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JUNE 1966
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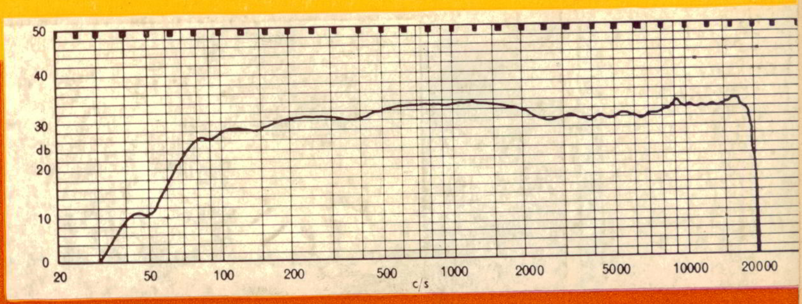

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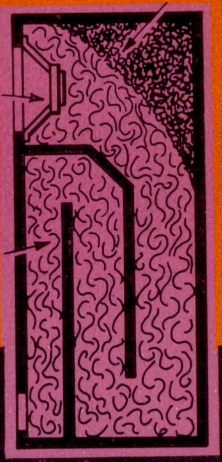

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"Hall Mark" for Instruments

A MODERN translation of Clause 35 of Magna Carta reads "There shall be standard measures of wine, ale, and corn (the London quarter), throughout the Kingdom. There shall also be a standard width of dyed cloth, russet, and haberject, namely two ells within the selvedges. Weights are to be standardized similarly."

Standardization of units of length, volume and weight have long been established and from time to time inspectors from the appropriate Government department check against transfer standards the "measures" used by tradesmen. But no such transfer standards are available to the U.K. manufacturer of, for instance, radio-frequency measuring instruments, in fact, it is true to say that some instruments made in this country have to be checked against standards at the U.S. National Bureau of Standards, or the Australian or German equivalents—there being no national reference standard. The writer of an article on the U.K. electronic instrument industry published in *International Commerce* in the States in July, 1964, stated: "British firms look to the United States for example and guidance.... Instrument producing factories often lack standard equipment. Most producers have inadequate or no standards laboratory or environmental facilities."

In the same month that this article appeared (was it coincidental?) the Scientific Instrument Manufacturers' Association set up an eleven-man Working Party under the chairmanship of R. H. C. Foxwell (chairman of Wayne Kerr) whose terms of reference included "the indentification of gaps in the availability of National Standards and the determination of industry's requirements for testing and certification." The Working Party's report in April, 1965, showed the magnitude of the problem which faces this country if it is to bring "its measurement integrity into line with its major competitors." Among its recommendations was the establishment of a national calibration and certification service; and a plan for the setting up of a British Standards Authority to do this was laid before the Government. These proposals have now borne fruit, for the Minister of Technology announced in Parliament on April 25th, that the Government is to establish a British Calibration Service. The actual calibration and certification of instruments will be carried out in existing public or private laboratories.

The National Physical Laboratory will remain responsible for the basic international standards of length, mass, time, electrical current, temperature and luminous intensity. One of the problems, however, will be the maintenance of the accuracy of the "tranfer standards" at the calibrating centres. One can also foresee the need for R. & D. support at Government level for continuously establishing new standards as techniques advance.

The primary need for the setting up of the calibration service is to increase our exports of measuring equipment in which there is an adverse balance of trade. Manufacturers find it increasingly difficult to export to such countries as the U.S.A., Australia, Sweden and Switzerland, who require a certificate of calibration on imported instruments.

Metrology and instrument technology enjoy high academic status in the United States and also in Germany where there are, we believe, eight chairs in metrology and the subject is now recommended as compulsory for all engineering degrees. It is good to learn that the Minister of Technology and the Secretary of State for Education and Science are discussing how the subject can be adequately covered in the curricula of universities and technical colleges but we would suggest that it is essential in a so fast developing science to establish a close partnership between the industry and colleges.

As we go to press, a series of meetings is being held by the National Conference of Standards Laboratories at Gaithersburg, Maryland, concerning the problems facing measurement standards laboratories. The problems in this country are well known; we hope the U.K. delegates will return with some solutions.

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Automatic Gain Control in Transistor Receivers

By K. R. STURLEY,* Ph.D., B.Sc., M.I.E.E.

A FULLY DETAILED DESIGN FOR SEMICONDUCTOR CIRCUITS

THE calculation of the automatic gain control (a.g.c.) characteristics of a valve receiver presents little difficulty because the control voltage is not required to supply current. The situation is more complicated in the transistor receiver because the a.g.c. source must supply current, and the maximum signal which can be accepted by a transistor is so much less than that by a valve. The controlled transistor acts in a manner similar to that of a non-variable-mu valve with short grid base, and some auxiliary form of a.g.c., such as a damping diode, is required before the controlled stage, to limit the input signal and prevent modulation envelope distortion.

The purpose of this article is to show that by making some normally justifiable assumptions it is possible to calculate the a.g.c. characteristics for a transistor receiver. There are two parts to the task; first to calculate the component values and then to determine the a.g.c. characteristics. We will consider the simplest type of a.g.c. circuit for which the control voltage is taken from the detector diode. The circuit in Fig. 1 shows two i.f. stages, the first of which is controlled and has a damping diode connected across the primary of the input transformer fed from the collector of a frequency-changer transistor. The second i.f. stage is not controlled because this would increase its input signal and could lead to modulation envelope distortion.

We will deal first with the conventional a.g.c. which exploits the variable gain characteristic of the first transistor. Since the mutual conductance (g_m) of a transistor is directly proportional to the collector current, we can use the $I_C V_{BE}$ characteristics as a measure of the variation in gain. Maximum and minimum collector currents will be about 1 mA and 30 μA respectively so that a gain control variation of $20\log_{10}(1/0.03)$, i.e. 30 dB, is possible. Typical values for I_C and V_{BE} are shown in Table 1.

A probable value of d.c. current gain β is 100, and the initial curvature of the detector diode will require it to be given a forward bias of 0.3 volt in order to achieve optimum detection of small signals. With these assumptions we can now begin the calculation of component values.

A resistance (R_4) is required in the emitter lead to achieve thermal stability and reduce the effect of transistor tolerances; a suitable value is 470Ω, producing an emitter-earth bias V_{EO} of 0.47 V for $I_C = 1$ mA. (I_E is very nearly equal to I_C when β is large). The base-earth bias $V_{BO} = V_{EO} + V_{BE} = 0.47 + 0.25 = 0.72\ \text{V} - V_{BE}$ is obtained from Table 1. Probable values for R_6 and R_7 in the detector circuit are 470 Ω and 5 kΩ respectively, and the current I_2 in R_2 must be such as to produce a forward bias of 0.3 V across R_7. Hence $I_2 = 0.3/5 \times 10^3 = 0.06$ mA, and this current must produce across R_2 a voltage of $V_{BO} - 0.3 = 0.42$ V. Thus $R_2 = 0.42/0.06 \times 10^{-3} = 7\text{k}\Omega$, the nearest preferred value to which is 8.2 kΩ. The higher value given to R_2 reduces the required value of I_2 if we are to maintain V_{BO} at 0.72 V, but for the moment we will ignore the effect until R_3 has been calculated.

The current through R_3 is the sum of I_2 and I_B, i.e. $0.06 + 0.01 = 0.07$mA, and

$$R_3 = (V_b - V_{BO})/(I_2 + I_B)$$
$$= 8.28/0.07 \times 10^{-3} = 118\ \text{k}\Omega.$$

The nearest preferred value is 120 kΩ, and we must next determine the effect on collector current of using preferred values in the base-bias circuit which will be increased. Let us try

$I_C = 1.1$ mA, $I_B = 11\ \mu$A, $V_{EO} = 0.516$ V, $V_{BE} = 0.2525$V (Table 1) and $V_{BO} \approx 0.77$ V.

$$I_2 = V_{BO}\,(R_2 + R_7) = 0.77/13.32 \times 10^3 = 57.6\ \mu\text{A}$$
$$V_{R1} = 68.6 \times 0.12 = 8.23\ \text{V} = V_b - V_{BO}$$

The transistor will therefore operate at 1.1 mA when the input signal is zero, and the forward bias on the detector is $I_2 R_7 = 0.288$ V.

* Chief Engineer, External Broadcasting, B.B.C.

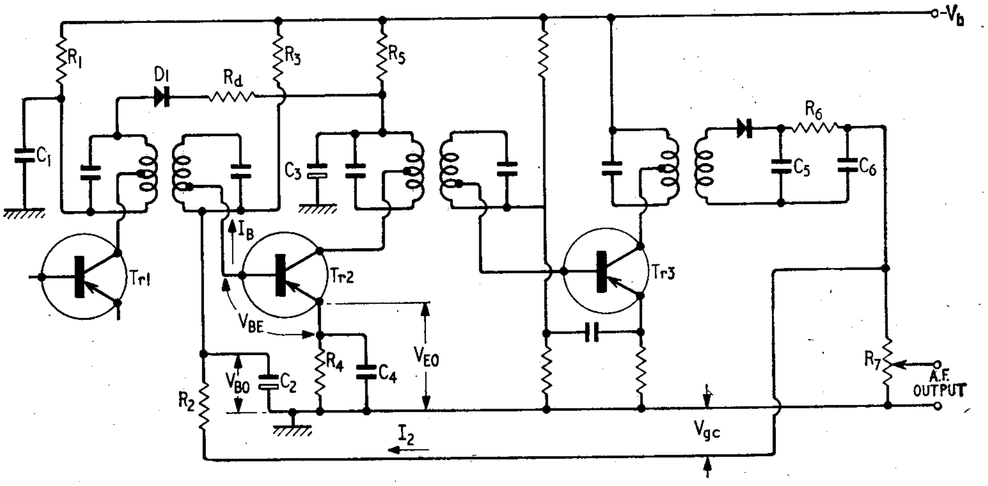


Fig. 1. A.G.C. in a transistor receiver.

Table 1

I_C (mA) ..	1.2	1.1	1	0.9	0.8	0.6	0.4	0.2	0.1	0.05	0.03
V_{BE} (volt) ..	0.255	0.2525	0.25	0.247	0.243	0.235	0.225	0.2	0.17	0.15	0.14

Table 2

I_c (mA)	0.03	0.05	0.1	0.2	0.4	0.6	0.8	0.9	1	1.1
20 log 10 I_c/0.03 (dB)	0	4.4	10.5	17.5	22.5	26	28.5	29.5	30.4	31.3
V_{BE}	0.14	0.15	0.17	0.2	0.225	0.235	.0243	0.247	0.25	0.2525
V_{EO}	0.014	0.0235	0.047	0.094	0.188	0.282	0.376	0.423	0.47	0.516
V_{BO}	0.154	0.1735	0.217	0.294	0.413	0.517	0.619	0.67	0.72	0.77
1.07 V_{BO}	0.164	0.185	0.232	0.314	0.44	0.552	0.66	0.715	0.768	0.825
$8.2 \times 10^3 I_B$ ..	0.0027	0.0041	0.0082	0.0164	0.0328	0.049	0.065	0.0735	0.082	0.09
V_{gc}	0.448	0.426	0.375	0.282	0.142	0.014	−0.11	−0.173	−0.245	−0.3
Output E_o	0.748	0.726	0.675	0.582	0.442	0.314	0.19	0.127	0.055	0
$20 \log_{10} \frac{0.748}{E_o}$ (dB)	0	0.2	0.9	2.2	4.6	7.6	11.9	15.4	22.6	—
Input E_i (dB) ..	0	4.6	11.4	19.7	27.1	33.6	40.4	44.9	53	—

We can now determine the a.g.c. bias (V_{gc}) for any given collector current I_c by using the following voltage-current relationships

$$V_b + V_{gc} = (I_2 + I_B) R_3 + I_2 R_2 \quad \text{.. .. ..} \quad (1)$$

$$I_2R_2 - V_{gc} = V_{BO} \quad \text{.. .. ..} \quad (2)$$

Solving for I_2 in (2)

$$I_2 = (V_{gc} + V_{BO})/R_2$$

and replacing in (1)

$$V_b + V_{gc} = (V_{gc} + V_{BO})(R_2 + R_3)/R_2 + I_BR_3$$

$$V_{gc} = V_b R_2/R_3 - V_{BO}(R_2 + R_3)/R_3 - I_BR_2 \quad (3)$$

Replacing the resistances in (3) by their preferred values given above and assuming a supply voltage of 9 V.

$$V_{gc} = 0.615 - 1.07\, V_{BO} - 8.2 \times 10^3 I_B \quad \text{..} \quad (4)$$

The values of V_{gc} for selected collector currents are given in Table 2, together with the ratio variation (dB) of transistor gain ($20 \log_{10} I_c/0.03$), of output signal (assumed to be the diode forward bias, plus the a.g.c. bias), and of input signal (sum of transistor gain and output signal variations).

When calculating the output signal, no allowance has been made for the curvature of the detection characteristic; this will tend to have a greater effect than in a valve receiver because the maximum signal is so much less, but even so, its influence at the point where a.g.c. is beginning to operate is not very considerable.

The a.g.c. characteristic, represented by the last two rows in Table 2, is plotted as curve 1 in Fig. 2. If the minimum current of the transistor is reduced below 30 μA, modulation envelope distortion begins to be appreciable, and the a.g.c. characteristic turns up, as shown by the dotted extension of curve 1.

Some idea of the signal voltages prevailing at various parts of the circuit can be gained as follows: The effective power gain of each i.f. transistor will be of the order of 30 dB, so that for a detector load of 2.73 kΩ [$\frac{1}{2}(R_6 + R_7)$], and a transistor input conductance of 1250 μmho, we have power in load.

$$P_o = \frac{E_o^2}{2.73 \times 10^3} = 10^3 \times E_1^2 \times 1250 \times 10^{-6}$$

$$\therefore \frac{E_o}{E_1} = (2.73 \times 1250)^{\frac{1}{2}}$$

$$= 58.5 \approx 60 = \text{voltage gain of } T_r3.$$

The voltage gain from base of Tr2 to base of Tr3, assuming 30 dB power gain and equal base conductances, is approximately 32. The detector output with no a.g.c. is about 50 mV giving an input voltage to Tr2 of 50/60 × 32 = 26 μV. With maximum a.g.c., the gain of Tr2 falls nearly to unity, the output voltage from the detector is 0.75 V, and the input to Tr2 base is 0.75/60 = 12.5mV. This represents about the maximum permissible carrier voltage that can be accepted at $I_c = 30\ \mu$A, and a damping diode is needed in order to achieve a wider range of a.g.c. without exceeding an input carrier of 12.5 mV.

The damping diode, D_1 in Fig. 1, provides a shunt load in the collector circuit of Tr1 to reduce the gain of this stage and prevent overload of succeeding stages. It has the secondary effect of reducing the selectivity of the i.f. transformer, across which it is connected, as the input signal increases. Fortunately this is no disadvantage—it may even be an advantage—because a strong signal will tend to suppress a weaker adjacent channel.

The a.g.c. action of the diode is quite easily calculated since the diode acts as a resistance in parallel with the transistor a.c. resistance, and the load resistance presented by the primary of the i.f. transformer. If the two latter are represented by R_o and r_d is the effective a.c. resistance of the diode, the gain of Tr1 is changed from g_cR_o to $g_cR_or_d/(R_o + r_d)$, and the attenuation due to the diode

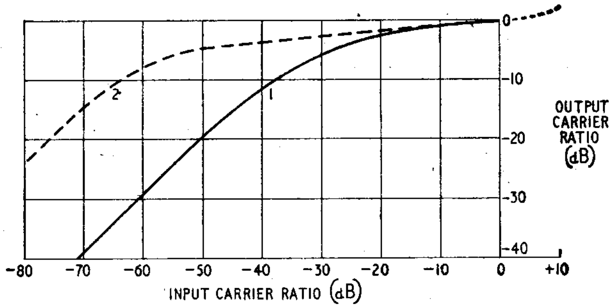


Fig. 2. A.G.C. characteristics of a transistor receiver. Curve 1. Without damping diode. Curve 2. With damping diode.

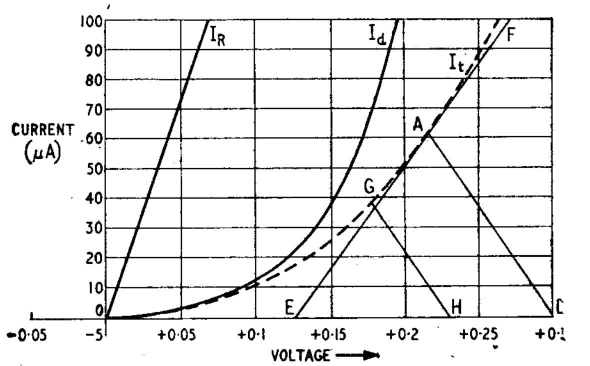


Fig. 3. IV characteristics of the damping diode.

Table 3

I_c (mA)	0.03	0.05	0.1	0.2	0.25	0.3	0.4	0.5	0.6	0.8	0.9	1
$V_{r1} - V_{r5}$	0.3	0.28	.23	0.13	0.08	0.03	−0.07	−0.17	0.27	−0.47	−0.57	−0.67
rd (kΩ)	1.45	1.55	1.9	3.9	6.8	14.5	100	750	—	—	—	—
$20 \log_{10} \frac{(rd + Ro)}{rd}$ (dB)	25.4	24.8	23.2	17.6	13.5	8.8	2	0.3	—	—	—	—
Gain (dB)	0	0.6	2.2	7.8	11.9	16.6	23.4	25.1	25.2	25.3	25.4	25.4
Output (dB)	0	0.2	0.9	2.2	—	—	4.6	—	7.6	11.9	15.4	22.6
Input (dB)	0	5.2	13.6	27.5	—	--	50.5	—	58.8	65.7	70,2	78.3

is $20 \log_{10} (R_o + r_d)/r_d$. Thus if $R_o = 25.5$ kΩ and $r_d = 25.5$ kΩ there is a loss of 6 dB. We have already noted that the maximum signal applied to the base of Tr2 is about 12.5 mV and this is stepped up to about 125 mV

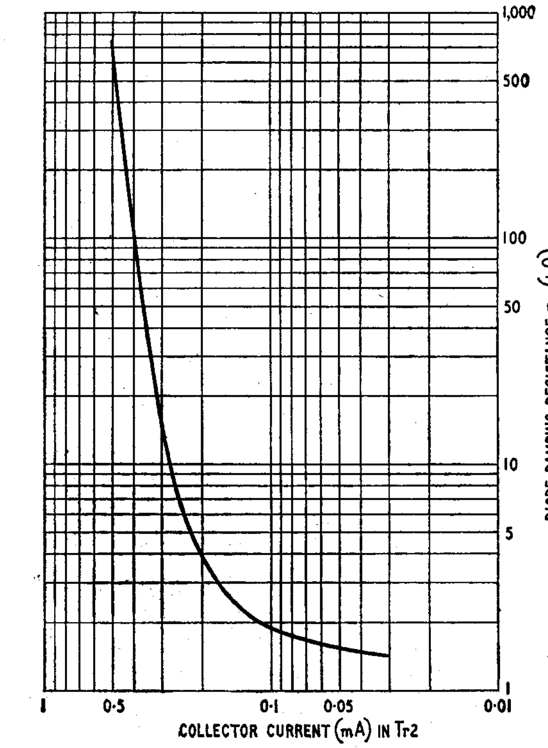


Fig. 4. Change of diode damping with in collector current of Tr2.

across the primary of the i.f. transformer. This is quite a small signal and the diode will tend to be continuously conducting with its a.c. resistance producing the damping. Fig. 3, curve I_d shows a typical *IV* characteristic of a damping diode over the range of bias voltages likely to be used. The relationship is non-linear and reverse current flows so that there is some slight damping with reverse bias. Since the damping is non-linear a degree of modulation envelope distortion occurs and is generally a maximum at a particular bias value. The non-linearity effect can be reduced by adding a resistance (R_d in Fig. 1 about 680 Ω) in series with the diode. Some sacrifice of a.g.c. action occurs because the maximum slope of the combined characteristic, I_t in Fig. 3, is reduced. The curve I_t is the sum of the voltages at given current from the diode curve I_d and the resistance line I_R.

The decoupling resistance R_5 (1 kΩ) in the collector of the controlled transistor Tr2 provides the control bias voltage, and reverse biases the diode to about 1 V at maximum I_c, and R_1 (330 Ω) forward biases to 0.33 V if the collector current of Tr1 is 1 mA. The diode is forward biased when the collector current of Tr2 falls below 0.3 mA.

The resistance values of R_1 and R_3 as well as the voltages due to the collector current of Tr1 and Tr2, determine the actual bias applied to the diode. The actual bias is found by drawing a resistance line of inverse slope equal to $R_1 + R_5$ (1.33 kΩ) from a voltage equal to the difference across R_1 and R_5 due to the collector currents. Thus if the collector current of Tr2 is 30 μA, the voltage across R_5 due to this is 0.03 V (reverse) and if that across R_1 is 0.33 V forward, the true bias is given by the intersection A (Fig. 3) of the combined I_t characteristic with the resistance line *AD* of 1.33 kΩ drawn from a forward bias of 0.3 V.

Fig. 3 shows that the true bias is 0.215 V forward. The inverse slope of the combined characteristic at *A* is given by the tangent *EF*, and is about 1.45 kΩ; this is the value of the damping resistance across the i.f. transformer primary.

This procedure can be repeated for selected values of collector current in Tr2; line *GH* gives the result for $I_c = 0.1$ mA, i.e. a forward bias of 0.23 V. The damping resistance of the diode at various collector currents in Tr2 is plotted in Fig. 4. Using the diode attenuation expression $20 \log_{10} (r_d + R_o)/r_d$ the loss due to the diode damping may be calculated; the reference is a collector current of 30 μA. These gain variations must be added to the input variations in Table 2 to obtain the overall input variations (see last column in Table 3). This has been done in Table 3 above on the basis of $R_o = 25.5$ kΩ.

The overall a.g.c. characteristic is shown by curve 2 in Fig. 2 and the damping diode is seen to have extended the a.g.c. range by about 25 dB, giving an input variation of about 56 dB for a 6 dB change of output. No account has been taken in curves 1 and 2 of the decrease of input conductance of Tr2 with decrease of collector current. The effect is small but it does reduce slightly the effectiveness of the a.g.c.

The biasing resistances R_1 and R_5 should have large capacitors C_1 and C_3 (about 8 μF) so that the a.f. voltage components due to rectification of the i.f. signal by the non-linear diode damping characteristic are negligible and will therefore have little effect on the bias applied to D_1.

Dr. K. R. Sturley, *has been chief engineer, external broadcasting, in the B.B.C. for the past three years having joined the Corporation in 1945 as head of the engineering training department. He graduated from Birmingham University and did postgraduate research on electrothermal storage problems which lead to his Ph.D. In 1936 he joined the staff of Marconi College, Chelmsford, as lecturer and was assistant principal when he left to join the B.B.C.*

A LOOK AT SOME OF THE EXHIBITS AT THE AUDIO FAIR AND AT THE CONCURRENT AMERICAN SHOW

AUDIO 1966

THE Audio Festival and Fair in London held a few surprises for the unsuspecting visitor—for a number of well known manufacturers had diversified their interests. Goodmans introduced an integrated stereo amplifier, Leak have come up with a pickup arm, Sonotone (Technical Ceramics) have produced a loudspeaker and enclosure and there has been a surge in the number of small enclosures on the market, such now being available from Goodmans, Richard Allan, Truvox, Braun, Saba and Pioneer (Japan). Many new amplifiers, tape recorders, microphones, pickup arms, turntables and loudspeakers were seen and it is not proposed to deal with them all since space does not allow.

This year many visitors to the Fair could commit "monolithic dual avicide"* by visiting the show of some American equipment at the U.S. Trade Center, which enabled visitors to draw their own comparisons between British and American equipment. From the viewpoint of performance, the importance attached to the various aspects is different, making direct comparison difficult, but the appearance of the U.S. equipment was more professional and business-like. The attention and assistance given to the visitor was also more professional and business-like; the performance of many staffing the stands and demonstration rooms at the Audio Fair seems to leave something to be desired.

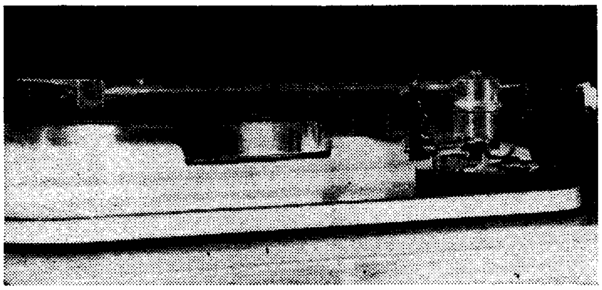
Pickup arms

Three of the new arms shown were illustrated in the preview (Leak, Goldring and the Japanese Micro). The Leak arm is noticeably unconventional in design—the unipivot is at disc height, the intention being to reduce the effects of warped records. The use of a single pivot leads to the low bearing friction which is given as 10 mg. The cartridge is a variable reluctance type fitted with all elliptical stylus (to minimize tracing distortion) and gives an output of 1.2 mV per cm/sec. Leak were one of the few manufacturers to quote the tip mass on their literature—which in this case is less than 1 mg.

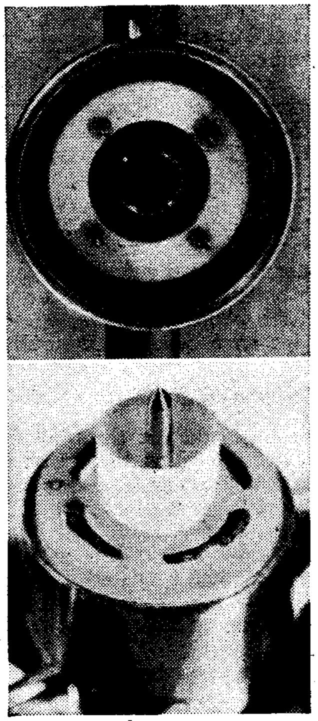
The Japanese Micro Seiki arm, imported by Medley, and exhibited jointly by Medley and Living Sound, has a low pivot friction and reference to the Fair preview shows the arm has some similarity to the SME arm.

An interesting arm was exhibited by Audio & Design (Fig. 1). The unipivot arrangement, which was slightly misrepresented in the preview, consists of a hardened needle upon which the arm with a miniature ball race rests. The top hub and pivot are illustrated in Fig. 2 and damping, to avoid torsional resonance, may be achieved by use of the cup around the needle. The pivot friction is not measurable, the abolition of the lead out wires contributing to this. Electrical connection is made with four nickel-plated electrodes which rest in four mercury baths when the arm is placed on its pivot. Two other features are the closeness of the counterweight to the pivot, giving a low moment of inertia, and the magnetic side-thrust compensation. The arm was designed for minimum distortion due to tracking error, rather than minimum tracking error†. The arm is undoubtedly a high-quality precision instrument and deserves its title of a "laboratory" pickup arm.

* According to a *Times* critic, killing two birds with one stone!



Above: Fig. 1. Pickup arm by Audio & Design. The pivot arrangement is shown in Fig. 2.



Right: Fig. 2. Pivot assembly of the arm of Fig. 1 showing needle, damping cup, mercury baths (bottom), ball-race cup and electrodes (top).

Semiconductor cartridge

At the business end of arms, perhaps the most significant development in recent years has been the introduction of a cartridge using a semiconducting element as a pressure sensitive transducer. The cartridge on show at the Fair was developed by the Euphonics Corpn. (U.S.A.)—represented in the U.K. by A. C. Farnell. The construction of an element of a stereo pickup is shown in Fig. 3. The silicon element is either compressed or elongated by movement in the plane shown and responds by changing its conductivity. A "bias" current must be passed through the elements so that this becomes modulated, the output voltage being taken across the cartridge. The output impedance is around 800 Ω and output voltage (no load) can be between 12 and 40 mV depending on the voltage supply available and the series resistor. With a series resistor of 2 kΩ,

† Such an approach is outlined by J. K. Stevenson in this and the May issue.

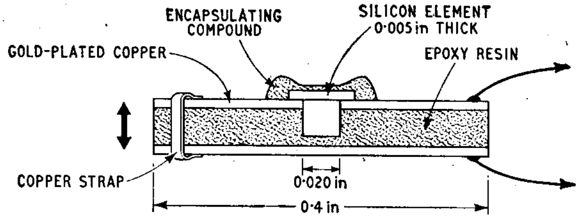


Fig. 3. One element of the Euphonics semiconductor cartridge.

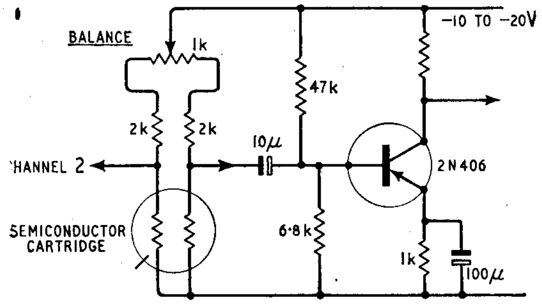


Fig. 4. Typical input circuit and a convenient method of balancing for the semiconductor cartridge.

a supply voltage of 20 V and feeding a load of 2 kΩ the output will be about 18 mV with a signal-to-noise ratio of better than 80 dB. For reproduction from an R.I.A.A. recording no external equalization is necessary, this being achieved mechanically. Some low frequency attenuation may be necessary with the cartridge since the output extends down to d.c. A convenient method of balancing is shown in Fig. 4.

