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Commentary

WITH the prospect that the world's reserves of coal and oil are within a measurable period of exhaustion it is to be expected that considerable efforts would be made to find alternative sources of energy. In Great Britain, the problem has been particularly acute although the nuclear power stations which are now being built will go a long way towards providing an abundant supply of power for practically all industrial and domestic purposes when our coal supplies are no more.

There are however many circumstances where energy is required in varying amounts and where the demand cannot be supplied from an electricity network, and there exists therefore the need for a self-contained energy producing source which does not utilize the shrinking reserves of fossilized fuels.

Great interest has therefore been aroused by the recent announcement by the National Research Development Corporation that a "fuel cell," capable of delivering useful amounts of electrical energy, has been developed by a team of Cambridge scientists led by Mr. F. T. Bacon who has been intimately associated with the project for more than twenty years.

The fuel cell is however not new. It was originally envisaged by Sir William Grove as far back as 1839 and the first practical cell was actually made to work at the turn of the century but the lack of suitable materials and technology prevented its further exploitation.

However, interest has quickened in the post-war period and a ten-year research programme at Cambridge University sponsored by the Electrical Research Association and the Ministry of Power led to the National Research Development Corporation placing a development contract in 1957 with Marshall of Cambridge Electronics Ltd.

The United States have shown a considerable interest in the development of the fuel cell in this country and already an American company has become a licensee of the Corporation and has now manufactured fuel cells of the Bacon type for use with the American Air Force.

The fuel cell is an electro-chemical device in which the free energy of combustion of a fuel is converted directly into electrical energy. The type developed at Cambridge uses hydrogen and oxygen as its fuel, producing about 5kW at 24V with an efficiency of about 80 per cent. This hydrogen-oxygen cell, as it is called, operates in fact by a simple reversal of the process of electrolysis of water.

The Bacon fuel cell consists of 40 unit cells each of 10in diameter giving an output of 2½kW at 32V or 5kW at 24V under maximum load conditions.

Each unit cell consists of two electrodes, one for hydrogen and the other for oxygen, separated by electrolyte. The electrodes are made of porous sintered nickel with a thin layer of a smaller pore size on the liquid side. The oxygen electrode is treated with lithium and pre-oxidized to prevent corrosion. The electrolyte is strong caustic potash, and working conditions are around 200°C and 400lb/in^2. The electrolyte soaks into the sintered metal but, on the application of the gas under pressure from the back of the plate, is expelled from the larger pores; the gas cannot bubble through the smaller-pored surface owing to the surface tension of the liquid. There is thus a very large surface of wetted sintered metal in contact with gas in each electrode—about forty square metres.

Oxygen molecules on the positive electrode combine with water to form negatively charged hydroxyl ions which each remove an electron from the oxygen electrode. The hydroxyl ions migrate through the electrolyte to the negative electrode, where they combine with the hydrogen to form water, depositing an electron in the process. Thus the hydrogen electrode becomes negatively charged with respect to the oxygen electrode and a current flows in the external circuit.

In order to remove the water formed, the hydrogen is circulated past the back of the hydrogen electrodes and the mixture of hydrogen and steam is cooled externally, so that the condensate can be released as required.

The pressure of the two gases must be very evenly balanced if damage is to be avoided and this requires a very accurate control system. The system now evolved is to keep the oxygen pressure constant and to control the hydrogen pressure against it by a very accurate differential pressure meter, actuating a power-operated inlet valve controlled by a servomechanism. Other problems of initial heating and subsequent cooling have also had to be solved.

It can be said that the hydrogen-oxygen cell has completed the second stage of development and is now approaching the commercial proposition of a completely automatic and reliable battery, capable of producing power at a moment's notice in a practical installation. Among the advantages claimed for the fuel cell is its ability to withstand large overloads at reduced efficiency without damage. It is silent and free from vibration in operation, it has few moving parts and the "exhaust" is only water, advantages which make it particularly attractive for certain types of transport applications.

Charging of the fuel cell is a simple process and consists merely of refilling with the two gases, a step which will be much simplified when new methods have been evolved of storing hydrogen and oxygen either in liquid form or as compressed gas at a very low temperature.

Production of the gases will, it is supposed, be obtained by the electrolysis of water so that in this regard the fuel cell may also be looked upon as an electrical accumulator, for power for the electrolysis process can be obtained during the off peak periods of the electricity supply.

It is unlikely, however, that the fuel cell battery could be competitive with existing types of accumulators in small sizes owing to the high cost of control gear in comparison with the overall cost of the plant. Power outputs of less than 100W are not likely to be economic.

A Domestic F.M. Receiver using Diffused Base Mesa Transistors

By M. Applebaum* and E. Midgley*

This article discusses briefly the characteristics of the diffused base mesa transistor and describes its use in the r.f. and i.f. stages of a domestic f.m. receiver. Performance details for the complete receiver are given.

(Voir page 505 pour le résumé en français: Zusammenfassung in deutscher Sprache Seite 509)

TRANSISTORS with alpha cut-off frequencies considerably greater than 50Mc/s are now becoming available. An increase in the upper limit of operating frequency of a transistor requires a reduction in the transit time across the base region. This can be achieved not only by decreasing the base width, but also by grading the impurity content through the base region. The graded base structure produces a field which accelerates the injected carriers across into the collector region and can decrease transit time by a factor of 3 or 4.[1] In addition, since the collector side of the base is lightly doped, the depletion layer at the collector junction extends well into the base and gives a low junction capacitance. Typical values for the collector junction capacitance are 0·5 to 2pF.

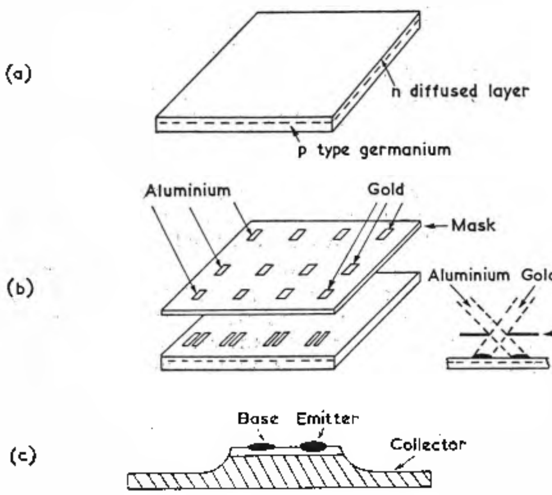


Fig. 1. Simplified manufacturing process
(a) Collector body
(b) Evaporation process
(c) After alloying, lacquering and etching

The Diffused Base Mesa Transistor

Construction

Transistors in which the base region is formed by solid state diffusion of an impurity into a previously formed collector region have the graded base structure previously mentioned. The germanium high frequency diffused base mesa transistor is such a device[2] and its construction will now be briefly described. The basic steps in the fabrication of the device are illustrated in Fig. 1. An n-type impurity is diffused from the vapour state into a wafer of p-type germanium, which forms the collector body of the transistor (Fig. 1(a)). The diffusion process permits close control of the base width. Emitter and base contacts with dimensions of only a few thousandths of an inch are then deposited on the n-type surface by evaporating aluminium and gold through a masking gratule (Fig. 1(b). The temperature of the wafer is then raised until the aluminium contact alloys into the n-type germanium. Since the aluminium is a p-type impurity the emitter-base p-n junction is formed on re-crystallization. A large number of potential transistors can be formed in one germanium wafer in this manner. After the alloying process the individual transistors are masked and the wafer is etched to remove excess n-type material and reduce the area of the collector junction. This leaves the emitter and base contacts on a raised surface or plateau, giving rise to the descriptive term "mesa", which derives from the Spanish word for a plateau.

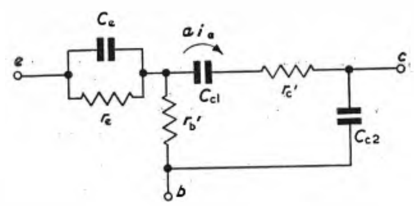


Fig. 2. Equivalent circuit for diffused base transistor

The units are then separated and mounted on "headers", and wires are pressure bonded to the emitter and base stripes.

Characteristics and Equivalent Circuit

A simple high frequency equivalent circuit is given in Fig. 2. It has been demonstrated that this equivalent[3,4] circuit is valid at frequencies which are sufficiently high for the collector junction resistance to be negligible compared with the reactance of the collector junction capacitance, and up to the alpha cut-off frequency. The resist-

TABLE 1

Characteristics of Type A and Type B Transistors at $I_e=2mA$, $V_{cb}=6V$.

TYPE A		TYPE B	
$f_{\alpha co}$	450Mc/s	$f_{\alpha co}$	90Mc/s
α_0	0·95	α_0	0·95
r_b'	75Ω	r_b'	75Ω
$C_{c1}+C_{c2}$	1pF	$C_{c1}+C_{c2}$	4pF
M.A.G. at 100Mc/s		M.A.G. at 10·7Mc/s	
Calculated	9dB	Calculated	27dB
Measured	10dB	Measured	25dB
y parameters at 100Mc/s (mmhos)		y parameters at 10·7Mc/s (mmhos)	
y_{11}	12·5 j50	y_{11}'	5+j3·7
y_{22}	·90+j2·5	y_{22}'	0·2+j0·5
y_{21}	7·6−j19·5	y_{21}'	39·4−j23·4
y_{12}	0·2+j1	y_{12}'	0·8+j0·26

* *Texas Instruments Ltd.*

ance r_c' in this circuit is the bulk resistance of the collector body. In transistors with larger junction areas this resistance is negligible for most applications. The capacitance C_{c2} is due to the header. At the frequencies to be considered here the characteristics of the transistor can be conveniently represented by the y-parameters of a four-terminal network, and maximum available power gain (m.a.g.) can be calculated in terms of these parameters

Assuming that the generator and load admittances (y_g, y_L) are finite

$$\text{m.a.g.} = \frac{4|y_{21}|^2\, G_g\, G_L}{[(y_{11}+y_g)(y_{22}+y_L) - y_{12}y_{21}]^2} \qquad \text{(1a)}$$

where $y_g = G_g + jB_g$

and $y_L = G_L + jB_L$

For a unilateralized stage, i.e. where $y_{12} = 0$, and for conjugate matching $G_{11} = G_g$, $B_{11} = -B_g$, and $G_{22} = G_L$, $B_{22} = -B_L$, this reduces to,

$$\text{m.a.g.} = \frac{|y_{21}^2|}{4G_{11}\, G_{22}} \qquad \text{(1b)}$$

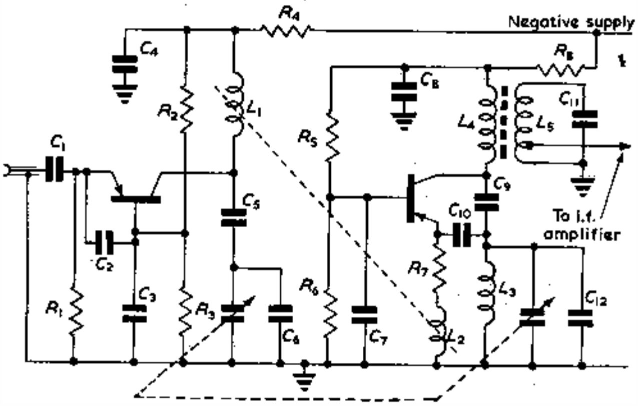


Fig. 3. Circuit diagram of tuner

In the receiver described, two experimental types of diffused base transistors are used. These are referred to as 'Type A' and 'Type B', and typical characteristics are given in Table 1.

Typical Specification for a Domestic F.M. Receiver

A survey of current domestic f.m. valve receivers indicates that the following specification might be acceptable for a transistorized receiver.

Tuning range	88 to 100Mc/s
Intermediate Frequency	10·7Mc/s
Input Sensitivity (50mW output)	5μV
Sensitivity for 10dB signal-to-noise ratio	10μV
Second channel rejection	30dB
Adjacent channel rejection	18dB

This specification was used as a criterion in the design of the receiver.

General Description of the Receiver

The receiver operates from a 12V battery and uses nine transistors. The tuner has an r.f. stage and a combined mixer-oscillator, both using the type A transistor. The i.f. strip has two stages, with a third acting as driver for the ratio detector; all three use the type B transistor. The audio output from the ratio detector is amplified by two class-A stages and a class-B push-pull output stage; the former use 2N185 transistors and the latter uses 2N291 transistors.

Assuming that the input is 5μV into 75Ω the power gain necessary for 50mW audio output is 112dB. An audio gain of 75dB was arbitrarily specified, requiring 38dB from tuner and i.f. stages. The authors have constructed a three-transistor tuner using separate mixer and oscillator stages, but for reasons of economy the two-transistor circuit described is preferred. A single-transistor tuner is possible. However, this would require a high Q input circuit, (with high insertion loss causing reduction of signal-to-noise ratio) to reduce oscillator radiation and improve second channel rejection. The two transistor circuit gives a conversion gain of 20dB leaving a net gain of 18dB to be obtained from the i.f. stages. This does not allow for the conversion loss of the detector and its driver stage, which may be as much as 20dB. Thus the i.f. amplifier was designed to give a net gain of at least 38dB.

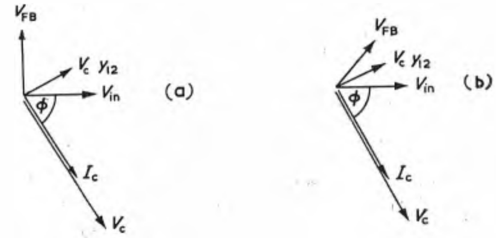


Fig. 4. Vector diagrams of r.f. amplifier

V_c = *Collector voltage*
I_c = *Collector current*
V_{in} = *Signal input voltage*
y_{12} = *Feedback admittance*
V_{FB} = *Feedback voltage at input* = $\dfrac{V_c y_{12}}{y_{11} + y_g}$
ϕ = *Phase-angle of* y_{21}

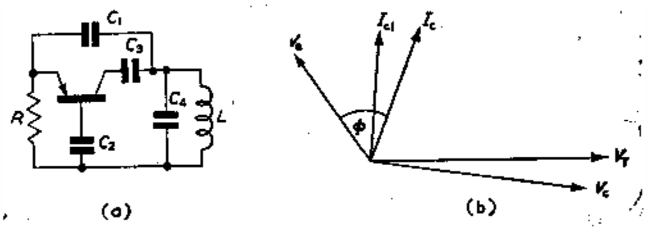


Fig. 5. Oscillator functional circuit and vector diagram

V_c = *Collector voltage*
V_T = *Tank circuit voltage*
I_c = *Collector current*
I_{c1} = *Current through* C_1
V_e = *Emitter voltage*

R.F. Amplifier

The circuit of the r.f. amplifier and mixer-oscillator are given in Fig. 3. Calculation showed that at 95Mc/s the grounded base configuration gave the greater gain and this was confirmed in a practical circuit.

Feedback Consideration

Referring to Fig. 3, the tuned circuit, consisting of L_1, C_5, C_6 and the tuning capacitor, in the collector circuit cancels the imaginary part of $(y_{22} + y_L)$ in the formula for maximum available power gain. However, the term $(y_{11} + y_g)$ can still have an imaginary part and has, in fact, a slightly negative susceptance at 90Mc/s. The inclusion of capacitor C_2 tunes out some of this susceptance. Substituting typical values in equation (1) shows that both these factors increase the m.a.g.

The effect of C_2 is probably best illustrated by considering the vector diagram. Fig. 4(a) shows the conditions when C_2 is disconnected. I_n lags on V_{in} by an angle ϕ. At resonance, V_c and I_c are in phase, and the feedback current due to y_{12}, i.e. $V_c y_{12}$, leads V_c by nearly 90°, since y_{12} is almost entirely capacitive. As $(y_{11} + y_g)$ is inductive the feedback voltage, $V_c y_{12}/(y_{11} + y_g)$, leads the feedback current and is almost 90° out of phase with the signal input voltage, V_{in}.

When C_2 is introduced, the feedback voltage is brought nearer in phase with the input voltage. This is illustrated in Fig. 4(b). In this way a measure of positive feedback is achieved and, in fact, this resulted in an increase of 5dB in the power gain of the tuner. The change of feedback phase on mistuning the collector load does not call for neutralization of the circuit, since y_{12} is low enough to ensure stability at Band II frequencies.

Aerial Matching

For a batch of type A transistors the real part of the input impedance averaged 75Ω. Thus using a 75Ω aerial no matching transformer is needed.

Selectivity and Bandwith

The collector tuned circuit should introduce minimum insertion loss. The unloaded Q of the coil (Q_U) was 130, and the loaded Q (Q_L) was 25. This corresponds to a bandwidth of 3·75Mc/s at the centre of the frequency band, which is quite adequate for good second channel rejection while it introduces no difficulties in tracking. The insertion loss of the tuned circuit is 1·86dB.

Local Oscillator and Mixer

A functional circuit of the oscillator, and the associated vector diagram, are given in Fig. 5.

The Oscillatory Circuit

The L/C ratio of the tank circuit, formed by L_3C_{12} and the tuning capacitor, is low to ensure stability of oscillation. At the same time the risk of spurious responses being generated is reduced since the dynamic impedance of the circuit is also low. The use of a large capacitance in the tuned circuit has the advantage of making the frequency of oscillation almost independent of supply voltage, since it masks any variation of output capacitance of the transistor with collector voltage. In practice the L/C ratio was chosen to give a dynamic impedance of about 3kΩ. If a lower value is used oscillations are likely to cease when the supply voltage falls below 8V. Drift of oscillator frequency with temperature in a transistorized tuner is almost entirely due to changes of ambient temperature. Correction for this can be achieved by using a capacitor for C_{12} with appropriate temperature coefficient.

The Mixer

The signal is injected into the emitter circuit of the oscillator/mixer by inductive coupling. This method was chosen to reduce oscillator radiation. The emitter circuit is at a voltage antinode, and a node of oscillator current, so that very little current at oscillator frequency is fed into the r.f. amplifier. The resistor R_7 is obviously essential if there is to be feedback from the oscillator tank circuit to the emitter, but it unfortunately causes some attenuation of the signal. Hence R_7 is kept as small as the maintenance of adequate oscillator feedback allows. Transistor mixing circuits are of the square law type; the working point is moved over the low-level non-linear part of the base-emitter characteristic. The transistor will therefore operate as a mixer for any signal frequency at which the base-emitter junction exhibits diode characteristics and provided that the transistor will amplify at the i.f.[6]. The primary winding of the i.f. transformer and its stray capacitance have a dynamic impedance of 60kΩ at 90Mc/s, and the winding is critically coupled to the secondary, L_5, at 10·7Mc/s. The transformer has an insertion loss of 8dB, and the total power gain of the tuner is 20dB with a signal-to-noise ratio of 10dB at 2μV.

The I.F. Amplifier

Assuming a peak deviation of 75kc/s and a maximum modulating frequency of 15kc/s the minimum bandwidth needed for 1 per cent distortion of sidebands is approximately 250kc/s[7]. It would, however, be difficult, and uneconomic, to get adequate adjacent channel rejection with this bandwidth and a compromise solution must be adopted. An acceptable bandwidth is 200kc/s with 18dB adjacent channel rejection. This can be readily achieved using double-tuned circuits.

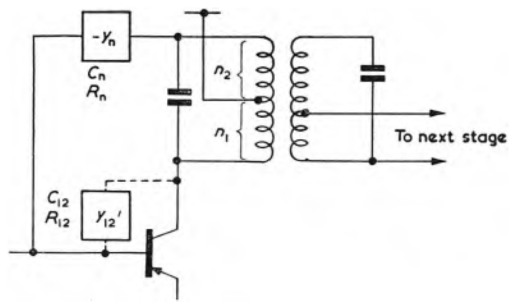


Fig. 6. Simplified diagram of neutralization used

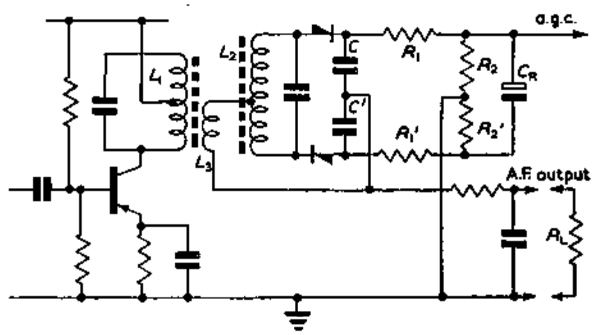


Fig. 7. Typical balanced ratio detector stage

The net i.f. gain requirement was stated earlier as 38dB. To this figure must be added the insertion losses due to the coupling circuits. Depending on how the bandwidth is defined, the insertion loss of a double-tuned circuit is generally greater than that of a single-tuned circuit[8]. In practice two i.f. stages were used, one with a single-tuned circuit load and the other a double-tuned transformer. The latter, together with the double-tuned transformer in the tuner gives the required selectivity. If greater selectivity is desired both i.f. stages may use double-tuned circuits.

Insertion Losses

The insertion loss (i.l.) of a single-tuned circuit is given by:

$$\text{i.l.} = (1 - (Q_L/Q_U))^2 \qquad (2)$$

where $Q_L = f_o/2\Delta f$, $2\Delta f$ being the 3dB bandwidth.

For a double-tuned circuit[8]:

$$\text{i.l.} = \left[\frac{4K^2 Q_{LS} Q_{LP}}{(1 + K^2 Q_{LP} Q_{LS})^2}\right](1 - (Q_{LP}/Q_{UP}))\,(1 - (Q_{LS}/Q_{US})) \qquad (3)$$

where K is the coupling factor.

Q_{LP} is the working uncoupled Q of the primary winding,

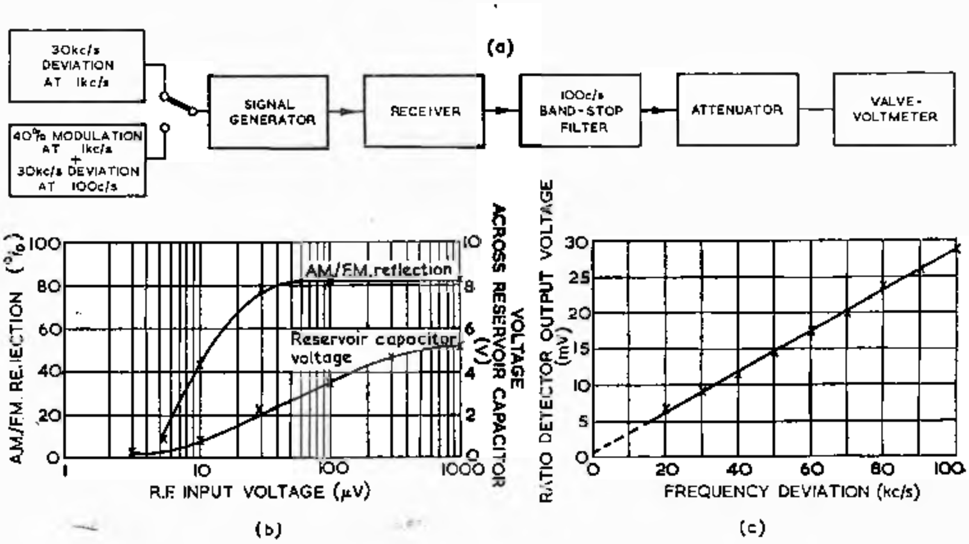
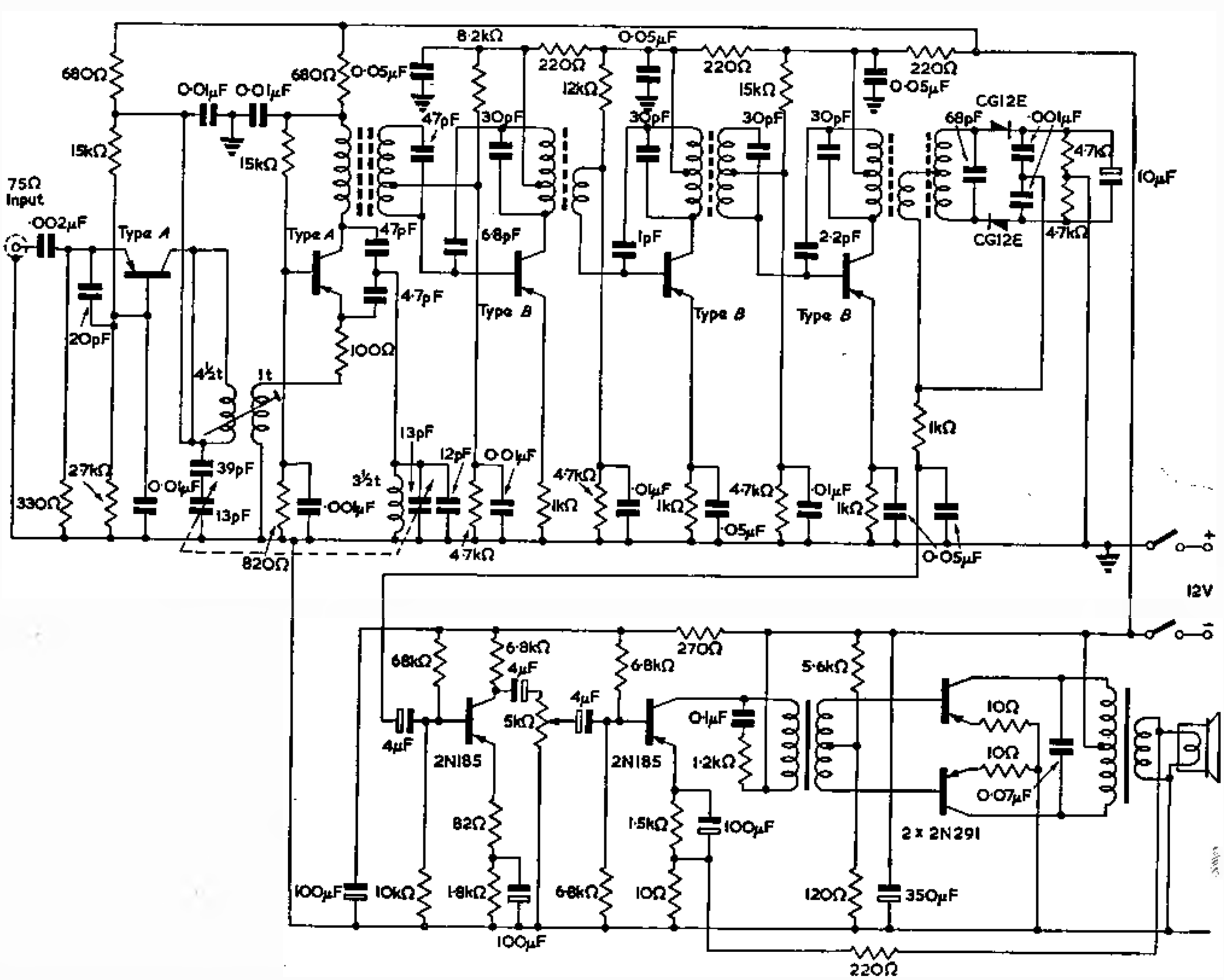


Fig. 8. (a) Measurement of a.m./f.m. rejection
(b) A.M./F.M. rejection ratio and reservoir capacitor voltage as a function of input voltage
(c) Transfer characteristics of ratio detector with r.f. carrier 8μV at 95Mc/s

Fig. 9. Complete receiver



Q_{UP} is the unloaded uncoupled Q of the primary winding,

and Q_{LS} and Q_{US} are similarly defined for the secondary winding.

If the coupling is critical, i.e. $K\sqrt{(Q_{LP}Q_{LS})} = 1$, and the Q factor of the primary and secondary winding are equal, equation (3) reduces to:

$$\text{i.l.} = (1 - (Q_{LP}/Q_{UP}))^2 \quad \ldots\ldots\ldots\ldots (4)$$

Furthermore, the transfer function of a double-tuned circuit is:

$$A_o/A = \left[\left(1 - \frac{Q_{LP}^2Y^2}{1 + K^2Q_{LP}^2}\right)^2 + \left(\frac{2Q_{LP}Y}{1 + K^2Q_{LP}^2}\right)^2\right]^{\frac{1}{2}} \ldots (5)$$

where $Y = (f/f_o) - (f_o/f) = 2\Delta f/f_o$

When the coupling is critical, equation (5) becomes:

$$A_o/A = \left[1 + \frac{Q_{LP}^4(2\Delta f/f_o)^4}{4}\right]^{\frac{1}{2}} \ldots\ldots\ldots\ldots (6)$$

So that the 3dB bandwidth is given by:

$$2\Delta f = \sqrt{2} f_o/Q_{LP}$$

Suppose that the bandwidth of the single-tuned circuit is arbitrarily taken to be 1Mc/s, and that $Q_U = 100$. The bandwidth formula gives $Q_L = 10{\cdot}7$ and from equation (2):

$$\text{i.l.} = (1 - (10{\cdot}7/100))^2$$
$$= 1\text{dB}$$

If the total bandwidth for the two double-circuits is to be 200kc/s, the bandwidth of each must be 250kc/s.

Therefore:

$$Q_{LP} = \sqrt{2} \times (10{\cdot}7/0{\cdot}25) = 60{\cdot}5$$

If the unloaded Q is again taken to be 100, equation (4) gives:

$$\text{i.l.} = (1 - (60{\cdot}5/100))^2 = 8\text{dB}$$

The total insertion loss in the i.f. amplifier is therefore 9dB, so that the total power gain provided by the transistors must be 47dB. This can be readily achieved using two type B transistors.

Neutralization

When reactive components are used in the collector and base circuits of a transistor the denominator of the expression for maximum available power gain can become zero; the gain is then infinite and the stage oscillates. Various methods have been suggested for unilateralization[10,11] and one which is convenient when double-tuned transformers are used is given in Fig. 6.

For complete unilateralization:

$$C_n = (n_1/n_2) \times \text{(reactive part of } 1/y_{12}')$$
$$R_n = (n_2/n_1) \times \text{(real part of } 1/y_{12}')$$

However, it has been found that y_{12}' is predominantly a susceptance, and the neutralization network, can conveniently consist of a single fixed capacitor.

The tapping points on the primary and secondary windings of the i.f. transformer were chosen to give the calculated Q values when loaded by the output and input admittances of the type B transistor. The complete i.f. amplifier is given in the circuit diagram of the receiver (Fig. 9). The amplifier stability was good and the gain and bandwidth were in agreement with the predicted values. The spread of transistor parameters gave a maximum change in gain of 6dB, but the effect on bandwidth and centre frequency was negligible.

The Detector and its Driver Stage

The comparative merits of the ratio detector and the Foster-Seeley discriminator have been discussed in the literature[12]. In this instance, a ratio detector (Fig. 7) was chosen, mainly because the circuit is self-limiting to a.m. so that no separate limiter stage is necessary. In addition, the circuit also gives a convenient a.g.c. output. No attempt was made to obtain selectivity in the driver stage, since the damping of the tuned circuit by the detector varies with the signal level.

The circuit gives maximum a.f. output when the d.c. voltage developed across the reservoir capacitor is maximum[13]. The optimum condition is when the load across the tertiary winding, L_3, of the driver transformer is matched to the output impedance of the driver stage. The load

Fig. 10. A rear view of a prototype receiver

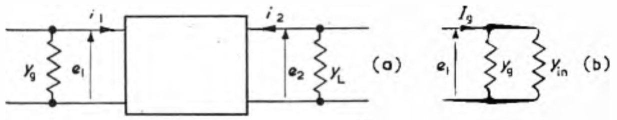


Fig. 11 Derivation of power gain

across the tertiary winding is given approximately by $(R_1 + R_2) R_d/4 (R_1 + R_2 + R_d)$, where R_d is the dynamic impedance of the tuned secondary. The total conversion loss of driver stage and detector in the actual receiver was less than 15dB at 30kc/s deviation. The a.m./f.m. rejection ratio is plotted as a function of input voltage in Fig. 8(b); the test procedure outlined in Fig. 8(a) was used for these measurements. The calculated value given by the expression $R_d/(R_d + R_1 + R_2)$ is 80 per cent which compares well with the measured value of 82 per cent. The transfer characteristics of the detector is given in Fig. 8(c). In practice it was found that the forward resistance of the diodes was sufficient to allow R_1 to be omitted.

Audio Amplifier

The specified gain of 75dB can be obtained with three a.f. stages and still leave sufficient gain to allow negative feedback to be applied. The circuit of the a.f. amplifier is shown incorporated in the complete receiver; the design is conventional and only a brief description is given.

Negative feedback is incorporated in the first stage by the inclusion of an unbiased resistor in its emitter, and a further 6dB of feedback is applied from the secondary of the output transformer to the emitter of the driver stage. Push-pull output is used giving a maximum power output of 500mW at 5 per cent distortion.

Performance Figures of the Complete Receiver

The circuit diagram of the complete receiver is given in Fig. 9. The performance figures given below were measured with a modulation frequency of 1kc/s at 30kc/s deviation.

Sensitivity (50mW output)	3μV
Sensitivity for 0·5W output	10μV
Sensitivity for 10dB signal-to-noise ratio	2μV
Input for 10dB noise quieting	6μV
Second channel rejection	28dB
Adjacent channel rejection	17dB
I.F. bandwidth	198kc/s
Total current consumption at 0·5W output	100mA
Quiescent current	20mA
Total harmonic distortion at 400mW output	3 per cent
3dB bandwidth (a.f.)	100c/s to 15kc/s

A photograph of the complete receiver is given in Fig. 10.

Brief Constructional Details

Printed circuit construction was used throughout the receiver, the tuner being fully screened to keep oscillator radiation to a minimum.

A 7in × 4in loudspeaker was used. Since this is located in the vicinity of the tuning capacitor, the latter was mounted on rubber shock absorbent fittings to reduce acoustic feedback from the loudspeaker to the vanes of the tuning capacitor.

No attempt has been made to miniaturize the receiver but clearly a considerable reduction in size is possible depending on the choice of loudspeaker and batteries.

Conclusions

The circuit described demonstrates that an f.m. receiver can be designed using diffused base mesa transistors to give a performance in line with that of valve domestic f.m. receivers. At this time, however, the cost of these devices precludes their use in entertainment applications.

However, the mesa type of construction is potentially a low-cost manufacturing technique. There is no doubt that as manufacturing experience increases, the reduction in the price of these devices will be considerable and will lead to their being extensively used in receivers for the entertainment market.

Acknowledgments

The authors wish to express their thanks to the Directors of Texas Instruments Limited for permission to publish this article. Thanks are also due to Mr. J. S. Walker for his encouragement and constructive criticism.

Appendix

Derivation of Maximum Available Power Gain in Terms of y-parameters

Referring to Fig. 11(a):

$$i_1 = e_1 y_{11} + e_2 y_{12} \quad \ldots\ldots (7)$$

$$i_2 = e_1 y_{21} + e_2 y_{22} \quad \ldots\ldots (8)$$

$$i_2 = -e_2 y_L \quad \ldots\ldots (9)$$

Substituting equation (9) in (8):

$$-(e_1/e_2) = \frac{y_{22} + y_L}{y_{21}} \quad \ldots\ldots (10)$$

The source generator sees an admittance:

$$Y = y_g + y_{in} \text{ (see Fig. 11(b)}$$

so that:

$$e_1 = \frac{y_{21}}{y_g + y_{in}} - \frac{e_2(y_{22} + y_L)}{y_{21}}$$

Therefore:

$$e_2 = \frac{-I_g y_{21}}{(y_{in} + y_g)(y_{22} + y_L)} \quad \ldots\ldots (11)$$

Substituting equation (10) into (7):

$$i_1 = e_1\left(y_{11} - \frac{y_{12}\ y_{21}}{y_{22} + y_L}\right)$$

Therefore:

$$y_{in} = y_{11} - \frac{y_{12}y_{21}}{y_{22} + y_L} \quad \ldots\ldots (12)$$

Hence:

$$e_2 = \frac{-I_g y_{21}}{(y_{11} + y_g)(y_{22} + y_L) - y_{12}y_{21}} \quad \ldots\ldots (13)$$

The output power is given by:

$$P_o = e_2^2 G_L$$

where G_L is the load conductance.

Therefore:

$$P_o = \frac{I_g^2 y_{21}^2 G_L}{[(y_{11} + y_g)(y_{22} + y_L) - y_{12}y_{21}]^2} \quad \ldots\ldots (14)$$

The input power to the transistor is:

$$P_{in} = e_1^2 G_{in}$$

where G_{in} is the real part of the input admittance, y_{in}

Therefore:

$$P_{in} = \frac{I_g^2 G_{in}}{(y_g + y_{in})^2}$$

and the power gain:

$$A_p = \frac{y_{21}^2 G_L (y_g + y_{in})^2}{[(y_{11} + y_g)(y_{22} + y_L) - y_{12}y_{21}]^2\ G_{in}} \quad (15)$$

This is a maximum for conjugate matching of y_g and y_{in}, i.e., if

$$y_{in} = G_{in} + jB_{in} \text{ and } y_g = G_g + jB_g$$

then:

$$G_{in} = G_g \text{ and } B_{in} = -B_g$$

Under these conditions equation (15) becomes:

$$\text{m.a.g.} = \frac{4 y_{21}^2\ G_g G_L}{[(y_{11} + y_g)(y_{22} + y_L) - y_{12}y_{21}]^2} \quad \ldots\ldots (16)$$

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An Electrical Analogue for Heat Flow Problems in Semiconductors

By N. L. Potter*, B.Sc., D.L.C.

In semiconductor protection problems it was found necessary to be able to estimate the temperature of the semiconductor junction.

This article shows the similarity between heat flow and current flow in a resistance capacitance network and hence the theory leading to the construction of an analogue for the study of junction temperature.

A typical analogue is shown with oscillograph displays obtained from it.

Other uses of the analogue are suggested and the possibility of three dimensional work is discussed.

(Voir page 505 pour le résumé en français: Zusammenfassung in deutscher Sprache Seite 509)

IN the study of protection systems and the rating of semiconductors it was found that an important factor to be considered was the temperature reached by the junction during a conduction cycle. If a certain temperature is exceeded during overloads the semiconductor would be destroyed; also the operating temperature of the junction at full load has been shown to have a considerable effect on the life of the semiconductor.

Some method of determining the temperature of the junction used was needed. It is not possible to use a thermocouple, as due to the low thermal capacity of the semiconductors the temperature almost follows the current waveshape.

In the United States and elsewhere an electrical analogue has been used to solve thermal flow problems on steam piping and furnace walls, and it was decided to apply this system to semiconductors.

Theory of the Electrical Analogue

The analogue is based on the similarity between the laws governing the flow of current in an electrical circuit and the flow of heat in a solid body. These are:

and for heat flow:—

$$R_e = V/I \text{ and } C_e = Q_e/V \quad \ldots\ldots\ldots \quad (1)$$

and for heat flow:—

$$R_t = T/J \text{ and } C_t = Q_t/T \quad \ldots\ldots\ldots\ldots \quad (2)$$

In this article heat flow in one linear direction only will be considered, but it is equally true for all three dimensions. However, as will be shown later, a three-dimensional analogue becomes rather complicated and it has been found easier in practice to use an electrolytic tank to give the heat flow pattern, and then calibrate this with results from a uni-directional analogue.

Consider heat flow along a solid bar (see Fig. 1). Assuming no loss of heat from the surface

Heat flow at X is

$$J_x = -KA\, dT/dx$$

Heat flow at Y is

$$J_y = -KA\, d/dx\, (T + dT/dx \quad x)$$

Difference is $\quad KA\, d^2T/dx^2\, \Delta x \quad \ldots\ldots\ldots\ldots \quad (3)$

This is the heat flow into the layer, or the gain of heat by the layer in one second. Before steady state is reached, the heat is employed in raising the temperature of the layer.

Heat required to raise temperature of layer by dT is

$$A\rho c\, \Delta x\, dT$$

or heat/sec $= A\rho c\, dT/dt\, \Delta x \quad \ldots\ldots\ldots\ldots\ldots\ldots \quad (4)$

Therefore, equating equations (3) and (4)

$$KA\, d^2T/dx^2\, \Delta x = A\rho c\, dT/dt\, \Delta x.$$

$$\therefore K/\rho c\, d^2T/dx^2 = dT/dt \ldots\ldots\ldots\ldots \quad (5)$$

Now consider a cable that has uniform distribution of resistance and capacitance but no inductance. (See Fig. 2).

Let R_e = Resistance per unit length

Let C_e = Capacitance per unit length

Consider a section Δx.

Voltage drop across section $= \Delta V = R_e\, \Delta x\, (i - \Delta i)$

$$\therefore \Delta V/\Delta x = R_e i - R_e \Delta i$$

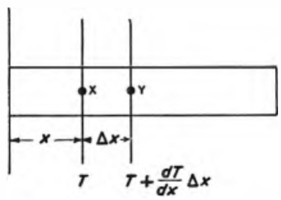


Fig. 1. Heat flow in a solid bar

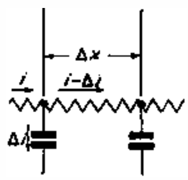


Fig. 2. Current flow in a resistance-capacitance cable

SYMBOLS USED

R_e	=	Electrical resistance (Ω)
R_t	=	Thermal resistance
I	=	Electric current (A)
J	=	Heat flow (cal/sec) = J' = 4·2J joules/sec
V	=	Electric potential (V)
T	=	Temperature (°C)
C_e	=	Electrical capacitance (F)
C_t	=	Thermal capacitance
Q_e	=	Electrical charge
Q_t	=	Stored heat
c	=	Specific heat (cal/gm/°C)
K	=	Thermal conductivity (cal/cm²/cm/sec/°C) (The quantity of heat passed through a centimetre cube in one second with a temperature difference of 1°C between the two faces)
ρ	=	Density (gm/cm³)
t	=	Time (sec)
A	=	Area (cm²)
r	=	Thickness (cm)

* *English Electric Co. Ltd.*

Now capacitance of segment $= C_e \Delta x$.

$$\therefore \Delta i = dV/dt \; C_e \; \Delta x.$$

$$\therefore \Delta V/x = R_e i - R_e \; C_e \; dV/dt \; \Delta x$$

Differentiating with respect to x:

$$d/dx \; (\Delta V/\Delta x) = - R_e \; C_e \; dV/dt \frac{d(\Delta x)}{dx}$$

Taking this to the limit, this becomes

$$d^2V/dx^2 \qquad R_e \; C_e \; dV/dt$$

or: $$dV/dt = \frac{1}{R_e C_e} d^2V/dx^2 \; \dots\dots\dots\dots \quad (6)$$

It will be seen that equations (5) and (6) are analogues with V equivalent to T, and $1/R_e C_e$ equivalent to $K/\rho c$, which equals $1/R_t C_t$ (unit length being considered).

This means that it is possible to represent a problem involving thermal conduction with a perfect resistance-capacitance cable. In practice it is sufficiently accurate to consider the cable in blocks, each represented by a resistance and capacitance.

Therefore, on the analogue 1 second would be represented by z seconds.

This now gives the following relationships between $R_t C_t$ and $R_e C_e$.

$$R_e = mzR_t \qquad C_e = (z/m) \; c_t$$

In semiconductor work to date it has been found sufficient to use a value $z = 1$, but in some applications where the true time-constant is say 30min, this can be reduced to a few seconds on the analogue.

Calculation of Thermal Resistance and Thermal Capacitance

Thermal resistance, like electrical resistance, is proportional to thickness and inversely proportional to area.

$$\therefore R_t = r/KA \text{ °C.sec/cal}$$

Thermal capacitance is proportional to the volume, and the specific heat of the substance concerned.

$$\therefore C_t = c\rho rA \text{ cal/°C.}$$

Therefore, to represent layers of materials with an electrical analogue, the following table must be completed.

	SPECIFIC HEAT	THERMAL CONDUCTIVITY	DENSITY	THICKNESS	AREA OF CROSS-SECTION	R_t	C_t	R_e	C_e
Properties of the Material	e	K	ρ	r	A	$\frac{r}{KA}$	$c\rho rA$	$\frac{m3r}{KA}$	$\frac{3c\rho rA}{m}$
Units	Cal/g/°C.	Cal/cm² cm/sec/°C	g/cm³	cm	cm²	°C. sec cal.	cala/°C.	Ω	Farad

Calibration of the Analogue

The fact that $R_e C_e$ is equivalent to $R_t C_t$ means that it is possible to multiply R_t by some factor m, providing that C_t is divided by that same factor m. In this way an analogue can be set up using easily obtained values of R_e and C_e.

Let $R_e = mR_t$ and $C_e = 1/m \; c_t$

Now $T = JR_t$ and $V = IR_e$ (Equations (1) and (2)).

$$\therefore T/V = \frac{JR_t}{IR_e} = J \,.\, R_t/R_e \,.\, 1/I = J'/4{\cdot}2 \,.\, R_t/R_e \,.\, 1/I$$

Let n milliamps represent 1 watt (1 joule/sec) of heat input.

Substituting $J' = 1$, $I = n \times 10^{-3}$A and $R_e = nR_t$ gives:

$$T/V = 1/4{\cdot}2 \; R_t/mR_t \; 10^3/n$$

$$\therefore T/V = 238/mn$$

$$\text{or } T \approx 238/mn \text{ volts} \quad \dots\dots\dots\dots \quad (7)$$

This means that if an analogue is set up with an average input current of nmA, it will give temperature rises equivalent to 1W average heat input, with 1°C being represented by $238/mn$ volts.

A suitable value for these constants in semiconductor work has been found to be $m = 10^4$ and $I = 0{\cdot}02$ (i.e. 2mA equivalent to 100W), which gives 1°C equivalent to 1·19V. However, any values can be adopted for these constants to suit the particular problem investigated.

