Wireless World

ELECTRONICS, RADIO, TELEVISION

DECEMBER 1964

Communications and/or Electronics

IN his inaugural address to the Electronics Division of the I.E.E. the Chairman for 1964-65, Mr. G. G. Gouriet, chose as his title: "Through communications to electronics" and in so doing he not only revived old arguments but reminded us that the meaning of the term "electronics" has changed and broadened over the last decade to describe a virile and growing branch of electrical science. Indeed, at the present time, the Electronics Division of the I.E.E. comprises no fewer than fifteen professional groups.

Like sunspots the arguments seem to reach a maximum every 11 years and it was in 1953 that we last had a vintage crop of proposals for redefining electronics. Before this electronics was confined to valve-like devices in which the common factor was conduction by electrons through a vacuum or a gas at low pressure. The advent of the transistor called for a further extension of the definition to include conduction by holes as well as electrons in semiconductors. At this point in time B. E. Noltingk (letter in *Proc. I.R.E.*, May, 1953) suggested that instead of adding to the definition we should do better to make it exclusive so that electronics covered all devices "in which electronics travel otherwise than along normal conductors." The whole of the "wireless" world in fact?

Sweeping as this proposal seemed to be it still dealt only with devices and did not embrace applications like measurement, control and computation which everyone tacitly accepted as coming within the meaning of the term. Prof. W. L. Everitt (*Proc. I.R.E.*, Aug. 1952) had taken the opposite approach and defined electronics as "the science and technology which deals primarily with the supplementing of man's senses and his brain power by devices which collect and process information, transmit it to the point needed and there either control machines or present the processed information to human beings for their direct use." This was pretty comprehensive and, as Everitt pointed out, would accommodate all future new devices. Unfortunately, it was too broad and could not be held to exclude optics and fluid mechanics. It could also be regarded as a good definition of instrumentation. Some brave attempts at less cumbersome definitions also appeared at this time, for example: "Electronics: the study and application of submolecular phenomena as elements of control" (R. A. Fairthorne, *W.W.* Sept. 1953) and the Brit I.E.R.E. definition: "Electronics is the radio valve, or kindred devices, at work in ways other than communications." A distinction between electronics and communications, incipient in some other definitions, was here made clear and positive. It is a good definition and one which ideally epitomizes the origins and growth of the Brit. I.E.R.E.; but the distinction was not drawn in the International Electrotechnical Commissions' definition (1953) which still stands as glossary item No. 11029 in British Standard 204:1960, "Electronics. That branch of science and technology which deals with the study of the phenomena of *conduction* of electricity in a vacuum, in a gas, and in semiconductors, and with the utilization of devices based on these phenomena." Communications arising from the movement of charges, is not excluded by this definition.

Mr. Gouriet in his address favoured the following definition: "Electronics. The branch of science and technology that is concerned with the communication of information by electrical means as distinct from the transmission of power." Note the use of the word "electrical" which neatly avoids identification with any particular current carriers (and also sustains the breadth of outlook of a great Institution). The definition is compact and to that extent seemed an improvement on the international standard until it came to our notice that the Americans are now using microwave power generated by giant klystrons to levitate aerospace vehicles (see p. 620 of this issue)!

So Prof. Everitt should have the last word. In 1953 he said: "Like other dynamic words 'electronics' will need to be redefined as its expansion continues."

TRANSISTOR R.F. POWER

THE advent of the planar technique for manufacturing junction transistors made it possible for high-frequency transistors to be produced with more consistent characteristics than with earlier processes.

The alloyed junction transistor is limited in high-frequency performance owing to the difficulty with which the necessarily consistent narrow base width and small base area are achieved. Modifications to this process give such devices as the micro-alloy diffused (MADT) and post alloy diffused (PADT) transistors, but these are satisfactory only for low-power applications.

The double-diffused mesa transistor was the forerunner of the planar device, but while it was possible to achieve the narrow base width, the small base area was achieved only by etching away the bulk material by corrosive agents which could contaminate the final transistor.

The planar technique obviates the trouble of the etching agent, since an oxide layer protects the junctions through the entire manufacturing process. The presence of the oxide layer over the junctions gives a further advantage, namely the protection against other contaminants that cause leakage currents across the junctions, to produce a more reliable transistor.

Germanium oxide does not have the property of masking the penetration of impurities and the planar technique is used, therefore, on silicon devices. There are, of course, many advantages in using silicon, such as the lower inherent leakage currents and higher breakdown voltages that can be attained. A limitation of the simple double-diffused planar structure is the high series collector resistance caused by the bulk of the material.

This disadvantage has been removed by the introduction of an epitaxial layer in the basic slice from which the transistor is manufactured. This enables a lower resistivity material to be used for the bulk of the transistor, producing low series resistance, while the epitaxial layer, being of a higher resistivity, enables the high breakdown voltages to be maintained.

Thus the double-diffused epitaxial planar transistor has been evolved. Its suitability and application as a radio-frequency power amplifier will be discussed in the following sections.

Transistor Ratings

There are three classes of maximum ratings which must be considered when designing a transistor circuit. They are those of current, voltage and power.

Current ratings:- The current rating of most importance in transistor power amplifiers is the maximum collector current rating (I_{CM}), since I_B is not likely to be of the order of I_{BM}. Also, $I_C \approx I_E$. In power switching circuits where the transistor is required to saturate heavily, I_{BM} can be of importance, but in power amplifiers this is very unlikely to be the case.

Two I_{CM} ratings are frequently given, one of these being $I_{CM(mean)}$ which is self-explanatory, the other being $I_{CM(pk)}$ which is normally stated for a pulse length and repetition frequency of greater than certain values. In double-diffused transistors the I_{CM} ratings are often imposed by the fine wires which are normally used for the internal connection to the emitter. Thus the ratings are linked to the current-handling capacity of the wires.† This does not necessarily mean that for a short-duration pulse the $I_{CM(pk)}$ rating may be exceeded, because there are heating time-constants within the transistor structure which might result in localized over-dissipation.

Also, the fall-off in h_{FE} at high values of I_C would limit the usefulness of an amplifier having I_C comprised of short pulses (such as in a class C amplifier).

Voltage ratings:- There are three voltage ratings associated with transistors, all of which must be considered in power amplifier applications.

In a single-ended stage, the emitter-base voltage rating (V_{EBM}) may be exceeded if the drive voltage is increased too much. Where a push-pull configuration is adopted, the base-emitter junction of one transistor tends to act as a catching diode for the other and therefore the configuration offers some protection in this respect.

V_s must not exceed V_{CEM} if $2 V_{CEM} \leqslant V_{CBM}$

V_s must not exceed $\frac{1}{2} V_{CEM}$ if $2 V_{CEM} > V_{CBM}$

SYMBOLS (IN ORDER OF APPEARANCE)	
I_{CM}	Maximum collector current rating
I_{BM}	Maximum base current rating
V_s	Supply voltage
V_{EBM}	Maximum reverse emitter-base voltage rating (collector open-circuited)
V_{CEM}	Maximum collector-emitter voltage rating (base open-circuited)
V_{CBM}	Maximum collector-base voltage rating (emitter open-circuited or reverse biased)
P_{CM}	Maximum collector power dissipation
h_{FE}	Large signal common emitter current gain
h_{FB}	Large signal common base current gain
f_T	Gain-frequency product
$|h_{fe}|$	Magnitude of small signal current gain
f_1	Frequency at which $|h_{fe}| = 1$
f_{max}	Maximum frequency of oscillation, i.e. frequency at which power gain equals unity
f_o	Frequency of operation
$c_{b'e}$	Collector to base feedback capacitance
$r_{bb'}$	Extrinsic base series resistance
R_L	Load resistance
$V_{CE(sat)}$	Collector-emitter saturation voltage

† A new manufacturing technique recently introduced by S.T.C. obviates the need for the fine wires, so that the current ratings are limited only by the total device dissipation and the fall in current gain at high currents.

AMPLIFIERS

By M. V. BOND*

SOME TYPICAL DESIGNS USING SILICON PLANAR TRANSISTORS

The latter will normally be the case and should therefore be adhered to for any type of amplifier. An amplifier stage should not be allowed to saturate very heavily, since the collector current waveform will tend to become square. The back e.m.f. from the inductor in the collector circuit may then produce voltage transients far in excess of the voltage ratings. If this cannot be avoided, a catching diode must be provided, although diodes of the necessary current rating and switching speed are few and far between.

It should be appreciated that, provided a sufficiently high series impedance is used, breakdown will not cause excessive dissipation, and any voltage rating may be exceeded. However, in power amplifiers where high conversion efficiencies from d.c. to r.f. are usually required, impedances high enough in value to prevent excessive power dissipation cannot be used owing to their adverse effect on conversion efficiency.

Power ratings:-The power rating which has most control over the design of power amplifiers is the collector power dissipation (P_{CM}). This is linked to the maximum permissible junction temperature, which is generally between 150°C and 200°C in the case of silicon transistors. The rise in junction temperature when electrical power is dissipated is due to the thermal resistance. For a typical transistor in a TO-3 encapsulation without a heat sink this is of the order of 40°C/watt. On an infinite heat sink the thermal resistance may be as low as 5°C/watt. Thus it may be seen that a large heat sink is essential when attempting to realize the maximum power handling capabilities of a transistor.

In order to eliminate P_{CM} as a limiting rating, the d.c. to r.f. conversion efficiency must be as high as possible, so that the minimum amount of power is dissipated in the transistor.

Transistor Parameters

Most of the parameters affecting the performance of a low level low-frequency transistor amplifier affect an r.f. power amplifier to a greater or lesser extent. Broadly, transistor parameters influence circuit performance in two ways, i.e. they influence stage gain and stage efficiency.

For a low-level amplifier the stage gain, input and output impedances and other performance details may be readily calculated by the use of " y parameters ". Unfortunately these parameters assume little or no change in bias conditions and transistor characteristics throughout the input voltage swing, a fact which is essentially true for low-level amplifiers, but far from true in the case of power amplifiers. Attempts have been made to relate y parameters measured under specific conditions to expected circuit performance. While this practice may give an approximate answer in some isolated cases, it is not safe to assume that it will be so for the majority of circuit designs. The answer to the problem is at present to design circuits empirically or to rely on functional performance information as given by some transistor manufacturers.

Effects on stage gain:-In a low-frequency amplifier, be it high or low power, h_{FE} (or h_{FB}) is the most important parameter. Whether the gain resulting from a given value of h_{FE} may be realized will be dependent on the input and output impedances of the transistor. As the frequency of operation is increased, so the gain will gradually fall off owing to the effect of the input and output shunt capacitances. The effect of these may be eliminated by placing LC circuits at the input and output, resonant with the capacitances at the operating frequency. As the frequency is raised still further, tuning the stage no longer maintains the gain since the cut-off frequency then becomes important. There are a number of ways of defining the cut-off frequency of a transistor, one of these being the gain-frequency product (f_T). It is assumed that h_{fe} falls off at a rate of 6 dB per octave, hence if $|h_{fe}|$ is measured at a frequency (f_m) known to occur on this slope, a gain-frequency product ($f_T = f_m.|h_{fe}|$) may be found. This approximates very closely to f_1, the frequency at which $|h_f| = 1$. This only applies to the current gain of a transistor. When considering the power gain one must take into account the input and output impedances. The power gain is influenced mainly by the maximum frequency of oscillation (f_{max})

It can be shown that

$$f_{max} = \frac{f_T}{8\pi c_{b'c} r_{bb'}} \text{ also, power gain} \propto \frac{f^2_{max}}{f_o}$$

Hence, power gain $\propto \dfrac{f_T}{c_{b'c} r_{bb'} f_o}$

It has been assumed that $c_{b'c}$ and $r_{bb'}$ are constant for a given transistor, but both these vary in value with frequency. For example, in a diffused transistor $r_{bb'}$ has no discrete value, since it is a complex quantity comprised of bulk base resistance and diffused capacitance.

So far it has been assumed that small signal conditions prevail, which is not true in a power amplifier. However, from the above it can be seen that at low frequencies h_{FE} is the main factor controlling the stage gain and that at higher frequencies f_{max} becomes the more important controlling factor.

Effects on stage efficiency:-As has already been stated, the maximum power available from a power amplifier is limited by the conversion efficiency if the maximum power dissipation is not to be exceeded. This is dependent on the most efficient utilization of the supply voltage, e.g. resistance in series with the collector current should be kept to a minimum in order to allow for maximum power to be available in the load. For this reason the collector saturation resistance should be as low as

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WIRELESS WORLD, DECEMBER 1964

595

possible. In a silicon epitaxial transistor such as the S.T.C. BLY12 high-frequency power type, the maximum saturation resistance is only $\frac{1}{2}$ ohm at a collector current of 2 amperes, i.e. the maximum saturation voltage at that current is only 1 volt.

Another factor which is of prime importance is the shape of the h_{FE} *versus* I_C curve. This should show a reasonably constant value of h_{FE} maintained to the peak collector current at which the transistor is to be used. When used at higher frequencies (where f_{max} is the more important factor) the f_T *versus* I_C curve should have a similar form if the transistor is to be used as a power amplifier at these frequencies.

The reason for this requirement is the dependency of the efficiency on the shape of the collector current waveform. This waveform tends to appear clipped due to the reduction in h_{FE} (or f_T) at high values of I_C. For a given value of $I_{C(pk)}$, this results in a higher mean value of I_C than would be obtained if the h_{FE} reduction did not take place. As the clipping effect becomes more severe, the fundamental signal current component amplitude increases at a slower rate than I_0 mean. In the case of the clipped I_C waveform the higher mean value of I_C results in a higher d.c. collector input power than in the ideal case where no reduction in h_{FE} (or f_T) takes place. This effect is shown below for a class B amplifier in the ideal case where the I_C waveform is a uni-directional half sine wave, and in the extreme case of clipping where I_C tends to be a square wave.

Half sine wave I_C

Mean $I_C = \dfrac{I_{C(pk)}}{\pi}$ amperes

$\therefore$ D.c. power input $= \dfrac{V_{CC}\, I_{C(pk)}}{\pi}$ watts

Fundamental a.c. power output $=$
$\dfrac{V_{CC}\, I_{C(pk)}}{4}$ watts

$\therefore$ Efficiency $= \dfrac{\dfrac{V_{CC}\, I_{C(pk)}}{4}}{\dfrac{V_{CC}\, I_{C(pk)}}{\pi}} \times 100 = 78 \cdot 5\%$

Square wave I_0

Mean $I_C = I_{C(pk)}$ amperes
$\therefore$ D.c. power input $= V_{CC}\, I_{C(pk)}$ watts
For a square wave, the peak fundamental
current $= \dfrac{4}{\pi} I_{C(pk)}$

Fundamental a.c. power output $=$
$\dfrac{2V_{CC}\, I_{C(pk)}}{\pi}$ watts

$\therefore$ Efficiency $= \dfrac{\dfrac{2V_{CC}\, I_{C(pk)}}{\pi}}{V_{CC}\, I_{C(pk)}} \times 100 = 63 \cdot 7\%$

Although the above instances are for an untuned amplifier, a similar result will be obtained when the amplifier is tuned. This merely selects the fundamental component and passes it on to the collector load.

Resulting from the above, it will be apparent that if I_C is too large it will put the transistor into saturation and the efficiency will suffer. If, owing to circuit tolerances, this cannot be avoided, another factor can reduce the efficiency still further. This factor is known as minority carrier storage, and it has the effect of increasing the mean value of I_0 without any increase in fundamental output power.

The effects of transistor ratings and parameters on performance have now been discussed, and with these effects in mind the design of typical amplifier circuits will now be considered.

Circuit Design

The choice of circuit configuration, be it common base, common emitter or common collector, is dependent on the application. For example, if a power amplifier is required to operate with consistent gain over a wide range of frequency, common base is the most suitable. However, if the maximum possible gain is required, common emitter would normally be more suitable, although it may be necessary to neutralize the stage to prevent regeneration in the latter case.

The use of the common collector configuration is only to be recommended in special cases. A tuned circuit in the emitter will be capacitive at some frequency and can result in a negative impedance appearing at the input to the stage. If the input source impedance is greater than this negative input impedance, regeneration may occur.

A typical amplifier may operate in class A, B or C. Class A is normally suitable only for low power stages, and the ideal efficiency under this condition is 25%. Class B is the most attractive amplifier form, since it combines reasonably high efficiency and moderate gain without the necessity for high values of $I_{C(pk)}$. Class C is very useful where a transistor maintains its h_{FE} (or f_T) to the high values of $I_{C(pk)}$ which result from using this class. There is, of course, a reduction in stage gain (c.f. class B), but in some cases this is acceptable. When considering the use of class C, it is important to note that the degree to which the transistor base-emitter junction may be reverse-biased is limited by V_{EBM}. The sum of the reverse bias voltage and reverse signal peak voltage must not exceed V_{EBM}.

Where a linear power amplifier is required, for instance when amplifying an amplitude-modulated signal, some form of class AB would normally be used since this gives good linearity whilst maintaining the power gain and efficiency of a class B stage.

Power matching:- When attempting to transfer the maximum available power from a source having appreciable internal impedance, it is customary for the load to have the conjugate value of the source impedance, i.e. if $Z_L = R + jX$ then $Z_S = R - jX$.

In the case of a load having a non-linear impedance it is obviously not possible to match a linear impedance source at all values of instantaneous current (or voltage). This is the case with a transistor power amplifier so that a compromise load impedance must be selected. This results in the maximum available source power not being utilized and the effect on the performance of the amplifier is a loss in power gain. This loss is not appreciable in the case of a class B push-pull amplifier, but can be appreciable for a single-ended stage operating in class B or C. This is because the transistor does not have another to balance the load and make it appear symmetrical for both positive and negative half-cycles of input current. For this reason, a push-pull circuit is recommended for use whenever possible. Another

factor which makes the use of the push-pull configuration more desirable is that the base-emitter diodes of the transistors act as catching diodes for one another and prevent the V_{EBM} rating being exceeded (as mentioned previously).

When designing a linear amplifier the most linear results will be obtained if a previous transistor is matched to the lowest value of input impedance of the next stage, i.e. at that part of the cycle when the base is conducting most heavily.

Owing to the fact that high currents flow in power amplifiers for relatively low applied voltages, the input and output impedances are low, say between a few ohms and tens of ohms. These low values of impedance present problems which probably are the most severe to be encountered during the design (and construction) of transistor r.f. power amplifiers.

These problems are due to two main causes, one of them being lead inductance. The inductive reactance of a lead becomes comparable to the low input and output impedance of the transistor when operating at radio frequencies, hence the transfer of power is inhibited unless the reactance can be tuned out. If amplifier interconnections are made as short as possible, or of flat strip, the above effect can be minimized.

The other problem is caused by the difficulty in achieving sufficiently high values of coupling coefficients in interstage transformers. Owing to the low impedances to be matched the required inductances are low in value and in fact become lower than the transformer leakage inductances. Ways of overcoming this problem are, for example, the winding of transformers on ferrite pot cores or ferrite ring formers. Even these may not be sufficient, and bifilar winding, or the twisting together of primary and secondary prior to winding, may be necessary to achieve efficient power transfer. The high self-capacitance of such transformers is of little consequence, again owing to the low impedance levels.

For a push-pull class B output stage:

$$\text{Load resistance } (R_L) = \frac{V_s - V_{CE(sat)}}{I_{C(pk)}}$$

and

$$\text{Output power} = \frac{V_s - V_{CE(sat)}}{2R_L}$$

Note that $V_{CE(sat)}$ is the value at $I_{C(pk)}$

Little has yet been said with regard to the types of matching circuit that are suitable, although they have been inferred by the above comments. It will be obvious that for reasonable values of working Q the transformers must have low inductances and the tuning capacitors must be large. In fact the values necessary are fractions of a microfarad instead of the tens of picofarads normally expected at radio frequencies greater than, say, 5 Mc/s. The use of high value capacitors is unattractive owing to the high dielectric losses which accompany them. They also tend to resonate or become inductive at these frequencies. The alternative would appear to be the use of tertiary windings of higher inductance in order to reduce the tuning capacitances to more practical values. However, this presents the same problem of leakage inductance as previously mentioned.

The most satisfactory method of obtaining reasonable harmonic rejection is by the use of ferrite-cored matching transformers followed by low-pass filters. It would be rarely necessary to achieve good harmonic rejection except when transferring the power from the output stage to the load, which may be an aerial. If capacitors of the required type were available, the problem of obtaining high values of working Q and good transfer efficiency would be considerably reduced.

Wherever possible, the ratio of output to input impedance should be low, since r.f. transformers having low turns ratios generally have lower leakage inductances, resulting in more efficient power transfer, than transformers having high turns ratios. This would seem to preclude the use of the common base configuration and, in practice, the common emitter configuration is found to be the more satisfactory of the two.

Neutralization:- If the stage gain is appreciably affected by the feedback component of output voltage, neutralization is generally required. The use of neutralization is well known in both v.h.f. valve amplifier stages and in low-level h.f. transistor amplifier stages. Its use is normally restricted to fixed-frequency amplifiers, since the neutralized feedback impedance varies in value with change in frequency. In a transistor power amplifier the feedback impedance not only varies with frequency, but also varies considerably throughout the signal cycle. Thus, a stage that may be neutralized for one value of instantaneous collector current may be unstable for another. This instability can result in self-oscillation of the stage. From this it is apparent that the use of neutralization is restricted to amplifiers where the signal swing is small.

Fortunately, the need for neutralization does not arise frequently. When it is necessary, the use of mismatching is an alternative solution. This is a very effective method for eliminating instability, and takes the form of presenting to the transistor a lower value of load impedance than is necessary for maximum power gain. This results in a lower value of feedback voltage and slightly lower power gain.

One effect that can cause trouble in a practical circuit is a momentary instability which occurs when the following stage ceases to conduct during part of the signal cycle. This can be cured by shunting the load with a resistance low enough in value to prevent the instability occurring. The power dissipated in this resistance generally will not be an appreciable proportion of the power output.

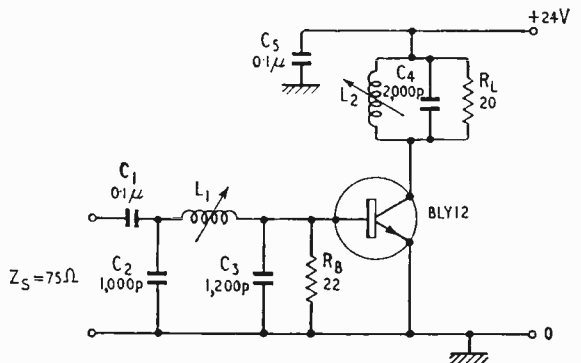The factors controlling the design of r.f. power

Fig. 1. 12-watt, 10 Mc/s power amplifier. L_1, L_2 are dependent on circuit layout and adjusted to resonate at 10 Mc/s with circuit capacitances.

amplifiers have been discussed. In the following sections typical circuits will be shown.

Typical Power Amplifier Circuits

12 watts output at 10 Mc/s:- The circuit (Fig. 1) uses a BLY12 transistor as a single-ended common-emitter class C amplifier. The resistor R_B is used to derive the reverse base-emitter bias voltage necessary for the amplifier to operate in class C. If class B operation is required, R_B should be replaced by a low-resistance radio-frequency choke of approximately $25\mu H$.

The input circuit takes the form of a π-network which transforms the 70Ω source impedance to a value suitable for matching to the base of the transistor. Both this transformation and R_B assist in keeping $V_{EB(pk)}$ below the V_{EBM} rating.

The output circuit is conventional and has a working Q value of 10. Where a load of other than 20Ω is to be used it is advisable to use a ferrite ring as a former for the collector transformer. It will be preferable, when the load is higher than 20Ω to tune the secondary of the transformer, since the required tuning capacitor will be of a lower value than that shown.

In practical cases it may be desirable to use a ferrite transformer of higher inductance and to follow this by a low-pass filter for harmonic rejection.

Performance

At 10 *Mc/s:*	Power output:	12 watts
	Stage gain:	10 dB
	Collector efficiency:	67%
	Harmonic output:	—20 dB on fundamental
At 5 *Mc/s:*	Power output:	12 watts
	Stage gain:	17 dB
	Collector efficiency:	70%
	Harmonic output:	< -20 dB on fundamental

For operation at 5 Mc/s: $C_1 = 2200$ pF; $C_2 = 0.01\mu F$; $C_3 = 2000$ pF; L_1 and L_2 to tune to 5 Mc/s. Other component values are similar to those of the 10 Mc/s circuit.

By using two BLY12 transistors in push-pull, it should be possible to obtain 25 watts output at frequencies up to at least 10 Mc/s.

9 watts output at 7 Mc/s:- Fig. 2 shows a complete, practical transmitter circuit.

The first stage is a crystal oscillator using a BSY24 transistor. If an inductive collector load is used the emitter capacitor, C_1, causes a negative impedance to appear between base and ground. When a quartz crystal is connected between these points, the circuit will oscillate at a frequency between the series and parallel resonant frequencies of the crystal.

The output of this stage is coupled to the driver stage by the transformer, T1.

The driver stage utilizes a BUY10 transistor in a class B, grounded-collector, common-emitter stage. The use of the grounded-collector connection gives the advantage of good heat dissipation, without the need for isolating the case from ground by the means of an insulating washer and bushes. The output from this stage is coupled to the final stage by the transformer, T2. The transformer is bifilar

wound to obtain a high coupling coefficient and, for convenience, has a Q of less than unity.

The output stage uses two BUY10 transistors in push-pull connection, operating in a similar manner to that of the driver stage. The load is coupled to the stage by the transformer, T3.

When operating in push-pull, with the output stage not saturating, further harmonic attenuation was found unnecessary.

The output is intended to be coupled to an antenna via a 75Ω coaxial cable.

Performance at 7 Mc/s

Power output:	9 watts
Output stage efficiency:	56%
Harmonic output:	—20 dB on fundamental

1.5 watts output at 27 Mc/s:- This circuit (Fig. 3) is similar to that of the 7 Mc/s transmitter. The power output is lower owing to the reduction in the gain of each stage at the higher frequency. It is possible to obtain greater than 6 watts output at 27 Mc/s by using the oscillator and driver stages of Fig. 3 to drive similar driver and output stages to those in the 7 Mc/s circuit. The latter stages would require BUY11 transistors for this application.

The first stage uses the BSY24 transistor in a negative resistance oscillator circuit similar to that in the 7 Mc/s transmitter.

The driver stage uses a BSY25 transistor in a single-ended class B configuration, which drives the push-pull BSY25 output stage. The BSY25 transistors must each be mounted on small isolated heat sinks (effective radiating area greater than 3in²) so that the P_{CM} rating of the transistor is not exceeded.

Performance at 27 Mc/s

Power output:	1,5 watts
Output stage efficiency:	50%
Overall d.c. to r.f. efficiency:	37%

1 watt output at 27 Mc/s:- This circuit (Fig. 4) while giving a lower power output than the previous circuit, is more economical and operates from a 12 volt supply.

The method of interstage coupling is more conventional than that utilized in the previous 27 Mc/s circuit and it is also easier to adjust for optimum results.

The first stage is a negative-resistance crystal oscillator using a BSY26. The output from this stage is coupled from a tap on the collector tuned circuit inductor, via a capacitor, to the base of the driver stage which uses a BSY25. The radio frequency choke from the base to the negative side of the supply is necessary to provide a d.c. path, in order to prevent the base-emitter junction from becoming reverse biased. The 10Ω resistor in the emitter circuit of the stage improves the current gain linearity and thereby assists in increasing the efficiency.

