THE INTERNATIONAL JOURNAL FOR RECREATIONAL ELECTRONICS

October 1990

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Elektor Electronics USA

Medium Power Audio Amplifier

Digital Car Engine Lock

400W Lab Power Supply

Microprocessor Telephone PBX

UHF TV Preamplifier

Guitar Tuner

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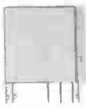
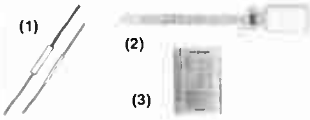
Infrared Module

Complete, Easy-to-Use IR Detector



Heart of a remote-control project! Combines detector, limiter, bandbass, demod, integrator and comparator in a compact 3-lead module. #276-137 .... 3.49

Neater Wiring



(1) **Heat-Shrink Tubing.** 1/16 to 1/2" sizes. #278-1627 ...... **Pkg. of 7/1.79**

(2) **Cable Tie/Markers.** For cables up to 5/8" dia. #278-1648 ... **Pkg. of 10/2.49**

(3) **Wire Markers.** 630 "peel and stick" characters. #278-1650 ...... **Set/1.79**

Tantalum Caps

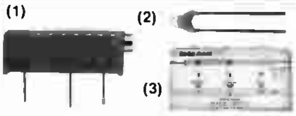
High capacitance in a tiny size.

µF	WVDC	Cat. No.	Each
0.1	35	272-1432	.59
0.47	35	272-1433	.59
1.0	35	272-1434	.59
2.2	35	272-1435	.69
10	16	272-1436	.79
22	16	272-1437	1.19

"Snap" RFI Choke

Snap-Together Toroid Cores. Effective "roadblock" to radio-frequency interference. Just wind on AC, phone, computer, coax or audio cable, then snap choke together. #273-104 ..... **Set of 2/7.95**

Resistance Items



(1) **15-Turn Trimmers. 1k,** #271-342. **10k,** #271-343 **20k** #271-340, Each **1.49**

(2) **Precision Thermistor.** Resistance changes in proportion to temperature. −50 to +110° C. #271-110 ...... **1.99**

(3) **Color-Code Decoder.** #271-1210 .... **79¢**

Shielded D-Sub Hoods

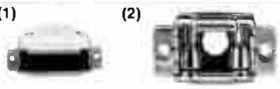


Fig.	Description	Cat. No.	Each
1	9-Position Metal	276-1508	2.19
...	25-Position Metal	276-1510	2.79
2	9-Pos. Metalized	276-1513	1.49
...	25-Pos. Metalized	276-1536	1.99

"Ding-Dong" Chime

Great Entry Alerter

This IC and mini-speaker combo has a pleasant chime output and it's ideal for a customer-entry alert, doorbell or event indicator. Operates from 6 to 18 VDC. Produces 80 dB sound pressure at 12 VDC. 5" leads. #273-071 .... **8.99**

Rod Antennas

For projects or replacing damaged antennas on cordless handsets and bases, walkie-talkies and radios. We have over a dozen styles in stock. For example:

Sections	Extended	Cat. No.	Each
5	13"	270-1407	2.79
6	17¾"	270-1409	2.59
4	34¾"	270-1402	3.79

TNC and Mini-UHF Connectors

(1) **TNC Twist-On Male** for RG-58 Cable. #278-140 .... **1.99**

(2) **TNC Double-Female.** #278-142, **1.99**

Fig.	Accepts	Fits TNC	Cat. No	Each
3	PL-259	Female	278-118	3.49
4	BNC Male	Female	278-145	4.99

(5) **Mini-UHF Double-Female.** Joins two male mini-UHFs. #278-172 .... **2.49**

(6) **Mini-UHF T Adapter.** Female-male-female power divider. #278-173 ... **3.99**

(7) **UHF Adapter.** Accepts PL-259. Fits female Mini-UHF. #278-174 ...... **3.99**

Solderless IDC-Type Connectors

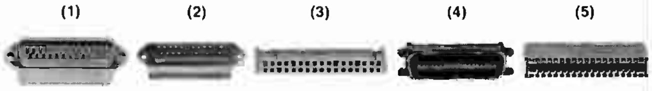


Fig.	Description	Cat. No.	Each
1	25-Position D-Sub Male	276-1559	3.99
2	25-Position D-Sub Female	276-1565	3.99
3	34-Position Header	276-1525	2.49
4	36-Position Male Printer Connector	276-1533	4.99
5	34-Position Card-Edge Connector	276-1564	1.99

Electronic Counter Module

Prewired! Great for a traffic or event counter. Counts up to 99,999. Advanced (or reset) by external switch closure or digital circuit. Accepts count rate up to 7 Hz. Has 1/2" liquid crystal display. Requires "AA" battery. With data. #277-302 .... **16.95**

Foldup Autoranging DVM

Select the function and this precision Micronta® meter sets the range. Has automatic power-off when shut, detented hinge for best viewing angle. Measures to 1000 VDC, 500 VAC, 10A AC/DC. 2 megohms resistance. Requires 2 "AA" batteries. #22-193 .... **69.95**

Dual-Power Soldering Iron

Go from 15 to 30 watts with the flick of a switch. Has replaceable tip. 8¼" long. UL listed AC. #64-2055 .... **9.49**

Replacement Tip. #64-2065 .... **1.19**

Chisel Tip. #64-2056 .... **1.19**

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Elektor Electronics USA

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In our next issue:

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- Receiver without PSU
- Active cross-over filter for sub-woofer
- The miser's T / R loop
- Smoke detector
- 1-of-N decoder

Front cover

Blind people will soon be able to' read' or listen to instant digital versions of daily newspapers if a new system unveiled at the Royal Institute for the Blind (RNIB) in London is successful.

The Institute's technology department is testing a system that gives blind people access to daily news within hours of publication. At present they must wait for weekly extracts on cassette tape or ask relatives or friends to read the newspaper to them.

In a trial project with The Guardian, text is transmitted over the television network and received in the homes of blind people with access through an authorized screen decoder card in a personal computer. This allows them to 'read' the latest news with the use of a speech synthesizer, as shown, or a transient braille display. The latter option is particularly useful to deaf-blind people whose access to any kind of information is very restricted.

RNIB, 224 Gt Portland St, London W1N 6AA

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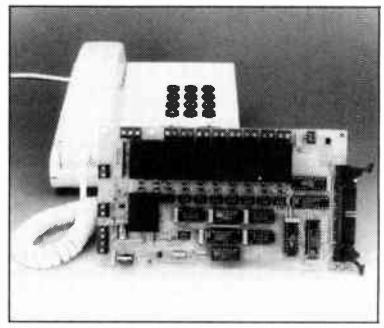

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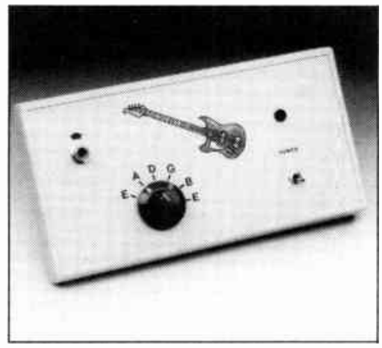

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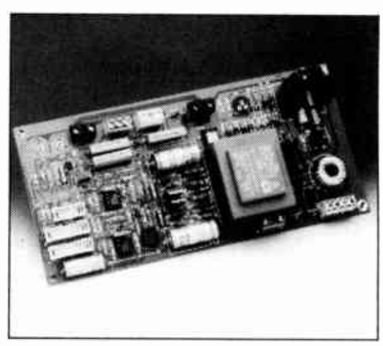

400-watt laboratory power supply p. 49

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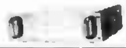
Special Offer

Amplifier	TA-3600	TA-477	TA-802	TA-1000A
+ Metal Cabinet	LG-1925	LG-1925	LG-1924	LG-1924
+ Transformer	#007	#003	#001	#001
Kit Set	$ 138.60	$ 115.20	$ 91.75	$ 104.12
Assmb. Set	$ 160.20	$ 130.50	$ 104.15	$ 122.92


Transformer


Metal Cabinet

- Due to this special offer & low price, we can only exchange or repair any of these returned units (TA-3600, TA-477, TA-802 & TA-1000A) within 15 days of your receipt.
- The above prices are for the complete sets. For separate unit price, please see aside.

300 W HI-FI POWER AMPLIFIER (MONO) TA-3600 ▲▲▲

The TA-3600 is an extremely high power amplifier specifically designed to reproduce the high dynamic range available on cc..pact discs.

It has low noise, high stability low distortion, extended frequency range and high efficeincy in a compact package. Two of these units with appropriate power supply and pre-amplifier are required for stereo reproduction.

Complete kit: $79.00
Assembled & tested: $103.00
Transformer: $42.00
10,000 µ/F 80V E. Cap: 22.00

SPECIFICATIONS

Power output 300 watts sine wave into 8 ohms, 540 watts music power into 8 ohms • Frequency response. 10Hz to 20 KHz • Total harmonic distortion Less than 0.05% • Intermodulation distortion Less than 0.05% • Sensitivity. 1 V ms at 47K • Power requirements: 60 to 75 VDC at 8 amp • Dimensions: 8" x 7¼" x 2¼"

120W MOSFET POWER AMPLIFIER (MONO) TA-477 ▲▲▲

This amplifier is designed specifically for low TIM distortion. It uses parallel MOSFET output transistors. This is a very popular kit because of its superior sound reproducing ability and high power reserve. BUILD TWO KITS FOR A PROFESSIONAL STEREO AMP!

SPECIFICATIONS

T. H. D. Less Than 0.007% • Power output: 120 watts into 8 ohms • Frequency response: 8 Hz to 20KHz, +0-0.4 dB • Sensitivity. 1V • Power requirements. 55 VDC @ 3 amps.

Complete Kit $68.00
Assembled & Tested $85.00

100W DYNAMIC CLASS A MAIN POWER TA-1000A ▲▲▲

This well designed and stable amplifier features DC coupling throughout and a fully complementary symmetry circuit topology. It operates in class A at low and medium levels and gradually shifts to class AB operation at high levels. A pair of these amplifiers provide 200 watts of super high fidelity sound reproduction.

SPECIFICATIONS

Power output: 100 watts into 8 ohms, 125 watts into 4 ohms • Frequency response 10Hz to 100 KHz • Total Harmonic Distonon: Less than 0.008% • Signal to noise ratio : Better than 80 dB • Sensitivity : 1V • Power requirements: 35 to 45 VDC @ 3 amp • PCB Dimensions 5.38" x 3.38" x 1.2" • Heat sink: 5.13" x 2.63 3" (Each).

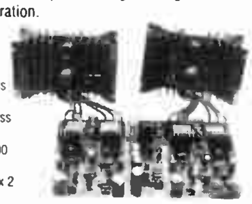
Complete Kit $59.69
Assembled & Tested $80.58

80W + 80W PURE DC STEREO MAIN POWER AMPLIFIER TA-802 ▲▲▲

Modern circuit design utilizes full complementary power output and differential cascode input stage. DC coupling provides superb transient response. A special output circuit guards against short circuits and overload operation.

SPECIFICATIONS

Power output:: 80 watts per channel into 8 ohms
Total harmonic distortion: Less than 0.05% at rated power
Intermodulation distortion. Less than 0.05% at rated power
Frequency response. DC to 200 KHz, -0 dB, -3dB, @ 1 watt
Power requirements. 30 VAC x 2 @ 6 amp. May use Mark V model 001 transformer
P.C. Board: 8-1/2" x 5" x 1-1/8"
Heat sink: 5-1/8" x 2-5/8" x 3" (Each)

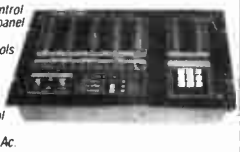


Complete Kit $45.94
Assembled & Tested $59.72

PROFESSIONAL COLOR LIGHT CONTROLLER SM-328 ★

The SM-328 professional color light controller is keyboard programmable for ease of use. It allows full control of intensity and flash rate. It has four separate channels with capacity of 1170 watts per channel. Total wattage capability is 4.68 killowatts. This is equivalent to 46 pcs. 100watt ligh bulbs or 936 pcs. 5-watt colored bulbs and is sufficient for the largest halls and auditoriums.

- *Independent input signal control*
- *Professional styled control panel*
- *4 independent outputs*
- *4 independent dimmer controls*
- *Chaser speed controls*
- *Automatic chaser operation*
- *4 preset chaser programs*
- *Clockwise chaser control*
- *Anti-clockwise c'laser control*
- *Match with any step-down transformer from 12 to 24V Ac.*



SPECIFICATIONS

Input sensitivity (music mode) 100mV (music + program) : 2V • Output power 1170 W per channel MAX, 4680 W total • Power requirements. 105 - 120 V, 60 Hz • Dimensions. 14.32" wide, 9" high, 3.19" deep

Assembled & Tested: ~~$150.00~~
Now: $139.00

Summer Sale

VIDEO/AUDIO SURROUND SOUND PROCESSOR SM-333 ▲▲▲★

It has inputs for VCR, Video Disc and Compact Disc and can also be used with FM tuners, tape decks and LP discs - in fact anything that can be connected to your preamplifier.

It also has front panel control of dynamic noise reduction DNR*, surround level, effect and time delay up to 50 milliseconds.

Rear panel outputs for front and rear stereo speaker amplifiers. The unit may be purchased as a kit or fully assembled ready to plug in.

SPECIFICATIONS

Frequency response. 20 Hz to 20KHz • Total harmonic distortion, Front channel. 0.05%, Rear channel. Less than 0.25% • Input signal voltage: 0.1 to 3.5V • Output. Front channel: 0.1 to 3.5V • Rear channel: 6.6V • Delay time: 5 to 50 milliseconds • Input impedance: 47 K • Power requirments: 100-120VAC, 60Hz • Dimensions: 14.2" wide, 4.82" deep, 2.1" high.

- TM OF NATIONAL SEMICONDUCTOR CORP.

Summer Sale

Complete Kit: $62.00
Assembled & Tested: ~~$83.00~~
Now: $70.00

ELECTRONIC ECHO AND REVERBERATION AMPLIFIER TA-2400A ★

The TA-2400 provides controlled reverberation for special sound effects. It has variable delay and depth. Great for use with tape recorders, discos, rock bands. Features LED display.

SPECIFICATIONS

Sensitivit 150 mV Input impedance: 50K Frequency response 10 Hz to 50 KHz, ± 1 dB Total harmonic distortion: Less than 0.5% Maximum input level 2 V Signal to noise ratio: 90 dB Reverberation time: 0 to 1.5 sec max Output level: 150 mV Power requirements 110 V 60 Hz, 13 W Dimensions 19"W x 2-3/8" H x 8-1/8" D

Summer Sale

Assembled & Tested: ~~$118.00~~
Now: $96.00

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AMPLIFIERS

MODEL	DESCRIPTION	KIT	ASSEMB.
TA-001	1W Mini-Amplifier ▲	$ 5.87	$ 8.50
TA-006	6W Mini-Amplifier ▲	7.50	9.85
TA-007	12W Mini-Amplifier ▲	12.05	15.30
TA-10	Stereo Pre Amp. w/Magnetic Mic Amp. ▲	8.86	11.25
TA-28MK2	Digital Voice Memo ▲▲	30.00	
TA-50A / B	Multi-Purpose Melody Generator ▲	12.84	17.20
TA-50C	Multi-Purpose Melody Generator ▲	13.65	18.71
TA-120MK2	35W Class "A" Main Power Mono Amp. ▲▲	28.50	39.80
TA-300	30W Multi-Purpose Single Channel Amp. ▲	20.00	29.00
TA-302	60W Stereo Power Booster (w/case)★		70.00
TA-323A	30W X 2 Stereo Pre-main Amp. ▲	29.50	38.35
TA-377A	Hi-Quality FET Stereo Pre-Amp. ▲▲▲	59.95	75.00
TA-400	40W Solid State Mono Amp ▲	28.00	34.93
TA-477	120W Mosfet Power Mono Amp. ▲▲	68.00	85.00
TA-800	80W + 80W DC Pre Main & Power Amp. ▲▲	60.92	79.20
TA-802	80W + 80W DC Stereo Main Power Amp. ▲▲	45.94	59.72
TA-820A	60W + 60W OCL DC Pre-Main Stereo Amp. ▲▲	40.39	49.37
TA-1000A	100W Dynamic Class "A" Main Power Mono Amp. ▲▲	59.69	80.58
TA-1500	100W X 2 Class "A" DC Stereo Pre-Main Amp. ▲▲▲★	73.70	95.81
TA-2200	Fet Super Class "A" DC Pre-Amp. ▲▲▲	47.70	58.24
* TA-2400A	Electronic Echo & Reverberation Amp. ★		96.00
* TA-2500	HD Pre-Amp w/10 band graphic equalizer ★		68.80
TA-2800	HI-FET IC Pre-Amp w/3 way tone control ▲▲	48.90	63.57
* TA-3000	Stereo Simulator (For Mono TV or Any Mono Source)▲▲	27.00	38.50
TA-3600	300W HD Hi-Fi Power Mono Amp ▲▲▲	79.00	103.00

MISCELLANEOUS

MODEL	DESCRIPTION	KIT	ASSEMB.
TY-1A	Battery Flourescent Light Driver ▲	7.50	9.50
TY-7	Electronic Touch Switch ▲	8.20	10.20
TY-8	Electronic Lotto ▲	15.00	21.00
TY11A	Multi-Functional Control Switch ▲	6.19	8.79
TY-12A	Digital Clock w/Timer ▲	17.95	26.00
TY-13	Color Led Audio Level Meter ▲	20.15	26.20
TY-14	Electronic Shock ▲	7.25	
TY-18	High Precision Sound Control Switch ▲	10.22	14.50
TY-20	V Shape Color Led Level Meter ▲▲	21.45	
* TY-23B	3 Channel Color Light Controller ▲▲▲★	51.20	65.00
TY-25	Stereo Loudspeaker Protector ▲	13.85	19.85
TY-35	FM Wireless Microphone ▲	10.22	
TY-36	AC/DC Quartz Digital Clock ▲	19.00	26.20
TY-38	Sound / Touch Control Switch ▲	12.00	
TY-41MK V	Infared Remote Control Unit (w/Case) ▲▲▲	20.00	35.00
TY-42	Bar / Dot Level Meter ▲▲	24.15	33.81
TY-43	3 Digital Panel Meter ▲	29.00	38.00
TY-45	20 Steps Bar / Dot Audio Level Display ▲▲	38.45	46.14
TY-47	Superior Electronic Roulette ▲▲	19.46	27.24
SM-222	7 Band HI-FI Graphic Equalizer ▲ ▲ ▲	26.80	38.80
* SM-328	4 Channel Professional Color Light Controller▲		139.00
* SM-333	Audio/Video Surround Sound Processor ▲▲▲★	62.00	70.00
SM-666	Dynamic Noise Reduction ▲	26.00	34.00
10,000 µ/f 80V Capacitor (for-TA-3600; TA-477, TA-1000A & TA-802)			22.00

POWER SUPPLIES

MODEL	DESCRIPTION	KIT	ASSEMB.
TR-100A	0-15V 2A Regulated DC Power Supply (W/Case)▲▲▲★		69.50
TR-355A	0-15V 5A Regulated DC Power Supply ▲	14.55	20.76
TR-355B	0-30V 3A Regulated DC Power Supply ▲	14.55	20.76
TR-503	0-50V 3A Regulated DC Power Supply ▲ ▲	15.75	22.65

INSTRUMENTS

MODEL	DESCRIPTION	KIT	ASSEMB.
SM-43	3½ Multi-Functional Led D.P.M. (w/ABS plastic case)▲▲	34.50	43.00
SM-48	4½ Hi-Precision D.P.M.▲▲▲	38.00	48.00
SM-48A	4½ Hi-Precision D.P.M. (w/APB plastic case)▲▲▲	41.20	52.00
SM-49	3½ Multi-Functional LCD D.M.P. (w/Hold Function)▲▲	36.00	44.50
SM-100	150MC Digital Frequency Counter▲▲▲	79.00	90.00
FC-1000A	1 GHz Frequency Counter ★		179.00

METAL CABINETS WITH ALUMINUM PANEL

MODEL	H" x W" x D"	MATCHING	PRICE
LG-1273	3" 12" 7"	TA-2800, TA-377A, TA-2200	$22.16
LG-1684	4" 16" 8"	TA 323A, TA-377A, TA-2200	26.64
LG-1924	4" 19" 11½	TA-802, TA-820A TA-1500, TA120MK 2, TA-800, TA-1000A	32.00
LG-1925	5" 19" 11½	TA-477, TA-800, TA-1500, TA-1000A TA 3600	35.00
LG-1983	2¾" 19" 8"	TA-377, TA-2800, TA-2200, TA-120MK 2	28.50

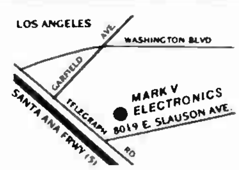
POWER TRANSFORMERS

MODEL	DESCRIPTION	MATCHING	PRICE
#001	56VCT, 60VCT 6A	TA-800, TA-802, TA-820A, TA-1000A, TA-1500	$26.00
#002	72VCT 3A	TR-503, TA-323A, TA-400, TA-300	21.00
#003	80VCT 6A	TA-477	27.00
#004	48VCT 6A	TA-120 MK 2	21.00
#005	52VCT 3A	TR-355B	15.00
#006	36VCT 5A	TR-355A	14.50
#007	112VCT 8A	TA-3600	42.00

CLOCKS / THERMOMETERS

MODEL	DESCRIPTION	PRICE
T-1	LCD Thermometer Clock w/ In-Outdoor Sensor	$15.40
T-2	LCD Thermometer Clock w/ F & C Measurement	14.40
#8501	Parrot Talking Clock	12.50
#8504	Mynah Talking Clock	12.50

* Summer Sale for limited quantities



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J. GORDON HOLT

If you had to invent a new language, where would you begin?

Back when high quality sound reproduction was a new idea and **J. Gordon Holt** was a staffer at *High Fidelity* magazine, manufacturers and journalists alike depended on the simple technical quality tests which everyone accepted as the yardsticks for performance. As the industry grew, equipment got better, competition fiercer, and technical reviewing became more crucial to sales managers. Before long, **J. Gordon** began to realize that reviewing was becoming more and more accommodating, and where the reviewers continued to rely on the standard tests, the measurement data began to look more and more alike.

Finally, in frustration, **Holt** left Great Barrington and headed for home in Pennsylvania where he founded *Stereophile* magazine in the spare room of his mother's house. He became convinced that although equipment tests and measurements were important, they no longer accounted for the differences he could hear. Two devices could easily measure the same and yet sound quite different.

Holt abhorred the tendency of the larger magazines to depend almost entirely on measurements, which he saw as a safe way to review without disturbing the manufacturer with any bad news. Not only that, he realized that not one of the US audio publications was publishing reviews that were critical of equipment. In fact, in some cases they were ignoring some flaws.

However, if the reviewer wishes to review how equipment sounds, he faces a severe problem. Our sense of hearing has the smallest vocabulary of any of our five senses. Thus, **Gordon** faced the difficulty of describing sound differences with all too few words with which to do it. He not only had to invent the techniques and disciplines of what has become known as "subjective reviewing" but also the language with which to do it.

Today, the magazine he founded has become a major force in audio quality judgments around the world. And almost all the vocabulary definitions are his work.

Seldom will you have the opportunity to purchase a reference work backed by so much primary research and experience. Few reviewers have spent more time and energy in an honest search for a defined, factual account of what matters in good sound reproduction techniques. In audio equipment reviewing, **J. Gordon Holt** is not only a pioneer but a master.

THE AUDIO GLOSSARY is the first publication of not only a vocabulary for sound description, but also a comprehensive overview of over 1,900 audio terms: technical and subjective. Your copy is waiting—at your favorite dealer—or order direct using the coupon below.

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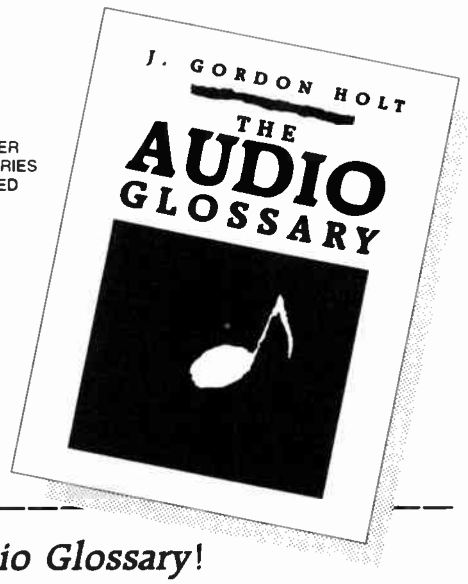

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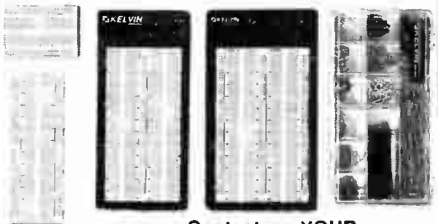

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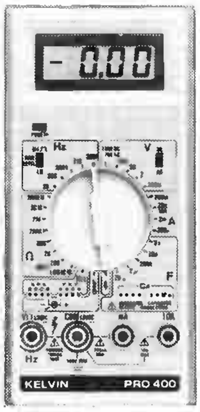

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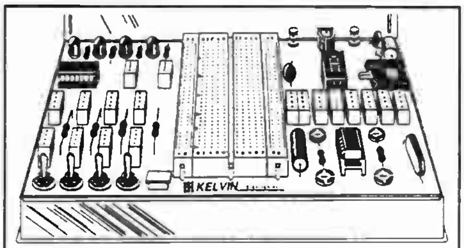

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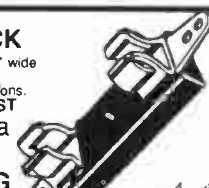

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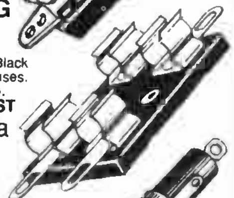

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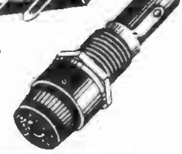

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KH-2: SPEAKER SAVER AND OUTPUT FAULT DETECTOR. [3:77] This basic two-channel kit includes board and all board-mounted components for control circuitry and power supply. It features turn-on and -off protection and fast opto-coupler circuitry that prevents transients from damaging your system. The output fault detector has additional board-mounted components for speaker protection in case of amplifier failure. **$65**

KF-6: 30Hz RUMBLE FILTER. [4:75] This kit implements a 1975 design for a low frequency garbage filter. The filter knee is set to 30Hz. Roll-off below that knee is the 18dB/octave characteristic of its three pole design. Gain for the filter is unity (0dB) but can be simply adjusted for up to 12dB of gain. The reprint of the article explores the use of the filter with other components in crossovers (see kits SBK-C1A, C1B). It shows how to obtain slopes of 6, 12 or 18dB in high and low pass filters. The kit contains all parts for building a two channel HPF including a board (3" × 3"), quad op amp IC, precision resistors and capacitors. Requires a bipolar supply of ± 15V—the KE-5 is suitable. **$30**

AIDS & TEST EQUIPMENT

KK-3: THE WARBLER OSCILLATOR. [1:79] This unit will produce a swept signal covering any ⅓-octave between 16Hz and 20kHz. The total harmonic distortion at the output is less than 1.5%. The output voltage is adjustable from 0 to 1V. When used with a microphone it is as effective as a pink noise source in evaluating speaker system performance. It also reveals the listening environment's effect on sound through reflection and absorption. The sweep rate is set at about 5Hz. The kit includes 3¼" x 3⅞" circuit board, transformer, all parts and article reprint. **$70**

KH-7: GLOECKLER PRECISION 101dB ATTENUATOR. [4:77] All switches, 1% metal film and 5% carbon film resistors to build prototype. Chassis, input/output jacks are not included. **$65**

KC-5: GLOECKLER 23-POSITION LEVEL CONTROL. [2:72] All metal film resistors, shorting rotary switch and two boards for a two-channel, 2dB per step attenuator. Choose 10k or 250kΩ. **$48**

KL-6: MASTEL TIMERLESS TONE BURST GENERATOR. [2:80] All parts with circuit board. No power supply. **$24**

KP-2: TWO TONE INTERMODULATION TEST FILTER. [1:82] This filter is designed to isolate the two high frequency tones at an amplifier's input from low frequency intermodulation products present at the output. The high pass filter corners at 2kHz and rolls off at 24dB/octave. A 5kHz signal at the low pass input will be down at the output by 80dB. An article reprint detailing design and use is included with the kit. All parts are supplied, including quad op amp IC, circuit board and precision resistors and capacitors. **$26**

SBK-D2: WITTENBREDER AUDIO PULSE GENERATOR. [SB 2:83] All parts, board, pots, power cord, switches and power supply included. **$80**

SBK-E4: MULLER PINK NOISE GENERATOR. [SB 4:84] All parts, board, 1% MF resistors, capacitors, ICs, toggle switches included. No battery or enclosure. **$35**

CROSSOVERS

KC-4A: ELECTRONIC CROSSOVER, KIT A. [2:72] Single channel, two-way. All parts including C-4 board and LF351 ICs. Choose frequency of 60, 120, 240, 480, 960, 1920, 5k or 10k. KE-5 or KF-3 supplies are suitable. **$14**

KC-4B: ELECTRONIC CROSSOVER, KIT B. [2:72] Single channel, three-way. All parts including C-4 board & LF351 ICs. Choose two frequencies of 60, 120, 240, 480, 960, 1920, 5k or 10k. **$18**

KK-6L: WALDRON TUBE CROSSOVER LOW PASS. Single channel, 18dB/octave, Butterworth [3:79], includes three-gang pot. Choose 1: 19-210; 43-465; 88-960; 190-2100; 430-4650; 880-9600; 1900-21,000 hertz. **$60**

KK-6H: WALDRON TUBE CROSSOVER HIGH PASS: Single channel, 18dB/octave, Butterworth [3:79], includes three-gang pot. Please specify 1 of the frequencies in KK-6L. No other can be supplied. **$62**

KK-7: WALDRON TUBE CROSSOVER POWER SUPPLY. [3:79] Includes board, transformer, fuse, semiconductors, line cord, capacitors to power four tube crossover boards (8 tubes), 1 stereo bi-amped circuit. **$110**

SBK-A1: LINKWITZ CROSSOVER/FILTER. [SB 4:80] Three-way crossover/filter/delay. 24dB/octave at 100Hz and 1.5kHz and 12dB/octave below 30Hz, with delayed woofer turn-on. Use the Sulzer supply KL-4A with KL-4B or KL-4C.
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SBK-C1A: ELECTRONIC TWO-WAY CROSSOVER. [SB 3:82] 30Hz filter with WJ-3 board & 4136 IC adapted as one channel crossover. Can be 6, 12 or 18dB/octave. Choose frequency of 60, 120, 250, 500, 1k, 2k, 5k or 10k. The KL-4A/KL-4B or KW-3 are suitable supplies. **$32**

SBK-C1B: THREE-WAY, SINGLE CHANNEL CROSSOVER. [SB 3:82] Contains 2 each SBK-C1A. Choose high & low frequency. **$60**

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KW-3: BORBELY IMPROVED POWER SUPPLY. [1:87] This single channel, low impedance supply was designed for the exacting requirements of Erno Borbely's moving-coil preamp [2:86, 1:87]. The design utilizes polypropylene caps and 1% metal film resistors. LM317/337s are used in the preregulator and Signetics NE5534 in the op amp regulator. The kit includes a low profile 24V toroidal transformer, 4¼" × 5½" circuit board and all board-mounted components. Chassis and heatsink are not included. **$135** Two or more, **$128**

KE-5: OLD COLONY POWER SUPPLY. Unregulated, ± 18V @ 55mA. **$20**

KF-3: GATELY REGULATED SUPPLY. ± 18V or ± 15V @ 100mA. **$52**

KL-4A: SULZER POWER SUPPLY REGULATOR. **$40**

KL-4B: SULZER DC RAW SUPPLY. ± 20V @ 300mA. **$60**

KL-4C: SULZER DC SUPPLY w/ toroidal transformer. **$85**

KH-8: MORREY SUPER BUFFER. [4:77] All parts, 1% metal film resistors, NE531 ICs, and PC board for two-channel output buffer. **$22**

SBK-E2: NEWCOMB NEW PEAK POWER INDICATOR. [SB 2:84] All parts & board, new multicolor bar graph display; red, green & yellow LEDs for one channel. No power supply needed. **$14** Two for **$22**

KL-2: WHITE DYNAMIC RANGE & CLIPPING INDICATOR. [1:80] One channel, including board, with 12 indicators for preamp or crossover output indicators. Requires ± 15V power supply @ 63 mils.
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KX-1A: DISC STABILIZER. Set of 3 Sorbothane feet, 3 Tiptoes and Mod Squad's Disc Damper with 15 centering rings. **$70**

KY-1: BEERS' BUDGET CD MOD. [1:89] Kit provides POOGE-4 improvements without additional wiring or circuit boards. Complete parts for assembling amplifier modules and replacing DAC components. Article reprint included. Soldering skills required: not recommended for beginners. **$95**

What's Included? Kits include all the parts needed to make a functioning circuit, such as circuit boards, semiconductors, resistors and capacitors. Power supplies are not included in most cases. Unlike kits by Heath, Dyna and others, the enclosure, faceplate, knobs, hookup wire, line cord, patch cords and similar parts are not included. Step-by-step instructions usually are not included, but the articles in *TAA* and *SB* are helpful guides. Article reprints are included with the kits. Our aim is to get you started with the basic parts—some of which are often difficult to find—and let you have the satisfaction and pride of finishing your unit in your own way.

A HAND UP FOR HANDS-ON

If the 1990s are distinguished for nothing else, they may at least be regarded by future historians as the decade when do-it-yourself was reborn in the United States. For some time now, a sea-change about doing it yourself has been evident in both print and electronic media.

Time-Warner is eager to sell you manuals for repairing or remodeling your house. PBS fills larger and larger portions of prime time with programs on both house refurbishing, cooking, art, and learning the basics about your car—and how to make a wise decision in buying your next one.

During the mid-sixties magazine managements began to drop do-it-yourself articles, with the exception of a few clearly committed to hands-on technology such as *Popular Mechanics* and others.

Such articles began to disappear from some of the most popular and largest of the electronics magazines. When Billboard, Inc. bought *Audiocraft* it effectively killed the latter whose electronic construction articles disappeared without a trace within two issues of *High Fidelity*. *Audio* magazine, which had begun life as almost exclusively an audio construction journal, evolved in the seventies into a predominantly consumer-type publication.

Ziff-Davis dropped *Electronics World* in the early seventies. *Popular Electronics* announced in the early seventies they would no longer publish construction articles—but relented later due to reader outcry. In time, they turned to computer technology, attempted to turn the magazine into a full-blown computer journal and were promptly deserted by their advertisers. Gernsback recently purchased and resurrected the *Popular Electronics* title.

But times have changed. Americans have become more than a little disenchanted with manufactured products. Older technologies and crafts have resurfaced. Elegant articles, hand-made in small quantities, have become collector's items with prices to match. Americans are once again rediscovering their own hands and finding, to their surprise, that they are not, as the magazines attempted to convince them, "inept klutzes" whose work would inevitably be inferior to what a professional company could manufacture for them.

We have also seen the electronics industry reach the point where many engineers, who began their careers as designers and relished the pleasure of seeing a design become a prototype, are sitting behind desks today, shuffling papers and doing something called "administration." The creative urge to solve a problem, wrestle with the impossibilities, and see the right set of compromises succeed is sorely missed by the majority of those I talk with. Building a project fulfills their unique need.

Building a device is the acid test of its viability. And we seldom acquire a full understanding of the device until we build it. One of the primary definitions of our uniqueness as humans is our ability to fashion tools, which, happily for our humility, we share with some of the higher animals. But if we leave the making of things to a smaller and smaller proportion of the race, we will inevitably impoverish ourselves, individually and collectively.

Our company is very proud to add *Elektor Electronics USA* to its roster of publications. For more than 20 years we have been committed to the value and the pleasure of recreational electronics. In my view, this North American edition of a distinguished European publication is another step in the same direction we have been headed since 1970.

I believe every publication functions as an information interchange point rather than merely a one-way conduit for knowledge. Therefore, I welcome your correspondence and commentary on what we publish. I welcome your personal involvement with the magazine in the matter of letters and classified ads which we are happy to publish. And although *Elektor Electronics USA* is primarily a reader-supported publication, we cannot build good equipment for ourselves without good suppliers who make easily accessible the components and tools we need.

We also welcome your suggestions about ways in which this edition of *Elektor* can better serve your needs. Although a high percentage of the articles are written by European staff writers, authors who wish to write for *Elektor* are welcome to submit a precis of the proposed article and request our outline of style and procedure.

We welcome you to this first issue of *Elektor Electronics USA* and look forward to a long and mutually satisfying relationship. ■

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ELECTRONICS SCENE

BATTERY-POWERED DIRECTIONAL WATTMETER

Bird Electronic Corp. is offering the Model 4410A portable, battery-powered THRU-LINE® RF directional wattmeter with seven power ranges per element. Accurate measurement of signal power is available within ±5% of reading. Standard elements provide frequency ranges from 0.2-2,300MHz and power ranges from 0.002-10,000W. Special elements provide measurements at frequencies as low as 50kHz.

It's ideal for field-service work, laboratories where high accuracies and low power levels are required, and any application where accurate measurements at milliwatts, watts or kilowatts must be performed simply, quickly, and economically. The 4410A includes a standard 9V alkaline battery.

Other versions are available: Model 4411, battery-powered portable or 115/230V AC (50/60Hz) operation; Model 4412, rechargeable Nicad battery or 115/230V AC (50/60Hz) operation; and for fixed location use, Model 4410P, 19″ rack panel-mounted configuration with battery or 115/230V AC (50/60Hz) operation.

The 4410A contains an amplifier employing a self-balancing measurement technique. Its patented bridge circuit has four legs divided between the base and each of the proprietary plug-in elements. The bridge circuit allows optimum reading accuracies, with a 5,000-to-1 dynamic element range. It is unaffected by temperature extremes.

Elements for the 4410A plug into the element socket and rotate for forward or reflected measurements. The seven overlapping power levels provided by each element cover 0.002-10W, 0.02-100W, 0.2-1,000W or 2-10,000W. The desired range is instantly selectable by turning a rotary switch on the wattmeter cover. A battery test position is located on the switch.

For information, contact Bird Electronic Corp., 30303 Aurora Rd., Cleveland, OH 44139, or call (216) 248-1200.

2-WATT PORTABLES FROM MIDLAND LMR

Midland LMR has released two-channel portable radios. These crystal-controlled, FM, two-way radios are available in either VHF (model 70-132) or UHF (model 70-232) and feature pre-set automatic noise squelch and optional CTCSS (tone-coded squelch).

The new Midland portables meet or exceed MIL SPEC 810C/D specifications for shock and vibration. Size and weight have been reduced through micro-circuit technology and the small new high-energy batteries. They weigh less than 12 ounces, including battery pack, and provide about eight hours service per battery charge.

The portables offer belt-clip backs, and accessories include external speaker/microphone, battery chargers, and screw-on stubby helical antennas.

For more information, contact Midland LMR, Marketing Department, 1690 N. Topping, Kansas City, MO 64120, or call 800/MIDLAND (800-643-5263), Ext. 1690.

FIELD SERVICE PARTS TESTER

B&K-Precision is offering a new hand-held parts tester with digital readout, designed for field service or general industrial applications. Model 815 tests capacitance and resistance in a variety of components, as well as transistors, SCRs, diodes, LEDs and batteries in 26 ranges.

Designed to withstand a five-foot drop, the 815 is also water and overload resistant. Its case seals out grease, dirt and other contaminants.

The Model 815 tests capacitance from 0.1pF to 20μF in capacitors, cables, switches, and other components, with accuracies ranging from 0.75% to 1.5%, resistance measurement spans from 0.1Ω to 20mΩ. Transistors are tested by measuring h_{FE} values (gain) and I_{CEO} (leakage). SCRs, diodes, and LEDs are tested for forward junction voltage. Batteries are tested under load for voltage output.

Other features include a 3½ digit 0.8″ LCD readout, tilt stand for bench use, test leads, and component insertion sockets.

The B&K-Precision parts tester has numerous applications in electronic and appliance service, industry, and education, and is suited for the hobbyist as well. The 815 is available at local B&K-Precision distributors at a suggested price of $99.

Contact B&K-Precision, Division of Maxtec International Corp., 6470 W. Cortland St., Chicago, IL 60635. Telephone: (312) 889-9087.

DIGITAL SIGNAL PROCESSING BOARD

Dalanco Spry has developed the Model 250 Digital Signal Processing Board with analog and digital I/O for the PC/AT, and bus compatible microcomputers. Typical applications of this product include instrumentation, machine control, data acquisition, and digital audio.

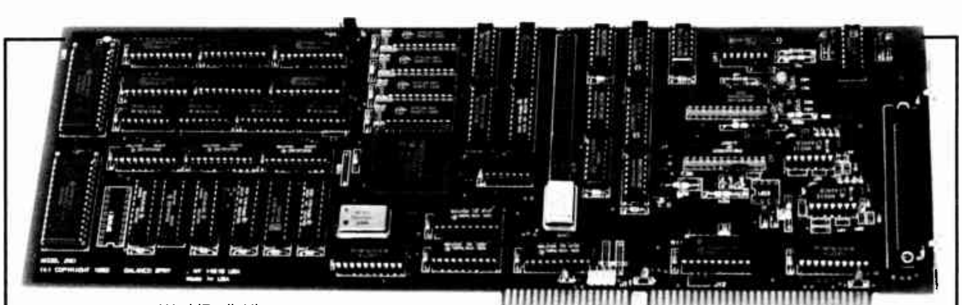
The Model 250 is based upon the 40MHz Texas Instruments TMS320C25 DSP, and can accommodate the faster TMS320C25-50, the EPROM based TMS320E25, and the newer TMS320C26. It may be operated in stand-alone mode in embedded systems when equipped with the TMS320E25 DSP.

The model provides data acquisition for eight single-ended channels at 12-bit resolution and a maximum 300kHz sampling rate. Two analog output channels are provided, as are a buffered digital I/O expansion connector, and the serial (codec) interface of the TI DSP.

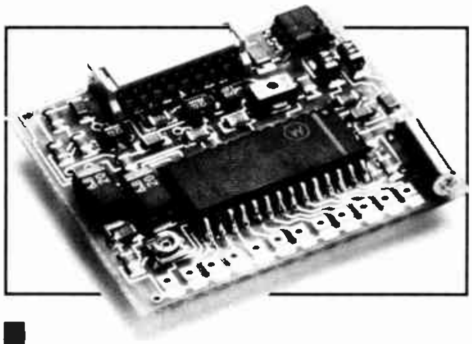
The board may be populated with up to 64K words of 0 wait state Program RAM and 128K words of 1 wait state Data RAM. The Data RAM is simultaneously available to the PC and to the TMS320 DSP through an on-board memory controller.

The following software is bundled with the Model 250: Assembler, Debugger, FFTs, Signal Spectrum Display, Digital Filter Examples, Record and Playback to/from Disk, and a Waveform Editor.

The Model 250 is priced from $1,095. (40MHz TMS320C25, 4K words of Program RAM, 32K words of Data RAM). For additional information, contact David Langmann at (716) 473-3610, or write Dalanco Spry, 89 Westland Ave., Rochester, NY 14618.



ELECTRONICS SCENE

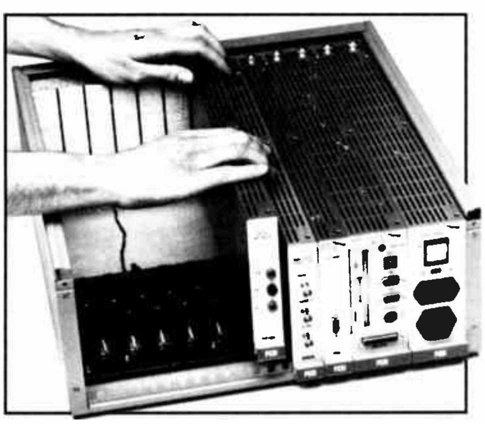


DIGITAL ENCODER-DECODER

A microminiature digital coded squelch encoder-decoder is now available from **Communications Specialists, Inc**. The DCS-23 is the latest in a line of add-on tone signaling products. It is compatible with all DCS systems such as: Digital Private Line, Digital Channel Guard, Digital Call Guard, among others. The new board uses surface mounted technology and measures 1.36″ × 1.18″ × 0.25″ which permits installation in all mobile and most portable radios.

All industry standard digital codes are field programmable using simple PCB jumpers. The board's design uses a crystal controlled CMOS microprocessor, which permits operation on low 6 to 20V DC at 8mA. All connections are made with color-coded jumper wires connected to a microminiature plug and socket. The DCS-23 price is $59.95, and is covered by a one-year warranty. Same day delivery is available from local stock. An illustrated brochure and instruction sheet are available at no cost.

For more information and a complete product list contact: Communications Specialists, Inc., 426 West Taft Ave., Orange, CA 92665-4296, or call (800) 854-0547 or (714) 998-3021. 24-hour FAX (714) 974-3420.



PCXI: NEW MODULAR INDUSTRIAL PC

Rapid Systems has developed a modular PC designed for industry. PCXI makes the PC a tool for demanding test instrumentation. It accepts any PC instrumentation, data acquisition, or control cards manufactured to the Industry Standard Architecture (ISA). These cards are mounted in metal shielded modules, which plug into a passive backplane.

PCXI specifications allow for a wide range of test system configurations, such as one consisting of a power supply, CPU, and video modules, plus up to eight modules containing instrumentation cards.

PCXI encloses the cards in EMI/RFI shielded modules. The benefits include greater reliability, and increased immunity to noise and crosstalk between modules. With PCXI, instrumentation cards can share and collect data based on measurements from several instruments. The modules are fully upgradeable, and interchangeable.

Applications include: OEM systems, calibration lab, ATE systems, field service data acquisition, vibration analysis, educational lab, portable test systems, rack mount test systems, factory automation, and data collection.

For more information contact Rapid Systems, Inc., 433 N. 34th St., Seattle, WA 98103, or call (206) 547-8311.

WIRELESS APPLICATIONS WHITE PAPER

Vega, the leading manufacturer of high-end wireless systems, is offering a white paper discussing wireless microphone application techniques.

