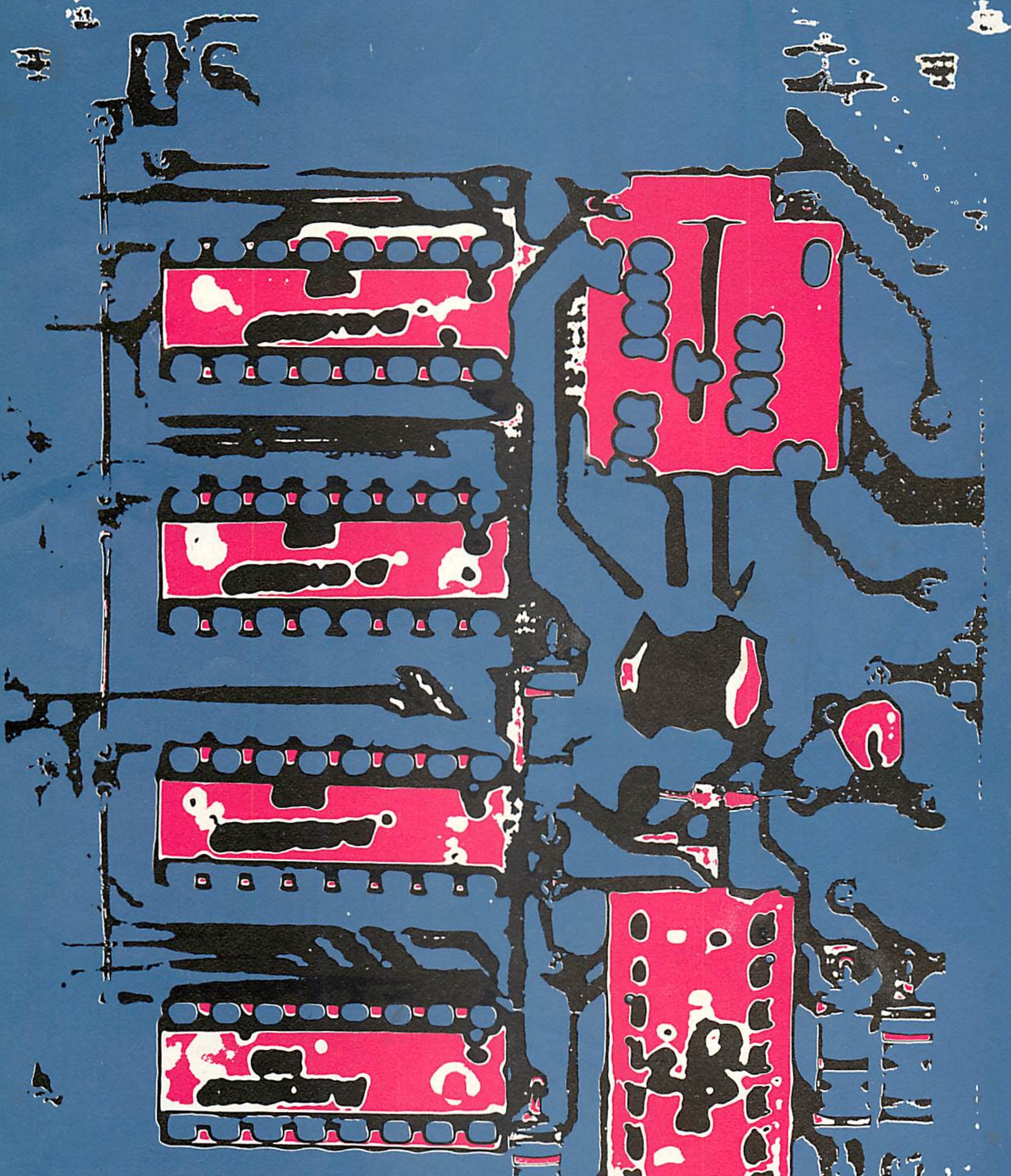
up-to-date electronics for lab and leisure

elektor 14



June 1976 40p

eps print service

Many Elektor circuits are accompanied by designs for printed circuits. For those who do not feel inclined to etch their own printed circuit boards, a number of these designs are also available as ready-etched and predrilled boards. These boards can be ordered from our Canterbury office. Payment, including £ 0.15 p & p, must be in advance. Delivery time is approximately three weeks.
Bank account number: A/C No. 11014587, sorting code 40-16-11 Midland Bank Ltd, Canterbury.

circuit	number	issue	price	% VAT
AUDIO				
austereo 3-watt amplifier	HB11	5	1.15	(12.5)
austereo power supply	HB12	5	0.55	(12.5)
austereo control amplifier	HB13	5	1.55	(12.5)
austereo disc preamp	HB14	5	0.65	(12.5)
edwin amplifier	97-536	6	1.25	(12.5)
miniature amplifier	1486	6	0.55	(12.5)
equa amplifier	1499	1	1.55	(12.5)
equin	9401*	13	2.15	(12.5)
tap preamp	4003	4	1.95	(12.5)
tap power	9072*	7	2.–	(12.5)
tap preamp front panels:				
power	1626A	7	1.65	(12.5)
input	1626B	4	1.65	(12.5)
volume	1626C	4	1.65	(12.5)
tone	1626D	4	1.65	(12.5)
width	1626E	4	1.65	(12.5)
730/740 (IC control amplifier)	9191*	8	1.15	(12.5)
disc preamp 76131	4040A	3	1.05	(12.5)
preco, preamplifier	9398*	13	2.–	(12.5)
preco, control amplifier	9399*	13	1.15	(12.5)
dnl	1234	11	0.85	(12.5)
electronic loudspeaker	1527	2	0.55	(12.5)
compressor	6019A	3	1.50	(12.5)
RF				
coilless receiver for MW and LW	3166	5	0.85	(12.5)
super-plam, main p.c.b.	6012-1b	11	2.05	(12.5)
super-plam, detector a	6012-2a	11	1.10	(12.5)
super-plam, detector b	6012-3a	11	1.25	(12.5)
ssb receiver	6031*	11	2.25	(12.5)
mini MW receiver	9369*	9	0.70	(12.5)
FM/TV				
pll stereo decoder MC1310P	1477	9	0.75	(12.5)
CA3090AQ stereo decoder	9126*	5	0.80	(12.5)
tv sound	6025	2	1.75	(12.5)
tv sound, front-end	9357*	11	1.30	(12.5)
ota pll	6029	7	1.15	(12.5)
feedback pll fm receiver (3 boards)	9356*	9	4.05	(12.5)
aerial amplifier	1668	1	1.15	(12.5)
integrated indoor fm aerial	9423*	13	0.80	(12.5)
CARS				
car power supply	1563	4	1.50	(8)
digital rev counter (control p.c.b. only!)	1590	1	1.35	(8)
car anti-theft alarm	1592	4	1.70	(8)
GAMES				
beetle	1492	4	2.35	(8)
tv tennis, main pcb	9029-1A*	7	4.40	(8)
tv tennis, modulator/oscillator	9029-2*	7	1.05	(8)
tv tennis, 5-volt supply	9218A*	7	0.90	(8)
tv tennis extensions	9363*	13	5.15	(8)
die	9169*	8	0.80	(8)
RHYTHM AND SOUND				
minidrum gyrator	1465A	2	0.95	(12.5)
minidrum mixer/preamp	1465B	2	0.60	(12.5)
minidrum noise	1465C	2	1.25	(12.5)
minidrum tap	1621A	2	0.90	(12.5)
minidrum ruffle circuit	1621B	3	1.15	(12.5)
automatic bassdrum	1621C	3	0.90	(12.5)
microdrum	1661	2	1.05	(12.5)
rhythm generator (M252AA)	9110*	5/12	0.90	(12.5)
rhythm generator (M253AA)	9344-3*	12	1.20	(12.5)
ic drummer, instruments mother board	9344-2*	12	1.70	(12.5)
ic drummer, instruments daughter board	9344-1*	12	0.30	(12.5)
screening for master oscillator	4011A	10	0.30	(12.5)
master oscillator AY10212	4011B-2	10	1.30	(12.5)
ancillary supply for m.o.	4011B-13	10	0.95	(8)
big ben 95	5028	2	1.50	(12.5)
7400 siren	9119*	5	0.80	(12.5)

circuit	number	issue	price	% VAT
TIME-KEEPING				
mos clock 5314 clock circuit	1607A	1	1.65	(8)
mos clock 5314 display board	1607B	1	1.20	(8)
mos clock timebase	1620	4	0.85	(8)
versatile digital clock	4414B	6	1.40	(8)
clamant clock, alarm	4015-13	7	1.45	(8)
clamant clock, time signal	4015-16	7	1.–	(8)
clamant clock, striking system	4015-27	8	1.35	(8)
car clock (2 boards)	7036	6	2.20	(8)
car clock front panel (transparent red plastic)	7036-3	6	1.15	(8)
kitchen timer	9147*	5	0.85	(8)
digital watch (date only)	9397-1*	10	0.80	(8)
digital watch (day/date)	9397-2*	10	0.80	(8)
DISPLAYS				
twin minitron display	4029-1	2	1.95	(8)
twin led display	4029-2	2	1.95	(8)
twin decade counter	4029-3	2	1.95	(8)
maxi display	4409	2	2.–	(8)
UAA170, 270° meter, basic board	9392-1*	12	1.40	(8)
UAA170, 270° meter, front panel	9392-2	12	1.90	(8)
TEST EQUIPMENT				
universal frequency reference	HD4	5	1.40	(8)
distortion meter	1437	1	2.20	(8)
a/d converter	1443	3	1.05	(8)
recip-riaa	4039	2	0.90	(8)
dil-led probe	5027A+B	2	2.45	(8)
frequency counter, control logic	9033*	7	1.50	(8)
frequency counter, minitron display board	9312*	7	1.30	(8)
frequency counter, led display board	9313*	7	1.30	(8)
frequency counter, counter/latch/display driver	9314*	7	1.05	(8)
frequency counter preamp	9031-1*	8	1.30	(8)
frequency counter, −5 V supply	9031-2*	8	0.85	(8)
tup/tun tester	9076*	4	2.05	(8)
tup/tun tester front panel	9076/2A	4	2.30	(8)
p.c.b. and wiring tester	9106*	5	0.60	(8)
fm test generator	9155*	10	0.85	(8)
capacitance meter	9183*	5	0.85	(8)
versatile logic probe	9329*	13	1.05	(8)
UAA170, LED voltmeter, basic board	9392-3*	12	1.–	(8)
UAA170, 16 LED display board	9392-4*	12	0.85	(8)
MISCELLANEOUS				
tap sensor	1457	1	0.85	(8)
mostap	1540	2	1.30	(8)
light dimmer	1487	6	0.55	(8)
ttl +5 V supply	4046	7	1.05	(8)
integrated voltage regulator	7043b	11	0.80	(12.5)
automatic call-sign generator	9017*	10	2.85	(8)
stylus balance	9343	12	0.40	(12.5)
polaroid timer	9379*	12	1.20	(12.5)
NEW				
digibell	9325*	14	1.90	(8)
led light show	9403*	14	2.55	(8)
fet front, preamp	9413*	14	0.80	(8)
fet front, probe	9427*	14	0.70	(8)

* with solder mask

All prices include VAT at the rate shown in brackets.

elektor decoder

What is a TUN?
What is 10 n?
What is the EPS service?
What is the TQ service?
What is a missing link?

Semiconductor types

Very often, a large number of equivalent semiconductors exist with different type numbers. For this reason, 'abbreviated' type numbers are used in Elektor wherever possible:

- '741' stands for μA741, LM741, MC741, MIC741, RM741, SN72741, etc.
- 'TUP' or 'TUN' (Transistor, Universal, PNP or NPN respectively) stands for any low frequency silicon transistor that meets the specifications listed in Table 1. Some examples are listed below.
- 'DUS' or 'DUG' (Diode, Universal, Silicon or Germanium respectively) stands for any diode that meets the specifications listed in Table 2.
- 'BC107B', 'BC237B', 'BC547B' all refer to the same 'family' of almost identical better-quality silicon transistors. In general, any other member of the same family can be used instead. (See below.)

For further information, see 'TUP, TUN, DUG, DUS', Elektor 12, p. 458.

Table 1. Minimum specifications for TUP (PNP) and TUN (NPN).

$V_{CEO,max}$	20V
$I_{C,max}$	100 mA
$h_{fe,min}$	100
$P_{tot,max}$	100 mW
$f_{T,min}$	100 MHz

Some **'TUN'**s are: BC107, BC108 and BC109 families; 2N3856A, 2N3859, 2N3860, 2N3904, 2N3947, 2N4124.
Some **'TUP'**s are: BC177 and BC178 families; BC179 family with the possible exception of BC159 and BC179; 2N2412, 2N3251, 2N3906, 2N4126, 2N4291.

Table 2. Minimum specifications for DUS (silicon) and DUG (germanium).

	DUS	DUG
$V_{R,max}$	25V	20V
$I_{F,max}$	100mA	35mA
$I_{R,max}$	1μA	100μA
$P_{tot,max}$	250mW	250mW
$C_{D,max}$	5pF	10pF

Some **'DUS'**s are: BA127, BA217, BA218, BA221, BA222, BA317, BA318, BAX13, BAY61, 1N914, 1N4148.
Some **'DUG'**s are: OA85, OA91, OA95, AA116.

BC107 (-8, -9) families:
BC107 (-8, -9), BC147 (-8, -9), BC207 (-8, -9), BC237 (-8, -9), BC317 (-8, -9), BC347 (-8, -9), BC547 (-8, -9), BC171 (-2, -3), BC182 (-3, -4), BC382 (-3, -4), BC437 (-8, -9), BC414

BC177 (-8, -9) families:
BC177 (-8, -9), BC157 (-8, -9), BC204 (-5, -6), BC307 (-8, -9), BC320 (-1, -2), BC350 (-1, -2), BC557 (-8, -9), BC251 (-2, -3), BC212 (-3, -4), BC512 (-3, -4), BC261 (-2, -3), BC416.

Resistor and capacitor values

When giving component values, decimal points and large numbers of zeros are avoided wherever possible. The decimal point is usually replaced by one of the following international abbreviations:

p	(pico-)	= 10^{-12}
n	(nano-)	= 10^{-9}
μ	(micro-)	= 10^{-6}
m	(milli-)	= 10^{-3}
k	(kilo-)	= 10^{3}
M	(mega-)	= 10^{6}
G	(giga-)	= 10^{9}

A few examples:
Resistance value 2k7: this is 2.7 kΩ, or 2700 Ω.
Resistance value 470: this is 470 Ω.
Capacitance value 4p7: this is 4.7 pF, or 0.000 000 000 004 7 F . . .
Capacitance value 10 n: this is the international way of writing 10,000 pF or .01 μF, since 1 n is 10^{-9} farads or 1000 pF.

Mains voltages

No mains (power line) voltages are listed in Elektor circuits. It is assumed that our readers know what voltage is standard in their part of the world!
Readers in countries that use 60 Hz should note that Elektor circuits are designed for 50 Hz operation. This will not normally be a problem; however, in cases where the mains frequency is used for synchronisation some modification may be required.

Technical services to readers

- **EPS service.** Many Elektor articles include a lay-out for a printed circuit board. Some – but not all – of these boards are available ready-etched and predrilled. The 'EPS print service list' in the current issue always gives a complete list of available boards.
- **Technical queries.** Members of the technical staff are available to answer technical queries (relating to articles published in Elektor) by telephone on Mondays from 14.00 to 16.30. Letters with technical queries should be addressed to: Dept. TQ. Please enclose a stamped, self addressed envelope; readers outside U.K. please enclose an IRC instead of stamps.
- **Missing link.** Any important modifications to, additions to, improvements on or corrections in Elektor circuits are generally listed under the heading 'Missing Link' at the earliest opportunity.

elektor

14 Volume 2
Number 6

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All prices include p & p.

Subscribers are requested to notify a change of address four weeks in advance and to return envelope bearing previous address.

Letters should be addressed to the department concerned:
TQ = Technical Queries; ADV = Advertisements;
SUB = Subscriptions; ADM = Administration;
ED = Editorial (articles submitted for publication etc.);
EPS = Elektor printed circuit board service.
For technical queries, please enclose a stamped, addressed envelope.

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contents

I wouldn't give a dime for an elektor circuit

You don't have to

on average they cost about half a dime (2 p) each!

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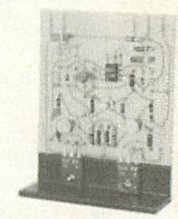

150	50
POWER OUTPUT: 170 W R.M.S.	**POWER OUTPUT:** 65 W R.M.S.
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BSY95A Transistors	6 for **50p**
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BCY72 Transistors	4 for **50p**
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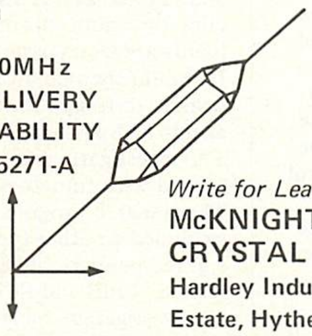

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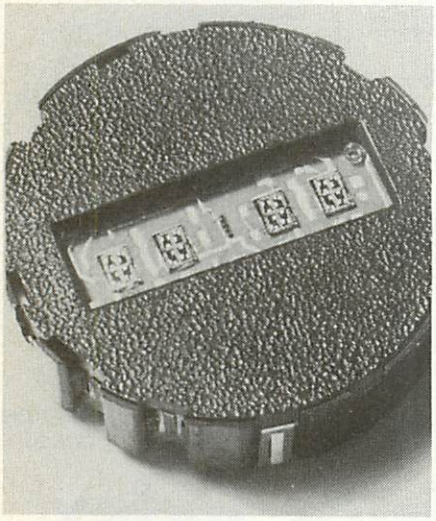
SELEKTOR

Build a Digital Watch – the easy way

For those who want to make a digital watch but are deterred by the degree of miniaturisation involved, a range of preassembled watch modules is available from Litronix. Prices have fallen so rapidly in recent months that it is now cheaper to buy one of these modules than to purchase the individual components for a watch, and it is only necessary to fit the module into a case and install batteries to obtain a fully-working watch.
The timekeeping function is integrated into a single ion-implanted CMOS chip and the display drivers are contained in two silicon bipolar IC's. These, together with the LED display, 32.768 kHz



crystal, capacitors and resistors, are mounted on a ceramic substrate using hybrid assembly techniques. This in turn is mounted in a high impact plastic frame to provide mechanical protection for the module. Gold-plated switch contacts are provided around the periphery of the module to activate the display and timesetting functions.
The module is powered by two RAY-0-VAC RW-44 silver oxide batteries (or equivalent) and battery life is said to be one year with up to twenty interrogations per day.
The cheaper version of the watch module, the LWM-6531, has an hours and minutes display with a flashing colon for seconds. One-off price is around £ 15.
The more expensive version, the LWM6560, retails for about £ 18 and displays hours and minutes, seconds, day-of-week and date. It also incorporates a light sensor to adjust the display brightness to suit ambient light and thus extend battery life.

Litronix House,
593 Hitchin Road, Stopsley,
Luton, Bedfordshire.

Vegetable Growing by Computer

Researches in Naaldwijk, Holland, wish to know exactly how fast tomatoes ripen and how well cucumbers, for example, grow in greenhouse conditions. Assistance of a computer has been called on at the research and experimental institute for the cultivation of fruit and vegetables. At the present time this is the largest research project of its kind in the world. A Siemens 330 process computer monitors and controls the environmental conditions in 24 growing compartments within the large greenhouse.
In addition to the central processing unit with a main memory of 64 K words (16 bits each), the computer system comprises a series of data input/output devices. These include, process interfacing devices for connection to the measuring and control systems of the individual climatic chambers. The task of this process periphery is to collect 800 analog and 80 digital signals, while also handling 450 digital output signals. Twenty-six analog signals are recorded in each of the 24 climatic chambers (each 56 m^2 in area). These include temperature values for the air and soil, relative humidity, CO_2 concentration of the air, temperature of the water in the heating system, signals indicating window positions and the various values for the irrigation and drainage systems and the heating system. Sixteen digital outputs in each climatic chamber control the motors of the valve drives and window adjusters. In addition, the weather station parameters can also be included in the calculations and control operations.
The signals issued from each sensor are scanned once every minute. The process computer calculates the manipulated variables and the setpoints of the secondary analog backup controllers by direct digital control so that a smooth changeover to analog control is possible if required.
Every day an estimated one million

measured values, signals and commands are exchanged between the process computer and its peripherals. All relevant data can be logged in tabular form, using the typewriter. In addition, the case history of 256 values over the the previous 96 hours can be graphically represented on the colour curve display station.
The aim of the investigations in Naaldwijk is not only to research the optimum conditions for growing fruit and vegetables but also to determine the climatic conditions under which the plants are least susceptible to diseases. It should then be possible, to a certain degree, to forego the use of chemical agents and pesticides.
The investigations are at present concerned with tomatoes and cucumbers. The research program is later to be extended to other types of vegetable, e.g. red peppers, aubergines, beans and lettuce. Fruit and flowers are also to be investigated at a later date. At present, tomatoes grown in Dutch greenhouses have annual production value of approximately £ 9.6 million.

Siemens AG
Zentralstelle für Information
Postfach 3240, D-8520 Erlangen 2
Federal Republic of Germany

'Noiseless' Discs and tapes

A new process which completely eliminates surface and background noise from disc recordings and captures for the first time the full dynamic range of the music is announced by dbx, Incorporated, manufacturer of noise reduction systems for the professional studio and the audiophile.
The process permits commercial discs as played in the home to equal the performance quality of studio master tapes. This is accomplished by electronically compressing the recorded

selektor

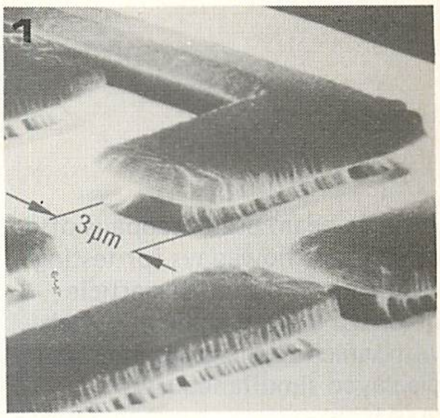
ignal by a factor of 2:1 at the time the naster disc is cut, and expanding the ignal by a complementary factor of :2 at the point of playback. If a dbx encoded master tape is used, full lynamic range and freedom from noise vill be realized upon playback. Master apes produced with other types of noise reduction systems, or with no noise at all, may be used and the played back disc will sound equal to the naster tape. Ordinary discs are limited o a dynamic range of some 60 dB, whereas the dbx encoded disc has a ange well in excess of 100 dB.
The dbx process compresses the lynamic range of the music to a dynamic range envelope' which fits conveniently within the inherent imitations of the record medium, then expands the music to its full original lynamics at the point of playback.
This compression and subsequent expansion reduces record surface noise and other unwanted background noise to inaudibility, so that when the musical program stops, no sound of any kind is heard from the playback system.
The complete absence of background noise also makes the quiet portions of the music easier to hear, and the definition of individual voices and instruments in ensemble music is dramatically improved.
A significant advantage of the dbx disc encoding process is that it does not obsolete any existing manufacturing technique or equipment presently in use in the recording industry, nor does it increase actual product cost in any way. On the contrary, dbx encoding offers numerous opportunities for reducing the cost of recorded music without compromising quality. For example, with dbx encoding, record grooves may be placed closer together, increasing the amount of music on a record by up to 30%.
Electronic expansion circuitry, similar in cost and complexity to Dolby B and quad matrices, is required at the point of playback to properly decode the dbx compressed signal. The decoder circuitry is now available to audio equipment manufacturers for inclusion in consumer audio components and systems on a license basis. Also, many audio component dealers are now selling the 120 consumer series of compressor/expander noise reduction systems which have disc decoding circuitry built in.
The 120 series uses the same 2:1 linear decibel compression/expansion principle used in dbx professional studio equipment, but the sensing circuits have been optimized to best complement the bandwidth requirements of consumer grade reel-to-reel, cassette and cartridge tape recorders. The 120 series is not compatible with dbx professional format tape noise reduction systems used in recording studios.
The new systems allow consumer grade recorders with signal-to-noise ratios as low as 45 to 50 dB to produce full dynamic range tapes which are audibly free from tape hiss and background noise. In excess of 30 dB noise reduction can be realized with the 120 system, along with 10 dB headroom improvement which reduces the likelihood of tape saturation.
The 120 noise reduction format is also used by record manufacturers for processing of dbx encoded records, and a 120 family decoding device is required for playing the dbx encoded discs.
Two models are initially available in the 120 family of noise reduction units.
Model 122 is a two-channel record or playback unit. That is, it will either record or playback, and is switched from one function to the other.
Model 124 is a four-channel record or playback unit suitable for the full range of quadraphonic activities, and having the added feature that when used in a two-channel system it can record and playback at the same time, permitting the recordist to monitor the decoded or normalized signal during recording.
Commercial record labels currently releasing stereo discs in dbx encoded format include Klavier and Creative World, and negotiations are in progress with a number of other lables to release material in dbx encoded format.
Both Models 122 and 124 will decode presently available dbx encoded stereo discs. dbx disc encoding is equally applicable to the production of quadraphonic releases as well, regardless of whether they are produced in discrete (4-4-4) or matrixed (4-2-4) format.

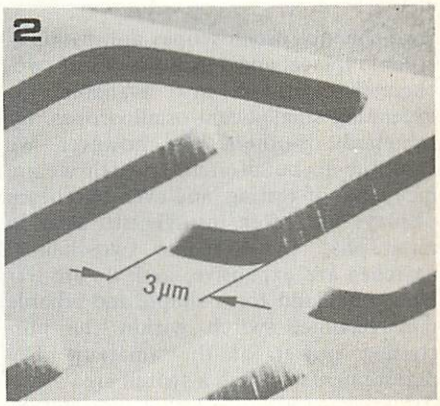
dbx Incorporated,
296 Newton Street,
Waltham, Massachusetts 02154.

Ion beams etch chip structures

New etching process for superfine structures on semiconductor chips.

Semiconductor chip structures have become unbelievably small, and this trend will continue. The wavelength of the light beam, used to create component contours on chips using the photomasking processes, is nowadays often not short enough. This is why electron beams with their markedly shorter wavelengths must be used. Such beams are capable of producing structure spacings of 1 μm and less. However, it is then necessary to use a different method to create the structures, which are defined by a photomask. This is because the chemical etching processes that are being used now have an undercutting effect – whereby the side walls of the structures are washed away from underneath. This just cannot be tolerated in the case of superfine structures in the submicrometer range. A process is at the present time being worked on at the Siemens laboratories, this process uses an ion beam, working mechanically like a sand blasting jet. It cleanly cuts out even the very finest chip structures.
The fast argon ions required for this job are produced in a plasma chamber, then are directed onto the photoresisted silicon chips. The structures are transferred practically without change in dimensions independently of the resist adhesion. The particular advantage of this process is that the side walls of the chip structures are smooth in contrast to chemical etching, and the angle of the side walls has a uniform value of around 65° (see photo).





Comparison of chemical and ion beam etching. The chemically etched structures in the left-hand photo are characterized by undercutting below the photo-mask and the curved side walls. The ion-etched structures in the right-hand photo are characterized by smooth, straight side walls.

Siemens Photo

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West Germany

Dipl. Ing. H. Weidner

channel quadrupler

A single-beam oscilloscope is often insufficient nowadays for testing electronic circuits. This article describes a multi-channel switch with which four signals may be displayed simultaneously.

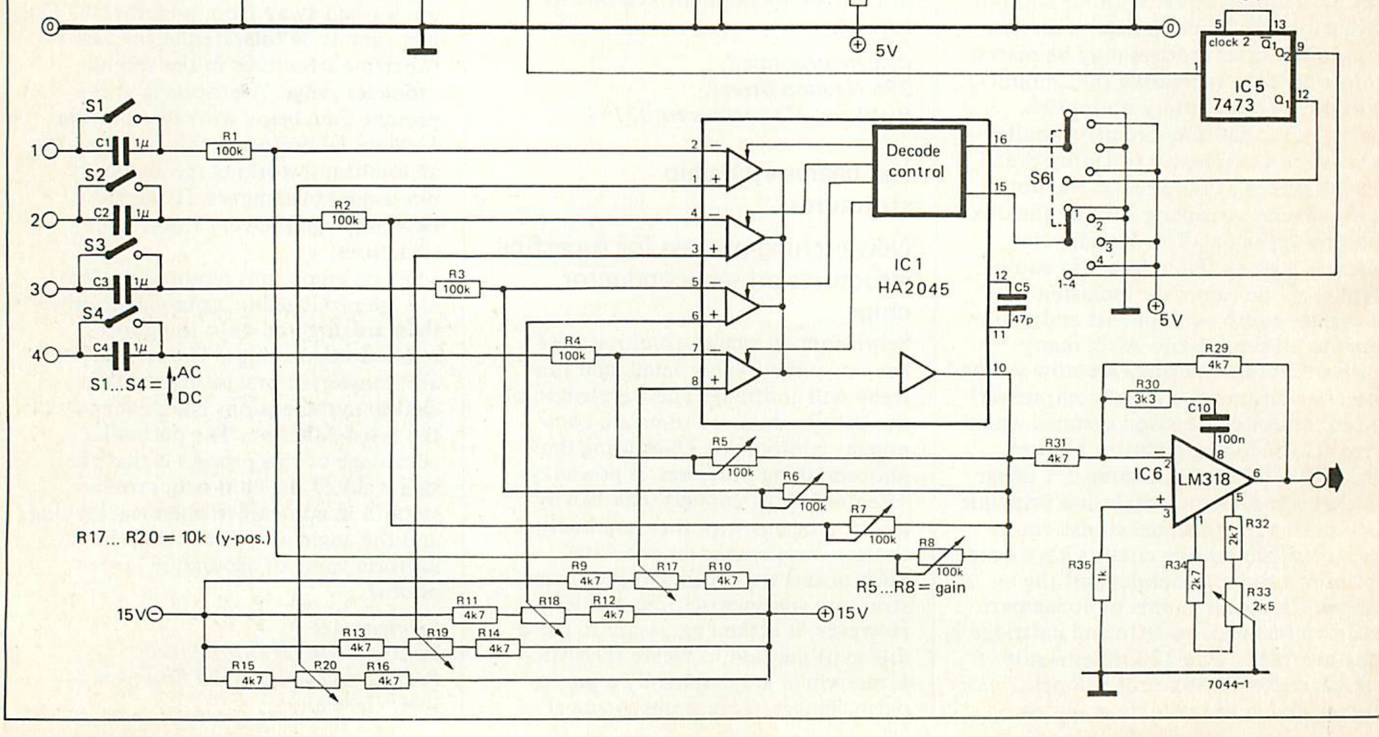
Two-channel oscilloscopes are, by now, commonplace, and various types of two-channel switches are available for extending single-beam oscilloscopes. By somebody-or-others law, however, we continue to build circuits which we are incapable of testing, and even two-beam displays are often insufficient. Bear in mind that the available two-channel switches are expensive, and we are left with a demand for a simple and reliable multi-channel switch within the purchasing power of the amateur. The requirements of such a switch are:

1. four channels;
2. unity gain and a facility for attenuation;
3. Y-position separately adjustable for each channel;
4. both chopped and alternating switching modes;
5. facility for selecting each channel separately.

This article describes a switch to meet these requirements, with the design being kept as simple as possible by starting off with a suitable integrated circuit.