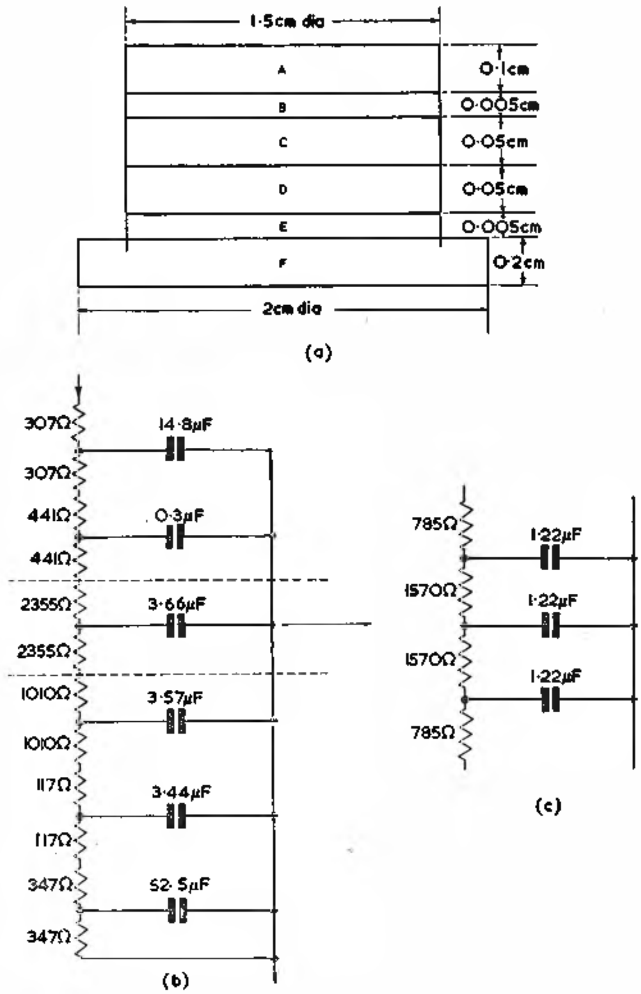
Variation of Time-Constant of Analogue

It is possible, using this analogue technique, to condense a long time-constant into a short period, or vice-versa.

The time-constant of the electrical circuit is given by $R_e C_e$ seconds.

Hence, by introducing a factor z the electrical analogue can be built up with a time constant of $zR_e C_e$ seconds.

Fig. 3. The analogue circuit for a series of layers of material



EXAMPLE

Consider the block shown in Fig. 3(a), which is made up of a number of layers whose properties are as shown below.

LAYER	DENSITY ρ	THERMAL CONDUCTIVITY K	SPECIFIC HEAT c	THICKNESS (cm) r	AREA (cm²) A	R_t $\left(\frac{r}{KA}\right)$	C_t $(c\rho rA)$
A	8·9	0·918	0·094	0·1	1·77	0·0615	0·148
B	10·5	0·032	0·032	0·005	1·77	0·0883	0·00297
C	7·28	0·060	0·057	0·05	1·77	0·471	0·0366
D	5·46	0·140	0·074	0·05	1·77	0·202	0·0357
E	9·73	0·120	0·400	0·005	1·77	0·0235	0·0344
F	8·9	0·918	0·094	0.2	3·14	0·0695	0·525

Now let $m = 10^4$ and $z = 1$.

Then the electrical analogues become:

LAYER	R_t	C_t	$R_e = mzR_t$ Ω	$C_e = z/m\ C_t \times 10^{-6}$ (μF)
A	0.0615	0·148	615	14·8
B	0·0883	0·00297	883	0.297
C	0·471	0·0366	4,710	3·66
D	0·202	0·0357	2,020	3·57
E	0·0235	0·0344	235	3·44
F	0·0695	0.525	695	52.5

Hence, the block may be represented by the electrical circuit shown in Fig. 3(b), with layer F being considered as the base temperature. To get greater accuracy, each element can be replaced by a circuit, as shown in Fig. 3(c). In practice, the thermal capacitance is distributed over the whole of the layer and not concentrated in a block at the centre.

The Three-Dimensional Analogue

Consider a 1mm cube of copper (Fig. 4(a)).

Thermal conductivity of copper = 0·918cal/cm²/cm/sec/°C

Specific heat of copper = 0·0936cal/g/°C

Density of copper = 8·9g/cm³

Cross-sectional area = 0·01cm²

Thickness = 0·1cm

$$\therefore \text{ Thermal resistance} = r/KA = \frac{0{\cdot}1}{0{\cdot}918 \times 0{\cdot}01} = 10{\cdot}9$$

$$\therefore \text{ Thermal capacitance} = cr A\rho = 0{\cdot}0936 \times 8{\cdot}9 \times 0{\cdot}1 \times 0{\cdot}01 = 0{\cdot}00834$$

Now let $m = 10^4$ and $z = 1$

Then $R_e = 10^4 R_t = 109\,000\Omega$

$C_e = 10^{-4} C_t = 0{\cdot}0834\mu$F.

This means that each cubic millimetre of a block of copper can be represented by the network shown in Fig. 4(b), where R is 54 500Ω and C is 0·08μF.

By linking similar networks together, a complete three-dimensional analogue of a block can be built up. However, for a large block this network will become rather complicated, and as already mentioned it is simpler to use an electrolytic tank and calibrate it with a uni-dimensional analogue.

Estimation of the Junction Temperature of a Semiconductor Using the Analogue

When considering heat flow in a semiconductor, three basic assumptions have to be made:

(1) Perfect jointing between different layers.

(2) Linear heat flow within the area considered.

(3) The position of the junction within the actual semiconducting material.

Fig. 5 shows the build-up of a typical semiconductor. Layers A and B may be of various materials, their main purpose being to give strength to the assembly. Each of the layers is joined to the adjacent one with a layer of solder a few thousandths of an inch thick. Perfect jointing over the whole of the surface is assumed, but in practice this may vary down to 90 per cent. To allow for this, it has been the general practice to increase the temperature rise obtained from the analogue by 10 per cent.

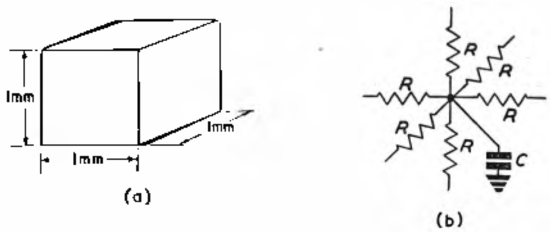


Fig. 4. The three dimensional analogue

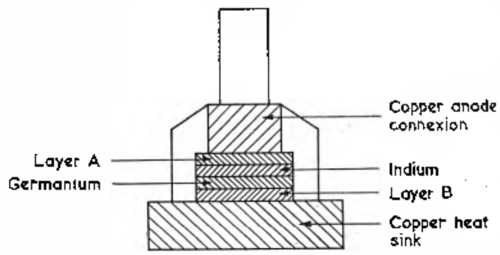


Fig. 5. A typical semiconductor

With regard to the actual junction position, this is usually assumed to be midway through the semiconductor slice, but this may vary with different manufacturers, as it will depend to some extent on the penetration of the impurity into the semiconductor materials.

In most cases it is possible to measure the temperature of the heat sink, and this usually has a large enough thermal capacitance to damp out any temperature fluctuations. Hence the base of the sink can be used as a reference temperature.

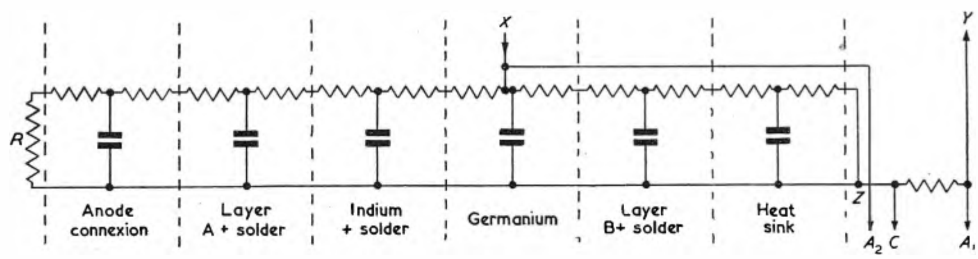


Fig. 6. Basic analogue circuit

R = large resistance to represent heat loss from anode lead

A_1, A_2, C = oscilloscope connexions

Fig. 6 shows a basic analogue circuit to represent the semiconductor shown in Fig. 5. In practice, to get more accurate results it is usual to represent the various layers by more than one element on the analogue, to simulate the fact that the thermal capacitance is distributed right through the layer.

By applying a current of the correct waveshape at x and Y, a trace representing the temperature of the junction above the base (i.e. voltage between X and Z can be obtained.

Typical oscilloscope displays are shown in Fig. 7(a) and 7(b).

The trace A_1 in each case is the current being fed into the analogue, which represents the heat being produced at the junction. The trace A_2 is the voltage appearing between X and Z, representing the temperature of the junction above the reference temperature.

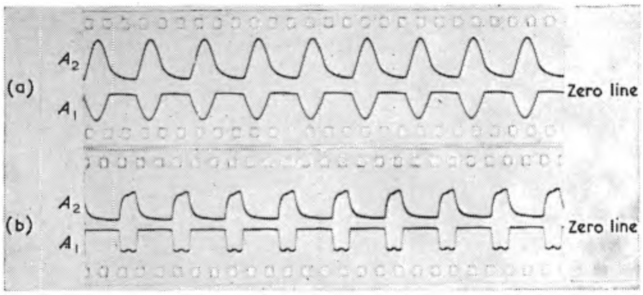


Fig. 7. (a) Single-phase condition
A_1 = heat input, A_2 = junction temperature (scale T°C/cm)
(b) Three-phase bridge condition
A_1 = heat input, A_2 = junction temperature (scale 2T°C/cm)

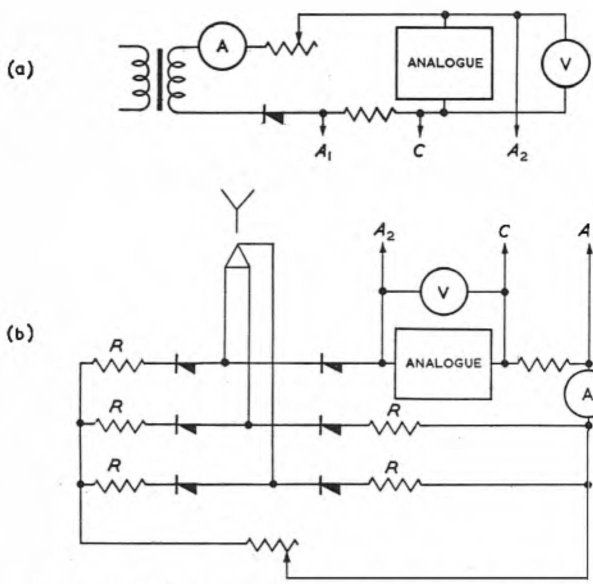


Fig. 8. (a) Single-phase condition (b) Three-phase condition
R = Total resistance of analogue

In the three-phase case the temperature scale in Fig. 7 is half that of the single-phase case. It will be seen that this gives the junction temperature at any instant over a complete cycle.

The circuits used for obtaining the current waveshapes are shown in Fig. 8.

Now considering these oscillograms, it is seen that the temperature of the junction follows very closely the current waveshape, and that the peak junction temperature, which is higher in the three-phase condition, is considerably higher than the average junction temperature. For this reason, when semiconductors are being rated it is important to consider the peak junction temperature and not the average junction temperature.

Another point to be borne in mind when rating the semiconductors is the variation in forward drop. This analogue gives the thermal conditions for a certain heat input in watts. In practice, this heat input is a product of the current and the forward drop, and therefore if an average temperature for a particular type of cell is to be given, some allowance must be made for the variation of this forward drop.

Other Applications of the Analogue

It will be seen that an analogue of this type can be used to investigate almost any heat flow problem.

Two other known applications are the study of heat loss from insulated steam pipes[1] and the study of steady-state heat transfer in power transistors[2]. In this second case, use was only made of thermal resistance; thermal capacitance not being considered.

Some other possible applications are: Cooling conditions in casting materials, loss of heat through building and furnace walls, drying of cores and heat flow in internal combustion engines.

Conclusion

In the thermal analogue there is a simple and accurate method of studying heat flow in situations where theoretical methods would become extremely complicated.

It is particularly useful where there are considerable temperature fluctuations. In semiconductors where temperatures tend to follow the current waveshape (e.g. 50c/s) passing through the junction, the method offers a simple thermal analysis of a situation that would be difficult to handle in theory.

Acknowledgment

The author is indebted to The English Electric Company Ltd for permission to publish this article.

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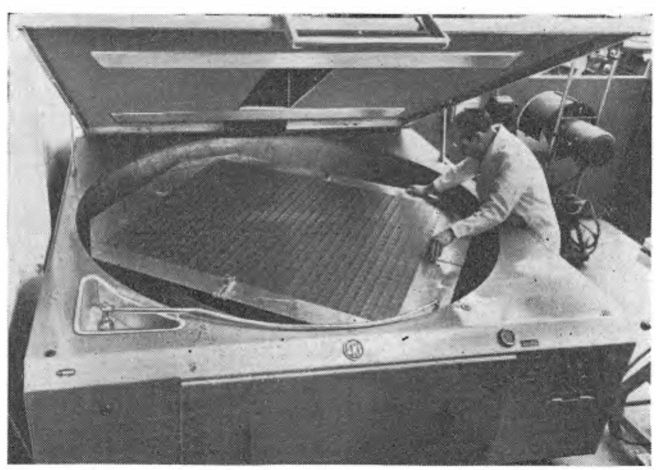
LARGE SIZE PRINTED CIRCUITS

Printed Circuits Ltd, a company within the A.E.I. Group, are currently producing what are claimed to be the largest printed circuit boards ever made, thereby opening up new fields of application as well as improving the production rate of smaller units.

The photograph illustrates part of the automatic equipment recently installed in the factory at Borehamwood. In this instance, 1 100 separate circuits, incorporating inductors, are being processed on one board 5ft by 5ft.



A Multi-Range, Low Frequency Band-Pass Filter

By J. B. Bratt*, B.A., B.Sc.

The band-pass filter described was developed as part of an amplitude control circuit used with apparatus designed for research in the field of unsteady aerodynamics. A frequency range from 16c/s to 175c/s is covered in nine bands, the ratio of the upper to the lower frequency of each pass-band being 1·38. The basis of the filter circuit is the frequency selective RC valve amplifier modified to give zero output at frequencies of 0 and ∞. With two such amplifiers in cascade the response in the pass-band is flat within ±5 per cent, and attenuation of the 3rd harmonic of the lower frequency limit is 33 times. The use of a special feedback network in the amplifiers enables a change of pass-band to be made by switching two capacitors in each network.

(Voir page 506 pour le résumé en français: Zusammenfassung in deutscher Sprache Seite 510)

THE filter described forms part of amplitude control circuits used with apparatus for the measurement of aerodynamic forces on oscillating models in wind tunnels. A self-excited spring-constrained vibratory system is used in this apparatus, in which electro-mechanical vibration generators derive their input from a displacement pick-up on the system, and a stable amplitude of oscillation is obtained by making use of a constant amplitude reference signal derived from the pick-up output by squaring and filtering. The input to the vibration generators is made proportional to the difference between the reference signal and a smaller signal obtained from the pick-up output.

It is essential to remove the harmonics from the square waveform to avoid complications in the measurement of power, which forms part of the test procedure. Since the frequency of the vibratory system alters for different conditions of aerodynamic loading, a filter with a reasonably flat topped response is required, and the use of a band-pass characteristic is desirable to avoid excitation of the system in lower frequency modes. The filter must also be multi-range in order to cover the frequency range of the tests, obtained by variation of the elastic constraint in the system.

Suitable characteristics for a filter for this purpose are:

(a) Frequency range 16c/s to 175c/s covered in nine stages.

(b) For each stage the variation in output in the pass-band to be not more than ±5 per cent.

(c) The third harmonic of the lowest frequency in any pass-band to be attenuated to 3 per cent or less of its value.

Condition (c) implies that with a square-wave input third harmonic content of the output signal will be not more than 1 per cent and higher harmonics proportionally less. Even harmonics are absent with a square-wave input.

Phase shift in the filter is of no importance in the present application since a special circuit is provided for controlling the phase of the current in the vibration generators.

General Characteristics of Filter

The basis of the filter circuit is the frequency-selective negative feedback RC amplifier[1] in which the feedback network is frequency dependent. If the transmission of the feedback circuit is zero at one particular frequency the amplifier exhibits a resonance characteristic, the gain being a maximum at that frequency. Two such amplifiers in cascade give a band-pass characteristic.

For each range in the present filter the two amplifiers are tuned to frequencies in the ratio 1·0 to 1·38, and adjacent ranges are overlapped by giving the higher resonance frequency of the lower range and the lower resonance frequency of the higher range a ratio of 1·38 to 1·30. Change of range is effected by switching components in the feedback circuits.

Some preliminary experiments were made with a circuit based on a filter described by Morris and Dawe[2] in which

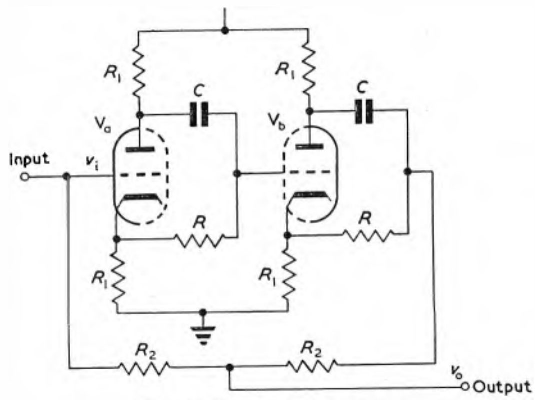


Fig. 1. The feedback network

positive feedback is employed via a zero-phase-shift network. This was abandoned in favour of a negative feedback circuit after considerable difficulty had been experienced in maintaining the shape of the transmission characteristic on switching from range to range. Some consideration was also given to the symmetrical, balanced twin-T RC filter[3] for the feedback network, but in view of the fact that switching of three accurately matched components for each amplifier would be required in order to maintain a constant Q-factor for all frequency ranges, the circuit described below was developed, in which variation of only two components is required.

Feedback Network

The basic circuit is illustrated in Fig. 1. Each half of the double triode V_a, V_b forms a phase-splitter with equal anode and cathode loads R_1. The RC circuit connected between anode and cathode gives an output which is constant in magnitude but varies in phase from 0° to 180° as the frequency varies from 0 to ∞. The output from V_a is directly coupled to the grid of V_b, and the final output is taken from the fixed potential divider R_2R_2 connected between the output of V_b and input of V_a.

* National Physical Laboratory.

If the gain of the phase-splitters is assumed to be unity and the shunting effect of R, C and R_2 is negligible, the output of V_a may be written

$$v_1 = -v_i \frac{1 + jx}{1 - jx} \quad \ldots\ldots\ldots\ldots\ldots\ldots (1)$$

where $x = 1/\omega RC$, and ω is the angular frequency. The output of V_b is thus

$$v_2 = v_i\left(\frac{1 + jx}{1 - jx}\right)^2 \quad \ldots\ldots\ldots\ldots (2)$$

and the final output v_o is given by

$$v_o = \tfrac{1}{2}(v_i + v_2) \quad \ldots\ldots\ldots\ldots\ldots\ldots (3)$$

The transmission of the network $\beta = v_o/v_i$ may thus be written

$$\beta = \frac{1 - x^2}{1 - x^2 - 2jx} \quad \ldots\ldots\ldots\ldots\ldots\ldots (4)$$

This is similar to the transmission of a symmetrical, balanced twin-T network[3] with zero input and infinite output impedance, but the cut-off is sharper as shown by the curves of $|\beta|$ in Fig. 2. The transmission is zero when $\omega RC = 1$, which corresponds to a shift of 90° in each of the triode circuits.

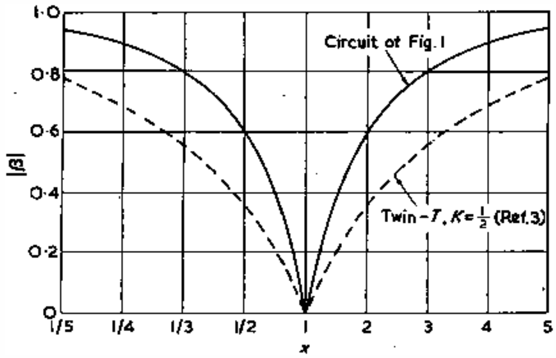


Fig. 2. Transmission of feedback networks

In a practical case the gains of the phase splitters will be somewhat less than unity, and this can be allowed for by making one of the arms of the fixed potential divider variable. This is adjusted until zero transmission is obtained at one particular frequency on sweeping through a range of frequencies.

Basic Tuned Amplifier

The gain of an amplifier with negative feedback is given by the expression

$$A = \frac{A_o}{1 + \beta A_o} \quad \ldots\ldots\ldots\ldots\ldots\ldots (5)$$

where β is the fraction of the output fed back, and A_o is the gain without feedback. In the present case A_o is real and includes the modulus of the gain of the feedback network; thus β is given by equation (4) which relates to unit gain. Equations (4) and (5) lead to the relation

$$|A/A_r| = \frac{1 + x^2}{[(1 + A_o)^2(1 - x^2)^2 + 4x^2]^{\frac{1}{2}}} \quad \ldots\ldots\ldots (6)$$

where A_r is the gain at resonance corresponding to $x = 1$. In terms of the resonance frequency ω_r, equation (6) may be written

$$|A/A_r| = \frac{\omega/\omega_r + \omega_r/\omega}{\left\{(1 + A_o)^2(\omega/\omega_r - \omega_r/\omega)^2 + 4\right\}^{\frac{1}{2}}} \quad \ldots\ldots\ldots (7)$$

which is shown plotted in Fig. 3 for $A_o = 4$. It is seen from equation (7) that when ω tends to zero or infinity the curve tends asymptotically to $1/(1 + A_o)$.

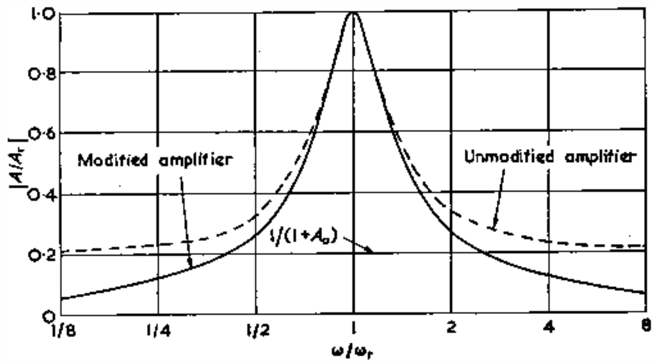


Fig. 3. Feedback amplifier characteristics (A_o = 4)

The transmission of a filter formed by two such amplifiers in cascade is given in Fig. 5. The resonance frequencies are taken in the ratio 1 to 1·38, and the value of A_o is chosen to attenuate the third harmonic of the lower resonance frequency to 3 per cent of the transmission at either resonance. It will be seen that the response over the pass-band between the peaks varies by ±25 per cent of the mean level. This could be made flatter by introducing additional amplifiers tuned to intermediate frequencies; however, this additional complexity can be avoided and sufficient improvement obtained by a modification to the basic amplifier.

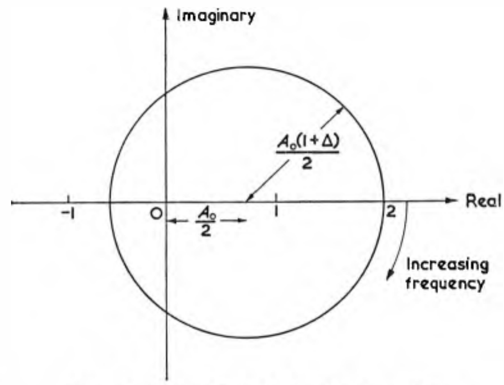


Fig. 4. Plot of βA_o for feedback amplifier

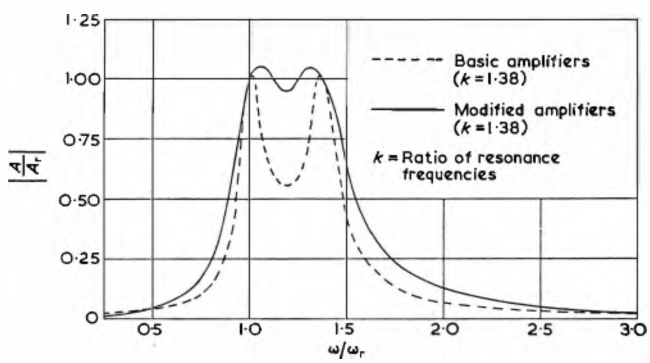


Fig. 5. Theoretical transmission characteristics of band-pass filter

Modified Tuned Amplifier

The response curve given by equation (7) may be modified to give zero output when ω tends to zero or infinity by subtracting from the final output a fraction γ of the input voltage, where γ is real. The advantage of this is that attenuation of the harmonics of the resonance frequency is increased while the sharpness of tuning is hardly affected. Thus, in the band-pass filter, amplifiers with flatter tuning may be employed while the same attenuation is maintained at the third harmonic of the lower resonance.

Equation (5) may now be written in the modified form

$$A = \frac{A_o}{1 + \beta A_o} - \gamma \quad \ldots\ldots\ldots\ldots\ldots \quad (8)$$

When ω tends to zero or infinity β tends to unity, and thus for zero transmission

$$\gamma = \frac{A_o}{1 + A_o} \quad \ldots\ldots\ldots\ldots\ldots\ldots \quad (9)$$

Substitution of equations (9) and (4) in equation (8) leads to

$$|A/A_r| = \frac{2x}{[(1 + A_o)^2(1 - x^2)^2 + 4x^2]^{\frac{1}{2}}} \quad \ldots\ldots\ldots \quad (10)$$

$$= \frac{2}{\left\{(1 + A_o)^2(\omega/\omega_r - \omega_r/\omega)^2 + 4\right\}^{\frac{1}{2}}} \quad (11)$$

which is compared with the unmodified characteristic in Fig. 3.

The effect of the modified amplifiers on the transmission of the band-pass filter is indicated in Fig. 5 and shows a very marked improvement over the unmodified filter, the variation in response between the peaks being now not more than ± 5 per cent of the mean level. The values of A_o to give the required attenuation of the third harmonic of the lowest resonance frequency are 19·12 and 8·91 for the unmodified and modified amplifiers respectively. The equation for the transmission of the filter may be derived from equation (11), and when normalized to give unity response at the resonance frequencies of the amplifiers, takes the form

$$|A/A_r| = \left[\frac{Y\left\{(1/k - k)^2 + Y\right\}}{\left\{(\omega/\omega_r - \omega_r/\omega)^2 + Y\right\}\left\{(\omega/k\omega_r - k\omega_r/\omega)^2 + Y\right\}}\right]^{\frac{1}{2}} \quad \ldots\ldots\ldots\ldots\ldots\ldots \quad (12)$$

where the resonance frequencies are ω_r and $k\omega_r$, and $Y = 4/(1 + A_o)^2$.

Adjustment of γ

In practice the fraction of the input signal subtracted from the final output in the modified amplifier is adjusted by means of a variable potentiometer. It is clearly impossible to adjust this directly to give zero output at frequencies of zero and infinity, and thus some indirect method of obtaining the correct setting must be employed. The method adopted makes use of the fact that, for a given input at fixed frequency, the output from the amplifier has a minimum value for a particular setting of the γ-control, and if the frequency is sufficiently remote from resonance this setting will be close enough to the theoretical value $\gamma = A_o/(1 + A_o)$ for all practical purposes.

The value of γ at the minimum setting may be derived from equations (8) and (4), which lead to the result that, when $\partial|A|/\partial\gamma = 0$,

$$\gamma = \frac{A_o}{1 + A_o} \cdot \frac{1 + 4x^2/(1 - x^2)^2(1 + A_o)}{1 + 4x^2/(1 - x^2)^2(1 + A_o)^2} \quad \ldots\ldots\ldots\ldots\ldots\ldots \quad (13)$$

For the practical circuit described later the settings were made with the filter switched to band 7 and at frequencies equal to the sixth harmonic of the resonance frequencies of the amplifiers. The value of γ obtained is given by substitution of $x = 1/6$, $A_o = 8{\cdot}91$ in equation (13), which leads to

$$\gamma = 1{\cdot}0106 \frac{A_o}{1 + A_o} \quad \ldots\ldots\ldots\ldots\ldots \quad (14)$$

This differs from the theoretically required value by approximately 1 per cent which in practice is of no importance. The use of still higher frequencies in making the γ-setting was avoided on account of the effect of stray shunting capacitances, while for suitable frequencies below resonance RC couplings in the amplifier circuits tended to become significant.

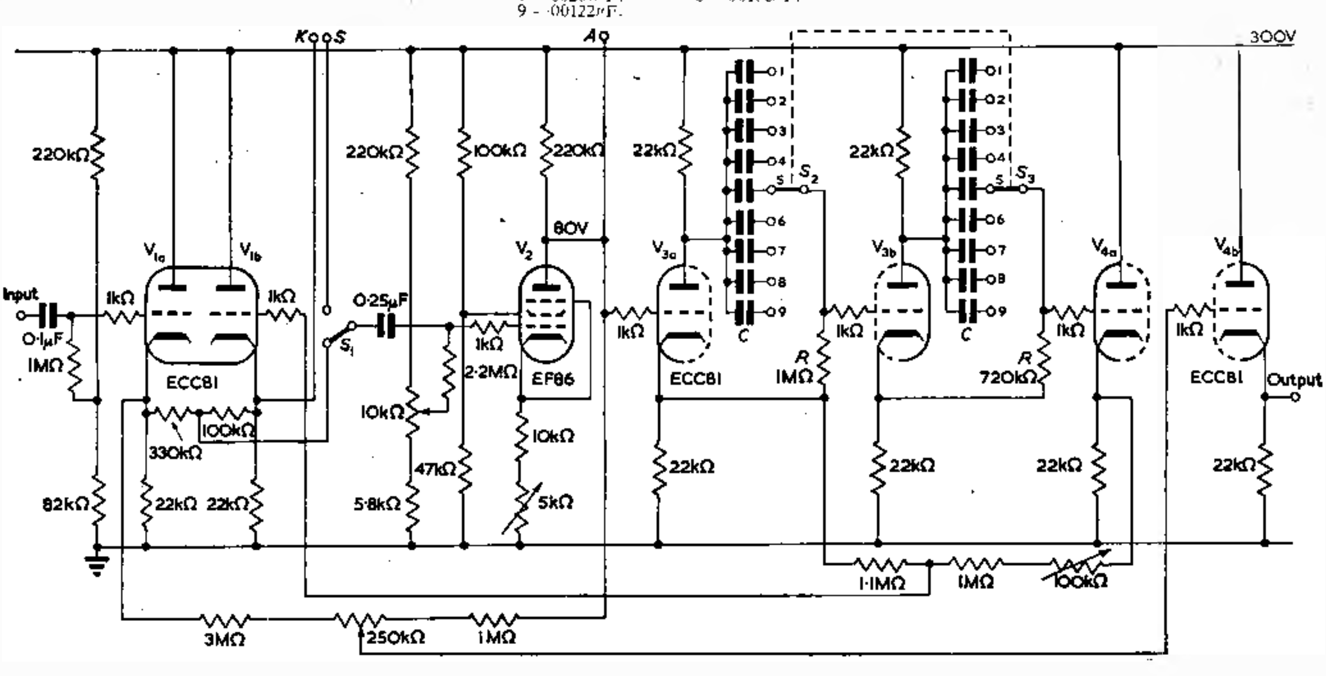


Fig. 6. The feedback amplifier

Values of C

1 - 0·0101μF. 2 - ·00759μF.
3 - ·00593μF. 4 - ·00458μF.
5 - ·00353μF. 6 - ·00270μF.
7 - ·00207μF. 8 - ·00156μF.
9 - ·00122μF.

Practical Circuit

The basic amplifier circuit is shown in Fig. 6. A somewhat lavish use of cathode-followers is made in order to avoid shunting effects of impedances where these would be likely to influence the shape of the transmission characteristic, and in the interests of stability a large amount of negative feedback is introduced into the amplifier valve V_2 and the phase-splitters V_3 of the feedback network via their respective cathode loads. The feedback ratios for V_2 and each half of V_3 are approximately 21 and 35 respectively.

The potential divider supplying the output from the feedback network is connected between the cathodes of the first phase-splitter and the first half of the double triode V_4, which forms a cathode-follower taking the output from the second phase-splitter. The output from the potential divider is then fed back to the input of V_2 via a cathode-follower formed by the second half of the double triode V_1. Subtraction of the required fraction of the input voltage from the output is effected by the potential divider connected between the anode of the amplifier valve V_2 and the cathode of the input cathode-follower V_{1a}. The output from this potential divider is taken directly to the control grid of the final output cathode-follower V_{4b}. With these arrangements shunting effects are kept to a minimum.

Change of frequency range is effected by switching capacitors in the RC circuits of the feedback network rather than by changing the resistors, since this maintains the same shunting effect on the phase-splitter loads for all frequency ranges and keeps the shunting loads at a high value. Capacitors with polystyrene dielectric were used and were selected to have capacitances lying within 0·5 per cent of the calculated values.

The second amplifier is identical with the one described above with the exception of the resistors in the RC circuits of the phase-splitters which are given a value of 72kΩ instead of 1MΩ.

The h.t. supply for these amplifiers is supplied by a 300V regulated power pack.

Setting-up Procedure

Initially the anode of V_2 is set to 80V by means of the adjustment on the control grid potential. The feedback network can then be balanced by switching the input of V_2 to a signal source at S and observing the output at K. When the 100kΩ variable resistor is correctly set a null reading is obtained at K at one particular frequency, which is the resonance frequency of the amplifier with feedback. The gain of V_2 can then be adjusted by means of the 5kΩ variable resistor in its cathode circuit, while maintaining the anode voltage at 80V. The gain from the input of V_2 to K at some convenient frequency is made equal to the value calculated from the loop gain βA_0 by making allowance for the potential divider between the cathodes of V_1. In the present case it was convenient to make the adjustment at three times the resonance frequency, for which the required gain was 9·29.

When these adjustments have been completed, the input of V_2 is switched back to the fixed potential divider associated with V_1 and the setting of γ then made by means of the 250kΩ variable potentiometer in the manner described earlier. In practice small readjustments were made to the loop gains in order to improve slightly the final characteristic of the two amplifiers in cascade. By noting the final loop gains required it becomes a simple matter to repeat the characteristic after replacing faulty valves.

Measured Transmission Characteristics

Experimental transmission characteristics for amplifier resonance frequencies f_r = 20·26c/s and kf_r = 27·75c/s in one case and f_r = 110c/s and kf_r = 151c/s in the other

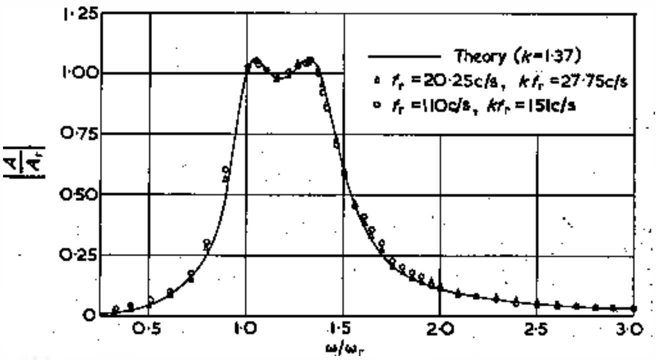


Fig. 7. Measured compared with theoretical transmission characteristics

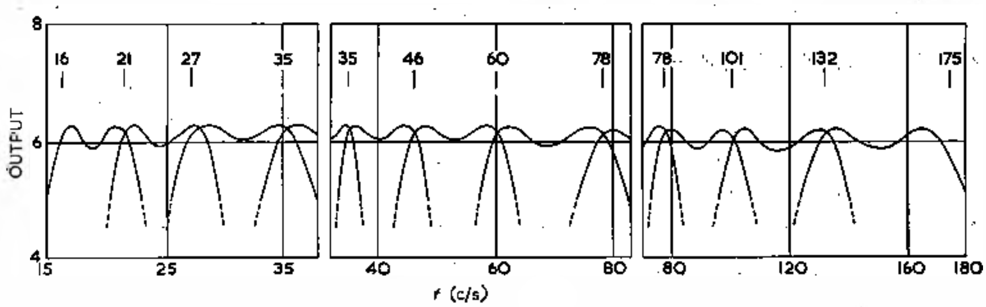


Fig. 8. Measured characteristics for the multi-range filter

are compared in Fig. 7 with a theoretical curve based on equation (12). Since an unknown factor was involved in the voltage measurements, the experimental results have been multiplied by a factor to make the mean level between the resonance frequencies equal to that for the theoretical curve.

The experimental output curves for each pass-band of the final filter are shown in Fig. 8 for an input of approximately 4V peak. The output scale here is arbitrary, the overall gain of the filter being approximately unity. Suitable working limits for the frequency ranges of the pass-bands are indicated above the curves. The third harmonic (not plotted) of the lower frequency limit in any pass-band is not more than 3 per cent of the average output between the frequency limits of the band.

Circuit Stability

The feedback network employed is somewhat at a disadvantage compared with a passive network since a change in the gain of the phase-splitters will throw the circuit out of balance. To examine the magnitude of this effect it may be assumed that the network is balanced for unity gain of the phase-splitters, which subsequently becomes a gain of $1 + \Delta$. Equation (2) now becomes

$$v_2 = v_1\left(\frac{1 + jx}{1 - jx}\right)^2 (1 + \Delta) \qquad \text{.......... (15)}$$

which gives a transmission $\beta + \delta\beta$, where

$$\delta\beta = (2\beta - 1)\,\Delta/2 \qquad \text{............ (16)}$$

The fractional change in gain of the feedback amplifier is expressed by $\delta|A|/|A|$, which may be written

$$\delta|A|/|A| = |1 + (\delta A/A)| - 1 \qquad \text{........ (17)}$$

and from equations (8) and (9)

$$A/A_r = \frac{1 - \beta}{1 + \beta A_o} \qquad \text{............ (18)}$$

whence

$$\delta A/A = \partial A/\partial\beta \,.\, \delta\beta/A = \frac{(1 + A_o)(1 - 2\beta)}{(1 + \beta A_o)(1 - \beta)} \,.\, \Delta/2 \qquad (19)$$

Thus

$$\delta|A|/|A| = \left| 1 + \frac{(1 + A_o)(1 - 2\beta)}{(1 + \beta A_o)(1 - \beta)} \,.\, \Delta/2 \right| - 1 \quad \text{.. (20)}$$

where A_o is maintained constant. If $\Delta = 1$ per cent the change in gain at the resonance frequency amounts to 5 per cent and this rapidly falls to 0·7 per cent at 1·3 times resonance and to 0·1 per cent at 3·0 times resonance. These changes correspond to an alteration in the intrinsic gain without feedback of the phase-splitter circuit amounting to 34 per cent.

The effect of a change in A_o with γ and Δ constant may be obtained from equations (8) and (9) which lead to

$$\delta A/A = \partial A/\partial A_o \,.\, \delta A_o/A = \frac{1 + A_o}{A_o(1 - \beta)(1 + \beta A_o)} \,.\, \delta A_o/A_o \qquad \text{............ (21)}$$

and thus from equation (17)

$$\delta|A|/|A| = \left| 1 + \frac{1 + A_o}{A_o(1 - \beta)(1 + \beta A_o)} \,.\, \delta A_o/A_o \right| - 1 \qquad \text{............ (22)}$$

For a 1 per cent change in A_o this gives a change in gain of 1 per cent at resonance, which falls off to 0·2 per cent at 3·0 times the resonance frequency.

A plot of the complex loop gain βA_o derived from equations (4) and (16) is given in Fig. 4 for the general case where the gain of the phase-splitters is $1 + \Delta$, the feedback network having been balanced at unity gain. The curve is a circle of radius $A_o(1 + \Delta)/2$ with its centre on the real axis at the point $A_o/2$, 0. If $\Delta = 0$ the circle passes through the origin and is tangential to the imaginary axis. In accordance with the Nyquist criterion instability would occur when the circle encloses the point -1, 0, and for this to arise Δ must be greater than $2/A_o$, which in the case of the present filter is equivalent to an increase in the gain of the phase-splitter circuit amounting to 22·4 per cent. Since in practice gains will tend to fall with time, the circuit may be regarded as inherently stable.

Acknowledgments

The author would like to thank Mrs. N. C. Woodgate for assistance in the construction and testing of this equipment.

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A High Power Silicon Rectifier Equipment

A silicon rectifier equipment, which it is claimed is one of the largest of its kind to be made in this country, has recently been demonstrated by its manufacturer, Standard Telephones and Cables Ltd at their Rectifier Works, Harlow, Essex.

This equipment has an output of 20 000A at 130V and will be shortly installed at the Electrolytic Metals Corporation Plant in Transvaal, South Africa for the electrolytic production of manganese.

The input to the rectifiers is taken from a 6·6kV, three-phase, 50c/s supply and a single 3 150kVA regulator controls the entire equipment, which is split into four transformer-rectifier sections. A.C. and d.c. isolators allow the removal of one section from service for maintenance purposes, while leaving the plant operating at 75 per cent full load.

The four rectifier cubicles are paralleled on the d.c. output side and supply the complete power requirement for the electrolytic process. The total output current is maintained to within 2 per cent of any required value between 15 000 and 20 000A by a constant current control loop to the main input regulator; the required current setting being made by the plant operator at the control panel in the electrolytic plant area. The regulator, in addition, is able to control the output down to 65V at no load.

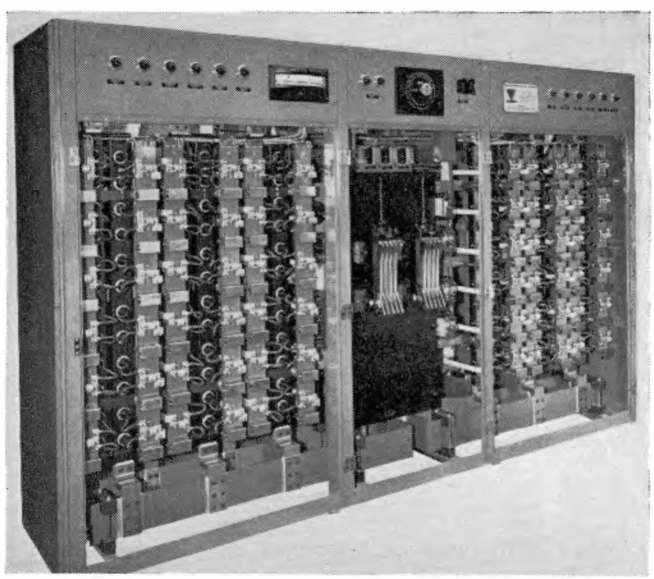
Each of the four rectifier cubicles contains two separate three-phase bridge connexions of RS8 series silicon rectifiers, supplied from common a.c. busbars and feeding via separate isolators into the main 20 000A busbar.

These rectifiers have a continuous rating of 100A with a surge rating of 1 000A and can operate at 300V p.i.v. with a case temperature of 100°C.

Six hollow aluminium busbars form the arms of each three-phase bridge and provide the cooling, by a circulatory oil system, to the twelve paralleled rectifiers in each arm. In every arm of a cubicle there are thus twenty-four parallel operated silicon rectifiers.

Special high-speed fuses are used to protect each pair of rectifiers and the rectifier cubicles are fitted with a pyrometer to indicate busbar temperature during operation.

A cubicle containing two separate three-phase bridge connected rectifiers. The isolating d.c. switch is shown in the centre panel. No d.c. circuit-breaker is provided as overload protection is arranged by an a.c. circuit-breaker tripping from current transformers.



A Voltage Operated Logarithmic Amplifier

By R. F. Mathams*, B.Sc.(Eng.)

A circuit is described which accepts an input from a voltage source and produces an output proportional to the logarithm of the input voltage over a range of inputs from 0·1V to 100V or more. The input voltage is chopped by a vibrating relay and differentiated. The time taken for the resulting exponential pulses to decay to a reference voltage is proportional to the logarithm of the input voltage.

(Voir page 506 pour le résumé en français: Zusammenfassung in deutscher Sprache Seite 510)

CURRENT operated logarithmic amplifiers have been in use in reactor instrumentation for a considerable time. In this application they accept a current input from an ionization chamber and produce a voltage proportional to the logarithm of this current over about 7 decades. The current is fed on to the grid of a pentode operating in the region where $i_g = i_o e \exp. (V/kT)$. The resulting changes in anode voltage are amplified and overall negative feedback is applied to the screen of the pentode to keep the total cathode current and hence the space charge constant. A logarithmic amplifier used for other applications, in particular in a reactor computer, might be required to work with a voltage input.

The logical method of producing a voltage operated logarithmic amplifier would, at first sight, appear to be to connect a high resistance to the grid of the pentode of the current operated amplifier described above. This, however, is not a practicable solution for the following reason. The input resistance of the pentode grid forms a variable potential divider with the high resistance connected to it. A signal proportional to the logarithm of the voltage input appears across the grid and earth terminals. This signal is about 0·2V per decade of input current and, of course, opposes the applied voltage. Hence, for small input voltages below about 10V, the input current varies considerably from the ideal virtual earth value, resulting in large errors.

This difficulty may be overcome by a feedback arrangement shown diagrammatically in Fig. 1. The grid voltage is held practically constant by amplifying any error voltage with a chopper stabilized computing amplifier and driving the cathode of the pentode in such a sense as to reduce the grid error voltage to negligible proportions.

While the latter method is reported to be satisfactory, it has the disadvantage of requiring an auxiliary amplifier. An alternative method capable of covering a range of at least three decades of input voltage is described below.