The paper, written by Vega president Gary Stanfill, presents wireless system information, and also offers solutions to many common wireless problems. Topics covered in the 20-pages include frequency selection, interference control, antenna systems, to name a few.

For a free copy of the Vega Wireless Microphone Application Techniques white paper call (800) 827-6701.

Vega is a manufacturer of wireless systems to professional markets worldwide and a subsidiary of Mark IV Industries. For information contact Vega, 9900 Baldwin Pl., El Monte, CA 91731-2204, or call (818) 442-0782, or (800) 877-1771.

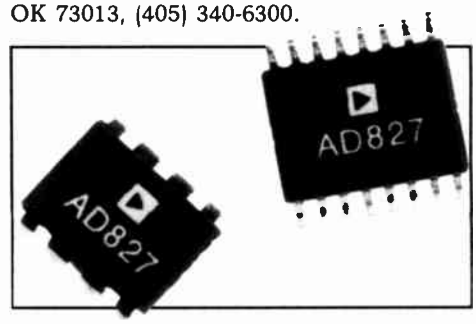
PAIA REINTRODUCES A CLASSIC EFFECT

The resurgence of interest in analog sound processing has prompted **PAIA Electronics** to again offer its popular Hyperflange and Chorus Unit in its line of sound processing kits.

Designed by Craig Anderton, the Hyperflange and Chorus is useful in the studio and for live processing. A hyper triangular control oscillator allows linear time sweeps, and the exponential time sweeps the human ear prefers, both over a range of 71:1.

The unit mounts in a single 1¾″ standard rack space, and requires a ±12–15V, 200mA power supply.

The Hyperflange and Chorus kit includes printed circuit board, all components, controls and hardware and, an illustrated step-by-step assembly and user's manual for $139.95, plus $2 shipping. An optional black anodized front panel with printed control designations is sold separately at $15.95, plus $2 shipping. Both are available from PAIA Electronics, 3200 Teakwood Lane, Edmond, OK 73013, (405) 340-6300.



DUAL VIDEO-SPEED OP AMP DRIVES HIGH CAPACITIVE LOADS

A general purpose dual 50MHz unity-gain bandwidth op amp, the AD827, from **Analog Devices, Inc.**, is stable driving *any* capacitive load and features 85dB channel separation. Well-suited for multi-channel video applications, differential phase and gain errors are typically 0.19° and 0.04°, respectively, measured according to PAL and NTSC standards.

Similar high-speed, low-power applications can benefit from the device's 300V/μs slew rate, 120ns settling time to within 0.1% for ±5V swings,and 100mW power dissipation. In 1,000s, 8-pin plastic mini-DIP and 16-pin SOIC-packaged devices cost $4.50.

The AD827's DC performance is suitable for buffering fast 8- and 10-bit analog-to-digital converters. With ±5V supplies, the AD827 has 2mV of maximum offset voltage (guaranteed) and typical open-loop gain of 3.5V/mV (500Ω load). Gain is maintained at 1.6V/mV even into loads as low as 150Ω. With ±15V supplies, open-loop gain increases to 5.5V/mV (1kΩ), with only 4mV maximum input offset voltage. From either supply, the output can drive current in excess of 30mA (minimum).

The monolithic AD827 incorporates a pair of AD827 high-speed op amps,and joins a family of high-performance op amps designed and built with Analog Devices' proprietary complementary bipolar (CB) process. Package options include 8-pin plastic mini-DIP and cerdip, and a 16-pin small-outline integrated circuit (SOIC). Operating temperatures range from 0 to +70°C, −40 to +85°C, and −55 to +125°C. Prices in 1,000s begin at $4.50. Contact Analog Devices, Inc., 181 Ballardvale St., Wilmington, MA 01887, (617) 937-1428.

NEGATIVE RESISTANCE

by Dr. Ir. A.H. Boerdijk

A portion of the current-voltage characteristic of certain devices, such as the thyristor, the tunnel diode and the magnetron, has a negative slope, that is, the current decreases with increasing applied voltage or the voltage drops when the current increases. This behaviour is, of course, opposite to that of an ohmic resistance. Whereas an ohmic (positive) resistance consumes power, a negative resistance appears to supply power. Negative resistance may be simulated electronically as described in this article.

From a pure arithmetic point of view, negative resistance remains resistance with the only difference that it is preceded by a minus sign. Figure 1 shows a conventional ohmic resistance and a negative resistance with an identical voltage applied across them. The difference in behaviour of the two is clear: the currents through them flow in opposite directions.

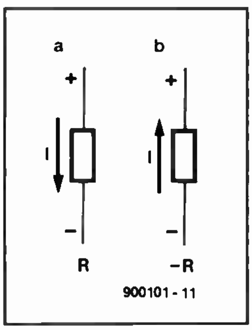


Fig. 1.

When a positive and a negative resistance are connected in series or parallel as shown in Fig. 2, the results are very interesting. The

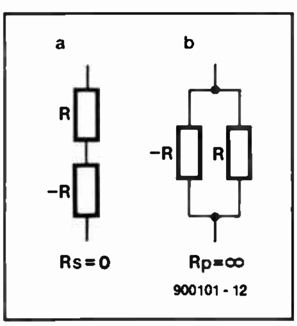


Fig. 2.

series combination (Fig. 2a) yields a short circuit:

$R_s = R + (-R) = 0.$

The parallel network (Fig. 2b) yields

$R_p = -R^2 / [R + (-R)] = -R^2/0 = -\infty,$

that is, a perfect insulator.

To confuse you further, in Fig. 3 a 10-V potential is connected across a series combination of a positive resistance of 1001 kΩ and a negative resistance of 1000 kΩ.

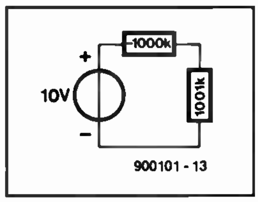


Fig. 3.

The total resistance in the loop, ignoring the internal resistance of the voltage source, is 1 kΩ. The current flowing in the loop is therefore 10 mA, and this causes a drop of 10.01 kV (!) across the 1001 kΩ resistance.

This does not indicate a new way of generating very high voltages, of course, as a quick consideration of the power distribution shows.

In Fig. 1a, the positive resistance **dissipates** a power $P = I^2 R$ or $P = U^2/R$, whereas in Fig. 1b the negative resistance **delivers** power to the voltage source. This means that negative resistance is not just a passive component and also that it can not exist by itself (since the power delivered to the voltage source must come from somewhere).

In fact, a negative resistance may be simulated by an electronic network as shown in

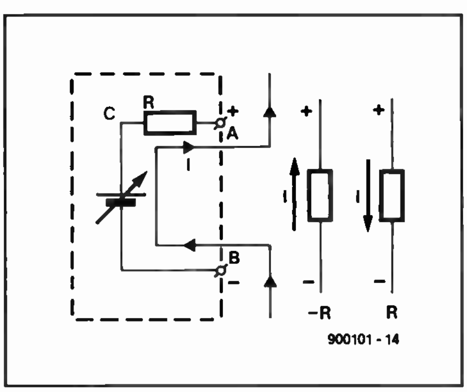


Fig. 4.

Fig. 4, where it exists between A and B. Terminal A is connected to a variable voltage source between terminals C and B that generates a voltage $U_{CB} = 2U_{AB}$. If the potential at A is positive with respect to B, the voltage at C is so, too. A current, $I = U_{CA}/R$ flows through R in the direction indicated, that is, from B (–) to A (+). In other words, the resistance between terminals A and B is negative.

When considering the operation of this network, it is important to pay attention only to terminals A and B: the circuitry hidden behind them is of no consequence here.

In practice

The circuit in Fig. 5 is constructed from an opamp and three resistors, while a negative resistance of –1 kΩ is simulated between terminals A and B. The operation may be checked by connecting a 4.7 kΩ resistor in series with terminals A and B. The total re-

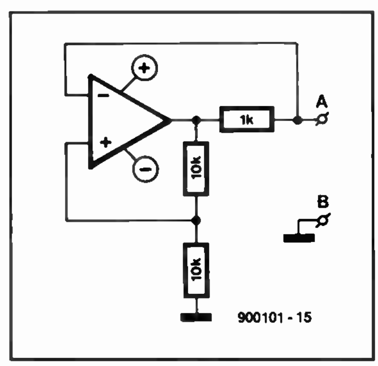


Fig. 5.

sistance measured with a standard ohmmeter is 3.7 kΩ, which shows that the effect of a negative resistance can be measured. The value of it depends on the value of the output resistor used in the simulation circuit and the ratio of the other two resistors. Replacing the fixed 1 kΩ resistor by a variable type enables a wide range of negative resistance values to be obtained.

Another fairly simple method is to connect a conventional resistor in series with the negative resistance. In this example, this resistor should have a value not exceeding 1 kΩ to prevent the negative resistance from disappearing.

If a resistor (fixed or variable) greater than 1 kΩ is connected in parallel with terminals A and B, the negative resistance increases (becomes more negative). The circuit in Fig. 5 is very suitable for experimenting with negative resistance.

When the circuit of Fig. 5 is translated into a practical design, a certain load, R_V, will exist between terminals A and B. This load has an effect on the operation of the circuit and its value must therefore be higher than the absolute value of $-R$, that is, in this circuit greater than 1 kΩ.

If the load across terminals A and B is always smaller than $-R$, the circuit is still usable, but the connections to the inputs of the opamp must be reversed (this maintains the required feedback).

Although the circuit in principle becomes unstable only when the numerical values of R_V and $-R$ are identical, it will be found in practice that it does not function satisfactorily when the values are close .o another.

It will have become clear that the maximum potential drop across the negative resistance is highly dependent on the voltage source used for the simulation circuit. This also explains why the circuit of Fig. 3 does not generate a very high voltage, although it works satisfactorily: the supply voltage is not high enough.

The output characteristic of the opamp determines the maximum current that can flow through the negative resistance. If larger currents are wanted, the output of the opamp must be provided with an additional stage. It is, of course, also possible to use an opamp that handles larger currents.

Applications

In practical electronics, negative resistance is used to compensate (ohmic) losses. A typical example is an *LC* circuit as shown in Fig. 6. The resonant frequency of this is 800 Hz and the *Q*-factor is 5.4. The value of *Q* is low, because it is heavily affected by the (loss) resistance of the inductor. It may be improved considerably by adding a variable negative resistance in parallel with the circuit. This is accomplished as shown: the fixed negative resistance is connected between A and B, and the potentiometer enables the losses caused by the resistance of the inductor to be compensated.

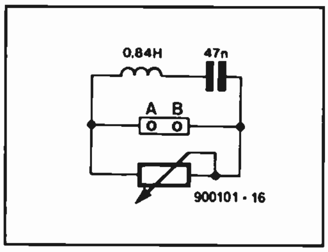


Fig. 6.

It is even possible to set the circuit into oscillation by making the negative resistance sufficiently large, that is, by reducing the value of the parallel resistance. The frequency range of the circuit will then be restricted, however, by the bandwidth of the simulation circuit.

Another application is the improvement of the control range of small d.c. motors. The rotating speed of such motors, especially at the low end of the range, is heavily dependent on the load moment. In fact, at a given point the motor just stops abruptly.

This behaviour may be improved greatly with the aid of the circuit in Fig. 7, which contains not only a variable negative resistance but also a variable supply for the motor. Potentiometer P1 controls the rotating speed of the motor, while P2 sets the value of the negative resistance.

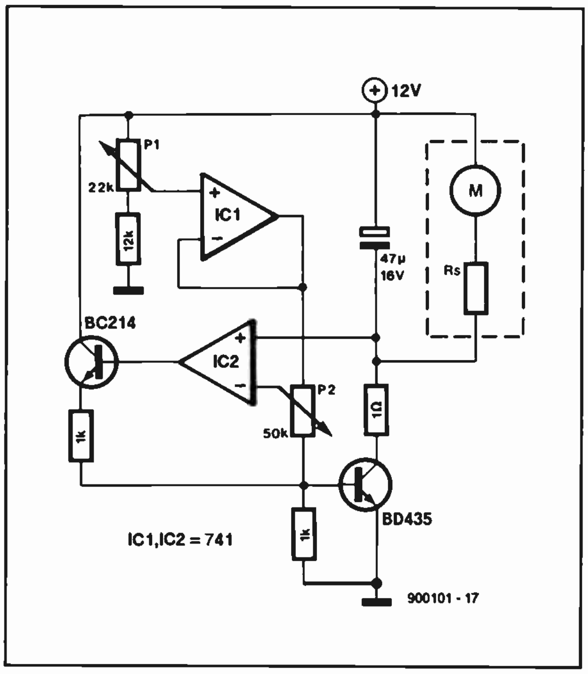


Fig. 7.

Experiments with a small d.c. motor showed that the deviation of the moment vs speed characteristic from the ideal could be improved by a factor of 2.7.

A final application is the use of a 3-Ω negative resistance to charge a battery. Connected to a 12-V battery, the charging current is 4 A; connected to a 6-V battery, the charging current is 2 A.

Such a negative-resistance charger has some peculiar properties: the connections to the battery terminals may be reversed with impunity and the short-circuit current amounts to nought. ■

IN QUEST OF A PANGRAM *Continued from page 29*

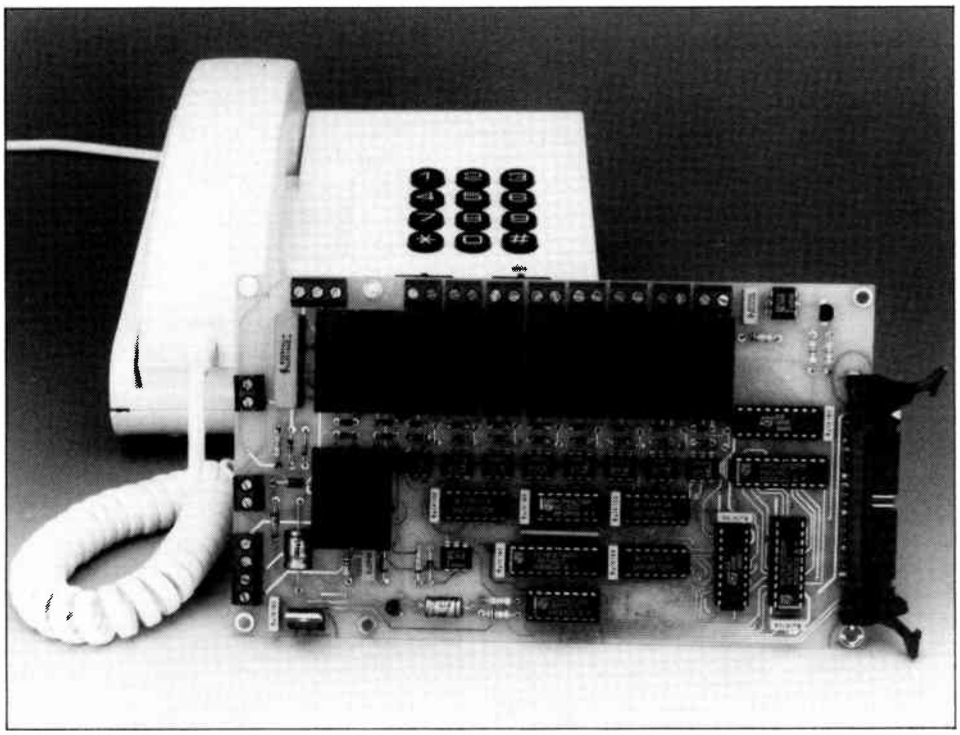
o's,' for instance, implied a range running from nine up to eighteen (or perhaps ten up to nineteen). The values actually settled ·upon—on the basis of pencil-and-paper trials with near-autograms—may be seen in Fig. 2. Ranges for each of the sixteen critical letters are represented as vertical scales with numbers (standing for number-words) indicating their starting and finishing totals. Within these ranges fall the hand-produced near-solution sums tracing out a histogram silhouette. In most cases these are, by definition, situated roughly in the middle of the range. For the low totals l, *g*, and *u*, however, this is impossible: in a pangram all letters must occur at least once; the range cannot extend below *one* (see Fig. 2.).

The second part of this article, reproduced by kind permission of Springer-Verlag, Heidelberg and New York, will apear in the September issue of *Elektor Electronics*.

MICROPROCESSOR-CONTROLLED TELEPHONE EXCHANGE

The telephone exchange presented here allows up to eight pulse-dialling telephone sets to be connected, and has an option for connecting calls to or from an external (trunk) telephone line. The unit is controlled by the popular 8052-based BASIC computer we introduced a few years ago.

A. Rigby



MAIN FEATURES

- 8 internal lines
- 1 external line
- memory for 10 numbers
- internal through connections
- versatile computer control
- automatic hold for external line
- simple-to-extend
- can be interfaced to a PC
- selective external call acceptance
- shortcut dial codes for external number
- works with pulse-dialling telephone sets
- one optional relay for extra switching function

Since the telephone exchange is controlled by a computer, it is relatively easy to add or change certain features simply by extending or changing the control program. The nice thing about the 8052-based computer used here (Refs. 1 and 2) is that it can be programmed in BASIC, a computer language familiar to many. In the present application, the BASIC computer runs its application program from an on-board EPROM. All that is required to modify this program is a terminal or a PC running a communications program, and a three-wire link to the BASIC computer. With these tools, the user is at liberty to edit and extend the existing control program in order to 'customize' the telephone exchange. The terminal or PC is no longer required once the control program has been tested and found to work all right. If you have no intention to change the 'standard' control program, or lack the ability to program in BASIC, simply use the ready-programmed EPROM available for this project. In most cases, this standard control program will provide all the necessary functions users of a telephone exchange for a small network in the home or small office or workshop have come to expect.

Telephone: the basics

Before discussing the operation of the telephone exchange, it is useful to look at the basic operation of the telephone system. In the following discussion, it is assumed that pulse-dialling telephone sets are used. The operation of tone-dialling (DTMF) is not covered. Details on this system may be found in Ref. 3.

Figure 1 shows the general lay-out of a telephone connection. When the receiver is on the hook, the bell inside the telephone set is connected to the telephone line. When the receiver is lifted, the voice circuit of the set is connected to the telephone network, and a direct current flows through the microphone. The telephone extensions connected to the network receive their supply voltage from the local telephone exchange. All sets are connected to two lines and operate free from the earth line. The use of balanced lines is a simple, yet effective, way, to eliminate noise in the network. Since any noise induced on the network is, in principle, equally strong and of equal phase on the 'a' and 'b' lines, it is effectively inaudible.

Outgoing calls

The timing diagrams in Fig. 2 show the switching sequences during a telephone call. Again, only the 'a' and 'b' lines are involved in establishing the call. Normally, a voltage of 50 to 60 V exists between these lines. The exchange detects that a receiver is lifted when the line voltage drops to about 10 V, and a microphone current of about 20 mA is established. Next, the exchange sends the dial tone to the calling extension to indicate that a number may be dialled. In the pulse-dialling system, the current loop is interrupted repetitively. The pulse rate usually lies between 9 and 11 pulses per second. The 'break' period is called 'pulse', and the 'connect' period is called 'pause'. The pulse length is generally defined as 61.5% ±3% of the period. Assuming that the period is 100-ms, the current is interrupted for periods of 58.5 to 64.5 ms. The pause allowed between successive numbers is 0.7 to 1 s.

The local exchange starts to call up the wanted extension with the aid of a ringing signal after the complete number has been received from the calling extension. When the call is answered, the exchange starts to put a cost count signal on the 'a' and 'b' lines. This signal is a sine-wave burst with an amplitude of about 50 V. Since it is the same on the 'a' and 'b' line, it is inaudible to the calling as well as to the called party. A cost counter, however, is connected asymmetrically to the network to allow it to detect the pulses. When either party rings off (puts the receiver down), the voltage between the 'a' and 'b' line reverts to the 'standby' level of 50 to 60 V.

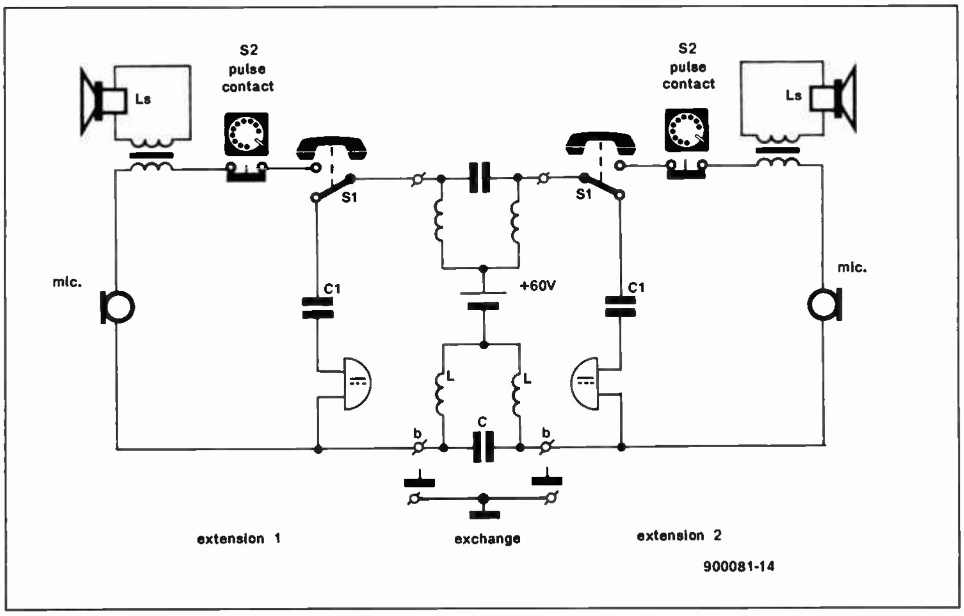


Fig. 1. Illustrating the basic operation of the two-wire telephone system.

Incoming calls

The operation of the telephone system in the case of incoming calls is illustrated in Fig. 3. An incoming call is detected by the ringing signal produced by the telephone set. The exchange calls up the extension by putting an alternating voltage of about 50 V_{pp} on the 'a' and 'b' lines. The fact that the signals on 'a' and 'b' are in anti-phase allows the telephone to detect the ring signal and actuate a sounder device (usually a small bell or buzzer). The ringing continues until the called party lifts the receiver to answer the call. If the call is not answered after a predetermined number of rings, the connection is broken (in the exchange discussed here, the maximum number of rings is set to 13). When the called party lifts the receiver before the last ring, the previously mentioned direct current flow is established, enabling the exchange to detect that the call is answered. The telephone conversation can begin!

Electronics at work

The signal sequences shown in Figs. 2a and 2b are generated and processed by the interface board of the telephone exchange, while the control functions are carried out by the BASIC computer. The function of the interface board, of which the circuit diagram is shown in Fig. 3, is to convert the digital signals supplied by the computer board to telephone network signals, and vice versa.

The eight interfaces that establish the connections with the telephone extensions are shown at the top of the circuit diagram. The extensions are connected either to the WAIT line or to the VOICE line. Extensions used for a telephone conversation are always connected to the VOICE line, which provides the necessary supply voltage. Extensions not involved in the conversation are connected to the WAIT line, and produce the 'engaged' tone when the receiver is lifted. A number can only be dialled when the exchange is back in the wait state with all extensions connected to the VOICE line.

The interface board is linked to the BASIC computer via connector K14, which carries all the necessary signals for proper communication between these units. The $\overline{Y7}$ signal supplied by the address decoder on the BASIC computer board is used to select the logic on the interface board. The line is actuated in the address range between $E000_H$ and $FFFF_H$, which is split into three parts with the aid of address lines A10, A11 and A12, giving buffer devices IC15, IC16 and IC17 their

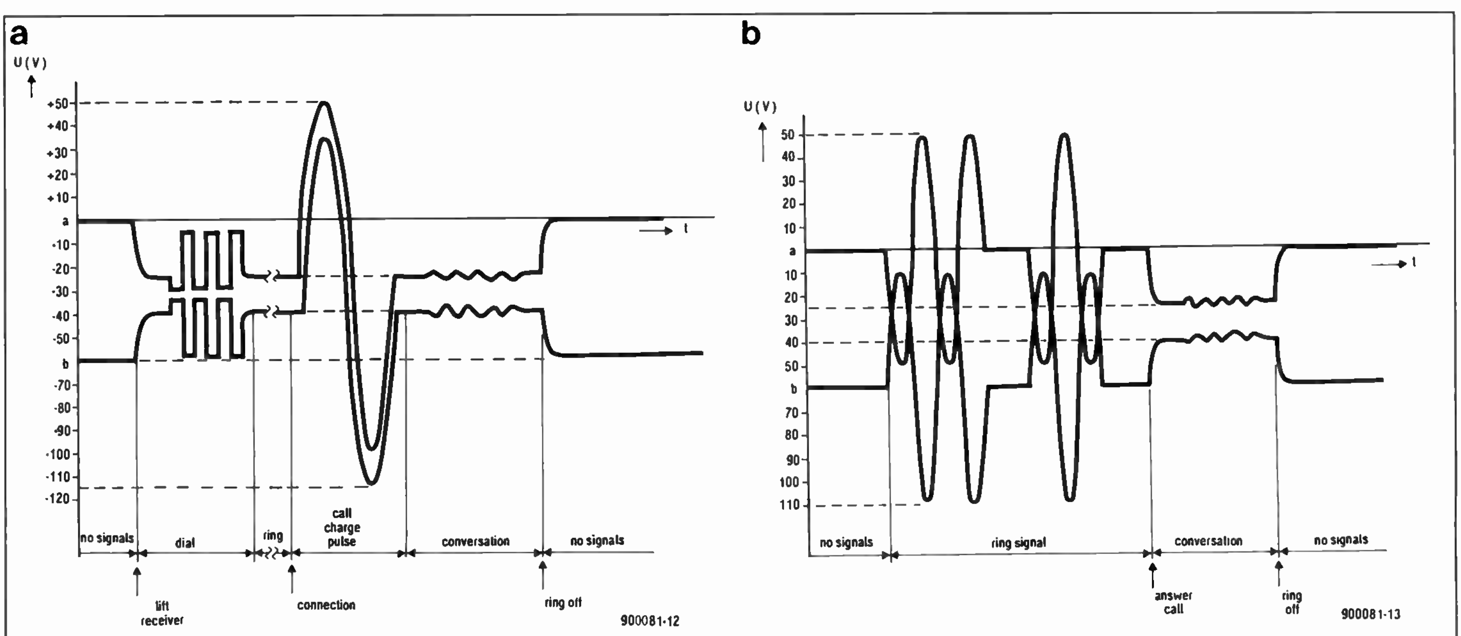


Fig. 2. Waveform sequences on the telephone lines, showing the call charge pulses (Fig. 2a) and the ring signal (Fig. 2b).

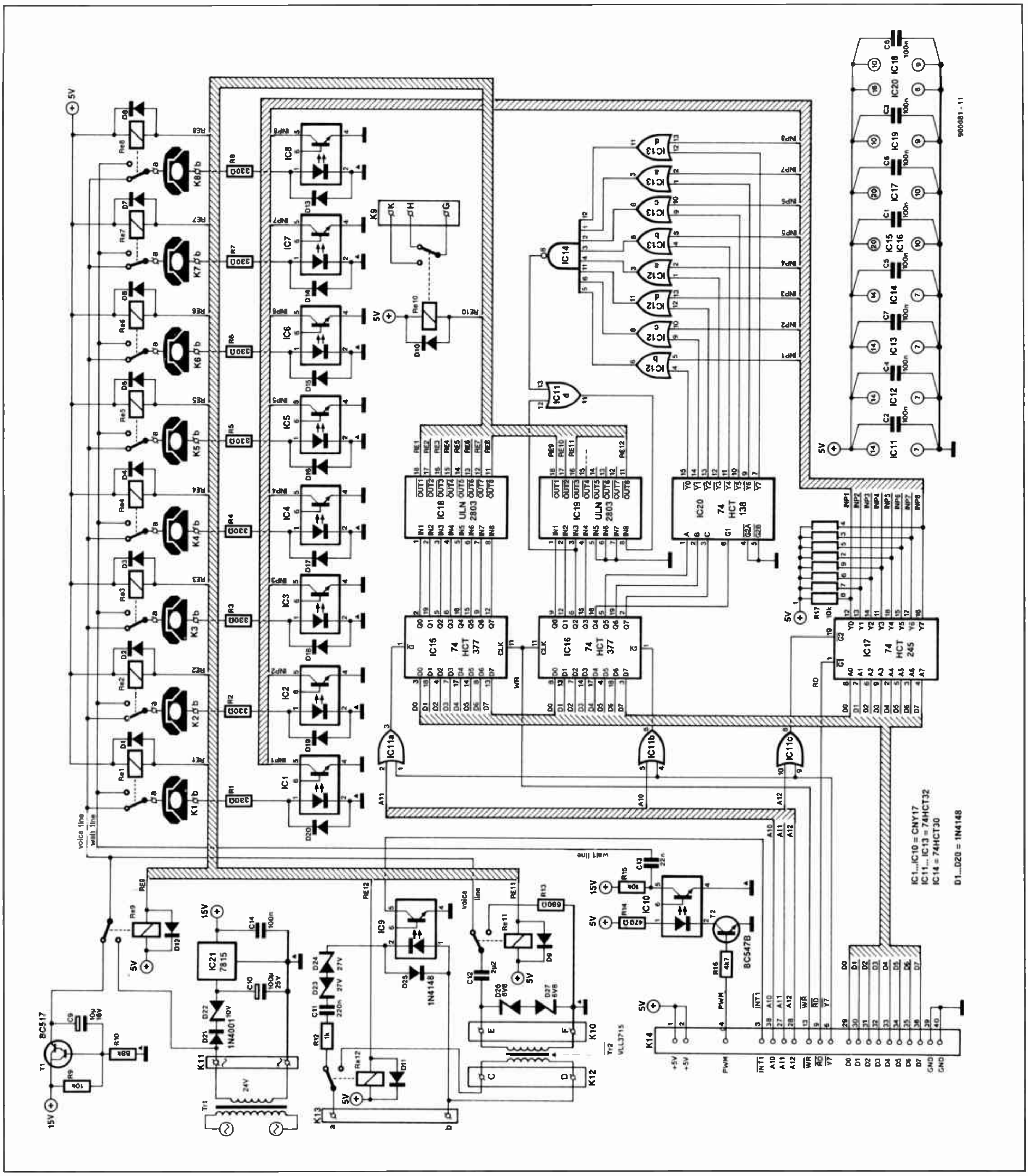


Fig. 3. Circuit diagram of the telephone exchange. This circuit is connected to the 8052-based BASIC computer via connector K14.

proper location in the memory map. The $\overline{\text{INT}}$ signal supplied by the interface board serves to wake up the BASIC processor from its stand-by state when a ringing signal is detected on the external line. The 8052 processor generates the 'engaged' tone on the PWM line. A dial tone is not generated—the network is free for dialling an extension when the receiver is silent upon being lifted. The remaining lines on K14 carry data signals, read and write signals, and the supply voltage.

Circuits IC15 and IC16 are latches that function as additional I/O registers for the control of the switching functions available in the exchange. Relays are used for the actual switching actions. Eight-bit register IC15 controls relays Re1–Re8 via the power drivers contained in IC18. These relays are used to switch the associated telephone sets between the VOICE and the WAIT line. The three least-significant datalines on IC16 switch relays Re9, Re10 and Re11. The first, Re9, is used to generate the ringing signal. In the stand-by state, transistor T1 is connected to the VOICE line, and provides all telephone sets with their supply voltage via the VOICE line. The gyrator configuration of the transistor prevents the supply short-circuiting voice signals from being superimposed on the direct voltage. When Re9 is switched, the full transformer voltage is applied to the VOICE line. As a result, the bell in the extension connected to the VOICE line starts to ring. The calling party hears the ringing signal as a series of buzzing tones.

Relay Re10 is intended for optional extensions, such as a telephone-controlled door opener, and can be controlled by appropriate modifications to the BASIC control program.

Relay Re11 is used to transfer a call received on the external line to another extension in the network. By switching Re11, the external line is terminated at the required impedance. As a result, the line is held while the exchange is being used for internal calls.

Making a call

When the receiver on any of the extensions is lifted to make a call, a current starts to flow that causes the LED in the associated opto-coupler to light. This results in the relevant $\overline{\text{INP}}$ line being pulled low. The processor identifies the calling extension by reading the logic 0 it produces in IC17 at address $EFFF_H$. Next, a write command is issued to IC15 and IC16 (at addresses $FBFF_H$ and $F7FF_H$ respectively) to connect all other extensions to the WAIT line. These extensions are effectively disabled and produce the 'engaged' tone when the receiver is lifted.

The processor counts the dialling pulses produced by the calling extensions via IC17. The dialled number determines what happens next. When a 0 is dialled, relay Re12 is actuated, and the external line is selected to establish a connection to another telephone network or another exchange. The line transformer, Tr2, is connected to the external line, and all dialling pulses that follow the 0 are fed to the external line by Re12 being actuated in their rhythm. The relay contact switches between a low impedance (the line transformer) and a high impedance (the ring pulse detector). The dialling pulses are fed out of the exchange via IC11, IC12, IC13 and IC14, after the right OR gate (IC12a–IC12d or IC13a–IC13d) has been enabled via IC20. Gate IC11d ensures that dialling pulses produced by one of the internal extensions are not passed to the external line while this is on hold. This is an important feature when an call received via the external line is being transferred to another extension served by the exchange.

Receiving external calls

Calls that reach the exchange via the external line are detected by the ring pulse detector based around D23, D24, D25, R12, C11 and IC9. When a ringing pulse is detected, IC9 pulls the $\overline{\text{INT1}}$ line of the BASIC processor logic

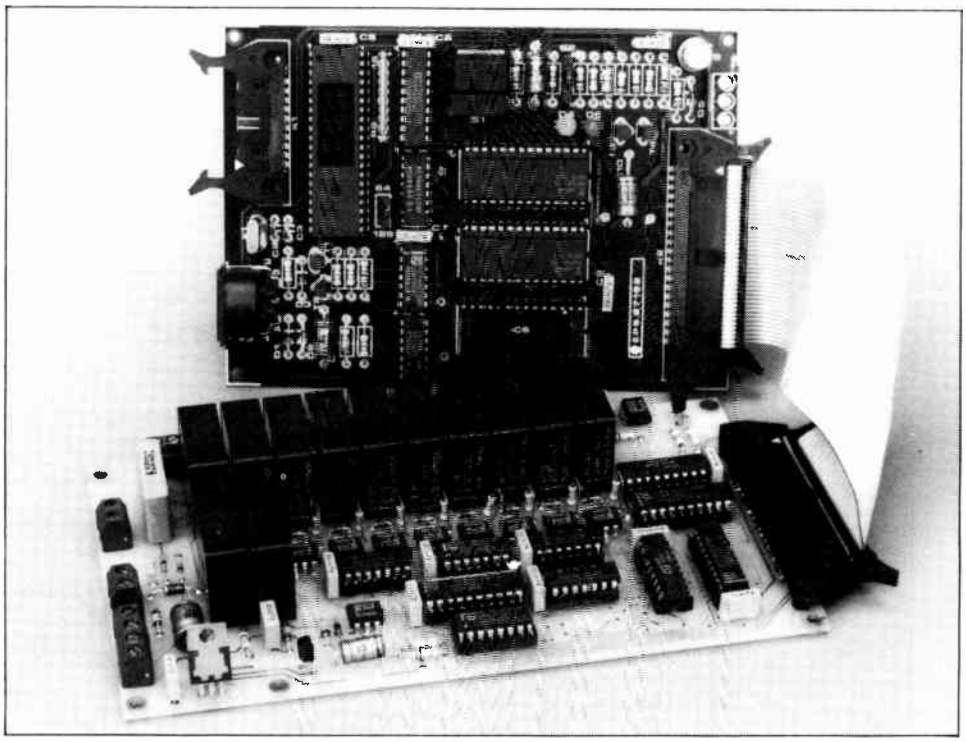


Fig. 4. Ready for use: completed BASIC computer and telephone exhange boards.

low. Only those extensions allowed to accept calls from the external line remain on the VOICE line; all others are connected to the WAIT line. A ringing signal is placed on the VOICE line with the aid of Re9. A total of 13 rings with 2.5-second pauses is allowed. The first extension that answers the call is connected to the external line. Once again the telephone conversation can begin!

After it has been answered, the external call can be transferred to another internal extension. To do this, the active extension puts the receiver down and dials the number of the wanted extension. The external line is not disconnected until any receiver has been on the hook for more than five seconds. While the external caller is on hold, the answering extension dials another extension. The external line is connected to whichever extension remains on the line when the other puts the receiver down. If the wanted extension does not answer the call, another one may be tried. In all cases, however, the total time the receiver is down must not exceed five seconds. If none of the other extensions answers, the external caller may be connected again by dialling your own number.

Construction

Figure 5 shows the track lay-outs and the component mounting plan of the double-sided, through-plated printed circuit board for the telephone exchange. The board has been designed to form a compact unit together with the BASIC computer. The greater part of the board space is reserved for the relays and the opto-couplers. Assuming that the ready-made board is used, the actual construction is unlikely to present problems if carried out with the necessary care. Accurate soldering is a must, though, to prevent short-circuits.

The two transformers are fitted as external parts on separate pieces of veroboard or stripboard to keep the overall size (and with it the cost) of the interface board as small as possible. Be sure to observe the necessary safety precautions because of the presence of the mains voltage on the mains transformer board. The BASIC computer is powered by a separate, regulated, 5-V supply.

The telephone sets and the transformers are connected to plastic or ceramic terminal blocks fitted on the PCB. The contacts of (optional) relay Re10 are available on connector K9 for experimental purposes.

The construction and operation of the BASIC computer is not covered here—for details, please consult Refs. 1 and 2. A small modification must be made to the existing circuit in regard of signals PWM, $\overline{\text{Y7}}$ and $\overline{\text{INT1}}$, which are not available on the expansion connector of the computer board. Three wire links are fitted to overcome this problem: connect pin 3 of K2 ($\overline{\text{INT1}}$ signal) to pin 10 of K1. Next, connect pin 4 of K2 (PWM signal) to pin 15 of K1. Finally, connect pin 8 of K2 ($\overline{\text{Y7}}$ signal) to pin 7 of IC3 (74HCT138). Since these wires go to previously unused pins of K2, they do not affect the normal operation of the BASIC computer.

The two boards are connected via a short length of flatcable fitted with IDC sockets that connect to K2 at the BASIC computer side and K14 at the interface board side. After

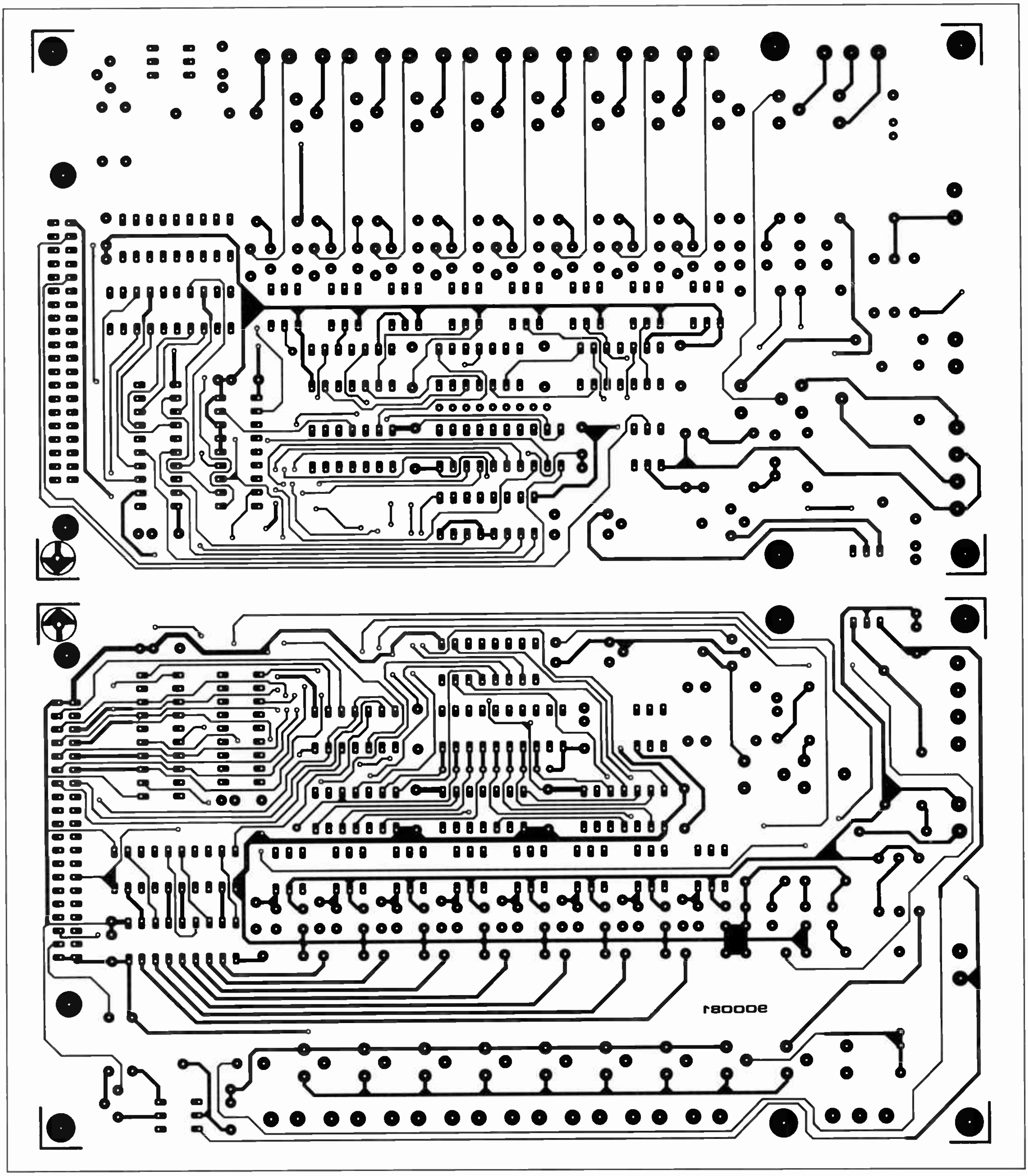


Fig. 5a. Track layouts of the double-sided, through-plated printed-circuit board.

fitting the system EPROM into its socket on the BASIC computer board, and resetting the system, the exchange is ready for use.

Control software

Software is essential for any microprocessor-based system. The control program for the telephone exchange is written in BASIC with plenty of comment in the listing to explain the operation. As already stated, the program is supplied in the form of an EPROM. Those of you who want to change it may get out their terminal or PC, connect it to the BASIC computer, and suspend the program by typing control-C. Next, LIST the program, and edit it as required. RUN the program to check that it does what you want. The syntax requirements of the 8052 BASIC interpreter are covered in the relevant Intel manual, while possible problems with the communication between the terminal or PC and the BASIC computer are tackled in Refs. 1 and 2.

A short description is given of the function of the main routines in the control program:

the internal numbers start with a '1', i.e., the extensions in the network have numbers 11 up and including 18.

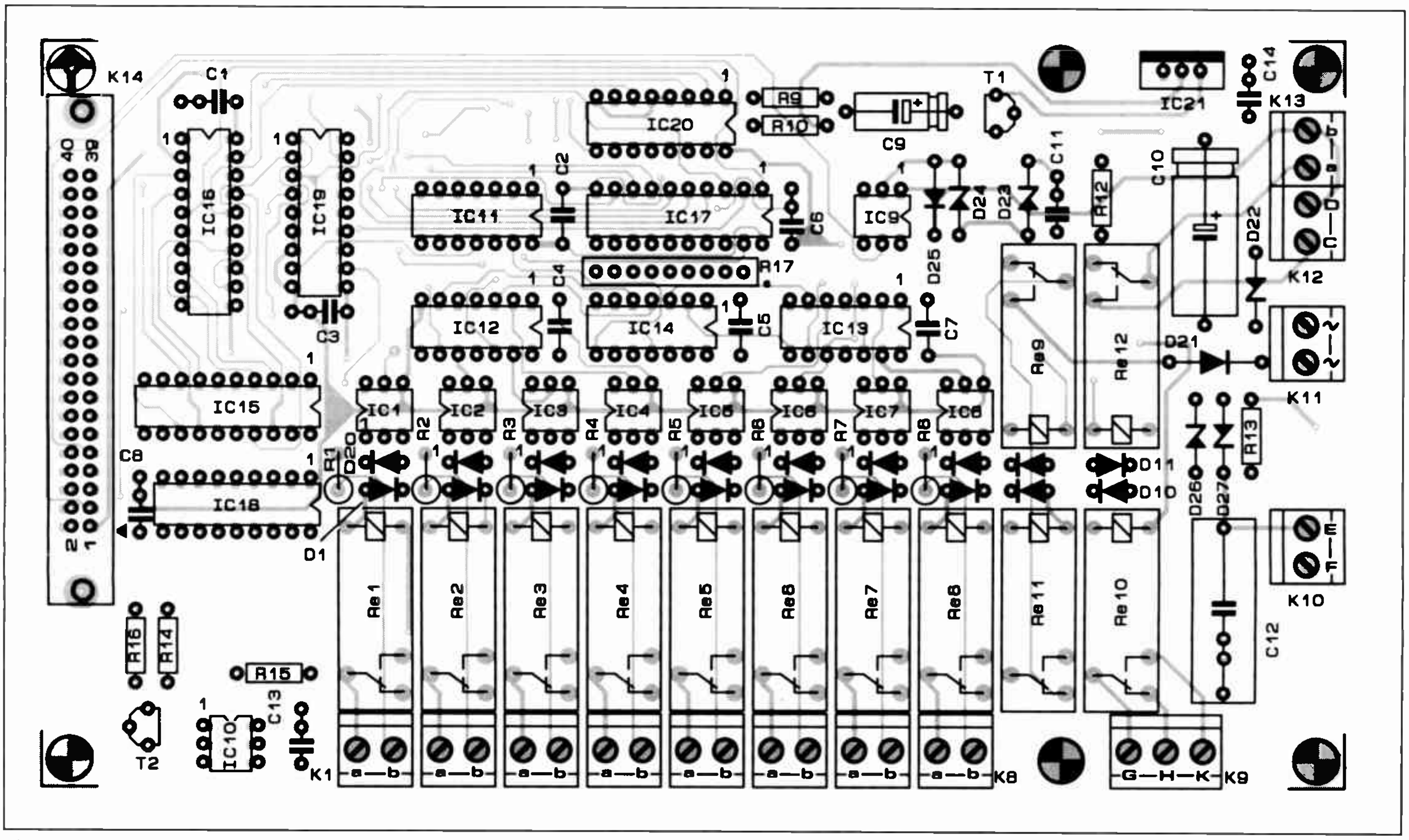


Fig. 5b. Component mounting plan.