The output from the driver stage is coupled via an LC network, similar to that used between the oscillator and driver stages, to the base of the BSY25 output transistor. It should be noted that the driver stage may oscillate when the output transistor is not conducting. This can be prevented by shunting the collector load with a resistance.

The load is matched to the output transistor by

598

Fig. 2. 7 Mc/s transmitter.
T1 Primary: 15 turns 28 s.w.g. Secondary: 5 turns 28 s.w.g. on 0.3in dia. slug-tuned former (Neosid Type 5000A).
T2 Primary: 2 turns 18×47 s.w.g. litz. Secondary: 1 turn each 18×47 s.w.g. litz on "Stanferite" ring WP3810/SF7.
T3 Primary: 8 turns 36×47 s.w.g. litz, centre-tapped. Secondary: 7 turns 36×47 s.w.g. litz on "Stanerite" ring, WP3809/SF7.
N.B. T2 and T3 are bifilar-wound.
Crystal X: 7 Mc/s fundamental STC Type 4044/BT.
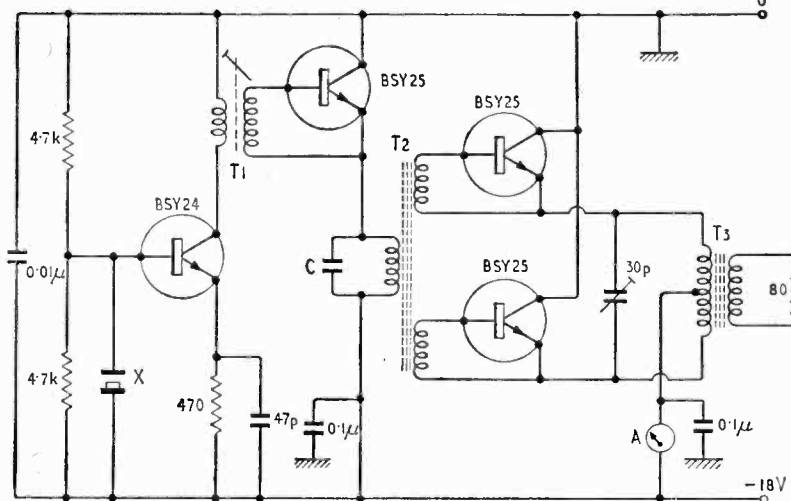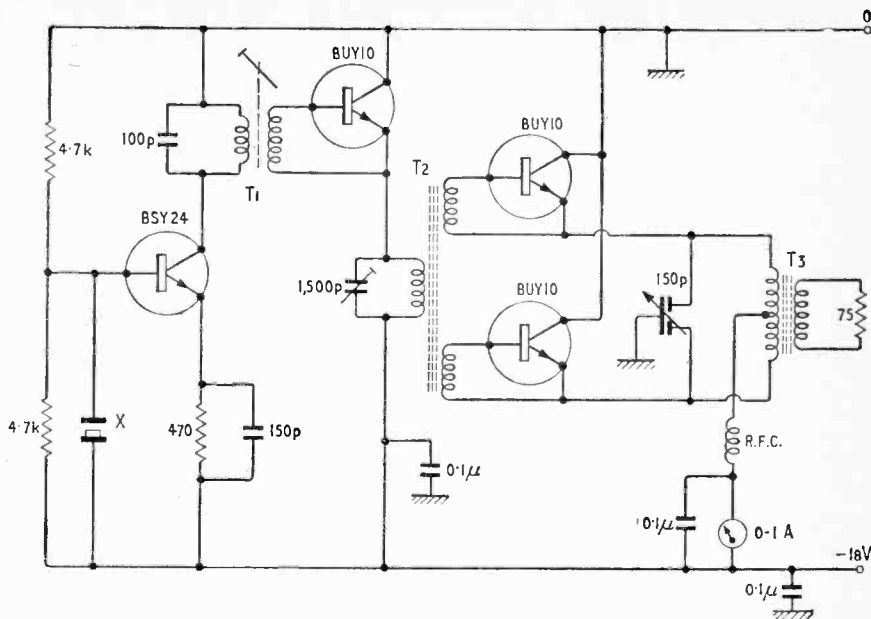R.F.C.: 30 turns 20 s.w.g. on WP3809/SF7 "Stanferite" ring.

Fig. 3. 1.5 Watt 27 Mc/s Transmitter.
T1: 7 turns/2 turns of 30 s.w.g. on 0.3in dia. slug-tuned former (Neosid Type 5000A).
T2: 4 turns/2 + 2 turns of 18×47 s.w.g. litz, bifilar-wound on a "Stanferite" ring Type WP3810/SB500.
T3: 4 turns centre-tapped/2 turns of 36×47 s.w.g. litz, bifilar-wound on a "Stanferite" ring, Type WP3809/SB400.
C: Dependent on circuit layout $\approx$ 20 pF.
Crystal X: 27-Mc/s 3rd overtone crystal, STC Type 4046.

Fig. 4. 1 watt 27 Mc/s transmitter.
Crystal X: 27 Mc/s quartz crystal. STC Type 4046.
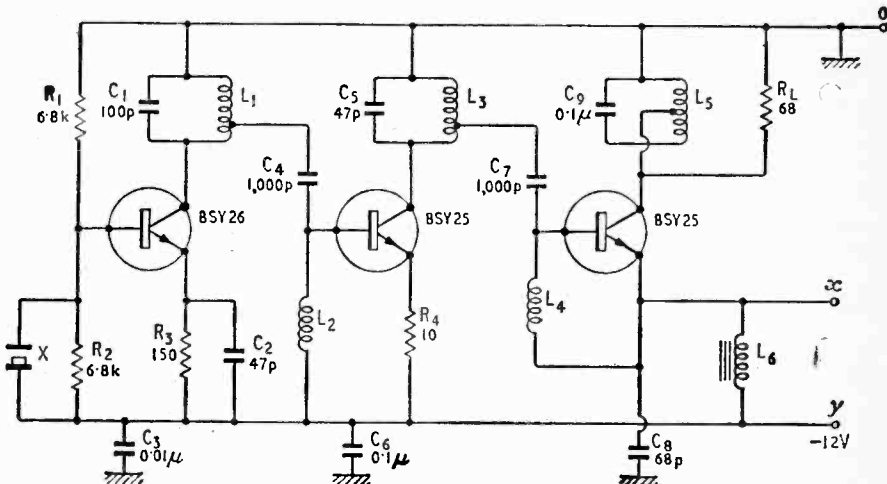L1: To resonate with circuit capacitance at 27 Mc/s.
L2: Radio-frequency choke—5μH.
L3: To resonate with circuit capacitance at 27 Mc/s.
L4: Radio - frequency choke—5μH.
L5: To resonate with circuit capacitance at 27 Mc/s.
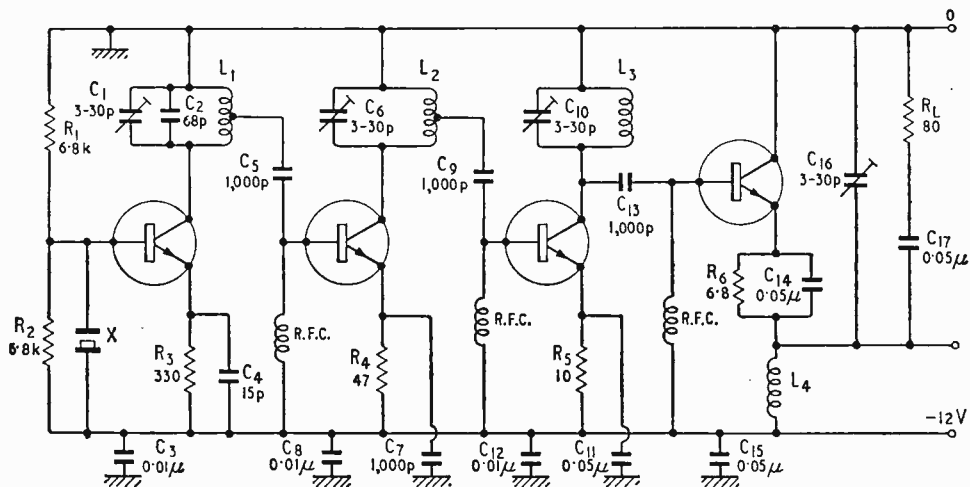L6: Audio - frequency choke—1 henry.

Fig. 5. 100-Mc/s ½-watt transmitter. All transistors are experimental type similar to 2N708. Crystal X: 50-Mc/s 3rd overtone crystal. L1, L2, L3, L4: 4 turns of 18 s.w.g., ½in long, ¼in dia. L1: tapped 1 turn from collector. L2: tapped 1½ turns from earth.

tapping the collector tuned circuit inductor at the point which gives maximum power output.

Performance at 27 Mc/s

Power output:	1 watt
Output stage efficiency:	58%
Overall d.c. to r.f. efficiency:	35%

Amplitude modulation has been attempted on this circuit. The most satisfactory method is to collector-modulate the output stage by connecting an audio frequency choke between points *x* and *y* on Fig. 4 and applying the modulating voltage to point *x*. It is possible to modulate the 1 watt carrier to a depth of 85% without the distortion exceeding 10%. The BSY25 in the final stage should be mounted on a 3in² (effective) heat sink.

0·5 watt output at 100 Mc/s:–The final circuit (Fig. 5) shows a design capable of giving 0·5 watt output at 100 Mc/s. All the transistors used in this circuit are S.T.C. type BSY28.

The first stage utilizes the negative resistance oscillator described in an earlier section, in this case using a 50 Mc/s overtone crystal. The 50 Mc/s output is coupled, via an LC network, to the base of a class C common-emitter stage operating as a frequency doubler. The reverse bias for class C operation is derived from the d.c. voltage developed across the emitter resistor which is decoupled. The collector circuit is tuned to 100 Mc/s and is coupled to the base of the driver transistor through a capacitor. The d.c. return path from the base is by way of a radio frequency choke to the negative supply line, as in the previous stage. The emitter resistor is decoupled as in the previous stage, so that the transistor operates in class C. Until this point, the design has been quite conventional, but owing to the difficulty of efficiently coupling the output of the driver stage to the base of the output transistor, the common collector configuration has been chosen for the final stage. This results in a higher load impedance being presented to the driver stage than would be so if the output stage were connected in common emitter. Even with this configuration the effective gain of the output stage is only 3 dB, compared to the 7dB gain of the driver stage. The reduction in gain is due mainly to the reduction in f_T at the peak current at which the transistor must operate. The output stage is intended to operate with an 80 Ω load.

Performance at 100 Mc/s

Power output:	500mW
Output stage efficiency:	65%

A low-pass or band-pass filter should be used in order to reduce the harmonic output to a satisfactory level.

Several circuits have been described illustrating the use of silicon double-diffused epitaxial planar transistors in practical power amplifiers.

The current trend, as would be expected, is towards higher output power at higher frequencies. Already transistors for this purpose are becoming available but are, as yet, more expensive than thermionic valves capable of giving a similar performance. However, it is likely that these transistors will be utilized in mobile transmitters where the low voltage supply is readily available.

"WIRELESS WORLD"

Change in Size and Price

PRODUCTION costs of this journal have been rising steadily over the past few years, and have reached a stage where they can no longer be absorbed. However, we have been reluctant to increase the price of *Wireless World* without giving the reader some benefit in return.

It has now been found possible to increase the size of the journal slightly to improve the appearance and readability, and still keep the price increase down to 6d.

As from the January, 1965 issue, therefore, the price will be 3s per copy, with a subscription rate of £2 6s (home) and £2 15s (overseas).

In the January issue we shall be publishing the first part of a constructional article on a high-quality transistor audio amplifier designed by J. Dinsdale, and also the start of a new series by T. D. Towers on laboratory instrument practice, written specially for those about to begin work in an electronics laboratory.

MANUFACTURERS' PRODUCTS

NEW ELECTRONIC EQUIPMENT AND ACCESSORIES

Digital Clocks

THE American Tymeter range of 12 and 24 hour digital clocks is now available from Spectrum Electronics Ltd., of Deneway House, Potters Bar, Middx. These clocks, which are suitable for operation from 50 or 60 c/s 110 or 230 volt supplies, are driven by 450 r.p.m. fully enclosed synchronous motors. The time is registered on three polysided drums and a 1 r.p.m. wheel, calibrated in ten second intervals. The hours and minutes of the 12 hour clocks use $\frac{5}{8}$ in high characters while the 24 hour clocks use $\frac{7}{16}$ in characters. The clocks can be supplied for mounting in users equipment or in attractively styled cases.
15WW 305 for further details

Count-down Relay

A SIMPLE count-down relay made by Rodene for either auto-reset or manual-reset applications is now available from the company's sole distributors D. Robinson & Co. Ltd., 5-7 Church Road, Richmond, Surrey. Developed from their 700 series, this new relay uses a twenty-position solenoid-operated ratchet wheel and has a maximum stepping rate of five per second. A life expectancy of about three million operations is quoted.

Any pulse length of over 25 milliseconds will operate the new relays, which incidentally can be cascade connected for decade counting. Auto-resetting is effected by means of a separate coil, which when energized releases the ratchet and allows the return spring to reset the relay. This operation takes less than 50 m sec.
15WW 306 for further details

Broadband Coaxial Diodes

AN extension to the range of low-noise broadband coaxial mixer diodes is announced by Microwave Associates Ltd., of Cradock Road, Luton, Beds. The peak power rating of the three new devices is three watts with a c.w. rating of 250 mW. The MA-490E covers the frequency band 12.5 to 17.5 Gc/s; the MA-491E covers 10.5 to 15.5 Gc/s; and the MA-492E covers 7.0 to 12.0 Gc/s. All three are hermetically sealed and feature high temperature operation (150°C.). Centre band noise figures are quoted as 8.0, 7.5 and 7.2 respectively. The i.f. impedance for the diodes is between 400 and 565 ohms and the r.f. impedance is 65 ohms.
15WW 307 for further details

Gearheads and Speed-reducers

A RANGE of small precision gearheads and speed-reducers for electromechanical servo applications is soon to go into production at Ancillary Developments Ltd. factory on Blackwater Station Estate, Blackwater, Camberley, Surrey. Five international frame sizes are to be available, with gear ratios of up to 1,000: 1 for the three smaller sizes (08, 10 and 11) and up to 5,000 : 1 for the size 15 and 18 frames. Higher ratios can be achieved by cascading.
15WW 308 for further details

Adhesive-backed Expanded Rubber

THE range of expanded rubber produced by Expanded Rubber and Plastics Ltd., of Mitcham Road, Croydon, under the trade name Rubazote, is now available with an adhesive backing. The company has been supplying similar materials to the industry for the past three years and it can now supply coils of adhesive-backed Rubazote in widths of up to two inches, in lengths from 18 to 50 ft, this being dependent upon the thickness of the rubber. Wider strips are available in lengths of about three feet, and gaskets can be punched to order.
15WW 309 for further details

Wide-range Microwave Sources

THREE klystron signal sources Types PCLL, PCLS and PCLX are available, covering the frequency ranges 0.75 to 3 Gc/s, 1.2 to 5 Gc/s and 4 to 12 Gc/s respectively, from Flann Microwave Instruments Ltd., Old Bridge Street, Kingston-upon-Thames, Surrey. All the models use plug-in klystrons in coaxial line cavities and the reflector voltage is automatically adjusted for the frequency of operation. Frequency is displayed on a direct reading dial to an accuracy of better than 1% for low loading conditions. The maximum output power available throughout the stipulated frequency ranges
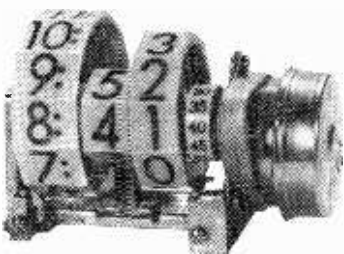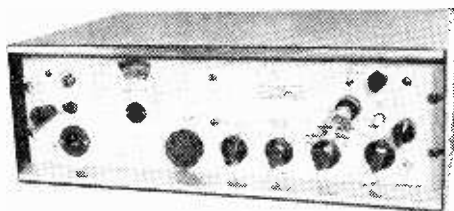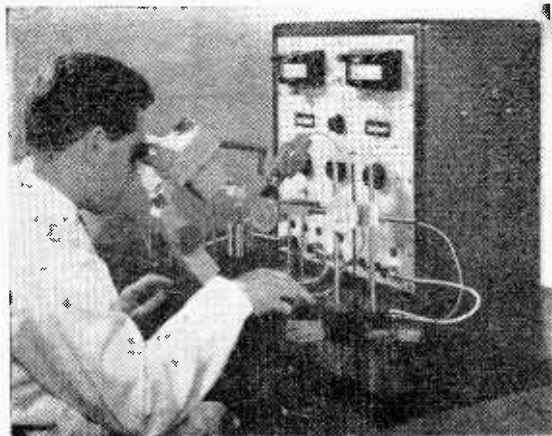

The Tymeter 12 hour digital clock.

Rodene count-down relay.

One of the new range of servo gearheads and speed-reducers shortly to be made by Ancillary Developments Ltd.
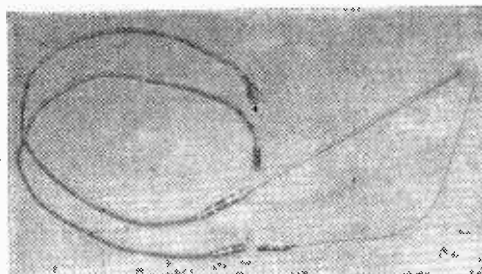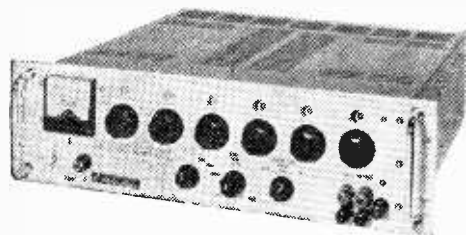

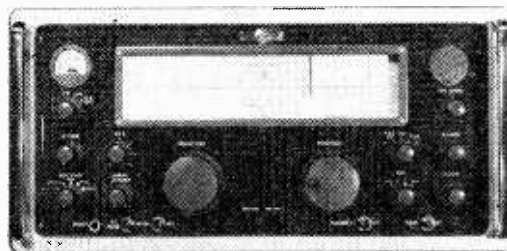Microwave signal source from Flann Microwave Instruments Ltd.

Thermocompression bonder from G. V. Planer Ltd.

Two standard thermocouples from Spembly Technical Products; one being 0.125 and the other being 0.040 in in diameter.

Type H2301 h.f. receiver from Marconi.

Fluke Model 382A voltage and current calibrator.

varies from 10 to 180 mW for the PCLL, 10 to 80 mW for the PCLS and 2 to 60 mW for the PCLX. Stabilized power supply units are featured in all of the instruments and drive both the klystron oscillator and the modulator circuits. Square-wave modulation of variable amplitude and rate, d.c. restored, or a sawtooth waveform can be switched to the reflector if required by means of a switch. Alternative versions of the instrument can be made to include a precision attenuator, and some variations to the frequency coverage can also be effected.

15WW 310 for further details

Themocompression Bonder

PARTICULARLY suited for bonding leads to thin film microcircuits, semiconductor materials and devices, etc., is the recently introduced thermocompression bonder by G. V. Planer Ltd., of Windmill Road, Sunbury-on-Thames, Middx. Precision micromanipulators are incorporated in this bonder, which has an independent thermostatically controlled bonding chisel and substage. Two micro-manipulators are used, one to carry the chisel and the other to carry the wire feed assembly, which is compressed-air assisted and capable of accepting wires of between 0.0005 to 0.01in in diameter. The chisel and substage

can be maintained at temperatures of up to 600°C; with the provision for the issue of inert gas over the working area, if required. A "stereozoom" binocular microscope with continuously variable magnification of 7 to 30 or 14 to 60 and adjustable illumination is incorporated.

15WW 311 for further details

Voltage and Current Calibrator

SOLID-STATE circuitry is featured in the Fluke 382A calibrator which can be used as an accurate constant voltage or a constant current source. Outputs of up to 50 volts and 2 amps, set to an accuracy of $\pm 0.01\%$, are available, and the load and the line regulation is quoted to be in the order of

five parts in 10^6. Continuous voltage and current limiting may be set to operate simultaneously. Known as the Model 382A, this instrument is made in the United States by Fluke Manufacturing and is available in this country, at £638 (subject to Treasury direction), from Livingston Laboratories Ltd., 31 Camden Road, London, N.W.1.

15WW 312 for further details

Mineral-insulated Thermocouples

STANDARD units consisting of a thermocouple with an insulated hot junction, a connector and a compensating lead are now being manufactured by Spembly Technical Products

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INFORMATION SERVICE FOR PROFESSIONAL READERS

To expedite requests for further information on products appearing in the editorial and advertisement pages of *Wireless World* each month, a sheet of reader service cards is included in this issue. The cards will be found between advertisement pages 16 and 19.

We invite readers to make use of these cards for all inquiries dealing with specific products. Many editorial items and all advertisements are coded with a number, prefixed by 15WW, and it is then necessary only to enter the number(s) on the card.

Readers will appreciate the advantage of being able to fold out the sheet of cards, enabling them to make entries while studying the editorial and advertisement pages.

Postage is free in the U.K., but cards must be stamped if posted overseas. This service will enable professional readers to obtain the additional information they require quickly and easily.

www.americanradiohistory.com

Ltd., of New Road Avenue, Chatham, Kent. Nickel chromium and nickel aluminium conductors are used in the thermocouple and are housed in a stainless-steel sheath, the outside diameter of which ranges from 0.020 to 0.125in according to type. The normal maximum operating temperature of these units is 800°C.

The conductors are terminated and sealed in a special two-pin "plug and socket" type of connector, which is fitted with a flange nut for bracket or bulkhead mounting. These thermocouples are supplied with test certificates and can, if required, be calibrated.

15WW 313 for further details

High-stability H.F. Receiver

A GENERAL-PURPOSE communications receiver covering the frequency band 500 kc/s to 30.5 Mc/s is now available from the Marconi Company. Known as the Type H2301, it is an Eddystone receiver modified to meet the Marconi specification for s.s.b. and f.s.k. operation. The frequency range is covered in 30 steps of 1 Mc/s with an overlap of 100 kc/s. The dial calibration, which is within 1 kc/s, is linear and presented on separate Mc/s and kc/s scales. Calibration can be checked by means of an internal 100 kc/s crystal oscillator.

Double conversion techniques are used in this receiver, with crystal control of the first local oscillator to provide an above average stability figure for this class of receiver. (Within ± 20 c/s with constant mains supply and ambient temperature.) The second i.f. operates at 500 kc/s and a cathode follower is incorporated to provide an output at this frequency for connection to ancillary equipment. The receiver has an internal loudspeaker as well as a number of separate audio outputs with independent gain controls. An all transistor f.s.k. adaptor is available for this receiver and caters for shifts of from 200 to 850c/s at telegraph speeds of up to 100 bauds.

15WW 314 for further details

Illuminated Push-button Switches

A NEW twin illuminated push-button switch containing two independent push-buttons in a single housing is announced by Highland Electronics Ltd., of 26-28 Underwood Street, London, N.1. This unit matches and is physically interchangeable with the existing Highland range of twin lamp illuminated push-button

switches and twin lamp annunciators. It is available with both switches impulse action, both switches step action or one switch step and one impulse action. There is one changeover contact per button rated at 6 amps at 230 volts a.c. They can be supplied to operate lamps from 6, 12, 24, 36, 48, 60, 110 or 240 V supplies (the last two voltages with a series resistor).

15WW 315 for further details

Small Rotary Switch

A HALF-INCH rotary switch, available with or without a water seal, is announced by Diamond H Controls Ltd., of Gunnersbury Avenue, London, W.4. The diameter over the contact lugs is only $\frac{11}{16}$ in and the complete unit is claimed to be one of the smallest in the world, accommodating up to five sections of switching. It incorporates the "Oak" double wiping, self-cleaning contact which features low contact noise and low contact resistance throughout its rated life. Indexing up to 12 position, 30° throw, is provided and each contact is rated at 100 mA at 50 V d.c. The minimum insulation resistance is quoted at 10 MΩ.

15WW 316 for further details

Automatic Quartz Cutting Machine

A PRECISION automatic slicing machine made by Meyer and Burger, of Switzerland, is being marketed in the United Kingdom by G. Zwicky (London) Ltd., of Zwicky House, 241-247 High Street, Acton, London,
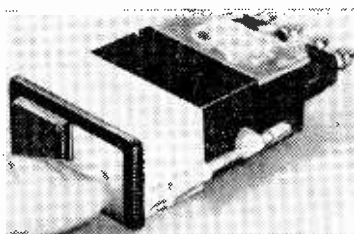
New twin illuminated push-button switch from Highland Electronics.

Small rotary switch from Diamond H Controls Ltd.
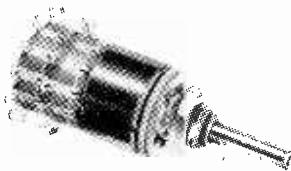
W.3. It is primarily designed for cutting quartz, but is equally suitable for other materials such as optical glass, ceramic, etc. Known as the Model QS 2, this machine features an electro-hydraulically controlled cutting feed and can be run at three different speeds; 1570, 1880 and 2190 r.p.m. Cutting blades with outside diameters from 10 to 14in can be used on this machine. The cost of the machine is in the region of £1,200.
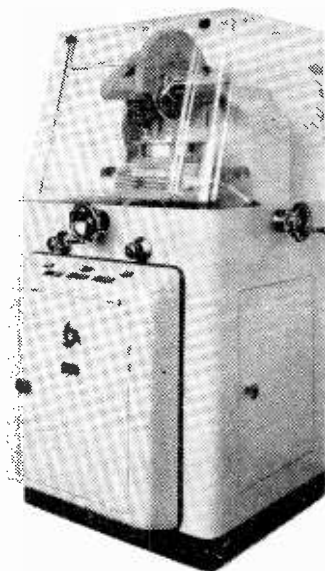
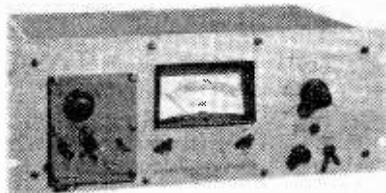15WW 317 for further details

Capacitor Electrometers

TWO new ultra-sensitive dynamic capacitor electrometers, one in the form of a portable instrument and the other suitable for standard rack mounting, are announced by the Victoreen Instrument Company, of Cleveland, Ohio. While differing in physical appearance, both models have the same electrical characteristics. The portable unit is designated 475A and the other 475R.

Transistors are used throughout and the minimum input impedance when the instrument is used for voltage measurements is quoted to be 10^{16} ohms. There are nine voltage ranges from 3 mV to 30 volts f.s.d. and on the current side there are forty-five ranges from 3×10^{-15} to 3×10^{-7} amps f.s.d.
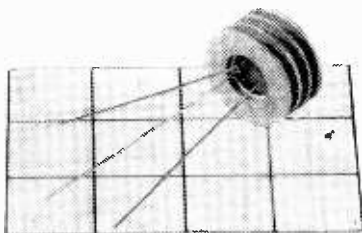
The basic element in these instruments, the dynamic capacitor assembly, comprises a vibrating capacitor, coupling capacitor, and magnetic drive and pickup coils. The dynamic capacitor is torsionally mounted and

Meyer and Burger automatic quartz cutting machine.

Bendix Type 399 digital d.c. voltmeter.

Transipack 3021 miniature stabilized power supply unit from Industrial Instruments Ltd.