List of Symbols	
i_g =	Grid current of pentode (A)
i_o =	Constant dependent on pentode characteristic (A)
k =	Boltzmann's constant ($eV. (°C)^{-1}$)
K =	Arbitrary constant (V)
t =	Time (sec)
T =	Absolute temperature (°K)
V =	Voltage (with respect to earth) (V)
V_{in} =	Input voltage (with respect to earth) (V)
V_{out} =	Output voltage (with respect to earth) (V)
V_2 =	Reference voltage (with respect to earth) (V)

General Principles and Analysis

Providing $t \gg CR$, the response of a quasi-differentiating circuit to a square wave input is a series of decaying exponentials of the form:

$$V = \pm V_{in}\, e^{-t/CR}$$

where V is the voltage appearing across the resistor, and V_{in} is the peak-to-peak value of the input square wave.

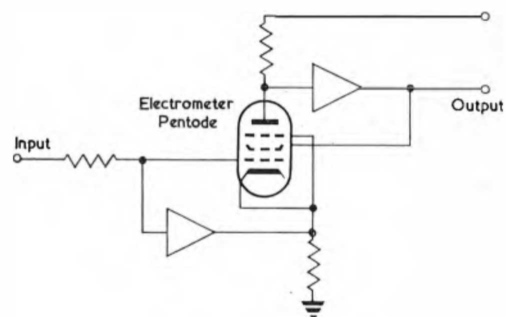


Fig. 1. Current operated logarithmic amplifier modified to accept a voltage input

The time taken for the output voltage V to decay to a reference voltage V_2 is t_2 where:

$$V_2 = V \exp(-t_2/CR)$$

Hence $t_2 = CR \ln V/V_2$.

Thus by measuring the value of t_2 a signal proportional to $\ln V/V_2$ may be obtained.

The principle of operation is therefore to chop the input voltage at a suitable frequency of the order 50 to 500c/s and apply the resultant square wave to a quasi-differentiating circuit and measure the time taken for the decay of the output voltage to a reference level V_2. Referring to the block diagram Fig. 2, the d.c. voltage is chopped by a high quality chopper relay and applied to the RC differentiating circuit. The resulting waveform is amplified by a direct-coupled amplifier with overall negative feedback and the following characteristics:

(a) High input impedance.

(b) Good gain stability.

(c) Good zero stability.

(d) Ability to accept input voltages ranging from 10mV to 100V without suffering damage and without any blocking or hysteresis effects. The ampli-

* *Associated Electrical Industries Ltd.*

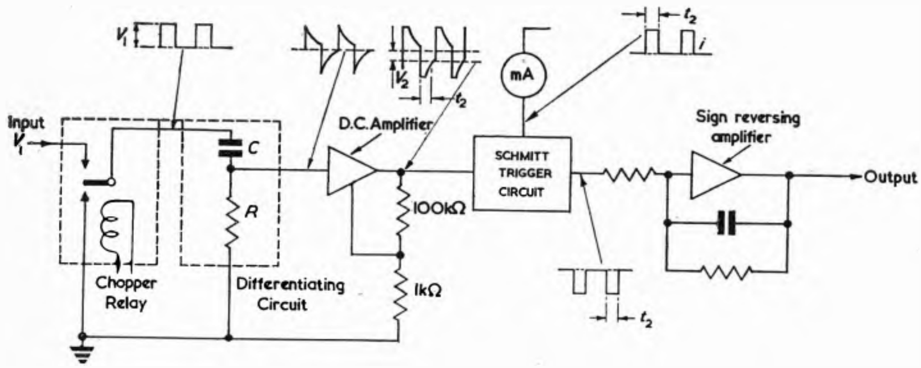


Fig. 2. Block diagram of logarithmic amplifier

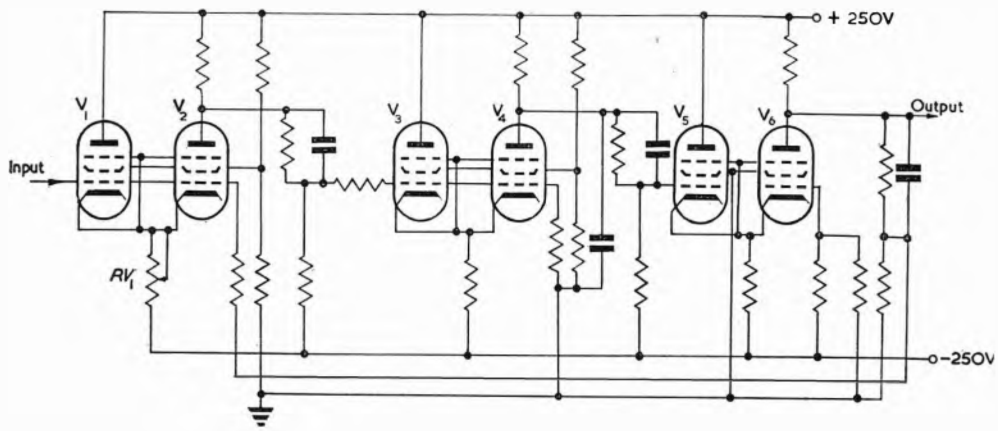


Fig. 3. D.C. amplifier with feedback for amplifying exponential pulses

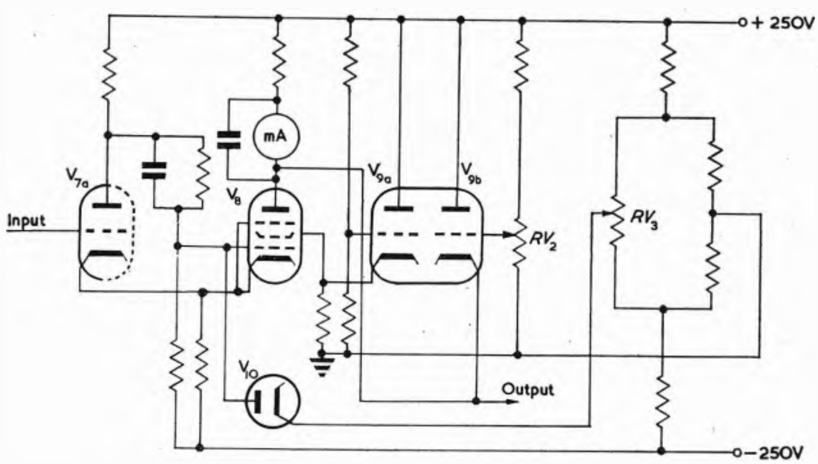


Fig. 4. Schmitt trigger circuit producing pulses of width t_2

fier cannot of course work linearly over this range of inputs, but this is not required. It is merely necessary that the amplifier should work linearly when the input voltage falls to the reference level of about 10mV.

The amplifier output is applied to a Schmitt trigger circuit designed to trigger when the amplifier output voltage rises above a level corresponding to the reference input voltage V_2. Hence, pulses of width t_2 and constant repetition frequency are available.

The d.c. component of these pulses is clearly proportional to t_2 and hence the logarithm of the input voltage. The pulses are therefore amplified and smoothed to give a d.c. output proportional to the logarithm of the input.

Detailed Description of Circuits

Chopper Relay and Differentiating Circuits

The chopper relay is a changeover device specially designed for drift corrected amplifier operation with short transit time between contacts and symmetrical dwell times. The operating coil is electrostatically screened to reduce pick-up by the contact leads. In this design the relay is driven at 50c/s.

The *RC* differentiating circuit comprises a polystyrene dielectric capacitor and a high stability resistor. For operation at 50c/s over a range of three decades the time-constant of this circuit must be less than 1·45msec. A polystyrene capacitor is necessary to keep dielectric hysteresis effects as small as possible. It is also desirable that the differentiating circuit should be of high impedance to avoid loading any previous circuit. However, this tends to increase electrostatic pick-up and for this reason the circuit is enclosed in an electrostatic shield.

Amplifier

The requirements of high gain and zero stability and absence of blocking effects suggested the use of a direct-coupled amplifier comprising three cathode coupled pairs with overall feedback as shown in Fig. 3. The absence of any large capacitors in the design helps to ensure that no hysteresis effects will occur, and the use of cathode coupled pairs reduces the effects of heater drifts. The high input impedance is maintained by taking the negative feedback to the second valve of the first stage. Stability is achieved by means of a dominant lag capacitor in the anode circuit of the second stage. The closed loop gain is 40dB with a bandwidth of over 5kc/s.

Trigger Circuit

This is a Schmitt circuit as shown in Fig. 4 with a diode clamp on the grid of the pentode to define the triggering potential. Valve V_{7a} is normally conducting until the input voltage falls below the triggering potential whereupon the pentode V_8 conducts. The amplitude of the current pulse through the moving-coil meter and the anode load is limited by the triode V_{9b} which defines the voltage pulse at the anode. The moving-coil meter therefore passes a mean current proportional to the pulse width and hence to the logarithm of the input voltage with a scaling factor determined by the current pulse amplitude which is variable by means of RV_2.

SMOOTHING CIRCUIT

The negative pulses at the anode of V_8 are applied to a single stage d.c. amplifier (Fig. 5) with anode-follower feedback which performs the triple function of phase inversion, smoothing and providing a low impedance d.c. output. The time-constant of the feedback circuit determines the degree

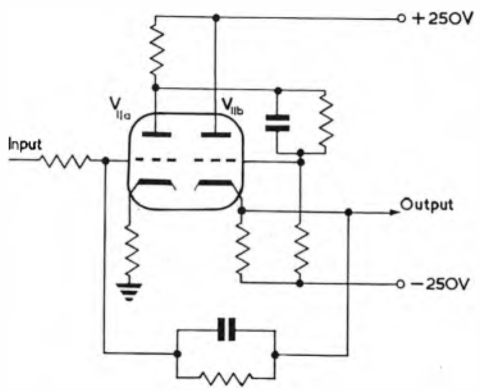


Fig. 5. Smoothing circuit to give low impedance d.c. output

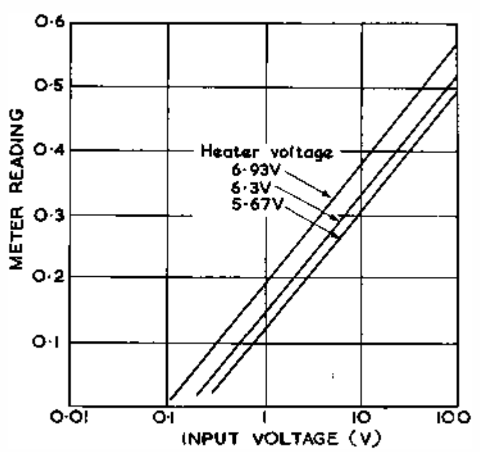


Fig. 6. Calibration variation with heater voltage

of ripple in the output waveform and the value chosen is a compromise between speed of response of the circuit to rapidly changing inputs and an acceptable ripple factor.

Performance

The general performance of the amplifier was in accordance with the design calculations. The equation describing the ideal performance of the device is:

$$V_{out} = K \log (V_{in}/V_2) \text{ for } V_2 \leqslant |V_{in}|$$

Fig. 6 shows typical calibration curves and the effects of changes of heater voltage. It can be seen that this effect is to change the value of V_2 without affecting K. This is almost certainly due to changes in the operating voltage of the trigger circuit with cathode temperature. The nominal value of V_2 was 0·16V and K was 11·5 for the output voltage and 0·18 for the reading of the meter, M in Fig. 2.

A change of mains voltage from −10 per cent to +10 per cent produces a change in meter reading corresponding to a factor of 2·4 in the input voltage. Assuming that a constant voltage transformer would stabilize normal variations of mains voltage to within ±1 per cent, the use of a c.v.t. to stabilize the heater voltage would reduce the errors to ±5 per cent referred to the input voltage of the amplifier. If further stability is required, valves may be fed from the stabilized h.t. supply or from a stable voltage oscillator. No unbalance of the amplifier was observed during these changes of heater voltage.

A zero stability test was carried out with the valve heater voltages stabilized by means of a constant voltage transformer. During the first hour, the meter reading changed by 3μA and during the following five hours the reading remained within ±1μA of its final value. Similar results were observed for the output voltage drifts. If these drifts after the first hour's running, are referred back to the input of the device, they correspond to a change of input voltage of +1·5 per cent. This percentage is, of course, independent of magnitude.

The logarithmic amplifier has since been used with a reactor computer incorporating a commercial logarithmic amplifier of the type shown in Fig. 1. Comparing the performance of both amplifiers with input voltage ranging from 0·1V to 100V (representing practical limits to computing voltages) there was no marked difference in performance between the two types.

Conclusions

Reasonable success was achieved in obtaining a coverage of three decades by means of this logarithmic amplifier without stabilization of the heater voltages. The upper limit of input voltage is set mainly by the available h.t. supplies. In these experiments no attempt was made to exceed 100V, although no difficulty is anticipated up to about 150V.

The lower limit of input voltages, namely V_2, is set by noise and drifts in the d.c. amplifier, and contact potentials and hum pick-up in the chopper relay. Difficulties are also encountered in the trigger circuit at low input voltages. If the peak input to the trigger circuit comes within its backlash region, the trigger circuit acts as a high gain amplifier to the exponential tail of the input waveform. This causes distortion of the logarithmic law at the low voltage end. This difficulty may partly be overcome by increasing the closed loop gain of the d.c. amplifier. It is desirable to employ a high chopping frequency so that the time-constant of the output smoothing circuit necessary to give an acceptable ripple factor is also small enough to allow a fast transient response. However, the necessary amplifier bandwidth to avoid pulse distortion is proportional to the chopper frequency and difficulty in stabilizing the d.c. amplifier at the higher bandwidths is increased.

An analysis of the effects of grid current at the input of the d.c. amplifier shows that providing the negative half of the input waveform is used for triggering the Schmitt circuit, these effects are completely negligible.

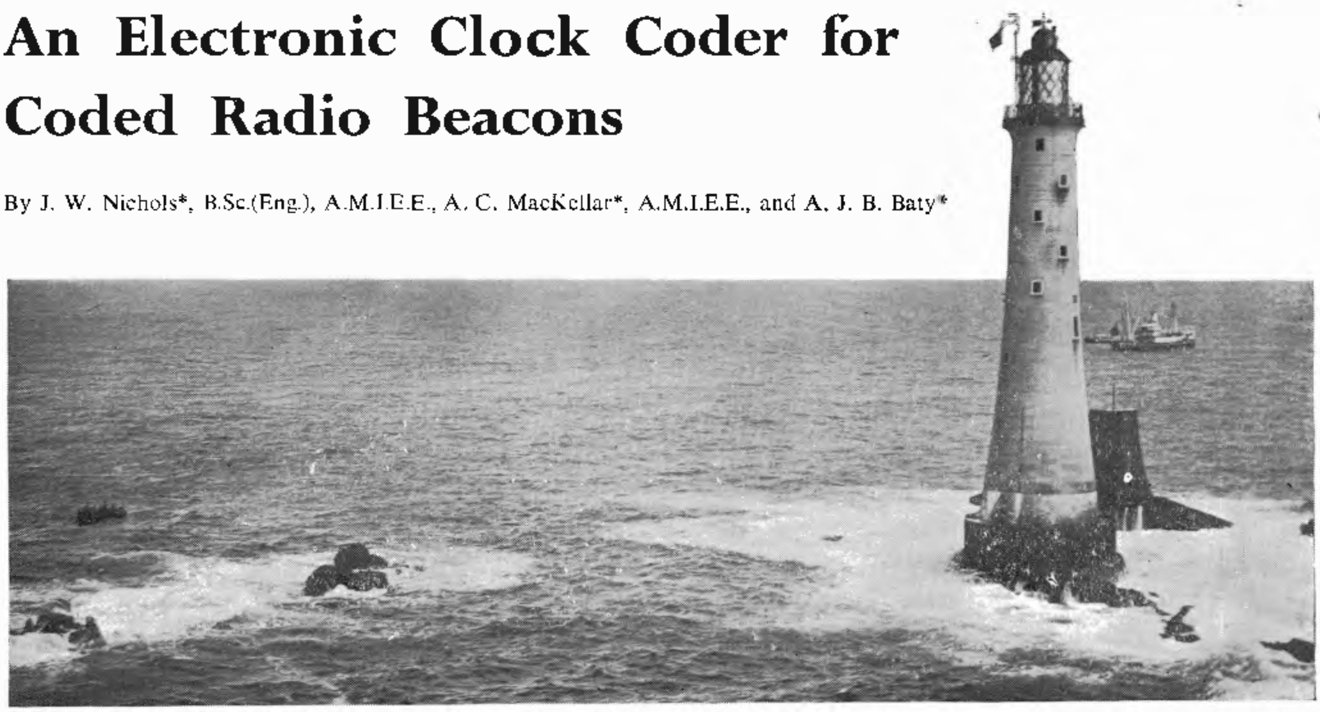
With further development, it is anticipated that this type of logarithmic amplifier could be made to cover over four decades from 0·01V to over 100V.

Acknowledgments

The author is indebted to Dr. T. E. Allibone, F.R.S., Director of the Research Laboratory, Associated Electrical Industries Ltd, for permission to publish this article. Thanks are also due to Mr. B. Lovejoy who made the prototype and assisted in the testing.

An Electronic Clock Coder for Coded Radio Beacons

By J. W. Nichols*, B.Sc.(Eng.), A.M.I.E.E., A. C. MacKellar*, A.M.I.E.E., and A. J. B. Baty*



A crystal controlled clock and electronic coder is described which is capable of originating the coded signal for marine radio beacons in conformity with the requirements of the Paris International Agreement, and also of regulating the time of transmission to a high degree of accuracy. Facilities have been provided for remote or manual control together with a fault warning system designed to monitor the timing and coding of the equipment.

(Voir page 506 pour le résumé en français: Zusammenfassung in deutscher Sprache Seite 510)

A NETWORK of grouped radio beacons has been established around the coast of England and Wales by the Corporation of Trinity House to provide direction-finding facilities to ships using these waters. There is a maximum of six beacons in each group, operated on a time sharing basis, the normal transmission period being of one minute duration, comprising a two letter station identification signal repeated either three or four times followed by a steady tone for 25sec and terminating in the identification signal repeated once or twice. The equipment described in this note is sufficiently versatile to meet other equal or less stringent requirements.

As can be seen from the foregoing description, it is necessary to determine with a high degree of accuracy, the time of transmission of each beacon to prevent overlap. At present this is accomplished on marine radio beacons in the United Kingdom by means of pendulum clocks, on shore and rock stations, and chronometers on light vessel stations. The equipment described below has been designed to operate at any site and to give a timing accuracy better than either the pendulum clock or chronometer, and at the same time to be sufficiently rugged to withstand the handling operations encountered at such sites, a problem which is particularly serious with pendulum clocks.

Coding on existing radio beacons is carried out by means of rotary cam actuated coder mechanisms which are started immediately prior to the appropriate transmission period. Experience with these devices has indicated that the inclusion of rotary cam operated elements has been a cause of unreliability and the equipment described below has been designed to eliminate all moving parts.

Equipment built to the designs described has been in operation under laboratory conditions for more than a year. A manufactured prototype has been built by an electronics firm licensed by the National Research Development Corporation to whom patent rights have been assigned.

Design Considerations

An analysis was made to determine the maximum number of dot elements present in any set of code letters used by the radio beacons under the control of Trinity House, the signal being made up as follows:

(1) A dot of one dot element.

(2) A dash of 3 dot elements.

(3) Space between letters, 3 dot elements.

(4) Space between 'words' of 5 dot elements.

The result obtained shows that all existing codes could be sent by a signal of 30 dot elements or less. If then a complete transmission is built up to conform to the international standard on radio beacon code transmission, it can be shown that the maximum number of dot elements required during the one minute transmission is 240, an allowance being made for a silent period of at least 3sec. Thus a dot unit is of 0·25sec duration and for this a pulse

* Corporation of Trinity House.

output at 4p/s is required. The crystal oscillator which is used both for the derivation of the dot element and for the time keeping function is chosen for convenient division to 4p/s and is kept as low as possible to reduce the scaling necessary and to bring it within the limits of cold-cathode dividing valves. In this case 8kc/s has been considered the most convenient frequency meeting these requirements.

In the coder unit a dot is the interval between one pulse and the next and a dash the interval between one pulse and the arrival of the fourth pulse, the intervals between letters and words being similarly built up to the duration of 3 dot elements and 5 dot elements respectively. The signal group, i.e. identification signal, is repeated at least three times before the 25sec d.f. signal and once after, although if the number of dot elements in the identification signal is small it may be repeated up to four times before the d.f. signal and twice after.

The most important factors influencing the design of the electronic clock coder are as follows:

(1) High reliability.

(2) Complete flexibility.

(3) Provision for remote control.

As the equipment is required to operate both ashore and afloat it is considered essential that a high degree of reliability be obtained to reduce to a minimum the amount of maintenance required. To this end the equipment is designed around the use of gas discharge tubes of the scaling and triggering variety which are considered to be considerably more reliable than hard valves.

As mentioned earlier existing radio beacons are controlled by either pendulum clocks or chronometers depending on the site, both of these timing elements will be replaced by the new electronic unit giving complete standardization at all sites, the character of transmission being controlled by a suitably connected multi-way plug. Thus complete flexibility is achieved as all sites will employ exactly identical equipment resulting in a considerable saving in maintenance effort.

All control functions necessary for the day-to-day running of the equipment, such as time correcting and changeover to continuous transmission during fog or for calibration of ships d.f. can be performed from a remote station over landlines or radio link. A transistorized warning system has also been designed to give an alarm should the clock or coder fail.

General Layout

The equipment comprises four chassis as follows:

(1) Power unit.

(2) Crystal oscillator.

(3) Electronic time-keeping element.

(4) Electronic coding element.

The power supply chassis and crystal oscillator chassis are mounted adjacent to each other behind a standard 19in panel leaving sufficient space to take the power supply and crystal oscillator chassis of a second clock, as normally there will always be a main and standby equipment provided at each site. A second 19in panel is used for mounting both the electronic time-keeping element and the electronic coding element so that the two clock coders required at each radio-beacon station are mounted on three 19in panels.

The only manual controls brought out to the front panel are for resetting and changeover to continuous transmission for fog or calibration service; visual indication is given of the coded signal and of the precise minute reached in the six minute cycle.

Description of the Apparatus

POWER SUPPLY

As the whole equipment includes a time keeping system it is essential that it should operate continuously and be unaffected by even a short interruption in the mains electricity supply. This feature has been accomplished by utilizing the normal 50V station battery together with a transistorized d.c. convertor with outputs provided at +510V d.c., +300V d.c. and −100V d.c., the total power required for each clock coder being 12W. It was early decided during development that the transistor d.c. convertor should operate from the full battery voltage in order to equalize the load on the battery and at the same time provide a system of high efficiency. However, the terminal voltage of a nominal 50V battery of 24 cells can vary from 45V to 67V depending on the state of charge. To compensate for these variations and at the same time utilize the full

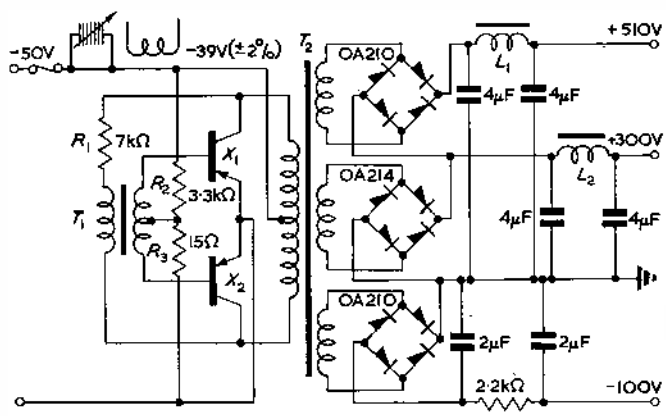


Fig. 1. Transistorized power supply for electronic clock coder

capabilities of the chosen transistors a carbon pile series regulator has been provided with an output of 39V ±2 per cent.

With a battery voltage of 39V the transistors in the d.c. convertor must be capable of withstanding a peak collector voltage of 78V and have a collector current of half an ampere or more. The circuit selected for the d.c. convertor is of the saturable transformer type using two transistors, this circuit having the advantage of a low output impedance necessary for inherently stable output voltages. Normally this type of circuit employs a single transformer which is designed to saturate at a collector current of something less than the peak permissible current of the transistor, this saturation of the transformer results in the loss of drive to the 'on' transistor which comes out of bottoming and a rapid switching action takes place. This method of operation, however, results in rather high current peaks in the transistor together with high transient voltages on the transistor collectors, both these undesirable features constitute a danger to the transistor in the circuit, especially when the transistors are operating at or near the limit of collector current or voltage.

In order to overcome the particular difficulty of excessive collector voltage in this application a second transformer is used to control the switching of the transistors and thus avoid the necessity of saturating the output transformer. Referring now to Fig. 1 which shows the circuit of the transistorized power supply, transformer T_1 is the feedback and switching transformer the action of which is as follows; assume transistor X_1 to have just started conducting and X_2 to be cut off, then the primary magnetizing current of T_1 will rise linearly causing a steady negative potential to be applied to the base of X_1 which will be held in the bottomed state, X_2 on the other hand will be cut off by a positive potential on its base. The primary current in T_1 will continue to rise until the core goes into saturation causing a reduction in the primary inductance and a corresponding increase in primary current, the increase in primary current results in a greater power being dissipated in resistor R_1 and a reduction in the power fed to the bottomed transistor X_1 which now moves out of the bottomed state and is rapidly cut off while X_2 is bottomed. This process is repeated at a frequency determined by the design of transformer T_1 and the value of resistor R_1, the usual frequency range betwen 400c/s and 1kc/s. Starting of the oscillator is ensured by the potential divider R_2 and R_3 which biases both transistors into conduction.

Transformer T_2 is quite conventional in all respects and designed in accordance with normal transformer practice to deliver the required outputs at the particular operating frequency selected. Rectification of the a.c. output is carried out by silicon power rectifiers connected in bridge configuration; final smoothing by standard smoothing filters. The overall d.c. to d.c. efficiency of the power unit is in excess of 80 per cent.

Crystal Oscillator

The circuit of this unit is shown in Fig. 2, the fundamental frequency of 8kc/s being derived from an XY cut flexural mode crystal connected in the feedback path of a two-stage directly coupled grounded emitter amplifier. Connected in this way the crystal presents a high impedance in the feedback path and it is desirable to shunt the crystal by a further resistance-capacitance network R_1, C_2 for reliable starting. Fine frequency control is achieved by adjustment of the pre-set capacitor C_1 in series with the crystal. The output of the crystal oscillator is fed to the phase-splitting transformer T_1 which drives two OC72 transistors in the Dekatron drive stage designed to deliver 80V

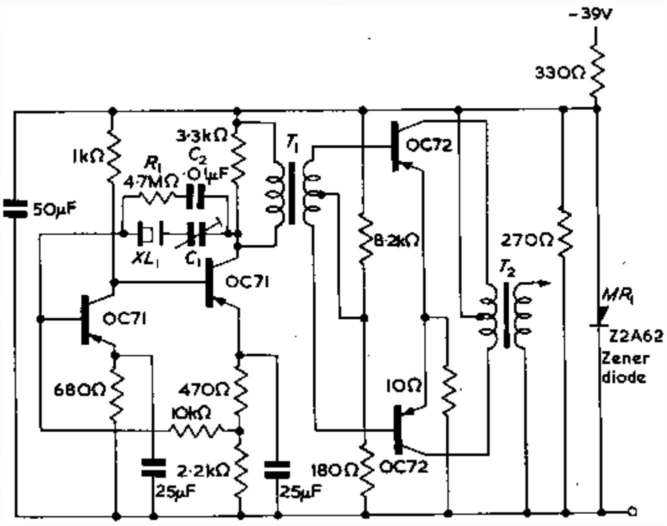


Fig. 2. Transistor crystal oscillator and Dekatron drive unit

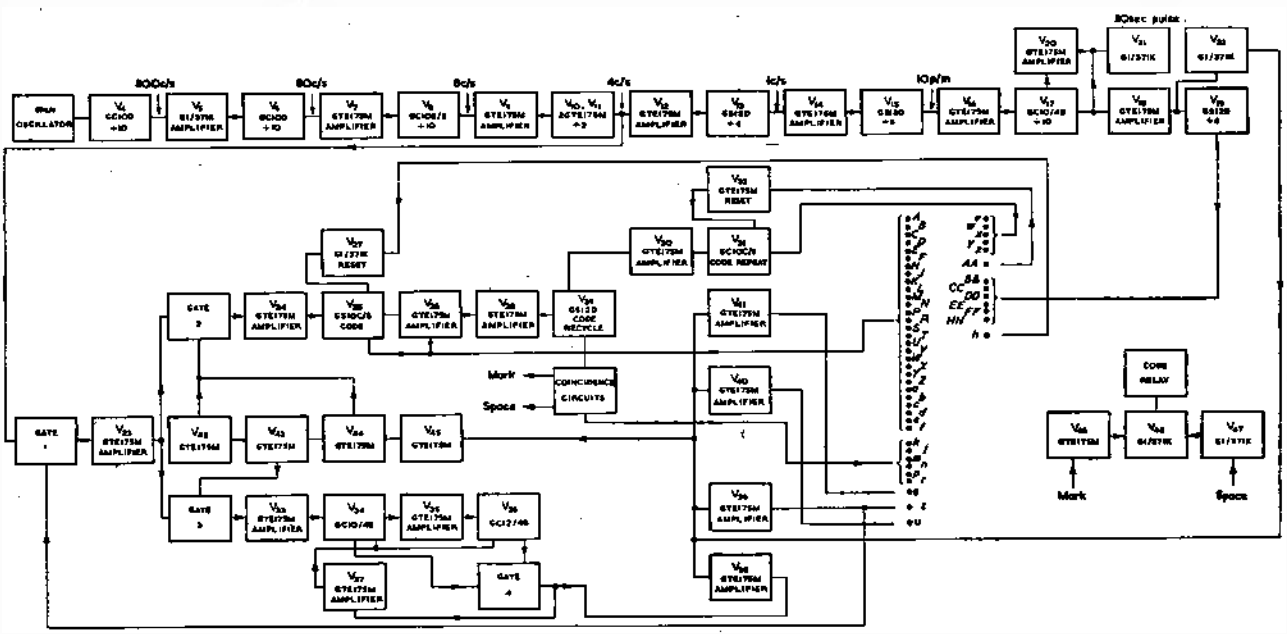
r.m.s. to the first Dekatron divider stage. Stabilization of the voltage applied to this unit is achieved by the use of a Zener diode MR_1 which provides a 6·5V supply from the −39V input voltage

In the particular application for which this equipment is designed, namely the control of radio beacons in coastal waters, it is not considered necessary to provide temperature control of the crystal as the required degree of accuracy can be obtained without this complication. The normal practice being to reset the clock coder daily as a routine measure. If, however, it is necessary to operate the equipment on an unattended basis without time resetting at, say, weekly intervals it may be desirable to provide thermostatic control of the crystal temperature by an oven operated from the d.c. supply.

Electronic Time-keeping Element

This unit, Figs 3 and 4, employs cold-cathode tubes throughout, the 8kc/s input frequency is divided down in

Fig. 3. Block schematic electronic clock coder



successive stages by valves V_4 to V_{17} inclusive, the final output pulse being at 1 pulse per minute. From Figs. 3 and 4 it can be seen that V_4 is a conventional Dekatron divider stage giving an output at 800c/s, V_5 being an interstage pulse forming valve as the output pulse of 40V from the Dekatron is insufficient in amplitude and of opposite phase to that required to operate a further Dekatron stage. The output of V_5 is about 150V negative going and is applied to V_6, another divide by 10 stage, with V_7 as the interstage pulse forming valve. Another stage of division by 10 is performed by V_8 resulting in an 8p/s output which is once again pulse formed by V_9 before being fed to V_{10} and V_{11} which are two trigger tetrodes forming a divide by two stage whose output is at 4c/s; this output is used as the fundamental dot element for subsequent coding and will be discussed later. The circuit configuration of V_{10} and V_{11} is similar to that used in ring counters the method of operation being as follows. Assume V_{10} to be switched on, its cathode will therefore be at a positive potential of about 80V as also will be the trigger of V_{11} through a 1MΩ resistor. The positive going input pulse to the two valves is fed to the trigger of V_{10} and V_{11} simultaneously. As the trigger of V_{11} has already been primed to 80V positive the positive input pulse of 150V applied to both triggers will cause V_{11} to strike. Current drawn through the common anode load of V_{10} and V_{11}, will cause the anode voltage to drop below the level necessary to keep V_{10} struck so it will extinguish leaving V_{11} struck. The cathode of V_{11} is now at 80V positive, thereby priming the trigger of V_{10} in readiness for the next incoming pulse; it can therefore be seen that the output pulses on the cathode of V_{11} are at half the frequency of the input pulses. After being pulse formed in V_{12} the 4c/s is divided by 4 in V_{13} which is a 12-way selector tube whose cathodes are joined to give division by 4 while V_{15} is a similar tube strapped to give division by 6, the resulting output of V_{15} being at 10 pulses/minute. Further division by 10 in V_{17} gives an output of 1 pulse/minute which is also taken to the coding chassis. Provision has also been made for operating a 30sec impulse clock, V_{17} being connected such that outputs are taken from cathodes *B* and *D*, the output of cathode *B* being used to trigger valve V_{29} whose cathode is connected to give a positive output to fire V_{21}, cathode *D* of V_{17} firing V_{21} direct. Isolating rectifiers are included in the common feed to the trigger of V_{21} to reduce to a minimum the mutual loading. Cold-cathode tube V_{20} is needed to reduce the total load on the cathodes of V_{17} to ensure that pulses of adequate amplitude are available for operating the 30sec impulsing valve V_{21} and also V_{18} the pulse forming stage for valve V_{19}. Valve V_{21} has in its cathode relay $T/4$ which operates each time the valve strikes, contact T_2 also in the cathode circuit being a 'Y' contact breaks after all other contacts have completed their operation, thus releasing $T/4$ by extinguishing V_{21}. Impulsing of the T_4 contact is used to step the 30sec impulse clock mounted on the front panel; contacts T_1 and T_3 are employed in the clock coder fault warning system.

Final divivsion in the timing element is performed by V_{19} a 12-way selector tube connected to divide by 6, cathodes 1 and 7, 2 and 8, 3 and 9, 4 and 10, 5 and 11, 6 and 12, being connected together so that each complete revolution of V_{19} divides by 6 twice. Outputs from V_{19} are therefore available at each minute of the 6 minute period which is the normal characteristic period employed in the beacon network. For convenience of wiring, V_{19} is mounted on the coder unit in such a way as to be visible through the front panel thus giving a visual indication of the precise minute reached in the 6 minute period. When continuous transmission is required relay *CC*/8 is energized, the complete operation being discussed later.

Electronic Coding Element

Before proceeding to describe the operation of the coding unit in detail it is necessary to understand the cross connexions in a 50-way plug which is used to determine all the coding details for the required station. Thus, substitution of the 50-way plug is all that is necessary to change the code and sequence minute of the equipment. This plug is mounted on the coder chassis, Figs. 3 and 4, to provide the following functions:

(1) The appropriate minute in the six minute cycle during which the code is sent is determined by connecting:

t to either *BB* for 1st minute
CC for 2nd minute
DD for 3rd minute
EE for 4th minute
FF for 5th minute
HH for 6th minute

(2) The letters comprising the code word by connexion of pins *A-L*, *M-X* and *Y-f* to pin *j* or *k*, *m* or *n*, *p* or *r* respectively as is explained later.

(3) The number of repeats of the code word before and after the long dash by connexion of pins

s to *v*	giving	2	repeats	before
w	,,	3	,,	,,
x	,,	4	,,	,,

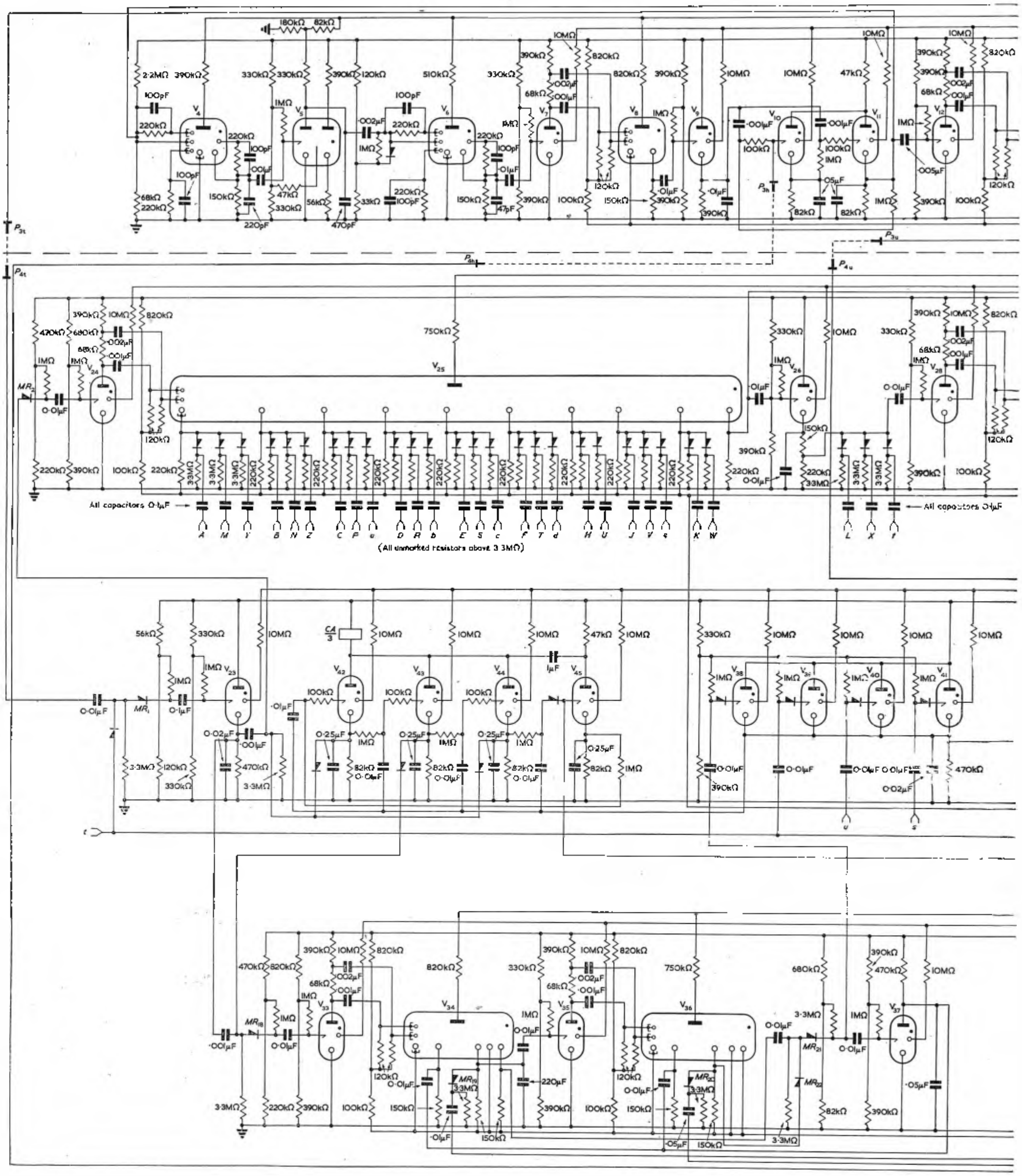
and

u to *w*	giving	2	before	1	after
x	,,	3	,,	1	,,
y	,,	4	,,	1	,,
z	,,	4	,,	2	,,

(4) The length of the code word in dot elements if it is between 20 and 30 dot elements long by connexion of any of the pins.

Y-f to pin *h*.

In order to understand the action of the coder unit it is helpful to use both the block schematic Fig. 3, and the theoretical circuit diagram, Fig. 4. The transmission of a coded signal is initiated by V_{19} stepping to the appropriate minute in the 6 minute period, this causes a d.c. voltage of about 40V to be extended through the appropriate contact (one of the group *BB* to *HH*) on the 50-way plug to pin *t*; this voltage step performs two functions, the first to open gate 1 by applying 40V positive via a blocking rectifier to the junction of a 0·01μF capacitor and a 3·3MΩ resistor, this voltage being sufficient to overcome the standing bias on rectifier MR_1 which then passes the 4p/s to the trigger of V_{23} and hence to gate 2 and gate 3. The second action of the pulse on pin *t* is to fire V_{30} a normal tetrode tube which in turn extends a pulse to the closed count of 4 formed by valves V_{42} to V_{45}, the discharge normally resting on V_{45} in the quiescent state, the discharge then moves round one stage to V_{42} thus opening gate 2 by applying a d.c. potential to MR_2 from the cathode of V_{42} through an isolating rectifier. Gate 1 is provided to limit to the actual minute of code transmissions the period during which the 4p/s is extended to the coding unit thereby preventing the transmission of any beacon from lasting more than one minute due to a fault condition in the coder which might cause such interference with the remaining beacons in the group as would make direction-finding impossible.


All capacitors 0·1µF
(All unmarked resistors above 3·3MΩ)

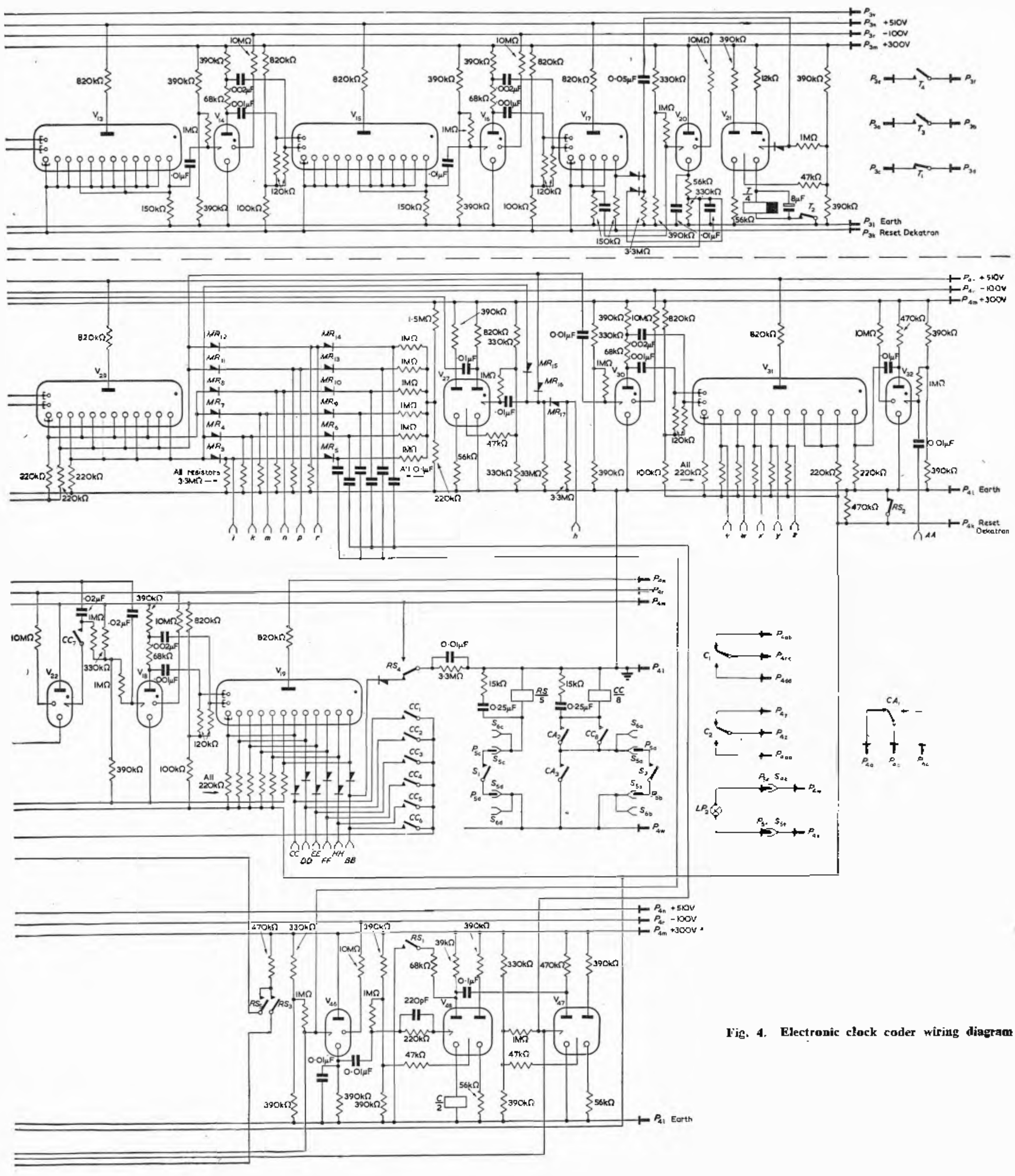


Fig. 4. Electronic clock coder wiring diagram

Gate 2 having been opened, pulses at 4p/s are extended to V_{25}, a 10-way selector tube which provides the identification signal, 3 outputs being taken from cathodes 1 to 6 and 8, two outputs from cathodes 7 and 9 with cathode 0 having one output which fires V_{26} for the reasons described below. This has been done to conserve the number of outputs required as it can be shown that all code letters including the break before the long dash require an even number of dot elements, 3 outputs are not required on cathodes 7 and 9, but they are required on cathodes 1, 3 and 5 as they provide dot elements 21, 23 and 25 in the code, all of which may be required. The outputs from each cathode of V_{25} are taken through isolating capacitors and rectifiers and connected to the 50-way socket pin *A* to pin *f*. In order to provide the 28 outputs required V_{25} is caused to rotate three times; during the first revolution the outputs are taken from pins *A* to *L*, during the second revolution from pins *M* to *X* and during the third revolution from pins *Y* to *f*. V_{26} is triggered from the 10th cathode of V_{25} and in turn fires V_{28} which steps V_{29}, a 12-way selector tube connected to give a count of 3 by strapping cathodes, 1,4,7 with 10, and 2, 5, 8 with 11, and 3, 6, 9 with 12. The discharge on V_{29} normally rests on cathode 3, or any of the cathodes commoned to cathode 3, a d.c. voltage being extended to the coincidence circuits through rectifiers MR_3 and MR_4 to bias 'on' rectifiers MR_5 and MR_6 respectively. Leaving this point of the circuit for a moment and returning to V_{25} which in its normal state rests on cathode '0' and assuming that a start mark signal is required, the first pulse from V_{24} will step V_{25} round one to cathode 1 and a pulse of some 40V will appear at pins *A*, *M* and *Y*, if then pin *A* is connected to pin *j*, as MR_5 is now conducting a pulse will pass via a 0·1 capacitor which is connected to the mark trigger line. Similarly to terminate the mark the output from cathode 2, if a dot is required, or from cathode 4, if a dash is required, is taken from pin *B* or *D* respectively and connected to pin *k* which feeds the pulse through MR_6 to the space trigger rail. It can therefore be seen that outputs can be taken from any of the cathodes of V_{25} and connected to the appropriate mark or space trigger rail. When the discharge in V_{25} steps on to the '0' cathode, V_{26} and V_{28} are fired and V_{29} steps round one position, MR_5 and MR_6 are biased off as the d.c. voltage on MR_3 and MR_4 has been removed, but MR_9 and MR_{10} are biased on by the d.c. fed through MR_7 and MR_8 from cathode 2, 5, 8 or 11 of V_{29}. Valve V_{25} now commences its second revolution with outputs taken from the appropriate cathodes via pins lettered *M* to *X* and connected to pin *m* for mark and pin *n* for space, MR_9 and MR_{10} now biased on extends these pulses to the mark or space rail as selected.

