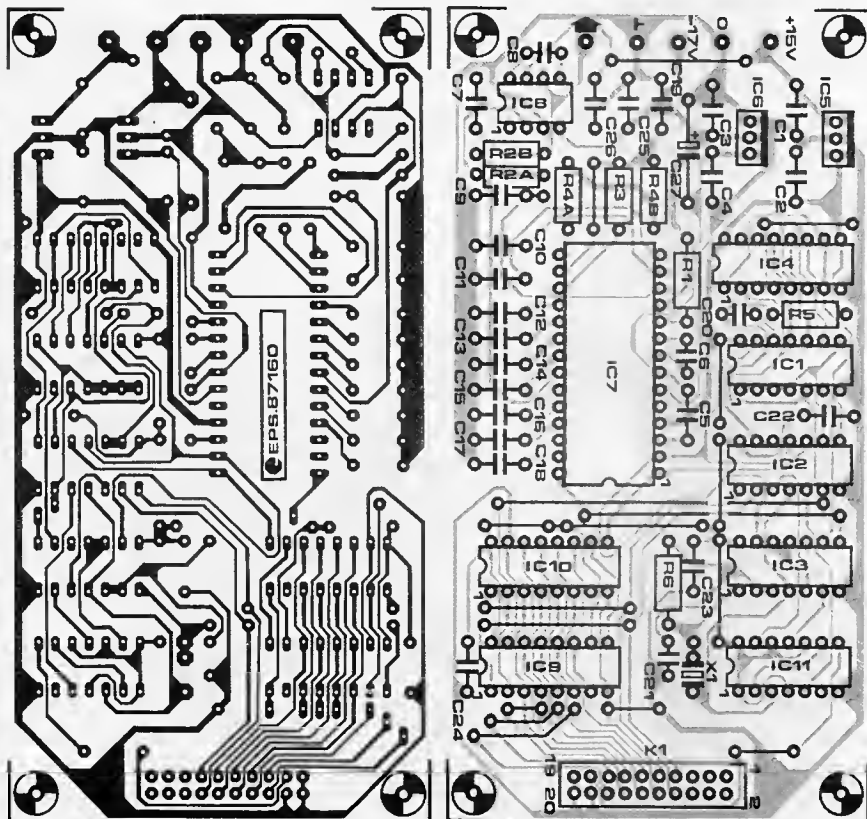


Fig. 8 Printed circuit board for building the 14 bit DAC.

for the slope of the low-pass filter, and makes it possible to use a third-order filter with Bessel or Butterworth characteristics and yet attain more than adequate performance.

Software

Table 2 lists a program for generating a direct voltage, a sawtooth voltage, or a sine-wave with the aid of an MSX computer. A similar program for the Commodore C64 is shown in Table 3. It should be noted that the addresses of the input/output devices may have to be altered as required. In the MSX program, the loading of the D-A converter is effected in lines 470...510, in the C64 program in lines 3040...3080.

D

References:

- (1) Computerscope - 2. *Elektor Electronics*, October 1986, p. 44.
- (2) MSX Extensions - 4. *Elektor Electronics*, January 1987.

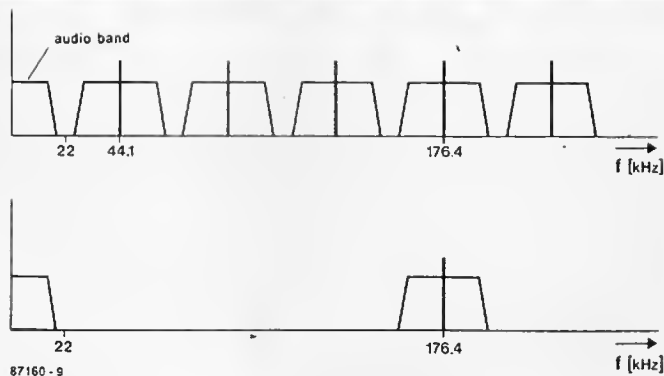
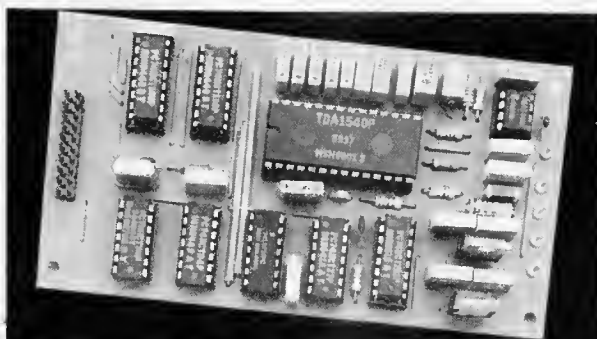


Fig. 9 Output spectrum of a DAC at a sample rate of 44.1 kHz (a) and 176.4 kHz (b).

Parts list

Resistors ($\pm 5\%$):

- R1 = 2K7
- R2 = 2K5 (1K0 + 1K5)
- R3 = 82RF
- R4 = 620RF (470RF + 150RF)
- R5 = 4K7
- R6 = 2K2

Capacitors:

- C1...C4 incl.; C11; C22...C26 incl. = 100n
- C5; C14...C18 incl. = 10n
- C6 = 820p
- C7 = 1n0
- C8 = 22p
- C9 = 470n
- C10 = 220n
- C12 = 47n
- C13 = 22n
- C19 = 1n5
- C20; C21 = 100p
- C27 = 2n2; 16 V; axial

Semiconductors:

- IC1 = 74HCT27
- IC2 = 74LS132
- IC3 = 74HCT04
- IC4 = 74HCT161
- IC5 = 7805
- IC6 = 7905
- IC7 = TDA1540*
- IC8 = NE5534
- IC9; IC10 = 74HCT165
- IC11 = 74HCT74

Miscellaneous:

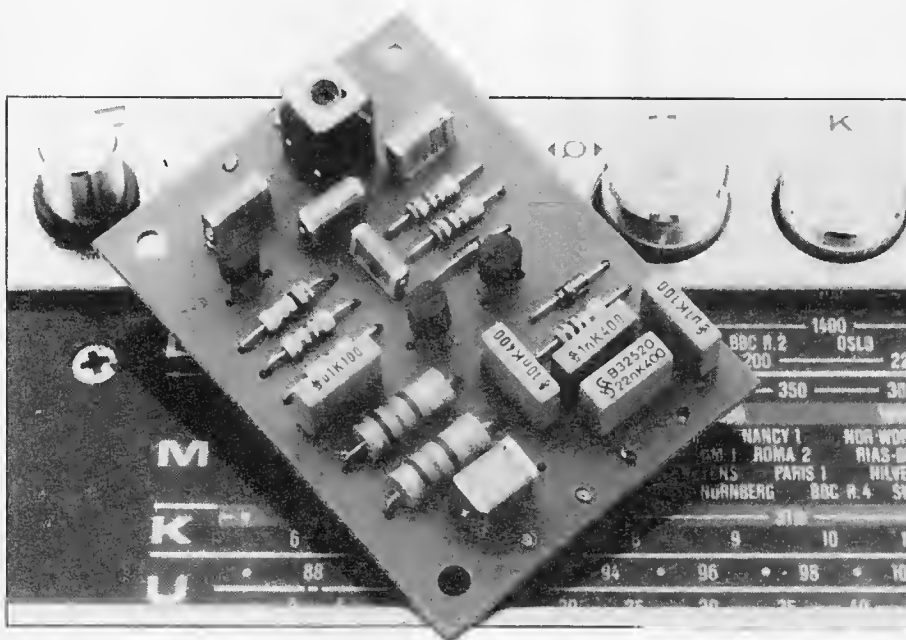
- X1 = 10.000 MHz quartz crystal, 30 pF (AT) series resonant; HC18/U enclosure.
- K1 = 20-way, double row, male header.
- PCB Type 87160 (available through the Readers services).

* Signetics-Philips • Mullard House • Torrington Place • London WC1E 7HD. For UK distributors see In-Card 500 (EE May 1987).

It is regretted that software for computers other than the C64 and those in the MSX series is not available.

SSB ADAPTER

A low-cost add-on unit that enables single-sideband reception on virtually any AM short-wave receiver.



Every experienced short-wave listener knows that single sideband (SSB) transmissions can not be received unless a special detector is installed in the receiver. Unfortunately, however, an SW receiver suitable for SSB reception is generally far more expensive than an AM/FM general coverage, radio set having adequate sensitivity and selectivity. To the dedicated SW listener, amateur and utility stations transmitting SSB signals are often more interesting than (broadcast) AM stations in view of their uniqueness, and the larger distance covered. SSB transmitters are more economical than AM transmitters as regards bandwidth and power consumption, due to the absence of the carrier and the second sideband.

Carriers and sidebands

It can be shown that the RF power contained in the carrier and one sideband of an AM modulated RF signal is redundant, because it is not, strictly speaking, needed to convey in-

formation to the receiver. For this purpose, one sideband suffices. The RF output signal of an AM transmitter modulated with a single, sinusoidal frequency is shown in Fig. 1. The instantaneous amplitude, U , of the RF carrier is a function of the amplitude of the modulating AF tone, which can be reconstructed by drawing a line along the peak excursions of the RF voltage (the *envelope*

waveform). Mathematically simplified, this AM signal is described by the expression

$$U_{AM} = U_c(f_c) + mU_c(f_c + f_m) + mU_c(f_c - f_m).$$

It is seen that the amplitude of the AM signal, U_{AM} , is the sum of 3 terms. The first, $U_c(f_c)$, is the amplitude of the carrier with frequency f_c . The second and third term are of equal am-

plitude, mU_c , but denote signals adjacent to the carrier, i.e., below and above f_c . The factor m represents the relative amplitude of the modulating signal with frequency f_m . Figure 2a shows an analysis of the AM signal in the frequency domain (*spectrum analysis*). The carrier is modulated with a single tone that gives rise to 2 side tones, each having a lower amplitude than the carrier. Modulating the AM transmitter with a composite AF signal, e.g. music or speech, causes two *side bands* rather than side tones adjacent to the carrier, and so increases the overall bandwidth occupied by the signal. Returning to the above formula, it is readily seen that the terms $mU_c(f_c + f_m)$ (upper sideband, USB) and $mU_c(f_c - f_m)$ (lower sideband, LSB) convey the same intelligence, namely the modulated signal, while the carrier, $U_c(f_c)$, does not convey any intelligence. Evidently, the carrier and one sideband are not needed to convey information from the transmitter to the receiver, and this forms the basis of the SSB modulation method, which is sometimes—more properly—

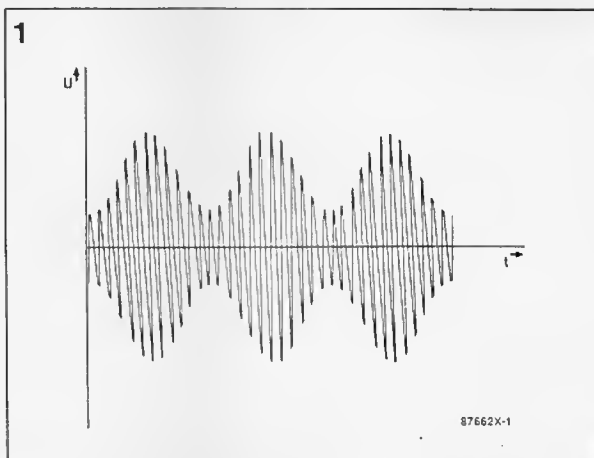


Fig. 1 RF output signal of an AM transmitter modulated with a single tone.

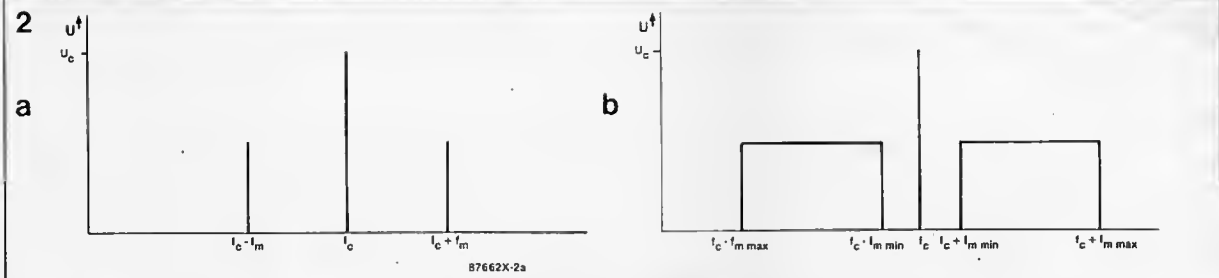


Fig. 2 Spectral analyses of AM signals. Fig. 2a: single tone modulation. Fig. 2b: modulation with a composite AF signal.

Parts list

Resistors ($\pm 5\%$):

- R₁ = 10K
- R₂; R₃ = 1K0
- R₄ = 100R
- R₅ = 4K7
- R₆ = 6K8
- P₁ = 100R linear potentiometer

Capacitors:

- C₁ = 200p variable capacitor
- C₂ = 6n8
- C₃; C₄ = 39n
- C₅ = 2n2
- C₆; C₇ = 100n
- C₈ = 1n0
- C₉ = 10n
- C₁₀ = 22n
- C₁₁ = 470n

Inductors:

- L₁ = Neosid Type 7A1S* inductor assembly (see text)
- L₂; L₃ = 270µH

Semiconductors:

- D₁; D₂ = 1N4148
- T₁; T₂; T₃ = BF494

Miscellaneous:

- S₁ = miniature SPST switch.
- PP3 battery (9 V) and clip-on connector.
- PCB Type 87662X (not available through the Readers Services).
- Suitable metal enclosure.

* Neosid inductor assemblies are available from

- Neosid • Eduard House • Brownfields • Welwyn Garden City • Hertfordshire AL7 1AN. Telephone: (0707) 325011. Telex: 25423.

or from

- Bacton Inductive Components Limited • Unit 8b • Cambridgeshire Business Park • Angel Drove • Ely • Cambridgeshire CB7 4DT.