COMPONENTS LIST

Resistors:

8	330Ω	R1–R8
2	10kΩ	R9;R15
1	68kΩ	R10
1	1kΩ	R12
1	680Ω	R13
1	470Ω	R14
1	4k7	R16
1	SIL array 8×10kΩ	R17

Capacitors:

9	100nF	C1–C8;C14
1	10µF 16V	C9
1	100µF 25V	C10
1	220nF	C11
1	2µ2 100V	C12
1	22nF	C13

Semiconductors:

10	CNY17	IC1–IC10
3	74HCT32	IC11;IC12;IC13
1	74HCT30	IC14
2	74HCT377	IC15;IC16
1	74HCT245	IC17
2	ULN2803	IC18;IC19
1	74HCT138	IC20
1	7815	IC21
1	BC517	T1
1	BC547B	T2
21	1N4148	D1–D20;D25
1	1N4001	D21
1	10V 0.4W zener diode	D22
2	27V 0.4W zener diode	D23;D24
2	6V8 0.4W zener diode	D26;D27
1	EPROM with control program for BASIC computer. Order code ESS 5941.	

Miscellaneous:

12	2-way PCB terminal block	K1–K8; K10–K13
1	3-way PCB terminal block	K9
1	40-way PCB-mount plug with eject headers	K14
12	5-V SPDT PCB-mount relay, Siemens type V23127-B0001-A101	Re1–Re12
1	24V 3.3VA mains transformer	Tr1
1	telephone line transformer type VLL3715T	Tr2
1	printed-circuit board	900081

- a total of ten shortcut codes is allowed for external numbers. These codes start with a '2', i.e., 20 up to and including 29 are available.
- the codes used for shortcut dialling are stored by first dialling '3'. Next, dial the code (0–9), followed by the number of the external connection. The processor stores the shortcut code and the associated number when the receiver is put down. All codes are available to all extensions in the network served by the exchange, and they may be changed at any time by any extension.
- a particular extension can be disabled from receiving external calls by dialling '5'. This can be undone by dialling '6'.

The function of dial numbers '4', '7', '8' and '9' is not fixed, although the software has built-in routines to intercept them. Number '8', for instance, could be used to switch on relay Re10, and number '9', to switch it off again. To be able to do this, you have to include the appropriate write command in the number interception subroutine, test the option, and program a new EPROM. ■

References:

1. BASIC computer. *Elektor Electronics* November 1987.
2. ROM-copy for 8052-BASIC computer. *Elektor Electronics* September 1990.
3. Dual-tone multi-frequency (DTMF) decoder. *Elektor Electronics* May 1989.

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GUITAR TUNER

by T. Giffard

A guitar is one of those instruments that needs to be tuned daily. The number of tuning aids on the market indicate that many players do not find this tuning all that easy. The electronic tuner presented here will make tuning easier for the beginner as well as for the advanced player.

For almost three centuries guitarists have used a tuning-fork (invented in 1711 by John Shore) or a pitch-pipe (which appeared later that century) as an aid to tuning their instrument. Between the two world wars, electrical aids to tuning became popular. Typical of these is the resonoscope, introduced almost simultaneously in the USA and the UK in 1936, which gives a visual indication at the correct tuning pitch. Some drawbacks of them are that they are not easily made and that beginners never learn how to tune the guitar properly. The tuning-fork and pitch-pipe (also known as flue) are not ideal either. The tone of a tuning-fork is barely audible and lasts but for a few moments. The pitch-pipe has a better volume and lasts for as long as the breath of the tuner allows, but its accuracy often leaves much to be desired (depending on its quality, of course).

The present tuner may be considered as a very accurate electronic pitch-pipe. Accuracy of pitch is ensured by a quartz crystal, since even a mediocre crystal does not deviate from its nominal frequency by more than 100 p.p.m. (0.01%).

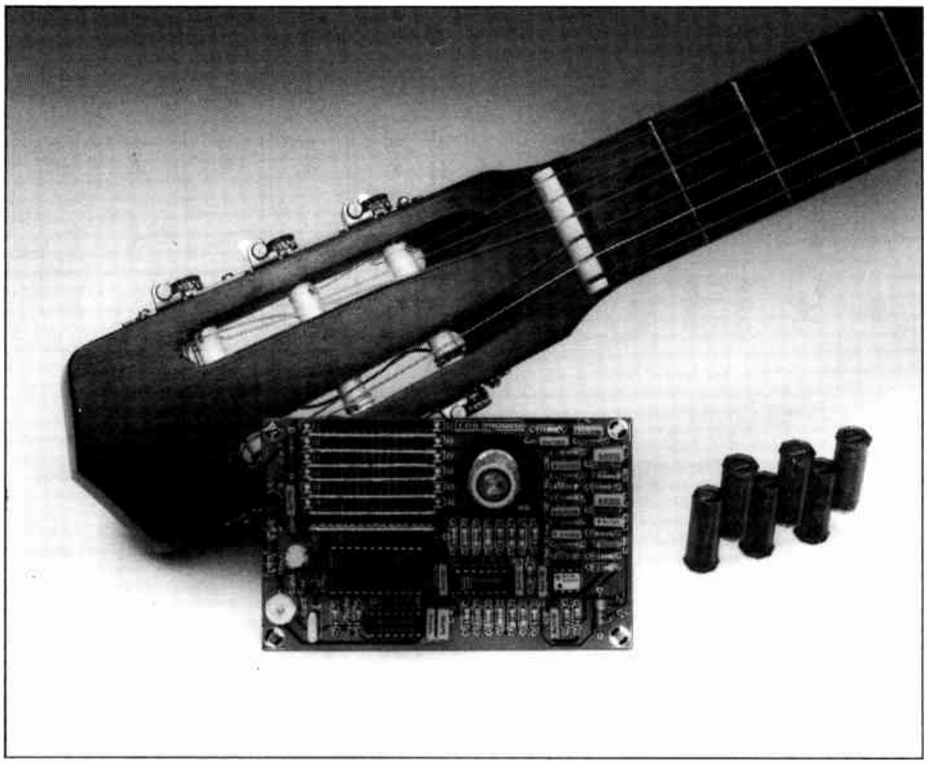

Fig. 1. General view of the electronic guitar tuner.

The six tones (US: notes)* required for tuning a guitar are obtained by dividing the crystal frequency with a presettable divider. Once the divider settings have been established, the six tones are always correct with respect to one another. Tuning is further simplified by the tone being a pure sine wave at constant loudness level. And, of course, the best acoustic measuring instrument available is used: the human ear. This has the additional advantage that the tuner's ear becomes trained in distinguishing between different tones: practice makes perfect.

Methods of tuning

Instruments may be tuned to the '**natural scale**' (which is deducible by physical laws); to **mean-tone temperament** (which gives a close approximation to natural tuning); and to **equal temperament** (in which musical intervals are moved away from the natural scale to fit them for practical performance). In equal temperament tuning, each semitone is made an equal interval. In other words, the twelve tones in an octave are equidistant on a logarithmic frequency scale, that is, each tone has a frequency that is $2^{1/12}$ (=1.05946) times greater than the preceding one. The advantage of this is that the instrument may be played in virtually any key. The disadvantage, however, is that the tones do not sound 'natural', which is especially

Table 1

			Harmonic tuning	Equal temp. tuning
$4f_{E2} = 3f_{A2} = 330.00$ Hz	$\therefore f_{E2}$	=	82.50 Hz	82.41 Hz
$4f_{A2} = 3f_{D1} = 440.00$ Hz	$\therefore f_{A2}$	=	110.00 Hz	110.00 Hz
$4f_{D1} = 3f_{G1} = 586.67$ Hz	$\therefore f_{D1}$	=	146.67 Hz	146.83 Hz
$4f_{G1} = 3f_{B1} = 782.22$ Hz	$\therefore f_{G1}$	=	195.56 Hz	196.00 Hz
$4f_{B1} = 3f_{A2} = 990.00$ Hz	$\therefore f_{B1}$	=	247.50 Hz	246.94 Hz
$4f_{E2} =$	f_E	=	330.00 Hz	329.63 Hz

*In the United Kingdom, and most other English-speaking countries outside North America, a **tone** means a "musical sound consisting of a 'pure' note". In the USA, this is called a **note**. Similarly, "a single sound of given pitch and duration" is called a **note** in the UK and a **tone** in the USA.

no. of harmonic	1	2	3	4	5	6	7	8	9
length of string vibrating	entire	1/2	1/3	1/4	1/5	1/6	1/7	1/8	1/9

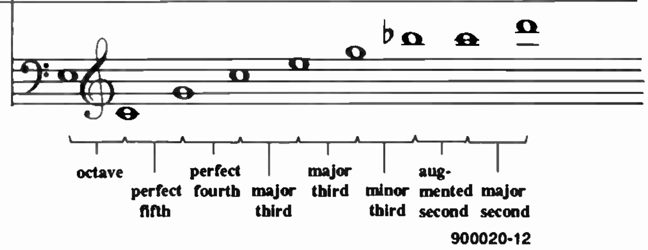

Fig. 2. Relation between harmonic frequencies and the length of string that is vibrating.

noticeable when concords are played.

In general, most guitarists prefer to tune their instrument to flageolet-notes, which means **harmonics**. The name derives from the supposed resemblance of these thin-sounding notes to those of the flageolet, an obsolete 6-holed wind instrument.

The name 'harmonic' is an abbreviation for 'harmonic tone', that is, one of the so-called harmonic series. The lowest of such tones, the 'fundamental', is called the 'first harmonic', the next lowest, the 'second harmonic', and so on. However, 'playing on harmonics' on a guitar really means 'harmonics without the first', since the fundamental is the 'normal' sound.

It is fairly simple to make the strings of a

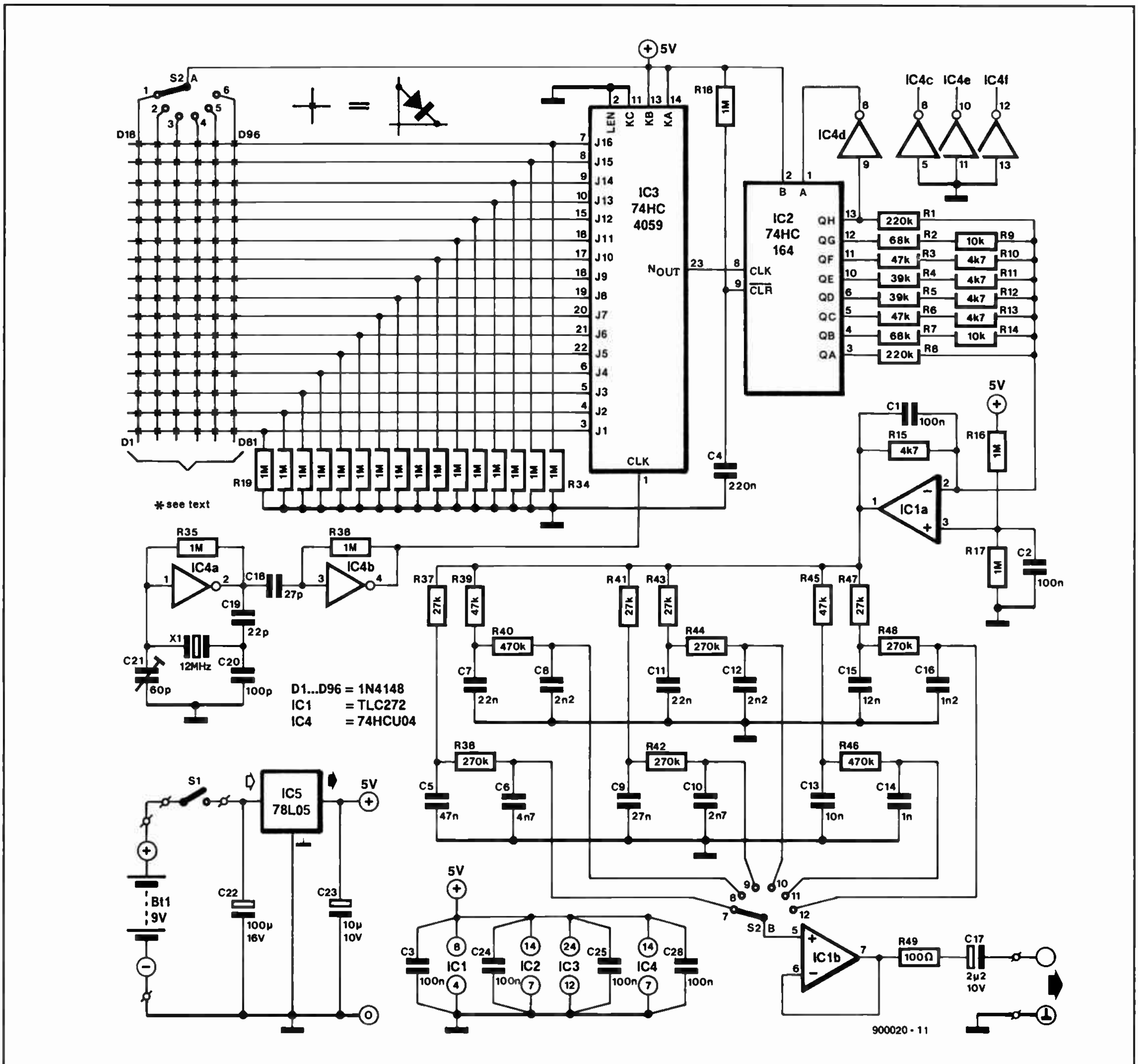

Fig. 3. Circuit diagram of the guitar tuner. Tuning frequencies are determined by the diode matrix D1–D81.

guitar vibrate at the second, third and fourth harmonic. This is done by setting the string vibrate not as a whole length but in fractional parts of its length. To obtain the second harmonic, the string is vibrated over half its length, for the third harmonic over a third of its length, and so on (see Fig. 2).

On a correctly tuned guitar, equal harmonic frequencies can be found on two different strings. For instance, the fourth harmonic of the lower E-string has the same frequency as the third harmonic of the A-string. In technical terms, $4f_{E2} = 3f_{A2}$. Relations between harmonics on other strings are shown in Table 1; if these are as shown, the guitar is tuned correctly. The frequencies are based on the International Concert Pitch, A = 440 Hz, whose second lower harmonic is A_2 = 110 Hz. From this, the other frequencies can be calculated. Note that the subscript or exponent of the tone following the string name indicates the number of octaves the tone is above or below Middle C respectively. The number 0 is traditionally omitted.

For completeness' sake, the table also shows the frequencies when the guitar is tuned to equal temperament. The choice is yours! Equal temperament is normally preferred for playing in a group, but for solo playing most guitarists choose harmonic tuning because that gives a 'smoother' sound. See also Table 2.

The tuner

The design of the tuner is far simpler than the theory behind the tuning. Its circuit, see diagram in Fig. 3, may be divided into four, excluding the power supply: a crystal oscillator, IC4a; a presettable frequency divider, IC3; a sine-wave shaper, IC2a-IC1a; and a six-stage *RC* output filter.

The oscillator is (and must be) an unbuffered inverter, and IC4a is therefore an HCU type. If you have a frequency counter, adjust C21 to give an oscillator frequency of exactly 12 MHz; otherwise, just set the trimmer to the centre of its travel. The oscillator is coupled to the clock (CLK) input of the divider.

The divider has four groups of four BCD (binary-coded decimal) inputs, J1–J16, with each of which one digit of the divisor, *k*, is set (max. 9999). Group J1–J4 sets the units; J5–J8 the tens; J9–J12 the hundreds; and J13–J16 the thousands. The setting is accomplished with the aid of a diode matrix, D1–D96 (not all of which are required).

The presence or absence of diodes determines the divisor for each of the six tones selected with S2. The presence of a diode causes a logic 1, and the absence a logic 0, at the relevant J-input of IC3. See also Table 2.

The output signal of the divider has a fre-

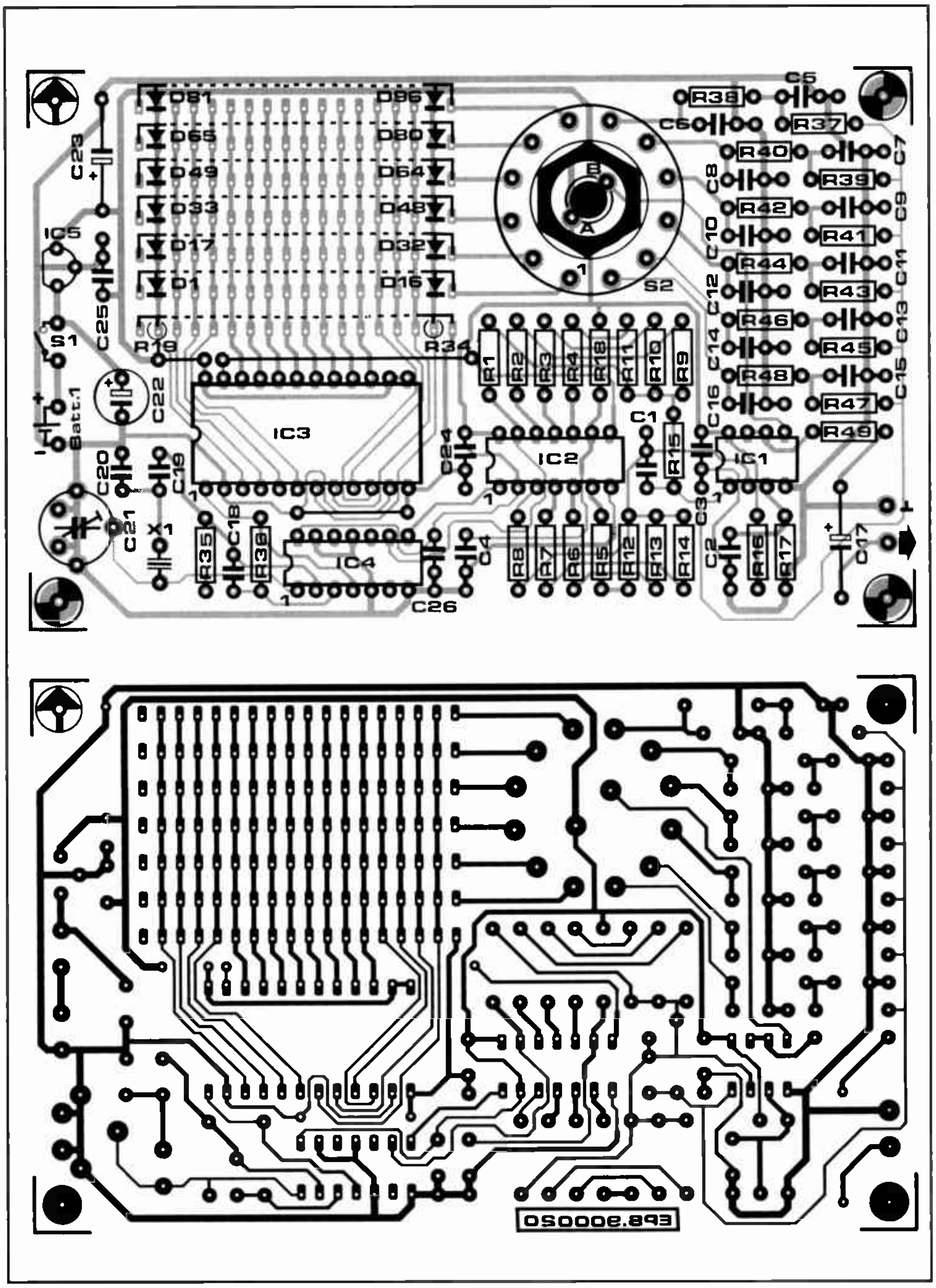

Fig. 4. The printed-circuit board for the tuner is single-sided to allow the diode matrix and the pull-up resistors to be mounted upright.

PARTS LIST

Resistors:
R1, R8 = 220 k
R2, R7 = 68 k
R3, R6, R39, R45 = 47 k
R4, R5 = 39 k
R9, R14 = 10 k
R10–R13, R15 = 4k7
R16–R36 = 1 M
R37, R41, R43, R47 = 27 k
R38, R42, R44, R48 = 270 k
R40, R46 = 470 k
R49 = 100 Ω

Capacitors:
C1,–C3, C24–C26 = 100 n
C4 = 200 n
C5 = 47 n
C6 = 4n7
C7, C11 = 22 n
C8, C12 = 2n2
C9 = 27 n
C10 = 2n7
C13 = 10 n
C14 = 1 n
C15 = 12 n
C16 = 1n2
C17 = 2µ2, 10 V, axial
C18 = 27 p
C19 = 22 p
C20 = 100 p
C21 = trimmer, 60 p
C22 = 100 µF, 16 V, radial
C23 = 10 µF, 10 V, axial

Semiconductors:
D1–D96 = 1N4148 (number required depends on tuning - see Table 2)
IC1 = TLC272
IC2 = 74HC164
IC3 = 74HC164
IC4 = 74HCU04
IC5 = 78L05

Miscellaneous:
S1 = single-pole switch
S2 = 2-pole, 6-position rotary switch for PCB mounting
X1 = 12 MHz quartz crystal
Bt1 = 9-V (PP3) battery and associated clip.
Enclosure, preferably ABS, 190×100×28 mm(approx.)
PCB Type 900020

quency that is 16 times higher than required. Clocked by this signal, the sine-wave shaper produces a pure sinusoidal signal at the right frequency at the output, pin 1, of IC1a.

Circuit IC2a is an eight-bit shift register that has been arranged to accept logic 1s when QH is low and logic 0s when QH is high. A reset at switch-on, provided by R18-C4, ensures that at the onset of operation always eight 0s are input first to the QA–QH inputs of the divider and then eight 1s.

The logic levels at the output of IC2 are translated into a resistance that, in conjunction with R15, determines the gain of IC1a.

The shape of the output of IC1a is nearly sinusoidal, mainly because of C1 in the feedback loop. Nevertheless, the 15th and 17th harmonics are still fairly strong but, since these are, in frequency, a good way from the fundamental, a simple second-order *RC* filter is sufficient to suppress them. However, the distance (in frequency) between the fundamental and these harmonics is not so large as to make one filter suffice for all six tones: each of the tones requires a separate filter.

The six filters are switched into circuit by S2b. The waveform at the pole of this switch is a good sine wave that has less than 0.04% harmonic distortion. Even so, the signal is still buffered by IC1b. The cut-off frequency of the filters is about 60% of the frequency at the –3 dB point.

The output of the tuner, which is protected against short-circuits by R49, is suitable for connecting to a variety of amplifiers. Its level depends to some degree on the frequency and lies between 450 mV and 600 mV r.m.s. Because of this dependency on frequency, the loudness appears to remain more constant than when the output level is kept constant.

Table 2

S2 posn.	Tone	Frequency	Diodes to be used	Divisor
		(Equal temperament tuning)		
1	f_{E2}	82.41 Hz	1; 9; 13; 16	9101
2	f_{A2}	110.00 Hz	20; 21; 28; 30; 31	6818
3	f_{D1}	146.83 Hz	36; 41; 45; 47	5108
4	f_{G1}	196.00 Hz	49; 50; 51; 54; 60; 61; 62	3827
5	f_{B1}	246.94 Hz	65; 66; 67; 69; 70; 77; 78	3037
6	f_E	329.63 Hz	81; 83; 85; 86; 87; 90; 94	2275
		(Harmonic tuning)		
1	f_{E2}	82.50 Hz	5; 8; 13; 16	9090
2	f_{A2}	110.00 Hz	20; 21; 28; 30; 31	6818
3	f_{D1}	146.70 Hz	33; 34; 37; 41; 45; 47	5113
4	f_{G1}	195.60 Hz	49; 51; 53; 54; 60; 61; 62	3835
5	f_{B1}	247.50 Hz	69; 70; 77; 78	3030
6	f_E	330.00 Hz	81; 82; 85; 86; 87; 90; 94	2273

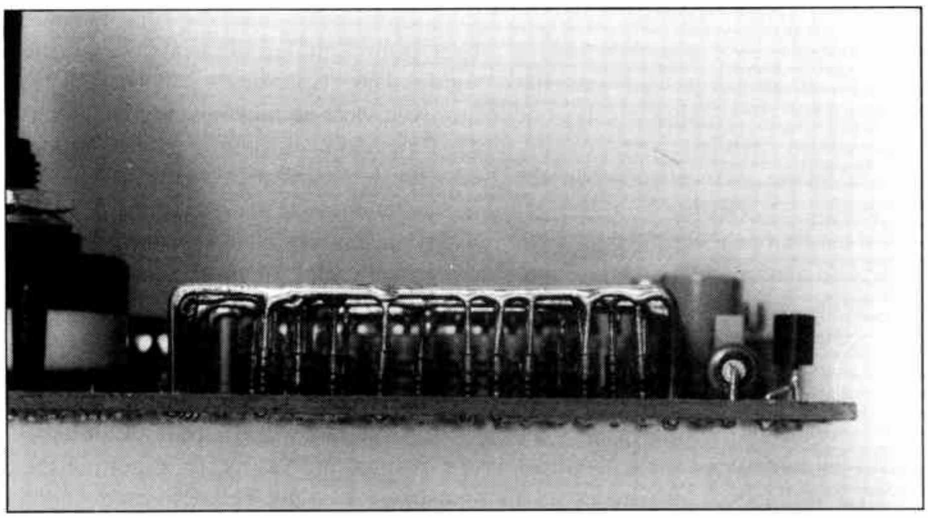

Fig. 5. All diodes used and the pull-up resistors must be mounted upright between the board and a bridge of stout circuit wire.

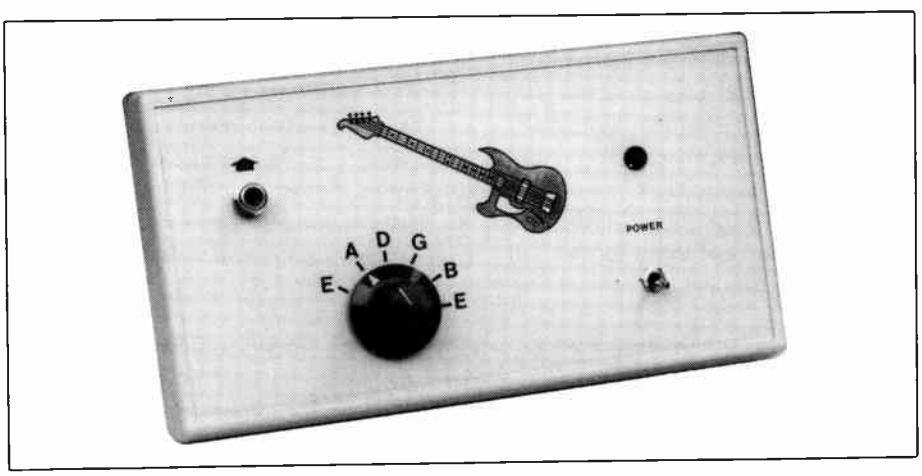

Fig. 6. Suggested front panel layout for the tuner.

Construction and setting up

The tuner is best built on the printed-circuit board shown in Fig. 4 and then fitted in a suitable enclosure of about 190×100×28 mm. Mount all pull-down resistors and the required diodes (34 for equal temperament tuning and 33 for harmonic tuning – see Table 2) upright as shown in Fig. 5.

The divisors are obtained as briefly explained earlier. For instance, the divisor, *k*, for frequency f_{G1} (equal temperament tuning) is obtained by placing diodes D49, D50 and D51 on to J1, J2, and J3 respectively and no diode to J4 to give binary number 0111 = decimal 7; diode D54 on to J6 and no diodes to J5, J7 and J8 to give binary number 0010 = decimal 2; diode D60 on to J12 and no diodes to J9, J10 and J11 to give binary number 1000 = decimal 8; and diodes D61 and D62 onto J13 and J14, and no diodes to J15 and J16 to give binary number 0011 = decimal 3. The divisor is thus 3827.

The divisor may be calculated from $k = 12 \times 10^6 / 16f$. If frequencies different from those in this article are used, bear in mind that the filters must be adapted accordingly.

The most practical power supply is a 9-V (PP3) battery. The average operational current drain is 12 mA. The battery will last a long time as shown by the prototype still working satisfactorily when the battery voltage had dropped to 4 V. ■

IN QUEST OF A PANGRAM

by Lee C.F. Sallows

The pangram problem

Some years ago, a Dutch newspaper, the *Nieuwe Rotterdamse Courant*, carried an astonishing translation of a rather tongue-in-cheek sentence of mine that had previously appeared in one of Douglas Hofstadter's *Scientific American* columns ("Metamagical Themas", January 1982). Both the translation and an article describing its genesis were by Rudy Kousbroek, a well-known writer and journalist in Holland. Here is the original sentence:

Only the fool would take trouble to verify that his sentence was composed of ten *a*'s, three *b*'s, four *c*'s, four *d*'s, forty-six *e*'s, sixteen *f*'s, four *g*'s, thirteen *h*'s, fifteen *i*'s, two *k*'s, nine *l*'s, four *m*'s, twenty-five *n*'s, twenty-four *o*'s, five *p*'s, sixteen *r*'s, forty-one *s*'s, thirty-seven *t*'s, ten *u*'s, eight *v*'s, eight *w*'s, four *x*'s, eleven *y*'s, twenty-seven commas, twenty-three apostrophes, seven hyphens and, last but not least, a single *!*

Complete verification is a tedious task: unsceptical readers may like to take my word for it that the number of letters and signs used in the sentence do indeed correspond with the listed totals. A text that inventories its own typography in this fashion is what I call an *autogram* (*auto* = self, *gramma* = letter). Strict definition is unnecessary, different conventions giving rise to variant forms; it is the use of cardinal number-words written out in full that is the essential feature. Below we shall be looking at some in which the self-enumeration restricts itself to the letters employed and ignores the punctuation.

Composing autograms can be an exacting task, to say the least. The process has points in common with playing a diabolically conceived game of patience. How does one begin? My approach is to decide first what the sentence is going to say and then make a flying guess at the number of occurrences of each sign. Writing out this provisional version, the real totals can be counted up and the initial guess updated into an improved estimate. The process is repeated, trials and error leading to successively closer approximations. This opening soon shades into the middle game. By now all of the putative totals ought to have been corrected to within two or three of the true sums. There are, say, nine *f*'s in fact but only *seven* being claimed, and 27 real *t*'s where *twenty-nine* are declared.

An English explorer's self-referent account of his hybrid machine for solving a challenging word puzzle.

Switching *seven* with the *nine* in *twenty-nine* to produce *nine f*'s and *twenty-seven t*'s corrects both totals at a single stroke. Introducing further cautious changes among the number-words with a view to bringing off this sort of mutual cancellation of errors should eventually carry one through to the final phase.

The end game is reached when the number of discrepancies has been brought down to about four or less. The goal is in sight but, as in a maze, proximity is an unreliable guide. Suppose, for instance, a few days' painstaking labour have at last yielded a near-perfect specimen: only the *x*'s are wrong. Instead of the *five* claimed, in reality there are six. Writing *six* in place of *five* will not merely invalidate the totals for *e*, *f*, *s*, and *v*, the *x* in *six* means that their number has now become seven. Yet, replacing *six* by *seven* will only return the total to six. What now?

Paradoxical situations of this kind are a commonplace of autogram construction. Interlocking feedback loops magnify tiny displacement into far-reaching upheavals; harmless truths cannot be stated without disconfirming themselves. Clearly, the only hope of dehydrating this Hydra and getting every snake-head to eat its own tail lies in doctoring the text accompanying the listed items. In looking at the above case, for example, only a fool will fail to spot instances where style has been compromised in deference to arithmetic. Short of a miracle, it is only the flexibility granted through choice of alternative forms of expression that would seem to offer any chance of escape from such a labyrinth of mirrors.

This is what made Kousbroek's translation of my sentence so stunning. Number-words excepted, his rendering not only adhered closely to the original in meaning, it was simultaneously an autogram in Dutch!

Or at least, so it appeared at first sight. Counting up, I was amused to find that three of the sums quoted in his sentence did not in fact tally with the real totals. So I wrote to the author pointing out these discrepancies. This resulted a month later in a second article in the same newspaper. Kousbroek wrote of his surprise and dismay in being caught out by the author of the original sentence, "specially come over from America, it seems, to put me right." The disparities I had pointed to, however, were nothing new to him. A single flaw had been spotted in the supposedly finished translation on the very morning of submitting his manuscript. But a happy flash revealed a way to rectify the error in the nick of time. Later, a more careful check revealed that this 'brainwave' had in fact introduced even more errors elsewhere. He'd been awaiting 'the dreaded letter with its merciless arithmetic' ever since. The account went on to tell of his titanic struggle in getting the translation straight. The new version was included; it is a spectacular achievement.

The tail concealed a subtle sting, however. At the end of his story, Kousbroek threw out a new (letter-only) autogram of his own:

Dit pangram bevat vijf *a*'s, twee *b*'s, twee *c*'s, drie *d*'s, zesenveertig *e*'s, vijf *f*'s, vier *g*'s, twee *h*'s, vijftien *i*'s, vier *j*'s, een *k*, twee *l*'s, twee *m*'s, zeventien *n*'s, een *o*, twee *p*'s, een *q*, zeven *r*'s, vierentwintig *s*'s, zestien *t*'s, een *u*, elf *v*'s, acht *w*'s, een *x*, een *y* en zes *z*'s.

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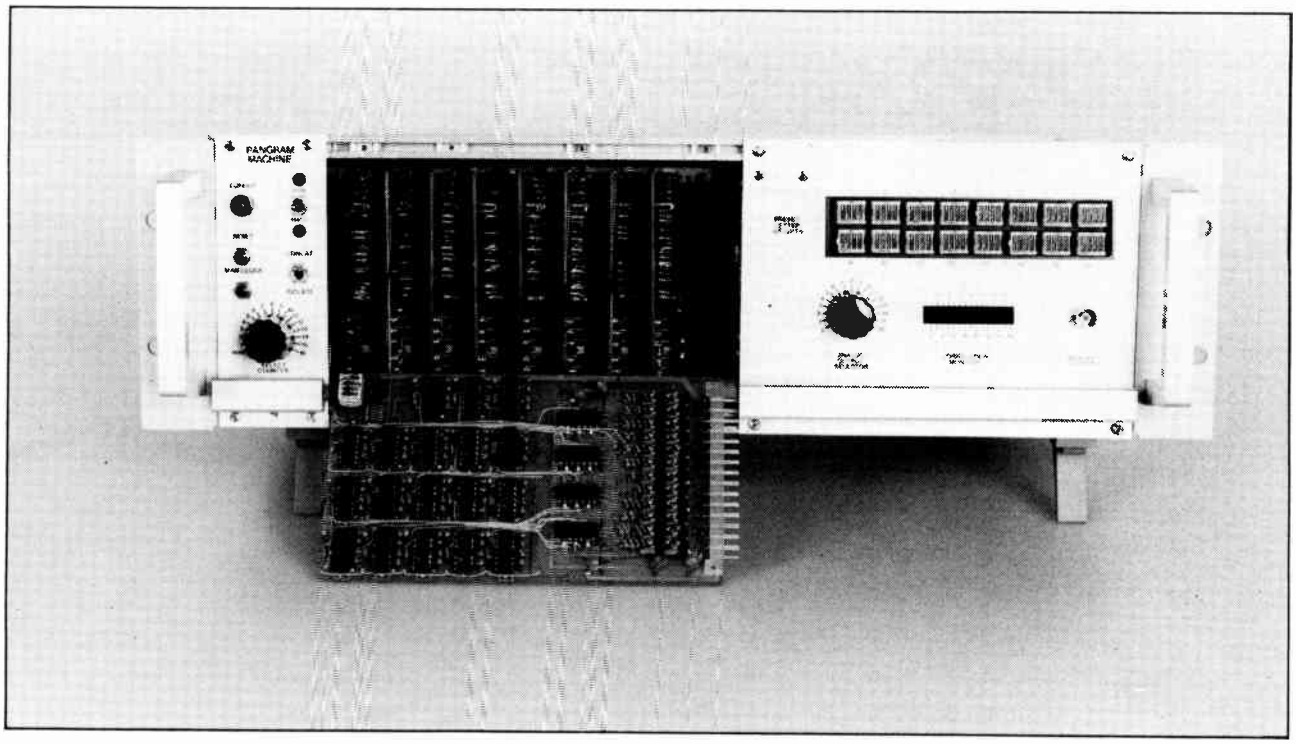


The automatic number-word selector board that transformed the original pangram machine into the Mark II version. On the left, 18 window-detector chips determine the number of *g*'s, *l*'s, *x*'s, and *y*'s. At right, four more integrated circuits and 24 transistors switch in the appropriate PROFILES on the resistor-bearing cards above.

A finer specimen of logological elegance is scarcely conceivable. The sentence is written in flawless Dutch and couldn't possibly be expressed in a crisper or more natural form. In ordinary translation, it says, "This pangram contains five *a*'s, two *b*'s, two *c*'s ... one *y*, and six *z*'s." [A *pangram*, it is necessary to explain, is simply a phrase or sentence containing every letter of the alphabet at least once (*pan* = all, *gramma* = letter). This article is about self-enumerating pangrams, that is, pangrams that are simultaneously autograms. In such pangrams, some letters will occur only at the point where they themselves are listed (look at *k*, *o*, *q*, *u*, *x*, *y*).] Following this pangram came a devilish quip in my direction: "Lee Sallows will doubtless find little difficulty in producing a magic English translation of this sentence," wrote Kousbroek.

Needless to say, I didn't manage to find any errors in *this* sentence of his!

Autograms by computer

Rudy's playful taunt came along at a time when I had already been looking into the possibility of computer-aided autogram construction. Anyone who has tried his hand at composition will know the drudgery of keeping careful track of letter totals. One small undetected slip in counting can later result in days of wasted work. At first I had envisaged no more than an aid to hand-composition: a program that would count letters and provide continuous feedback on the results of keyboard-mediated surgery performed on a sentence displayed on screen. Later I began to wonder what would happen with a program that cycled through the list of number-words, checking each against its corresponding real total and making automatic replacements where necessary. Could autograms be evolved through a repetitive process of selection and mutation? Several such LISP programs were in fact written and tested; the results were not unpredictable. In every case, processing would soon become trapped in an endless loop of repeated exchanges. Increasing refinements in the criteria to be satisfied before a number-word was replaced would win only temporary respite from these vicious circles.

What seemed to be needed was a program that could look ahead to examine the ramifications of replacing *nineteen* by *twenty*, say, before actually doing so. But how is such a program to evaluate or rank prospective substitutions? Goal-directed problem solving converges on a solution by using differences between intermediate results and the final objective so as to steer processing in the direction of minimizing them. The reflexive character of autograms frustrates this approach. As we have seen, proximity is a false index. 'Near-perfect' solutions may be anything but near in terms of the number of changes needed to correct them, while a sentence with as many as eight discrepant totals might be perfected through replacing a single number-word. If hand-composition is obliged to rely on a mixture of guesswork, word-chopping, prayer, and luck, how can a more intelligent strategy be incorporated into a program?

I was pondering this impasse when Rudy Kousbroek's challenge presented itself, distracted my attention, and sent me off on a different tack. The sheer hopelessness of the undertaking caught my imagination. But was it actually impossible? What a comeback if it could really be pulled off! The task was to complete a letter-only autogram beginning, "This pangram contains ...". A solution, were it discoverable, must in a sense already exist 'out there' in the abstract realm of logological space. It was like seeking a number that has to satisfy certain predetermined mathematical conditions. And nobody—least of all Kousbroek—knew whether it existed or not. The thought of finding it was a tantalizing possibility. Reckless of long odds, I put aside programs and launched into a resolute attempt to discover it by hand-trial.

It was a foolhardy quest, a search for a needle in a haystack without even the reassurance of knowing that a needle had been concealed there in the first place. Two weeks' intermittent effort won only the consolation prize of a near-perfect solution: all totals correct save one; there were 21 *t*'s instead of the 29 claimed. With a

The SUMPROFILE

LABEL	e	f	g	h	i	l	n	o	r	s	t	u	v	w	x	y	NUMBER-WORD	LETTER
27	3	0	0	0	0	0	2	0	0	1	2	0	1	1	0	1	twenty-seven	E
6	0	0	0	0	1	0	0	0	0	1	0	0	0	0	1	0	six	F
3	2	0	0	1	0	0	0	0	1	0	1	0	0	0	0	0	three	G
5	1	1	0	0	1	0	0	0	0	0	0	0	1	0	0	0	five	H
11	3	0	0	0	0	1	1	0	0	0	0	0	1	0	0	0	eleven	I
2	0	0	0	0	0	0	0	1	0	0	1	0	0	1	0	0	two	L
20	1	0	0	0	0	0	1	0	0	0	2	0	0	1	0	1	twenty	N
14	2	1	0	0	0	0	1	1	1	0	1	1	0	0	0	0	fourteen	O
6	0	0	0	0	1	0	0	0	0	1	0	0	0	0	1	0	six	R
28	2	0	1	1	1	0	1	0	0	0	3	0	0	1	0	1	twenty-eight	S
29	2	0	0	0	1	0	3	0	0	0	2	0	0	1	0	1	twenty-nine	T
3	2	0	0	1	0	0	0	0	1	0	1	0	0	0	0	0	three	U
6	0	0	0	0	1	0	0	0	0	1	0	0	0	0	1	0	six	V
10	1	0	0	0	0	0	1	0	0	0	1	0	0	0	0	0	ten	W
4	0	1	0	0	0	0	0	1	1	0	0	1	0	0	0	0	four	X
5	1	1	0	0	1	0	0	0	0	0	0	0	1	0	0	0	five	Y
	7	2	2	2	4	1	10	11	2	24	7	1	2	5	1	1	INITIAL TEXT CONSTANTS	
	27	6	3	5	11	2	20	14	6	28	21	3	6	10	4	5	SUMPROFILE	

Fig. 1. A stack of PROFILES and initial text constants are added to produce a SUMPROFILE. The example shown is the hand-produced near-perfect pangram. All SUMPROFILES and label numbers coincide except that for *T*.

small fudge, it could even be brought to a shaky sort of resolution:

t t t t t
t
t
t

this pangram contains five *a*'s, one *b*, two *c*'s, two *d*'s, twenty-seven *e*'s, six *f*'s, three *g*'s, five *h*'s, eleven *i*'s, one *j*, one *k*, two *l*'s, two *m*'s, twenty *n*'s, fourteen *o*'s, two *p*'s, one *q*, six *r*'s, twenty-eight *s*'s, twenty-nine *t*'s, three *u*'s, six *v*'s, ten *w*'s, four *x*'s, five *y*'s, and one *z*.

To the purist in me, that single imperfection was a hideous fracture in an otherwise flawless crystal. Luckily, however, a promising new idea now suggested itself. The totals in the near-solution must represent a pretty realistic approach to what they would be in the perfect solution, assuming it existed. Why not use it as the basis for a systematic *computer search* through neighbouring combinations of number-words? Each of the near-solution totals could be seen as centred in a short range of consecutive possibilities within which the perfect total was likely to fall. The number of *f*'s, say, would probably turn out to lie somewhere between two and ten, a band of nine candidates clustered about 'six'. With these ranges defined, a program could be written to generate and test every combination of twenty-six number-words constructible by taking one from each. The test would consist in comparing these sets of potential totals with the computed letter frequencies they gave rise to, until an exact match was found, or until all cases had been examined. Blind searching might succeed where cunning was defeated.

Profiles

It isn't actually necessary to deal with all twenty-six totals. In English there are just ten letters of the alphabet that never occur in any number-word between *zero* and *hundred*, the one too low and the other too high to appear in the pangram. These are *a*, *b*, *c*, *d*, *j*, *k*, *m*, *p*, *q*, and *z*. The totals for these letters can thus be determined from the initial text and filled in directly:

This pangram contains five *a*'s, one *b*, two *c*'s, two *d*'s, ? e's, ? *f*'s, ? *g*'s, ? *h*'s, ? *i*'s, one *j*, one *k*, ? *l*'s, two *m*'s, ? *n*'s, ? *o*'s, two *p*'s, one *q*, ? *r*'s, ? *s*'s, ? *t*'s, ? *u*'s, ? *v*'s, ? *w*'s, ? *x*'s, ? *y*'s, and one *z*.