Digital voltmeter from Digital Measurements Ltd., of Salisbury Grove, Mytchett, Aldershot, Hants.

Victoreen Fetometer dynamic capacitor electrometers.
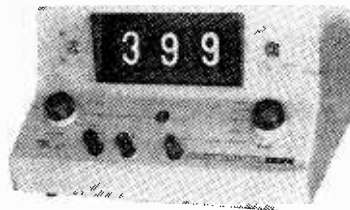

Heat sink from Jermyn Industries.

has a mechanical resonance of 300 c/s, ±30 c/s. During operation, a charge on the dynamic capacitor produces a 300 c/s sinusoidally varying voltage. This voltage is amplified and then compared with an output from an internal oscillator in a phase detector. The resultant d.c. is then applied to a 50 μA meter movement.

These instruments are available in the United Kingdom from A.E.P. International Ltd., of 443 Great West Road, Hounslow, Middx.

15WW 318 for further details
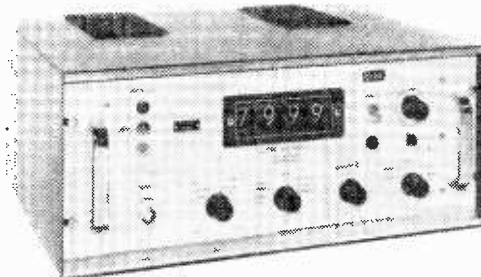
Printed Circuit Heat Sinks

TWO heat sinks, one for transistors encased in Type T05 cans and another for Type T018 cans, have been introduced by Jermyn Industries, of Vestry Estate, Vestry Road, Otford Road, Sevenoaks, Kent. Made of aluminium, these heat sinks comprise two metal parts that screw together on the welded flange of the transistor and thus require no fixing to printed circuit boards for which they are designed. Heat is radiated through fins on the outside of the

heat sink and a figure of 38° C per watt thermal resistance is given for the unit that fits T05 cans and 37° C per watt for the other.

15WW 319 for further details

Miniature Stabilized Power Supply

A FULLY stabilized power supply unit of $4\frac{1}{4} \times 3 \times 2\frac{1}{2}$ in and able to produce 0.5 A at any voltage up to 30 V has been announced by Industrial Instruments Ltd., of Stanley Road, Bromley, Kent. Called the Transipack 3021, it is protected against overload and has been primarily designed to be built into equipment, although it may be used as a separate unit if desired. It operates from any 100-250 V, 45-65 c/s supply, with a maximum fluctuation of ±10%, and the output voltage may be preset between 6 and 30 V. Output resistance (d.c.) is quoted to be within 0.01 to 0.05 Ω, impedance less than 0.5 Ω and ripple approximately 5 mV p. to p. Ambient temperature range is from −10 °C to +40 °C and a packaging density of 2 watts per cubic inch of volume is stated. The price is £24 10s.

15WW 320 for further details

Digital Voltmeter

A THREE-DIGIT instrument capable of measuring voltages of either polarity to a maximum of 1,000 volts d.c. has been added to the range of measuring and test instruments made by Bendix Electronics Ltd., of High Church Street, New Basford, Notts.

Known as the Type 399, it has five ranges, the lowest of which has a f.s.d. of 100 mV d.c. Resolution, linearity and accuracy are each quoted at 0.1% full scale. Projection type digital indicators are used in the display which incidentally, can be manually controlled or arranged to automatically re-cycle at five second intervals. Transistors are used throughout.

15WW 321 for further details

High-impedance Digital Voltmeter

AN INPUT impedance of greater than 25,000 MΩ on the two lowest ranges and of 10MΩ on the higher ranges is quoted for Digital Measurements latest digital voltmeter, the DM2005. This instrument, priced at £500, has an accuracy of 0.0125 % f.s.d. (±0.02 % of reading) and a sensitivity of 10 μV.

Six ranges are provided on this instrument covering zero to 2 kV and the input can be isolated from earth to reject common mode voltages sometimes present at the signal source. A four-digit display is featured in the DM2005 which incidentally retains the facilities featured in earlier DM digital voltmeters.
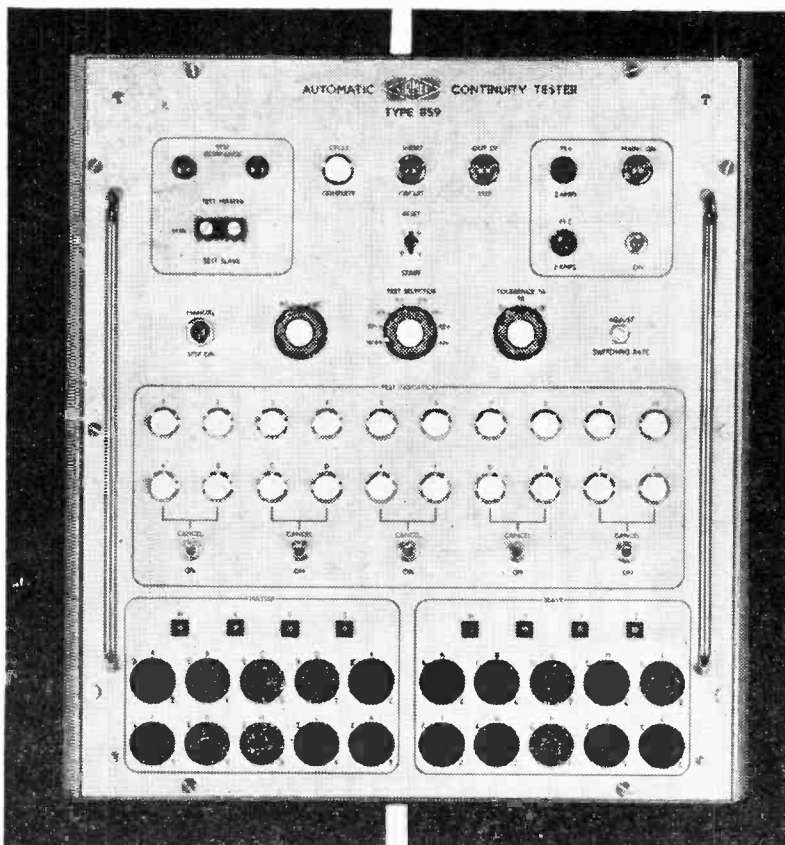
15WW 322 for further details

Ultrasonic Delay Lines

STANDARDIZATION of piece parts in a new range of ultrasonic delay lines announced by the M.E.L. Equipment Company, of Crawley,

One of the new ultrasonic wire delay lines from the M.E.L. Equipment Company.
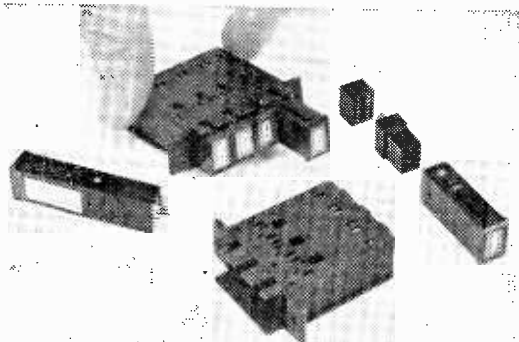
Teldix MA-11 micro-indicators with cassettes for three-, four- and five- decade displays. The unit on the right is unplugged to show its three component sub-units—a socket, lamp unit and display unit.

Mark II step and repeat camera from Watson for use in the semi-conductor industry.

Canadian "Imperial" soldering iron, that features interchangeable bits, elements, and handles, is being distributed in the United Kingdom by Lugton & Co.

has considerably reduced unit cost. Designated "Delay Modules YL2108", units in the range are available with delays from 100 microsec to 3.2 millisec, each type of unit being adjustable over a range of 20 microsec.

These units will operate from any 12 volt supply, which may be either positive, negative or symmetric, and are therefore, suitable for most types of n-p-n and p-n-p transistor logic systems, and are fully compatible with the M.E.L. Series 1 digital circuit blocks. They will operate within the temperature range $-10\,^{\circ}\text{C}$ to $+60\,^{\circ}\text{C}$, with bit rates of up to 350,000 per sec. The lines are contained in steel cases measuring $9\frac{3}{4} \times 6\frac{1}{4} \times 1\frac{1}{2}$ in.

15WW 323 for further details

Step and Repeat Camera

PRIMARILY intended for the manufacture of photographic masks on maximum resolution material for use in the semiconductor industry is the Mark II step and repeat camera from W. Watson & Sons Ltd., of Barnet, Herts. The camera is, however, equally suited to the making of other graticules with repeated patterns within a 1×1 in field. The reduc-

tion lenses in this instrument are based on microscope objective lens designs and the fine focusing adjustment is pneumatically monitored by a sensitive air gauging system. Exposure is controlled by an automatic timer.

15WW 324 for further details

Micro-indicator Displays

A SMALL illuminated digital indicator, manufactured by Teldix—a German subsidiary of Telefunken and Bendix—is now available in the United Kingdom through Kynmore Engineering Co. Ltd., 19 Buckingham Street, London, W.C.2. Known as the Model MA-11, the standard unit contains the figures 0 to 9 and one set of colour bars, which when energized add a red bar to both sides of the displayed numeral. White-on-black characters, of 0.11 in wide by 0.22 in tall, are used and individually beamed on to a frontal screen by separate lamps. Any number of these units, which measure $0.34 \times 0.51 \times 1.82$ in, can be mounted side by side and stacked to make up decade displays. Standard black anodized alloy cases are made up for mounting up to six units, but as many as 400 units have been mounted together for a
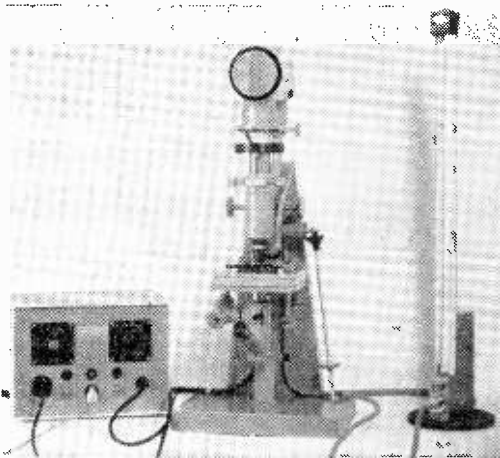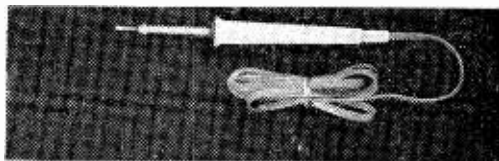
digital radar display. Special units can be provided to display letters, symbols, or any required sign.

These indicators operate from $1\frac{1}{2}$ volts and have a power consumption of 0.1 watt. The characters are visible at angles up to 30° from the normal in daylight and the average lamp life is quoted to be 500 hours. The units cost £8 each.

15WW 325 for further details

Soldering Irons

THE range of "Imperial" soldering irons made by Eldon Industries of Canada Ltd. are being imported by Lugton & Co. Ltd., of 209-212 Tottenham Court Road, London, W.1. These irons operate from any 250 volt supply and feature interchangeable bits, which incidentally screw directly on to the heating element, three interchangeable elements (25, 30 and 40 watt) and three different coloured interchangeable handles. A wide selection of iron-clad, gold-plated bits and tellurium copper bits are available for this range of irons. The cost of the irons is from £2 17s 6d to £3 1s 0d according to the bit selected. All parts are sold separately if so required.

15WW 326 for further details

Balanced Diode Video Attenuator

Transistor A.G.C. Circuit for 2 kc/s—20 Mc/s

By J. P. BÉNÉTEAU* and G. M. RIVA*

AUTOMATIC gain control (a.g.c.) with transistor circuits is always a thorny problem. The forward and reverse a.g.c. methods commonly used with transistors suffer from the disadvantages of relatively narrow gain control range, frequently severe signal distortion and bandwidth variation.

This paper discusses a simple, balanced, wideband video attenuator using the silicon planar epitaxial diodes FD-600. Some of the aforementioned difficulties are removed using this technique. An attenuation change of over 50 dB is possible, the distortion is kept below a few percent, while the bandwidth variation difficulty remains.

Circuit Descriptions

The diode attenuator is shown in Fig. 1. A balanced attenuator is preferable due to the constancy of d.c. level with a.g.c. application and harmonic distortion reduction. The attenuator results from the variable diode resistance in series with R.

The diodes are fed from current sources, transistors Q1 and Q2, which in turn are controlled by the transistor Q3.

The potentiometer R_5 controls the voltage on the base of Q3 which controls the collector and emitter voltages of Q3, which in turn control the diode current and thus the attenuation. In a practical circuit, the a.g.c. voltage would be applied directly to the base of Q3.

The equations governing the circuit are:

$$V_A = V - I_{e1}R_1 + V_{BE1} = V - I_{c3}R_3 \ldots (I_{e1} \simeq I_{c1})$$
$$V_B = V + I_{e2}R_2 + V_{BE2} = V + I_{c4}R_4 \ldots (I_{e2} \simeq I_{c2})$$
$$I_{c1} = \frac{I_{c3}R_3 + V_{BE1}}{R_1}, \quad I_{c2} = \frac{I_{c4}R_4 - V_{BE2}}{R_2}$$

If
$$R_1 = R_2, R_3 = R_4$$
and
$$R_3 \ll R_{IN} = r_{bb}' + (r_e + R_1)(h_{fe} + 1)$$
$$I_{c1} - I_{c2} = \frac{V_{BE2} + V_{BE1}}{R_1} \simeq 0, \quad \ldots \quad \ldots \quad (1)$$

(Since Q1 is a p-n-p and Q2 is a n-p-n)

From equation (1), it can be seen that substantially
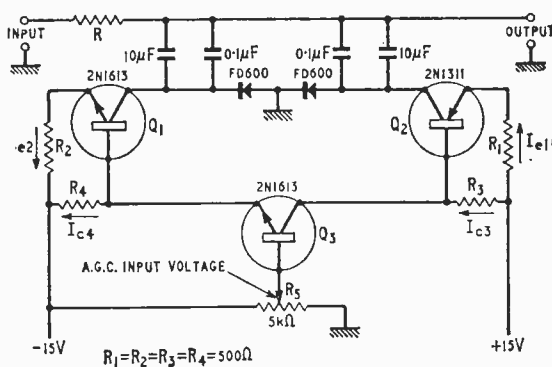
Fig. 1. Circuit diagram of video attenuator.

equal currents flow in each diode and if the diode characteristics are very similar as is the case with planar diodes, the system, is balanced. An epitaxial diode becomes useful because the series resistance at high current is smaller than in a conventional device and hence the range of attenuation becomes larger.

Performance

The performance of the attenuator is indicated in Figs. 2 to 5. The variation in attenuation obtained is shown in Fig. 2; for R = 100 ohms, the attenuation is about 35 dB in the current range of 0—40 mA, while for R = 1 kΩ, the attenuation is about 55 dB in the same current range. The frequency response for low and high frequencies are shown in Figs. 3 and 4. It can be seen that for attenuations of 20 and 40 dB, the response is within +1 dB from 170 c/s to 5.5 Mc/s and from 2 kc/s to 20 Mc/s respectively. The low-frequency response is a function of the input capacitance and is calculated in the appendix. Though the frequency response still varies with attenuation, it is important to note that it is sufficiently wide for many applications and that the effect of the variation may not be as serious as it appears.

The overall curve of ± 1 dB bandpass as a function of attenuation is shown in Fig. 5.

The total harmonic distortion at $f = 20$ kc/s is shown in Fig. 6 and is, of course, a function of the attenuation and output voltage. For small output signals (50 mV r.m.s.) it is in the range of 2-5%.

Conclusion

A simple and practical diode attenuator suitable for video a.g.c. systems has been presented. It is expected that this circuit may be of some usefulness in radar and TV circuits.

APPENDIX
Low-frequency Response of the Attenuator

See Fig. 7 which represents the equivalent circuit of the attenuator.
The transfer function represented in Fig. 8, can be written after a little manipulation:

$$\frac{e_2}{e_1} = \frac{1 + R_D Cp}{1 + (R_D + 2R)Cp} = \frac{1 + T_b p}{1 + T_a p}$$

where
$$T_b = R_D C$$
$$T_a = (R_D + 2R) C \qquad \text{and } p = j\omega$$

It can be seen from Fig. 8 that $1/T_b$ is the frequency below which the attenuation is no longer flat.

From Fig. 4, $T_b = 1/(2\pi \times 800) = 2 \times 10^{-4}$, while the calculated value using $C = 10\mu F$ and $R_D = 20$ (calculated from the assymptotic attenuation behaviour), also gives $T_b = 2 \times 10^{-4}$. This verification is a little startling when the tolerances of the electrolytic capacitor used are considered. $1/T_a$ can be calculated to be approximately 8 c/s.

*S.G.S. Società Generale Semiconduttori Agrate, Milano.
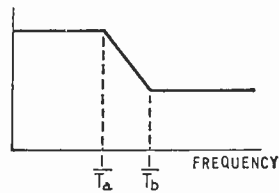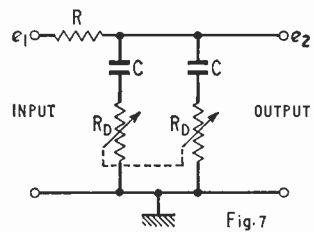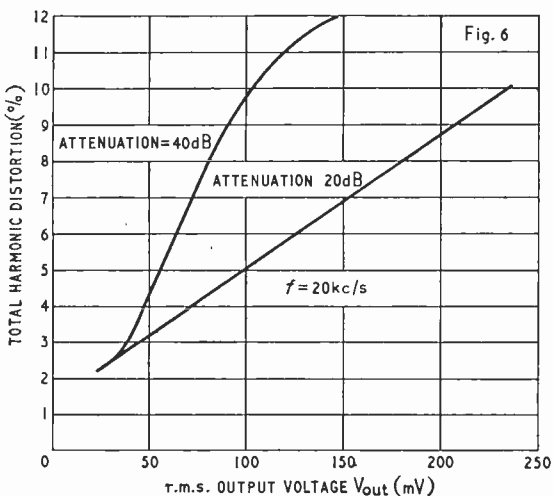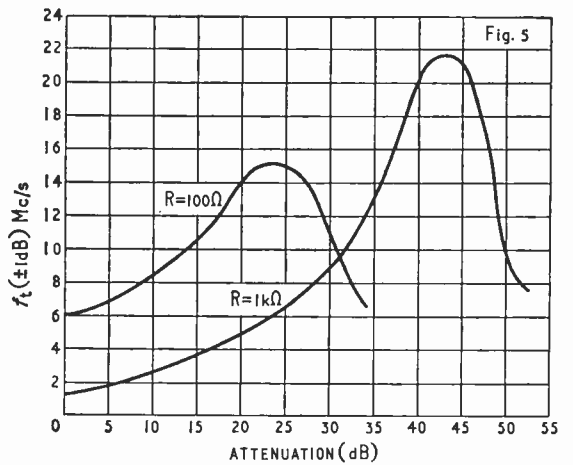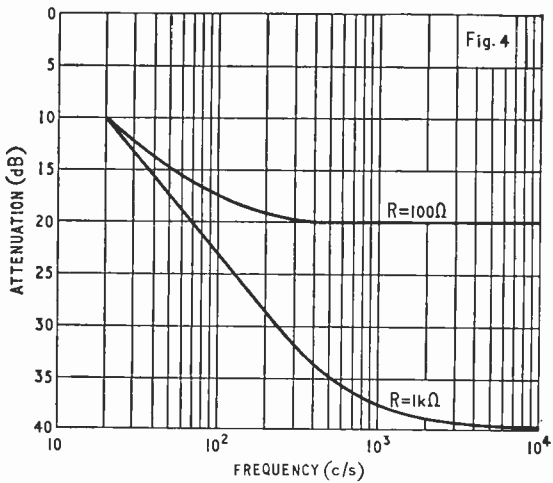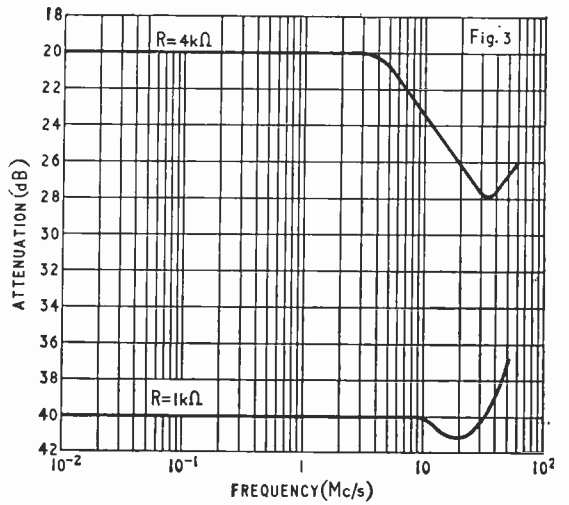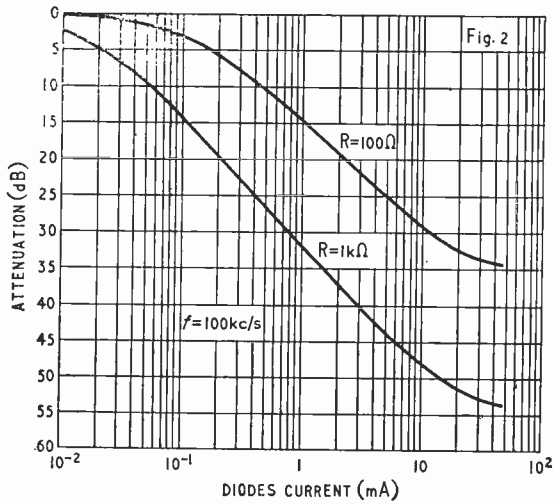
Fig. 2. Attenuator performance. Fig. 3. High-frequency performance of attenuator. Fig. 4. Low-frequency performance of attenuator. Fig. 5. Bandwidth/attenuation curve for different values of R. Fig. 6. Distortion of the attenuator. Fig. 7. Equivalent circuit of attenuator. Fig. 8. Transfer function of attenuator.

POST OFFICE TOWER AND ITS RADIO LINKS
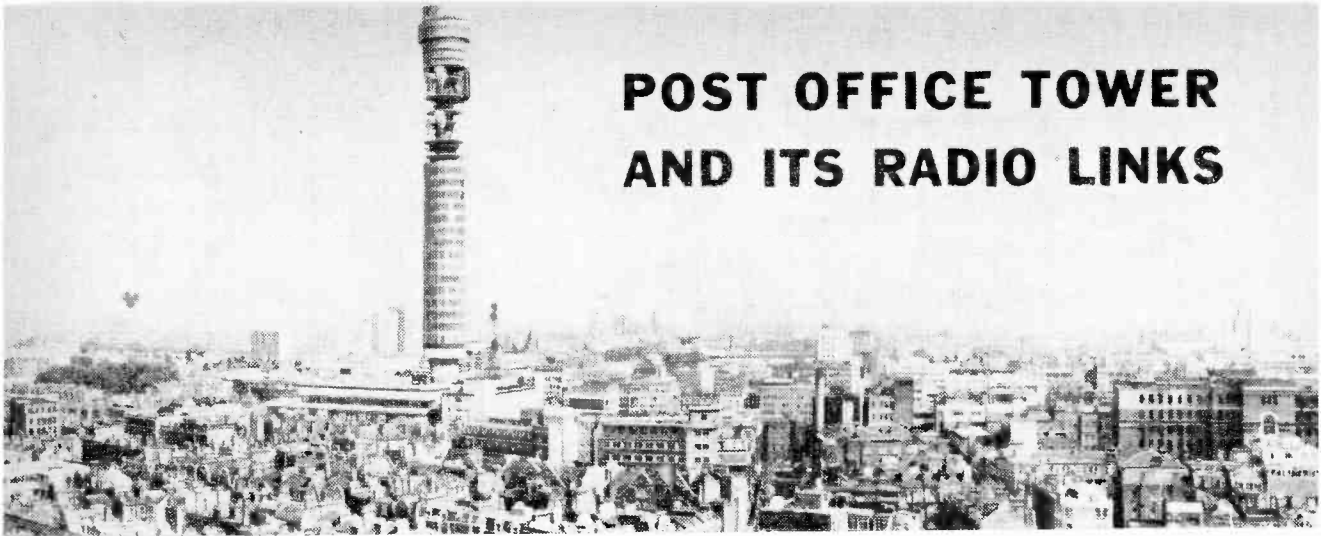
The network of radio links which the Post Office will need by 1970.
(Courtesy "Post Office Telecommunications Journal")
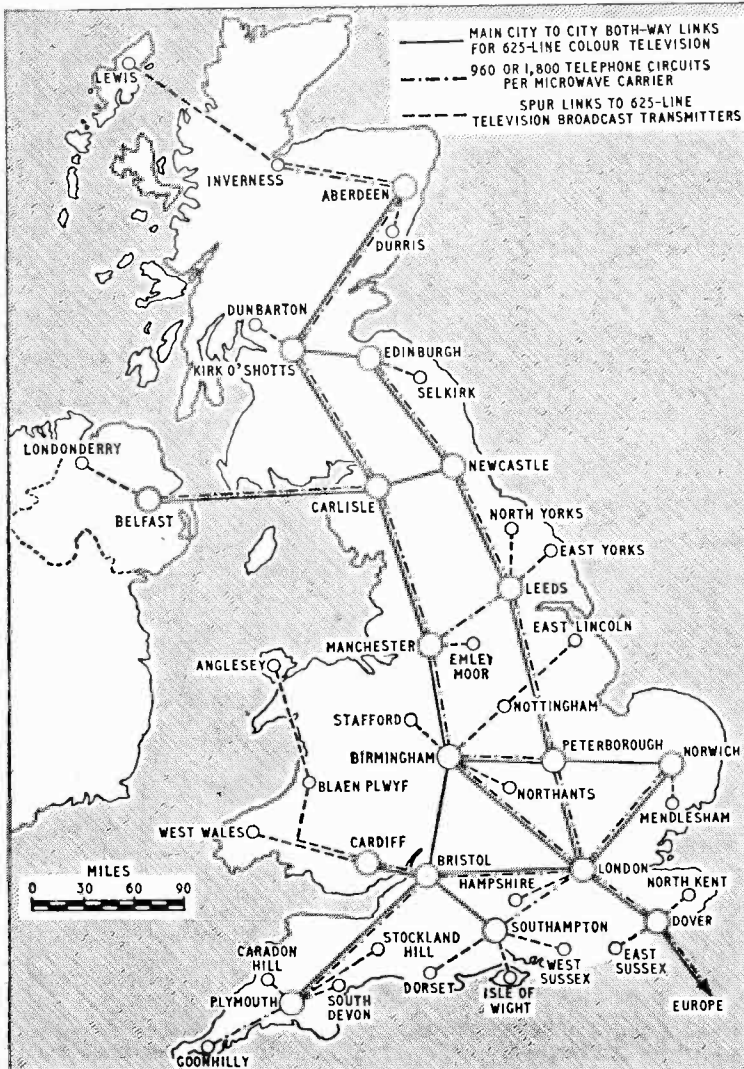
THE introduction of the 625-line 8 Mc/s television service, for which a network of 64 main stations is to be built, and the continued growth of the country's trunk telephone network, has necessitated the building by the Post Office of a country-wide network of point-to-point microwave links. At the hub of this network will be the 580ft Radio Tower near Tottenham Court Road, London, which is scheduled to be brought into service in January. The cylindrical tower, which is 50ft in diameter, will house radio transmitting and receiving equipment on most of the 16 apparatus floors and above these, in the section between 355 and 475ft above the ground, are four open cantilever galleries. Two of these, each 36ft high, carry the horn radiators and two 18ft galleries carry the paraboloids. The Tower, replacing a 150ft lattice tower, is considerably taller than any other building in London and can, therefore, provide an uninterrupted line-of-sight radio path to relay stations outside the metropolis. The top section of the Tower houses observation galleries and a revolving restaurant.