Four versions are available, two having elliptical styli (0.0002 and 0.0009 in radii) and a high vertical and horizontal compliance of 25×10^{-6} cm dyne^{-1} (U-15-LS) and two having styli with spherical tips and a compliance of 15×10^{-6} cm dyne^{-1} (U-15-P)—the prices differ by about £10. Separation at 15 kc/s is 10 dB for the U-15-LS; 5 dB for the cheaper model and 25 dB at 1 kc/s for both models. The cartridges are available for standard mounting or in plug-in heads for use with a Euphonic TA-15 arm. An effective tip mass of between 0.3 and 0.6 mg can be achieved. (It was noted that on some of the manufacturer's literature, the necessity for a phase inverter for one channel was stressed, but on other literature no mention was made of this. This would be explained if the cartridge and supply unit have been modified since introduction.)

Silicon transistor amplifier

Probably the most notable development in the amplifier field is the introduction of a silicon transistor amplifier by Goodmans. The amplifier is known as the Maxamp 30 and is the physical companion to the Maxim loudspeaker enclosure. The photograph (Fig. 5) shows the two sides with printed boards hinged for easy servicing. The amplifier gives 15 W per channel into an 8 Ω load or 10 W per channel into a 4 or 15 Ω load. With an 8 Ω load and at full power the distortion is given as 0.4% at 1 kc/s.

The power amplifier circuitry is similar to that of the Tobey and Dinsdale, and Goodmans are by no means alone in adopting this configuration. The use of silicon n-p-n transistors (and consequently the positive supply rail) is the main difference. The preamplifier circuit is also similar, using feedback around the first two stages to increase the input impedance. (These are indirectly coupled, incidentally.) The Baxandall circuit is also used for the tone controls but the filter circuits differ —an *LC* type being used for the low pass filter giving a slope of 12 dB/oct with an 8 kc/s turnover frequency. An input for ceramic cartridges is provided with a sensivity of 50 mV and an input impedance of 100 kΩ. One difference throughout is that most of the resistors in the Dinsdale amplifier are 5% tolerance whereas 10% types are used in the Goodmans amplifier and more critical resistors are 2% and 5% respectively.

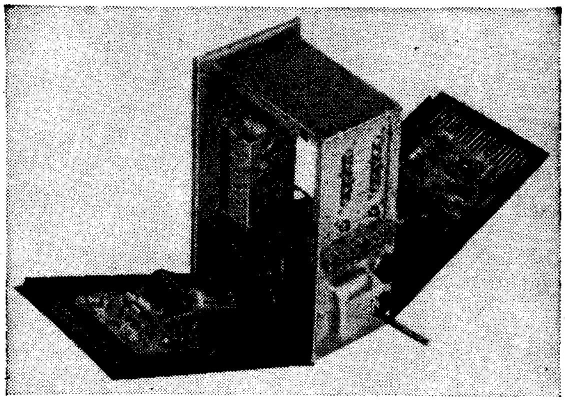


Fig. 5. Hinged panels for easy servicing on the Goodmans silicon transistor Maxamp.

Mullard took a back seat this year (for most visitors) in a trade-only room, but Ferranti were out in front and suggested circuits for items of audio equipment using silicon transistors. Literature is available describing 7 W and 15 W amplifiers, and the 7 W version has been described in the article on the silicon transistor tape recorder*.

A circuit of a silicon transistor f.m. tuner was shown which had a novel form of tuning indicator. The circuit of this is shown in Fig. 6 and two gallium phosphide lamps are used in a similar manner to the two neon

* July and August 1965 issues.

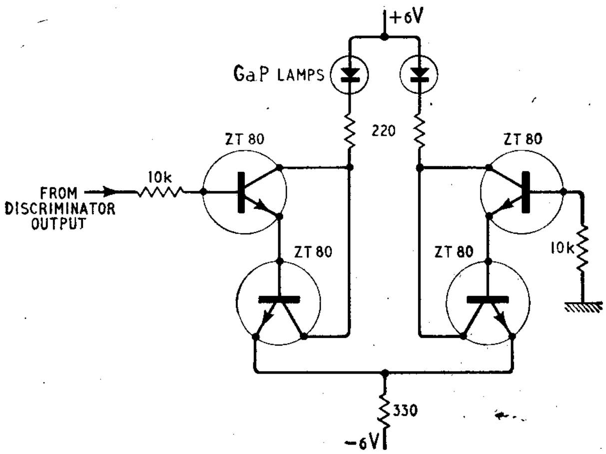


Fig. 6. Novel tuning indicator demonstrating the possibilities of gallium phosphide semiconductor light sources.

lamps of the Quad tuner.† This circuit, though, is obviously unecomonic, and its purpose is merely to demonstrate the potential of semiconductor light sources.

Titanium cone loudspeaker

Development in loudspeakers since last year appears to have been largely a matter of small improvements obtained by the use of materials new to drive unit construction. Neoprene roll surrounds are being adopted more widely for front suspension of cones, for example (in the Mk 2 version of the B139 bass unit made by KEF), as they give a more linear suspension and consequently a reduction in waveform distortion. E. J. Jordan's work in developing wide-range drive units using small metal cones is well known, and the latest phase, shown by Audio & Design, is the use of 4-in titanium alloy cones with plastics/metal laminate surrounds. The high strength/weight ratio of titanium allows the cone to be thin and light but strong. Because of this and the high velocity of sound in titanium, cone break-up starts at a higher frequency than with other materials, and the behaviour of the cone as a transmission line is more predictable for design purposes. The laminate cone surround—a plastic material coated with metal—has been adopted to cope with the conflicting requirements of acoustic sealing, flexibility, cone centring and high-frequency termination, and it has an annular structure giving a "negative" stiffness which tends to cancel the stiffness of the cone and the rest of its suspension. The -3dB frequency response of the D30/20 loudspeaker using these materials is stated to be 20 c/s to 22 kc/s. Maximum output is 115 dB relative to 0002 dyne cm^{-2}.

A new name amongst the loudspeaker exhibitors was Technical Ceramics Ltd. who demonstrated the Sonotone "Solent," incorporating a 6-in bass unit and a 3$\frac{3}{8}$-in tweeter in a 14×9×8 in cabinet. (The frequency response is shown on our front cover.)

Some American equipment

One of the components in a transistor power amplifier which is often thought of as undesirable is the loudspeaker coupling capacitor. The obvious way to eliminate this is to use a power supply with +ve and −ve rails. But the capacitor does not quite disappear; it is in effect moved to the power supply, since generally extra smoothing will be required. This method is, however, adopted in an amplifier manufactured by Lansing and a simplified circuit of the output stage is shown in Fig. 7. Another Lansing product is a stereo pre-amplifier ("graphic controller") and it can be seen that the appearance is somewhat unconventional (Fig. 8), the rotary controls being replaced by sliders and illuminated push-buttons. Other controls appear behind a drop-door, a practice common with TV receivers at one time. A 1-kc/s oscillator provides a test tone for balancing stereo channels.

Both in the U.K. and in the U.S.A. the last strongholds where valve amplifiers and tuners have remained supreme are beginning to fall. Quad have now introduced a 50 W transistor amplifier and McIntosh (U.S.A.) are producing transistor pre-amplifiers but are still clinging to valve power amplifiers and tuners. The amplifiers use an output stage similar to that of the Quad—the primary of the output transformer forming part of both cathode and anode loads.

The MR 67 and MR 71 McIntosh stereo tuners feature

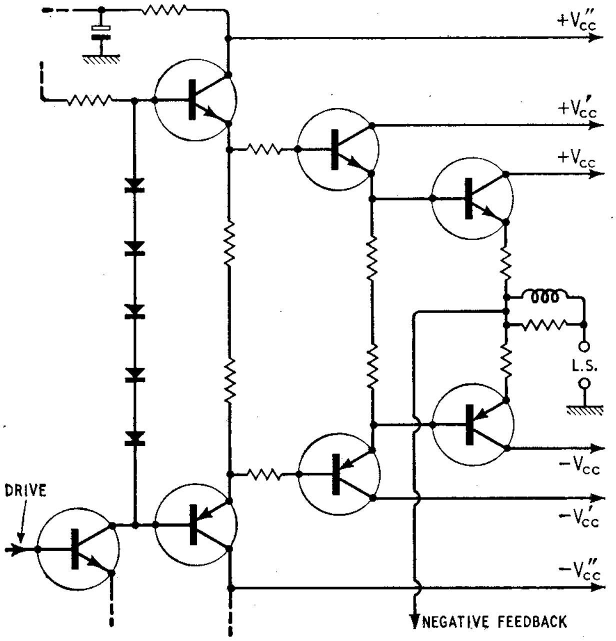


Fig. 7. Elimination of the loudspeaker coupling capacitor by using + ve and − ve supply rails.

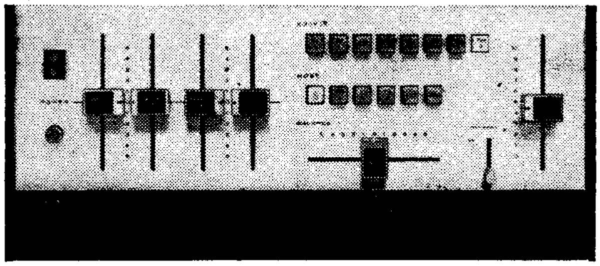


Fig. 8. The unconventional appearance of an American stereo pre-amplifier eliminating rotary controls (Lansing).

multipath indicators as well as the usual multiplex indicators. The MR 71 in fact has four indicators—apart from the frequency scale—the other two being tuning and signal strength meters. On the MR 67 the signal strength and multipath indicators are combined and switched. The indicator is fed from a limiting i.f. stage and signal variations due to multipath reception causes the indicator to fluctuate. The aerial is then adjusted for minimum variation. Signal strength is registered by switching in a shunt capacitor to smooth out the variations.

Marantz introduced a combined multipath and tuning indicator using a 3 in cathode ray tube some time ago and this is featured on their 10B stereo tuner. The ordinate represents signal strength, the signal being derived from a limiter, and the abscissa represents deviation, the signal being taken from the discriminator. With this method of indication, stereo balance, separation and phase may also be visually represented. The valve tuner incorporates 34 diodes and a number of neon-l.d.r. switches for muting and automatic stereo/mono switching.

The Marantz turntable unit was also seen—using radial tracking in order to give zero tracking error and side thrust.

† September 1955 issue, p. 428.

Matching the C.R.T. Display to the Viewer

By D. W. KAHAN, M.A., B.Sc.

There is no point in straining to achieve optimum electronic design in a television receiver if the picture display device does not give optimum transmission of visual information to the viewer in his normal environment. This article shows how the problem is tackled in the design and correct operation of cathode-ray tubes

MANY of the television receivers now coming into the shops have much darker-looking screens than has been usual. The effect of this is to alter the appearance of the receivers and the contrast in their pictures, for the better in both cases—although this is perhaps a matter of opinion as far as the appearance of the cabinet is concerned.

A good television picture is required to reproduce the original scene with adequate contrast between the light and dark parts of the picture. The idea of "black" in a real scene or in a picture is not an absolute one, and in different circumstances the level measured as the "black" may be very different.[1] This is a difficulty with all methods of displaying pictures where it is necessary to reproduce the appearance of a scene originally at a quite different level of luminance. Fortunately, it is not usually necessary for each part of a reproduced picture

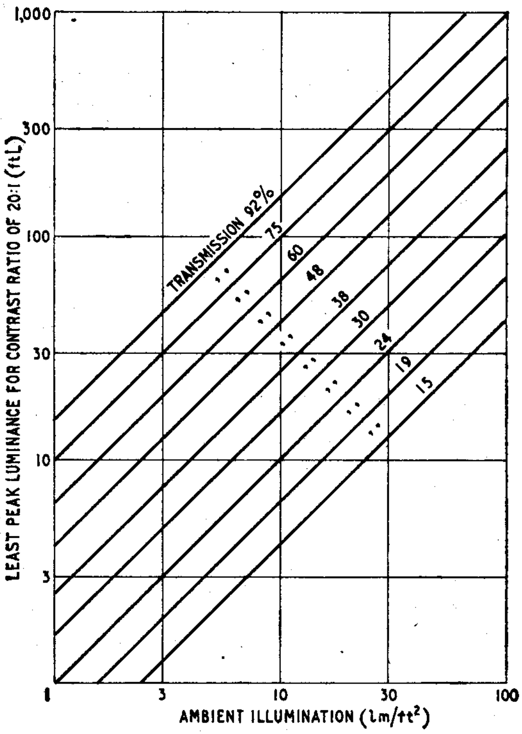


Fig. 1. Luminance needed on c.r.t. screens for 20 : 1 contrast with different levels of illumination and faceplate transmissions.

David W. Kahan, after taking the Natural Sciences Tripos at Downing College, Cambridge, spent two years with Ultra working on the design of radio receivers. He then studied physics at Birkbeck College, London, where he received his B.Sc. He subsequently joined the applications laboratory of Edison Swan Electric Company (now Thorn-AEI Radio Valves & Tubes) where he has since been concerned with the development of cathode-ray tubes. He is 39.

to be identical in brightness (a subjective quantity) to the original for an acceptable picture to be obtained.

Outdoor scenes will often include luminance levels differing by many hundred times—perhaps by thousands if the view is from a window and some of the interior can be seen at the same time. The eye is able to make use of the whole of this range, but quite good pictures can be produced with less. For example[2] the range in projected colour transparencies is at best 125 : 1, and reflection prints give a range of rather less than 35 : 1.

In television studios and when making films intended for television, it is understood that the range of luminance in a picture should be quite limited and that the larger ratios that are sometimes unavoidable will result in the scale of tonal values being compressed at the ends. The latest versions of the television Test Cards D and E include a column of five squares, intended for checking the grey scale on the receiver, in which the range of contrast is 30 : 1. A slightly different value is used for films for television, where a maximum contrast (transmission) range of 40 : 1 is suggested,[3] and this is considerably exceeded in films which may be shown on television although originally intended for the cinema. However, the range of ratios that can be reproduced well is likely to be between 20 : 1 and 50 : 1.

The darkest parts of a television screen are never totally black, since they are illuminated by light scattered from the brighter areas of the picture and by the general lighting of the surroundings. As the screen itself is made of a mixture of phosphors whose crystals are themselves fairly highly reflecting, the level of "black" in the picture depends on the illumination reaching the screen. When only small areas of the picture are bright so that the ambient illumination determines that on the screen, the luminance of the dark parts L is set by the illumination E, the reflectance of the screen material P and the fraction of the light transmitted by the glass T. The incident light

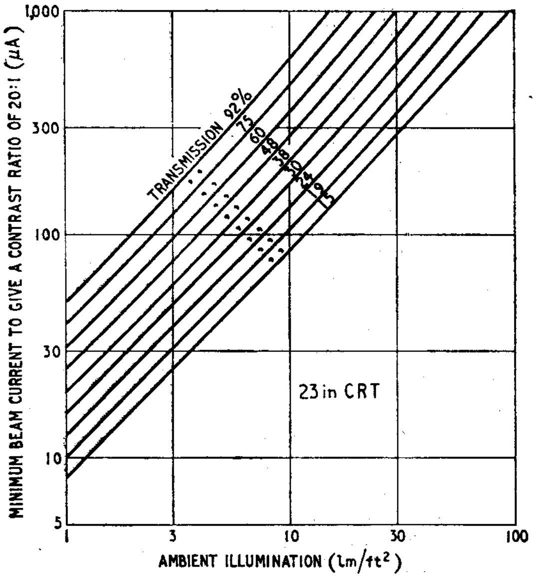


Fig. 2. Beam currents needed to give highlight levels in Fig. 1 on 23-inch c.r.ts with faceplate transmission shown.

traverses the glass and is attenuated twice, before and after its reflection, so that $L = EPT^2$. If a particular ratio of contrast $C : 1$ between the luminance levels of the lightest and darkest parts of the picture is required, the highlights must be seen at a luminance of $CL = CEPT^2$. We can therefore say what the luminance of the highlights must be for the picture to be a satisfactory one, but only if the ambient illumination is known.

There are a few special cases where this calculation gives a more definite result. If the surroundings are well lit, say between 10 and 50 lm/ft², it may be desirable to see the picture at levels the same as if the scene had been in the room; this might be useful for monitor screens in television studios which could be set up so that the studio scene and its television image could be seen at the same luminance and under the same conditions of adaptation of the eye. In studios where it is the practice to adjust the overall gamma of the system to unity, this would give similar luminance and subjective brightness for the scene and the picture at all levels of luminance. In this case, the contrast ratio is known and the transmission of the glass can be set.

A second case is where the surroundings are less brightly lit and can be thought of as a background to the picture. There is very little published information on the way in which ordinary television viewers set up their pictures, although it is fairly certain that there are large variations between them. In an account[4] of experiments in which people could choose the luminance and width of an illuminated frame surrounding a picture, it seemed that the frame was preferred with luminance levels at or below the mean level of the picture. This is what one might expect from the accepted practice in visual measurement work where it is usual to arrange a surround at the mean level of the photometric field of view, for the most comfortable conditions for the observer.

It is reasonable to suppose that a similar preference applies if the ambient illumination is fixed, and that the mean luminance of the picture should be set at least to that of its surroundings. The ratio of peak to mean luminance on the screen varies from one picture to another, and values between 3 : 1 and 5 : 1 are fairly typical. In a recent report[5] describing the measurements taken during two complete evening television transmissions, this ratio was found to be about 2.5 : 1 for the average video voltage signal. This corresponds to a luminance ratio of 10 : 1 when we allow for the non-linear relationship between the drive applied to the c.r.t. and the luminance produced on the screen. The best filter can now be specified for any particular ratio of contrast required in the picture.

In general, if the phosphor screen on clear glass has a reflectance P and the general surroundings of the tube have an average reflectance of S, the ratio of peak (highlight) to mean luminance is R and contrast ratio C, the required filter transmission factor T is:—

$$T = 0.92\sqrt{\frac{R.S}{C.P}}$$

(The figure of 0.92 appears because a " clear " glass loses 4% of light at each face by reflection.) The table below shows some of the values of transmission calculated for a possible case where the screen and surrounding reflectance are equal.

Ratio of highlight to surround luminance	1 : 1	3 : 1	5 : 1	10 : 1
Filter transmission for contrast of 20 : 1	21%	36%	47%	67%
Filter transmission for contrast of 50 : 1	12%	23%	27%	38%

The graphs in Fig. 1 show the luminance needed for the highlights in a picture to have a contrast of 20 : 1 at any particular level of illumination, for a range of transmission values between that of clear glass and 15%, at intervals of 20%. These have been extended to show the peak currents needed to give these highlight levels on modern 23-inch screens, if made with faceplates having these transmissions. Fig. 2 shows the least current needed in the electron beam for a tube run at 16 kV to give a contrast range of 20 : 1, and Fig. 3 shows the similar characteristics for the higher range of 50 : 1. The graphs show how tubes used in well-lit surroundings can be operated at lower beam currents to give pictures with satisfactory contrast if the filter is made with lower transmission. This is an advantage for two reasons; the current available from the usual high-voltage supply derived from the line timebase is quite limited, and it is not always appreciated

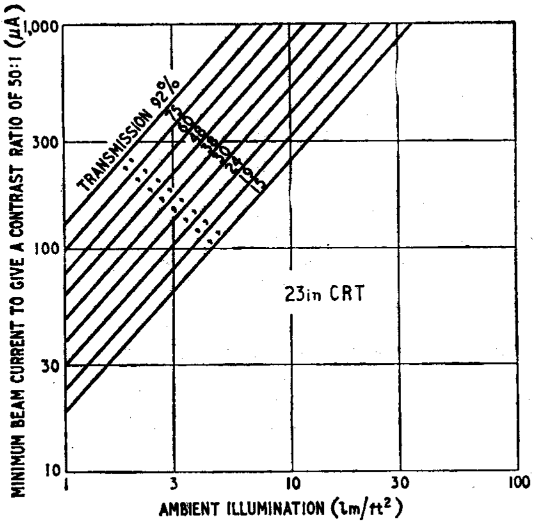


Fig. 3. Beam currents needed for 50: 1 contrast with different levels of illumination and faceplate transmissions.

that the focus characteristics of all c.r. tubes begin to deteriorate as the beam current increases.

There is a limit to the absorption of light in a filter that can be used in this way, since the reflections from the outside surface cannot be ignored entirely. These reflections will not normally be less than 4%, whether the surface is polished or if it is slightly roughened to reduce the specular reflection at the cost of a corresponding increase in the diffuse reflectance. As a result, there is little point in reducing the filter transmission below about 20%, when the luminance of the unexcited parts of the phosphor seen through the filter equals that of the outer surface. The use of surface coatings of the sort used to reduce reflections in camera lenses would allow lower transmissions to be used in special cases where the extra expense could be justified. The obvious idea of using phosphors whose surfaces are themselves black has not proved to be practicable, since such screens are found to be of such low efficiency that the beam current required is too high for the c.r. tubes to be run satisfactorily.

In particular cases it may be possible to avoid the effect of the surface reflections by making sure that lamps and windows are not seen by direct reflection in the surface and that the wall opposite the screen is a dark one. These effects are less significant as the transmission increases, so that for 48% a contrast ratio of 20 : 1 is reduced to 17 : 1 by the surface reflections of light from the walls of the room.

When a separate filter is used with a tube faceplate of relatively clear glass, the extra pair of surfaces between the filter and the screen will also give reflections from the bright parts of the picture, tending to reduce the contrast. It is better to use a filter optically bonded to the faceplate (as in the twin-panel construction) or to use a darker glass in manufacturing the cathode-ray tube itself. This will also help to reduce the effect of light reflected within the glass from the brighter parts of the picture on to otherwise dark parts of the screen.

Acknowledgement.—The author wishes to thank the management of Thorn-AEI Radio Valves & Tubes Ltd., for permission to publish this article, which is based on work done in the Applications Laboratory.

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Hill-climbing in Control Systems

By K. C. Ng*, Ph.D., B.Sc.

In the past few years various techniques have been developed for automatic adjustment of system parameters to obtain the best possible performance from a system—a procedure known as " self-optimization." Few of these techniques have been put into practice but the idea looks promising. The Warren Spring Laboratory of the Ministry of Technology, for example, are working on a system, using electronic digital computing techniques, for achieving maximum economy in the operation of a chemical manufacturing process by automatically adjusting a number of the process variables. This article explains the electronic principles of a method known as " hill-climbing " widely used in self-optimizing systems

AUTOMATIC control is introduced into a system partially or completely to replace the human operator in the control of a complicated process. Simple control systems can be designed, using conventional synthesis procedures, to perform the task to within specified accuracy. Large complex systems are not so easily designed. The design may be further complicated by lack of knowledge of the exact behaviour of system components or by the presence of non-linearities. The system may be working under environmental conditions which vary in a completely unknown manner.

To understand the problem more fully, consider the radar tracking of a target. If the target is assumed to be moving at uniform velocity, the control mechanism can be designed to track the target accurately. Having " locked on " to the target and measured its velocity, and assuming the wind velocity is known, the system can execute the positioning and firing of the gun. The problem is obviously grossly simplified here. Conditions are never ideal; the aircraft may be taking evasive action; atmospheric disturbances will affect the tracking operation; the wind velocity may be vastly different at varying altitudes. If such disturbances are known *a priori* or are measurable, they can usually be taken into account in the design stage. This is obviously not possible in this example. One must therefore resort to more sophisticated schemes to obtain good performance. The homing missile is one approach to this problem.

This much simplified example is one of many practical cases where the system is working under conditions varying in an unpredictable manner, or under unknown conditions. The variations or disturbances may be external to the system or may arise from within the system—for example, change in gain of a valve in an

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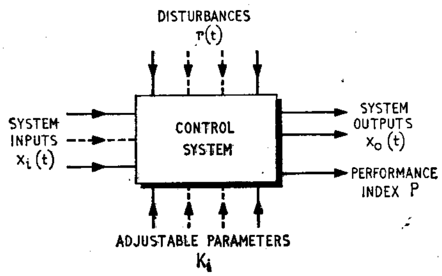


Fig. 1. A general control system with a performance measure, P.

amplifier. In most cases, these variations are not only unpredictable but they cannot be measured directly. Here an indirect method must be used in which the effect of these disturbances on a suitably chosen quantity in the system is measured and appropriate corrective action is taken. If this quantity is used to indicate the "goodness" of the system, then it is obviously desirable that the system should function with the best or optimum value of this quantity at all times, despite variations in system operating conditions.

The "hill-climbing" technique.—Consider the general control system shown in Fig. 1. The variables $x_i(t)$ and $x_o(t)$ are the normal input and output of the system. The signals $r_n(t)$ are unknown disturbance signals. P is a measure of the performance of the system. This may be the accuracy of control, efficiency or profit. There may be practical limitations on the maximum control power available or restrictions on the quality of the product. The different and possibly conflicting requirements, e.g. maximum profit with a minimum guaranteed product quality, can normally be combined into a single figure-of-merit or performance measure.

In general P will be a function of the system variables, that is the system inputs, the disturbances and the settings of the parameters K_i of the system, some of which are adjustable. The performance can be controlled by adjusting these parameters, and a maximum value of P can usually be found for a particular setting of the parameters. A typical relationship between performance and parameter is shown in Fig. 2, curve A, where it is assumed that only one parameter is adjustable. The shape and height of the "hill" depends on the system inputs and disturbances. As the system operating conditions change, the characteristic may change to the curve B. Provided K can be adjusted to be at the optimum

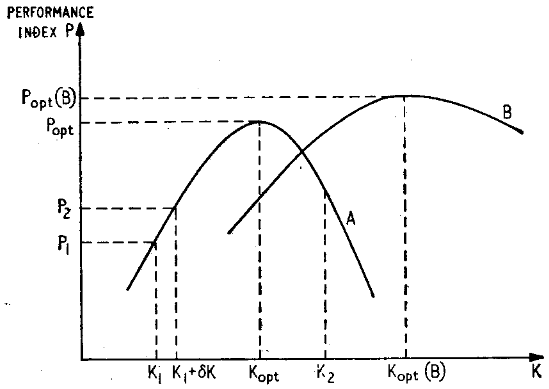


Fig. 2. Performance index vs. parameter characteristic.

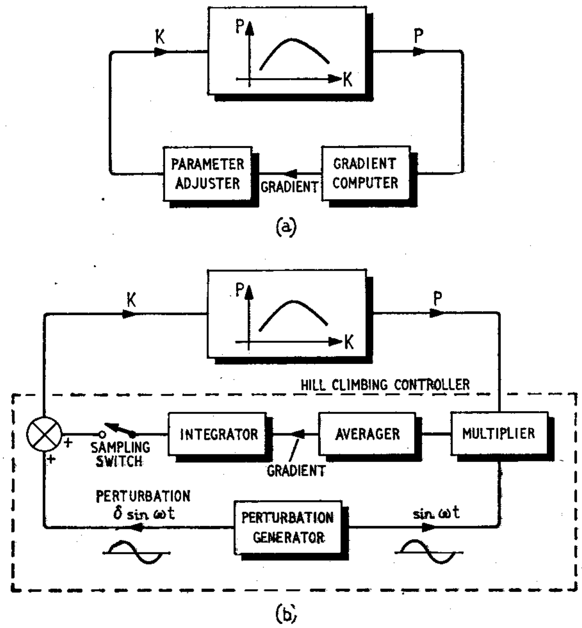


Fig. 3. (a) Basic hill-climbing system; (b) hill-climbing system with sinusoidal perturbation

setting as the characteristic curve moves, then peak performance can be maintained.

Assume that K is at the non-optimum setting K_1. One way of adjusting K to K_{opt} is to measure the slope or gradient of the performance characteristic at K_1 and then, using this information, adjust K in the direction which improves the performance. The gradient at the working point may be obtained by trial-and-error method. The parameter is displaced by a small trial step δK and the corresponding change in performance measure, $\delta P = P_2 - P_1$, is observed. If increasing K by δK improves the performance, as at K_1, then K must be adjusted in this direction. If the change in P is negative (point K_2), then the parameter setting must be decreased. The process is repeated at each new parameter setting. The technique thus involves determining the gradient of the hill and climbing up the gradient to the peak: hence the term "hill-climbing."

The same method can be applied if the object is to minimize the cost of operating a plant. In this case the characteristic is a valley or trough and the parameter is adjusted in the direction of negative gradient.

Practical system.—In practice the hill-climbing technique is made automatic by perturbing the parameter with a periodic signal of amplitude δK. The gradient is obtained by phase-sensitive rectification of the performance measure with respect to the periodic perturbation. The output of the gradient computer is then used to adjust the parameter K. The hill-climbing system now takes the form shown in Fig. 3(a), where the original system has been represented by the steady-state performance characteristic.

Fig. 3(b) shows schematically the system with a sinusoidal perturbation. The phase-sensitive rectifier consists of a multiplier and an averaging circuit, the output of which is the gradient. The parameter is adjusted once every perturbation period. The system within the dashed line comprises the hill-climbing controller.