The outputs that were previously connected to pins *j* and *k* are no longer effective as these pins are isolated from the mark and space rails by rectifiers MR_5 and MR_6 which are biased off. When the discharge reaches cathode '0' of V_{25} for the second time V_{29} is once again stepped on, biasing off MR_9 and MR_{10} and opening MR_{13} and MR_{14}, the output from V_{25} being taken from pins *Y* to *f* which are connected to pin *p* for mark and pin *r* for space. During the third rotation of V_{25} it may be necessary to step from any cathode to cathode '0' if the number of dot elements in the identification signal is less than 30. This forced stepping is performed by V_{27}, the action being as follows. Cathodes 1, 4, 7 and 10 and 2, 5, 8 and 11 of V_{29} are connected to the trigger line of V_{27} by MR_{16} and MR_{15} respectively which during the first two rotations of V_{25} cause a steady d.c. voltage to be applied to MR_{17} which is biased off, during the third revolution, however, no biasing voltage is extended to MR_{17} so any incoming pulse on pin *h*, which is connected to that cathode of V_{25} which delivers the last space pulse in the code signal will fire V_{27} and result in its anode potential dropping by 140V, this negative pulse is connected to cathode '0' of V_{25} and forces the discharge to jump to this cathode by virtue of its being so much more negative than the other cathodes. When the discharge in V_{25} reaches cathode '0' for the third time V_{29} once again steps and MR_{17} is biased off, simultaneously a pulse is extended to fire V_{30} which in turn steps V_{31} from cathode '0', its home position, to cathode 1. The function of V_{31} is to determine the number of transmissions of the identification signal both before and after the long dash d.f. signal, the process by which this is done will now be described. On each occasion when the discharge in V_{29} reaches its normal home cathode, i.e. after every three revolutions of V_{25}, V_{31} is stepped on one position and by taking an output after the requisite number of steps the number of code transmissions is selected. Assume that three transmissions are required before the long dash d.f. signal then the output on pin *w* would be cross connected in the 50-way plug to pin *s* to fire V_{41} which in turn steps the ring counter V_{42} to V_{45} round one to V_{43}, this causes gate 2 to close and gate 3 to open as the standing bias on MR_{18} is overcome by the d.c. voltage fed from V_{43}.

Pulses at 4p/s are now fed to V_{33} and hence to V_{34} in order to commence the long dash d.f. signal. A bi-directional 10-way computing tube with intermediate outputs is used for V_{34} which is arranged to send a mark pulse through MR_{19} as soon as it steps from its home cathode and also to fire V_{35} which in turn steps V_{36}, a 12-way bi-directional computing tube with intermediate outputs; the stepping of V_{36} is carried out by V_{35} which acts as the coupling trigger tube between V_{34} and V_{36}. This process is repeated for 9 revolutions of V_{34} with a mark pulse produced each time but as the code relay is already on mark this pulse has no effect. On reaching the 10th revolution in V_{34}, V_{36} is stepped to the 11th position and a space pulse is sent from V_{36} through rectifier MR_{24} and at the same time the d.c. bias on MR_{21} is overcome by the cathode voltage of V_{36} fed through isolating rectifier MR_{22}. V_{34} continues to step at 4p/s for a further 5 pulses when an output is taken to perform two functions, firstly to fire V_{37} which resets to their respective home positions both V_{34} and V_{36} in the manner previously described and secondly to fire V_{38} which now moves the discharge in the set of 4 ring counter V_{42} to V_{45} round one position to V_{44}, thus closing gate 3 and reopening gate 2. It is necessary to continue to count 5 dot units after completion of the long dash d.f. signal in order to give the requisite break before once again transmitting the identification signal.

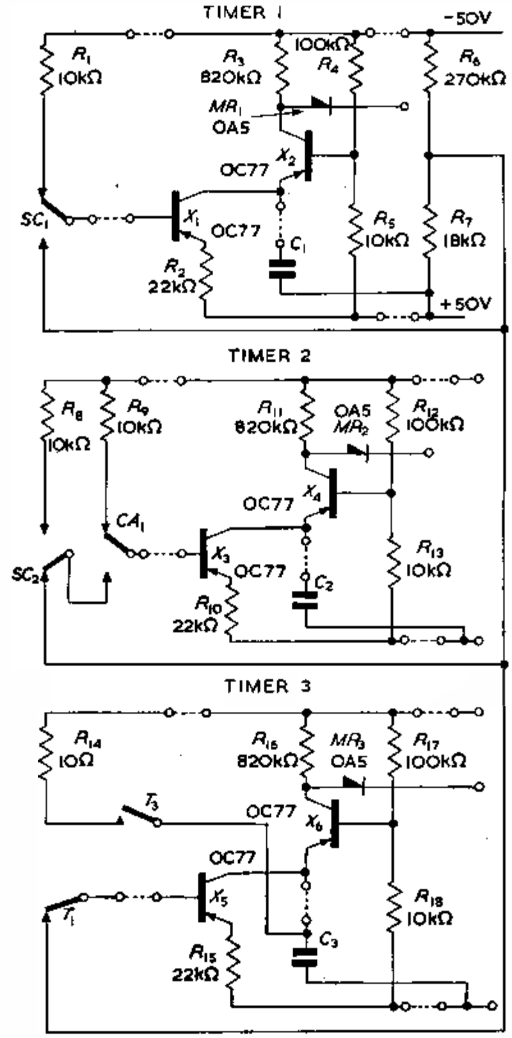
Returning now to that part of the coder which is concerned with the identification signal it will be remembered that V_{25} discharge is on cathode '0', V_{29} discharge on cathode 3 or any of the other cathodes that are linked to 3, V_{31} discharge is at cathode 3 having so far counted 3 transmissions of the code letters. With gate 2 once again open V_{25} will proceed to step round as before and assuming that only one transmission of the identification signal is required after the long dash d.f. signal then pin *x* on the 50-way plug will be cross connected to pin *u* as also will pin *AA*. The pulse on pin *u* fires V_{40} which once again steps the ring counter round one position to V_{45}, thus shutting gate 2, the pulse on *AA* fires V_{32}, which resets V_{31} to its home cathode. When V_{19} steps on receipt of the next minute pulse gate 1 is closed and all valves in the coder circuit are in their normal waiting state ready for the next initiation pulse

Operation of the code relay *C*/2 from the mark and space rails is carried out by V_{46} and V_{47} the code relay being connected in the cathode of V_{48}. The mark rail is connected to the trigger of V_{46}, a normal tetrode trigger tube which fires each time a mark pulse appears on its trigger, the circuit being arranged to be self quenching. When V_{46} fires its cathode potential rises by 150V and the resulting pulse is fed to the trigger of V_{48} the code valve. This code valve is a trigger tetrode of slightly different design to most of the other trigger tubes used and takes a high current which is needed in this case to operate the polarized code relay *C/2*. The circuit of V_{48} is such that once it has fired it stays conducting until extinguished by an external stimulus, in this case a space pulse. Pulses on the space rail fire V_{47}, again a high current tetrode, thus causing its anode voltage to drop by 150V; this negative pulse is extended to the anode of the code valve V_{48} which is extinguished due to the sudden reduction of anode voltage which makes it impossible for the discharge to maintain itself.

As mentioned earlier provision has been made in this equipment for continuous transmission during fog or for calibration purposes, relay *CC/8* being used to perform this function. Contacts CC_1 to CC_6 are arranged to common all the cathodes of V_{19}, thus keeping gate 1 permanently open, contact CC_7 brings valve V_{23} into circuit to give a 'code start' pulse at the start of each minute as it is now no longer possible to obtain an output from V_{19} for this purpose as all its cathodes are commoned. Contact CC_{18} is provided to lock relay *CC/8* over its own contact so that the release of *CA/3* relay does not affect relay *CC/8* during a continuous transmission period. Operation of the manual switch S_2 prepares the circuits for relay *CC/8* which operates when contact CA_2 makes, that is when the equipment transmits its normal code transmission which could be anything up to 5 minutes later, CC_8 locks in CC_8 relay and continuous transmission follows. On release of S_2 contact CA_3 holds relay *CC/8* until the termination of the current transmission period when relay *CA/3* releases in turn releasing *CC/8* and returning the equipment to normal, ready to take up transmission during the normal minute period.

Re-setting

It will be appreciated that when the equipment is switched on, the discharge in each multi-way tube will be positioned at random between any cathode and anode, consequently it is necessary to provide a re-setting circuit which will ensure that the discharge on each tube is taken to the zero cathode. Provision must also be made for striking those tetrode trigger tubes, in divide by two stages or in ring counters, that normally have the discharge resting on them. For re-setting purposes relay *RS/5* has been provided and will normally be operated by the reset switch on the front panel although it could equally well be operated by radio link or by land line connexion. Contact RS_1 is arranged to reduce the anode voltage of V_{48}, the code valve, to a level below the maintaining voltage, thus extinguishing V_{48}. The resetting of all the multi-way cold-cathode tubes is done by contact RS_2 which is normally made and short-circuits the resetting resistor; during reset, however, RS_2 breaks and the resetting resistor is connected in series with all the cathodes of the multi-way tubes except the zero cathode which remains connected to earth. The 470kΩ resetting resistor is also connected in series with the guide potential divider and consequently a positive potential exists across it which ensures that all cathodes connected to the reset line are sufficiently positive to prevent a discharge between any one of them and anode. The only path able to sustain a discharge being between zero cathode and anode with the result that the discharge jumps to the zero cathode as required.



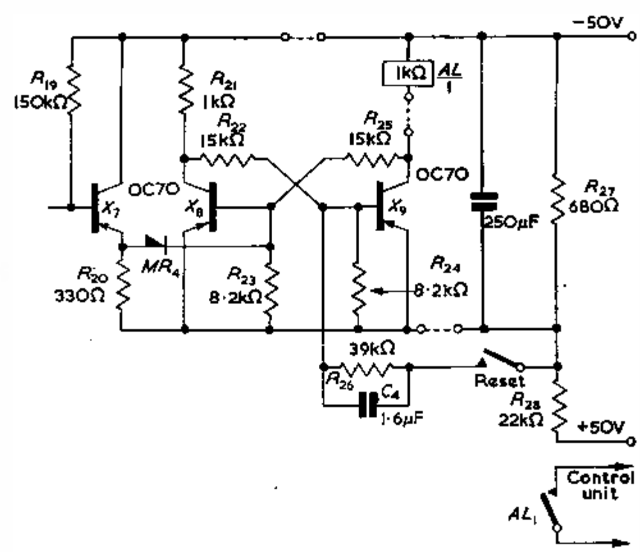


Fig. 5. Clock coder warning system

The set of ring counters V_{32} to V_{45} and the divide by two stage V_{18} and V_{21} have to be primed on resetting the equipment as the operation of the circuits require that a discharge shall rest in one of the tubes. As this is not achieved on switch-on contacts RS_3 and RS_5 are used to connect the trigger of V_{19} and V_{46} to the h.t. line through a 470kΩ resistor thereby causing these tubes to strike, the circuits are then primed and await the next input pulse before stepping on. The last contact RS_4 on the reset relay is employed to provide a start impulse on cathode '0' of V_{19} when the equipment is reset; a 0·01μF capacitor being charged up to h.t. voltage and then discharged to cathode '0' when RS_4 releases, this pulse provides the start signal to the equipment should a code transmission be required during the first minute.

Transistorized Clock Coder Warning System

The operation of relay $CA/3$ during the actual period of transmission of the code has already been described together with the functions of contacts CA_2 and CA_3. Contact CA_1 is used in a transistorized warning system shown in Fig. 5. This unit carries out the following functions:

(1) Ensures that there is no mark of duration longer than 30sec.

(2) Ensures that there is no space of duration longer than 3sec.

(3) Ensures that impulses are received from the time keeping element at intervals of less than 45sec.

Should any of the above conditions not be fulfilled a fault condition is indicated by relay $AL/1$.

The warning system consists essentially of three transistorized timing circuits followed by a bistable flip-flop circuit which operates the fault relay.

The transistor timing unit, Fig. 5, uses two Mullard OC77 transistors which operate from a 50V d.c. supply the method of operation being that the rate of discharge of the timing capacitor C_1 is set by means of transistor X_1 which acts as a constant current discharge path when relay contact SC_1, of the slave code relay, is operated, thereby connecting a negative d.c. bias from the junction of R_6 and R_7 to the base of transistor X_1. This bias together with the resistor R_2 determines the discharge current of C_1. As the discharge continues the voltage on the emitter of X_2 falls until such time as the emitter of X_2 is slightly positive to the base whose potential is set by resistors R_4 and R_5, when this occurs transistor X_2 takes current and an output is taken from the collector through diode MR_1. If, however, contact SC_1 is returned to normal before X_2 takes current capacitor C_1 is recharged to full h.t. voltage through R_1 and the base-collector path of X_1, R_1 being of suitable value to limit the charging current to a value that can safely be handled by X_1. It can therefore be seen that by making the effective discharge time of C_1 30sec, an output will be obtained if the code relay stays in the mark condition for longer than 30sec. Similarly with the second timer unit as CA_1 is operated during coding, if the code relay remains on space for longer than 3sec an output is once again received at MR_2. The third timer unit which is used to check the 30sec impulses from the clock is slightly different as the relay contact T_1 is only operated for a very short period during which time it is necessary to charge C_3 to the full 50V. To perform this function with safety an additional contact of relay T has been used to charge capacitor C_3 through a very low value resistor R_{14}, thereby safeguarding transistor X_5. When relay T operates C_3 is charged through R_{14} and on the release of relay T C_3 is discharged in the way previously described. If therefore contact T_3 remains in its normal position for longer than 45sec diode MR_3 delivers an output.

For indication of a fault condition use has been made of a transistor bistable flip-flop circuit with manual reset facilities, the circuit being shown in Fig. 5. This circuit employs standard low voltage transistors operating from a potential divider R_{27} and R_{28} across the 50V supply, transistor X_8 being normally conducting and transistor X_9 normally cut off. The receipt of a positive going pulse on the base of transistor X_7 from any of the diodes MR_1, MR_2, or MR_3, causes X_7 to cut off, which in turn gives a positive going pulse on diode MR_4. The action of transistor X_7 being to provide an impedance matching stage between the high impedance timer circuit and the low impedance flip-flop circuit. The positive going pulse on MR_4 cuts off transistor X_8 which by the well-known flip-flop action causes transistor X_9 to conduct, thereby operating relay $AL/1$ the fault relay. To reset the flip-flop circuit the base of X_9 is shorted to emitter through capacitor C_4, thereby causing X_9 to cut off and X_8 to conduct, the circuit then being in its normally waiting condition.

In the present unit the transistorized timing units and the flip-flop circuit have been encapsulated to provide small plug-in units $1\frac{1}{4}$in in diameter and 2in high, the timing capacitors being external to the encapsulated units so that by the addition of a suitable capacitor any desired time delay can be obtained up to about 90sec.

Acknowledgments

The authors wish to thank the Corporation of Trinity House for permission to publish this article.

Appendix I

Example of Connexions on 50-Way Plug

Signal required RY repeated three times
25sec d.f. signal
RY repeated once.

The transmission to take place during the second minute of the 6 minute period. Connect pin t to pin CC.

The identification signal RY is obtained by making the following connexions.

Revolution 1 of V_{25} connect j to pins A. C. H for mark
connect k to pins B, F, J. for space

Revolution 2 of V_{25} connect m to pins M. S. U for mark
connect n to pins R, T X for space

Revolution 3 of V_{25} connect p to pin Y for mark
connect r to pin b for space.

To reset V_{25} connect pin h to pin e.

As a space of 5 dot elements is required after each set of identification letters V_{25} is reset from the fourth output after the end of the code letters and allowed to remain on the home cathode '0' for one dot element before restarting to transmit the code letters, thus giving the required total space of 5 dot elements between groups of letters.

In order to transmit the identification signal three times the third cathode of V_{31} is connected to pin u, i.e. pin W to pin u; the first cathode of V_{31} is not brought out as a minimum of two sets of identification letters will be required.

The d.f. signal follows for 25sec with a space of 5 dot elements before the code letters are again transmitted once. As only one transmission is required the 4th cathode of V_{31} is connected to pin AA, i.e. pin x to pin AA and to pin s which closes gate 3 and leaves the equipment waiting for the next initiation pulse.

The Design of Thyratron Stabilized D.C. Supplies

By B. G. Higdon*, B.Sc., and M. E. Bond*

For stabilized h.t. supplies at currents greater than about 0·5A d.c. the grid controlled gas-filled rectifier is, in general, preferable to the series vacuum valve type of stabilizer.

In this article the design of the thyratron type stabilizer and the components incorporated in it is dealt with in some detail. Both a simple stabilizer and a more complex unit incorporating an amplifier in the feedback loop are discussed and performance details are given.

(Voir page 506 pour le résumé en français: Zusammenfassung in deutscher Sprache Seite 510)

THERE are frequent occasions when a stable direct voltage is required for use with electrical and electronic apparatus. When only small currents (<0·5A) are required the series vacuum valve method of stabilizing the output from a rectifier is commonly used. For currents greater than this, however, the series vacuum valve type of stabilized supply becomes increasingly bulky and inefficient. It is for currents greater than 0·5A, therefore, that the thyratron stabilized supply is to be preferred. In fact for large currents, up to approximately 70A, its high efficiency and low cost make it very attractive.

General Principles of Operation

Nearly all stabilized supplies rely for their stability on a continuous comparison between a fraction of the output voltage and a standard voltage.

The changing differences between these two are fed back to a control element which varies the quantity of electricity passing through it. Thus the complete network attempts to maintain the output voltage constant. Sometimes the feedback voltage is amplified before being applied to the control element.

In a thyratron stabilized supply the thyratrons are the control elements and are used as grid controlled rectifiers. The mean current passing through the thyratron is varied by changing the instant during the forward half cycle at which conduction starts. In a thyratron conduction starts when the control grid passes through the critical grid voltage from negative towards positive.

Control of the firing angle can be achieved in several ways but the one most convenient for a stabilized supply is to use the so-called 'vertical' method of control in which an alternating voltage (normally retarded in phase with respect to the anode voltage) is applied to the thyratron grid in series with the controlling direct voltage.

Thus as the control voltage is made more positive or more negative with respect to the cathode of the thyratron the instant in the cycle at which the grid potential passes from positive to negative through the critical grid voltage is advanced or retarded and the thyratron fires either earlier or later in the cycle

This is demonstrated in Fig. 1.

The way in which a grid controlled rectifier operates makes it impossible to use the controlling network to compensate for changes which occur at main or higher frequencies. The output from the rectifier has therefore to be smoothed with a suitable filter. Because of the peak current limitations of thyratrons and the ripple current limitations of capacitors a choke input filter will normally be used. The output capacitor will provide a low output impedance at frequencies above that controlled by the feedback circuits.

Although any type of rectifier circuit may be used with thyratron controlled stabilized supplies it is most convenient to use those in which the controlled valves have a common cathode connexion. All half-wave rectifiers can satisfy this condition and so do full-wave rectifiers in which only half the thyratron rectifiers are controlled. Certain difficulties arise in connexion with the latter and only the two common half-wave systems, two-phase and three-phase, are considered further here.

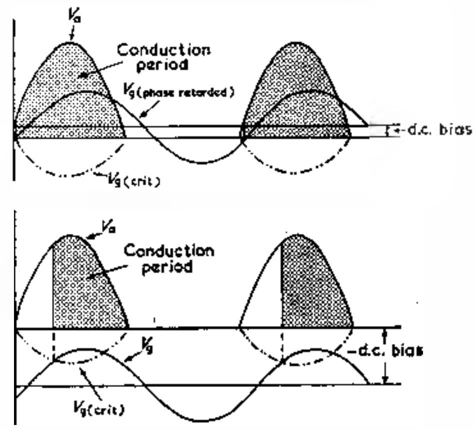


Fig. 1. Vertical shift control

Design Procedure

The design requirements of a stabilized direct voltage supply are:

(1) The required direct output voltage.
(2) The maximum output current.
(3) The minimum output current.
(4) The mains input voltage variations over which the unit will operate.
(5) The maximum ripple voltage tolerable in the output.
(6) The maximum allowable output voltage variation.

The first five of these points determine the rectifier and smoothing circuits required while the sixth determines the feedback and grid control circuits. The rectifier and smoothing circuits will be considered first.

In all the figures and graphs which are shown it is assumed that the main transformer secondary voltage may change by up to + 10 to − 20 per cent of its normal value due to supply mains variations and transformer regulation losses. It is also assumed that the voltage losses due to the resist-

* *Mullard Limited*

ances of the main transformer and choke and due to the arc-drop of the thyratrons will not exceed 35V except in Fig. 4 where it has been assumed that these losses do not exceed 25 per cent of the direct output voltage. If the voltage losses are greater than this the transformer voltage and choke inductance should be increased above those calculated by the following method.

Rectifier and Smoothing Circuit

The components used in the rectifier and smoothing circuits are:

(1) The thyratrons.

(2) The transformer.

(3) The choke.

(4) The smoothing capacitor.

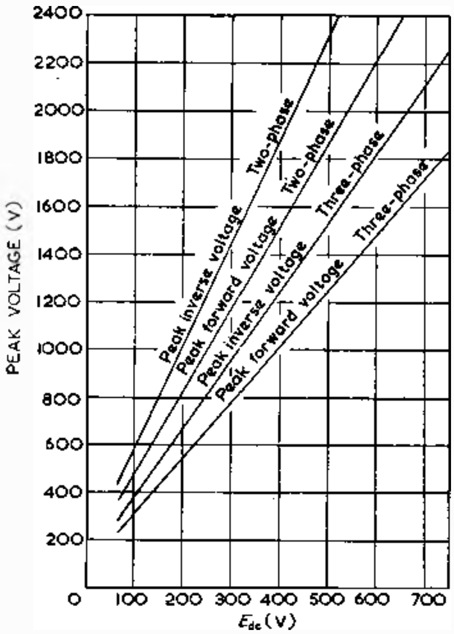


Fig. 2. Peak inverse and forward voltages

The Thyratrons

The thyratron ratings which are of primary importance are the maximum peak forward and inverse voltages and the maximum peak and mean currents The maximum peak forward and inverse voltages which can occur in these circuits for a given direct output voltage are shown in Fig. 2. This takes into account the supply variations and losses which have been mentioned before.

The mean current passing through the thyratrons is the maximum direct output current divided by the number of phases. The maximum peak current passing will, in the ideal case, equal the maximum direct output current, but because the inductance of the choke used is not infinite the peak current through the thyratrons may well exceed the maximum output current by some 25 per cent. The thyratrons should be chosen so that these ratings are not exceeded. In addition the final circuit should be chosen so that none of the other thyratron ratings is exceeded.

The Transformer

The transformer secondary voltage required can be obtained from Fig. 3. It should be noted that this is the voltage from phase to neutral. For three-phase operation this should be clearly stated in the transformer specification as it is more normal in three-phase work to specify the phase to phase voltage.

Because the effect of the firing angle on the secondary current form factor is small when the controlled rectifiers have inductive loading, the secondary current r.m.s. rating can be taken as that of a normal rectifier with the same maximum mean output current.

In this case $I_{rms} = 0{\cdot}707\, I_{dc}$ for a two phase circuit and $0{\cdot}577\, I_{dc}$ for a three-phase circuit. In order to limit the losses the resistance of the transformer should be small. If the transformer is wound so that it has a larger leakage inductance than normal it will be found easier to adjust the circuit so that the commutation factor rating for the thyratron is not exceeded.

The Choke

For a choke input system to operate satisfactorily in a rectifier circuit the inductance of the smoothing choke must

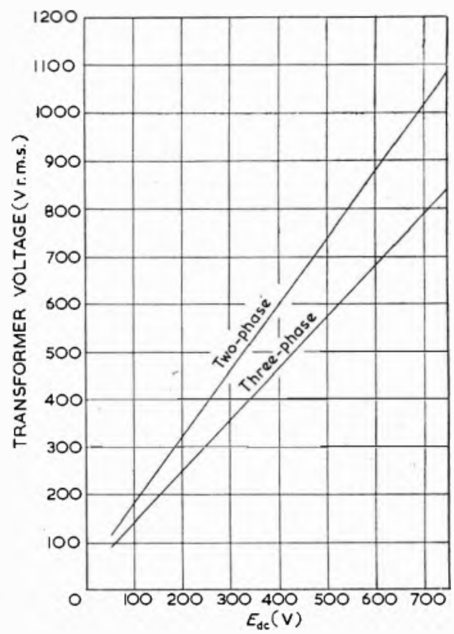


Fig. 3. Transformer voltage

be greater than a critical value. This value is the minimum that will ensure that each thyratron conducts for the full interval between its own ignition point and that of its successor.

The critical value of the smoothing choke is greater when the voltage output from the transformer secondary is at its maximum due to high mains voltage, etc., and when the load current is at its minimum.

If the choke inductance required for the maximum load current is much less than that required for the minimum current it is common practice to use a swinging choke rather than one of constant inductance because cost, bulk, and weight are reduced. When specifying a swinging choke it is essential to give the maximum and minimum currents and the inductances at these currents.

The value of the critical inductance can be obtained from Fig. 4. It is desirable to use a choke of inductance greater than this critical value because of the assumptions and approximations made in its calculation.

An increase of at least 50 per cent is recommended.

The resistance of the choke should be small in order to limit the circuit losses. The choke should also have adequate insulation to withstand the peak voltages from the rectifier and the specification should include the peak voltage across its terminals.

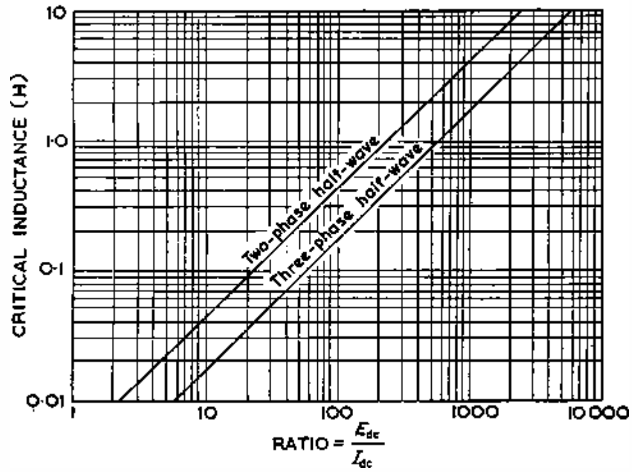


Fig. 4. Critical inductance value

The Smoothing Capacitor

The smoothing capacitor should be large enough to avoid resonance with the choke at the ripple frequency or at 50c/s (mains frequency) for all values of the smoothing inductance. Otherwise its value is solely determined by the ripple voltage that may be tolerated in the output. To a first approximation the value of capacitor required will be:

$$\frac{1{\cdot}33 \times 10^{-6} \times E_{dc}}{V_r L} \text{ Farads for a two-phase half-wave circuit}$$

and

$$\frac{2{\cdot}78 \times 10^{-7} \times E_{dc}}{V_r L} \text{ Farads for a three-phase half-wave circuit}$$

where V_r is the maximum acceptable ripple voltage, E_{dc} is the output voltage and L is the minimum smoothing choke inductance in Henries. An increase of 50 per cent over this value is recommended to cover the effects of firing angle change and the large tolerance usually found in these components. The voltage rating of the capacitor should be adequate for the direct output voltage and it should also have an adequate ripple current rating.

Feedback and Grid Control Circuits

The stability of the power supply with changes of mains voltage and output current will depend on the control and feedback circuits used. An appreciable degree of stabilization of output compared with the changes that occur in unstabilized supplies can be obtained with simple circuits. A much greater degree of stabilization can be obtained with increased complexity, i.e., by the introduction of an amplifier in the feedback network.

Two circuits, which approximately represent the practical limits of simplicity and complexity are described here. If the degree of stabilization required is intermediate between those of the circuits described then this will be achieved by a circuit complexity intermediate between those shown.

The circuits shown have been designed for a particular output voltage and current range and will need modification in order to operate satisfactorily at other output voltages and current ranges.

Simple Control Circuit

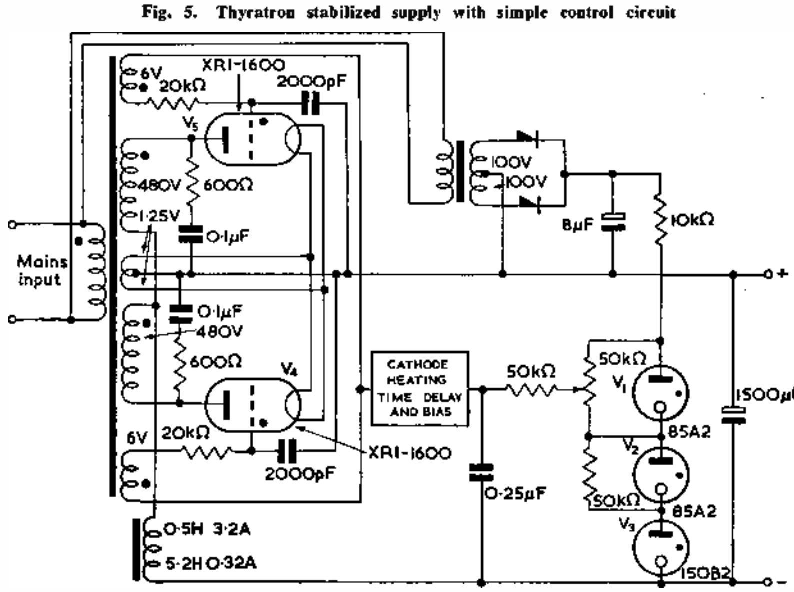
In the circuit shown in Fig. 5 feedback is obtained by the action of the voltage at the thyratron cathodes increasing or decreasing with respect to a standard voltage applied to the thyratron grids. In this case the standard voltage is provided by a potentiometer across a reference tube. The 6V a.c. bias is applied solely to eliminate any differences in the characteristics of the thyratrons. If this alternating voltage is increased then the sensitivity of this circuit will be reduced, as will the stability of the output voltage. If the a.c. bias is reduced below the 6V shown then it will be inadequate to cover the possible range of variations of the thyratron characteristics. The resistor, capacitor network connected between the transformer and thyratron grid has the effect of retarding the phase of the grid waveform and also of attenuating any transient which could be passed to the thyratron grid through the grid-anode capacitance. With this circut the change in output voltage (nominally 310V) which occurred with a load current change from 0·3A to 3·0A was 3V. The change in output voltage when the mains voltage changed from 180V to 240V was 8V. The maximum ripple voltage occurring was 2·0V peak-to-peak.

Amplifier Feedback Circuit

In order to achieve greater stability of the direct output voltage a method of feedback similar to that commonly used in the series vacuum valve type of supply can be used. In this circuit which is shown in Fig. 6 a fraction of the output voltage is compared with the voltage from an 85V reference tube V_2 using a double triode valve V_1 as a difference amplifier. The anode of V_{1b} is connected directly to the grid of the pentode V_3 which further amplifies the voltage difference signal. The voltage drop across the anode load of V_3 is used to provide the variable d.c. bias for the thyratrons in the rectifier circuit.

The grids and cathodes of V_1 are operating at the potential of the anode of the reference tube which is approximately 85V above the negative line. In order to provide an adequate range of voltage variation for the anode of V_{1b} the circuit has been arranged so that the anode potential of V_{1b} is normally 150V. Because, then, the grid of pentode V_3 is at 150V its cathode also needs to be at a similar potential. This is achieved with a minimum loss of gain due to cathode impedance by connecting the cathode of the pentode to the negative line through a 150V stabilizer V_4. This particular circuit will only operate if the

Fig. 5. Thyratron stabilized supply with simple control circuit



output voltage required from the power supply is greater than 250V. For lower voltages some re-arrangement is necessary in order to ensure satisfactory operation.

As in the simple type of control feedback circuit an a.c. bias is needed for the thyratrons in addition to the d.c. bias in order to achieve equal current distribution between them. In this circuit a much greater change in the d.c. bias occurs for a given change in direct output voltage and greater precision in the thyratron firing point can be obtained by using a larger a.c. bias.

As mentioned earlier, feeding back ripple voltage to the grids of the thyratrons cannot serve to reduce the ripple content of the output. In a thyratron stabilized supply the ripple voltage can only be reduced by the smoothing circuit. Therefore unless very large smoothing components are used, the ripple content will be large compared with the tolerated voltage stability of the supply.

a damped oscillation of 9V peak-to-peak in the output voltage. The frequency of this oscillation is about 1c/s and has fallen to negligible amplitude after 5 cycles (5sec). A similar damped oscillation of 12V peak occurs with a sudden mains voltage change from 180V to 240V.

TABLE 1

Performance and Stability

SUPPLY VOLTAGE (V r.m.s.)	LOAD CURRENT (A)	DIRECT VOLTAGE OUTPUT (V)	RIPPLE VOLTAGE (Vpk-pk)	MAX. DRIFT OVER 10Min(V)
240	0·32	309·9	0·5	0·07
240	3·0	309·9	2·0	0·05
220	0·32	310·0	0·5	0·02
220	3·0	310·0	1·8	0·02
180	0·32	309·9	0·5	0·04
180	3·0	309·9	1·5	0·02

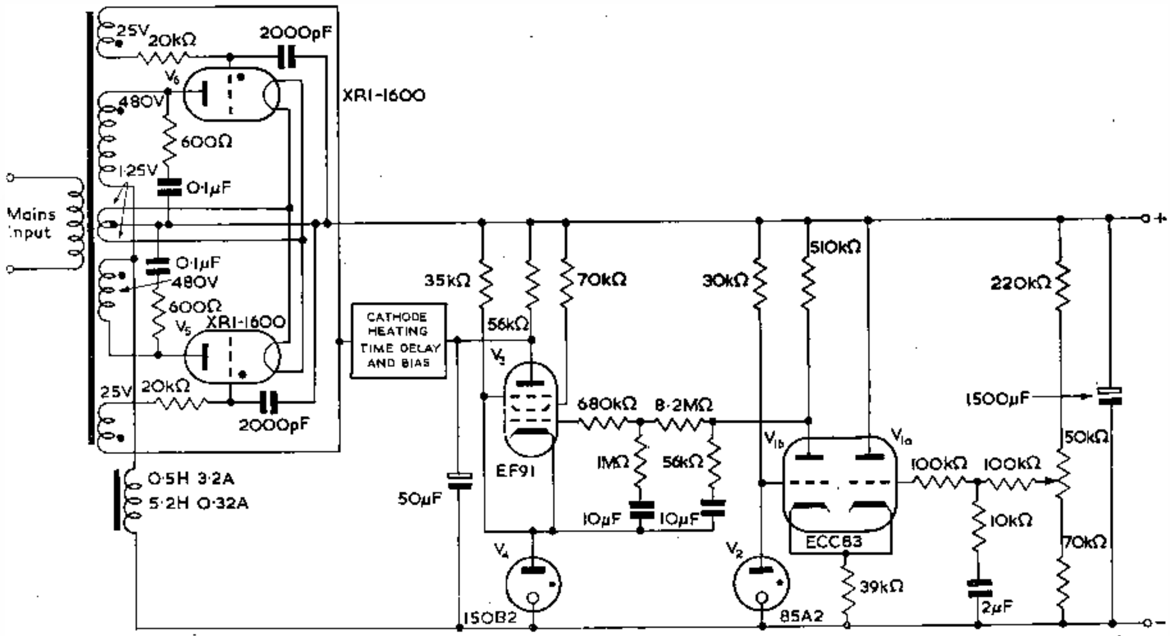


Fig. 6. Thyratron stabilized supply with amplifier feedback

This relatively large ripple content in the feedback circuit would be sufficient to overload the amplifier. This would cause erratic behaviour of the rectifiers. It is therefore necessary to filter out the ripple in the first stage of the feedback circuit to overcome this effect.

Unfortunately, the addition of the filter to remove the ripple in the feedback circuit causes a phase change in the voltage amplified. The choke and smoothing capacitor will also cause a phase change in the feedback loop with the result that there is a possibility of oscillation if the total phase change is 180° at a frequency where the loop gain is greater than 1.

If this occurs it is essential to introduce components to damp the circuit so that the oscillation is not self sustaining. The circuit shown in Fig. 6 has operated satisfactorily under various conditions. The response of the circuit is such that it takes a few seconds to settle down after an extreme change in either the mains voltage or load current.

With this circuit the changes of output voltage which occur with variations of mains input voltage and output current, together with the ripple voltage and the drift over 10 minutes are shown in Table 1.

The transient response of this circuit is such that a sudden change in load current from 0·32 to 3·2A produces

Smaller sudden changes of load current or mains voltage produce correspondingly smaller damped oscillations. It is believed that with more complex design this feature could be improved. In some applications this oscillation is unimportant. It has been found that a feedback network with reduced gain such that the change of direct output voltage is approximately 0·5 per cent for a change from 1·0 to 0·1 of the full output current, makes this effect negligible.

General Precautions

When stabilized power supplies using thyratrons are built it is essential that the final operating conditions of the thyratrons are measured and checked against the appropriate ratings in the published data in order to ensure satisfactory life.

In addition attention to the following points is recommended:

(1) Anode fuses are recommended for thyratrons, as for all rectifier circuits.

(2) The life of the thyratrons will be increased if snubbing circuits are used to reduce the voltage peaks across the thyratrons and the commutation factor.

(3) When each thyratron fires the relatively sudden change of potential across it and the smoothing choke causes a current to flow which charges the stray capacitances in the transformer and choke. In order to avoid voltage pulses in the output due to this it is essential that all earth connexions of both rectifier and feedback circuits are returned to a single point.

(4) It is good practice also to place a capacitor of small effective inductance in parallel with the main smoothing capacitor. This provides a low impedance at high frequencies.

General

This article contains the basic information necessary for the construction of thyratron stabilized direct voltage supplies when certain assumptions are made.

Several power supplies with different voltage and current ratings have been constructed according to these principles and are in regular use. One, built as the circuit shown in Fig. 6, but modified for 250V output, has had its output voltage monitored by a recording voltmeter and has proved to have a long-term stability equal to that of the best series vacuum valve types of supply available.

A type of thyratron stabilized power supply in which a constant output is stabilized against mains variation was previously described by L. Knight[4].

Acknowledgments

The authors are indebted to the directors of Mullard Ltd. for permission to publish this article.

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Medical Electronics

A Report on the Second International Conference in Paris

Following the First International Conference on Medical Electronics held in Paris last year, the Second Conference was held at the UNESCO building in Paris from June 24 to June 27 with over 400 delegates attending from seventeen countries. The conference was under the patronage of Monsieur Chenot, the Minister of Public Health with Dr. A. M. Monnier, Professor of Physiology at the Sorbonne as the President. The two Vice-Presidents of the Conference were Monsieur P. Abadie, Président de la Société des Radioélectriciens and Dr. M. Marchal, Directeur à l'Ecole des Hautes Etudes. Associated with the conference was an exhibition of commercial medical electronics apparatus together with demonstrations of scientific equipment.

At a meeting of the interim committee, formed at the first conference, it was agreed to form an International Federation for Medical Electronics, the members signing this declaration becoming founder members. As laid down in the statutes, the objects of the Federation are to encourage the development and dissemination of knowledge in Medical Electronics throughout the world and various functions are put forward to assist in achieving them. Membership is open to interested persons who satisfy the requirements of the Federation and such members will form the General Assembly. The control and direction of the policy is to be vested in the Council composed of two delegates for each country. The responsibility for the fulfilment of the scientific and business obligations is given to an Executive Committee elected by the General Assembly.

The conference was opened by M. Anjaleu of the Department of Public Health. National representatives then gave reports on the progress of medical electronics in their respective countries, the opening session being concluded with five special lectures by speakers from Germany, France, U.S.A. and Great Britain. In the following sessions, over 150 papers were presented under general headings of Biochemistry, Manometry, Physiology, General Medicine, Electroencephalography, Respiration, Computers, Cardiology, Oculography, Ultrasonics, Radiology and Obstetrics. These were covered in two parallel sessions, both having translation facilities for French and English. In addition two other rooms were available for symposia and discussion groups.

The proceedings of the conference, including all papers are to be published in French and English and will be issued to the members of the Conference.

At the closing session Dr. V. K. Zworykin was elected President of the Association with Professor A. M. Monnier, Dr. M. Marchal and Dr. C. N. Smyth as Vice-Presidents. The other officers elected were Dr. A. Rémond, Secretary and Mr. B. Shackel, Treasurer, with three other members of the Executive Committee, Dr. C. Berkeley, Mr. W. J. Perkins and Dr. R. C. G. Williams. The two delegates put forward by the British delegation to represent them on the Council of the Federation were Professor R. Woolmer and Dr. A. Nightingale. The General Assembly agreed to an annual subscription equivalent to one pound sterling per annum. The proposal to arrange a conference next year aroused considerable discussion, it being finally agreed to consider the possibility of running the third conference in London next year.

The UNESCO building was an ideal setting for the conference but it would appear that the use of slides during lectures is most unusual, the arrangements being quite inadequate. If criticisms are to be made of the organization, there were probably too many papers for the time allowed. This may have been partly due to the large backlog of information available and may not apply to future meetings. Cross-linkage between rooms showing the papers being read might have helped, but less main papers and more time for discussions between persons having common interests would appear to provide a better basis for a conference of the kind. Nevertheless in bringing so many people together the Conference was undoubtedly a success, and the prospects of disseminating information among members of the newly formed Federation are a step towards achievement of the aims of Dr. Zworykin in organizing the first Medical Electronics Conference.

Transistor Blocking Oscillator for Use in Digital Systems

By Agnes A. Kaposi*, Dipl.Ing.

A saturable transformer is used in a blocking oscillator circuit. It is shown that the pulse length is independent of the transistor parameters, of the temperature and of load current. Practical circuits have been made giving pulse lengths from 1·5μsec upwards. The circuit is particularly suitable for driving magnetic core counters.

(Voir page 506 pour le résumé en français: Zusammenfassung in deutscher Sprache Seite 510)

A PULSE generator for use in digital computer work must satisfy some general requirements. The pulse-length and amplitude must not depend on the trigger and must be reasonably independent of the transistor parameters and temperature. It must also be able to supply a range of load currents.

The Blocking Oscillator and Its Disadvantages

The grounded base blocking oscillator (Fig. 1) appears to satisfy a number of the above requirements, but it has

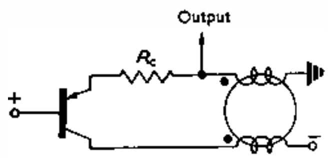


Fig. 1. Basic circuit of a grounded base blocking oscillator

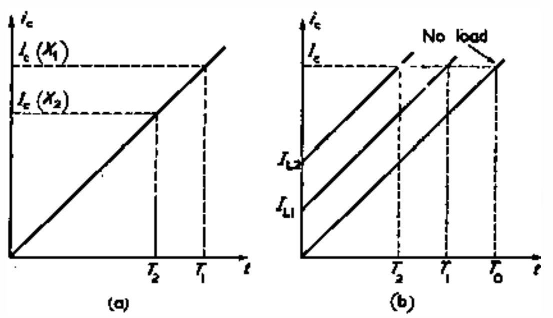


Fig. 2. Variation of pulse length
(a) With final transistor current from one transistor to another
(b) With parallel load

several disadvantages. Firstly the pulse-length itself is to a great extent a function of the final transistor current, that is, of α', r_e and r_{bb}'. Fig. 2 illustrates this effect, and gives an indication of the pulse-length variations which occurred in a circuit similar to that of Fig. 1, when a transistor (designated X_1) with a high value of α' was substituted for another (designated X_2) with a low value of α', both transistors being of the same type. Secondly there is a pulse-length reduction proportional to the load-current taken. (See Fig. 2(b)).

Application of a Saturable Transformer

The two above disadvantages may be overcome by the use of a saturable transformer. A hysteresis loop is shown in Fig. 3. Due to previous operations, the blocking oscillator transformer is, at the beginning of each pulse, in the state of remanent magnetism shown by W. The circuit is arranged so that when the blocking oscillator operates, the transformer is driven to saturation, i.e. in case of the X_1 transistor to point Z and in the case of X_2 to point Y.