As will be seen from the map the network of radio links, planned for completion by 1970, extends from the Outer Hebrides to the satellite ground station at Goonhilly and to Dover for the cross-channel link. At Birmingham a similar tower, but smaller than that in London, is being built and at intervals of some 25-30 miles along the various routes relay stations are being set up. Most of these will be equipped with a standard lattice tower a model of which is illustrated opposite. At six sites, at main route intersections, concrete towers are being used; they

are at Charwelton, Northants; Heaton Park, Manchester; Pye Green, Staffs; Stokenchurch, Bucks; Sutton Common, Cheshire; and Wotton-under-Edge, Glos.

The top 70ft or so of these 250-300ft concrete towers will carry three circular cantilever platforms for the horns and dishes. The transmitting and receiving equipment at these stations is not housed in the towers but in an adjacent single-storey building.

Neither the standard lattice tower, nor the concrete towers can be called things of beauty. One would have imagined some artistry could have been introduced even in so utilitarian a structure as is instanced by many Continental towers, particularly those in Germany.

The microwave network will operate in six frequency bands between 1.7 and 11.7 Gc/s. Within these bands a total of 64 carriers in each direction will be used and each carrier can accommodate a 625-line television transmission or up to 1,800 telephone circuits, and later possibly 2,700. The total main route capacity of the Tower could be 150,000 simultaneous telephone conversations if used exclusively for telephony or alternatively 100 television links but it is understood that the most likely division will be 40 television channels and 100,000 telephone circuits. When the Tower is brought into service the first link will be to Birmingham and will provide 5,400 telephone circuits and an 8 Mc/s television channel. It will operate in the 6 Gc/s band.

The horns being used on this route were supplied by G.E.C. and consist of an offset parabolic reflector fed by a pyramidal horn with its axis vertical. It has the advantages of a high back-to-front ratio, low side lobe radiation, and an excellent input match over a wide frequency band. The sides and back are enclosed by aluminium alloy sheeting, the transmission aperture is closed with a Hypalon coated Terylene window, and the whole is slightly pressurized to exclude moisture. The aerial has a radiating aperture area of 140 sq ft and a gain at 6 Gc/s of 45 dB relative to an isotropic aerial. There is a smaller version with an aperture of 70 sq ft and a gain of 42 dB. Up to four transmitters and four receivers can be connected to a single horn aerial by the use of two sets of three-port multiplexing circulators.

Incidentally, provision has been made for the installation of 24- and 50-volt batteries as future equipment will be solid-state.
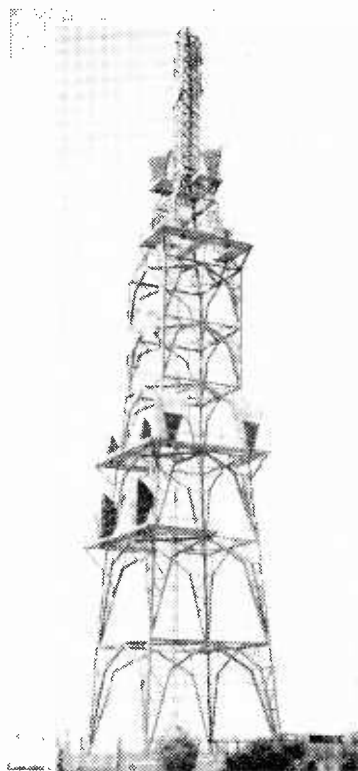
Model of the standard lattice tower to be used to carry the horns and paraboloids at repeater stations.

Six concrete towers, like this one at Stokenchurch, Bucks., are being built at the intersections of main routes.
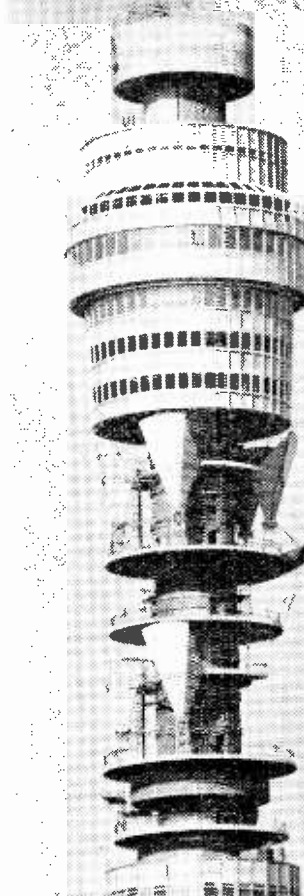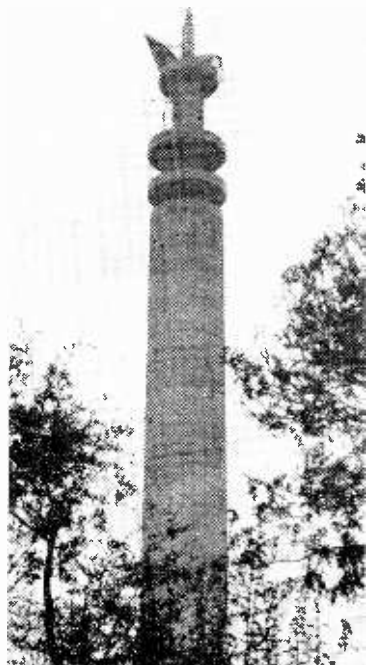
The four aerial galleries at a height of between 355 and 475ft on the London Tower. A 40ft mast atop the Tower will carry a storm warning radar scanner for the Meteorological Office.

A non-standard tower carries the horns and dishes at the Riddings Hill, Dumfriesshire, relay station.
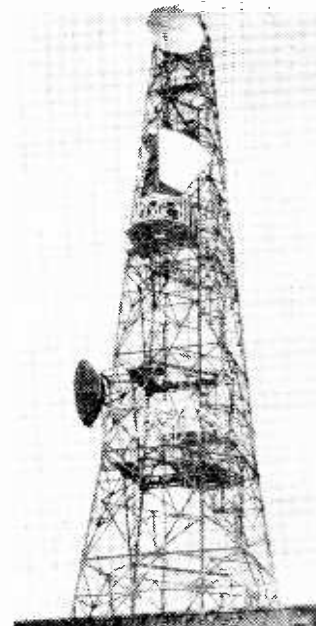
INTERNATIONAL RADIO

THIS year's exhibition, officially opened at the Seymour Hall, London, by Mr. E. D. Whitehead, M.B.E., Director of Electronics Production (Radar), M.o.A., was universally voted "the best ever" by the usually critical amateur radio fraternity. It was also notable for a major breakthrough in the "politics" of amateur radio. For the first time the G.P.O. gave blanket permission for any overseas visitor who holds a current transmitting licence in his own country to operate the exhibition stations GB3RS (160/80 metres) and GB2VHF (4/2 metres) under the supervision of the licensees. Permission has been given in the past, but only for "nominated" aliens. This new concession is regarded as a major step towards the introduction of full reciprocal licensing facilities in the U.K. which would allow holders of British licences to operate in all countries where reciprocal licensing is agreed between governments.

The first overseas visitor to take advantage of the new concession was Chet Lambert, W4WDR, from Chattanooga, Tennessee. No newcomer to reciprocal licensing, as he holds current licences F7CL and 9Q5WF in France and the Congo in addition to his U.S. licence, he was delighted to be first in the field. Hard on his heels, however, were DPØBM, PAØEO, W3FIU, E14Q and E19V, and dozens more overseas operators had been on the air by the time the exhibition closed.

New, too, was an increased emphasis on education. Mr. G. M. C. Stone, President of the R.S.G.B., in his opening address stressed this. "We feel that in the past our Society has not done enough to encourage amateur radio as a hobby among boys and girls and we are now looking at ways and means of taking amateur radio into secondary schools and other schools of similar level. In the first instance we plan to do this through schoolmasters who are members of our Society." He then spoke of the new Education and Training Committee of the R.S.G.B., the Christmas lecture for young people to be introduced, if possible, this year and drew particular attention to the Roding Boys' Club, who were exhibiting for the second year running, as an example of what young people can do under good leadership.

Despite criticisms that ham radio has gone "commercial"—and there was plenty of evidence for this—home-constructed equipment was shown in quantity and, generally, was of high quality. The u.h.f. tunnel diode amplifier for the 70-cm band, one of the two items which won the Horace Freeman Trophy for S. F. Weber, typifies the modern approach of the new generation of home constructors and the greatly increased interest in the higher frequencies. The amplifier has a loaded $\frac{1}{4}\lambda$ coaxial line with a 1N2969A tunnel diode across the end to form a parallel type amplifier matching a 75 ohm source and load. Mr. Weber is one of the new brigade of G3-three letter calls licensed for telephony only on 420 Mc/s and higher amateur frequencies.

Judging was so difficult that when the show opened no decision had been reached on the winner of the Exhibition Organizer's Silver Plaque. After a further day's deliberation the award went to H. Rogers, G3NHR, for his 1.8 and 144 Mc/s transmitter.

Between the whole-hog home constructors and those who buy off-the-shelf ready-mades or even tailor-mades lies an increasing body of enthusiasts who enjoy building equipment but haven't the time or, perhaps, the technical expertise to start from scratch. Daystrom Ltd. meet the needs of these in-betweens with their range of kits for the home constructor. Although these can be bought fully assembled for an extra charge, it would appear that their business with radio amateurs is almost 100 per cent home assembly for transmitters and receivers. With instruments, however, some 50 per cent are purchased already assembled, many of these going to industrial concerns.

K. W. Electronics also supply some of their equipment in kit form, if required. It seems, then, that although there is a marked increase in the use of commercialy built equipment, a hard core of radio amateurs, both newcomers and old-timers, still construct either a part or the whole of their stations.

The increasing use of single sideband telephony is, perhaps, the main driving force behind the boom in commercially built equipment. The additional initial technical requirements on both transmission and reception over simple a.m. are severe enough in themselves to daunt all but the most stout-hearted home constructor without the extra test gear needed for satisfactory adjustment and alignment, most of which would not be to hand in the average home workshop.

To meet the demand six companies showed a range of s.s.b. equipment at prices which would meet all pockets. Of the British contributors, K. W. Electronics Ltd., now veterans at the s.s.b. game, introduced their new 2000A transceiver. This is a higher-power version of the well-established 2000 of which 65 per cent of production has been exported to Europe, the Commonwealth and South Africa. The new model has 180 W
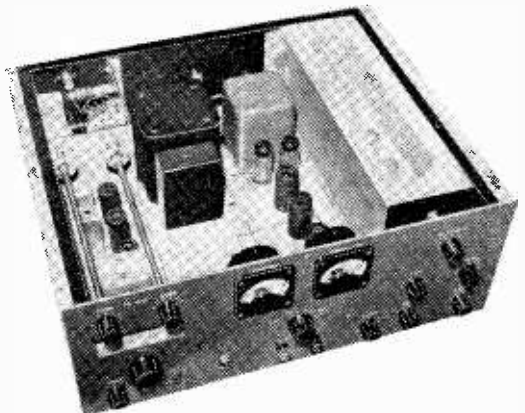
1.8 and 144 Mc/s transmitter by H. Rogers (G3NHR) which was awarded the Exhibition Organizer's Plaque for the best home-constructed exhibit.

COMMUNICATIONS EXHIBITION—1964

p.e.p. output, double that of the 2000, covers all amateur bands from 1.8 to 28 Mc/s, and is suitable for either fixed or mobile use. Another new line from K.W. is the KW600 linear amplifier designed to match the 2000. It employs a 572 A valve amplifier operating in class AB1 with grounded grid and screen and requires 35-40W drive from the transmitter/exciter. The built-in power supply produces 2000 V from hash-free silicon rectifiers to run the linear at over 500 W on c.w. and s.s.b.

For those who want something a little more economical K.W. have introduced the "Vespa" s.s.b. transmitter at a price of £97 10s. This transmitter has all the essentials for good performance, such as a mechanical filter, in the same circuit as the 2000. Three bands are provided, 1.8, 3.5 and 14 Mc/s, the most popularly used for s.s.b. The power amplifier runs at 90 W on s.s.b. and 75 W on c.w. Like the 2000, power is reduced on 1.8 Mc/s to meet the power restriction on this band. The cabinet matches the K.W. 77 receiver.

The itch that overcomes many h.f. band operators to get on v.h.f. is satisfied by the K.W. 201 Transverter. This is an add-on unit for use with an existing s.s.b. transmitter with separate receiving facilities. Basically it is a linear transmitting and receiving frequency converter. An input of 0.5 W s.s.b./a.m./c.w. at 28 Mc/s is converted by mixing to 144 Mc/s and then linearly amplified to 200 W p.e.p., s.s.b. and c.w., or 60W input a.m. On receive, the 144 Mc/s signal is converted to 28 Mc/s as in a conventional converter. One advantage of this equipment is that it is effectively v.f.o. controlled on the 2-metre band. Nuvistors in the receiving chain give a claimed 3 dB noise figure.

A newcomer to the show, s.s.b. Products of Edward Street, Derby, showed their Sphinx transmitter, now in production, and both the Sphinx receiver and Pyramid linear amplifier in prototype form. Three models of the Sphinx transmitter are available from a 1.8 Mc/s only unit at £48 to a 4-band unit 1.8 Mc/s to 14 Mc/s at £75, including built-in power supplies. A 7 Mc/s output is provided for c.w. and a.m. only but all modes are obtainable on the other bands. Power is 40-70W p.e.p. and speech quality on s.s.b. is stated by the manufacturers as "No robot-like sounds—we suppress the carrier, not the quality." A remark which makes one wonder what one's own transmission is like!

Back in the medium/high price range was the CTR-1 transceiver by Courier Communications. This £175 equipment (power supply extra) covers all six h.f. amateur bands with 200W p.e.p. and 150 c.w. The big feature of this equipment is the provision of split frequency operation over wide limits instead of the nominal few kc/s normal on other transceivers. Like the other s.s.b. equipment mentioned here it is suitable for mobile or fixed operation. In the mobile role the CCU-1 control unit incorporates a six-band field strength meter, main on-off switch, send-receive switch, loudspeaker socket and p.a. bias control.

The Heathkit s.s.b. equipments shown in the U.K. for the first time by Daystrom, Ltd., were the de luxe SB300E receiver and SB400E transmitter retailing in the U.K. at £133 14s and £165 4s (not including the recently imposed import surcharge). Both units tune all amateur frequencies from 3.5 to 30 mc/s, have linear master oscillators with 1 k/c dial calibration, 100 kc/s per dial revolution and a claimed frequency stability of better than 100 c/s drift per hour after a 20 minute warm-up period. The transmitter has an output of 180W p.e.p. on s.s.b. and 170W on c.w.

Among a number of other American s.s.b. equipments the new NCX-5 all-band transceiver by the National Radio Company, Inc., shown by Ad Auriema, Ltd., deserves comment. It has a linear solid-state v.f.o. with no warm-up drift, digital read-out of frequency to 1 k/c on all bands with an additional counter calibration to 100 c/s. Stability is 100 c/s in any ten-minute period from switch-on including ± 10 per cent variation of supply voltage. Among the features is an 8-pole crystal lattice filter. Power is 200W input s.s.b. and c.w., and 100W a.m. and receiver sensitivity is 0.5 mV for 10dB signal/noise ration on s.s.b. It costs £255 10s 1d in U.K.

Green & Davis, better known for v.h.f. equipment, have now entered the s.s.b. market with their ½ kW PGLA1 linear amplifier and 20-2 144 Mc/s s.s.b. transverter. The latter unit converts existing 14 Mc/s equipment to 144 Mc/s with a power greater than 120W p.e.p. on s.s.b. and with a receiver noise factor of 3dB. The exciter is tuned to exactly 14 Mc/s and then one of three available transmitting frequencies selected on the transverter. On receive the entire band can be tuned. Drive requirement is 20W from the 14 Mc/s transmitter or transceiver.

The PGLA1 linear amplifier includes a power pack which contains 12 silicon rectifiers. Output in excess of 250W peak is obtainable on all bands from 3.5 to
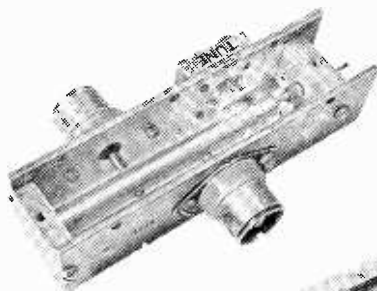
Left: 435-Mc/s tunnel diode amplifier which won the Horace Freeman Trophy for S. F. Weber.
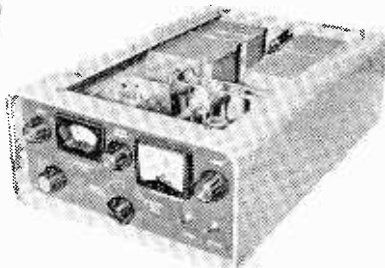
Right: Green & Davis transverter for transmission and reception on 144 Mc/s using basic 14 Mc/s equipment.
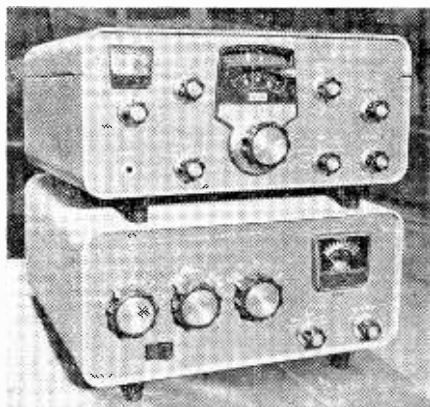
Two new items for the s.s.b. enthusiast from the Heathkit (Daystrom) American range, the SB 300 receiver and (below) SB 200 linear amplifier.

28 Mc/s. Forced air cooling is used and the unit is priced at £87 10s.

While still discussing h.f. equipment it is as well to make a few comments on commercial receivers. These are by no means getting lost among the spate of transceivers. Although there was great disappointment at the non-appearance of the all-solid-state National HRO 500, which could not be obtained in time for the show, it was appreciated that for most amateurs the U.K. price, over £600, would make it a curiosity more than anything else. But other HRO'S were on show though a little more costly than the old, pre-war faithfuls.

Also in the luxury class were Racal's RA17/RA71 receivers. No fewer than 10,000 of these superlative receivers have now been made and the 9,999th, a RA71, was being offered in a "guess the number of joints" competition, the proceeds going to cancer research.

The HQ17OA by Hammarlund (K.W. Electronics are agents) has solved those messy converter problems. In addition to full h.f. band coverage with triple conversion and all facilities, the new receiver has built-in 2 and 6 metre converters with separate Nuvistor front ends for both bands.

The new Eddystone EA12 amateur band receiver and companion EP20 panoramic display attracted great attention. Linear tuning is employed, limited to 600 kc/s, which entails four selections of the band switch to cover the 28 Mc/s band. The scale can be read to an accuracy of 1 kc/s on any band, stability is 100 c/s per hour after warm-up and short-term drift typically less than 20 c/s. Diecastings are used for the front panel and coil box and the general construction is the solid job expected from this long-established company. The 888A is still in demand as proved by Eddystone's trade-in offer of £55 allowance if you buy an EA12. Factory reconditioned 888As will be available from Webb's Radio.

Heathkit didn't have a new receiver on show. It seems that the RA-1 amateur band kit at under £40, plus optional plug-in crystal calibrator, exactly suits the lower-priced home-constructor market and for the general coverage short-wave listener the RG-1 and Mohican are still on offer and selling very well.
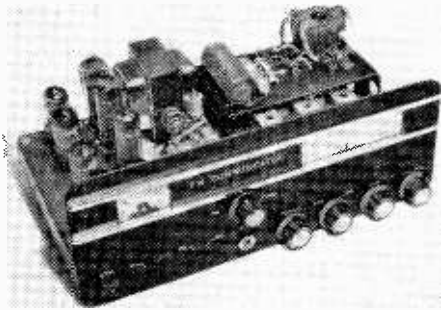
T.W. Communicator, an integrated station which won the Manufacturer's Silver Plaque for T. Withers (Electronics).

Apart from transceivers, completely packed stations consisting of basic units are coming into vogue. T. Withers (Electronics) won the Manufacturer's Silver Plaque with just such a unit. Called the TW Communicator, it is essentially the well-established TW2, 4 or 160 metre transmitter together with the appropriate transistorized receiver, power supply and aerial switching, all packed into a neat case 12 in × 4½ in × 7 in. The basic unit is designed for mobile or portable use from a 12 V d.c. supply but a mains power unit is available for fixed use. If you just want to listen the consumption at 12 V is only 25 mA. With 15 watts input of fully modulated (a.m.) carrier, power consumption is 5 A.

And the trend towards tiny equipment was also evident in new items from Labgear and also Codar Radio. The table-top transmitter of the immediate post-war years needed a table with reinforced legs to carry the weight. A 1964 table-topper from Codar or Labgear would grace and not overburden the most delicate of occasional tables. The Codar unit, measuring 8½ in × 5 in × 4 in, has 12 W input, plate and screen modulated, tunes the 1.8 and 3.5 Mc/s bands and comes ready to use for £16 10s. A matching power unit with control and aerial switching and flashing neon to indicate "standby" or "on-the-air" costs £8.

The Labgear One Sixty Mobile/Fixed transmitter, rushed to the show, is the major unit in their new One Sixty series. This is a four-valve unit with high-level modulation, compact enough to fit under the dash of the car. Special fitting brackets can be obtained for most makes of car. A companion 1.8 Mc/s converter adapts any car radio which operates on the m.w. band for 1.8 Mc/s reception. These two units, at £15 15s and £5 5s respectively, are an economical route to "getting mobile" without tears.

The mobile market is perhaps the fastest growing of all. The G.P.O. has now issued nearly 2,000 mobile licences and most manufacturers are catering for this demand. So is the Amateur Radio Mobile Society, whose "Project Papoose" starts this month. This will be a new home-constructors equipment using Brayhead and Newmarket transistor modules. The critical tuning bugbear of mobile aerials has been overcome by Labgear who showed for the first time a permeability-tuned whip giving continuous tuning from 1.8 to 14.4 Mc/s. It has been developed from their commercial h.f. mobile whip used in the Services and by many other organizations.

The whole aerial is of hollow glass fibre construction with the radiating element on the inside and the loading coil is copper tape wound on and fused into the glass fibre. Tuning is by a ferrite core driven by a Bowden cable from inside the car.

Other new aerials on view were the K.W. VEE VEE two- or three-element beam for 14 and 21 Mc/s and the redesigned range of J-Beam aerials. Minimitter (1964) Ltd. report they will soon be offering the well-known "Birdcage" aerial for 144 Mc/s operation.

The new G8-three letter calls as well as u.h.f. old timers are being served by new equipment from T. Withers (Electronics). The 70-cm converter has silver-plated on brass chassis and uses the GEC A2521 grounded-grid triode with a condensed trough line in the output circuit and a grounded-grid 6DS4 Nuvistor as the mixer rather than a diode to obtain conversion gain. The "TW 70" 70-cm transmitter gives an output of 12W and has tuned anode and grid lines.

One brand new equipment exciting attention was the Uniframe Modular Rack System developed by C & N (Electrical) Ltd. which enables instrument cases or six-foot racks to be assembled as required. The beauty of the system is that a unit, once assembled, is expansible at any time to a larger unit. The six-foot rack and cabinet, when dismantled, folds into a volume only 13% of its assembled size.

Spectrum Electronics Ltd., another new exhibitor, found their digital clocks, with either 12- or 24-hour presentation, a popular draw. Available either as basic clock movements or with a stylish case complete with station call sign, these were obviously being noted by radio amateurs as items for their Christmas present list. Spectrum also featured neat Italian-made wavemeter/grid-dip meters and a.m. signal generators. Both were fully transistorized with internal batteries, and so fully portable.

The Royal Navy, the Army—in the shape of the 65th Signal Regiment, T.A., and their sister regiment, the 92nd Signal Regiment, A.E.R.; the British Amateur Radio Teleprinter Group; the B.A.T.C. and many other exhibtors and exhibits not mentioned here all helped to make this exhibition the great success it unquestionably was. If any proof were needed it is in the fact that bookings have already been placed by many exhibitors for the 1965 show—and most are booking bigger stands.

DESIGN FOR A SINGLE-CHANNEL TRANSISTOR F.S.K. OSCILLATOR

By R. B. C. COPSEY,* A.M.I.E.E.

COMMUNICATION of information on both radio and line links is often achieved by moving the carrier frequency by a small amount in sympathy with the keying voltage. This is known as frequency shift keying (f.s.k.) and has an advantage over on/off keying in that the signal-to-noise ratio on both "mark" and "space" is similar. With on/off keying the signal-to-noise ratio on "space" is obviously zero.

It is customary to make the f.s.k. oscillator operate at a convenient frequency, and to change this frequency in a mixer to the frequency required to drive the transmitters. Many methods have been employed to produce a small shift in carrier fre-

quency. Variable reactance valves, "pulled" crystals and variable permeability inductors represent a few well-tried methods. More recently circuits have been devised using semiconductor junctions as voltage-dependent capacitors.

The oscillator described in this article is one which has been developed to give a constant centre frequency, irrespective of deviation. It was developed for use with a v.l.f. transmitter where the maximum frequency deviation required was $\pm$ 50 c/s, and for this reason a low initial frequency of 1000 c/s was chosen, but the circuit constants could be modified to give a much higher initial frequency and greater deviation.

Basic Oscillator Circuit

Fig. 1 shows the basic circuit of the oscillator, omitting, for simplicity, the frequency shifting circuit. In some respects it is similar to the valve oscillator described in Ref. 1. It consists of a three-stage amplifier with overall positive feedback. In the first stage R_2 is equal to R_3, and the transistor operates as an emitter follower with substantially unity voltage gain. As the collector load resistor R_2 is equal to the emitter load resistor R_3, there appears at the collector of Tr1 an a.c. voltage of similar magnitude to that at the emitter, but 180° out of phase.

In the vector diagram, Fig. 2, V_e represents the instantaneous a.c. voltage at the emitter of Tr1

Fig. 1. Circuit of basic oscillator.

*Redifon Limited, Crawley.
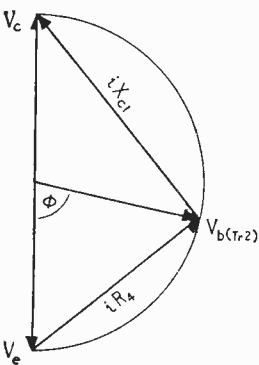
Above: Fig. 2. Vector diagram of voltage at emitter of Tr1 in Fig. 1.
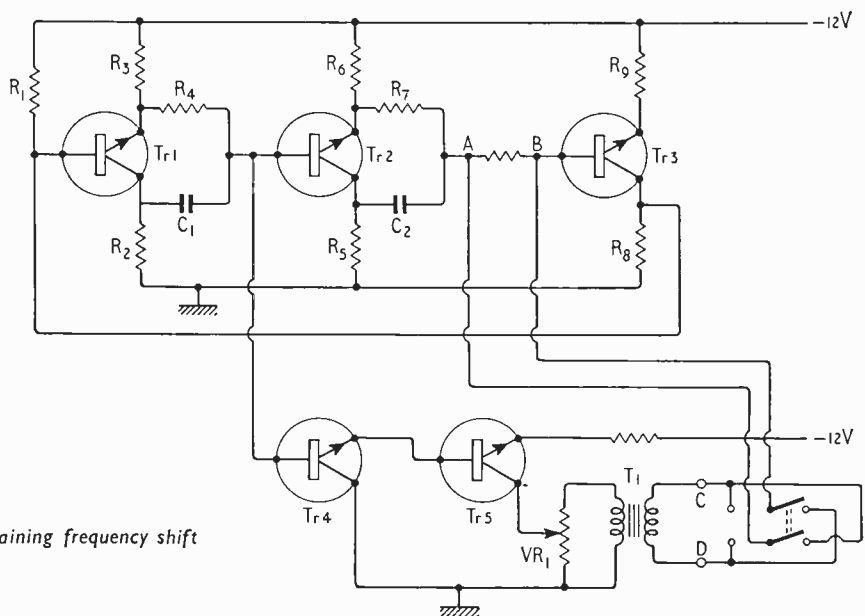
Right: Fig. 3. Method of obtaining frequency shift in the circuit of Fig. 1.