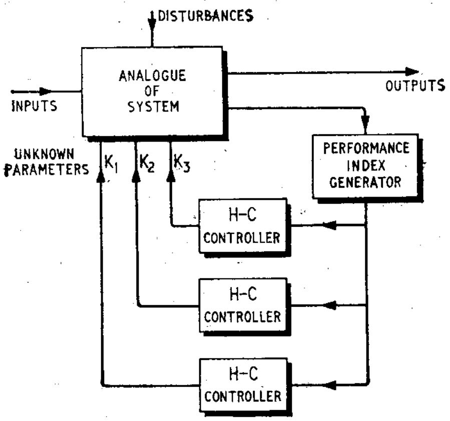


Fig. 4. Use of hill-climbing controllers to optimize the designer of a system.

Such a controller can be attached to any system to optimize the parameter K according to some suitably chosen criterion of performance.

Assume, for example, that the performance index P is related to the parameter K by a quadratic relationship:

$$P = P_o - (K - K_{opt})^2$$

where $K = K_1 + \delta K \sin \omega t$

The output of the phase-sensitive rectifier is given by:

$$G = \frac{1}{2\pi}\int_0^{2\pi} P \sin \omega t \, d(\omega t)$$

$$= \frac{1}{2\pi}\int_0^{2\pi}\Big[P_o - (K_1 - K_{opt})^2 + 2(K_1 - K_{opt})\,\delta K \sin \omega t - \delta K^2 \sin^2 \omega t\Big]\sin \omega t \, d(\omega t)$$

$$\doteqdot 2(K_1 - K_{opt})\,\delta K$$

since $(\delta K)^2$ is negligibly small.

The gradient of the $P - K$ characteristic is $2\,(K_1 - K_{opt})$

$$\therefore G = \text{grad.} \times \delta K.$$

At the end of each cycle the switch S is closed and the parameter is adjusted by an amount proportional to G.

i.e. $\triangle K = \alpha \int_0^T G\, dt = \alpha G T$

Thus the rate of adjustment $\frac{\triangle K}{T}$ is proportional to the gradient. At the optimum, the only movement will be due to the intentional perturbation. It is desirable, therefore, to keep the perturbation amplitude small, of the order of 10% of the maximum value of K.

Applications. — The hill-climbing technique described has been applied successfully to various engineering problems. One of the earliest applications was the optimization of internal combustion engines. The efficiency and power output of the engine depend, among other things, on the instant during a working cycle at which the air-fuel mixture is ignited. Draper and Lee used the hill-climbing technique for controlling the ignition-timing of the engine to optimize the engine performance. Similarly the combustion process in a gas burner which forms part of an industrial plant has been optimized by controlling the air supply. The air supply is subjected to the perturbations, and the performance criterion here is the completeness of the combustion and is determined by the amount of CO present in the combustion products.

The technique has been used to adjust more than one parameter simultaneously using several controllers of the type shown in Fig. 3(b), in parallel operation. Such controllers have been incorporated into an analogue computer. Such a computer has many useful applications. It has been used to study optimization of the performance of particular control systems.

It can be used in the *design* of control systems. If certain parameters of the design cannot be determined mathematically, then their values can be found in the following way. The design is simulated on the analogue. computer with the unknown parameters built in as adjustable variables. The system is subjected to normal input signals and expected disturbances, and a suitable performance measure is generated. Several hill-climbing controllers, one for each unknown parameter are then employed, as shown in Fig. 4, to adjust these parameters, thus yielding an optimum design under the given operating conditions.

A similar application is in system identification or model building. It is expensive and sometimes dangerous to carry out experiments on complex systems like chemical plants. Initial tests are best conducted on an electronic model or analogue of the plant. If an analogue is not available, it can be built by first setting up an approximate analogue with adjustable parameters. Recordings of the normal plant inputs are used as inputs to the analogue The output of the analogue is compared with the recorded output of the plant. The hill-climbing controllers then adjust the variable parameters until the difference in the outputs is minimized. The analogue will then be a very accurate dynamic model of the plant.

The hill-climbing technique can be applied to any system in which a performance measure can be formulated and where one or more parameters can be controlled. As a final example, the optimization of an electromechanical system will be described in detail.

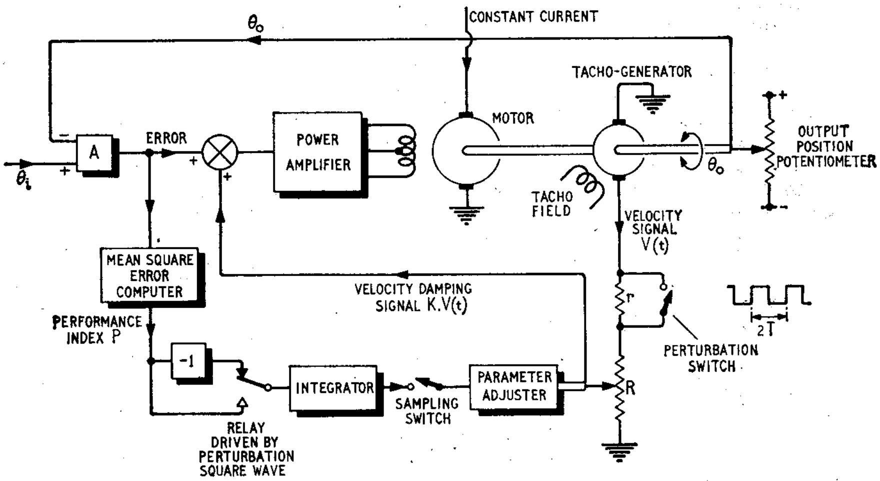


Fig. 5. A practical system based on a position-control servo.

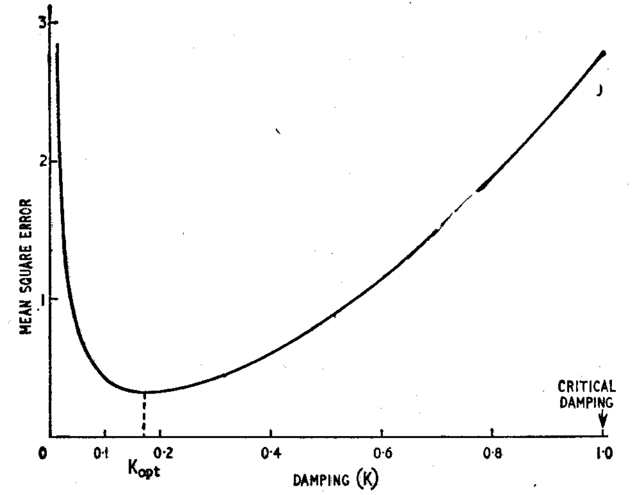


Fig. 6. P-K characteristic of the position control system of Fig. 5.

Fig. 5 is a system for controlling the position of an output shaft θ_o to a demanded position θ_i. The output position is measured using a rotary potentiometer. Amplifier A compares θ_i with θ_o. The difference or error $e = \theta_i - \theta_o$ is used to control the motor driving the output shaft. The motor will rotate until the error is zero. This system will oscillate continuously at a natural frequency ω_o radians per second unless some damping is introduced. Stabilization is achieved by using velocity damping, the velocity signal being obtained from a tachometer generator.

The accuracy in following a varying input demand is dependent on the amount of damping. Too little damping produces an oscillatory response, while too much damping produces a sluggish system. Also the amount of damping required depends on the type of input signal; for example, the optimum value of damping for minimum mean-square value of error in following a step change in position is half-critical damping, while that for a sinusoidal input of frequency equal to $\omega_o/2$ is zero.

Fig. 5 shows how the hill-climbing technique is implemented. The performance index used is mean square error. The damping is varied by controlling the potentiometer R. The damping is perturbed periodically by shorting out the small resistance $r(\doteqdot 0.1R)$ in series with R. This method introduces a constant percentage perturbation instead of a constant amplitude perturbation, and is generally preferable. The perturbation signal is a square-wave signal. Phase-sensitive rectification is performed by multiplying the performance signal by ± 1. This is achieved using a relay as shown.

The change in the integrator output in one cycle of the perturbation is given by:

$$\begin{aligned}\Delta e_o &= \frac{1}{T_i}\int_0^T f(K - \delta K)\,\delta t - \frac{1}{T_i}\int_T^{2T} f(K + \delta K)\,dt \\ &= \frac{T}{T_i}\left[f(K - \delta K) - f(K + \delta K)\right] \\ &= -\frac{2T}{T_i}\frac{df(K)}{dK}\,\delta K\end{aligned}$$

for small values of δK. This is proportional to the gradient $\frac{df(K)}{dK}$. The sampling switch closes once every perturbation cycle so that the parameter adjuster changes the damping potentiometer setting in steps proportional to Δe_o.

Fig. 6 shows the steady-state relationship between mean-square error and damping for a random input signal θ_i. It is observed that optimum here is a minimum; the characteristic is not a " hill " but a trough. Typical responses of the system as it approaches the optimum from an initial offset in damping are shown in Fig. 7. The gain of the optimizing loop is readily varied by adjusting the gain of the integrator and is set as high as possible consistent with reliable operation. In practice, it is possible to adjust the parameter to the optimum in a time interval equal to about twenty cycles of the perturbation.

Features and limitations.—The perturbation method of gradient measurement described is one of many methods of " hill-climbing ". It is not possible in this article to consider all the various techniques described in current literature on the subject. It is relevant, however, to mention certain features and limitations of the method described. The main advantage of the method is that it is easy to mechanize. Analogue studies have shown that it is applicable to a wide variety of control problems. Very little information about the control system to be optimized is required. Basically, we only need to know what performance measure to use, how to generate this signal and which parameter to control.

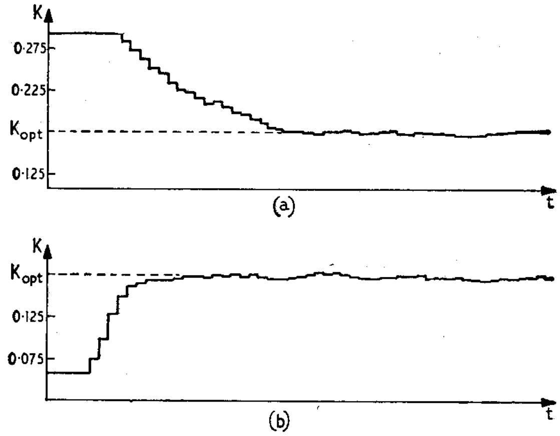


Fig. 7. Response of the hill-climbing system to initial misadjustment in damping; (a) $K > K_{opt}$; (b) $K < K_{opt}$.

The technique is easily extended to control several parameters simultaneously.

The main limitation is on the maximum speed of response, which is of the order of ten times the response time$\frac{(1)}{(\omega_o)}$ of the control system. If the performance curve has more than one " peak ", the technique is not capable of discriminating between these to find the highest peak. Starting from any initial point it will find the peak nearest to the initial point. These limitations are common features of other " hill-climbing " systems at present.

Current research in this field is directed mainly towards methods of gradient measurement which will increase the speed of optimization and of applying "hill-climbing" techniques to industrial plants.

WORLD OF WIRELESS

Record Number of Teleprinter Messages

IN April 1961 a Philips Type ES automatic switching system was installed in Kershaw House near London Airport at the S.I.T.A. (Société Internationale de Telecommunications Aeronautiques) Telegraph Centre which is operated by B.E.A. Since then the system has been in operation for 24 hours every day and at the end of the five year period the equipment has processed what is believed to be a record for the automatic routing of teleprinter messages—over 72 M messages. When first installed the system operated in a semi-automatic mode but was converted subsequently, without withdrawal from service, to fully automatic operation. High speed uniselectors stepping at 300 points per second are used for switching. All control and routing functions are electronic and re-transmission of incoming messages is commenced automatically as soon as sufficient information is obtained from the message to indicate the outgoing circuit required for re-routing. When the outgoing circuit is engaged messages are stored and then when the circuit becomes available transmitted in order of priority and waiting time. Storage is effected by ferrite core memories and magnetic tape.

Radio "Bugging"

AFTER seeing a film of American "bugging" equipment, the Postmaster General, the Rt. Hon. Anthony Wedgwood Benn, M.P., said "There is no doubt at all, having seen that film, that the menace of the micro-bug and eavesdropping equipment is a serious problem in other countries. Happily not many of them, so far as we know, are in use in this country." In discussing the legal position, the P.M.G. continued, "As far as the legal position is concerned, it is very simple: to attach anything to a telephone without permission is illegal and as far as radio-microphones are concerned we've tightened the conditions and made it a condition of the use of a radio-microphone that they can't be used for eavesdropping. So eavesdropping is illegal and we shall prosecute."

C.E.I. Common Examination

ONE of the functions undertaken by the Council of Engineering Institutions is to establish standards for the qualification of professional engineers. While the 13 constituent institutions* of the Council will remain responsible for the conditions of entry to their own membership, and some may require qualifications additional to those called for by the Council, the Council itself will set standards to which corporate members of the constituent institutions must in due course conform for the designation "Chartered Engineer."

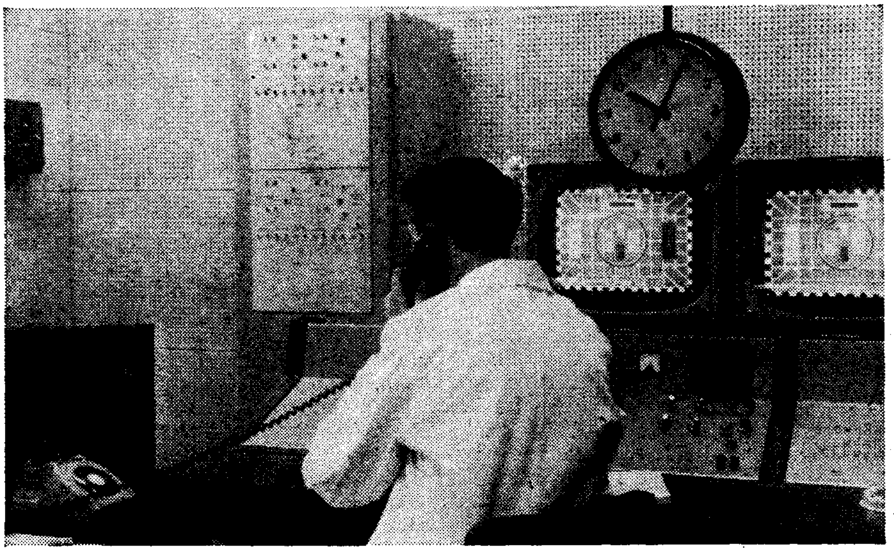
In accordance with the practice of the constituent institutions, the Council regards the qualification of a professional engineer as comprising three parts, namely academic education, training for the profession, and a period of responsible experience in the profession. The Council's plan for establishing the standard of academic education required of future Chartered Engineers is given in a booklet "Education and Training—Statement No. 1."

In this booklet, obtainable from the C.E.I., 2 Little Smith Street, London, S.W.1, details of Part I of the common examination are given with a syllabus and specimen papers. Part I of the examination will be held for the first time in October 1967, and and Part II (details of which are not yet published) in April 1968.

After a date still to be decided by the Privy Council it will be obligatory for all who wish to be registered as Chartered Engineers to pass Parts I and II of the C.E.I. examination or have an approved exempting academic qualification and, of course, have completed the requirements regarding training and experience for corporate membership of a constituent institution.

* Including the I.E.E. and the I.E.R.E.

Plumbicon Team Honoured.—The Television Society's Geoffrey Parr Award, which is made annually to an individual or team "for an outstanding contribution to television engineering or an associated science" has been presented to the Philips team concerned with the development of the



Remote Controlled TV Stations

Two unmanned I.T.A. television relay stations, Great Massingham, Norfolk and Hameringham, Lincolnshire are controlled from Belmont, Lincolnshire. The two stations, equipped with G.E.C. Telecode and Teleshift time and frequency division multiplex equipment, are linked by G.P.O. tie lines to the control centre. The illustration shows a mimic diagram (on the door of a console) which gives engineers continuous indication of conditions at the relay stations and enables the monitoring and controlling of numerous functions to be effected. Signals are transmitted over the G.P.O. lines which can also be used for speech communication when service engineers visit the relay stations.

Plumbicon colour television camera tube. The names mentioned in the citation are Dr. H. Bruijning, director of research of the Philips Research Labs., Aachen, Germany, Dr. E. F. de Haan, assistant director of the Research Labs., at Eindhoven, and Dr. L. Heijne, research physicist at Eindhoven.

Lady Fleming, widow of Sir Ambrose, was present at the 17th **Fleming Memorial Lecture** of the Television Society on April 21st at the Royal Institution, London, when Professor W. D. Wright, of Imperial College, was the lecturer. His subject was "The implications for television of modern thinking on the visual process." Lady Fleming presented the Society's 1964/5 awards. E. J. Gargini of Rediffusion Research received the *Electronic Engineering* Premium for his paper "Colour television by wire"; H. Steele (A.B.C. Television) the T.C.C. Premium for "The transcoding of colour television signals"; J. Weltman (I.T.A.) the E.M.I. Premium for "Television university"; J. E. F. Voss and C. J. Paton (B.B.C.) the Pye Premium for "Television coverage of the Tokyo Games"; Dr. N. Mayer (Inst. für Rundfunktechnik, Munich) the *Wireless World* Premium for "The N.T.S.C. colour television system using additional reference transmission"; and J. D. Last the Mullard Premium for "Varactor diode parametric amplifier and harmonic generators."

Magnetic Cores for Matrix Stores.—The British Standards Institution has published a "Guide to the specification of magnetic cores for use in co-incident current matrix stores," B.S. 4010. The publication defines terms used to specify the properties of magnetic cores intended for use in coincident-current matrix stores having a nominal 2:1 selection ratio. Measuring methods, conditions of test, recommendations for the specification of cores and the correct presentation of core performance data are also included. Copies, price 12s 6d, are available from B.S.I., Sales Branch, 2 Park Street, London, W.1.

Ten more u.h.f. transmitting stations for BBC-2 have been approved in principle by the P.M.G. Six of these will be installed on existing sites, these are indicated by an asterisk in the following list. All transmissions will be horizontally polarized and channel numbers are given in brackets. Belmont* (28), Sandy Heath* (27), Londonderry* (44), Caradon Hill* (28), East Lothian (27), Moel-y-Parc* (45), Staffordshire (26), Angus* (63), Sussex (55), and North Hampshire (45). These ten stations will serve about 5.25 M people and together with the 18 stations already approved will make BBC-2 available to about 77% of the population.

The Society of Electronic and Radio Technicians and the Wolverhampton College of Technology are organizing a three-day symposium on **radio and television maintenance** to be held at the College of Technology on June 14th, 15th and 16th. Subjects to be discussed will cover education and training of radio servicemen, Television picture quality, Maintenance problems, Test equipment, Colour television and Programme distribution systems. Further information, including registration forms, may be obtained from W. J. Anderson, College of Technology, Wulfruna Street, Wolverhampton.

Speaking at the jubilee luncheon of the **Scientific Instrument Manufacturers' Association,** Mr. Edmund Dell, Joint Parliamentary Secretary at the Ministry of Technology, referred to the industry's increased exports from £27.5M in 1958 to £62.7M in 1964. He added: "Nevertheless, it would be a mistake to disguise the fact that there are question marks over the industry's future. Imports have been increasing twice as rapidly as exports. Whereas in 1958 about 11% of apparent home consumption was supplied by imports, in 1964 this percentage has reached almost 30%. Although our export record in the field of scientific instruments has been good it is still true that our share of world trade in scientific instruments has shown some tendency to decline."

The **Baird Travelling Scholarship** for 1966 has been awarded by the Television Society to John D. Penney who is in the Department of Electrical Engineering at University College, London, doing research work for a Ph.D. His work is concerned with tunnel diode amplifiers and is being supported by a Science Research Council grant. With the Baird Scholarship grant of £200 he intends to visit America this year where he plans to visit companies and technical institutes to study tunnel diode amplifiers.

New Radio-telephone Facility.—Thames Radio came into operation recently and provides an improved Post Office radio-telephone service for ships using the Port of London. With aerials near Sevenoaks it will cover an area roughly from Tower Bridge to beyond Canvey Island, including the Medway. It will be available to handle telephone calls with ships at anchor, in port, or anywhere in the Thames area. The service operates at v.h.f. with frequency modulation. The frequency, 156.8 Mc/s, is used for establishing communication in either direction, after which the working frequencies are: ships 157.35 Mc/s and Thames Radio 161.95 Mc/s.

Two-way personal radio for Police Constables has been introduced in six Divisions of the Metropolitan Police Force. The equipment, which weighs only two pounds, is worn strapped across the chest and consists of two units, a transmitter-receiver and a combined microphone and speaker attached by a flexible lead. The aerial is incorporated in this lead. By pressing a button on the microphone a signal is transmitted. At the receiving station the receiver energizes a bell on a special telephone handset and conversation is then carried on in a simplex mode.

Thin Films Conference.—A joint I.E.R.E./I.E.E. conference on "Applications of thin films in electronic engineering" is to be held at Imperial College, London, from July 11th to 14th. There will be sessions on the preparation of thin films; general applications; thin film elements and integrated circuits; magnetic films; and cryoelectric films. Registration forms (fee £13) are obtainable from the I.E.R.E., 8-9 Bedford Square, London, W.C.1.

This summer the new **remotely controlled P.O. radio station** at Leafield, Oxfordshire, becomes fully operational. The station has cost over £1M and is equipped with six 85 kW and 12 30 kW h.f. transmitters. Transistors and motorized switches have been incorporated to provide a very high degree of reliability. At present the station is controlled by an "on site" operator but eventually the operation will be remotely controlled from London.

The **Northern Radio Societies Association** is to hold its second convention during September 3rd and 4th at Belle Vue, Manchester. Further details are available from I. D. MacArthur, 55 Langdale Road, Bramhall, Cheshire.

BBC-2 test transmissions from Black Hill, Central Scotland, will start on Channel 46 on June 11th. The u.h.f. transmitter will provide BBC-2 programmes to about 2,300,000 people in Central Scotland, including Glasgow, and part of Edinburgh.

A one-day symposium, **Computers in Medicine,** is to be held on July 6th at Enfield College of Technology. Demonstrations of equipment will be given by selected computer manufacturers during the day. Further details are available from the Academic Registrar, Enfield College of Technology, Queensway, Enfield.

Correction.—In the news item on page 174 of our last issue regarding the new mast and aerial for Winter Hill we inadvertently gave the incorrect frequencies for channel 9. These should read 191.25 Mc/s sound and 194.75 Mc/s vision.

"A.F. Cascode."—We regret that an ECC81 was indicated instead of an ECC83 in Fig. 7 of "A.F. Amplification with the Cascode" by G. A. Stevens in the last issue.

PERSONALITIES

H. E. Barnett, T.D., M.Sc., A.C.G.I., D.I.C., M.I.E.E, until recently assistant director of the Electrical Inspection Directorate of the Ministry of Aviation, has been appointed director of the British Calibration Service set up by the Minister of Technology (see "Editorial Comment"). For the past 11 years Mr. Barnett has been in charge of engineering services in E.I.D. which included the organization of the Inspectorate's electrical standards, metrology and materials laboratories and has been concerned with developing new methods of measurement at r.f. Mr. Barnett is also a member of the Radio and Electronics Measurement Committee of the Ministry.

Air Commodore J. C. Millar, D.S.O., M.I.E.E., has joined the London Office staff of the Marconi Company for special liaison duties with the Services, Government departments and other users, on behalf of the Marconi Aeronautical Division. Educated at Malvern and Trinity College, Cambridge, Air Cmdre. Millar served for 33 years in the R.A.F. He was for three years Command Signals Officer, Bomber Command and from June 1963 until his retirement was R.A.F. Provost-Marshal in the Ministry of Defence.

Air Cmdre. J. C. Millar

R. N. Barton, M.I.Mech.E., has joined Plessey as director and general manager of the Telecommunications Group. He was previously production director with Standard Telephones and Cables.

G. Ivor Thomas, senior design engineer of A. B. Metal Products Ltd. for the past 12 years, has been appointed quality manager, a newly created position covering all aspects of inspection and quality control.

W. A. Everden has been appointed head of the Passive Components Department of Mullard's Industrial Markets Division. Formerly the commercial product manager for ferrites, Mr. Everden, who is 37, joined the Mullard company in 1948. He spent six years in various production posts and a further year at Mullard Research Laboratories where he specialized in applications research on ferrites.

W. A. Everden

Francis Seely B.Sc., aged 39, has been appointed head of Market Departments in Mullard's Industrial Markets Division. The Market Departments are customer-orientated groups specializing in and serving various sectors of the electronics industry such as computing and telephone exchanges, telecommunications and radar, and instrumentation, control and power. Mr. Seely, who has been in the electronics industry for 21 years, was previously manager of the market department for instrumentation, control and power. The post will now be filled by **B. H. Penney,** Grad.I.E.E., formerly manager of the Division's Industrial Sales Department. Mr. Penney, has had seventeen years' experience in electronics, first with the G.P.O. and subsequently in the semiconductor industry. He joined Mullard in 1964. The new head of the Industrial Sales Department is **T. E. Days** who worked at the Mullard Research Laboratories on special types of valve from 1940 until he left in 1955. He rejoined Mullard in 1961 and has been concerned with the supply of specialized components to universities and research laboratories.

F. C. McCrea, after 37 years' service with the Dubilier Condenser Company, has relinquished his executive duties but is remaining as chairman of the Board. He is succeeded as managing director by **J. H. Cotton** who has been with the company since 1930 and joined the board as works director in 1947. The new assistant managing director is **J. Goodman.**

Colonel J. S. Vickers, B.Sc.(Eng.), A.M.I.E.E., who joined the British Standards Institution in 1961, has been appointed head of the Planning Group set up to co-ordinate the Institution's increasing volume of work. He trained as an electrical and mechanical engineer, serving his apprenticeship at the Rugby works of British Thomson-Houston. Early in 1939 Colonel Vickers took a Regular commission in R.A.O.C. from which R.E.M.E. was formed in 1942. During the past four years at B.S.I. Colonel Vickers has been primarily concerned with U.K. participation in the work of the International Commission on Rules for the Approval of Electrical Equipment.

John Lawson, who joined Feedback Ltd. in 1961 as a development engineer, has been appointed to the company's technical sales staff with specific responsibilities in exports. He served for several years in R.E.M.E. and was in the Test and Development Department of Servomex Controls before joining Feedback.

F. Seely

B. H. Penney

T. E. Days

G. C. Gaut, M.A., B.Sc., research director of the Plessey Company, has been appointed by the Minister of Technology as a part-time member of the National Research Development Corporation. Mr. Gaut has been with Plessey since graduating at University College, Oxford, in 1934, and has been an executive director since 1951. He was responsible for setting up in 1937 the company's first laboratory for research and development on technical processes for the manufacture of electronic components. The laboratory, which was started at Ilford, is now at Caswell, Towcester. Mr. Gaut is also a director of the Plessey subsidiary Semiconductors Ltd. The N.R.D.C. was set up in 1949 by the Board of Trade to secure "in the public interest, the development and exploitation of inventions resulting from public research or other inventions where these are not being sufficiently exploited." N.R.D.C., which has been the responsibility of the Minister of Technology since 1965, has eight part-time members and three full-time members including the managing director **J. C. Duckworth,** who was for some years with Ferranti and was in charge of the development of the guidance and control system for "Bloodhound" guided missile.

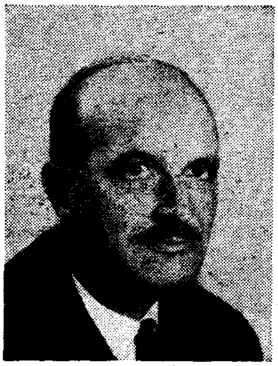
Ian D. Davie, A.M.I.E.R.E., has joined Kent Precision Electronics Ltd. as chief engineer. Mr. Davie, who is 31 and was awarded the City and Guilds of London Institute Full Technological

I. D. Davie

Certificate in Telecommunications Engineering in 1959, was formerly with Roband Electronics Ltd. as manager of the power supply department.

J. S. Lasenby has joined Kent Precision Electronics Ltd. as sales manager. He held a similar post with Everett Edgcumbe & Co. before joining K.P.E.

B. J. Nearn has been appointed group sales manager in the S.T.C. Components Group. A former Signals Officer in the Technical Branch of the Royal Air Force, he joined S.T.C. in 1948 and has held a number of executive posts. Initially in the Rectifier Division, he



B. J. Nearn

D. R. Salmon

G. Thornton

subsequently became marketing manager, Magnetic Materials Division, and later, marketing manager Rectifier Division. Since 1963 he has been general sales manager, Components Group. **D. R. Salmon** has become home sales manager in the company's Components Marketing Division. Aged 39, Mr. Salmon joined the company as a student in 1944. After National Service in the R.A.F., he joined the Capacitor Test Department (then at North Woolwich) in 1948. Becoming interested in the suppression of radio interference, he took part in setting up the first radio interference suppression service in 1950 and for some years was engaged on the design of interference suppression devices. He has been product sales manager, Capacitor Division, since the end of 1962. **G. Thornton,** who is 28 and joined S.T.C. in 1962, was initially engaged on production control of transistors, and has been head of market research since 1963.

R. B. C. Copsey, A.M.I.E.E., who joined Redifon in 1949, has been appointed chief engineer, Redifon Marine, at the company's Wandsworth headquarters. Mr. Copsey, who is a member of the Ships' Wireless Working Party Radio

R. B. C. Copsey

Technical Committee, will control both development and project engineering for Redifon's marine operations. Designer of many of Redifon's marine communications equipment, Mr. Copsey led the Redifon team which designed the drive and frequency synthesis equipment for the recently completed NATO high-power v.l.f. station at Anthorn in Cumberland. Prior to his present appointment he was engaged in research on aerial systems.