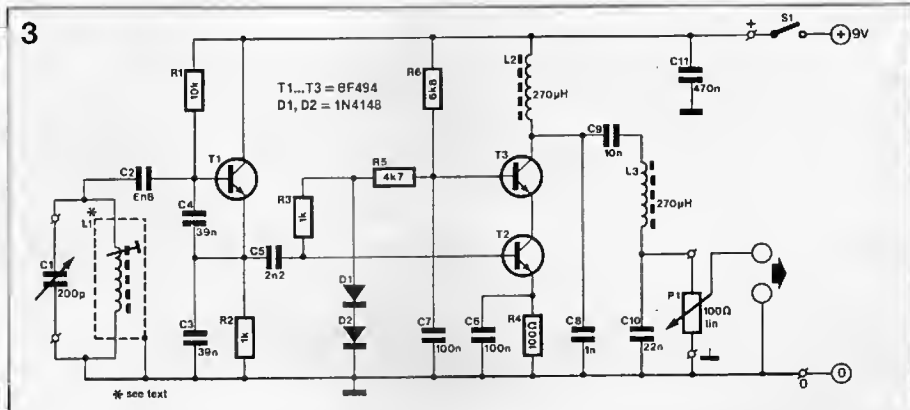


Fig. 3 Circuit diagram of the SSB adapter for AM receivers.

referred to as SSBSC (single sideband, suppressed carrier). In theory, an SSB transmitter uses only a quarter of the power of an AM transmitter for conveying the same information. In an AM transmitter, half the power is "wasted" in the carrier, the other half goes into the sidebands. The RF power of an (ideal) SSB transmitter drops to nought in the absence of a modulating signal. Hence an SSB transmitter has a far better power efficiency than an AM transmitter, and at the same time occupies less bandwidth; relatively low power SSB transmitters can, therefore, be used to cover considerable distances (maritime communications, radio amateurs, etc.)

without laying too heavy a claim on the available power source.

From SSB to AM

The operating principle of the present adapter follows from the previously discussed relationship between AM and SSB. An SSB signal can be converted into AM by adding a carrier and a sideband. Both are obtained with the aid of an oscillator tuned to the receiver's intermediate frequency (IF), which is usually 455 kHz. The externally generated carrier serves as the reference frequency against which the (upper or lower) sideband is demodulated. The second sideband is automatically obtained

in this process, so that a double sideband AM signal is available for demodulating. The combination of the AM receiver and the SSB adapter is, understandably, not up to a real SSB compatible receiver with its special, narrow band, IF section. None the less, the results obtained with the present add-on unit are satisfactory for relatively strong, interference-free, signals.

Circuit description

The SSB adapter is a simple circuit, shown in Fig. 3. Transistor T₁ oscillates at 455 kHz with the aid of parallel tuned circuit C₁-L₁. The oscillator signal is raised and filtered in a cascode amplifier set up around T₂ and

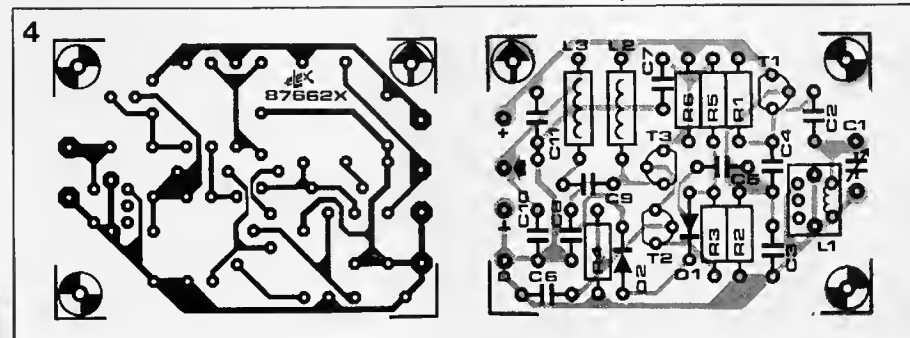


Fig. 4 Printed circuit board for the SSB adapter.

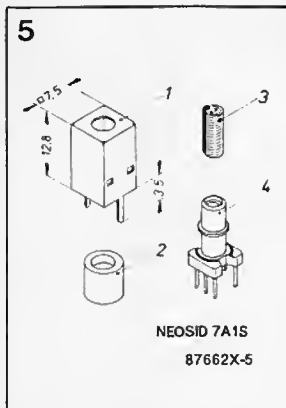


Fig. 5 The Type 7A1S inductor assembly from Neosid. 1: screening can. 2: ferrite cup. 3: iron dust core. 4: ABS former and base.

T₃. The amplitude of the output signal is made variable with P₁, enabling optimum performance with any receiver. The adapter's output signal is connected direct to a length of insulated wire, wound as 1 or 2 turns around the receiver (inductive coupling). The adapter is fed from a 9 V battery.

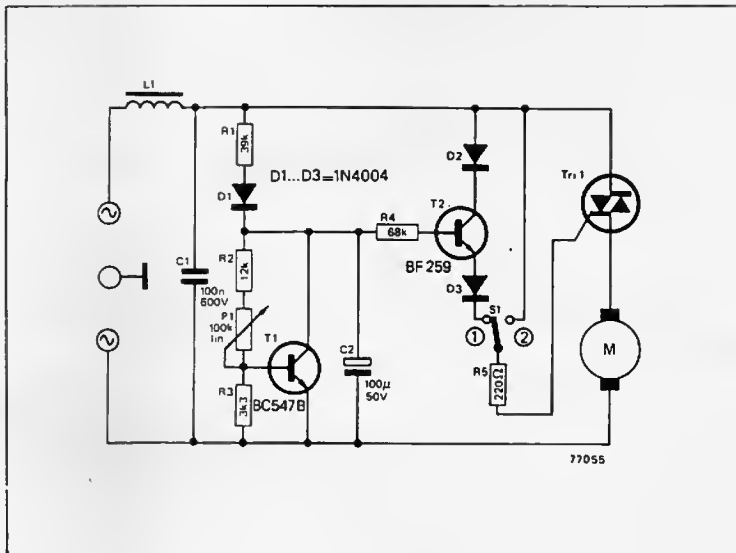
Construction and alignment

The printed circuit board for the SSB adapter is shown in Fig. 4. Construction is straightforward with the possible exception of inductor assembly L₁ — see Fig. 5. Viewed from underneath, the base of the ABS former in the Type 7A1S as-

sembly has 5 pins, 3 at one side and 2 at the other. Inductor L₁ is connected to the latter 2 pins. Close-wind 53 turns of $\varnothing 0.2$ mm (36 SWG) enamelled copper wire onto the 2 sections of the former, and make sure that the ferrite cup (part 2 in Fig. 5) can be fitted on top. Secure the winding with a piece of Sellotape. Check the continuity at the base, and fit the former onto the PCB. Carefully slide the screening can over the former, then push-fit and solder its mounting tabs in the holes provided. Make sure that the top end of the former fits snugly in the hole in the top of the screening can. Tuning capacitor C₁ and level control P₁ are fitted as external components.

It is recommended to fit the SSB adapter in a metal enclosure to prevent spurious radiation. Set variable capacitor C₁ to the centre position, and P₁ to maximum. Connect the coupling loop around the receiver to the adapter output. Tune the receiver to an AM broadcast station, and switch the adapter on. Adjust the core in L₁ with a non-magnetic trim tool until a whistle (*beat note*) is heard in the receiver. Lower the frequency of the beat note by adjusting L₁, until it is no longer audible (zero beat tuning). Switch off the adapter, and tune the receiver to an SSB station. Switch the adapter on again, and adjust C₁ and P₁ until the speech becomes intelligible. B

DRILL SPEED CONTROL



Most drill speed controllers suffer from one or more drawbacks. These include poor speed stability, excessive instability at low speeds, and high power dissipation in the series resistor used to sense motor current. The circuit described here suffers from none of these drawbacks, and in addition is extremely simple.

The mains input is rectified by D1 and dropped by R1. The current drawn by T1 can be controlled by means of P1, thus also controlling the DC voltage that appears across C2, and hence at the base of T2. T2 is connected as an emitter follower, and the voltage appearing at the cathode of D3 is about 1.5 V less than the base voltage of T2. Assuming that the motor is turning but that the triac is turned off, the back e.m.f. generated by the motor will appear at the T1 pin of the triac. So long as this voltage exceeds the cathode voltage of D3 the triac will remain turned off, but as the motor slows down this voltage will fall and the triac

will trigger. If the load on the motor increases, thus tending to slow it down, the back e.m.f. will fall more quickly and the triac will trigger sooner, thus bringing the motor back up to speed.

Since the triac can be triggered only on positive half-cycles of the mains waveform the controller will not vary the motor speed continuously from zero to full speed, and for normal full-speed running S1 is included, which turns the triac on permanently. However, the circuit exhibits good speed control characteristics over the important low speed range.

L1 and C1 provide suppression of r.f. interference generated by the triac. L1 can be a commercially available r.f. suppression choke of a few microhenries inductance. The current rating of L1 should be from two to four amps, depending on the current rating of the drill motor. Almost any 600 V 6 A triac can be used in the circuit.

SWITCH-MODE POWER SUPPLIES

Recent advances in power semiconductor technology and inductive components have boosted the use of compact, high efficiency power supplies of the switch-mode type. Now SMPSs of various power ratings are becoming widely available at reasonable prices, it seems timely to focus on their design principles and practical aspects.

Most electronic circuits can not work without a power supply of some kind. The basic mains supply consists of a transformer, a rectifier, a filter (smoothing/reservoir capacitor), and a linear control circuit (regulator) for adjusting the output voltage to the desired value. It may be argued that the basic power supply has a number of important disadvantages. For relatively high powers, the mains transformer is often bulky and expensive, and the same goes for the smoothing capacitor(s). Moreover, the product of the voltage drop across the regulator and the current consumption of the load forms dissipated, and therefore wasted, power, which results in a very low overall ef-

iciency, especially at relatively low output voltages. Not surprisingly, in the rapidly expanding world of microelectronics there arose a growing need for a high-efficiency power supply.

This need was met by the switch-mode power supply (SMPS), in which the output power is not regulated continuously, but pulsed at a relatively high frequency. An output filter is included for smoothing the supply voltage.

The filter components can be kept small thanks to the high frequency, and the same goes for the (toroidal) transformer if galvanic insulation is required.

Basic configurations

A switch-mode power supply is essentially a DC-DC converter. The 3 basic circuit configurations are shown in Fig. 1. The **flyback circuit** works as follows. A magnetic field builds up in the inductor as long as the switch remains closed. When the switch is opened, the inductor functions as an energy source. The voltage across the inductor is reversed, and the conducting diode passes the energy to the reservoir capacitor. Note that the output voltage is reversed with respect to the input voltage.

The **forward converter** does not reverse the polarity of the input voltage. The capacitor is charged via the inductor when the switch is closed. The difference between the input and the output voltage is available on the inductor. In contrast to that in the flyback converter, the switch is closed when the capacitor is being charged. When the switch is opened, the magnetic field of the inductor is weakened via the flyback diode. The switch is, of course, a power transistor, and the di-

ode affords protection against the induced voltage. In a forward converter, the input voltage is higher than the output voltage.

The third basic configuration is referred to as **boost or step up converter**. This circuit increases the input voltage, and is functionally similar to the flyback converter. Energy is stored in the inductor when the switch is closed. When the switch is opened, this energy is supplied to the load at the output via the diode.

The continuous and discontinuous mode

Two modes of operation can be distinguished, depending on the current in the inductor—see Fig. 2.

After closing the switch, the current in the inductor increases linearly up to a specific maximum value ($U_i = \text{constant}$). After opening the switch, the current decreases linearly. The

circuit operates in the *discontinuous current mode* if the current is nought in every period. The capacitor supplies the load current during the remainder of the period. The discontinuous mode is characterized by the good response of the closed regulation circuit to fluctuations in the input voltage (*line regulation*), and the output load (*load regulation*). There is no energy in the inductor at the start of each period, and regulation can, therefore, take place on a period-to-period basis. It can, in fact, be argued that the inductor is not present in the regulator circuit. The maximum phase shift of 90° in the buffer capacitor ensures the stability of the closed regulation circuit. A disadvantage of the discontinuous mode is the relatively high peak current carried by the power switch. Flyback and step up converters usually operate in the discontinuous mode.

In the *continuous current*

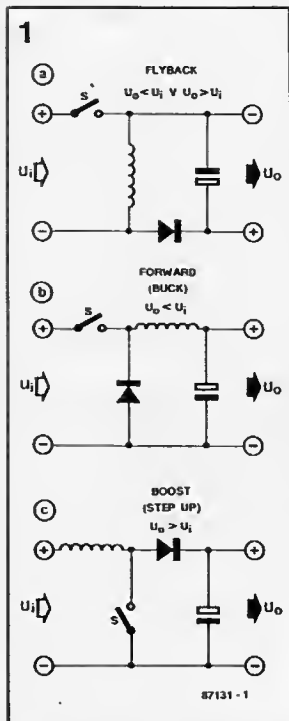


Fig. 1 The 3 basic configurations of a switch-mode power supply.

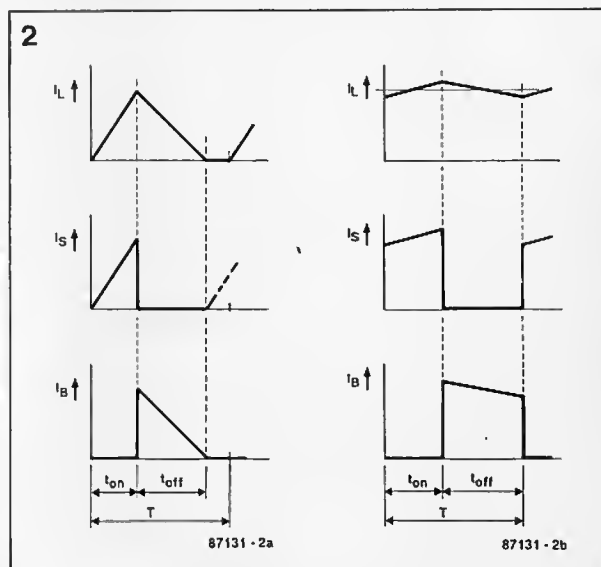


Fig. 2 The discontinuous (a) and the continuous current mode (b) differ in respect of the current carried by the inductor.

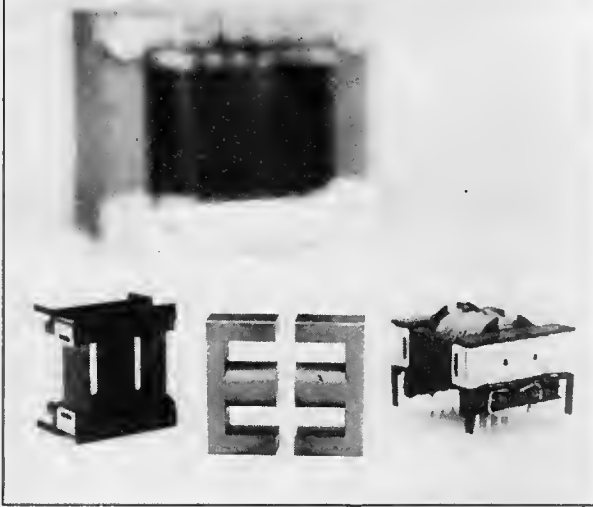


Fig. 3 A traditional soft iron core (background) and a modern ETD ferrite core of equal power rating.

mode, the current through the inductor does not drop to nought at the end of every period. The ripple current in the inductor is small relative to the load current, and this requires a fairly high self-inductance. The buffer capacitor, on the other hand, can be kept relatively small. The favourable shape factor of the current through the power transistor and the diode makes the continuous mode eminently suitable for high power applications. The response to load fluctuations is, however, worse than that of a circuit in the discontinuous mode. Each change in the output load current requires a corresponding change in the direct current through the (relatively large) self-inductance, and this process may take several periods to complete.