A survey of ICs available on the marke led to the selection of the HA240 made by Harris. This consists of fou operational amplifiers (opamps), only one of which is activated at any one time depending on the information a the two control inputs (pins 15 and 16) The outputs of the four opamps are internally connected to the commor output amplifier, so that the output o the activated opamp is available at the common output point (pin 10). Other wise, the IC behaves as expected for ar opamp. The maximum output voltage

1

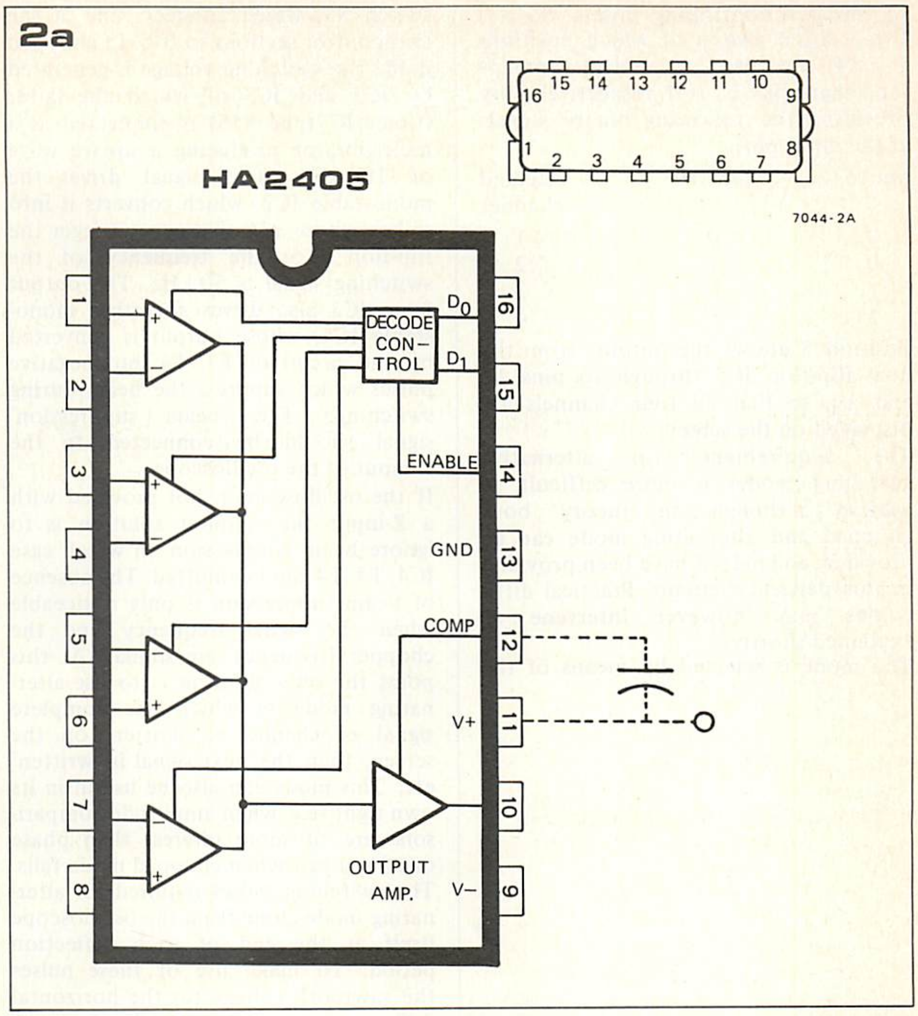


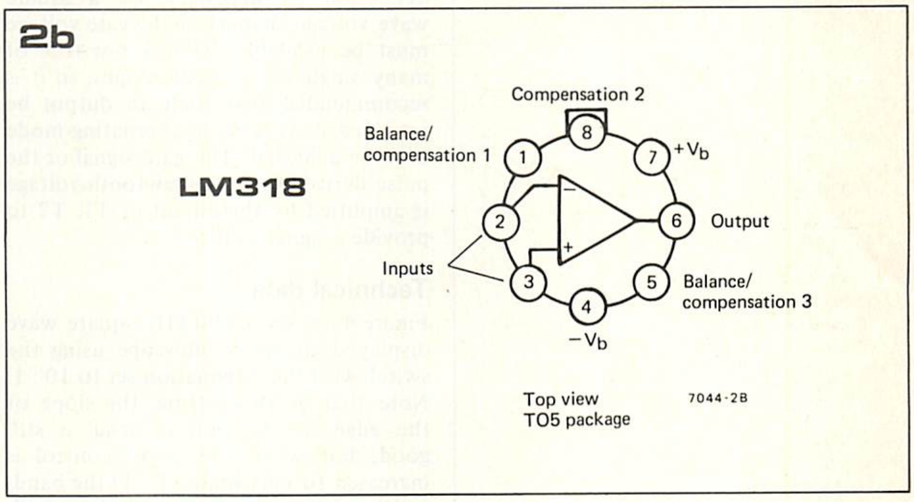
igure 1. Circuit diagram of the four-channel witch.

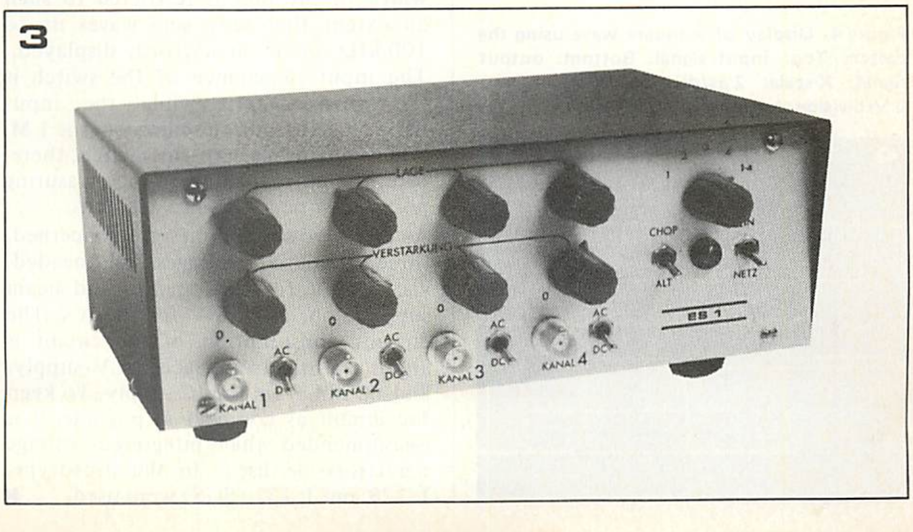
igure 2A. Pin connection diagram for the IA2405.

igure 2B. Pin connection diagram for the M 318.

igure 3. Front view of the (German!) prototype.







ariation is ± 10 V, the gain and attenuation being adjustable in the usual way with feedback resistors.

The circuit

Figure 1 gives the circuit diagram of the witch. As can be seen, the four opamps n IC1 provide the central part of the ircuit, with their gain and attenuation eing controlled by the input resistors R1 . . . R4 and the feedback resistors R5 . . . R8. The gain of the first opamp s controlled by the ratio of R1 : R5 similarly for the other opamps); the naximum gain in this case being unity 0 dB), and the attenuation being controlled by the variable resistors R5 . . . R8. Calibration of the attenuation has not been included (since phase comparisons are usually of more nterest) but it would be fairly simple o include calibration if required, either by fitting a calibration scale for the variable resistors, or by replacing them with multi-position switches between fixed resistors. The maximum input voltage should not exceed ± 10 V, which s sufficient for most applications, although the input voltage range could always be extended by using voltage-divider probes with an attenuation of 10 : 1.

The Y-position of the signals displayed on the screen is controlled by the potentiometers R17 . . . R20 which control the voltage to the non-inverting nputs of the opamps. The values of the resistors in series with these potentiometers have been chosen to give a full-screen display on the oscilloscope with the input sensitivity set to 1 V/cm. Other values may be substituted if other ratios are required.

Since the HA2405 performs an inversion between input and output, IC6 has been ncluded to invert the signal again so that the correct polarity is obtained at the output of the switch. Also included here is a feed-forward circuit R30, C10, which improves the slew rate.

Moving on to channel selection, this is achieved by the switch S6, through which digital information is supplied to the two controlling inputs. S6 is a five-position switch of which positions 1 . . . 4 are used to select opamps (i.e. channels) 1 . . . 4 respectively, by providing the following binary signals at the input pins:

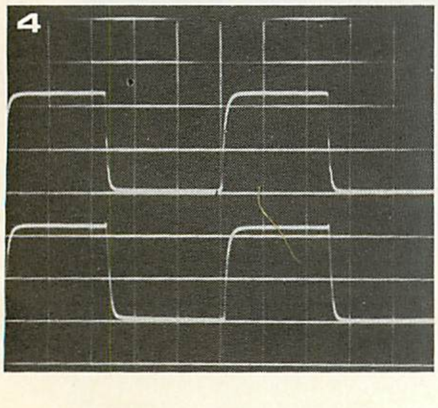
pin 15	pin 16	selected channel
0	0	1
0	1	2
1	0	3
1	1	4

Position 5 allows the outputs from the dual flip-flop IC5 through to pins 15 and 16, so that all four channels are displayed on the screen.
The requirement for alternative switching modes is more difficult to achieve although in theory both chopped and alternating mode can be provided, and indeed have been provided in this particular circuit. Practical difficulties may however intervene as explained shortly.
The mode is selected by means of the switch S5, which connects one of the two control sections to IC5. In chopped mode the switching voltage is generated by IC2 and IC3 of which the latter (timer IC type 555) is connected as a multivibrator producing a square wave of 100 kHz. This signal drives the monostable IC2, which converts it into spikes whose negative edges trigger the flip-flop IC5; the frequency of the switching signal is 50 kHz. The output from IC3 also drives a further monostable IC4, whose output is converted by the circuit of T3, T4 into negative pulses which suppress the beam during switching. The 'beam suppression' signal should be connected to the Z-input of the oscilloscope.
If the oscilloscope is not provided with a Z-input the simplest solution is to ignore beam suppression, in which case IC4, T3, T4 can be omitted. The absence of beam suppression is only noticeable when the signal frequency and the chopper frequency are similar. At this point the only solution is to use alternating mode in which one complete signal or channel is 'written' on the screen, then the next signal is 'written' etc. This mode can also be useful in its own right (e.g. when amplitude comparisons are of more interest than phase ones) not just when chopped mode fails. The switching pulses required for alternating mode come from the oscilloscope itself at the end of each deflection period. To make use of these pulses the sawtooth voltage for the horizontal deflection (if necessary via a square wave voltage shaper) or the gate voltage must be available. This is not true of many single-beam oscilloscopes, so it is recommended that such an output be provided if working in alternating mode is to be achieved. The gate signal or the pulse derived from the sawtooth voltage is amplified by the circuit of T1, T2 to provide a signal to IC5.

Technical data

Figure 4 shows a 100 kHz square wave displayed on an oscilloscope using the switch with the attenuation set to 10 : 1. Note that as this setting, the slope of the edge of the output signal is still good, but when the 'gain' control is increased to maximum (1 : 1) the bandwidth of the unit is restricted to such an extent that only sine waves up to 100 kHz can be satisfactorily displayed.
The input impedance of the switch is less than 100 k, while the input impedance of most oscilloscopes is 1 M. When measuring with this switch, therefore, the extra loading on the measuring point must be taken into account.
As far as power supplies are concerned, three stabilised voltages are needed, vis: ± 15 V for the opamps and beam suppression, and +5 V for the ICs. The current consumption of the circuit is about 25 mA from each 15 V supply, and 60 mA from the IC supply. To keep the circuit as compact as possible, it is recommended that integrated voltage regulators be used. In the prototype, L 129 and L 131 (SGS) were used.

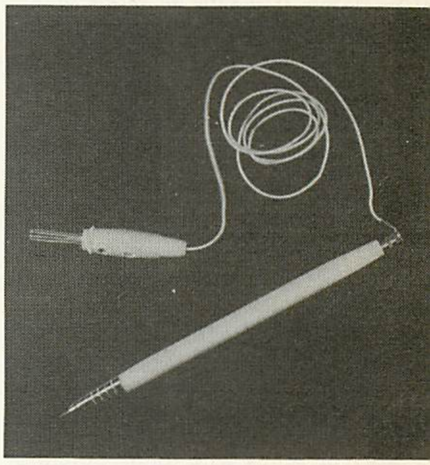
Figure 4. Display of a square wave using the switch. Top: input signal. Bottom: output signal. X-scale: 2 μs/division. Y-scale: top: 2 V/division; bottom: 0.2 V/division.



J. Hájek

measuring pencil

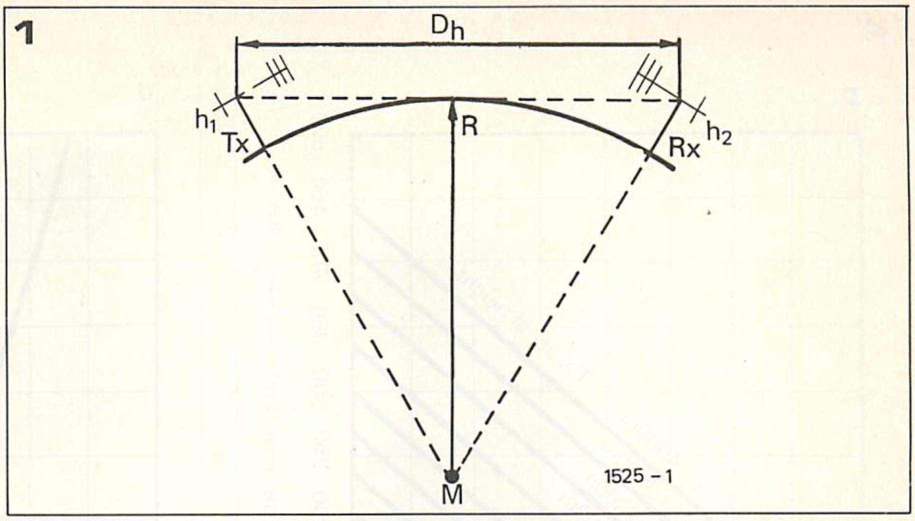
Here is a device to help with that of recurring requirement – an extra pai of hands. This measuring pencil can b used as a probe for inspecting voltages in addition to its conventional appli cation of writing down the results.
The measuring pencil consists of propelling pencil made of a syntheti resin. A flex is soldered to the push button such that the propelling mechan ism is unimpaired, and the other end o the flex is provided with a plug to fi the measuring instrument (e.g. a volt meter). To use the pencil, connect th circuit to be tested to the earth termina of the meter, and the pencil to the inpu socket. Using the pencil as a probe, th circuit voltages may now be measurec at various points and the results writter down with the same instrument.

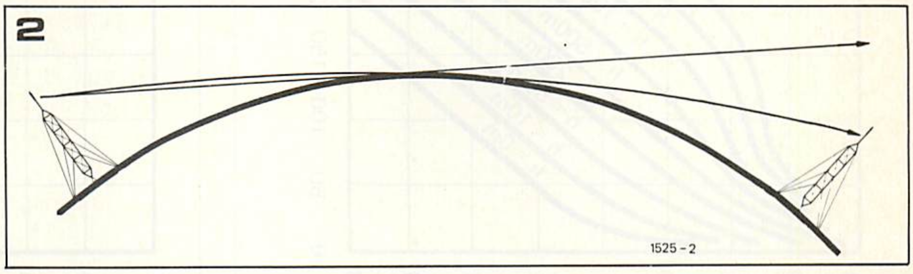


The measuring pencil is particularly suitable for circuits with low voltages (⩽ 42 V AC or ⩽ 60 V DC) as the insulation of the pencil is not then a problem. The contact resistance between the measuring point and the meter is less than 1 Ω in this design, and when used with modern high impedance equipment, the measuring error is negligible.

VHF FM reception

Dwners of FM tuners may often vish to know what sort of signal evel they can expect to receive in he locality in which they live. The simplest and most accurate nethod is to obtain a direct eading using a field strength neter, but of course very few people possess one of these nstruments. The charts produced by the BBC of the ervice areas of their various ransmitters provide a useful juide, but it is often possible with a sensitive receiver and good erial to obtain a usable signal outside the service area. This rticle investigates the rules joverning VHF propagation and hows how received signal trengths may be estimated with few simple calculations.





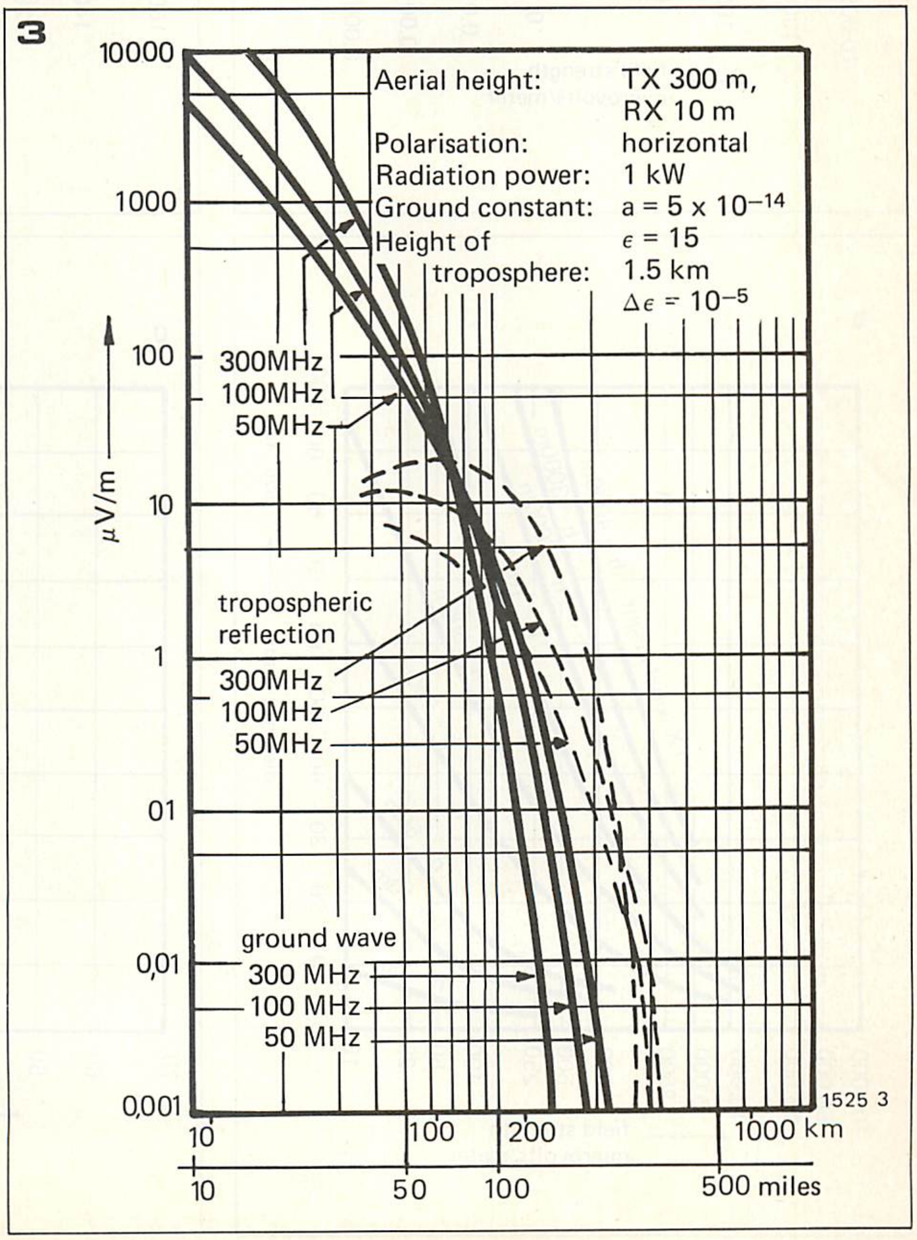
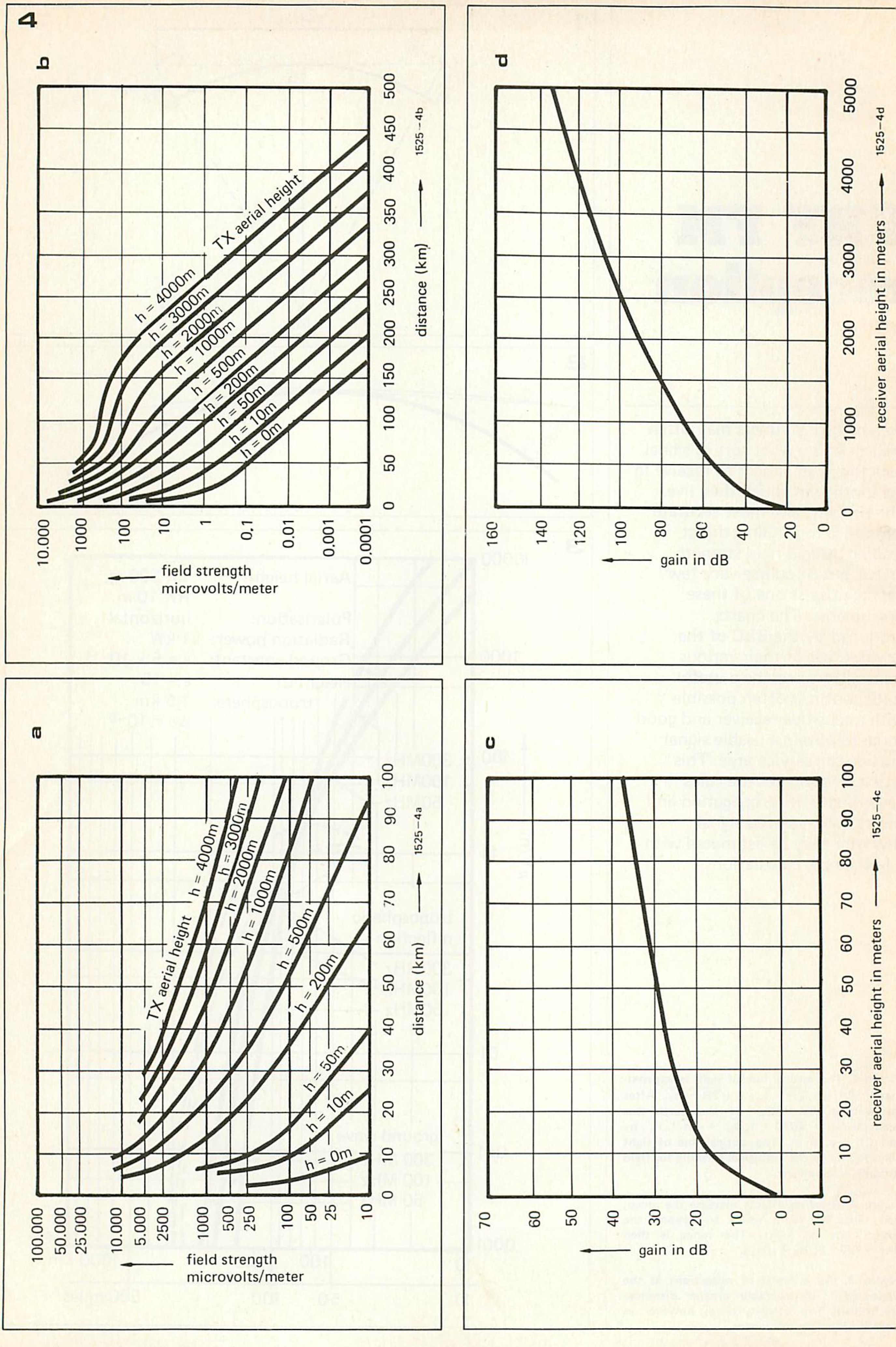
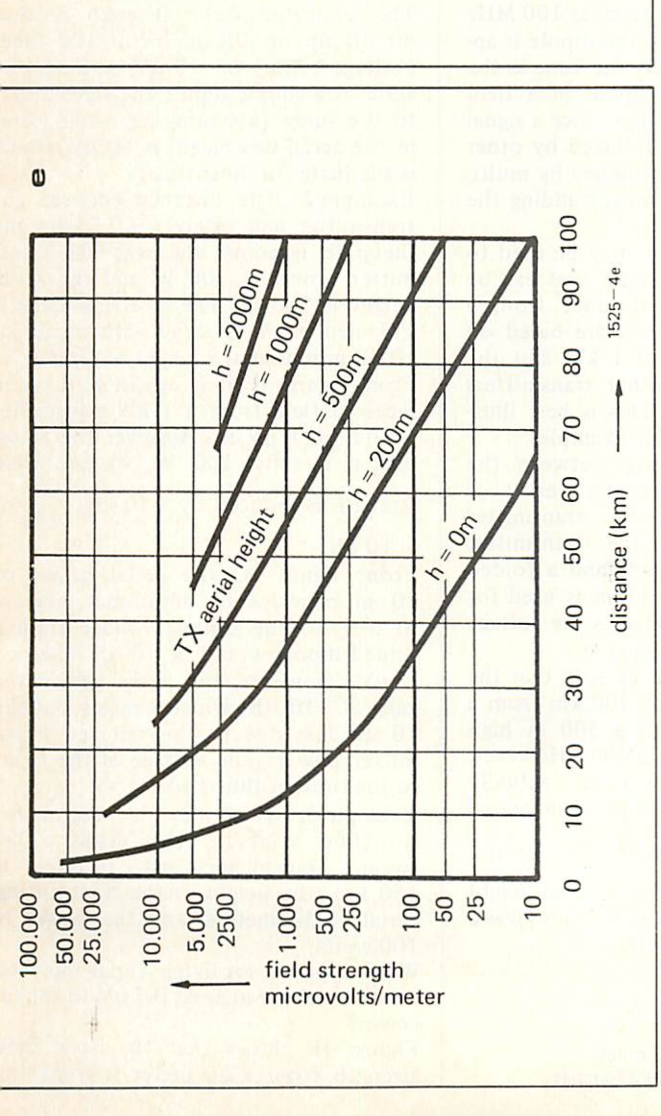


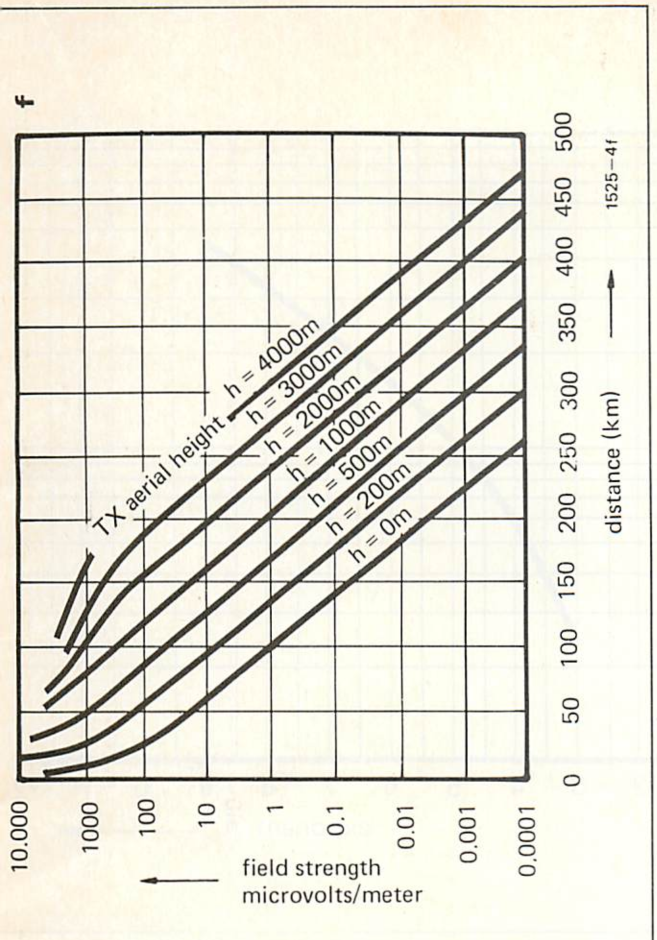
Figure 1. The optical line of sight is approximately $Dh = \sqrt{2R \cdot h_1} + \sqrt{2R \cdot h_2}$. After ntroducing the radius of the earth this becomes $Dh = 3570 \cdot (\sqrt{h_1} + \sqrt{h_2})$. h_1, h_2 and Dh are in m. The optical line of sight plays a part in the empirical formula for field trength calculations.

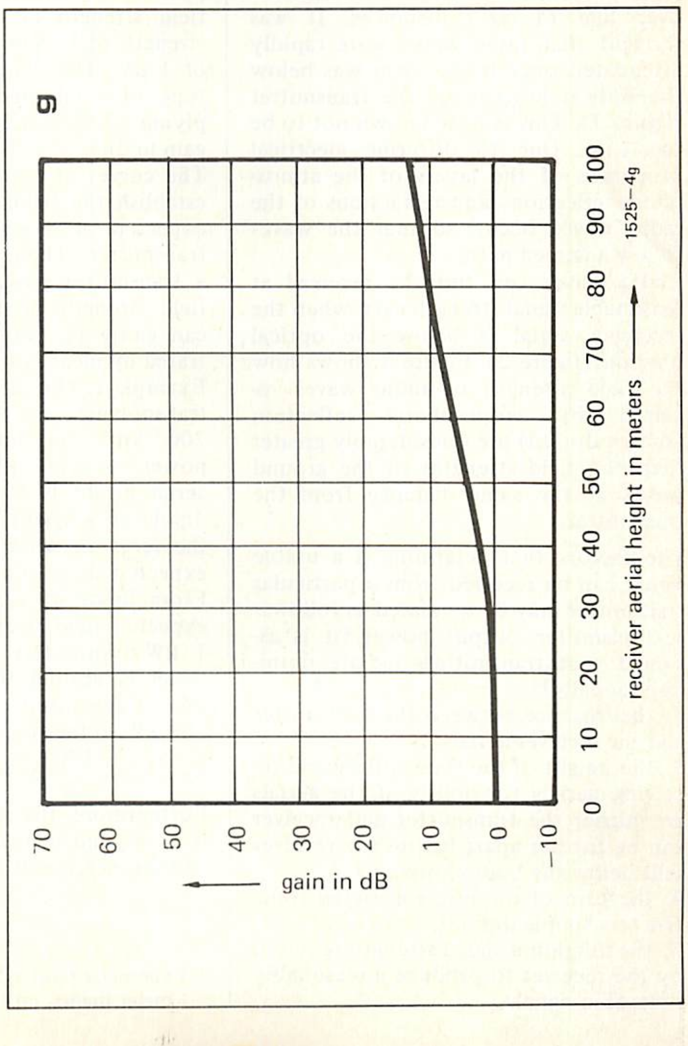
Figure 2. Bending effects increase the recepion range of radio waves to beyond the optical line of sight. The range is then $Dh = 4120 \cdot (\sqrt{h_1} + \sqrt{h_2})$.

Figure 3. As a result of reflections at the roposphere, considerably greater distances are bridged. The intensity varies, however, as result of cosmic influences.


4
b
TX aerial height
h = 4000m
h = 3000m
h = 2000m
h = 1000m
h = 500m
h = 200m
h = 50m
h = 10m
h = 0m
distance (km)
field strength microvolts/meter
1525 – 4b
d
gain in dB
receiver aerial height in meters
1525 – 4d
a
TX aerial height
h = 4000m
h = 3000m
h = 2000m
h = 1000m
h = 500m
h = 200m
h = 50m
h = 10m
h = 0m
distance (km)
field strength microvolts/meter
1525 – 4a
c
gain in dB
receiver aerial height in meters
1525 – 4c


e
field strength microvolts/meter
100.000
50.000
25.000
10.000
5.000
2500
1.000
500
250
100
50
25
10
distance (km)
0
10
20
30
40
50
60
70
80
90
100
TX aerial height
h = 2000m
h = 1000m
h = 500m
h = 200m
h = 0m
1525–4e


f
field strength microvolts/meter
10.000
1000
100
10
1
0.1
0.01
0.001
0.0001
distance (km)
0
50
100
150
200
250
300
350
400
450
500
TX aerial height
h = 4000m
h = 3000m
h = 2000m
h = 1000m
h = 500m
h = 200m
h = 0m
1525–4f


g
gain in dB
70
60
50
40
30
20
10
0
–10
receiver aerial height in meters
0
10
20
30
40
50
60
70
80
90
100
1525–4g

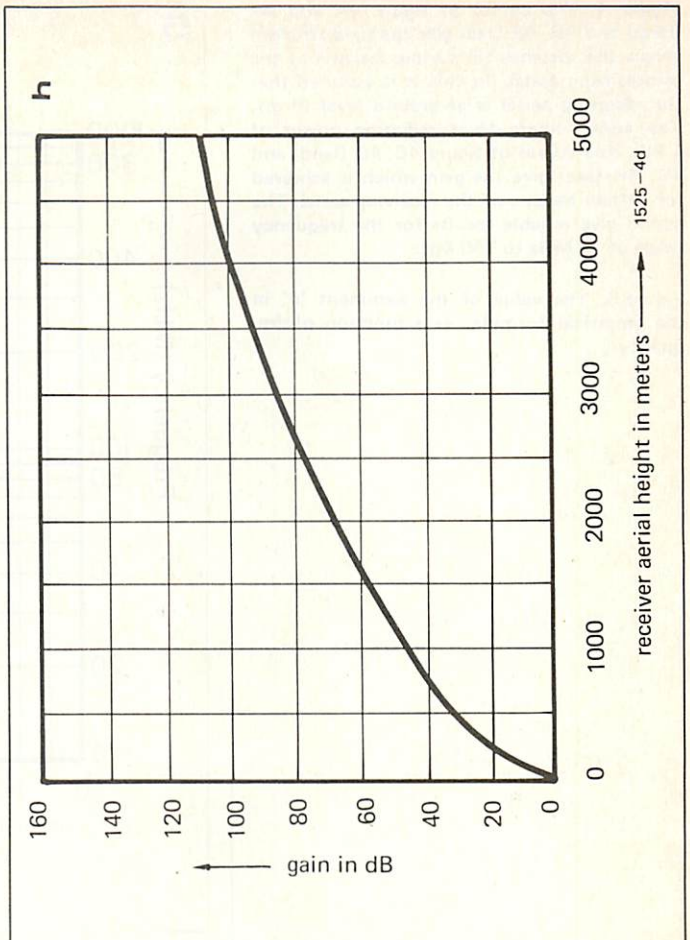
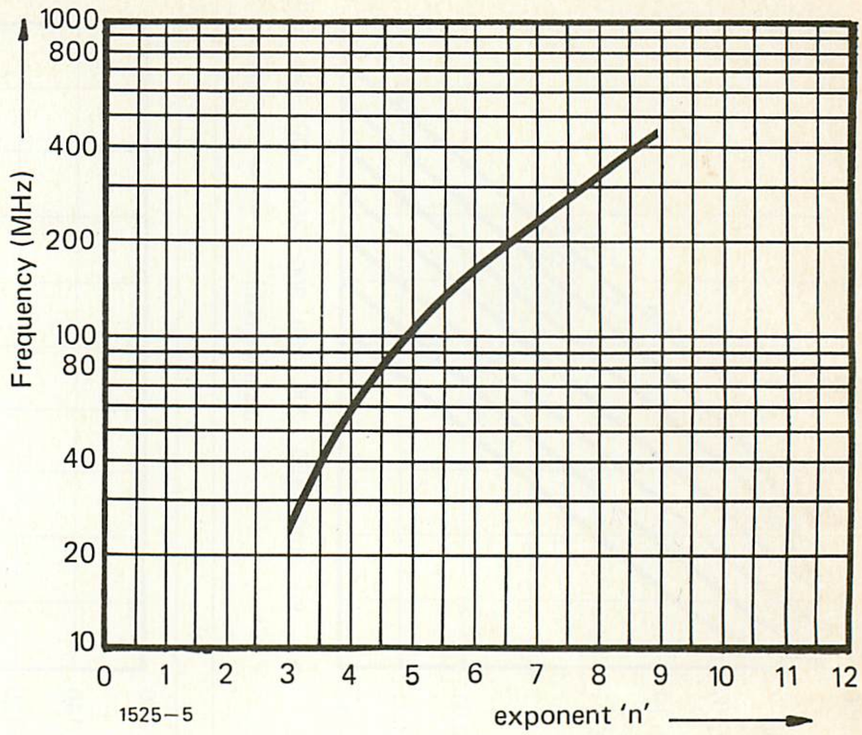

h
gain in dB
160
140
120
100
80
60
40
20
0
receiver aerial height in meters
0
1000
2000
3000
4000
5000
1525–4d

Figure 4. The curves of figure 4A and 4B (land) and 4E, 4F (sea) give the field strength versus the distance for various heights of the transmitting aerial. In this it is assumed that the receiving aerial is at ground level (0 m). The curves relate to a radiation power of 1 kW. The curves of figure 4C, 4D (land) and 4G, 4H (sea) give the gain which is achieved for certain heights of the receiving aerial. The curves give reliable results for the frequency range of 70 MHz to 150 MHz.