D

(which is, of course, in phase with the base voltage of Tr1), and V_o represents the voltage at Tr1's collector. Postulating a current i flowing from the emitter connection of Tr1 via R_4 and C_1 to Tr1's collector connection, gives the vectors iR_4 and iX_{c1}, the resultant of which is V_b(Tr2). The magnitude of V_b(Tr2) is independent of frequency, but its phase angle ϕ by which it leads V_o depends on the reactance of C_1. At the frequency at which the reactance of C_1 is equal to the resistance of R_4 there is 90° phase shift.

The second stage, Tr2, is identical to the first stage and produces a further 90° phase shift at the operating frequency, making a total of 180° phase shift in the first two stages. The third stage, Tr3, is a conventional collector-follower stage which produces a further 180° phase shift, and this feeds its output voltage back to the input of the amplifier,
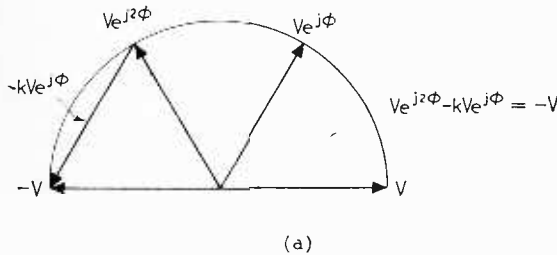
(a)
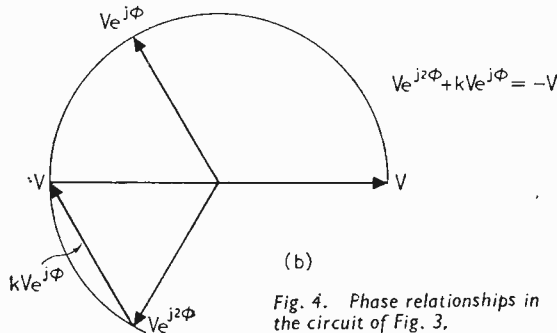
(b)

Fig. 4. Phase relationships in the circuit of Fig. 3.

Tr1's base. The resistance of R_8 is greater than that of R_9, so that Tr3 provides some voltage gain, and as the overall voltage gain is initially greater than unity, and the phase shift is 360°, the circuit oscillates. The build-up of oscillation amplitude is limited by the bottoming of Tr3.

Frequency Shift Circuits

Frequency shift is achieved by adding some fraction, k, of the output of the first stage to the output of the second stage (Fig. 3). This is done by means of the two stage amplifier Tr4 and Tr5, which feeds a signal via a transformer T1, and a " reversing switch " to the points A and B in Fig. 3. The injected signal is therefore in series with the input to Tr3. Its level is controlled by VR_1, and its phase may be reversed by the " reversing switch."

If V (Figs. 4(a) and 4(b)) represents the input signal at the base of Tr1, then $Ve^{j\phi}$ represents the input to the base of Tr2, since the voltage gain of Tr1 is substantially unity. Similarly, the output of the second stage of the oscillator Tr3 (point A in Fig. 3) is represented by the vector $Ve^{j2\phi}$. To the output of Tr2 is added a fraction, k, of the output of Tr1, so that the input to Tr3 (point B) is $Ve^{j2\phi} \pm kVe^{j2\phi}$. The circuit oscillates at the frequency which makes the circuit conditions such that there is 180° phase shift between the input to Tr1 and Tr3, i.e. such that

$$Ve^{j2\phi} \pm kVe^{j\phi} = -V \qquad .. \qquad .. \qquad .. \qquad (1)$$

The vector diagrams Figs. 4(a) and 4(b) illustrate this, and show that to satisfy equation (1), k must be equal to $-2\cos\phi$.

From Fig. 2 we see that $\phi = 2\tan^{-1}\dfrac{R}{v} = 2\tan^{-1}\omega CR$

$$\therefore \omega = \frac{\tan\phi/2}{CR}$$

When there is no deviation, k = 0, and $\tan\phi/2 = 1$,

$$\therefore \omega = \frac{1}{CR}$$

Fig. 5. Variation of arithmetic mean frequency with valve of k. expressed as a percentage of the deviation from nominal frequency.
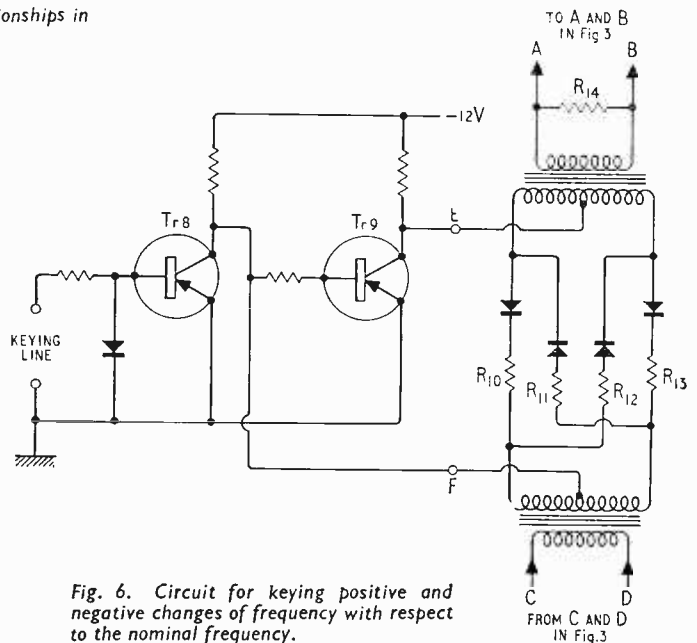
Fig. 6. Circuit for keying positive and negative changes of frequency with respect to the nominal frequency.
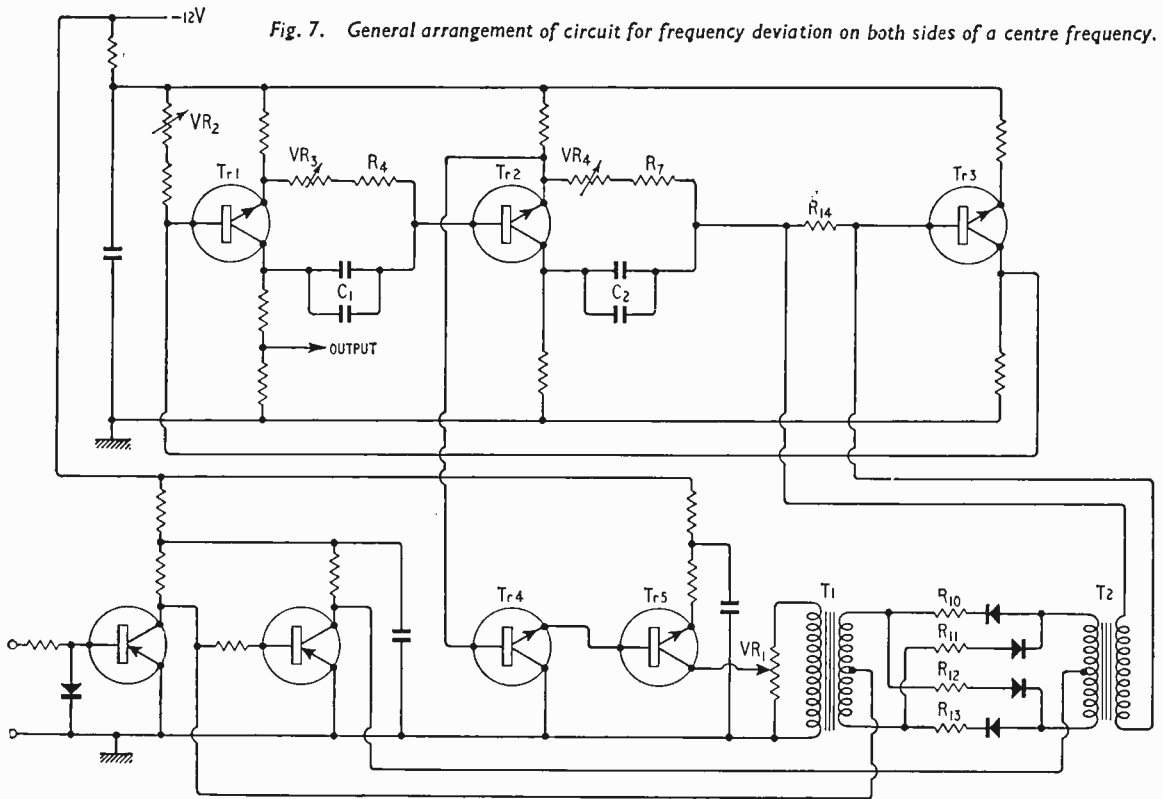
www.americanradiohistory.com

If we call the undeviated frequency ω_0, and the deviated frequency ω^d, then $\left(\dfrac{\omega_d}{\omega_o}\right)^2 = \tan^2\phi/2$

$$\therefore \left(\frac{\omega_d}{\omega_o}\right)^2 = \frac{1 - \cos\phi}{1 + \cos\phi}$$

Multiplying numerator and denominator by 2 gives
$$\left(\frac{\omega_d}{\omega_o}\right)^2 = \frac{2 - 2\cos\phi}{2 + 2\cos\phi}$$

Since $k = -2\cos\phi$
$$\frac{(\omega_d)^2}{(\omega_o)^2} = \frac{2 + k}{2 - k} \qquad .. \qquad .. \quad (2)$$

A positive value of k gives an increase in frequency and a negative value of k gives a decrease in frequency. It can be seen from equation (2) that for any value of k, if the sign of k is reversed, the geometric mean of the upper and lower frequencies will remain constant, but the arithmetic mean frequency will gradually increase as k is made bigger.

Fig. 5 shows the change of arithmetic mean frequency, expressed as a percentage of the deviation from nominal frequency, plotted against k. This shows that for values of k below, say, 0.2 the shift of mean frequency (5%) may be tolerable.

In cases where this change of mean frequency with deviation is not permissible, it is possible to reduce the change by arranging that the value of k in the positive sense (to increase frequency) is less than the value of k used in the negative sense (to decrease frequency).

Fig. 6 shows how this has been achieved. The keying input signal has two conditions, positive or negative with respect to earth. The resistor in series with the base of Tr8 limits the base current when the keying polarity is negative, and the diode

limits the reverse base-emitter voltage when the keying polarity is positive. Tr8 and Tr9 are p-n-p transistors with identical collector loads, and are arranged so that when Tr8 is bottomed, Tr9 is cut off, and *vice-versa*. This means that the potential between the points E and F reverses with the keying, point E being positive with respect to point F when the keying signal is positive, and E is negative with respect to F when the keying signal is negative. The four diodes in Fig. 6 operate as a reversing switch, and the attenuation due to R_{10} and R_{13} feeding the load R_{14} is arranged to be greater than that due to R_{11} and R_{12}.

Practical Considerations

Fig. 7 shows the relevant parts of the circuits as finally combined.

In the particular example described in this article the centre frequency of the oscillator is 1000 c/s and it is designed to provide up to $\pm$ 50 c/s deviation, i.e. from 950 to 1050 c/s, which requires k to have a maximum value of 0.1. VR_1 is the deviation control.

Tr1-Tr5 inclusive are high-beta silicon n-p-n diffused mesa transistors. V.h.f. transistors were chosen because of their negligible internal phase shift at low frequencies.

The frequency determining resistors are high-stability low-inductance wirewound resistors with a temperature coefficient of less than $\pm$ 0.0015% per °C.

The capacitors C_1 and C_2 are each made up of a silvered mica capacitor in parallel with a polystyrene capacitor. The mica capacitor has a positive temperature coefficient, the polystyrene capacitor has a negative temperature coefficient and their capacities

are proportioned so that the combination has a substantially zero temperature coefficient.

The feed from the output of the first stage to Tr4 has been taken from the emitter of Tr2 as this is a lower impedance point than its base.

VR_2 provides an adjustment of the overall gain, and VR_3 and VR_4 provide sufficient adjustment of the resistance of the frequency determining resistors to take up the manufacturing tolerances of C_1, C_2, R_4 and R_7.

The output of the oscillator is taken from a low impedance tap on the collector load of Tr1.

The switching currents flowing in the two halves of the centre-tapped windings of the switching transformers T1 and T2 must be carefully balanced to avoid transients at the instant of switching. The resistors R_{10} to R_{13} are made to a tolerance of $\pm$ 0.1%, and their resistance is high enough to swamp any differences in the forward resistance of the diodes. It will be noted that VR_1 has been connected so that the resistance across the primary winding of T2 is independent of the setting of the deviation control, VR_1, thereby avoiding any change in the phase shift in the transformer itself.

Performance:-The performance of this oscillator may be summarized as follows:—

(a) The maximum variation of the arithmetic mean of the mark and space frequencies from the nominal frequency of 1000 c/s is less than 0.2 c/s as the deviation is varied from 0 to 50 c/s.

(b) The variation in amplitude of oscillation between mark and space frequencies is less than 0.1 dB.

(c) The frequency stability with change in temperature is less than 0.25 c/s for a temperature variation from 15°C to 30°C.

Bandwidth:-The bandwidth occupied by an f.s.k. signal depends on three factors: (a) the frequency deviation, (b) the keying speed, (c) the time taken to change from " mark " to " space ".

The oscillator described in this article was designed for: (a) a maximum deviation of $\pm$ 50 c/s, (b) a maximum keying speed of 50 bauds, (c) a " rise time " of less than 2 milliseconds, to change from " mark " to " space ".

It is of interest to calculate the bandwidth occupied when the oscillator is keyed to its maximum deviation at its maximum rate. It is well known that a perfect square wave contains *all* the odd harmonics of the fundamental, and would therefore require an infinite bandwidth.

Fig. 8 shows the required variation of frequency with time, and this trapezoidal waveform has a rise time from zero to full amplitude of 1 millisecond.

The general expression for a square wave is:

$$v = \frac{4V}{\pi}(\sin \omega t + \tfrac{1}{3}\sin 3\,\omega t + \tfrac{1}{5}\sin 5\,\omega t + \text{etc.})$$
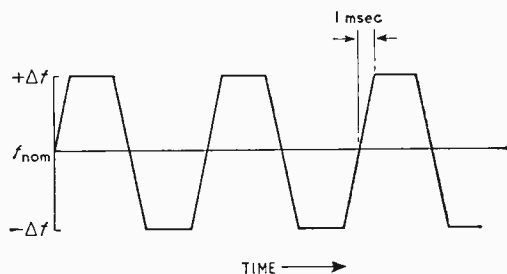
Fig. 8. Trapezoid waveform with a rise time of 1 millisecond.

In this case, the amplitude of v is required to be V, or greater, in a time $t = 1$ millisecond. At 50 bauds, the fundamental frequency is 25 c/s and 1 millisecond represents 9°. A little simple arithmetic shows that the first six terms in the series are necessary to achieve the desired rise time. The sixth term is the one dealing with 11 ωt, so that the highest necessary modulating frequency is 11 times the keying speed of 25 c/s, which is 275 c/s.

The problem may now be regarded as a frequency modulation one, in which the deviation is $\pm$ 50 c/s, and the modulating frequency 275 c/s, which gives a modulation index of $\frac{50}{275} = 0.182$. At this low modulation index[2] only the first-order sidebands have a significant amplitude, but it is essential that these sidebands should be present at their proper amplitude and phase relative to the carrier, in order to achieve the desired rise time.

The lower-order harmonics of the keying frequency produce sidebands at 25-c/s intervals on each side of the carrier. The fundamental component of the keying frequency, 25 c/s, has a modulation index of 2, and at this modulation index only the first four sidebands have a significant amplitude, requiring a bandwidth of $\pm$ 100 c/s. Similar calculations for the other harmonics of the keying frequency show that in no case is the bandwidth required greater than $\pm$ 275 c/s therefore the bandwidth which is necessary to pass the 11th harmonic of the keying frequency, i.e. 2 × 275 = 550 c/s, is the limiting factor.

The oscillator described in this article has a much faster rise time, and therefore occupies a much larger bandwidth, than is necessary. The output of this oscillator is modulated with 9 kc/s in a balanced modulator, and the sum frequency (nominally 10 kc/s) is selected by a filter which has a flat amplitude response over the range 9,725-10,275 c/s and has a very low differential phase shift. The effective rise time of the oscillator is entirely controlled by the bandwidth of this filter. More limiting of bandwidth occurs in the transmitter and aerial circuits, which results in further increase of " rise time."

Summary

The oscillator described above was designed with a low nominal frequency, and low deviation. However, the principle used would be quite suitable for much higher frequencies where one would probably require k to be less than 0.1. This type of oscillator has two chief advantages over those mentioned in the introduction. These advantages are (a) the mean frequency is substantially independent of deviation, and (b) the amplitude of oscillation is identical on both " mark " and " space " frequencies.

REFERENCES

1. "Electrically Tuned RC Oscillator or Amplifier," by Oswald G. Villard, Jr., and Frank S. Holman, *Proc. I.R.E.*, Vol. 41 (1953), p. 368.

2. " Radio Engineers Handbook " (first edn.), by F. E. Terman, p. 578.

LETTERS TO THE EDITOR

The Editor does not necessarily endorse opinions expressed by his correspondents

Logic Symbols

IN the current series of "Elements of Transistor Pulse Circuits," your contributor T. D. Towers is filling a long felt need very lucidly, and I for one read his articles as soon as I receive my copy of *Wireless World*. However, I was disturbed to see his comments about symbols for logic gates in the October issue.

There are certainly many different types of logic symbols at present in use in Great Britain, let alone those in use in Europe and in the U.S.A. etc., most of these logic forms have some advantages and many confirmed supporters. But in an article introducing the subject to new readers, surely it is worth while using the symbols described in Supplement 5 of the British Standard B.S.530, even if they are not all ideal? These symbols are almost mandatory in Service publications and are familiar to most of the adherents of other logic systems. In the B.S. symbols, circles are generally used for gates, and a circle can be scribbled on the back of an envelope at least as easily as a rectangle. Also, a negatory bar, or a negative symbol (−) is simpler and more explicit than the little half-moons indicating NOT of Mr. Towers' symbols. I expect that the artificial shading on Mr. Towers' gates was added for artistic effect, but to the uninitiated it is an added complication, and serves no useful purpose.

Also I deplore the use of UP and DOWN in reference to the signal inputs to the gate circuits. While these terms are fairly clear when UP refers to positive, as shown in Mr. Towers' circuits, this is by no means general; especially when p-n-p transistors, drawn with the supply line (−ve) at the top, are concerned. The terms Affirmative and Negatory have the advantage of being less ambiguous, but are long-winded, and most people would prefer the standard nomenclature of "1" and "0", provided that these terms are clearly defined at the start of the text matter.
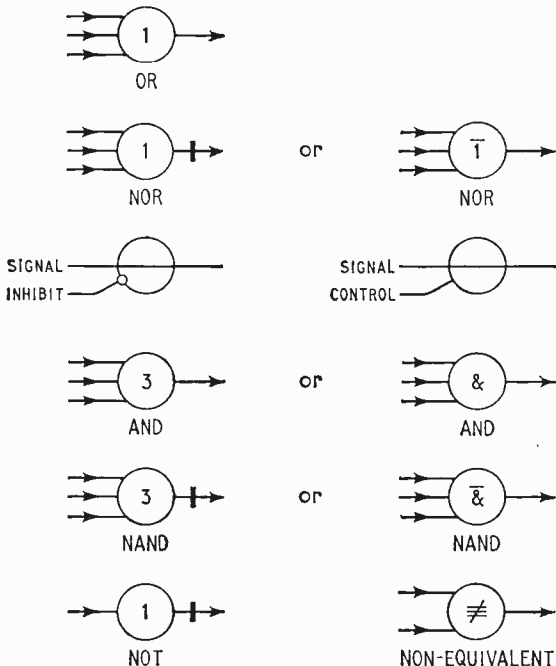
There is one other point that is worth mentioning and that is the complementary relationship of AND and OR gates. For example, an AND gate for positive (+ve) input signals can be used as an OR gate for negative signals, and an OR gate for positive signals becomes an AND gate for negative signals.

These are minor criticisms, however, and I look forward to the day when the whole series is published in book form—it will surely be one of the most useful books on transistors then available.

Welwyn Garden City.   R. B. GREEN

YOUR contributor T. D. Towers, in stating in his article in the October issue that "the normal NOR symbol can also be represented as an OR symbol with a NOT symbol in each input line" implies a logical equivalence in these two representations. In fact the NOR configuration is logically equivalent to an AND symbol with a NOT symbol in each input line.

Kilmarnock.   E. KEATING

The author replies:

Mr. Green's comments on my symbology are, alas, only too true. As a member of various VASCA & B.S.I. committees, I am not unaware of B.S.530, Supplement 5. My excuse for the symbols I use are that they were used by Millman and Taub in their classic "Pulse & Digital Circuits" on which I was teethed. If I can be excused a horrid mixture of metaphor and parodied cliché, "You can't teach a toothless cur new tricks"! (Also I draw my own sketch circuits and find a straight edge always to hand, where a compass may be difficult to find.)

As to my use of "up" and "down", I defend this a bit more vigorously. With some considerable experience of explaining and teaching digital circuitry, I have found "up" and "down" the most effective terms to use to get things over to the newcomer. However, I cannot but admit that standardization in this, too, would be most welcome.

Finally, I deliberately omitted the complementary relationship of "AND" and "OR" gates. It is an interesting logical fact (and invariably quoted *in extenso* in textbooks), but in practice I find it of little help in the practical approach to logical circuit design.

I must thank Mr. Keating for pointing out the error —a simple drafting mistake in Fig.89(b) on page 538 of the October issue. This should show an "AND" in the right-hand box, and the text a few lines above the diagram should be amended accordingly.

T. D. TOWERS

OR

NOR   NOR

SIGNAL / INHIBIT   SIGNAL / CONTROL

AND   or   AND

NAND   or   NAND

NOT   NON-EQUIVALENT

Elements of Transistor Pulse Circuits

MR. BILSTON'S method of design (November issue p. 560) is straightforward enough, but could not he still further simplify his formula by omitting the terms $V_{BE(sat)}$ and $V_{CE(sat)}$ except in the formula for R_2?

These voltages will normally be small enough to be neglected in comparison with the other voltages. For example, from the OC75 characteristics, the collector bottoming voltage in grounded emitter is given as about 100 mV. The base input voltage will presumably be even smaller. In this case the formula become very

simple indeed. I wonder if Mr. Towers agrees to these simplifications?

P. G. M. DAWE

Institute of Experimental Psychology,
University of Oxford.

Mr. Bilston replies:

Anyone having a hand in practical engineering will I am sure recognize the difficulty of generalization.

In my attempt to simplify the conclusions of others with regard to the Schmitt my goal was to enable the derived expressions to be applied without the necessity of specifying exclusion of certain types of transistor.

Using very cheap Ge devices the main factors will vary with leakage current and need to allow very low gains to be used; alternatively with high-gain Si devices the voltages across the transistor would show a large effect if the expressions ignored these parameters. It *is* possible, using my method, to derive a circuit which cannot work. However, using normal engineering intuition the expressions have the advantage of allowing for the worst practical case.

Mr. Dawe may be interested to consider the case of using silicon transistors when, although his " simplifications " would not apply, the base divider current of Q2 may be decreased by about an order of ten before the effect shows in the practical operation of the circuit.

Errata: In the circuit diagram which accompanied my letter (p. 561, November issue) the transistors should have been shown as n-p-n types and the expression for R_4 should have read $\dfrac{V_1 - V_{BE(sat)}}{\text{"}i_B\text{"}}$

D. R. BILSTON

I HAVE greatly appreciated the excellent series by Mr. T. D. Towers, but would like to comment on his design method for Schmitt trigger circuits in the August issue. It is very evident that something is amiss when it is realized that for a base current of $V_0/R_1 h_{FE min}$ (page 405, column 2) the minimum current through the cross-coupling resistor R_5 is required to be $11 \, V_0/R_1 h_{FE min}$, which in the example cited would be 1.32 mA. This would produce a voltage drop of 15.9 V in R_5, whereas the supply voltage is only 12V (or 10V, it is not clear which).

The simplified approach used by Mr. Towers can readily be modified to yield an unambiguous method, which includes a test as to whether the initial requirements are physically realizable. All the formulae given can easily be verified, using Mr. Towers' original assumptions. His Fig. 70, page 405, is included for reference.
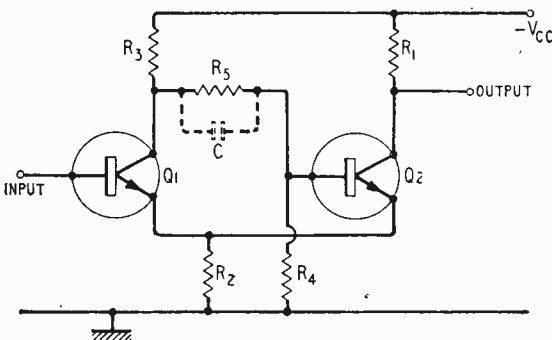
(a). Select a value of R_1 low enough to drive the output load resistance, and make $R_3 = R_1$. (It can be shown that the circuit will always be regenerative under these circumstances.) Select a suitable supply voltage V_{CC}, and establish the desired values of the upper trip voltage V_U and lower trip voltage V_L (this last value may or may not be physically realizable, as will appear as the design proceeds). Select a suitable transistor, and establish

its $h_{FE min}$ *when saturated*; this value is usually less than that quoted as $h_{FE min}$ for normal operation. It is then required to evaluate V_0 (the peak-to-peak output voltage) and R_2, R_4 and R_5.

(b) $$V_0 = V_{CC} - V_U \quad \dots\dots\dots\dots\dots\dots (1)$$

(c) $$R_2 = \frac{V_U R_1}{V_0} \quad \dots\dots\dots\dots\dots\dots (2)$$

Evaluate a parameter H, such that

$$H = \frac{V_{CC}}{V_L} - \frac{R_3}{R_2} - 1 \quad \dots\dots\dots\dots (3)$$

Provided that H is positive, but less than V_0/V_U, the design is physically realizable. Otherwise either V_L or the other original parameters selected must be changed. Assuming these conditions are met, R_5 can then be written in terms of R_4:

$$R_5 = H R_4 - R_3 \quad \dots\dots\dots\dots (4)$$

(e) Determine the value of nR_4 from the equation:—

$$n R_4 = R_1 h_{FE min} \left(\frac{1}{H} - \frac{V_U}{V_0} \right) \dots\dots\dots (5);$$

(n is actually the ratio $i_{B2\ actual}/i_{B2\ min}$ for saturation n_{min} is obviously 1).