It is not possible for a system to automatically switch from continuous to discontinuous operation, or vice versa, because this would cause a considerable change in the open loop transfer characteristics, giving rise to instability of the closed regulation system. This means that the load current of a system in the continuous mode should be higher than half the peak-to-peak value of the ripple current in the inductor. Forward converters usually operate in the continuous current mode.

Off-line operation

In many cases, the input voltage is a rectified and smoothed

voltage obtained direct from the mains. The direct voltage so obtained (approx. 335 V at a 240 VAC mains supply) is rarely used for converting down to, say, 12 V, in view of the resultant low duty factor, and the need for large self-inductances. Also, a direct connection to the mains is dangerous, and normally not permitted. This calls for a (ferrite) transformer, which, in an SMPS, has the advantage of being much smaller than a soft iron type used in the traditional 50 Hz mains supply (see Fig. 3). There are, however, a number of important considerations as to keeping the losses of the core material within acceptable limits. Ferrite is used instead of laminated iron, and offers a number of advantages. The construction of a ferrite core is relatively simple, and ferrite is a light and insulating material.

The turns ratio of the ferrite transformer enables converting the high input voltage down to a value close to the desired output voltage. The conversion increases the duty factor, and hence reduces peak currents in the power transistors.

Transformer circuits

The simplest configuration of an SMPS is the single transistor flyback converter shown in Fig. 4a. This circuit is well-known in low power supplies with an output rating up to about 250 W. In this application, the transformer is more properly referred to as a coupled self-inductance,

because it assumes the function of the inductor shown in Fig. 1a. The forward converter in continuous mode is more suitable for feeding relatively heavy loads. The most commonly found version is based on a single transistor and a demagnetization winding in the primary circuit—see Fig. 4b. The transistor must be able to handle twice the input voltage. The demagnetization winding can be omitted if the circuit is extended with a transistor and a flyback diode as shown in Fig. 4c. In this circuit, the transistors need only withstand half the voltage. They are, however, driven with respect to different potentials, just as in the bridge circuits to be discussed.

Half or full bridge circuits are used mainly for high power applications. The full bridge variant is suitable for very heavy loads thanks to the fact that the effective input voltage is doubled. The last variant, shown in Fig. 4g, is also a bridge circuit, based on a centre-tapped transformer that enables the transistors to be

driven with respect to a common reference potential.

Bipolar transistors as well as power FETs can be used in the primary circuit. Bipolar technology is suitable for switching frequencies up to 50 or 100 kHz. Power FETs are faster, and can be used at higher frequencies without running into excessive switch losses. Currently, the maximum usable frequency is about 1 MHz, and power FETs are expected to become predominant in SMPSs in view of the ever increasing switching frequencies. Power FETs for relatively high voltages are, however, still quite expensive, and more attractive for use in countries with a 117 V mains supply, such as the USA.

Core saturation

Any transformer winding forms a self-inductance, and the average voltage across it should, therefore, be nought. When this is not so, the remaining direct voltage causes an linearly increasing direct current until the core is saturated. The

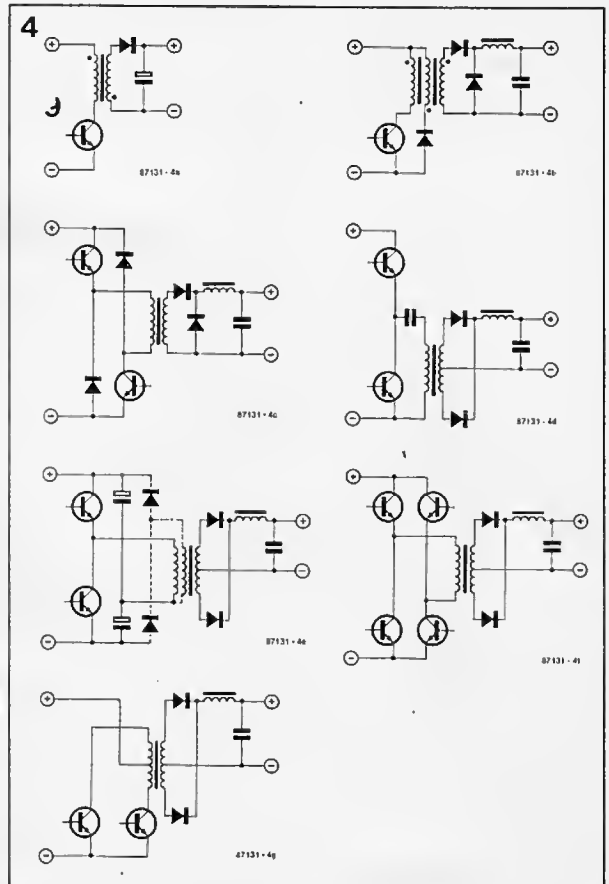


Fig. 4 Various configurations of the switching power stage in an SMPS.

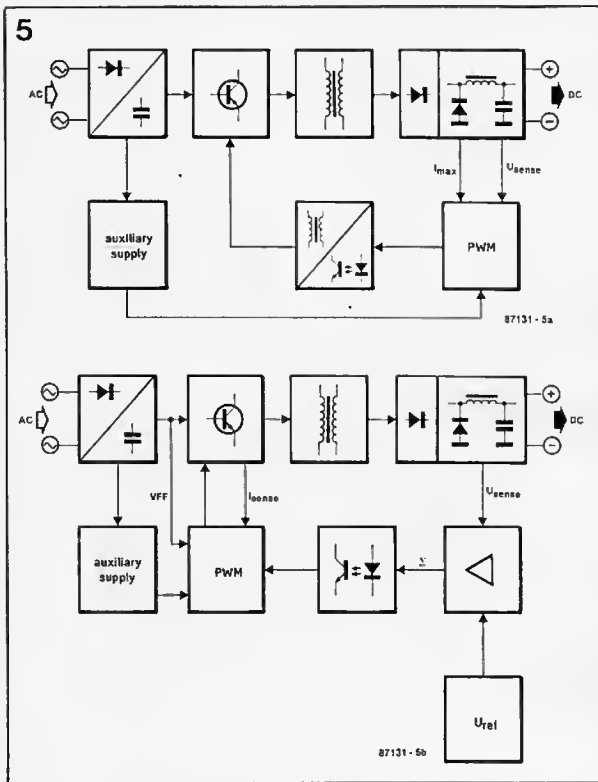


Fig. 5 Functional diagram of a SMPS with the control electronics located at the secondary side (a) or the primary side (b).

H-field, and with it the current, then increases exponentially, in accordance with Faraday's law of constant increase of the magnetic flux per unit of time. This effect must be prevented because it can lead to destruction of the primary circuit. In the circuits of Fig. 4b and 4c, the field in the transformer core is weakened with the aid of the flyback diode(s), but only as long as the duty factor remains below 50%. Problems owing to permanent magnetization are not expected to arise in the circuit of Fig. 4d, where the coupling capacitor ensures the absence of direct current through the primary winding. The situation is more complex in Fig. 4e. Although the primary winding is AC coupled, a direct voltage may still exist at the junction of the capacitors. This positive or negative potential may arise from less than perfect (i.e., unbalanced) driving of the power transistors, which may have different recovery times also. Unbalancing of the primary circuit can be prevented with the aid of a compensation winding and 2 flyback diodes. A coupling capacitor for blocking the magnetizing

direct current is rarely used in the circuit of Fig. 4f, because this is rated for very high power. Both the positive and the negative current in the primary winding are measured, and any difference between them is compensated by controlling the duty factor. This safety measure can be applied to the circuit of Fig. 4g also.

Voltage control

In a switch-mode power supply, the output voltage is measured, compared to a reference, and kept constant by controlling the duty factor of the drive signal applied to the power switches (i.e., transistors). The regulating effect of the control circuit depends on the open-loop characteristics of the system. The simplicity of the flyback converter makes it less suitable for many purposes, since the duty factor depends primarily on the load at a constant output voltage. Good voltage regulation requires a high amplification of the measuring and control circuit. In a forward converter, the voltage control circuit need not have a strongly regulating effect because in essence the output voltage

depends only on the turns ratio of the transformer (assuming a constant input voltage).

There are 3 basic types of voltage control system:

- Direct duty factor control.

The error signal is amplified, and drives a pulsewidth modulator, which in turn adjusts the duty factor as required. A high overall amplification is needed, at the cost of some stability—notably in the case of converters operating in the continuous mode.

- Voltage feedforward. This is the most commonly used system. Preregulation of the duty factor is implemented as a function of the input voltage, enabling the output voltage of the open loop system to be made independent of the input voltage. The control circuit is, therefore, only required for compensating load fluctuations. The preregulation system improves the line regulation, and so ensures sufficient suppression of hum.

- Current mode control. A second control circuit (inner loop) inside the voltage control circuit (outer loop) enables switching off the power transistor at a more or less fixed peak value of the current. The effect so obtained is the (quasi) disappearance of the inductor from the output filter. The whole system is then essentially a first order network with the capacitor in the output filter as the only phase changing element. The stability of the whole supply, as well as the response of the closed system to fluctuations in the input voltage and

the load current, is excellent. High power forward converters of the continuous type are often equipped with a current mode control for obvious reasons.

Location of the control circuit

The voltage control circuit can be located at the primary or the secondary side of the supply. A circuit at the primary side (Fig. 5b) makes it possible to drive the power stage direct from the IC, while it is a relatively simple matter to implement circuits for primary current monitoring and preregulation. A disadvantage of the primary location is the need for an insulating device in the control loop for transmitting an analogue signal from the primary to the secondary side. This is usually done with the aid of an optocoupler. Only the error signal is transmitted to rule out instability of the output voltage owing to ageing effects in the optocoupler.

A control circuit at the secondary side (Fig. 5a) enables direct coupling of the voltage control circuit. Also, it allows the use of the reference circuit built into many of the currently available integrated SMPS controllers. The PWM signal is fed to the power stage via a fast optocoupler, or a special pulse transformer. Differences in the drive applied to several power transistors are relatively simple to monitor and correct, but the primary location makes it difficult to keep tabs on the primary current.

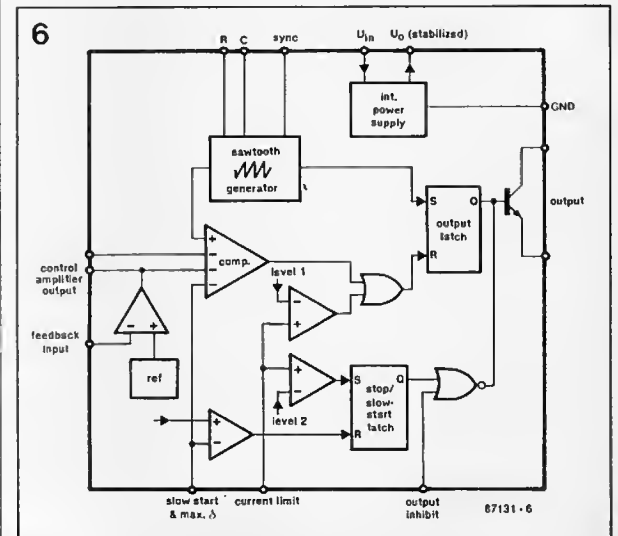


Fig. 6 General block diagram of an integrated SMPS controller.

Whatever control circuit is used, it must have its own (start) supply. The self-generated voltage is, of course, only usable after the supply is fully operational.

SMPS controllers

A wide variety of integrated circuits is currently available for controlling switch-mode power supplies. These ICs are essentially very similar, and a general block diagram is therefore given in Fig. 6 to explain their operation.

The pulsewidth modulator is composed of a sawtooth generator, a voltage comparator and a set-reset bistable. A second input on the comparator is connected to the output of an opamp that amplifies the difference between the real and the required (set) output voltage. An accurate, and temperature compensated, voltage reference is often included for adjusting a specific output voltage.

The remainder of the circuits in the chip have auxiliary functions, and serve for various types of protection. Voltage feedforward or current mode control is possible by changing the slope of the sawtooth signal. Special control inputs make it possible to set a duty factor of no more than 0.5. An analogue input is activated above a predefined voltage level, and can be used for making a 2-level short-circuit protection. When the output current reaches the first level, the duty factor is held constant, and the circuit supplies a constant, maximum, current. The duty factor is made no more than 0.5 above level 2. A digital input enables remote control of the supply (computer-controlled test sites, etc.).

An input for setting the maximum duty factor is standard on most SMPS controllers. Properly driven, it prevents saturation of the transformer core, and hence an exponentially rising primary current. This safety measure is especially useful for supplies operating in the continuous current mode. In these, the duty factor has a tendency to rise to the maximum value at each change in the output current, in an effort to correct the direct current through the inductor in the output filter in accordance with the new load.

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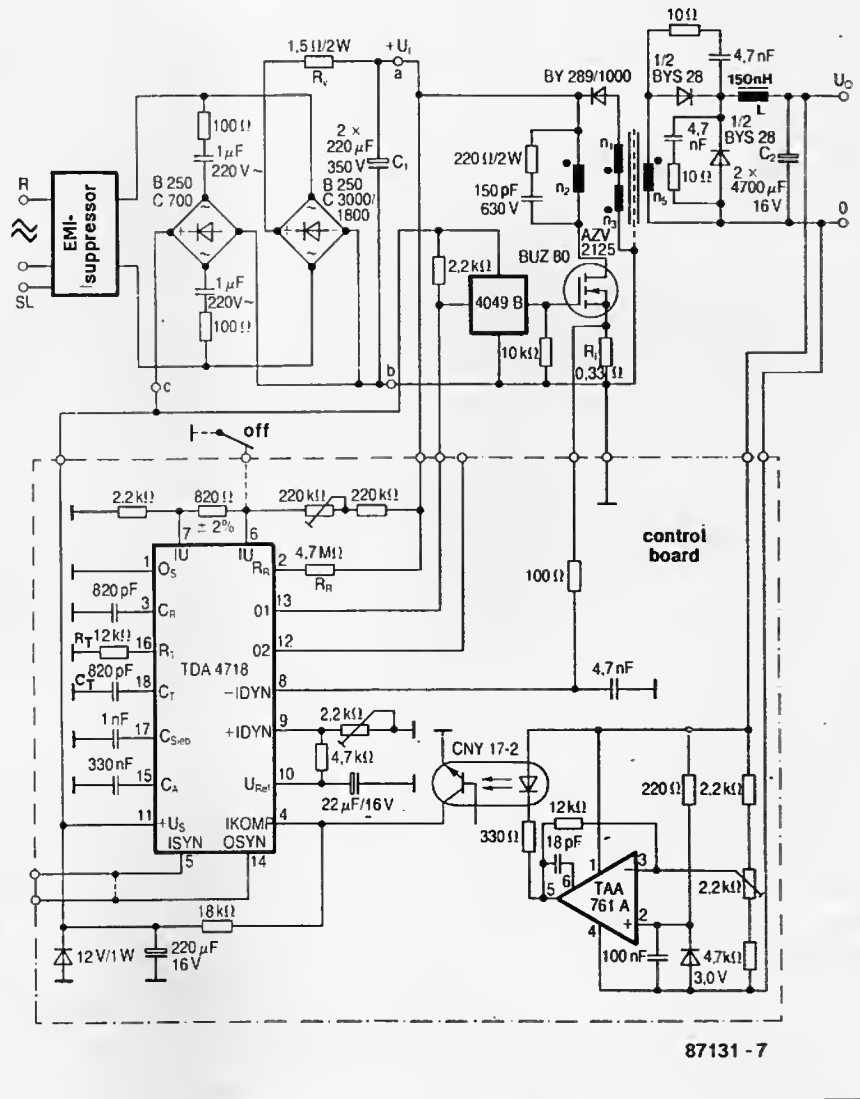


Fig. 7 Circuit diagram of a 5 V; 20 A; 50 kHz supply (courtesy Siemens).