Figure 5. The value of the exponent 'n' in the empirical formula, as a function of frequency.

5



In the early days of wireless it was believed that radio waves of short wavelength (VHF) could only be transmitted over line of sight distances. It was thought that radio waves were rapidly attenuated once the receiver was below the optical horizon of the transmitter (figure 1). This is now known not to be the case. Due to differing electrical properties of the layers of the atmosphere reflections and refractions of the radio waves occur, so that the waves follow a curved path.
Radio waves can thus be received at reasonable signal strength even when the receiving aerial is below the optical horizon (figure 2). Figure 3 shows how the field strength of radio waves received by tropospheric reflection (shown dotted) are considerably greater than the field strengths of the ground waves at the same distance from the transmitter.

The factors that determine if a usable signal can be received from a particular transmitter may be tabulated as follows:
1. transmitter output power (it is assumed that transmitters radiate omnidirectionally);
2. the distance between the transmitter and the receiver aerials;
3. the height of the transmitting and receiving aerials (obviously if the aerials are higher the transmitter and receiver can be further apart before the receiver falls below the 'radio horizon');
4. the gain of the receiving aerial (relative to a simple dipole);
5. the minimum signal strength required by the receiver to produce a reasonably noise-free signal.

Using a folded dipole aerial at 100 MHz the signal produced by the dipole is approximately numerically the same as the field strength i.e. a dipole in a field strength of 1 μV/m will produce a signal of 1 μV. The signal produced by other types of aerial can be obtained by multiplying by the aerial gain (or adding the gain in dB).
The curves of figure 4 may be used to establish the field strength that can be expected at a given distance from a transmitter. These curves are based on a transmitter power of 1 kW, but the field strengths with other transmitters can easily be found. This is best illustrated by means of some examples.
Example 1. The distance between the transmitting and receiving aerials is 200 km* on land, the transmitter power is 100 kW, the transmitter aerial height is 500 m*, and a folded dipole at a height of 15 m is used for the receiving aerial. What is the voltage expected at the tuner input?
From figure 4B it can be seen that the expected field strength 200 km from a 1 kW transmitter with a 500 m high aerial is around 0.1 μV/m. However, the transmitter power is actually 100 kW so the field strength is increased by 10 log $\frac{100}{1}$(power ratio!) or 20 dB.
Furthermore, the receiving aerial height is 15 m, and from figure 4C this gives a further increase of 20 dB.

* 1 kilometer (km) = 0.62 miles;
1 meter (metre, m) = 39.37 inches.

The expected field strength is thu 40 dB up on 0.1 μV/m or 100 time (voltage ratio!) i.e. 10 μV/m. Since th aerial is a simple dipole the signal inpu to the tuner (assuming negligible losse in the aerial downlead) is 10 μV, whic is adequate for most tuners.
Example 2. The distance between th transmitter and receiver is 175 km an the path is across the sea. The trans mitter power is 100 W and the aeria height is 500 m. The receiving aerial i a 4 element Yagi array with a gain o 10 dB mounted at a height of 10 m.
From figure 4F it is apparent that th received field from a 1 kW transmitte would be 1 μV/m. However the trans mitter is only 100 W, so the fiel strength is reduced by $-10 \log \frac{1000}{100}$ o – 10 dB.
From figure 4G the aerial height o 10 m provides no additional increas (0 dB), so the output voltage from folded dipole would be 10 dB down o 1 μV. However, the aerial provides gain of +10 dB, which cancels out th 10 dB loss due to the reduced trans mitter power. The voltage at the inpu to the tuner is thus 1 μV.
Example 3. This may be useful fo would-be spies . . . The distance be tween transmitter and receiver i 150 km, the height of the transmittin aerial is 50 meters, and the power i 100 watts.
What type of receiving aerial must b used to deliver at least 0.5 μV to the re ceiver?
Figure 4B shows that the basic fiel strength from a 50 meter transmittin

aerial at 150 km is 0.01 μV/meter. Since a voltage of at least 0.5 μV is required at the receiver input, the basic gain requirement is equal to

$$20 \log \frac{0.05}{0.01} = 34 \text{ dB}.$$

However, that's for a 1 kW transmitter but the power in this case is only 100 W. Therefore the field strength is reduced by a further $10 \log \frac{100}{1000} = -10$ dB.

Therefore, the receiving aerial will have to give a total gain of 44 dB (10 dB for the low transmitting power and 34 dB for distance). One possible way to achieve the 44 dB gain figure is to attach a folded dipole to a balloon floating at 250 meters. A more practical (?) solution is a 6-element yagi array with a gain of 14 dB mounted at 50 meters.

Example 4. This is of a more practical nature, and can be modified for individual circumstances. A 4-element aerial with a gain of 10 dB is mounted at a height of 20 meters, what receiving range can be expected for reasonably noise free FM stereo broadcasts?

First, some assumptions must be made: most FM transmitters have power outputs of 100 kW or more with aerial heights of over 200 meters; most receivers need approximately 100 μV for noise-free stereo reception.

Adding the basic receiving aerial gain (10 dB) to the gain due to receiving aerial height as derived from figure 4C (22 dB), a total aerial gain of 32 dB is found. Since the receiver requires 100 μV, the 32 dB aerial gain means that a field strength of 2.5 μV/m is required at the aerial. If the transmitter power was 1 kW a look at figure 4B would show the distance.

However, the transmitter power is 100 kW. This means that the field strength at the receiving aerial is increased by $10 \log \frac{100}{1} = 20$ dB for any given distance. 20 dB in field strength is a factor 10, so if a 100 kW transmitter gives the required 2.5 μV/m at a certain distance, a 1 kW transmitter would give a field strength of $\frac{2.5}{10} = 0.25$ μV/m at the same distance.

According to figure 4B, if the transmitting aerial height is 200 meters a 1 kW transmitter will give a field strength of 0.25 μV/m (so the 100 kW transmitter will give 2.5 μV/m) at a distance of 150 km.

Or, looking at it the other way round, using the 4-element aerial at 20 meters height with an ordinary receiver it should be possible to get good (stereo) reception from any transmitter within a range of 150 km over land.

Of course, the results obtained from the graphs of figure 4 are true only where the transmission path is over the sea or over relatively flat terrain. If the receiving aerial is on top of a hill then the received field strength will be greater since the effective height of the aerial is greater. Conversely, if the receiving aerial is in a valley the field strength will be less since the effective height of the aerial is smaller and the aerial will be screened by the walls of the valley.

Instead of using graphs, the received field strength may be calculated empirically from the equation given below, which takes account of the terrain by including a term for the distance to the optical horizon (obviously the distance to the optical horizon is greater for an aerial on flat terrain than for one mounted at the bottom of a valley).

The equation is as follows:

$$E = \frac{88 \cdot \sqrt{P} \cdot h_1 \cdot h_2 \cdot D_h^{n-2}}{\lambda D^n}$$

Where E is the r.m.s. value of the field strength in volts per metre;

P is the effective radiated power in watts;

h_1 is the height of the receiving aerial in metres;

h_2 is the height of the transmitting aerial in metres;

D_h is the distance of the optical horizon in metres (figure 1);

λ is the wavelength in metres;

D is the distance between transmitter and receiver in metres;

n is a frequency dependent exponent (see figure 5).

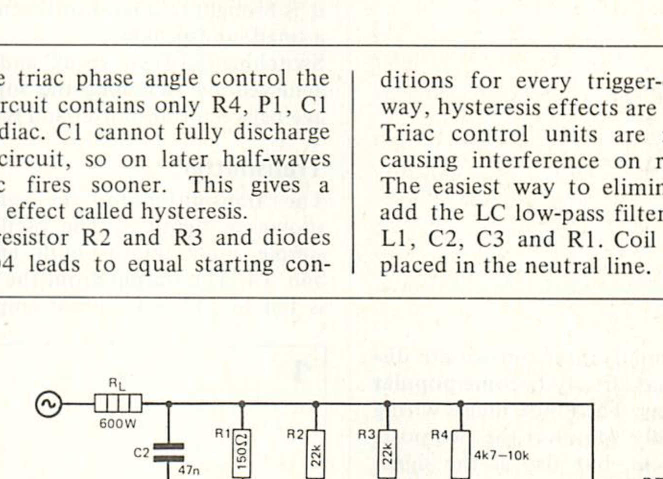
Both the curves of figure 4 and the empirical equation apply only for normal atmospheric conditions. During unusual weather conditions (such as inversion layers) VHF reception at distances of several thousand kilometres have been observed. This, of course, is not the norm, and is really of interest only to those interested in DX activities.

triac control

In simple triac phase angle control the trigger circuit contains only R4, P1, C1 and the diac. C1 cannot fully discharge in that circuit, so on later half-waves the triac fires sooner. This gives a 'snap-on' effect called hysteresis.

Adding resistor R2 and R3 and diodes D1 . . . D4 leads to equal starting conditions for every trigger-cycle. In this way, hysteresis effects are avoided.

Triac control units are notorious for causing interference on radio and TV. The easiest way to eliminate this is to add the LC low-pass filter consisting of L1, C2, C3 and R1. Coil L1 should be placed in the neutral line.



A.M. mains intercom

Mains intercoms of a more or less reasonable quality are still a bit expensive on the market. Consequently, there appears to be a fair demand for a super simple and cheap a.m. intercom which despite a modest performance will be useful in certain applications where mains interference is not excessive.

The type of mains intercom we are discussing here has already become popular as a babyphone. The house mains wiring is used not only to power the two posts of the intercom, but also as the signal connection. Each post is therefore plugged in and the exasperating task of laying out and rolling up lines is eliminated. Intercoms using the mains as their signal connection are, however, susceptible to mains-born interference. The unit described here uses amplitude modulation (a.m.), as do most of the commercially available units. This gives a reasonable compromise between simplicity and performance. It is worth noting here that a more complex design using frequency modulation (f.m.) to obtain high quality results will be published in a future issue of Elektor.

Arrangement

Every intercom post consists of a transmitter and a receiver. Figure 1 shows the block diagram of one such post. In the position 'speak' (or 'transmit') a simple oscillator produces a carrier which is amplified by an output stage. The resulting signal is amplitude modulated by the amplified microphone signal. The high frequency signal is then fed into the mains via a special transformer. In the position 'listen' (or 'receive') the high frequency signal transmitted from another post is picked up from the same transformer (at point A) and fed to a high frequency preamplifier. From there it goes to a modest low frequency amplifier where it is brought to a level sufficient to drive a small loudspeaker.
Switching between 'speak' and 'listen' is achieved by switching the supply voltage between transmitter and receiver.

Transmitter

The transmitter for the intercom is shown in figure 2. The oscillator is a simple multivibrator built around T3 and T4. The output from the oscillator is fed to a class C output amplifier T6, via a buffer stage T5. The collector voltage of this output stage is controlled by the amplifier T2/T7 which is adjusted to its maximum. This amplifier is in turn driven by the microphone amplifier T1, whose gain may be varied by the potentiometer P1 in order to vary the modulation depth. The final result is that an amplitude modulated high frequency signal is fed into the mains via transformer Tr1. Diodes D1 . . . D4 protect the output stage against voltage peaks at switch on. Capacitors C9 and C10 isolate the circuit from the mains.

Receiver

The receiver, which is very simple in design, is shown in figure 3. The transmitted signal is picked up at point A in the transmitter circuit and fed to the receiver. The received signal is amplified considerably by the circuit around T8 and T9, and then detection takes place in the simple demodulator D7, D8, C15. The automatic gain control (a.g.c.) circuit formed by R18, D5, D6 is designed to operate only on very high input levels so protecting the listener from high level mains-born interference.
The low frequency amplifier is simple but adequate. The output power is about 250 mW, which is sufficient for good intelligibility. The volume may be

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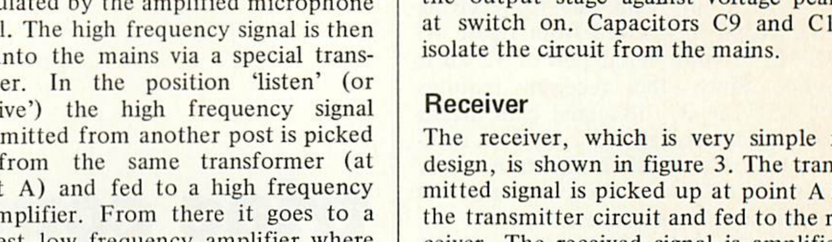

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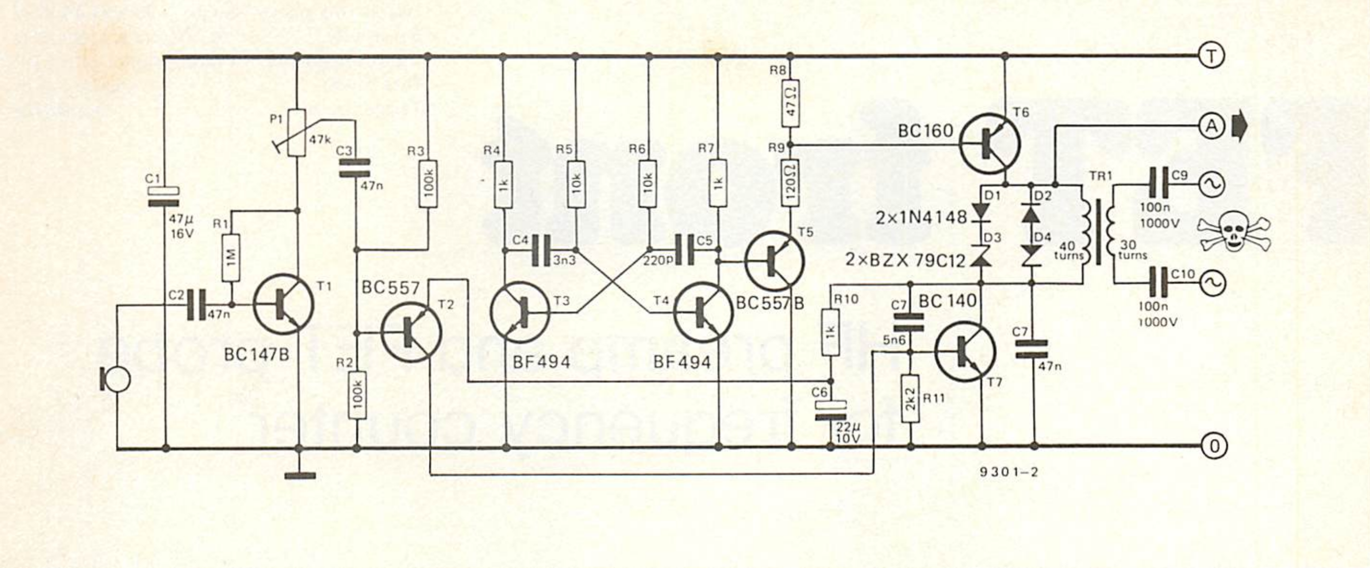

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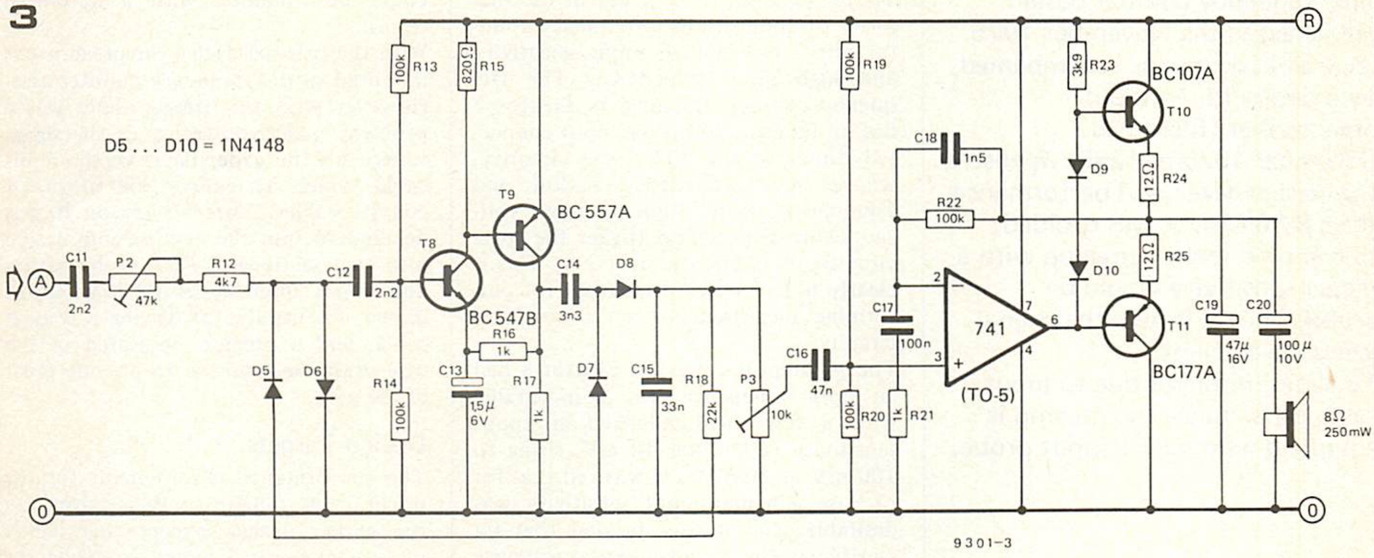

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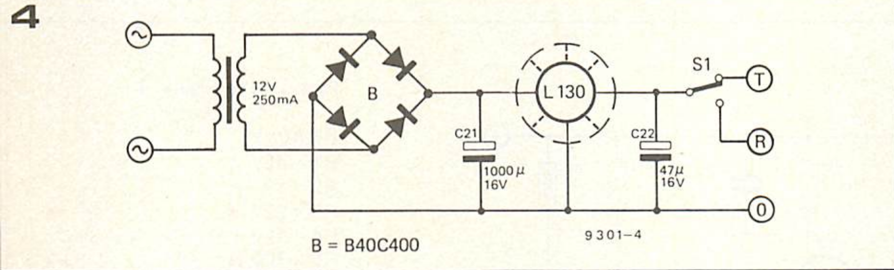

Figure 1. Block diagram of one post of the mains intercom. Each post is a combination of an a.m. transmitter and a conventional 'direct' receiver.

Figure 2. Circuit diagram of the a.m. transmitter. Either high or low impedance microphones may be used.

Figure 3. Circuit diagram of the receiver and the l.f. amplifier.

Figure 4. A suitable power supply for the intercom. S1 switches between transmitter and receiver.

controlled by P3.

It is advisable to set P2 as low as practicable (i.e. so that the modulation is at the audible threshold) as a considerable number of components will then remain below the detection level hence limiting the interference on reception as much as possible.

Conclusion

The simplicity of this design gives more than a hint of the quality if its performance. Since the transmitting power is fairly low (about 1 W) and interference suppression cannot reasonably be compared with that obtained with a narrowband f.m. system, good performance can only be expected in a conventional one-family house. For a number of applications, this will be sufficient.

Thanks to the low transmitting power though, the current consumption is very low, and a simple supply will do, figure 4 shows the circuit diagram of a suitable supply using an IC L130. The switch S1 changes the supply between transmitting and receiving circuits.

The circuit is not very critical and can therefore be built without too much difficulty. The only obstacle may be the transformer which must be wound. For the prototype, a potcore AL250 with a diameter of 18 mm was used. Various manufacturers (e.g. Siemens, ITT and Philips) supply these potcores in a range of versions and sizes.

Specification

Usable Frequency Range:	20 kHz-45 MHz
Sensitivity:	4 mV r.m.s. at 20 MHz
Input Impedance:	1 M in parallel with 5 pF
Rise Time:	approx. 5 ns
Trigger Level:	adjustable

FET front

HF preamp and FET probe for frequency counter

The frequency counter design published in the November 1975 issue of Elektor was accompanied by a design for an input preamplifier (Elektor 8, December 1975 p. 1235). Whilst this design gave good performance from 0-20 MHz it was decided that for r.f. work a preamp with a higher sensitivity would be useful, since it is here that signal levels are smallest.
To avoid problems due to input cable capacitance the preamp is equipped with a FET input probe.

To be of any practical use in the majority of applications a frequency counter must have a high input sensitivity and high input impedance. The frequency counter described in Elektor 7 has, in its basic form, the input connected direct to the TTL logic circuitry, whose input impedance is low and asymmetric. In addition the input voltage swing required to trigger the logic circuitry is of the order of 2 V. This is clearly not of much use except for performing measurements on other logic circuits.
The preamp described in Elektor 8 had an input impedance of 1 M in parallel with a few picofarads and an input sensitivity of around 40 mV, rising to 100 mV at 20 MHz. It was felt that for r.f. use a higher input sensitivity was desirable, and it was decided that by sacrificing the low-frequency response (for reasons explained later) a high gain could be obtained with a simplified circuit.
With the original design the preamp was mounted in the frequency counter case. However with this design there was a problem: reactive loading of the signal source by the capacitance of the input cable, which can be over 100 pF/m for coaxial cables. For this reason it was decided to split the new preamp design into two sections, a FET probe with a high input impedance and low output impedance capable of driving a coaxial cable, and a preamp, mounted in the case with the counter, to provide most of the gain.

Design Targets

The performance requirements for the preamp are similar to those given for the earlier design, except that higher input sensitivity is aimed at, while the low-frequency response is unimportant,

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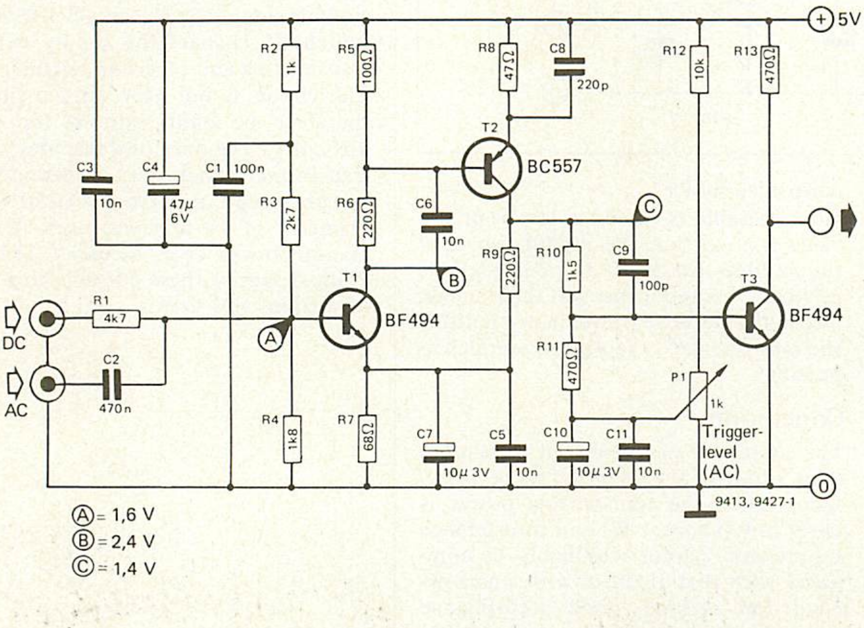


Parts List to Figure 1

Resistors
R1 = 4k7
R2 = 1 k
R3 = 2k7
R4 = 1k8
R5 = 100 Ω
R6,R9 = 220 Ω
R7 = 68 Ω
R8 = 47 Ω
R10 = 1k5
R11,R13 = 470 Ω
R12 = 10 k
P1 = 1 k

Capacitors
C1 = 100 n
C2 = 470 n
C3,C5,C6,C11 = 10 n
C4 = 47 µ/6 V
C7,C10 = 15 µ/3 V
C8 = 220 p
C9 = 100 p

Semiconductors
T1,T3 = BF494,BF194,BF195
T2 = BC557,BC157

since the circuit is specifically intended for r.f. work. The requirements are tabulated below.

1. Bandwith. It was decided that the usable frequency range of the preamp should extend from the top end of the audio band to above the upper frequency limit of the counter (18 MHz).

2. Input sensitivity. It was decided that 10 mV was a useful value and could be obtained without resorting to complex circuitry.

3. Input impedance. This should be as high as possible i.e. input resistance should be high and input capacitance low.

These design requirements are met, and in some cases exceeded, by the new design. The usable frequency range of the preamp + probe extends from below 20 kHz* to above 45 MHz. At 20 MHz the input sensitivity is around 4 mV, whilst at 45 MHz it is still only 17 mV, which is better than the l.f. sensitivity of the original design. These figures refer to the r.m.s. input voltage necessary to cause an output voltage swing that will reliably trigger the TTL Schmitt input of the frequency counter. The impedance of the probe input is 1 M in parallel with 5 pF.

Preamp Circuit

The preamp circuit is shown in figure 1 and consists of a simple, three-stage, direct coupled amplifier. To minimise the effect of transistor capacitances and stray circuit capacitance the resistor values around the circuit are kept low. As a consequence of this the low-frequency response is sacrificed, since to extend the l.f. response down into the audio band excessively large value electrolytics would have been required for C2 and C7 (especially C7). Quite apart from the size consideration the parasitic inductances of such large electrolytic capacitors can cause undesirable resonances at higher frequencies.

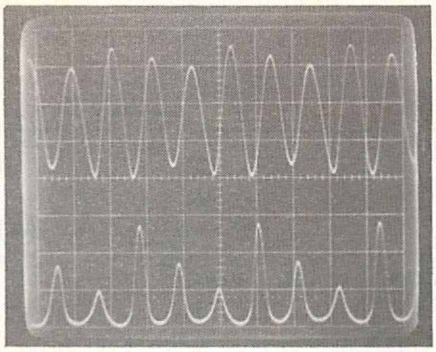
The preamp is provided with two inputs, an a.c. input, which is normally fed from the probe output, and a d.c. input intended principally for low-frequency measurements on logic circuits. At low frequencies the d.c. input sensitivity is compatible with TTL logic levels. The a.c. trigger level control P1 sets the d.c. operating point of T3 and hence determines the input level at which it will turn on. With this control it is possible, when measuring complex waveforms, to trigger the frequency counter from either the fundamental or one of the harmonics as required.

Figure 3 shows the effect of varying the trigger level control. The upper trace of each oscillograph is an 18 MHz input signal with a high 3rd harmonic content. The lower trace in each case is the output of the preamp, with different settings of the trigger level control.

The graph of input sensitivity versus frequency for the preamp is given in figure 2. This shows the input voltage (r.m.s. mV) required for a 4 V peak-to-peak output swing. As can be seen from

** Frequencies below 20 kHz can be measured, provided the rise time is sufficiently short – less than 10 µs.*

3a



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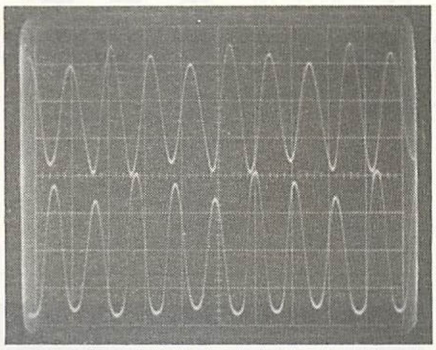

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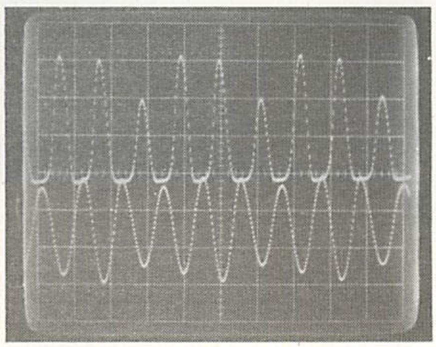

Figure 1. Circuit diagram of the preamplifier. P1 adjusts the triggering level.

Figure 2. Graph showing the preamplifier sensitivity as a fuction of the input signal frequency.

Figures 3a, b, c. Oscilloscope shows the effect of trigger level adjustment by control P1 (see text).

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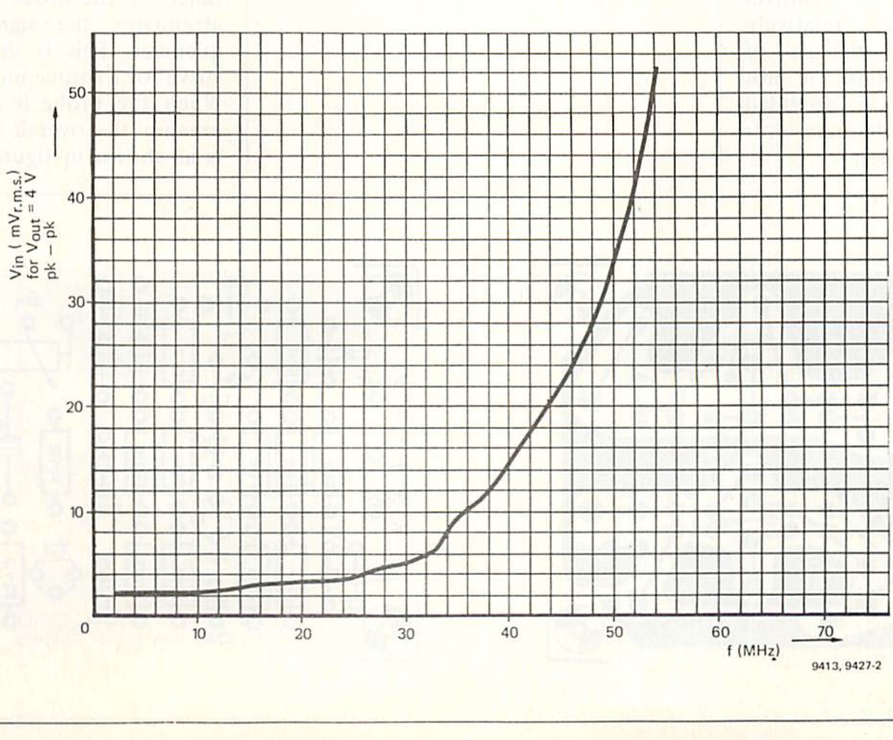

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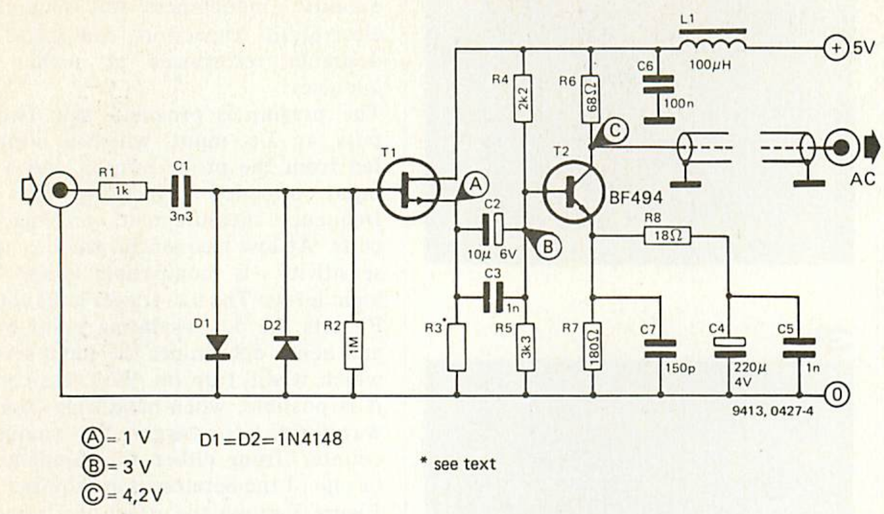


Parts List

Resistors
R1 = 1 k
R2 = 1 M
R3,R7 = 180 Ω
R4 = 2k2
R5 = 3k3
R6 = 68 Ω
R8 = 18 Ω

Capacitors
C1 = 3n3
C2 = 10 μ/6 V
C3,C5 = 1 n
C4 = 220 μ/4 V
C6 = 100 n
C7 = 150 p

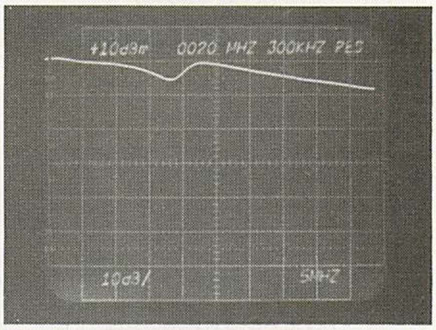
Semiconductors
D1,D2 = 1N4148
T1 = E300,BF245C
T2 = BF494,BF194,BF195

Inductors
L1 = choke 100 μH

the graph the required input voltage rises sharply above about 30 MHz, until at 45 MHz it is about 22 mV. Even so this is better than the original preamp design and is further improved by the addition of the probe, which also provides some gain.