(f) Determine the minimum usable value of R_4;

$$R_{4\ min} = R_3/H \quad \dots\dots\dots\dots (6)$$

Knowing the value of $n R_4$ from eqn. (5), this will give the value of n_{max}. Select a value of n between 1 and n_{max} (about 3 is most suitable), and from the value of $n R_4$, obtained above, evaluate R_4. Hence evaluate R_5 from eqn. (4).

Using this method, the conditions given by Mr. Towers are found to be unrealizable in a saturated circuit, but they become realizable for $V_{CC} = 10V$, $R_1 = R_3 = 1k\Omega$, $h_{FE min} = 50$, $V_U = 4\,V$, $V_L = 3V$. Using $n = 3.3$ ($n_{max} = 22$) this gives $R_4 = 8.2\,k\Omega$, $R_5 = 5.6\,k\Omega$.

Conversely, the method can be used to evaluate a given circuit. Given V_{CC}, $h_{FE min}$, R_1, R_2, R_3; R_4 and R_5, we can proceed as follows, assuming for the moment that Q2 is saturated.

$$V_0 = \frac{R_1}{R_1 + R_2} V_{CC} \quad \dots\dots\dots\dots (7)$$

$$V_U = V_{CC} - V_0 \quad \dots\dots\dots\dots (8)$$

$$H = \frac{R_5 + R_3}{R_4} \quad \dots\dots\dots\dots (9)$$

and for saturation this must be less than V_0/V_U.

$$V_L = \frac{V_{CC}}{H + \dfrac{R_3}{R_2} + 1} \quad \dots\dots\dots (10)$$

$$n = \frac{R_1 h_{FE min} \left(\dfrac{1}{H} - \dfrac{V_U}{V_0} \right)}{R_4} \quad \dots\dots\dots (11)$$

The value of n will determine by what margin Q2 is saturated, remembering that n must be greater than 1 for saturation.

These methods have been extensively tested by comparison of predicted with measured parameters, and have proved most satisfactory.

Parkville, N.2, Victoria,   D. J. DEWHURST
Australia.   University of Melbourne

Mr. Towers' comments:

I agree with Mr. Dawe that with germanium transistors you can ignore $V_{BE(sat)}$ and $V_{CE(sat)}$ in the first order design of a Schmitt trigger. You only need take them into account with some silicon transistors at low supply voltages.

I found Mr. Bilston's design formulation interesting. The simplified design I gave was determinedly practical —too determinedly in the event, as it will work only for a limited range of design parameters. Several correspondents have kindly gone to the trouble of writing in on this point. In particular I have to thank

Mr. M. D. Thomas, of Benson-Lehner, Southampton, for a very helpful critical examination of the factors involved. A particularly useful modified design is that suggested by Dr. D. J. Dewhurst.

T. D. TOWERS

Blocking Oscillators

WHILE Mr. T. D. Tower's useful article (Sept. issue) on blocking oscillators is still fresh in our minds, I should like to bring to notice a monostable blocking oscillator circuit which has only small dependence of pulse length on applied load. The idea is to use RC timing in the base circuit, but also to swamp the base current through the transistor with a fixed current from a supply rail.

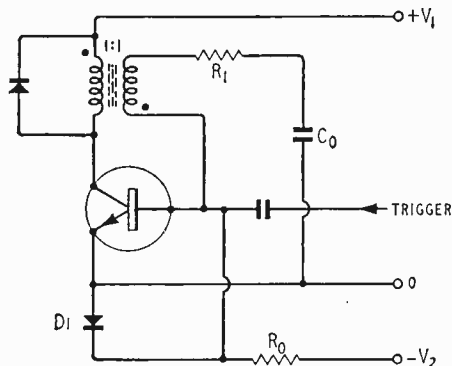Fig. 1 shows the simplest form, in which the transistor

Fig. 1

is normally held off by the forward voltage across the diode D1. When the circuit is triggered on, the condenser C_o is charged by a current determined by V_1 (across the base winding) and the resistance R_1. It is assumed that the initial charging current exceeds that through R_o by enough to bottom the transistor. As C_o charges, the current through R_1 falls until it is insufficient to drive the transistor as well as supplying R_o. The transistor switches off and C_o discharges through D1 and R_1. Provided the current V_2/R_o is several times the base current required by the transistor to drive whatever load is applied, the pulse duration is given approximately by

$$t = C_o R_1 \log_e \frac{R_o V_1}{R_1 V_2}$$

Termination of the pulse occurs at a well-determined point.

The need for two supply rails may be avoided by the arrangement of Fig. 2. C_1 is a large capacitance and R_3 is chosen so that the current through D2 during the pulse is easily replenished between pulses. This circuit requires less triggering current than that of Fig. 1 and the pulse length is reasonably independent of V_1 as well as of the applied load. An astable version can easily be made.
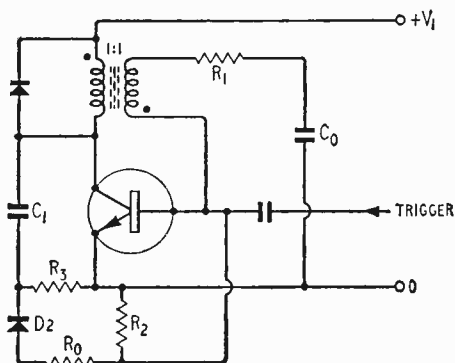
Fig. 2

I was going to say that the transistor involved was an OC42, but shall now have to admit that I have forced myself into drawing the circuit in the way I consider best, positive up (as for the oscilloscope), but have modified it for the benefit of those of you who prefer your collectors at the top

Derby.

R. O. RIDLEY
Rolls-Royce Ltd., Aero Engine Division

Practical Transistor Circuit Design

I WOULD like to comment at some length on the recent series of articles by Mr. G. P. Hobbs. They have so far been disappointing—his timidity in stating that it is "treading on the most dangerous ground" to put forward a design philosophy seems to expose his own lack of any such philosophy.

My main criticisms are:

(1) The reader is being taught to analyse circuits in a loose inaccurate way. (Note, *analyse* not *design*).

(2) It is not helpful to continually draw comparisons with valve circuits. Many (young) designers have never used valves. In any case it may inhibit the invention of new circuits and techniques more suited to transistors.

(3) The engineer does not "throw up his books in despair" when confronted with several equivalent circuits. On the contrary, he is grateful to the analysts for having derived them, and merely selects the one most appropriate to his needs. It is the mixing up of several sets of parameters which leads to confusion!

(4) The terminal conditions under which h-parameters are measured are not stated, nor are we told how to use them. It may be laudable for manufacturers to quote them, but as they are not fully explained by Mr. Hobbs, they can only cause confusion. The hybrid-π and equivalent-T circuit parameters are also mixed.

(5) Mr. Hobbs seems to have arrived at some erroneous conclusions. He states that the input resistance of the common emitter amplifier is $\beta r_e/2$ if the load is infinite. Under such conditions, no signal current can flow between collector and emitter, so all the emitter current flows in the base. This means that the mutual conductance and current gain are both zero, so one would expect the input resistance to be just $r_b + r_e$. These conclusions are supported by analysis of the equivalent-T circuit. Where does the discrepancy arise?

The hybrid π circuit turns out to be not very "rewarding," as its parameters are approximate and take no account of the variation of current gain with load. The theoretical maximum gain obtained from the equivalent-T network is $r_e/(r_e + r_b)$ which again, reasonably, does not depend on β, unlike the value Mr. Hobbs arrives at. The voltage feedback fraction is approximately $(1 + \beta)r_e/r_c$, which is *not* the reciprocal of the maximum gain figure.

(6) The equivalent-T network also provides different answers for output impedance. They are approximately r_c/β for an infinite source impedance, and $r_c(r_e + r_b/\beta)/(r_e + r_b)$ for zero source impedance. Whence come all these factors of two?

(7) The sample calculation appears to give an over-estimate of gain due to ignoring r_b (which is greater than r_{bb}'). Further, the standing base current of $80\mu A$ does need explanation as in general, d.c. current gain is not equal to small signal current gain r_{bb}'.

Camberley, Surrey.

M. TILEY

The author replies:
My attempt to resolve the equivalent circuit into something practical that an engineer can grasp obviously failed for Mr. Tiley. Because he is working from the equivalent-T network against my choice of the hybrid-π, he has a different set of parameters which cannot be immediately compared. My line has been to give magnitudes to the quantities whose significance is important in a circuit, and I opted for the common emitter hybrid-π equivalent circuit for very good reasons.

One reason is the simple way in which internal feedback in the transistor is represented, and another (despite Mr. Tiley's remarks) is the similarity that this model has to the valve equivalent circuit—and incidentally the author comes from the school of young designers who have had little to do with valves.

It was obviously not my intention to produce theory (nor to go into the full definitions of the h-parameters which were included to enable manufacturers' data on these to be used with the equivalent circuit) but it will now be necessary briefly to look at the equivalent-T circuit. Together with the equivalent-T circuit drawn for common emitter operation, I have set out the input and output resistances and voltage gain. I will deal with Mr. Tiley's factual criticisms shortly, but would first like to point out two pitfalls.

(1) The current generator is not across the output and the form of internal feedback is obscured.

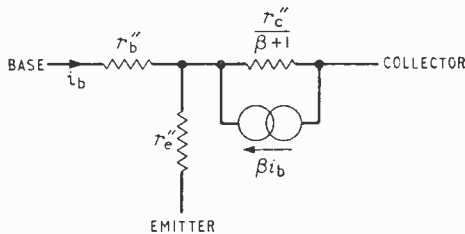(2) The resistor r_e'', is *not* the emitter resistance given by $\frac{25}{I_e}$ ohms. (The parameters are denoted by r_e'', r_b'' and r_c'' to avoid this confusion. The reader may like to refer to pages 81 and 82 of the "Mullard Reference Manual of Transistor Circuits.")

The equivalent-T network is, I believe, deceptively simple; it does give the same information as the hybrid-π circuit and is, of course, as accurate as the number of elements it contains.

The extrinsic base resistance is ignored and unity is neglected in comparison to β. The voltage feedback factor, $1/\mu$ is $\frac{\beta r_e''}{r_c''}$ approx.

Input Resistance, $R_{in} = r_b'' + r_e''. \dfrac{r_c'' + R_c}{r_c''/\beta + R_c}$

which for $R_c = 0$ becomes $r_b'' + \beta r_e''$
and for $R_c = \infty$ becomes $r_b'' + r_e''$.

Output Resistance, $R_{out} = \dfrac{r_c''}{\beta} + r_e''. \dfrac{r_c'' + R_s}{r_c'' + r_b'' + R_s}$

which for $R_s = \infty$ becomes
$r_c''/\beta + r_e'' \approx r_c''/\beta$
and for $R_s = 0$ becomes
$r_c''\dfrac{(r_e'' + r_b''/\beta)}{r_e'' + r_b''}$

Voltage Gain, $A_v = \dfrac{R_c}{r_e'' + \dfrac{r_b''}{\beta} + R_c\left(\dfrac{r_e'' + r_b''}{r_c''}\right)}$

which for $R_c = \infty$ becomes
$\dfrac{r_c''}{r_e'' + r_b''}$

It is not true to say that the parameters of the hybrid-π circuit take no account of the variation of current gain with load—this point was dealt with on page 435, middle of column 2, in the September issue.

To find the relations between the parameters of the two equivalent circuits we can examine R_{in} for zero load and R_{out} for zero R_s giving:

$$r_{b'e} + r_e = r_b'' + \beta r_e'' \text{ or } r_e = r_b''/\beta + r_e'' \ldots\ldots (1)$$

and $r_{ce} \approx \dfrac{r_c''}{\beta} \left(\dfrac{r_b'' + \beta r_e''}{r_b''}\right) = \dfrac{r_c''}{\beta}\left(\dfrac{r_e}{r_e - r_e''}\right) \ \ldots (2)$

We also have that the voltage feedback factor $\dfrac{1}{\mu} = \beta r_e''/r_c''$ which gives in eqn. (2), $r_{ce} = \mu r_e\left(\dfrac{r_e''}{r_e - r_e''}\right)..(3).$

So far this is quite general but if we take the value of r_e'' equal to r_{e2}, as given for example in equation 2.38a on page 44 of R.F. Shea's book "Transistor Circuit Engineering," we obtain $r_{ce} = \mu r_e$. The theoretical maximum gain becomes μ, the reciprocal of the feedback factor, and from (1) and (2) we also get $r_{b'e} = 2r_b''$ and $r_{ce} = 2r_c''/\beta$. According to a footnote in the above reference r_e'' can vary either side of the value $r_e/2$ whilst according to the derivation of the equivalent circuit from the basis of a transmission line it comes to exactly $r_e/2$ (in my opinion because the transmission line is assumed uniform). Whether or not the factor 2 is representative of an actual transistor depends on how closely the transistor fits the model chosen. My measurements indicate a good fit for most types. Of course, one can leave the equivalent circuit in its general form with r_{ce} indeterminate, as Mr. Tiley would prefer for his analyses, but as there is this close link between r_{ce} and μr_e given by equation (3), it is sensible for design purposes to adopt the simplest value. It is interesting to note that transistors made with a very small degree of internal feedback exhibit a very high output resistance and one would obviously adopt the hybrid-π circuit to describe these.

Perhaps we can ascribe my "erroneous conclusions" to the "mixing up of several sets of parameters"— on someone's part? Dealing specifically with item (5), R_{in} for $R_c = \infty$ is $r_b'' + r_e''$ which is approximately r_b'' or $\beta r_e/2$. The mutual conductance does not go to zero when R_c tends to infinity (the collector current remains fixed and obviously a limiting condition is being discussed). There is still, of course, a current gain, as distinct from β, is reduced to zero. There is still, of course, a current βi_b from the current generator passing through $r_c''/\beta + 1$.

On point (6) R_{out} for $R_s = \infty$ is r_c''/β which equals $r_{ce}/2$, and R_{out} for $R_s = 0$ must be r_{ce}. The first part of point (7) isn't clear to me; the second point is valid criticism although I didn't think it opportune to complicate the sample calculation with the difference between h_{fe} and h_{FE}.

G. P. HOBBS

Microwave Levitation:—*U.S. Air Force officials recently attended a demonstration by the Raytheon Company of a model helicopter driven by rectified current derived from an array of 4000 "dipole diodes" in a vertical beam of microwave energy. The power used is stated to have been 5 kW and the frequency is deduced to be of the order of 1,500 Mc/s. The project forms part of an aerospace programme to supply power to hovering vehicles.*

WORLD OF WIRELESS

I.T.A./B.B.C. Co-operation in Engineering and Research

THE section of the 1963/64 report of the Independent Television Authority covering engineering states, " The year has been noteworthy for the large measure of engineering and research co-operation between the I.T.A. and the B.B.C. in the planning and provision of joint transmitting facilities at existing and new television stations." This has arisen, it says, as the result of three separate, though related, factors. One was the Government decision to use the unallocated v.h.f. channels in Band III to extend and improve the coverage of the present 405-line services of both organizations. Another was the decision to use the u.h.f. bands for the provision on the 625-line standard of any additional television services. The third was the proposal, still under review, that the u.h.f. bands might also be used for transferring the 405-line services to 625 lines by duplicating their transmissions on the new national network of u.h.f. stations as it is progressively expanded. The Authority had to assume, therefore, that some time in the future it would be transmitting in the u.h.f. bands and, on this basis, technical and financial agreements were reached between the I.T.A. and the B.B.C. for the joint development of the u.h.f. station network both at new and at existing sites.

No new stations were brought into service by the Authority during the year under review, but work was begun on eight stations to be opened before the end of 1965 and on the modification of four other stations for u.h.f. operation. By the end of 1965 the Authority will then have thirty stations in operation. Six of the new stations are to be unattended relays—at Rumster Forest, Caithness (Channel 8, vertical polarization, 30 kW e.r.p.); Angus (11-V-50); Scarborough (6-H-1); Sandy Heath, Bedford (6-H-30); Membury, Berks (12-H-30); and Richmond Hill, Isle of Man (8-H-10). A new station is being built at Belmont, East Lincs (7-V-20) and a new transmitter to radiate the Welsh service (7-V-100) has been added at the St. Hilary, South Wales, station. Provisions are being made at each of these stations for their eventual use in the u.h.f. chain. The stations being modified for u.h.f. operation are Winter Hill, Lancs; Emley Moor, Yorks; Black Hill, Lanarks; and Moel-y-Parc, Denbigh. The report gives predicted service area maps for the new stations.

Electronics Industry Dinner

THE recently formed Conference of the Electronics Industry, an organization comprising nine trade associations which are concerned with the various sections of the electronics industry, held its first annual dinner in London on 10th November, at which the guest of honour was Earl Mountbatten of Burma, chairman of the National Electronics Research Council.

Earl Mountbatten spoke of events leading up to the formation of N.E.R.C. (see p. 449, September issue) and of its future plans. One of these is to tackle the " information explosion." To this end a committee has been appointed and has prepared a research project on the Scientific Dissemination of Information which has been placed before D.S.I.R. The system of information retrieval proposed by N.E.R.C. will use a computer to select from the hundreds of papers and articles published every week the references to those of interest to each particular research worker throughout the country. In addition, a weekly listing of all new material will be provided, and there will be facilities for retrospective searching by computer. It is believed that the project will take three years to develop to the full, and it is hoped to make a major break-through in tackling the information retrieval problem.

N.E.R.C. is also sponsoring an up-to-date survey of all electronics research that is being carried on throughout Great Britain, which will be made available to those in charge of research in Government establishments, universities, technical colleges and in industry. This should reduce unnecessary duplication and encourage the cross-fertilization of ideas.

Pay-TV Ballot

WHEN electors in the State of California went to the polls on November 3rd, in addition to voting for a president, and for government and local officials they also voted on the future of pay-television in their State. The opponents of pay-television, which was introduced over wire by Subscription Television Inc. in Los Angeles and San Francisco a few months ago, had a proposition included on the ballot paper which sought to outlaw pay-TV. We learn on good authority that the proposition, which was one of 17 on the ballot paper, including such questions as racial discrimination in real estate and state lottery, was successful. It is thought, however, that it may be held to be unconstitutional.

Refurbishing Rugby

GBR, the transmitter used for v.l.f. transmissions at the Post Office station at Rugby, is being modernized. The 16kc/s transmitter, which has been in service for 40 years, is used for some 22 hours a day for morse transmissions and for the radiation of time signals four times a day. The output stage consisting of three panels each containing 18 water-cooled valves is being replaced by three vapour-cooled valve stages, and the lower-power stages by three wide-band amplifiers. This will enable the output power of 210kW to be doubled and will permit the use of f.s.k. in addition to hand-speed morse. The aerial system, supported by twelve 820ft masts, was completely renewed a few years ago.

The winner of the competition at the recent R.S.G.B. Radio Communications Exhibition to guess the number of soldered joints in a Racal RA71 communications receiver was Mr. Buckby, of Corby, Huntingdonshire, who wins the set. There were actually 1,910 joints, and he guessed 1,911. Well over 1,000 tickets were sold for 2s 6d each, the proceeds from which are being given to the Imperial Cancer Research Fund.

Television by Wire.—Speaking at the annual luncheon of the Relay Services Association on 10th November, Sir Ronald German, director general of the G.P.O., said that 750,000 homes in the U.K. received television by wire and that there were 80,000 new subscribers last year. Alex. M. Lowe, chairman of the Association, predicted that in ten years time most of the homes in our cities will receive their sound and television programmes by wire.

Instrumentation and Control is the title of a new special subject in Part III of the examination of the Institution of Electrical Engineers. The first examinations will be held in June 1966.

Receiving Licences.—The number of combined television and sound licences in the U.K. increased during the first nine months of the year by 271,732, bringing the total at the end of September to 13,061,215. Sound only licences decreased by 164,050 during the nine months and now total 2,928,111, including 604,829 for sets fitted in cars—36,025 more than at the beginning of the year.

The latest addition to the **Printed Wiring Guides** prepared by the Electronic Engineering Association is on "The quality of copper clad epoxide bonded glass fabric laminate. suitable for precious metal plating." The guide is based on extensive work by several member companies of the Association and is obtainable free of charge from the E.E.A., 61 Green Street, London, W.1.

This year's **Christmas Holiday Lecture** for older schoolchildren arranged by the I.E.E. will be delivered by Dr. D. H. Parkinson, of the Royal Radar Establishment, whose subject will be superconductivity. It will be given at Savoy Place, London, W.C.2, on 30th December at 2.30 and will be repeated at the same time the following afternoon. Admission is free but application for tickets, stating for which afternoon they are required, should be made to the I.E.E., Savoy Place, London, W.C.2. As accommodation is limited, parents are asked not to accompany their children.

An "**Introduction to Industry**" course is to be held at the Chippenham Works of the Westinghouse Brake and Signal Company during the week 4th to 8th January, for senior pupils taking G.C.E. "A" level science subjects and leaving school in July 1965. Applications should be sent, by the end of November, to the Training Superintendent, Westinghouse Brake and Signal Co., Chippenham, Wiltshire.

Whitworth Foundation Awards, worth up to £1,500 per year plus dependants' allowances as appropriate and tenable for periods ranging from three months to two years, will again be offered by the Department of Education and Science in 1965 for projects of research or study in engineering or in the teaching of engineering. Candidates, who must hold a degree or diploma in engineering, H.N.D. or H.N.C. in engineering with at least two distinctions, are required to submit a 2,000 word paper describing the project which they would undertake if successful. Application forms are available from the Department of Education and Science (F.E.1, General), Curzon Street, London, W.1.

A one day symposium on the "**Silicon gate-controlled switch,**" or gate turn-off Thyristor, is to be held on 2nd December at the Enfield College of Technology, Queensway, Enfield, Middx. Registration forms are available from R. D. Kitchener, head of the Electrical Engineering Department. (Fee 2gn).

PERSONALITIES

Col. G. W. Raby, C.B.E., M.I.Mech.E., M.I.E.R.E., has been nominated president-elect of the Institution of Electronic and Radio Engineers for 1964/5 and will be giving his presidential address on December 9th. Trained at the Rugby works of the B.T.H. Company (now A.E.I.) he joined the English Electric Company in 1934 and became assistant general manager, on new developments and research. Col. Raby, who as a Territorial Officer in the Royal Engineers went to France with the B.E.F. in 1939, was attached to the Ministry

Col. G. W. Raby

of Supply in 1941 as superintendent engineer and was concerned with the development and production of radar. He later became chief superintendent, Signals Research & Development Establishments at Christchurch

and Woolwich. Col. Raby, who is 64, became deputy director (engineering) at the Atomic Energy Research Establishment, Harwell, in 1955 and is now chairman of Atomic Power Constructions Limited.

Sir Jules Thorn, chairman of Thorn Electrical Industries, has succeeded **C. O. Stanley,** chairman of Pye, as president of the British Radio Equipment Manufacturers' Association—the domestic sound and television equipment makers' association. Mr. Stanley who became the Association's first president in 1962 has relinquished the presidency at his own request but is continuing to serve as a member of the president's committee.

A half share in the Nobel Prize for Physics has been awarded to **Dr. Charles Townes,** of the Massachusetts Institute of Technology, and the other half to two Soviet physicists, **Professors Nikolai Basov and Alexander Prokhorov,** of the Academy of Sciences, Moscow, for their "basic research in the field of quantum electronics" which lead to the development of masers. Dr. Townes, who is 49, has been provost of M.I.T. since 1961. He has been a Fulbright lecturer at the Universities of Paris and Tokyo, was a consultant to President Eisenhower's Scientific Advisory Committee and, before joining M.I.T. in 1961, was a director of Research for Defence Analysis, a non-profit organization in Washington, run by nine universities.

J. M. Wilson, B.Sc., F.R.I.C., manager of the semiconductor division of Standard Telephones & Cables for the past year, has been appointed general manager of the radio division in succession to **K. P. Wood** who, as announced last month, is now marketing director. Mr. Wilson graduated with honours in chemistry at University College, London, in 1947 and that year joined Standard Telecommunication Laboratories subsequently becoming responsible for the semiconductor materials group. In 1958 he joined the rectifier division as head of the silane

J. M. Wilson

silicon department and in 1962 became manager of the division. Since 1963 he has been manager of the semiconductor division. Mr. Wilson was elected a fellow of the Royal Institute of Chemistry in 1955.

The Rt. Hon. **Anthony Wedgwood Benn**, M.A., the new Postmaster General, was at one time chairman of the Labour Party's Broadcasting Advisory Committee. Prior to his election to Parliament in 1950, Mr. Wedgwood Benn, who is 39 and is a graduate of New College, Oxford, was for two years a producer for the B.B.C. North American Service. The assistant P.M.G. is **Joseph Slater**, B.E.M., member for the Sedgefield Division of Durham since 1950. He is 60.

A. Frank Boff, B.Sc., originator of the Boff snap-off diode which is used as an harmonic generator, has joined Racal Instruments Ltd., of Crowthorne, Berks., as technical director. A graduate of London University Mr. Boff went to North America in 1950 after spending a few years with S.T.C. and Marconi. While in America he was first with Beckman Instruments, of California, develop-

A. F. Boff

ing electronic instruments and from 1954-57 with the Canadian Marconi Company, of Montreal, working on communication system design. In 1957 he went back to the U.S.A. as manager of research and development for the Dymec Division of Hewlett-Packard. Four years ago Mr. Boff, who is 40, returned to the U.K. and has been technical manager of Hewlett-Packard Ltd.

Gilbert H. Metson, M.C., D.Sc., M.I.E.E., deputy director of research at the Post Office since 1962, is to succeed **Reginald J. Halsey**, C.M.G., B.Sc.(Eng.), F.C.G.I., D.I.C., M.I.E.E., as director of research when he retires on December 31st. Dr. Metson, who joined the Post Office as a youth-in-training in 1925 at the age of 18, led the team which developed the long-life steep-slope pentode valves used in submerged repeaters. During the war he served with the Royal Signals and was awarded the Military Cross. He returned to the Dollis Hill Research Station in 1946. For his work on thermionic valves and particularly

the oxide-coated cathode he was awarded in 1958 the degree of doctor of science by Queen's University, Belfast, where he earlier received his M.Sc. and Ph.D. following research on magnetron oscillators. Mr. Halsey, who won a Royal Scholarship to the City & Guilds College after serving a five year apprenticeship at Portsmouth Dockyard, took an honours degree in engineering at London University in 1925 and two years later, at the age of 25, entered the engineering department of the Post Office. For the major part of his service he has been in the research branch and became director of research in 1958. He was at one time assistant engineer responsible for submarine cable systems and in particular the British contribution to the first transatlantic telephone cable. Mr. Halsey is also a director of Cable and Wireless.

E. D. Whitehead, M.B.E., B.Sc., M.I.E.E., director of electrical inspection, Ministry of Aviation, since 1960, has been appointed director of electronics production (radar) in the Ministry in succession to **E. C. Drew**, O.B.E., M.I.E.E., who has become director of technical costings. Mr. Whitehead, a graduate of London University, joined the Scientific Civil Service in 1937 after a few years in industry first with G.E.C. and then Pye. In 1940 he went to R.A.E., Farnborough, and after three years became technical assistant to the director general of communications equipment. He was for six years secretary of the Radio Components Standardization Committee. From 1951-57 he was assistant director electronic component production in the Ministry of Supply and for the following three years was in charge of the division of E.I.D. responsible for the inspection of electronic equipment. A photograph of Mr. Whitehead opening the Radio Communications Exhibition is on page 610.