The multi-voltage supply

Additional supply voltages are fairly simple to implement in a SMPS with the aid of appropriate auxiliary windings on the secondary of the transformer. In computer equipment, the usual combination is a powerful 5 V section, and auxiliary ± 12 V supplies. Voltage control is usually only effected on the +5 V rail. If the magnetic coupling between the secondary windings on the transformer is sufficiently tight—this is typical of a well-constructed transformer—the output voltage of the auxiliary windings is regulated along with the main supply, at acceptable accuracy.

Losses

Switch-mode power supplies are known mainly for their high efficiency. In spite of this, some power is, of course, wasted.

- Switching losses in the power transistors. Faster switching—e.g. with the aid of a speed-up capacitor—keeps these losses acceptable. The power loss incurred in the conductive transistor is relatively small, especially at high input voltages.
- Transformer losses can be classified as copper or core losses. At switching frequencies below 100 kHz, copper losses are the main consideration in finding the optimum specifications of the

transformer in the SMPS. Also, care should be taken to counter a considerable skin effect, which reduces the effective diameter of the copper wire as the frequency increases. Many SMPS manufacturers use litz wire for their transformers to prevent losses arising from the skin effect.

Core losses are due to eddy currents and hysteresis of the ferrite material. They depend on the so-called *flux density sweep*, and the frequency. Manufacturers of ferrite cores can supply graphs to establish the maximum permissible core losses as a function of the thermal resistance of the core, and other parameters. Core losses

form the crux in designing a transformer for use in an SMPS.

• Rectification and filter losses. These become more serious at relatively low output voltages (5 V), and are mainly due to the forward voltage drop across the diodes. Schottky diodes are often used in view of their low forward voltage drop and good switching characteristics. Some power is also wasted in the inductor as part of the output filter.

A practical circuit

The circuit diagram of a typical switch-mode power supply is shown in Fig. 7 (Siemens Application).

Buffer capacitor C_1 is fitted at the output of a bridge rectifier, which is fed from a mains filter. Power resistor R_v limits the peak charge current when the supply is switched on. The type of transformer used makes clear that the circuit is a forward converter. The primary and secondary winding (N_2 ; N_3) are in phase, as indicated by the dots. Auxiliary windings N_1 and N_3 serve to demagnetize the core. The dotted line in the core indicates the use of an electromagnetic shield. The configuration of the secondary output filter shows that the supply is designed for operation in the discontinuous current mode. A relatively small self-inductance is used in conjunction with a large buffer capacitance to ensure sufficient output power when the inductor carries no current. The output voltage is divided and compared to a 3 V reference. The error signal from the operational amplifier is fed to the TDA4718-based primary control circuit via an optocoupler. Components C_r and R_r define the switching frequency of 50 kHz. Provision has been made to set the maximum output current (pin 9). The maximum primary current is monitored with the aid of the network connected to pin 8. The capacitor fitted at pin 15 of the controller ensures a gradual increase of the duty factor after power-on (*soft start*). The input voltage is checked via pin 6 and 7. The duty factor is made nought when the input voltage is either too low or too high. Voltage feedforward is implemented with the aid of R_r , connected to pin 2. The switch

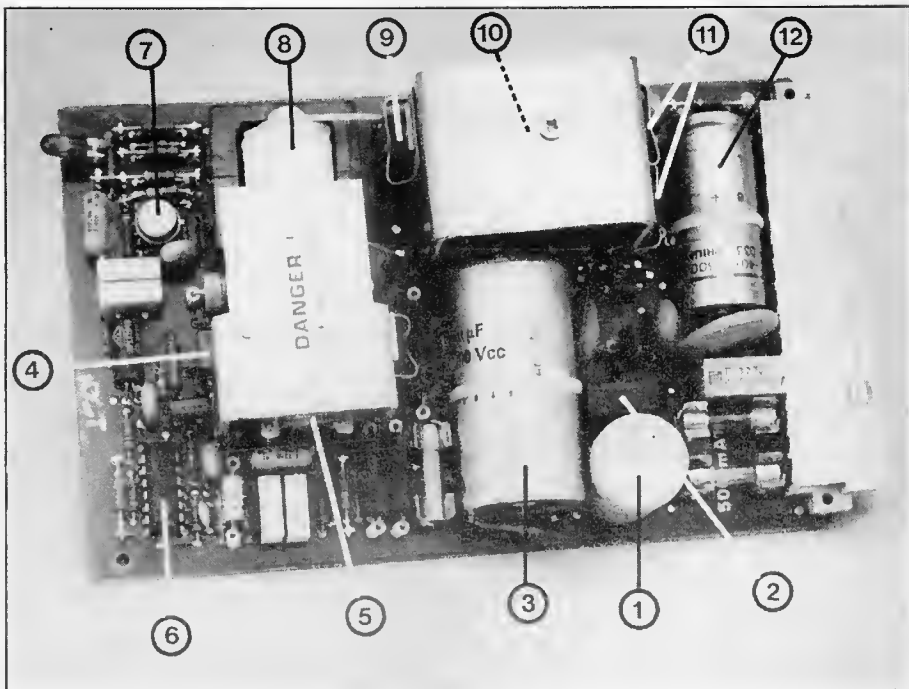


Fig. 8 A compact switch-mode power supply.

signal at the output of the controller is fed to the power MOSFET via a number of parallel CMOS buffers contained in a CD4049 package. The control circuit is fed from the mains via a capacitive voltage divider. Figure 8 shows a compact SMPS fitted on a printed circuit board. The arrows point to the following, essential, parts: (1) mains filter; (2) primary rectifier; (3) buffer capacitor for primary voltage; (4) switching transistor; (5) pulse transformer for base drive; (6) SMPS controller for pulsewidth modulation; (7) (secondary) voltage reference and error amplifier; (8) ferrite core transformer; (9) primary flyback diode for weakening the transformer field; (10) inductor in output filter; (11) secondary rectifier and flyback diode; (12) output capacitor.

Further developments

The scope of this introductory article does not allow a detailed discussion of all the technical considerations that go into designing a switch-mode power supply. In a forthcoming issue of *Elektor Electronics* the subject will be reverted to in the context of a construction project.

The theoretical aspects of the SMPS have been known for some time, but it was not until the coming of fast power tran-

sistors, integrated controllers and new ferrite materials that serious development of the SMPS was launched. Ever higher switching frequencies make it possible to reduce the size of the secondary filter, but at the same time pose a real difficulty as regards electromagnetic interference (EMI) due to the often large number of strong harmonics and other spurious products. It is with this in mind that there is a growing interest in free-running supplies, in which the currents are sinusoidal rather than rectangular. Meanwhile, supplies of the types discussed in this article are constantly reworked and enhanced to make the current consumption from the mains sinusoidal. This ensures less interference on the mains, a higher efficiency of the rectifier thanks to the more favourable shape factor of the current, and reduced peak currents in the buffer capacitor(s).

TW

For further reading:

- *High frequency power transformer and choke design. Part 1... 4 incl.*; Philips Technical publication, September 1982.
- *Electronic Components & Applications. Vol. 2; No. 1*; Philips, November 1979.

- *Unitrode Switching Regulated Power Supply Design Seminar Manual*; Unitrode, ed. 1986.
- *Schaltnetzeile (SNT), Technik und Bauelemente*; Siemens Technische Mitteilung No. B9-B3269.
- *Steuer und Überwachungsschaltungen fuer moderne Schaltnetzeile (SNT), TDA47xx Familie*; Siemens Technische Beschreibung No. B/3132.
- *Integrierte Schaltnetzeil-Steuerschaltungen, Funktion und Anwendung*; Siemens Technische Beschreibung No. B1-B3116.
- *SGS power supply application manual*; July 1985.
- Various databooks and application notes: Intersil; Mullard; SGS; Siemens; Thomson; Unitrode.

THE POSITIVE IMPEDANCE CONVERTER

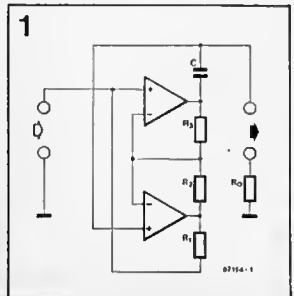
by A.B. Bradshaw

One of the very practical means of simulating inductance in electrical circuits is by the use of gyrators. The positive impedance converter is a member of this family. Its main use is to replace wound inductors in AF circuits, particularly where these have large values, or to simulate coils at very low frequencies.

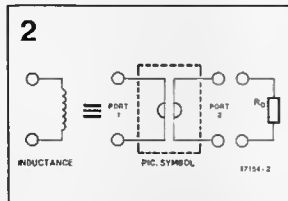
The positive impedance converter makes use of operational amplifiers: two are needed to simulate a grounded inductor; four are required to simulate a balanced inductor. But only four opamps are needed to simulate a node, containing balanced and grounded inductors, as will be shown later. Where simulation of grounded inductors is used, the opamps should be operated from balanced power supplies.

The writer has used the PIC as a circuit element for a number of years in the design of high-performance AF filters. The Q values obtained with these devices is very much higher than that of the wound equivalent. Instability problems are rare. Although the frequency response of the opamps usually limits the operating range of PICs to about 40 kHz, this is adequate for most AF and even a number of data filtering requirements.

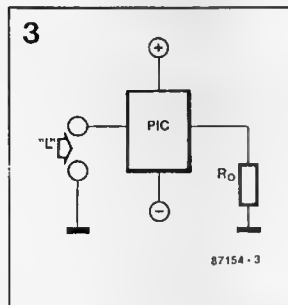
The circuit diagram of a typical PIC is shown in Fig. 1. The input impedance of this circuit appears as a pure inductance when the output is grounded through R_o . Resistors R_1 , R_2 , and R_3 , as well as capacitor C ,



are normally 1% types. Resistor R_o is used as the inductance setting component.



The general symbol of a PIC is shown in Fig. 2, but in practical circuits, when used as a circuit element, it is usually indicated as in Fig. 3.



Analysing the PIC

The analysis of gyrators is normally performed with the aid of matrix algebra and the formal nodal analysis of network theory. The formal approach has a lot to offer as regards generality, but it sometimes tends to obscure the practical operation of the circuit. Because of this, the writer has adopted an approach which will be familiar to most readers. Assuming that opamps are perfect, the analysis can be

simplified considerably. In an ideal opamp,

- the voltage gain is infinite: $A_{vo} = \infty$;
- the input resistance is infinite: $I_{in} = \infty$;
- the output resistance is zero: $I_{out} = 0$;
- the bandwidth is infinite: $BW = \infty$;
- there is zero input offset voltage: $E_o = 0$ if $E_{in} = 0$.

Since the voltage gain is infinite, any output voltage is the result of an infinitely small input voltage. In effect, therefore, the **differential input voltage is zero.**

The preceding assumptions are used as axioms in the following. For the purposes of examining the operation of a PIC, it is redrawn in Fig. 4. Some of the voltages and currents are shown twice to emphasize the circuit action.

$$(V_1 - V_2)/R_3 = I_3 \quad \text{Eq. 1}$$

$$(V_1 - V_4)/R_2 = I_5 \quad \text{Eq. 2}$$

$$(V_1 - V_4)/R_1 = I_2 \quad \text{Eq. 3}$$

Since the input resistance of the opamps is infinite, $I_o = 0$, whence

$$I_3 = I_5$$

and

$$I_2 = I_{in}$$

From Eq. 3:

$$V_1 - V_4 = R_1 I_2 = R_1 I_{in}$$

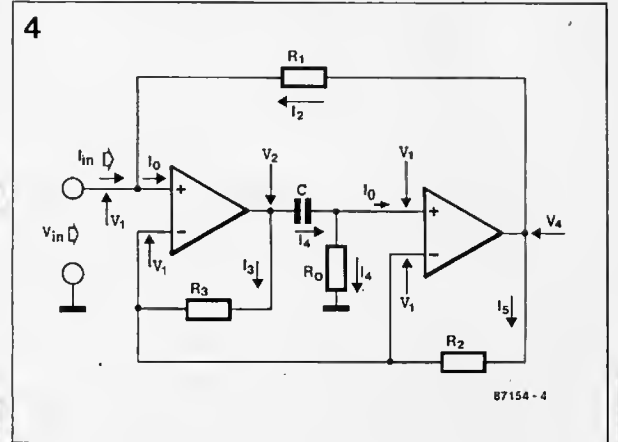
Dividing both sides by R_2 gives

$$(V_1 - V_4)/R_2 = I_{in} R_1/R_2 \quad \text{Eq. 4}$$

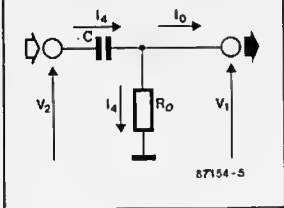
Since $I_3 = I_5$, Eq. 1 is equal to Eq. 2, so that

$$(V_1 - V_2)/R_3 = (V_1 - V_4)/R_2 = I_{in} R_1/R_2 \quad \text{Eq. 5}$$

To remove V_2 from Eq. 5, consider the middle section of Fig. 4, which for convenience's sake is reproduced in Fig. 5.