Probe Circuit

To reduce the cost and simplify the circuit it was decided to use a single FET as the input stage instead of the dual FET used in the original circuit. T1 operates as a source follower to provide a high input impedance, with T2 providing a gain of about 2 and an output impedance of 68 Ω to drive the coaxial cable. Diodes D1 and D2 clamp the input voltage to ± 0.6 V maximum to protect the FET. The equalization network in the emitter of T2 helps to maintain a relatively flat frequency response, though of course a 'ruler-flat' response is not important in this application, provided the input sensitivity is adequate over the required frequency range.



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The FET used in this circuit is the tried and trusted Siliconix E300. Other FET's, such as the 2N5397, 2N5398, BF245C and BF256C may also be suitable. If an alternative FET is used it should be selected for a zero gate voltage drain current of at least 10 mA. R3 should then be selected to give a drain current of 3 to 5 mA with the device in the circuit.

Frequency Response

The gain of the probe circuit versus frequency is shown in figure 5 with the probe fed from a 50 Ω source. At the low frequency end the gain is about 2, and the response exhibits a slight rise up to about 60 MHz, after which it falls off. If the probe is fed from a high source impedance then the shunt capacitance of the probe impedance quickly attenuates the signal at higher frequencies. This is shown in the dashed curve for a source impedance of 10 k.

When the probe is combined with the preamp the overall frequency response is as shown in figure 8. At 1 MHz the

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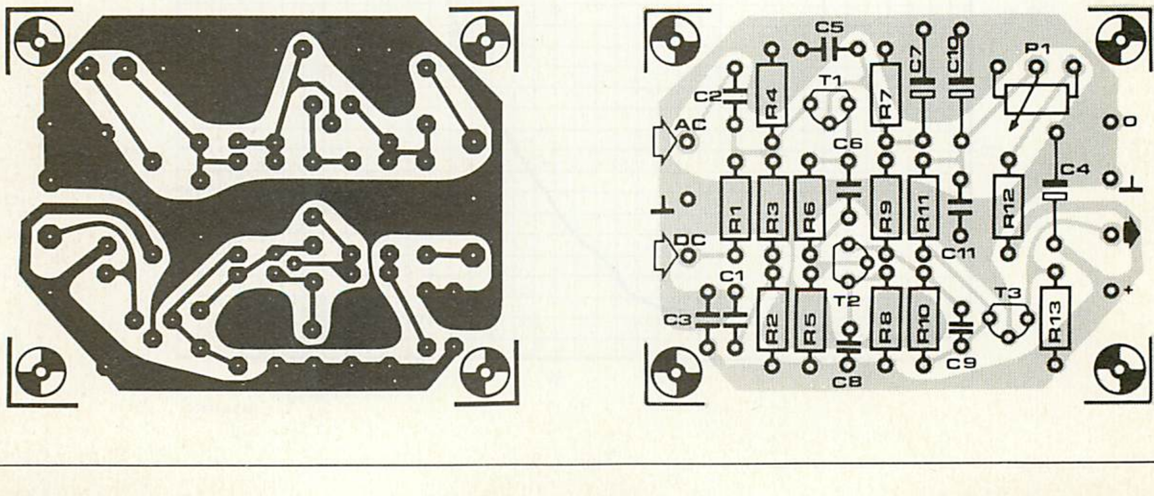

Figure 4. Circuit diagram of AC FET probe.

Figure 5. Fet probe gain as a function of frequency. The probe is terminated into 50 Ω. Spectrum analyzer display shows frequencies from 100 kHz to 50 MHz (left to right). The input signal to the probe was 0 dB. The gain of the probe by itself isn't of much importance, because its main job is as an impedance match.

Figure 6. Preamplifier p.c. board and component layout (EPS 9413).

Figure 7. FET probe p.c. board and component layout (EPS 9427).

Figure 8. Probe-plus-preamplifier sensitivity as a function of the input signal frequency.

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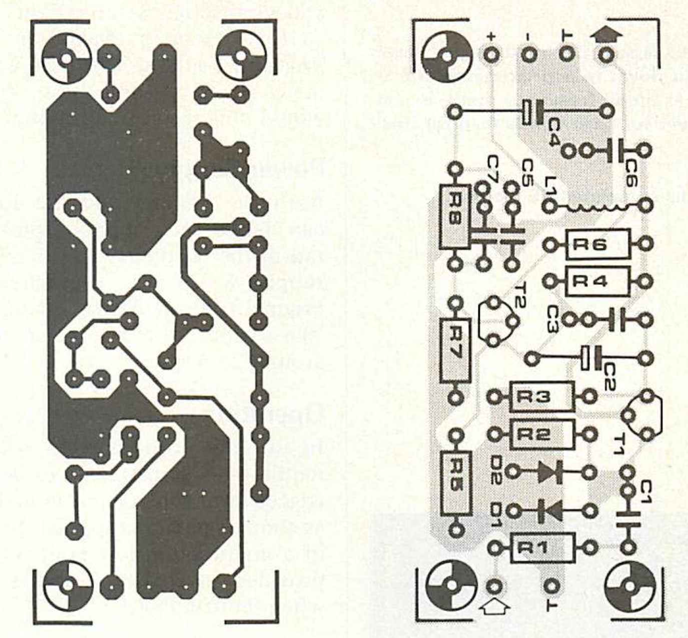

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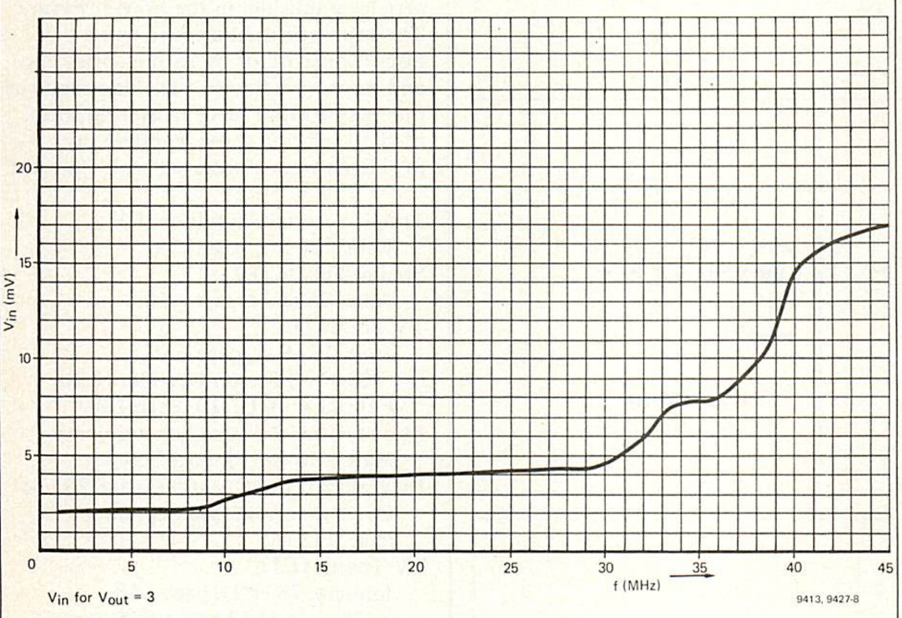

input sensitivity is around 2 mV, while at 45 MHz it is still only 17 mV, which is sufficient for many applications. Figure 9 shows an oscillograph of the preamp output to the probe at 40 MHz. The upper trace is the 40 MHz input signal (scale 20 mV/cm) whilst the lower trace is the preamp output (scale 1 V/cm). The timebase speed is 20 ns/cm.

Construction

A p.c. board and component layout for the preamp is given in figure 6, and for the probe in figure 7. Normal r.f. practice should be followed when mounting the components on the boards i.e. the component leads should be kept as short as possible, especially the transistor leads.

The preamp may be mounted in the frequency counter case, and is connected to the FET probe by a length of 50-75 Ω coaxial cable. The supply lead to the probe can be run alongside the cable. The housing for the probe is a matter of individual taste. The board is sufficiently small to fit in a small box folded from sheet aluminium. A test prod made from brass rod may be connected to the probe input through an insulating bush in the end of the box so that the probe may be used as a hand-held unit.

The ground connection to the probe input may be made with a crocodile clip on a flying lead. Alternatively the probe input connections may be made using 4 mm sockets, as shown in figure 11. Short flying leads ter-

Figure 9. Probe-plus-preamplifier response at an input signal frequency of 40 MHz (see text).

Figure 10. Excessive DC input signal levels should be cut down to approximately 0.5 V, since the maximum frequency limit is considerably lower for large signals than for small signal levels.

Figure 11. The completed FET probe.

minating in test prods or crocodile clips may be then be plugged into these and connected to the circuit under test so that the user's hands are left free. The unit could also be used with a prod made from 4 mm brass rod which would plug into the signal input socket.

Power Supply

Both the preamp and the FET probe can derive their supply from the +5 V rail in the frequency counter. The –5 V supply which was necessary with the original preamp design is not required. The total current consumption is around 25 mA.

Operation

In use the probe/preamp combination requires no adjustments except for the trigger level control and should function as soon as power is applied. In the event of a malfunction test point voltages are provided in figures 1 and 4 as an aid when faultfinding.

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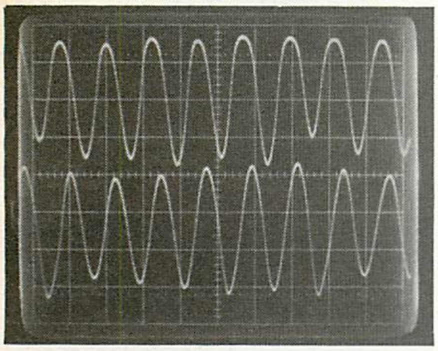

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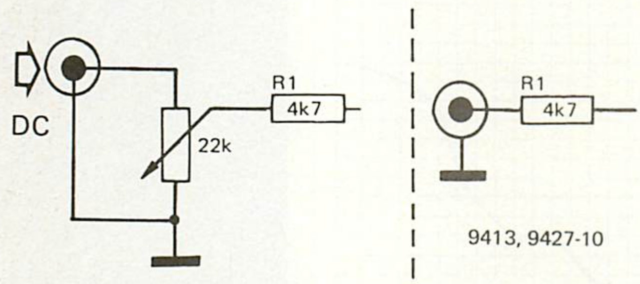

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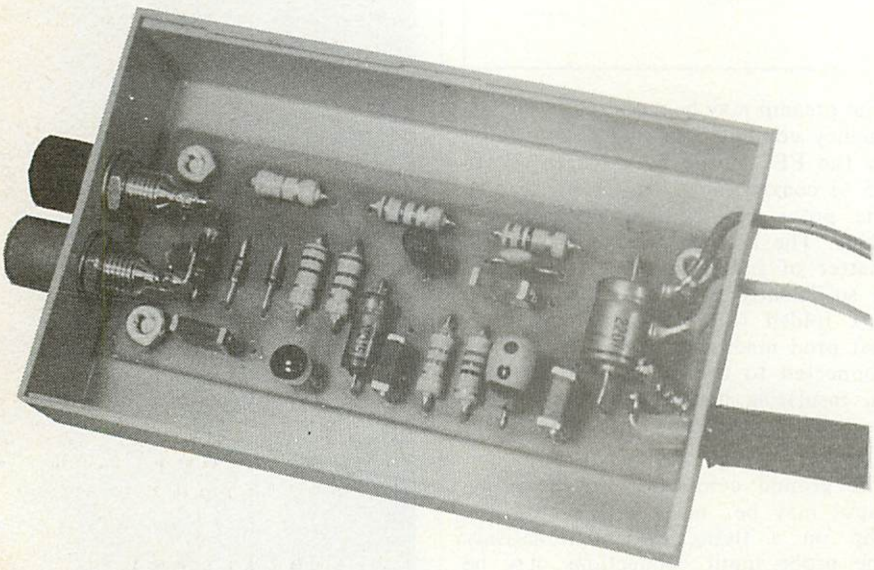


THE MISSING LINK

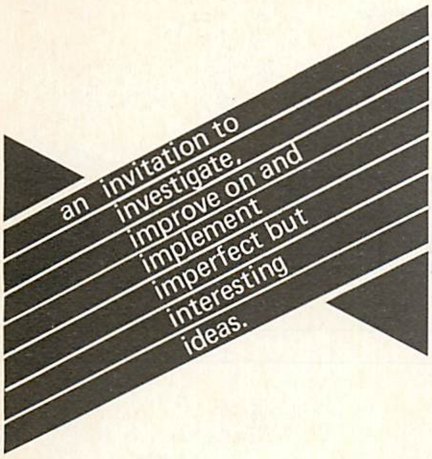
LINK 75

Cumulative index of 'Missing Links'.

The Link will appear each year in the June issue of Elektor. It contains an index to all Missing Links concerning articles published in the previous year. The intent of the link is to assist the home constructor by listing corrections and improvements to Elektor circuits in one easy to find place. A simple check of the Link will show whether any problems were associated with a project.

Missing Links concerning articles in volume 2:

EJEKTOR



These pages offer our design staff – and, we hope, our readers! – a long wished for opportunity to eject more-or-less wild ideas.
It often happens that promising ideas cannot be converted into practical circuits. The problem may be lack of specialised knowledge, lack of equipment or even simply lack of time.
Several examples of this kind of thing are still floating around the Elektor laboratories, such as a spot-sinewave generator and an OTA-gyrator. For the time being, development of these projects was stopped after unexpected technical problems arose. We simply cannot afford to spend any more development time on these projects at the moment.
Sometimes projects are put on ice at an even earlier stage, especially when it is obvious from the start that development will cost a disproportionate amount of lab time. In this case the design may never get past the block diagram stage, or it may be developed bit by bit in the course of (several) years. An example of this type of thing is the one-line intercom. That basic idea dates back to 1971, but it is only now nearing completion. It is rather frustrating to see a recent Philips Press release describing a very similar arrangement developed at their research laboratories . . . . . .
This means that our design staff are regularly coming up with interesting ideas that are only published several years later, if at all. Our readers also regularly submit circuits that contain an interesting idea, but are not (quite) suitable for publication because of technical imperfections. Usually our editorial staff can add the final touches, but this, too, costs development time and manpower – which is not always available.
Somehow, we want to eject these ideas. Somebody may be able to use them or carry on where the designer stuck.
From now on, interesting ideas which can not (yet) be implemented in practical circuits may be published in 'Ejektor'. It is not the intention to use these pages for publishing 'dud' circuits; on the contrary, the intention is to publish interesting ideas. This may, of course, include circuits that look as if they should work but don't. Also, some of the ideas may well prove completely impracticable on fundamental theoretical grounds. If so, we hope that the reader who discovers this will let us know . . .
Our editorial and design staff are quite enthousiastic about the new opportunities offered. It should also be quite a challenge to our readers.
We hope to be able at a later date to publish practical circuits based on the ideas presented here, after our readers or our design staff have found time to investigate them further.

OTA gyrator

Comparison of the basic gyrator formulae with the basic OTA formulae shows that the OTA should be an ideal active device in gyrator circuits.

Using an OTA gyrator it should be possible to construct a filter that can be swept through the whole audio band, while maintaining either constant bandwidth or constant Q. Such a filter could be used for spectrum analysis, electronic music (synthesiser!), equaliser, LF PLL, etc.

The basic principles of the gyrator were discussed in a previous article ('How to gyrate – and why', Elektor 2, p. 255). It was shown that when a gyrator is used to simulate a parallel LC tuned circuit (figure 1), the following formulae apply:

$$f_O = \frac{g}{2\pi C};$$

$$Q = \tfrac{1}{2} gR,$$

where: f_O = resonance frequency;
$g = g_1 = g_2$ = gyration constant;
Q = quality factor;
$C = C_1 = C_2$;
$R = R_1 = R_2$.

Note that both resonance frequency and quality factor are linear functions of the gyration constant (g).
The gyration constant is equal to the absolute value of the slope (or trans-

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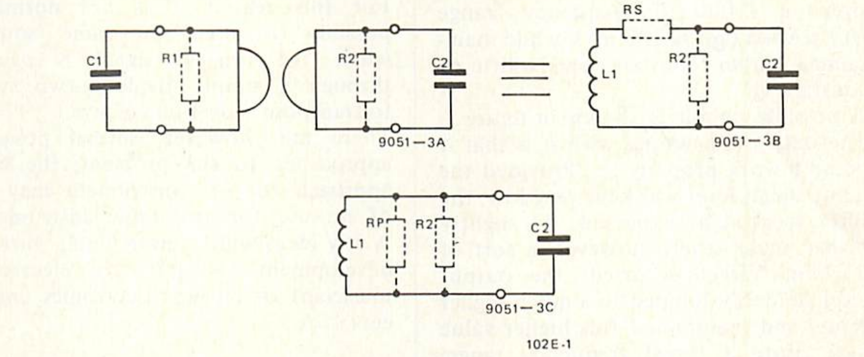

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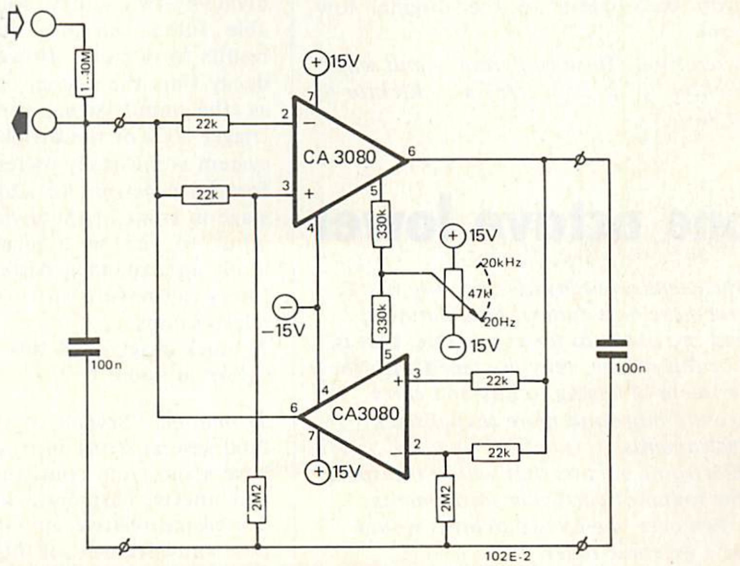

EJEKTOR

conductance) of the two amplifiers used in the gyrator, so for each amplifier:

$$i_{out} = \pm g.v_{in}$$

(the + sign for the non-inverting amplifier, the – sign for the inverting amplifier).
Compare this with the basic OTA formula:

$$i_{out} = \pm g_m.v_{in}.$$

The similarity is obvious!
A bias current sets the value of g_m. For a CA 3080, say, the transconductance equals:

$$g_m = 19.2 \times I_{ABC}.$$

This means that if OTAs are used in a gyrator circuit, the gyration constant is a linear function of the bias current (I_{ABC}).
From this it follows that the resonance frequency must also be a linear function of the bias current. The quality factor will also be proportional to the bias current, provided the input impedance of the OTAs is large compared to R. If, however, the OTA input impedance determines the value of R, the quality factor will be almost constant over the whole band.
This means that a simple DC adjustment will suffice to sweep such a filter over a 1000 : 1 frequency range ($0.1\ \mu A \leqslant I_{ABC} \leqslant 100\ \mu A$), while maintaining either constant bandwidth or constant Q!
A possible circuit is shown in figure 2. The only thing wrong with it is that it doesn't work properly . . . Provided the input signal level was kept very low, the filter worked as expected. At slightly higher signal levels, however, a sort of 'lock-on' effect occurred: the output level suddenly jumped to a much higher value, and maintained this higher value over quite a broad frequency range. Outside that range it would suddenly drop back again to the original low level.

Literature: 'How to gyrate – and why': Elektor 2, p. 255; 'OTA': Elektor 6, p. 927.

one octave lower

Musicians nowadays tend to use more and more bass guitars, bass clarinets, and the like – as far as possible, that is. For this reason, they are faced with the problem of having to buy and carry around more and more (expensive) instruments.
Electronic circuits that would transpose the sound of particular instruments down over one or two octaves would be a welcome relief.

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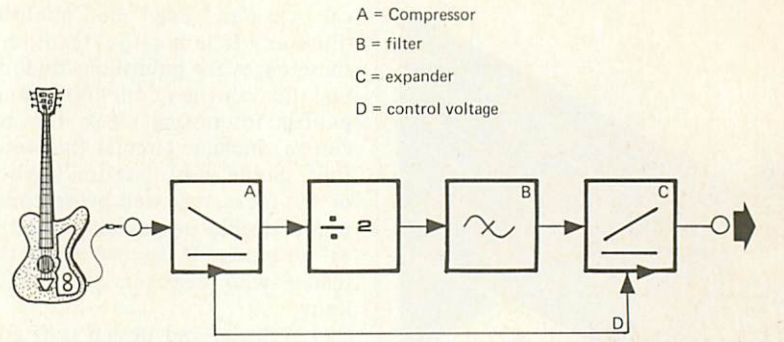

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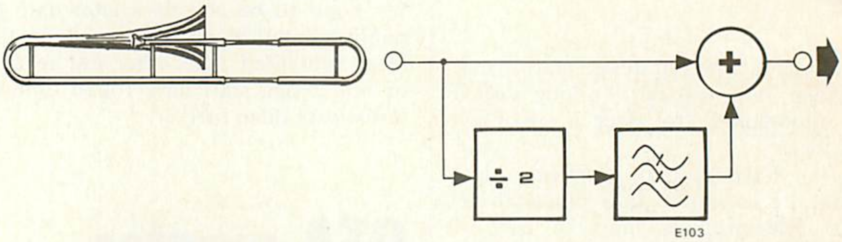

Several factors determine the 'sound' of a particular instrument: wave shape, attack, decay, non-harmonic sounds (e.g. wind noises), changes of harmonic content as a function of amplitude, etc. For this reason it is not normally possible to retain the same 'sound' when the original signal is passed through a simple divide-by-two stage to transpose it over one octave.
There are, however, several possible approaches to the problem; the best approach for one instrument may be of no use for any other instrument. A few ideas will be given here – further development is left to electronic musicians or musical electronics engineers . . . . .

Guitar. It has been found that a simple divide-by-two circuit followed by suitable filters can give quite reasonable results for a guitar. However, during the decay time the system tends to 'stutter' as the input signal drops below the trigger level of the divider, so this basic system is musically useless. It will therefore be necessary to add a compressor stage in front of the divider to keep the input to this at a relatively constant level; an expander before or after the filters can restore the original amplitude relationships.
A block diagram of this arrangement is shown in figure 1.

Trombone. Several brass instruments (and several string instruments as well!) have a spectrum consisting of both even and uneven harmonics with a gradually decreasing relative amplitude. To transpose an instrument of this type over one octave, it should be sufficient to add one 'sub-harmonic' to the original signal. This must be done in such a way that the original fundamental becomes the second harmonic of the new, added, fundamental – with the correct amplitude relationship.
An advantage of this system (sketched in figure 2) is that the original 'sound' is retained to a very large extent.

Clarinet. Several woodwinds, including the clarinet, have a spectrum consisting mainly of uneven harmonics with gradually decreasing relative amplitude. The even harmonics are at a much lower, and fairly constant, level of approximately –20 dB.
To transpose the sound of this type of instrument it should be possible to use the same basic system as that described for the trombone. The difference is that in this case the new 'sub-harmonic' fundamental must be one-third of the frequency of the original fundamental. This means that a divide-by-three stage will have to be used, and that the instrument will be transposed over one octave plus one quint . . . . this could be a nuisance!

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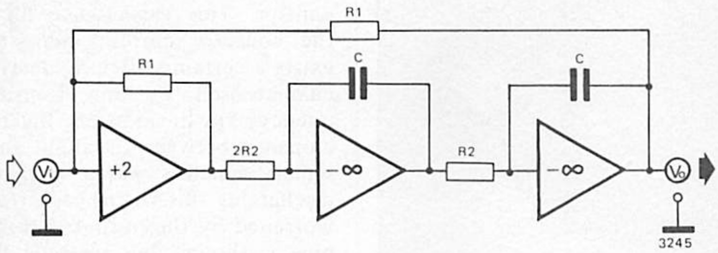


filter without phase-shift

A problem that occurs regularly in control systems using negative feedback is instability at high frequencies. If the system contains a non-minimum phase element, say, the total phase shift at high frequencies can become 360° while the total gain around the loop is still more than unity.
If a low-pass filter is added inside the loop to reduce the gain to a safe level at high frequencies, the law of conservation of misery dictates that the point where the phase shift is 360° will also move down to a lower frequency – where the gain is still more than unity, in spite of the additional filter!
What is needed is obviously either a completely new design, or else a filter that does not introduce phase shift in the frequency band that matters.

The basic principle of a filter of this kind was published in an earlier issue of Elektor: the 'Frequency dependent resistor' (Elektor 5).
The circuit is repeated here (figure 1). It should be made clear that this is only a basic block diagram; it is also assumed for the present that the amplifiers have infinite gain and zero phase shift.
In this case the total transfer function is:

$$\frac{v_o}{v_i} = (\frac{1}{j\omega\tau})^2$$

in which

$$\tau = R_2 \times C$$

The input current is therefore:

$$i_i = \frac{v_i - v_o}{R_1} + \frac{v_i - 2v_i}{R_1} =$$

$$= \frac{v_o}{R_1} = \frac{1}{R_1\,\omega^2\tau^2} \times v_i$$

which means that the input impedance is:

$$z_i = \frac{v_i}{i_i} = R_1\,\omega^2\tau^2.$$

This is a real resistance – with current and voltage in phase – but increasing with the square of the frequency.
If the output is left open, this frequency dependent input resistance can be used to construct a filter without phase shift – the limitation being the frequency where the amplifiers start to introduce phase shift. It's a fundamental law that there must be phase shift somewhere!
To give an example, if a signal is fed in via a resistor and the output is taken from the input of the frequency dependent resistor, the result is a high-pass filter without phase shift at the roll-off point.

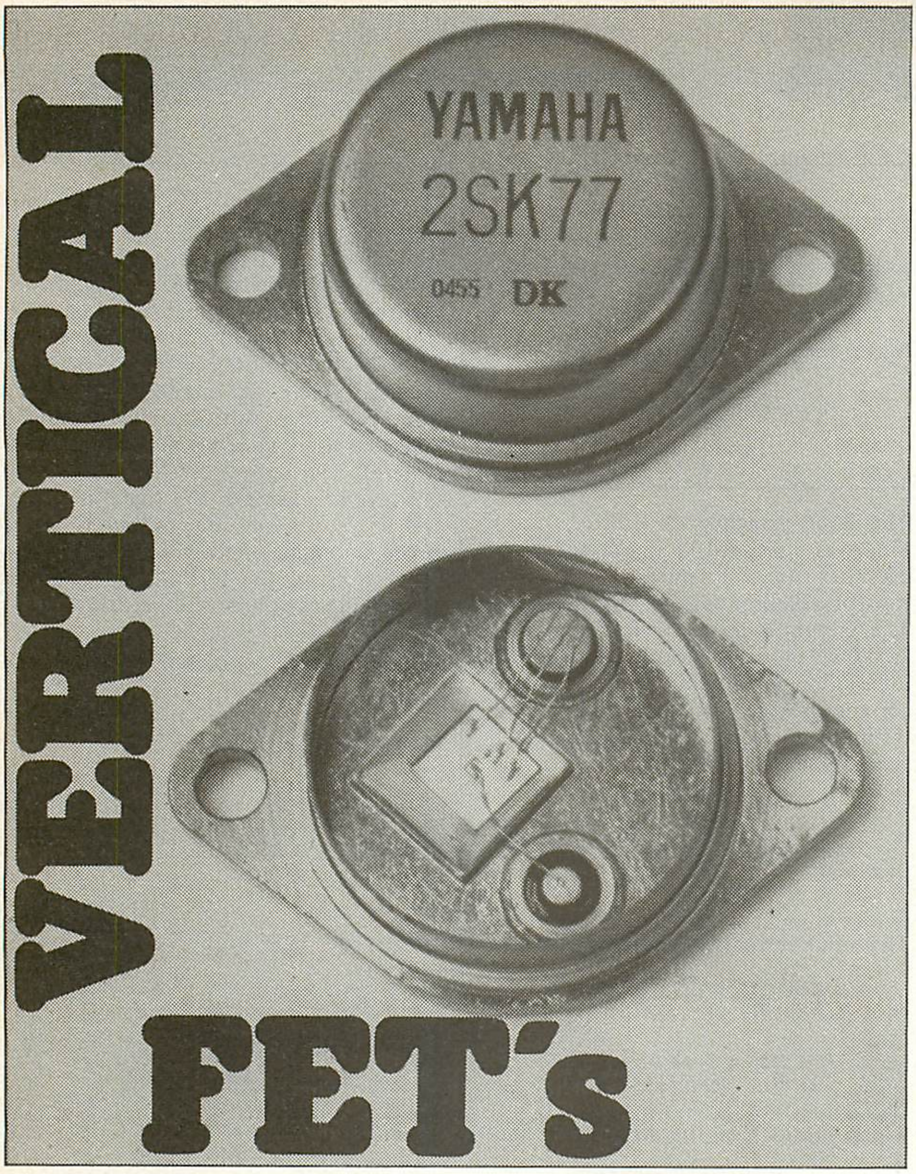
Literature: Elektor 5, p. 712.

COMING SOON

The next Elektor is the July/August 'Summer Circuits' issue. It contains over 100 projects and design ideas, from control units for solar heating panels to speech garblers.
Some circuits are basic design ideas, such as a monoflop using a single 7400. Others come with p.c. board layouts and are complete functional units, an example is a pulse generator with variable pulse width and repetition rate.
It should be made clear that this issue is not a review of circuits already published, nor is it a preview of designs that will be published in the coming year.

- dark room timer
- rain synthesiser
- car clock
- kettlestat
- current source
- battery indicator
- antenna amplifier
- logic tester
- wind machine
- min/max temperature indicator
- digital contrast
- power supply
- SSB adapter
- wideband frequency doubler
- headphone adapter
- pulse generator
- over 84 other circuits

VERTICAL FET's



There is always something new under the rising sun. The Japanese have now developed a semiconductor called 'vertical field effect transistor', intended for use in high power output stages. The V-FET performance and basic characteristics are vastly superior to the common bipolar transistors.
This article describes the V-FET's construction and operation, along with its application in commercial circuits.

Circuit Requirements

As unlikely as it may seem, a firm market seems to be developing for heavy, (i.e. 60-120 lbs) stereo output amplifiers capable of feeding some 150-350 watts to the 8 Ω load of each channel. Obviously, the quality of these 'audio power houses' is dependent on the characteristic properties of the active elements used in their output stages. Consider some of the difficulties which are associated with high output power. First, there is the problem of high heat dissipation in the final transistors. Then, the transistors themselves are subjected alternately to high collector potentials and currents, with the condition becoming more critical as the output power increases. It may even be necessary to arrange the output transistors in series-parallel, using the same principle as the coachman who replaces a single horse with a four-in-hand, so that the output is increased although the effort by each horse is less.
Further considerations for the output circuit designer are the switching properties of the transistors to be used. In the familiar class B final stage, the two halves alternate between a current-conducting and a current-blocking function which is, ideally, under the complete control of the signal from the driving amplifier.
However, this is hardly ever the case in practice, since power transistors are not capable of following fast input signal variations, due to the non-linear diffusion capacitance between the base and emitter. This capacitance increases as the collector current increases. There exists a certain switching delay which is characterised by the transition frequency, f_T. In addition, the phase discrepancy between the input and output signals, which is caused by charging and discharging this virtual capacitor, can be worsened by the output transistors clipping (although this clipping should be prevented in the preceding stage by limiting the driving swing to an amplitude that will not drive the final output signal against the power supply voltage). These transition imperfections cause a reduction in efficiency, i.e. an increase in the heat dissipated, so much so that, in some cases, it may be necessary to install a cooling fan.
High f_T transistors capable of following fast input signal variations at large amplitudes are more vulnerable than low f_T types under overload conditions, due to their construction and manufacturing technology. In particular, it is the second breakdown phenomenon, occurring at high collector voltages during high heat dissipation, that can lead to their complete destruction. This danger can be avoided by respecting the I_C-V_{CE} diagram that shows the Safe Operating Area (SOAR), which indicates the safe combinations of collector voltage and current. It can be seen that for high f_T types this area is considerably smaller than that for the more robust 2N3055 family, and good protective circuitry is badly needed.
Transistors respond inversely with the emitter current, i.e. as the emitter current rises, the response falls off. A water tap provides a suitable metaphor: between the order to turn off the water and the completion of the action, the water continues to flow, with the volume of the wasted water being dependent on the number of turns that the tap was turned on. Unlike electronic valves and FETs, the base of a bipolar transistor always draws some current, which can be of considerable magnitude in the case of high power transistors.
It follows from the above considerations that there is an urgent demand for an alternative high power active semiconductor with improved characteristics and higher safe ratings for heat dissipation, current and voltage. There are, admittedly, improved transistors such as the low emitter concentration (LEC) types, but at the moment they are only suitable for small signal applications. The demand for transistors for large signal applications still remains. However, it appears that it is in this region that the V-FET will be useful.