L. F. Broadway, O.B.E., B.Sc., Ph.D., M.I.E.E., head of the research division of electric and Musical Industries since 1956, has been appointed a divisional director. Dr. Broadway, who obtained a Ph.D. degree from both Bristol University and Clare College, Cambridge, joined the E.M.I. Research Laboratories in 1933 at the age of 28. In 1940 he became head of c.r.t. and valve research and later head of vacuum tube research.

Professor P. M. S. Blackett, F.R.S., who earlier this year resigned the headship of the Department of Physics at Imperial College, London, but continued as professor of physics and as pro-rector, has been granted leave of absence by the College to become deputy chairman of the Advisory Council being set up by

the newly formed Ministry of Technology. He has tendered his resignation as a member of the National Research Development Corporation and as consultant to E.M.I. Professor Blackett joined the staff at Imperial College in 1953.

E. W. Yeomanson has been appointed president of the Radio Society of Great Britain for 1965. He has been a member of the Council for the past six years and is at

E. W. Yeomanson

present executive vice-president. Mr. Yeomanson, whose call sign is G3IIR, also serves as a member of the Radio Amateur Emergency Network Committee. He has been in the telecommunications industry for 30 years and is sales office manager of the Private Telephone Department of A.E.I. at Grosvenor Place, London, S.W.1.

OUR AUTHORS

M. V. Bond, author of the article on transistor r.f. power amplifiers in this issue, served his apprenticeship with the Electrical Inspection Directorate of the Ministry of Aviation and joined the transistor division of Standard Telephones and Cables as an apprentice engineer in 1961. It was while at S.T.C. that he carried out the work on which his article is based. Since February this year Mr. Bond, who is 24, has been with Transitron Electronic Ltd., of London, W.5. He holds an amateur transmitting licence (G3NWF).

R. B. C. Copsey, a senior section leader in the Communications Laboratory of Redifon Ltd., at Crawley, Surrey, contributes the article on a single-channel transistor f.s.k. oscillator in this issue. During the war he was engaged on the development of transceivers for the Armed Services and from 1946-48 was a technical instructor, with the rank of lieutenant, at the R.E.M.E. Training Centre, Arborfield. Mr. Copsey then spent a year with Philips before joining Redifon in 1949.

Introduction to

Practical Transistor Circuit Design

4.—THE BASE-EMITTER VOLTAGE AND SOME APPLICATIONS FOR DIODES

By G. P. HOBBS.* B.A., Grad. I.E.E.

IN last month's article the influences of leakage current and current gain upon the working point of an amplifying stage were treated. Both influences occasioned the use of the stability factor for calculating the possible deviation in collector current, ΔI_c, away from a design centre. These two influences are by no means the only disturbances upon the operating current; certain other tolerances have so far been omitted to keep the story simple.

One must, for instance, take into account the variation of base-emitter voltage with temperature. Then there are possible rail supply and resistor tolerances to be considered. Again, leakage in electrolytic coupling capacitors can bring about a shift in the working point. In short, the sum total of all these influences will be matched by a grand total of ΔI_c; the grand total representing the worst combination of tolerances. The design procedure can only be taken as fully comprehensive when all these contributors to ΔI_c have at least been considered. It would be in the interests of reliability for more attention to be paid to the systematic design of biasing networks and it is hoped that the method described in these final two parts of the series will provoke some thought on the subject. The system described has the great merit that it can easily be broken into its component parts and moulded to an individual's requirements.

Resistor tolerances and rail supply variations happily fall into line with the design centre system as their tolerances are given as percentages about a mean value. Their proportionate effect upon the standing current is usually apparent from the circuit and the contribution to ΔI_c is simply a small fraction of the standing current, $\overline{I_c}$.

All electrolytic capacitors exhibiting leakage must disturb the working point of the stage to which they are connected, but where they are used as coupling capacitors to the high-impedance base circuit, the disturbance may be worth taking note of. A shift in the working point must be incorporated into the calculations for ΔI_c, if a coupling component is not to make nonsense of the design centre system.

Now the capacitor leakage current will pass both into the base of the transistor and into the biasing resistance R_B. In splitting between these paths, it behaves in exactly the same manner as the fundamental transistor leakage, I_{CBO}. Accordingly a change in the collector current S times as great as the capacitor leakage must result. It is a simple matter, therefore, to add on the maximum capacitor leakage to $I_{CBO\ max}$ wherever the latter occurs in the design calculations. In counting the capacitor leakage in with $I_{CBO\ max}$ account should be taken of the directions of the two leakage currents in finding the extreme limits of possible total leakage current. The troublesome side of the problem may be more likely to lie in finding out what, under the specific circuit conditions, the capacitor manufacturer believes the maximum leakage to be.

Temperature Dependance of V_{BE}:—The final factor is the base-emitter, voltage V_{BE}. Like all other transistor parameters, it is sensitive to the operating environment; notice needs be taken of the temperature and standing collector current when investigating its magnitude. For germanium transistors the size of V_{BE} is typically 250mV at 25°C and it has a negative temperature coefficient of between -2.5mV/°C and -1.8mV/°C (or even smaller). The numerically smaller values of the temperature coefficient apply to higher standing currents. At the same time, at the higher standing currents the magnitude of V_{BE} is increased; this, of course, is partly due to the slightly greater bias voltage necessary to set up the higher collector current, but it is also due to the increased voltage drop across the extrinsic resistance. The extrinsic resistance is temperature dependent and its increasing influence at higher currents explains the change in the temperature coefficient. The magnitude of the base-emitter voltage in a bottomed stage can be quite large—up to about 500mV in germanium— due to the predominance of the extrinsic resistance.

For silicon transistors V_{BE} is of the order of 600mV at 25°C and at a low standing current. The temperature coefficient, however, is the same as for germanium. Now the base-emitter voltage drop is directly represented in that fundamental diode equation which the physicists use (see previous article) and V_{BE} is greater for the silicon transistor because the binding energy of the electrons to the lattice is greater. The greater energy gap, it will be remembered, made the thermal generation of electron-hole pairs a far weaker phenomenon in silicon. It can now be added that it is no coincidence that as the maximum permissible junction temperature is approached, the base-emitter voltage (at low standing current) converges to zero. As silicon types start off with a higher V_{BE} at room temperature and the temperature coefficients are similar, the maximum operating temperature for silicon turns out much higher than for germanium. In practice the maximum temperature for silicon transistors may be limited not by this physical process, but by the mechanical construction of the device.

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* The Marconi Company Ltd., Chelmsford, Essex.

Having found, preferably from data, the range of possible V_{BE}, a mean value may be taken such that excursions $\pm\varDelta V_{BE}$ on either side of the mean cover the range. If then the base potential were held constant, this change $\varDelta V_{BE}$ would be inflicted directly on to the emitter terminal. This state of affairs is closely met in a well-stabilized circuit featuring a value of R_B several times smaller than the resistance presented by the base terminal, $(r_b{'}_e + \beta R_E)$. The change in collector current in the stabilized circuit is consequently $\dfrac{\varDelta V_{BE}}{R_E}$ and this will be the contribution to $\varDelta I_C$.

Even for less precise calculations, a rough value for $\dfrac{\varDelta V_{BE}}{R_E}$ shows the kind of influence that V_{BE} changes can have on the standing current. To keep the percentage contribution to $\varDelta I_C$ within reasonable limits, at least 1 volt ought to be dropped across R_E by the standing current.

V_{BE} as a Parameter for Measuring Junction Temperature:—This completes our discussion on the biasing of amplifying stages, but there are a few more things to be said about the base-emitter voltage. Happily there is a brighter side to the business of V_{BE} variation: it will also provide an indication of the operating junction temperature. Previously it has been emphasized that for all transistor parameters it is *junction* temperature which is pertinent, but up to the present no hint has been given as to how information on this temperature can be elicited. First, therefore, a method will be given for actually measuring it.

To begin with a calibration curve of base-emitter voltage *versus* temperature is obtained by heating the specimen transistor in an oven. It must follow that where there is no internal dissipation the junction temperature will equal the case temperature. A thermocouple is attached to the transistor to measure the case temperature. The base-emitter voltage is measured by applying a fixed base current and the collector is conveniently left open circuit. With the transistor reconnected in the equipment, it is simply necessary to arrange to switch it momentarily back into the test conditions for measuring V_{BE}. The junction temperature is then read off from the calibration curve. There must, of course, be some instantaneous cooling of the junction as the transistor is switched out of circuit, but an immediate reading of V_{BE} closely reveals the working junction temperature.

Obviously this cannot be a quick test to carry out and for more general use a method for rapidly estimating the junction temperature from available data is required. As it is internal dissipation that raises the junction temperature, it will be information on the power rating of the device which will be of assistance, and in particular the dissipation derating characteristic with temperature. Fig. 1 shows some typical characteristics.

The maximum permissible operating junction temperature for the example is 85°C. At an ambient temperature of 85°C, no power dissipation at all can be tolerated but at lower ambient temperatures the lower characteristic, marked $\theta_{j-amb} = 0.4°\text{C/mW}$, gives the permissible dissipation. The figure of 0.4°C/mW is the reciprocal of the slope of the characteristic and is the thermal resistance present between the area of the junctions and the air surrounding the transistor. In speaking of the junction temperature collectively we assume that there is little temperature gradient across the minute region embodying the junctions. The temperature gradient between this region and the surrounding air is on the other hand most significant. It takes place in two steps: from the junctions to the case and from the case to the surrounding ambient.

Now let us suppose that 50mW is being dissipated in the example of Fig. 1. It can immediately be read off from the curves that the maximum allowable case temperature is 79°C and the maximum ambient is 65°C. The significant point is that under either of these two conditions the junction temperature will actually be running at its top limit of 85°C. Because the derating curves are straight lines, it must follow that when the power input to the device is 50mW the excess of junction temperature over ambient will always prove to be 20°C, and of junction over case temperature, 6°C. For a given power input we are enabled to read off the excess temperature T_{j-amb} or T_{j-case}. Given the maximum anticipated
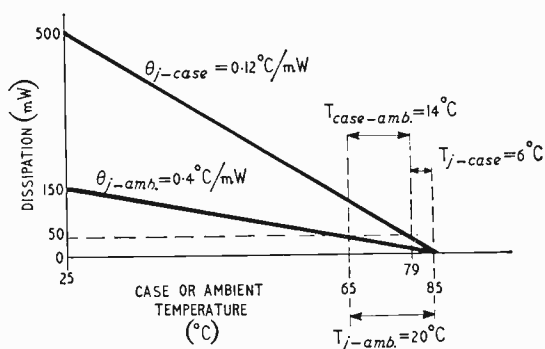
Fig. 1. Dissipation derating curves for a germanium transistor.

ambient or heat sink temperature, the maximum junction temperature can be obtained simply by adding T_{j-amb} or T_{j-case} respectively. In the example, it follows that a maximum ambient of 50°C will result in a maximum junction temperature of 70°C.

The method gives poor accuracy because the thermal resistance quoted by manufacturers is a maximum limit and usually a conservative one at that. Furthermore, the thermal resistance between the case and the surrounding air is very dependent on the packaging of the transistor within the equipment and on convection currents. A useful technique for checking on one's calculations here is to measure the operating temperature of the transistor case by means of a thermocouple.

Methods for Offsetting V_{BE} Drift with Temperature:—To return to the disturbing effects that the base-emitter voltage change has on the working point, we now come to consider some methods for compensating the drift. The decrease of base-emitter voltage with temperature proves to be a linear function over the small range around room temperature; the true function is exponential with absolute temperature, but the curvature of the characteristics will not show for, say, a 50°C temperature change.

To compensate for the drift, a device with a similar constant temperature coefficient is required. Positive and negative coefficient resistors have been used,

but a more basic solution is to employ a further semiconductor junction—a diode.

So long as R_2 is several times smaller than R_1 (Fig. 2), the forward biased diode, D1, lowers the base potential as the ambient temperature increases and tracks the V_{BE} change. However, although ideally the junction temperature of VT1 is in excess of ambient by a fixed amount, in practice it takes time for the final junction temperature to stabilize. On switching on there will be a warming-up period during which some drift between the diode voltage and the base-emitter voltage must occur. The most aggravating cases arise in power stages where there is a large temperature gradient between the junction and ambient; diode compensation can then only be considered as a partial corrective.

The diode employed may be either germanium or silicon as both have a similar coefficient. The coefficient will not necessarily precisely correspond to the V_{BE} coefficient and perfect compensation cannot be expected. A pair of transistors mounted within the same can and connected in a long-tailed pair arrangement offers perhaps the ideal solution to the drift problem. Again where a diode and transistor form part of an integrated circuit, excellent tracking is
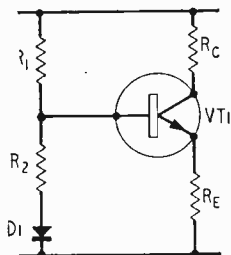

Fig. 2. Diode compensation for V_{BE} drift.

obtained; without doubt the technique of micro-miniaturization pays dividends here.

Compensation may also be effected with Zener diodes. Low voltage ones have negative coefficients, the coefficient is zero for diode voltages around 6, and the higher voltage Zener diodes have positive coefficients. Fig. 3 illustrates the use of a positive coefficient Zener diode connected in series with the emitter lead to offset the V_{BE} temperature variation. The Zener selected has a temperature coefficient of the order of $+ 2.5$ mV/°C at the operating current of the OC84 transistor. The circuit is based on a series regulator and is used to supply a constant current to the focus coil of a vidicon camera's yoke assembly. The OC84 transistor and the OAZ204 Zener diode are mounted together on the same heat sink to give good long-term stability.

Base-emitter Protection:—This is an opportune moment to bring up a further use for diodes in connection with the base emitter junction. In Fig. 4 diodes are incorporated to protect the junction from reverse voltage surges. Whereas the breakdown voltage, BV_{CBO}, of the collector junction will be rated in tens of volts, the base-emitter reverse rating, BV_{EBO}, will in some cases be comparatively low—around 4 volts or even down to 500 mV. As this junction is normally forward-biased, some sacrifice in reverse voltage rating is often made by the manufacturers to gain greater merit in other directions. Permanent damage can be caused by exceeding the reverse rating and nothing is more exasperating than
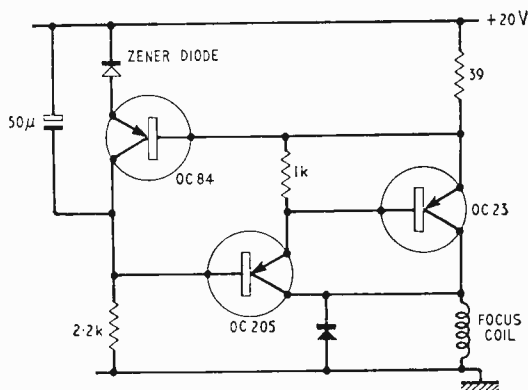
Fig. 3. Current regulator with temperature compensation.

fault finding for stages which have been accidentally "killed" by a surge. Typical causes for such transients are the sudden overload of a signal, or the flow of charging currents in electrolytic capacitors as the equipment is switched on or off. As capacitors charge at different rates in different parts of the circuit, it pays to fit protection diodes even where the need is not thoroughly proven.

The action of the diode in Fig. 4(a) is easily comprehended. Under normal operating conditions the diode is reverse-biased, but it becomes conducting when the base is driven negative; the maximum reverse voltage that the transistor can experience is the forward voltage drop of the protection diode. Fig. 4(b) is slightly more subtle and is of use where the input signal is known to swing below the emitter potential but where it is not desired that the input signal should be loaded by a forward-conducting diode. No direct protection against reverse bias is provided, but the reverse current is limited to that of the series emitter diode. Under forward-biasing conditions this diode is conducting, but on a negative-going pulse both the diode and the base-emitter

Fig. 4. Three methods of protecting the base-emitter junction from reverse voltages.

Fig. 5. Ideally cut-off stages can be disturbed by leakage current, but attention to the value of R_B can assure a leakage current no greater than I_{CBO}.

junction become reverse-biased. Of the two effective diodes, the base-emitter junction has the lower reverse resistance and it is therefore the actual diode which bears t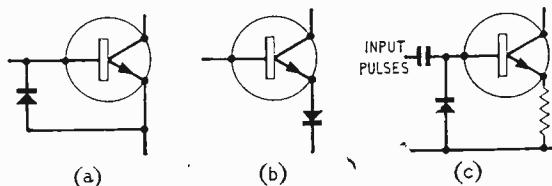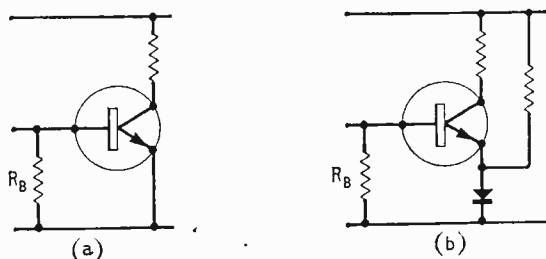he voltage drop. Should the breakdown voltage BV_{EBO} be approached, the ratio of reverse resistances will increase even further and it will be quite impossible for any further rise of the base-emitter voltage to occur.

Where a transistor is intended for switching stages, the BV_{EBO} rating will often be made greater than in the equivalent amplifying transistor. Nevertheless there is often a call for protection of the kind shown in Fig. 4(b). A further useful technique is demonstrated in Fig. 4(c) where a diode is used to d.c.-restore the input pulses to a level above which they can do no damage.

Leakage Current in Pulse Stages:—Now that the discussion is definitely on switching circuits, it will be as well to consider leakage current in a supposedly cut-off transistor. In a bottomed transistor, the collector potential is uniquely fixed, but some drift is possible in a cut-off stage because leakage can never be reduced fully to zero. In Fig. 5(a) the transistor is held cut-off by the base resistor, R_B, being returned to ground. Nevertheless the collector potential will be slightly below the upper rail voltage because of the leakage. Where an input signal actually drives the base negative with respect to the emitter, the leakage current under the reverse bias will simply equal I_{CBO}. Where, however, in Fig. 5(a) in the absence of a signal, I_{CBO} is allowed to build up a sufficient voltage drop across the resistor R_B to forward-bias the transistor, the total leakage will be able to increase to I_{CBO}. To prevent this highly undesirable state of affairs suddenly occurring at high temperatures, R_B must be chosen to satisfy the condition $R_B \lessdot \dfrac{V_{BE\ min}}{I_{CBO\ max}}$ where $V_{BE\ min}$ and $I_{CBO\ max}$ both correspond to the highest anticipated junction temperature. In the extreme case when R_B is zero, one is still left, of course, with the resistance
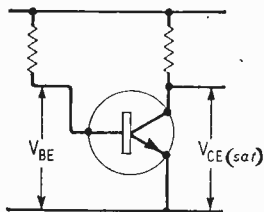

Fig. 6. Potential distribution around a bottomed transistor.

r_{bb}' between base and emitter; and whilst the leakage current I_{CBS} may be equivalent to I_{CBO} at room temperature, it may become more prominent at higher temperatures.

The voltage required to forward-bias the stage, V_{BEmin}, is small, particularly so at high temperatures, so that it is prudent, where practical, to take the lower end of R_B to a more negative supply. The alternative procedure is to raise the emitter potential, and Fig. 5(b) demonstrates a simple expedient for achieving this with a forward-biased silicon diode. A silicon diode is used in preference to germanium because of the greater voltage drop, although other lines of action might include employing a number of diodes in series or a Zener diode. The maximum value that may be used for R_B in Fig. 5(b) is greater than

in Fig. 5(a); the condition is modified by adding the minimum diode voltage to $V_{BE\ min}$.

The Bottomed Transistor:—The potential distribution around a bottomed transistor is illustrated by Fig. 6; V_{BE} is greater than $V_{CE(sat)}$. Both junctions are forward-biased and the device loses its amplifying properties. When the signal excursion of a common emitter amplifier suddenly carries the stage into bottoming, the output waveform is more than just clipped—it features a small superimposed signal that is not inverted; breakthrough of the input signal occurs directly through the forward-biased collector-to-base junction.

But that is an aside; the relevant point is that Fig. 6 shows two forward-biased diodes: base-to-emitter and base-to-collector. In both cases the base is at the more positive potential (n-p-n. transistors); $V_{CE(sat)}$ is actually a rather small voltage
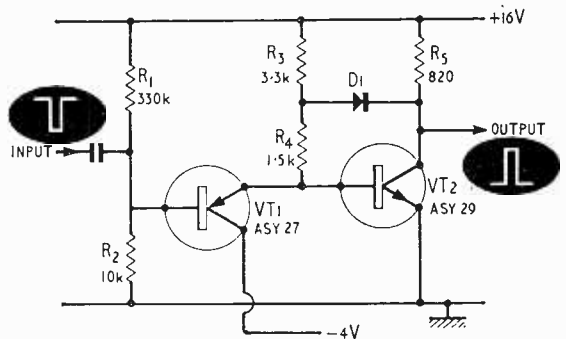
Fig. 7. Pulse shaper and buffer stage.

difference between the two forward-conducting junctions but includes the voltage drop across the emitter and collector extrinsic resistances. The extrinsic resistances cause $V_{CE\ (sat)}$ to increase with increase in bottoming current, but magnitudes of the order of 100mV are typical for either silicon or germanium. Again, because $V_{CE(sat)}$ is the difference voltage between two forward conducting diodes, its temperature coefficient will not be expected to follow the same pattern as for a single diode. The extrinsic resistances, however, are not without their influence on this matter and V_{CE} does have a temperature coefficient which is positive and rather variable between transistors.

To switch on a transistor it is normal to provide a reserve of base current, such that the stage is always fully bottomed—even where the current gain takes a minimum value. In providing ample base current, an excess of charge carriers is introduced to the base region and the removal of this charge causes some delay in switching the stage off again. The hole storage effect is always making its unwanted appearance, and some technique for the controlled bottoming of stages is worth carrying in mind. So finally in Fig. 7 an application for a diode is shown for limiting the base current of a fully on stage to the optimum value.

The switching transistor is VT2; VT1 is normally off and its presence may be ignored for the moment. In the absence of the diode D1, the two base resistors, R_3 and R_4, would supply more than sufficient current to bottom the transistor. However the diode D1 conducts as soon as the collector potential falls

627

below the tap on the bias supply. As the voltage drop across R_4 lowers, so, of course, does the base current until a condition is reached where the transistor is almost but not quite bottoming. The collector potential can never manage to fall below the base potential and so VT2 can never truly enter the bottomed state. There is no hole storage delay in switching off and VT2 base is simply driven negative from a low-impedance driving source (the emitter of VT1). Leakage current is limited to $I_{CBO\ max}$. The output waveform is positive-going for a negative input drive, is of fixed amplitude and generally has faster rise and fall times than the input pulse. The circuit of Fig. 7 has been applied as a buffer stage for " cleaning up " a television camera blanking waveform.

AGARD RADAR SYMPOSIUM

THE Eighth Avionics Panel Symposium was held in September at Church House, Westminster, on " Radar Techniques for Detection, Tracking and Navigation " under the auspices of the Advisory Group for Aeronautical Research and Development, NATO.

Of the 30 papers presented to nearly 400 delegates, some describing recent work others not so recent, those in the systems section proved to be most interesting and occupied more than half the agenda. The section included papers on MTI, Doppler systems, surveillance radars and space applications: this latter inevitably gives rise to radar astronomy, an aspect which was discussed in several papers.

Some of the least co-operative radar targets are the planets, but when the received signal is large enough, so that there is a margin left over from the requirements of straightforward detection, it is possible to gather much information about the target. R. M. Goldstein described methods of analysis (used for Venus) in one of the morning session papers. Total power and power distribution measurements give directly the radar cross section of the planet and can be used to determine the surface characteristics. If a sine wave is transmitted the echo is shifted and broadened in frequency by the Doppler effect. Measurements of the shift have been used to determine the astronomical unit and Venus and Earth ephemerides. The Doppler broadening is caused by apparent rotation of the planet and a spectrogram of the echo power is equivalent to scanning the disc with a narrow slit parallel to the rotational axis so that inferences can be made about the scattering properties of the surface. Information of the nature of the surface can also be revealed by setting the receiver to receive the opposite sense of polarization, hence only even reflections will be received (reflection reverses the sense of circular polarization). A set of range-gates will divide the signal power according to the rotational axis so that inferences can be made trum from each gate a two dimensional picture can be built up.

Moving target indication (MTI) systems using the technique of cancelling the stationary targets by delaying their echoes and using them to subtract the following stationary echo are unsatisfactory for slow target detection on the ground due to a fairly low sub-clutter visibility factor. A matched filter using five channels consisting of a range-gate to which the video-frequency signals are applied and a Doppler filter which is a band-pass filter representing that spectral band of the moving echo spectrum which is nearest to zero frequency has been used and was described by C. van den Broek (Pulse Doppler Radar). For fast targets, the filter covers a bandwidth of 120 c/s-2,000 c/s detecting targets with velocities of about 4 m.p.h.-120 m.p.h. For slow targets the filter has a narrower bandwidth and velocity range is 1.2 m.p.h.-5.5 m.p.h. When the radar is in use, a gating pulse selects a target scale by means of a handwheel, whilst another controls the antenna. The Doppler filter feeds into a loudspeaker so that the controls are operated for maximum volume and give azimuth and range directly. The Doppler sound is used to discriminate between classes of targets. It is hoped that by increasing the number of Doppler filters and strobing gates and by the use of processing systems it may be possible satisfactorily to solve the problem of detection of low-flying aircraft.

Area MTI, the name given to systems which detect targets by change in plan position from scan to scan, makes the least severe demands on a radar system. Storage tubes have been used for some time in area MTI where cancellation of permanent echoes can take place by reading out store information by a separate electron gun and subtracting from incoming radar signals. The direct view storage tube in which a previously written charge pattern can be erased by further " negative " writing should shortly find more application in echo suppression. Another aspect of MTI touched on by P. Bradsell in his paper " MTI, Survey of Developments since 1948 " was the conversion of the video signals into digital form for subtraction. Coherent video signals are quantized into a number of amplitude levels—100 for 40 dB cancellation—then coded into binary words and a computer used for subtraction.

A QUIZ

NO prizes are offered, but do you know the answers to the following questions?

What are the connections for an AF125 transistor?

What length should the radiating element be for a channel 23-33 aerial?

On what frequencies do transmitters for the radio control of models operate?

How many countries have adopted the 525-line standard?

What are the connections for an AF125 transistor?

What is the address of VASCA?

Is a licence needed for " inductive " paging systems?

What channels will be used by the B.B.C.'s first group of high-power u.h.f. television staions?

If you do not know the answers you should get a copy of the *Wireless World* 1965 Diary, the information section of which will answer these and 1001 other questions so often asked of the man in radio and electronics. The Diary, with a week-at-an-opening and an 80-page reference section, costs 7s 9d in leather and 5s 9d in rexine, including purchase tax. Overseas prices are 6s 9d and 5s, postage 4d.

An Introduction to

MICROWAVE TECHNIQUES

5.—MICROWAVE MEASUREMENTS

By K. E. HANCOCK*

IN this, the fifth and final article in this series, we will discuss microwave measurement techniques, using for the most part the theory covered in part two, and the components referred to in part four.

Most measurements made at microwave frequencies fall under three main headings. These are mismatch, attenuation and frequency. Each type of measurement will be discussed, first in its simplest form, then by more unusual or sophisticated methods covering specific measurement requirements.

Measurement of Mismatch

As mentioned in an earlier article in this series, any discontinuity in a waveguide run will give rise to a reflection and hence to a standing wave in the waveguide. The discontinuity, or mismatch, since it is the effect of a change of impedance, may be of any form. It may be a misalignment of flanges, a support in a piece of measuring equipment such as an attenuator, or a change of dielectric in the waveguide such as the glass of a pressure window in a high-power circuit.