A. W. Cross, B.Sc., who has joined Abbey Electronics and Automation Ltd., of Cheshunt, Herts., as sales manager, was previously senior engineer, then sales manager of W. H. Sanders (Electronics) Ltd. He gained a B.Sc., in mathematics and a B.Sc., honours in physics at Hull University and is author of the book, "Experimental Microwaves."

OBITUARY

William Henry Eccles, F.R.S., "the first of the radio physicists" died on April 29th in his 91st year. In our issue of last September we published a tribute to him on the occasion of his 90th birthday. In it our contributor "H.F.S." wrote "It would hardly be too fanciful to put forward Dr. Eccles as the grandfather of the transistor. In 1909 he demonstrated oscillating crystal detector circuits and developed the general theory 'that under certain conditions a rectifying detector could become a generator of oscillations and conversely a generator of oscillations could be used as a rectifier'." What was however, his most significant work was on radio wave propagation. In 1911 he published a Royal Society paper explaining and expanding the theory put forward ten years earlier by Heaviside. Dr. Eccles was for about two years with Marconi's but in 1901 went into the academic world and in 1916 was appointed to the chair of electrical engineering and applied physics at the City and Guilds of London Institute. For a short time during the First War he was director of the Admiralty Electrical Engineering Laboratory. He was elected a Fellow of the Royal Society in 1921.

THE MONTH'S CONFERENCES AND EXHIBITIONS

Further details are obtainable from the addresses in parentheses

LONDON

June 6-8 — Brunel College
Integrated Circuits in Electronic Equipment
(Brunel College, Woodlands Avenue, W.3)

June 6-8 — Savoy Place
Design and Construction of Large Steerable Aerials
(I.E.E., Savoy Place, W.C.2)

June 20-25 — Central Hall
Automatic Control (I.F.A.C. Congress)
(Congress Secretariat, U.K.A.C., c/o I.E.E., Savoy Place, W.C.2)

WOLVERHAMPTON

June 14-16 — College of Technology
Radio & Television Maintenance
(W. J. Anderson, College of Technology, Wulfruna St., Wolverhampton)

OVERSEAS

June 6-10 — Prague
Interkama Symposium
(Ing. J. Moravek, Plzenska 66, Prague-5-Smichov)

June 15-17 — Philadelphia
International Communications Conference
(A. E. Joel, Bell Telephone Labs, Holmdel, N.J.)

June 15-20
Scientific Congress on Electronics
(Rassegna Elettronica, via Crescenzio 9, Rome)

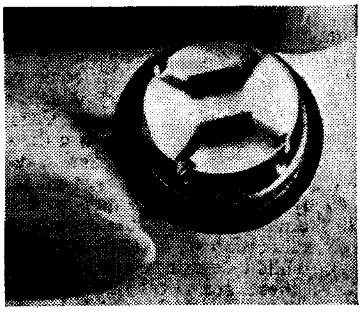
June 21-23 — Colorado
Precision Electromagnetic Measurements
(J. Cronland, Bureau of Continuation Education, University of Colorado, Boulder)

June 21-24 — Chicago
Data Processing Conference & Exhibition
(Data Processing Management Assoc., 524 Busse Highway, Park Ridge, Ill.)

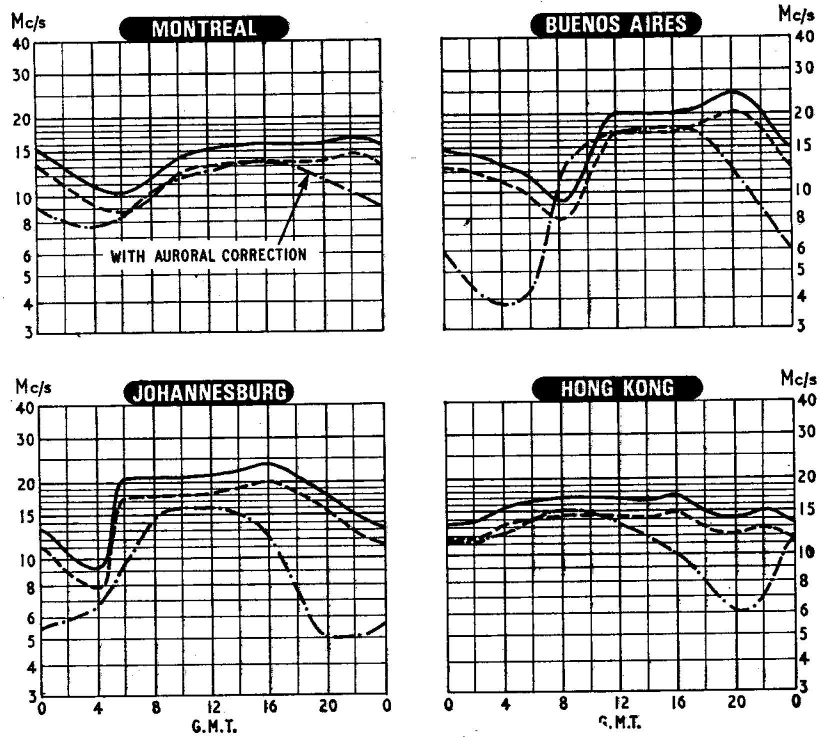
June 22-24 — Pasadena
Electron Devices Research Conference
(S. J. Buchsbaum, Bell Telephone Labs., Murray Hill, N.J.)

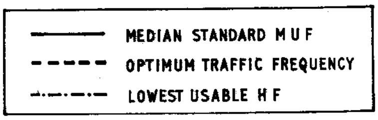
Quartz Bandpass Filter

CONVENTIONAL bandpass crystal filters normally require a network of capacitors, transformers and crystals. To replace these components, Bell Telephone Laboratories have developed a filter consisting of a single wafer of quartz. The device is made from a wafer of quartz 10 mm dia. and 1 mm thick on which four rectangular electrodes are deposited, two on each side. When the device is used as a filter, the crystal resonates at two different frequencies, above and below the normal quartz resonant frequency. This property of dual resonance is due to mechanical coupling between the resonators, the couplings depending on the mass of the electrodes and the distance between them.



The two resonant frequencies determine the pass band of the filter, the bandwidth being about 0·1% of the mid-band frequency (which can be between 1 and 150 Mc/s). Modifications can be made to the filter, e.g., more electrodes can be added giving greater selectivity and addition of a thin film capacitor would improve the loss characteristics. The filter exhibits a "constant-*k*" type impedance and will give up to 80 dB of attenuation. The filter has uses in many fields including narrow-band f.m. systems and carrier telephone systems (e.g., the new coaxial cable system referred to on page 283).





H. F. PREDICTIONS JUNE

A marked increase in sunspot activity was noted during the months of March and April. Several large active sunspot groups were apparent during this period. Solar flares resulted in a number of Dellinger-type fadeouts. Such a high measure of solar activity, during the early years of the sunspot cycle, is comparatively rare.

The prediction curves show the median standard MUF, optimum traffic frequency and the lowest usable frequency (LUF) for reception in this country. Unlike the standard MUF, the LUF is closely dependent upon such factors as transmitter power, aerials, and the type of modulation. The LUF curves shown were drawn by Cable & Wireless Ltd. for commercial telegraphy and assume the use of transmitters of several kilowatts and rhombic aerials.

LC NETWORKS IN TO-5 CASE

TUNED CIRCUITS JOIN THE COMPONENTS NOW AVAILABLE IN TRANSISTOR CASES

INTEGRATED circuits have been produced in TO-5 transistor cans for some time. More recently, relays and potentiometers have appeared in TO-5 cans, and have been described in previous issues. An American company, JFD Electronics Corporation*, has now developed a tunable *LC* filter network in a TO-5 enclosure, and was exhibited at the I.E.E.E. International Convention and Show in New York.

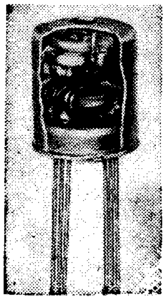


Fig. 1. A typical example of an LC network enclosed in a TO-5 case. The particular item is a phase detector circuit.

The construction is shown in Fig. 1 and a typical filter comprises a toroidal transformer, a fixed ceramic capacitor and a variable ceramic capacitor. A range of filters are available with centre frequencies from 3 to 250 Mc/s, which can be varied by about 10%. 3 dB bandwidths range from 40 kc/s to 4 Mc/s with unloaded Q's of about 60 to 100.

The tuning capacitors used in these networks measure 0.21×0.28×0.12 in and seven capacitance ranges are available from 1.6-9 pF to 8.5-50 pF.

*Represented in the U.K. by S.T.C., Capacitor Division.

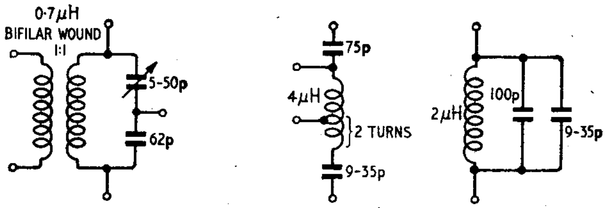


Fig. 2. Three of the many variations possible with the LC networks.

Some of the circuit configurations possible with these networks are shown in Figs. 2 and 3.

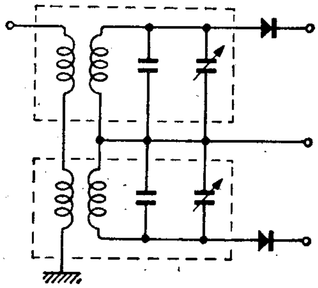


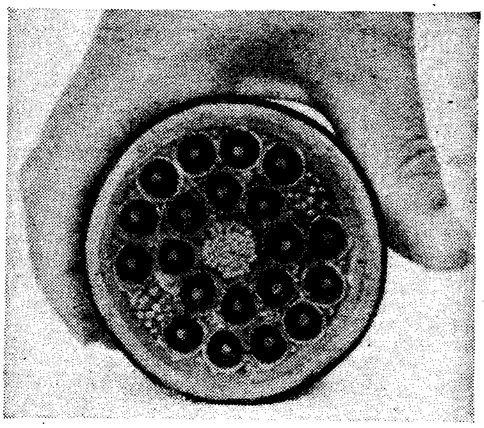
Fig. 3. A 5 Mc/s discriminator using two of the networks in TO-5 cases.

High-Capacity Coaxial Cable

FIELD trials are in progress in Ohio of a coaxial cable system with a capacity of 32,400 voice channels. The cables for the system have been developed by Bell Telephone Laboratories and contain 20 copper coaxial conductors (10 for each direction) each about $\frac{3}{8}$ in dia.

The capacity is much higher than normal cables because of the increased number of conductors and also because the frequency of operation is about 0.5-17.5 Mc/s, each coaxial pair carrying 3,600 voice channels. (A previous Bell cable system operated from about 0.3 Mc/s to 8.3 Mc/s with a capacity of 1,860 channels per pair.)

Three types of repeater are used in the system, which uses specially developed silicon transistors. In addition to the basic repeaters located at intervals of 2 miles, there are regulator and equalizing repeaters. Regulator repeaters are placed every 14 miles and their function is to compensate for cable losses caused by changes in temperature. The remotely controlled equalizing repeaters are at intervals of 50 miles and intended to compensate for changes in gain due to various unpredictable effects that occur in cables and equipment. Additional repeater stations occur at intervals of 160 miles. The repeaters are checked by test signals whose levels are monitored at a main repeater or receiving station.



Coaxial cable developed by Bell with a capacity of 32,400 voice channels.

Demonstrations at V.L.F.

SERIES CIRCUITS AND PHASE-SHIFT OSCILLATORS

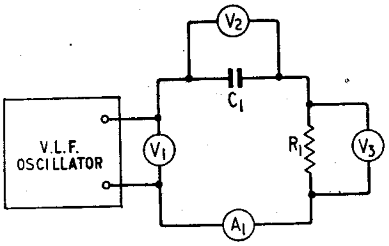
By T. PALMER,* B.A., Grad.I.E.E., Assoc.I.E.R.E.

A DEMONSTRATION of the phase difference between the currents flowing in parallel branches containing resistance and capacitance respectively, was described in *Wireless World,* p. 515, October, 1963. The present demonstration begins with the phase difference between the voltages across a resistor and across a capacitor connected in series to the output of a v.l.f. oscillator. In the final example of the phase-shifting properties of a series CR circuit, a ladder network of four CR sections is connected in a v.l.f. phase-shift oscillator circuit.

* Acton Technical College.

Fig. 1 shows the first part of the demonstration. The v.l.f. oscillator should have a frequency of the order of 5 cycles/minute. Suitable v.l.f. oscillators were mentioned in the previous article. In view of the use we shall make of the circuit later, it is convenient to let C1 be 27 μF and R1, 67 kΩ, but the values are not critical.

Voltmeters V2 and V3 should be high input impedance valve or transistor voltmeters giving readings on centre-zero meters. Suitable circuits are given in the Mullard Reference Manual of Transistor Circuits (Page 271, Fig. 4) and Application Note 8, issued by Texas Instruments (Fig. 3). A feature of the demonstration is that it is possible to make a mistake in the connection of the voltmeters: this draws attention to some of the conventions associated with phasor diagrams. If the meters are not connected according to a consistent pattern, the indications on them do not conform to our theories. Before connecting C1, it is worth while arranging a circuit with a resistor R8 in place of C1. When the voltmeters are correctly connected, the pointers of all of them should swing from side to side in synchronism. The pointer of the centre-zero microammeter A1 should swing in phase with them. Note that the sum of the voltages indicated by V2 and V3 equal that of V1. After this has been arranged, C1 in inserted in place of R8. We now see that the pointers of the voltmeters no longer swing in step.



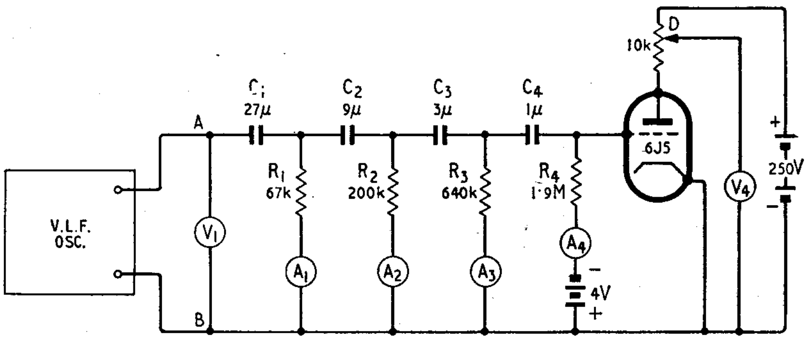
The pointers of the various meters indicate the phase relation in the circuit; V3, measuring the voltage across R1, leads V1; V2, measuring that across C1, lags V1, and the current indicated by the pointer of the microammeter A1 is in step with that of V3. Although the maximum readings on V2 and V3 added up to the maximum reading on V1 when we had R8 in series with R1, we now see that this is no longer so. Instead we have a relation of the form:—

$$(V_{2max})^2+(V_{3max})^2=(V_{1max})^2$$

We can next try the effect of varying the frequency of the oscillator; since the effects are in accordance with theory, we shall not mention them here. At a frequency of 5 cycles/minute, V3 leads V1 by about 45°.

The next step is to modify the circuit to the form shown in Fig. 2.

The values shown on the circuit give a phase shift of approximately 45° in each CR stage, at a frequency of 5 cycles/minute. It was decided to reduce the shunting effect of each CR stage on the resistor of the previous stage by increasing the impedance of the different stages as we go from the input towards the output. (See Summary note 2.) With the values chosen, at a frequency of 5 cycles/minute, the voltage at the output is about 0·15 of that applied across the input at AB. But again it is, of course, necessary to connect the meters according to a consistent pattern; here, also, it may be worth while, before connecting C2, C3, C4, to connect suitable resistors in their place, to show that all the meters, when connected according to a consistent pattern, swing in step. When we revert to capacitors, the phase shift in each stage is apparent. It can now be seen that, at a frequency of 5 cycles/minute, the pointer of A2 leads that of V1 by 90°; that of A4 leads by 180°. Since the valve introduces a phase shift of 180°, V4, measuring the voltage from anode to cathode. now swings in phase with



V1, when the input frequency is 5 cycles/minute. V4 is a moving-coil voltmeter of the usual type, with a full scale deflection of 300 V. It gives its highest reading at the instant when V1 shows that A has its maximum positive voltage with respect to B. When the oscillator generates frequencies other than 5 cycles/minute, V4 and V1 are not in phase.

The oscillator is now disconnected from the input AB and a lead from the output, D, of the triode amplifier is now connected to terminal A. If a sufficient fraction of the output voltage is fed back to the input, we now see that the system has finally become an oscillator working at 5 cycles/minute. If the fraction of the output voltage is too small, oscillations will not be maintained (they may take some time to die away in this circuit with its long time constants). If too large a fraction is fed back, the voltage at the output is not sinusoidal. By considering the swings of the pointers of A1, A2, A3 and A4 the phase shift in the different stages can still be appreciated. If the values of the resistors (or the capacitors) are multiplied by a suitable factor, students can see the effect this has on the frequency of the oscillation.

Summary

1. The oscillator I used in Fig. 2 gave an output of about 1 volt, peak, which gave small deflections on the microammeters. I amplified the output by a simple triode, as shown in Fig. 3. The arrangement has the advantage that, when we are considering whether V1 and V4 are in phase or not, we are concerned with the swing of pointers on similar instruments. The voltages are in phase if the two meters give maximum deflection at the same instant.

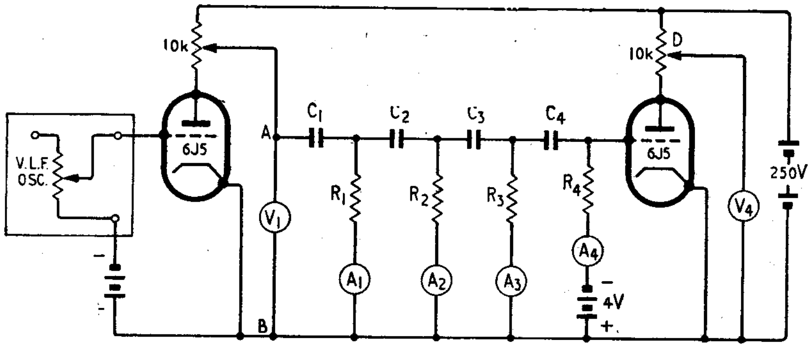
2. The advantage of increasing the impedance of the CR stages as we go from input to output, is that we do not have to take the loading of one stage on the previous stage very seriously. It has the snag, however, that with R4 equal to 1·9 MΩ, a very sensitive meter, rated at, say, 10-0-10 μA, is required to show the swing on A4, which may be of the order of 2 μA. If you prefer, all the CR stages can have the values of the first stage. (The frequency will not be 5 cycles/minute but it may be suitable; also, a larger fraction of the output voltage must be fed back to maintain oscillation.)

3. There is no reason why this technique should not be used with a transistor amplifier which turns into a v.l.f. oscillator when the output is returned to the input. The circuit of Fig. 15 of the article by Mr. F. Butler, (*Wireless World*, p. 588, December, 1962) is suitable for this demonstration. I prefer not to use it for an elementary class because:

(a) there is some uncertainty about the input resistance of the transistor amplifier; it is possible to measure this experimentally, but this detracts from the basic simplicity of the demonstration;

(b) it is more difficult for elementary students to understand how the output is fed back to the input.

4. When a class had seen the demonstration, a student who had seemed to understand how the system oscillated, wanted to know why anyone should want to make an oscillator.



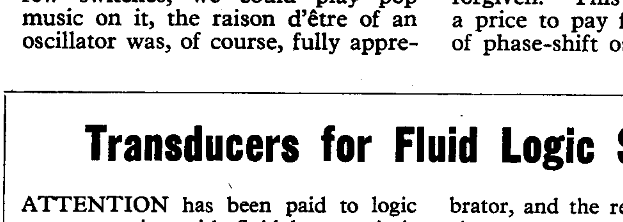
When I explained that if we chose suitable CR values and threw in a few switches, we could play pop music on it, the raison d'être of an oscillator was, of course, fully appreciated. If an epidemic of electronic organs breaks out, I hope I may be forgiven. This would be too high a price to pay for an understanding of phase-shift oscillators.

Transducers for Fluid Logic Systems

ATTENTION has been paid to logic gates operating with fluid because it is felt that such might form the basis of low-cost low-speed digital logic systems. To enable electronic signals to be converted to pneumatic signals devices are required which are capable of operating up to several hundred pulses per second. Often solenoids are used as transducers but these are limited in speed of operation. I.C.T. demonstrated a faster method using a pressure type loudspeaker at the Physics Exhibition held recently (reported in the May issue).

A fluid logic OR/NOR gate was shown operating with an input data rate of 300 pulses per second. The electrical input triggered an 8 kc/s multivibrator, and the resultant modulated a.f. signal was then amplified to 3 W and fed to the loudspeaker. A convergent cone coupled the acoustic output to the logic element, the presence or absence of the acoustic signal switching a fluid jet into one of two positions. Pressure signals at various points of a logic network could be displayed on an oscilloscope and small piezoelectric transducers were used to extract the signals.

Another demonstration showed that line matching is not only an electrical problem—interconnections in fluid systems exhibit characteristic impedance. A mismatch was caused by an open-ended tube and the pressure signals near the end of the tube could be compared to that of a matched line. By altering the size of the open-ended line between fully open and fully closed the signal could be seen to reach an optimum between the two extremes.



Transistor Electronic Organ—2

Semiconductors in Electronic Organs

By T. D. TOWERS, M.B.E.

THE first article in this series (May 1966 issue) examined in broad terms the principal features of electronic organs. The present article shows in some detail how such principles are reflected in actual circuits from what is now the commonest class of commercial instrument—the non-mechanical, transistorized, all-electronic organ.

Fig. 1 illustrates in block diagram form the main sections of a complete electronic organ. Transistors are

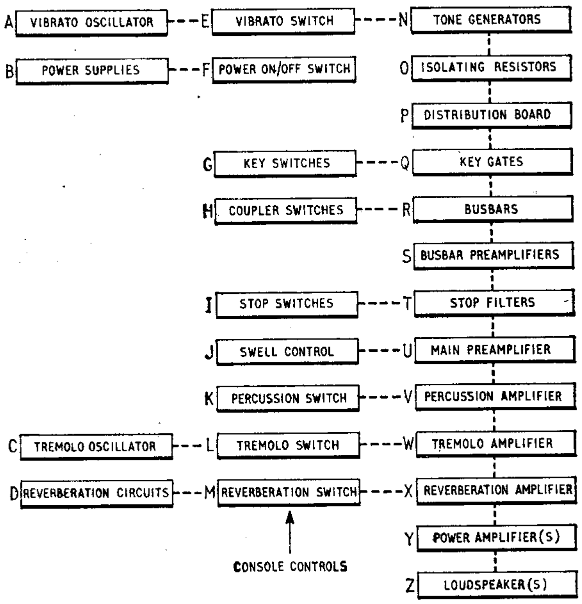


Fig. 1. Block diagram showing main sections of complete electronic organ.

often used in (1) *oscillators* as source for tones (N), vibrato (A) or tremolo (C); (2) *dividers* as sources for tones (N); (3) *gates* controlled by playing key switches for tone transmission (Q) or percussion (V); (4) *modulators* for applying vibrato (N) or tremolo (W) drive; (5) *preamplifiers* (S, U, W); (6) *power amplifiers* (Y); and (7) *power supplies* (B).

OSCILLATORS

The electronic organ uses two quite distinct classes of oscillators: (1) an audio-frequency type to generate tone signals, directly or indirectly, and (2) a subsonic type to provide drive for vibrato or tremolo effect in the generated tones.

Tone-generator oscillators.—Tone-generator oscillators are usually LC-tuned, because you can achieve the required frequency stability more easily with LC tuning than with, say, the resistor-capacitor network controlling an RC oscillator.

As to the level of frequency stability required, it is not commonly realized that there is a subjective element in this. In the equal-tempered scale (to which electronic organs are normally tuned) the frequencies of notes a semitone apart are in the ratio of $2^{1/12}:1$, i.e. 1.05946:1. For non-mathematicians this means that a semitone corresponds to about 6% difference in frequency. Hearing varies, but most people can distinguish a pitch difference of the order of $\frac{1}{2}$%, i.e. about a tenth of a semitone. (Although I once played in a group with a trumpet player who had to appeal to me to tell him when he had managed to set his tuning slide to bring him to the tuning "A." I reckoned he could not distinguish much better than a quarter tone!) The $\frac{1}{2}$% discrimination of normal subjects suggests that an organ should be capable of being tuned (and of holding its tuning) to something well down on this, say at 0.1%. However, experience has proved that too exact tuning can destroy to some extent the fine musical qualities of an organ, and that an instrument deliberately mistuned randomly at between one-fourth and one-half per cent of perfect pitch receives better customer acceptance on listening tests than one with perfect pitch. This still calls, however, for something of the order of 0.1% oscillator stability, if the deliberate mistuning is to be held over a period.

Another approach is to accept a less stable oscillator, tune it as nearly as possible to perfect pitch and be confident that the inevitable frequency wandering will produce the same effect as a deliberate controlled mistuning of a higher-stability oscillator.

Free-phase LC oscillators for direct tone-generation.—Historically, the first organs to use *LC* tone generators were the "free-phase" type. In these, a separate oscillator was used for each tone frequency required. Fig 2(a) gives a typical circuit. The oscillator is the Hartley type, normally used for this application because of its good frequency stability and its harmonic-rich sawtooth output. The inductor, *L*, is a ferrite pot-core with an adjustable slug permitting some ±4% variation of inductance. One standard pot-core assembly with a range of coil inserts can be used to cover the whole organ gamut, say from 64c/s to 4,080c/s. The playing key switch in the d.c. supply is the hallmark of the "free-phase" tone-generator oscillator. An uncommon refinement is that the d.c. supply can be set to 3V for "soft" or 9V for "loud." This illustrates the good frequency stability of the oscillator under changing supply voltage. The 100kΩ resistor in the output line isolates the oscillator from others connected to the same output busbar.

Master LC oscillators for divider organs.—The commonest type of LC oscillator in electronic organs nowa-

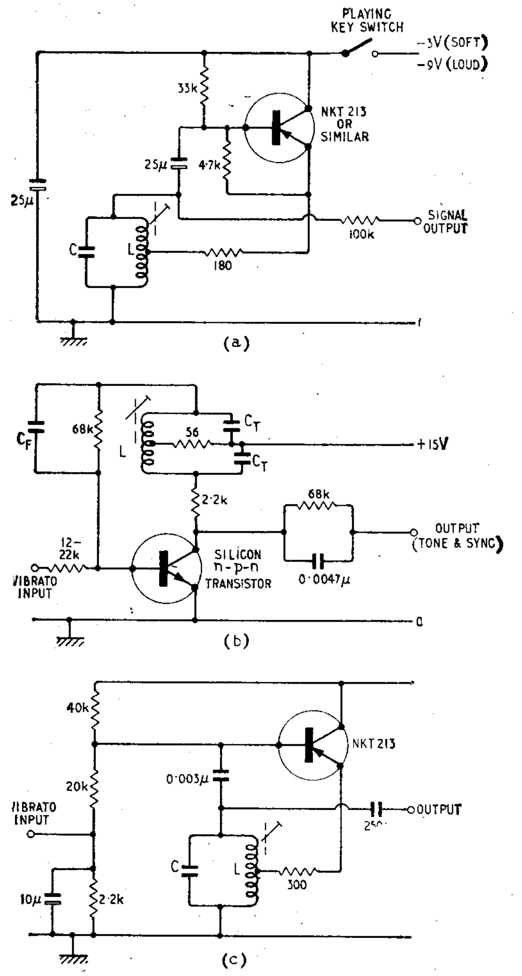


Fig. 2. Tone-generator LC oscillators: typical transistorised circuits used commercially for direct generation of tone signals ("free-phase") or for providing master synchronising frequencies ("divider"): (a) free-phase, emitter-coupled, Hartley (L, C = 3H, 2µF at 64 c/s to 30 mH, 0.05µF at 4,080 c/s) (Gulbransen); (b) master, collector-coupled, modified Hartley (L, C_T, C_F = 0.9 H, 0.1 µF, 0.047 µF at 740 c/s to 0.4 H, 0.066 µF, 0.033 µF at 1397 c/s) (Heathkit); (c) master, emitter-coupled, Hartley (L, C = 350 mH, 0.016 µF at 2 kc/s to 350 mH, 0.004 µF at 4 kc/s((Kawai).

days is the master oscillator in the divider type of organ. A set of twelve of these is required, each producing a frequency corresponding to one note in the chromatic scale in the highest octave of the instrument keyboard. The master oscillator not only generates a tone of its own frequency, but also provides synchronizing pulses to a string of divider circuits. Each divider produces in turn an output one octave lower than the one above it. Almost always the master oscillators are some form of Hartley circuit, with either collector-base or emitter-base feedback.

Fig. 2(b) gives a representative example of a collector-base feed-back master oscillator. This one is designed to provide chromatic frequencies from F ♮740 c/s to F 1,397 c/s by selecting suitable values of the split tuning capacitors, C_T, the feedback capacitor, C_f, and the inductance, L. For any selected frequency, the final exact tuning is effected by adjusting the slug inside the ferrite pot core. The rail voltage of 15 V is typical of the range of 15-18 V normally used with mains-operated instruments. The transistor used illustrates the trend towards silicon n-p-n devices. The overcoupled oscillator feedback gives a non-sinusoidal waveform rich in upper harmonics. (The "vibrato input" will be explained later below.)