Horizontal FETs

Before discussing the new V-FET, it is worth mentioning some aspects of the

Figure 1. Conventional (horizontal) FET construction. The hatched area around the gate indicates the depletion zone caused by the non-conducting state of the pn junction. With a constant V_{GS}, an increase in the drain current causes the depletion zone to grow as indicated by the dotted line, until it impinges on the edge of the substrate. At this point further increases in the drain potential V_{DS} fail to cause an increase in the drain current which may be regarded as saturated.

Figure 2. Drain current I_D expressed as a function of V_{DS} with parameter V_{GS} for the FET of figure 1. The dotted line shows the knee potential as a function of V_{GS}.

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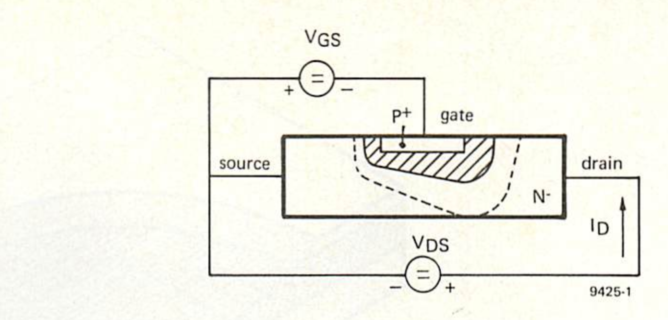

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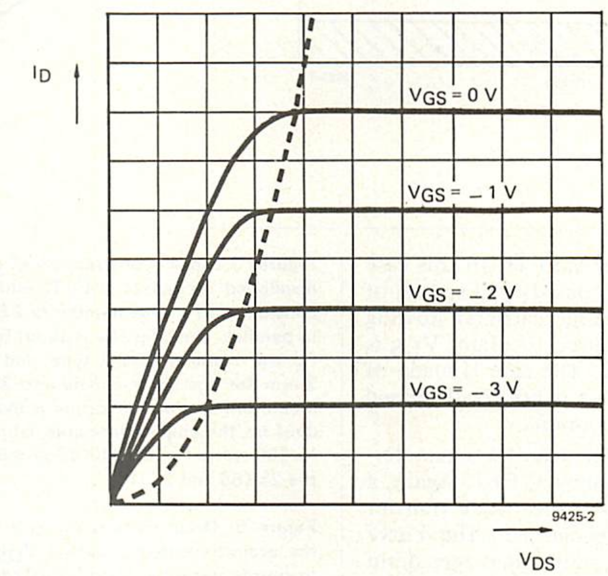

conventional (horizontal) FET, which is only suitable for low power applications such as circuits with high input impedance and low noise. Figure 1 shows the construction of an ordinary n-channel FET. A positive potential between drain and source causes electrons to flow from source to drain. The gate is made of p-type material, and when a negative potential with respect to the source is applied to it, the pn junction becomes non-conductive. Now the junction is surrounded by a depletion zone (hatched in the figure) which is completely empty of majority charge carriers. Figure 2 shows a family of output curves. For a given V_{GS}, the drain potential V_{DS} and the drain current I_D increase linearly at first. The current increase extends the depletion zone until the point at which the zone touches the opposite edge of the substrate (shown by the dotted line in figure 1). Any further increase in V_{DS} fails to increase I_D, so the FET virtually behaves as a constant current source for any higher values of V_{DS}. The output curves also show that more negative values of V_{GS} cause the 'knee' to occur at a lower V_{DS} so reducing the corresponding saturation current. The position of the knee for varying V_{GS} is shown by the dotted line.

Vertical FETs

Both Sony and Yamaha have now developed high power FETs whose properties are very promising. The construction and manufacturing technology of the Sony and Yamaha devices are similar in that both may be considered to be made of a large number of 'mini-FETs' working in parallel, but in other respects the two devices are quite different. Sony have developed both p-channel and n-channel V-FETs, whereas Yamaha have only developed an n-channel type. Consequently, the circuits utilising these new devices differ considerably in their design, as will be discussed later.

Figures 3 and 4 show the construction of the Yamaha n-channel device. Compared with the horizontal junction FET, the current flows vertically. The drain is located at the bottom of the crystal and its mechanical connection to the casing has a very low thermal resistance, which is vital since practically all the heat produced inside the device is developed in the drain and must be led away from there. The channel is made of N^- type material into which a grating of P^+ type material, the gate, has been embedded. In figure 4, each square represents a separate FET, at 5 to 10 micron spacings. The entire chip size is about 5 mm by 5.5 mm and it consists of tens of thousands of FETs, all working in parallel.

The output characteristics of these FETs are shown in figure 5, and a major difference from those in figure 2 is immediately obvious: there is no knee, or corresponding saturation voltage. Readers once familiar with the output curves of the old faithful triode will no doubt be struck by the resemblance, which has already been used in the publicity given to these new devices, as nostalgia is a great selling point. Those readers will remember the one-upmanship in the triode-fitted amplifiers against the pentode-equipped counterparts with their inherent current-saturation effects. However, the comparison is not really fair, as modern circuits with high negative feedback (made possible by the elimination of the output transformer) display hardly any distortion of this kind. On the other hand, it is much easier to trim an amplifier which is inherently devoid of clipping and other nasties, than one which is full of these distortions. V-FETs present other advantages than merely being free of these annoyances.

The family of output curves in figure 5 may also be used to describe the oper-

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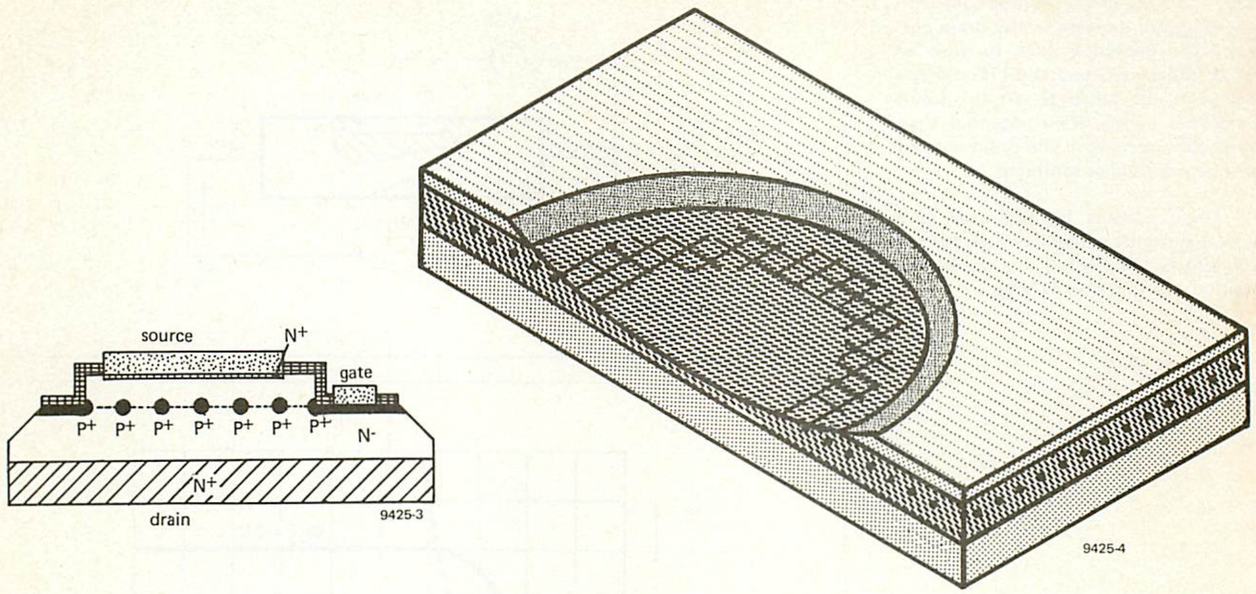

ation of p-channel V-FETs. In this case the drain current consists of a stream of holes (positive charge carriers) flowing vertically from source to drain. V_{DS} is therefore negative. The gate is made of n-type material and is positively biased with respect to the drain.

Figure 6 shows the transfer characteristics for an n-channel V-FET. Again, a close resemblance to the triode transfer curve may be recognised. The curve slopes much more gently near zero drain current than the steep exponential curve peculiar to the conventional bipolar transistor. This property of the V-FET is favourable for rounding the cross-over point in class B power stages. It must be admitted that the slope of the transfer characteristic can be adversely affected by fluctuations in the supply voltage or in the drain-source potential (such as would be caused by a drain load impedance). This is due to the negative feedback acting on the drain potential. The equivalent effect with thermionic triodes is the difference in transfer characteristics between the dynamic and the no-load output.

Performance Figures

The data sheet gives some interesting figures for different types of V-FETs. Particularly impressive are the maximum ratings for the Yamaha 2SK77 type, they are not likely to be equalled by any conventional power transistor. All types permit a very high maximum drain-to-gate potential, which is the highest potential found in a V-FET. All types are completely free from secondary breakdown effects, since the density of the drain current is the same throughout the channel. This is due to the absence of current crowding and also to the manufacturing technology which permits an extremely low level of contamination in the n- (or p-) channel. Although the V-FET is basically a voltage-controlled device, the gate does, nevertheless, draw some current, as the input impedance is not quite so high as in the horizontal FET. This input current consists of an inherent leakage current through the barrier between gate and channel, and a capacitive current caused by the charge and discharge of the source-gate capacitance and the virtual capacitance due to the Miller effect on the gate. Consequently, the currents required to drive the 2SK77 are so high that a source-follower driving stage is needed (see figures 7 and 8), for which the type 2SK75 has been developed (see table). In spite of this, V-FETs offer some important advantages over conventional power transistors. For one thing, the input capacitance is smaller and almost independent of the drain current. For another, the transition speed of all these V-FETs is 5 to 10 times faster and the power switched at these frequencies is 2 or 3 times higher than for the fastest bipolar transistor. High frequency distortion at the cross-over point is practically non-existent, especially with optimum setting of the quiescent current in the output stage.

An exclusive and very recommendable V-FET property is the negative temperature coefficient of the drain current, i.e. the current decreases as the crystal temperature increases. There is, therefore, no risk of thermal runaway in class B power stages in contrast to circuits with conventional transistors where, if there is insufficient thermal stabilisation of the standing current, the current rises as the temperature does, the temperature rises as the current does, which ever-increasing circle is only terminated when the transistors give up the ghost. Thanks to the absence of this cumulative effect, V-FETs need no preventative measures against thermal runaway.

Figures 3 and 4. Construction of the Yamaha developed n-channel V-FET, which may be considered as a large number of FETs working in parallel. The chip size is about 5 by 5.5 mm for the Yamaha 2SK77 type, and about 3 by 3 mm for the Sony 2SK60 and 2SJ18 types. Maximum dissipation rating is mainly dependent on the chip surface area, which gives rise to 200 watts for the 2SK77 and 63 watts for the 2SK60 and 2SJ18.

Figure 5. Drain current I_D as a function of the drain-to-source potential V_{DS} with gate-to-source potential V_{GS} as parameter. These output characteristics for the 2SK77 V-FET show similarity to thermionic triode characteristics.

Figure 6. Transfer characteristics for the 2SK77 V-FET: drain current I_D as a function of gate-to-source potential V_{GS} with drain-to-source potential V_{DS} as parameter. Any load in the drain circuit causes the slope of the dynamic transfer characteristic to drop, with the exception of the curved portion near the cut-off point. These characteristics show that the optimum quiescent current (i.e. a working point where the slope has half the gradient of the full swing slope for a class B stage) should be about 400 mA.

Figures 7 and 8. Stripped version (figure 7) and complete circuit diagram for one stereo channel of the Yamaha B-I output amplifier. This circuit clearly resembles the direct-coupled output circuit for thermionic valves and 800 Ω loudspeakers.

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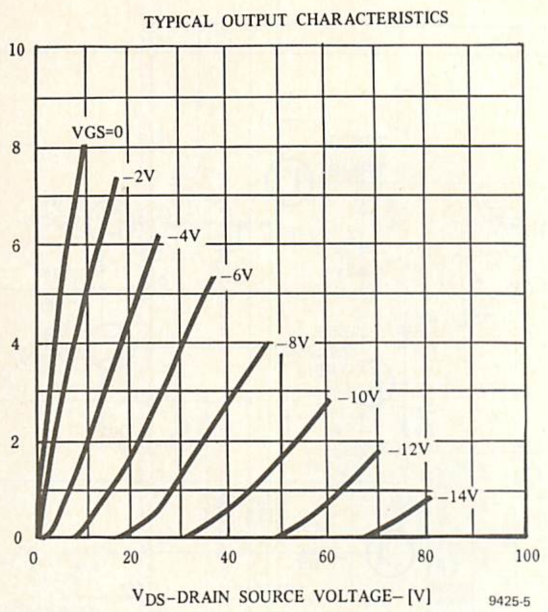

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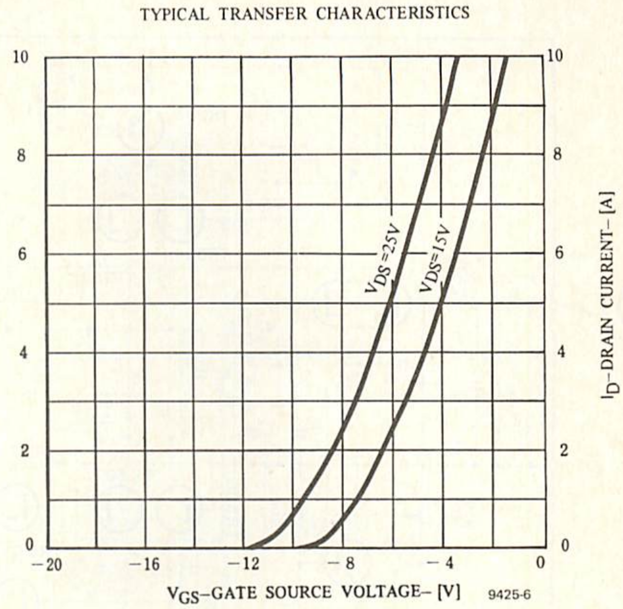

V-FET Circuitry in Practice

Two commercial circuits featuring V-FETs are discussed here, namely the Yamaha B-I type, which is a separate final stage, and the Sony TA 8650, which is an integrated amplifier containing both the pre-amplifier and the output stage. The discussion is confined to the final stages; power supply, filter and protective circuitry are not treated (the Yamaha circuit has in total 39 FETs, 113 transistors, 3 LEDs, 64 diodes and 7 zeners!). Both circuits feature class B output amplifiers, and the dynamic transfer characteristics involve a relatively high quiescent current (400 mA for both). This, in combination with the high supply voltage, results in fairly high quiescent heat dissipation (64 watts per channel in the B-I) in the final stage, but this is no problem for these V-FETs. The temperature protection in the B-I was intended originally to safeguard other components, for example the expensive computer-quality electrolytic buffer capacitors (rated at 80°C).

A condition not found in conventional transistor circuitry, and only revealed on close inspection of figure 5, is that the drain current will rise out of all proportion if power is supplied to the drain without sufficient bias (positive for n-channels, negative for p-channels) on the gate. Even for short durations, an excessive drain current of this magnitude will endanger not only the V-FET itself, but also the associated series resistors and probably the power unit as well. This condition only occurs at the moment of switching the power on or off. To prevent this, the circuit must be designed such that at switch on, the power is applied first to the pre-amplifying stages, so enabling the gate bias to build up before power is applied to the final stage; conversely, at switch off, the power is removed from the final stage before the pre-amplifier.

The requirement for a bias voltage of opposite polarity to the drain voltage calls for more than one stabilised (or unstabilised) power supply; if they were of the same polarity, some of the drain supply which is vital to obtain the full output voltage swing, would be diverted. To avoid this, the driver stage is supplied with a separate power supply.

Yamaha B-I Amplifier

This stereo output amplifier will continuously deliver 160 watts per channel into an 8 Ω load, with the harmonic and intermodulation distortions remaining well below 0.1%. Figure 8 gives the circuit diagram for one channel of the

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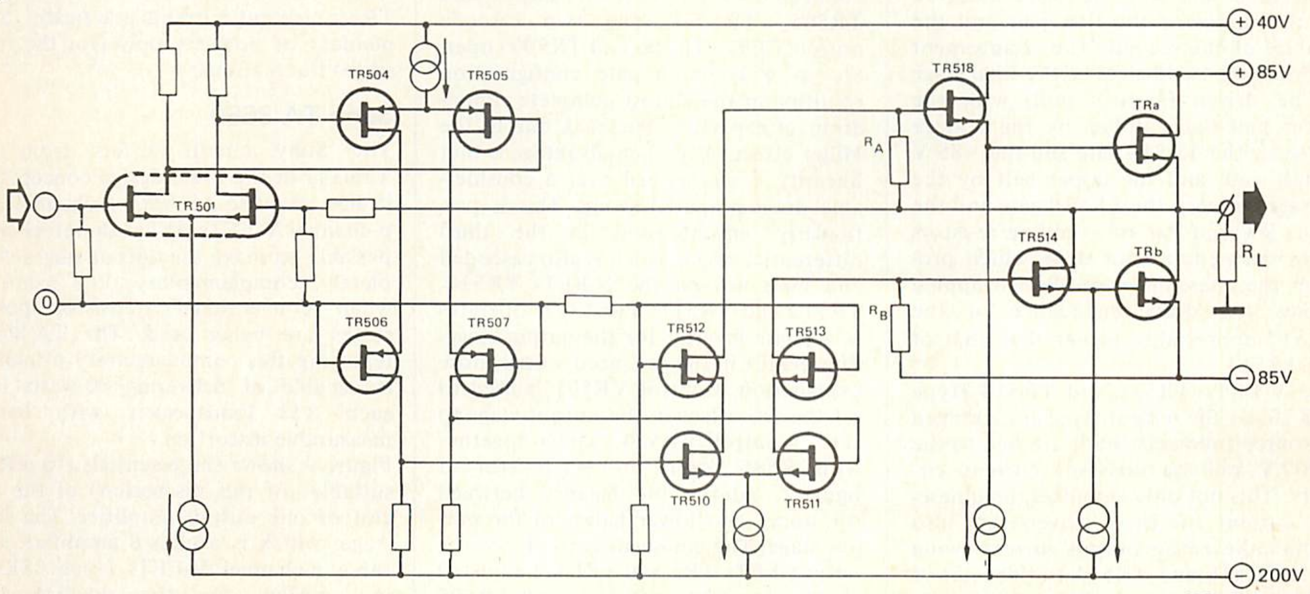

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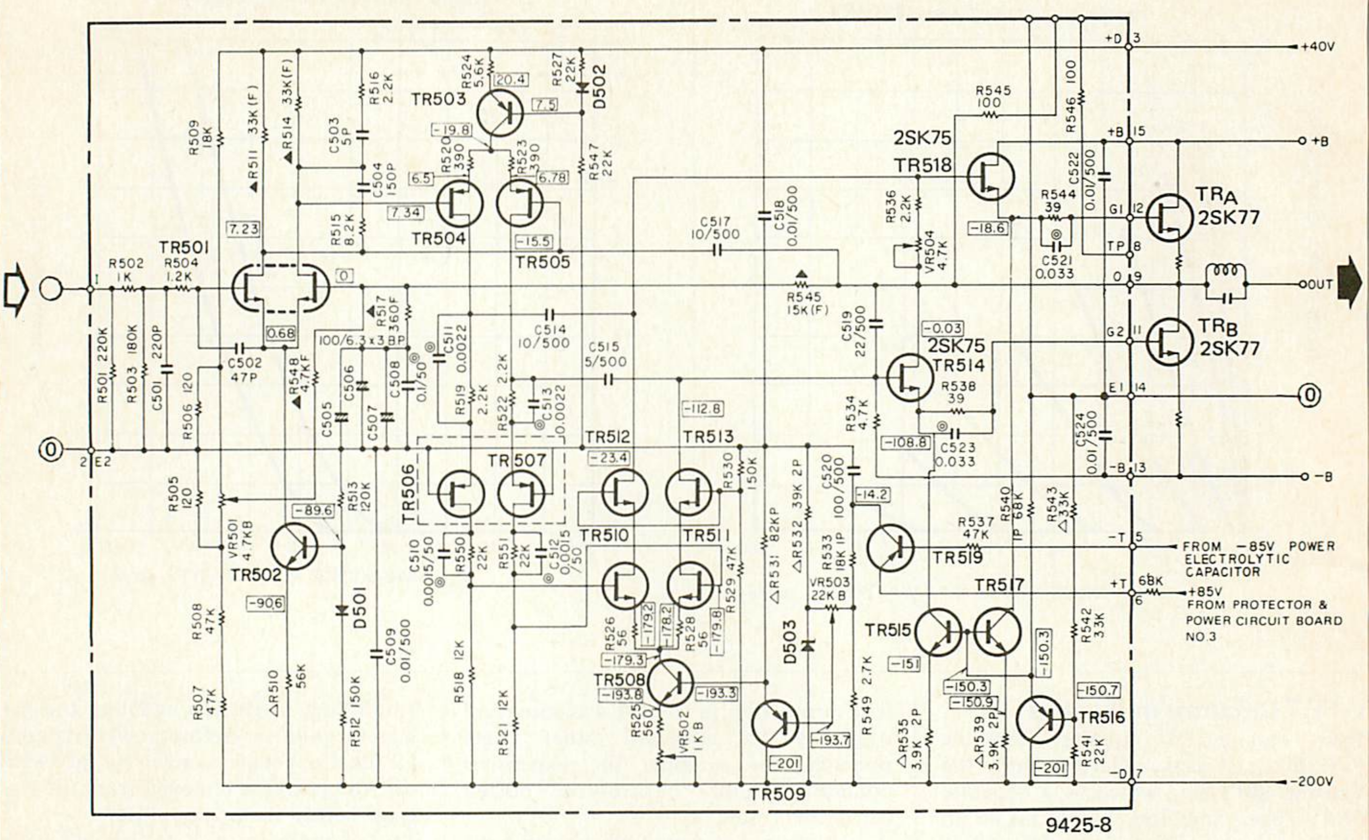

output stage, but the principles of operation can best be explained from the stripped diagram in figure 7. The basic design of the circuit may now be seen to bear some resemblance to the once-familiar, single-ended, push-pull power stage featuring two pentodes (EL 86) feeding an 800 Ω loudspeaker. This resemblance is due to the circuit being equipped in all signal handling stages with FETs and the single polarity n-channel V-FETs. In this Yamaha design, the power pair consists of two Darlingtons (TR518, TRa and TR514, TRb) series connected as far as DC is concerned, and with the load connected to the drains of the first pair and the sources of the second. The arrangement requires the two halves of the final stage to be driven in push pull, with the lower half being driven by the voltage between the TR514 gate and the –85 V supply rail, and the upper half by the voltage between the TR518 gate and the load. R_A and R_B are coupling resistors from the penultimate stage which provides the phase inversion. Bootstrapping makes the drain impedance of the TR512 appreciably higher than that of the TR513.

The V-FETs TR518 and TR514 (type 2SK75) in the output stage are arranged as source followers; both are tied to the –200 V rail via constant current circuits. This not only stabilises the quiescent current for these drivers, but also applies the entire output current swing from TR518 and TR514 to the gates of TRa and TRb respectively.

The B-I pre-amplifier stages are composed of three differential amplifiers in cascade with long tail pairs consisting of constant current circuits. The first stage is an n-type dual FET TR501. The input signal is applied to the left hand gate while the right hand gate receives a fraction of the output signal, so providing negative feedback (since the input and output are in antiphase). A dual FET with both elements on the same chip is necessary here since any difference in the DC characteristics of the two would produce an offset voltage at the output. Each half of the second differential stage consists of two cascode connected P-FETs (TR504, TR506 and TR505, TR507). The two cascode output FETs (TR506 and TR507) operate in a common gate configuration resulting in the almost complete elimination of capacitive feedback due to the Miller effect. A further advantage is that linearity is guaranteed over a considerable drain potential swing. This is particularly advantageous in the third differential stage, which is also cascoded and built around the N-FETs TR510, TR512 and TR511, TR513. It operates as a phase inverter for the output stage.

The details in figure 8 need a little more explanation. Control VR501 is used to set the DC offset in the output stage to zero. Control VR504, which together with R536 forms the R_A resistor of figure 7, adjusts the balance between the upper and lower halves of the output stage. The quiescent currents for the output FETs TRa and TRb are adjusted via the constant current circuit TR508 in the third differential stage, whereas the quiescent current for the drivers TR518 and TR514 is stabilised by the constant current circuits TR515 and TR517. Variation of the direct currents through TR510, TR512 and TR511, TR513 causes the TRa and TRb bias to vary, and with it the drain currents. TR508 is fed by a portion of the potential across zener diode D503, and the controls VR502 and VR503 are used to set the quiescent current. The base and emitter of TR519 are strapped via series resistors between the positive and negative power rails (±85 V): the collector is connected to the quiescent current circuit. This circuit makes the TRa and TRb quiescent currents practically independent of power supply (for the final stage) fluctuations.

Sony TA 8650

The Sony circuit differs from the Yamaha in the basic design concepts, as it has recourse to both n-channel and p-channel V-FETs. It has therefore been possible to make the output stages completely complementary, the same as when conventional transistor power stages are being used. The TA 8650, featuring this complementary principle, is capable of delivering 80 watts into each 8 Ω loudspeaker with barely measurable distortion.

Figure 9 shows the essentials (to a level suitable for this discussion) of the circuit of one output amplifier. The final stage, which is a class B amplifier, uses three n-channel V-FETs (type 2SK60) in parallel for the positive half (T717 . . . T719 in the figure), and three

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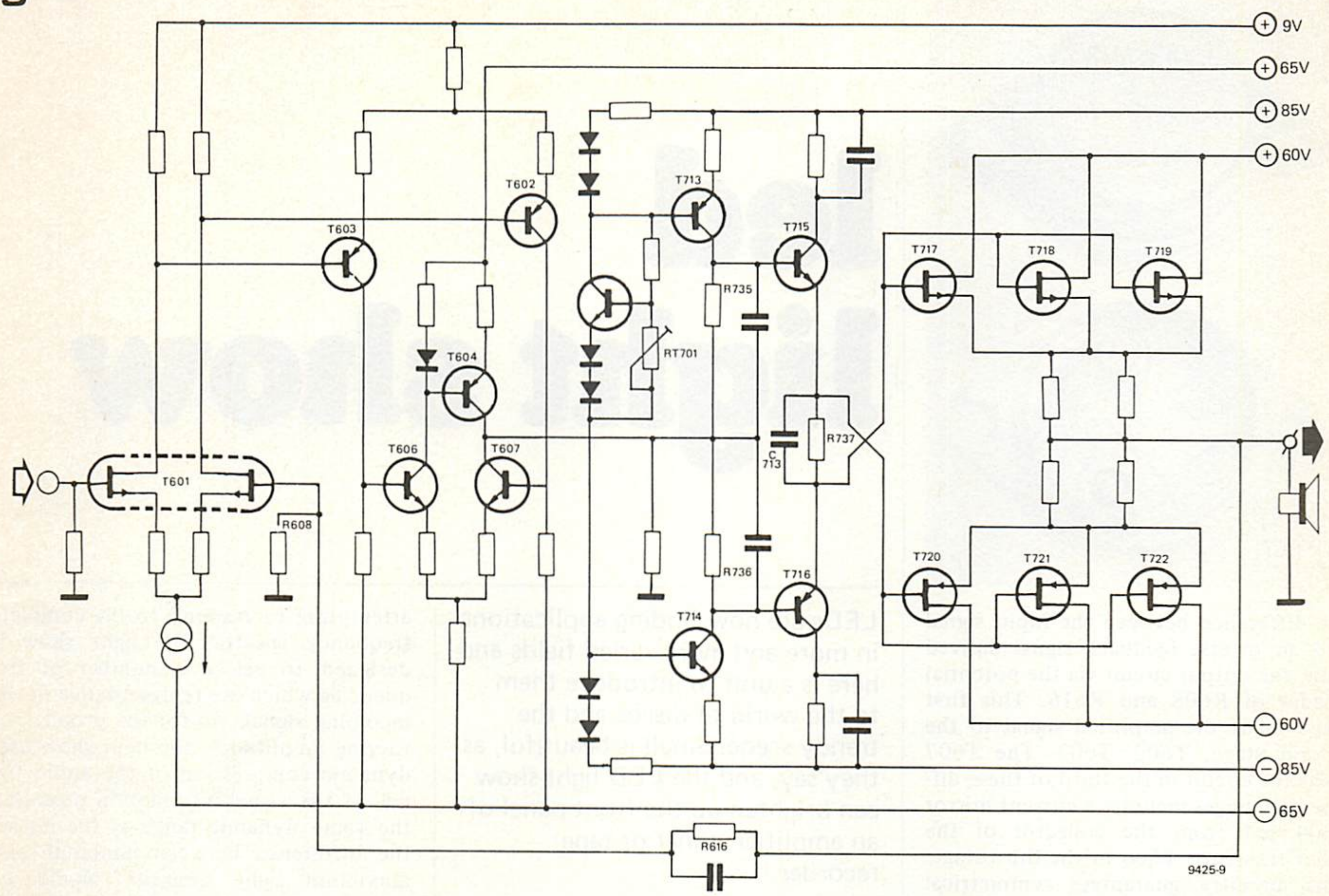
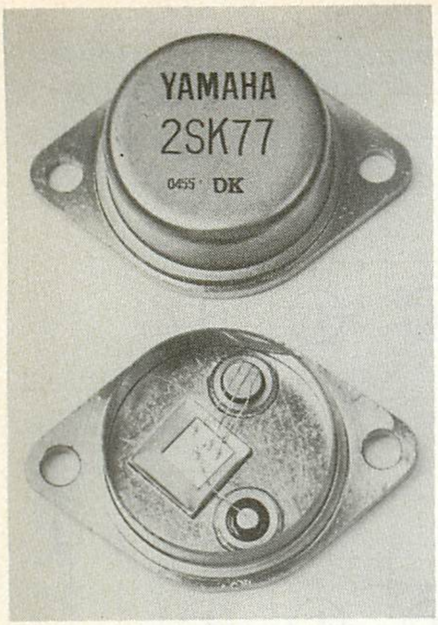


Figure 9. Simplified circuit diagram of one channel of the output stage in the Sony TA 8650 integrated amplifier. This stage is a complementary source follower with an inherently low internal impedance, even without negative feedback. The six output FETs must be selected to have a very small spread in the transfer characteristics between them, since the current distribution in the three parallel outputs in each half and in the two halves themselves must be equal.

p-channel (type 2SJ18, T720 . . . T722 in the figure) for the negative half. In this configuration they form a complementary source follower feeding the common load. The output stage is powered by ±60 volts. The parallel connection enables each half of the output stage to dissipate about 190 watts.
The V-FETs in the positive half are biased negatively, while the negative half has a positive bias. In the no signal state, the potential at the junction of the sources is zero. For this reason the respective gates are connected crosswise across resistor R737. The necessity of providing these gate-to-source bias potentials calls for a voltage for the penultimate stage (T711 . . . T716) exceeding that for the final stage. The required supply used in this case is ±85 volts.
The final stage is driven from a low impedance circuit: the emitter followers of T715 and T716. The sum of the drain-to-source bias potentials for the output FETs appears across the resistor R737. The emitter followers T715 and T716 are powered via the constant current circuits T713 and T714, which are in turn fed via a diode-resistor network from the ±85 V supply rails. Control RT701 is used to set the potential difference across R737 and thereby the quiescent current for the output stage. This complex circuitry eliminates the effect of power supply fluctuations upon this quiescent current.
N.B. The six output FETs are selected for accurate equality of $|V_{GS}|$ at a constant drain current, which is imperative for uniform current and dissipation distribution among the six power FETs. Unfortunately, this means that a failure of any FET in the final stage will mean replacing all six.
The driver stage T715, T716 is symmetrically controlled via resistors R735 and R736, by the pre-amplifier. The latter is composed of three differential stages, the first of which is designed as an n-type dual FET T601 operating on

Table. Characteristics of specific V-FETs		2SK75 n-channel Yamaha	2SK77 n-channel Yamaha	2SK60 n-channel Sony	2SJ18 p-channel Sony
maximum dissipation at 25°C	P_D	20 W	200 W	63 W	63 W
maximum crystal temperature	T_j	150°C	150°C	120°C	120°C
maximum drain-gate voltage	V_{DGO}	200 V	200 V	170 V	–170 V
maximum gate-source voltage	V_{GSO}	–30 V	–40 V	–30 V to –50 V	30 to 50 V
maximum drain current	I_D	0.5 A	20 A	5 A	–5 A
maximum gate current	I_G	10 mA	1 A	0.5 A	–0.5 A
gain	μ	40*	7.5**	4***	4***
slope	S	30 mA/V*	1.5 A/V**	0.25 A/V***	0.25 A/V***
internal resistance	R_i	1k3*	5**	16***	16***

* with V_{DS} = 80 V and I_D = 10 mA
** with V_{DS} = 30 V and I_D = 2 A
*** with $|V_{DS}|$ = 20 V and $|I_D|$ = 1 A



the difference between the input signal and an inverse feedback signal derived from the output circuit via the potential divider of R608 and R616. This first stage feeds the amplified signal to the second stage, T602, T603. The T607 collector circuit in the third of these differential stages includes a current mirror T604 fed from the collector of the other transistor T606 in the third stage. This circuitry guarantees symmetrical control of the driver and hence the output stage.