As the transistor is bottomed in this period, it is easy to see that the collector current follows the curve shown in

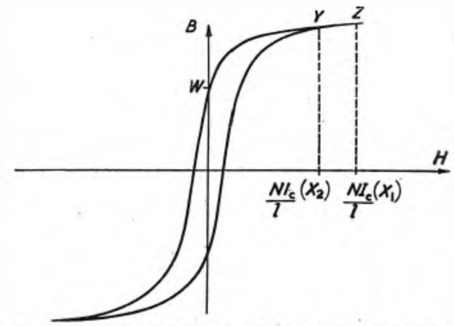


Fig. 3. Hysteresis loop of a transformer driven into saturation

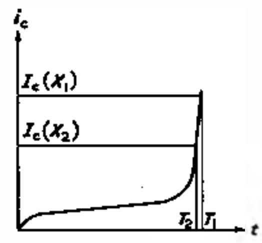


Fig. 4. Collector current waveform of a saturated-transformer blocking oscillator

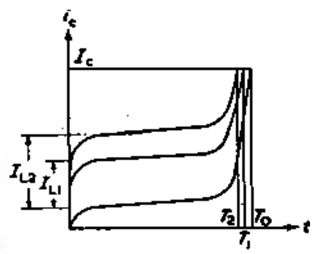


Fig. 5. Effect of load upon collector current waveform in a saturated-transformer blocking oscillator

Fig 4. The steep rise in the current at the end of the pulse leads to very little change in pulse-length with variations of transistor parameters.

Fig. 5 shows the effect of a parallel load on the pulse-length. Since the current rises very rapidly at the end of the pulse, the variation of the pulse-length as the load changes is small. A blocking oscillator with a saturating transformer lends itself admirably to driving magnetic core counters. If a square-loop core is switched by a constant voltage pulse, during switching the core presents a constant impedance; at the end of the switching operation the impedance decreases very rapidly. The corresponding current waveform is shown in Fig. 6.

In order to switch the core by a blocking oscillator, the latter is arranged to have an unloaded pulse length T greater than τ, the switching time of the counter core (Fig. 7). At the end of the period τ the counter core tends

* *Ericsson Telephones Ltd.*

to take an increasing current. This is limited by the collector current I_c available of the blocking oscillator transistor, and the oscillator switches off shortly after τ. Thus the switching time of the counter core controls the length of the pulse generated by the oscillator. If the counter core is in the switched state when the blocking oscillator is triggered, the load presented to the oscillator is nearly a short-circuit and the pulse is terminated immediately.

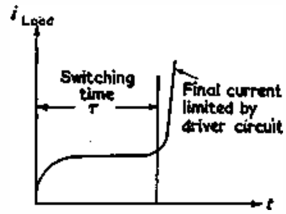


Fig. 6. Waveform of switching current of a square-loop magnetic core in constant voltage operation

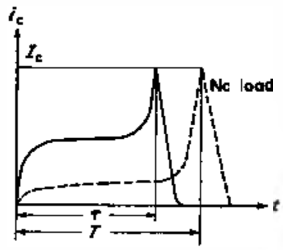


Fig. 7. Collector current waveform of a saturated-transformer blocking oscillator (full line) switching a square-loop magnetic core

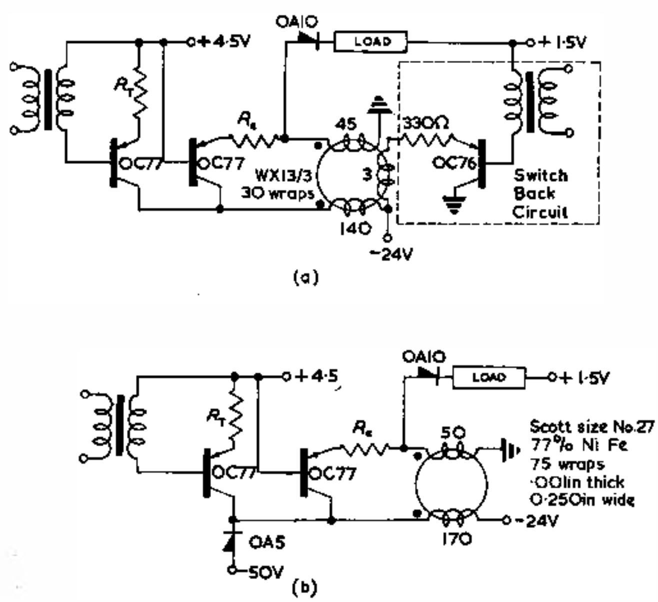


Fig. 8. Experimental circuits

Results with Practical Circuit

Two alternative types of transformer core were used in experiments: a high permeability material (77 per cent Ni-Fe alloy) and a rectangular loop material (WX 13: 80 per cent Ni, 8 per cent Fe, 12 per cent Mo alloy). The trigger was supplied by an emitter-follower which had its collector commoned with the collector of the blocking oscillator. The experimental circuits are shown in Fig. 8. Fig. 8(a) also includes the constant current circuit which resets the square-loop blocking oscillator core in between trigger pulses.

Measurements were made with the two blocking oscillators operating under similar conditions and the results were found to be very similar. Hence only those obtained with the 77 per cent Ni-Fe alloy blocking oscillator will be discussed in detail.

The results will be considered under the following headings:—

(1) The effect of transistor parameters on the pulse length and amplitude.

(2) The effect of resistive load on the pulse length and amplitude.

(3) The application of the blocking oscillator to switching a magnetic core counter.

(1) Effect of Transistor Parameters

The transistors used in this experiment had the following characteristics:

X_1: $\alpha' = 110$ with $V_c = -1{\cdot}5$V and $I_b = 1$mA

X_2: $\alpha' = 27$ with $V_c = -1{\cdot}5$V and $I_b = 1$mA

There was no measurable difference between the collector voltage pulses in the two cases. Damping was not applied, but the overshoot was caught at −50V (Fig. 9(a)).

The base current waveform shows that the transistors were heavily bottomed (Fig. 9(b)).

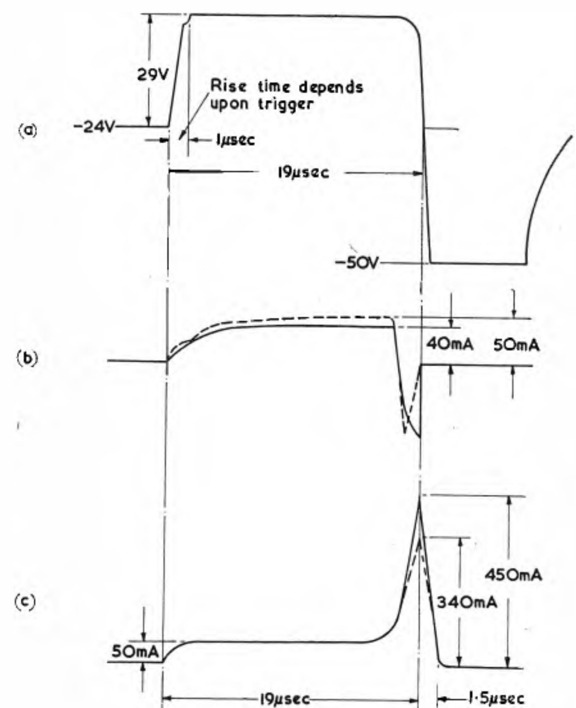


Fig. 9. Waveforms of experimental circuit of Fig. 8(b). Blocking oscillator on no load

——————— X_1 *Transistor*

— — — — — — X_2 *Transistor*

(a) Collector voltage with both X_1 and X_2 transistors

(b) Base current

(c) Collector current

Apart from the value of the final transistor current, the collector currents in the two cases are similar.

The value of the emitter control resistor, R_e, was 33Ω.

(2) Effect of the Load

When a load is connected, the transistor collector impedance is reduced, so that the transistor is not bottomed as heavily as in the unloaded case and the base current is reduced. As the base current must be sufficient to keep the transistor bottomed, this sets a limit to the load current available with a particular emitter control resistor.

Fig. 10 shows a range of results with R_e = 33Ω. This range can be extended if R_e is reduced, for example in Fig. 11 a set of results can be seen with R_e = 22Ω.

For a range of 68 to 197mA load current a drop of 4·5 per cent was measured in the output voltage. An addi-

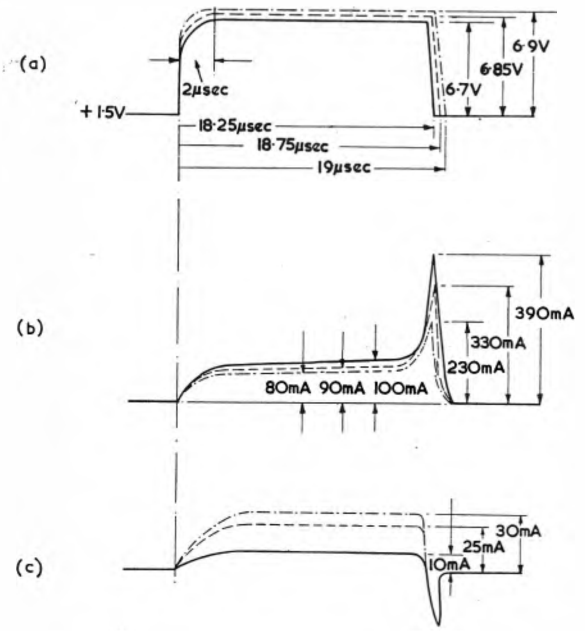


Fig. 10. Waveforms of experimental circuit of Fig. 8(b). Effect of resistive load with $R_e = 33\Omega$

——— ——— *33Ω load, I_{load} = 202mA*
— — — — — — — *62Ω load, I_{load} = 111mA*
— · — · — · — *100Ω load, I_{load} = 69mA*

(a) Output voltage
(b) Collector current
(c) Base current

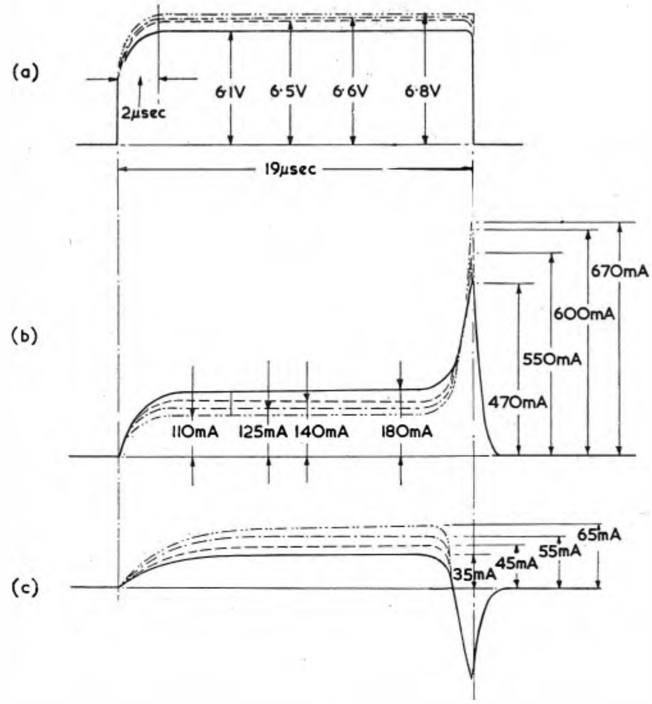


Fig. 11. Waveforms of experimental circuit of Fig. 8(b). Effect of resistive load with $R_e = 18\Omega$

——————— *22Ω load, I_{load} = 277mA*
— — — — — — — *33Ω load, I_{load} = 197mA*
— · — · — · — *62Ω load, I_{load} = 107mA*
— · · — · · — *100Ω load, I_{load} = 68mA*

(a) Output voltage
(b) Collector current
(c) Base current

tional drop of approximately 6 per cent was caused by a further 80mA increase of load current. No more than 4 per cent variation of pulse length was observable over a range of 69 to 202mA (Fig. 10).

(3) Blocking Oscillator Switches Counter

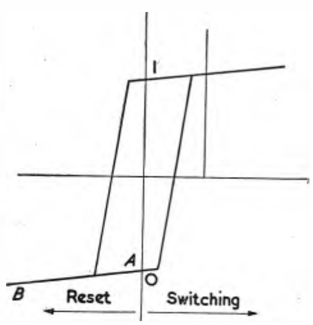
Fig. 13 shows an arrangement in which a counter employing ten square-loop cores is operated by blocking oscillators. Each counter core has three windings. A separate 'switching' blocking oscillator is connected to one winding on each core and the ten oscillators are triggered one at a time, switching the corresponding counter core from its '0' state to '1' (Fig. 12). The second winding of all ten cores are connected in series and joined to the output of a 'reset' blocking oscillator. When this oscillator is triggered, the core in the '1' state is shifted back to '0' while the other nine cores in series are only driven along the *A-B* portion of the hysteresis loop (Fig. 12). Some

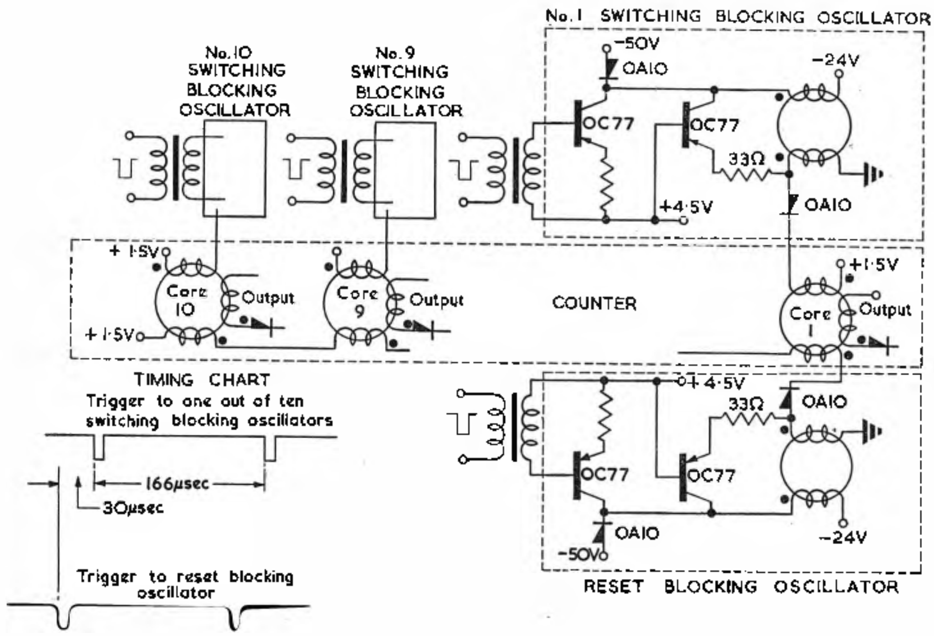
Fig. 12. (below) Idealized hysteresis loop of counter cores of Fig. 13

Fig. 13. (right) Blocking oscillators operating a counter of magnetic cores

Counter cores: Plessey S3 9mm
All windings 20 turns

Blocking oscillator transformer:
Scott 77% Ni-Fe alloy size 27
75 wraps, 0·001in thick
Collector windings: 170 turns
Emitter windings: 50 turns





of the output voltage of the 'reset' blocking oscillator is dropped across these cores and therefore the 'reset' operation takes a longer time than the 'switching' (Fig. 14(a) and (b)). When a counter core is reset, it supplies a load of 5·1V and 60mA in its third winding. The source of this energy is the 'reset' blocking oscillator which, therefore, delivers a larger current to the counter than the 'switching' oscillator (Fig. 14(a) and (b)). The two oscillators are switching off at the end of the switching of the counter core as predicted on Fig. 7.

Temperature Considerations

The Curie points of both 77 per cent Ni-Fe alloy and WX13 are above 150°C so that it can be assumed that the blocking oscillator transformer does not change its

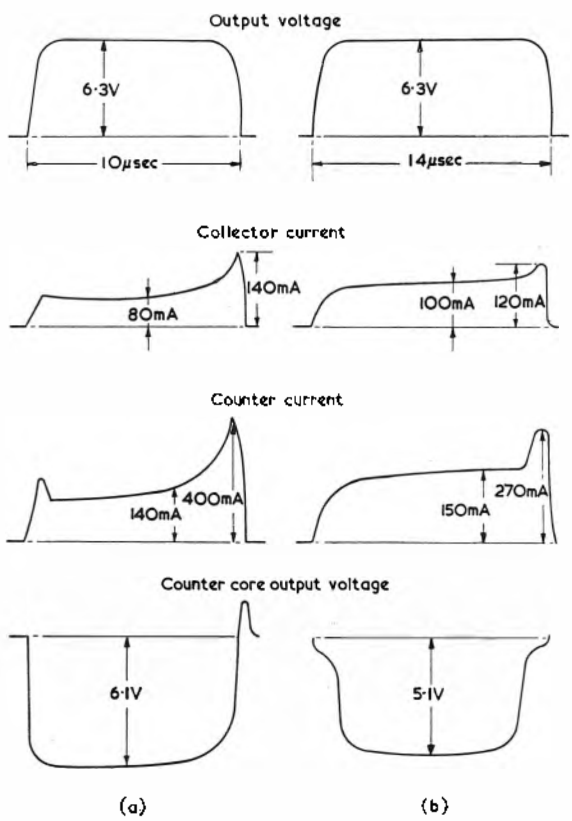


Fig. 14. Circuit waveforms of Fig. 13
(a) Switching
(b) Reset

magnetic properties over the temperature range in which it is normally used.

As the emitter control resistor is of carbon composition, its value will fall as the temperature rises. This will only cause an increase in the value of the final current; otherwise the operation of the blocking oscillator will not be affected.

Similarly, the increase of the transistor junction temperature will only affect the final current, and so the pulse amplitude and length is independent of temperature in the required temperature range.

Limitations of the Circuit

There seems to be no reason why a saturable blocking oscillator could not supply very high load currents indeed, provided the transistor itself was capable of handling large currents. However, the expression for the final collector current,

$$I_a = \alpha_0 \frac{V_2}{n[r_e + R_e + r_{bb'}(1 - \alpha_0)]}$$

where V_2 is the amplitude of the collector voltage pulse
n is the turns ratio of the transformer,

indicates that to obtain good control of the final current, R_e must be much greater than $[r_e + r_{bb'}(1 - \alpha_0)]$. This sets a limit to the load current available with a particular type of transistor.

$R_e = 7\Omega$ was chosen for the circuit of Fig. 8(b), and with this arrangement load currents up to 660mA could be supplied. With load currents as great as this, the voltage drop across the series output diode can be considerable

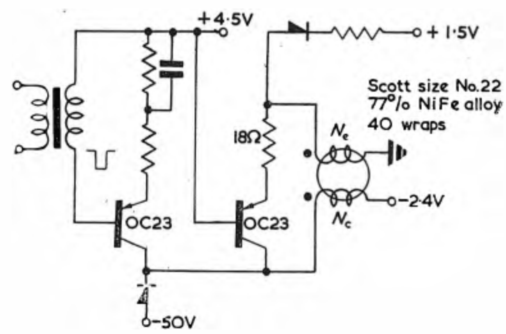


Fig. 15. Experimental circuit
2μsec pulse : N_e = 15T, N_c = 60T
1·5μsec pulse : N_e = 10T, N_c = 40T

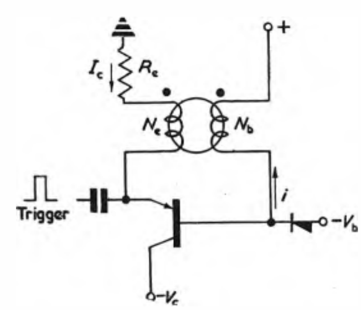


Fig. 16. Basic circuit of a grounded collector blocking oscillator

and the triggering may be difficult, but the blocking oscillator is fully capable of supplying the current without a significant reduction of the pulse width.

The dissipation of the transistor sets an upper limit to the repetition rate and the pulse length. The lower limit of the pulse length is determined by the frequency response of the transistor and the transformer material available. The windings of the transformer should be wound on top of each other to ensure maximum coupling.

The circuit arrangement of Fig. 15 provided pulses of 1·5μsec width and a load current of 150mA was taken without a significant reduction of the pulse-length.

Other Applications

The saturable blocking oscillator can be used as a constant current driver for magnetic core memories. The output is connected to the series combination of a current control resistor R_{out} and the memory core. If R_{out} is much bigger than the switching impedance of the memory core, the current definition is satisfactory.

Grounded Collector Blocking Oscillator

Some thought has been given to the application of the saturable transformer to grounded collector blocking oscillators. The basic circuit is shown in Fig. 16.

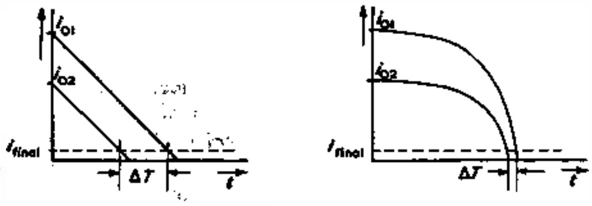


Fig. 17. Variations of pulse length with initial base current
(a) Linear transformer
(b) Transformer driven into saturation

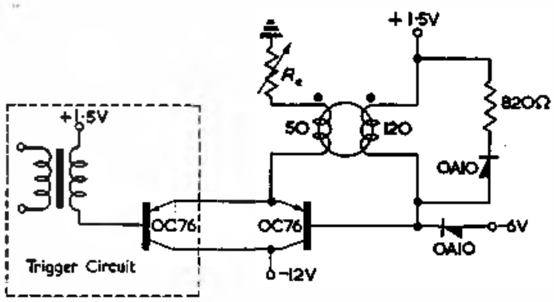


Fig. 18. Experimental circuit
Core: S.T.C. WP3311, Permalloy C Thickness 0·001in

Only a short description of the operation is given here; a more complete discussion can be found in Ericsson Research report No. 40279. During the pulse a constant emitter current flows. The transformer is magnetized by a field.

$$H_m = \frac{N_e\, I_e \, - \, N_b i}{l}$$

Initially $H_m = 0$; therefore, at the beginning of the pulse,

$$i = i_e = \frac{N_b}{N_e I_e}$$

i is decreasing during the pulse, at a rate determined by the transformer, to a value i_{final} which is fixed by the transistor. If the emitter resistor R_e is reduced from a value R_{e2} to R_{e1}, I_e increases, which means that i_o increases from i_{o2} to i_{o1}. In the case of a linear transformer the pulse-length will increase proportionally, but if a saturable transformer is used the variation of pulse-length will be negligible. (See Fig. 17).

The results obtained in a practical circuit (Fig. 18) are shown in Fig. 19 The dotted line represents the case of a linear transformer. The saturable transformer makes the grounded collector blocking oscillator suitable as a constant current driving stage for rectangular-loop magnetic

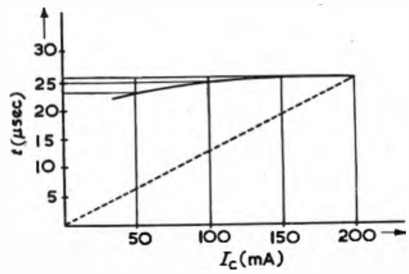


Fig. 19. Variations of pulse length with collector current in grounded collector blocking oscillator

core devices. Such a circuit can be used repetitively in a digital system. The switching core is connected in series with the collector. The transformer gives good control of the pulse length while the emitter resistor can be chosen to suit the current requirement of the core.

Conclusions

The use of saturable transformers enables the grounded base blocking oscillator to supply high load currents and the pulse amplitude and pulse-length is substantially independent of load and variation of transistor parameters. The emitter resistor R_e controls the load current available.

The circuit is particularly suitable for driving magnetic core counters. The blocking oscillator automatically supplies the required amount of current to the counter and switches itself off immediately after the counter core has finished switching. If transistors and transformer material of good frequency response are used, the advantages of the saturable transformer can be utilized in case of short pulses. The upper limit of the pulse length and repetition rate are determined by the transistor dissipation.

In grounded collector blocking oscillators the saturable transformer makes the pulse length independent of the transistor current over a wide current range which makes this circuit suitable as a constant current driving stage for rectangular loop magnetic cores.

Acknowledgments

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Television Engineering

A Report on the British Institution of Radio Engineers Convention

A convention was held at Cambridge by the British Institution of Radio Engineers from 1 to 5 July and had as its title 'Television Engineering in Science, Industry and Broadcasting'. Well over 300 delegates attended including representatives from thirteen countries, and some fifty papers were presented and discussed.

Included in the Convention programme was the Clerk Maxwell Memorial Lecture which was delivered this year by Dr. Vladimir K. Zworykin. In his lecture Dr. Zworykin dealt with applications of electronic engineering in spheres outside those normal to the electronic engineer. As examples he dealt with three subjects: medicine, public safety and public affairs. In the section of his lecture devoted to medicine Dr. Zworykin dealt with a number of applications, one of the most interesting of which was the evolution of 'radio pills'. These consist of an f.m. transmitter and a transducer of one sort or another; the whole being encapsulated and small enough to be swallowed. The transducer can be arranged to provide information on any one parameter, e.g. temperature or pressure, and this can be monitored as the 'pill' goes on its way through the tracts of the digestive system. Under the heading of public safety Dr. Zworykin dealt with the problem of accidents on long motorways and described a system for providing guidance to an automobile and also for providing information on the distance between itself and a vehicle ahead of it, irrespective of whether or not the vehicle ahead was fitted with the control system. The basis of the system was a buried cable carrying r.f. current and divided into sections to provide the fore and aft information. By means of lantern slides Dr. Zworykin showed that a considerable amount of work has already been carried out on this project; the perimeter track of a disused airfield has been equipped with the cable system and a special steering-wheel-less car has been built and fitted with the receiving equipment. For his example of the intrusion of electronic engineering into public affairs, Dr. Zworykin described a system whereby truly democratic decisions could be made by a community in a very short time. The system envisaged requires a central computer and a modification of the telephone system. The questions to be answered by the population are propounded by radio, television or some other means and a time limit is set for answering. Each telephone would be equipped with push-buttons by means of which the answer would be given and, probably, the code number of the voter sent in. This information would be routed to the computer input by short duration pulses or such other means as would not interfere with the normal operation of the telephone system. The computer itself would be programmed to gather in the data, process it and provide the decision of the voters. With a medium sized town this could be carried out in a period of one hour or so. Although the two latter examples tend to suggest electronics run riot it is well to remember that the fantasies of today are often the realities of tomorrow and certainly any suggestion put forward by so eminent a personage cannot be dismissed lightly. Dr. Zworykin closed his address with the stimulating suggestion that with forethought, imagination and initiative the electronic engineer could well become the architect of the progress of humanity.

Of the papers presented those which caused most interest were probably the four presented by the Russian delegation. In one of these, dealing with television receiver production in the U.S.S.R., it was stated that in January 1959 three million sets were in use but that by 1965 it was hoped that production would be at the rate of three and a half million sets per annum and that by that time some eighty million sets would be in use. These receivers are produced in three classes; 21in luxury receivers and 17in and 14in types. It was stated that many of these sets have f.m. units incorporated and that transistors are being introduced. An all-transistor television receiver was produced by the U.S.S.R. in 1957 and it was stated that this type of receiver would be in production by 1961-2. On the transmitting side, stations with powers between 5kW and 200kW have been, or are being built (it was not clear whether these are actual transmitter powers or effective radiated powers) and horizontal polarization is used throughout. Translators (to receive and retransmit in fringe or difficult reception areas) have also been designed and these have receiver sensitivities of the order of $8\mu V/m$ and output powers between 20 and 100W; some of these are designed for unattended operation, on-off control being obtained via a monitor receiver which is left permanently on. Experimental work on colour television is also proceeding and one of the Russian delegation claimed that colour cathode-ray tubes have been produced with a performance comparable if not superior to their America counterparts. However, as far as one could gather the tubes in question are a copy of the shadow mask tube as produced by the Radio Corporation of America. A number of applications of both colour and monochrome television to science and industry were also described; these included applications in railway marshalling yards and the remote control of pouring steel into casting formes. Applications to astronomy were also described and it was claimed that gains of 10 to 20 times over photographic plates have been achieved: it was also stated that a three-dimensional astronomic telescope had been constructed consisting of two telescopes placed some distance apart—the distance was not stated! The overall impression from the Russian contributions was that a great deal of knowledge has been acquired, a considerable amount of development has been carried out and an ambitious programme has been set out for the future. But it seemed rather doubtful whether any novel contribution to knowledge has been made or indeed whether any of the well-known techniques have been improved upon.

The other papers presented covered almost every aspect of television engineering and it is possible to comment on only one or two.

In one A. Ciuciura of the Mullard Radio Valve Co. dealt with the problem of X-radiation from television receivers. In this he showed that, while almost all of the receivers currently offered for sale come well within the safety requirements, it is a matter which should receive some attention from the set designer. If e.h.t. voltages in excess of 16 to 17kV became common practice the problem could become acute and steps, such as screening, would have to be taken. The danger to the service engineer would appear to be greater than that to the viewer since the former frequently operates receivers out of the cabinet and consequently without the considerable screening which the implosion guard provides.

Colour television occupied a fairly prominent place in the Convention and it was apparent that there is a considerable amount of interest in single gun colour television display devices, particularly the 'Lawrence' and 'Apple' tubes. In a paper describing a gating circuit for single gun colour tubes by K. G. Freeman of the Mullard Research Laboratories it was concluded that 'in the absence of a revolutionary new technique the provision of cheap and simple gating circuits for domestic single gun colour tube receivers is only possible at the cost of severe degradation of picture quality'. It also seemed to be generally accepted that the present form of NTSC signal may not be the most suitable for use with this type of tube and it would, therefore, appear desirable to decide on the most likely form of presentation before deciding upon a colour standard.

Other subjects which provoked discussion were the desirability of departing from 'mains lock', and the maintenance or otherwise of the d.c. component.

On the former it seemed generally agreed that it would be desirable to depart from 'mains lock' and that the BBC would like to do so. In any case, this would probably be necessary if a change was made in the line standard or if colour television was introduced.

The arguments for and against the full restoration of the d.c. component produced a lively if somewhat inconclusive debate. It is, however, a problem which cannot be settled on purely theoretical grounds since its economic advantages is one of the main virtues of the mean level a.g.c. circuit.

The papers presented at the Convention together with the discussions will be published in the Journal of the British Institution of Radio Engineers in due course.

Developments in Automatic Machine Tool Control

A transistorized hydraulic continuous machine tool control system and a numerical positional machine tool control system have recently been developed by Ferranti Ltd, Edinburgh, and it is claimed that their simplicity of design, reliability and ease of maintenance will contribute to higher productivity and that their capital cost will place them within reach of the smaller engineering concerns.

The Transistor Hydraulic Continuous-Path Machine Tool Control System

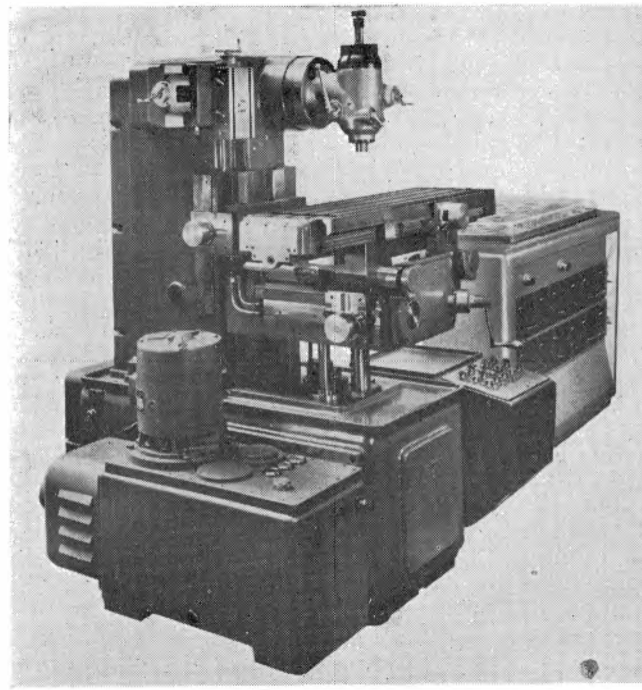
This system operates from the normal electricity supply which eliminates the use of the frequency convertor employed in the earlier Mark III Ferranti equipments. It is now fully transistorized and it incorporates hydraulic servomechanisms and a simplified electro-optical measuring system. Maximum machining time of the new system, compared with the old, has been increased from 40 minutes to 160 minutes.

The electrical servomotors on the original Ferranti system are now replaced by hydraulic piston motors, controlled by electro-hydraulic valves, resulting in an increase in the overall structure stiffness of the machine transmission, and peak output of the servo motors has been increased from 0·7 b.h.p. to 3 b.h.p. Simplified transistor circuits for each axis of the machine tool are contained in the control console in three easily replaceable or interchangeable trays, thereby simplifying maintenance in the field. Modifications have also been made to the electro-optical measuring system which has resulted in a cheaper yet equally accurate system.

The earlier Mark III control equipment utilized ½in magnetic tape running at a speed of 15in/sec. together with diode switching techniques, using sub-miniature thermionic valves as pulse amplifiers and requiring three-phase 400c/s electric servomotors controlled by magnetic amplifiers and supplied by a high-frequency motor-alternator set.

The limitation which necessitated the speed of 15in/sec was one of pulse packing density, which cannot be increased substantially beyond 150 pulses/inch if the tape is to be used under workshop conditions. To achieve an economy in tape speed, it was decided to change the system of recording so that the required resolution could be obtained at lower speeds by the use of a phase displacement carrier system, in which the control information is recorded as discrete displacements in phase between a series of control pulses and a pulse train recorded as a reference channel. The frequency of recording is 208p/s and, after reading, the waveforms are reshaped to give square-waves with a frequency of 104c/s.

If no movement is demanded and the system is at datum, the signals on all four tracks of the tape are identical and in phase; should movement be demanded on any channel, the signals on this channel undergo a progressive positive or negative displacement in phase depending on the direction and speed of motion. These signals can be recorded and resolved satisfactorily at a tape speed of 3·75in/sec so that considerable economy in tape can be made and a 3 600ft tape now has a playing time of 190 minutes.



The Hayes/Ferranti Tapemaster

Measuring System

This type of recorded information demands feedback information in similar form if control simplicity is to be preserved, and this required the development of a new method of using the diffraction grating measuring system. It will be recalled that this system is based on the fact that a pair of crossed diffraction gratings, moving relatively to one another, can be made to generate a multi-phase electrical system in which the total number of cycles is a measure of the distance moved, the frequency a measure of the velocity and the direction of phase rotation a measure of the direction of the movement, the carrier frequency being zero. If now the carrier frequency is made not zero but 104c/s, by arranging that the grating which was previously fixed scans steadily across the field at a speed of 104 lines/second, the system behaves much as before, except that the change in frequency when external movement occurs is above or below the carrier instead of on either side of zero. This frequency may be compared with a reference generated by the addition of a further section of fixed grating also scanned by the continuously moving grating, the light variations from this system being read by a separate photocell. Thus, if the external displacement is zero and at datum, both photocells give outputs of the same frequency and phase; when external movement occurs, the signal output acquires a progressive phase displacement with respect to the reference, these signals being of the same nature as those derived from the command signals recorded on the tape. The servomechanism has to lock these two sets of signals together to ensure that the table displacement obeys the tape commands.

Scanning is accomplished by a rotating disk which has a multi-start spiral recorded on it as a back pattern on clear glass. This disk is driven from the reference signal on the tape by a synchronous motor, the combination of motor poles and spiral starts forming an electro-optical gear to give the correct scanning frequency in lines per second. To eliminate any possible error due to departures from perfection of the scanning system and its drive, the reference signal from the scanning unit is utilized in a manner which compensates for such errors. The opportunity has been taken of using reflection gratings engraved on stainless steel strip which has a coefficient of thermal expansion very close to that of the materials used for machine tool construction. This gives a more compact system which is easily fitted to machines and which eliminates the elaborate mounting procedures necessitated by temperature changes when glass gratings are used on very long machines.

The advantages of this system of measurement are that there are substantially no manufacturing limitations on the grating pitch, so that a flexible choice can be made between accuracy of phase comparison and fundamental grating pitch. It is comparatively easy to phase-synchronize two square-wave signals to an accuracy of 1 or 2 per cent. If, therefore, a measurement accuracy of 1 or 2 ten thousandths of an inch is required, a suitable grating pitch would be 100 lines/inch.

The square-wave signals from the control tape and the grating measuring system are phase-synchronized in a specially designed phase comparator, which has an accuracy of one part in 500 and gives an output error signal to the servo amplifier and thence to the electro-hydraulic servo-valve.

Servomechanisms

Hydraulic servomechanisms have been used on account of their superior performance and reliability. A standard 3 b.h.p. peak motor and gearbox unit has been designed, which is adaptable to fit many types of machine tool. This unit has a very low compliance and zero backlash and can extract all the performance of which the machine tool structure is capable in terms of acceleration and accuracy of position.

Control Console

The use of transistors and semiconductors as circuit elements in the control equipment has enabled the size of the control unit to be greatly reduced and a complete control channel is now contained in a single drawer 7in by 7in by 18in. On the front of each control channel drawer is a red light which is operated should a fault occur in that channel and simultaneously zero speed is demanded from the servo and the machine is shut down. "Reset" and "Test" buttons are provided but if the fault persists the drawer is interchanged with a spare contained in the console, and the work resumed. The defective drawer is returned to Ferranti for servicing. In addition to the three control drawers, a reference drawer is provided containing the equipment to synchronize the scanning motors from the reference track on the magnetic tape. This includes provision to switch over automatically to the internally generated reference signal when the tape is stopped, and allows full manual control to be obtained over the servomechanism and measuring loop, enabling the slides to be moved under the control of push-puttons or vernier control knobs as required.

The power supply for the whole equipment consists of two 12V lead-acid accumulators, fitted with Catylators which reduce to negligible proportions the loss of water in the form of gases, as the hydrogen and oxygen emitted are combined and returned as water.

Automatic Checking

A fourth checking track on the Mark III control equipment was provided on the magnetic tape and used to check for tape drop-outs and the whole operation of the control equipment. The methods used in the new equipment are such that this is no longer required. The system is inherently insensitive to tape drop-outs because of the continuous cyclic nature of the recorded signals. Also, in the event of a deviation from the true position occurring, the absolute position is not lost until the error has exceeded half a grating pitch.

The Numerical Position Control Equipment

This is an entirely new equipment designed for rapid point-to-point positioning of machine tools and allied equipment in Cartesian or polar co-ordinates. Measurement of position is by means of the diffraction grating measuring system fitted to the machine tool slides or tables. The electronic control equipment is designed around semiconductors, and is completely error-checked and fail-safe.

Control Console

The tape reader, manual co-ordinate-setting dials, numerical readout indicators if provided, all electronic equipment and power supplies are assembled in a floor-standing console. The size of this unit is 29in by 24in by 47in high. All controls and the tape reader are grouped for maximum accessibility at the top of the cabinet. The tape reader is provided with tape spooling equipment, the feed and take up spools accommodating 200ft of tape, sufficient for 600 three-dimensional position movements. Tape or dial input can be selected at will. Information space is allocated on the tape for controlling machine functions other than positioning so that a complete operating cycle such as drilling can be controlled.

The electronic equipment is built in the form of plug-in "cards", accessible behind a drop-down panel at the front of the console. Only four types of "card" are used and servicing is rendered extremely simple, an indicator lamp being actuated by the automatic error-check system to indicate the faulty channel. Only semiconductor circuits are used in the equipment, which is based on a newly developed high-speed transistor decimal counter provided with a continuous ternary checking circuit. This counter keeps a continuous record of the position of the machine slide from an arbitrary datum, which can be changed at any time and set to any point on the table. The input position is selected by tape or dials, a direction discriminator decides whether the co-ordinate in the counter is higher or lower than the demanded position and selects the direction of movement to bring the table position into coincidence with the setting.

The decimal counter is continuously checked by means of the ternary logic, so that in the event of a failure the equipment will stop automatically and will not function again until the defective circuit card has been replaced or repaired; thus there is no possibility of an incorrect setting due to failure. In addition a similar check is applied to the tape-reading equipment to obviate misreading the punched tape.

Measuring System

Gratings, either linear or circular, of pitch suitable for the accuracy required are fitted to each axis of the machine, rendering the accuracy independent of the lead-screws or other transmission elements.

Servomechanisms

The servopacks utilize a 50c/s induction motor and two control clutches and gearing to give a two-speed drive. Fast approach speed up to 200in/min can be provided, depending on the resolution required, and a final slow setting speed brings the table to rest at the correct setting two or three seconds after the changeover is made. At this point the slides are automatically clamped.

The Theory of Broadband Coupled Circuits

By W. Dougharty*

This article gives the exact transfer admittance formulae for various types of parallel-tuned coupled circuits. No approximations are made and there is no limitation as to bandwidth. From these formulae are derived expressions for such characteristics as the peak separation, peak-to-valley ratio etc., which are sufficiently accurate for bandwidths approaching the value of the centre frequency. Formulae for the phase response are also given for various types of coupling.

It is shown that with a certain degree of inductive coupling a circuit results which not only has good phase response but also has a response curve showing a remarkable degree of symmetry with frequency deviation.

The treatment throughout is restricted to those cases in which the circuits are synchronous and have equal dissipation factors.

(Voir page 506 pour le résumé en français: Zusammenfassung in deutscher Sprache Seite 510)

THE theory of coupled circuits has been presented by many writers such as Aiken[1], Tellegen[2], Dishal[3] and Rideout[4], but all the above, with the exception of Rideout make the assumption that the ratio of bandwidth to resonant frequency is small. This leads to considerable simplification of the theory, but at the same time introduces inaccuracy (not always serious) in cases where the bandwidth is broad.

Rideout makes no restrictions as to bandwidth but confines his treatment to the case of zero transmission loss (i.e., coupling only slightly removed from critical) and considers mutual inductance coupling only. In this way he also obtains some simplification in the theory but without any sacrifice of accuracy.

When the common assumption is made that the ratio of bandwidth to resonant frequency is small so that certain terms in the theory can be neglected, it will also be found that the resulting formulae, being expressed in terms of coupling coefficient and circuit dissipation factor, are independent of the mode of coupling used.

The exact theory as developed below shows that this is not so. The frequency and phase response of a coupled pair will differ according to the type of coupling used. The differences involved are admittedly negligible for narrow bandwidths but are considerable for cases in which the bandwidth is an appreciable fraction (20 per cent or more) of the centre frequency.

In this article the exact expression for the transfer admittance of a coupled pair is given for parallel-tuned circuits and for various types of coupling.

From this exact expression the various design formulae are obtained. It is in some of these formulae that approximations may with justification be made in the interest of simplicity. The minor errors thereby introduced do not in general invalidate the formulae for either broad or narrow bandwidths.

The phase response for the various circuits is also considered. Here also the formulae are not independent of the type of coupling used.

In order to avoid developing expressions which are so involved as to be useless, the work here is based on two assumptions:

(1) That the circuits are tuned to the same frequency.

(2) That they have equal dissipation factors which are constant over the band considered.

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The Transfer Admittance

The formula for the transfer admittance is of fundamental interest since from it formulae characteristic of the circuit are readily obtained. It is obtained by straightforward manipulation of the network equations. This is sometimes rather involved, but the work is considerably simplified if the general form of the solution is known in advance.

Types of Coupling

Five types of coupling are considered; three of series coupling:

(1) Mutual inductance

(2) Inductive-T or 'bottom' inductance

(3) Capacitive-T or 'bottom' capacitance

and two of parallel coupling:

(1) Inductive-π or 'top' inductance

(2) Capacitive-π or 'top' capacitance.

Definition of Resonant Frequency

When it is stated that two coupled circuits both resonate at a particular frequency it becomes necessary to define the conditions under which the resonant frequency is determined

The definitions adopted here are the normal ones, being those most suited to practical work; that for series coupling the resonant frequency is that which obtains when the other circuit is open-circuited, while for parallel coupling it is that which obtains when the other circuit is short-circuited.

The above difference in the definitions of resonant frequency accounts for the differences between the formulae for series and parallel coupling. If the conditions under which circuit resonance is defined were the same in all cases, there would be two transfer admittance formulae only, namely one for inductive and one for capacitive coupling. That this must be so is clear from the following consideration, that, for every T or π coupling network considered, a corresponding π or T network can be derived which is equivalent at all frequencies. Hence the cause of the differences in the formulae for series and parallel coupling does not lie in the form of the coupling used (i.e., whether T or π); it lies in the way in which the circuits are tuned.

Circuit Constants used in the Formulae

The following constants are used in the formulae to be given below.

L_1, C_1, g_1 are the inductance, capacitance and conductance of the primary circuit

L_2, C_2, g_2 apply to the secondary circuit

M is the mutual inductance between primary and secondary coils

k is the coefficient of coupling

d is the circuit dissipation factor assumed the same for both primary and secondary

δ is a constant which appears instead of $d = d\sqrt{(1-k^2)}$ in some of the formulae

$\nu = \omega/\omega_0$ is the normalized frequency

$\nu_1 = \omega\sqrt{(1-k^2)}/\omega_0 \qquad \nu_2 = \omega/\omega_0\sqrt{(1-k^2)}$

It will be noticed that the terms δ, ν_1 and ν_2 appear only in the formulae for series coupling. This is because, as has already been mentioned, it is a practical convenience with this form of coupling to tune one circuit when the other is open-circuited.

Formulae for the Transfer Admittance

The modulus of the transfer admittance in all cases has the form:

$$|i/e| = 2\sqrt{(g_1g_2)}\left\{1 + [f(\nu)]^2\right\}^{\frac{1}{2}}$$

but the expression for $f(\nu)$ will be seen to differ according to the type of coupling used. If the formulae for inductive and capacitive-π coupling are compared, it will be seen that the terms in the expression for $f(\nu)$ would be the same if, in one case, ν was defined as ω_0/ω (instead of ω/ω_0). A similar relation is seen to exist between the formulae for inductive and capacitive-T coupling.