5



$$V_1 - V_2 = I_4 = j\omega C$$

Also, since $I_0 = 0$, $I_4 = V_1 / R_0$

Combining these expressions gives

$$V_1 - V_2 = V_1 / j\omega C R_0 \quad \text{Eq. 6}$$

Dividing both sides of Eq. 6 by R_3 gives

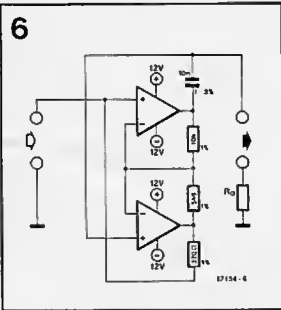
$$(V_1 - V_2) / R_3 = V_1 / j\omega C R_0 R_3$$

in which the left-hand term is also that of Eq. 5, whence

$$I_{in} R_1 / R_2 = V_1 / j\omega C R_0 R_3$$

from which V_2 has been removed.

6



Cross multiplying this last expression yields

$$V_1 / I_{in} = j\omega C R_0 R_1 R_3 / R_2$$

Since $V_1 = V_{in}$,

$$V_{in} / I_{in} = Z_{in} = j\omega (C R_1 R_3 / R_2) R_0$$

which is the expression for a pure inductance in which

$$L = (C R_1 R_3 / R_2) R_0$$

$$\text{If } K = C R_1 R_3 / R_2,$$

$$L = K R_0$$

K is called the conversion factor of the PIC.

This completes the basic analysis of the PIC without the need of anything more than elementary AC theory and algebraic manipulation.

Practical applications

The numerical values of R_1 , R_2 , and R_3 affect the signal handling capabilities of practical opamps and are, therefore, a compromise. Typical values are:

$$R_1 = 270R \quad 1\%$$

$$R_2 = 5k6 \quad 1\%$$

$$R_3 = 10k \quad 1\%$$

$C = 10 \text{ nF } 1\%$ silver mica (two 5000 pF types in parallel).

These values enable the computation of K :

$$K = C R_1 R_3 / R_2 = 270 \times 10^4 \times 10 \times 10^{-9} / 5.6 \times 10^3 =$$

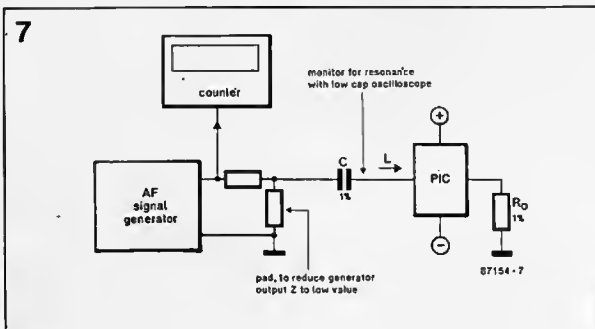
$$= 4.8214 \times 10^{-6}$$

In practice, the author used an MC1458 operating from a $\pm 12 \text{ V}$ power supply, which has a $K = 5.7155 \times 10^{-6}$. The departure from the ideal K value is due to the approximations used with the opamps: this is not a real problem in practice. The complete circuit of the PIC with the values stated is given in Fig. 6.

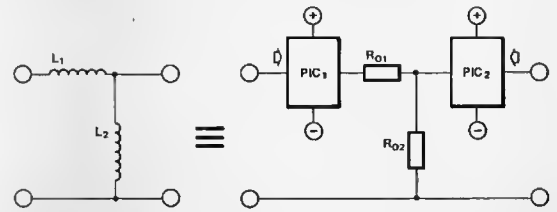
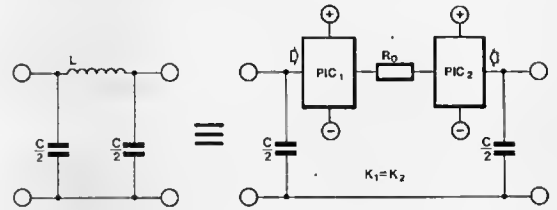
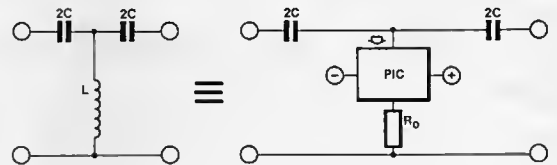
Checking the value of K in practice

Build the circuit of Fig. 6 and bring it to series resonance with the aid of a test set-up as

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8



L_1 is simulated by PIC1 + PIC2 (R_{01})
 L_2 is simulated by PIC2 (R_{02})

shown in Fig. 7. If the generator equivalent impedance is small and C is known accurately, the frequency, f , can be measured. Since

$$f = 1 / 2\pi \sqrt{L_{eff} C},$$

and

$$L_{eff} = K R_0$$

K can be found to 1-2% accuracy once R_0 is given.

The conversion factor can be trimmed by small adjustments to R_3 .

Inductances are simulated by PICs as shown in Fig. 8, where (a) is grounded coil simulation; (b) is balanced coil simulation; and (c) is grounded and balanced coil simulation. In (c), L_1 is simulated by PIC1 + PIC2 (R_{01}), and L_2 by PIC2 (R_{02}).

In conclusion, the author would emphasize that the PIC is a very useful circuit element and should not be left shrouded in mystery. It is hoped that this article will help it find much larger appreciation and application.

RESEARCH COLLABORATION TO BOOST INFORMATION TECHNOLOGY

by Kenneth Owen

The fast moving world of information technology has witnessed two significant developments recently in Britain. Much closer links have been forged between research and product development, and plans have been made to exploit the results more actively in the world's markets.

In 1981, Japan launched a ten year national programme to develop what are known as fifth generation computer systems. This ambitious project aims to bring together all the necessary elements of "clever" computers which, in the 1990s, will be able to demonstrate artificial intelligence and simulate human hearing, speech and vision.

Britain's response to the Japanese challenge was to mount its own five year programme of advanced IT research known as the Alvey Programme. It started in 1983, and will cost eventually about £350 million, of which £200 million will come from public funds and £150 million from participating companies.

Wide ranging study

The programme consists of pre-competitive research projects which fit into national strategies in key areas of technology. They involve collaboration between universities, industry and research establishments; and between different companies within industry.

The Alvey Programme originally aimed to cover four main technologies:

- * Very large scale integration (VLSI) in integrated circuits.
- * Software engineering, the development of standard methods for writing computer programs and systems.
- * The man/machine interface

(MMI), software and hardware aspects of making computers easier to use.

* Intelligent knowledge based systems (IKBS), exhibiting artificial intelligence.

Another topic, advanced computer architectures (overall design frameworks) has since been added.

In parallel with the Alvey Programme, a similar one was started by the European Community. This is known as the European Programme of

Research and Development in Information Technology

(ESPRIT). Twelve major European electronics and computer companies played a key role in defining ESPRIT, including Britain's GEC, ICL and Plessey.

Collaboration between companies and universities in different countries is a feature of the ESPRIT research and development. The hope is that it will eventually lead to Europe-wide market exploitation.

Innovative products

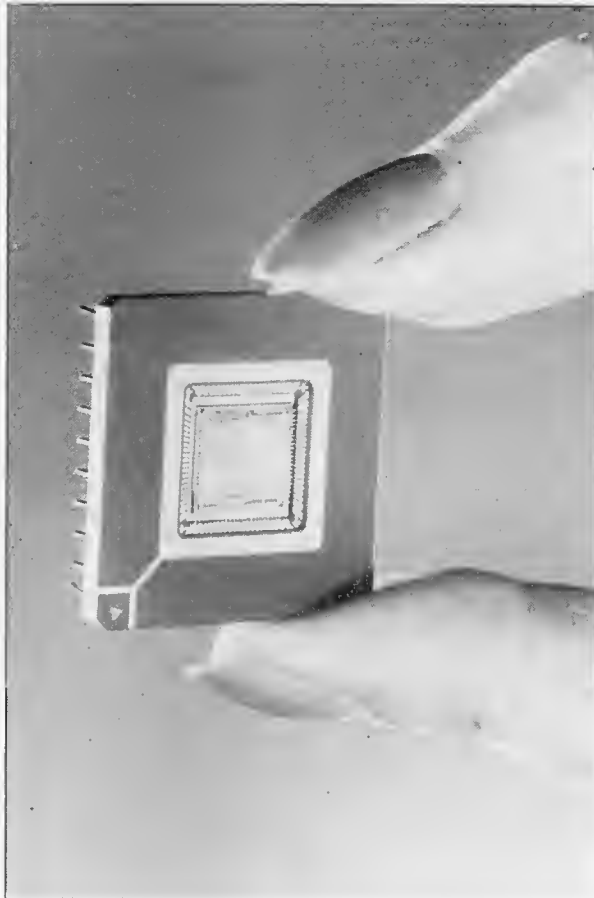
So British IT firms' own research and development activity has three elements: Alvey projects, ESPRIT projects, and individual in-house work. Together, these various elements have strengthened the industry's research base and the projects are leading to innovative products, systems and services in areas of key importance for the future. They reflect an important feature of the information technology business: the predicted convergence of different technologies, computers, communications and so on into integrated systems. This has happened much more rapidly than many people expected.

The new IT techniques, products and processes are being applied throughout the manufacturing and service industries to automate existing functions and introduce new services and facilities that were once impossible.

An ingenious British IT invention now being applied in powerful systems is the INMOS transputer, a computer on a chip. A single one can be used as a high performance conventional microprocessor. Even more significantly, groups of them can be used in multiple transputer systems to provide extremely powerful supercomputers.

All involved

Planning for the exploitation of products and systems in the world's markets forms part of Britain's national IT effort that will follow the Alvey Programme. The aim is to stimulate the development and use of advanced IT systems, and so will involve both producers and operators in addition to the academic research sector.(LPS)



Claimed as the world's first computer on a chip, the INMOS transputer, developed in Britain, combines the functions of processing, storage and communications. It can be used either as a conventional microprocessor or in multiple transputer systems to provide extremely powerful supercomputers.

THE DESK-TOP SUPERCOMPUTER

by Staniforth Webb

Modular concepts of computer structure using large numbers of transputers, each as powerful as a mainframe installation, are bringing the vast computing power so far used only for such applications as meteorology or high-technology design down to the domain of the desk-top device. They will cut computing costs by some 90 per cent.

Electronics engineers at Southampton University plan to unveil a supercomputer this year to match the most powerful machines now in service, but which could be produced and sold at about only one-tenth of their price. Designed by Dr Chris Jesshope and Dr Denis Nicole in the University's Department of Electronic Engineering, it has been developed as part of the ESPRIT programme, the European initiative at government level which aims to keep advanced technology in Western Europe well abreast of that in other countries.

The main reason for the low cost of the Southampton computer is that it is built on a modular principle. It is assembled from 350 so-called transputers, each of which is a complete computer in miniature with its own memory and an appropriate set of connections to link it to other transputers or to other computers, all on a single silicon chip of about 100-mm² area. This contrasts with the design of the Cray supercomputer, for example, which is composed of four linked mainframe computers.

Transputers sell at about \$500 apiece. A commercial computer built on the Southampton design could be sold for around \$800 000; a Cray costs somewhere in the region of \$6 million.

So far, such enormously powerful computers are used in only a few specialised applications such as weather forecasting, aircraft design and some areas of scientific research. But reducing the cost of supercomputing power by a factor of 10 would obviously open up far more applications and bring supercomputing power within the reach of many more potential users. In the words of Chris Jesshope, "It could put yesterday's supercomputer on today's posh office desk."

Mainframe Power

The transputer has been developed and is being marketed by the British silicon chip company INMOS. Each carries a processor that can be programmed to perform various tasks, a memory that contains a large part of the information the processor needs for its job, and all the connections for linking into other devices.

A single transputer is as powerful as an average full-sized mainframe computer. It is able to perform one-and-a-half million operations every second. Its circuitry is as complex as a complete street map of London with all the gas mains, electricity cables and sewers superimposed. It works faster than comparable processors, is easier to programme and is more compact. But from the point of view of assembly into supercomputers, its most important advantage is the ease with which it can be linked to other transputers or to other computers with no need for extra electronic circuitry.

Because of the Southampton project's significance for the future of the company, INMOS supplied the University with the first of a new generation of transputers, the IMS T 800. This

model can handle decimal points or fractions as well as integers; it is the first microprocessor to incorporate a floating-point processor capable of dealing rapidly with decimal digits on the same piece of silicon as a conventional processor handling integers.

In technical language, the IMS T 800 includes a 32-bit integer processor which is the world's fastest, with special instructions to support graphics operations, a 64-bit floating-point processor, four kilobytes of fast on-chip RAM and four standard INMOS communications links, all on a single chip.

As Peter Cavill, Director of the INMOS Microcomputer Products Division says, "Incorporating all this on the same chip significantly increases its performance, by eliminating all the delays associated with multi-chip solutions. The result is higher reliability, reduced board space and lower overall system costs, as well as being considerably faster."

Development Potential

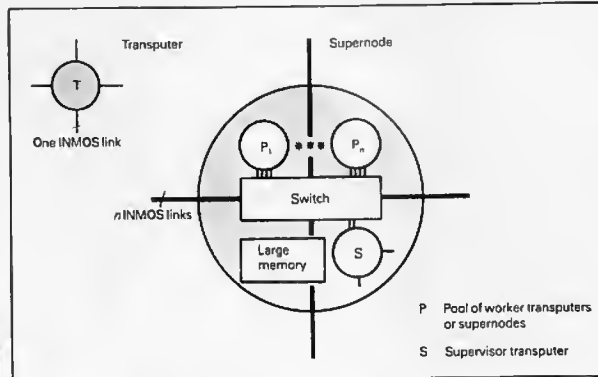
Jesshope and Hey see no problems in linking as many as 1000

transputers into a single computer. Beyond that, radical redesign will be needed because of the complexities of communication between so many modules. But there is clearly a vast amount of development potential in the present design.

The cost of the transputer is confidently expected to come down as European companies now begin to make it a mainstay of their bid for sales in new areas. The Southampton project is being backed by two French companies as well as by the UK Royal Signals and Radar Research Laboratory. Half of the development costs have been born by the ESPRIT programme. French and British companies (TELMET and Thorn-EMI) are expected to build computers based on the Southampton design. When these and the transputers they incorporate come into wide use and their cost advantages become apparent, it is predicted that large sales will reduce transputer prices still further.

Because of their ability to work co-operatively in parallel on a number of different but related tasks, transputers are well suited for use in so-called parallel processing. By designing computers which work on a number of tasks simultaneously, instead of doing everything in sequence, designers aim to mimic more closely the workings of the human brain.

Transputers are also being assigned to less futuristic applications, including desk top supercomputers, laser printers and what have been nicknamed turbochargers where the transputer is used as an add-on unit to an existing system to upgrade its performance. High-performance graphics, engineering workstations and robotics are other areas where the transputer is already beginning to make an impact.