Conclusion

The object of this article has been to provide our readers with some insight into the properties and possible applications of these new beasts. The high quality of the commercial circuitry described goes without question, especially as regards the absence of cross-over distortion (although this has been achieved at the cost of a quiescent dissipation that equals the maximum output power of some small conventional power amplifiers!)

Despite the impressive figures the relationship between quality and price is not in all cases more favourable than that for equipment using conventional bipolar transistors with a few FETs thrown in. In this respect, it is worth noting that Sony offers two interesting alternatives in the shape of amplifiers TA 4650 and TA 5650, which are considerably lower in price and have an output power only 2 or 3 dB down.

led light show

LEDs are now finding applications in more and more varied fields and here is a unit to introduce them to the world of discos and the trendy scene. Small is beautiful, as they say, and the LED light show can brighten up the front panel of an amplifier, tuner or tape recorder.

The light show described here is fairly inexpensive and simple to build, and is distinguishable from conventional light shows, not only by its use of LEDs (light-emitting diodes), but also by its filtering system and the lack of a (hazardous) high voltage. It is intended as a decorative addition to an audio set-up by installing the LED light show in the front panel (say) of the amplifier or tuner, perhaps. This will give an attractive (even if somewhat miniaturised) visual rendering of the music; the display is enhanced by the use of three different colours.

Design philosophy

Consider the requirements of a light show: its task is to accentuate audio information with a visual display. The audio information will be from various sources (e.g. musical instruments), each of which has a characteristic frequency spectrum and amplitude. Rather than attempting to respond to the complete frequency spectra, the light show is designed to select a number of frequencies which are representative of the incoming signal. As for the problem of varying amplitudes, the light show uses dynamic compression of the audio signal. If the visual information possessed the same dynamic range as the music, the difference between minimum and maximum light intensity would be annoying. Another consideration is that visual selectivity is much less than acoustic selectivity, so the frequency bands chosen for visual representation are deliberately restricted. It is generally accepted that audio frequencies are divided into three groups for display: low, medium and high. However the boundaries which lie between these groups are not so well defined. Many applications leave overlaps between these groups, but in this case it has been found that leaving gaps gives a better solution (see figure 1).

Figure 2 is a block diagram of the light show. The compressor which is used to reduce the dynamic range of the audio signal can be of a much lower quality than one for use in an audio circuit. The only restriction is that the harmonic distortion produced by (say) the middle channel should not be visible on the high channel. The characteristic of the compressor should not be such as to amplify noise or hum which would then be displayed during quieter passages. Even so, the visual effect is still impaired by peaks, but since the response of the amplifiers A1 . . . A3 is unimportant, they are designed to limit on a fairly low input signal to mask nonlinearities in the compressor. The filters feeding A1 . . . A3 are low-pass, bandpass and high-pass, respectively. In the circuit described here, use is made of double-T filters. This facilitates the early-limiting design of the amplifiers. The double-T filters are of simple design and give good selectivity, but they have the disadvantage of being peaked. This can be remedied by damping the filters, bearing in mind the restriction, still, that the frequency bands must not overlap.

In designing the amplifiers A1 . . . A3 it must be remembered that the programme material upon which this type of device is most commonly used contains a predominance of the low frequencies. The gains of A1 . . . A3 must therefore be such as to overcome this disparity.

The point at which the light show is connected into the audio circuit can vary between a pre-amplifier and a 100 W output amplifier. The compressor should not only be resistant to the very high levels which may well occur, but must also continue to function normally when they do. In addition, the input impedance must be sufficiently high so that the circuit to which it is connected is not loaded by it.

In most light shows, the loads of the amplifiers A1 . . . A3 are driven by triacs or thyristors, either via an isolating transformer or an optocoupler. No relationship between light intensity and amplitude can be obtained, since the condition controlling the lamp state is binary, i.e. the lamp is on if the amplifier output is above the trigger voltage, and off if it is not. How the relationship is obtained in this particular design is explained shortly.

The Circuit

Figure 3 gives the circuit diagram for the light show. The compressor is formed by the circuit around IC1 (741), which is connected as a non-inverting amplifier whose gain may be varied from 20 dB to 60 dB by means of P1. When the emitter voltage of T1 becomes greater than +6.5 V D1 and D2 start conducting, thus reducing the input signal and hence the output signal until equilibrium is restored. Since the time constants in the control circuit are fairly large, it takes some time for equilibrium to be restored. Asymmetrical compressors require a fairly slow control rate because otherwise motorboating (low frequency oscillations) could easily occur.

The output of the compressor stage is fed to the three double-T filters, as shown. These filters have enough output current to drive the LEDs through 220 Ω resistors, allowing the relationship between amplitude and light intensity to be retained.

Construction and Operating

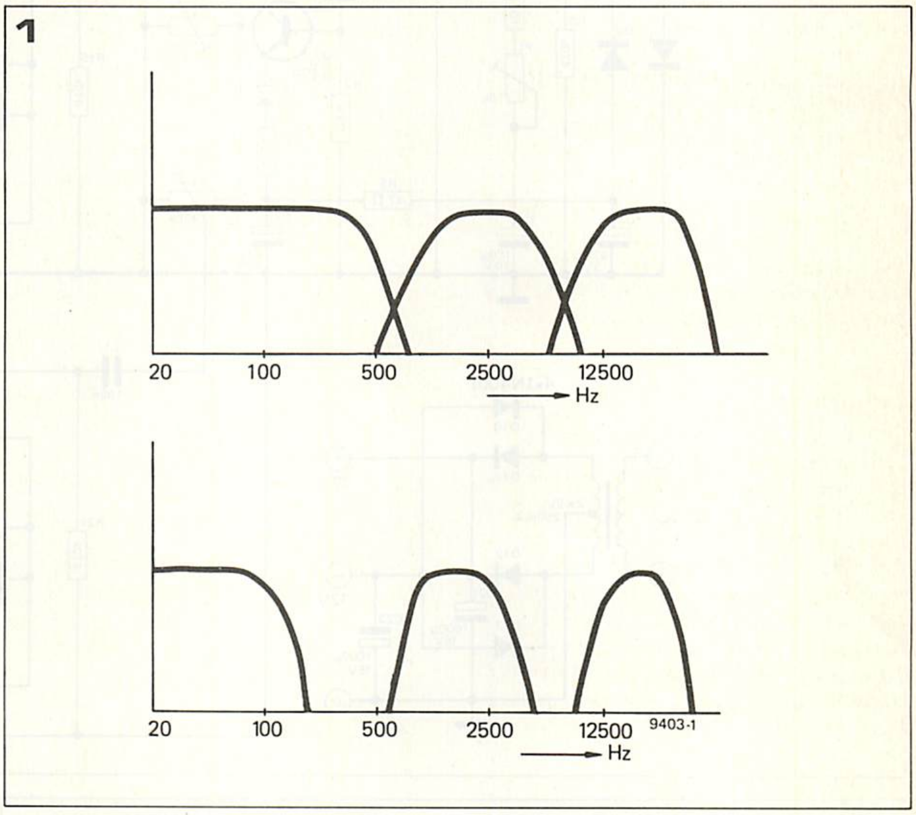
The circuit should preferably be given a metal housing, and a screened lead should be used between the input socket and the compressor input, with the screening connected to the housing at the input socket. The connections from the filters to the LEDs can be made with ordinary connecting wire.

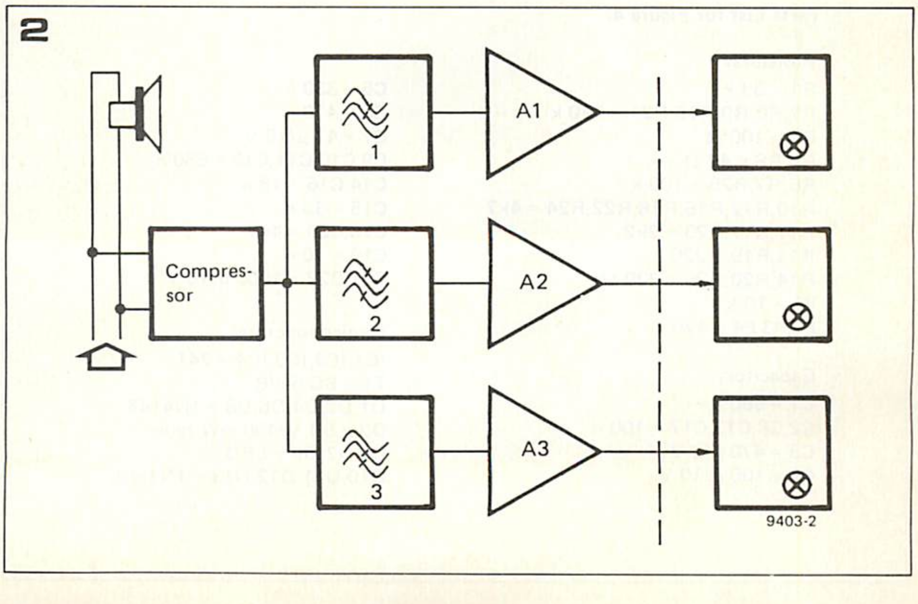
After construction and careful inspection of the completed circuit, the light show is set up in the following way: adjust P1 to give maximum gain (slider against R4) and then P2 . . . P4 (sliders against C5). Connect the compressor input to the loudspeaker or to the pre-amplifier output so that the audio signal is fed to the light show (it is better to avoid classical and choral music while setting up). Now adjust P1 . . . P4 so that the light display gives the best (subjective) match with the music. The setting is a matter of personal taste and it may be found that taste demands different settings for different types of music. If instabilities occur in any of the channels, a lower value must be selected for the corresponding resistor, R13, R19 or R25.

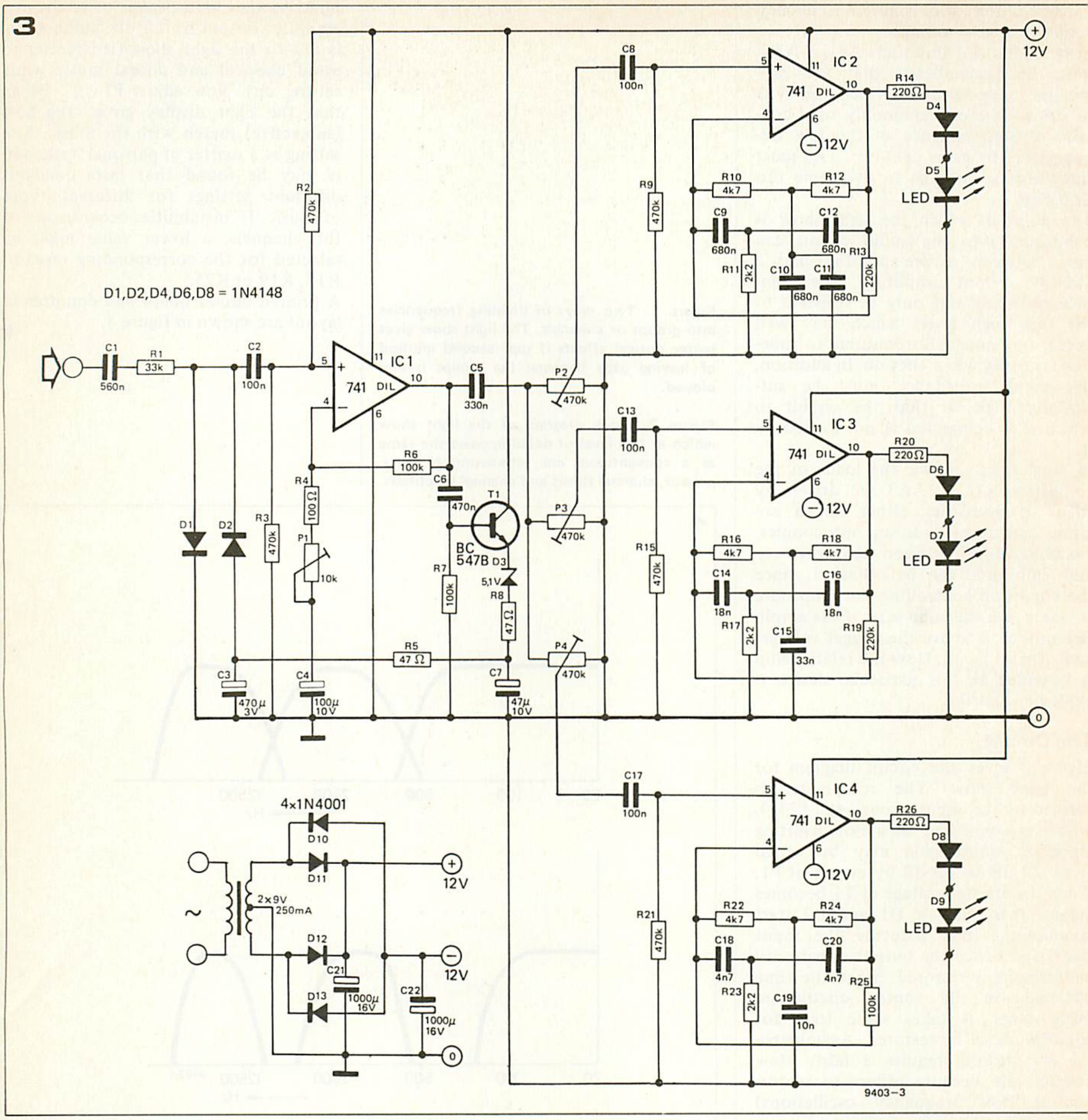
A printed circuit board and component layout are shown in figure 4.

Figure 1. Two ways of dividing frequencies into groups or channels. The light show gives better optical effects if the second method of leaving gaps between the groups is employed.

Figure 2. Block diagram of the light show which at this level of detail appears the same as a conventional one, consisting of compressor, channel filters and channel amplifiers.







Parts List for Figure 4.

Resistors:
R1 = 33 k
R2,R3,R9,R15,R21 = 470 k
R4 = 100 Ω
R5,R8 = 47 Ω
R6,R7,R25 = 100 k
R10,R12,R16,R18,R22,R24 = 4k7
R11,R17,R23 = 2k2
R13,R19 = 220 k
R14,R20,R26 = 220 Ω
P1 = 10 k
P2,P3,P4 = 470 k

Capacitors:
C1 = 560 n
C2,C8,C13,C17 = 100 n
C3 = 470 μ/3 V
C4 = 100 μ/10 V
C5 = 330 n
C6 = 470 n
C7 = 47 μ/10 V
C9,C10,C11,C12 = 680 n
C14,C16 = 18 n
C15 = 33 n
C18,C20 = 4n7
C19 = 10 n
C21,C22 = 1000 μ/16 V

Semiconductors:
IC1,IC2,IC3,IC4 = 741
T1 = BC 547B
D1,D2,D4,D6,D8 = 1N4148
D3 = 5.1 V/400 mW zener
D5,D7,D9 = LED
D10,D11,D12,D13 = 1N4001

Figure 3. The LED light show makes use of the cheap 741. Channel separation is achieved by means of damped double-T filters. One advantage of this filtering method is that when connected to an FM receiver the stereo pilot tone gives no trouble.

Figure 4. Printed circuit board (EPS 9403) and component layout for the light show circuit of figure 3. The printed circuit has been designed so that either TO-5 or dual-in-line packages may be used.

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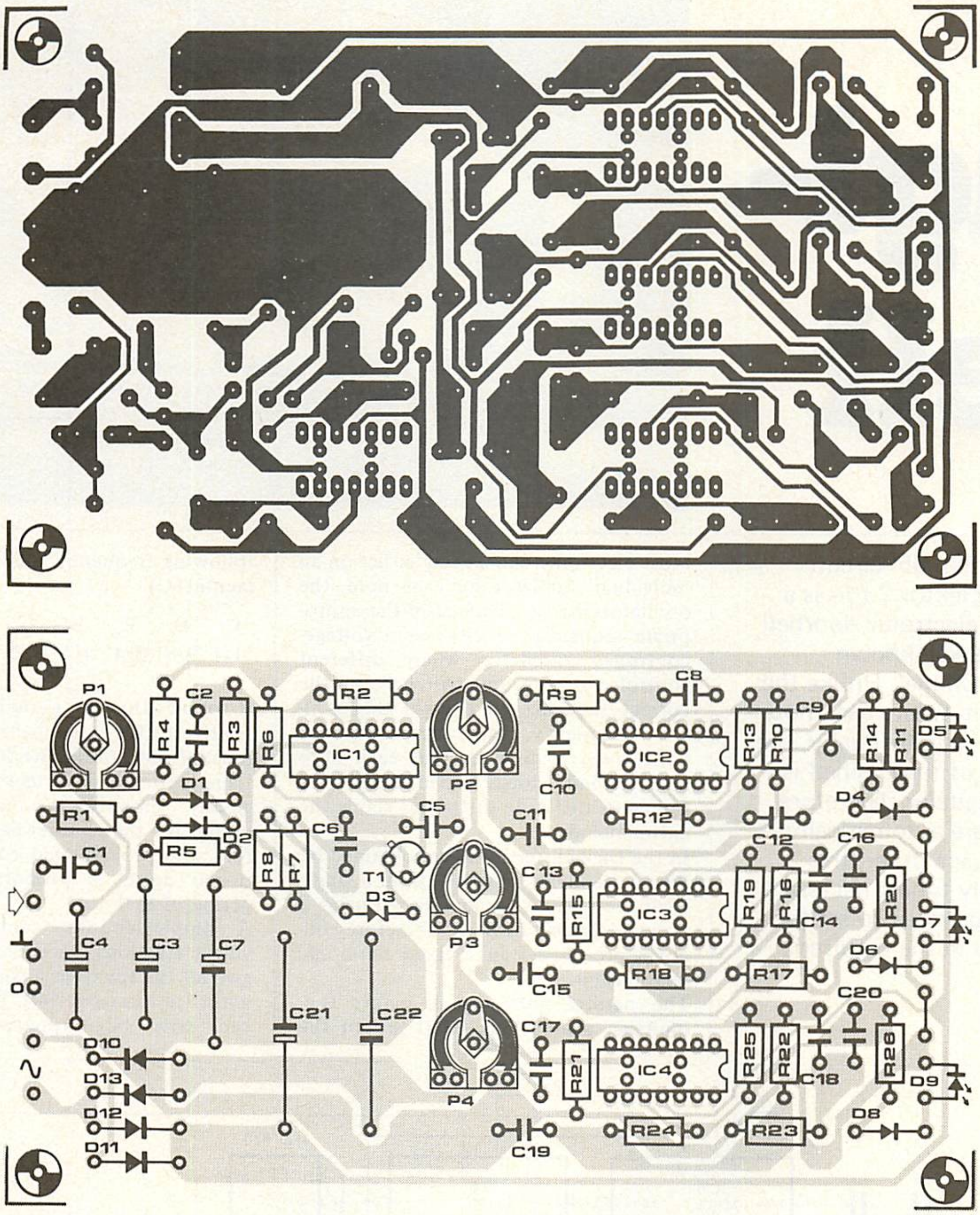
P1
R4
C2
R3
R6
R2
IC1
P2
R9
C8
IC2
R13
R10
C9
R14
R11
D5
C10
D4
D1
R1
D2
R5
C1
R8
R7
C6
C5
T1
D3
C11
R12
C12
C16
C13
P3
R15
IC3
R19
R16
C14
R20
D7
D6
C4
C3
C7
0
C15
R18
R17
C21
C22
C20
C17
P4
R21
IC4
R25
R22
C18
R26
D9
D10
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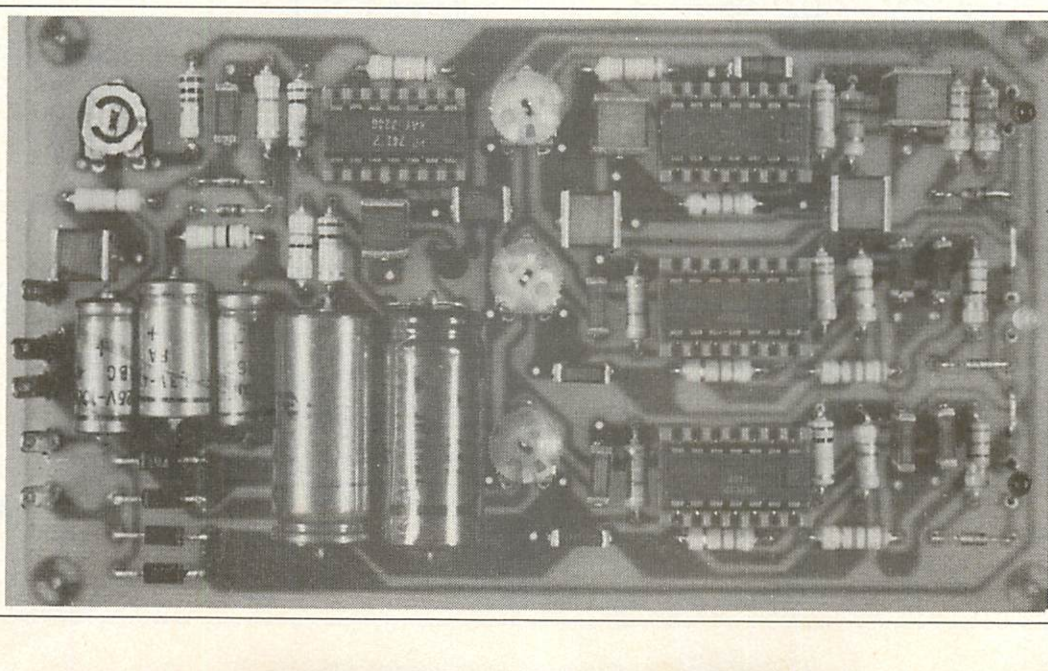
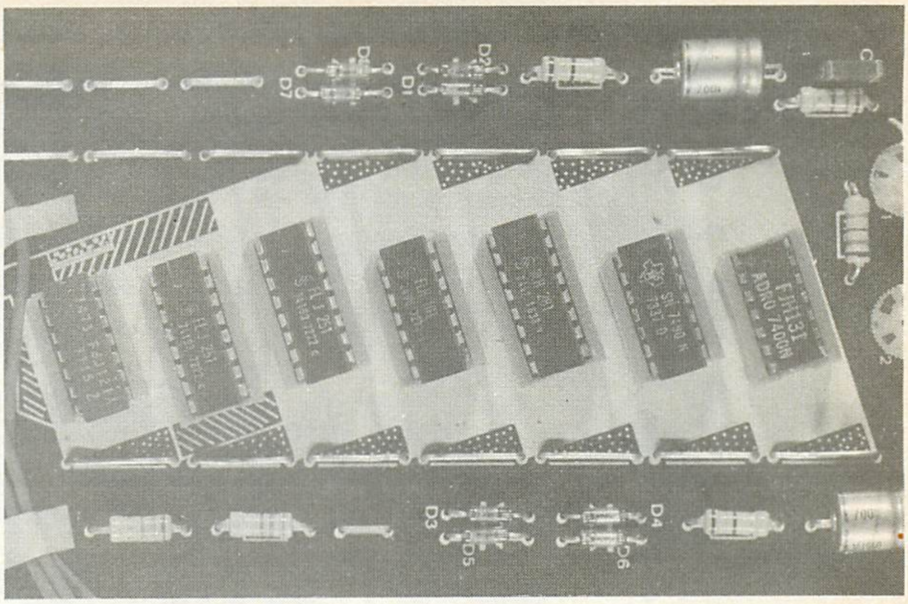
C19
R24
R23
D8



R. Janssen

digi-bell



Like the 'Big Ben 95' circuit published in Elektor 2 this is a design for an electronic doorbell that plays the well-known 'Westminster Chime'. Unlike the Big Ben circuit, where each note had to be tuned individually, the programming of the Digibell is carried out digitally. Therefore all the notes are automatically in tune with each other, this means the only tuning necessary is a single adjustment to set the entire melody in the required key.

Most electronic chimes rely either on an individual oscillator for each note, the oscillators being switched in the appropriate sequence, or else use a voltage-controlled oscillator where different control voltages are switched in the correct sequence to control the oscillator frequency. The disadvantage of both these systems is that each note must be individually tuned, and if the tuning of one note drifts this spoils the entire melody.

In the Digibell the notes are obtained by digital frequency division from a single frequency. This is accomplished using a programmable divider. Therefore the notes are always in a fixed harmonic relationship.

The natural notes in an octave (i.e. omitting sharps and flats) are in the following frequency ratios to the fundamental (C):

C	D	E	F	G	A	B	c
1/1	9/8	5/4	4/3	3/2	5/3	15/8	2/1

It follows that the period of each note, relative to the fundamental, is the reciprocal of the appropriate frequency ratio. Using period rather than frequency will greatly simplify the calculations required for the counter programs. Therefore, the programmed count for a particular note is proportional to the period of that note.

A simple counter cannot deal with vulgar fractions, so the next step is to give all the fractions a common denominator (in this case 180). The numerators can now be expressed as integral

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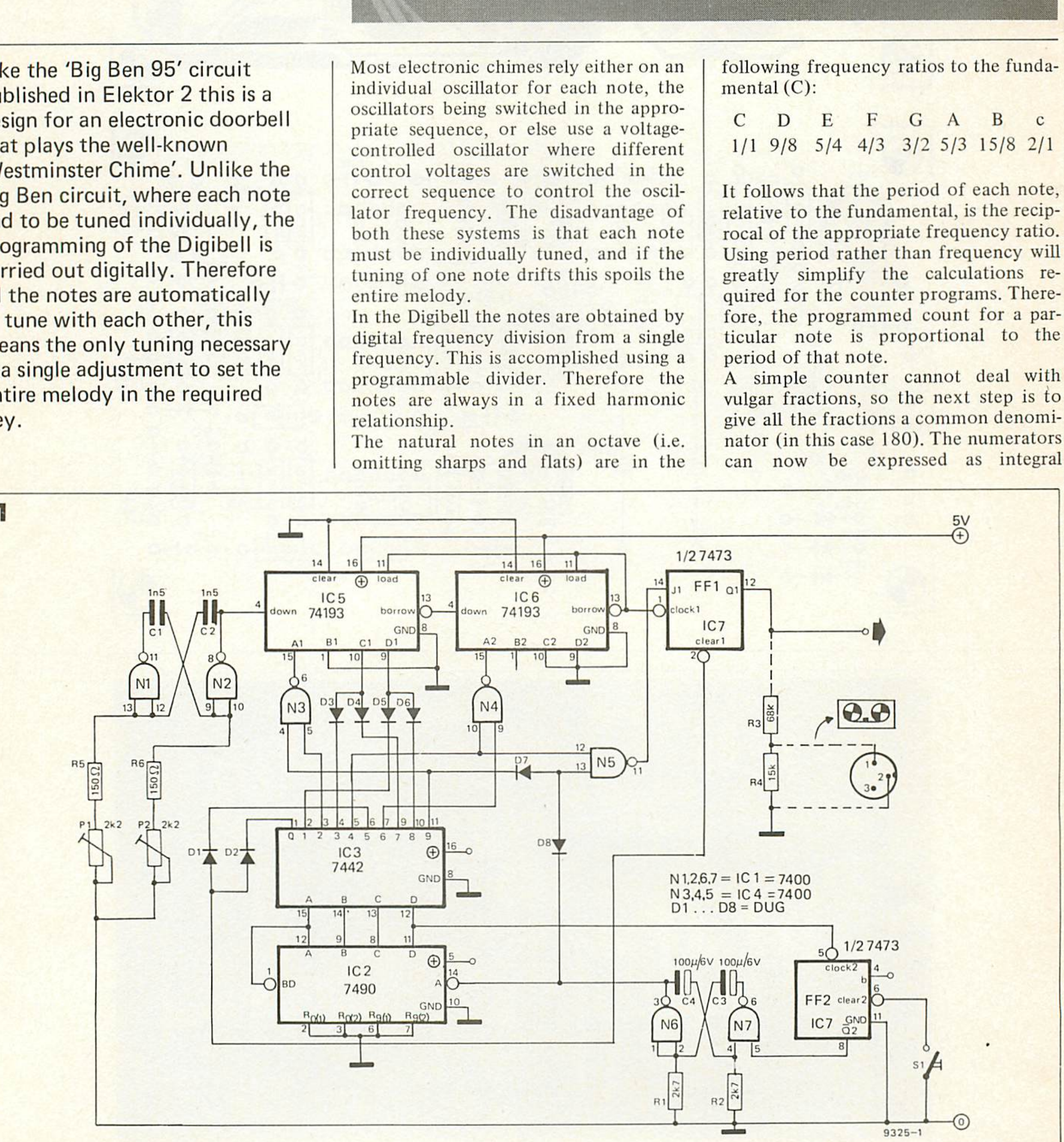

numbers and converted into binary code (since this is what the counter will be programmed with).

This results in the following table:

Note	Decimal	Binary
c'	90	01011010
b	96	01100000
a	108	01101100
g	120	01111000
f	135	10000111
e	144	10010000
d	160	10100000
c	180	10110100
B	192	11000000
A	216	11011000
G	240	11110000

It is evident, from the above table, that if the programmable counter is set to count to 90 and is fed with a clock frequency 90 times that of c' then the output frequency will be c'. If it is set to count to 180 and is fed with the same clock frequency then the output will be c, one octave below c'. This is how each note is synthesised from a single clock frequency. Since each note bears a fixed frequency ratio to all the other notes it is obvious that the only tuning necessary is to adjust the clock frequency until the melody is in the required key.

The Westminster Chime uses only the notes G, c, d and e in the sequence e, c, d, G, G,d, e, c, so at first sight it appears that division ratios of 240, 180, 160 and 144 are required. By coincidence however, it happens that these numbers are all divisible by four, so the division ratios can be reduced to 60, 45, 40 and 36. This means that the programmable counter can be of shorter length, and that the programming is simplified.

In addition to getting the notes right it is also important to achieve the correct tempo. The first three notes each have a duration of one crotchet, while the fourth note has a duration of one minim (two crotchets). This is followed by a rest of one minim duration. The fifth, sixth and seventh notes are each of one crotchet duration, while the final note has a duration of one minim. The total duration of the tune is thus 11 crotchets. When designing the circuit that performs the sequencing this must be taken into account.

The Circuit

The circuit of the Digibell is given in figure 1. The clock pulse generator consists of two NAND gates N1 and N2. They are connected as an astable multivibrator whose frequency and duty-cycle can be adjusted with P1 and P2. The programmable counter consists of two presettable up-down counters type 74193. These are connected to count down from a preset number to zero, the number being loaded into the data inputs A1-D1, A2-D2 before the start of each count. The operational sequence for the presettable counter is as follows:

initially the borrow output of IC6 is low. This takes the load inputs of IC5 and IC6 low, so the data on the inputs A1-D2 is loaded and the count commences. During the count the borrow output is high, but when the count reaches zero it again goes low, the data is reloaded, the count recommences and so on.

Since the borrow output is low for only a small proportion of each count the output waveform is very asymmetric and is not suitable for use as an audio tone. For this reason FF1 is connected to the borrow output and produces a square wave with a 1:1 mark-space ratio (50% duty-cycle) at half the frequency (i.e. one octave below) the borrow output.

To produce the Westminster Chime melody the programming numbers corresponding to the four required notes must be fed to the data inputs of the presettable counter in the correct sequence. This is controlled by a second counter (type 7490), and a 7442 BCD-to-decimal decoder.

When the bell-push S1 is pressed the $\overline{Q}$ output of FF2 goes high, enabling the astable multivibrator N6/N7, which feeds clock pulses at about a 2 Hz rate into the A input of IC2. These are counted by the 7490 and the BCD outputs of the 7490 are decoded by the 7442. The 10 outputs of the 7442 go low in turn, at each step feeding a different number into the data inputs of IC5 and IC6 via the encoder comprising N3, N4 and D3 to D6. (Note that the 7442 has active low outputs, i.e. outputs are normally high and go low when enabled by the appropriate input code).

On the tenth clock pulse the D output of IC2 goes low, clocking FF2 back to its original state ($\overline{Q}$ output low) until the next time the bellpush is pressed. The correct tempo of the melody is achieved in the following manner. If each note were sustained until the next clock pulse occurred then the notes would simply slur into one another without a pause. This is avoided, and the rest in the middle of the tune is included, by means of N5 and diodes D1, D2, D7 and D8.

At the start of the sequence the 0 output of IC3 is low so FF1 is held in the

Figure 1. Circuit diagram for Digibell.

Figure 2. Truth table for the counter program.

Figure 3. Timing Diagram which shows sequencing of the Digibell.