The formulae for the various types of coupling are listed below.

(1) Mutual Inductance Coupling

The circuit is given in Fig. 1.

$$\omega_0^2 = 1/L_1C_1 = 1/L_2C_2 \qquad d = g_1/\omega_0C_1 = g_2/\omega_0C_2 \qquad \delta = d\sqrt{(1-k^2)}$$

$$k = M/\sqrt{(L_1L_2)} \qquad \nu_1 = \omega\sqrt{(1-k^2)}/\omega_0$$

$$i/e = \frac{-j\sqrt{(g_1g_2)}}{\delta k}\left\{\nu_1^3 - \nu_1(2+\delta^2) + \frac{1-k^2}{\nu_1} - 2j\delta(\nu_1^2-1)\right\}$$

$$|i/e| = 2\sqrt{(g_1g_2)}\left\{1 + \frac{1}{4\delta^2k^2}\left[\nu_1^3 - \nu_1(2-\delta^2) + \frac{1-k^2}{\nu_1}\right]^2\right\}^{\frac{1}{2}}$$

$$\tan\phi = \frac{\nu_1^4 - \nu_1^2(2+\delta^2) + 1 - k^2}{2\delta\nu_1(\nu_1^2-1)}$$

(2) Inductive-T Coupling

The circuit is given in Fig. 2.

$$\omega_0^2 = 1/C_1(L_1+L_3) = 1/C_2(L_2+L_3) \qquad d = g_1/\omega_0C_1 = g_2/\omega_0C_2$$

$$\delta = d\sqrt{(1-k^2)}$$

$$k = \frac{L_3}{\sqrt{[(L_1+L_3)(L_2+L_3)]}} \qquad \nu_1 = \omega\sqrt{(1-k^2)}/\omega_0$$

The formulae are exactly as for mutual inductance coupling.

(3) Capacitive-T Coupling

The circuit is given in Fig. 3.

$$\omega_0^2 = \frac{(C_1+C_3)}{L_1C_1C_3} = \frac{(C_2+C_3)}{L_2C_2C_3} \qquad d = g_1\omega_0L_1 = g_2\omega_0L_2$$

$$\delta = d\sqrt{(1-k^2)}$$

$$k = \sqrt{\left(\frac{C_1C_2}{(C_1+C_3)(C_2+C_3)}\right)} \qquad \nu_2 = \omega/\omega_0(1-k^2)$$

$$i/e = \frac{j\sqrt{(g_1g_2)}}{k}\left\{1/\nu_2^3 - (1/\nu_2)(2+\delta^2) + \nu_2(1-k^2) - 2j\delta(1-1/\nu_2^2)\right\}$$

$$|i/e| = 2\sqrt{(g_1g_2)}\left\{1 + \frac{1}{4\delta^2k^2}\left[1/\nu_2^3 - \frac{(2-\delta^2)}{\nu_2} + \nu_2(1-k^2)\right]^2\right\}^{\frac{1}{2}}$$

$$\tan\phi = \frac{1 - \nu_2^2(2+\delta^2) + \nu_2^4(1-k^2)}{2\delta\nu_2(\nu_2^2-1)}$$

(4) Inductive-π Coupling

The circuit is given in Fig. 4.

$$\omega_0^2 = \frac{L_1+L_3}{C_1L_1L_3} = \frac{L_2+L_3}{C_2L_2L_3} \qquad d = g_1/\omega_0C_1 = g_2/\omega_0C_2$$

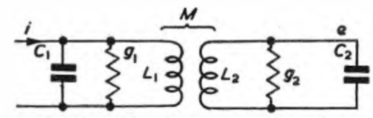

Fig. 1. Mutual inductance coupling

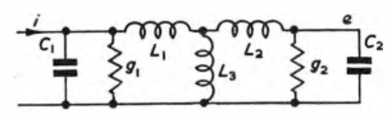

Fig. 2. Inductive-T coupling

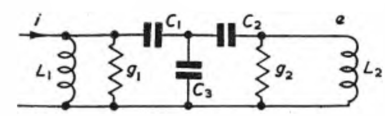

Fig. 3. Capacitive-T coupling

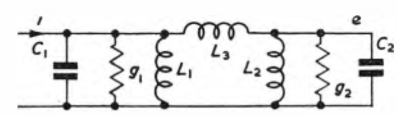

Fig. 4. Inductive-π coupling

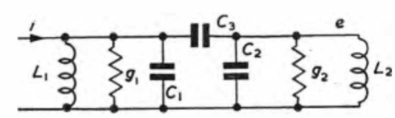

Fig. 5. Capacitive-π coupling

$$k = \sqrt{\left(\frac{L_1L_2}{(L_1+L_3)(L_2+L_3)}\right)} \qquad \nu = \omega/\omega_0$$

$$i/e = \frac{-j\sqrt{(g_1g_2)}}{kd}\left\{\nu^3 - \nu(2+d^2) + (1/\nu)(1-k^2) - 2jd(\nu^2-1)\right\}$$

$$|i/e| = 2\sqrt{(g_1g_2)}\left\{1 + \frac{1}{4d^2k^2}\left[\nu^3 - \nu(2-d^2) + \frac{1-k^2}{\nu}\right]^2\right\}^{\frac{1}{2}}$$

$$\tan\phi = \frac{\nu^4 - \nu^2(2+d^2) + 1 - k^2}{2d\nu(\nu^2-1)}$$

(5) Capacitive-π Coupling

The circuit is given in Fig. 5.

$$\omega_0^2 = \frac{1}{L_1(C_1+C_3)} = \frac{1}{L_2(C_2+C_3)} \qquad d = \frac{g_1}{\omega_0(C_1+C_3)} = \frac{g_2}{\omega_0(C_2+C_3)}$$

$$k = \frac{C_3}{\sqrt{[(C_1+C_3)(C_2+C_3)]}} \qquad \nu = \omega/\omega_0$$

$$i/e = \frac{j\sqrt{(g_1g_2)}}{kd}\left\{1/\nu^3 - (1/\nu)(2+d^2) + \nu(1-k^2) - 2jd(1-1/\nu^2)\right\}$$

$$|i/e| = 2\sqrt{(g_1g_2)}\left\{1 + \frac{1}{4d^2k^2}\left[1/\nu^3 - \frac{(2-d^2)}{\nu} + \nu(1-k^2)\right]^2\right\}^{\frac{1}{2}}$$

$$\tan\phi = \frac{1 - \nu^2(2+d^2) + \nu^4(1-k^2)}{2\nu d(\nu^2-1)}$$

TABLE 1

Characteristic Formulae for Parallel Tuned Coupled Circuits

CHARACTERISTIC	(1) MUTUAL INDUCTANCE AND (2) INDUCTIVE—T	(3) CAPACITIVE—T	(3) INDUCTIVE—π	(5) CAPACITIVE—π
(1) The frequencies of the peaks .. $f_p=$	$f_o\left[\frac{1-\delta^2/_2\pm\sqrt{(k^2-\delta^2+\delta^4/_4)}}{1-k^2}\right]^{\frac{1}{2}}$	$f_o[1-\delta^2/_2\pm\sqrt{(k^2-\delta^2+\delta^4/_4)}]^{\frac{1}{2}}$	$f_o[1-d^2/_2\pm\sqrt{(k^2-d^2+d^4/_4)}]$	$f_o\left[\frac{1-d^2/_2\pm\sqrt{(k^2-d^2+d^4/_4)}}{1-k^2}\right]^{\frac{1}{2}}$
(2) Geometric mean of peak frequencies $f_a=$	$\frac{f_o}{(1-k^2)^{\frac{1}{4}}}$	$f_o(1-k^2)^{\frac{1}{4}}$	$f_o(1-k^2)^{\frac{1}{4}}$	$\frac{f_o}{(1-k^2)^{\frac{1}{4}}}$
(3) The peak separation	$\frac{f_a\sqrt{(k^2-\delta^2)}}{(1-k^2)^{\frac{1}{4}}}$		$\frac{f_a\sqrt{(k^2-d^2)}}{(1-k^2)^{\frac{1}{4}}}$	
(4) Condition for coincident peaks (flat response)	$k=\delta$		$k=d$	
(5) The frequency of the valley	$f_a(1+\frac{k^2-\delta^2)}{8}$	$f_a(1-\frac{k^2-\delta^2)}{8}$	$f_a(1+\frac{k^2-d^2)}{8}$	$f_a(1-\frac{k^2-d^2)}{8}$
(6) The separation of points of equal height to the valley	$\frac{f_a\sqrt{[2(k^2-\delta^2)]}}{(1-k^2)^{\frac{1}{4}}}$		$\frac{f_a\sqrt{[2(k^2-d^2)]}}{(1-k^2)^{\frac{1}{4}}}$	
(7) The peak to valley ratio .. $e_p/e_v=$	$\frac{\delta^2+k^2}{2\delta k}$		$\frac{d^2+k^2}{2dk}$	
(8) Bandwidth at various points—				
(a) $1/n$ of peak amplitude $f_1\sim f_2=$	$\frac{f_a}{1-k^2)^{\frac{1}{4}}}\left[k^2+2\delta k\sqrt{(n^2-1)}-\delta^2\right]^{\frac{1}{2}}$		$\frac{f_a}{\sqrt{(1-k^2)^{\frac{1}{4}}}}\left[k^2+2dk\sqrt{(n^2-1)}-d^2\right]^{\frac{1}{2}}$	
(b) 1dB below peaks .. ,, —	$\frac{f_a}{(1-k^2)^{\frac{1}{4}}}\left[k^2+\delta k-\delta^2\right]^{\frac{1}{2}}$		$\frac{f_a}{\sqrt{(1-k^2)^{\frac{1}{4}}}}\left[k^2+dk-d^2\right]^{\frac{1}{2}}$	
(c) 3dB below peaks .. ,, =	$\frac{f_a}{(1-k^2)^{\frac{1}{4}}}\left[k^2+2\delta k-\delta^2\right]$		$\frac{f_a}{\sqrt{(1-k^2)^{\frac{1}{4}}}}\left[k^2+2dk-d^2\right]^{\frac{1}{2}}$	

NOTE.—f_o = *Tuning frequency of both circuits under conditions described in the text*
k = *Coefficient of coupling*
d = *Dissipation factor of both circuits*
$\delta = d\sqrt{(1-k^2)}$

TABLE 2

Characteristic Formulae for the Critically Coupled Condition

CHARACTERISTIC	(1) MUTUAL INDUCTANCE AND (2) INDUCTIVE—T	(3) CAPACITIVE—T	(4) INDUCTIVE—π	(5) CAPACITIVE—π
(1) The centre frequency .. .. $f_a =$	$\frac{f_0}{\sqrt{(1-k^2)^{\frac{1}{2}}}}$	$f_0\sqrt{(1-k^2)^{\frac{1}{2}}}$	$f_0\sqrt{(1-k^2)^{\frac{1}{2}}}$	$\frac{f_0}{\sqrt{(1-k^2)^{\frac{1}{2}}}}$
(2) Bandwidth at various points—				
(a) 1/n of response at centre $f_1 \sim f_2 =$	$\frac{f_a k}{(1-k^2)^{\frac{1}{2}}}\left[2\sqrt{(n^2-1)}\right]^{\frac{1}{2}}$ For all types of coupling			
(b) 1dB below response at centre ,, $=$	$\frac{f_a k}{(1-k^2)^{\frac{1}{2}}}$,, ,,			
(c) 3bB below response at centre ,, $=$	$\frac{f_a\sqrt{^2}k}{(1-k^2)^{\frac{1}{2}}}$,, ,,			

TABLE 3

Characteristic Formulae for the Undercoupled Condition

CHARACTERISTIC	(1) MUTUAL INDUCTANCE AND (2) INDUCTIVE—T	(3) CAPACITIVE—T	(4) INDUCTIVE—π	(5) CAPACITIVE—π
(1) The peak frequency	$f_0\left(1+\frac{3k^2-\delta}{8}\right)$	$f_0\left(1-\frac{3k^2-\delta^2}{8}\right)$	$f_0\left(1-\frac{k^2+\delta^2}{8}\right)$	$f_0\left(1+\frac{k^2+\delta^2}{8}\right)$
(2) Bandwidth at various points—				
(a) $1/n$ of response at peak $f_1 \sim f_2 =$	$f_0\left[\frac{k^2-\delta^2+\sqrt{[n^2(\delta^2+k^2)^2-4\delta^2k^2]}}{1-k^2}\right]^{\frac{1}{2}}$	$f_0[k^2-\delta^2+\sqrt{[n^2(\delta^2+k^2)^2-4\delta^2k^2]}]^{\frac{1}{2}}$	$f_0[k^2-d^2+\sqrt{[n^2(d^2+k^2)^2-4d^2k^2]}]^{\frac{1}{2}}$	$f_0\left[\frac{k^2-d^2+\sqrt{[n^2(d^2+k^2)^2-4d^2k^2]}}{1-k^2}\right]^{\frac{1}{2}}$
(b) 3dB below response at peak ,, $=$	$f_0\left[\frac{k^2-\delta^2+\sqrt{[2(\delta^4+k^4)]}}{1-k^2}\right]^{\frac{1}{2}}$	$f_0[k^2-\delta^2+\sqrt{[2(\delta^4+k^4)]}]^{\frac{1}{2}}$	$f_0[k^2-d^2+\sqrt{[2(d^4+k^4)]}]^{\frac{1}{2}}$	$f_0\left[\frac{k^2-d^2+\sqrt{[2(d^4+k^4)]}}{}\right]^{\frac{1}{2}}$
(3) Ratio of response at peak to that given by critical coupling	$\frac{2\delta k}{\delta^2+k^2}$		$\frac{2dk}{d^2+k^2}$	

The Characteristic Formulae

The characteristics of a coupled circuit which it is useful to know may be listed as follows:

(1) The frequencies of the peaks.

(2) Their geometric mean frequency.

(3) The peak separation.

(4) The condition for coincident peaks (flat response).

(5) The frequency of the valley.

(6) The separation of points of equal height to the valley.

(7) The peak-to-valley ratio.

(8) The overall bandwidth at various points on the curve.

The expressions for the above quantities are listed in Table 1 for all the circuits mentioned above.

It is also useful to have available formulae for the special case of critically coupled circuits. These are readily obtained from the above by applying the condition for coincident peaks. The resulting expressions are given in Table 2 for the following characteristics:

(1) The centre frequency.

(2) The overall bandwidth at various points on the curve.

In the undercoupled condition the peaks disappear and what was previously the valley becomes the frequency of maximum response. The fall-off in response each side of resonance is now to be measured relative to the new maximum and not to the previous peaks since these are non-existent.

This complicates somewhat the formulae for bandwidth and it seems that no simpler expressions exist other than those given. The formulae suited to the undercoupled condition are given in Table 3.

A few words may be in order as to the derivation of these formulae.

The formulae for the transfer admittance is of the form

$$|i/e| = 2\sqrt{(g_1 g_2)}\left\{1 + [f(\nu)]^2\right\}^{\frac{1}{2}}$$

where g_1 and g_2 are the conductances of the two circuits and ν is the normalized frequency.

This expression will have minimum and maximum values at the frequencies of the peaks and the valley respectively. Accordingly by differentiating the expression for $|i/e|$ with respect to ν and equating to zero

$$2\sqrt{(g_1 g_2)}\left\{1 + [f(\nu)]^2\right\}^{-\frac{1}{2}}[f(\nu)]\,[f'(\nu)] = 0$$

Dividing through by the terms which are not zero

$$f(\nu)\ .\ f'(\nu) = 0$$

By inspection the equation $f(\nu) = 0$ contains the frequencies of the peaks (minimum transfer admittance) and $f'(\nu) = 0$ must, therefore, contain the frequency of the valley.

From the formula for the peak frequencies the expressions for the peak characteristics (formulae (1) to (4)) can be obtained exactly. In practice considerable simplification can be had, at the price of errors of a few per cent only, by neglecting terms which involve the fourth or higher powers of d and k where this is mathematically justified. This follows Rideout[4] whose 'flat-topped' transformer ($k = 0{\cdot}5$) has actually a slight dip in the centre of 0·005dB due to his use of simplified formulae.

The valley frequency derived from the equation $f'(\nu) = 0$ cannot be exactly expressed. As a result, formulae (5) to (7) are close approximations with errors of not more than a few per cent for values of d and k less than about 0·5.

The equation from which (8) is derived is insoluble unless approximations are made. Here again Rideout's approximate expression for the bandwidth of his transformer at 1dB down is seen to be a particular case of the general formula given here.

All the above approximations, although theoretically objectionable, seem in practice to be entirely justified. They do not, as do the usual 'narrow band' formulae, obscure important characteristics of broad band coupled circuits and the errors introduced are almost always less than those due to inexact knowledge in practice of the basic circuit constants of capacitance and conductance.

While none of the formulae given is exceptionally complicated, those for the critically coupled condition are delightfully simple, especially in cases where k is small enough to make $(1 - k^2)^{\frac{1}{2}}$ substantially unity.

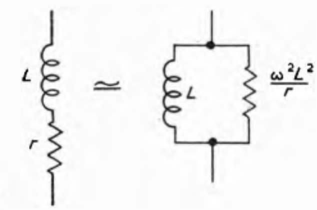

Fig. 6. Equivalent parallel circuit for series combination of r and L when $\frac{r}{\omega L}$ is small

Fig. 7. Mutual inductance coupling with series coil loss

Calculations for the overcoupled condition are scarcely less simple. For the undercoupled condition the formulae are not so straightforward, but unfortunately this condition, with the exception of the phase-linear circuit treated later on, is not usually of much interest.

The Effect of Series Loss in the Coils

It is not necessary to treat the effect of series coil loss very exactly for two reasons, firstly because in a parallel tuned circuit the value of series loss is designed to be small, and secondly its value is not normally very accurately known. It will be sufficient for most purposes to observe qualitatively the effect of series loss in the various cases of coupling already considered.

If an inductor having small series resistance is represented in its equivalent parallel form, one obtains substantially the same value of inductance shunted by a resistance whose value varies as the frequency squared (Fig. 6). If this substitution is made for all types of coupling except mutual-inductance and inductive-T, it will be seen that with increase of frequency the total parallel damping is reduced and so the response curve will tend to rise with frequency.

In the case of mutual-inductance or inductive-T coupling (which are, of course, exactly equivalent to one another) it can be seen from an inspection of Fig. 7 that the presence of series loss will cause the output to fall with frequency.

The above conclusions can readily be corroborated by an exact analysis of the network.

Since, in an r.f. amplifier, the output often falls at the high frequency end due to increased input admittance of the valves, it is usually unwise to use mutual-inductance or inductive-T coupling between stages since the above effect will be accentuated by their use; it is better to use one of the other forms of coupling so that the effect of series loss and valve input admittance will oppose one another. It is also always possible to introduce extra resistance if necessary in series with the coils to neutralize the effect of a large value of valve admittance.

The Phase Response

The phase response of a coupled circuit is conveniently expressed as a power function of frequency deviation, the deviation being measured from the resonant frequency of the circuits or from some closely adjacent frequency.

The series takes the following form:

$$\phi = \pm(\pi/2) + C_1(p/k) + C_2(p/k)^2 + C_3(p/k)^3 + C_4(p/k)^4 + C_5(p/k)^5 + \dots$$

where ϕ is the angle by which the ouput voltage lags the input current and p is the deviation Δf divided by a frequency of phase symmetry f_θ which is approximately that of circuit resonance.

The positive and negative signs in front of $\pi/2$ apply to cases of inductive and capacitive coupling respectively.

The values of the coefficients C_1, C_2 etc., for the various types of coupling are given below.

(1) Mutual Inductance Coupling

$$C_1 = \frac{4n}{1+n^2} \quad C_2 = -\frac{2kn(1-3n^2)}{(1+n^2)^2} \quad C_3 = \frac{-2n}{(1+n^2)^3}\left[(8n^2/3)(1-3n^2) - k^2(1-6n^2+n^4)\right]$$

$$C_4 = \frac{2kn}{(1+n^2)^4}\left[(1-10n^2+5n^4)(4n^2-k^2)\right]$$

$$C_5 = \frac{n}{(1+n^2)^5}\left[(64n^4/5)(1-10n^2+5n^4) - k^2n^2(3-51n^2+65n^4-9n^6) + k^4(1-15n^2+15n^4-n^6)\right]$$

where

$n = k/\delta$, $p = \Delta f/f_\theta$ and $f_\theta = f_0/\sqrt{(1-k^2)} = f_a/(1-k^2)^{\frac{1}{2}}$

The deviation Δf is measured from the frequency f_θ.

(2) Inductive-T Coupling

The coefficients are exactly as for mutual-inductance coupling.

(3) Capacitive-T Coupling

$$C_1 = \frac{4n}{1+n^2} \quad C_2 = \frac{-2kn(1+5n^2)}{(1+n^2)^2} \quad C_3 = \frac{-2n}{(1+n^2)^3}\left[(8n^2/3)(1-3n^2) - k^2(1+2n^2+9n^4)\right]$$

$$C_4 = \frac{2n}{(1+n^2)^4}\left[4kn^2(1+6n^2-11n^4) - k^3(1+4n^2+n^4+14n^6)\right]$$

$$C_5 = \frac{2n}{(1+n^2)^5}\left[(32n^4/5)(1-10n^2+5n^4) - 2k^2n^2(3+13n^2+65n^4-73n^6) + k^4(1+5n^2+11n-5n^6+20n^8)\right]$$

where

$n = k/\delta$, $p = \Delta f/f_\theta$ and $f_\theta = f_0\sqrt{(1-k^2)} = f_a(1-k^2)^{\frac{1}{2}}$

The deviation Δf is measured from the frequency f_θ.

(4) Inductive-π Coupling

The coefficients are exactly as for case (1), except that n has the value k/d and the frequency of phase symmetry $f_\theta = f_0 = f_a(1-k^2)^{\frac{1}{2}}$.

(5) Capacitive-π Coupling

The coefficients are exactly as for case (3), except that n has the value k/d and the frequency of phase symmetry $f_\theta = f_0 = f_a(1-k^2)^{\frac{1}{2}}$.

The power series for ϕ will not be of much value unless it can be shown to be convergent. Unfortunately it is not at all easy to state the general conditions under which it will be so. By inspection, however, it is almost certainly convergent for values of n and (p/k) less than unity.

Bearing the above in mind, it will be seen that with inductive coupling the phase response becomes more nearly linear as the coupling is reduced below critical ($n < 1$), until the maximally linear condition is reached when $n = 1/\sqrt{3}$. This fact is already known[5]. It may not be generally known, however, that with capacitive coupling, since the coefficient C_2 can never vanish, the phase response can never be as good as and in general will always be inferior to that given by inductive coupling.

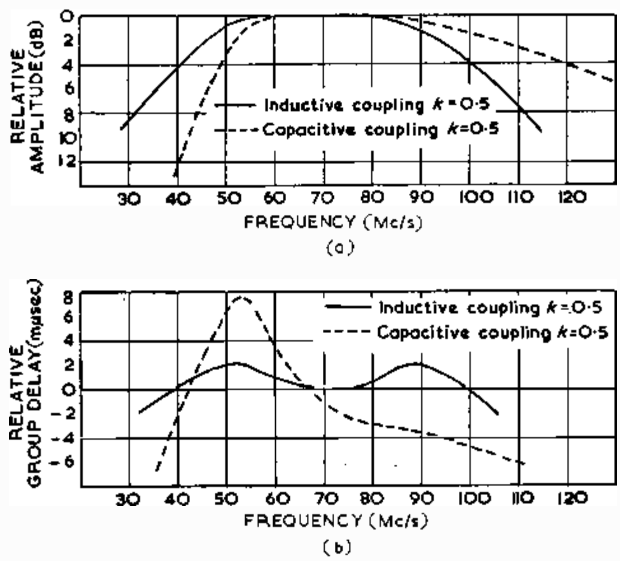

Fig. 8. Curves of frequency response and group delay for inductive and capacitive coupling

Group Delay

While the expression for phase-angle is informative in that it clearly shows the condition for maximally linear phase, a more generally useful concept is that of group delay ($d\phi/d\omega$). The expression for the group delay of a coupled circuit is readily shown to be

$$(2/\omega_0 d)\left[\frac{1 + (d/2k)\,f(\nu)(\nu + (1/\nu))}{1 + [f(\nu)]^2}\right]$$

for inductive-π coupling,

where

$$f(\nu) = (1/2dk)\left[\nu^3 - \nu(2-d^2) + \frac{1-k^2}{\nu}\right]$$

and

$$(2/\omega_0 d\nu^2)\left[\frac{1 + (d/2k)\,f(\nu)(\nu + (1/\nu))}{1 + [f(\nu)]^2}\right]$$

for capacitive-π coupling,

where

$$f(\nu) = (1/2dk)\left[1/\nu^3 - \frac{(2-d^2)}{\nu} + \nu(1-k^2)\right]$$

For T coupling the expressions are similar except that ν, d is replaced by ν_1, δ or ν_2, δ as in the expressions for the transfer admittance.

The difference in the delay contributed by inductive and capacitive coupling becomes very marked when the coefficient of coupling becomes at all large. Consider for instance Fig. 8 in which is shown the amplitude response and group delay for two pairs of critically coupled circuits

TABLE 4
Characteristics of the Phase-Linear Circuit

CHARACTERISTIC	MUTUAL INDUCTANCE AND INDUCTIVE — T	INDUCTIVE — π
(1) The peak frequency	f_0	$f_0\sqrt{(1-k^2)}$
(2) Bandwidth at various points—		
(a) $1/n$ of response at peak .. .. .. $f_1 \sim f_2 =$	$f_0k\left[\dfrac{2(\sqrt{(4n^2-3)}-1)}{1-k^2}\right]^{\frac{1}{2}}$	$f_0k[2(\sqrt{(4n^2-3)}-1)]^{\frac{1}{2}}$
(b) 1dB below response at peak .. .. .. $n =$	$\dfrac{0{\cdot}91\, f_0k}{\sqrt{(1-k^2)}}$	$0{\cdot}91\, f_0k$
(c) 3dB below response at peak .. .. .. $n =$	$\dfrac{1{\cdot}57\, f_0k}{\sqrt{(1-k^2)}}$	$1{\cdot}57\, f_0k$
(3) Ratio of response at peak to that given by critical coupling with same circuit damping	·866	

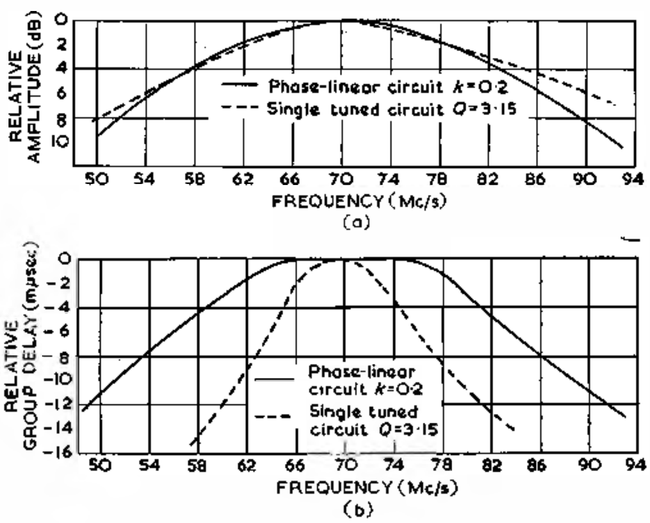

Fig. 9. Curves of frequency response and group delay for the phase-linear and single-tuned circuits

centred on 70Mc/s with $k = 0{\cdot}5$ one inductively and the other capacitively coupled.

It will be seen that capacitive coupling owing to its poor phase response is normally quite unsuitable for broad-band coupling circuits. Since it has already been pointed out that mutual-inductance and inductive-T coupling are inferior to inductive-π coupling as regards series loss in the coils, it appears that the most generally satisfactory form of coupling for use in broad-band amplifiers will be inductive-π.

The Phase-Linear Circuit

If the case is considered of an inductively coupled pair with maximally linear phase response ($n = 1/\sqrt{3}$), one arrives at a circuit of considerable interest.

This circuit, which for want of a better name may be called the phase-linear circuit, presents the condition for maximally linear phase response, but also exhibits a remarkable degree of symmetry in the shape of the response curve when plotted to a base of frequency. This makes the circuit potentially useful for such applications as wide-band discriminators, and filters where arithmetic symmetry of phase and amplitude with frequency are desired.

The derivation of the transfer admittance formula for this circuit is given in the Appendix and a few characteristic formulae are listed in Table 4. The curve of amplitude response and group delay are shown in Fig. 9 together with those of a normal tuned circuit for comparison.

Coupled Circuits in Practice

The problem of realizing a given design practically is not a difficult one. It resolves itself into making a sufficiently accurate measurement of the primary and secondary capacitances and a fair estimate of the circuit losses. It is also necessary to ensure firstly, that any stray coupling is either negligible or else is allowed for, and secondly, that the coupling coil in inductive coupling is so far removed from self-resonance that its effective inductance can be considered constant over the band.

Appendix

The Transfer Admittance Formula for the Phase-Linear Circuit

Consider the transfer admittance formula for mutual-inductance coupling.

$$|i/e| = 2\sqrt{(g_1g_2)}\left\{1 + \frac{1}{4\delta^2k^2}\left[\nu_1^3 - \nu_1(2-\delta^2) + \frac{1-k^2}{\nu_1}\right\}^2\right]^{\frac{1}{2}}$$

Put $\delta^2 = 3k^2$ ($n = 1/\sqrt{3}$)

and

$$\nu_1\,[\equiv(\omega/\omega_0)\sqrt{(1-k^2)}] = (1+p)\sqrt{(1-k^2)}$$

where $p = \Delta\omega/\omega_0 \equiv \Delta f/f_0$.

Then

$$|i/e| = 2\sqrt{(g_1g_2)}\left\{1 + \frac{1-k^2}{12k^4}[(1+p)^3(1-k^2) - (1+p)(2-3k^2) + (1+p)^{-1}]^2\right\}^{\frac{1}{2}}$$

$$= 2\sqrt{(g_1g_2)}\left\{1 + \frac{1-k^2}{3}[1 + (2p^2/k^2)(1-(3k^2/4)) - p^3/2 + p^4/2k^2]^2\right\}^{\frac{1}{2}}$$

Now $p^2 (\equiv \Delta f^2/f_0^2)$ will always be small around the frequency of maximum response, and if k^2 is small also the above expression will differ very little from

$$|i/e| = 2\sqrt{(g_1g_2)}\left\{1 + \tfrac{1}{3}[1 + 2(p/k)^2]^2\right\}^{\frac{1}{2}}$$

$$= (4/\sqrt{3})\sqrt{(g_1g_2)}\left\{1 + (p/k)^2 + (p/k)^4\right\}^{\frac{1}{2}}$$

This is an even function of (p/k) and to the extent that the above approximations hold will give a response curve which is absolutely symmetrical on both sides of the centre frequency.

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Short News Items

A British European Airways helicopter has recently been used to make 22 airlifts of electronic equipment to the 2 754ft summit of Broad Law, on the Selkirkshire-Peebles border, in the hills above Tweedsmuir where a party of technicians from the Telecommunications Engineering Establishment of the Ministry of Transport and Civil Aviation has just completed site trials for the new Talla radio beacon. A v.o.r. beacon is to be installed as part of the re-organization of the Scottish Control Area of the National Airways System, recently announced, to mark a possible future extension of the Airways from Dean Cross (Cumberland) to Edinburgh (Turnhouse) Airport. It will operate in addition to the existing Non-Directional Beacon, at present situated near the junction of the Moffat-Talla roads at Tweedsmuir village but which will later be transferred to the mountain-top site. A road convoy of three lorries, laden with Wilcox v.o.r. equipment in packaged form and the prefabricated building to house it, left the Telecommunications Engineering Establishment at Yeading, Middlesex on 20 July and took three days to make the journey to Talla. A Bell helicopter (which was the one used by the BEA in their recent 51 minute dash in the London to Paris Air Race) of the BEA Experimental Helicopter Unit was chartered by the Ministry to carry the equipment 2 000ft up to the summit of the mountain which has no access by road. Some of the smaller items of equipment were carried inside the aircraft, but many of the crates were slung below the fuselage in a nylon net. The first lifts took to the summit the party of six M.T.C.A. telecommunications technicians from Yeading, who camped on Broad Law under canvas for a fortnight, erecting and testing the equipment on site. The operation had barely got under way when the radio station at the base camp was struck by lightning and put out of action for some time. Conditions were far from comfortable for the men camping on the summit and equipment frequently gave trouble due to excessive moisture. However, the ten days of tests were completed successfully, and flight testing proved the site very satisfactory indeed both for range and accuracy of radials. On 18 August, work started on the dismantling of equipment, and the helicopter carried out the airlift in reverse, bringing men and equipment down to the base camp. After analysis of the results, if it is finally decided to locate the v.o.r. on Broad Law, a road will probably be built to the summit from the Talla Reservoir road. The Talla v.o.r. on Broad Law will then be Scotland's highest radio station.

A stereophonic sound-reproducing system which, it is claimed, is not only the world's largest but the first open air system in Europe, has recently been installed by Telefunken at the Federal Horticultural Show in Dortmund, West Germany. In view of the complete lack of experience in open-air stereophony, Telefunken made trial sound reproduction tests on a building site in Hannover. Based on the results obtained from these series of trials, the first system was then installed on the ground of the Federal Horticultural Show. A power of 300W was provided for each stereophonic channel and the output is fed into sound columns each containing 48 loudspeakers. These sound columns are spaced about 100 yards apart and permit stereophonic sound reproduction over an area with space for about 6 000 listeners. The installation permits not only the reproduction of stereophonic records but also the reproduction of live orchestral music.

The National Bureau of Standards is now publishing its Journal of Research in four separate sections, namely (a) Physics and Chemistry, (b) Mathematics and Mathematical Physics, (c) Engineering and Instrumentation, and (d) Radio Propagation. Each of these sections may now be subscribed for individually.

Standard Telephones and Cables Ltd are to supply a quantity of airborne i.l.s./v.o.r. equipment for installation in Aeroflot TU104 jet aircraft, flying the Moscow-London route. It is similar to S.T.C. equipment already in service with B.O.A.C. Britannias and B.E.A. Viscount aircraft. The airborne equipment operates in conjunction with ground facilities installed on air routes and major airfields all over the world.

A Sectional Meeting of the World Power Conference is to be held in Madrid from 5-9 June, 1960. The theme of this conference will be Methods of Solving Power Shortage Problems and arrangements have been made for the presentation of seventeen or eighteen papers by the British National Committee. The general programme, containing membership and other application forms together with particulars of six alternative Post-Conference Tours, is announced for publication by the Spanish National Committee towards the end of November 1959. Copies will be distributed as soon as they have reached the British National Committee. In the meantime preliminary notifications of intention to participate should be sent to: The Secretary, British National Committee, World Power Conference, 201-2 Grand Buildings, Trafalgar Square, London, W.C.2.

The Development and Workshop Production sections of the Engineer-in-Chief's Department of Cable and Wireless Ltd has recently moved to new premises in Kirby Street, London, E.C.1. The new building has been named Smale House after J. A. Smale, Engineer-in-Chief of Cable and Wireless Ltd from 1948 to 1957 and who has been closely associated with much of the research and development work of the company. At Smale House, facilities will be provided to connect it to any of the overseas stations of the company's system so that new apparatus developed at Smale House may be tested over long distances and under service conditions.

Two British scientists have been named among the 1959 winners of the John Scott awards for developing inventions for the benefit of mankind. Dr. Henry Boot, of the Services Electronic Research Laboratory at Baldock, Hertfordshire, and Professor John Randall, Wheatstone Professor of Physics at the University of London, were cited jointly for their invention of the cavity magnetron. These awards amount to $1 000 (about £357) each. The magnetron was the basis for the development of narrow beam radar which aided precise location of aircraft targets during World War II. These awards have been presented for the past 143 years and were first established by John Scott, a Scottish scientist who died in 1816.

Mullard Limited announce that a course of ten special lectures dealing with modern developments in Solid State Physics will be held at the Enfield Technical College, Queensway, Enfield. The lectures, which will be delivered by members of Mullard Research Laboratories, will be given at 7 p.m. on successive Monday evenings, commencing on 5 October 1959. The course is suitable for physics graduates or others with comparable background knowledge, and the fee is £1. Application forms may be obtained from the Head of the Department of Pure and Applied Science at the Enfield Technical College.

The Middlesex Education Committee announce that the next entry to the full-time course in Electrical Engineering and Applied Electronics will take place in September. Evening courses, commencing on September 28, will also be held on Advanced Mathematics, Analogue Computers, Digital Computers, Transistors, Servo-Mechanisms, Pulse Techniques, Industrial Electronics, Radio Telemetry, Practical Numerical Analysis and Automatic Process Control. Details can be obtained from the Principal, Technical College, Beaconsfield Road, Southall, Middx.

LETTERS TO THE EDITOR

(We do not hold ourselves responsible for the opinions of our correspondents)

Reducing Ripple in Stabilized Supplies

Dear Sir,—Some of your readers may be interested in a simple transistor circuit for the reduction of hum and ripple content of a stabilized power supply. It is well known that a transistor, when used as emitter-follower with a battery or a reference stabilizer valve in the base circuit, gives a low output impedance supply almost as stable as the reference source. If instead of a d.c. voltage, a capacitor were connected as in Fig. 1

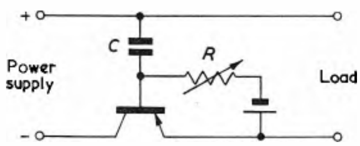

Fig. 1. Basic ripple reducing circuit

the circuit will reduce the ripple by a factor approaching β of the transistor, the actual improvement depending on the time-constant of the circuit. The base current is provided in the figure by a battery and resistance R. It could be provided by a rectifier and smoothed low voltage winding such as a heater winding on the mains transformer. The ripple in the base could be kept low by using very large capacitance which is easily obtainable for low voltages. However, the basic advantage of this circuit lies in its simplicity and from this point of view it would be difficult to improve on a 1·5V battery. The battery consumption can be kept low by using a second and if necessary a third transistor as for example in Fig. 2. The unit could be used as an ex-

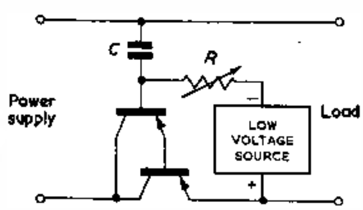

Fig. 2. Alternative, low consumption and improved ripple performance circuit

ternal attachment to an existing stabilized power supply whose output ripple is excessive. It is independent of output voltage and only the base resistance R has to be adjusted so that the variations in output circuit do not cause the transistor to be either bottomed or overloaded.

Since most popular transistors in this country are of the pnp type the control circuit has to be placed at the negative end of the supply. npn transistors could be used on the positive side.

A particularly useful adaptation of this circuit is to klystron power supplies. Here the positive end is earthed and so it is advantageous to use the pnp transistor. Also the current is virtually constant hence the resistance R can be adjusted once and for all for optimum operation. The idea was tried out here recently when it was found that the 20mV ripple from a Q-band klystron power pack was excessive. A single V 10/15A audio transistor (β = 23) with 1·5V cell, 2μF capacitor and 2·7kΩ resistor reduced the ripple to below 1mV.

Other methods of biasing the transistor could be utilized such as a resistance between collector and base[2] with a suitable protective circuit (Zener diode) to avoid switching overload, but none appears as simple, trouble free and easy to apply as the one described.

Yours faithfully,

M. H. N. Potok,

Department of Physics,
University of Cambridge.

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A Crystal Controlled Induction Heater

Dear Sir,—Referring to the correspondence on the above subject published in the December 1958 issue, the working frequencies stated by Mr. Ashford are correct. In the Federal Republic of Western Germany they are enforced under penalty of a fine by a special "High Frequency Law" (passed in 1949 by order of the Allied Military Government).

In practice, however, the frequency of 13·56Mc/s is hardly used at all, while the higher frequencies are applied exclusively for dielectric generators.

I am quoting from "Die induktive Wärmebehandlung"[1]:

"Three important requirements must be mentioned: frequency stability, freedom from feedback into the main and a very appreciable reduction in generation, inductive propagation and radiation of harmonics. How feasible it is to implement these requirements must be considered under three headings: physical-technical, economical and operational.

"The frequency allocation under the law is completely insufficient for induction heating, even if no account is taken of the fact that the lowest frequency band is really too high for many requirements. In this respect physical characteristics are paramount, e.g. the dimensions of the work, depth of penetration, electrical and thermal conductivity of the material. Account must also be taken of changes in these characteristics during operation; particularly the values σ and μ are very temperature dependent. The generator must, therefore, be designed to take care of these changes and to deliver the maximum output into the work during the whole of the operation. According to today's knowledge only a self-excited generator with automatic frequency control would meet this requirement, more particularly as the specification requires the whole operation to be completed in a very short time. The latter is the basic advantage of induction heating, allowing a relatively large energy concentration per unit of surface and volume of the work. That is why the use of stabilized frequency obtained by tuning adjustments of the oscillatory circuits and the coupling of the output stage seems hardly possible with the means at our disposal today, particularly as the necessary measures nearly always result in a considerable worsening of efficiency. Such arrangements also lead to a considerable increase in complexity of the equipment and are not desirable from the reliability point of view."

The working frequencies of nearly all inductive generators are one to two decades lower than the three Atlantic-City-Frequencies—The German "High Frequency Law" permits any desired frequencies for screened generators, provided that the following defined field intensities for the fundamental wave and the other frequencies are not exceeded:

The figures shown in Table 1 might be of interest to your readers.

Yours faithfully,

N. C. Weyss,

Brown, Boveri & Cie AG.,
Mannheim, Germany.

REFERENCE

1. Brunst, Kegel, Weyss. Die induktive Wärmebehandlung. p.189. (Springer-Verlag Berlin 1957).

TABLE 1.

		MEASURING DISTANCE	FIELD INTENSITY
All frequencies, (41..68) 174..223Mc/s	Urban area Urban area	d= 100m d= 30m	E=<45μV/m E=<30μV/m
All frequencies,	Industrial area	d=1500m	E=<10μV/m

Current Creep in Transistors

DEAR SIR,—With reference to Mr. R. E. George's discussion in the March issue of ELECTRONIC ENGINEERING, we have on various occasions been confronted with a similar problem.

Various transistor manufacturers state that with this circuit configuration I_{co} is composed of two components, namely surface leakage component which arises from surface contamination, which is directly proportional to applied voltage, and a thermal component which arises as a result of diode action, which varies exponentially with temperature and is independent of applied voltage. According to our investigations the creep in I_{co} was found due to a region of depletion adjacent to the junction.

This depletion or barrier region is free of mobile charge and is essentially a dielectric of permittivity e similar to that found in diodes biased in reverse direction. Due to this effective capacitance, it is possible for transistors with slight junction impurities to indicate an I_{co} of exponential characteristic; I have found this more so in certain drift transistors which have ended in thermal run-away even at constant ambient temperature, and stabilized voltage.

Yours faithfully,

R. FELDMAN,

Allis-Chalmers Manufacturing Co., Milwaukee, Wisconsin.

DEAR SIR,—I am pleased to confirm Mr. George's observations in ELECTRONIC ENGINEERING (March 1959 page 172). Last year an attempt was made to calibrate a transistor as a temperature sensing element, using a small oven to raise the temperature at a rate of approximately 1°C/min with frequent pauses. I_{co} was

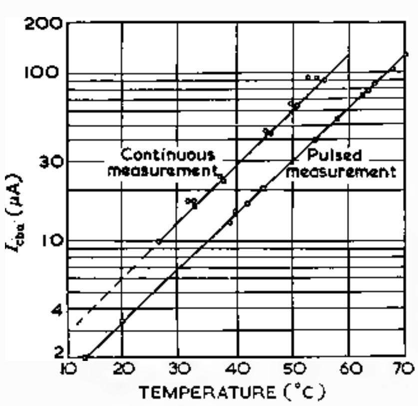


Fig. 1. Current/temperature measurements

measured using 10V applied continuously. The results, which were rather variable, are shown in Fig. 1.

Pulsed measurements, made by suddenly applying voltage and reading the meter (AVO type 8) the moment the pointer was steady, were much more consistent, but differed from the continuous measurement by a factor of approximately half (see graph). This pulsed measurement of I_{co} was constant, at all temperatures, for applied voltages between 2 and 10V i.e., it was a normal saturation current.

Observations of the time rate of increase of current made at 10V and 61°C gave similar results to those reported, although after 1 000sec continuous rise, 15sec 'off period' was needed to return the measured I_{co} to its 'pulsed' level.

The transistor measured was an OC71 which is still available for further tests.

Yours faithfully,

M. R. NICHOLLS,

Decca Radar Ltd, Chessington.

Electrometer Amplifiers

DEAR SIR,—The electrometer amplifier described in your March issue (letters: page 173) closely resembles one used by Electronic Instruments Ltd since 1947. It appears in several of their instruments and forms the basis of over 4 000 pH-meters manufactured by them. It has often been described[1,2].

The original circuit has two advantages omitted in your correspondent's version which may be of interest; here it is shown in skeleton form (Fig. 1). By

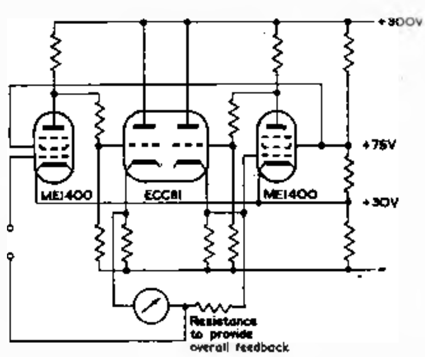


Fig. 1. Basic amplifier circuit

returning the cathodes of the voltage amplifier pentodes to a low resistance point about + 30V above base, heavy d.c. locking is provided. This closely 'pegs' working conditions so that the input stages can be arranged as starved amplifiers with very high stage gain, up to 1 000 times using ME1400 electrometers. Secondly, potential divider coupling between the voltage amplifiers and the triode output stages avoids the need for high resistance in the upper or cathode arms of the meter bridge. Such resistance is harmful in that it reduces output stage transconductance heavily, by a factor roughly equal to its value multiplied by the mutual conductance of each triode valve. Resistance below the meter arm is unimportant because it is paralleled by the very low incremental resistance of the amplifiers.