This leaves exactly sixteen critical totals. Counting up shows that there are already 7 *e*'s, 2 *f*'s, 2 *g*'s, 2 *h*'s, 4 *i*'s, 1 *l*, 10 *n*'s, 11 *o*'s, 2 *r*'s, 24 *s*'s, 7 *t*'s, 1 *u*, 2 *v*'s, 5 *w*'s, 1 *x*, and 1 *y*: sixteen constants that

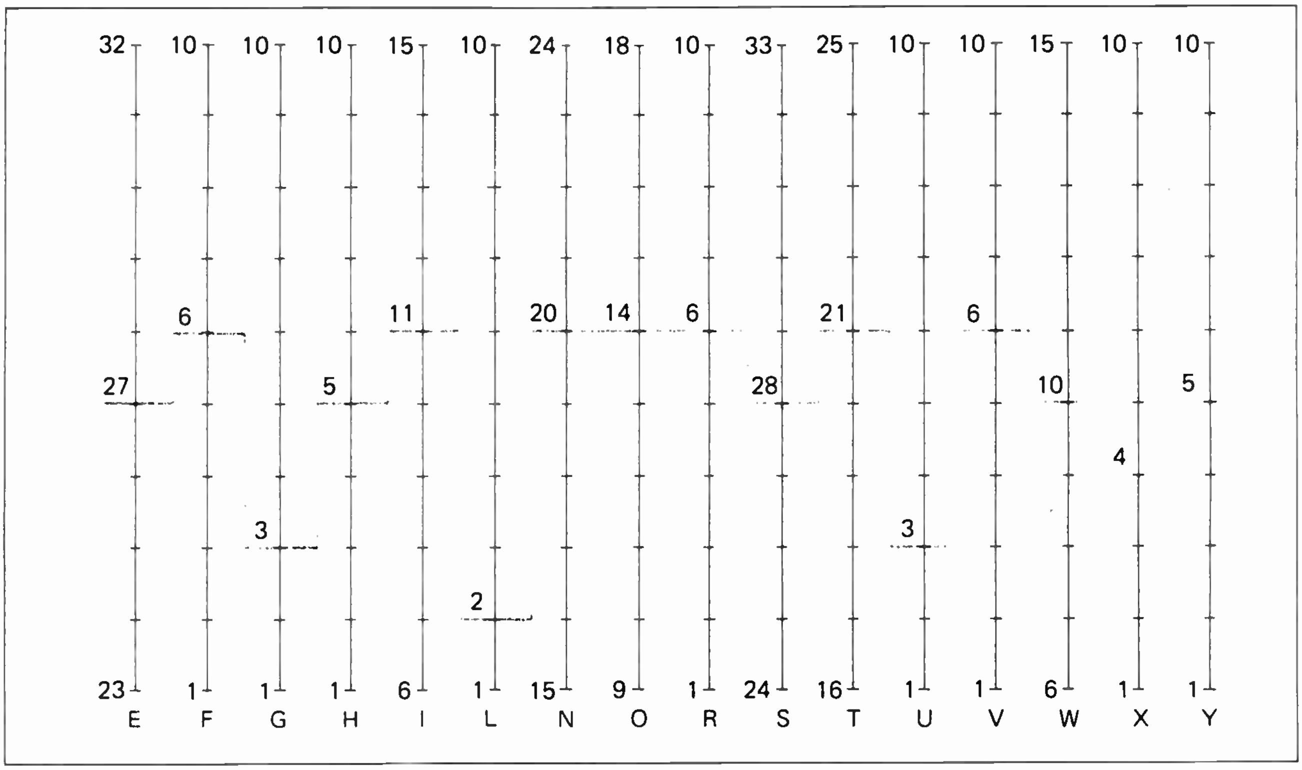


Fig. 2. The range of frequency values to be considered for each letter that appears in number-words.

must be added to those letters occurring in the trial list of sixteen number-words.

Though straightforward in principle, the program I then set out to write carried its practical complications. Number-words lack the regularity of numerals (in whatever base notation), still less the harmony of the numbers both stand for. An obvious step was to replace number-words by PROFILES: alphabetically ordered sixteen-element lists representing their letter content. The PROFILE for *twenty-seven*, for instance, would be:

e f g h i l n o r s t u v w x y
(3 0 0 0 0 0 2 0 0 1 2 0 1 1 0 1)

The letters above the list are for guidance only, and form no part of the PROFILE itself. A special case was the PROFILE for *one*, which provided for the disappearance of plural *s* ('one *x*, two *x*'s') by including −1 in the *s* position. PROFILES for all number-words up to *fifty* (anything higher than *forty* was unlikely ever to be needed) were stored in memory, and a label associated with each. These labels were chosen to coincide with the number represented. The label for the PROFILE of *twenty-seven*, for example, would be the decimal number 27.

Starting with the lowest, a simple algorithm could now generate successive combinations of labels, that is, numbers, drawn from the 16 pre-defined ranges. We shall return to these shortly. Each set of lables would be used to call up the associated set of PROFILES. These 16 PROFILES would be added together element for element, and the resulting sums in turn added to the above-mentioned constants so as to form a SUMPROFILE—see Fig. 1. The SUMPROFILE would thus contain the true letter frequencies for the presently activated sentence (the 16 number-words represented by the current combination of labels plus residual text). All that remained was for the program to check whether the numbers in the SUMPROFILE coincided with the present set of PROFILE labels. If so, the candidate combination of number-words agreed with the real totals and the pangram had been found. If not, generate the next combinations and try again ... .

The simplicity of this design conveys no hint of the uncounted alternatives reconnoitered before reaching it. The 'obvious' PROFILES were not quite so conspicuous as suggested, being in fact a later improvement over a previous look-up table. Weeks were spent in exploring a quite different approach that sought to exploit the mutual-cancelling technique formerly used in hand-composition. By the time the final version of the program had come into focus, half a dozen prototypes lay behind and several months had slipped by. In the mean time, cheerful enthusiasm had given way to single-minded intensity as the problem wormed its way under my skin. Neither was I working entirely alone. Word of the pangram puzzle had spread among colleagues, discussion sprang up and contending design philosophies were urged. At one stage, complaint of "excessive CPU-time devoted to word games" came in from the University of Nijmegen Computing Centre, whose facilities had been shamelessly pressed into service. This was when rival programs were running simultaneously. It was bad enough to be in search of a Holy Grail that might not even exist; the thought of someone else finding it first added a sticky sense of urgency to the hunt.

The question of determining the exact ranges of number-words to be examined seemed to me an essentially trivial one, and I put it off until last. The important thing was to get the program running. For the time being it was enough to decide what the lowest combination was going to be, and to let the algorithm generate all possibilities up to, say, ten higher for each number-word. In terms of software it was convenient for ranges to be of equal length; ten might be unnecessarily high, but better the net be too large than that the fish should escape. Since the totals in the near-solution were to define the midpoint of these ranges, their lower limits would commence at about five less. 'Fourteen

Continued on page 15

SPEED CONTROL FOR 3-PHASE MOTORS

by F. P. Zantis

Three-phase motors, particularly asynchronous types, are very reliable and consequently popular prime movers. A drawback of these engines is, however, their fixed speed of rotation which depends on the frequency of the applied voltage. This article describes means of varying that frequency and thus controlling the speed of the motor over a wide range.

Until not so long ago, the only way of varying the frequency of the voltage driving a three-phase motor was the use of a rotary converter . However, the advent of power semiconductors has made possible the development of static frequency changers that convert an alternating voltage of a given frequency into one of a different frequency. These devices may also change the number of phases, for instance, single-phase current into three-phase current.

Apart from their application with prime movers, such frequency changers are also very useful in standby and emergency power supplies and in the control circuit of oscillators. In these applications, the widely varying direct output voltage of a battery is converted to an alternating voltage of steady level and frequency.

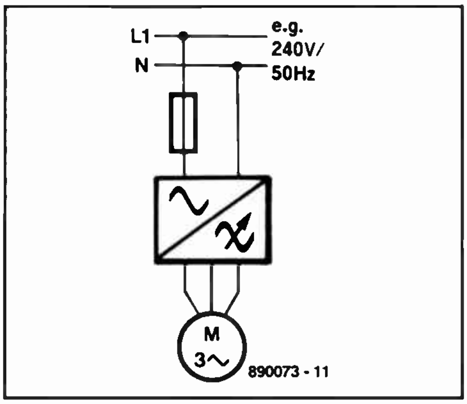


Fig. 1. Schematic representation of how a 3-phase motor may be operated from a single-phase supply under the control of a frequency changer.

A frequency changer to control the speed of a three-phase motor must have an output voltage whose level and frequency are both variable. A proportional change in the level and frequency of the voltage applied to the motor enables the speed of rotation of the motor to be varied at constant moment as is shown schematically in Fig. 1.

Speed control offers many advantages, such as a saving in energy, reduced maintenance costs, and optimalization or greater flexibility of operation.

Frequency changers

All frequency changers work on the same principle: the drive voltage (normally 240 V or 415 V, 50 Hz mains) is rectified, smoothed (filtered) and applied to the motor via an inverter (= dc-to-ac converter) as shown schematically in Fig. 2.

The construction of the various sections depends on the type of frequency changer. For instance, the power section of a small frequency changer is shown (greatly simplified) in Fig.3.

The **rectifier** is normally a conventional bridge type. The voltage source is loaded by the required current only, independent of the motor rating. That results in a good power factor which is virtually constant (typically, cos $\phi = 0.97$) over the entire load and speed ranges.

The **smoothing circuit** is normally an *LC* low-pass filter, whose inductance protects the mains against transients. The large values of capacitance required with high loads are obtained by series and parallel connection of a number of high-voltage electrolytic capacitors as shown in Fig. 4. Matched resistors ensure correct division of the voltage.

The **inverter** consists of three pairs of transistors that are arranged in a star configuration. The three motor phases are connected cyclically at 120° intervals to, respectively, the positive and negative terminal of the smoothing filter, which results in a rotating field being induced in the motor. Appropriate control of the inverter enables the smooth, precise control of the output frequency. In general, three-phase motors may be operated at up to twice their rated speed. This means that, for example, a four-pole

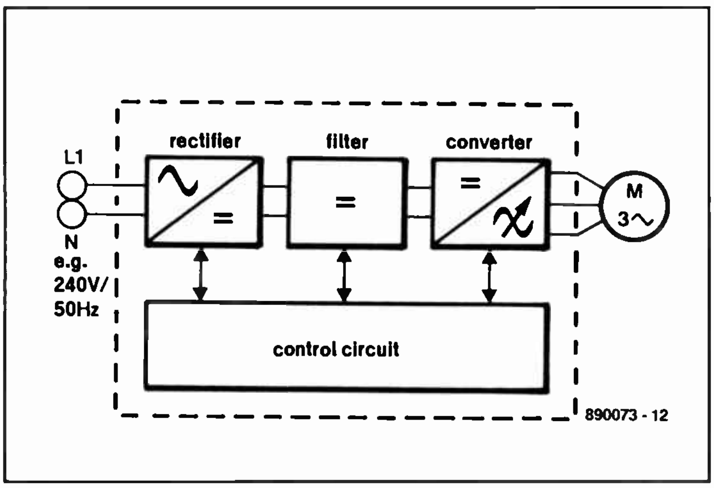


Fig. 2. Block diagram of a basic frequency changer.

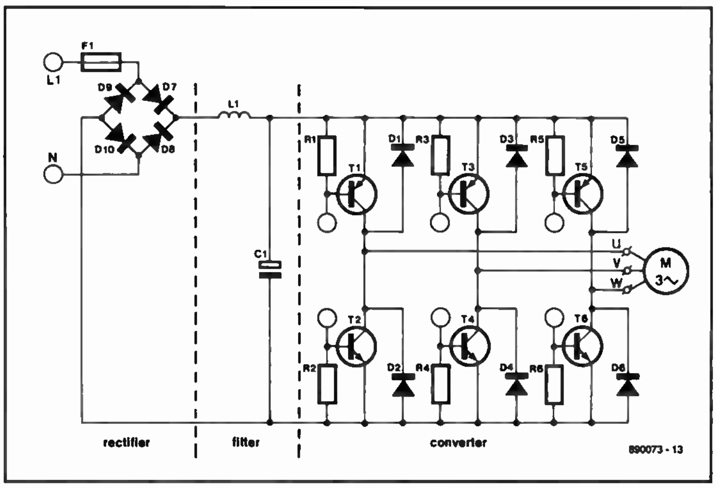


Fig. 3. Basic circuit of the power section of a small frequency changer.

three-phase motor may be operated at up to 3000 rev/min. Such control can not, however, be achieved by varying only the frequency.

To obtain a constant moment, the magnetic flux in the stator of the motor must be related to the set frequency. To that end, the motor voltage must be increased or reduced, as the case may be, in direct proportion to the frequency. When the frequency is higher than the rated frequency of the motor (50 Hz), the voltage can not rise, since the frequency changer can not generate a potential that is higher than the applied voltage.

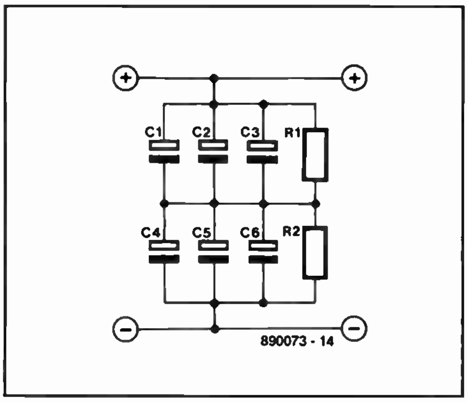


Fig. 4. The filter capacitance may consist of series and parallel arrangements of small capacitors.

This results in a reduction of the magnetic flux and thus of the moment. Operation above the rated frequency is thus not possible at the rated moment.

The basic relation between supply frequency, f_s, the ratio of actual motor speed to rated motor speed, n/n_r, and the ratio of the actual motor voltage to the rated motor voltage, U/U_r, is given by the characteristic in Fig. 5, while the relation between n/n_r and the ratio of the actual moment and rated moment, M/M_r, is shown in Fig. 6.

The power section of the inverter is preceded by a control stage which, among others, contains the electronics for the generation of the variable-frequency rotating field. Two basic circuits for the generation of this field are shown in Fig. 7: both are digital phase advancers.

Figure 7a provides a presettable frequency to control a ring counter. The outputs of this counter are connected to a binary-coded-decimal (BCD) decoder. The signals that are shifted by 120° with respect to one another, and which are required for driving the power semiconductors, appear at the output of the three bistables.

The circuit in Fig. 7b is rather simpler in that it does not include a BCD decoder. The AND gate obviates any unacceptable state, but has no other effect on the operation of the phase advancer.

For a rotating field frequency of 50 Hz, the clock of both designs must be 300 Hz.

Some semiconductor manufacturers provide special ICs for the generation of the rotating field, but in state-of-the-art frequency changers microprocessors are used for this purpose.

As already stated, the output voltage must rise in direct proportion to the frequency. There are a number of ways in which the output voltage may be varied: the two most important of these, pulse-amplitude modulation—PAM—and pulse-duration modulation—PDM—will be described below.

Pulse-amplitude modulation

In the PAM method of varying the output voltage, a chopper (electronic switch), following the smoothing filter, opens and closes at a rate determined by the control stage (see Fig. 8). This results in a variable direct voltage that is in direct proportion to the duty factor of the control signal.

The level of the voltage can be adjusted

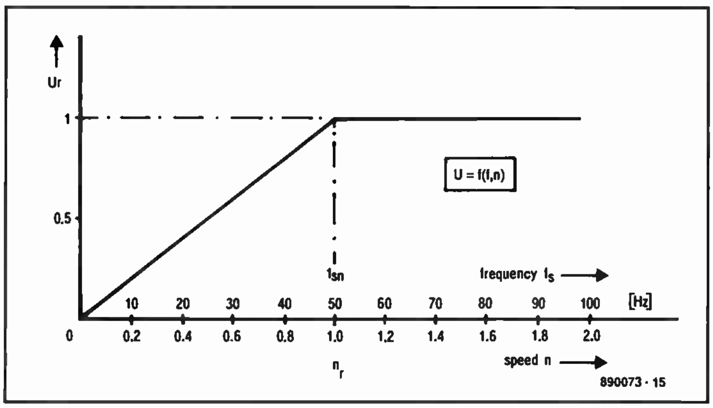


Fig. 5. Characteristic curve of the motor voltage vs rotational speed vs source frequency.

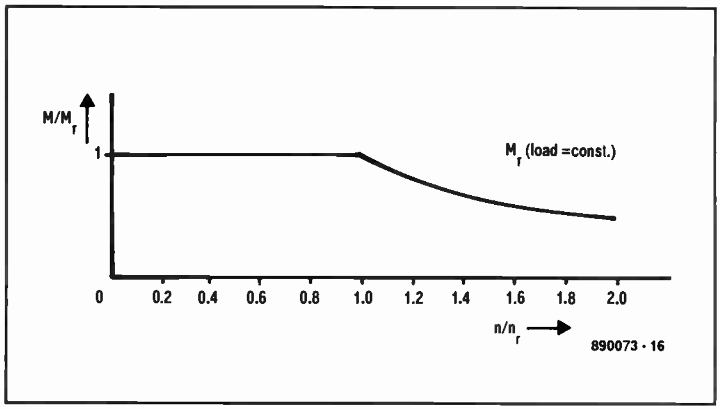


Fig. 6. Moment characteristic.

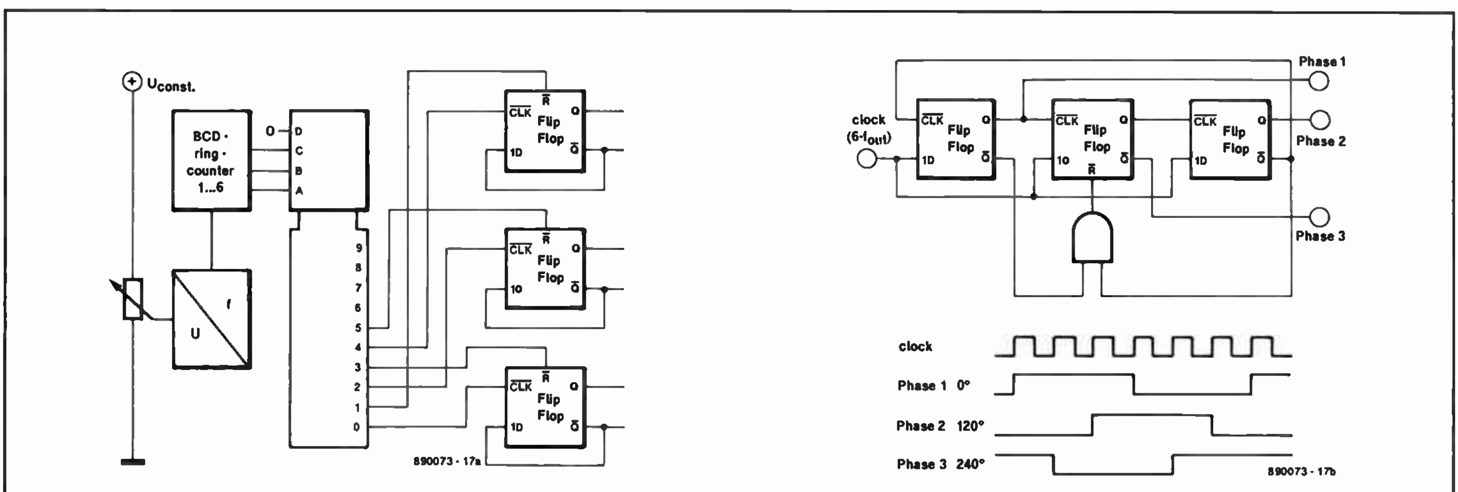


Fig. 7. Two possible designs for use with three-phase generators: both generate three signals, the phase of each of which is shifted 120° with respect to the other two.

in proportion to the frequency in a manner which ensures that the ratio of output voltage to output frequency remains constant. For a 400 V motor, this ratio must be 400 V:50 Hz = 8 V per Hz.

In the type of frequency changer described here, the output voltage is built up from six pulses per period (see Fig. 9). That voltage is not sinusoidal: it contains, apart from the fundamental frequency component, a number of harmonics of which the 5th, 7th, 11th and 13th are the most important. These harmonics cause the motor to produce spurious moments and result in additional losses in energy.

Because of the inductance of the stator, the current through the motor is rather more sinusoidal than the voltage. The waveform becomes better when the voltage is built up from a larger number of pulses per period. For instance, when there are 18 pulses per period, the output voltage and current are shaped as shown in Figure 10. The amplitude of the harmonics is then much smaller so that the motor runs rather more smoothly.

The PAM frequency changer may also be designed with a controlled rectifier instead of a chopper as shown in Fig. 11. The amplitude of the voltage across the smoothing filter can then be varied by controlling the phase angle. The main drawback of this method is the feedback on to the source (mains), which manifests itself in a poor power factor and relatively high-level harmonics. These harmonics must be filtered out, otherwise the user might run into trouble with the relevant electricity supply authority.

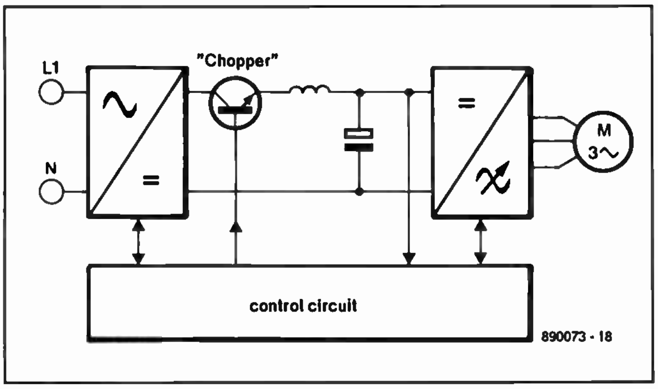


Fig. 8. Frequency changer using pulse-amplitude modulation.

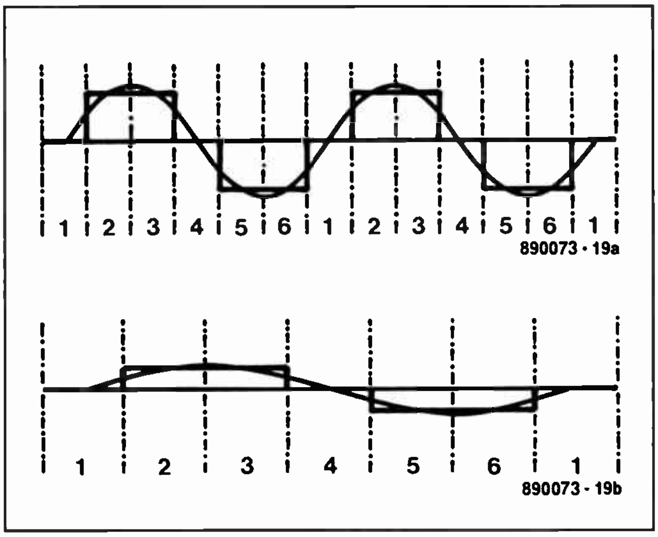


Fig. 9. Output voltage of a PAM frequency changer built up from 6 pulses per period: (a) at maximum source voltage and frequency and (b) at half the maximum source voltage and frequency.

Pulse-duration modulation

Frequency changers using pulse-duration modulation have no chopper or controlled rectifier, so that the voltage across the smoothing filter assumes a fairly constant value that is dependent only on the supply voltage and the load.

The level and shape of the output voltage are determined solely by the width of the output pulses of the inverter.

The basis of this type of modulation is a comparison between a triangular and a sinusoidal signal. The frequency of the triangular signal is much higher than that of the sine wave (see Fig. 12). The resulting pulse train is the control instrument.

As before, because of the integrating action of the stator inductance,

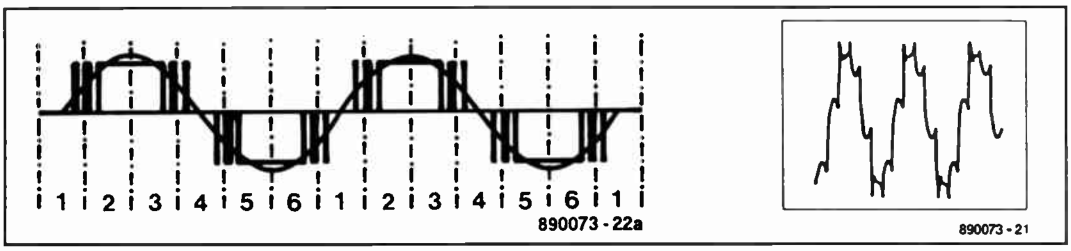


Fig. 10. Output voltage of a PAM frequency changer built up from 18 pulses per period (a) and corresponding current (b).

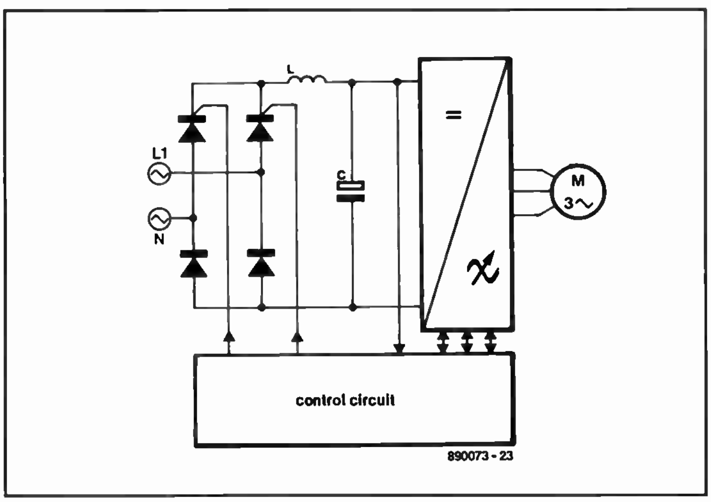


Fig. 11. The PAM frequency changer with controlled rectifier.

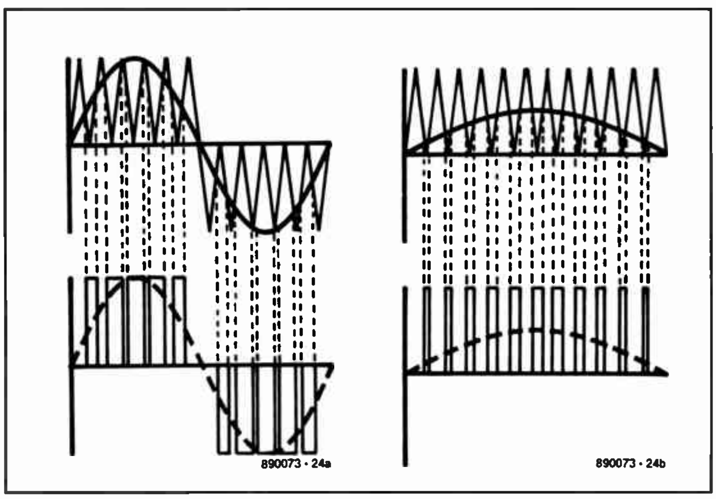


Fig. 12. Output voltage of a PDM frequency changer: (a) maximum source voltage and frequency; (b) half the maximum source voltage and frequency.

the current is sinusoidal for all practical purposes.

The design of the control circuitry is much simplified by the availability of special ICs that contain the comparator, sine wave generator and triangular-wave generator. Again, the design may also be based on the use of microprocessors.

The frequency of the sinusoidal voltage determines the output frequency of the frequency changer and thus the speed of the motor. When the frequency of the triangular signal is constant, pulses are generated that depend on the output frequency. At low output frequencies, the number of pulses per period is relatively high, which results in a near-sinusoidal current through the motor. When the output frequency is high, the waveform of the current deteriorates.

High-voltage switching transistors may be used in the power section of inverters as long as the output power does not exceed 50 kVA. Modules are available from some manufacturers that contain all six transistors required for a three-phase bridge circuit.

When the output power exceeds 50 kVA, thyristors must be used. Figure 13 shows one phase circuit of the power section of a thyristors-based inverter; the other two circuits are identical. Since thyristors can not be switched off in a simple manner, some additional components are required.

Each phase circuit needs four thyristors, of which two, Th1 and Th2, switch the current through the motor. The other two, Th3 and Th4, are required for the so-called commutation process, which is described below.

Commutation

During forward conduction all the junctions of a thyristor are forward biased. To be able to turn off the device, the charge carriers must be removed and this is usually done by applying a reverse voltage across the thyristor, a process known as commutation. The process wil be described with the aid of the circuit shown in Fig.14, which is part of that in Fig. 13.

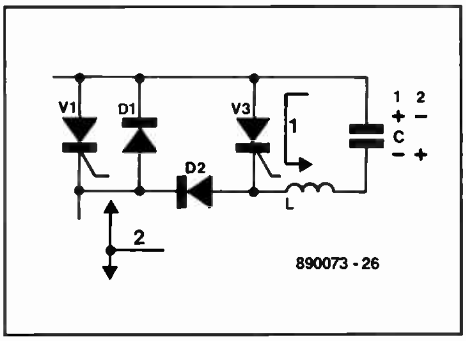


Fig. 14. Simplified commutation circuit (detail of Figure 13).

The simplified circuit in Fig. 14 consists of two thyristors, a commutation *LC* network, which forms a series resonant circuit, and two diodes.

In order to turn off Th1, Th3 must be fired. At first, capacitor C is charged as shown (1). This results in a sinusoidal current in the *LC* circuit. The first half period of that current flows through Th3, which then fires. When the current reaches a zero crossing, Th3 is turned off and the polarity of C is reversed (2). The next half period of the current flows through the diode and Th1 or the motor. For a brief instant, the motor current flows through the *LC* circuit: no current then flows through Th1 and it turns off.

The remaining components in Fig. 14 protect the power semiconductors or are needed to prepare the commutation process.

The operation of this type of inverter depends on the voltage across the motor and the inverter is, therefore, called a voltage inverter. The level of the current through the motor depends entirely on the load on the motor.

It is also possible to base the frequency changer on the processing of the current, in which case the load on the motor determines the voltage across the motor. This type of frequency changer is called a current-based frequency changer

The design of the power section of a current-based frequency changer is shown, greatly simplified, in Fig. 15. Note that the smoothing filter does not contain a capacitor. The controlled rectifier at the input determines the r.m.s. value of the current through the smoothing filter. The inductor smoothes the direct current. Before it is fed to the stators of the motor, the current through the filter is divided in the inverter in a manner to ensure that a rotating field is induced. The current through, and the voltage across, the motor are shown in Fig. 16. Note that in this type of frequency changer the current waveform is rectangular, whereas the voltage is near-sinusoidal.

Because of the controlled rectifier, the power factor at the source end of a current-based frequency changer is not constant, but varies in accordance with the load. Furthermore, this type of frequency changer can not be used in parallel operation of motors, because the commutation capacitor in the inverter must be adapted to the rating of the motor.

Since in a current-based frequency changer the motor is part of the commutation network, such a frequency changer, in contrast to a voltage-based changer, can not be operated without the motor. On the other hand, the design of the inverter for a current-

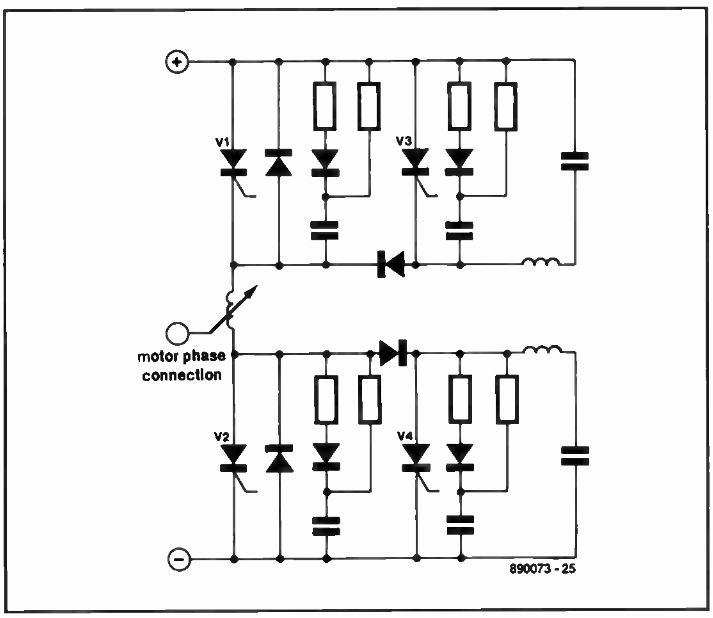


Fig. 13. Diagram of one phase circuit of a frequency changer based on thyristors, also called voltage-based frequency changer.

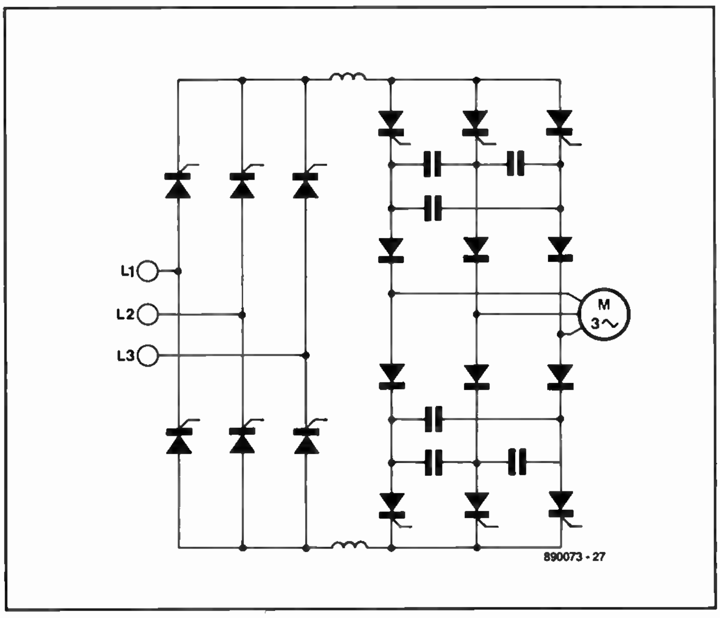


Fig. 15. Circuit diagram of the power section of a current-based frequency changer.

based frequency changer is much simpler than that for a voltage-based model. Furthermore, the commutation process is simply effected by the inverter and the rotating field of the motor without the need of additional components.

However, the great advantage of a current-based frequency changer is not its simpler design, but the possibity of feeding back the brake horsepower into the source without the need of additional components. When the motor thus operates as a generator, the current through it is reversed and fed to the controlled rectifier via the smoothing filter. The rectifier then functions as an inverter, which returns the energy to the source.

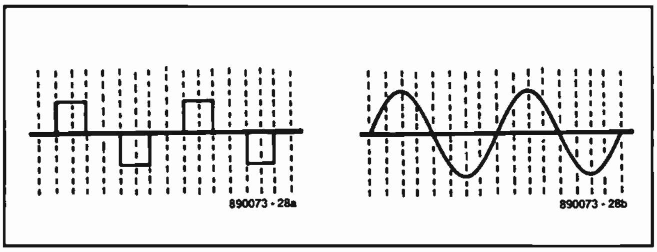


Fig. 16. Current through, and voltage across, a three-phase motor operated with the aid of a current-based frequency changer.

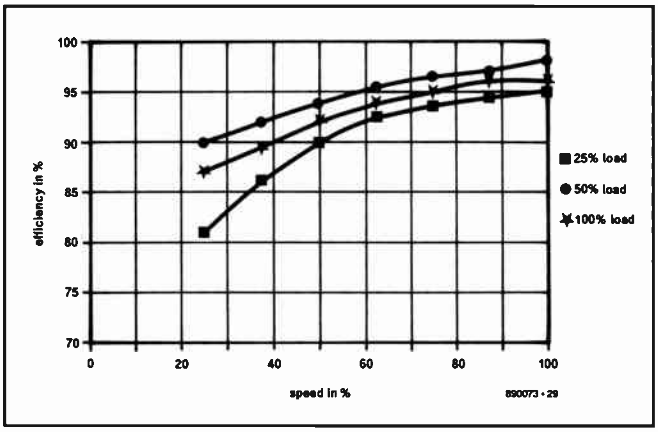


Fig. 17. Typical efficiency curves of a voltage-based frequency changer.

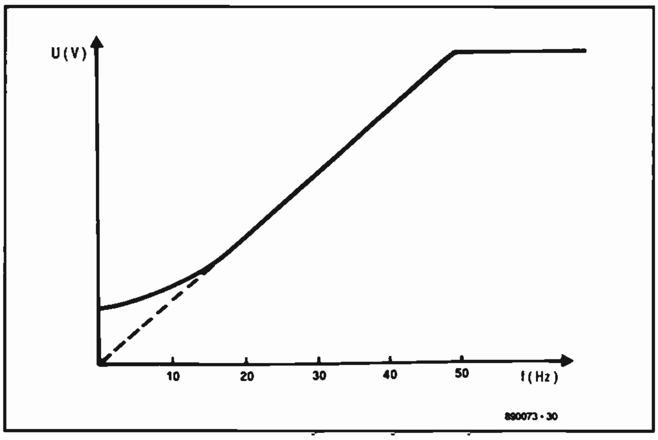


Fig. 18. Output voltage vs frequency characteristic when the voltage is raised at low frequencies.

Efficiency

The efficiency of the modern frequency changers discussed in this article is pleasantly high: depending on the design and technology employed, it lies somewhere between 93% and 97%. Typical efficiency curves of a voltage-based frequency changer operating with various loads are given in Fig. 17.

The overall dissipation includes losses in diodes, thyristors, inductors and control sections. Most of the dissipation, however, occurs in the commutation process. The overall dissipation is, therefore, highly dependent on the number of commutations per unit time. Thus, although high pulse rates reduce the harmonic content, they increase the dissipation.

At frequencies below 10 Hz, the potential drop across the resistance of the motor winding reduces the flux and the moment. In these circumstances, this may be compensated by a higher than proportional rise in the output voltage as indicated by the characteristic in Fig. 18.

Miscellaneous

When adapting your own requirements to the observations and discussions in this article, bear in mind that a number of components are shown for very large loads. In no circumstances, for instance, must the large capacitors shown in the smoothing filters be connected directly across the mains or other high-voltage source since at the moment of first switch-on they form a short circuit. Such high-value capacitors must be charged via a suitable resistor. Only when they have been charged can they be connected to the input source and even then via a suitable protection circuit.

The potential across the smoothing filter rises to the peak value of the source voltage. The capacitor(s) will retain a lethal charge even after the source has been disconnected.

The commutation process causes high voltage peaks, which should be borne in mind during test and measurement.

As a rule, all components in the power section carry a potential to earth. Only very small frequency changers can be isolated from the source (mains) by a suitable transformer. Again, this should be borne in mind during test and measurement. Also, test instruments whose 'zero potential' is connected to their case (as in a number of oscilloscopes) must be connected to the mains via an isolating transformer. There exists then, however, a potential between the case and earth. **BE WARNED** that in certain circumstances it may be dangerous to touch the case.

The voltage peaks in the inverter may reach values that exceed the isolation breakdown voltage of the test instrument. It is therefore important that for each instrument at least the maximum input voltage and the highest permissible voltage with respect to earth or to case are known and strictly observed. Furthermore, in order to judge whether a given instrument is suitable for certain measurements, its principle of operation must be known. This is particularly so when alternating voltages and currents are measured whose waveform does not match that with which the instrument has been calibrated. ■

FURTHER READING:

Power Electronics Handbook
by F.F. Mazda
ISBN 0 408 03004 6
Butterworth Scientific Ltd
Westbury House,
Bury Street
GUILDFORD GU2 5BH
England
(*Elektor Electronics* June 1990)

Power Electronics
by M.H. Rashid
ISBN 0 13 686619 0
Prentice-Hall
66 Wood End Lane
HEMEL HEMPSTEAD HP2 4RG
England
(*Elektor Electronics* October 1988)

Electric Machinery
by Peter F. Ryff
ISBN 0 13 248691 1
Prentice-Hall
Wood End Lane
HEMEL HEMPSTEAD HP2 4RG
England
(*Elektor Electronics* January 1989)

Solving Problems in Electrical Power and Power Electronics
by H.F.G. Gwyther
ISBN 0 582 28644 1
Longman Scientific & Technical
Longman House
Burnt Mill
HARLOW CM20 2JE
England
(*Elektor Electronics* March 1989)

S-VHS/CVBS-TO-RGB CONVERTER

PART 1: INTRODUCTION

Although the technical advantages of the Super-VHS video system are well proven, many owners of an S-VHS video recorder balk at the expense of a compatible monitor or TV set with separate luminance and chrominance inputs. This article describes an obvious missing link in the apparently ever-incompatible field of video equipment connections. An advanced circuit is discussed that converts S-VHS or CVBS (composite) video signals into RGB components. The upshot is that you can use your existing monitor with an RGB input (i.e., with a SCART or Euro-AV connection) to benefit from the improved picture resolution offered by an S-VHS video recorder. This month we discuss the basics of the video standards involved.

H. Reelsen

The compatibility issue has played a significant role in the development of both the NTSC and the PAL TV transmission systems. In both cases, there were two conflicting aspects: on the one hand, existing monochrome TV sets were not to be affected by colour transmissions; on the other hand, existing bandwidths of about 5 MHz for the luminance (brightness) signal were to be maintained.

The compatibility requirement automatically dictates that the black-and-white information (luminance or 'Y' signal) must also be conveyed in colour transmissions. The Y signal forms the sum of all basic colours, red (R), green (G) and blue (B), but only as far as their relative brightness is concerned. From perception experiments, the brightness appears to determine the overall sharpness of the picture. Hence, the luminance bandwidth must be as large as possible (up to 5 MHz) for monochrome as well as colour TV sets. However, this raises the problem of where to put the colour information.

Colour components and transmission

Any colour can be reproduced on a picture tube by actuating in the correct proportion the basic colours it is composed of. The final colour is obtained by controlling the intensity at which the RGB pixels at the inside of the picture tube light up. To the human eye, the three individual basic colours in a pixel group appear as one, composite, colour or hue at a particular saturation.

The need to convey R, G and B, is, therefore, obvious. Since the sum of the equivalent luminance values of all three is already contained in the Y signal, only two further signals, R–Y and B–Y, are generated by means of a differential operation with the Y signal. R–Y and B–Y are therefore referred to as the colour difference signals. Before these signals are transmitted, they are given relative brightness factors. The resulting chrominance signals may be written as

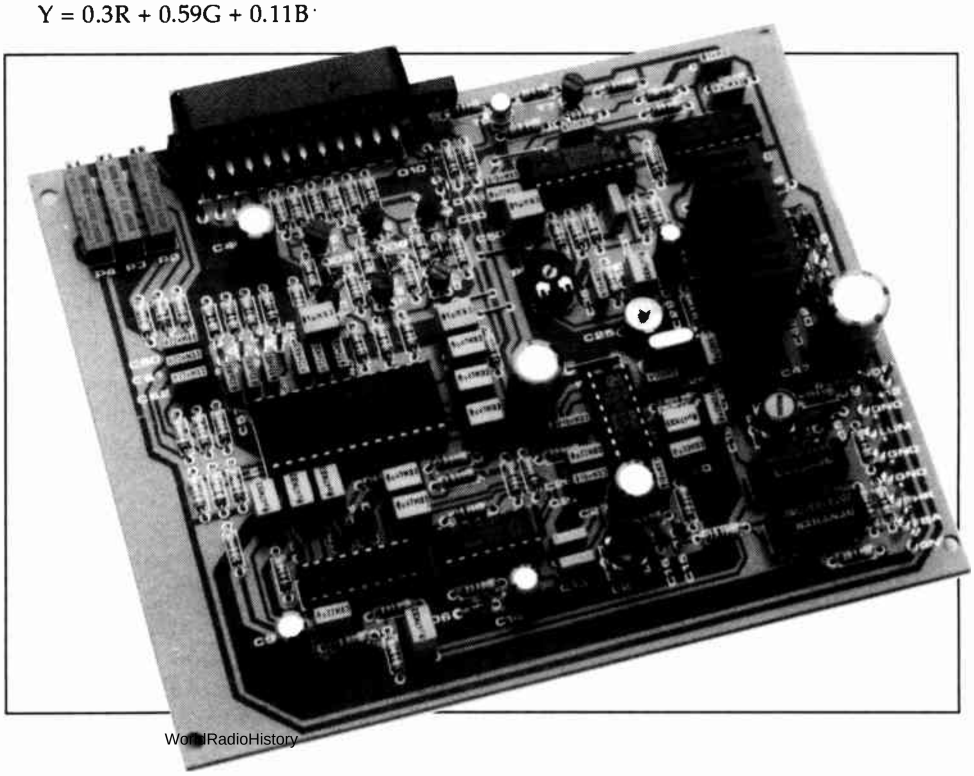
$$U = 0.49(B-Y)$$
$$V = 0.88(R-Y)$$

and the luminance, Y, as

$$Y = 0.3R + 0.59G + 0.11B$$

The RGB intensity information required to control the respective electron guns in the picture tube is obtained from the R–Y, B–Y and Y information with the aid of an addition operation in a matrix circuit.

A problem that remains to be solved is how to include the colour difference signal in the bandwidth already occupied by the Y signal, without causing interference on monochrome TV sets, or reducing the picture sharpness on colour sets. At this point, design engineers are in a position to profit from a characteristic of human eye, namely its reduced ability to resolve colour contours as compared to brightness values. This means that the colour infor-



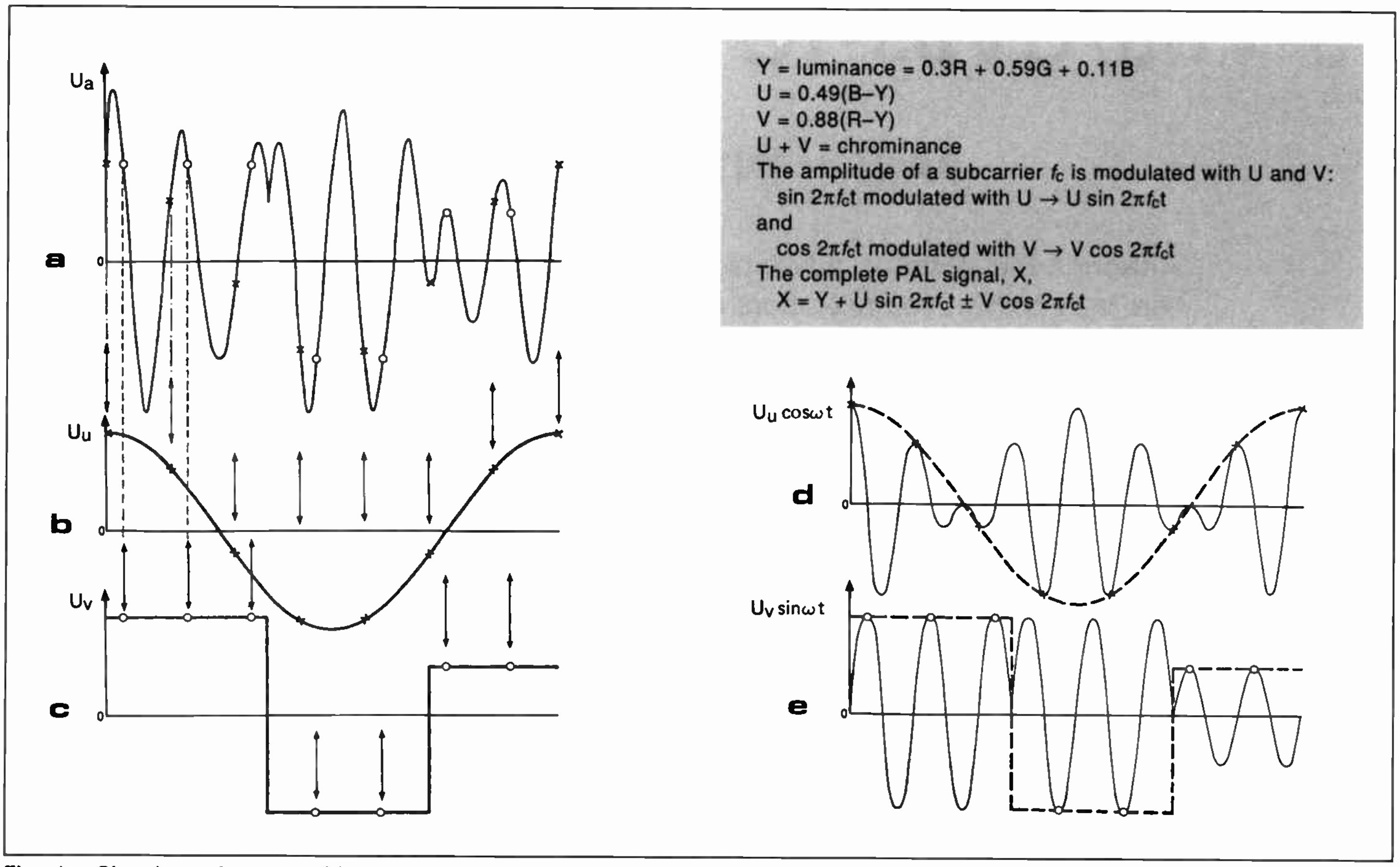


Fig. 1. Signal waveforms resulting from quadrature modulation of the colour difference signals U_u = 0.49(B–Y) and U_v = 0.88(R–Y). Drawing 'a' shows the quadrature-modulated signal U_a, while 'b' and 'c' show the modulation signals U_u and U_v, which for clarity's sake are formed by a sinusoidal waveform and a rectangular waveform respectively. Drawings 'd' and 'e' illustrate how these signals are modulated on to the 90-degrees shifted carriers. The waveform shown in drawing 'a' is the result of adding the signals in 'd' and 'e'.

mation may be transmitted at a relatively low bandwidth without significantly degrading the overall sharpness of the picture. In the PAL system, the colour (or chrominance) bandwidth is about 1 MHz.