The basic transputer (T) is incorporated into a pool to make up a 'supernode', which in turn is linked into an array to give a multistage switch with hierarchical control.

BACKGROUND TO HOLLOW EMITTER TECHNOLOGY

by Sue Cain and Ray Ambrose*

Hollow emitter technology falls into the area of power transistor technology lying between bipolar and power mos, providing an economic solution to high voltage switching. Ideally suited to switching frequencies which cause problems to both bipolar and power mos devices, hollow emitter technology improves the performance of multi-epitaxial mesa power transistors. This article describes the features of hollow emitter technology, and discusses the characteristics of some currently available devices.

Within the field of power transistor technology, the area between bipolar and power mos is occupied by high voltage switching. An economic solution is provided by hollow emitter technology.

Hollow emitter technology is ideally suited to switching frequencies which cause problems to bipolar and power mos devices. Products such as SGS's Fastswitch range provide rugged operation at very high switching speeds, and high levels of efficiency which allow more compact designs with smaller heatsinks.

When compared to industry standard high voltage devices, hollow emitter types provide faster switching times and a lower saturation voltage. The term 'hollow emitter' refers to the missing centre region in the emitter area, which reduces the

charge crowding effect at this point, as well as the storage time required to remove the excess charge.

A thinner intermediate N layer, present with hollow emitter devices, reduces the collector resistivity and the saturation voltage.

Standard versus hollow emitter technology

Hollow emitter technology has developed out of the standard high voltage multi-epitaxial mesa technology, the main difference being that, with the new technology, the emitter is not diffused over the normal area but only on the normal emitter edge.

This difference has been implemented to overcome the

charge-crowding effect in the centre of the emitter during the off transition found in industry-standard high voltage devices. The result is much faster switching times.

Standard high voltage

During the ON transition, the charge in any normal high voltage device is easily built up on the edge of the emitter. At the same time, the increase in base resistance towards the centre of the emitter reduces the charge level at the device's centre.

While undergoing the OFF transition, the base extraction current of the device can rapidly eliminate the charge at the edge of the emitter, but the charge at the centre is removed with some difficulty owing to

the transversely distributed resistance in the base under the emitter finger.

While the edge is turned off, an amount of charge remains at the centre of the emitter diffusion which decelerates turn-off time. These conditions are illustrated at Figures 2a and 2b.

Hollow emitter

In the case of transistors such as the SGS Fastswitch range, diffusion at the centre of the emitter is prevented by masking, as appropriate. The resulting hollow in the emitter prevents the accumulation of charges that would slow down the turn-off time.

Removal of the central region, with consequent reduction of the emitter area has a negligible effect on V_{CEsat} . There are two reasons for this: first, the centre region of the standard device carries less charge, and second, the intermediate N layer is thinner, reducing the resistivity of the collector, producing a fast switching, highly efficient device.

Comparison of switching times

Comparison of a standard multi-epitaxial mesa device with a hollow emitter device of the

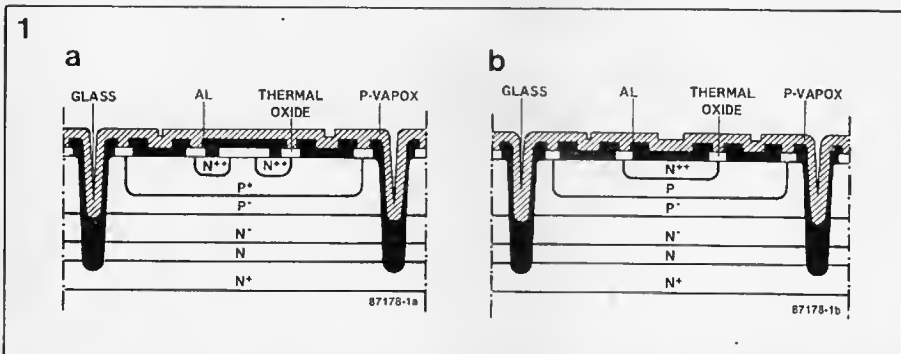


Fig. 1 Structure of the hollow emitter technology (a) and the standard multi-epitaxial mesa technology (b).

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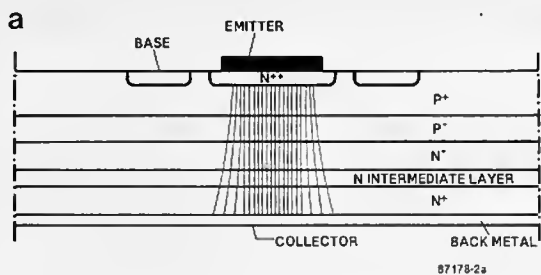


Fig. 2 Charge condition during conduction — normal high voltage technology.

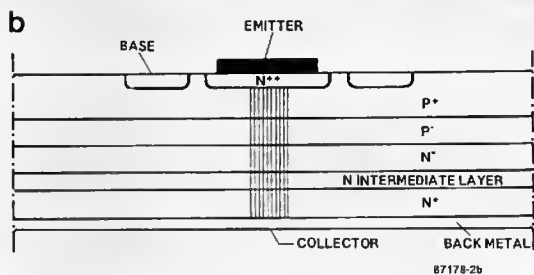


Fig. 2b. Charge remaining after turn-off.

same dimensions and similar static characteristics gives a clear indication of the advantages provided by the latter. The fall time of the hollow emitter device is clearly superior to that of the standard device, while storage time is also slightly better. Figure 3 compares the switching characteristic curves for two typical devices, the BUX48 standard multi-epitaxial mesa device, and the SGSD00032 hollow emitter, showing that with resistive loads the results produced by the hollow emitter are better than those for the standard device over a wide range of collector current.

Shorter fall and storage times automatically give rise to reduced dissipation energy while the device is turned off. Furthermore, both base and emitter switching times are almost always equally low in the case of hollow emitter devices. For example, the SGSD00035 t_{fall} is 50ns for both switching conditions.

Reverse bias safe operating area (RBSOA)

Figure 4 compares two comparable devices, from the standard and the hollow emitter technologies, comparing their reverse bias safe operating areas. Although the RBSOA characteristic curves are better for the standard device at 400V, the hollow emitter device has been optimised to give an improved RBSOA at higher voltages to suit switch mode power supply applications.

It can be readily demonstrated that the hollow emitter device

gives more protection at higher voltages, and this is balanced by adequate protection in areas of lower voltage and high current.

Applications

Hollow emitter technology has been devised in order to produce devices with extremely short switching times as regards their high voltage and current capabilities. As a result, hollow emitter devices are highly efficient, and are therefore well suited for applications demanding fast switching without high energy consumption.

Typical applications are those which require the use of inverters, switching regulators, fluorescent lighting and deflection circuits with high definition displays.

*Sue Cain is with BA Electronics and Ray Ambrose with SGS

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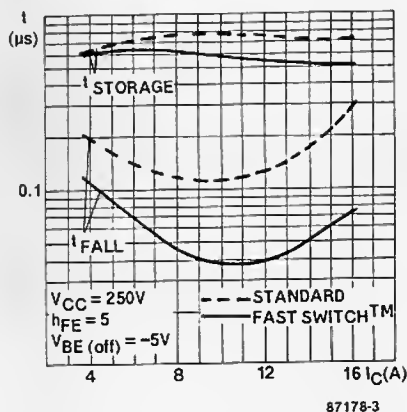


Fig. 3. Storage and fall times for standard and high voltage types.

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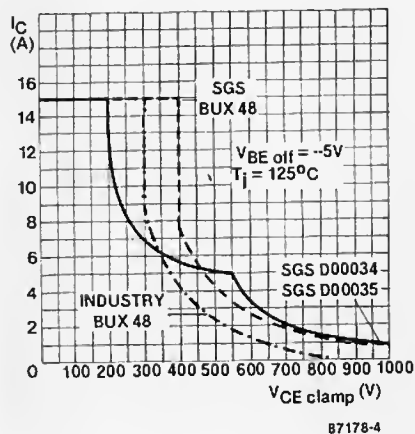


Fig. 4. Reverse bias safe operating areas for standard and hollow emitter types.

LONG LIFE BULB

How frequently do you have to change the light bulbs in your house? It depends entirely on how you use them. If you have a habit of switching the bulb ON and OFF frequently to "save" electricity, you are actually reducing the life of the bulb. You can even end up paying much more for the replacement of the bulbs than you are saving on your electricity bills.

A normal bulb in normal vibration free operation is expected to give a life of about 1000 hours. The useful life of the bulb can, however, be extended by an electronic trick upto about 100,000 hours! The relation between the applied voltage across the bulb and the life of the bulb is shown in figure 1. From this graph you can see that if you use the bulb at about 70% of its

rated voltage, it can "live" a hundred times longer.

The circuit which does this, is shown in figure 2. The reduction in voltage across the bulb due to this circuit is about 30%. So if you are using a 40 W bulb, you can now use a 60 W bulb at 70% of the rated voltage and get a very very long life out of it. How is this achieved? The filament inside the bulb is made of

tungsten. When the bulb glows, the filament becomes white hot, and continuously some tungsten keeps on evaporating. But if we use it at a lower voltage, the filament temperature is a little less and the loss of the tungsten due to evaporation reduces dramatically. As the life of tungsten filament increases, so does the bulb life.

The reduction in working voltage is achieved by the simple circuit shown in figure 2, and costs much less than a regular lamp dimmer circuit. It can be assembled on a small lug strip of two rows of 8 lugs on each side. The component layout is shown in figure 3.

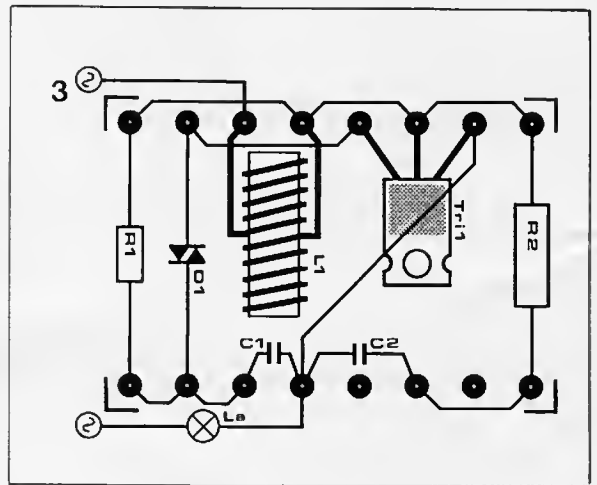
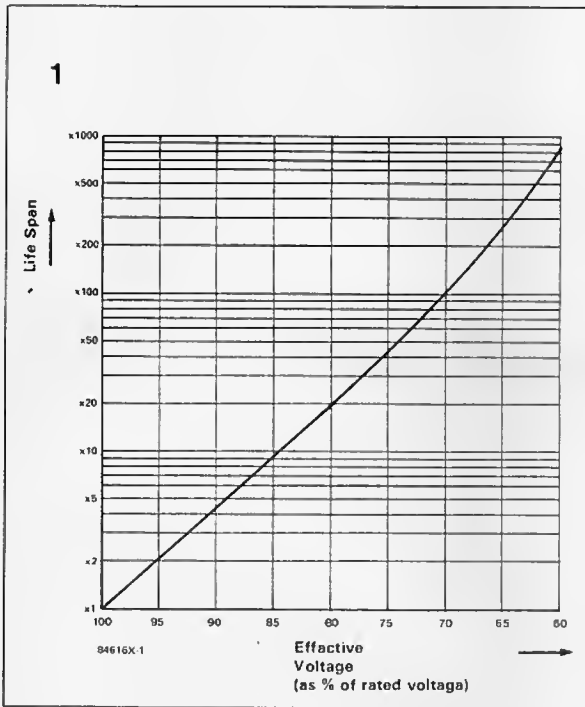
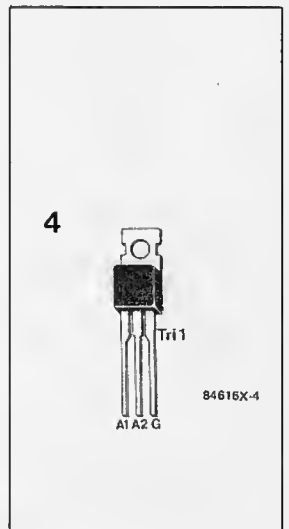
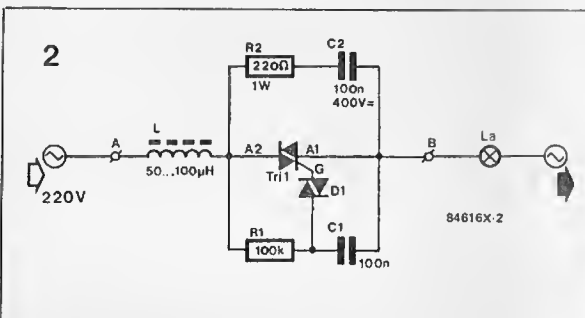


Figure 1: The graph showing the relation between the average life span of the bulb and the effective voltage across it.

Figure 2: The circuit for reducing the effective voltage to about 70%

Figure 3: Component layout of the circuit on a lug strip.

Figure 4: Pin assignment of a Triac



The Circuit

The main components of the bulb life extender circuit are a Diac and a Triac. The resistance R1 and capacitor C1 decide the percentage of voltage applied across the bulb. R2 and C2 (400V rating) protect the Triac against the rapid changes in voltages, which can happen due to the harmonics in the supply voltage waveform after switching through the

Triac. Coil L1 acts as a suppressor for radio frequency disturbance.

Construction:

Do not use a SELEX PCB for construction of this circuit because the tracks on the SELEX PCB are not capable of carrying heavy currents. A lug strip of two rows with 8 lugs each can be used for this purpose. Figure 3

shows the component layout of the circuit on a lug strip. The pin assignments of the Triac are shown in figure 4.

The circuit can be housed in a small plastic box, and installed inside the switch board if there is sufficient space. It can also be mounted near the lamp holder. Don't forget to switch off the mains supply completely when you are

installing this circuit. The circuit must be inserted in the power supply. LINE. Bulbs upto 100 W can be connected on this circuit.

- Notes: 1. Do not use this circuit with gas filled lamps.
2. Do not replace R1 with a potentiometer.

In all types of digital instruments and circuits, a seven segment indicator is a very essential part. Seven segment indicators are available in many different forms, like LED, LCD or Fluorescent.

The most commonly used seven segment LED indicator is discussed here. A small circuit has been built on the SELEX PCB. The control of the display segments is handled by just one IC, 74LS247, known as BCD to Seven Segment Decoder.

A four bit binary number can be presented at the four inputs of the 74LS247 IC, and at the output we directly get the decoded seven segment signals. These seven outputs decide which number will be displayed on the seven segment LED indicator. The segment designation and 16 possible combinations of the display are shown in table 1.

Figure 1 shows the actual connections between the IC and the display.