Fig. 2(c) gives a typical example of the emitter-coupled master oscillator, which tends to be more common than the collector-coupled one. The example shown is designed to work in the frequency range 2-4 kc/s. For traditional C-to-C organ manuals this is the most common range for master oscillators to be set in. In economy organs, however, the oscillators may be found set from 1-2 kc/s, and in some specialized organs with wider tonal range 4-8 kc/s is used.

Although most organ designers now use some form of Hartley, there is wide variation in tank circuit L, C values. At 2 kc/s, the bottom end of the normal master oscillator compass, inductance values from 20 mH to 350 mH will be found. This wide design variation can be explained by the conflicting requirements of the tank circuit. First, to keep the organ reasonably in tune, it is best to have as high a Q as possible; but not so high as to make it difficult to frequency-modulate the oscillator for vibrato. The tendency nowadays is towards lower L and higher C. The higher capacitance tends to swamp out the effects of varying transistor junction shunt capacitances and makes the oscillator stability relatively independent of the transistor. Miniature, low-voltage polystyrene capacitors are becoming standard in the tank circuit, because they have a negative temperature coeffi-

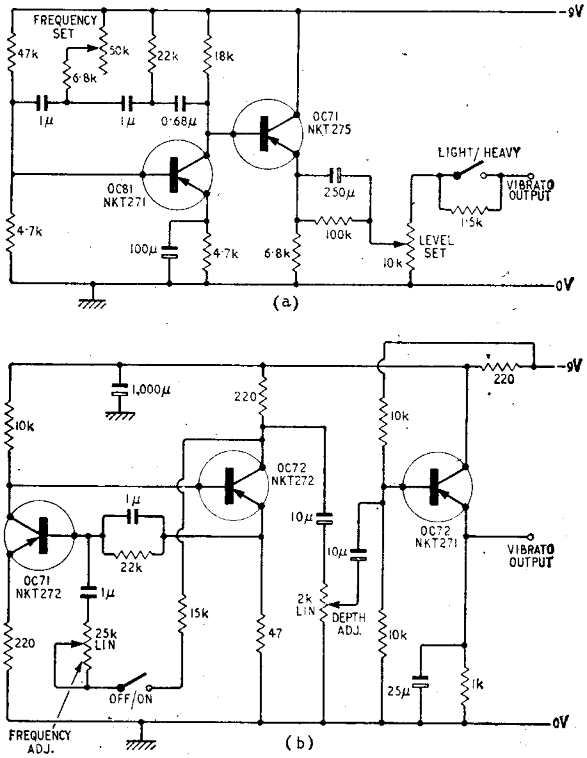


Fig. 3. Vibrato-tremolo RC oscillators: typical transistorised circuits used to provide modulating voltage for vibrato (frequency modulation) or tremolo (amplitude modulation) variation of tones at approximately 6 c/s: (a) RC phase-shift circuit (Watkins Electric Music); (b) Wien-bridge (Jennings Musical Industries).

cient offsetting the positive coefficient of the ferrite pot-core inductance.

Oscillators for vibrato or tremolo drive.—In a pipe organ, the "tremulant" stop induces a cyclic variation of both the pitch and the loudness of a speaking pipe at a rate of about 6 c/s. This is simulated in electronic organs by applying "vibrato" (subsonic modulation of the tone frequencies) or "tremolo" (modulation of the amplitude), or both combined. The modulation drive is usually provided by some form of *RC* oscillator which we will in future refer to as the vibrator oscillator, although tremolo drive is provided by the same sort of oscillator). This oscillator has to provide several volts r.m.s. of pure sinewave from an impedance low enough to be able to drive a number of loads. The pure sinewave is necessary because distortion can give rise to an unpleasant roughness in tone.

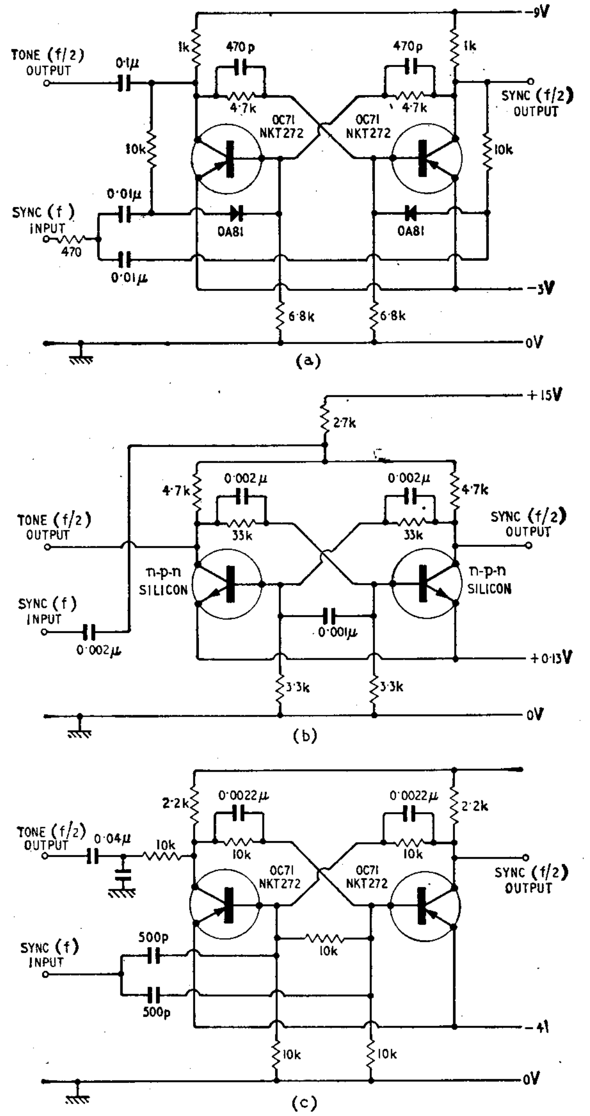


Fig. 4. Square-wave dividers: selection of commercial transistorised bistables: (a) diode-triggered (Watkins Electric Music); (b) collector-resistance-triggered (Heathkit); (c) base-resistance triggered (Jennings Musical Industries).

For low distortion, most vibrato oscillators use either an *RC* phase-shift or a Wien-bridge type of oscillator. For low output impedance, they usually have a buffer stage following the oscillator proper. Fig. 3(a) illustrates a typical *RC* phase-shift circuit employing only two transistors, while Fig. 3(b) shows a Wien-bridge circuit using three. Both include an emitter-follower output buffer stage and have potentiometer provision for presetting frequency in a range around 6 c/s and for adjusting the output level.

DIVIDERS

In divider-type organs, the master oscillators discussed earlier control a string of dividers, each giving a frequency one-half of that above it in the string. There are two main types of such dividers: "square-wave" and "sawtooth," characterized by their output waveshapes. Sawtooth dividers are more versatile in terms of their harmonic content, but nowadays most commercial organs use square-wave dividers. This is because they are cheap and convenient in manufacture, since the divider circuits are all identical wherever they lie in the divider string, and do not require any setting up by adjustment or selection of component values. With sawtooth dividers, component values change as frequency changes along the string, and most designs call for setting up each divider individually.

Square-wave dividers.—Almost without exception, the square-wave divider used in an electronic organ is some form of bistable multivibrator. The main variation between different designs lies in the method of trigger pulse steering. The normal computer-type diode steering is illustrated in the example given in Fig. 4(a). Fig. 4(b) illustrates the other main trigger-steering method used, the common collector resistor. A third variant also used is the divider of Fig. 4(c), where the trigger pulses are steered to the appropriate base with the aid of a 10 kΩ cross-coupling resistor between the bases. (Interested readers will find explanations of these and other pulse circuits described in this article set out in full in "Elements of Transistor Pulse Circuits" by T. D. Towers, Iliffe Books, Ltd., 1965.)

Sawtooth dividers.—Designers aiming at sawtooth output from their dividers usually adopt a blocking-oscillator divider of the type shown in Fig. 5 (a). An alternative sometimes used is to start with a square-wave divider and convert its output into a sawtooth with the circuit of Fig. 5 (b). Both the direct sawtooth divider and the converted square-wave circuit call for setting up by selection of the integrating capacitors used for different frequencies as indicated in the caption under Fig. 5.

Organ designers divide themselves firmly into "square-wave" or "sawtooth" proponents. An interesting compromise is the Heathkit (Thomas) arrangement whereby the square-wave output at any frequency is mixed with 50% of the octave above before passing to the key-switches and tone filters. This produces a staircase waveshape which partakes of some of the character of both square-wave and sawtooth.

GATES CONTROLLED BY KEYSWITCHES

As semiconductor devices get cheaper, designers are making more and more use of diode or transistor gates to route tone signals after they leave the generators.

Transmission gates.—When you use an ordinary key-switch to make and break an a.c. signal path from the

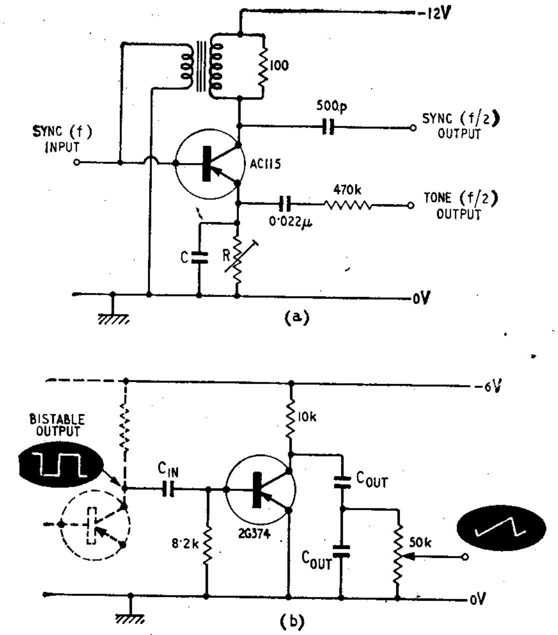


Fig. 5. Sawtooth dividers: transistorised circuits used commercially: (a) monostable blocking-oscillator triggered to run at half-frequency of next higher frequency oscillator ($C = 0.01$ μF at high to 0.047 μF at low frequency) (Harmonics); (b) integrator used to convert square-wave output of bistable to sawtooth at same frequency (C_{in}, C_{out} = 0.001, 0.022 μF at high to 0.047, 1.0 μF at low frequency) (Livingston-Burge).

generators, you run head on into the problem of objectionable keyclicks. The tendency nowadays is not to feed the signal to the keyswitch but to a semiconductor transmission gate which is controlled by the keyswitch. Click filter networks can be included quite simply in such gates, as indeed can arrangements for varying the rate of turn-on and turn-off of the signal. Moreover, such gates can easily be paralleled so that any number of signal lines can be switched on and off with one switch contact.

The best known example of this type of transmission gate is that given in Fig. 6 (a). In this, for ease of illustration, the single keyswitch is shown controlling only two signal lines, f_1 and f_2, but there is virtually no limit to the number of lines that can be so controlled. Consider the f_1 signal line. When the keyswitch is off, the key busbar is at zero volts d.c., and diodes D2, D4, being silicon diodes and not forward biased, present a high impedance in the signal line, so that there is no f_1 output. When the keyswitch is closed, the voltage on the key busbar rises towards +15 V with a time constant approximately equal to $R_1 C_1$ (7.5ms in this case). A positive voltage on the key busbar forward-biases the diodes D2, D4, and permits the f_1 input to pass through to the output. When the keyswitch is released, the key busbar voltage drops towards zero with a time constant roughly equal to $R_2 C_1$, if the decay busbar is set to its +OV "normal" position. This delays the switch-off of diodes D2, D4, according to a 50ms time constant in this case. If the decay busbar is set to +15 V, however, the diode D1 remains reverse biased and the capacitor C_1 can discharge only through the network of 47 kΩ resistors across the signal lines. This gives a length "sustain" before diodes D2, D4 eventually cut off and the signal output ceases.

Percussion gates.—The other type of key-controlled gate circuit is the percussion amplifier, of which a typical example is given in Fig. 6 (b). In this, when the key switch is open (off), transistor Tr1 is on and Tr2 off, with the result that Tr3, Tr4 are biased off and no tone passes through to the output. When the keyswitch is closed, the charged capacitor C applies a +ve pulse to the base of Tr1 which switches over the monostable circuit Tr1-Tr2, and biases Tr3, Tr4 on for a length of time set by the monostable recovery time. At the end of this time the monostable switches back and the amplifier is cut off again. A variety of percussion effects can be achieved by switching in different values for the monostable timing capacitor C_T, or by cross-coupling feedback to make the monostable run as an astable.

MODULATORS

The gates described above can be regarded as forms of signal modulators. Semiconductor signal modulators

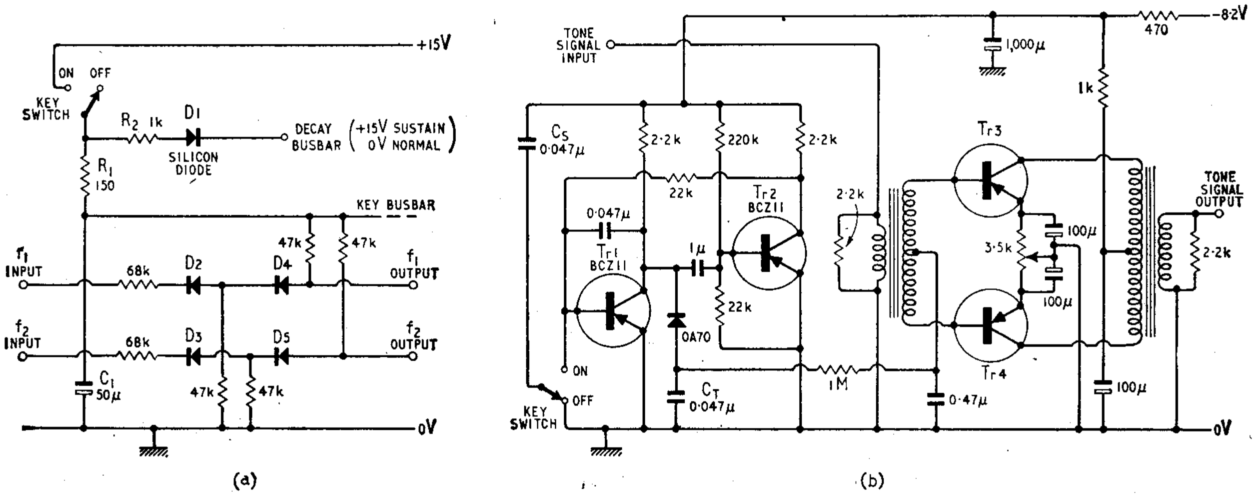


Fig. 6. Semiconductor gating circuits controlled by keyswitches: (a) diode transmission on gate controlling tone generator outputs (Heathkit); (b) percussion transmission gate controlling composite tone transmission through a transformer-coupled amplifier after tone forming (Jennings Musical Industries).

are also used for various other purposes in electronic organs.

Swell control.—The simplest way to control the volume output of an organ is with a foot-pedal controlled potentiometer, but this gives rise to such difficulties of hum and noise that many designers use something more refined. With present day low-voltage organs, the tendency is to use a light-dependent-resistor (l.d.r.) circuit. Fig. 7(a) gives a typical example of this. The l.d.r., type ORP12, is inserted in the signal line and illuminated by a lamp through a shaped shutter. The shutter is controlled by a foot pedal. As the pedal is depressed,

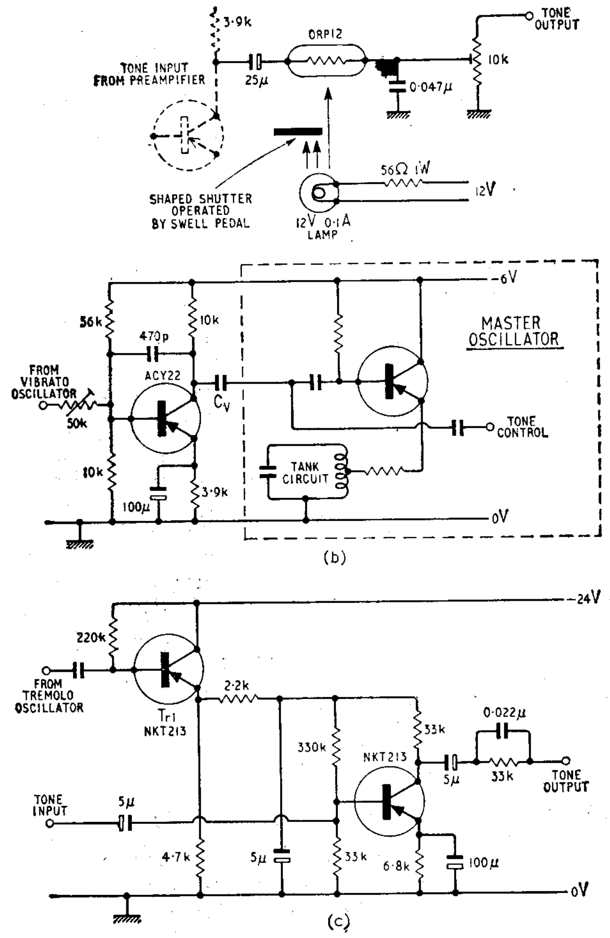


Fig. 7. Transistorised swell, vibrato and tremolo control circuits: (a) swell (volume) control by varying resistance of light-dependent-resistor with light beam (Livingston-Burge); (b) vibrato frequency-modulation of tone generator oscillator by means of special "reactance" modulator transistor stage (Livingston-Burge); (c) tremolo amplitude-modulation of amplifier stage gain by modulating supply voltage to stage (Ormston-Burns).

the shutter allows more and more light to fall on the l.d.r., and its resistance falls from several megohms to several hundred ohms. Also the series resistance of the ORP12 falls, the attenuation on the signal line decreases and the signal output rises correspondingly.

The other common way to use the l.d.r. is to set it at a fixed distance from the lamp and vary the light intensity

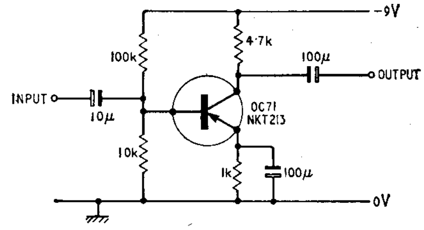


Fig. 8. Transistor general-purpose low-level preamplifier. (Jennings Musical Industries).

by varying the voltage across the lamp with a foot-controlled potentiometer.

Vibrato modulation.—Earlier we discussed the design of the vibrato oscillator. Now we consider how it frequency-modulates the tone signal. In the simplest arrangement, the vibrato drive voltage is applied through an isolating resistor to some point in the oscillator circuit, usually the transistor base (as in Figs. 2(b) and 2(c) above). Some designers interpose a modulator amplifier between the vibrato oscillator and the tone-generator oscillator; Fig. 7(b) gives an example of this.

Tremolo modulation.—Where tremolo (amplitude modulation) is used, some such arrangement as Fig. 7(c) is common. The d.c. supply to an amplifier stage in the main signal path is varied at the tremolo oscillator subsonic frequency through an emitter-follower modulator stage, Tr1. This causes the gain of the amplifier stage, Tr2, to vary correspondingly, and gives a tremolo effect in the tone output.

AMPLIFIERS

Apart from the special circuits discussed above, electronic organs use fairly conventional preamplifiers, power amplifiers and power supplies.

Preamplifiers.—To make up the signal losses incurred in the various switching and tone-forming circuits, most electronic organs use a number of preamplifiers. Fig. 8 shows a typical design.

Power amplifiers.—Electronic organ power amplifiers generally give between 10 W and 100 W output. They fall into two classes, mains and battery driven. Mains amplifiers tend to work across high (24-48 V) positive and negative d.c. rails, with a driver transformer and direct-coupled transformerless output. Battery/mains types tend to work from a lower voltage single-sided supply, usually 12 V, with both a driver transformer and an output transformer or choke to enable a standard 15-ohm speaker to be used with the low voltage d.c. supply.

FUTURE DEVELOPMENTS

The various circuits described above are taken from current models of commercial organs, but recent developments are pointing to the likelihood that many of these will be superseded soon by integrated circuits, and new devices such as field-effect transistors. The time is not far distant when a five-stage divider, which can now occupy a 5in × 3in printed circuit board, will be produced commercially in a block circuit smaller than a sixpence.

LETTERS TO THE EDITOR

The Editor does not necessarily endorse opinions expressed by his correspondents

Attenuation in Coaxial Cables

MR. WADE'S article in the April issue was both interesting and informative. However, the graph he uses in Fig. 5 gives a rather pessimistic picture of the attenuation on a lossy line. A reflection coefficient of 0.6 measured at the transmitter end of a short-circuited line, would indicate a total attenuation over the forward and return paths, of 2.2 dB. This figure is actually twice the cable attenuation, and the values for attenuation given by Fig. 5 should be halved.

Chippenham, Wilts. D. P. GRAY

The author replies:—

May I draw upon Reference 4 mentioned in the article (i.e. "VHF Line Techniques" by C. S. Gledhill, pp. 14-17), where the theory associated with the graph of Fig. 5 is shown. Summarizing this theory, and modifying it for convenience to yield equations in terms of line *currents* instead of voltages (because we are considering a short-circuit termination), we can state the following.

If the forward current at the transmitting end of a cable of electrical length l is I_F, it will be attenuated by a factor $e^{-\alpha l}$ at the short-circuit end, so that the current flowing through the short-circuit is $I_F e^{-\alpha l}$. This current will be totally reflected, and again attenuated by $e^{-\alpha l}$ on its return to the transmitter. Hence the reflected current at the transmitter will be $I_R = I_F e^{-\alpha l} . e^{-\alpha l} = I_F e^{-2\alpha l}$.

The reflection coefficient k measured at the transmitter end of the cable is defined as the ratio I_R/I_F, from which it will be seen that

$$k = e^{-2\alpha l}$$

$$\therefore \alpha l = \tfrac{1}{2} \ln \frac{1}{k} \text{neper}$$

$$\text{or } \alpha l = 10 \log \frac{1}{k} \quad \text{dB}$$

This last equation is the one from which Fig. 5 was plotted. It should be noted that αl is the attenuation over a length l, and not over $2l$; the fact that the signal has travelled over both the forward and return paths before the reflection coefficient is measured is allowed for in the index $-2\alpha l$.

If Mr. Gray still needs to be convinced on this point, let us calculate from first principles the reflection coefficient to be expected at the transmitter for the particular example quoted. If the cable has an attenuation of 2.2 dB and the magnitude of the forward current at the transmitter is, say, 100 mA, this current will be attenuated by 2.2 dB to 77.6 mA at the short-circuit termination, and after reflection the wave will be attenuated by a further 2.2 dB to 60.2 mA at the transmitter. Hence the reflection coefficient is 60.2/100 = 0.602, which to all intents and purposes is equal to the value of 0.6 stated in the article.

I understand Mr. Gray to imply that the cable attenuation for a reflection coefficient of 0.6 should be only 1.1 dB. If we now repeat the above example, assuming an attenuation of 1.1 dB, the short-circuit load current will be 88.1 mA, and the reflected current at the transmitter will be 77.6 mA, in which case $k = 0.776$; not 0.6 as stated by Mr. Gray. However, the value $k = 0.776$ (or shall we say 0.78) is in agreement with Fig. 5, and accordingly I maintain that this graph is correct. I wonder if Mr. Gray has confused current with power when using equations involving dB, remembering that the reflection coefficient is expressed in terms of a current or voltage ratio, and not a power ratio?

May I take this opportunity of drawing attention to three small errors which did find their way into the article. Two are concerned with the graph shown in Fig. 9. The word **LOAD** towards the bottom right-hand corner has become detached from **S.W.R.** at the top, i.e. each curve represents the s.w.r. at the load. Secondly, the horizontal and vertical dotted lines (corresponding to the numerical example given in the text) should intersect on the **LOAD SWR = 10** curve, and not just above it as shown. Finally, my call-sign is G3NRW.

Chelmsford, Essex. A. I. H. WADE

Temperature Indicator

IN recent times it has become common to refer to all voltage regulator diodes which use reverse voltage breakdown as Zener diodes. A look at the graph in Fig. 1 will show that only those diodes which operate at the lower voltages are truly Zener diodes. Zener breakdown occurs in relatively highly doped junctions where

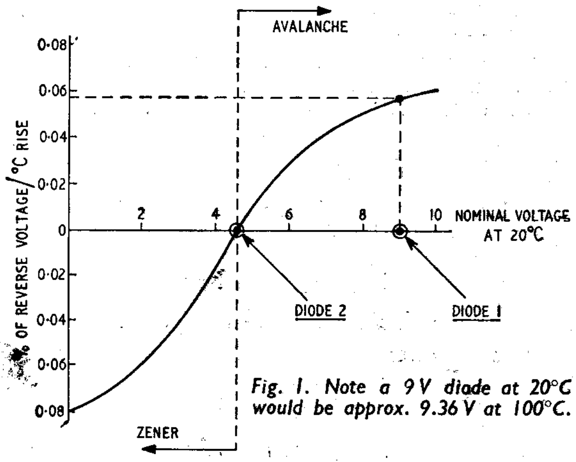


Fig. 1. Note a 9 V diode at 20°C would be approx. 9.36 V at 100°C.

the depletion layer is very thin, the electric field stress at this junction can be in the order of 500 kV/cm, this being the main cause of the breakdown, this occurs at lower voltages as temperature increases.

Avalanche breakdown occurs within the depletion layer when the current carriers forming the reverse saturation current reach a sufficiently high velocity to create other carriers by collision. In this type of breakdown the breakdown voltage rises as the temperature increases.

Readers may be interested in a temperature gauge developed in an attempt to use this increasing voltage as an indication of diode temperature (Fig. 2).

One side of the meter is held at 4.7 V by diode 2. The voltage of this diode being independent of ambient tem-

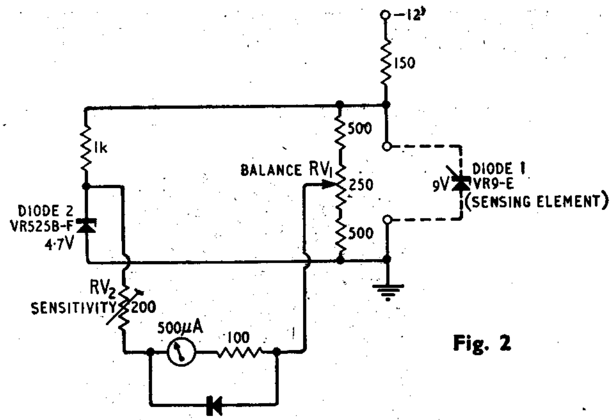


Fig. 2

perature changes. Diode 1, an avalanche diode, forms the temperature sensing element and as the temperature rises unbalances the bridge. The original gauge was required to read from 0 to 100°C. In setting up the gauge diode 1 was reduced in temperature to 0°C. With RV2 set for maximum sensitivity, the balance control RV1 was set for zero reading on the meter. Diode 1 was then increased in temperature to 100°C. and the sensitivity control adjusted for f.s.d.

The gauge has been used for several months as a water temperature gauge for a petrol engine.

The avalanche diode used was a 5 W stud mounted type, screwed into the thermostat housing by its 2 B.A. thread. Diode 3 was used for meter protection during starting conditions, where the battery voltage could possibly fall to a low value.

The gauge has been found to be accurate and linear, and also free from ambient temperature changes. The device has many possibilities, it can be set to give a positive or negative output about any pre-set temperature and should work well in a control system.

Preston, Lancs. J. B. LEIGH

Information Explosion

I AM deeply interested in the problem which you raise in your April editorial—"Can the Information Explosion be Controlled?" I nevertheless doubt whether this question is as crucially important as many people think. Let me elaborate.

An explosion is a transient phenomenon: this is not. Human society has been developing a long time, and the rate at which it has been changing has also apparently been steadily increasing. We have become worried because the rate of change is now severely taxing human adaptability and it may not be long before we cannot cope with it. This is the wider context in which the rapid proliferation of technical information must be considered: information of this kind is, in fact, only a single feedback path in the far larger and more complex closed-loop relationship between man and his culture. This overall system is the only one that really matters. To concentrate on the information problem alone is rather like tackling a runaway horse and cart by putting roller bearings on the wheels.

In these circumstances I am not clear what "control" really means. Of course we must try to *accommodate* this expansion, but control can only be effectively applied to our total, increasingly technically based culture. To restrain cultural development would, however, thwart the drive of human imagination and enterprise and we would be sure to suffer one way or another—perhaps in frustration and bloodshed, or less dramatically in complacency and decadence. If we are, in fact, more interested in reins than in roller bearings, we must shape our culture into a continuously expanding phenomenon which has intrinsically safe, self-adaptive, characteristics. Such a culture will, I believe, become an increasingly important concept in the next decade or so; but it will be a long time before we can realize a practical model.