High sensitivity is valuable because it can permit high degeneration. The overall trans-conductance of the circuit shown can be as high as 300mA/V. This means that in a pH-meter covering from 0 to 14 (814mV at 20°C) and using a 100μA meter, overall feedback factor can exceed 2 400 times. Heavy degeneration is desirable, to stabilize gain, to raise input resistance, and to reduce effective capacitance across the input circuit, in all cases to an extent numerically equal to the feedback ratio.

As Mr. Allenden points out, degeneration cannot improve zero stability nor can it reduce drift due to grid current. In all commercial applications of my circuit, heater supply stabilization has been included.

Yours faithfully,

G. I. HITCHCOX,

Philips Electrical Industries Ltd, Wellington, New Zealand.

REFERENCES

1. HITCHCOX, G. Electronic Instrumentation Applied to the Chemical Industry with Special Reference to pH Measurement. *J. Brit. Instn. Radio Engrs.* *13*, 401 (1953).
2. HITCHCOX, G. Extending the Limits of Resistance Measurement using Electronic Techniques. *J. Brit. Instn. Radio Engrs.* *16*, 299 (1956).

The Author Replies:

DEAR SIR,—I studied Mr. Hitchcox's amplifier with great interest, as another example of the very same principle which I quoted in my letter. The question of originality is beyond dispute; balanced d.c. amplifiers or the application of overall feedback are probably as old as the history of electronics.

There is only one thing I should like to refer to, and that is the 'difference' between the two circuits as pointed out by Mr. Hitchcox. I am afraid I am not able to discover this 'difference'.

The d.c. compensation of the anode voltage—as that is the only significance I can attach to the +30V cathode bias—is 16V in my circuit, obtained from the heater-dropping resistor chain, as it is easy to see on the circuit diagram.

Secondly, I cannot find any difference in output impedance between the two circuits. (See Fig. A.)

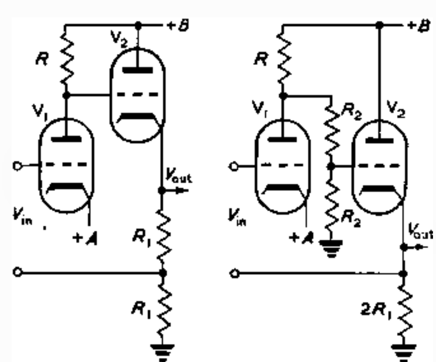


Fig. A. The two circuits discussed

The definite disadvantage of the second version is the $2R_1$ load on the voltage amplifier, as its value is limited by the grid current in V_2.

Yours faithfully,

A. D. BORONKAY,

Unicam Instruments Ltd, Cambridge.

BOOK REVIEWS

Fundamentals of Radio and Electronics

Edited by W. L. Everitt. 792 pp. 595 figs. Medium 8vo. 2nd Edition. Constable & Co. Ltd. 1959. Price 57s. 6d.

THIS is a large and somewhat expensive book claiming to be 'eminently suitable for self study' which can 'be used by people at all technical levels' and 'has a coverage to include the general field of electronics'. These are big claims and the value of the book in this country has been judged in the light of them. There is little doubt that the general field of electronics is covered but so superficially as to be of little value to the student.

In Chapter I, mathematics is revised from the addition of '21 + 12 = 33' to the differentiation of exponential series in 47 pages and includes one page which has log scales to be 'cut out and pasted on cardboard or wood to make a model slide rule'. Surely the student can be persuaded to go out and buy himself a slide rule or advised not to proceed further. This sort of treatment pervades the book, sometimes the material is treated exhaustively, usually when it is very simple, and at others results are assumed, e.g. in Chapter III the energy of a magnetic field associated with a coil is assumed to be equal to $\frac{1}{2}LI^2$ and no units are mentioned. No statement was found anywhere of any system of units employed and it is difficult to guess the intention. The ampere is defined in terms of its electrochemical action and axes on the hysteresis curves quoted are labelled 'Lines per square inch' against 'ampere turns per inch'.

Chapter IV deals with valves and transistors and takes us from de Forest to the klystron, magnetron, camera tube and transistor. Incidentally, it was surprising to see in Chapter IV that the image orthicon 'employs an electron-beam-scanned mosaic plate' although a good description of the image orthicon is given in the later chapter on monochrome television. In explaining transistor action, one feels sympathy for the student who suddenly meets the expression 'It is known that the elements of the Fourth Group of the Periodic table . . . have four valence electrons in their outermost shell . . .' and one feels that the addition of fractions in Chapter I might have been omitted and such a remark explained—if it is considered necessary to the understanding of transistor action.

Throughout the book American terminology would be confusing to the British student. For instance, it is no use looking up n.f.b. in the index—it is called inverse feed back; integrating and differentiating circuits are not referred to by these names but their action is called 'rate of change action'. No distinction is made between d.c. restoration and clamping. Some of the phrasing, too, while possibly good American, is poor English, e.g. 'After a wave has passed through an R.C. amplifier and lost its own d.c. axis . . .' and elsewhere 'the gas tube cannot turn on and off too rapidly' meaning, as appears later, that it cannot turn on and off very rapidly.

The chapters on monochrome and colour television are of course dealing entirely with the American system, and British students looking for an introduction to the subject would be advised to read books dealing with our own standards first. In connexion with the treatment of television in this book, it is a little surprising to find no mention of photo-conductive tubes and the recording and transmission of film gets very slight treatment. No mention is made of the magnetic recording of video.

Some chapters are quite good, particularly the chapters on 'Electromagnetic Waves' and 'Transmission and Reception of Signals by Radio' although all the equipment referred to is American.

In summary, it may be said that this book sets out to lead the student through the electronic field and in this it largely fails. It might be interesting to technical readers in that it gives a general, superficial and almost entirely non-mathematical review of many possibly unfamiliar electronic applications but superficiality is ultimately annoying when information is expected. The authors have among them a mass of first-rate material which in the process of condensation and simplification has lost a good deal of its original value.

H. HENDERSON

Radio Engineering Formulae and Calculations

By W. E. Pannett. 200 pp. 160 figs. Demy 8vo. George Newnes Ltd. 6th Edition. 1959. Price 17s. 6d.

IN this book, examples have been selected from everyday practice in design, installation and operation and are shown fully worked and generously illustrated. Although it is intended primarily for the practising engineer, it will be found equally valuable to the technical assistant and the student about to embark on an engineering career.

The introductory section shows how the common mathematical tools can be best used, with simple examples. Then follows, in concise form, radio engineering formulae and reference notes, grouped under subject headings, and accompanied by many worked examples of problems typical of those encountered by field, installation, maintenance and design engineers. The final section of reference material includes much useful data and the mathematical tables commonly required in practical work.

Punched Cards: Their Application to Science and Industry

Edited by R. S. Casey, J. W. Perry, A. Kent & M. Berry. 697 pp. 191 figs. Medium 8vo. 2nd Edition. Chapman & Hall Ltd. 1959. Price 120s.

THIS well-known book on punched card systems has been thoroughly revised and brought up to date, giving a second edition which is in some respects supplementary to the first, in that few chapters remain unaltered, and many new ones have been added.

Those not familiar with the earlier edition must not expect a book on calculating and computing machines. Its scope is strictly limited to the use of cards for recording, sorting and correlating data, and especially for indexing literature. Although of particular value to librarians and information workers, all engineers and scientists should know at least as much about punched cards as is contained in the chapters which are recommended in the preface as an introductory survey.

The book is in five parts.

Part I describes the various types of hand and machine-sorted cards, and gives detailed instructions for the manipulation of hand-sorted cards. For completeness the 'Peek-a-boo' system should have been described here, instead of with the applications in Part II.

The variety of new applications given in the second part indicates the growing importance of punched cards for documentation and record sorting. Examples range from chemical to linguistic analysis, and from report writing to patent searching. The Uniterm system is included, but does not make use of punched cards, and so is rather out of place.

Classification, coding and searching operations are analysed in Part III. This is valuable, since the efficiency of a system in practice depends to a large extent on these operations. Symbolic logic is introduced, but not to an extent that need deter even the casual reader.

Part IV is entitled 'Future Possibilities', inappropriately since its two chapters are on advanced, but current, applications. The chapter on mechanized documentation is too long, as are some other chapters in the book.

The bibliography, which forms the last part, is a supplement to the bibliography in the first edition, and brings the references up to date by adding a further 401 to the original 276 items. There are also lists of references for the individual chapters.

It would have been useful to include a section giving a comparison, from a practical viewpoint, of punched cards with other methods of indexing, and a clear explanation of the factors to be considered when selecting systems for various applications. The book is nevertheless a comprehensive survey of the available equipment, and gives extensive and valuable examples of its use. Although the equipment described is primarily American, the same, or similar types of equipment are available in this country.

G. G. BOYT

Basic Electricity (in five parts)

By Van Valkenburgh. Nooger & Neville Inc. 600 pp. 620 figs. Royal 8vo. Technical Press Ltd. 1959. 12s. 6d. per part. 55s. per set.

THE series of books is listed as a course of training developed for the U.S. Navy, where many recruits with no knowledge of electricity were to be trained rapidly as technicians—operators and maintenance men—but not to be trained further as electrical engineers. A modified form of the original Navy manual was introduced for American schools, technical institutions, etc., and was apparently well received. In this edition the terminology has in general been altered to conform to British practice, but it is most unfortunate that there are many lapses.

Teaching is in colloquial style, with much repetition; most of the illustrations are sketches of the apparatus as seen, but in some cases followed by a circuit diagram of the type the user will later meet in practice. The general level is somewhat similar to that of a boys' book in this country, but many bright boys might consider much of this as 'kid's stuff'. The appeal is to the dimmest student, which is presumably intentional.

Loose thinking is apparent in various ways; for example e.m.f. and p.d. appear to be regarded as alternative terms because they are both measured in volts. The almost alternate use of meter 'pointer' and 'needle' shows that the parts dealing with measuring instruments (termed meters throughout) were not written by anyone familiar with instruments. Among examples that irritate are the continual use of 'D.C. Current'. Do the writers not know what d.c. stands for?

Although some literal and graphical symbols are used there are some departures from B.S. practice; literal symbols such as k instead kΩ, for example, are all quoted as abbreviations, which they are not in British practice.

Because of this lack of sound editing the work has less application to British students, although the simple style may be of the level required for mass-produced technicians with limited capabilities.

E. H. W. BANNER

Loudspeakers

By G. A. Briggs. 336 pp. 50 figs. Demy 8vo. 5th Edition. Wharfedale Wireless Works Ltd., Idle, Bradford. 1959. Price 19s. 6d.

THE statement on the reverse of the title page, 'Fifth Edition, Completely Revised and Enlarged,' is by no means an understatement. It looks all new. There is no doubt that Mr. Briggs works on his books, as a result of which each offering is more informative, interesting and entertaining than its predecessor. This one represents tremendous all-round advance from the previous edition in these respects.

Some in the publishing business have commented that Mr. Briggs' previous offerings have been somewhat 'amateurish'. If that means failure to adhere to traditional rules for writing this type of book, he still is refreshingly so. But from the 'mechanical' viewpoint—make-up, general presentation, and concise informative content, this is no amateur offering.

His coverage includes many historical and personal notes that give the reader a feeling of being personally acquainted with many of the people who have contributed to 'high fidelity's' progress. The early part of the book treats the various parts that make a loudspeaker. It then goes into various aspects of performance characteristics. Finally come essentially related matters, such as room acoustics, crossovers, audio fairs, concert halls and various other details.

His chapter on stereo is typical of his general philosophy in the book. There are so many controversial areas in such a subject that the 'experts' often disagree, so it is difficult to be impartially authoritative. In this particular area, Mr. Briggs has taken four representative gentlemen 'whose opinions we value, but do not necessarily share', provided them with a comprehensive demonstration, and then given a full report of their individual findings and opinions.

In his introduction, Mr. Briggs predicts that the first two chapters ('Looking Back' and 'Books') are calculated to exasperate any technical reviewer. Far from being exasperated, this reviewer derived considerable enjoyment from reading them. Many must have asked themselves, 'What makes a man in the loudspeaker manufacturing business take up publishing? What's he up to?' Mr. Briggs tells us his story in very pleasant style.

Altogether, this book is strongly recommended for the bookshelf of anyone with any interest whatever in either loudspeakers or high fidelity.

N. H. CROWHURST

Théorie et Pratique des Circuits Electroniques et des Amplificateurs

By J. Quintet. 256 pp. 160 figs. Vol. I. Demy 8vo. Dunod, Paris. 1958. Price 1,960 fr.

THE title of the book, of which this is the third edition, may be somewhat misleading unless it is noted that this is Volume I, devoted to Complex Algebra and Networks.

Apart from a specific circuit (pentode integrator) and as an application of Thevenin's theorem the thermionic valve only enters into the subject in the last chapter where the negative resistance effect is dealt with, no doubt as an introduction to oscillators which form part of Volume II.

This is not a book for experts. About one quarter of the contents deals with complex algebra and its application to simple a.c. circuits. Two useful chapters then follow: One on network theorems and circuit transformations and the other on mutual inductance. The general transformer equations are then deduced but in quoting the approximate equations the author draws particular attention to the absence of resistance but fails to insist on the equally important approximation of zero leakage.

There remains almost half the text for the circuits which enter into amplifiers and oscillators, transients in *RL*, *RC* and *RLC* circuits, series and parallel resonance and coupled circuits.

The author has made every effort to present a clear understanding of his subject and if some of the calculations appear elementary the text is singularly free from such aggravating expressions as "it is evident that . . .".

Students will find much of value in the book particularly in the chapters dealing with mutual inductance and coupled circuits. As would be expected, however, the choice of material is ill fitted to most British engineering syllabuses, but it might well fit simpler more specialized courses. The detailed arithmetic is calculated to give confidence to a student by showing him not simply that he has gone wrong (which is all that a numerical answer to a problem can do) but where he has gone wrong.

In the reviewer's opinion the author has tried to spread his net too widely by stating in his preface that the book is eminently suitable to, among others, engineers and that radio amateurs could profit from it. Surely specialization in the training of the engineer has not reached such a pitch that a qualified man needs to be taught complex algebra. On the other hand it is doubtful if the radio amateur would have the patience to reach the important aspects. To those in between the book has attractions.

G. ASHDOWN

THE INTERNATIONAL TRANSISTOR EXHIBITION

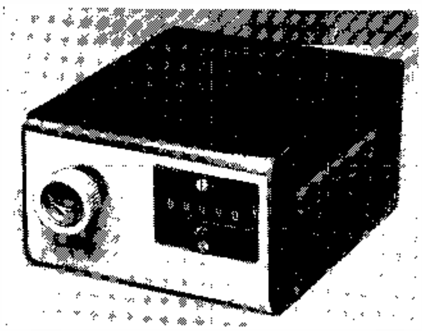
A description, compiled from information supplied by the manufacturers, of a further selection of the exhibits at the International Transistor Exhibition.

(Voir page 503 pour la traduction en français: Deutsche Übersetzung Seite 507)

DEPENDABLE RELAY CO. LTD.
8a Ainger Road, London N.W.3
MINIATURE HIGH SPEED COUNTERS
(Illustrated below)

The standard model of this miniature high speed electromagnetic counter will count at speeds of up to 1 500 counts per minute. It is complete in a box with an outlet for a photo-electric cell or for ultra-sensitive contacting units such as moving-coil relays. A selector switch is provided for selection of photocell or normally open or normally closed contact operation. The current carrying capacity of the contact need only be 500μA. The unit measures 5in × 4in × 2½in and can be supplied for a.c. mains or battery operation. Units with batch counting and electrical reset can also be supplied and a model capable of counting at 3000 counts/minute is also available.

EE 12 751 *for further details*



FERRANTI LTD
Hollinwood, Manchester, Lancashire
SILICON TRANSISTORS

General work is being carried out on high-speed silicon transistors in the range 50 to 100Mc/s, the technique of solid state diffusion being applied to their production. In these transistors both the base layer and the emitter are formed by diffusion into a silicon wafer which forms the collector electrode. These transistors are at present being produced in small quantities. The Wythenshawe laboratories are also working on alloy signal transistors working at frequencies up to 10Mc/s which are particularly suitable for fast switching applications. It is anticipated that both types will be commercially available towards the end of the year. Ferranti Ltd have announced that they will concentrate on silicon alloy and diffused transistors for military applications and for industrial applications, where high temperature operation would rule out the use of germanium transistors.

Parametric diodes intended for high frequency use are under development at present and work on four layer diodes has been progressing rapidly with the result that these are now in pilot production. This is a pnpn diode which can act as a switching device for use in computers and pulse circuits. It is basically the semiconductor version of a gas discharge tube. Striking voltages range from 80 to 120V and the 'on impedance' is a few ohms. Holding on currents range from 1 to 50mA. It is about the same physical size as the ZS10 series diodes.

EE 12 752 *for further details*

SILICON SOLAR CELLS

Another interesting device made by the diffusion method is the silicon solar photovoltaic cell, which will convert energy from the sun or from incandescent lamps into electrical energy. It has a response time of a few microseconds and operates without deterioration under conditions of high temperature and humidity.

Under such conditions light falling upon a silicon photovoltaic cell will produce an increase in the temperature of the cell; fortunately, however, silicon cells will work at high temperatures without being damaged and with very little reduction in their output voltage. Connecting an external load causes a minor reduction in this temperature due to the tranfer of energy from the cell to the load. With cells of this type power conversion efficiencies of over 11 per cent can be achieved.

These cells can be supplied in various shapes and sizes, the maximum area for a single cell being two square inches. It is possible to use them in series, parallel, or in series parallel combinations to obtain higher output voltages and currents.

Quite substantial amounts of current can be obtained from the solar cell. For example, a cell with an area of 1in² can deliver a current of 20mA when exposed to white light of moderate brightness.

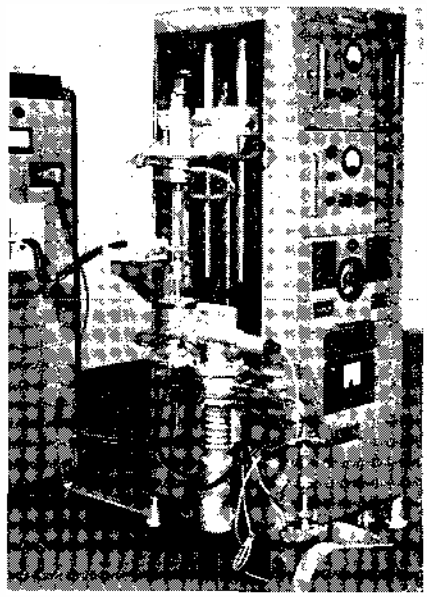
EE 12 753 *for further details*

RADIO HEATERS LTD
Eastheath Avenue, Wokingham, Berkshire
SILICON REFINING UNIT
(Illustrated above right)

The Radyne floating zone silicon refining unit has been specially designed to meet the requirements of the semiconductor industry.

The unit will handle silicon rods varying from 1 to 2½cm in diameter and 15 to 56cm in length. The principle of operation used is to traverse the coil over the silicon rod while the latter is rotated. The plant is complete with all vacuum equipment and control gear and is designed to work in conjunction with either 5, 10 or 15kW, 450kc/s, 3 to 4Mc/s Radyne r.f. generators.

The quartz tube refining assembly is carried on two heavy cast aluminium brackets. Vacuum seals are provided on the top and bottom bracket assemblies. 'O' rings are used between bolted flanges, modified Wilson shields on the rotating shafts and a Radyne pinch seal of neoprene rubber for the quartz tube seals. Quartz tubes ranging from 20 to 40mm internal diameter can be accommodated. Adjustable chucks are provided for sup-

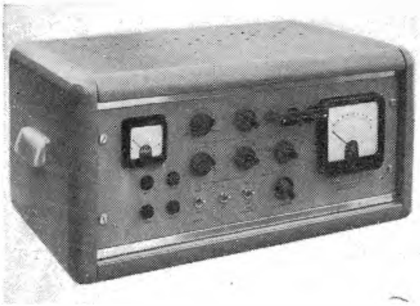


porting the silicon rod and the rotation speed can be varied from 2 to 60rev/min. The rod can be driven from both chucks or either chuck independently. The speed of rotation is continuously variable and controlled by a Servomex amplifier. The top bracket supporting the quartz tube can be adjusted to accommodate lengths of tube varying from 8 to 24in (203 to 610mm). In addition, provision is made for adjustment of the zone length during processing and a simple collet release is provided on the bottom chuck so as to accommodate various lengths of silicon rod.

EE 12 754 *for further details*

RIVLIN INSTRUMENTS LTD.
Doman Road, Camberley, Surrey
TRANSISTOR TESTER
(Illustrated above right)

This instrument is designed to indicate changes in the characteristics, or the failure of a pnp junction transistor. It enables measurement of the grounded emitter current gain the open-circuit base-to-collector leakage current and the collector turnover voltage to be made



over wide ranges of current and voltage.

All readings are taken on a 4½in meter, having a knife edge pointer and mirror scale.

A 2½in meter is provided for monitoring the low voltage supply to the transistor under test.

Collector voltage can be measured from 0 to 20V, collector current from 0 to 2·5A, current gain from 0 to 100 and collector leakage current from 0 to 2·5A in six ranges. The measurement of turnover voltage is dependent on the maximum dissipation of the transistor under test but three ranges of 2·5, 25 and 100mW give voltages up to 85V.

The accuracy of measurement of current gain and turnover voltage is ± 5 per cent and of leakage current ± 2 per cent.

The low voltage supply to the transistor is variable in five steps from 0·4V to 19V with a fine adjustment provided on the 0·4V position.

All controls and the connectors for the transistor under test are mounted on the front panel.

EE 12 755 *for further details*

ROBAND ELECTRONICS LTD

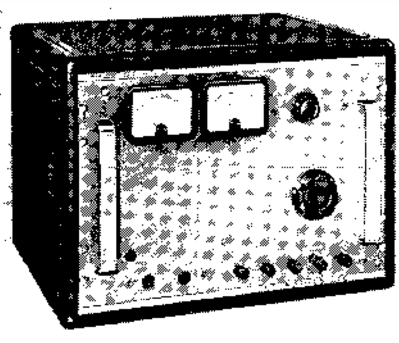
33 Mountgrove Road, Highbury, London N.5

REGULATED POWER SUPPLIES

(Illustrated below)

A complete range of transistorized regulated power supplies was shown extending from open chassis units of 25V max at 1A to variable supplies of 0 to 30V at 20A. A typical example, the T111, a 0 to 50V 5A unit is illustrated. Among advantages claimed are adequate protection against accidental overload, 4 terminal output networks to eliminate the resistance of interconnecting leads, and relay protection to avoid current surges during switching-on conditions. All sub-units can be preset to operate at any voltage at the maximum quoted current.

EE 12 756 *for further details*



SIEMENS EDISON SWAN LTD

155 Charing Cross Road, London W.C.2

BETA TESTER

(Illustrated below)

This simple test set provides a quick, direct, indication of current gain and collector leakage current of pnp transistors under common emitter conditions.

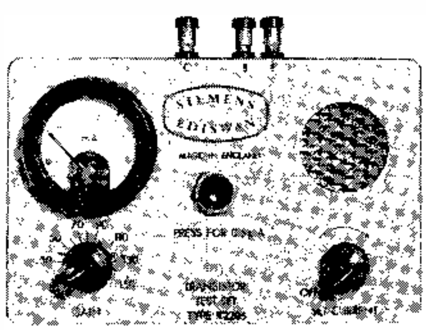
Current gain (β or α'), within a range 10 to 150, can be read directly on a calibrated dial by adjusting until an audible note just ceases. This measurement can be carried out at a collector current of 0·5 to 4mA, the set collector current being read on a meter.

Leakage current is measured on the meter, at a fixed collector voltage of 9V. A push-button switch alters the range of the meter to 500μA f.s.d. for this test.

Accuracy of measurement: ± 5 per cent or ± 5 whichever is the greater.

The test set is a battery operated transistorized unit housed in a robust die-cast box. The transistor under test is connected by means of quick-release terminals, and both measurements can be carried out by unskilled personnel in a few seconds.

EE 12 757 *for further details*



TELEFUNKEN G.m.b.H.

Ulm/Donau, Germany

TRANSISTOR TESTER

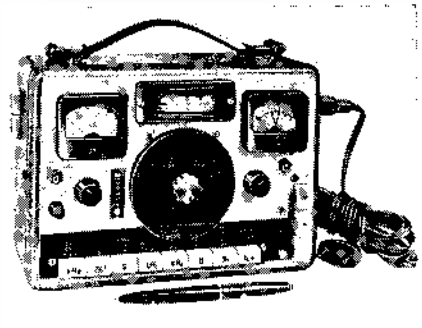
(Illustrated above right)

This transistor tester, known as the 'Teletrons I', is a multi-purpose measuring equipment for npn and pnp transistors. It will make five static tests and carry out seven dynamic measurements at 1kc/s for various adjustable operating points; all dynamic measurements being effected in the emitter circuit.

The tester consists basically of a bridge circuit, a 1kc/s generator, a null indicator (amplifier and meter) and a mains power supply unit.

All *h* and *y* parameters can be measured as a.c. characteristic quantities in the emitter circuit. The characteristic quantities are: input resistance (short-circuit output), current amplification, transconductance, output resistance, (input-shorted), inverse voltage transfer ratio, and inverse transconductance, three measuring ranges having been provided for each of these parameters.

By means of calibrated resistors in the measuring bridge an exact determination of the individual measuring values can be determined, as required, from the short-circuit and open-circuit parameters



measured This is effected by applying simple conversion formulae included in the instructions for use.

The operating points of the transistor to be measured can be adjusted continuously between 0 and 30V collector voltage and 0 and 5mA emitter current. The incorporated indicator instruments are each provided with two measuring ranges, which can be switched over by pressing a button on the instrument, enabling the exact setting of operating points (voltage ranges 0 to 6V and 0 to 30V, current ranges 0 to 1mA and 0 to 5mA).

In addition to these a.c. characteristic quantities the most important d.c. checking measurements can be effected on the transistor collector junctions. These include measurements of the collector cut-off current I_{co}, I_{co}', I_{ck} and the measurement of the emitter cut-off current I_{oe} and the associated breakdown voltage in so far as they lie in the range between 0 and 30V. Finally the base d.c. voltage V_{be} can be read off on the zero indicator for the operating point concerned. Knowledge of the base d.c. voltage is important, for instance, for stabilizing the transistor in the circuit.

EE 12 758 *for further details*

TRANSITRON ELECTRONIC CORPORATION

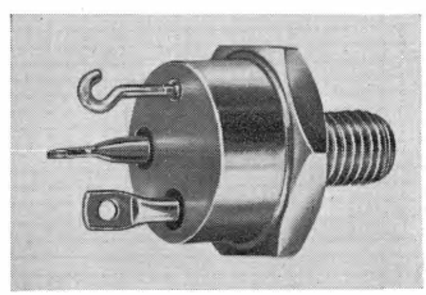
168-182 Albion Street, Wakefield, Massachusetts, U.S.A. Represented by: G.E. Industrial Supplies Ltd, 357 Euston Road, London N.W.1

SILICON CONTROLLED RECTIFIERS

(Illustrated below)

These silicon controlled rectifiers are pnpn power switching devices and have current ratings of up to 10A at a case temperature of 100°C and peak inverse voltage ratings ranging from 50 to 400V.

Within their power capabilities Transitron controlled rectifiers can be used to replace thyratrons, ignitrons, switches, relays, power transistors, magnetic amplifiers and contactors.



They are mounted on a standard 11/16in hexagonal base which permits easy mounting by drilling one hole only.

EE 12 759 *for further details*

VENNER ELECTRONICS LTD

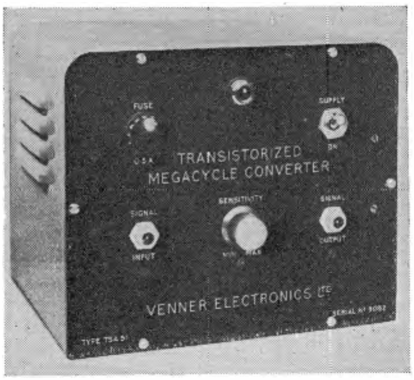
Kingston-By-Pass, New Malden, Surrey

MEGACYCLE DIGITAL CONVERTOR

(Illustrated below)

This 'add-on' unit is designed for use with Venner frequency meters type TSA3 and TSA53. It enables measurements to be made up to a frequency of 1Mc/s on 100 kc/s instruments, or up to 500kc/s on 50kc/s instruments. The convertor may also be used in conjunction with frequency measuring equipment of other manufacture, but intending users should verify that this applies to their particular instrument before ordering.

The convertor consists essentially of a scale-of-ten transistor megacycle divider preceded by a megacycle pulse shaper



and emitter-follower stage presenting a high input impedance to the signal. An input attenuator and a well stabilized mains power supply are included.

To use, the output of the convertor should be connected to the input socket on the frequency meter, which should be set up for frequency measurement in the usual way. The input attenuator should be set to maximum sensitivity. The signal to be measured should be fed to the coaxial input socket of the convertor and the convertor input attenuator suitably adjusted.

If a gating period of 1sec is employed the reading displayed on the frequency meter should be multiplied by a factor of ten to give an answer in cycles per second. The sixth significant figure will not be shown, but can now be obtained by employing the 10sec gating period, e.g. a frequency of 962 115c/s would appear as 96211 with the 1sec gate, or as 62115 with the 10sec gate.

EE 12 760 *for further details*

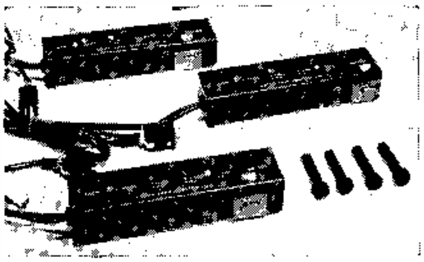
WAYNE KERR LABORATORIES LTD

Roebuck Road, Chessington, Surrey

TRANSISTOR ADAPTORS

(Illustrated above right)

The Q601 series of transistor adaptors is designed to measure transistor para-



meters under small signal conditions over a frequency range of 15kc/s to 5Mc/s. The adaptors are used with the Wayne Kerr radio frequency bridge type B601 and a suitable signal source and detector.

Two series of adaptors have been designed; one for pnp transistors and the parameters which can be measured are as follows:

Y_{11}', the input admittance in the grounded-emitter configuration with the output short-circuited.

Y_{11}, the input admittance in the grounded-base configuration with the output short-circuited.

Y_{22}, the output admittance in the grounded-base configuration with the input short-circuited.

α, the current gain in the grounded-base configuration with the output short-circuited.

Y_{21}', the transfer admittance in the grounded-emitter configuration with the output short-circuited.

EE 12 761 *for further details*

WESTINGHOUSE BRAKE & SIGNAL CO. LTD

82 York Way, King's Cross, London N.1

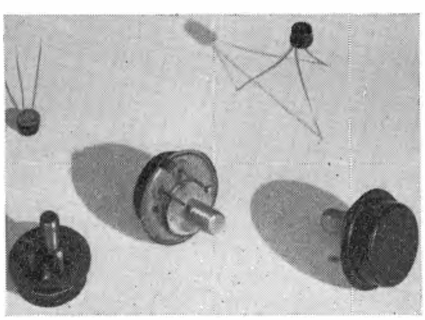
SWITCHING DEVICES

(Illustrated below)

The Dynistor switch is a multi-layer pnpn germanium device having a breakover voltage characteristic which enables it to change from a high resistance (blocking) condition to a low resistance (conducting) condition in a fraction of a microsecond. This switching time is considerably shorter than that of switching transistors of comparable power handling capabilities.

The breakover may be initiated by a short-duration trigger pulse of sufficient magnitude to raise the Dynistor voltage above the breakover level. This trigger pulse can be of less than one microsecond duration.

In operating practice the circuit voltage



across the Dynistor is kept below the bottom voltage of its category, and a trigger pulse of sufficient magnitude to raise the voltage above the top voltage of that category is used. For example, type DG11 should be used at an operating voltage of less than 50V and a trigger pulse sufficient to raise the voltage above 75V would be required.

To turn the Dynistor off it is necessary to reduce the current below a minimum value known as the sustaining current which is of the order of 20 to 50mA.

If the polarity of the supply is reversed the Dynistor will offer a low resistance at all times, there being no breakover characteristic in this direction on account of the pnpn structure.

The Westinghouse Trinistor is a three-terminal multi-junction silicon npnp switch whose breakover voltage can be controlled by the base current. Electrically it is similar to a high power silicon rectifying diode to which a third (control) electrode has been added. In the forward (conducting) direction, with no input signal to the base (control) terminal, the Trinistor will block up to the full rated voltage, and will only breakover into a conducting condition if the rated breakover voltage is exceeded.

If the base is made positive with respect to the cathode, the breakover voltage is progressively reduced with increasing base drive and will fall to a few volts with a base drive of approximately 2 to 5V at 25 to 150mA. Full forward current can now flow with a forward voltage drop similar to that of a normal silicon diode. However, once forward current has commenced to flow, the base drive may be discontinued if desired, since (apart from certain special circumstances) the base no longer has any effect on the characteristics.

To turn the Trinistor off it is in general necessary to reduce the current below a certain minimum value known as the sustaining current although turn-off by a base control signal is possible at low currents.

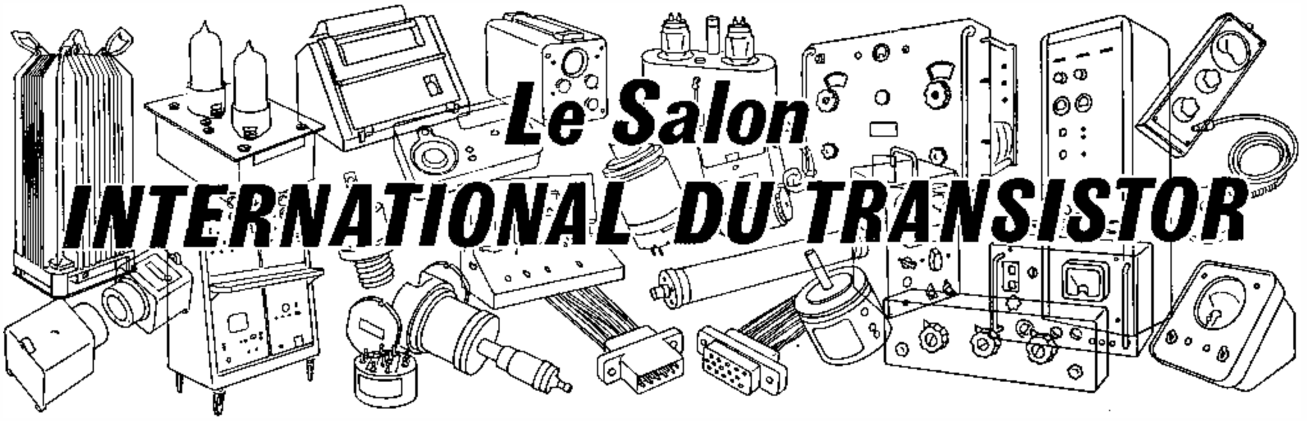
In the reverse direction the characteristic is similar to that of a normal silicon diode, being capable of blocking up to the full rated voltage. No breakover or control is possible in this direction, and the rated reverse voltage must not be exceeded.

A range of high power npn fused silicon transistors is also available and these are intended for high power switching and amplifier applications. These transistors are characterized by a very low saturation resistance and this, together with the voltage and current ratings available, permits the handling of high switching powers with a minimum of losses. Encapsulation is by a welded stainless steel case designed for mounting on an external heat sink by means of a copper stud.

They are available in 2A and 5A types with V_{ce} (max) up to 300V.

EE 12 762 *for further details*

(Pages 503 to 510 French and German Section)



Le Salon INTERNATIONAL DU TRANSISTOR

Une description complémentaire, basée sur des renseignements fournis par les fabricants, de certaines des réalisations présentées au Salon International du Transistor.

Traduction des pages 500 à 502

DEPENDABLE RELAY CO. LTD

8a Ainger Road, London, N.W.3

COMPTEURS MINIATURE À ACTION RAPIDE

(*Illustration à la page* 500)

Le modèle courant de ce compteur miniature électromagnétique à action rapide peut atteindre des vitesses de comptage allant jusqu'à 1500 coups par minute. C'est un instrument complet, enfermé dans un boîtier avec une ouverture pour cellule photoélectrique ou pour contacteur ultra-sensible, tel que les relais à bobine mobile. Un sélecteur permet de choisir soit le fonctionnement par cellule photoélectrique soit par contacteur normalement ouvert ou normalement fermé. Le courant nécessaire au contacteur peut se réduire à 500 μA. Les dimensions de l'instrument sont de 127 mm × 102 mm × 64 mm et il fonctionne sur secteur alternatif ou sur batterie. Il existe également des modèles pour comptage groupé à réenclanchement électrique, ainsi qu'un modèle pouvant compter à la cadence de 3000 coups par minute.

EE 12 751 *pour plus amples renseignements*

FERRANTI LTD

Hollinwood, Manchester, Lancashire

TRANSISTORS AU SILICIUM

La Société Ferranti se consacre actuellement à la mise au point de transistors à action rapide dans la gamme de 50 à 100 MHz, appliquant à leur production la méthode de diffusion à l'état solide. Dans ces transistors, tant la couche d'embase que l'émetteur sont formés par diffusion dans un support plat de silicium qui constitue l'électrode collectrice. Ces transistors sont actuellement produits en petites quantités. Les Laboratoires Wythenshawe travaillent également à des transistors de signaux en alliage, fonctionnant à des fréquences allant jusqu'à 10 MHz, qui conviennent particulièrement aux applications de commutation rapide. On s'attend à ce que les deux modèles soient mis sur le marché vers la fin de l'année en cours. La Société Ferranti a annoncé qu'elle portera son effort principal sur l'alliage au silicium et sur les transistors diffus pour les applications militaires et industrielles où le fonctionnement à des hautes températures exclurait l'emploi de transistors au germanium.

Les diodes paramétriques pour les hautes fréquences sont actuellement en cours de développement et des travaux sur des diodes à quatre couches progressent rapidement, de sorte qu'elles sont déjà en production pilote. Il s'agit de diodes PNPN pour calculateurs et circuits d'impulsion, représentant essentiellement la version en semi-conducteurs de tubes à décharge gazeuse. Les tensions d'amorçage vont de 80 à 120 volts et l'impédance en circuit est de quelques ohms. Le courant de maintien va de 1 à 50 mA. Enfin, ces diodes sont à peu près de même grandeur que les diodes de série ZS10.

EE 12 752 *pour plus amples renseignements*

CELLULES SOLAIRES DE SILICIUM

La cellule solaire photovoltaïque au silicium représente un nouvel élément intéressant fabriqué par la méthode dite de diffusion. Cette cellule transforme l'énergie solaire ou l'énergie de lampes à incandescence en énergie électrique. Elle a une durée de réponse de quelques microsecondes et fonctionne sans se détériorer dans des conditions de grande humidité et de température élevée. En pareilles conditions, la lumière tombant sur la cellule photovoltaïque au silicium produit une augmentation de température de la cellule. Les cellules au silicium fonctionnent, heureusement, à de hautes températures sans risque d'endommagement et avec une très faible diminution de leur tension de sortie. En reliant une charge extérieure, on provoque une légère réduction de cette température en raison du transfert d'énergie de la cellule à la charge. Des transformations d'énergie de plus de 11% peuvent être réalisées avec des cellules de ce genre.

Ces cellules sont fournies en différentes formes et grandeurs, la surface maximum d'une seule cellule étant de 12,9 cm². On peut les employer en série, en parallèle ou en combinaisons série/parallèle, afin d'obtenir des tensions de sortie et des courants plus élevés.

On peut obtenir de la cellule solaire des courants d'une puissance assez importante. Ainsi, une cellule de 6 cm² de surface pourra fournir un courant de 20 mA si elle est exposée à une lumière blanche d'un éclat moyen.

EE 12 753 *pour plus amples renseignements*

RADIO HEATERS LTD

Eastheath Avenue, Wokingham, Berkshire

ÉLÉMENT D'AFFINAGE DE SILICIUM

(*Illustration à la page 500*)

L'élément d'affinage de silicium à zone flottante RADYNE a été spécialement conçu pour répondre aux besoins de l'industrie des semi-conducteurs.

L'élément peut traiter des barres de silicium dont les dimensions peuvent varier de 1 à 2,5 cm de diamètre et de 15 à 56 cm de longueur. Le principe de fonctionnement consiste à faire passer la bobine le long de la barre de silicium pendant la rotation de cette dernière. L'élément est complet avec tous les accessoires à vide et l'appareillage de commande, et il a été prévu pour être utilisé conjointement avec des générateurs haute fréquence RADYNE de 5, 10 ou 15 kW, 450 kHz, 3 à 4 MHz.

L'assemblage d'affinage à tube de quartz est porté par deux supports d'aluminium moulé. Des scellements à vide sont fixés à l'assemblage à supports supérieur et inférieur. Des anneaux "O" sont employés entre les brides boulonnées. Les arbres rotatifs portent des scellements Wilson modifiés et les scellements des tubes de quartz sont munis de joints RADYNE en caoutchouc néoprène. Des tubes de quartz d'un diamètre intérieur allant de 20 à 40 mm peuvent être logés. L'appareil comprend aussi des mandrins réglables pour soutenir la barre de silicium, dont la vitesse de rotation peut être modifiée de 2 à 60 tours/minute. La barre peut être actionnée à partir des deux mandrins ou de l'un d'eux séparément. La vitesse de rotation est à variation continue et elle est contrôlée par un amplificateur Servomex. Le support supérieur soutenant le tube de quartz peut être réglé pour recevoir des longueurs de tube variant de 203 à 610 mm. On peut, en outre, ajuster la longueur de zone en cours de traitement et le mandrin inférieur est doté d'un déclencheur simple à pince de serrage, permettant de recevoir différentes longueurs de barres de silicium.

EE 12 754 *pour plus amples renseignements*

RIVLIN INSTRUMENTS LTD

Doman Road, Camberley, Surrey

CONTRÔLEUR DE TRANSISTORS

(Illustration à la page 501)

Cet appareil a été prévu pour indiquer les changements de caractéristiques ou les pannes de transistors à jonction PNP. Il permet de mesurer dans des gammes étendues de courant et de tension, le gain de courant de l'émetteur à la terre, le courant de fuite de la base au collecteur d'un circuit ouvert ainsi que la tension de renversement au collecteur.

Toutes les lectures s'effectuent sur un compteur de 114 mm, ayant une aiguille à couteau et un cadran à miroir.

Un compteur de 64 mm assure, en outre, le contrôle de l'alimentation en basse tension du transistor en cours d'essai.

La tension au collecteur peut être mesurée de 0 à 20 volts, le courant au collecteur, de 0 à 2,5 A, le gain de courant, de 0 à 100, et le courant de fruite, de 0 à 2,5 A dans six gammes. Les mesures de tension de renversement dépendent de la dissipation maximum du transistor en cours d'examen, mais trois gammes de 2,5 mV, 25 mV et 100 mV donnent des tensions jusqu'à 85 volts.

La précision de la mesure du gain de courant et de la tension de renversement est de ±5% et celle de la fuite de courant de ±2%.

L'alimentation du transistor en basse tension peut être effectuée, en cinq étapes, de 0,4 V à 19 V, avec réglage précis prévu à la position de 0,4 V.

Toutes les commandes, ainsi que les connecteurs du transistor en cours d'essai, sont fixés sur le panneau avant.

EE 12 755 *plus pour amples renseignements*

ROBAND ELECTRONICS LTD

33 Mountgrove Road, Highbury, London, N.5

ALIMENTATION RÉGLÉE

(Illustration à la page 501)

Une série complète de blocs d'alimentation réglés transistorisés a été exposée au Salon. Cette série s'étendait des modèles à châssis ouvert de 25 Volts à 1 A, aux modèles à alimentation variable de 0 à 30 volts à 20 ampères. Un spécimen typique, le modèle T111, allant de 0 à 50 V 5 A est montré dans notre gravure. Les avantages qu'il comporte comprennent la protection contre la surcharge accidentelle, quatre réseaux de sortie terminale pour éliminer la résistance des câbles d'interconnexion, et la protection de relais pour éviter les surtensions durant les mises en circuit. Tous les sous-assemblages peuvent être préréglés pour fonctionner à n'importe quelle tension au courant nominal maximum.

EE 12 756 *pour plus amples renseignements*

SIEMENS EDISON SWAN LTD

155 Charing Cross Road, London, W.C.2

CONTRÔLEUR β

(Illustration à la page 501)

Ce simple appareil de mesure assure une indication rapide et directe du gain en courant et du courant de fuite collecteur de transistors PNP dans des conditions ordinaires d'émission.

Le gain en courant (β ou α'), dans une gamme de 10 à 150, peut être lu directement sur un cadran étalonné par réglage jusqu'à ce qu'une note audible cesse. Cette mesure peut être effectuée à un courant collecteur de 0,5 à 4 mA, le courant collecteur calé étant lu sur un compteur.

Le courant de fuite est mesuré sur le compteur à une tension collectrice fixe de 9 volts. Un commutateur à bouton-poussoir modifie la gamme du compteur à 500 μA de déviation totale pour cet essai.

La précision de mesure est de ±5% ou ±5mA, selon celle qui est la plus grande.