The colour difference signals are readily embedded in the frequency spectrum of the Y signal by making use of the fact that the spectral lines of the Y signal occur at even multiples of the line frequency (15,625 Hz). Also, the amplitude of these spectral lines decreases with frequency.

The colour difference signals modulate a subcarrier of which the frequency, f_c, is an odd multiple of the line frequency divided by four, plus the picture refresh frequency (see Ref. 1):

$$f_c = 1135 \times (15{,}625/4) + 25 \quad \text{(Hz)}$$

This causes the spectral lines of the colour difference signal to be slotted in between those of the Y signal. The colour subcarrier frequency is set at 4.43361875 MHz, and the colour difference signals are quadrature-amplitude modulated (QAM). The B–Y and R–Y components modulate the amplitude of the colour subcarriers of 0 degrees and 90 degrees respectively (see Figs. 1d and 1e). The carrier itself is suppressed, so that it has an amplitude of nought in the absence of a colour difference signal. This is done to keep the picture free from interference caused by the otherwise continuously present subcarrier.

In order to eliminate the risk of phase shifts in the transmission path, the phase of the R–Y component is inverted every other picture line. Details of this operation peculiar to the PAL system may be found in Refs. 2 and 3.

The use of amplitude modulation with suppressed carrier requires a phase- and frequency-synchronized subcarrier at the receiver side. In a TV set, the modulated R–Y and B–Y components are recovered from the chrominance subcarrier with the aid of a 4.433-MHz quartz crystal oscillator whose phase and frequency are corrected every 64 µs by a 2-µs long burst signal slotted into the rear porch in the blanking period at the end of every picture. The burst consists of 8 to 11 cycles of the colour subcarrier frequency and follows the line sync pulse as shown in Fig. 2. A phase comparator is used to keep the crystal oscillator synchronized to the received burst, which also contains the PAL switch signal for the line-by-line R–Y phase reversal. This arrangement ensures that the R–Y signal in the receiver is inverted in synchronism with that at the transmitter side to ensure that the demodulation operation can work correctly.

Pitfalls...

In practice, the 'packaging' of the luminance and the chrominance information into a single CVBS (chrominance-video-blanking-synchronisation) signal is not without problems. Since the colour subcarrier falls in the spectrum of the luminance signal, it causes a finely patterned type of interference known as moiré. Luminance circuits in all modern TV sets are therefore fitted with a 'colour trap', which is a relatively simple filter that removes most of the moiré effects with the exception of those occurring at areas with sharp colour transitions. Here, large phase jumps give rise to subcarrier sidebands that fall outside the stop band of the 4.43-MHz colour trap. Unfortunately, Y signals in this stop band are also suppressed, which results in reduced picture resolution because some of the high-frequency components disappear. Incidentally, most monochrome sets also contain a colour trap to eliminate moiré.

The (possible) interference between chrominance and luminance also works the other way around: since the lumin-

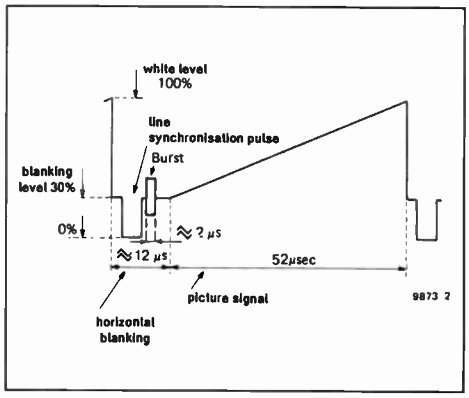


Fig. 2. Structure and timing of a composite video signal (PAL standard).

ance band includes the frequency range for the colour subcarrier, high-frequency Y signals can cause interference in the frequency range around 4.43 MHz. The result is a quasi-random type of patterning and colouring in and around picture areas of fine detail. Notorious examples of this happening can be seen virtually every evening in jackets, shirts or ties of people on television.

Standard VHS video recorders

Some 15 years ago, during the development of the VHS video system, a luminance bandwidth of 3 MHz was deemed satisfactory for VCRs considering the technical limitations imposed by the drum head speed and the tape consumption. In the original VHS system, the colour subcarrier is mixed down to 627 kHz to keep it well way from the lower end of the spectrum of the Y information, which is recorded as a frequency-modulated (FM) signal (see Fig. 3)

The FM recording improves the signal-to-noise ratio of the Y signal and makes it largely independent of amplitude variations of the tape signal. The frequency sweep ranges from 3.8 MHz to 4.8 MHz.

Returning to the colour information, this is recorded as an analogue signal in 'helical scan' mode (Ref. 3). The different frequencies used allow ready separation of the two signals. However, the bandwidth of the colour information is inevitably reduced to about 500 kHz. The result is noticed as 'smeared' colour transitions, to which the reduced (3-MHz) luminance bandwidth adds an impression that the picture is blurred.

These imperfections of the original VHS system were soon recognized by VCR manufacturers. Their answer, the HQ video recorder, was based on small improvements to the recording method and a better edge definition of the Y signal. The resultant picture quality improvement was marginal and not really a step forward. It was, however, the best that could be achieved given the need for continued compatibility. Clearly, real improvements to the picture quality offered by VCRs could be achieved only by changing some of the standards.

The Super-VHS system

The bandwidth of the recorded video signal was increased significantly (at the existing relative speed of 4.85 m/s between the tape and the head) by virtue of two technological developments. First, new metallurgic techniques allowed the size of the air gap of the video head to be reduced. Second, tapes with a very high magnetic particle density became available.

To maintain compatibility with older VHS recorders, the S-VHS system is based on the same method of colour recording (see Fig. 3). However, the frequency

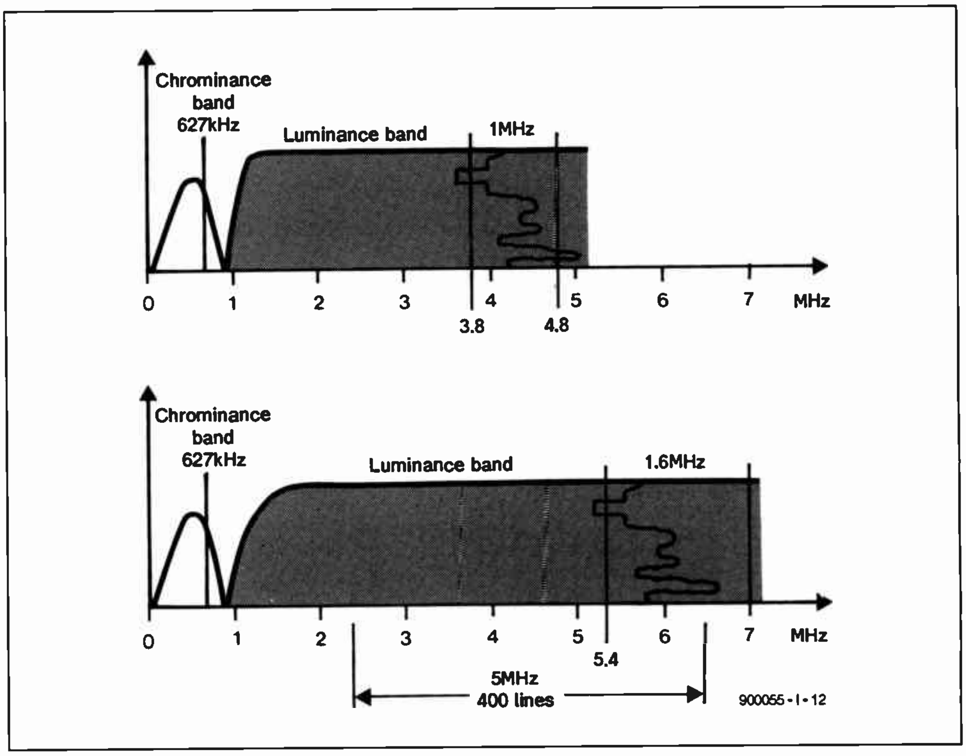


Fig. 3. Typical standard-VHS and S-VHS spectra. In both cases, the quadrature-modulated colour signal is recorded with the aid of a carrier which is mixed down to 627 kHz, while the luminance signal (Y) is recorded in FM. S-VHS recorders use a luminance carrier frequency of 5.4 MHz and a frequency sweep of 1.6 MHz. This offers a bandwidth of 5 MHz for the Y signal, as opposed to about 3 MHz for the standard-VHS video recorder.

sweep of the Y signal is shifted up to a band from 5.4 MHz to 7.0 MHz to give a much higher noise margin. At the same time, the frequency of the FM subcarrier allows the luminance signal to be recorded at its full bandwidth of about 5 MHz.

In most standard VHS video recorders, the chrominance and luminance signals are processed separately until they are combined to give a CVBS signal with all the previously mentioned risks of running into trouble with interference.

By contrast, the S-VHS system is based on separate chrominance and luminance signals right up to the two associated outputs on the VCR. Evidently, this separation is not perfect when, for instance, a TV programme is recorded, since then the chrominance and luminance components must be extracted from the composite signal before they can be recorded, played back and fed separately to a monitor. The process of extracting the components has pitfalls as described before. Not so, however, with video sources that do supply the components separately. Examples include S-VHS cameras, some prerecorded S-VHS video tapes and MAC decoders.

Connection problems

So far, so good. A look at the rear panel of the TV set, however, reveals that there is at best a SCART connector, which does not allow luminance and chrominance signals to be taken in separately. The TV set is, therefore, not S-VHS compatible. This unfortunate discovery forces owners of S-VHS recorders to connect the monitor and the recorder via a CVBS link, forgoing most of the advantages of better picture reproduction offered by the new video system.

Considering the cost of an S-VHS compatible monitor, the only way to benefit from the separate chrominance and luminance signals supplied by S-VHS recorders and other video sources is to convert these to RGB signals that can be applied to the existing monitor or TV set via its SCART input. Next month's second instalment of this article discusses a circuit to accomplish this. In addition, the circuit provides a colour transition improvement (CTI) function, and is capable of converting CVBS to RGB.

From composite to RGB

Although most standard video recorders have a SCART socket, this rarely supplies RGB signals. Likewise, most set-top TV tuners and indoor units for satellite TV reception supply a CVBS (composite video) signal only. A problem arises when this equipment is to be connected to a high-resolution colour monitor with analogue RGB inputs, or a TV set with a SCART (Euro-AV) input. In both cases, the converter to be described next month can link this equipment and ensure optimum picture quality. ❑

References:

1. Chrominance-locked clock generator. *Elektor Electronics* July/August 1988.
2. Video line selector. *Elektor Electronics* April 1990.
3. *Video Handbook* (second edition). by R. van Wezel. Published by Heineman Newnes, ISBN 0 434 92189 0.

DIGITAL CAR ENGINE LOCK WITH ALARM

The circuit described here is a car theft deterrent that locks the starting motor until a pre-programmed code is recognized.

P.U. Mahesh

The code fed into the memory of the car engine lock is retained in a memory until it is intentionally cleared by the rightful owner of the car.

The operation of the circuit is relatively simple. Bistable IC1A-IC1B forms a debounce circuit for the clock pulses generated by S1 while the code is keyed in. The preset code is latched in memory IC2 and the code entered is decoded by IC3. If the preset code matches the code entered, and the ignition key is switched on, thyristor Th1 is provided with gate current, and fires so that the starting motor is powered. When no code or a wrong code is keyed in with the ignition switch on, Th2 fires and actuates the horn.

The operation of the circuit may be illustrated by assuming that code 0101 (example) is to be entered. The sequence in which the switches are pressed is as follows:

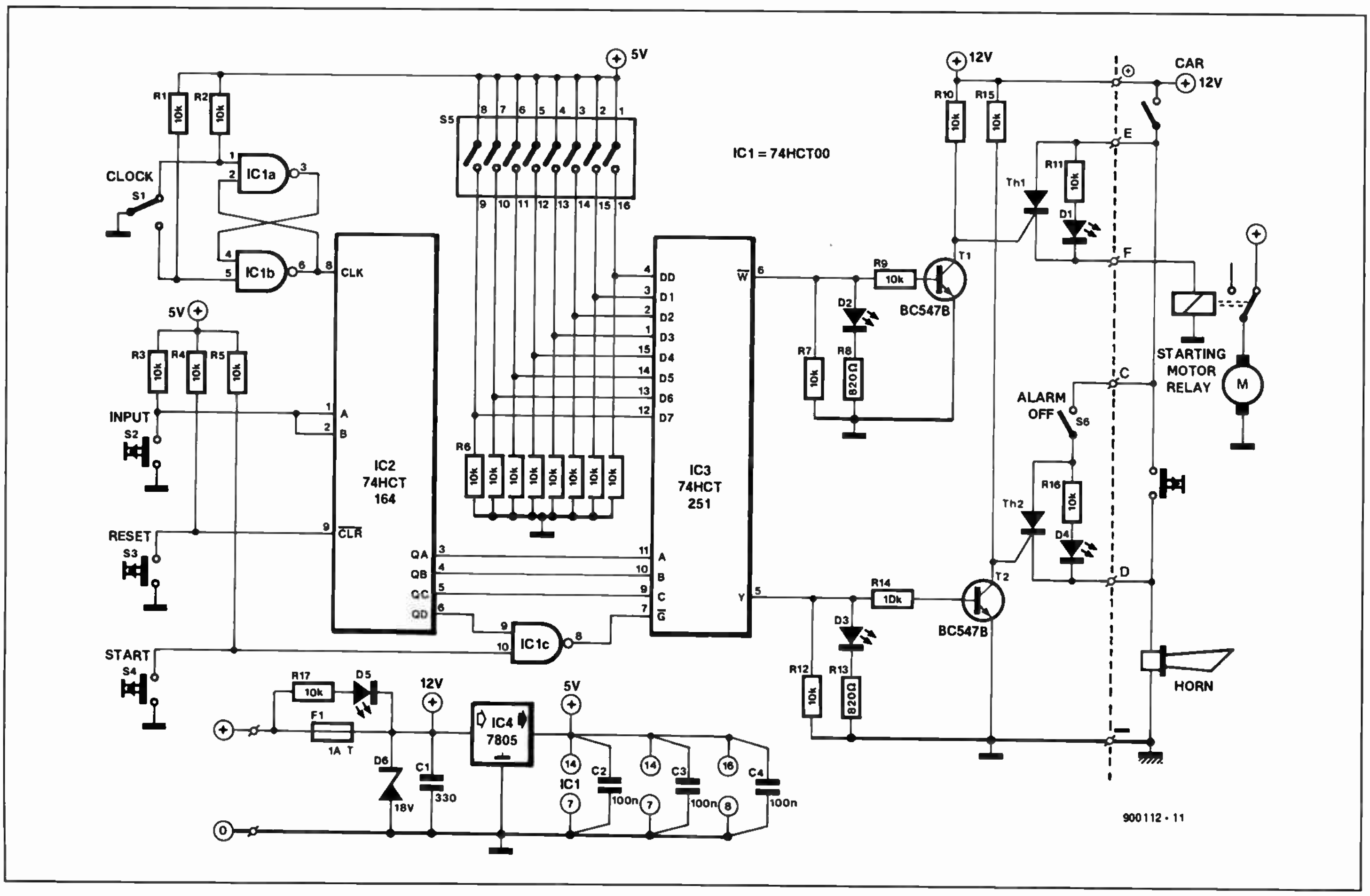
(S1) → (S1) → (S2, S1) → (S1) → (S4 , ignition)→ start

Here, (S2, S1) means that S2 is pressed, S1 is pressed, S1 is released, and S2 is released in that order. Note that pull-up resistor R3 ensures that a '1' is loaded when only S1 is actuated. The least-significant bit (LSB), which is keyed in first, is not used in IC4, so that the data is actually 010. Assuming that dataline D4 of IC3 is logic high because the associated switch in S5 is closed, the preset code matches the code entered. When the START switch, S4, is pressed while output QD of IC2 is high, multiplexer IC3 is enabled via its $\overline{G}$ input by a low level supplied by NAND gate IC1C. Since the code is right, the Y output (pin 5) goes high, and the $\overline{W}$ output goes low. A green LED, D3, lights to indicate that the correct code has been entered. Transistor T1 conducts and keeps the gate of Th2 low. At the same time, the low level at $\overline{W}$ of the multiplexer turns off T1 so that Th1 is fired via R10.

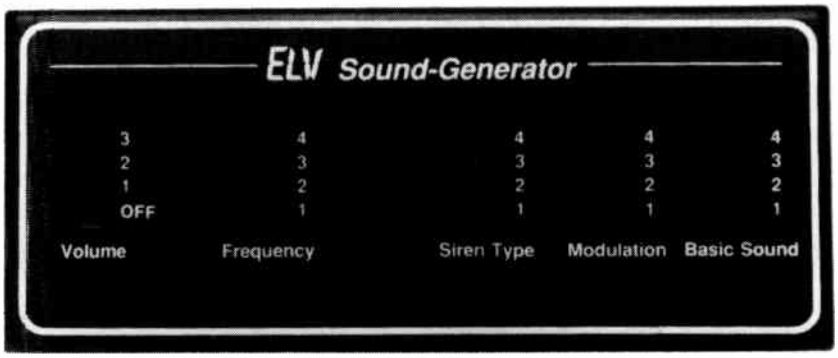
When the wrong code is keyed in, outputs $\overline{W}$ and Y of the multiplexer are high and low respectively. Upon turning the ignition key, Th2 is fired, and the horn sounds to alert passers-by and the owner of the car that someone is attempting to steal the vehicle.

Upon leaving the car, the owner must actuate the lock and the alarm by pressing S3. A standard 5 V regulator is incorporated into the circuit. LED D5 lights when the associated fuse, F1, blows as a result of a short circuit.

The complete circuit is easily built on a piece of veroboard. It is recommended to check the operation of the digital circuitry before connecting the transistors and the thyristors. ■



SOUND GENERATOR



The sound generator described here, designed and marketed as a kit by ELV, is capable of producing up 256 different siren-like sounds, including the popular Kojak-, FBI-, and Hawaii-Five-0 types. Compact, easy-to-build and suitable for use in conjunction with alarm systems in and on premises as well as on vehicles, the unit is complete with an on-board 20-watt amplifier.

The type of sound is selected with four slide switches on the front panel of the sound generator. Since each slide switch has four positions, a total number of 256 (4×4×4×4) different sounds are available. An output stage is included in the circuit to provide a solid 20 watts of audio power at a supply voltage of 12 V to 15 V. The slide switch at the extreme left on the front panel functions as a three-level volume control and as an on/off control.

Circuit description

Circuit IC1, a Type NE556, contains two multivibrators. One of these, IC1b, generates the basic siren sound. Switch S4 allows four different basic sounds to be generated by selecting one of four timing capacitors C7–C10. The output of IC1b, pin 9, drives the power output transistor, T1, direct via resistor R14. Depending on the position of volume switch S5, the loudspeaker is either disconnected ('off'), connected direct to the collector of T1 (volume level 3), or connected via series resistors R15 or R16 (volume levels 2 and 1).

Evidently, a single oscillator does not make a siren, let alone one capable of producing up to 256 different sounds. Circuit IC1b, is, therefore, frequency-modulated by applying a signal to its control voltage input, pin 11. This modulation signal is supplied by a second oscillator, formed by the parts to the left in the circuit diagram.

The second multivibrator in the circuit, IC1a, operates at a much lower frequency than IC1b. The oscillation frequency is determined by one of four capacitors C1–C4 connected to IC1a via the 'frequency' switch, S1. The other frequency-determining parts are R1 and R2, which set the charge and discharge periods respectively.

When S2 is set to the position shown in the circuit diagram, R3 is connected in parallel with R2, so that the input of buffer opamp IC2 receives a sawtooth signal. In the other extreme position, i.e., when S2 is set to the top position, R3 is not connected so that a triangular waveform is produced. The two centre positions of the switch produce a rectangular waveform and a combined rectangular/logarithmic waveform (as shown inset in Fig. 1). The latter is obtained with the aid of components C6, R7 and R3.

Opamp IC2 forms a buffer between the modulation waveform generator, IC1a, and the tone generator, IC1b. The level of the modulation signal fed to IC1b is determined by the position of switch S3, which connects one of four series resistors R8–R11 between the output of IC2a and pin 11 of IC1b. Switch S3 thus determines the modulation intensity.

Summarizing the above, the functions of the slide switches in the circuit are as follows (front panel marks in brackets):

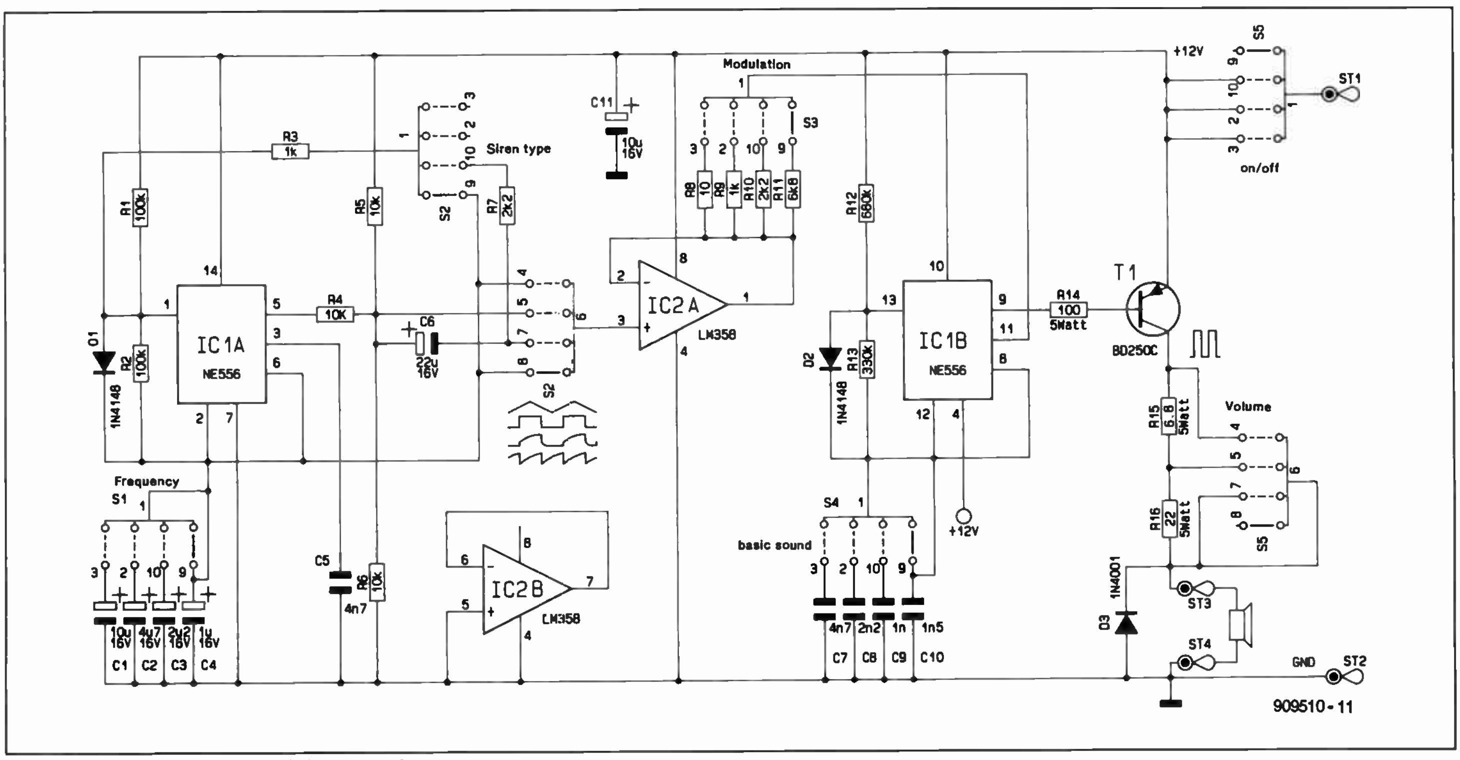


Fig. 1. Circuit diagram of the sound generator.

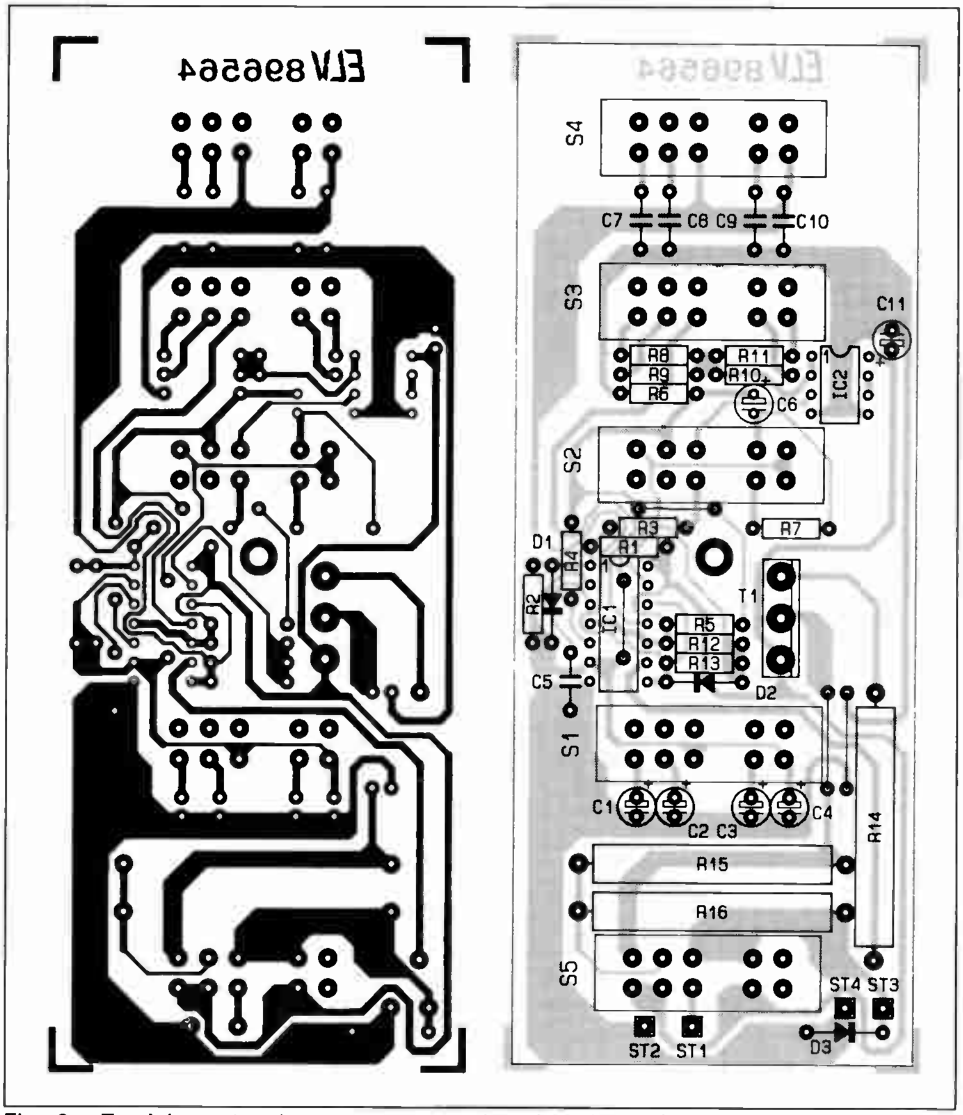


Fig. 2. **Track lay-out and component mounting plan of the PCB for the sound generator.**

COMPONENTS LIST

content of kit supplied by ELV France

Resistors:		
1	6.8Ω 5W	R15
1	10Ω	R8
1	22Ω 5W	R16
1	100Ω 5W	R14
2	1k	R3;R9
2	2k2	R7;R10
1	6k8	R11
1	9k1	R4
3	10k	R4;R5;R6
2	100k	R1;R2
1	330k	R13
1	680k	R12
Capacitors:		
1	1nF	C9
1	1n5	C10
1	2n2	C8
2	4n7	C5;C7
1	22nF	C11
1	1µF 16V	C4
1	2µ2 16V	C3
1	4µ7 16V	C2
1	10µF 16V	C1
1	22µF 16V	C6
Semiconductors:		
1	NE556	IC1
1	TLC271	IC2
1	BD250C	T1
1	1N4001	D3
2	1N4148	D1;D2
Miscellaneous:		
5	2-pole 4-way slide switch	S1–S5
4	solder pin	
1	printed-circuit board	
1	enclosure	

- S1 (frequency): modulation frequency
- S2 (siren type): modulation waveform
- S3 (modulation): modulation intensity
- S4 (basic sound): fundamental siren frequency
- S5 (volume): sound level and on/off control

The four switches S1–S4 allow 4^4=256 different sounds to be generated at three volume levels.

For the highest possible sound level (particularly in alarm systems), it is recommended to use a pressure-chamber type loudspeaker with a sufficiently high power rating (≥20 W). For other applications, standard loudspeakers may be used with good results. The minimum loudspeaker impedance is 4 Ω.

Construction

The sound generator is a relatively simple circuit which should not present difficulties in assembling. Moreover, the unit is supplied in kit form, which obviates problems with obtaining certain components.

Start the construction by fitting and soldering the low-profile parts, followed by the higher parts, on the single-sided printed circuit board shown in Fig. 2. The overlay printed on the component side of the board indicates the position of the parts mentioned in the parts list.

To assist in their cooling, the 5-W power resistors are mounted at a small distance above the printed-circuit board.

The use of a relatively flat enclosure makes it necessary to bend the power transistor, T1, towards the PCB surface as shown in the photograph of the completed board. By virtue of its low internal resistance, and the fact that it is driven at a fairly high level, T1 dissipates relatively little heat, even at full output power. Consequently, the transistor does not require a heat-sink.

After a careful visual check of the completed board, this may be fitted into the enclosure supplied with the kit. Connect the supply voltage to PCB terminals ST1 (+12 V to +15 V) and ST2 (ground). Connect the loudspeaker to terminals ST3 and ST4. Drill holes in the enclosure to pass the supply wires and the loudspeaker wires. Make knots in the wires at the inside of the enclosure to provide strain reliefs. Finally, fit the top half of the enclosure and secure it with the screws supplied.

Practical use

When a 4-Ω loudspeaker is used, the unit draws a peak current of up to 4 A. When used in a switched circuit, e.g., as a horn, the sound generator may be powered via a push-button or a relay with a suitable contact current rating. Use as a horn is possible because the siren starts to sound the moment is it powered. It should be noted, however, that in many countries the use of a siren as a sound actuator device in or on vehicles, and in some cases in or on premises as well, is restricted to emergency services. The use of a siren in general may also be subject to special licenses, rules or regulations as regards on-time, sound type and sound level. ■

A complete kit of parts for the sound generator is available from the designers' exclusive worldwide distributors:

ELV France
B.P. 40
F-57480 Sierck-les-Bains
FRANCE
Telephone: +33 82837213
Fax: +33 82838180

MEASUREMENT TECHNIQUES (1)

By F.P. Zantis

There are many electronics/electrical practitioners who have implicit faith in their measuring instruments and believe anything displayed on the these, particularly if the display is digital. There is more to measuring electrical quantities, however, such as knowledge of the instrument being used and the purpose of the measurement. With those in mind, it is possible to select the most suitable instrument for a given measurement. Judging and interpreting the measurements are the next important steps. Basically, each and every measuring instrument produces errors. The magnitude of such errors and how to arrive at meaningful results is the object of this new series of articles.

In all measurements a number of errors may occur that can, however, be avoided or at least minimized by a proper knowledge of them. Such errors fall under four headings.

1. **Systematic errors.**

Systematic errors are in direct proportion to the quality of the instrument and are the result of inaccuracies in the design, construction, final assembly and calibration of the instrument. The more care the manufacturer has taken in these areas, the smaller the errors will be (and the higher the price of the instrument).

2. **Environmental errors.**

Environmental errors are caused by geographical position, temperature, humidity, and electric and magnetic fields. One of the most common is the electro-magnetic field around a mains transformer (see Fig. 1),

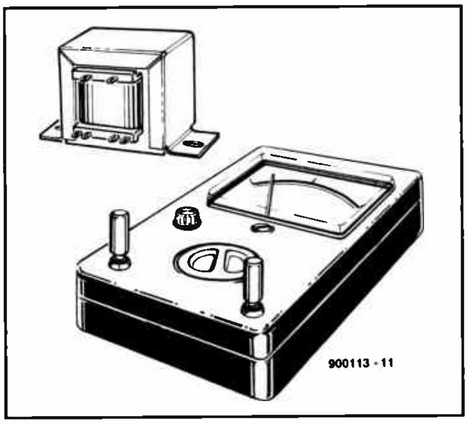


Fig. 1.

which may make a measurement at best meaningless and at worst impossible. Even the earth's magnetic field can distort a measurement: the electron beam in an oscilloscope, for instance, is deflected by it. The degree of deflection is determined by the position of the oscilloscope in the laboratory or workshop and is evidenced by the horizontal trace not being exactly parallel with the horizontal graticule lines. Screening the oscilloscope does not help and a good-quality oscilloscope is therefore provided with a trace rotation control.

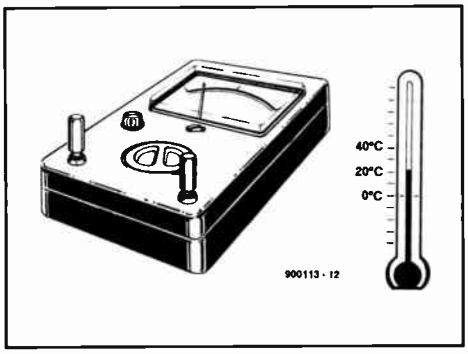


Fig. 2.

The effect of ambient temperature should also not be underestimated (see Fig. 2.). Test equipment is normally calibrated at +20 °C. When the ambient temperature differs from this by more than ±10 °C, temperature-dependent errors are likely to occur.

With analogue instruments, it may also be important whether it is used upright or lying down (see Fig. 3).

3. **Human errors.**

Human errors occur when the user is not quite clear about the purpose of the measurement or does not know enough about the test equipment he is using, or both. Typical are not knowing the input resistance when voltage measurements are carried out (error caused by incorrect loading of the voltage source) and incorrect setting of the instrument. Carelessness in taking or recording readings is another cause of errors.

4. **Applicational errors.**

Applicational errors occur when the equipment used is not suitable for the measurement to be carried out. For instance, the use of an instrument whose internal resistance is comparable to the input resistance of the circuit under test.

Also, reading errors are very common, particularly when analogue instruments are used. Such instruments should always be read with the eye at right angles to the pointer to avoid parallax errors. Some instruments have a mirror directly below the meter scale to ensure that the pointer is read correctly (it should not be seen in the mirror).

Errors in practice

Most measuring instruments are still based on analogue techniques and have analogue displays. This is so for very good reasons as

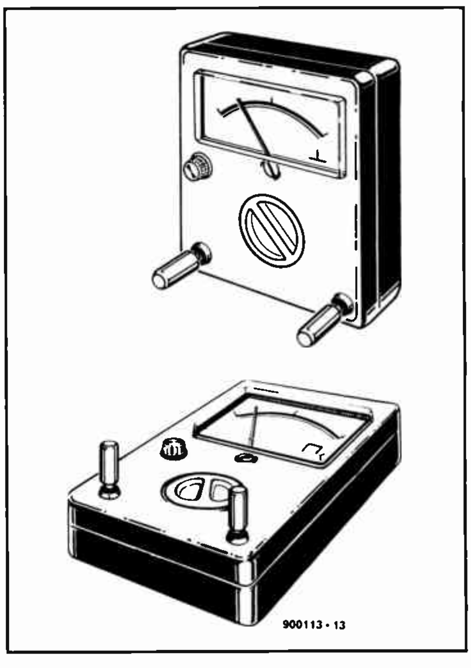


Fig. 3.

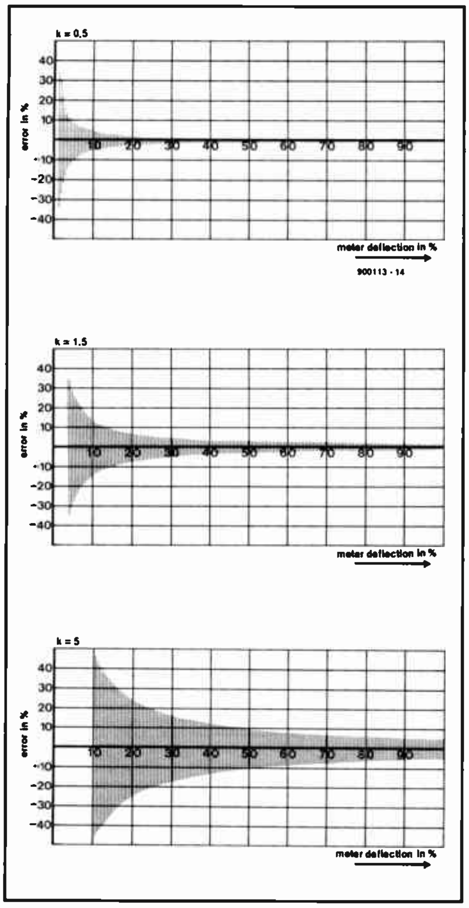


Fig. 4.

will be seen later in the series.

The accuracy of analogue instruments is normally given as a percentage (absolute error) of full-scale deflection (FSD). Note that the a.c. and d.c. ranges of multimeters usually have different accuracies. This means, for instance, that the pointer of an instrument with 1.5% accuracy and set to its 10-V range may at any given reading be off the true value by as much as 150 mV.

If, therefore, a voltage is measured as 1 V, the true value may lie between 0.85 V and 1.15 V. That represents a maximum relative error of 15%! If, in the same range, a voltage of 9 V is measured, its true value may lie between 8.85 V and 9.15 V. The maximum relative error is then 0.15/ 9×100% = 1.67%.

From these two examples it is seen that at the lower end of the range measurements are guesswork, in the middle of the range they are merely unreliable, and only at the upper end of the range does the relative error compare with the stated accuracy. This means that for greater accuracy the proper range (that is, the one where the greatest deflection is obtained) should be selected. Figure 4 shows the relative error in per cent for three instruments that have absolute errors of 0.5%, 1.5% and 5%, respectively, of full-scale deflection. In all cases, note the rapid rise of the relative error for small pointer deflections.

Matching measurand and measurement range

The importance of selecting the correct measurement range will be shown by the following example. Assume that an alternating voltage of about 20 V is to be measured and that two measuring instruments are available. Of these, one has an absolute accuracy of 0.5% and FSD ranges of 1 V, 10 V and 100 V, while the other has an absolute accuracy of 1% and FSD ranges of 3 V, 30 V and 300 V.

The first instrument must be set to the 100 V range and the relative error will be 0.5/20×100% = 2.5%. The second instrument is set to its 30 V range, where the relative error wil be 0.3/20×100% = 1.5%. It is thus seen that the less accurate instrument in this case gives the more accurate result! This is an important point to bear in mind when buying an instrument: the ranges available should be compatible with the expected measurands. When you are normally engaged in power electronics, you need different ranges from those required in radio/TV engineering.

Digital instruments

In digital instruments, the accuracy is normally related to displayed value. Furthermore, because of the analogue-to-digital conversion, the accuracy of the final digit must be added. Normally, the accuracy of the displayed value is ±1 digit, although there are instruments with an accuracy of ±2 digits and more.

The overall accuracy of a digital multimeter is typically

- for direct voltage measurements: 0.1% of measurand ±1 digit;
- for alternating voltage measurements: 2% of measurand ±7 digits;
- for direct current measurements: 0.35% of measurand ±1 digit;
- for alternating current measurements: 0.9% of measurand ±3 digits.

If the measurement of the alternating voltage of around 20 V in the earlier examples had been carried out by this digital instrument, the readings would have been

20×1.02 + 7×0.1 = 21.1 V

and

20×1.02 – 7×0.1 = 18.9 V

respectively. The relative error would have been

1.1/20×100% = 5.5%.

This shows that digital instruments are not necessarily more accurate than analogue ones as quite a few people believe. Note that the graphs in Fig. 4 apply to digital instruments as well.

Measurand and instrument

The measurand may have any one of a multitude of waveforms. Because of that, different instruments, or different methods of measuring, will give different results when such quantities are measured. Only when the basis of the measurement is known can the reading have any significance. For instance, a moving coil meter without rectifier will show the arithmetic mean of a quantity. However, the artihmetic mean of a pure alternating voltage (or current) is 0, so that there is no deflection of the pointer (see Fig. 5). If that meter is used to measure a com-

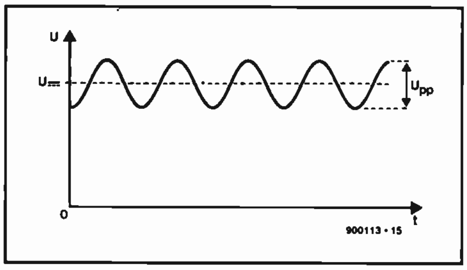


Fig. 5.

posite voltage (or current), it will only show the value of the d.c. component Care should be taken, however, because if the level of the superimposed alternating voltage (or current) is high, the meter may be overloaded in spite of that quantity not affecting the reading.

To measure an alternating voltage (or current) with a moving coil meter, a rectifier should be connected in series with the meter. This is the case, for instance, when an analogue multimeter is set to one of its a.c. ranges. The instrument will then display the mean d.c. value of the rectified quantity (see Fig. 6). The arithmetic mean of a rectified sinusoidal voltage, $U_a = 0.318U_{rms}$, and this

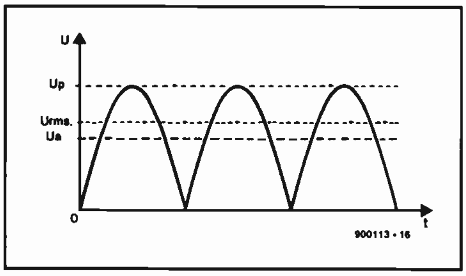


Fig. 6.

is the value indicated by the meter. However, the meter scale is calibrated in r.m.s. values. There is a fixed relation between the r.m.s. value of an alternating voltage or current and its rectified mean value, which depends on the waveform. The value given above is only true for a sinusoidal voltage or current. If the voltage or current is rectangular, the meter reading must be multiplied by 0.89 to find the r.m.s. value. When a triangular quantity is measured, the error is negligible: the true value is then only 0.36% larger than that displayed.

It might be expected that all this is true for digital instruments as well, but unfortunately that is not always so. Depending on the method of test, deviations may occur that can not really be foreseen. Some (relatively expensive) of these instruments can display the true r.m.s. value of a voltage or current. These instruments compute the r.m.s. value independently of the waveform.

Another factor that should be borne in mind is that, apart from the waveform, the frequency of a voltage or current plays a role. Most digital multimeters measure alternating quantities correctly only if the frequency is not higher than 400 Hz. Analogue instruments of about the same price perform rather better: right into the kHz range.

The r.m.s. measurement in many standard d.c. instruments is made possible by a variety of special circuits. These compute the r.m.s. value from the applied alternating voltage and convert it into a direct voltage, which is read in the usual way. These circuits are normally based on special ICs: a typical one is shown in Fig. 7 (see also Ref. 1 and Ref. 2).

When measurements are carried out with an oscilloscope, the instantaneous value of the measurand can be seen at a glance, so that at least the peak value of a voltage or current can be determined quickly. Determining the r.m.s. value is, however, possible only if the waveform is regular and the appropriate conversion factor is known. ■

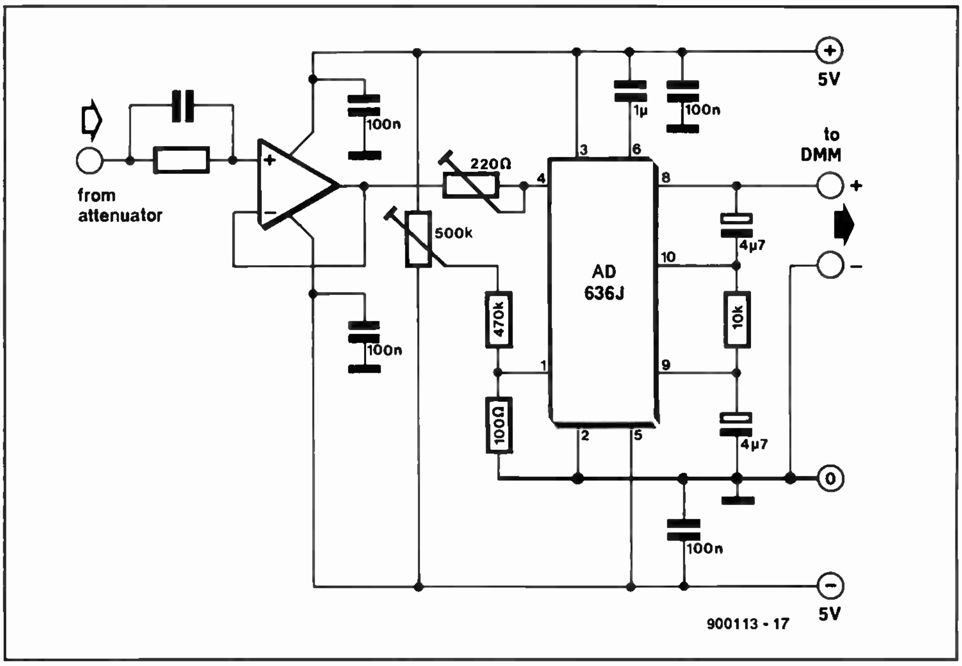


Fig. 7.