I am sure that, in the detailed problems of information dissemination, computers and verbal techniques will, as you have suggested, become increasingly important. But literature of one kind or another has been with us a long time, and we have become used to the presentation of the printed word. In their own field engineers are now skilful at flipping pages, jumping paragraphs, extracting key data, and perhaps referring back to check earlier qualifications. In the 12th Graham Clark Lecture, Lord Snow pointed out that engineers as engineers can get on very comfortably without words, and that as students it doesn't matter if they mis-spell the vocabulary they use. All these considerations suggest to me that improved information paths may increase the quantity of material handled, but that the quality may seriously deteriorate. Perhaps, after all, our culture has some inherently stable characteristics. It may be able to sense when it is wise for the roller bearings to become clogged with sand!

Welwyn Garden City, Herts. CRAWFORD ROBB

Amplifier Noise Level

WHEN at the recent Audio Fair I requested a very well known person in the audio industry if he would adjust the volume and treble controls of a transistor amplifier he was demonstrating to maximum, and allow the audience to estimate the noise level it produced.

I was told that "this is meaningless" and that the volume control would never normally be in the maximum position.

I would like to suggest, in your columns, that there are very good reasons for this subjective test to be of value. Unless the noise figures are very good (better than 70 dB below 30 W r.m.s.), two comparable transistor amplifiers may give very different subjective impressions if in one, the effect is concentrated in the low frequencies, due to excess or flicker noise from transistors. And not all gramophone records have the ideal dynamic range (70 dB), or they may be undermodulated, or they may be recordings of solo instruments, played pianissimo with silent intervals (that is except for amplifier and tape noise), all requiring a higher setting of the volume control to maintain the same average listening level. If these conditions are not allowed for in design, then a higher power amplifier is not a higher quality amplifier as I believe it also should be. A maximum acceptable noise power delivered to the loudspeaker should be the aim, more difficult perhaps to achieve than a reasonable signal-to-noise ratio.

Decca Radio & Television, London, S.W.8. R. C. DRISCOLL

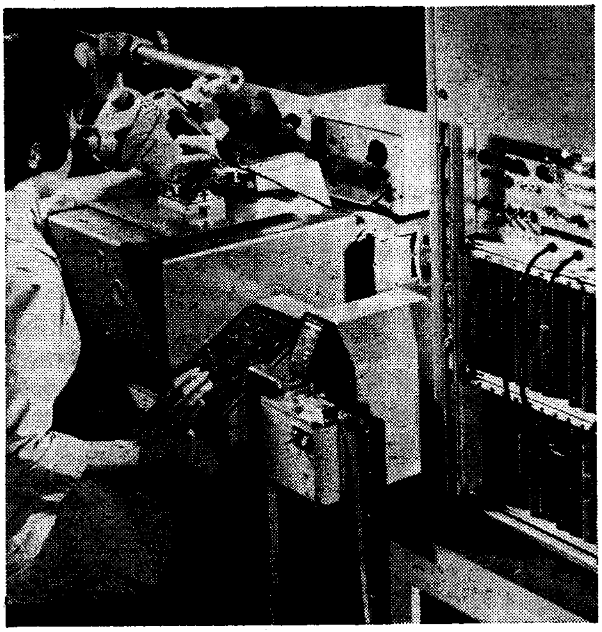
HANOVER FAIR

We regret it has not been possible for us to include in this issue a report on the Hanover Fair as we had planned.

A SPARK MICRO-ENGRAVING TECHNIQUE FOR THIN-FILM CIRCUITS

A SPARK micro-engraving technique for use in the production of thin film circuits has been developed by Standard Telephones and Cables Limited. The engraving equipment is positioned controlled from a digital punched paper tape programme, and a significant feature is that all the scaled-up draughting processes normally associated with thin-film circuit production are entirely eliminated because the programme can in fact be prepared from a simple dimensioned sketch drawn by the circuit design engineer on squared paper. The machine can engrave tracks down to 0.002in wide with a positional accuracy of 0.0003in. An important aspect of the use of punched tape is the possibility of tape preparation at a remote location—the digital information can be sent from one place to another over the telex system. The use of punched tape also suggests another interesting possibility—feeding circuit design parameters directly into a computer which will calculate the necessary dimensions of the circuit elements and control the action of the micro-engraving equipment.

Spark micro-engraving is a technique for cutting lines in thin films of electrically conductive material. Although a spark engraving probe or stylus, traversed while in contact with the film surface, does produce the required erosion effect the process applied in this way is unreliable because of the tendency to short-circuit the stylus to the film surface, which stops further erosion taking place. The method used by S.T.C. is to vibrate the stylus perpendicularly to the plane of the surface being engraved, so as to make and break contact with the film and produce momentarily critical gap conditions for spark breakdown during each half-cycle of movement. The spark energy is provided by a small capacitor placed across the gap and charged from a low voltage d.c. supply through a resistance which is large enough to prevent thermal damage to the film due to flow of d.c.



The illustration shows the equipment for micro-engraving of thin-film circuits. The control console for positioning of the table is shown on the right. In the centre is the tape reader and control box. The table with engraving head is on the left.

Optimum voltage and component values are dependent on type and thickness of the material being engraved. For example, typical values for nichrome films less than 100 Å thick are: 24 volts, 50 pF and 100 kΩ. The engraving is carried out under the surface of a dielectric fluid which provides high breakdown strength and restricts the breakdown to an area closely defined by the stylus diameter. Vibration of the stylus is effected by a piezo-electric element which is caused to vibrate in a longitudinally bending mode by the application of a driving voltage to metallized electrode areas. When energized in a resonant mode, e.g. 4.5 kc/s, a tip movement of approximately 25 μm is obtained with 100 volts a.c. input. One end of the transducer is rigidly clamped to a balanced arm and the stylus is mounted at the free end. The tip diameter is precisely controlled, and although any lateral movement of the tip affects positional accuracy, and hence the edge definition of the engraved line, some resilience is necessary in the stylus mounting to avoid deformation of the tip by continuous hammering action. The stylus is housed in jewelled bearings and a 1 μm radial clearance provides free longitudinal movement. The engraving speed which can be used is dependent on the volume of material to be removed. Typical values for a 50 μm wide track are: 1 cm/s for 70 Å thick nichrome and 1 mm/s for 3000 Å thick gold film.

Relative movement between the stylus and thin film substrate is effected by a two-axis numerically controlled table similar to that of a conventional machine tool. An open-loop control system is used incorporating a stepping motor as the drive element and precision leadscrews as the reference lengths. One electrical input pulse fed to the motor drive circuitry causes the motor shaft to make one step of rotation (equivalent to 1.8°). This rotates a leadscrew and moves the table a distance of 0.0005 in. (This is the resolution of the table.) The stepping motors have a strong in-built detent action, and provide an accurate step movement provided the frictional forces opposing the motion of the table are not larger than the detent force of the motor.

The data tape control system for the table consists of two registers, one containing the co-ordinates of the actual table position and the other containing the co-ordinates of the position to which the table must move. The motors and position register are pulsed until the two registers contain identical co-ordinates. This indicates that the table has reached the required position.

Preparation of the tape programme consists of punching the co-ordinates of the various change points in relation to a suitable origin. The origin must be in the extreme position in each axis to which the table is required to move during the engraving process because the control unit cannot handle negative co-ordinates as would be required if the table were to be expected to move past the origin. The origin must also be an easily identified point on the substrate so that the engraving head can be set in relation to the substrate before engraving is commenced. The data consists of groups of nine numbers punched in standard G.P.O. telex code; the first numeral controls the machining speed and energy input. The next four represent the Cartesian ordinate expressed in units of 0.0005 inch. The last four numerals in the group represent the Cartesian abcissa in the same units.

At present the process has two main applications to thin-film circuits. These are the machining of passive components to close tolerance values after deposition and the manufacture of photolithographic masters.

By direct machining, component tolerances which can be obtained are 0.1% for resistors and 0.5% for capacitors. For photolithographic masters, accurate artwork or photography is eliminated and a master can be engraved in a hard adherent metallic layer on a standard 2 in × 1 in glass slide in about 10 to 15 minutes.

PICKUP ARM DESIGN—2

DESIGN AND MOUNTING OF ARMS FOR MINIMUM DISTORTION DUE TO LATERAL TRACKING ERROR

By J. K. STEVENSON,
B.Sc., Grad.Inst.P., Grad.I.E.E.

Concluded from page 218 of May 1966 Issue

In part 1, the author maintained that whilst distortions in disc reproduction are gradually being reduced, tracking error distortion has not been. It was pointed out that it is possible to design and mount an arm so that distortion due to lateral tracking error is less than 1%. After discussion of distortion due to tracking error, design formulae are derived for offset angle and overhang for minimum tracking error distortion.

In part 2, two designs are presented, one for 7in discs and one for 7, 10 and 12in discs. Tracking error is shown to be critically dependent on mounting errors, and in view of this, the author outlines two mounting procedures. Pickup arm shape and optimum tracking mass are also considered.

RECORD players may be considered as being of two types, those suitable for all diameter records, and those suitable for only 7in records. Design values are now given for both types.

Design values for 7in discs

x was measured at the start and finish of a number of records of different makes and the values obtained are given in Table 1. x_{outer}, which was measured for 7in and 12in discs, was found to be fairly constant and varied at the most by 1/32in. The values for x_{inner} are representative values for the minimum distance from the record centre and x_{min} denotes the minimum distance in exceptional cases, being the minimum value obtained for x_{inner} from measurements on a batch of records of different makes and different types of music.

Pickup arms have been designed using equations 6, 7, and 8. In the design of an arm restricted to 7in records, x_o and x_2 were chosen as follows.

$$x_o = x_{inner} = 2.125\text{in},$$
$$x_2 = x_{outer} = 3.281\text{in}$$

x_m^2, x_p, x'_o and x_3 were then determined and the values are given in Table 2. Maximum distortion of a given modulation occurs at the start of a record, and also at $x = 2.49$in, and $x = 2.00$in. The distortion becomes zero at $x = 3.00$in and $x = 2.13$in. In this design, the distortion is set to zero at the normal finish of a 7in disc (a desirable feature) and increases to the maximum value at $x = 2.00$in. However, it is unusual for the modulated section of a 7in disc to continue as far as $x = 2$in.

Design values for all discs

In the case of a record player suitable for 7in, 10in and 12in discs, the situation is more involved. In considering the distortions from 7in discs, it must be remembered that the turntable speed for these records is different to that for 10in and 12in discs. Three designs were obtained as shown in Fig. 6, and as the offset angle and overhang vary fairly linearly between them, the extreme designs 1A and 1C provide limits for overhang and offset angle between which any value of either of these parameters may be chosen, and the value for the other immediately given. With the smaller offset angle (design 1C), the maximum harmonic distortion for 7in discs tracking at 45 rev/min is less than that for long-playing records, as seen from Fig. 7. As the angle increases,

TABLE 1
Extreme values of x obtained from measurements on discs (x in inches)

x_{min}	x_{inner}	x_{outer}	Discs considered
$2\frac{1}{2}$	$2\frac{5}{8}$	$5\frac{23}{32}$	10″, 12″ ($33\frac{1}{3}$ rev/min)
2	$2\frac{1}{8}$	$3\frac{9}{32}$	7″ (45 rev/min)

TABLE 2
Values of x used in pickup arm design (x in inches)

Design	Design values		Zero distortion	Maximum distortion (also at x_2)		Maximum tracking error		Application
	x_0	x_2	x'_0	x_3	x_f	x_m	x_m^2	
1A	2·500	5·719	4·684	2·279	3·260	3·422	11·710	7″, 10″, 12″ discs
1B	2·375	5·719	4·606	2·158	3·134	3·307	10·939	
1C	2·250	5·719	4·522	2·038	3·005	3·190	10·175	
2	2·125	3·281	3·001	2·004	2·488	2·525	6·379	7″ discs

In a design, the distortion of a given modulation is maximum at the largest value of x (x_2) and set to zero at x_o. Then:—

x'_o, x_o are the values at which the distortion is zero

x_2, x_p, x_3 are the values at which the distortion is maximum, where $x_2 > x'_0 > x_p > x_o > x_3$, as shown in Fig. 5 (last month).

then provided the overhang is adjusted accordingly, the maximum distortion for long-playing records gradually reduces, and the maximum distortion for 7in discs increases.* For the larger offset angle (design 1A), the maximum distortion for 7in discs is about $2\frac{1}{2}$ times as great as that for long-playing records. However, it is still less than 2% second harmonic even for an 8in pickup, and in view of the fact that other forms of distortion will be lower for 7in discs at 45 rev/min than for 10in or 12in discs at $33\frac{1}{3}$ rev/min, the design is still suitable. The values used in these designs are given in Table 2. Note that for designs 1A to 1C the tracking error changes from positive to negative for a long-playing record, and changes from negative to positive for a 7in disc. If we had disregarded 7in discs and obtained a design with minimum distortion for long-playing records, i.e. one for which the distortion changes from positive to negative and back to positive again, the improvement would be very small but the distortion from 7in discs would by comparison be enormous.

The intermediate design 1B is recommended although, as mentioned earlier, provided that the offset angle lies within the given range, the mounting may be considered as optimum with the smaller angle slightly favouring 45 rev/min records and the larger angle favouring long-playing records.†

45 rev/min records are slightly favoured insofar as most forms of distortion are inversely proportional to the groove speed or a power of the groove speed i.e.:—

$$\text{distortion} \propto \left(\frac{1}{u}\right)^n \propto \left(\frac{1}{sx}\right)^n, \text{ where } n \geqslant 1.$$

The maximum distortion of a given modulation (the total

*By maximum distortion, we mean the maximum distortion of a given modulation, and in our calculations of distortion, we consider an effective recorded velocity of 10 cm/sec, a typical maximum value. Values as high as 20 cm/sec corresponding to a peak recorded velocity of 28 cm/sec occasionally occur but only for brief periods, e.g. a clash of cymbals. The average recorded velocity is usually greater for standard 7in discs than for extended play (7in) and long-playing records. However, standard 7in discs are usually restricted to popular music in which harmonic distortions are less objectionable.

†One design may be best for one record and another best for a second record simply because the most heavily modulated passages, at which the largest distortion is most likely to occur, are at different values of x. However, the overall best design is clearly the one for which the maximum distortion of a given modulation is least.

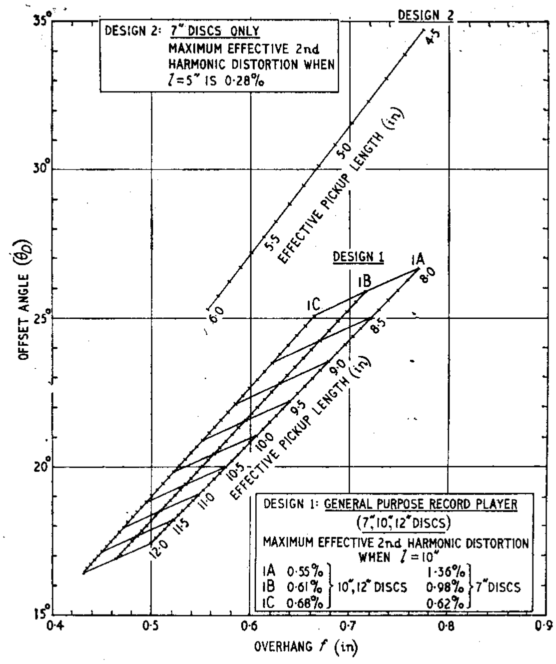
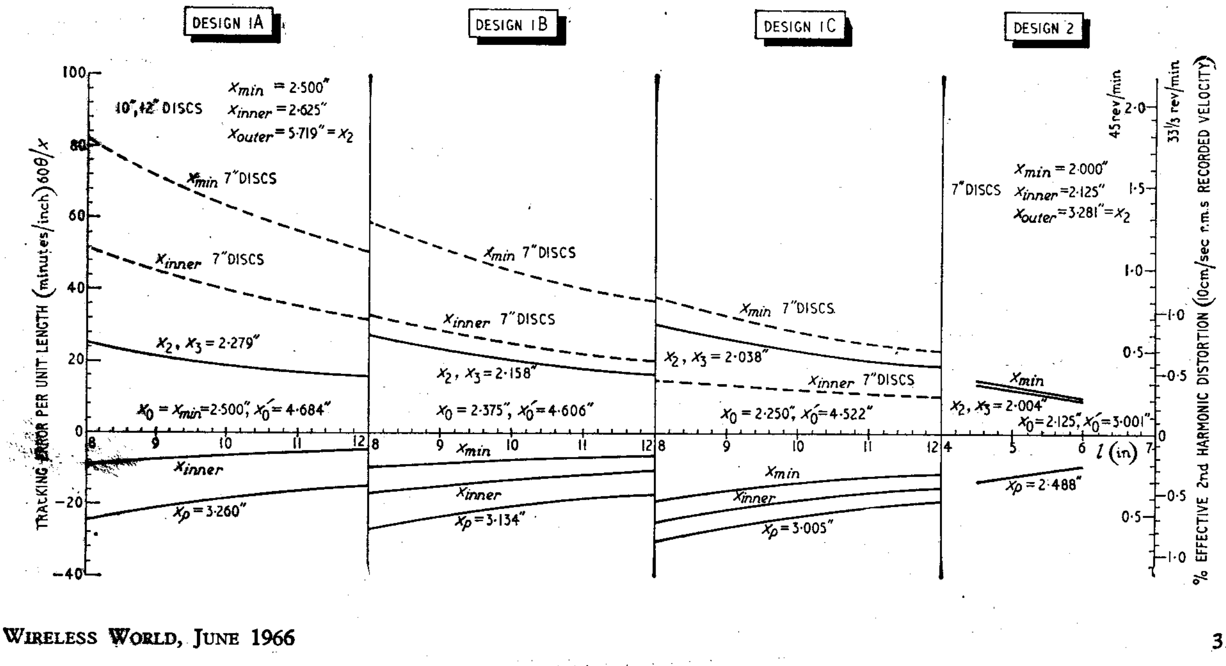


Fig. 6. Design values. The values of distortion quoted correspond to a 10 cm/s r.m.s. recorded velocity. Of the three designs 1A, 1B and 1C, 1B is recommended. The slightly greater harmonic distortion for 7in discs is counteracted by a reduction in distortion from other causes, as a result of the faster turntable speed. For an 8in pickup arm mounted as suggested, the distance between the edge of a 12in disc and the centre of the arm pivot is $1\frac{1}{4}$ in. A shorter distance corresponding to a shorter arm is impractical if adequate compensation is to be made for side-thrust.

Fig. 7. Variation of 2nd harmonic distortion with x. The values of distortion are typical maximum values, as distinct from occasional peak values due to exceptionally heavy modulation.



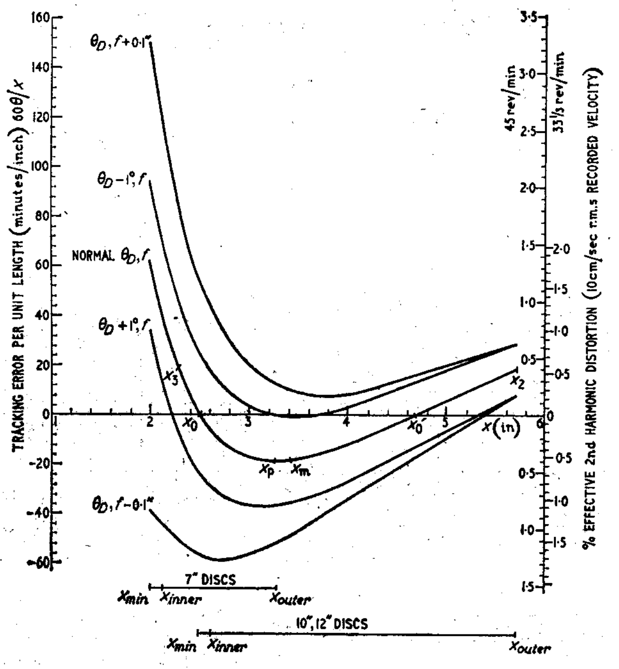
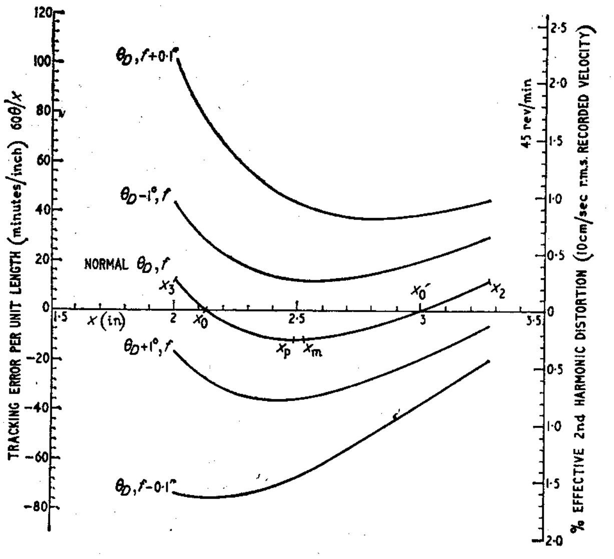


Fig. 8. Variation of 2nd harmonic distortion with changes in θ_D and f from the values given for design 1A in Fig. 6 (l = 10in). For other values of l, multiply harmonic distortion by 10/l (approximate). The same variations are applicable to designs 1B and 1C. It is seen that if f is too small (bottom line), $\phi/x(=[\theta-\theta_D]/x)$ is negative over most of the range of x. The tracking error may therefore be reduced by reducing θ_D as suggested in Fig. 6.

Fig. 9. Variation of 2nd harmonic distortion with changes in θ_D and f from the values given for design 2in Fig. 6 (l = 5in). Design 2 is for 7 in discs only. For other values of l, multiply harmonic distortion by 5/l (approximate).



from all causes) occurs at x_{inner} and in extreme cases at x_{min}. Hence,

$$\left.\begin{aligned}\frac{1}{sx_{inner}} &= 0.0105\\ \frac{1}{sx_{min}} &= 0.0111\end{aligned}\right\}\text{ 7in records, 45 rev/min}$$

$$\left.\begin{aligned}\frac{1}{sx_{inner}} &= 0.0114\\ \frac{1}{sx_{min}} &= 0.0120\end{aligned}\right\}\text{ 10 and 12in records, } 33\tfrac{1}{3}\text{ rev/min}$$

The maximum value for $1/sx$ is therefore slightly less for 7in records at 45 rev/min than for long-playing records.

Effect of errors in mounting

Figs. 8 and 9 give the tracking error per unit length (ϕ/x) for typical values of l, 10in using design 1A and 5in using design 2. Also plotted are the tracking errors for a pickup arm mounted imperfectly so that the values of θ_D and f are not as recommended. From Fig. 8 it is evident that an error in f of ± 0.1in or an error in θ_D of $\pm 2°$ will more than double the maximum distortion due to tracking error for long-playing records. The tracking error at a given value of x varies fairly linearly with changes in θ_D and f so that the error corresponding to θ_D, $f + 0.05$in for example, is given by the value lying mid-way between the lines (θ_D, f) and ($\theta_D, f + 0.1$in.) The changes in tracking error due to variations in θ_D and f are additive so that the error per unit length corresponding to $\theta_D - 0.5°$, $f - 0.05$in is given by a line lying mid-way between the lines $\theta_D - 1°$, f and θ_D, $f - 0.1$in. These changes also apply to designs 1B and 1C (and intermediate designs). Fig. 7 should be consulted as this gives the maximum distortion of a signal of 10 cm/sec recorded velocity, for a pickup arm mounted as suggested, and the values of x at which it occurs. It is fairly clear from Fig. 8 that too large an overhang, f, or too small an offset angle, θ_D, will have a considerable effect on the distortion at the inner grooves of 7in discs. An increase in f of 0.1in will increase the maximum harmonic distortion by almost 2% at a radius of 2in.

For a record player restricted to 7in discs, the maximum distortion need not exceed 0.3% which is negligible. An error in f of 0.03in or an error in θ_D of 0.5° will more than double the maximum distortion as shown in Fig. 9.

When constructing and mounting a pickup arm, it is therefore clear that unless the values for the offset angle and overhang are very closely adhered to, the distortion introduced is liable to be as great as the maximum value of distortion given in the design.

Mounting procedure

In view of the considerable effect of errors in mounting pickup arms, two alternative methods of mounting are suggested for which an alignment protractor is required.

For a given design, the values of x at which the tracking error is zero are fixed and independent of the arm length.

(*a*) *Optimum mounting* (*design* 1B)

The distance of the pickup arm pivot from the turntable centre should be fairly accurately measured and the pickup mounted in such a manner that about 1/20in movement towards and away from the turntable centre is possible. Small slots should replace the normal mounting holes. Similarly, the offset angle which is initially adjusted as closely as possible to the design value should be

(*Continued on page 317*)

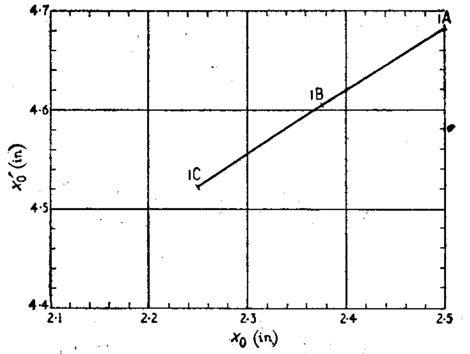


Fig. 10. Values of x at which tracking error is zero. As an aid to mounting for a design between 1A and 1C, values of x are given at which the tracking error is zero. A point one-third of the distance between 1A and 1B in Fig. 6, for example, corresponds to a point one-third of the distance between 1 A and 1B in this Fig. Tracking error is zero at x_0 and x'_0.

variable by about 1°. If the head is mounted to the arm with two nuts and bolts, then one of the holes need only be made fractionally wider. Unless the fit is very close, such a movement may already be possible. Clearly it is best to check the mounting as given below before widening any holes. Values of x are given to three decimal places for the benefit of precision engineers. Using an alignment protractor and a metal rule, the average constructor should be able to measure x to 1/50in without undue difficulty.

(*i*) Set x to 2.375in and adjust the overhang by moving the pickup arm pivot towards or away from the turntable centre, until the tracking error is zero.

(*ii*) Set x to 4.606in (4.6in) and observe whether the tracking error is positive or negative. If it is uncertain which is which, set x to $5\frac{1}{2}$in and the indicated tracking error is positive.

(*iii*) If the tracking error at $x = 4.606$in is positive, then both the offset and overhang are too small: the offset angle should be increased slightly. Similarly, if the tracking error is negative, then both the offset and overhang are too large: the offset angle should be decreased slightly.

The steps (*i*) to (*iii*) are then repeated until the tracking errors at both values of x are negligible (less than $\frac{1}{4}$°).

(*b*) *Alternative mounting* (*design between* 1A *and* 1C)

Although 1B is considered by the writer to be the best design, any design between 1A and 1C may be considered as good and will be optimum insofar as no other method of mounting can result in lower distortion (as a result of lateral tracking error) from *both* long-playing and 45 rev/min records. As seen from Fig. 7, design 1A favours long-playing records and design 1C favours 45 rev/min 7in discs.

It may not be considered possible, or it may be felt undesirable, to alter the offset angle, particularly if different pickups are used in the same arm. In this case, provided that the offset angle for a given value of l is within the range of values in Fig. 6, the following method should be used.

(*i*) Set x to 2.375in (x_o for design 1B), and adjust the overhang until the tracking error is zero.

(*ii*) Set x to 4.606in (x_o' for design 1B), and observe whether the tracking error is positive or negative.

(*iii*) If positive, then both the offset and overhang are too small for this design: move towards 1C which requires a smaller offset angle and overhang. If negative, move towards 1A.

(*iv*) To move towards 1C, x is set to a smaller value of x_o and f is reduced until $\phi = 0$. From Fig. 10, the value of x_o' corresponding to the new x_o is obtained. x is then set to the new x_o' (smaller than before) and the tracking error observed. To move towards 1A, a larger value for x_o is chosen and ϕ set to zero as before. The new x_o' is given from Fig. 10 and the tracking error observed.

(*v*) Return to stage (*iii*) and repeat until the tracking error at x_o' is negligible (less than $\frac{1}{4}$°).

If θ_D for a commercial pickup arm is greater than the 1A value, it has most likely been designed for minimum tracking error. The mounting can be improved by using a value of f slightly smaller than the manufacturer's suggested value so as to reduce the large positive tracking error at the inner grooves. An alignment protractor should be used and ϕ/x at $x = 2\frac{1}{8}$in reduced until equal in magnitude to the largest negative value of ϕ/x which will occur between $2\frac{1}{2}$in and 4in.* If θ_D is less than the 1C value, allowance has probably not been made for the faster turntable speed of 7in discs in which case, provided that θ_D is less than $\frac{1}{2}$° from the 1C value, Fig. 10 may be used with the line slightly extended. Otherwise ϕ/x at $x = 2\frac{1}{8}$in ($\phi/2.125$) should be adjusted until equal to the value at $5\frac{7}{10}$in ($\phi/5.7$). Both should then be greater than the maximum value occurring between $2\frac{1}{2}$in and 4in.†

Therefore, if a commercial pickup arm is to be mounted and the recommended methods given earlier are not used, ϕ/x at $x = 2\frac{1}{8}$in should be adjusted (by varying f) to equal the other maximum value between $2\frac{1}{2}$in and $5\frac{7}{10}$in. This will occur between $2\frac{1}{2}$in and 4in if θ_D is too large and at $5\frac{7}{10}$in if θ_D is too small.

Shape of pickup arm

Fig. 11a gives the shape of a conventional pickup arm. In order to reduce friction, the most suitable method of mounting is a single pivot (unipivot). A line joining the pivot and stylus tip should be horizontal otherwise the pickup cannot move vertically upwards when tracking a warped record. The centre of gravity of the assembly must lie below this line for stability and directly beneath it with the pickup untilted.