Behind every segment, there is an LED. The anodes (plus poles) of all seven LEDs are connected together and brought to the pin marked CA (Common Anode). This pin is connected to the positive voltage supply of +5V. The

SEVEN SEGMENT DISPLAY

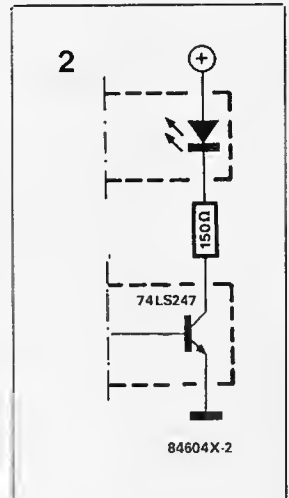
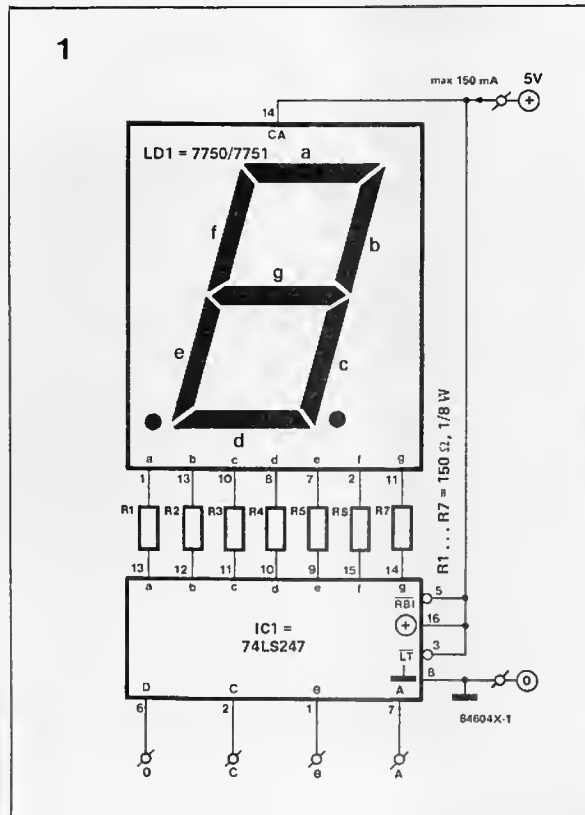
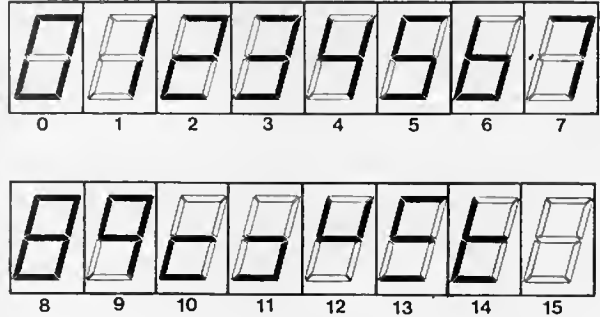
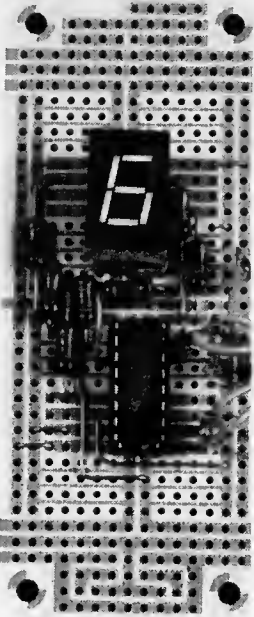


Figure 1: The circuit is very simple and compact, as all the decoding circuit is internal to the IC 74LS247. (subsequent version of 7447)

Figure 2: Transistors inside the IC switch the current through LEDs ON or OFF depending on the desired combination of seven segments to be lighted up. The resistor is used for limiting the current through the LED and the transistor.



seven cathodes are brought to the pins marked a,b,c,d,e,f,g and are connected to the seven outputs of the IC 74 LS 247 through 150Ω resistors.

As shown in figure 2, the output pins of the IC are nothing but collectors of the driving transistors inside the IC. The transistors are switched ON internally depending on the desired combination of the seven segments to draw current through the LED. The 150Ω resistor is placed in series to avoid excessive current through the LED and the transistor. (approx. 20 mA)

The decimal point is not controlled by the IC, and must be turned on or off directly. Pin 3 of the IC is used for testing all the segments of the display, by connecting it to ground. It should be connected to +5V in normal operation. Pin 5 is used for suppressing zeros, when used in multidigit indicators. Connecting pin 5 to ground suppresses the indication of zero on the display.

For construction of the circuit, the component layout shown in figure 3 must be followed. Use an IC socket for the 74 LS 247.

The Seven Segment LED display can be any common anode display with 1/2" digit height. An alphabet after the type number generally indicates the light emitting strength of the LEDs. The LEDs can be of different colours like Red, Green, Orange and Yellow depending on the type number.

With an operating voltage of +5V, the single digit circuit consumes about 25 to 150 mA current. It takes the minimum current when number 1 is displayed and maximum current when number 8 is displayed.

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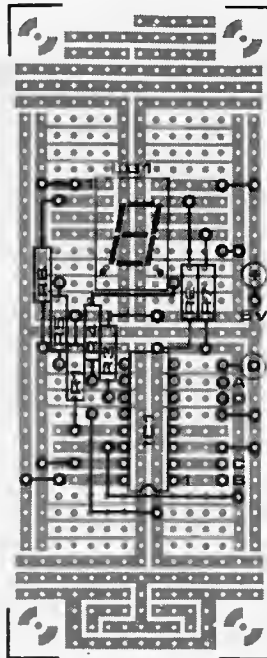


Figure 3:
The Component layout of the single digit display. Use IC socket for 74 LS 247.

- Component List:**
 R1-- R7 = 150Ω (1/8 W)
 IC1 = 74 LS 247
 Seven Segment Display =
 DL 7750, DL 7751
 HP 6082 - 7751,
 TIL 312, TIL 314, TIL 315
 TIL 339 or equivalent.
 One SELEX PCB (40 x 100 mm)
 One 16 Pin IC socket.

All the multimeters have at least four different measuring ranges. DC Voltage, DC Current, AC Voltage and Resistance. An audio enthusiast would always be happier with a frequency measuring range on a multimeter.

We have done just that for you in this article. A simple circuit which can convert your multimeter into a frequency meter has been described. As the multimeter dial is not calibrated for frequency, you have to be satisfied with the voltage scale for reading the frequency. Because what the circuit does is, in fact, to convert the frequency into a proportional voltage. It can be used with any make of multimeter, and does not cost much to construct.

The Circuit

The main task of the circuit of our "frequency meter" is to convert the frequency

FREQUENCY MEASUREMENT WITH A MULTIMETER

into a proportional DC voltage, so that the multimeter can read the DC voltage directly and thus indicate the frequency.

This requires some signal processing to be done. As we may not have any control over the amplitude of the signal at the frequency input, we must take care of the limiting of

the input signal amplitude in our circuit.

Circuit shown in figure 2 takes care of the amplitude limiting as well as signal processing. The input signal reaches IC1 through C1, R1 and C2 This IC is an operational amplifier (op amp) which then converts the input waveform into a rectangular wave with the

same frequency. The amplitude of the signal reaching IC1 is limited by the two diodes D1 and D2. The conversion of the signal to a suitable DC Voltage signal for the multimeter is accomplished by IC2 which is a timer IC 555 configured as a monoshot. The monoshot is triggered by the output of the

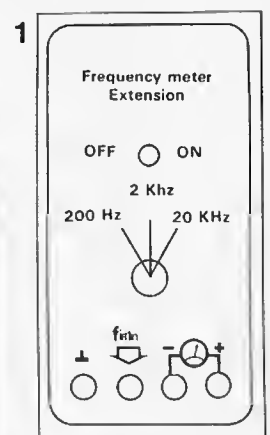
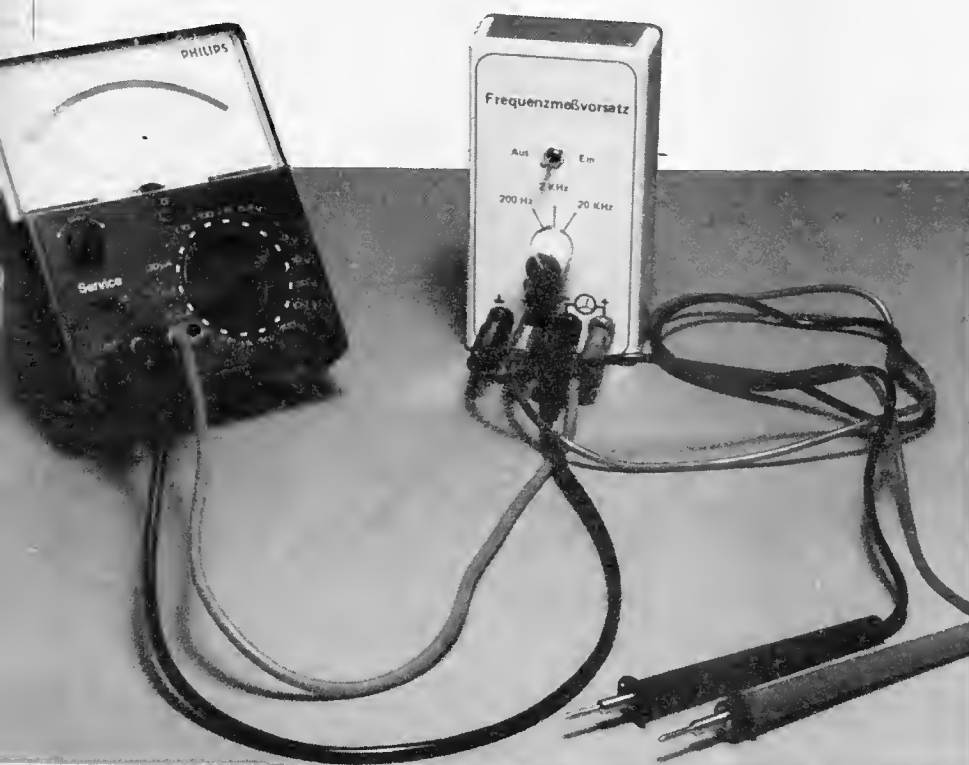
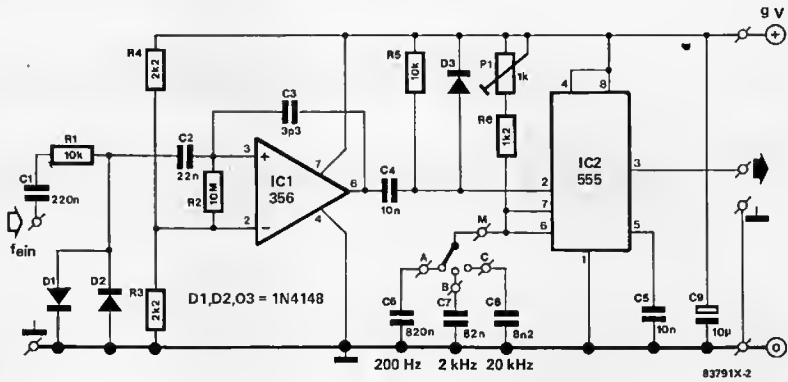


Figure 1: Suggested design of the front panel for the frequency meter extension. The terminals must be clearly marked with + and - for the multimeter.

2



operational amplifier. A pulse of fixed duration is produced at the output of the monoshot for every cycle of the input signal. These DC pulses can be given to the multimeter. The movement of the multimeter cannot respond quickly to each pulse and effectively the DC voltage indicated by the multimeter depends upon the frequency at which these fixed duration, fixed amplitude pulses arrive at the input of the multimeter. Thus the reading on the multimeter is directly proportional to the frequency of the input signal.

A more detailed description of the functioning of the circuit is given below:

The Comparator:

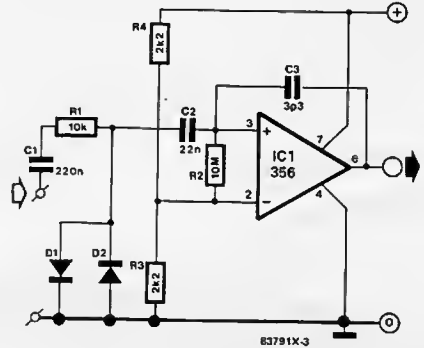
Figure 3 shows the first part of our circuit which consists of the Op Amp IC 356. It is used here as a comparator. A comparator compares two signals on the two inputs, (pins 2 and 3 of IC1) and produces an output depending on the difference between the two input voltages. In this case, the inverting input pin 2 is connected to the potential divider R4/R3 and thus lies at half the supply voltage.

Figure 2: Only two ICs and a few components are required for converting the input audio frequency into a dc pulse train of same frequency.

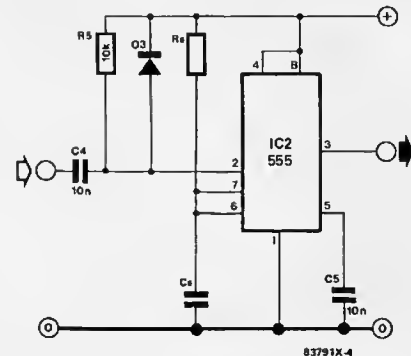
Figure 3: First part of the signal processing circuit is the comparator. It converts the AC signal into a DC rectangular waveform with same frequency and fixed amplitude.

Figure 4: The second part of the signal processing circuit is a monostable multivibrator which converts the rectangular wave into a pulse train of fixed amplitude and fixed duration pulses at the same frequency as that of the input signal.

3



4



The input resistance of the Op Amp is very high compared to R2 and the effective resistance between pins 2 and 3 can be assumed to be equal to R2.

The component values are such that the Op Amp output voltage switches between the supply voltage and zero volts depending on whether there is a positive difference or negative difference between the two input voltages.

It is not necessary to consider the Offset voltage of the Op Amp in this case, as we are interested only in a rectangular wave with same frequency as that of the input signal, and not the actual amplitude of the difference voltage.

Capacitor C3 is connected between the output and the input, so that the switching is steep. The input signal must have an amplitude of at least 50 mV so that the Offset voltage of the Op Amp plays a negligible part in the switching at the output. The voltage rating of

the capacitor must be sufficient to match the maximum amplitude of the input signal. If a capacitor with 100V rating is used for C1, it should be able to take care of all types of audio signals. Voltages above 60V will almost never be encountered in audio frequency measurements.

The actual voltage applied at C2 is limited to about 0.7 V by the two diodes D1 and D2. The presence of C2 at the input allows only AC voltages to pass through to the Op Amp.