Further reading:

1. "RMS-to-DC converter", *Elektor Electronics*, July 1986, p. 38.

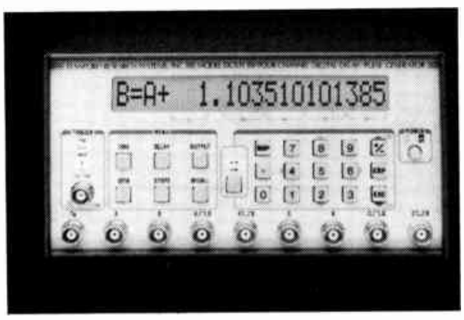
2. "True RMS DVM", *Elektor Electronics*, December 1989, p. 40.

3. *Electronic instruments and measurement techniques*" by F.F. Mazda, ISBN 0 521 26873 7, Cambridge University Press (*Elektor Electronics*, November 1987)

4. *Modern Electronic Test Equipment*, by Keith Brindley, ISBN 0 434 90567 4, Heinemann-Newnes (*Elektor Electronics*, November 1986).

FOUR-CHANNEL DIGITAL DELAY / PULSE GENERATOR

The Precision Pulse Generator and Digital Delay Generator recently announced by Fieldtech offers high accuracy and precision, a wide range and low jitter.



As a digital delay generator, the four outputs of the unit may be programmed for any interval between 0 s and 1000 s with 5 ps resolution. The standard timebase provides 25 ppm accuracy, but 1 ppm is available as an optional extra. The jitter of any output is less than 50 ps plus 1 part in 10^8 of the programmed delay. All outputs return to their pre-trigger levels about 800 ns after the longest delay.

As a precision pulse generator, the four time intervals define two pulses for applications that require precisely controlled pulse widths. The position and width of each pulse may be programmed from the front panel or via the GPIB. Front panel BNCs provide fast outputs at TTL, NIM, ECL or continuously adjustable levels. These pulses and their complements are available from separate front panel outputs. The outputs may be set to drive either 50-ohm or high-impedance loads.

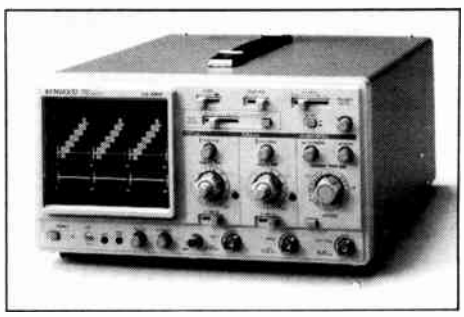
Fieldtech Heathrow Ltd, Huntavia House, 420 Bath Road, Longford, WEST DRAYTON UB7 0LL.

20 MHZ OSCILLOSCOPE WITH FULL BANDWIDTH AND HIGH SENSITIVITY

The Kenwood CS4025 oscilloscope is a low-cost 20 MHz model with a 80×100 mm (8×10 div.) screen. Its sensitvity is adjustable between 5 mV/div. and 5 V/div. over the full bandwidth; high-sensitive positions of 1 mV/div. and 2 mV/div. are available up to 5 MHz.

The sweep time can be varied from 0.5 µs/div to 0.5 s/div; a maximum sweep speed of 50 ns/div. can be achieved with the use of the ×10 magnifier.

Cross-talk is specified as at least–40 dB for a 1 kHz sine wave.



A useful vertical amplifier signal output (channel 1) is available, which gives an output of 50 mV/div. over a bandwidth of 100 Hz to 10 MHz. This enables, for instance, a frequency counter to be connected for accurate measurement of a waveform at low frequency.

Thurlby-Thandar Ltd, 2 Glebe Road, HUNTINGDON PE18 7DX, Telephone (0480) 412451.

TUNED UHF TV PREAMPLIFIER

based on a design by K. Kraus

For the reception of weak UHF-TV signals a good antenna is, of course, indispensable, but by itself it may not be enough. If that's the case, the preamplifier described in this article may be just what the doctor ordered.

IT is fairly straightforward to couple the antenna to a wideband amplifier, but that could give troubles if the weak signals to be received are close in frequency to strong signals. Even if the amplifier is a good-quality type, the most likely result is a fair dose of cross modulation and all that goes with it. This may be prevented by making the amplifier tunable over a relatively narrow range of frequencies. Even a strong transmitter in a channel adjacent to that of the weak signal can then be suppressed to a fair degree, but the present preamplifier can cope only to some degree if the wanted weak signal is surrounded by a number of strong signals.

The circuit

Basically, the circuit in Fig. 1 consists of two tuned circuits and a dual-gate MOSFET. The signal from the antenna is fed to the MOSFET via a tap on the input circuit, which consists of a 30×3.5 mm stripline. This arrangement ensures that the input impedance is 50–75 Ω. The input circuit is tuned with C1.

MOSFET T1 is arranged as a grounded-source amplifier in which C3 forms the ground connection for HF signals. The drain impedance is formed by the second tuned circuit, which also consists of a stripline, and a small trimming capacitor, C5. The gain of T1 is a maximum at the resonant frequency of the second tuned circuit. Capacitor C4 prevents the DC supply to the MOSFET being short-circuited by the tuned circuit. At the same time, inductor L1 prevents the HF signals being short-circuited by the power supply. The output of the units is taken from a tap on the second tuned circuit to obtain an output impedance of 50–75 Ω.

The DC operating point, determined by the voltage between gate 1 and the source of T1, is set by R2. Owing to the spread of parameters of the MOSFET, the operating point may differ from type to type, and it may therefore be necessary to alter the value of R2 to make certain that the source current is about 10 mA. In the prototype, a value of 150 Ω resulted in a source current of around 12 mA. The value of R2 will be somewhere between 100 Ω and 220 Ω.

Technical data

Frequency range	400–750 MHz*
	400–800 MHz†
Bandwidth at f_c = 500 MHz	492–513 MHz
Gain at 500 MHz	15 dB
Max. attenuation outside pass band	about 40 dB
Output noise (measured with $Z_{in} = Z_{out}$ = 50 Ω)	≤ –80 dBm

* C1, C5 = 1.5–5 pF
† C1, C5 = 0–5 pF

Network R1-C2 ensures that gate 2 is at ground potential for HF signals, so that these signals are processed via gate 1 only.

Power supply

There are various ways of providing power to the amplifier. If it is mounted near the antenna, the supply may be connected via the coaxial antenna feeder cable—see Fig. 2.

The direct voltage is applied to voltage regulator IC1 via inductor L2. The output of the regulator is fed to the cable via the circuit shown in Fig. 2.

If the supply is not via the coaxial cable, L2 may be omitted and an unregulated voltage applied to A.

In case a regulated voltage of 8 V is available, IC1 and C8 may be omitted. The wire bridge between B and C must then be replaced by one between A and B.

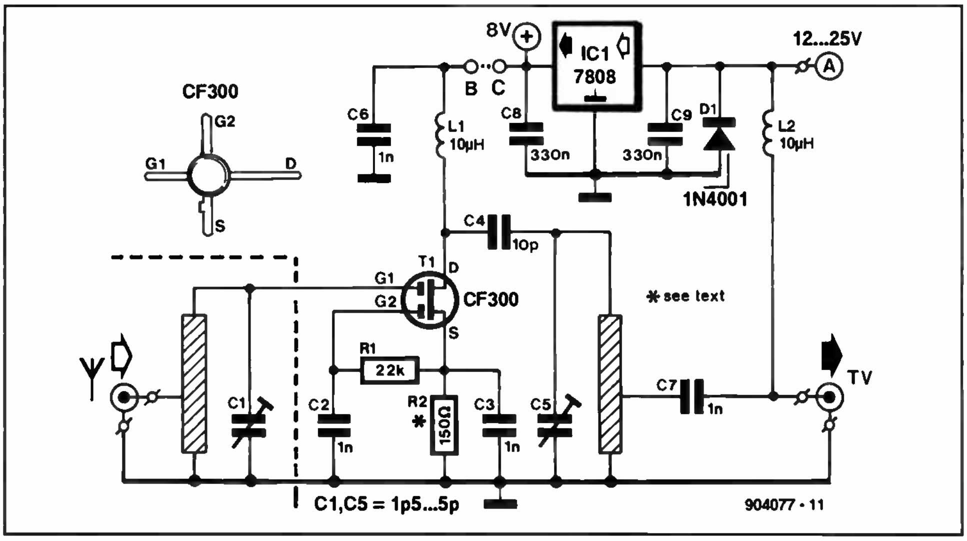


Fig. 1. Circuit diagram of the tuned UHF-TV preamplifier.

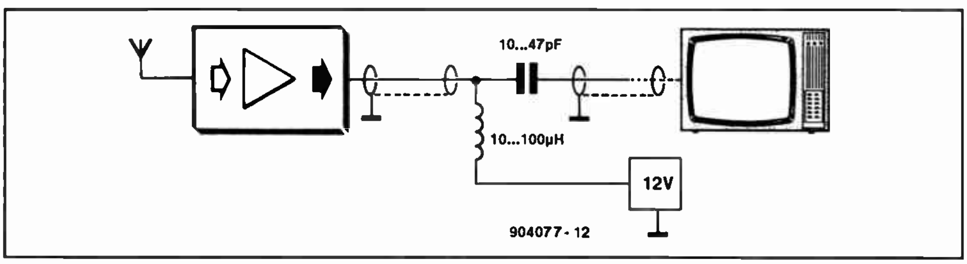
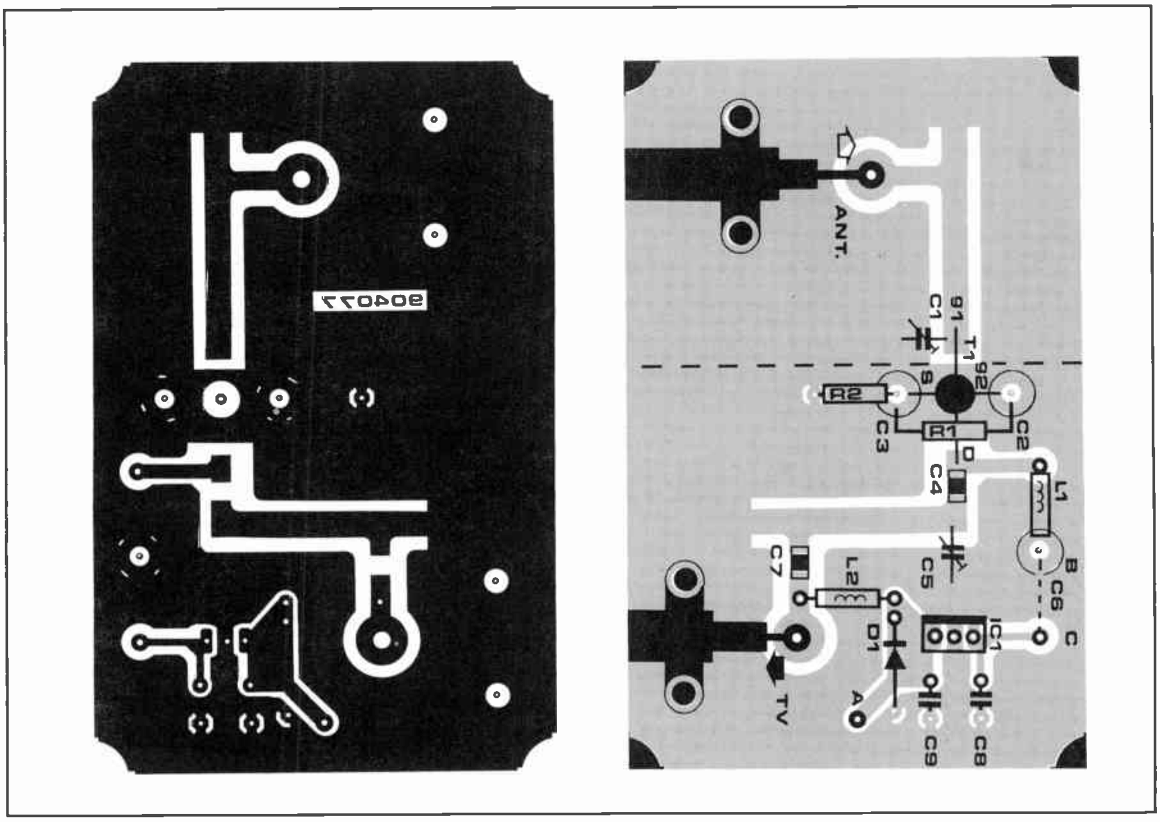


Fig. 2. If power is supplied to the preamplifier via the coaxial antenna feeder, the TV receiver and the power supply must be connected as shown.



PARTS LIST

Resistors:
R1 = 22 k
R2 = 150 Ω (see text)

Capacitors:
C1, C5 = 0–5 pF trimmer (Murata); 1.5–5 pF is permissible (see Technical Data)
C2, C3, C6 = 1 n disk
C4 = 10 p*
C7 = 1 n*
C8, C9 = 330 n

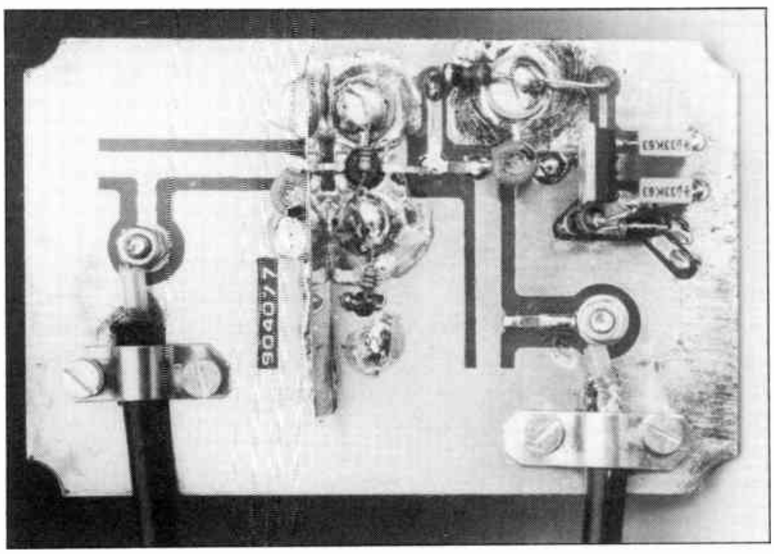
Inductors:
L1, L2 = 10 µH

Semiconductors:
D1 = 1N4001
T1 = CF300 (Telefunken)
IC1 = 7808

Miscellaneous:
PCB Type 904077

* surface mount type

Fig. 3. The printed circuit board for the preamplifier is (or must be) made of epoxy resin. When the board is made, rather than bought ready-made through our Readers' Services, great care should be taken in the dimensioning of the striplines.



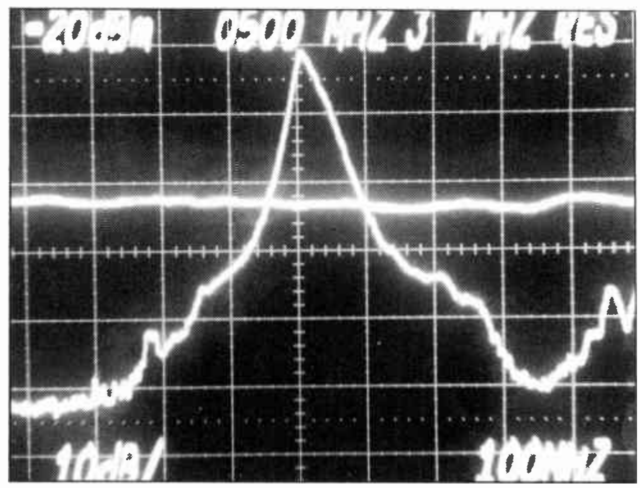


Fig. 4. Selectivity curve of the preamplifier. Horizontal scale 100 MHz/div; vertical scale 10 dB/div.; centre of graph at 500 MHz.

Construction and alignment

The preamplifier must be constructed on the PCB shown in Fig. 3. Note that the component side is also the track side. If you have the board made, make sure that the dimensions of the striplines are exactly right.

Mount disk capacitors C2, C3 and C6, followed by surface-mount capacitors C4 and C7. Next, install the MOSFET between C2 and C3. Solder gate 2 and the source of this transistor to the top terminal of C2 and C3 respectively. Solder gate 1 and the drain to the tracks underneath them.

Solder R1 between C2 and C3 above the MOSFET (although Fig. 3 shows it alongside the transistor).

Next, solder the remaining components in place. Connect R2, L1 and wire bridge B–C (or A–B: see under "Power supply") to the top terminal of the relevant disk capacitor. Mount capacitor C1 as far away as possible from T1, C2 and C3 to ensure that the screen—shown by the dashed line—fits neatly between them.

Then, solder a 45×20 mm screen, made of thin tinplate, in the position shown by the dashed line.

Finally, connect the coaxial cables to the relevant terminals and secure them in place with cleats as shown in the photograph of the completed prototype board.

The alignment is pretty straightforward: turn C1 and C5 until the wanted signal is at maximum level and any unwanted ones at minimum level; in other words, until the picture on your TV screen and the associated sound are at an optimum.

If the preamplifier is mounted near the antenna, it may still be aligned as discussed, since the input and output impedance it 'sees' at either end of the feeder cable are equal (at least, if everything is all right).

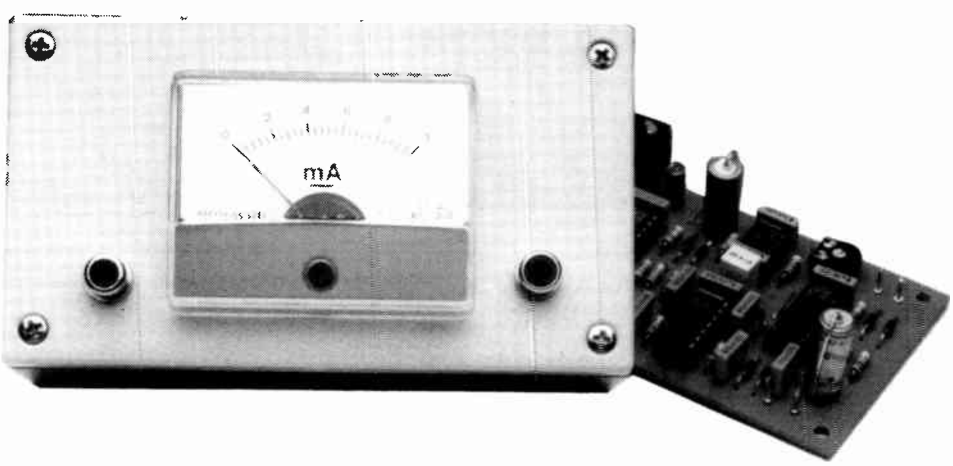
Finally, a word about the trimming capacitors. The parts list shows two possible types: the 1.5–5 pF is an inexpensive type, but its use makes tuning to the higher channels in the UHF-TV band impossible. If operation in that part of the band is required, there is no choice but to buy the more expensive 0–5 pF Murata type. ■

INTERMEDIATE PROJECT

A series of projects for the not-so-experienced constructor. Although each article will describe in detail the operation, use, construction and, where relevant, the underlying theory of the project, constructors will, none the less, require an elementary knowledge of electronic engineering. Each project in the series will be based on inexpensive and commonly available parts.

PHASE METER

Phase, phase difference and phase shift are commonly used terms in electronics theory. Where an oscilloscope or a vectorscope are not available, a special test instrument is required to measure phase shift so that it can be checked against the calculated value. Such an instrument is described here.



J. Bareford

THE term phase is used when describing two or more alternating voltages of which the frequency is equal, while their zero-crossings occur at different instants. An alternating voltage crosses the zero line when its instantaneous voltage is nought. An example of two alternating voltages of different phase is shown in the oscilloscope photograph in Fig. 1. The phase difference is about one sixth of the period in this case. The difference is, however, more commonly expressed as an angle because a sine-wave is periodical, so that its period, T, may be described as a circle, or 360°. This means that any part of the period may be expressed in degrees: a period of $0.25T$, for instance, equals 90°, $0.10T$ equals 36°, etc. The phase shift between the two sine-waves in Fig. 1 is, therefore, roughly 60°.

What causes phase shift?

Since the phase difference between two separate voltage sources is hardly ever constant because of small frequency deviations, it makes little sense to design a test instrument for this purpose. The meter described here, however, is intended to locate signal changes that occur in a circuit. Take, for instance, an inverting amplifier. The term 'inverting' implies that the input and output signals are 180° out of phase. In some cases, it is useful to be able to check this.

Transistors and opamps are not the only components that can cause phase shift. Capacitors and inductors introduce a phase shift of 90° between voltage and current. In the case of the capacitor, the voltage lags the current by 90° (see Fig. 2). Inductors have the opposite behaviour: the current lags the voltage by 90°.

Capacitors and inductors are typically used in filters. An example is shown in Fig. 3. In this circuit, one would expect the output voltage to be in phase with the current through the capacitor, i.e., there exists a phase shift of 90° between the input and the output. Unfortunately, this is not so because the phase shift depends on the frequency (the actual relation is an inverse one). The dependency is caused by the ratio of the capacitor's reactance to the resistor value. Since the reactance, x_c, is inversely related to the frequency, it is large with respect to the (fixed) resistor value at low frequencies. Hence, the phase difference is roughly 90° at low frequencies and 0° at high frequencies.

Phase difference: what it does

In many case, the phase relation between alternating voltages is of little or no importance. There are, however, situations in which the phase relation is all-important. Take the case of two alternating voltages that are to be added. If they are 180° out of phase, the summing operation would result in total cancellation of the two, i.e., an output of 0 V. The equalizing current that flows as a result of this operation is wasted.

In large power plants, phase differences

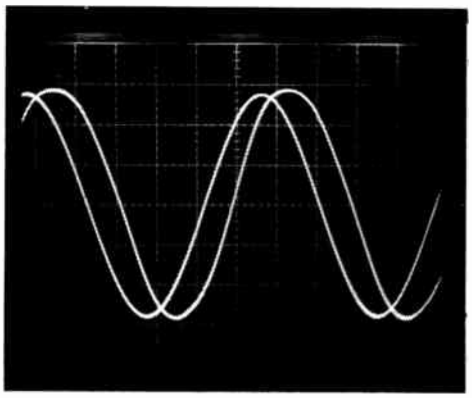


Fig. 1. Two sine-waves with a phase shift of about one sixth of the period, or about 60 degrees.

and the resulting equalizing currents must be kept to a minimum. Generators are, therefore, electrically coupled to maintain synchronicity. Furthermore, many power plants are linked by a special network, which often extends beyond country borders.

Fortunately, problems with phase differences in power plants need not bother us here, and can be safely left to power electricity engineers. The stereo equipment in your living room may, therefore, be a better example to illustrate the importance of steady phase relations. Since any amplifier in the system is bound to contain a filter of some kind, it may be expected that there is some phase shift between the input source, e.g., a CD player, and the output, the loudspeaker. Evidently, it is desirable for this shift to be roughly the same over the full frequency range of the signal. If this is not so, the sound is not perfect, particularly when the phase differences between the left and the right channel is large.

A final example of a circuit in which phase shift is all-important is the oscillator, which functions only when the amplified output signal is fed back to the input at a phase shift of 0°.

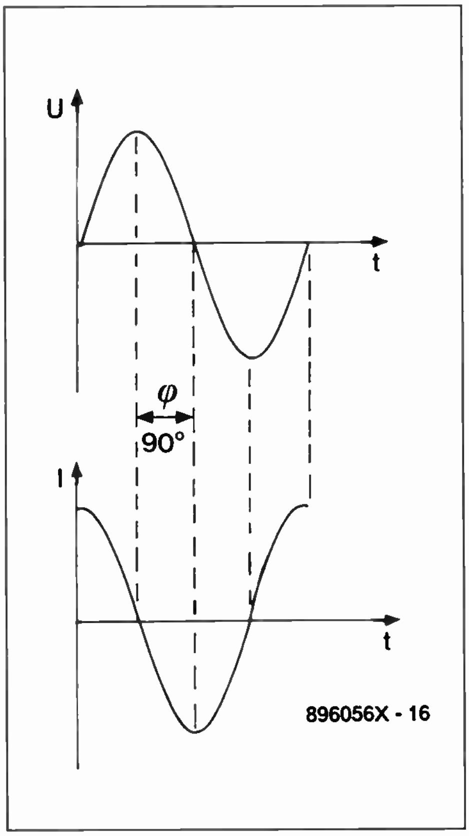


Fig. 2. The voltage on a capacitor lags the current by 90 degrees.

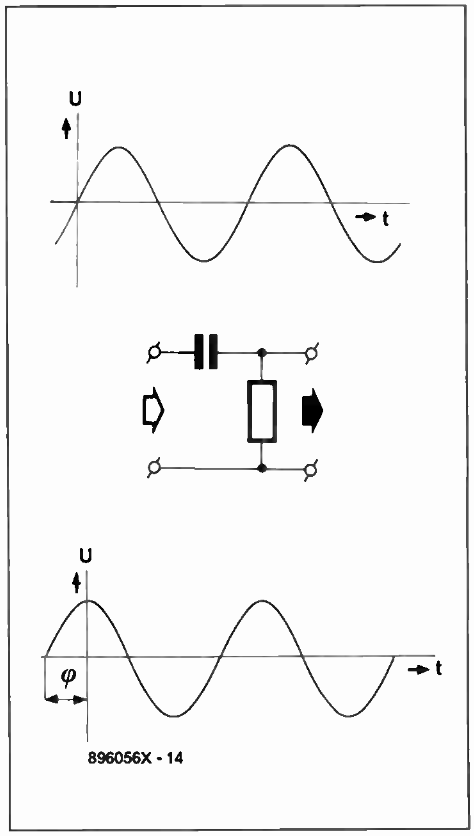


Fig. 3. Phase shift introduced by an *R-C* network

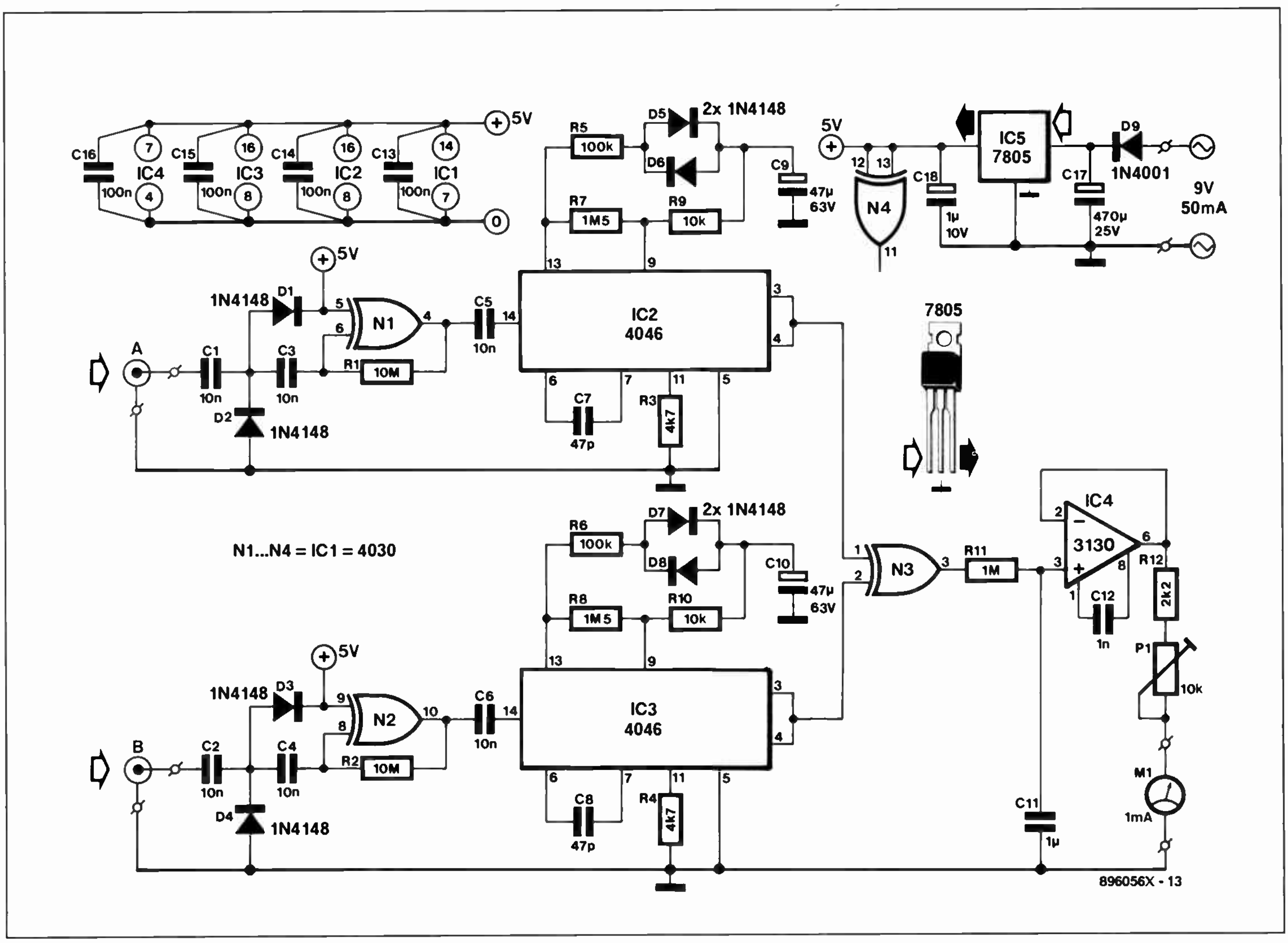


Fig. 4. Circuit diagram of the phase meter. The crucial part is XOR gate N3 which produces a voltage as a function of the phase difference that exists between the signals applied to the A and B inputs of the meter.

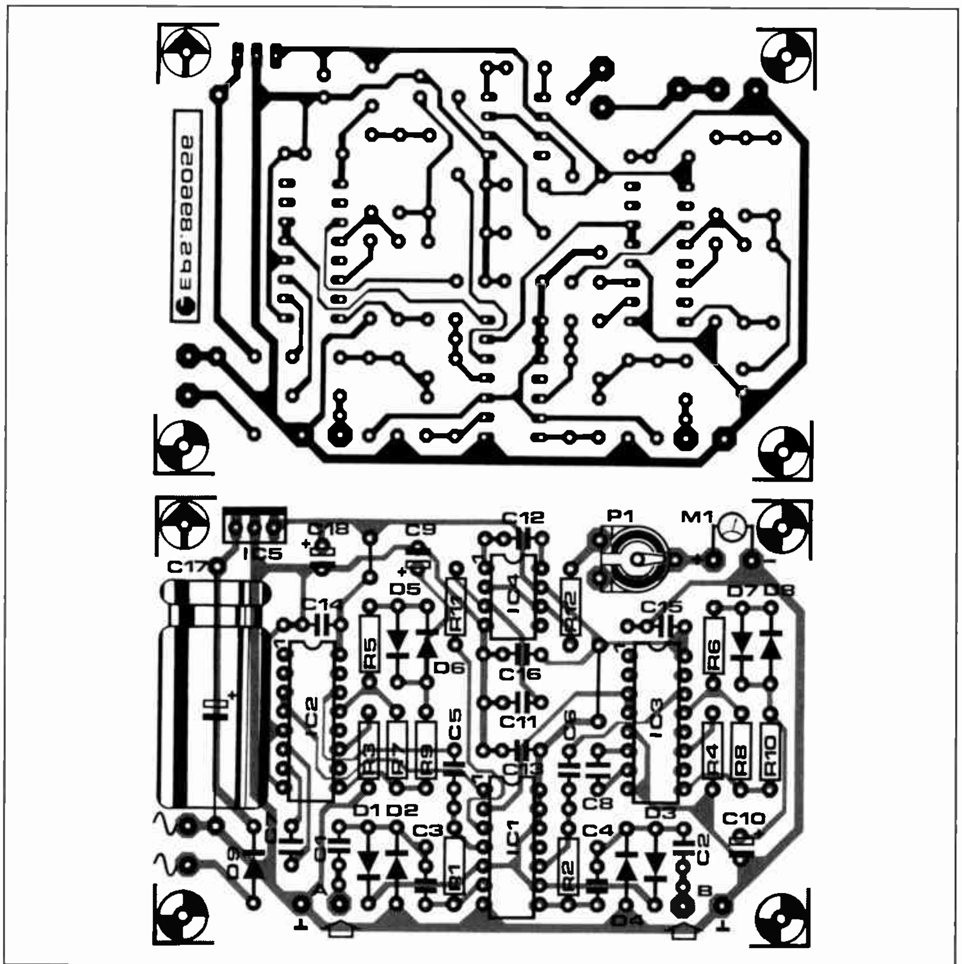


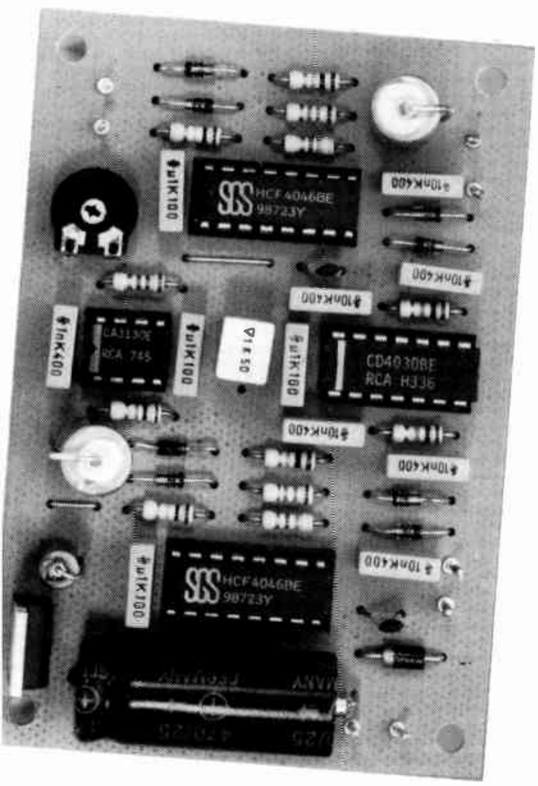
Fig. 5. **Track layout and component mounting plan of the PCB for the phase meter.**

COMPONENTS LIST

Resistors:		
2	10MΩ	R1;R2
2	4kΩ7	R3;R4
2	100kΩ	R5;R6
2	1MΩ5	R7;R8
2	10kΩ	R9;R10
1	1MΩ	R11
1	2kΩ2	R12
1	10kΩ preset H	P1
Capacitors:		
6	10nF	C1 – C6
2	47pF	C7;C8
2	47µF 63V radial	C9;C10
1	1µF	C11
1	1nF	C12
4	100nF	C13 – C16
1	470µF 25V	C17
1	1µF 10V radial	C18
Semiconductors:		
8	1N4148	D1 – D8
1	1N4001	D9
1	4030B	IC1
2	4046B	IC2;IC3
1	CA3130	IC4
1	7805	IC5
Miscellaneous:		
1	1 mA moving-coil meter	M1

The phase meter

As illustrated in Fig. 1, a two-channel oscilloscope is perfect for measuring phase differences. Unfortunately, such an instrument is not available to every one, and this is where a special phase meter comes in.



The circuit diagram of the two-channel phase meter is shown in Fig. 4. The capacitors at inputs A and B block any d.c. components in the signals applied to the circuit. Diodes D1 – D4 protect amplifiers N1 and N2 against negative or too high voltages. The amplified signals are subsequently fed to IC2 and IC3, two phase-locked loops (PLLs) that track the input frequencies and convert them to digital signals with a fixed duty factor of 0.5 and a swing of 5 Vpp. The low-pass filters of these PLLs are special types to enable the 4046s to keep track of fast frequency changes in the input signals. This important characteristic is provided primarilyby the diodes in in the network connectd to pins 13 and 9.

The actual phase comparator is formed by a single XOR (exclusive-OR) gate, N3. The output signal of the gate is averaged by integrator network R11-C11. The larger the phase difference between the two input signals, the larger the voltage supplied by buffer IC4. Moving-coil meter M1, resistor R12 and preset P1 together form a meter that indicates the phase shift. Since the voltage indicated by the meter lies between 0 V and the positive supply level, the circuit must be powered from a regulated supply. A three-terminal fixed voltage regulator Type 7805 is used here for this purpose. The minimum unregulated or alternating input voltage is 9 V.

Construction and setting up

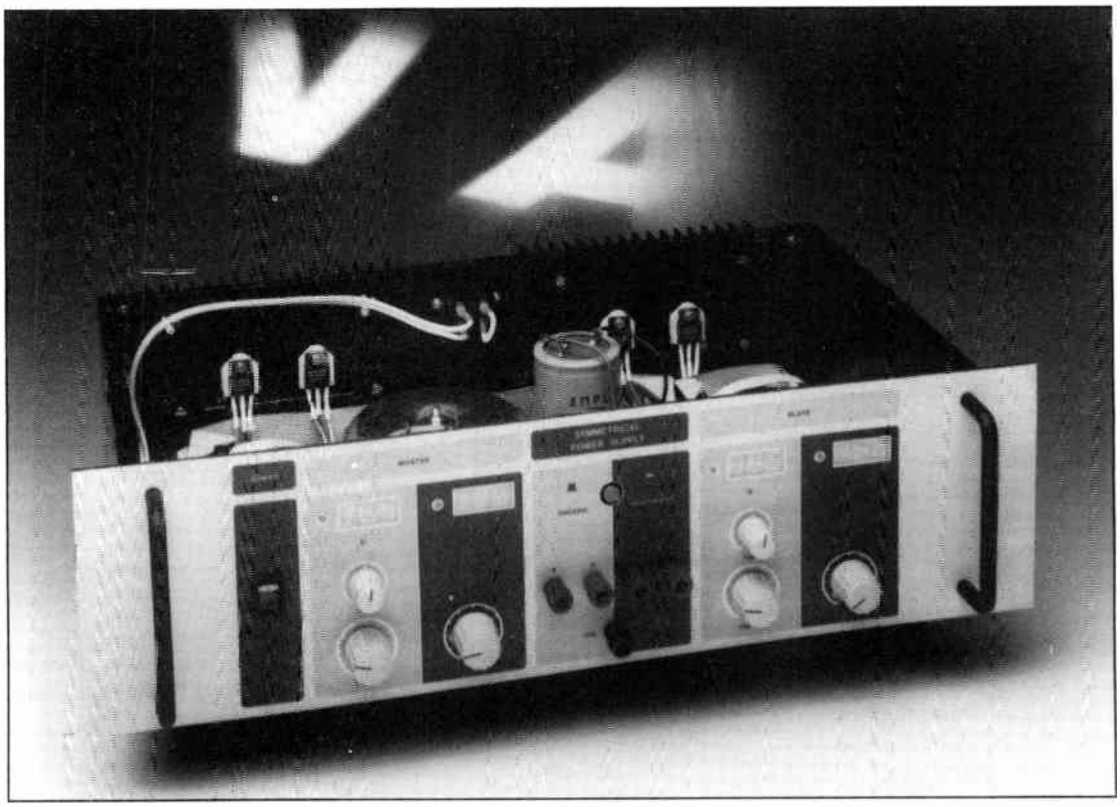
The circuit is best constructed on the printed-circuit board shown in Fig. 5. First, fit the two wire links so that they are not forgotten later. Pay attention to the orientation of the polarized components (diodes, transistors, ICs and electrolytic capacitors). Fit IC sockets for all ICs.

To align the meter, first remove IC2 and IC3 from their sockets. Connect pin 3 of the socket for IC2 to pin 5 (ground) via a 10 kΩ resistor. Next, connect pin 3 of the socket for IC3 to pin 16 (+5 V) via a 10 kΩ resistor. Switch on and check that pin 1 of N3 is at about 0 V, and pin 2 at about +5 V. The output of the XOR gate, pin 3, should be at +5 V. Since this represents the maximum phase shift that can be measured between the two channels, adjust preset P1 until a meter indication of 180° is obtained.

Remove the two 10-kΩ resistors and fit the two 4046s. At this stage, the circuit is ready for fitting into an enclosure.

Although the indication range of the meter is, in principle, limited to 180°, it is possible to measure greater phase shifts. A shift of 270°, for instance, is indicated as 90° because the meter can not detect which channel, A or B, was high first. ■

400-WATT LABORATORY POWER SUPPLY



PART 1: CIRCUIT DESCRIPTION

Here is an all-purpose d.c. power supply for symmetrical as well as asymmetrical use, and capable of supplying high output currents and voltages. An all-analogue design based on discrete parts only, this 400-watt PSU deserves a prominent place on your work bench.

G. Boddington

The problem with power supplies in an electronics laboratory or workshop is that their application range is often limited because of their specifications. Any one who has been involved in practical electronics will admit that finding a suitable power supply for a particular test is not at all easy, when none of the available ones (say ±15 V/2 A, 0–60 V/100 mA and 5 V/10 A types) seem to be fully geared to the job. Obviously, what is needed is a supply that combines the most frequently used voltage and current ratings, both symmetrical and asymmetrical, while offering a properly operating overload protection.

Although the user manual with many an inexpensive, ready-made power supply will confidently inform you that the power transistors are protected against overloads, this type of protection has an inherent disadvantage. True, the supply will happily supply the maximum output current at the maximum output voltage, but it will typically shut down the moment the voltage is reduced just one volt or so. The reason is clear: the overload protection is actuated because the extra dissipation, which is the product of the output current and the voltage difference across the series transistors, exceeds the cooling capacity of the heat-sink, or the power rating of the (expensive) series transistors.

The present power supply puts an end to

MAIN SPECIFICATIONS

- **Mode: Single**
  - one adjustable power supply with current and voltage controls.
  - Output: 0 – 40 V at 0 – 5 A
- **Mode: Independent**
  - two identical, electrically separated, power supplies.
  - Outputs: 2 × 0 – 4 0 V at 2 × 0 – 5 A
- **Mode: Tracking**
  - two identical, series connected, power supplies.
  - Outputs: ±0 – ± 40 V at 0 –5 A; 0 – 80 V at 0 – 5 A
  - Voltage and current of *slave* follow *master*.
- **Mode: Parallel**
  - two identical, parallel connected, power supplies.
  - Outputs: 0.6 – 39.4 V at 0 – 10 A
- **Maximum output voltage:** **40 V (at full load)** **48 V (no load)**
- **Maximum output current:** **5 A**
- **Ripple:** **10 mV (no load)** **50 mV (at full load)**
- **Voltage difference in tracking mode:** **50 mV**

these problems. It can be set up to supply either 2 × 40 V/2 × 0 – 5 A, ±0 – 40 V/0 – 5 A, or 0 – 80 V/0 – 5 A, and is capable of supplying the maximum output current at low voltage settings. Special ICs or microprocessors are not used: just straightforward analogue electronics based on readily available components. The result is a power supply with an excellent price/performance ratio.

Block diagram

The instrument consists of two identical, electrically isolated, power supplies, which may be connected in a number of ways to give different operating modes. The block diagram in Fig. 1 shows relatively many functional blocks, which together form three partly 'interwoven' regulating circuits. The first of these, the outer circuit, is a transformer preregulator that serves to keep the voltage drop across the series transistors (T4-T5) constant at about 10 V, so that the maximum dissipation remains smaller than 50 W (or 25 W per transistor). The other two regulation circuits are for the output voltage (*U*) and current (*I*). These circuits are almost identical, the only difference being that the current control obtains its control information from a series resistor, and the voltage control from a potential divider fitted across the output terminals. In contrast to the transformer preregulation, the *U* and *I* control circuits allow the range of the regulating action to be adjusted manually. Interestingly, the series transistors, T4 and T5, function in all three regulation circuits.

The block diagram shows a second power supply, which provides auxiliary ±12 V rails for use in the main circuit. The ground line of this symmetrical supply is connected to the positive output terminal of the main supply. This means that all references to '+12 V' and '–12 V' in the following text, and in the circuit diagram, are actually '+12 V and –12 V with reference to the positive output terminal'. The auxiliary power supply also functions as a voltage reference.

Finally, the block marked 'current limit' stands for a circuit that keeps the output current of each supply below 5 A. This circuit may be fitted with an optional temperature monitor to prevent overheating.

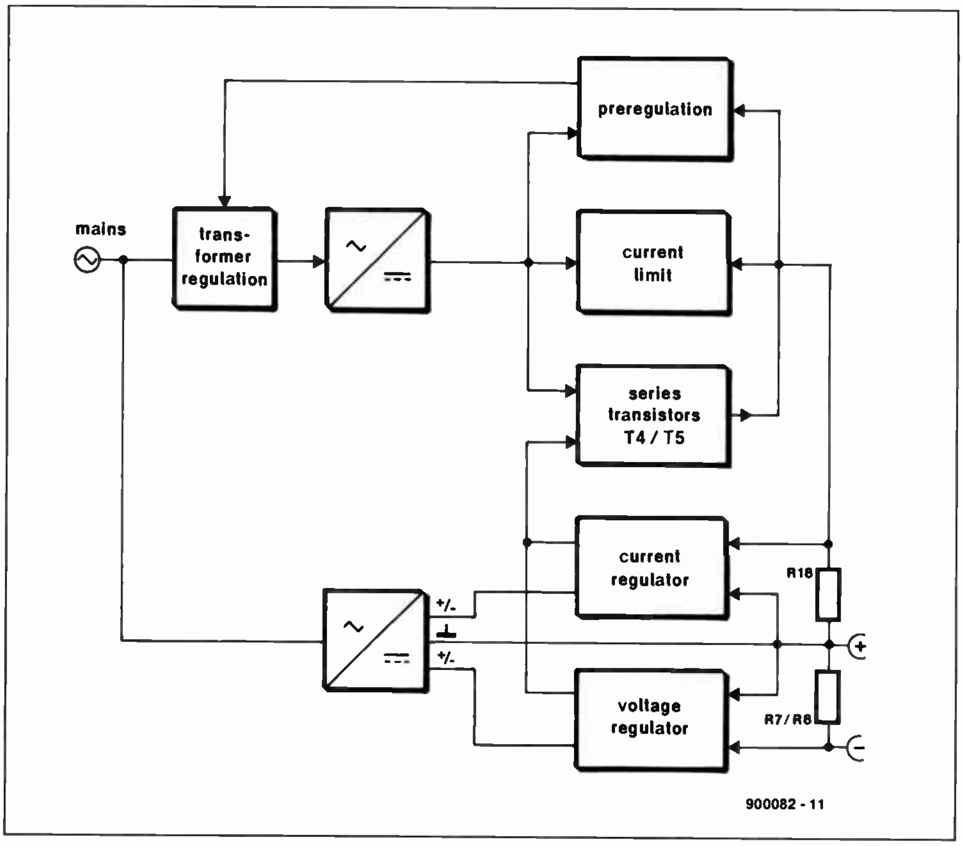


Fig. 1. Block diagram of the power supply. The design is based on three interactive control circuits: (1) transformer preregulation, (2) current control and (3) voltage control.