A means of lateral adjustment is usually provided and, in general, this consists of a device clamped to the arm at the pivot, being movable in the direction AA′ so as to apply a moment to the arm to remove any tilt. If a larger tracking mass is required, the counterbalance, or part of it, is moved towards the pivot. This decreases the anti-clockwise moment due to this counterbalance as seen from B, and a pickup which was previously correctly adjusted will tilt slightly in a clockwise direction causing a stylus tip to press on the inner wall of a record groove.

A suggested shape for a pickup arm is given in Fig. 11b. With the centre of gravity always lying along the length of the arm, lateral adjustment is no longer necessary when altering the tracking mass. Also, the pickup arm is slightly shorter for a given value of l, and will therefore be lighter.

A further advantage is that the lateral adjustment at

*Remember that the largest negative value of ϕ occurs at x_m and the largest negative value of ϕ/x at x_p, where x_m is just over $\frac{1}{8}$in greater than x_p. The difference in ϕ/x is very small (as seen from Fig. 8) in which case we may disregard x_p and divide the largest negative tracking error by the value of x at which it occurs (x_m).

†Note that if the largest value of ϕ/x between $2\frac{1}{2}$in and 4in is equal to the values of ϕ/x at $2\frac{1}{8}$in and $5\frac{7}{10}$in, the design will be optimum and lie about $\frac{1}{4}$ of the way between designs 1B and 1C (as seen from Fig. 7 with $x_s = 2\frac{1}{8}$in).

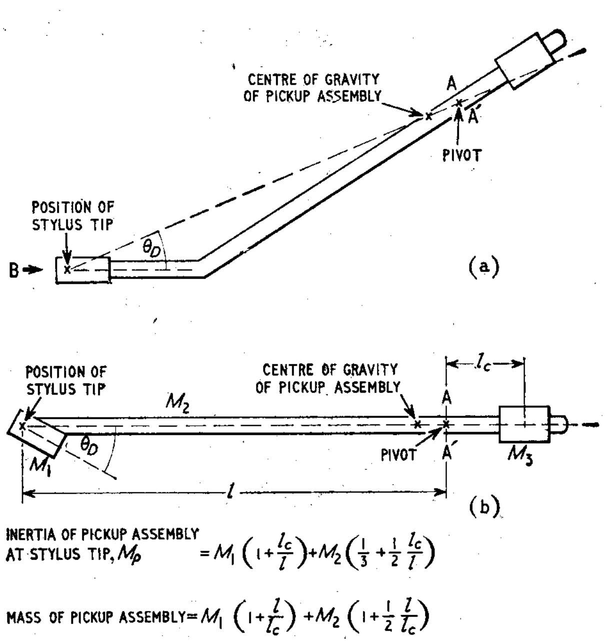


Fig. 11. (a) Shape of conventional pickup arm. (b) Preferred shape of pickup arm.

AA′, which is necessary to compensate for the offset of the pickup (unless the stylus tip is central), is much less than that for a conventional arm.

Compensation should be made for side-thrust, the clockwise moment which a moving record applies to a stylus. A practical explanation of this effect and how to reduce it has been given by J. Crabbe in one of a series of articles on pickups.[2]

Inertia of pickup assembly

A pickup arm should be constructed so that its inertia or effective mass at the stylus tip is as small as possible.† The inertia of the pickup assembly at the stylus tip is given by the moment of inertia of the pickup arm and counterbalance about the axis of rotation (the arm pivot) divided by the square of the effective arm length i.e.

$$M_p l^2 = M_1 l^2 + M_3 l_c^2 + \tfrac{1}{3} M_2 l^2$$

as given in Fig. 11b where M_p, M_1, M_2, and M_3 denote the inertia of the pickup assembly and masses of the pickup (including the mounting shell), arm, and counterbalance, respectively. The mass of the arm is considered to be evenly distributed along the length and, for simplicity, M_3 is assumed to be concentrated at its centre of gravity, and M_1 concentrated at the stylus tip. Taking moments about the pivot with the pickup balanced,

$$M_1 l + \tfrac{1}{2} M_2 l = M_3 l_c$$

Hence, the mass of the counterbalance is given by

$$M_3 = (M_1 + \tfrac{1}{2} M_2) l / l_c$$

and the inertia is as follows

$$M_p = M_1(1 + l_c/l) + M_2(\tfrac{1}{3} + \tfrac{1}{2} l_c/l)$$

M_1 is reduced as much as possible by manufacturers, and constructors should ensure that the pickup arm is as light as possible. The inertia of the assembly may be further reduced by using a heavier counterbalance nearer the pivot to reduce l_c. However, the mass of the pickup assembly is given by

$$M_1 + M_2 + M_3 = M_1\left(1 + \frac{l}{l_c}\right) + M_2\left(1 + \frac{1}{2}\frac{l}{l_c}\right)$$

and M_3 should not be too great otherwise unnecessary friction and wear are liable to occur in the pivot bearings.

Tracking mass for optimum reproduction with minimum record wear

The tracking mass, M_t, must be sufficient for the stylus to remain permanently in contact with the groove walls. At low frequencies we require the compliance of the stylus suspension, in particular the lateral compliance, C_l, to be as large as possible to enable the stylus to cope with modulations of large amplitude. At high frequencies we require the effective tip mass, M_{etm} (the inertia of the stylus and elastic support at the stylus tip) to be as small as possible so that the force on the stylus resulting from the considerable accelerations is much less than that due to the tracking mass. Also, we require the inertia of the pickup assembly, M_p, to be as small as possible so that forces on the stylus caused by tracking warped records are reduced to a minimum.*

To enable the stylus to follow the most difficult modulations, it is suggested that the following should be satisfied

$$M_t \geqslant M'_t$$

where

$$M'_t = \frac{5}{10^6 C_l} + 1000 M_{etm} + \frac{M_p}{40} \quad .. \quad .. \quad 10$$

M is given in gm and C in cm/dyne, e.g. if the compliance is 15.10^{-6}cm/dyne the effective tip mass $\frac{1}{3}$ mg, and the inertia of the pickup assembly $13\frac{1}{3}$ gm, the minimum tracking mass is 1 gm.

If the tracking mass is smaller than M'_t, then for heavily modulated grooves the contact between the stylus and the groove walls is liable to be intermittent, damaging the walls in addition to increasing the noise content of the resulting signal.

Usually the vertical compliance, C_v, need not be considered, for although the value may be as low as half the lateral compliance, the vertical groove modulation only results from the difference (in amplitude and phase) between the two stereophonic signals: such a modulation seldom exceeds half the lateral modulation especially at low frequencies where the phase differences are normally very small. However, if for a stereophonic pickup $C_v < \frac{1}{2}C_l$, $2C_v$ should replace C_l in equation 10. Even a monophonic pickup requires a vertical compliance to track the modulation resulting from the pinch effect. The second harmonic vertical modulation due to a spherical stylus tracking a laterally cut groove is likely to reach 20%. In this case we require $C_v > 1/5C_l$ for a monophonic pickup, otherwise $5C_v$ should replace C_l in equation 10.

Unfortunately, the maximum lateral displacement of the stylus occurs at the same instant as a maximum vertical displacement due to the pinch effect. As a result, it appears preferable for $5/10^6 C_l$ in equation 10 to be replaced by

$$\frac{5}{10^6 C_l}\left(1 + \frac{C_p^2}{25 C_v^2}\right)^{\frac{1}{2}}.$$

However, the difference is usually very small except for

[2] J. Crabbe, *Hi-Fi News*, April 1963, p. 797-800.
† See equation 10 and Appendix III.

* Normally, a faster groove speed is an advantage: this is one of the rare instances where the converse is true.

monophonic pickups from a designer who has disregarded vertical compliance.

Equation 10 is obtained from the following

$$M'_t g = \frac{z}{C_l} + M_{etm}a_s + M_p a_a \qquad .. \qquad 11$$

where z, a_s, and a_a denote the maximum displacement of the stylus, the maximum acceleration of the stylus tip, and the maximum acceleration of the arm, respectively. Representative values[3] are $z = 0.005$ cm, $a_s = 1000g$ $a_a = 0.025g$, where $g = 980$ cm/sec^2.

Clearly, the downward force due to the stylus mass should be greater than the sum of these three forces which try to prevent the stylus remaining in perfect contact with the groove walls. The forces resulting from resistive damping are by comparison very small for modern arms and have not been included: this is permissible as we have considered the worst possible case, although it is most unlikely that all these forces will be maximum at the same time. Equation 11 is based on three equations given by Professor F. V. Hunt[3] who divided the downward force into three equal parts, requiring each to equal or exceed the three individual forces on the right of equation 11 (with the force due to arm damping added to $M_p a_a$). This constraint, although useful for determining suitable design targets, is unnecessary. Note that side-thrust can amount to as much as 20% of the downward pressure so that the maximum lateral deflection due to side-thrust, z_l is given by $z_l = 1/5\ M_t g\ C_l$. Hence, if the lateral compliance of the stylus suspension is very large in relation to the effective tip mass, compensation for side-thrust is especially important to avoid large stylus deflections.

An important point to consider is that a stylus cannot track frequencies above the stylus-groove resonant frequency, f_{rg} where

$$f_{rg} = \frac{1}{2\pi\sqrt{M_{etm}C_g}}$$

C_g, the compliance of the groove material, is given by

$$C_g = \frac{0.00406}{r^{1/3}M^{1/3}}$$

where r is the radius of the stylus tip (in): a value of 3.76. 10^{10} dynes/cm^2 has been used for the plane-stress elastic modulus of the vinylite record material. It is necessary for the stylus to be able to track the highest frequencies which occur on a record otherwise the record will be permanently damaged. With an upper audio limit of 15 kc/s, f_{rg} must not be less than 30 kc/s, the second harmonic of a lateral signal of 15 kc/s; 30 kc/s occurs as a vertical pinch effect modulation. Hence, $M_{etm} \leqslant 0.00693 r^{1/3}M^{1/3}$. To track at 3 gm with stylii of 0.001in, 0.0007in and 0.0005in, the effective tip mass must not exceed 1.00, 0.89 and 0.79 mg, respectively: tracking at 1 gm, these values become 0.69, 0.62 and 0.55 mg.

The stylus-groove resonance is usually sufficiently excited to introduce audio noise by intermodulation. Scanning loss, the high frequency loss due to the finite size of the stylus in relation to the groove modulations, will prevent this excitation if f_{rg} is sufficiently large. The condition which must be satisfied is $M_{etm} \leqslant 0.197 r M_t$ Hence, to track at 3 gm or 1 gm with stylii of 0.001in, 0.0007in, and 0.0005in, the effective tip mass for a 'low noise' signal must not exceed 0.59, 0.41, 0.30 mg, and 0.20, 0.14, and 0.10 mg, respectively.

[3]F. V. Hunt, *J.A.E.S.*, October 1962, p. 274-289.

To summarise, M_t should satisfy the following conditions:—

$$M_t \geqslant \frac{5}{10^6 C_l} + 1000\ M_{etm} + \frac{Mp}{40}$$

$$M_t \geqslant \frac{k}{10^6 C_v} + 1000\ M_{etm} + \frac{Mp}{40}$$

$$M_t \geqslant \frac{3.10^6\ M^3{}_{etm}}{r}$$

where $k = 2.5$ for a stereophonic pickup and 1 for a monophonic pickup. Also, for a low noise signal,

$$M_t \geqslant \frac{5.1\ M_{etm}}{r}$$

If the above conditions allow, M_t should be set to a value within the range 1-3 gm if $r = 0.0005$in or 1-4 gm if $r \geqslant 0.0007$ in. With M_t less than 1 gm, dust in record grooves becomes a tracking problem: if greater than the upper limit, record wear will occur as a result of the groove deformations no longer being within the elastic limit or the record material.

Much useful advice on pickups is given in a recent book by J. Walton[4] who stresses the importance of a low effective tip mass, the most important consideration when choosing a pickup.

To conclude, there is no practical advantage in a properly mounted pickup arm being longer than 9in or 10in. A 12in arm as well as being heavier usually has a larger inertia at the stylus tip: it also requires more space for mounting. The importance of mounting accurately is usually underestimated. The distortion from a pickup with a 12in arm and an error in mounting of $\pm 1/20$in is greater than the distortion from a correctly mounted 9in pickup. In view of this, an alignment protractor should be considered as essential when mounting a pickup.

APPENDIX I

According to E. R. Madsen[5], intermodulation in a lateral cutting appears as modulation of the even harmonics, and in a vertical cutting as modulation of the odd harmonics. The percentage distortion, E_{im}, due to an incorrect vertical tracking error is given by

$$\epsilon_{im} = 100 \left[\sqrt{\frac{\cos(\alpha-\phi)}{\cos(\alpha+\phi)}} - \sqrt{\frac{\cos(\alpha+\phi)}{\cos(\alpha-\phi)}} \right]$$

$$\text{where } \tan\alpha = \frac{\sqrt{2}\ V_{rms}}{u} = 5{\cdot}32\ \frac{V_{rms}}{xs}$$

Therefore, when as in our case $\phi < 5°$, this expression for distortion may be reduced to

$$\epsilon_{im} = 200 \sin\phi \tan\alpha \approx 200\left(\frac{\pi}{180}\phi\right)\tan\alpha = 18{\cdot}6\ \frac{V_{rms}\phi}{xs}$$

Note that the distortion is proportional to ϕ/x. Distortion due to lateral tracking error is similarly proportional to ϕ/x.

If an elliptical stylus is used, lateral tracking error will cause the stylus to sink slightly further into an unmodulated groove. It can be shown[6] that

$$p = \frac{2a\left(1-\frac{b^2}{a^2}\right)\sin\phi}{\left(1+\frac{b^2}{a^2}\tan^2\phi\right)^{\frac{1}{2}}} \approx 2a\left(1-\frac{b^2}{a^2}\right)\sin\phi \approx \frac{\pi\left(a-\frac{b^2}{a}\right)}{90}\cdot\phi$$

where a and b denote the major and minor radii, respectively,

[4]J. Walton, "Pickups—The key to Hi-Fi" (Pitman).

[5] E. R. Madsen *Audio*, November 1962, p. 21-24.

[6] Private communication to C. Dineen, July 1965.

of the horizontal cross-section through the points of contact with the groove walls, and p the distance in the direction of record motion between these points of contact. The distortion, peculiar to elliptical styli, due to the points of contact not being perpendicular to the groove walls, depends on the time difference, t, i.e. the time taken for the groove to move a distance p.

$$t = \frac{p}{2\pi x} \cdot \frac{60}{s} = \frac{1}{3}\left(a - \frac{b^2}{a}\right) \cdot \frac{\phi}{xs}$$

where a, b, and p are in inches and t in seconds. The distortion depends on ϕ/x and may be reduced by reducing this quantity.

When tracking with styli of both circular and elliptical cross section, it is therefore evident that the maximum distortion due to tracking error is least when ϕ/x is a minimum, as in the given designs.

APPENDIX II

An examination of pickup arm manufacturers' recommended values of offset angle and overhang reveals considerable discrepancies. Some manufacturers have clearly determined these values on a trial and error basis: others have minimised the angular tracking error forgetting (or not knowing) that the distortion resulting from tracking error is inversely proportional to the distance from the turntable centre. The most suitable values appear to have been obtained from a graph by Bauer[7] who (using our notation) derived the following formulae.

$$\phi = \frac{57{\cdot}3\, x_3 \left(1 + \frac{x_3}{x_2}\right)}{l\left[\frac{1}{4}\left(1 + \frac{x_3}{x_2}\right)^2 + \frac{x_3}{x_2}\right]}$$

$$f = \frac{x_3^2}{l\left[\frac{1}{4}\left(1 + \frac{x_3}{x_2}\right)^2 + \frac{x_3}{x_2}\right]}$$

Bauer set x_2 and x_3 to the radii of the inner and outer grooves, and therefore x is maximum at the extreme limits of the stylus movement. Bauer's values are x_2=5.75in, x_3=1.75in.

Although suitable at the time, x_3 is too small for a modern record player. The tracking error which changes from positive to negative and back to position between x_2 and x_3 is still negative when x=2in. Replacing these constants with the values used in the present design 1C (x_2=5.719in, x_3=2.038in) reduces the maximum distortion due to lateral tracking error of a given modulation by 16% for an 8in pickup and 18% for a 10in pickup. However, unlike the present designs, the maximum negative of x is less than the values x_2 and x_3. If our design 1C is used, the improvements become 29% and 27%, respectively. Bauer made two approximations: in an expression for $\sin\theta$, $2l-f$ is replaced by $2l$, and $\sin\theta$ itself is replaced by $\pi\theta/180$. These approximations have not been used in the present analysis which, as a result, is more extensive. The reduction in distortion is fairly small and it may not be considered worthwhile to modify an existing pickup arm. However, if a new arm is being designed, values of offset angle and overhang have to be chosen, and as no extra work is involved in using the values given here in preference to Bauer's, it would be foolish to disregard these improved values on the grounds that the reduction in distortion is very small. It is a step in the right direction although two or three such steps may have to be made before the improvement is audible.

APPENDIX III

The movement of a stylus relative to a pickup produces the required electrical signals. If the stylus moves very slowly from side to side (or up and down), the pickup will follow these movements and result in a negligible signal. Maximum signal is obtained at the transition frequency f_{ar}, the frequency of 'arm resonance' given by:—

$$f_{ar} = \frac{1}{2\pi\sqrt{M_pC}}$$

where M_p is the inertia of the pickup assembly and C the compliance (or inverse stiffness) of the stylus suspension.

We require the lateral arm resonant frequency to be greater than 1 c/s so that the effect of the stylus moving towards and away from the turntable centre as a result of eccentric record-mounting holes is negligible. Similarly, for stereophonic pickups we require the vertical arm resonant frequency to be greater than about 10 c/s so that signals are not obtained from ripples and warps. Although sub-audio, these signals are liable to overload the amplifier. Both resonant frequencies should be less than 20 c/s to permit a flat frequency response down to 30 c/s.

The optimum vertical resonant frequency is about 15 c/s. A stereophonic pickup with a vertical compliance of 5.10^{-6} cm/dyne restricts the inertia of the pickup assembly to the range 13 to 51 gm corresponding to resonant frequencies of 20 and 10 c/s respectively (an easy requirement). However, many modern pickups have a compliance of 10.10^{-6} cm/dyne corresponding to an inertia of 6 to 25 gm and some pickups require an even smaller inertia. Since the pickup is liable to weigh about 10 gm, it is evident that the effective mass of a present-day arm (and counterbalance) as seen at the stylus tip should be as small as possible; this is also suggested by equation 10.

The lateral compliance is usually equal or up to twice as large as the vertical compliance so that the lateral resonance occurs at the same or a slightly lower frequency than the vertical resonance. Since the lower limit for lateral resonance is 1 c/s, it is clear that if the frequency of vertical resonance is suitably fixed, the lateral resonant frequency requirement is automatically satisfied. In these circumstances, only a small amount of damping is necessary. Any additional damping required should be associated with the arm pivot and not the stylus suspension and preferably of a viscous type.

If a pickup which has been lowered until the stylus just touches a record is released, it will sink slightly lower at the same time compressing the stylus suspension. The compliance is a measure of the 'springiness' of this suspension and denotes the distance relative to a stationary pickup that the stylus will move as a result of a given force acting on it. A larger compliance implies that the stylus can move more easily relative to the pickup, and therefore the force on the stylus tip due to a groove modulation is smaller. Hence the minimum tracking mass is smaller. The stylus is driven by the groove laterally and upwards and relies on the vertical compliance of the stylus suspension and gravity for downward movement; vertical compliance above the vertical arm resonant frequency and gravity below this frequency. The downward force resulting from the tracking mass becomes the force on the effective tip mass due to the vertical compliance of the stylus suspension acting as a spring. Therefore, the smaller the effective tip mass, the smaller the force required by the stylus to follow a high frequency modulation requiring a large acceleration. Hence the minimum tracking mass is smaller.

[7] B. B. Bauer *Electronics*, March 1945, p. 110-115, and quoted by "Sound Recording and Reproduction", J. W. Godfrey and S. W. Amos (Iliffe) and "Disc Recording and Reproduction", P. J. Guy (Focal).

Correction:—We regret the error which occurred on p.216 of the May issue. The 20th line from the bottom of the second column should start with $\sin\theta_D\approx$ and not $\sin\theta_D n$. In the caption to Fig. 4 $x_2<x<x_0$ was used, but the caption should have conveyed "as x varies from x_2 to x_0 . . . . ." On the left hand side of the first equation on p.217 (second column), θ_D should have read θ_O. But since θ_O, which corresponds to x_O, is equal to θ_D the design value, θ_O may be taken as θ_D.

The Root-locus Technique

1—INTRODUCTION

By W. TUSTING

The root-locus technique is a largely graphical method which enables the stability conditions of a feedback amplifier or closed-loop control system to be determined rather easily. Its great merit is that it enables an approximate solution to be found quickly and easily. This series of articles explains how to use the technique and in this first article the preliminary concepts are discussed.

THE designer of any feedback system is faced with the problem of predicting whether or not a proposed system will be stable. If, as often happens, it turns out that it will not be stable, he has the further problem of finding out what to do about it. The system with which he is concerned may be an electronic amplifier or it may be an automatic control mechanism, which may not embody electronics at all, but the basic problems and the methods of solving them are the same.

The deliberate controlled use of negative feedback in electronics started something like 40 years ago when it was used to improve the performance of valve amplifiers in telecommunications. It is still very widely used in electronic amplifiers and in this field it is still called by that name. It is just as widely used in automatic control, in closed-loop control systems, and here the mechanical applications of the principle go right back to the 17th century. In both cases the purpose is the same: to make the output of a system or device follow the input as closely as possible.

There exists an enormous amount of literature on negative feedback and a great deal of it is concerned with the problem of achieving stability. This is a good indication that the problem is a difficult one. There are, too, a considerable number of different ways of tackling the problem.

Whether one does so consciously or not, the prediction of stability involves expressing the gain of the proposed system in the form of an equation. This is usually quite easy. It is then necessary to find out whether the equation satisfies certain stability criteria. This is the difficult part in all but the simplest cases and the difficulties are entirely mathematical ones. All the different methods that exist are basically different mathematical ways of solving the problem.

The root-locus method is comparatively recent and its use is largely confined to the control field, although it is equally applicable to any other feedback problem. Its use is explained in many books on control but most explanations suffer from being either too terse or too complex for the beginner. Often the "explanations" are comprehensible only when one has gained some familiarity with the method.

The purpose of these articles is to explain how to use the method and as far as possible this will be done in a succession of steps which form a kind of drill. As with most of the other available methods, the root-locus technique is limited to linear circuits.

Feedback amplifier.—Fig. 1 shows the block diagram of a feedback amplifier. It comprises an amplifier of voltage gain A and a feedback path of voltage gain β. Usually $A \gg 1$ and $\beta \ll 1$, but there are cases where $\beta = 1$ and then A is rarely very large. The product $A\beta$ is quite often of the order 10–30. It is commonly called the open-loop gain.

From Fig. 1 $A = V_0/V_i$. The amplification of the feedback amplifier as a whole is $A_f = V_0/V'_i$. Now

$$V'_i = V_i + \beta V_0 = V_i(1 + A\beta)$$

consequently

$$A_f = A/(1 + A\beta)$$

and $A_f =$ is the closed-loop gain.

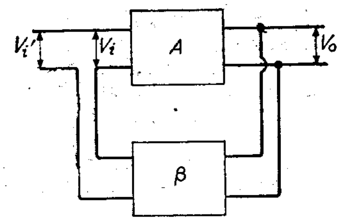


Fig. 1. Block diagram of feedback amplifier.

Temperature control.—Now consider a control problem. Suppose that we have a room which we desire to keep at a temperature θ_r and we have in it a heater delivering heat power P. The outdoor temperature is θ_0 and the room loses heat through the thermal resistance R of its walls. In the steady state the heat lost equals the heat supplied and so

$$P = \frac{\theta_r - \theta_0}{R}$$

If we measure P in watts and temperature in °F, thermal resistance is in °F/watt.

The control problem is so to vary P that θ_r stays constant, or nearly so, despite changes of θ_0. The first step is to derive a control signal proportional to θ_r. We use a set of elements, which we need not now specify, which produces a d.c. output signal of voltage $V_r = k_1\theta_r$, where k_1 is a constant. We compare this with a reference voltage $V_s = k_2\theta_s$, where θ_s is called the set value of the room temperature. For example, if we want the room temperature θ_r to be 70 °F we turn a control on the apparatus so that its calibrated scale reads 70 °F; we set the apparatus to produce this temperature. It may not in fact do so, which is why we must distinguish between θ_r and θ_s. A block schematic of the control system is shown in Fig. 2.

By comparing V_r and V_s we form an error signal

$$\epsilon = V_s - V_r = k_2\theta_s - k_1\theta_r$$

Under some given mean conditions, say, $\theta_r = 70$ °F,

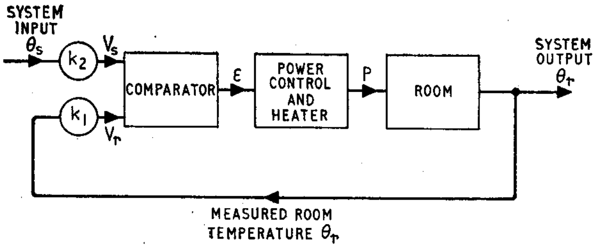


Fig. 2. Block diagram of temperature control system, for comparison with feedback amplifier in Fig. 1.

$\theta_0=40$ °F, a certain power P_0 is supplied. Under these conditions, $\theta_s=70$ °F and $\epsilon=0$, and if $k_1=k_2=k$, $P_0=(70-40)/R=30/R$. To cope with other conditions we arrange for the error signal to control the power supplied to the room so that

$$P=P_0+K\epsilon$$

Combining these equations

$$\frac{\theta_r-\theta_0}{R}=P_0+Kk(\theta_s-\theta_r)=\frac{30}{R}+Kk\ (\theta_s-\theta_r)$$

or $\quad \theta_r-\theta_0=3\theta+KkR(\theta_s-\theta_r)$

and so $\quad \theta_r=\dfrac{30+\theta_0+kKR\ \theta_s}{1+KkR}$

or $\quad \theta_r-\theta_s=\dfrac{30+\theta_0-\theta_s}{1+KkR}$

Quite clearly whatever the values of θ_0 and θ_s, the difference between θ_r and θ_s can be made as small as we like by making KkR large enough. Of course, this presupposes the availability of unlimited power for heating the room, but that is inherent in the assumption of a linear system. Even with limited power, conditions will be the same but for a limited range of θ_0 and θ_s.

The point of importance is the similarity of the equation with that for the closed-loop gain of an amplifier. The term $1+KkR$ of the one is analogous to the $1+A\beta$ of the other.

Symbols.—Instead of the $A\beta$ symbolism usual for amplifiers a GH symbolism is common in control theory. Thus $1+A\beta$ of the one is the same as $1+GH$ of the other, with $G=A$, $H=\beta$. We shall here adopt the GH symbolism because it is the one most used in expositions of the root-locus technique. Its use here thus facilitates comparisons with the literature.

To use the root-locus technique it is necessary to write the expression for the open-loop gain GH in terms of the Laplace transform or Heaviside operator. It does not matter which is used, for the result will be exactly the same. Some people use the symbol s for the Laplace operator and p for the Heaviside and this is helpful as an indication of which system is being used. Here we shall use p.

No knowledge at all of these operational systems is needed for the root-locus method. All that is necessary is that the equations should be written in terms of p. All this means is that the "reactance" of an inductance is written as pL instead of $j\omega L$ and of a capacitance as $1/pC$ instead of $1/j\omega C$. Indeed, any steady-state equation written in terms of $j\omega$ can be translated into the proper forms merely by making the substitution $\omega=p/j$.

Also an equation written as a differential equation can be put into p form just as easily, since $p=d/dt$, $p^2=d^2/dt^2$, $p^n=d^n/dt^n$ and $1/p=\int_0^t ..dt$.

When the equation for GH has been obtained it will be found that it is always in the form of a constant multiplied by the ratio of two polynomials in p. It is necessary that both numerator and denominator be factored into the products of simple factors. In some cases this factorization may be difficult, but in most cases the equations turn out automatically in factors and there is no difficulty at all.

An ordinary valve amplifier stage has a coupling resistance R with shunt capacitance C. With a valve of mutual conductance g_m, the stage gain is $A=g_m\ R/(1+j\omega CR)$, as is well known.

Writing $G_0=g_mR$, $T=CR$ and $\omega=p/j$ we have

$$A=G=\frac{G_0}{1+pT}$$

For the root-locus technique it is necessary that the coefficient of p should be unity. We thus divide numerator and denominator by T and so get

$$G=G_0K\ \frac{1}{p+1/T}$$

with $K=1/T$. This is the proper form for the root-locus technique and all equations must be brought into it.

For a three-stage RC amplifier having different time constants in each stage we should clearly have

$$1=GH=1+G_0H_0K\frac{1}{(p+1/T_1)\ (p+1/T_2)\ (p+1/T_3)}$$

where $K=1/T_1T_2T_3$.