L'appareil de mesure est un élément transistorisé fonctionnant sur batterie et logé dans un boîtier coulé en coquille. Le transistor qu'on examine est relié au moyen de bornes à déclenchement rapide, et les deux mesures peuvent être effectuées par un personnel non-entraîné en quelques secondes.

EE 12 757 *pour plus amples renseignements*

TELEFUNKEN G.m.b.H.

Ulm/Donau, Germany

CONTRÔLEUR DE TRANSISTORS

(Illustration à la page 501)

Le Contrôleur de transistors "TELETRANS 1" est un appareil de contrôle universel pour transistors NPN et PNP. Il peut effectuer cinq essais statiques et sept mesures dynamiques à 1 kHz pour différents points de fonctionnement réglables, toutes les mesures dynamiques étant effectuées dans le circuit émetteur.

Le Contrôleur comporte essentiellement un circuit à pont, un générateur de 1 kHz, un indicateur de zéro (amplificateur et élément de mesure) et un bloc d'alimentation secteur.

Tous les paramètres h et y peuvent être mesurés comme quantités caractéristiques de courant alternatif dans le circuit émetteur. Les quantités caractéristiques sont: la résistance d'entrée (sortie de court-circuit), l'amplification de courant, la conductance de transfert (entrée ouverte), la résistance de sortie (entrée court-circuitée), le facteur de transmission de la tension inverse et la conductance de transfert inverse, trois gammes de mesure ayant été prévues pour chacun de ces paramètres.

Les valeurs de mesures individuelles peuvent être déterminées avec précision au moyen de résistances étalonnées dans le pont de mesure. De plus, les valeurs de fonctionnement peuvent être déterminées, selon les besoins, par les paramètres mesurés de court-circuit et circuit-ouvert. Ceci s'effectue en utilisant de simples formules de conversion données dans les instructions d'emploi.

Les points de fonctionnement du transistor devant être mesuré peuvent être réglés continûment entre 0 et 30 volts de tension collectrice et 0 et 5 mA de courant émetteur. Les instruments indicateurs incorporés à l'appareil sont tous munis de deux gammes de mesure, la commutation s'effectuant en pressant un bouton sur l'appareil ce qui permet d'obtenir un réglage précis des points de fonctionnement (gammes de tension de 0 à 6 volts et de 0 à 30 volts; gammes de courant de 0 à 1 mA et de 0 à 5 mA).

En plus de ces quantités caractéristiques de courant alternatif, les plus importantes mesures de contrôle de courant continu peuvent être effectuées aux jonctions collectrices des transistors. On peut ainsi contrôler le courant de coupure collecteur Ico, I'co', I ck, le courant de coupure émetteur I eo et la tension de rupture connexe dans la mesure où ils sont compris dans la gamme entre 0 et 30 volts. Enfin, la tension continue de base peut être lue sur l'indicateur de zéro pour le point de fonctionnement correspondant. Il est important de connaître la tension continue de base afin de pouvoir, par exemple, stabiliser le transistor dans le circuit.

EE 12 758 *pour plus amples renseignements*

TRANSITRON ELECTRONIC CORPORATION

168-182 Albion Street, Wakefield, Massachussetts, U.S.A.

Représentée par: G. E. Industrial Supplies Ltd., 357 Euston Road, London. N.W.1

REDRESSEURS CONTRÔLÉS AU SILICIUM

(Illustration à la page 501)

Ces redresseurs contrôlés au silicium sont des dispositifs de commutation PNPN ayant des caractéristiques de courant allant jusqu'à 10A à une température de boîtier de 100°C et des caractéristiques de tension inverse de crête allant de 50 à 400 volts.

Dans la limite de leur puissance, les redresseurs controlés Transitron peuvent être utilisés pour remplacer des thyratrons, ignitrons, commutateurs, relais, transistors de puissance, amplificateurs magnétiques et contacteurs.

Ces redresseurs sont montés sur une base hexagonale standard de 17,46 mm qui permet un montage facile en ne perçant qu'un seul trou.

EE 12 759 *pour plus amples renseignements*

VENNER ELECTRONICS LTD

Kingston-By-Pass, New Malden, Surrey

CONVERTISSEUR ARITHMÉTIQUE DE MÉGACYCLES

(Illustration à la page 502)

Ce bloc additionnel a été prévu pour l'emploi avec les fréquencemètres Venner, Type TSA3 et Type TSA53. Il permet d'effectuer des mesures jusqu'à une fréquence de 1 mégacycle avec des instruments de 100 kHz ou jusqu'à 500 kHz avec des instruments de 50 kHz. Le convertisseur peut également être employé avec des fréquencemètres d'autres marques, mais les utilisateurs devront s'assurer, avant de commander le convertisseur, qu'il s'adapte à leur appareil.

Le convertisseur comprend principalement un diviseur de mégacycles à transistors démultiplicateurs de rapport 10/1, précédé d'un "façonneur" d'impulsions de mégacycles et d'un étage suiveur d'émetteur présentant une impédance d'entrée élevée au signal. Il comporte également un atténuateur d'entrée et une alimentation secteur bien stabilisée.

Pour l'emploi de l'appareil, il faut relier sa sortie à la prise d'entrée du fréquencemètre qui doit être réglé pour la mesure de fréquence. L'atténuateur d'entrée doit être réglé à la sensibilité maximum. Le signal devant être mesuré doit alimenter la prise d'entrée coaxiale du convertisseur et l'atténuateur d'entrée du convertisseur devra être ajusté de façon appropriée.

Si on emploie un déclenchement périodique d'une seconde, l'indication donnée par le fréquencemètre sera multipliée par un facteur de di— pour donner une réponse en cycles par seconde. Le sixième chiffre significatif ne sera pas indiqué mais pourra alors être obtenu en employant un déclenchement périodique de 10 secondes. Ainsi, par exemple, une fréquence de 962 115 Hz apparaîtra comme 96 211 avec le déclenchement de 1 seconde ou comme 62 115 avec celui de 10 secondes.

EE 12 760 *pour plus amples renseignements*

WAYNE KERR LABORATORIES LTD

Roebuck Road, Chessington, Surrey

ADAPTATEURS DE TRANSISTORS

(Illustration à la page 502)

La série Q601 d'adaptateurs de transistors a été prévue pour mesurer les paramètres de transistors dans des conditions de petits signaux dans une gamme de fréquences de 15 kHz à 5 MHz. Ces adaptateurs sont employés avec le pont haute fréquence Wayne Kerr type B601 ainsi qu'avec une source de signaux et un détecteur appropriés.

Deux séries d'adaptateurs ont été conçues, soit une série pour les transistors PNP et l'autre pour les transistors NPN. Les paramètres pouvant être mesurés sont comme suit:

Y_{11}', l'admission d'entrée dans la configuration d'émetteur à la masse, avec la sortie court-circuitée.

Y_{11}', l'admission d'entrée dans la configuration de base à la masse, avec la sortie court-circuitée.

Y_{22}, l'admission de sortie dans la configuration de base à la masse, avec l'entrée court-circuitée. a, le gain en courant dans la configuration de base à la masse, avec la sortie court-circuitée.

Y_{21}, l'admission de transfert dans la configuration d'émetteur à la masse, avec la sortie court-circuitée.

EE 12 761 *pour plus amples renseignements*

WESTINGHOUSE BRAKE & SIGNAL CO. LTD

82 York Way, King's Cross, London N.1

DISPOSITIFS DE COMMUTATION

(Illustration à la page 502)

Le commutateur Dynistor est un dispositif au germanium pnpn multicouches ayant une caractéristique de tension de commutation qui lui permet de passer d'un état de haute résistance (arrêt) à un état de faible résistance (conductrice) dans une fraction de microseconde. La durée de commutation est considérablement plus réduite que celle de transistors de commutation d'une puissance comparable.

La commutation peut être effectuée par une impulsion de déclenchement de courte durée, d'une force suffisante pour élever la tension du Dynistor au-dessus du niveau de commutation. Cette impulsion de déclenchement peut être d'une durée inférieure à une microseconde.

Durant le fonctionnement, la tension de circuit traversant le Dynistor est maintenue au-dessous de la tension minimum de sa catégorie, et on emploie une impulsion de déclenchement d'une force suffisante pour élever la tension au-dessus de la tension maximum de cette catégorie. Par exemple, le type DG 11 doit être employé à une tension de fonctionnement de moins de 50 volts et une impulsion de déclenchement suffisante pour élever la tension au-dessus de 75 volts serait requise.

Pour mettre le Dynistor hors circuit, il faut baisser le courant au-dessous d'une valeur minimum appelée courant d'entretien, qui est de l'ordre de 20 à 50 mA.

Si la polarité de l'alimentation est inversée, le Dynistor présentera toujours une faible résistance, due à l'absence de caractéristiques de commutation dans cette direction par suite de la structure pnpn.

Le Trinistor Westinghouse est un commutateur multi-jonctions au silicium, à trois bornes, dont la tension de commutation peut être commandée par le courant de base. Du point de vue électrique, il est semblable à une diode redresseuse au silicium de grande puissance à laquelle on a ajouté une troisième électrode (de commande). Dans la direction avant (conductrice) et en l'absence de signal d'entrée vers la borne de base (commande), le Trinistor arrêtera la tension jusqu'à sa pleine valeur nominale et ne commutera dans un état conducteur que si la tension de commutation nominale est dépassée.

Si la base est rendue positive par rapport à la cathode, la tension de commutation sera progressivement réduite avec l'entraînement basique croissant et tombera à quelques volts avec un entraînement de base d'environ 2 à 5 volts pour 25 à 150 mA. Le plein courant avant pourra circuler alors, avec une chute de tension avant similaire à celle d'une diode au silicium normale. Toutefois, dès que le courant avant a commencé à circuler, l'entraînement de la base pourra être interrompu, si l'on veut, puisque (en dehors de certaines circonstances spéciales), la base n'a plus aucun effet sur les caractéristiques.

Pour mettre le Trinistor hors de circuit, il est en général nécessaire de réduire le courant à un niveau inférieur à une certaine valeur minimum appelée courant d'entretien, quoiqu'on puisse effectuer la mise hors de circuit à de faibles courants par un signal de commande de la base.

Dans la direction inverse, la caractéristique est analogue à celle d'une diode de silicium normale, pouvant arrêter la tension jusqu'à sa pleine valeur nominale. On ne peut effectuer de commutation ou de commande dans cette direction, et la tension inverse nominale ne devra pas être dépassée.

Une gamme de transistors npn au silicium et à fusibles, de grande puissance, est également livrable. Ces transistors sont prévus pour la commutation de grande puissance et les applications amplificatrices. Ils se caractérisent par une très basse résistance de saturation ce qui, ajouté aux valeurs existantes de tension et de courant, permet d'effectuer des commutations de grande puissance avec un minimum de pertes. Enfin, ils sont enfermés dans des étuis en acier inoxydable pour le montage sur une source froide extérieure au moyen d'un goujon de cuivre, et s'obtiennent en modèles à 2A et 5A avec V_{ce} (max.) jusqu'à 300 volts.

EE 12 762 *pour plus amples renseignements*

Résumés des Principaux Articles

Un récepteur d'intérieur à modulation de fréquence utilisant des transistors mésa à base diffuse par M. Applebaum et E. Midgley

Résumé de l'article aux pages 448 à 453

Cet article traite brièvement des caractéristiques du transistor mésa à base diffuse et décrit son emploi dans les étages haute fréquence et moyenne fréquence d'un récepteur d'intérieur à modulation de fréquence. Les auteurs donnent aussi des détails du rendement du récepteur complet.

Un dispositif analogique pour les problèmes de flux de chaleur dans les semi-conducteurs par N. L. Potter

Résumé de l'article aux pages 454 à 457

Dans les problèmes de protection des semi-conducteurs, il est nécessaire de pouvoir évaluer la température de leurs jonctions.

Cet article montre la similitude entre le flux de courant dans un réseau résistance-capacité et *indique, partant, la théorie servant de base à la construction d'un dispositif analogique pour l'observation de la température des jonctions.*

La gravure montre un "analogue" type, avec les images oscillographiques qu'on en obtient.

L'auteur suggère d'autres usages de l'appareil et envisage la possibilité de travaux à trois dimensions.

Un filtre passe-bande basse fréquence à plusieurs gammes par J. B. Bratt

Résumé de l'article aux pages 458 à 462

Le filtre passe-bande décrit dans cet article a été conçu comme un élément de circuit de contrôle d'amplitude utilisé dans des appareils prévus pour les recherches dans le domaine de l'aérodynamique instable. La gamme de fréquences du filtre va de 16 à 175 Hz et s'étend sur neuf bandes, le rapport entre la fréquence supérieure et inférieure de chaque passe-bande étant de 1,38. La base du circuit du filtre est un amplificateur à lampes à sélection de fréquence, modifié pour donner une sortie zéro aux fréquences 0 et 100. Avec deux de ces amplificateurs en cascade, la réponse dans le passe-bande est horizontale jusqu'à ±5% et l'atténuation du troisième harmonique de la limite de fréquence inférieure est de 33 fois. L'emploi d'un réseau à rétroaction spécial dans les amplificateurs permet d'effectuer un changement de bande passante en commutant deux condensateurs dans chaque réseau.

Un amplificateur logarithmique de tension par R. F. Mathams

Résumé de l'article aux pages 463 à 465

Cet article décrit un circuit qui reçoit une tension d'entrée d'une source d'énergie et produit une sortie proportionnelle au logarithme de la tension d'entrée sur une gamme d'entrées de 0,1 V à 100 V ou davantage. La tension d'entrée est interrompue par un relais vibrant et differenciée. Le temps mis par les impulsions exponentielles qui en résultent pour décroître à une tension de référence est proportionnel au logarithme de la tension d'entrée.

Une montre électronique a modulateur pour radiophares à occultations codées par J. W. Nichols, A. C. MacKellar et A. J. B. Baty

Résumé de l'article aux pages 466 à 474

Cet article décrit une montre pilotée par quartz avec modulateur électronique produisant des signaux codés pour radiophares de marine, conformément aux prescriptions des Accords Internationaux de Paris, et réglant également, avec un haut degré de précision, le moment de l'émission. Elle peut être commandée à la main ou à distance et elle est munie d'un système avertisseur de défauts pour le contrôle du minutage et du codage.

L'étude d'alimentations c.c. stabilisées par triodes à gaz par B. G. Higdon et M. E. Bond

Résumé de l'article aux pages 475 à 479

Le redresseur à gaz, à grilles commandées, est, en général préférable au stabilisateur du genre à tube à vide, pour l'alimentation h.t. stabilisée à un courant supérieur à environ 0,5A c.c.

Les auteurs de cet article traitent en détail du mode de construction des stabilisateurs du type à triodes à gaz et des pièces qui en font partie. Ils décrivent un stabilisateur de conception simple, ainsi qu'un modèle plus compliqué, incorporant un amplificateur à réaction, et ils concluent par des données de performances.

Oscillateur surcouplé à transistor, à l'usage des systèmes arithmétiques par Agnes A. Kaposi

Résumé de l'article aux pages 480 à 484

Un transformateur saturable est utilisé dans un circuit à oscillateur surcouplé. L'auteur démontre que la durée de l'impulsion ne dépend pas des paramètres du transistor, de la température et du courant de charge. Des circuits pratiques ont été réalisés dont la durée d'impulsion s'étend à partir de 1,5 microsecondes. Ces circuits conviennent particulièrement au fonctionnement des compteurs à noyaux magnétiques.

La théorie des circuits couplés à large bande par W. Dougharty

Résumé de l'article aux pages 488 à 494

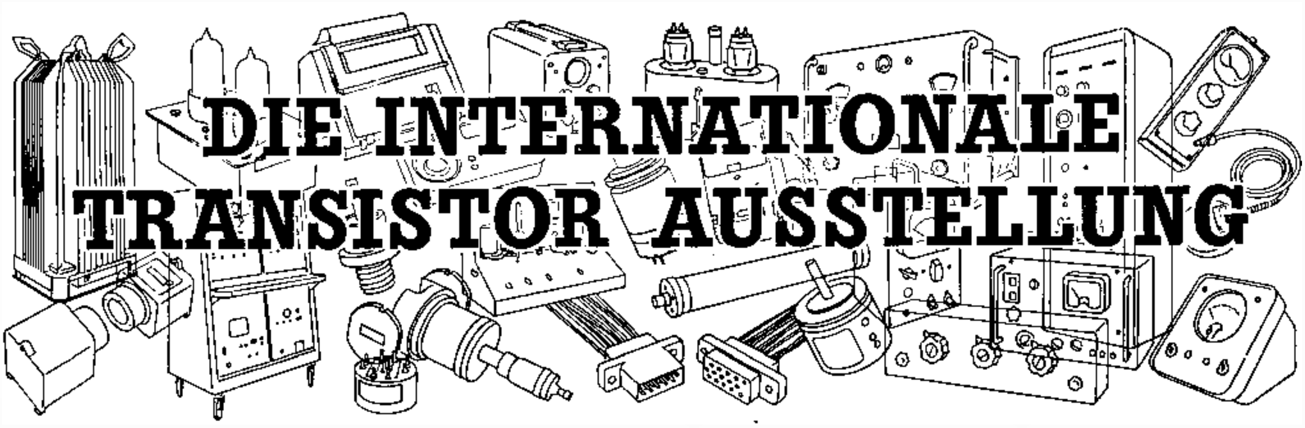
Cet article donne les formules exactes d'admission de transfert pour divers types de circuits couplés à accord parallèle. Il n'y a pas d'indications approximatives, pas plus que des limites à la largeur de bande. De ces formules découlent des expressions pour des caractéristiques telles que la séparation de crête, le rapport de crête à creux, etc. qui sont suffisamment précises pour des largeurs de bande approchant la valeur de la fréquence nominale.

L'auteur indique également des formules de réponse de phase pour divers types de couplages.

Il démontre qu'avec un certain degré de couplage inductif, on obtient un circuit ayant non seulement une bonne réponse de phase mais aussi une courbe de réponse indiquant un degré remarquable de symétrie avec la déviation de fréquence.

L'article ne traite que des cas de circuits synchrones et à facteurs de dissipation égaux.

DIE INTERNATIONALE TRANSISTOR AUSSTELLUNG



Beschreibung einer weiteren Auswahl der auf der Internationalen Transistor Ausstellung gezeigten Geräte, die auf Grund der von Herstellern gemachten Angaben zusammengestellt wurde. *(Übersetzung der Seiten 500 bis 503)*

DEPENDABLE RELAY CO. LTD

8a Ainger Road, London, N.W.3

MINIATUR-SCHNELLZÄHLGERÄT

(Abbildung Seite 500)

In Normalausführung zählt dieser Miniatur-Schnellzähler bis zu 1500 Impulse/min. Er ist komplett im Gehäuse mit Anschluss für eine Fotozelle oder höchstempfindliche Fühler, wie z.B. Drehspulrelais. Durch einen Schalter kann der Betrieb auf die Fotozelle, normalerweise offene, oder normalerweise geschlossene Kontakte umgestellt werden. Die erforderliche Kontakt-Höchstbelastung ist nur 500 μA. Das für Wechselstrom oder Batteriebetrieb lieferbare Gerät hat Aussenabmessungen von 127 × 102 × 64 mm. Auf Wunsch auch als Abzählgerät und mit elektrischer Nullstellung, sowie als Sonderausführung für Zählgeschwindigkeiten bis zu 3000 Impulse/min lieferbar.

EE 12 751 *für weitere Einzelheiten*

FERRANTI LTD

Hollinwood, Manchester, Lancashire

SILIZIUM TRANSISTOREN

Allgemeine Entwicklungsarbeiten an Silizium-Transistoren mit Schaltgeschwindigkeiten zwischen 50 und 100 MHz sind im Gange, wobei Festkörper-Diffusionsmethoden für die Fertigung Anwendung finden. Sowohl die Basis, als auch der Emitter werden durch Diffusion in ein Siliziumplättchen hergestellt, das selbst als Kollektorelektrode dient. Im Augenblick werden diese Transistoren nur in kleinen Mengen hergestellt. Die Wythenshawe Laboratorien arbeiten auch an legierten Flächentransistoren für Frequenzen bis 10 MHz, die besonders für hohe Schaltgeschwindigkeiten geeignet sind. Voraussichtlich werden beide Typen Ende des Jahres aus laufender Produktion lieferbar sein. Ferranti Ltd hat bekannt gegeben, dass sie sich auf Legier- und Diffusions-Silizium-Transistoren für militärische und industrielle Zwecke konzentrieren wird, für die Germanium-Transistoren der hohen Temperatur wegen nicht in Frage kommen.

Entwicklungsarbeiten an parametrischen Dioden für HF Verwendung werden durchgeführt. Besonders gute Fortschritte wurden für Dioden mit vier Schichten erzielt, die schon in Pilotenproduktion sind. Diese pnpn-Diode ist als Schaltvorrichtung für Rechner und Impulsschaltungen gedacht. Im Grunde genommen ist es eine Halbleiter-Ausführung, die der Gasentladungsröhre entspricht. Die Zündspannungen liegen zwischen 80 und 120 V und die „an"-Impedanz beträgt wenige Ohm. Der Betriebsstrombereich liegt zwischen 1 und 50 mA. Die Abmessungen sind fast dieselben wie die der ZS 10 Diodenreihe.

EE 12 752 *für weitere Einzelheiten*

SILIZIUM-SONNEN-ZELLEN

Eine andere interessante Vorrichtung ist die nach dem Diffusionsprinzip hergestellte fotoelektrische Silizium-Sonnen-Zelle, die die Energie der Sonne oder einer elektrischen Glühlampe in elektrische Energie umwandelt. Die Ansprechzeit beträgt wenige μsec, und die Zelle kann ohne Beeinträchtigung in hoher Temperatur und Feuchtigkeit arbeiten.

Unter diesen Bedingungen wird das auf die Silizium-Fotozelle fallende Licht eine Temperaturerhöhung in der Zelle hervorrufen; glücklicherweise arbeitet die Siliziumzelle bei hohen Temperaturen ohne Schaden und mit sehr geringer Beeinträchtigung der Ausgangsspannung. Belastung der Zelle durch einen Aussenwiderstand ruft eine geringe Temperatursenkung durch Energieübertragung hervor. Ein Wirkungsgrad von über 11% kann für die Leistungsumwandlung mit diesen Zellen erreicht werden.

Die Zellen können in verschiedenen Formen und Grössen geliefert werden, haben jedoch eine Höchstfläche von 12,9 cm² für die Einzelzelle. Sie können in Serien-, Parallel-, oder Serien-Parallelschaltung Verwendung finden, um höhere Spannungen und Stromstärken zu erreichen.

Wesentliche Stromstärken können von Sonnenzellen abgegeben werden. Weisses Licht mittelmässiger Stärke kann z.B. auf einer Fläche von 6 cm² einen 20 mA Strom hervorrufen.

EE 12 753 *für weitere Einzelheiten*

RADIO HEATERS LTD

Eastheath Avenue, Wokingham, Berkshire

SILIZIUM-VEREDELUNGSANLAGE

(Abbildung Seite 500)

Die Radyne Silizium-Veredelungsanlage mit „gleitender Zone" wurde besonders entwickelt, um den Forderungen der Halbleiterindustrie gerecht zu werden.

Die Anlage kann Siliziumstangen von 25,5 bis 63,5 mm Durchmesser und 150 bis 560 mm Länge verarbeiten. Im Arbeitsablauf wird die Spule über die sich drehende Siliziumstange vorgeschoben. Zur Anlage gehören auch alle notwendigen Vakuumvorrichtungen und Steuerorgane. Sie ist für Betrieb mit Radyne HF-Generatoren für 5, 10 oder 15 kW konstruiert, die auf 450 kHz, 3 oder 4 MHz arbeiten.

Die Quartzröhre der Veredelungseinheit wird zwischen zwei schweren Armen aus Aluminiumguss gehalten und liegt vakuumdicht im oberen und unteren Arm. Verwandt werden O-Ringe zwischen verschraubten Flanschen, abgeänderte Wilson-Dichtungen für die sich drehenden Spindeln und eine Radyne Neopren-Klemmdichtung zum Abdichten der Quartzröhre. Quartzröhren mit Innendurchmessern von 20 bis 40 mm können Verwendung finden. Die in einstellbaren Futtern gehaltene Siliziumstange läuft mit 2 und 60 U/min, wobei der Antrieb gleichzeitig durch beide Futter, oder unabhängig durch eines der beiden erfolgen kann. Die Umdrehungsgeschwindigkeit ist kontinuierlich einstellbar und wird durch einen Servoverstärker geregelt. Die veränderliche Höhe des oberen Armes ermöglicht das Einsetzen von Quartzröhren mit 203 bis 610 mm Länge. Die Dicke der Veredelungszone kann während des Arbeitsablaufes geändert werden. Das untere Futter hat eine einfache Spannzangenvorrichtung, um verschieden lange Siliziumstangen aufnehmen zu können.

EE 12 754 *für weitere Einzelheiten*

RIVLIN INSTRUMENTS LTD
Doman Road, Camberley, Surrey
TRANSISTOR-TESTER
(*Abbildung Seite 501*)

Das Gerät zeigt Kenngrössenabweichungen und das Versagen von pnp Flächentransistoren an. Folgende Messungen können für grosse Strom- und Spannungsbereiche durchgefürt werden: Stromverstärkung in Emitterschaltung, Kollektorreststrom in offener Schaltung und Kollektor-Durchbruchsspannung.

Ablesungen erfolgen auf einem 114 mm Messgerät mit Messerzeiger und Spiegel.

Niederspannungen für den Transistor-Prüfling werden mit Hilfe eines 64 mm Voltmeters überwacht.

Die Kollektorspannung kann von 0-20 V, der Kollektorstrom von 0-2,5 A, Stromverstärkung von 0-100 und der Kollektorreststrom von 0-2,5 A in sechs Bereichen gemessen werden. Das Messen der Durchbruchsspannung hängt von der Höchstverlustleistung des Transistor-Prüflings ab, aber drei 2,5, 25 und 100 mV Bereiche ergeben Spannungen bis zu 85 V.

Die Messgenauigkeit liegt innerhalb ±5% für Stromverstärkung und Durchbruchsspannung, innerhalb ±2% für Reststrom.

Niederspannungen für den Transistor sind in fünf Stufen von 0,4 bis 19 V, mit Feinregulierung für die 0,4 V Stellung, variabel.

Alle Bedienungsknöpfe und -klemmen für den Transistor-Prüfling sind auf der Frontplatte angebracht.

EE 12 755 *für weitere Einzelheiten*

ROBAND ELECTRONICS LTD
33 Mountgrove Road, Highbury, London N.5
(*Abbildung Seite 501*)
STABILISIERTE NETZGERÄTE

Eine komplette Serie transistorenstabilisierter Netzgeräte als Einbauchassis von 25 V max bei 1 A bis zu variablen Leistungen von 0-30 V bei 2 A wurde ausgestellt. Die Abbildung zeigt ales typisches Beispiel den T 111, der bei 0-50V 5A abgibt. Unter anderem werden folgende Vorteile genannt: Ausreichender Schutz gegen zufällige Überlastung, ein Ausgangsvierpol, um den Widerstand der Verbindungsschnüre auszuschalten, und ein Schutzrelais zur Vermeidung von Stromstössen während des Einschaltens. Alle Bausteine können für jede Spannung bis zur Spitzennennspannung eingestellt werden.

EE 12 756 *für weitere Einzelheiten*

SIEMENS EDISON SWAN LTD
155 Charing Cross Road, London W.C.2
BETA-TESTER
(*Abbildung Seite 501*)

Dieses einfache Prüfgerät gibt eine schnelle und direkte Anzeige für Stromverstärkung und Kollektor-Reststrom von pnp Transistoren in Emitter-Schaltung.

β und α' Stromverstärkungen von 10 bis 150 können von einer geeichten Skala abgelesen werden, die eingeregelt wird, bis ein abgegebener Ton gerade nicht mehr hörbar ist. Diese Messungen können für Kollektorstromstärken von 0,5 bis 4mA durchgeführt werden, wobei der Kollektorstromwert auf einem Messer abgelesen werden kann.

Der Reststrom wird für eine feststehende 9V Kollektorspannung gemessen und am Messer abgelesen. Eine Drucktaste ändert den Messbereich zu 500 µA, aber nur für diese Prüfung.

Die Messgenauigkeit ist ± 5% oder ± 5, wobei immer die grössere Toleranz gültig ist.

Das batteriegespeiste Gerät ist mit Transistoren bestückt und wird in einem robusten Spritzgussgehäuse geliefert. Der zu messende Transistor wird mittels schnelltrennender Klemmen angeschlossen, und beide Prüfungen können von ungelerntem Personal innerhalb weniger Sekunden ausgeführt werden.

EE 12 757 *für weitere Einzelheiten*

TELEFUNKEN G.m.b.H.
Ulm/Donau, Germany
TRANSISTOR-TESTER
(*Abbildung Seite 501*)

Der als „Teletrons I" bekannte Transistor Tester ist ein Mehrzweck-Messgerät für npn und pnp Transistoren. Fünf statische Tests werden vorgenommen und sieben dynamische Messungen an verschiedenen, einstellbaren Arbeitspunkten bei 1kHz ausgeführt; alle dynamischen Messungen erfolgen in Emitterschaltung.

Der Tester besteht hauptsächlich aus einer Brückenschaltung, einem 1kHz Oszillator, einem Nullpunktanzeiger (Verstärker und Messer) und einem Netzgerät.

Alle „h" und „y" Parameter werden als Wechselstrom-Kenndaten in Emitterschaltung gemessen. Die Kenndaten sind: Eingangswiderstand (Ausgang kurzgeschlossen), Stromverstärkung, Steilheit, Ausgangswiderstand (Eingang kurzgeschlossen), Spannungsrückwirkung und Rückwirkungsleitwert. Für jeden dieser Parameter sind drei Messbereiche vorgesehen.

Mit Hilfe geeichter Widerstände in der Messbrücke ist eine genaue Bestimmung der einzelnen Messwerte möglich Weiterhin können mit Hilfe der gemessenen „Kurzschluss" und „offenen" Parameter Arbeitspunkte bestimmt werden, falls es wünschenswert ist. Das geschieht durch Anwendung einfacher Umwandlungsformeln, die in der Gebrauchsanweisung enthalten sind.

Die Arbeitspunkte für den Transistor-Prüfling können kontinuierlich eingestellt werden und zwar zwischen 0 und 30 V für die Kollektorspannung und 0 bis 5mA für den Emitterstrom. Jedes der einzelnen Anzeigegeräte hat zwei Messbereiche, die durch Betätigung einer Drucktaste am Instrument umgeschaltet werden können und die genaue Einstellung der Arbeitspunkte ermöglichen. (Spannungsbereiche 0 - 6V und 0 - 30V, Strombereiche 0 - 1mA und 0-5 mA.)

Ausser diesen Wechselstrom-Kenndaten können auch die wichtigsten Gleichstromprüfungen an Flächentransistoren vorgenommen werden. Dazu gehören Messungen des Kollektorreststromes I_{co}, I_{co}', I_{ck} und des Emitterreststromes I_{eo} und der damit zusammenhängenden Durchschlagsspanungen, soweit sie im Bereich 0-30V liegen. Schliesslich kann die Basisgleichspannung für den betreffenden Arbeitspunkt am Nullanzeiger abgelesen werden. Kenntnis der Basisgleichspannung ist z.B. für die Stabilisierung des Transistors im Schaltkreis wichtig.

EE 12 758 *für weitere Einzelheiten*

TRANSITRON ELECTRONIC CORPORATION
168-182 Albion Street, Wakefield, Massachusetts, U.S.A. Represented by: G.E. Industrial Supplies Ltd, 357 Euston Road, London N.W.1
GESTEUERTE SILIZIUM-GLEICHRICHTER ...
(*Abbildung Seite 501*)

Diese gesteuerten Silizium-Gleichrichter sind pnpn Leistungsschaltelemente und haben Nennstromwerte bis zu 10A bei einer Gehäusetemperatur von 100°C und Spitzensperrspannungen zwischen 50 und 400V.

Gesteuerte Transitron Gleichrichter können innerhalb ihrer Nennleistung Thyratrons, Ignitrons, Schalter, Relais, Leistungstransistoren, magnetische Verstärker und Kontaktgeber ersetzen.

Die Gleichrichter werden mit einer genormten, sechseckigen Basis geliefert, die die Montage erleichtert, da nur ein Loch gebohrt zu werden braucht.

EE 12 759 *für weitere Einzelheiten*

VENNER ELECTRONICS LTD
Kingston-By-Pass, New Malden, Surrey
MHz-DIGITALWANDLER
(*Abbildung Seite 502*)

Dieser Zusatzbaustein wurde für Verwendung mit den Frequenzmetern TSA 3 und TSA 53 entwickelt. Mit seiner Hilfe können Messungen bis 1 MHz auf 100 kHz-Geräten durchgeführt werden. Der Wandler kann mit Frequenzmetern anderer Hersteller eingesetzt werden, vor der Bestellung muss aber bestätigt werden, dass er für das vorhandene Gerät geeignet ist.

Der Wandler besteht hauptsächlich aus einem MHz-Transistor-Untersetzer Masstab 10 : 1, dem ein MHz-Impulsformer und eine Emitter-Verstärkerstufe vorausgehen, die für das Signal einen hohen Eingangsscheinwiderstand bilden. Ein Eingangsdämpfungsglied und ein gut stabilisiertes Netzgerät sind eingebaut.

In Gebrauch wird der Wandlerausgang mit dem Frequenzmetereingang verbunden und das Frequenzmeter wie üblich für Frequenzmessung eingestellt. Der Eingangsdämpfer wird auf höchste Empfindlichkeit gestellt. Das zu messende Signal wird der koaxialen Eingangsbuchse des Wandlers zugeleitet und der Eingangsdämpfer des Wandlers entsprechend angepasst.

Um die Antwort bei 1 sec Austastzeit in kHz zu erhalten, muss die vom Frequenzmeter angezeigte Zahl mit zehn multipliziert werden. Die sechste geltende Zahl wird nicht angezeigt, kann aber durch Umschalten der Austastzeit zu 10 sec sichtbar gemacht werden. Die Frequenz 962115 Hz würde z.B. mit 1 sec Austastzeit als 96211 und mit 10 sec Austastzeit als 62115 erscheinen.

EE 12 760 *für weitere Einzelheiten*

WAYNE KERR LABORATORIES LTD

Roebuck Road, Chessington, Surrey

TRANSISTOR MESSZUSATZ

(*Abbildung Seite 502*)

Transistor Messzusätze der Reihen Q 601 sind zum Messen von Transistorparametern mit kleinen Signalen über den Frequenzbereich 15kHz bis 5MHz bestimmt. Sie werden mit der Wayne Kerr HF-Brücke B601, einer geeigneten Signalquelle und einem Anzeigegerät eingesetzt.

Zwei Zusatzreihen wurden entwickelt: eine für pnp Transistoren, die andere für npn Transistoren. Folgende Parameter können gemessen werden:

Y_{11}', der Eingangsleitwert in Emitter-Schaltung mit kurzgeschlossenem Ausgang.

Y_{11}, der Eingangsleitwert in Basis-Schaltung mit kurzgeschlossenem Ausgang.

Y_{22}, der Ausgangsleitwert in Basis-Schaltung mit kurzgeschlossenem Eingang

α, die Stromverstärkung in Basis-Schaltung mit kurzgeschlossenem Ausgang.

Y_{21}', Steilheit in Basis-Schaltung mit kurzgeschlossenem Eingang.

EE 12 761 *für weitere Einzelheiten*

WESTINGHOUSE BRAKE & SIGNAL CO. LTD

82 York Way, King's Cross, London N.1

SCHALTELEMENTE

(*Abbildung Seite 502*)

Der Dynistor Schalter ist ein mehrschichtiges pnpn Germaniumelement mit einer Überschlagspannungs - Kennlinie, die es ermöglicht, innerhalb eines Bruchteiles einer Mikrosekunde vom hochohmigen Sperrzustand zum niederohmigen Durchlasszustand umzuschalten. Die Schaltzeit ist bedeutend kürzer, als die mit Schalttransistoren gleicher Leistung erzielbare.

Der Überschlag kann durch einen kurzen Triggerimpuls eingeleitet werden, der ausreichend genug bemessen sein muss, um die Dynistor Spannung über den Überschlagpegel zu bringen. Dieser Triggerimpuls kann kürzer als 1 Mikrosekunde sein.

In Praxis wird die in der Schaltung an den Dynistor angelegte Spannung kleiner gehalten, als die für diese Type angegebene untere Spannung, und der Triggerimpuls muss gross genug sein, um die Spannung über die obere Spannung für diese Type zu bringen Sollte z.B. Type DG 11 mit einer Betriebsspannung unter 50V benutzt werden, so muss der Triggerimpuls genügend gross sein, um die Spannung auf über 75V zu bringen.

Der Dynistor wird durch Herabsetzung des Stromes abgeschaltet. Der Strom muss genügend reduziert werden, um unter dem als Dauerstrom bekannten Mindestwert zu liegen, dessen Grössenordnung zwischen 20 und 50 mA liegt. Bei Umpolung der Spannung liefert der Dynistor immer einen sehr geringen Widerstand, da auf Grund der pnpn Ausbildung in dieser Richtung keine Überschlagcharakteristik besteht.

Der Westinghouse Trinsistor ist ein Silizium-Mehrflächen-npnp-Schalter mit drei Anschlüssen, dessen Überschlagspannung durch den Basisstrom gesteuert werden kann. Elektrisch ist er einem Silizium-Hochleistungsgleichrichter ähnlich, dem eine dritte (Steuer-) Elektrode zugefügt wurde. Ohne Eingangssignal am Basis-(Steuer)-Anschluss wird der Transistor in Durchlassrichtung bis zur vollen Nennspannung sperren und nur überschlagen, wenn die Überschlagspannung überschritten wird.

Wenn die Basis gegenüber der Kathode positiv gemacht wird, wird die Überschlagspannung fortschreitend herabgesetzt. Wenn die Basis mit 2-5V bei 25-150 mA getrieben wird, fällt sie auf wenige Volt. In diesem Zustand kann der volle Durchflussstrom fliessen, und der Durchflussspannungsabfall ist dann dem einer üblichen Siliziumdiode ähnlich. Wenn es wünschenswert erscheint, kann der Basistrieb abgeschaltet werden, sobald der Durchflussstrom fliesst, da (ausser unter gewissen besonderen Umständen) die Basis keinen weiteren Einfluss auf die Kenndaten hat.

Der Trinistor wird durch Herabsetzung des Stromes unter die als Dauerstrom bekannte Grösse abgeschaltet. Es ist jedoch auch möglich, die Abschaltung bei geringen Stromstärken durch ein Basis-Steuersignal vorzunehmen.

In der umgekehrten Richtung ist die Kennlinie ähnlich der einer üblichen Silizium-Diode mit Sperrung bis zur vollen Nennspannung. In dieser Richtung sind weder Überschlag noch Steuerung möglich, und die Spitzensperrspannung darf nicht überschritten werden.

Eine Reihe von Silizium-Hochleistungs-npnp-Transistoren ist auch lieferbar, Diese Typen sind für Hochleistungsschalten und Verstärkung geracht Ein Merkmal dieser Transistoren ist der sehr niedrige Grenzwiderstand, der, zusammen mit den lieferbaren Spannungs- und Stromnennwerten, Hochleistungsschalten mit geringsten Verlusten ermöglicht. Der Transistor wird in einem geschweissten, rostfreien Gehäuse mit Kupferbolzen zur Montage auf Kühlflächen geliefert.

In Typen für 2 und 5A mit U_{ce} max bis zu 300 V lieferbar.

EE 12 762 *für weitere Einzelheiten*

Zusammenfassung der wichtigsten Beiträge

FM-Rundfunkempfänger mit Diffusions-Mesa-Transistoren von M. Applebaum und E. Midgley

Zusammenfassung des Beitrages auf Seite 448-453

Dieser Beitrag behandelt kurz die Kenngrössen der Diffusions-Mesa-Transistoren und beschreibt deren Verwendung in den HF- und ZF- Stufen eines FM-Rundfunkempfängers. Leistungsangaben für den Empfänger werden gegeben.

Ein elektrisches Analog für den Wärmefluss in Halbleitern von N. L. Potter

Zusammenfassung des Beitrages auf Seite 454-457

Die Kristalltemperatur eines Halbleiters muss oft geschätzt werden, wenn Schutzfragen behandelt werden.

Die Ähnlichkeit des Wärmeflusses und des Stromflusses in einem RC Netzwerk werden in diesem Beitrag gezeigt und davon die Theorie für die Konstruktion eines Analogs zur Untersuchung der Kristalltemperatur abgeleitet.

Ein typisches Analog wird zusammen mit den erzielten Sichtanzeigen beschrieben.

Anwendung dieses Analogs für andere Zwecke wird vorgeschlagen und die Möglichkeit dreidimensionaler Untersuchungen diskutiert.

Ein NF Bandfilter für mehrere Frequenzbereiche von J. B. Bratt

Zusammenfassung des Beitrages auf Seite 458-462

Das beschriebene Bandfilter wurde für eine Amplitudensteuerschaltung entwickelt, die in Geräten für Forschung auf dem Gebiete der instabilen Aerodynamik Verwendung findet. Der Frequenzbereich von 16 Hz—175 Hz ist in neun Bereiche unterteilt, wobei für jeden Durchlassbereich das Verhältnis der oberen zur unteren Frequenz 1,38 ist. Die Basis für die Filterschaltung bildet ein abgeänderter frequenzselektiver RC Röhrenverstärker, der Nulleistung für die Frequenzen 0 und ∞ liefert. Zwei dieser Verstärker in Kaskade haben eine Durchlasskurve, die innerhalb ± 5% flach ist, wobei für die untere Frequenzgrenze die dritte Oberschwingung 33fach abgeschwächt ist. Ein besonderes Gegenkopplungs-Netzwerk im Verstärker ermöglicht Durchlasskurvenwahl durch Umschalten zweier Kondensatoren in jedem Netzwerk.

Ein spannungsgesteuerter logarithmischer Verstärker von R. F. Mathams

Zusammenfassung des Beitrages auf Seite 463-465

Dem Eingang eines Verstärkers werden Spannungen von einer Spannungsquelle mit einem Bereich von 0,1 bis 100 V zugeleitet, die mit Hilfe eines beschriebenen Schaltkreises als Ausgangsspannungen abgegeben werden, die zur Eingangsspannung in einem logarithmischen Verhältnis stehen. Die Eingangsspannungen werden mittels eines Unterbrecherrelais zerhackt und differenziert. Die Zeit, die der entstehende Exponentialimpuls benötigt, um auf eine Bezugsspannung abzuklingen, ist dem Logarithmus der Eingangsspannung proportional.

Elektrische Uhr mit Signalgeber für Kodefunkbaken von J. W. Nichols, A. C. MacKellar und A. J. B. Baty

Zusammenfassung des Beitrages auf Seite 466-474

Eine quartzgesteuerte Uhr mit elektronischem Signalgeber wird beschrieben, die Kodesignale für Seefunkbaken in Ubereinstimmung mit den Anforderungen des Pariser Internationalen Abkommens abgeben kann und auch die Zeiteinstellung mit hoher Genauigkeit reguliert. Fernsteuerung und Handbedienung, sowie ein Fehlermeldenetz zur Uberwachung der Zeiteinstellung und Verschlüsselung sind vorgesehen.

Die Konstruktion von Gleichstromquellen mit Thyratron-Stabilisierung von B. G. Higdon und M. E. Bond

Zusammenfassung des Beitrages auf Seite 475-479

Das Stromtor wird im allgemeinen dem seriengeschalteten Vakuumröhrenstabilisator für Hochspannungsquellen vorgezogen, denen mehr als ungefähr 0,5 A Gleichstrom entnommen wird.

In diesem Beitrag wird die Konstruktion der Thyratron-Stabilisatoren und eingebauten Bauelemente ausführlich behandelt. Sowohl ein einfacher Stabilisator, als auch ein Gerät mit Verstärker in der Gegenkopplungsschleife werden beschrieben und Einzelheiten ihrer Leistung angegeben.

Transistor-Sperrschwinger für Digitalsysteme von Agnes A. Kaposi

Zusammenfassung des Beitrages auf Seite 480-484

In einem Sperrschwinger wird ein Ubertrager eingesetzt, der gesättigt werden kann. Es wird gezeigt, dass die Impulsbreite von den Parametern der Transistoren, der Temperatur und dem Belastungsstrom unabhängig ist. In Versuchsschaltungen wurden Impulsbreiten über 1,5 μs praktisch erreicht. Die Schaltung ist besonders für den Antrieb von Zählern mit magnetischen Kernen geeignet.

Die Theorie breitbandgekoppelter Kreise von W. Dougharty

Zusammenfassung des Beitrages auf Seite 488-494

In diesem Beitrag werden genaue Formeln für den Ubertragungsleitwert der verschiedenen Kopplungsparallelschaltungen gegeben und dabei keine Annäherungen oder Bandweitenbegrenzungen zugelassen. Von diesen Formeln werden Ausdrücke für den Spitzenabstand, das Maximum-zu-Minimum Verhältnis und andere derartige Kennzeichen abgeleitet, die genügend genau für Bandbreiten sind, die dem Wert der Mittelfrequenzen naheliegen.

Formeln für die Phasenempfindlichkeit bei verschiedenen Kopplungsschaltungen werden gegeben. Es wird gezeigt, dass sich bei einer gewissen induktiven Kopplung eine Schaltung ergibt, die nicht nur einen guten Phasengang hat, sondern deren Phasenkurve auch in beträchtlichem Masse mit dem Frequenzhub symmetrisch verläuft.

Der Beitrag beschäftigt sich durchgehend mit Fällen, in denen die Kreise synchron und von gleicher Resonanzschärfe sind.