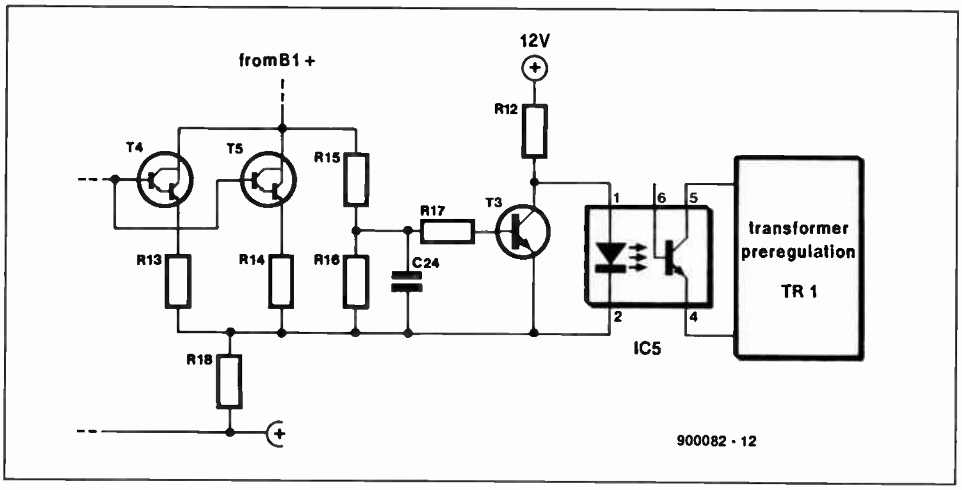


Fig. 2. Basic diagram of the circuit that controls the transformer preregulation.

The preregulation circuit

The basic operation of the preregulation circuit is best explained with reference to Fig. 2. The current flows from the positive connection of the bridge rectifier to the positive output terminal via two parallel-connected darlington transistors, T4 and T5, and resistors R13, R14 and R18. The regulation circuit tries to maintain a constant drop of 10 V across the series transistors and their emitter resistors. Transistor T3 is driven via potential divider R15-R16 and network C24-R17. The network introduces a small delay to eliminate the effect of noise spikes in the preregulation. The current through the LED in optocoupler IC5 is inversely proportional to the voltage across R15-R16.

The power fed to mains-connected ohmic loads is relatively simple to control. Usually, an adjustable *R-C* network connected across the mains terminals supplies the trigger voltage for a triac. The timing of the trigger (or firing-) pulse with respect to the start of the half-cycle is determined by the *R-C* delay. After being fired, the triac conducts until the mains voltage drops to a level below the minimum hold current. This happens close to the zero-crossing. The triac remains blocked until it receives another trigger pulse at a particular phase angle during the next half-cycle of the mains voltage. The current supplied to the load is inversely related to the phase angle, i.e., to the delay of the trigger pulse following the zero-crossing. This principle of phase-angle control works as long as voltage and current are in phase, i.e., as long as the load is a pure resistance.

Unfortunately, the mains transformer in the power supply forms an inductive rather than an ohmic load, so that the mains voltage and the load current are out of phase. Hence, a simple 'dimmer' with conventional triac control as described above will not do as a preregulation circuit. With an inductive load, it may happen that although the instantaneous voltage is high enough to fire the triac, there is no current to 'hold' the device. Therefore, the firing may take place only when the load current is sufficiently high to keep the triac conductive. However, since the load current in a power supply is variable, the phase shift between voltage and current is also variable. This means that the width of the trigger pulse rather than the position must be controlled. If the pulse were simply shifted, the result would be an asymmetrical output current with a high d.c. component, causing rapid saturation of the transformer winding. When the pulse is stretched, however, due care must be taken to prevent it from extending over the zero crossing of the mains voltage.

The circuit section in Fig. 3 stretches the first pulse by means of pulse sequence triggering, an approach which is particularly suited to applications with load currents that are prone to variation. The *R-C* network connected between the live and neutral lines of the mains serves to delay the trigger instant. The network consists of C1, potential divider R29-P1-R30 (branch 1), series resistor R31 and bridge rectifier D20–D23 (branch 2). The combination of the bridge rectifier and the optocoupler it powers simply forms an adjustable resistor for alternating voltages, so that both branches have the same function: making the trigger delay, ϕ, variable (see Fig. 4a). The basic delay is determined by P1.

When the power supply is switched on, C1 is charged. When the trip voltage of the diac is reached, both Di1 and Tri1 are fired. When a trigger current flows from C1 to Tri1, resistor R32 drops a voltage which is high enough to trigger a smaller triac, Tri2. The result is that the discharge time is no longer determined by the two branches, but by (R33+R29)C1. When C1 can no longer supply the hold current for Tri2—which happens fairly quickly because of the small resistors R33 and R29—the triac blocks and C1 charges again. This sequence is repeated until just before the zero-crossing, when the mains voltage can no longer charge C1 (see Fig. 4a). The waveform across the thyristor is shown as a dashed line in Fig. 4b. Figure 4c, finally, shows the waveform of the current shifted by an angle φ as a dashed line, and the waveform produced by the dimmer as a solid line. Asymmetry of the waveform occurs during the first half-cycle only. The triac conducts up to instant 'B', when the load current falls to zero.

The function of the remaining parts in this section of the circuit is quickly explained: the zener diodes limit the voltage across Tri2 to about 66 V whilst providing a stable reference voltage for the trigger circuit. Components D12, D13, D14, R27 and R28 ensure that C1 discharges during the zero-crossing. Inductor L1 serves to eliminate current surges and thus prevent HF interference. Network C2-R34 short-circuits spikes generated by the switching sequences, and so prevents erroneous triggering.

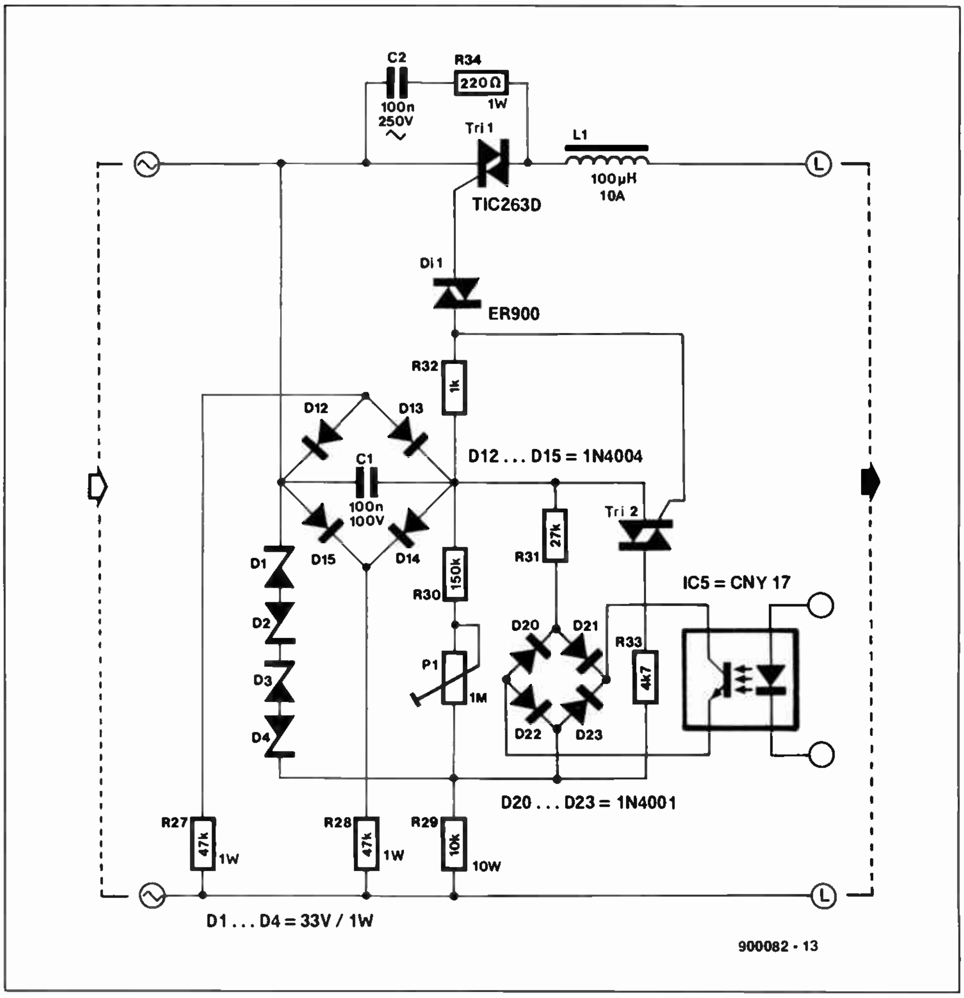
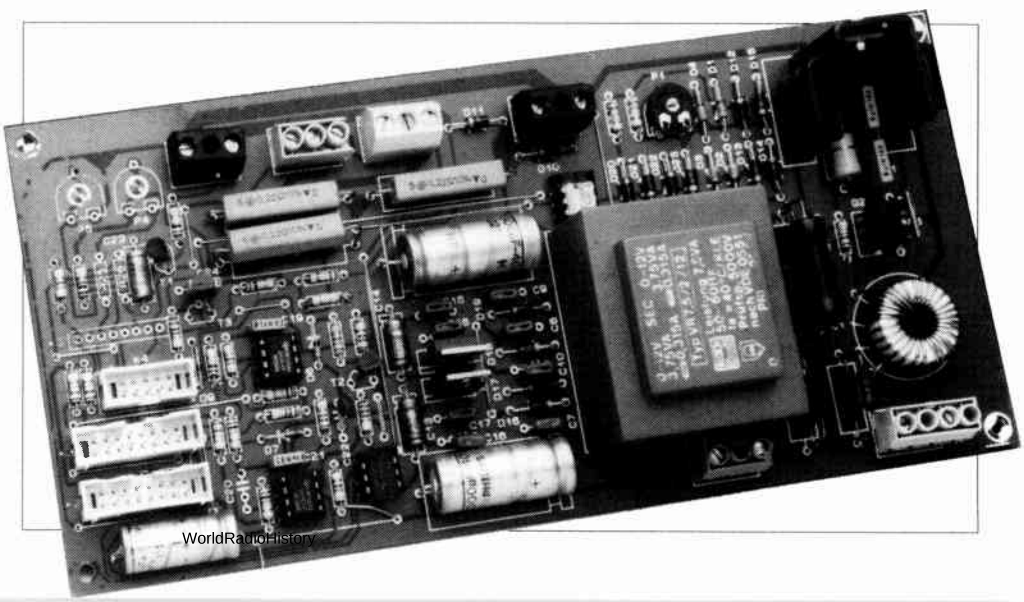


Fig. 3. Circuit diagram of the transformer dimmer. The trigger delay is controlled by the circuit in Fig. 2 via an optocoupler, IC5, and an adjustable bridge rectifier, D20–D23.

Voltage and current control

The operation of the voltage control circuit is illustrated in Fig. 5. Potential divider P3-R9 allows a reference voltage of 0 to −10 V to be set between ground (the positive output terminal) and −12 V. A second potential divider, R7-R8, at the output terminals supplies about 20% of the output voltage, i.e., about 0 to −9 V (with respect to the positive output terminal). The voltages supplied by the two potential dividers are compared by opamp IC4, which, with the aid of T4-T5, will attempt to keep the voltage difference between its two inputs as small as possible. When a higher output voltage is required, the wiper of potentiometer P3 is turned towards the −12 V potential. The voltage at the



non-inverting input of IC4 drops, so that the output voltage of the opamp rises. Conversely, when a lower output voltage is set either by the user turning P3, or by the actuation of the voltage limiting circuit, the inverting input is at a higher potential than the non-inverting input, so that the opamp output voltage drops.

The current control circuit (Fig. 6) operates in a similar manner. Like IC4, opamp IC3 will attempt to keep its output voltage at 0 V. The main difference with the voltage control circuit, however, is that the reference voltage for the opamp (applied to the non-inverting input) is permanently grounded via R1, while the current is measured as the drop (max. 1.1 V) across series resistor R18. Potential divider P2-R3 is arranged so that its junction carries a voltage between −1.1 V and +1.1 V with respect to the positive output terminal. When no current flows through R18, the positive side of P2 is at ground potential. When P2 is advanced to the 5-A position, i.e., to its full resistance of 2.2 kΩ, the inverting input of IC3 is at a voltage of −1.1 V. Consequently, the voltage at the opamp output rises.

When a current of 5 A flows, R18 supplies 1.1 V. When P2 is turned to the other extreme position (i.e., a resistance of 0 Ω), the voltage at the inverting input is higher than that at the non-inverting input, so that the opamp output voltage drops.

As shown by Figs. 5 and 6, and also by the complete circuit diagram in Fig. 7, the anodes of D8 and D24 share a common connection, R23, where the opamp outputs of the current and voltage control circuits are joined. This means that the opamp that supplies the lower output voltage determines the base voltage of the current booster, T4-T5. Resistor R23 serves to hold the bases of T4-T5 at about +11.5 V. Diodes D7 and D9 decouple the opamp outputs, preventing current from flowing between them. One of the series-connected LEDs lights when the voltage at the associated opamp output drops to a level below 11.5 V minus two diode voltages (D24-D7 or D8-D9). This happens when the relevant limiter (current or voltage) starts to operate.

During the switching-on sequence the

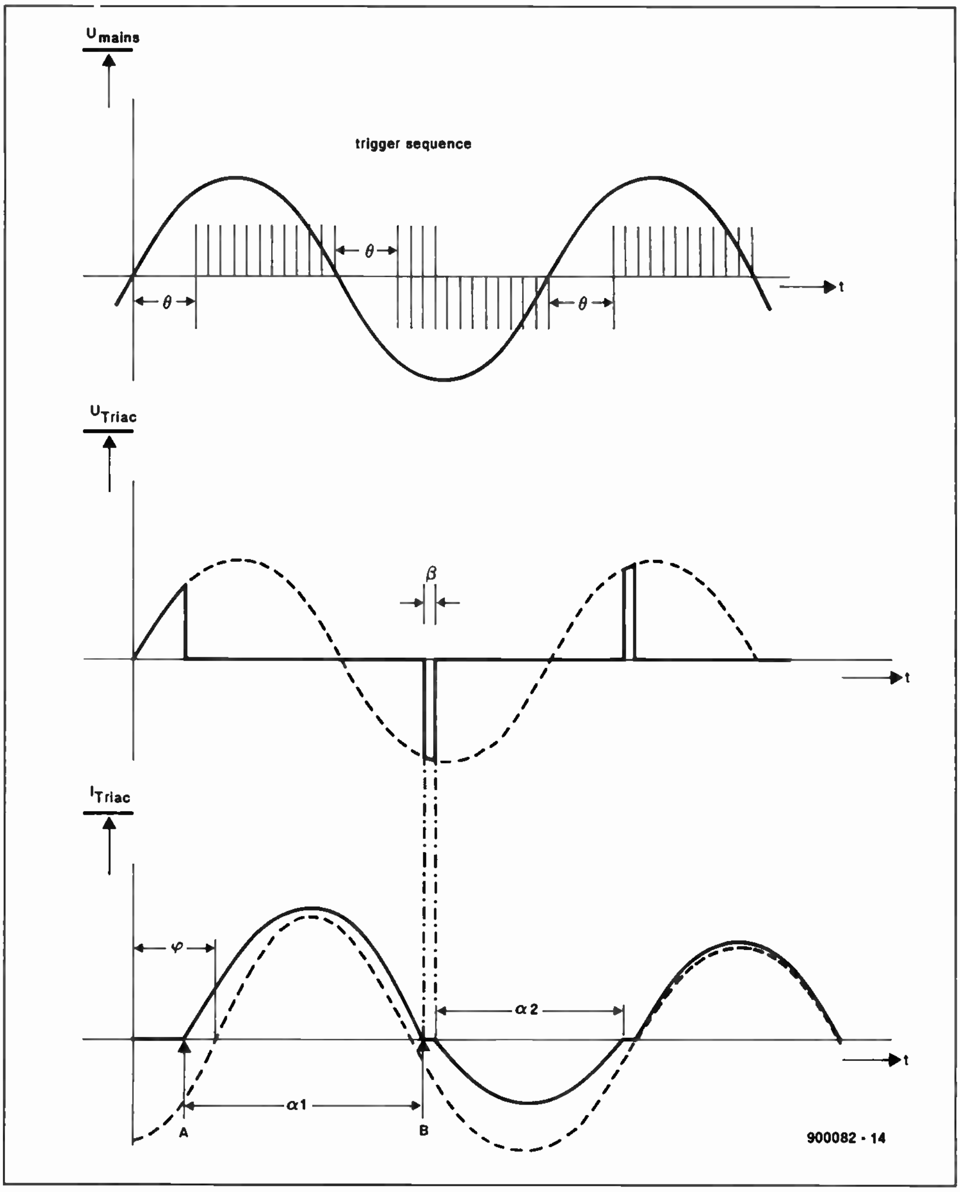


Fig. 4. Illustrating the basic operation of the dimmer for inductive loads, applied here for the purpose of transformer preregulation. Fig. 4a shows the position of the trigger pulses with respect to the mains voltage. The voltage across triac Tri1 as compared to the mains voltage (dashed line) is shown in Fig. 4b. Fig.4c, finnaly, shows the current shifted by an amount of φ, without (dashed line) and with (solid line) phase angle control.

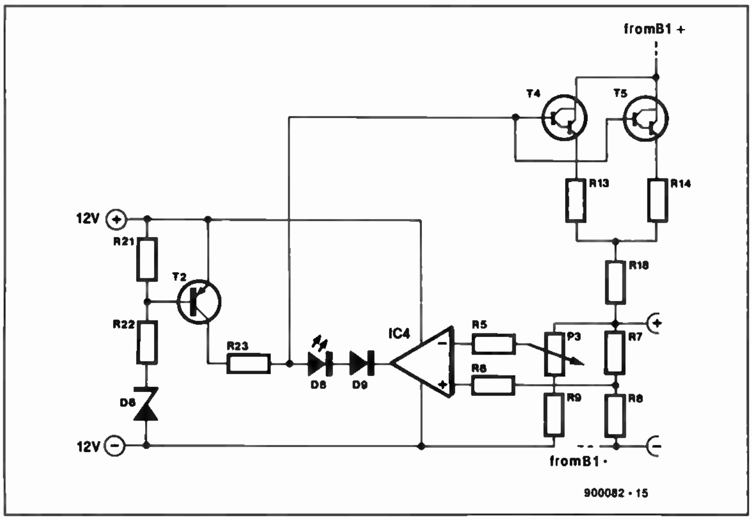


Fig. 5. Basic voltage control circuit.

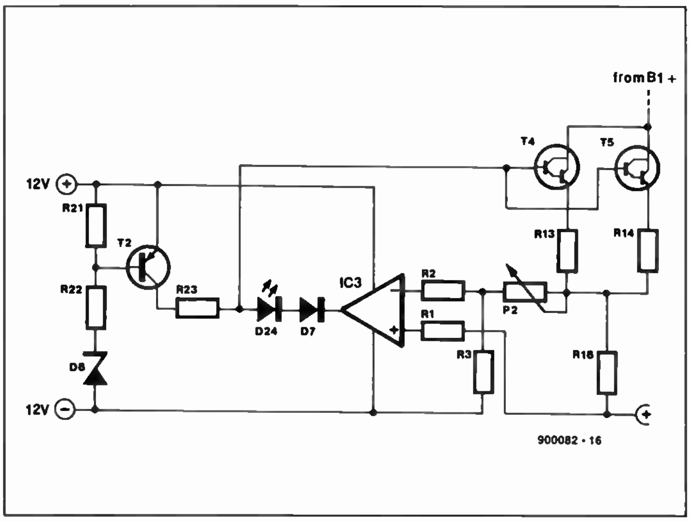


Fig. 6. Basic current control circuit.

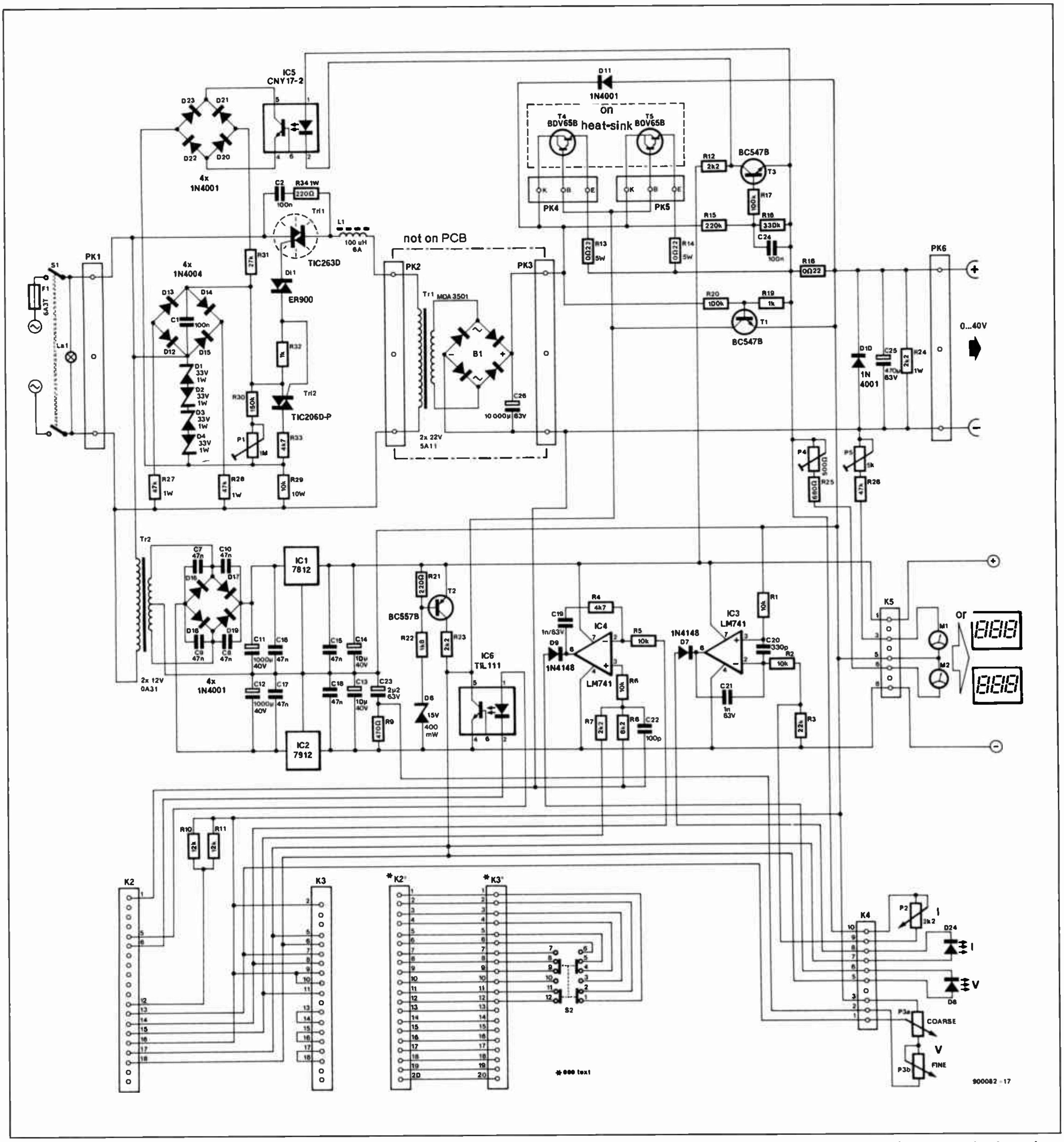


Fig. 7. Circuit diagram ofthe laboratory power supply. Two of these circuits are required for the parallel, series and symmetrical modes.

circuit around T2 keeps the series transistors off until the zener voltage of D6 is reached. This happens when the negative supply voltage of the opamp is sufficiently high. In this way, the voltage peak at switch-on is limited to about 2.5 V above the set output voltage, which is available after a few milliseconds. Although the switch-on peak is not likely to cause damage to most equipment powered by the supply, it is recommended to first switch on the PSU and then connect the load.

The current limiting function of the PSU is provided by the circuit in Fig. 8. As long as the pre-regulation circuit operates correctly, there exists a constant voltage difference across T4-R13 and T5-R14. In a fault condition of any kind (overcurrent, overvoltage), T1 is switched on via potential divider R19-R20. This reduces the base voltage of the darlington transistors, so that the output current is limited. To implement a combined current/temperature overload function, replace resistor R20 by a 100-kΩ NTC (negative temperature coefficient) resistor which is bolted on to the heat-sink, close to T4-T5. ❑

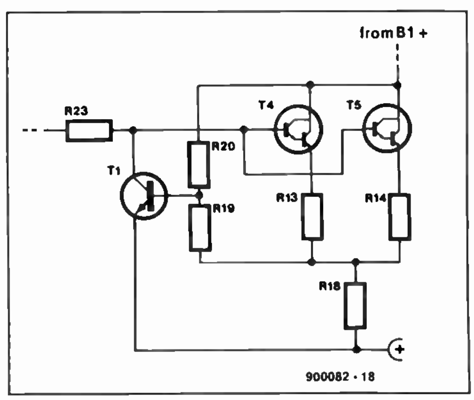


Fig. 8. Basic layout of the current limiter.

MEDIUM POWER A.F. AMPLIFIER

by T. Giffard

For the many readers who do not need hundreds of watts output from their audio hi-fi installation here is a modest 60-watt a.f. amplifier that is a match for even the best quality loudspeakers. Loads down to 2 ohms may be driven without any problem and with very low distortion. The design is entirely symmetrical and, apart from an input capacitor, direct coupled.

WHAT should be the power rating of a domestic hi-fi installation? Twenty watt, 50 watt, 200 watt? It is a vexed question that will never be answered to the satisfaction of every hi-fi buff. Psychology may play a role here, too: the difference in sound pressure between a 20-watt and a 200-watt system is only (!) 10 dB and that sounds a lot less than the difference of 180 watt in output powers. Be that as it may, a continuous power of 50–70 watt is more than adequate for at least 95% of all domestic hi-fi installations. It's far better to have a good-quality 50-watt system than a mediocre 200-watt one. Readers who take music reproduction seriously will, no doubt, have found this out themselves a long time ago. Having said that, it is admitted that there are a few, and fortunately only a few, loudspeakers that need at least a hundred watts to come to live.

Go to almost any hi-fi retailer and ask for a good-quality, medium power (say, 50 W) amplifier and you'll find that there are not many, if any. Until recently, the quality of almost all proprietary power amplifiers was in direct proportion to their power rating and that is the reason that many people buy a system with too high a power output for their requirements. Fortunately, some clever manufacturers have realized this anomaly and are doing something about it, so that amplifiers rated at 50–70 watt with a good specification are slowly becoming available. Nevertheless, we felt that a good-quality design for the enthusiastic DIY-er would be a welcome addition to the couple of high-quality preamplifiers we have published over the past few years.

Symmetrical design

Top priority in the design of the amplifier was quality, and this has resulted in a unit with a signal-to-noise ratio of not less than 100 dB at 1 W, harmonic distortion of not greater than 0.006% (8 Ω/50 W) and a slew rate of 30 V/μs.

If you imagine a horizontal line through the centre of the simplified circuit diagram in Fig. 1, the part above that line is a mirror image of that below the line (ignoring the DC correction and protection circuits).

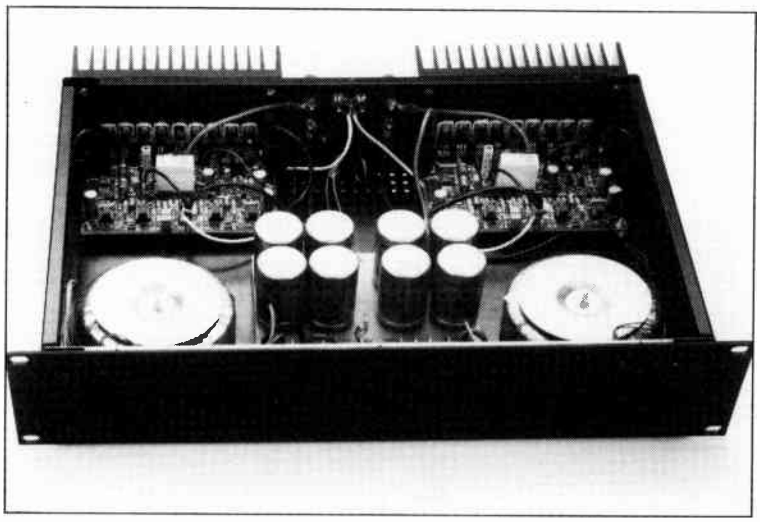
The active DC correction circuit ensures that in no circumstances will direct voltage appear at the output, which is important in view of the direct coupling between all stages.

The protection circuit provides a delay on power-on, monitors whether there is any direct voltage at the output, and measures the current that is drawn by the output transistors. A mechanical relay is used, because we could not devise an electronic switching method that would satisfactorily limit the current without audible side-effects.

Although the design cannot be called revolutionary, we feel that parts of it are pretty original and combine some of the advantages of a number of other, standard designs.

The open-loop gain has been kept low, so that the amount of feedback can be kept small, which is all to the good of the overall quality. After all, the various stages then need contribute a smaller part of the overall amplification, which helps in keeping the distortion in each stage very small.

The input is formed by a differential amplifier, T1-T5, whose gain is limited to about 40 dB. The input transistors are cou-

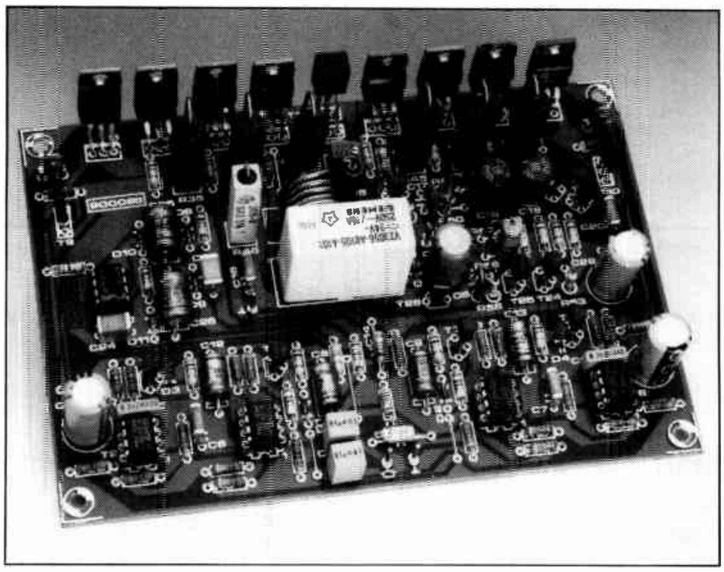


TECHNICAL DATA
(Power supply 225 VA; buffer capacitor 20 000 μF per line)

Continuous power (THD = 0.1%)	60 W into 8 Ω		
	110 W into 4 Ω		
	170 W into 2 Ω		
Music power	200 W in to 2 Ω		
	(1 kHz; 20 ms on, 80 ms off)		
Harmonic distortion	50 W / 8 Ω	100 W / 4 Ω	150 W / 2 Ω
100 Hz	<0.0006%	<0.008%	<0.01%
1 kHz	<0.006%	<0.008%	<0.01%
10 kHz	<0.015%	<0.025%	<0.06%
Intermodulation distortion	<0.013% (1–50 W into 8 Ω at 1 kHz)		
	<0.05% (1 W into 4 Ω at 250 Hz–4 kHz)		
Slew rate	>30 V/μs (with input filter)		
Peak output current	15 A		
Signal-to-noise ratio	>100 dB at 1 W		
Output impedance	<0.02 Ω at 100 Hz and 1 kHz		
	<0.04 Ω at 10 kHz		
Input impedance	47 kΩ		
Input sensitivity	1 V r.m.s.		
Quiescent current	100 mA (nominal)		

pled to another pair of differential amplifiers, T2 and T6 respectively, whose gain is about 27 dB. These are followed by controlled current sources T9 and T10. Setting of the quiescent current for the output stages is provided by a variable zener diode consisting of three transistors, which are connected between the collectors of T9 and T10.

The output stages consist of drivers T13 and T17, and two pairs of three parallel-connected power transistors, each forming a super emitter follower.



Choice of components

In a symmetrical design, equality of the transistors in the input stages is of paramount importance. In our first design, self-paired BC transistors were tried, but these gave problems, particularly with thermal stability. It was therefore decided to use proprietary dual transistors, although these are rather more expensive. However, taking into account the necessary reliability of reproduction of the design, and looking at the near-perfect parallel operation of the dual transistors and their excellent thermal behaviour, the expense is well worth it.

Some of the problems a designer often has to cope with are concerned with the availability of components. The output transistors in the present design are a typical example. Although they were listed as standard types in a Philips catalogue, the ones we ordered had not been delivered six months later in spite of numerous telephone calls. Perhaps this explains why Philips has not been doing too well of late. In the end, the order was cancelled and replaced by one for SGS-Thomson devices (delivered in a few weeks). These are in most respects as good as the Philips types but lack somewhat in bandwidth, although in the prototypes that was not noticeable. On the other hand, the characteristics of the complementary n-p-n and p-n-p types are for all practical purposes identical.

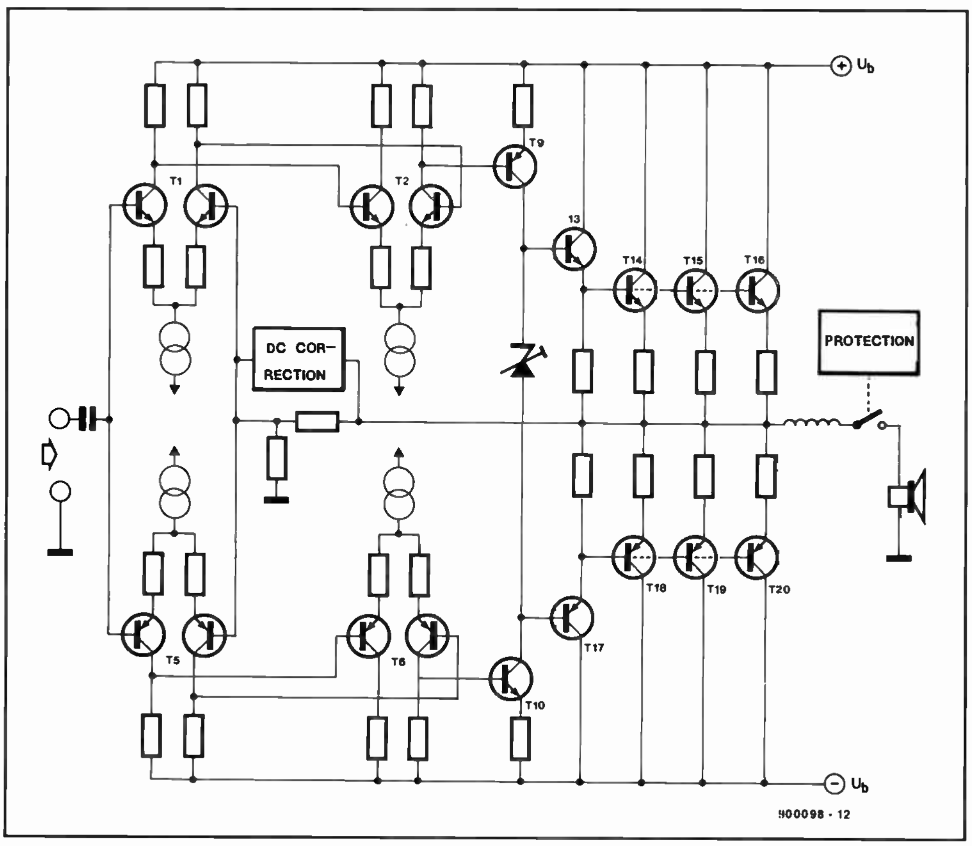


Fig. 1. The simplified circuit diagram of the amplifier clearly shows the symmetrical design.

The final design

The input of the amplifier—see Fig. 2—is protected against direct voltages by capacitors C1 and C2, which may be omitted if the preamplifier to be used already has a capacitor at its output. Be careful in future when connecting a different preamplifier which may not have a capacitor at its output.

The input capacitors are followed by a low-pass filter, R1-C3. This network limits the bandwidth of the input signal to obviate any transient intermodulation distortion (TIM). The cut-off frequency of the filter is about 180 kHz, assuming that the output impedance of the preamplifier is 50 Ω.

Dual transistors T1 and T5 form the symmetrical input stage, which is controlled by current sources T3 and T7. The current through each of the dual transistors is set at just above 1 mA. The gain of the differential input amplifiers is determined by the ratio of the values of their collector and emitter resistors. Networks R5-C6 and R14-C7 set the open-loop bandwidth at 500 Hz.

The second stage consists of dual transistors T2 and T6, which are controlled by current sources T4 and T8. This stage not only provides 27 dB gain, but als impedance matching between the input stage and the following current amplifier.

Current sources T4 and T8 use LEDs to obtain a voltage reference. Bear in mind that these diodes have a potential drop of about 1.6 V across them.

The quiescent current through current amplifiers T9 and T10 is some 20 mA. This level of current is necessary to ensure that drivers T13 and T17 provide sufficient current in all circumstances. The amplification of these amplifiers is determined by the value of their emitter resistors and the input impedance of T13 and T17.

The collectors of the current amplifiers are intercoupled via a presettable 'zener' network, which serves to set the quiescent operating point of the output stages. This network normally consists of just one transistor, but it was found that the thermal behaviour of this was not sufficient to compensate the thermal conduct of the output stages correctly and quickly enough. The three transistors finally chosen work almost perfectly and have the further advantage that they form a virtually ideal current source which ensures that current variations through them hardly affect the zener voltage.

The drivers and associated power transistors are, of course, of the same type. Three

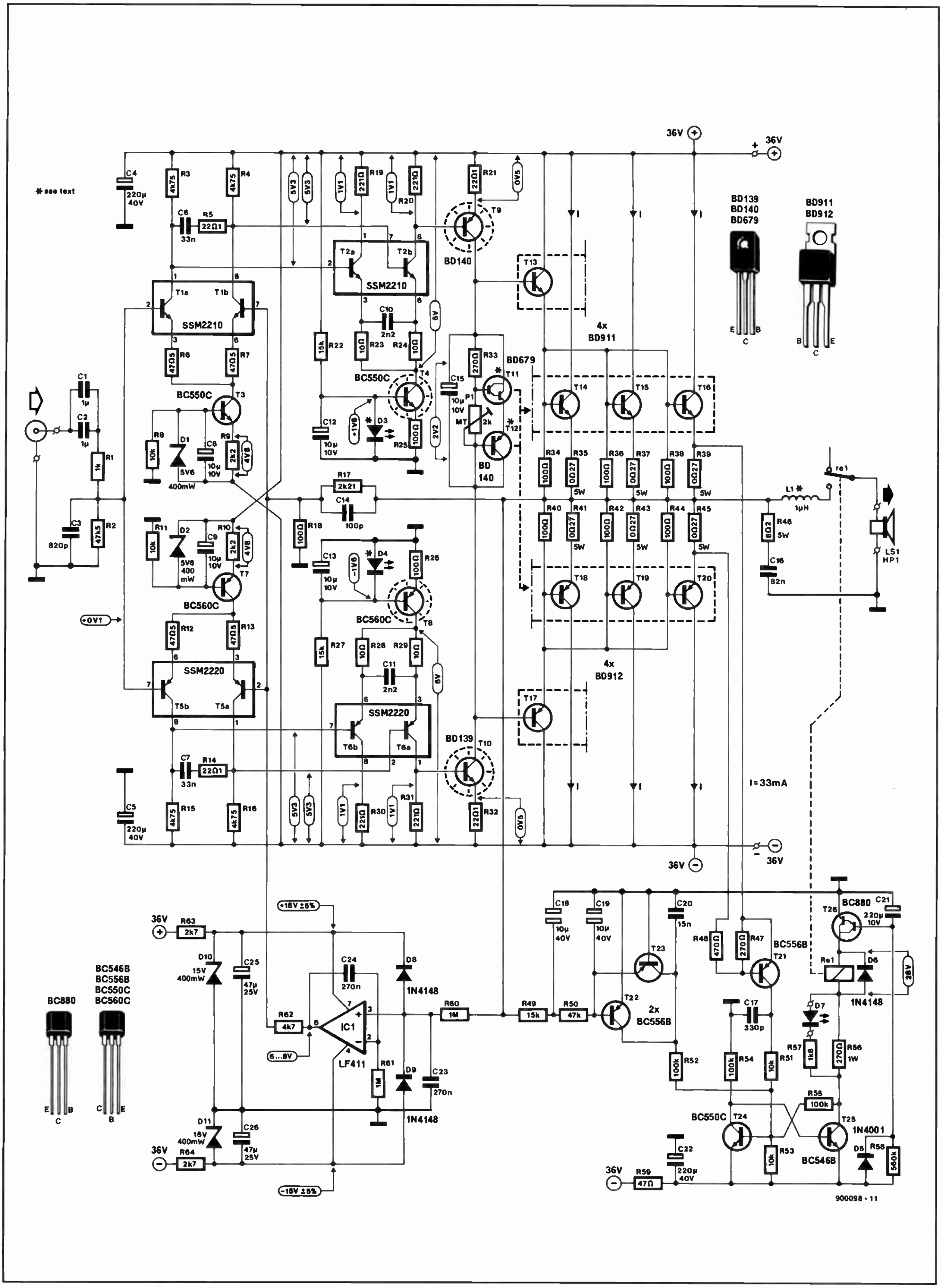


Fig. 2. Circuit diagram of the medium-power A.F. amplifier.

output transistors instead of one make for more efficient cooling; moreover, low-power transistors normally have better electrical properties than high-power ones.

A Boucherot-type network, R46-C16, at the output ensures that the amplifier remains adequately loaded at high frequencies. Inductor L1 limits the rise time of the output signal when the load is capacitive. This results in slightly less damping at high frequencies, but even so the damping factor at 10 kHz and an output impedance of 8 Ω remains greater than 100.

Feedback is provided via resistors R17 and R18, and an important role in this is played by DC correction circuit IC1. Since the feedback loop does not include coupling capacitors, direct voltages at the input are amplified to the same degree as AC signals. Owing to the slightly asymmetric setting of the direct voltages at the differential input stages, which is unavoidable because of differences between n-p-n and p-n-p transistors and also in base resistance, it is essential that the output voltage of the amplifier is kept at 0 V, and this is effected by IC1.

The output of the amplifier is applied to integrator IC1 via low-pass filter R60-C23. The supply for the integrator is derived from the main 36-V supply with the aid of resistors R63 and R64, and zener diodes D10 and D11.

The protection circuit is formed by transistors T21–T26. When the supply is switched on, T26 is off, so that bistable T24-T25 is reset, which results in T25 being switched on. Transistor T26 cannot conduct until its base-emitter voltage is high enough and this does not happen until C21 has been charged via R58. It is this action that delays the actuation of the amplifier.

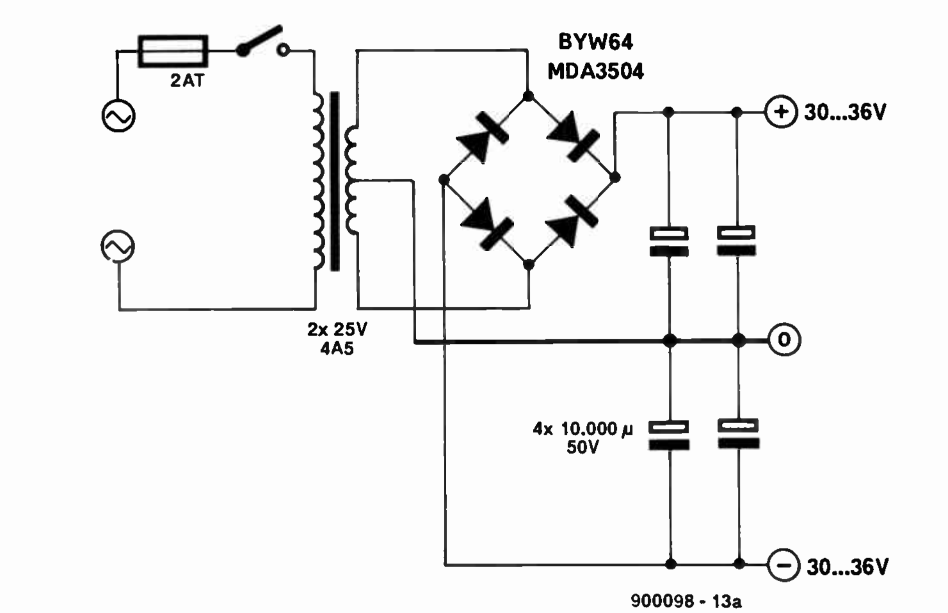
Once T26 is switched on, relay Re1 is energized and diode D7 lights to indicate that all is well.

Transistor T21 monitors the output current of the amplifier by measuring the potential drop across the emitter resistors of T16 and T20 via voltage divider R47-R48. If that current exceeds 5 A (that is, a total current throught the three output transistors of 15 A), T21 switches on which results in bistable T24-T25 toggling after which the relay is deenergized within 5 ms.