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COUNT	IC3 OUTPUTS 0123456789	D2	C2	B2	A2	D1	C1	B1	A1	DECIMAL	NOTE
		IC5, IC6 DATA INPUTS BINARY									
0	0111111111	0	0	1	0	1	1	0	0	44*	–
1	1011111111	0	0	1	0	0	1	0	0	36	e
2	1101111111	0	0	1	0	1	1	0	1	45	c
3	1110111111	0	0	1	0	1	0	0	0	40	d
4	1111011111	0	0	1	1	1	1	0	0	60	G
5	1111101111	0	0	1	0	1	1	0	0	44*	–
6	1111110111	0	0	1	1	1	1	0	0	60	G
7	1111111011	0	0	1	0	1	0	0	0	40	d
8	1111111101	0	0	1	0	0	1	0	0	36	e
9	1111111110	0	0	1	0	1	1	0	1	45	c

* Not important as output of FF1 disabled during rest

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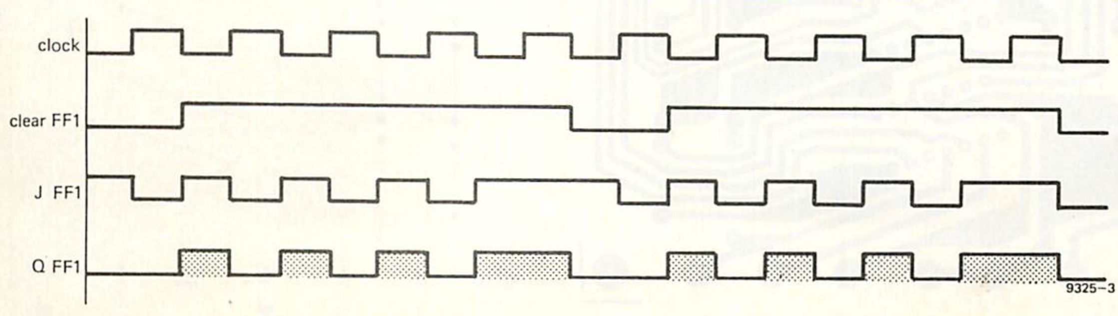

clear state and there is no output. During notes 1, 2 and 3, pin 12 of N5 is high and pin 13 is switched alternately high and low by the output of N6. The output of N5 thus gates the J input of FF1 so that there is an output only when the clock pulse (output of N6) is low. The first three notes thus have a duration of half a clock pulse.

On the fourth note pin 12 of N5 goes low, so the output remains high whatever N6 does. The J input of FF1 is thus high and the fourth note has a duration of one clock pulse. On the fifth step output 5 of the 7442 goes low, so FF1 is held in the clear state via D1 and there is no output. This is the rest.

The next three notes are all of half a clock pulse duration, but on the final note pin 13 of N5 is held low, via D7, the output of N5 holds the J input of FF1 high and this note has a duration of one clock pulse.

To make the sequence of operations clearer, a truth table for the counter programming and a timing diagram for the sequencing are given in figures 2 and 3. Figure 4 shows a p.c. board and component layout for the Digibell. For use as a doorbell the output of FF1 must be amplified to a level suitable to drive a loudspeaker. The output attenuator R3/R4 may or may not be required, depending on the sensitivity of the amplifier used, or they may be replaced by a potentiometer of between 10 k and 100 k to provide a volume control.

Parts list for figure 1

Resistors:
R1,R2 = 2k7
R3 = 68 k
R4 = 15 k
R5,R6 = 150 Ω
P1,P2 = 2k2 adjustable

Capacitors:
C1,C2 = 1n5
C3,C4 = 100 μ/6 V

Semi-conductors:
D1 . . . D8 = DUG
IC1, IC4 = 7400
IC2 = 7490
IC3 = 7442
IC5,IC6 = 74193
IC7 = 7473
S1 = Push button switch

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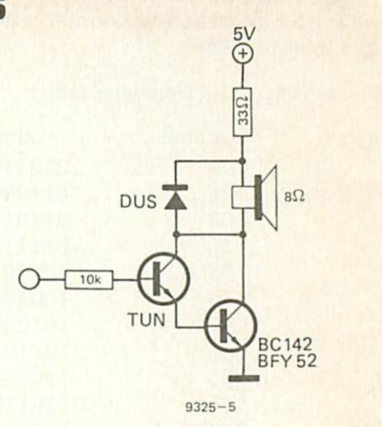

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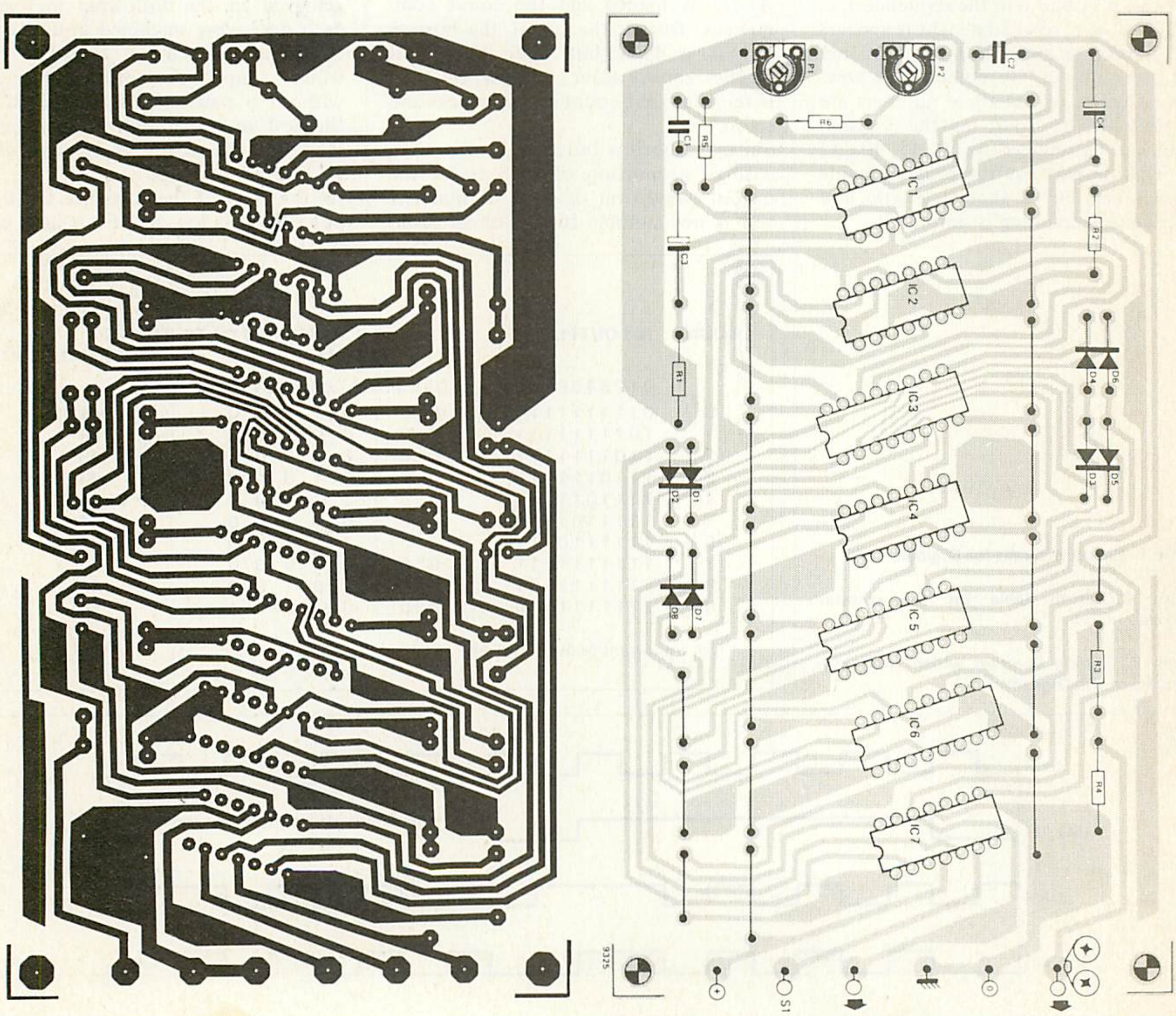


Figure 4. p.c. board (EPS 9325) and component layout.

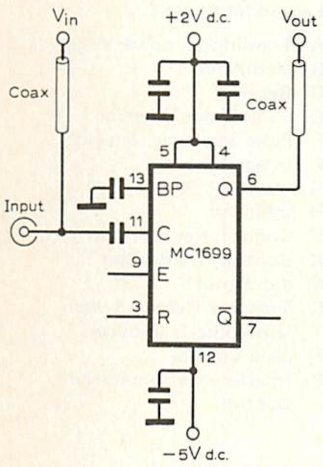
Figure 5. A possible amplifier for use with the Digibell.

MARKET

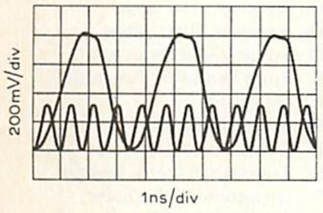
Divide-by-Four Gigahertz counter from Motorola Semiconductors

Communications engineers will be interested in a new integrated circuit from Motorola known as the MC1699 divide-by-four gigahertz counter.
This is a very high speed device for prescaler applications. The clock input requires an a.c.-coupled driving signal of 160 mVpp amplitude (typical). A sine-wave signal is acceptable for frequencies from 50 MHz to 1.2 GHz. Below 50 MHz wave-shaping is recommended. With pulses which have good rise and fall times (in the order of 1 to 2 nsec), the MC1699 has no lower limit on clock frequency. The clock toggles two divide-by-two stages and the complementary outputs (50% duty cycle) are taken from the second stage.

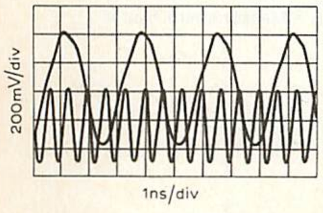
DIVIDE-BY-FOUR GIGAHERTZ COUNTER



COUNT FREQUENCY TEST CIRCUIT



OPERATING CHARACTERISTICS AT 1·2GHz



OPERATING CHARACTERISTICS AT 1·5GHz

The MC1699 includes clock enable and reset inputs both compatible with MECL111 voltage levels. The reset operates only when either the clock or the enable is high and provides increased flexibility for counter and time measurement requirements.
The MC1699 is supplied in a flat ceramic package (F Suffix) for compact assemblies and soon will be available also in a DIL ceramic package for easier mounting.

Motorola B.V.
Emmalaan 41, UTRECHT,
The Netherlands

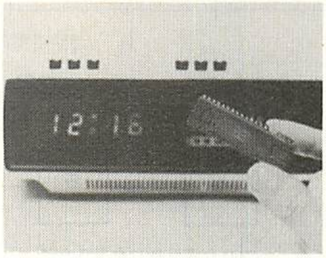
Microwave Power Transistor

Motorola have just announced a state-of-the-art microwave power transistor, the MRF 835. Characteristics specified at 870 MHz using a 12.5 V DC supply are 15 watts output power, 7.0 dB minimum gain and 50% efficiency.
High gain, high power transistors actually consist of a few hundred transistors in parallel. Current to and from these doped regions is carried by metal conductor stripes deposited on top of the die. These stripes must withstand high current densities. In order to achieve good preformance at frequencies as high as 950 MHz Motorola make these 'fingers' extremely narrow. However, in conventional designs, aluminium in narrow lines with high current density migrates, thus causing early failure of the device. In the MRF 835 Motorola have overcome these problems by using a gold metallisation system – because the automatic weight of gold is higher than that of aluminium the migration is greatly reduced and the mean-time-between-failure figure is increased 1,000 to 10,000 times.
The MRF 835 is designed specifically for mobile radio applications at frequencies around 900 MHz and incorporates Motorola's 'Controlled Q' built-in matching network which ensures broadband performance.

Motorola B.V.
Emmalaan 41, UTRECHT
The Netherlands

High current digital clock

AMI Microsystems have introduced a new digital clock module which offers a high current output for direct drive of large LED displays as used in clocks, clock radios, and timers/ elapsed-time counters.
Designated S1998A, it provides

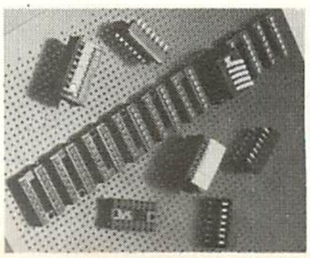


more than 8 mA per segment, and can be directly substituted for the industry-standard S1998 in high current, low voltage display applications. The 1998A directly interfaces with both solid-state LED displays, and fluorescent/ gas discharge displays. The timekeeping function operates on 50 Hz or 60 Hz inputs, and the display output is available with either AM/PM indicators or 24-hour format options.
Other outputs include timed radio turn-off, and radio/alarm enable. A power failure indication is provided to inform the user of an incorrect time display. The S1998A also incorporates a presettable 59-minute count-down timer, an alarm with snooze feature, and unlimited snooze repeat.
Clock input noise rejection circuitry eliminates the need for external filtering of the line frequency input. Reset-to-zero circuitry is included for timer/ elapsed time applications, and blanking control also allows the use of several circuits in parallel with a single display for multiple event timing.
The S1998A, which can operate from power supplies between 8 and 26 V, is pin compatible with the S1998, MM5316, EA5316, and FCI3817.

AMI Microsystems Ltd.
108A Commercial Road,
Swindon, Wiltshire, England.

Low profile IC socket

Molex have announced the availability of a range of low profile dual-in-line integrated circuit sockets.
Designated the 6197 series, they consist of a 94 V-0 black polyester polarized housing containing either 14 or 16 discrete pin sockets. These utilize side-wiping contacts, which offer significantly

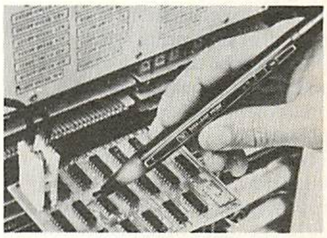


greater contact surface than the more conventional edge-bearing type. Contacts are of 60Cu 30Zn cartridge brass, and are available in a variety of finishes including tin plate and gold over nickel. Optional contact materials such as phosphor bronze, and alternative special contact finishes, are also available.

Molex Electronics Ltd.,
1 Holder Road, Aldershot,
Hants GU12 4RH.

Multi-family logic probe

Designed to simplify and speed logic circuit testing, this new $125 model 545A logic probe from Hewlett-Packard indicates digital states and pulses in both high level (CMOS) logic and low level (TTL) logic. An unambiguous single lamp indicator displays high or low level or detects bad level and open circuit conditions. CMOS and TTL operation is selected with a slide switch. CMOS logic threshold levels are variable and set automatically. Now, nearly all positive logic up to +18 volts dc can be sensed using one probe. These families include: TTL, DTL, RTL, CMOS, HTL, HiNIL, NMOS and MOS.
Another feature of the model 545A is a built-in pulse memory which, along with the display, will catch intermittent pulses. When



a logic change occurs, the indicator lamp turns on and remains lighted until the memory is reset.
Pulse stretching is provided so the operator can see fast pulses as short as 10 nanoseconds with the blinking display. Pulse trains to a frequency of 80 MHz are detected in TTL logic, and to 40 MHz in CMOS logic.
Light and rugged, this hand-held model 545A is fully protected against voltage overload. Power required for TTL operation is 4.5 to 15 volts DC, and for CMOS operation is 3 to 18 volts DC.
To use the 545A, the operator connects the probe to the circuit's highest level power supply, sets the slide switch to the appropriate logic family, then probes. Open, pulsing, or stuck nodes and gates are quickly detected.

Hewlett Packard, P.O. Box 349,
CH-1217 Meyrin 1 Geneva,
Switzerland

MARKET

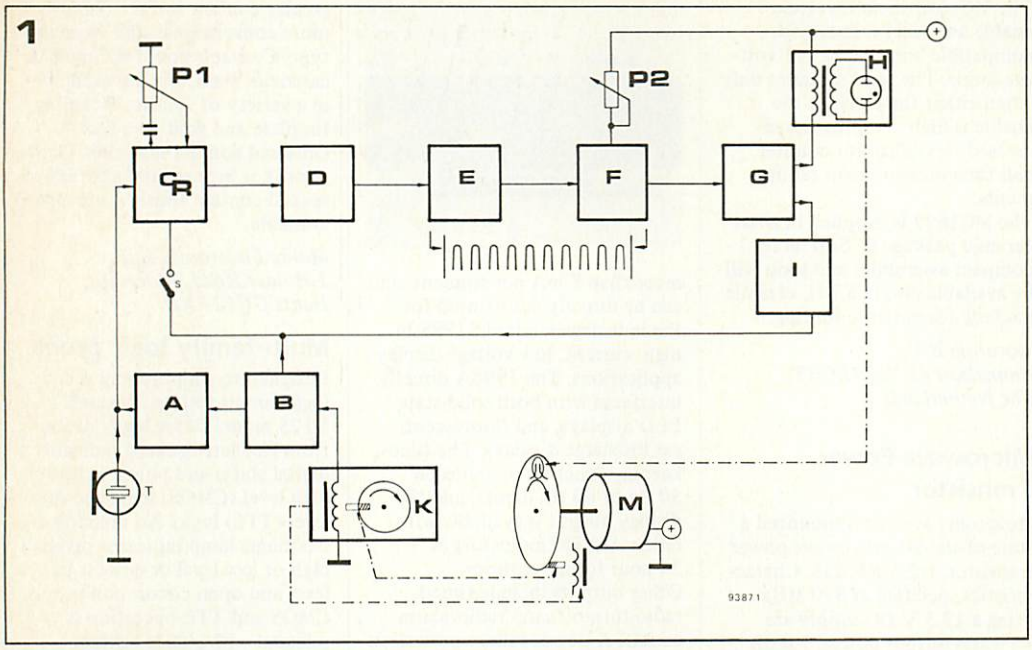


Figure 1. Block diagram of an Echo Sounder. The ultrasonic transducer T operates both as transmitter and as receiver.

Figure 2. The circuit of an Echo Sounder constructed around LM1812. The transmission frequency for underwater distance measurement is about 200 kHz.

Figure 3. This block diagram is intended to complement figure 2 to make clear at which stages the external components are connected to the IC.

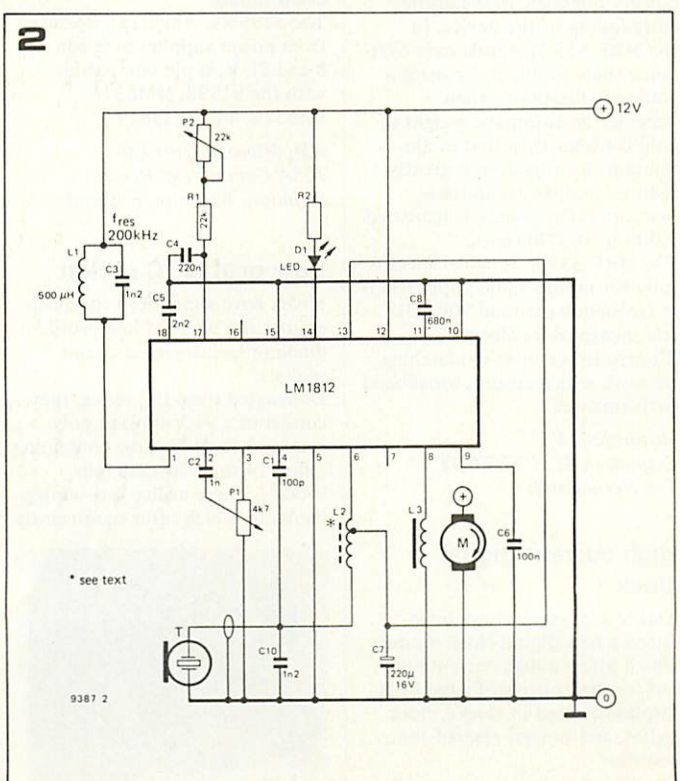
Figure 4. A Sodar equipment with the LM1812 for operation in air. It operates with a transmitting frequency of 40 kHz. The efficiency is improved by the stage enclosed on the right of the figure, the purpose of which is to extend the internally produced 1 μs pulse to 5 μs.

LM1812 Universal Ultrasonic Transceiver

The National Semiconductor LM1812 IC consists of a 12 W ultrasonic transmitter, a selective receiver and indicator drive circuitry on a single chip. It is available in an 18 pin DIL plastic package. The circuit was developed primarily for use as an underwater echo sounder (Sonar). As well as measuring water depth it can also be used to locate the position of shoals of fish and other immersed or sunken objects.
The IC may also be used with ultrasonic transducers in air, which opens up a completely different range of applications. For example, it is possible to measure the level of corrosive liquids where a sensor cannot be dipped in the fluid. Other possibilities include collision warning systems and intruder alarms (ultrasonic radar –Sodar). Finally, it is also possible to transmit data over the ultrasonic link so that communication or remote control systems are possible, both in air and underwater. One application would be the remote control of model submarines, which is virtually impossible by any other means.
In order to understand the operation of the circuit a block diagram of an echo sounder is given in figure 1. The transmitter (block A) feeds the ultrasonic transducer T with a burst of 1 μs pulses for the duration of the transmitting phase (about 800 μs). In under water applications the pulse repetition frequency is about 200 kHz. During this period the receiver (blocks C to F) is switched off to avoid damage by the high transmitter power. This is indicated symbolically by switch S, but is of course accomplished electronically within the IC.
The ultrasonic pulses applied to the transducer are coupled into the water and spread out as spherical wavefronts. When the sound waves strike the bottom or some other objects they are ultimately received by the transducer and converted back into electrical impulses. These are amplified and detected by the receiver which is now activated.

The time duration between transmission and reception of the burst of pulses is proportional to the distance between the transducer and the reflecting object.

In the common type of commercially available echo sounder the timing is carried out electromechanically using a constant speed motor (M). The motor has a disc attached to its shaft the periphery of which are mounted a small neon lamp and a magnet, 180° out of phase. Once every



Legend for figure 1

- A Transmitter, power stage
- B Modulator
- C Receiver
- D Surge Value Detector
- E Pulse Sequence Detector
- F Integrator
- G Indicator Driver
- H Indicator
- I Control, Keying Ratio
- K Control, Transmitter
- M d.c. Motor
- S Transmit/Receive Switch
- T Ultrasonic-Transducer
- P_1 Gain Control
- P_2 Interference Suppression Control

Legend for figure 3

- A Transmitter, power stage
- B Modulator
- C_a Receiver, 1st. stage
- C_b Receiver, 2 nd. stage
- D Surge Value Detector
- E Pulse Sequence Detector
- F Integrator
- G Indicator Driver
- I Control, Keying Ratio
- S Transmit/Receive Switch
- T Ultrasonic-Transducer
- P_1 Gain Control
- P_2 Interference Suppression Control
- a = Transmit Pulse Sequence
- b = Measurement Range

market

revolution the magnet induces a pulse in a pickup coil (block K). This pulse triggers the transmitting sequence by controlling the modulator (B).

The ultrasonic wave sent out by the transducer is of course subject to the inverse square law, and since it must traverse the distance between the transducer and the reflecting object twice it is attenuated very severely by the time it returns to the transducer. Energy absorption by the reflecting object and the efficiency of the transducer must also be taken into account, and it is evident that the receiver must be very sensitive.

The receiver consists of a multistage amplifier (block C), with gain control provided by the potentiometer P1. This is followed by a threshold gate (block D), which allows only signal exceeding a certain amplitude to pass. This ensures that noise and delayed echoes from very distant objects cannot cause a spurious indication.

The signal is further processed in blocks E (pulse sequence detector) and F (integrator). These together form a pulse width detector, which performs two functions. It is first determined whether a valid echo has been received by ensuring that the received signal is of the same duration as the transmitted signal. If more than five consecutive pulses in the sequence are missing then no indication of depth is given. This ensures that any interference pulses strong enough to pass the threshold gate will still not cause a spurious indication. The degree of interference suppression may be adjusted by P2.

If a valid echo has been received then the output driver (block G) is triggered and produces an output pulse that is stepped up by a transformer to strike the neon. The angle through which the motor shaft (and hence the neon) has turned before the neon strikes depends on the distance between the transducer and the reflecting object. The rotating disc is mounted behind a circular scale with transparent divisions marked off in fathoms, feet or metres.

Figure 2 shows a practical circuit for an echo sounder using the LM1812, while figure 3 shows how the external components associated with figure 2 tie in with the internal circuitry of the LM1812. The pulse induced in the pickup coil L3 by the rotating magnet activates the transmitter. The 200 kHz sinewave oscillator is tuned by L1 and C3 (it also tunes the receiver gain stages. See figure 3). The output of this oscillator is amplified and limited to provide a 200 kHz squarewave that is used to trigger a monostable that produces the 1 μs pulses. After amplification by the transmitter output stage the pulses appear across the transducer T.

L2, C2 and the capacitance of the screened transducer lead form a parallel resonant circuit that can be tuned to the transmitting frequency by adjusting the core of L2. In this circuit an LED D1 is used rather than a neon. If additional receiver gain is required then an extra stage of amplification may be connected between blocks C_a and C_b.

This system relies for its accuracy on the speed stability of the d.c. motor. Slip rings must be used to make connection to the neon unless a rotating transformer is used, and the system inevitably suffers from mechanical wear. However, it is cheap and provides an easily interpreted analogue readout, and echo sounders using these principles are popular with small boat owners.

A completely electronic system can be constructed by replacing the motor arrangement with an electronic digital counter. The transmitter can be activated by a rectangular pulse of about 1 ms duration. This also opens a gate to the clock input of the counter, which is fed by pulses from a

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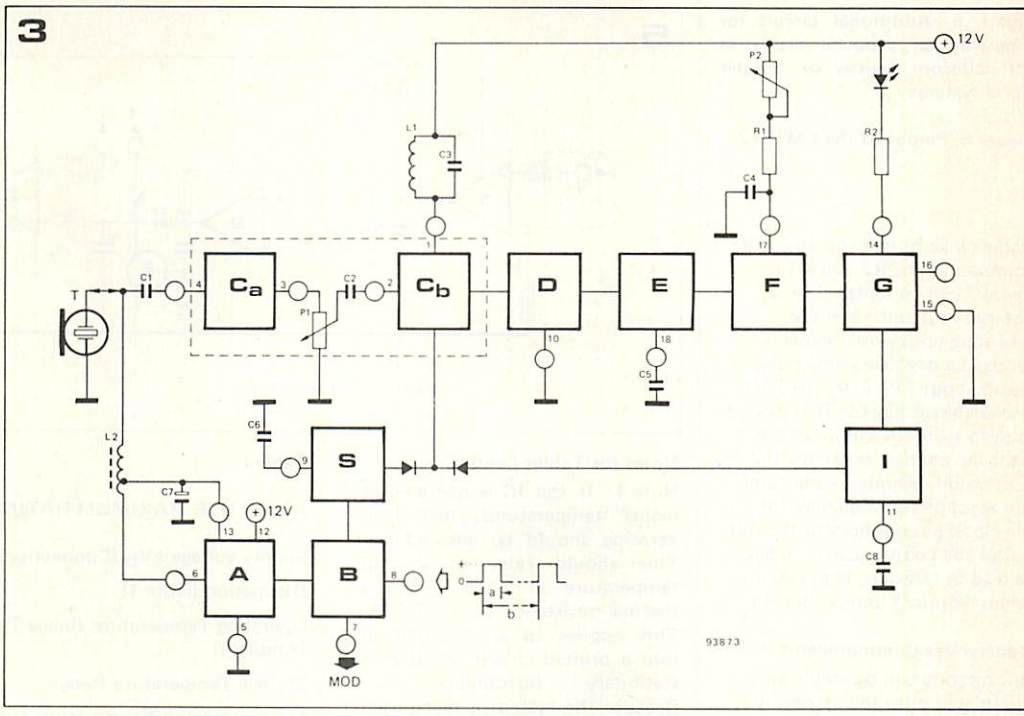

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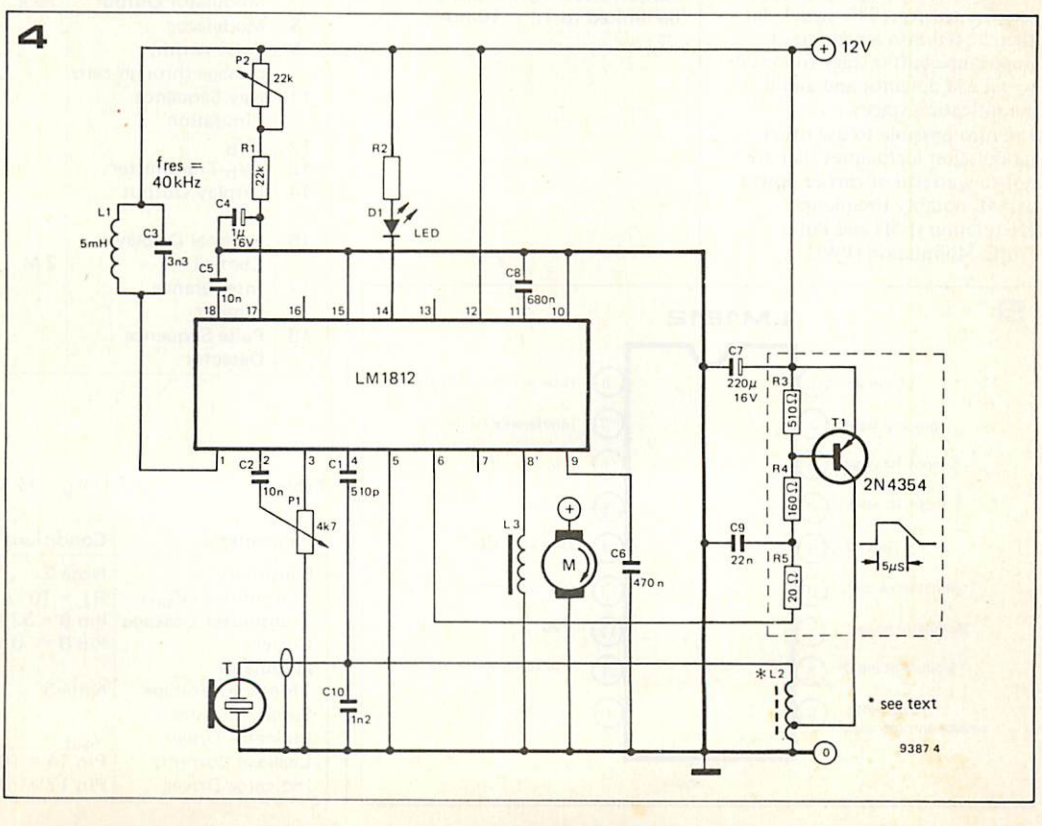

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Figure 5. Additional circuit for triggering an acoustic alarm in anti-collision devices or burglar alarm systems.

Figure 6. Pinout of the LM1812.

stable clock pulse generator. The combination of R2 and D1 in figure 2 can be replaced by a 5k6 resistor, and when the returning unltrasonic signal is received a negative going pulse (from about $+V_b$ down to 1 V) is available at pin 14. This can be used to store the count and to reset the counter ready for the next count. Simply by choosing the appropriate frequency for the clock pulse generator the output of the counter can be made to read in fathoms, feet, metres or any required units of length.

Underwater Communications

Information can be amplitude-modulated onto the carrier by feeding a low-frequency (e.g. audio) signal into the modulator input pin 8, rather than simply switching the transmitter on and off with digital signals. The received signal must be taken out of the receiver at a point before the signal processing stages (since after signal processing it is no longer an amplitude modulated carrier). A convenient point to take off the signal is at the LC circuit L1/C3, connected to pin 1, which tunes both the transmitter and the receiver. The signal can then be fed into a high input impedance buffer stage followed by an AM detector and audio amplification stages.
It is also possible to use other modulation techniques that are not so wasteful of carrier power as AM, notably Frequency Modulation (FM) and Pulse Width Modulation (PWM).

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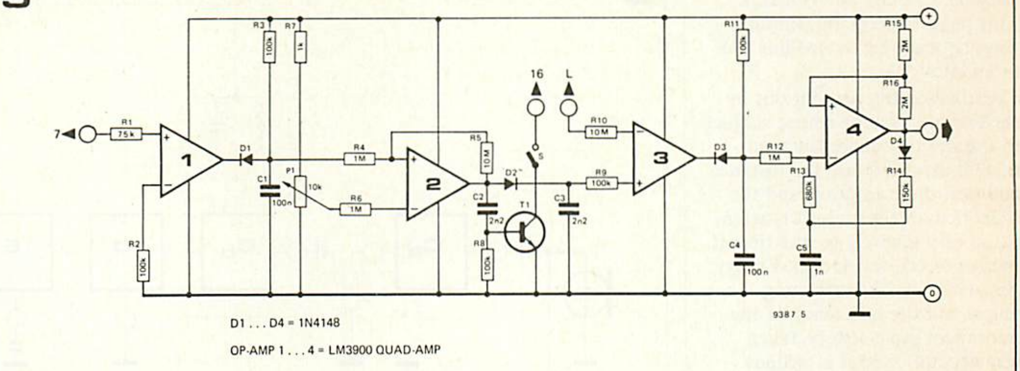

Notes for Tables I and II

Note 1: If the IC is operated at higher temperatures, then load derating should be allowed for. This should refer to a chip temperature of +125°C and a thermal resistance of +167°C/W. This applies to an IC soldered into a printed circuit board with stationary surrounding air. Because the system is being used for pulsed operation, the heat resulting from the dissipation in the enclosure is only slight.