The Monoshot:

The rectangular wave generated by the comparator has the same frequency as that of the input signal, but still it is not suitable for the multimeter. The monostable multivibrator (Monoshot) shown in figure 4 is used for converting this rectangular wave to a series of fixed amplitude, fixed duration pulses.

For improving the triggering action, first the rectangular wave is converted into a series of sharp spikes by C4 and R5. These spikes trigger the monoshot at every positive to negative crossing on the input signal, thus maintaining the correlation between the input frequency and the output frequency.

The duration of the pulses at the output of the monoshot is decided by R4 and C4. It should be of such a value that the pulse is over before the next triggering spike comes. If the pulse overlaps two cycles of input frequency, the relation between input and output frequencies will be lost.

The pulse duration must be such that at the highest frequency in the measuring range, the pulse ends before the next triggering spike comes from the input.

In the circuit of figure 2, this is achieved by using three different values of C4 for three different ranges of

input frequencies (200Hz, 2KHz, 20KHz), R4 is also made variable by using a potentiometer in series (R6 + P1).

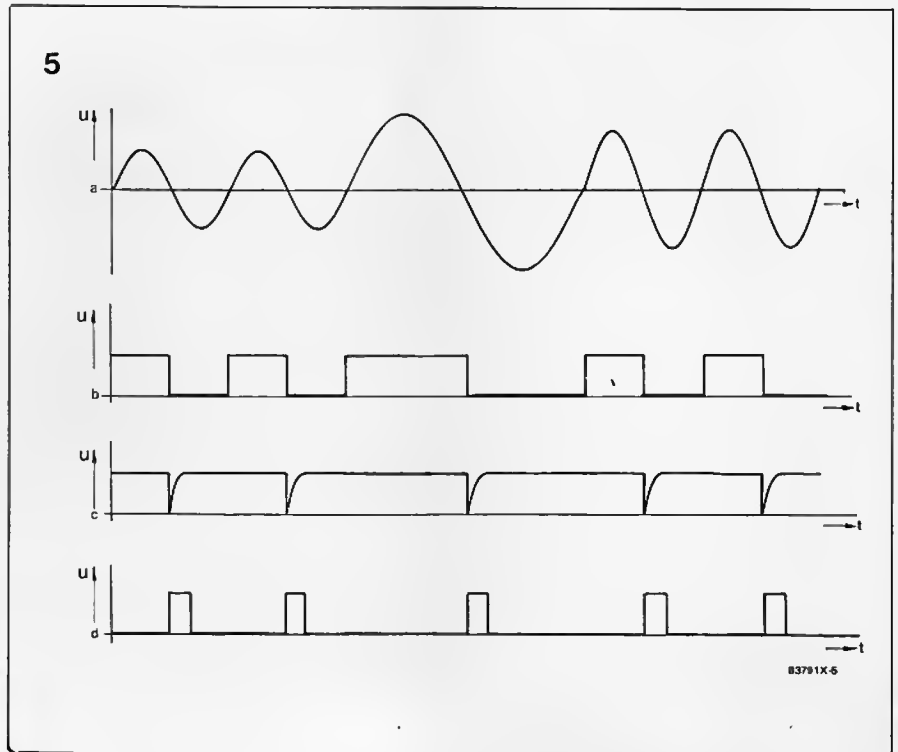
Pin 6 and 7 of IC2 (555) are shorted together and connected to C6, C7 or C8 through a selector switch. C6 corresponds to a range of 200 Hz, C7 takes care of frequencies upto 2KHz and C8 is for the 20KHz range.

Figure 5 shows the various waveforms from input to output. Waveform 5a is the input waveform. The frequency of the waveform 5a is shown with a varying frequency to illustrate the one to one correspondence between the input and output frequencies. Figure 5b-waveform shows the output of the comparator which is a rectangular wave which remains at zero level during negative half cycles of the input and switches to positive supply voltage during the positive half cycles of the input waveform.

Figure 5:
Various waveforms encountered during the processing of the input signal.

- a) Input frequency waveform
- b) Comparator output — rectangular waveform
- c) Spikes for triggering the monoshot.
- d) Output pulse train from monoshot.

Note that the frequency is same from input to the output.



Waveform at 5c is the differentiated waveform generated by R5 and C4. These spikes then trigger the monoshot made of 555 IC and the Ra, Ca combination. 5d is the waveform at the output of the monoshot. These are fixed duration fixed voltage pulses—one for each cycle of the input signal.

The duration of the pulses is so small that the needle of the multimeter cannot physically jump forward and backward on each pulse. What happens is the averaging effect of these pulses on the multimeter needle. The reading shown by the multimeter is thus proportional to number of pulses arriving at the input of the multimeter per second, which is nothing but the input frequency.

Construction :

Fig 6 shows the component layout of the circuit on a small SELEX PCB of size 40 x 100 mm. As usual, the soldering must start with jumper wires, followed by resistors, capacitors and then the ICs. Both the ICs are 8 pin ICs in plastic

package, and preferably an IC socket should be used for each of them. Marking for pin 1 must be correctly observed. In both the ICs the notch near pin number 2 is facing towards the trimpot P1. Correct polarity of the diodes must also be ensured for proper functioning of the circuit.

The wires for input signal, supply voltage, rotary switch and for the output to the multimeter are directly soldered to the PCB. Soldering terminals can also be provided for these wires on the PCB for ease of soldering and future maintenance and trouble shooting.

For the supply voltage, a 9V miniature battery is enough. It can be connected through a clip onconnector. An ON/OFF switch must be inserted in the battery supply line and mounted on the front face of the enclosure for easy access.

Before turning on the supply to the circuit, all connections and polarities of components must be thoroughly checked to avoid

- R1, R5 = 10 K Ω
- R2 = 10 M Ω
- R3, R4 = 2.2 K Ω
- R6 = 1.2 K Ω
- P1 = 1 K Ω (Trimpot.)
- C1 = 220 μ F
- C2 = 22 μ F
- C3 = 3.3 μ F
- C4, C5 = 10 μ F
- C6 = 820 μ F
- C7 = 82 μ F
- C8 = 8.2 μ F
- C9 = 10 μ F/16V
- D1, D2, D3 = IN 4148
- IC1 = LF 356 (Op amp with FET input)
- IC2 = NE 555 (Timer)

Other parts:

- 3 position rotary switch with knob.
- 9V miniature battery
- 1 Clip on connector for battery
- 1 ON/OFF toggle switch
- 1 SELEX PCB (100 x 40 mm)
- 10 Soldering terminals
- 1 Suitable enclosure
- 4 banana pins & sockets (2 Red, 2 Black)
- flexible hook up wire etc.

Part list for figure 8:

- 1 Step down transformer 230V to any voltage between 3V to 12V.
- 4. Diodas IN 4001

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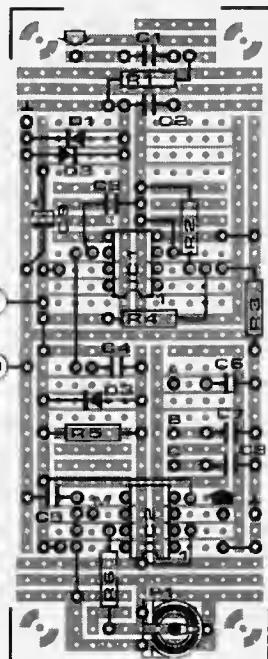
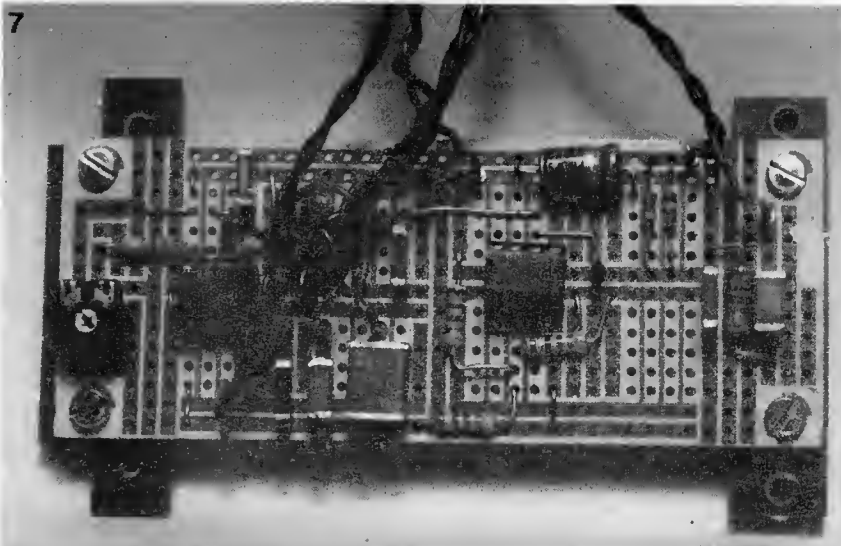


Figure 6: Component layout for the frequency meter extension circuit. The orientation of the ICs and the polarity of capacitor C9 as well as the diodes is very important.

Figure 7: Photograph of the assembled circuit.

7



any damage to the circuit. Care should be taken to look for bad solder joints as well as unwanted shorting of tracks due to overflow of solder material. After switching on the power to the circuit the three voltages must be first checked:

1. Supply voltage (9V)
2. Voltage at the junction of $R_2/R_3/R_4$ which must have half the supply voltage value (4.5V)

Voltage at the non inverting input of the IC1, which should be about 2.4 Volts. Deviations within $\pm 10\%$ can be accepted.

After all the above checks are found to be satisfactory, we can start the calibration procedure.

However, before attempting to take up the calibration work, a few more things must be done. First and most important, we must

have a known frequency source. An audio signal generator would be ideal, but if one cannot afford that luxury, there are other means to get over the calibration problems.

Figure 8 shows a very simple source of a known frequency signal. Just a small 230V to 3V step down transformer (even 6V, 9V or 12V transformer can be used.) and four diodes are

required. If the voltage is large, it is to be reduced by using a voltage divider, R_1/R_2 . The rectified signal gives a reference source of 100 Hz frequency.

This 100 Hz input signal can be given at the input of our circuit under calibration. As we know the frequency is 100Hz, we must set the rotary switch on 200Hz range. The multimeter is set on 3V DC range and connected across the output of our circuit.

The trimpot P_1 should now be adjusted in such a manner that the multimeter indicates 1.5 V. This corresponds to a frequency of 100Hz, naturally, because we have a 3V full scale reading which must indicate 200 Hz. This setting of the trimpot will be valid even for other two ranges; because we are changing the values of the capacitance C_a in the same proportion for each range.

It is not a must that the multimeter should have a 3V range. a 2V or 2.5V range can also be used. P_1 must be suitably adjusted for a half scale reading at 100 Hz.

Any suitable plastic box can be used as the casing. Suggested construction is shown in the photograph. A front panel layout is also given in figure 1.

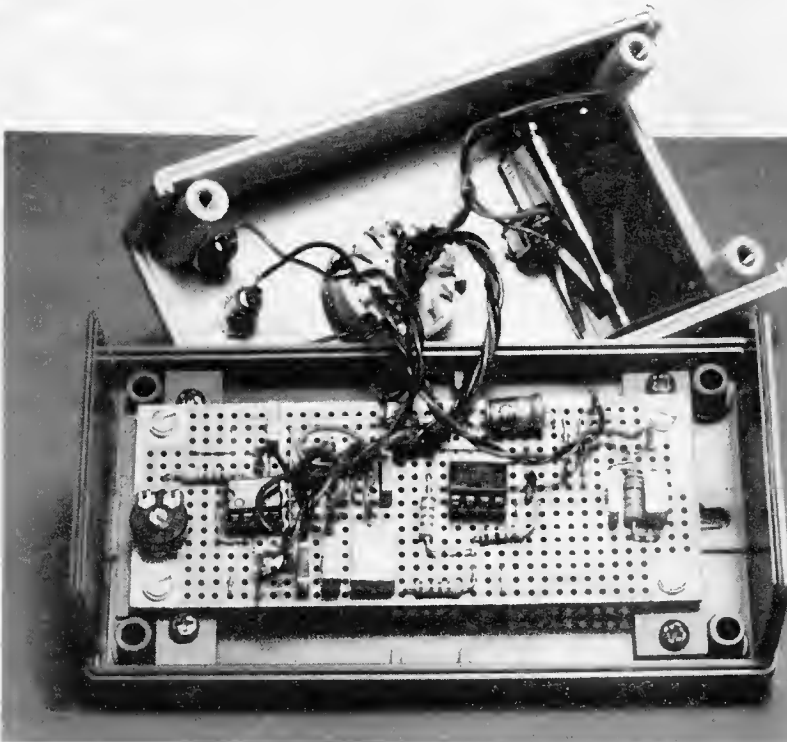
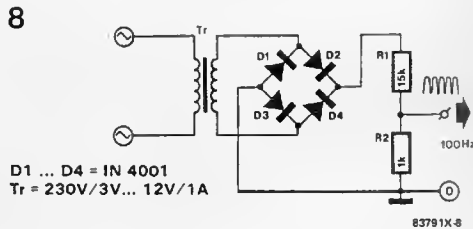


Figure 8:
Simple circuit for a reference frequency source of 100 Hz. This can be used for calibration of the circuit if an audio generator is not available.

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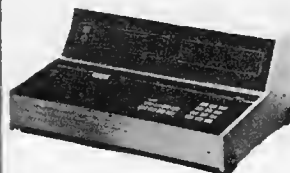


For further details please contact:

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Bombay 400 031
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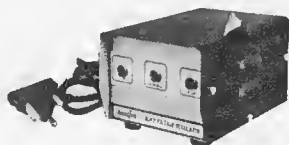
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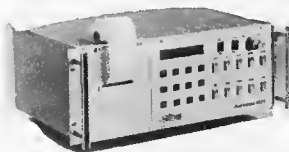
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Bombay 400 058
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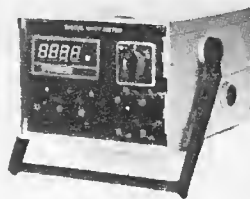


For further information please contact:

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For further information please contact:

JIVAN ELECTRO
INSTRUMENTS
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Makarapura,
Baroda 390 010.

CORRECTIONS

Preset extension for function generator

May 1987 p.5 28

The circuit diagram should be amended as follows:

The outputs of IC₃ are QA = pin 12; QB = pin 9; QC = pin 8; QD = pin 11. The connections of the BCD coder to the inputs of IC₄ are as follows: A to pin 13; B to pin 10; C to pin 1, D to pin 4.

Pins 3 and 4 of IC₂ should be connected to ground.

16 Kbyte CMOS RAM for C64

October 1987 p. 10-53

Pull-up resistor R₁ should be connected to junction D₁-D₂-R₃, not to +5 V. It is recommended to fit 22K resistors between the OE terminals of the RAM chips and the above junction.

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