Any direct voltage at the output is measured via low-pass filter R49-R50-C18-C19. If there is a direct voltage at a level of more than 1 V, T22 will switch on of the voltage is negative and T23 if it is positive. Again, the relay is deenergized via the bistable.

When the relay is denergized owing to too high a current or voltage, it will remain so until the supply is switched off. When, after a few seconds, the supply is switched on again, the relay will be energized if the fault condition has been removed.

The nominal supply voltage is 2×30 V, which may rise under no-load conditions to 2×37 V. Each amplifier may have its own power supply, but it is, of course, possible to power a stereo system (two amplifiers) from one supply only as discussed below.



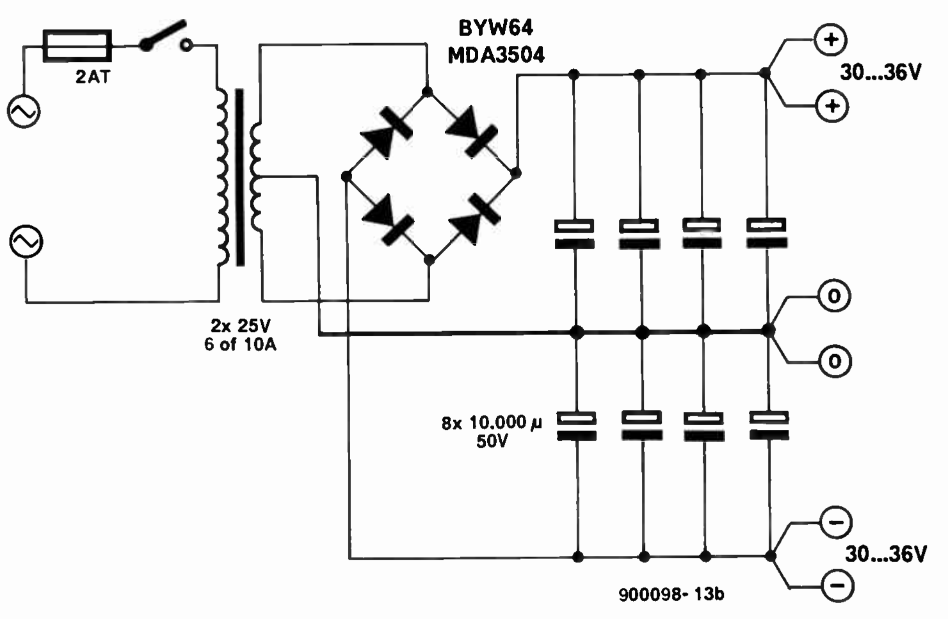


Fig. 3. Two possible power supplies: the one at the top is for a monaural amplifier, while the one below is a single supply intended for a stereo amplifier.

Power supply

In principle, there are three ways of providing the amplifier with power: a single supply for a monaural amplifier; a stereo amplifier with an independent supply for each amplifier; and a stereo amplifier with a single supply. The first two obviously provide the best possible channel separation.

The circuit of a power supply for a monaural amplifier is shown in Fig. 3a. The mains transformer specified provides sufficient power to allow the amplifier delivering continuous power into a 4 Ω load. It was not thought necessary to specify it for continuous power into 2 Ω. After all, the nominal resistance of many loudspeakers is 4 Ω, although there may be dips to 2–3 Ω. However, the electrolytic capacitors have sufficient capacitance to ensure adequate current during peaks in music reproduction. This explains why a total capacitance of 40 000 µF is specified.

The single supply for a stereo amplifier —see Fig. 3b—has a higher rated mains transformer. For normal use, a 6-A type will suffice. However, if low-impedance loudspeakers are used, it is better to use a 10-A type.

The electrolytic buffer capacitors are 50-V types; if these cannot be obtained, 63-V types may be used, although these are somewhat larger.

The construction of the amplifier will be described in next month's issue. ■

DUBBING MIXER EV7000

PART 1: CONNECTION AND CIRCUIT DESCRIPTION

This sound mixer, designed and marketed as a kit by ELV, allows a variety of fading, sound dubbing and voice-over effects to be realized. The voice channel can override the music channel either automatically (by voice control) or manually. A total of fourteen controls and two toggle switches bring out all features of this easy-to-operate unit.

A dubbing mixer like the EV7000 is often used at parties and film or slide presentations, when a voice channel occasionally overrides the (background) music to provide announcements or comment. A smooth transition between the music and the voice channel requires a fader such as the one described here. To prevent different sound qualities on the music and the voice-over, the EV7000 has separate tone controls for each channel. In addition, separate volume and balance controls are provided.

Operation and controls

As shown by the above photograph, all indicators and control elements of the dubbing mixer are arranged on the front panel. The input and output connectors are located on the rear panel.

Before it is taken into use, the dubbing mixer is connected to the power supply and the available audio equipment. A small mains adapter with an output of 12 V at about 300 mA is connected to the 3.5-mm adapter socket on the rear panel. The 'Mic On' and 'Line on' LEDs on the front panel light to indicate that the mixer is on.

The dubbing mixer is best connected between the preamplifier and the power amplifier. The stereo output signal supplied by the preamplifier is connected to the phono (RCA-type) input sockets on the rear panel of the dubbing mixer. The outputs of the mixer are connected to the inputs of the power amplifier. When a mono preamplifier is used, its output signal is fed to the left input channel of the dubbing mixer, while the right input channel is not used.

The dubbing mixer has an internal ampli-

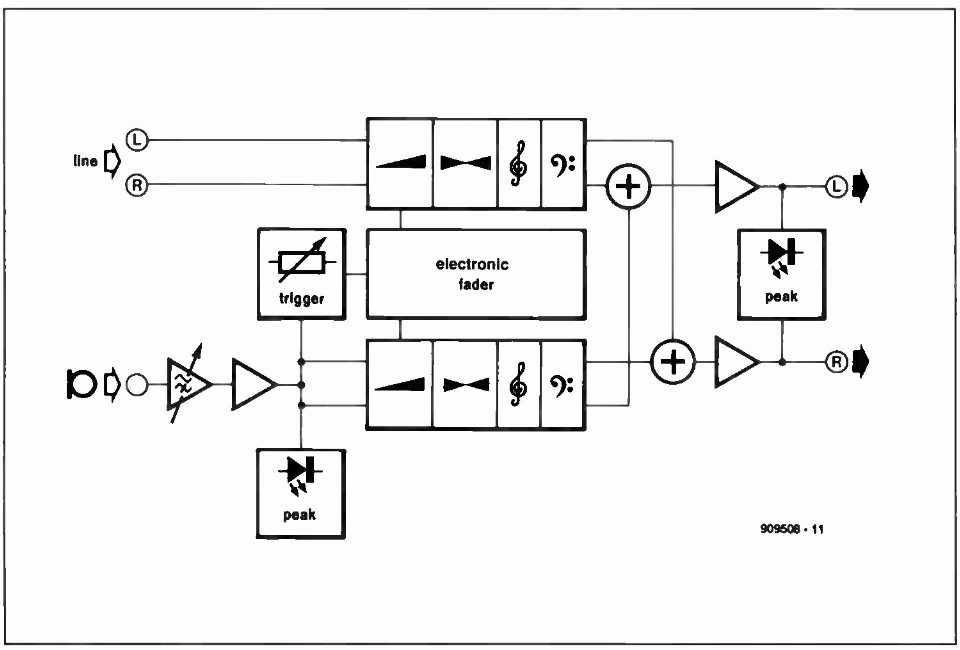


Fig. 1. Block diagram of the dubbing mixer.

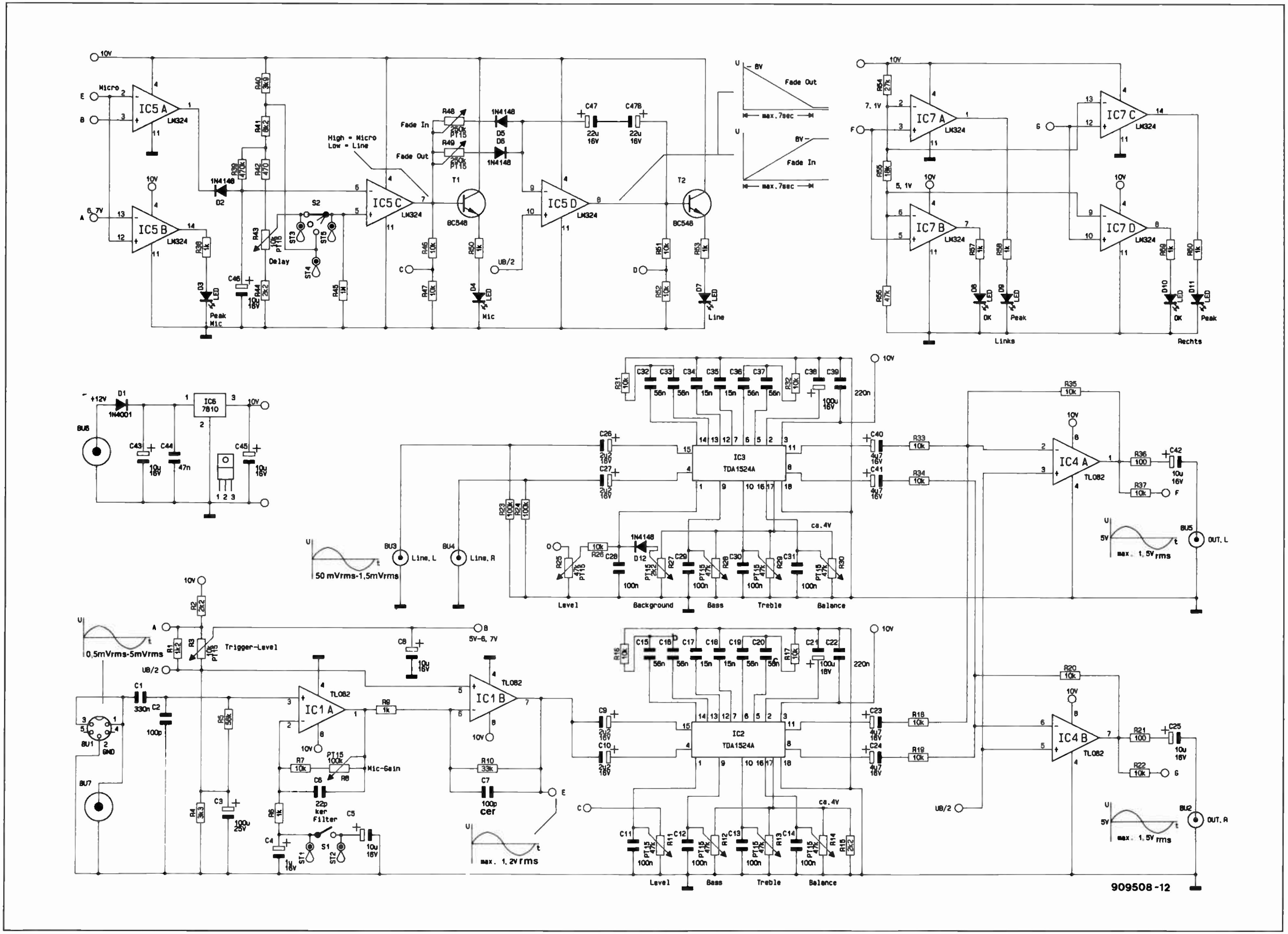


Fig. 2. Circuit diagram of the dubbing mixer. The central parts are two audio signal processing ICs Type TDA1524A (IC2; IC3).

fier capable of handling input levels between about 50 mVrms and 1.7 Vrms, so that virtually any (line-) signal source can be connected.

The stereo output signal of the dubbing mixer has a maximum level of about 1.7 Vrms, and can be set to the required volume within a range of 100 dB. This allows the mixer to drive almost any type of power amplifier.

The mixer has two parallel-connected microphone input sockets: one for a 3.5-mm jack plug and one for a DIN plug.

The controls on the front panel are arranged into three areas, marked by white lines. Much of what follows below on the basic operation of the mixer is illustrated by the block diagram in Fig. 1.

- The controls for the **microphone** channel are located in the top left-hand area on the front panel. The LED at the extreme left indicates the peak microphone level. The tone and volume controls are only effective for the microphone channel. The toggle switch marked 'function' allows three different modes to be selected:
  - 'off' to disable the microphone channel;
  - 'on' to enable the microphone channel;
  - 'auto' to enable the automatic voice control (VOX) function. When selected, this function provides an automatic fade-in when a certain (preset) microphone level is exceeded. The microphone is faded out automatically when its signal level drops below the preset level. The state of the VOX is indicated by the 'Mic on' LED.
- The controls for the **line** channel are found in the lower left-hand area on the front panel. The four LEDs at the left are used as level indicators for the two output channels. The 'Pe' (peak) LEDs should preferably remain off, while the 'OK' LEDs light when the signals are at a sufficiently high level. The minimum recommended signal level is indicated by the 'OK' LEDs flashing irregularly. Like the microphone channel, the line channel has separate tone controls and a level control, marked 'background'. The latter sets the background level of the music signal while the voice channel is actuated. When set fully counter-clockwise, the music is totally suppressed during the voice-over.
  The 'Line on' LED indicates the fade-in and fade-out actions. The intensity at which the LED lights is a rough indication of the background music level during the voice-over. When the LED lights at its full intensity, the music channel is at maximum level, i.e., the microphone channel is off.
- The **basic functions** are set by the controls in the right-hand area on the front panel. The 'Mic gain' potentiometer determines the amplification in the microphone channel. It should be set to a position at which the microphone signal is loud enough when the 'Mic level' control is in the last one-third of its travel. The toggle switch marked 'Mic filter' allows the low side of the frequency range to be limited to about 200 Hz. This mode is particularly suited to the suppression of floor noises, rumble and other low-frequency interference. When the filter is switched off, the frequency range starts at about 20 Hz, which makes the microphone channel usable even for music signals.
  The 'trigger level' control sets the threshold of the previously mentioned VOX function. Turning this control clockwise results in a higher switching threshold, i.e., a higher microphone signal level at which the dubbing mixer switches to voice-over. The 'trigger level' control is, however, enabled only when the 'function' switch is set to the 'auto' position.
  The three timing controls in the lower right-hand corner allow the speed of the fade-in and fade-out effects to be changed to requirement. The 'fade-in' control has a range of 0 to about 7 seconds, independent of the time set with the 'fade-out' control. The 'delay' control determines the time between the end of the voice-over and the start of the music channel fade-in. The range of this control is 0 to about 5 seconds. Like the VOX level control, the 'delay' control is active in the 'auto' mode only. It is intended mainly to prevent the music channel being faded in during every short pause in the voice channel.

Circuit description

The crucial parts in the circuit diagram in Fig. 2 are two audio processor ICs Type TDA1524A. All signal parameters (volume, balance, and tone) are set by control voltages.

The left and right line signals are applied to sockets BU3 and BU4 respectively and fed to input pins 4 and 15 of the TDA1524A (IC3) via coupling capacitors C26 and C27. Since all active parts are contained in the TDA1524A, only a handful of external capacitors and resistors is required to achieve the signal conditioning functions. The bass level is determined by R31, C32 and C33 for the left channel, and R32, C36 and C37 for the right channel. The treble controls require one capacitor only: C34 (left) and C35 (right). The bass and treble settings are controlled by electronic potentiometers in the TDA1524A. These potentiometers, in turn, are controlled by externally applied direct voltages. The volume is set via control input pin 1, the bass level via pin 9, the treble level via pin 10, and the balance via pin 16. At a supply voltage of 10 V, the range of the control voltage is about 0.25 V to 4.0 on all these inputs. The level of the supply voltage hardly affects the settings, however, since the potentiometers that set the sound parameters, R27 – R30, are connected to the reference voltage supplied by pin 17 of the TDA1524A. Capacitors C28 – C31 serve to suppress contact noises as the potentiometers are turned.

The two volume controls that operate at pin 1 of IC3 form a special configuration, in which diode D12 is an important component. The control voltage at the wiper of potentiometer R25 is fed to pin 1 of IC3 via R26 and C28. The positive supply voltage for the potentiometer is provided by the output of IC5 and potential divider R51-R52. In the 'line' mode, R25 is supplied with about 4 V, which allows the full volume range to be covered. When the microphone is switched on, the potentiometer supply voltage drops to about 0 V, which would normally result in total suppression of the music signal. Diode D12, however, keeps the minimum volume control voltage at 0.7 V below the level set with R27, the 'background' level control. The operation of the driver circuit around IC5 will be reverted to in due course.

Electrolytic capacitor C38 forms a buffer for the internal supply voltage of the TDA1524A, while C39 filters the supply voltage applied to the circuit.

The output signals at pin 11 (left) and 8 (right) of the TDA1524A are fed to the summing inputs of inverting amplifiers IC4A and IC4B via C40-R33 and C41-R34. The summing operation involves the two line signals (left and right) and the microphone signal supplied by IC2. Components with values equal to those used for the line signals take the microphone signal from output pins 8 and 11 of IC2 to input pins 2 and 6 of IC4. The two opamps, IC4A and IC4B, serve to invert and to buffer the signals.

The left channel signal reaches the output socket, BU5, via the output of IC4A, pin 1, and R36-C42. Similarly, the right channel signal arrives at BU2 via pin 7 of IC4 and R21-C25.

The microphone signals are treated by IC2 and potentiometers R11 to R14, in a manner similar to that described for the line amplifier, IC3. The supply voltage of the volume control potentiometers is provided by pin 7 of IC5C and potential divider R46-R47. The two inputs of IC2 are driven in parallel by the output of opamp IC1B.

(to be continued next month)

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Project	No.	Price $	Issue
Digital model train	109	11.50	2/89—4/90
Logic analyzer for Atari ST (b/w only)	111	20.00	10/89
Computer-controlled Teletext decoder	113	20.00	10/89
Plotter Driver	117	11.50	5-6/88
FAX interface, IBM PCs	119	14.00	6/90
RAM extension for BBC-B	123	10.00	7/89
EPROM simulator	129	11.50	12/89
RS-232 splitter	1411	11.50	4/90
Centronics ADC/DAC	1421	11.50	5/90
Transistor characteristic plotting (Atari ST b/w)	1431	13.00	5/90
ROM-copy for BASIC	1441	13.00	9/90

PRINTED CIRCUIT BOARDS

Project	No.	Price $
MARCH 1990		
Digital model train	87291-1	7.00
IC monitor	896140	15.00
Power line monitor	900025	9.50
Replacement for TCA280A	894078	11.00
Video mixer (3) (issue 1-4/90)	87304-3	71.00
APRIL 1990		
BBD sound effects unit	900010	15.50
Digital model train	87291-10	8.00
Q meter	900031	12.00
RS-232 splitter	900017-1	14.50
	900017-2	9.00
Video line selector	900032	13.00
MAY 1990		
Acoustic temperature monitor	UPBS-1	4.00
Budget sweep/function generator	900040	14.00
Centronics ADC/DAC	900037D	30.50
Transistor characteristic plotting	900058	9.50

PRINTED CIRCUIT BOARDS—Continued

Product	No.	Price $
JUNE 1990		
Electronic load simulator	900042	24.00
Mini EPROM viewer	900030	36.00
Power zener diode	UPBS-1	4.00
JULY-AUGUST 1990		
Compact 10A power supply	900045	23.00
Intermediate projects	UPBS-1	4.00
Mini FM transmitter	896118	8.50
Sound demodulator for satellite-TV receivers	900057	7.50
Audio power Indicator	904004	7.50
Four-monitor driver for PCs	904067	10.50
SEPTEMBER 1990		
High current h_{FE} tester	900078	11.00
Infrared remote control	904085/86	3.50
OCTOBER 1990		
Guitar tuner	900078	Not Avail.
µP-controlled telephone exchange	900081	36.00
Medium power audio amplifier	900098	18.00
Tuned UHF TV preamp		Not avail.
Dubbing mixer EV7000		*
S-VHS/CVBS-to-RGB converter	900055	24.50
Phase meter	896056	Not avail.

*Available from ELV France, BP-40-F.-57480 Sierck-les-Bains, France.

TERMS OF BUSINESS

PRICING
Prices, except as noted, include shipping in the continental USA. The minimum order is $10. A $2 service charge will be added to orders of less than $10. Prices are subject to change as our costs change. WE RESERVE THE RIGHT TO MAKE PRICE CHANGES WITHOUT NOTICE.

ORDERING
Please use the ORDER FORM provided and include part number, quantity and price. Please print clearly and include your Magic Number if you are a subscriber. Please include a ZIP code in your address. The ORDER FORM is designed to help us quickly and accurately process your order. We reserve the right to refuse mail orders not made using the ORDER FORM or orders under the $10 minimum.

TELEPHONE ORDERS
Telephone orders are welcome between the hours of 9:00 a.m. and 4:00 p.m. EST, Monday-Friday. Orders for parts are to be made on the ORDER FORM provided. Payment for telephone orders is by MC/VISA only. Our TELEPHONE ORDER NUMBER IS (603) 924-6371 or 924-6526. Our answering machine for MC/VISA orders during non-business hours is (603) 924-6371. FAX: (603) 924-9467 (24 hours).

PAYMENT
All remittances **MUST BE IN US FUNDS**. Payment for telephone orders must be made by MasterCard or VISA. Mail order payment may be made by MC/VISA, money order, cashier's check or personal check. If payment is made by personal check, allow up to two weeks for check clearance before your order is shipped. Should a check be returned for any reason, there will be a $10 service charge.

SHIPPING
Unless noted, freight is prepaid by Old Colony. All orders shipped USPS first class mail or by UPS. Allow two to three weeks for delivery. Please inquire about appropriate funds for special handling (UPS next day or second day). UPS requires a street address. If you cannot receive UPS delivery, please include an extra $2 for insured service via Parcel Post. We cannot accept responsibility for safety or delivery of uninsured Parcel Post shipments. **Absolutely no COD shipments**.

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The shipping/handling charge is $1.75, plus 50¢ each additional item. **No returns.** According to the Connecticut Dept. of Revenue Services, Regulation #27, Section 12-426-27, in order to sell software to residents of Connecticut, we must register with the department as well as collect taxes for the state from the purchaser. We regret to inform our customers that we will not sell software to anyone from Connecticut, as it will prove to be too costly and burdensome.

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CIRCUIT BOARDS

Old Colony's circuit boards are made of top quality epoxy glass, two ounce copper and reflowed solder-coated material for ease of constructing projects which have appeared in *Audio Amateur*, *Glass Audio* and *Speaker Builder* magazines. Many also have the component layout printed right on the board!

The builder needs the original article (indicated by the date in brackets) to construct the project. Articles are not supplied but are available through *Audio Amateur Publications*.

AUDIO AMATEUR

A-1: WILLIAMSON TWIN 20 POWER AMPLIFIER. (RW-8) 3 x 5" [1:70] Each **$8.35**

A-2: WILLIAMSON TWIN 20. Power supply board. (RW-9:2) 2½ x 3" [1:70] Each **$6.85**

B-2: WILLIAMSON-WATLING 4+4 MIXER. Staked terminals. (RW:AW) 3½ x 5" [2:71] Each **$10.00**

B-5: WILLIAMSON TWIN 20 PREAMP. Stereo. (RW-11) 3 3/8 x 8" [2:71] Each **$16.00**

B-7: V.U. METER. (DG-7-A) 1¾ x 3" [3:71] Each **$8.00**

C-4: ELECTRONIC CROSSOVER. Board takes 8 pin DIPs, ten eyelets for variable components. (DG-13R) 2 x 3" [2:72] Each **$10.00**

C-5: GLOECKLER VOLUME CONTROL. 23 position wafer. (FG-1) 3 x 3" [2:72] Each **$5.50**

D-1: HERMEYER ELECST AMP II. 4½ x 4⅞" [3:73] Each **$12.00**

E-2: REGULATED LAB POWER SUPPLY. ±15V. (Ref. 1) 4 3/8 x 3 1/8" [4:74] Each **$8.00**

F-3: GATELY ±18V POWER SUPPLY. Regulated. (EG-2) 2¼ x 4" [2:75] Each **$8.00**

F-6: 30Hz FILTER/CROSSOVER. High pass or universal filter or crossover. (WJ-3) 3 x 3" [4:75] Each **$10.00**

G-1: GATELY PEAK DETECTING OVERLOAD INDICATOR. Two channel. (EG-3) 1½ x 2½" [2,3:76] Each **$6.40**

H-2: SPEAKER SAVER. (WJ-4) 3¼ x 5¼" [3:77] Each **$13.25**

H-4: GATELY MICROMIXER. Input. 15 pin plug-in gold edge. (MIC-10S) 8¾ x 3" [3:77] Ea. **$17.00** Five or more, Each **$15.00**

H-5: GATELY MICROMIXER. Output. 15 pin plug-in gold edge. Two channel. (MIC11-005) 12¾ x 3" Each **$26.20**

H-7: GLOECKLER 101dB ATTENUATOR. (FG-3) 2 x 2" [4:77] Each **$8.00** Five for **$35.00**

H-8: MORREY SUPER BUFFER. Two channel. (WM-3) 1½ x 2½" [4:77] Each **$8.00**

J-4: DIDDEN AUDIO ACTIVATED POWER SWITCH. (J-4) 3 x 4⅛" [3:78] Each **$7.55**

J-5: PASS A-40 POWER AMP. One channel. 3 x 3" [4:78] Each **$6.00**

J-6: SCHROEDER CAPACITANCE CHECKER. (CT-10) [4:78] 3¼ x 6" Each **$9.95**

K-3: CRAWFORD WARBLER. 3¼ x 3 3/8" [1:79] Each **$11.20**

K-6: WALDRON TUBE CROSSOVER. (Two needed per 2-way channel.) 2 x 4½" [3:79] Each **$12.00** Four **$40.00**

K-7: WALDRON TUBE CROSSOVER POWER SUPPLY. 5 x 5⅝" [3:79] Each **$12.95**

K-11: WILLIAMSON 40/40 POWER AMP. One channel. 3 x 5" [4:79] Each **$7.00**

L-1A: BOAK POWER AMP REGULATED SUPPLY. Plus or minus supply for power. 2¾ x 4⅛" [1:80] Each **$8.00**

L-2: WHITE LED OVERLOAD & PEAK METER. One channel. 3 x 6" [1:80] Each **$18.70**

L-4: SULZER OP-AMP PREAMP POWER SUPPLY. ±15V supply for preamps. 4 3/8 x 4" [2:80] Each **$12.00**

L-6: MASTEL TONE BURST GENERATOR. 3½ x 6 5/8" [2:80] Each **$15.75**

L-9: MASTEL PHASE METER. 6 5/8 x 2 3/8" [4:80] **$11.25**

M-1: MULLER-CARLSTROM. Sweep Generator-Oscillator. (Two required.) (CM-2) 2 9/16 x 5" [2:81] Each **$8.50** Pair **$14.00**

M-2: MULLER-CARLSTROM. Log Sweep Board. (CM-4) 2 x 2 1/8" [2:81] Each **$5.00**

M-3: MULLER-CARLSTROM. Sweep Power Supply. (CM-5) 2 5/8 x 3 5/8" [2:81] Each **$6.50**

M-4: MULLER-CARLSTROM. Logger Board. (CM-3) 3½ x 4" [3:81] Each **$9.25**

M-5: MULLER-CARLSTROM. Logger Power Supply. (DG-12B) 2½ x 2¾" [3:81] Each **$5.00**

M-6: CARLSTROM IM FILTER. Intermodulation Filter. 2 5/8 x 3¾" [1:82] Each **$6.50**

P-3: BORBELY 60W POWER AMP. (EB-60) 3 3/8 x 6 1/8" [2:82] Each **$11.75**

P-5: SWEEP MARKER ADDER. 3½ x 2¾" [2:82] Each **$6.20**

P-6: ADVENT MIKE PREAMP UPDATED. (K5) 3 7/8 x 2 3/8" [3:82] Each **$18.75**

R-2: BORISH DIGITAL DELAY. 5¾ x 9" [1,2:83] Each **$79.80**

R-4: DIDDEN MAIN PWR AMP. 4 5/8 x 6 3/8" [4:83] Each **$30.00**

S-1: BORBELY SERVO 100 AMP. 4 1/8 x 6½" [1:84] Each **$16.00**

S-3: BORBELY DC 100 AMP. 6½ x 4 1/8" [2:84] Each **$16.00**

S-5: KRUEGER MOD FOR MORREY IG-18. 2 11/16 x 2 1/8" [3:84] Each **$7.80**

S-6: CURCIO VACUUM TUBE PRE-PREAMP AMP/REGULATOR. 4¾ x 2¾" [5:84] Each **$12.35**

S-6A: CURCIO VACUUM TUBE PRE-PREAMP MASTER POWER SUPPLY. 4¾ x 3¾" [5:84] Each **$10.35**

S-8: DIDDEN FAN CONTROL. 6¼ x 1⅝" [4:84] Each **$11.25**

T-1: CURCIO VACUUM TUBE PREAMP/AMP REGULATOR. 10 11/16 x 6 15/16" [2:85] Each **$36.00**

T-1A: CURCIO VACUUM TUBE PREAMP MASTER POWER SUPPLY. 3 3/16 x 2 11/16" [2:85] Each **$8.00**

T-2A: BORBELY R1AA-1. 3⅞ x 3⅞" Each **$10.50**

T-2B: BORBELY R1AA-2. 3⅞ x 5⅞" Each **$15.00**

T-2C: BORBELY TAPE BUFFER. 1⅝ x 3⅞" Each **$5.45**

T-2D: BORBELY LINE BUFFER. 3⅞ x 5⅛" Each **$9.50**

T-2S: BORBELY PREAMP BOARD SET. Eight boards. [4:85, 1:86] Set **$90.00**

T-2F: BORBELY PREAMP POWER SUPPLY. (Two required.) 3⅞ x 4⅞" Each **$12.00**

V-2: LANG CLASS A MOSFET AMP. 5 11/16 x 6 1/16" [2:86] Each **$21.40**

V-3A: CURCIO AUTO MUTE. 1½" x 2" Each **$8.00**

W-3: BORBELY IMPROVED POWER SUPPLY. 4¼ x 5½" [1:87] Each **$16.00**

X-3: CHATER 40W MOSFET AMP. Two sided, one channel. 4 x 6¾" [2, 3:88] Each **$26.00**

X-3A: CHATER AMP POWER SUPPLY. 3½ x 6" [2, 3:88] Each **$14.00**

X-4A: VIKAN CAR AMPLIFIER. 4 x 5" [4:88, 1:89] Each **$23.20**

X-4B: VIKAN PWR SUPPLY. 4⅛ x 5⅛" [4:88, 1:89] Each **$17.00**

Y-2: RYAN ADCOM GFA-555 POWER SUPPLY REGULATOR. (One per channel required.) 3 x 6¼" [4:89] Each **$28.50**

SPEAKER BUILDER

SB-A1: LINKWITZ CROSSOVER. 5½ x 8½" [4:80] Each **$25.50**

SB-D2: WITTENBREDER PULSE GENERATOR. 3½ x 5" [2:83] Each **$11.85**

SB-E2: NEWCOMB PEAK PWR INDICATOR. 1 x 2" [2:84] Each **$3.90**

SB-E4: MULLER PINK NOISE GENERATOR. 4⅛ x 2 3/16" [4:84] Each **$9.40**

GLASS AUDIO

GB-1A: CURCIO ST-70 POWER SUPPLY. 5 x 9" [1:89] Each **$27.00**

GB-1B: CURCIO ST-70 DRIVER BOARD. 3¼ x 7" [1:89] Each **$17.00**

CIRCUIT BOARD ORDER FORM

Please print *in clear block capitals* quantity needed, board number and price. Total the amounts and **REMIT IN US $ ONLY** by MC/VISA, check or money order. Postpaid in US; in Canada, please add 10%; overseas, add 20%.

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QTY		PRICE
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SESCOM INC.
PARTS DIVISION
2100 WARD DR.
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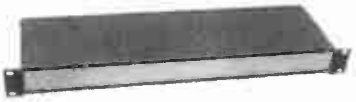
THE "FIRST CHOICE" FOR CONSTRUCTORS HARDWARE

Rack Chassis



MODEL #.	DESCRIPTION (Width x Depth x Height)		PRICE $
1RU5	19 x 5 x 1.75 in.	(482.6 x 127 x 44.45mm)	28.00
1RU7	19 x 7 x 1.75 in	(482.6 x 177 x 44.45mm)	30.00
1RU10	19 x 10 x 1.75 in.	(482.6 x 254 x 44.45mm)	32.00
2RU5	19 x 5 x 3.50 in.	(482.6 x 127 x 88.9mm)	30.00
2RU7	19 x 7 x 3.50 in.	(482.6 x 177.8 x 88.9mm)	32.00
2RU10	19 x 10 x 3.50 in.	(482.6 x 254 x 88.9mm)	34.00
3RU5	19 x 5 x 5.25 in.	(482.6 x 127 x 133.35mm)	38.00
3RU7	19 x 7 x 5.25 in.	(482.6 x 177 8 x 133.35mm)	40.00
3RU10	19 x 10 x 5.25 in.	(482.6 x 254 x 133.35mm)	42.00

The above rack chassis are made of .063 aluminum.
The front and rear panels are clear brushed anodized.
All hardware is included. Assembly Required.

Metal Cabinets

MODEL #.	DESCRIPTION (Width x Depth x Height)		PRICE $
MC-1A	4 x 3 x 2 in.	(101.6 x 76.2 x 50.8mm)	15.00
MC-2A	6 x 3 x 2 in.	(152.4 x 76.2 x 50.8mm)	17.00
MC-3A	8 x 3 x 2 in.	(203.2 x 76.2 x 50.8mm)	19.00
MC-4A	4 x 5 x 3 in.	(101.6 x 127 x 76.2mm)	17.00
MC-5A	6 x 5 x 3 in.	(152.4 x 127 x 76.2mm)	19.00
MC-6A	8 x 5 x 3 in.	(203.2 x 127 x 76.2mm)	21.00
MC-7A	4 x 7 x 4 in.	(101.6 x 177.8 x 101.6mm)	19.00
MC-8A	6 x 7 x 4 in.	(152.4 x 177.8 x 101.6mm)	21.00
MC-9A	8 x 7 x 4 in.	(203.2 x 177.8 x 101.6mm)	23.00

The above cabinets are made of .063 aluminum.
The front and rear panels are clear brushed anodized.
All hardware and rubber feet are included. Assembly Required.

Punch Kits and Punches

Model #.	Description	Price $
HP-1	5 JR TOOL KIT	49.00
HP-3	BENCH MOUNT	15.00
PD-1	ROUND 1/16"	6.00
PD-2	ROUND 5/64"	6.00
PD-3	ROUND 3/32"	6.00
PD-4	ROUND 7/64"	6.00
PD-5	ROUND 1/8"	6.00
PD-6	ROUND 9/64"	6.00
PD-7	ROUND 5/32"	6.00
PD-8	ROUND 11/64"	6.00
PD-9	ROUND 3/16"	6.00
PD-10	ROUND 13/64"	6.00
PD-11	ROUND 7/32"	6.00
PD-12	ROUND 15/64"	6.00
PD-13	ROUND 1/4"	6.00
PD-14	ROUND 17/64"	6.00
PD-15	ROUND 9/32	6.00
PD-16	SQUARE 1/8"	66.90
PD-17	SQUARE 5/32"	66.90
PD-18	SQUARE 3/16"	66.90
PD-19	REC. 1/8 x 3/16"	66.90
PD-20	REC. 1/8 x 7/32"	66.90
PD-21	REC. 1/8 x 15/64"	66.90

Model #.	Description	Price $
HP-2	XX TOOL KIT	129.75
HP-4	BENCH MOUNT	15.00
PD-30	ROUND 1/16"	9.50
PD-31	ROUND 5/64"	9.50
PD-32	ROUND 3/32"	9.50
PD-33	ROUND 7/64"	9.50
PD-34	ROUND 1/8"	9.50
PD-35	ROUND 9/64"	9.50
PD-36	ROUND 5/32"	9.50
PD-37	ROUND 11/64"	9.50
PD-38	ROUND 3/16"	9.50
PD-39	ROUND 13/64"	9.50
PD-40	ROUND 7/32"	9.50
PD-41	ROUND 15/64"	9.50
PD-42	ROUND 1/4"	9.50
PD-43	RONND 17/64"	9.50
PD-44	ROUND 9/32	9.50
PD-45	ROUND 19/64"	9.50
PD-46	ROUND 5/16"	9.50
PD-47	ROUND 21/64"	9.50
PD-48	ROUND 11/32"	9.50
PD-49	ROUND 23/64"	9.50
PD-50	ROUND 3/8"	9.50
PD-51	ROUND 25/64"	9.50
PD-52	ROUND 13/32"	9.50
PD-53	ROUND 27/64"	9.50
PD-54	ROUND 7/16"	9.50
PD-55	ROUND 29/64"	9.50
PD-56	ROUND 15/32"	9.50
PD-57	ROUND 31/64"	9.50
PD-58	ROUND 1/2"	6.00
PD-59	ROUND 33/64"	6.00
PD-60	ROUND 17/32"	6.00
PD-61	SQUARE 1/8"	66.90
PD-62	SQUARE 5/32"	66.90
PD-63	SQUARE 3/16"	66.90
PD-64	REC. 1/8 x 3/16"	83.10
PD-65	REC. 1/8 x 1/4"	83.10
PD-66	REC. 1/8 x 5/16"	83.10
PD-67	REC. 1/8 x 3/8"	83.10
PD-68	REC. 1/8 x 11/32	75.10

Rub-On Letters

MODEL #.	DESCRIPTION	PRICE $
LT-1	CAPITAL LETTERS HELVETICA 14 pt.	3.95
LT-2	LOWER CASE LETTERS HELVETICA 14 pt.	3.95
LT-3	NUMBERS HELVETICA 14 pt.	3.95
LT-4	CAPITAL LETTERS HELVETICA 10 pt.	3.95
LT-5	LOWER CASE LETTERS HELVETICA 10 pt.	3.95
LT-6	NUMBERS HELVETICA 10 pt.	3.95
LT-7	CAPITAL LETTERS HELVETICA 6 pt.	3.95
LT-8	LOWER CASE LETTERS HELVETICA 6 pt.	3.95
LT-9	NUMBERS HELVETICA 6 pt.	4.95
LT-10	DIAL MARKING (DOTS)	4.95
LT-11	DIAL MARKING (LINES)	3.95
LT-12	WORDS (AUDIO) HELVETICA 10 pt.	3.95
LT-12	WORDS (AUDIO) HELVETICA CAPS 6 pt	3.95
LT-12	WORDS (AUDIO) HELVETICA UPPER AND LOWER CASE 6 pt.	3.95

TERMS AND CONDITIONS

MINIMUM ORDER: The minimum order is $10.00.

TERMS: Terms of payment on open accounts are NET 30 days from date of invoice.

C.O.D. ORDERS: Payment must be made in cash, money order or certified check. A company check will be accepted only if previously approved by the factory.

OPEN ACCOUNTS: SESCOM, INC. extends credit to government agencies and industrial accounts with a good published rating. Firms may apply for an open account by requesting a credit application from the Accounts Receivable department. All orders are shipped C.O.D. until credit is approved. (Please allow 4 to 6 weeks to open a new account.) We accept Mastercard and Visa.

PRICES: Catalog prices are NET, FOB destination shipped UPS ground. UPS BLUE and RED LABEL service available at an additional charge: UPS BLUE is $10.00 and UPS RED is $20.00. PRICES ARE SUBJECT TO CHANGE WITHOUT NOTICE. Orders are invoiced at prices in effect at the time of shipment unless specifically quoted otherwise. SESCOM, INC. shall add any existing or future Federal, State, or Local taxes that may apply. We reserve the right to apply the above if your Tax Exemption Certificate is not on file with our office.

SHORTAGE OR DAMAGE: All claims for shortage or shipment error must be made within 10 days after receipt of the shipment. Our liability is limited to the material value on the invoice.

MERCHANDISE RETURN: A prior written authorization from SESCOM, INC. must be issued before return of any material. Instructions for returning the material will accompany the authorization. Material that is returned without authorization or is over 90 days will carry a restocking fee of a minimum of 25% to a maximum 100%. Special order items will not be accepted for return.

SUBSTITUTIONS: We reserve the right to substitute higher quality (but identical part) at our option, unless you specify no substitution at the time of the order.

Sheet Metal Punches

MODEL #.	HOLE SIZE		PILOT HOLE	PRICE $
PUNCH 1	3/8"	ROUND	3/16"	8.95
PUNCH 2	7/16"	ROUND	7/16"	8.95
PUNCH 3	1/2"	ROUND	7/32"	8.95
PUNCH 4	9/16"	ROUND	7/32"	11.95
PUNCH 5	5/8"	ROUND	7/32"	11.95
PUNCH 6	11/16"	ROUND	7/32"	11.95
PUNCH 7	3/4"	ROUND	7/32"	11.95
PUNCH 8	13/16"	ROUND	7/32"	11.95
PUNCH 9	7/8"	ROUND	7/32"	11.95
PUNCH 10	1"	ROUND	5/16"	12.95
PUNCH 11	1-1/16"	ROUND	5/16"	12.95
PUNCH 12	1-1/8"	ROUND	5/16"	12.95
PUNCH 13	1-3/16"	ROUND	5/16"	12.95
PUNCH 14	1-1/4"	ROUND	5/16"	13.95
PUNCH 15	1-3/8"	ROUND	5/16"	13.95
PUNCH 16	1-1/2"	ROUND	5/16"	15.95
PUNCH 17	1-5/8"	ROUND	5/16"	20.95
PUNCH 18	1-3/4"	ROUND	5/16"	23.95
PUNCH 19	2-5/8"	ROUND	1/2"	59.95
PUNCH 20	11/16"	SQUARE	1/2"	31.95
PUNCH 21	3/4"	SQUARE	1/2"	35.95
PUNCH 22	1"	SQUARE	1/2"	45.95
PUNCH 23	21/32 x 5/16"	REC.	1/2"	45.95
*PUNCH 24	*THURST RACES FOR PUNCHES (FOR EASIER PUNCHING) 1" thru 1-3/4"			9.95

Component Parts

TRANSISTORS:

PART NO.	DESCRIPTION	1-9	10-99	100 +
2N3904	NPN T092	.12	.10	.08
2N3906	PNP T092	.12	.10	.08
2N2102	PNP T039	.80	.67	.54
2N4037	PNP T039	.90	.75	.60
PN2219A	NPN T0237	.40	.34	.28
PN2905A	PNP T0237	.40	.34	.28

DIODES:

PART NO.	DESCRIPTION	1-9	10-99	100 +
1N34A	GE DIODE	.17	.15	.13
1N4001	SI DIODE	.13	.11	.10
1N4148	SI DIODE	.09	.07	.06

LINEAR INTEGRATED CIRCUITS:

PART NO.	DESCRIPTION	1-9	10-99	100 +
LF351N	SINGLE OP-AMP	.60	.54	.48
LF353N	DUAL OP-AMP	1.05	.95	.84
NE5534N	SINGLE OP-AMP	1.28	1.15	1.02
NE5534AN	LOW NOISE SINGLE OP-AMP	1.80	1.62	1.44
NE5532N	DUAL OP-AMP	1.85	1.67	1.48

I C SOCKETS:

PART NO.	DESCRIPTION	1-9	10-99	100 +
SOC-8	8 PIN TIN	.11	.10	.09
SOC-14	14 PIN TIN	12	.11	.10
SOC-16	16 PIN TIN	.13	.12	11
SOC-18	18 PIN TIN	.15	.14	13
SOC-20	20 PIN TIN	.19	.16	15

CARBON FILM RESISTORS:

PART NO.	DESCRIPTION	1-9	10-99	100 +
CF + VALUE	5 % 1/4W	.05	.02	.01

ALL STANDARD VALUES 1.0 OHM THRU 1.0M OHM

POTENTIOMETERS:

PART NO.	DESCRIPTION	1-9	10-99	100 +
10KA	10K AUDIO TAPER	2.00	1.80	1.60

CONNECTORS:

PART NO.	DESCRIPTION	1-9	10-99	100 +
XLR F CA	3 PIN CABLE FEMALE	3.20	2.88	2.56
XLR M CA	3 PIN CABLE MALE	2.75	2.48	2.20
XLR F CH	3 PIN CHASSIS FEMALE	3.60	3.24	2.88
XLR M CH	3 PIN CHASSIS MALE	2.60	2.40	2.08
1/4 ST	1/4" STERIO JACK	97	.87	.78
3.5 ST	3.5mm STERIO JACK	.97	.87	78
RCA	RCA PHONO JACK CHASSIS	.65	.59	.52

CAPACITORS: Axial Lead Electrolytic

PART NO.	DESCRIPTION	1-9	10-99	100 +
AC 1/50	1uf - 50V	.38	.35	.30
AC 4.7/25	4.7uf - 25V	.39	.35	.30
AC 10/25	10uf - 25V	.39	.35	.30
AC 47/25	47uf - 25V	.46	.42	.37
AC 100/25	100uf - 25V	.69	.62	.55
AC 220/35	220uf - 35V	.97	.87	.78
AC 470/35	470uf - 35V	1.34	1.21	1.07
AC 1000/35	1000uf - 35V	2.22	2.00	1.78

CAPACITORS: Radial Lead Electrolytic

PART NO.	DESCRIPTION	1-9	10-99	100 +
RC 1/50	1uf - 50V	.26	.24	.21
RC 4.7/25	4.7uf - 25V	.28	.26	.23
RC 10/25	10uf - 25V	.31	.28	.25
RC 47/25	47uf - 25V	.39	.35	.28
RC 100/25	100uf - 25V	.41	.36	.33
RC 220/35	220uf - 35V	.61	.55	.49
RC 470/35	470uf - 35V	.99	.90	.80
RC 1000/35	1000uf - 35V	1.27	1.15	1.02

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Function	Freq. Period Ratio,Interval, Avg. Prescale	Frequency	Frequency	Frequency	Frequency	Frequency	RF Indicator
Range	10Hz-2.4GHz	10MHz-2.4GHz	10Hz-2.2GHz	1MHz-1.3GHz	10MHz-2.4GHz	10MHz-550MHz	10MHz-1.8GHz
Display	10 Digit LCD w/Function Annunciators	10 Digit LCD	8 Digit LED	8Digit LED	8 Digit LED	8 Digit LED	•
RF Signal Strength Indicator	16 Segment Adjustable Bargraph	16 Segment Adjustable Bargraph	•	•	•	LED with Adjustable Threshold	10 Segment Adjustable Bargraph
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