Note 2: During the measurement of the sensitivity, an attenuator of 500:1 was fitted, to ensure reliable measurements at higher input levels.

Note 3: The 'Modulator Threshold Voltage' is the voltage which has to be applied to pin 8 to bring the system into the 'Transmit' condition. The current flowing in pin 8 should be limited to 1 . . . 10 mA.

Table I

ABSOLUTE MAXIMUM RATINGS

Supply voltage $+V_b$ (Connections 12, 6 and 4)	18 V=
Dissipation (Note 1)	600 mW
Operating Temperature Range T_{amb} (Ambient)	0°C . . . +70°C
Storage Temperature Range	−65°C . . . +150°C
Maximum Lead Temperature during soldering (Solder duration 60 s max.)	+300°C

Pin	Function	R_{ext} (min)	V_{max} (Instantaneous value V_s)	I_{max}
1	LC-Circuit		30 V	
2	Input 2nd. stage		18 V=	50 mA
3	Output 1st. stage		18 V=	
4	Input 1st. stage			50 mA
6	Transmitter Output		36 V (Transmitter 'Off')	1 A for 1 µs
7	Modulator Output	75 k	18 V	
8	Modulator			50 mA
9	Pulse Width, passage through zero		7 V	
11	Key Sequence Limitation			50 mA
12	$+V_b$		18 V=	
13	$+V_b$-Transmitter		18 V=	
14	Display Output		25 V= (if 'Off')	1 A for 1 ms
16	external Display Control	2 M	18 V	
17	Interference Limiter			50 mA
18	Pulse Sequence Detector			50 mA

6

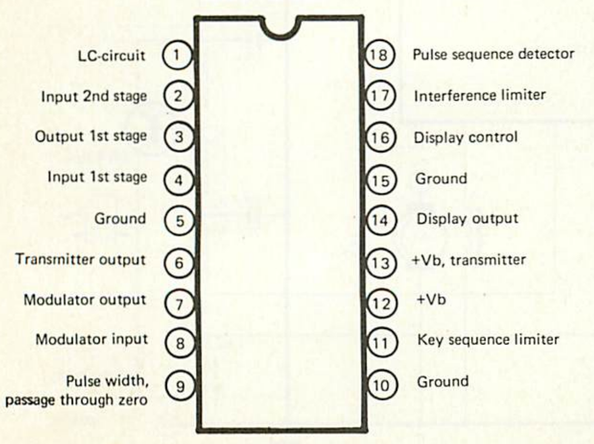

Table II ($+V_b$ = 12 V, T_{amb} = +25°C)

Parameter	Conditions	Min.	Typical	Max.	Unit
Sensitivity	Note 2		200	600	µV (V_{ss})
Transmitter (V_{sat})	R_L = 10 Ω		1.3	3.0	V
Transmitter Leakage Current	Pin 6 = 32 V, Pin 8 = 0 V		0.01	1.0	mA
Modulator Threshold Voltage	Note 3	0.55	0.7	0.9	V (V_s)
Supply Current		5	8.5	20	mA
Indicator Driver	V_{sat}		1.5	3	V
Leakage Current, Indicator Driver	Pin 14 = 16 V, Pin 17 = 0 V		0.01	1	mA

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Operation in Air

Figure 4 is a circuit for an echo sounder or distance measuring instrument operating in air. This is almost indentical to the circuit of figure 2, with two exceptions. The transmission frequency is lowered to 40 kHz by altering L1 and C3, and the pulse length of the internal monostable is extended to about 5 μs by the circuitry around T1 (shown in dotted box). These modifications give a better transducer efficiency for operation in air.
This type of circuit could be used, as mentioned earlier, in vehicle collision warning systems or liquid level sensors. Other applications include intruder alarms. The circuit would be set up so that echoes from the area to be protected would not trigger an alarm, but any movement of an intruder in the vicinity of the transducer would cause the reflected signal to be received earlier. Another posibility would be to set up the circuit so that the received signal level from the protected area would not trigger the alarm. Any movement of an intruder within this area would alter the signal level and this could be detected and used to trigger an alarm. An alarm control circuit is given in figure 6.

Literature
T. M. Frederiksen, W. M. Howard. 'A Single-Chip Monolithic Sonar System'. IEEE Journal of Solid-State Circuits, December 1974, Vol. SC9 NO. 6. pp. 394-403.
'LM1812 Ultrasonic Transceiver'. March 1975, National Semiconductor GmbH.
National Semiconductor Limited, The Precinet, Broxbourne, Herts.

Automatic LCR Bridge

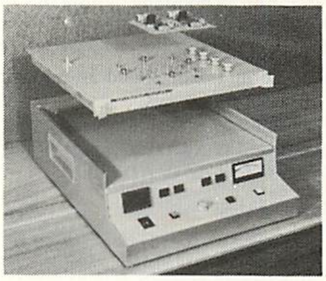
With this new automatic LCR Bridge, the user need only select the function L, C, or R — the instrument selects the measurement range and equivalent circuit. Besides producing 3½digit LED readouts in capacitance, inductance, and resistance, it also provides readouts in D, C/D and L/D with the accuracy of manual bridges at rates as high as 1 reading per second. Accuracy is typically 0,2% of reading.
Parameters of components such as semiconductors, pulse transformers, filter coils, electrolytic and film capacitors, or the internal resistance of a dry cell can be quickly and easily determined in production and in the laboratory with this new Hewlett-Packard Model 4261A Digital LCR Meter.
Measurements ranges are 0.1 picofarads to 1.900 microfarads and 0.1 microhenries to 190 henries at 1 kHz measurement frequency. Resistance is measured from 1 milohm to 19 megohm in 8 ranges and dissipation factor from 0.001 to 1.900. At a measurement frequency of 120 Hz, capacitance can be measured from 1 picofarad to 19 millifarads in 8 ranges; inductance can be measured from 1 microhenry to 1900 henries in 8 ranges.
Two test signal levels, 1 volt and 50 millivolts, are available. Three internal bias voltages or external bias can be selected as well as internal or external triggering. A range hold function can be used where a series of repetitive

measurements must be made.
In addition to autoranging, the Model 4261A will automatically select the appropriate equivalent circuit depending upon the value of the component under test. However the user may select a parallel or series equivalent circuit manually if he wishes.
Measurements are made with the five terminal method — easily converted to four, three or two terminals, depending upon the measurement requirement and the value of the device under test.
Three test fixtures are available as accessories — one direct-coupled for five terminal measurements, a four terminal fixture and a three terminal fixture.
A pull-out instruction card on the front panel is a ready reference for measurement procedures.

Hewlett Packard, King Street Lane, Winnersh, Workingham, Berkshire RC11 5AR.

Low-cost production

A low-cost, automatic test unit for electronic components, circuit boards or other assemblies, with a test capacity from 40 to 100 test points has been introduced by Ancom Ltd.
Designated the 'Sentitest' production test monitor, it incorporates several features that the latest experience in this field has shown to be desirable. It is a bench-top unit, built to a modular construction and designed for simple programming and 'instant' servicing.
Intended for use by unskilled operators, the Sentitest is basically an automated multimeter, which switches its ranges and function according to a simple pre-set programme requiring just the wiring-up of resistors and diodes.
Testing can be carried out at low signal levels to avoid damage in the event of wiring or component faults.
Operation is by a single push-button with plain language, the read-out is by go/no-go coloured lamps, and there is an analogue meter to help analyse faults when required. Additional output points are available for data-logging or alarm purposes.
In the automatic mode it will scan up to 100 test points in about 10 seconds; it can also be set to re-cycle for life testing, or to locate an intermittent fault.
In the manual mode, the test programme is sequenced at the operator's own pace by pressing the button.
A digital indicator numbers the test points as they are scanned and, for subsequent fault analysis, any test point can be recalled by dialing its number on a manual dial.
The item under test is held by a suitable jig, to make solderless contact with the required test points, on top of a box section carrier which plugs into the top of the test unit. The carrier contains all the jigging and programme logic for the particular item.
Programme changes are made simply by interchanging carriers, so that with a library of suitable carriers, a single Sentitest unit can cover a wide range of test needs and quickly be changed from one to another.
The test unit itself is built up from plug-in modules comprising a power supply, one analogue circuit board, one logic board and 'decade extender boards' (DEB's).



The basic unit incorporates one 'master' and four 'slave' DEB's for up to 40 test points. Capacity can be extended in steps of ten test points at a time by up to six more DEB's (i.e. a total of 100 test points).
With the increasing acceptance of the cost benefits of ATE within the electronics industry, which now extends to very sophisticated computerised digital equipment costing many thousands of pounds, Ancom believes that there is an equally growing need for low-cost ATE.

Ancom Ltd, Denmark House, Devonshire Street, Cheltenham, Glos. U.K.

$ 15.00 Microprocessor

For the designer who doesn't really need the performance of a general-purpose microprocessor, a specialized unit is now available for $ 15.00 in quantity lots.
Built by National Semiconductor, the unit, called SC/MP (Simple Cost-effective Applications Microprocessor), will be good for simple control and timing functions usually taken care of by Random logic.
SC/MP is expected to find application in appliance controls, small building security monitors, fuel-injection units for cars as well as traffic-signal control, word-processing terminals and scales and electronic toys — "Anything that doesn't require speed or too much computation".
The 8-bit PMOS microprocessor operates at 2 μs cycle time, it requires a single 12-V supply, with a comfortable margin of ± 2 V. And it generates its own timing right on the chip, as opposed to a need for other chips to handle this function.
Multiple SC/MPS can communicate with one another when they all share a common bus. Logic built on the chip allows each SC/MP to sense when the bus is in use. Only when one SC/MP stops transmitting or receiving. the one next to it can take over. If it declines, the one adjacent to it is given a chance.
A simple control system can be configured using only the SC/MP and a program memory, which can be selected from a wide range of standard memory parts. This system can access up to 4 kilobytes of memory to provide the control logic for almost anything previously controlled by sheet-metal logic: electronic games, small-intersection traffic control signals, simple industrial systems, appliances, and vending machines.
A five-chip system, composed of the SC/MP, a two-chip bidirectional transceiver, address latch, and buffer element, can interface to 65 kilobytes of memory for more complex control functions, as in credit-card verification, business and accounting machines, text-editing typewriters, intelligent stand-alone terminals, and measurement systems.

National Semiconductor GmbH
D 808 Fürstenfeldbruck
Industriestraße 10
West Germany

market

Synthesized signal generator

A new synthesized microwave signal generator, Model 8672A from Hewlett-Packard, covers the range 2 to 18 GHz in one solidstate package only 5 1/4" high. With AM/FM and calibrated output usually associated only with signal generators, 8672A also offers the resolution, spectral purity, stability, and programmability of a high-quality synthesizer. Yet its price is well below equipment commonly used in the past to synthesize microwave signals over comparable ranges. A companion instrument, Model 8671A, covers the range 2 to 6 GHz with FM capability only. Both machines are programmable via the HP Interface Bus.

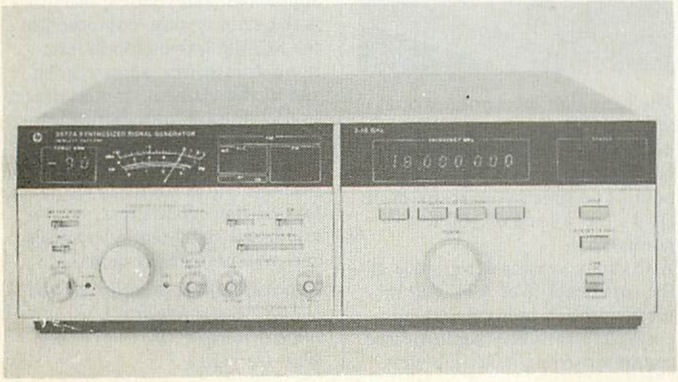
Frequency resolution of the 8672A is 1 kHz in the range from 2 to 6.2 GHz, 2 kHz from 612 to 12.4 GHz, and 3 kHz above 12.4. Frequency stability is better than ± 5 in 10^{10} parts per day with the internal frequency standard. Spurious signals are more than 70 dB below the carrier at 6 GHz, more than 60 dB down at 18 GHz. Single-sideband phase noise (in a 1-Hz bandwidth) is more than 78 dBc 1 kHz away from a 6-GHz carrier; 100 kHz away, it's −109 dBc. Calibrated output is from +3 to −120 dBm; attenuation is displayed digitally on an LED readout. In overrange, power levels to +10 dBm are typically available below 6.2 GHz, and to +7 dBm at other frequencies. Power can be internally leveled ± 1.25 dB; connections are provided for leveling externally.

Modulation signals are externally supplied but internally monitored. AM 3-dB bandwidth is more than 500 kHz at 6 GHz, more than 100 kHz at 18 GHz. AM depth is selectable in two ranges, 30% per volt and 100% per volt, controlled linearly by varying the input signal. FM input rates to 10 MHz are possible up to peak deviations of 10 MHz. Six ranges of peak deviation are offered, from 30 kHz per volt to 10 MHz per volt of input. Peak deviation may also be displayed on the front panel meter. FM and AM are entirely independent, and may be applied simultaneously. Without exception, all front panel controls are remotely programmable. Frequency changes typically settle within 1 kHz of command in less than 15 milliseconds. RF output can be programmed over its full amplitude range in 1-dB steps as well as OFF (without unpowering the instrument). AM or FM, or both at once, may be remotely applied.

The lower-priced 8671A has the same spectral-purity, resolution, and switching-speed performance as the 8672A; it is likewise totally programmable via the HP-IB, and it has the same FM specs. 8671A lacks AM capability, output levelling, and calibrated output attenuator. Its frequency range is 2 to 6.2 GHz.

First customer deliveries are expected in June.

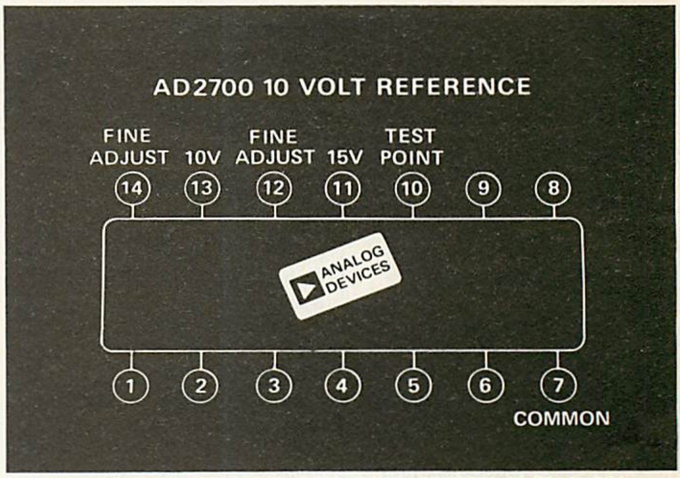
Hewlett Packard
7, rue du Bois-du-Lan,
P.O. Box 349,
CH-1217 Meyrin 1 Geneva,
Switzerland



Precision Voltage References

Analog Devices Ltd., announce the availability of a new ± 10 V and −10 V precision voltage reference, together with an improved performance version of the recently introduced 10 V reference. Designated AD2702, AD2701 and AD2700/L, respectively, the new units are ideally suitable for application in 12-bit A/D and D/A converter circuits, combining high performance thin-film technology, automatic laser trimming and volume hybrid assembly, with small size and hermetically sealed 3-terminal terminal references.

Designed to now operate over the extended temperature range of −55°C to 125°C, AD2700/L features ± 0.03% total maximum error guaranteed from −25°C to 85°C, and load regulation over the 0 to 20 mA range of ± 0.004%. Military versions – AD2700/U and AD2700U/883 – feature screening to MIL-STD-883A, 5004.2, Class B, ans +0.03%, −0.05% total maximum error over the full −55°C to 125°C temperature range.



ANALOG DEVICES' AD2700 10.000 ±0.001V REFERENCE FEATURES LOW PRICE AND ±0.03% MAXIMUM OUTPUT ERROR AT +70°C

The new Analog Devices AD2701 and AD2702 offer identical specifications to the AD2700 but having −10 V and ± 10 V ouputs, respectively.

Available in 14-pin DIP packages, AD2700/L, AD2701 and AD2702, each feature an accuracy to within ± 0.001 V and minmize voltage noise (0.1 to 10 Hz) to below 50 μ Vp-p. Maximum output current is 20 mA and short circuit protection is provided.

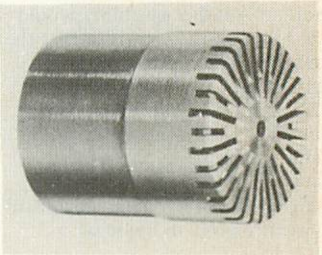
Analog Devices Ltd.,
Central Avenue,
East Molesey,
Surrey. 01-941 0466

High Sensitivity Microphones

Two new ½" Measuring Condenser Microphones with improved sensitivity have been developed by Brüel & Kjaer. These microphones have the same sensitivity as 1" condenser microphones. Both are intended for general and low level sound measurements and are delivered with individual calibration chart.

The 4165 is a free-field microphone with a linear 0° incidence frequency response from 4 Hz to 18 kHz ± 1 dB and a sensitivity of 50 mV/Pa. It fulfils the requirements to microphones in IEC R 123, IEC R 179 and ANSI S 1.4-1971, Types 2 and 3.

The 4166 has a linear random incidence frequency response from 4 Hz to 8 kHz ± 1 dB and a sensitivity of 50 mV/Pa. It fulfils the requirements to microphones in ANSI S 1.4-1971, Types 1, 2 and 3.



Brüel & Kjaer
23 Linde allé
DK-2850 Naerum Denmark

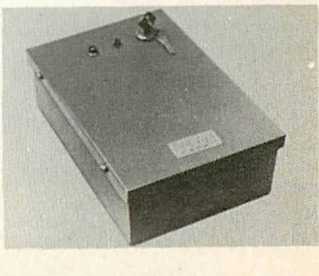
Burglar Alarm

ANTEX has launched what is believed to be the first D.I.Y. burglar alarm kit which complies with the strict standards of B.S.4737.

The illustration shows the control box of the battery-operated version (Model A 1 B). The mains-operated Model A 1 MB is suitable for the average 3-bedroomed house.

The electronic circuitry which provides the owner with a variety of useful facilities is housed in a strong, tamper-proof steel box. Special precautions were taken to prevent false alarms, the bane of so may alarm systems.

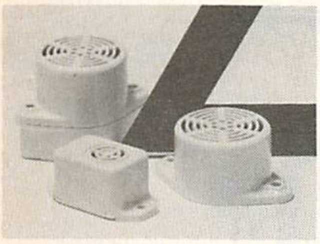
ANTEX Ltd.,
Mayflower House,
Plymouth, Devon.



MARKET

Audible warning modules

Roxburgh Electronics Ltd have released a series of micro-electronic buzzers capable of providing audio frequency outputs in excess of 90 dB at 20 cm. They have been designed for ease of installation, and for direct connection to associated integrated circuitry.
The SMB series, which requires a power supply of 1.5Vd.c., and the MB series, which operate from 3 V, 6 V or 12Vd.c., each

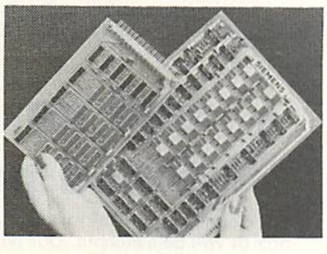


provide a continuous alarm output. The MBF series is available for applications requiring an intermittent alarm, and operates from 6 V or 12Vd.c.
These two ranges are complemented by the RMB-24 series, which operates from a power supply between 20Vd.c. and 28Vd.c., and similarly provides an audio frequency output exceeding 90 dB at 20 cm.

Roxburgh Electronics Ltd,
22 Winchelsea Road,
Rye, Sussex

Memories that plug in

For small- and medium-scale data systems Siemens is now supplying two memories designed for word lengths of 4 bits and 8 bits. These semiconductor memories are mounted on Euroboards fitted with plug connectors and are organized on the 4K by 4 bit (static) and 8K by 8 bit (dynamic) basis. Bit prices of .55 p for the 16K memory and .3 p for the 64K memory make these plug-in memory systems a good choice for such equipment as point-of-sales terminals, data display terminals, metering and regulation systems and particularly microprocessors of all kinds.
The 4K by 4 bit matrix of the smaller memory system is constructed from 16 static 1K MOS RAMs of the SAB 2102 type. A start pulse and two control levels (address mode, read-write) are used as control signals. So as not to overload the address lines capacitively, a register is provided for driving the memory with only brief information (around 150 ns). The cycle and access times are 1.10 μs and 1.05 μs respectively. Job control, input and output registers and a decoder complete the memory system, which is plugged in via a 31-point pin connector.
The memory section of the 8K by 8 bit system comprises 16 dynamic 4K MOS RAMs of the HYB 4060 type each with a capacity of 4096 bits. The total capacity is thus organized into 8192 words by 8 bits. Due to the dynamic character of the modules, one of the 64 rows in each of the memory modules is automatically refreshed at intervals of 32 μs. During these refresh periods no operating cycles are performed by the memory; any signals that arrive are buffered. The cycle and access times are 650 ns and 450 ns respectively. All input and output signals transmitted via the 60-point plug connector are TTL-compatible as with the 4K x 4 bit system.
Monolithic 16K memories are no longer dreams of the future, 64K memories will not be long in coming either. Nevertheless, there will always be certain cases where such memory capacities are best built up from several individual modules on a discrete basis. With such systems, word lengths consisting of several bits can also be processed directly, if the appropriate number of individual memories are connected in parallel. The total memory capacity of 16K is then the result of 4K by 4 bit (8K by 8 bit for 64K). If a 16K single-chip memory were used, a word length of, for example, 4 bits would fix the capacity at a minimum of 64K, but this represents a capacity in excess of that required for most of the envisaged applications. Quite apart from this, the distribution of a capacity among several chips on a plug-in

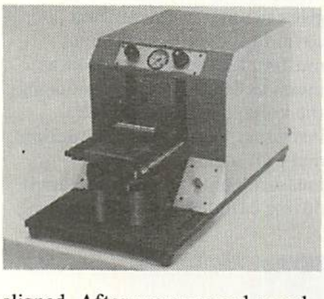


board offers advantages with regard to manufacture and operation.

Siemens A G,
Zentralstelle für Information,
D-8520 Erlangen 2,
Postfach 3240,
West Germany

Safer wafer breaker

Using a patented technique, the WB610 actually separates chip segments after the first and prior to the second break in the scribed wafer. This exclusive feature reduces chip damage to a minimum.
First step in operating the WB610 wafer breaker is the application of an adhesive film to the back of the scribed wafer. The film-backed wafer is placed on the unit's break table and properly

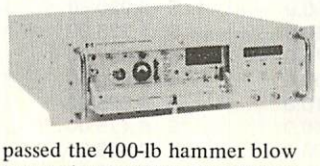


aligned. After pressure and speed have been adjusted to the correct level, the breaker roller is passed over the wafer.
The first break having been completed, the break table is rotated by 90°. Then, the wafer is stretched slightly on the film, thereby separating the chip segments. The final step is passing the roller over the separated chip segments to complete the second break.
With other wafer-breaking techniques most chip damages occurs during the second break. Because of its exclusive chip separation feature, the WB610 eliminates the major cause of damage by separating chips before the second break.
When operated properly, the machine should reduce chip loss by as much as 40 to 60 per cent of normal.

Laurier Associates, Inc.,
Executive Drive, Hudson,
N. H. 03051

Cesium standard

The Hewlett-Packard 5062C Cesium Beam Frequency Standard offers both the precision of the best lab standard with the ruggedness of military hardware in a compact package. It maintains 3×10^{-11} accuracy over a wide operating temperature range and requires only 20 minutes of warm-up time even from −28°C.
Operating temperature range: −28°C to +65°C. Ruggedness:

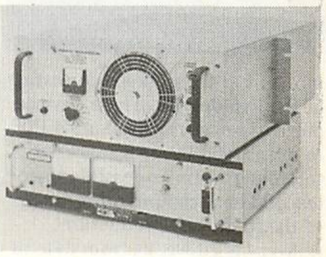


passed the 400-lb hammer blow test under operating conditions. With a calculated MTBF of 25,000+ hours, the 5062C is highly serviceable. Twelve critical circuits are monitored by the front panel meter. The unit is 5¼" high and will fit into a standard 19" rack. The basic 5062C weighs 50 lbs.
This new frequency standard is ideally suitable for navigation, communication, guidance systems, among other on-line system applications where high performance in field environments is required.
Optional digital display clock and standby battery available at extra cost.
Basic unit costs around £ 10,000.

Hewlett Packard, King street lane,
GB-Winnersh,
Wokingham RG11 5AR,
Berkshire

Troposcatter amplifier

MPD announces the availability of a 250 W solid state Tropo amplifier operating in the frequency range 755-985 MHz. The Model PWA7598-251/2726 delivers 250 W cw saturated power with an input power level of 0.5 W. The amplifier is essentially transparent to NPR noise loading.
Applications for this amplifier is as a troposcatter transmitter on off-shore oil riggs such as those located in the North Sea, the Persian Gulf and any place else where troposcatter over-the-horizon communications are utilized.
Features of the amplifier include circulator protected output for



load protection against open or short circuits, B^+ reversal protection, thermal overload protection, and the unit also has alarms (visual and remote) for RF input failure and low power output reduction. The unit comes with built-in test equipment (BITE) to determine if an RF transistor has failed.

Specifications:

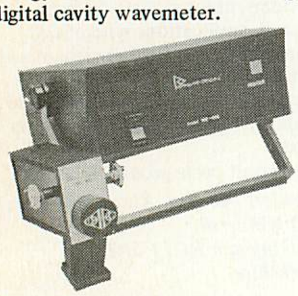
Frequency range:	755-985 MHz
Bandwidth:	230 MHz @ 1 dB
Power output (50 Ω):	250 W cw saturated
RF input range:	0.5-3.5 W
Input VSWR:	2 : 1
Load VSWR:	Circulator protected
Harmonics:	−30 dB minimum

Microwave Power Devices, Inc.
Adams Court, Plainview,
N. Y. 11803

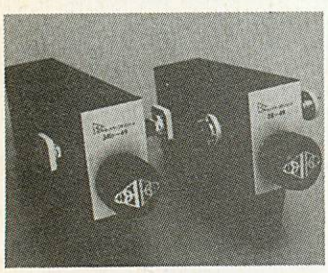
market

Digital Cavity Wavemeter

Baytron Company, Inc. manufacturers and developers of millimeter wave components and systems, is proud to introduce the most destinctive contribution to waveguide passive device technology in a decade-direct reading digital cavity wavemeter.



Some of the features of this device are as follows:
Digital readout – 8 mm LED characters; high resolution – 1 MHz; accuracy – 0.06% absolute error; excellent repeatability – zero backlash; and no spurious resonances – TM_{110} and TM_{010} modes eliminated.
The Wavemeter will carry the part number 10E-49 followed by BAYTRONS 'band' designation, eg. Ka thru D. The digital processor and readout is always

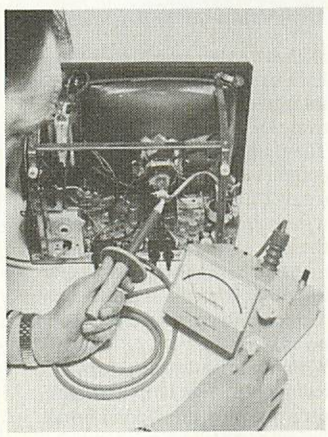


included in the package consisting of the two devices. These devices range in price from under $ 4,000 to over $ 10,000. It is suggested to order as soon as possible since demand has accounted for much of the first years production. Delivery is 20 weeks from date of order.

Reine Electronics B.V.,
Postbus 6730,
Den Haag 2040,
The Netherlands.

direct reading up to 40 kV

The latest high-voltage meter from Brandenburg Limited, the Model 109M, offers direct reading of e.h.t. voltages up to 40 kV. The instrument's very high impedance of 60 GΩ means that the current drain on the circuit under test is very small. This gives the meter a significant advantage over conventional multimeters which can draw as much as 50 μA, thereby reducing the voltage under test. The instrument's rugged construction makes it ideal for routine measurements taken, for example, during cathode-ray-tube servicing.



The instrument uses the latest semiconductor operational amplifiers to give a readout with a linear mirror scale, with divisions of 1 kV. The instrument is battery operated, and incorporates a battery test facility; battery life is 600 h. The high-voltage probe, with a special attachment for c.r.t. measurements, is available as an accessory.
The model 109M costs £ 100 (plus V.A.T.) and the probe £ 5 (plus V.A.T.).

Brandenburg Limited,
High-Voltage Engineering Division,
939, London Road,
Thornton Heath,
Surrey. CR4 6JE.

Model number cavity & processor	Operating frequency-GHz	Suggested list price F.O.B. fact
10E-49 Ka	26.5- 40.0	$ 3,650.00
10E-49 B	33.0- 50.0	$ 3,950.00
10E-49 Q	40.0- 60.0	$ 4,225.00
10E-49 V	50.0- 75.0	$ 4,500.00
10E-49 E	60.0- 90.0	$ 5,050.00
10E-49 R	75.0-110.0	$ 5,625.00
10E-49 N	90.0-140.0	$ 6,475.00
10E-49 T	110.0-170.0	$ 7,325.00
10E-49 G	140.0-220.0	$ 8,450.00
10E-49 Y	170.0-260.0	$ 9,850.00
10E-49 D	220.0-325.0	$ 11,250.00

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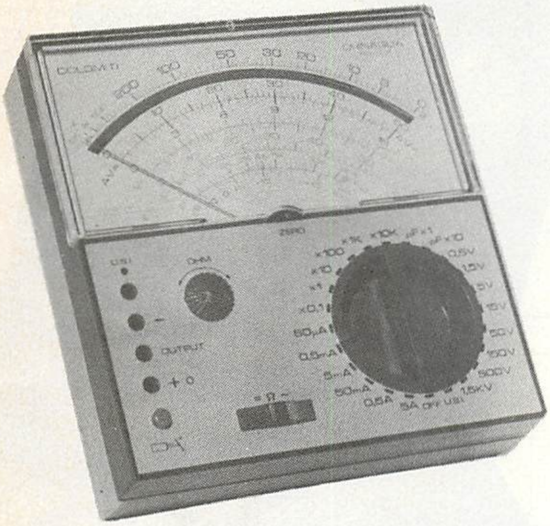

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- modulation systems
- how to gyrate

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publishers announcement

The government has announced that as from Monday 12th of April the higher rate of VAT is reduced from 25 per cent to 12½ per cent. Some of the advertisements in this issue of Elektor may still show VAT at 25 per cent due to lack of time for amendment. We recommend that readers should check with the advertiser regarding the rates quoted before ordering. The publishers can not be held responsible for any error in VAT rates quoted.

As our Editorial Staff will be at the HEDA and Sound and Vision exhibitions in Birmingham from May 23rd to 31st, the Technical Queries 'phone-in' will not operate on Monday May 24th and 31st. We shall, however, be pleased to talk with you in person and answer your queries at our exhibition stand:

Hall No. 3,
Stand No. 5015,
National Exhibition Centre,
Birmingham.

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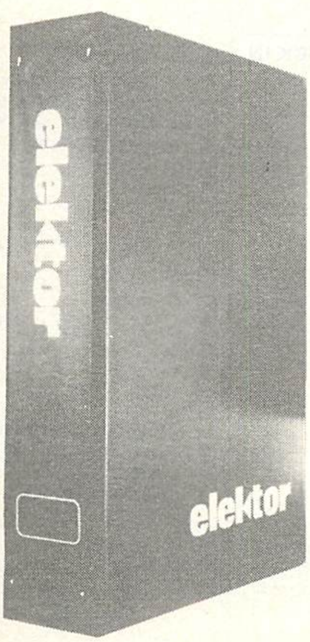

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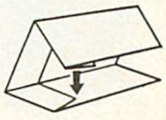

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   Example: The M6900 series is now being defined to meet defined customers' requirements.
3. **Very efficient programme code.**
   Wide instruction repertoire, including seven addressing modes.
4. **Sub-function devices already available.**
5. **Single power rail.** 5 volt.
6. **Interfaces easily** with TTL and CMOS.
7. **Second sourced** by AMI across Europe.

Here's the M6800 family today:—

MC6800	Microprocessor.
MC6820	Peripheral Interface Adapter.
MCM6810	Static RAM.
MCM6830	ROM.
MC6850	Asynchronous Communications Interface Adapter.
MC6860	Low Speed Modem.

Alternative N-Channel Si Gate RAMs for large systems:—

MCM68102	1K x 1	Static	16-pin.
MCM6814	4K x 1	Dynamic	16-pin.
MCM6815	4K x 1	Dynamic	22-pin.

Recent new devices include:—

Dynamic Memory Refresh Controller.

MCM68112A	256 x 4	Static RAM.	16-pin.
MCM68317	16K	Static ROM.	24-pin.

An 8K x 1 erasable and electrically reprogrammable ROM (MCM68708) was introduced in the first quarter of 1976. And there's more to come!

MOTOROLA Semiconductors

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