The critical condition for stability is $0=1+GH$. The system is stable if $|GH\ |<1$ even if it is negative but is unstable if $|GH\ |>1$ and is negative. To determine the critical condition it is necessary in some way to solve the equation $0=1+GH$. This means finding its roots. For stability it is necessary that all the roots should have negative real parts. The critical condition occurs when the real parts of one pair of complex conjugate roots become zero; that is, when there is a pair of imaginary roots.

This may sound rather complicated. In fact, the only difficulty is the mathematical one of finding the roots of an equation. In the case of the above expression for a three-stage RC amplifier a general algebraic solution is easy. For more complicated expressions the difficulties grow rapidly. The root-locus technique enables solutions to be obtained by a graphical method.

There are three terms in constant use, the meanings of which must be clearly understood. These terms are root, pole and zero. Bearing in mind that in general GH stands for a fraction which may have terms in p in both its numerator and denominator, a *root* is a value of p which makes $1+GH=0$, a *pole* is a value of p which makes the denominator of GH equal to zero, and a *zero* is a value of p which makes the numerator of GH equal to zero. Poles and zeros are thus respectively the roots of the denominator and numerator of GH, but in the root-locus technique the word "root" is in the main reserved for the special values of p which make $1+GH$ equal to zero.

As an example, suppose

$$GH=G_0H_0\ \frac{p+2}{(p+3)\ (p+4)} \quad \ldots\ldots\ldots\ldots\ldots\ldots\ (1)$$

then there is one zero, $p=-2$, and there are two poles, $p=-3$ and $p=-4$. The roots of $1+GH=0$ can easily be found algebraically in this instance. Multiplying out we get

$$p^2+p(7+G_0H_0)+12+2G_0H_0=0$$

whence

$$p = \frac{-(7+G_0H_0) \pm \sqrt{[1+6G_0H_0+G_0^2H_0^2]}}{2}$$

so that there are two roots which in this instance are always real.

As an example of the root-locus technique consider a three-stage valve amplifier at high frequencies. The relevant equation is

$$GH = G_0H_0 \frac{1}{(1+pT_1)(1+pT_2)(1+pT_3)}$$

To bring it into the proper form we divide numerator and denominator by $T_1T_2T_3$ to get

$$GH = G_0H_0K \frac{1}{(p+1/T_1)\ (p+1/T_2)\ (p+1/T_3)}$$

with $K=1/T_1T_2T_3$.

We must now assign numerical values to the terms in T. Let $T_1=1$ μsec, $T_2=2.5$ μsec and $T_3=10$ μsec, then $K=1/25$ and

$$GH = G_0H_0K \frac{1}{(p+1)(p+0.4)(p+0{\cdot}1)} \quad \ldots\ldots\ldots\ldots \quad (3)$$

and since T is in microseconds, frequencies will be in megacycles per second.

In this particular example there are no zeros and there are three poles viz, -1, -0.4 and -0.1.

Before we continue and deal with the way in which a root-locus diagram is prepared, it is useful to consider the final diagram itself. Fig. 3 shows the root-locus plot for our example.

The diagram is one in what is called the complex p-plane. It is a plot of all the values of p which make $1+GH=0$; some of these values of p are real, and $p=-\rho$ say. Other values of p are complex, and $p=-\rho \pm j\omega$.

The root-locus is a plot of the roots of $1+GH$ as G_0H_0 is varied from zero to infinity.

In Fig. 2 the three poles of our example are plotted and marked by crosses. These points are also roots for $K_0G_0=0$. As K_0G_0 is increased the roots move away from the poles. One moves to the left from $p=-1$ along the real axis and tends to infinity when $K_0G_0 \rightarrow \infty$. Of the two other roots one moves to the left from $p=-0.1$ and the other to the right from $p=-0.4$.

For some particular value of K_0G_0 the two roots meet, in this case at $p=p_b$ -0.235 approximately. For a further increase of K_0G_0, these roots become complex. They move away from the real axis along the curve one upwards and the other downwards.

There is a critical condition for stability when the roots become purely imaginary. This is where the curved part of the locus crosses the $j\omega$ axis, for this is the transition from complex roots with negative real parts (stability) to complex roots with positive real parts (instability).

In our example this occurs at $p=\pm 0.735$ megaradians per second. If our amplifier becomes unstable it will thus be at a frequency of $0{\cdot}735/6.28 = 0{\cdot}117$ Mc/s.

Now what we really do want to know is the value of G_0H_0 at this point; that is, we want to know the critical value of the open-loop gain. We can find this very easily. In our example, we measure the distance between the critical point $p=j0.735$ and each of the poles and then form the product of these distances. This is the value of G_0H_0K. The distances are 0.74, 0.835 and 1.24, so $G_0H_0K = 0.775$. Now $K = 1/25$ so $G_0H_0 = 19.4$.

One can find out other things about the amplifier. For example, suppose it is required that the transient response be non-oscillatory. This requires that there shall be no complex roots, so the critical condition is with a pair of equal real roots at $p= -0.235$. The distances to the poles from this point are $0.235-0.1=0.135$, $0.4-0.235=0.165$ and $1-0.235=0.765$. The product is 0.01435 and, dividing by K, we get $G_0H_0=0.358$. In this case the response becomes oscillatory for a very small open-loop gain.

This has been something of a digression, for we have been showing the end before the middle. Our aim has been to show how easy it is to find out important things about the amplifier from the root-locus plot. We shall now show how to produce the plot. Parts of this are easy; other parts can be tedious.

We shall necessarily have to give the whole procedure and it will seem lengthy and complicated. There is no doubt that in some cases the production of an accurate root-locus plot can be laborious. However, in most practical cases it is unnecessary to do so. If we are mainly interested in finding the critical gain for stability, for example, we need find only one point with any pretensions of accuracy. This is the one at which the locus crosses the $j\omega$ axis.

What we usually do is to draw the curved part of the locus free-hand! There are certain guides which enable us to do this and it enables us to find the crossing point within perhaps 10 or 20%. We can then explore that region and find the right point as accurately as we want. However, in the initial stages of design we usually need to know G_0H_0 only quite roughly, for it will often turn out to be a lot smaller than we want and we shall have to modify the amplifier to obtain the required stable gain.

It is the great merit of the root-locus method that it enables rough values to be determined very easily and quickly.

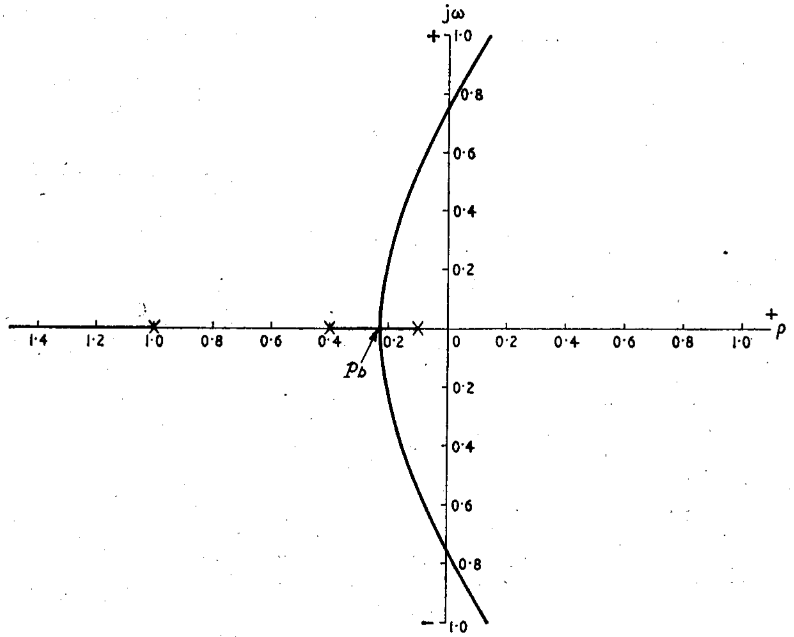


Fig. 3. Root-locus plot for a three-stage RC amplifier.

(*To be continued*)

NEWS FROM INDUSTRY

TRANSMITTERS WORTH £500,000 FOR ONGAR RADIO STATION

S.T.C. Radio Division is to supply and instal 23 high-frequency transmitters and ancillary equipment for use at the G.P.O. radio station at Ongar, Essex. [With this equipment the station will then be able to handle radio-telephone traffic.] Important features will include remote control of tuning and other functions. There will also be facilities for self-monitoring and automatic fault location and service restoration in the event of a failure.

This new system of automatic control will provide for the pre-selection of all the facilities required to work a number of services which can then be selected by a single switch for each service frequency. Facilities are provided for "dualling" during periods of frequency change.

An operator at a console in the control centre will be able to select any one of five frequencies particular to any service. He will also be able to monitor and test equipment without disturbing traffic.

QUEEN'S AWARD FOR INDUSTRY

IN 1965 the Prime Minister announced that the Duke of Edinburgh had consented to chair a committee which would draw up a scheme for awards to industry, to be made by the Sovereign for outstanding achievement either in increasing exports or in technological innovation. Recipients of the Award will be entitled to display the emblem; both the Award and the title to display the emblem expires after five years. However, an industrial concern in possession of an Award may apply for a fresh Award. The first list of recipients contains 115 names, these include British Insulated Callender's Cables Ltd., Decca Radar Ltd., Derritron Electronic Vibrators Ltd., Elliott-Automation Ltd., English Electric Co. Ltd., Garrard Engineering Ltd., General Electric Co. Ltd., Hilger & Watts Ltd., George Kent Ltd., Morganite Resistors Ltd., Multicore Sales Ltd., Pye of Cambridge Ltd., Redifon Ltd. and Smiths Industries Ltd.

The design of the emblem symbolizes Royal recognition of technological and export achievements. The crown signifies the Royal connection, the arrows symbolize exports to the four corners of the earth and the cog-wheels have been chosen as a symbol common to most industrial processes.

As a result of an order from Iraqi Airways world sales of the **Marconi Doppler Navigator** now total over £15 M. The equipment being supplied is the latest Sixty Series version AD560, a civil Doppler system that is being fitted to the Iraqi Airways fleet of new Trident aircraft.

£25M of a recently announced £102M **Saudi Arabian air defence** contract, the largest order of its kind received in Britain, is for electronic equipment. AEI Electronics, who are providing Type 40 radar control stations, have subcontracted part (over £5M) of the £25M order to the Marconi Company for data handling, display and communications equipment. The data handling system is based on the Marconi Myriad micro-electronic computer. Among the displays provided will be a large screen presentation of the air situation shown by three-colour synthetic radar projectors.

A new company, **Devices Implants Ltd.**, has been formed to specialize in the development and manufacture of implanted electronic stimulators, that control human body functions. The new company, taking over the manufacture of the St. George's cardiac pacemaker, will be established in a newly acquired factory, with specially equipped research laboratories. Their products will be distributed by P. J. Reynolds Ltd., of Enfield, Middlesex.

The electronic timing system installed at Coventry's new £1,300,000 international standards swimming baths by **Hadley Telephone & Sound Systems Ltd.**, gives individual times in six lanes to 0.01 sec., with an accuracy of 0.01 sec. This is one of four systems being installed as the bath's communications and signalling network at a total cost of £16,000.

Solbraze Ltd. announce a change of address to Lakedale Road, London, S.E.18. Tel. PLUmstead 3428/9.

Livingston Transistor Agreement.—An agreement has been signed by which Livingston Components Ltd. of North Watford, Herts., will market exclusively in the U.K., the range of field effect, dual, n-p-n and other special types of transistors manufactured by Union Carbide of U.S.A. There will be off-the-shelf delivery of these components.

Zambia's first radio manufacturing plant is being built on the outskirts of Livingstone. The factory is expected to be completed within weeks and will be used to produce transistor radios and radiograms for the Zambian market and for export. The plant, which is costing £250,000, is owned by Supersonic Radio Zambia Ltd., a subsidiary of Standard Telephones and Cables Ltd., of London.

General Technology Corp.—Racal Instruments Ltd. are now able to offer —through their marketing agreement with Tracor Inc.—the range of frequency standards manufactured by the General Technology Corporation, of California. The marketing agreement covers all countries except the U.S.A., Canada and France.

The latest **Soviet industry** five-year plan (1966-70) calls for considerably increased outputs of certain consumer goods. The 1965 production of radio receivers and radiogramophones was 5.2M; by 1970 it is planned to be as high as 8M. Television receivers are to rise from 3.7M to 7.7M in the five years.

Cleveland Electronics Inc. has appointed T. J. Sas and Son Ltd., of Victoria House, Vernon Place, London W.C.1, as U.K. distributors for their Audio Division. The company manufacture hi-fi and p.a. equipment.

A new company, **B. & K. Instruments Ltd.**, has been formed within the B. and K. group to market electronic instruments from manufacturers located chiefly in the U.S.A. B. & K. Laboratories Ltd. will continue marketing the internationally established Bruel and Kjaer instruments from Denmark.

Australian telecommunications exports in 1964-65 amounted to £2.75M, this included £323,000 worth of radio transmitters compared with £417,000 the previous year.

Racal Electronics Ltd. announce that the group profit (before taxation) for the year ended 31st January 1966 amounted to £731,000 compared with £611,000 for the previous year.

The **Instrument Division of Claude Lyons Ltd.** has moved to Valley Works, Ware Road, Hoddesdon, Herts. All enquiries, orders and correspondence in regard to sales and service should be sent to this address.

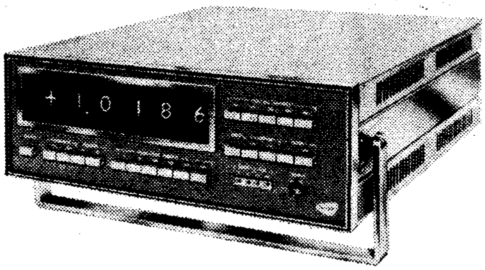
NEW
PRODUCTS

equipment systems components

Digital Voltmeter 301A

THE Fenlow Digital Voltmeter 310A is a precision instrument using a new strobe locked integration technique. The oscillator which produces pulses to be counted during the signal and reference integration periods, is servo controlled to be 20,000 times the mains frequency by means of an early and late strobe principle often used in radar tracking systems. When the mains frequency deviates the strobe locked integration reduces the effect of series mode signals (without the use of filters) by about 100 times over fixed integration period systems.

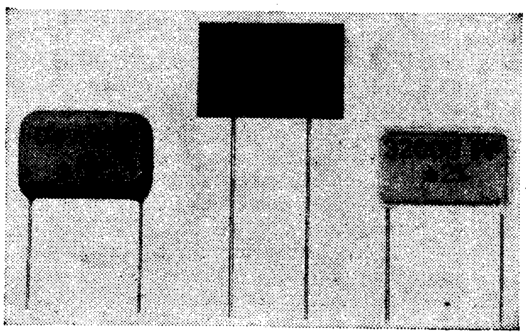
The input and integrator amplifiers make use of field effect transistors for chopping, thus reducing the drift and giving very low input currents and high input impedances. The whole of the input system is floating, permitting differential operation even when used for driving a printer. The instrument contains a programme unit facilitating operation from external commands. Voltage ranges are: 0 to 100 mV, 0 to 1 V, 0 to 10 V, 0 to 100 V, and 0 to 1,000 V. Calibration is achieved by means of a Muirhead reference cell with current and voltage zero balance by front panel controls. Data output consists of decimal coded information, sign, range, and overload by means of a 54-way connector to the rear. For any b.c.d. outputs, a decoder unit plugs into the rear of the instrument. When the decoder is used, the levels and signs correspond with the N.P.L. standard interface. Size is 15 × 5¼ × 18 in and weight 30lb. Fenlow Electronics Ltd., Springfield Lane, Weybridge, Surrey.



WW 301 for further details

Silver Mica Capacitators

LEMCO have now added to their range of sintered silver mica capacitors by the use of mica blades 25 mm × 15 mm in a sintered construction. Pure silvered ruby mica plates are stacked, compressed, and fired, bonding the plates into a stable, robust block. Available moulded or dipped in synthetic resin or with a wax finish, with capacities from 5,000 pF to 0.05μF, and operating voltages of 200 and 350 V d.c. London Electrical Manufacturing Co. Ltd., Bridges Place, Parsons Green Lane, London, S.W.6.



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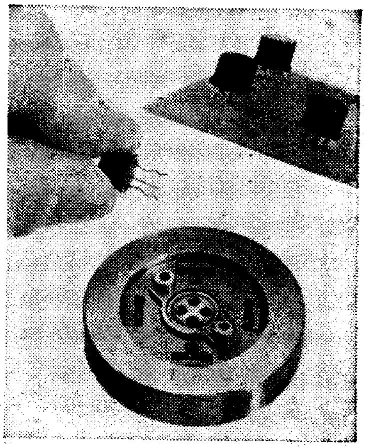


Transistor Joggling Die

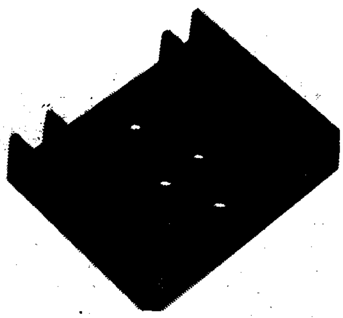
THIS lead cutting and joggling die manufactured by Pico Crimping Tools Co. of the U.S.A. eliminates the need for transistor pads between transistor and printed circuit board, since the die cuts three or four leads to predetermined length, and joggles in a single operation. Transistors are then snapped into standard hole patterns, and remain in place with a ⅛ to 5/32 in air gap, regardless of p.c. board position. For use with the manufacturer's 300, 300B or 300BT air power unit. Available from Kingham Electronics Ltd., 17 Briary Wood Lane, Welwyn Heath, Welwyn, Herts.

WW 303 for further details

Heat Sink

A FLAT based heat sink (A1057) is introduced by Jermyn Industries, for power transistors, s.c.r.s and G.E. triacs. Of black anodized aluminium extrusion, standard units are drilled for T03 cans, or with a single hole which accepts the press-fit version of the G.E. triac. Thermal resistance is better than 8°C/watt. Jermyn Industries, Vestry Estate, Vestry Road, Sevenoaks, Kent.

WW 304 for further details



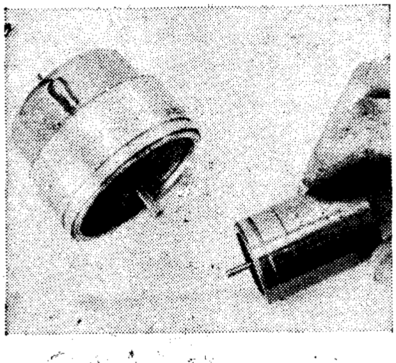
AUTOMATIC PROGRAMMABLE TIMEBASE

THIS unit by Tektronix, known as the 3B5, will, when used with the 3A5 automatic programmable amplifier, provide an automatic oscilloscope display. Programmable functions include time/div, magnifier range, trigger mode with coupling, and trigger slope, by contact closure to ground. By using a 263 programmer, which accepts 6 plug-in type programme cards, the oscilloscope system can be externally pre-set for a given measurement. There is a selection of 11 different programmable functions from plug-in units, and the combination eases the problems involved in many measurements such as production line testing systems check-outs, and also simplifies "away from the oscilloscope" tests, where manual manipulation of the front panel controls would be inconvenient. Further information from Tektronix UK Ltd., Beaverton House, Station Approach, Harpenden, Herts.

WW 305 for further details

Angular-motion Transducers

The Nilson Manufacturing Co., of Florida, U.S.A., have produced a range of angular motion transducers that utilize the principle of phase modulation, and claim that these Variogon transducers overcome the difficulty of phase-shifting accurately in the r.f. range.



This electrostatic phase-shifting transducer has an offset dielectric rotor, providing a variable capacitance between fixed input and output plates. No electrical coupling to the rotor is required. The input plate has four segments which are fed by a quadrature network, the amplitude of the four inputs being equal. These four voltages are capacitively coupled to the output plate through the dielectric rotor, without which the algebraic sum of the resultant output voltage would be zero. Since the rotor is present however, their sum does have some value, the phase of which varies as a linear function of the angular position of the rotor shaft. Variogon transducers are available from the U.K. agents Kynmore Engineering Co. Ltd., 19, Buckingham Street, London, W.C.2.

WW 492 for further details

Portable Gas Torch

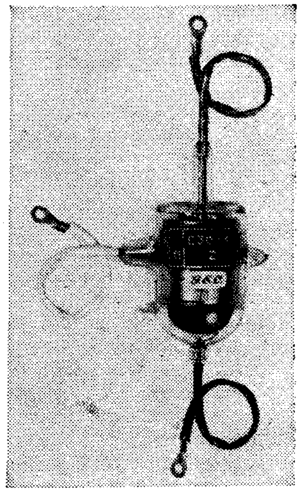
A MINIATURE, portable butane gas torch for soft and hard soldering of small objects is marketed by the Southern Watch and Clock Supplies Ltd., 48 High Street, Orpington, Kent. The Flamidor, as it is called, is $7\frac{1}{4}$ in long, $1\frac{5}{8}$ in diameter and weighs $4\frac{1}{2}$ oz. It is stated that the pencil-sharp flame reaches a temperature of 1,500° C. Complete unit costs 38s 6d and replacement gas containers cost 6s. A slip-on soldering bit is included in the unit.

WW 306 for further details

Triggered Discharge Device

DESIGNED for protection purposes, or for the discharge of capacitor banks the triggered cold cathodes gas discharge device E 3073 is offered for high current, single shot applications such as nuclear research, laser and photographic flash uses, and discharge welders. The working voltage range in 500 to 1200 V, and the peak anode current is 2000 A. Total discharge per operation is 0.5 coulomb. Two valves can be used in series to give 24 kV operation, and in this configuration only one valve needs to be triggered. The E 3073 will operate in any position through an ambient temperature range of −40 to +70°C. The M-O Valve Co. Ltd., Brook Green Works, London, W.6.

WW 307 for further details



4MM MAGNETRON

THE Elliott 4 mm magnetron Type 4MA can maintain stable frequency and efficiency characteristics for operating periods in excess of 200 hours. It is claimed that a useful life of 500 hours can be expected. It has a power rating of 5 kW, a minimum pulse duration of 5 ns, and a rise time of about 1 ns. Typical operating characteristics include peak anode voltage 10 kv, peak anode current 6.4 A, heater current 3.2 A. Overall dimensions are $8\frac{3}{4}\times7\frac{1}{4}\times3\frac{1}{4}$ in and it weighs 14 lb. Cooling of the anode and also the cathode bushing is carried out by forced air. Information from Elliott Electronic Tubes Ltd., Elstree Way, Boreham Wood, Herts.

WW 308 for further details

Mercury Relay

A HERMETICALLY sealed mercury relay, the Euroswitch-M utilizes the principle of breaking the arc on mercury-to-mercury contacts, and it is claimed that a life expectancy in excess of 10 million operations is a result. Since operation is by plunger action, external moving parts are eliminated. The stainless steel mercury container will withstand severe shocks resulting from current surges, arcing, etc.

A current of 25 A at 440 V a.c. resistive, can be disconnected at a rate up to 2,000 per hour. Standard coil voltages are available, while operating power required is 4 W on d.c. voltages, and 6 VA on a.c. voltages.

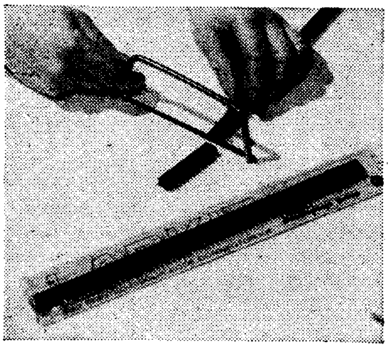
The relay operates to within 25° of vertical. Size 1.5 in diameter × 3.25 in long. Distributed by Techna (Sales) Ltd., 47 Whitehall, London, S.W.1.

WW 309 for further details

EDGE CONNECTOR KITS

THESE Plessey 603 printed circuit edge connectors to be marketed as Labkits, comprise a 78-way strip, four pairs of mounting feet and four polarizing keys. The two basic kits, Nos. 1 and 2 with contact spacing of 0.156 in, and 0.150 in, respectively, are intended for use by prototype and development engineers in industry, universities and technical colleges. The connectors can be cut to size depending on the number of terminations required. Manufactured in glass-loaded polycarbonate, with brass, gold flash or silver contacts, the connectors are rated at 750 V a.c. r.m.s. Operating temperature range is −40° to +85°C, with current rating at 5 A continuous. Available from the Plessey Blue-Arrow Service, Wiring and Connectors Division, Cheney Manor, Swindon, Wilts.

WW 310 for further details



Electronics and Shipping

A REPORT ON THE GLASGOW I.E.E./I.E.R.E. SYMPOSIUM

THE main aims of the recent symposium on Electronics, Measurement and Control in Ships and Shipbuilding which took place in Glasgow during mid-April was to bring to the attention of shipbuilders and shipowners the advantages that electronic techniques would have from the point of view of building and operating ships. However, an impression gained during the lively discussions which followed some of the sessions was that the shipping industry appears to be wary of the "black box," especially as regards data logging. In addition, an overall mutual appreciation of the problems facing engineers from both industries is required. In general, the ship should be treated by both industries as a basic unit to be automated.

Working within the electronics industry, we were perhaps biased but a general example quoted by one speaker who was involved in the installation of data logging equipment was where the ship builder considered the installation as three separate entities—the transducers, the cables and the data logger, instead of a complete installation—and tried to dictate the siting of these three "separate entities." Accordingly where the cables are routed through an electrically noisy environment, error signals are introduced into the system and cause a detrimental effect on performance. In fact, another speaker quoted an example where interference was so bad that it simulated a starting pulse for a generator normally operated by a push-button. As a result of these effects, the ship owner is obviously dissatisfied and this has produced a great deal of scepticism and almost hostility towards the equipment. The logical attitude of one speaker was that a great deal of basic electronic experience gained from other industries is available to the shipping industry, but this has to be studied in a different perspective before it can be applied to an "aboard ship" environment.

Although most of the papers were concerned with shipping there were several of general electronic interest.

In his paper "Modern Marine Radar Systems," C. J. Collingwood (A.E.I.) described current types of marine radar presentation. The form of presentation was a most important factor from the point of view of the navigator and the author felt that collisions of vessels in radar contact were due to ineffective or delayed interpretation by the navigator of radar information.

When considering the relative motion type of p.p.i. display which is available with ship's head or compass stabilization, the author thought that it was adequate for long range navigation and detection of targets at medium and short ranges, but in congested shipping areas where avoiding action was necessary the relative movement of echoes had to be resolved graphically by plotting methods before an avoidance course could be decided. Time required for manual plotting was not always available.

Another form of presentation considered was the true motion or chart-plan presentation on which the progress of the ship is shown against a stationary background, i.e., coastline and fixed navigational marks. With this method the navigator is provided with the means to assess simultaneously the position and movement of the ship relative to stationary objects, and all moving targets tracking on their true courses and at their true speeds. Mention of a computer used in conjunction with chart-plan presentation was also made by the author, and this included the feature of tide correction applied to the ship's course and speed. The author also described a reflection plotter which can be used so that direct plotting can be made on a form of display. Combination of the plot and radar display provide the navigator with past, present and predicted information.

Basically the reflection plotter consists of a half silvered mirror midway between the face of the c.r.t. and a plotting surface which has a surface curvature equal but opposite to that of the c.r.t. The author proved that a point or mark on the plotting surface appears to be on the face of the c.r.t. Thus plotting lines appear as if they were drawn directly on the face of the c.r.t.

From this type of display the author continued with a discussion on photoplot. With this method, radar information is displayed with high brilliance on a 3½ in c.r.t. and photographed on special 16 mm film which is processed in a few seconds, and then projected on the underside of a translucent screen, large enough to allow observation by several viewers at the same time. Pictures can be projected successively at intervals of 15 seconds, three minutes or six minutes. Targets appear black against a white background. As the author pointed out, during the long time intervals, successive radar paints of moving targets become integrated and are then automatically shown on the display as lines of varying length and direction which indicate the track and distance travelled.

The importance of radar as part of a ship's equipment was stressed by the author who pointed out that during the past 15 to 20 years the number of ships in the world has almost doubled and much more sea room is required by tankers and carriers which have increased in size by a factor of five, but, surprisingly, radar aboard ship is not compulsory.

Collision avoidance of aircraft has been discussed in *Wireless World* recently, and it is interesting to note that a paper, by P. G. Tarnowski (A.S.W.E.) "Radar Computer for the Closest Point of Approach" dealt with the collision avoidance of ships. An experimental computer has been developed which can be connected to any radar set of the Merchant Navy. During operation, the bearings of the tracking ship and the other ship are displayed in p.p.i. form. The operator locks a bearing and range marker to the echoes from the possible hazard and this marker then follows the movement of the hazard. The echoes are fed to a computer which processes the received bearing and range information in rectangular co-ordinates and computes the relative course of the hazard. Information in the form of the x and y co-ordinates is stored by the computer and after half a mile decrease in range a second set of x and y co-ordinates is obtained, stored in a second store and subtracted from the first to give the first value of relative track. Alternating voltages corresponding to this information are amplified and fed to a c.r.t. where the relative track is displayed as a line on the screen from which the closest point of approach (C.P.A.) can be read. After a second half-mile decrease in range, the first store is cleared to receive the latest x and y co-ordinates from which relative courses over the second half mile can be obtained. The stores are cleared alternately during the overall track. If the true course of the hazard is required provision has been made for subtraction of the velocity of the tracking ship. In addition, when the tracking ship alters course to avoid collision the computer will predict the result of the altered course.