The measurement of the amplitude of any mismatch may be achieved by measuring the standing wave caused by it with a standing wave detector or other means, and expressing the mismatch in terms of voltage standing wave ratio or reflection coefficient. This is the type of measurement that will be discussed here, and is sufficient for a large number of applications. It must be remembered, however, that it is in fact only half the story, giving as it does only amplitude information and no indication of phase. The phase angle is of particular importance when we wish to cancel out a standing wave by producing another of equal amplitude but opposite phase, and in several other applications. For this type of work the reader is referred to more advanced reading—see partial bibliography.

First let us consider the measurement of v.s.w.r. up to around 3:1. The setting-up of the circuit will be described in detail, and certain precautions pointed out. It will be apparent that many of these precautions apply to the setting-up of all microwave circuits, and mention of them will be omitted in further circuit descriptions.

The circuit is connected as shown in Fig. 1. This circuit applies to any size of waveguide and hence to any microwave frequency, although at the short millimetre wavelengths some of the components are likely to have become so sophisticated as to bear little resemblance to those illustrated in the previous article. Their funcion, however, is precisely the same.

It should be noted that the symbols used in the circuits, whilst in common use, are not necessarily standard or B.S. approved symbols.

When connecting the circuit, care should be taken that the correct bolts, and dowel pins if called for, are used on the flanges to give precise alignment of the waveguide. Each waveguide component should be checked to ensure that no nuts, bolts and other odds and ends have got into the waveguide. Many most peculiar results have been obtained due to this precaution being overlooked. The writer knows of an installation crew who, whilst fitting large-sized waveguide on an outside site, arrived one bright spring morning to find a couple of birds industriously building a nest in a length of horizontal guide they had fitted the day before! So always look in the waveguide before connecting. Care should also be taken to see that isolators and directional couplers are installed in the correct direction.

*Canadian Marconi Company.

Fig. 1. Basic v.s.w.r. measurement circuit.
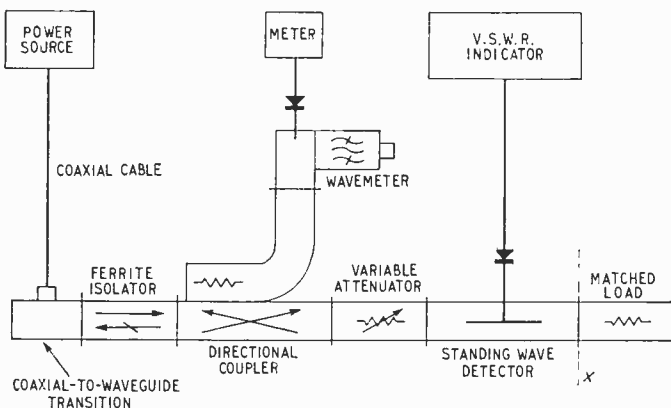
The wavemeter is set to the required frequency and the power source switched on, power source modulation being set to that required and the power source attenuator being set to zero. The power source is tuned until an output is obtained on the wavemeter indicating meter, the power source attenuator being adjusted to reduce power until a sharp precise indication of frequency is obtained.

The variable attenuator, v.s.w.r. indicator attenuator, and standing wave detector matching probes are adjusted until an output is obtained on the v.s.w.r. indicator. The standing wave detector matching probes should now be set for optimum output, and the probe depth, if adjustable, set for minimum depth consistent with adequate output. If possible, a reasonable amount of attenuation should be set on the v.s.w.r. indicator attenuator as there is usually a tendency towards fluctuation of output level if these instruments are used on maximum gain.

The probe carriage of the standing wave indicator is now moved to a point giving maximum output and the v.s.w.r. indicator attenuator adjusted to give full-scale output, that is to show a v.s.w.r. of unity. The probe carriage is now moved to a point of minimum output and the v.s.w.r. read from the scale on the indicator. It should not exceed around 1.01:1. If it does, check such things as the load, probe depth and flange connections until a good match is obtained. The v.s.w.r. obtained at this time will give an indication of the possible maximum error in the final measurement. The component to be measured is now inserted at (xx) in Fig. 1, and the v.s.w.r. of the component measured as shown above, making adjustments only to the v.w.s.r. indicator attenuator and, of course, moving the probe carriage from maximum to minimum output points.

Due to the fact that the scale of the v.s.w.r. indicator follows a square law, and at low power levels a crystal detector gives a linear rather than square law output, measurements of v.s.w.r.'s above about 3:1 tend to be inaccurate with the method shown above.

A modification of the method using the same equipment will, however, allow us to measure high v.s.w.r.'s with a good degree of accuracy. It will be remembered from part two of this series that v.s.w.r. is the ratio of minimum to maximum voltage of a standing wave. Let us then measure this ratio in decibels by means of the calibrated attenuator and convert to voltage ratio by applying the formula:

$$\text{Attenuation (dB)} = 20 \ \log_{10} \ \frac{V_{max}}{V_{min}}$$

where V_{max} = maximum voltage point of standing wave

V_{min} = minimum voltage point of standing wave

With the same circuit as before, with the component to be measured at (xx), set the power source and calibrated attenuators to minimum attenuation.

The probe carriage is now set to a minimum point of the standing wave, and the v.s.w.r. indicator attenuator set to give a convenient output level, say mid-scale on the indicator. The probe carriage is now moved to a maximum point of the standing wave, and the attenuator adjusted to give exactly the same output as before.

If an attenuation of 12.0 dB is measured between minimum and maximum points of the standing wave—

$$12 = 20 \ \log_{10} \ \frac{V_{max}}{V_{min}}$$

Therefore

$$0.6 = \log_{10} \ \frac{V_{max}}{V_{min}}$$

$$\frac{V_{max}}{V_{min}} = 3.981$$

$$\text{V.s.w.r.} = 3.981 : 1$$

Measurement of Attenuation

The measurement of the loss, or attenuation of a signal through a component, is, in its basic form, a very simple one, being a substitution of the component by a calibrated attenuator adjusted to give the same output.

The circuit is set up as shown in Fig. 2. The power source is switched on and the frequency adjusted as in the previous measurement. The rotary vane attenuator is set to minimum attenuation, and the screws of the stub tuner set flush with the inside wall of the guide. The latter component is a piece of waveguide with a number of tapped holes, often three, set on the centre line of one broad wall. In the holes silver-plated screws are set. By suitable adjustment of these screws, a standing wave of any amplitude and phase can be set up. A mismatched waveguide circuit may thus be matched by producing with a stub tuner a standing wave of equal amplitude but opposite phase to that already existing is the circuit, thus cancelling it out. The function of the stub tuner in this circuit is to cancel out the mismatch of the crystal detector, as this is often
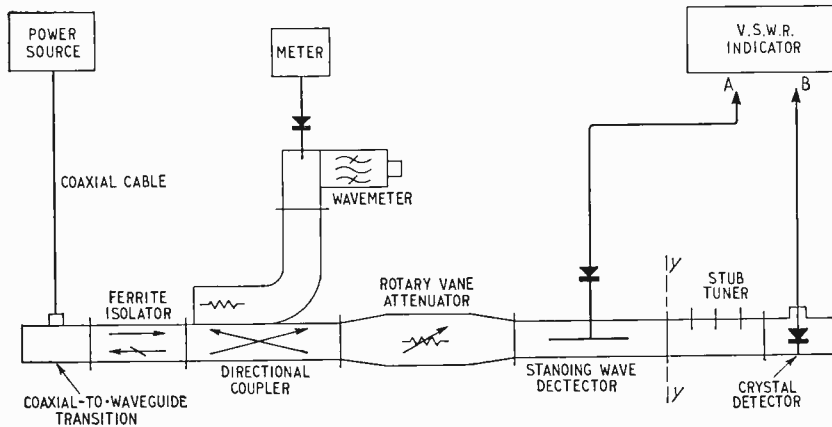

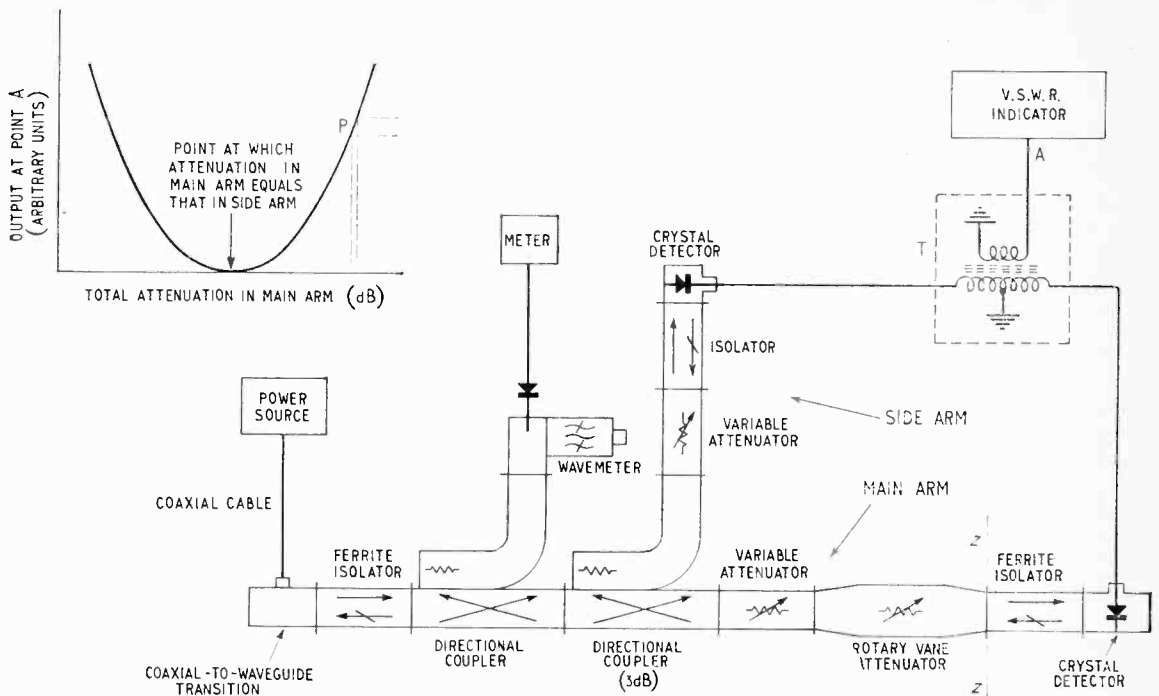Fig. 2. Basic attenuation measurement circuit.

Fig. 3. Bridge method of attenuation measurement.

high. A bad mismatch in the circuit will cause multiple reflections and will substantially reduce the accuracy of the attenuation measurement, particularly when the attenuation to be measured is small.

The v.s.w.r of the circuit is therfore measured by the method shown previously, the output (A) from the standing wave detector being connected to the v.s.w.r. indicator. The screws of the stub tuner are now adjusted, the v.s.w.r being measured at each adjustment until the match is good, say better than 1.05:1. There is no set procedure for adjustment of the stub tuner, the various screws being adjusted in or out in turn, gradually reducing the v.s.w.r.

An isolator which, of course, needs no adjustment, may be used in place of the stub tuner, provided its own match is good.

The output lead (B) from the crystal detector is now connected to the v.s.w.r. indicator in place of the lead (A). The component to be measured is then connected at (yy) in the circuit, care being taken to make tight, well-aligned connections.

The v.s.w.r. indicator attenuator is now adjusted to give a convenient level, perhaps, as before, the mid-point on the scale. The device being measured is now removed and the circuit reconnected. The removal of the instrument being measured will increase the output by an amount equal to loss. The output is therefore attenuated with the rotary attenuator until it is exactly at the previous level, and the attenuation of measured component read from it. As mentioned in the previous article, the scale of the rotary vane attenuator is very wide at the low attenuation end, so this type of attenuator is very useful for the measurement of low values of attenuation. Unfortunately another problem arises with measurements of this type. For small variations

of attenuation the movement for the indicating needle from its reference point will be small, and therefore the effect of not resetting the needle exactly on the reference point, or of any drift in the output, will be large. What is required is some method of increasing the sensitivity of the measurement to ensure greater accuracy.

Fortunately this can be done fairly simply. A circuit is connected as shown in Fig. 3. It is essential, for this circuit to operate in the form shown, that the power source be modulated, preferably at a few kilocycles. The main arm is essentially similar to the basic attenuation measurement circuit, with the addition of a further directional coupler, and a variable attenuator. The second directional coupler, having a 3 dB coupling factor, splits the power equally between the main arm, and the side arm. The action of the circuit is as follows. Assuming the attenuation, from the second directional coupler to the crystal, is equal in each arm, the output in each case will be the same. If the crystal in one detector is of opposite polarity to the other, however, the modulation signal appearing at one output will be in antiphase to that at the other, and when the two outputs are fed together through the balanced split primary transformer (T), give zero output into the v.s.w.r. indicator at point (A). As the attenuation in one arm, say the main arm, varies with respect to the attenuation in the other arm, the output will vary as shown in the graph in Fig. 3. If the attenuation in the main arm is adjusted such that the output is on a steep part of the curve, say at point (P) any small variation of the attenuation in the main line will produce a large variation in the output as shown, thus giving an increased sensitivity over the basic circuit.

To align the circuit, the frequency is checked in

Aligning a 2,000 Mc/s radio link waveguide duplexer.

the normal way and the rotary vane attenuators in the main arm set to zero. The variable attenuator in the main is set to about 1 dB and the component to be measured inserted at (zz).

The attenuator in the power source is set to zero, and that on the v.s.w.r. indicator adjusted to give maximum output consistent with that of stable output. The variable attenuator in the side arm is first set to maximum attenuation, and then slowly varied towards minimum until a null position is indicated on the v.s.w.r. indicator.

Reduce the attenuation in the side arm a little further until a convenient output is obtained. By slight adjustment of the rotary vane attenuator an estimate of the increased sensitivity can be made. The attenuator should then be returned to zero. It will be noted that as attenuation in the main arm is increased, the output increases. By suitable adjustment of the two variable attenuators, any required sensitivity below that obtained with both attenuators in the main arm reading zero may be achieved.

Once the required sensitivity is obtained, and the output adjusted to a convenient level, the two variable attenuators should be locked and not adjusted further. The component to be measured is now removed from (zz) and the circuit reconnected carefully. The rotary vane attenuator is now adjusted until the output is returned to the reference point, and the loss of the component under test obtained. Care should be taken to return the output to the correct side of the null, i.e. output should increase as the attenuation of rotary vane attenuator is increased.

This is a particularly useful measurement technique for measuring small increments of attenuation, and one readily adaptable for swept frequency techniques.

Measurement of Frequency

We have already covered the simple determination of frequency by the use of a wavemeter as part of the previous measurement. Unfortunately, the best wavemeter will only give an accuracy of few parts in ten thousand. For many applications, such as the design of precision bandpass filters, this is not sufficient.

The method commonly used when greater accuracy is required utilizes a crystal-controlled counter and a transfer oscillator. These instruments will give an accuracy of one or two parts in one hundred million. As a counter and transfer oscillator are proprietary instruments, rather than microwave circuits, the detailed method of use varies from maker to maker. A general description of principle and method of operation only can therefore be given here, precise details of any particular make being found in the manufacturers' handbook.

An electronic counter is a device that actually counts the input signal over a precise period of time. The signal is converted to an approximate square wave, and the counting carried out by standard digital methods, the output being displayed optically on cold cathode tubes or similar devices.

The heart of the counter is the gate, the pulse that allows the counter to operate. The gate time is obtained from a highly stable, temperature-controlled crystal oscillator.

Using a counter alone the accuracy of frequency measurement depends on the stability of the oscillator and the inherent plus or minus one count error of any counter.

Unfortunatly counters are practicable only up to around 10 Mc/s, at least at the present. Some way of extending the frequency range, whilst maintain-

ing accuracy, must therefore be devised. The first stage of this process is usually the use of a frequency converter. These units heterodyne the signal to be measured with a harmonic of a fixed-frequency oscillator. The fixed oscillator operates at the highest counter frequency, often 10 Mc/s, thus providing a signal within the range of the counter, whilst maintaining accuracy. Frequency converters normally operate up to the 300 Mc/s range.

The next stage of extending the frequency range is the use of a transfer oscillator. This is a variable frequency device operating within the range of the frequency converter. The transfer oscillator is tuned so that one of its harmonics beats with the unknown signal. The frequency of the transfer oscillator is then measured by the counter and frequency converter and multiplied by the harmonics used.

The beating of the transfer oscillator is usually carried out visually on a cathode ray tube. Error can occur here particularly if the unknown signal is frequency modulated or noisy. However, with care high accuracies can still be obtained.

When a counter and transfer oscillator are used in conjunction with microwave circuit, the input is usually taken via a directional coupler and waveguide-to-coaxial transition. Care should be taken in selecting the coupling factor of the directional coupler so that the maximum input of the transfer oscillator, usually a few tens of milliwatts, is not exceeded.

Conclusion

The aim of this series of articles has been to stimulate the interest of newcomers to the microwave field, and to provide background knowledge for electronic engineers working in other areas.

The coverage was of necessity very general and elementary, and for those interested in pursuing the subject further the partial bibliography given below is recommended.

The books listed have been selected as being those most likely to be of interest to the comparative newcomer to the field, although many are also standard references.

Acknowledgement:—The author thanks Dr. H. A. Hamilton for his valuable suggestions in relation to these articles and is indebted to many of his colleagues for their encouragement and constructive criticism.

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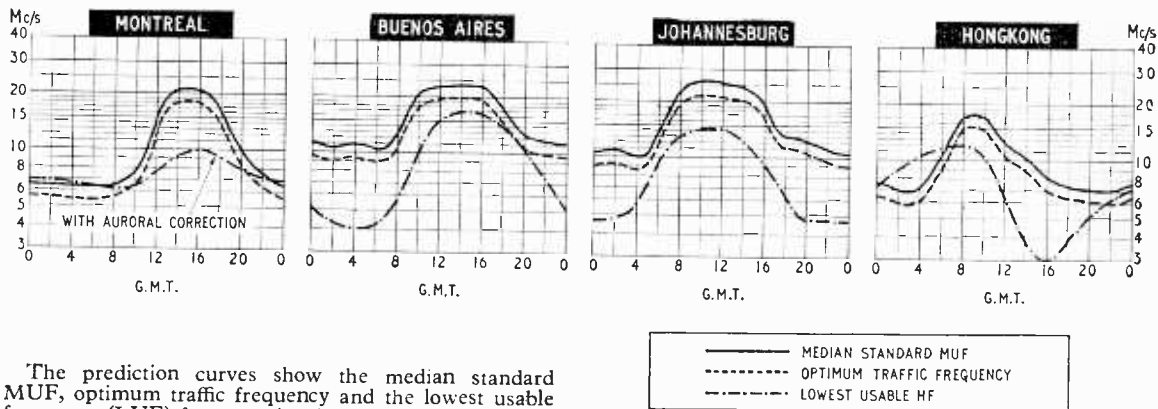
The prediction curves show the median standard MUF, optimum traffic frequency and the lowest usable frequency (LUF) for reception in this country. Unlike the standard MUF, the LUF is closely dependent upon such factors as transmitter power, aerials, and the type of modulation. The LUF curves shown were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

The sunspot number has for several months been hovering about 8. The predicted value of the solar activity index, IF2, has been zero or slightly negative for some time. The lowest running average value of IF2 in 1954 was -6.5. These facts would suggest that the sunspot minimum cannot be far away.

Elements of Transistor Pulse Circuits

11.—TIMEBASES (SWEEP GENERATORS)

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

"MILLER integrators," "bootstrap circuits," "phantastrons" — as soon as the budding engineer starts to take an interest in time bases he runs into a discouraging array of jargon used by the *cognoscenti*. He oftentimes gives up despairingly in what is not really a very difficult area of pulse circuits. This article is an attempt to lead him gently to some understanding of the "arcana."

If you turn to O. S. Puckle's classic "Timebases" (Chapman and Hall Ltd.) you will find a fascinating detailed account of many different timebases likely to be met with in valve practice. With transistors, we will simplify matters by confining our attention to what are called "linear timebases," i.e. circuits producing a waveform that rises steadily with time until it reaches a certain amplitude and returns in a much shorter time to the point from which the linear rise can start again. The forward linear rise is usually called the "sweep" and the sharper return the "flyback."

Different authors use different names to describe this type of generator, e.g. "timebase," "sweep generator," "ramp generator," and "sawtooth generator." Now a timebase output can be either a current or a voltage. Again it can be single-sided or symmetrical push-pull; i.e. it can be a unidirectional sawtooth or can be made up of two sawtooths of equal and opposite polarity. The second type is often required for balanced driving of c.r.t. deflection circuits.

Timebases find their principal use in sweeping the

Fig. 100. Simple transistor sweep generator (driven by square-wave input): (a) without feedback, (b) with feedback.

beam of an oscilloscope across the face of the cathode ray tube to make possible the display of a time-varying signal. In this context, we find them widely used in radar and television circuits, computers, time-measuring devices, and time-modulation equipment. They also find a place in analogue and digital equipment, e.g. in some digital voltmeters.

Driven Non-regenerative Timebases

The simplest type of linear timebase is the circuit which takes in a square wave and gives out a sawtooth, the forward sweep of the output occupying the same time as the input pulse. When the input pulse ceases, the flyback occurs and the circuit returns to its quiescent condition. This sort of circuit is commonly non-regenerative. It really just reshapes the input square wave into a ramp voltage. A whole family of elementary circuits can be used as the basis of this type of non-regenerative timebase.

"Transistor Switch" Timebase

The most primitive form of non-regenerative linear timebase uses a transistor as a simple switch to control an RC relaxation network. The basic circuit arrangement is shown in Fig. 100(a). Here the transistor Q is set up so that, in the absence of any signal at the input, it is bottomed. This means that the current through the base resistor, R_B, holds the transistor fully switched on, and the collector output potential is close to zero volts. In this condition, the capacitor C_T is discharged. If now a positive pulse of duration T is applied to the input *via* the large coupling capacitor C_B, the transistor base is immediately driven positive and the transistor cuts off for the duration of the pulse. With Q cut off, the capacitor C_T begins to charge negatively through resistor R_T. This is shown in the output waveform in Fig. 100(a) for the interval T. At the end of the pulse, the positive voltage is removed from the base, and the transistor is driven on again through R_B. The transistor bottoms and presents a low impedance across the capacitor C_T, causing it to discharge rapidly, and bringing the collector potential quickly back close to zero volts again as shown. The output sweep voltage during the interval T is exponential in form, but, provided the pulse time T is small compared with the time constant $C_T R_T$, the sweep will be a reasonable approximation to a linear waveform.

Another circuit which on paper looks very similar to the last one is given in Fig. 100(b). This too is driven by a positive square-wave input pulse, but

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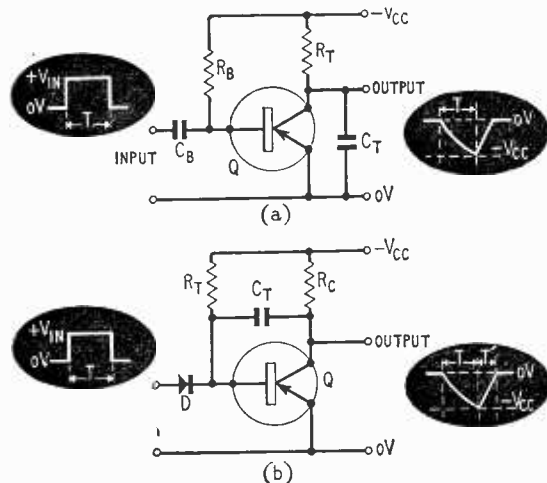
the action is somewhat different due to the negative feedback capacitor C_T, connected between the transistor collector and base. As before, in the quiescent condition, the transistor is held hard on by the base current through the resistor R_T from the negative supply rail. When the pulse arrives, it takes the base positive and cuts the transistor off. The capacitor C_T then begins to charge up through R_C with a time constant $C_T R_C$ as shown for the interval T in the output waveform of Fig. 100(b). When the input pulse ends, the positive voltage is removed from the base, and the transistor base drive current through R_T is restored. The transistor does not, however, return instantaneously to the bottomed condition where its potential is close to zero volts, because its collector potential is held up initially by the charge on C_T. C_T then discharges through R_T with a time constant proportional to $C_T R_T$, as shown in the output waveform for the second interval marked T′. This linear " secondary " waveform can be used as the forward sweep voltage of a timebase, and its rise time is a function of C_T, R_T and the h_{FE} of the transistor.

This circuit is a sort of " phantastron." (In valve practice the " phantastron " was a circuit in which the time base was triggered off by a short pulse at the input, and then the circuit supplied its own gating to continue the linear sweep. The pulse merely initiated the sweep, which then " runs down " without any further help from the trigger input. The circuit of Fig. 100(b) really belongs more correctly to the family of Miller integrators to be described below.

Miller Integrator Timebase

The circuit of Fig. 101(a) is the Miller integrator sweep generator. This widely-used circuit works as follows. It is biased so that when the switch Sw is closed (as it normally is) the transistor Q is held cut off, because the base is short-circuited to earth, and no base drive current flows into the transistor. Thus the collector is virtually at the negative rail voltage and the base at 0V. As a result, C_T is negatively charged to $-V_{CC}$. If now the switch is opened at time T_1 and closed at time T_2, the voltage at the collector of the transistor takes the waveform shown. Immediately the switch is opened, the base goes a few hundred millivolts negative, when the input current is switched into the base through R_T. This small negative step voltage is transmitted *via* C_T to the collector, and appears as a small negative-going jump at the point T_1 in the output waveform (which is characteristic of the basic Miller integrator). This can be shown to be due to the finite output resistance of the transistor, and can be reduced to zero by including a suitable resistor in series with C_T. In the simple version shown, once Sw has been opened, C_T begins to discharge through R_T, with a time constant approximately proportional to $C_T R_T$. However, the ramp output waveform corresponding to this is not exponential but linear. This is because of the feedback action explained in an earlier chapter on operational amplifiers. During the process of discharge, the current through C_T remains constant. When the switch Sw is closed again at time T_2, the transistor is switched off again and C_F discharges through R_C, to give a waveshape returning finally to $-V_{CC}$ as shown. R_C is usually much smaller than R_T so that the recovery or flyback time is much shorter than the forward sweep time.
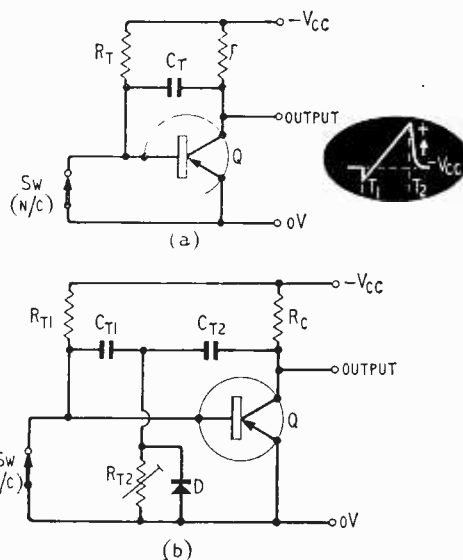

Fig. 101. Miller sweep generator: (a) simple basic circuit, (b) with additional linearizing network.

Although the Miller integrator circuit produces a ramp waveform much more linear than that which would be achieved merely by discharging a capacitor through a resistor, its linearity can be even further improved by additional feedback circuits. A commonly used one is illustrated in Fig. 101(b). Here the feedback timing capacitor C_T has been split into two capacitors C_{T1} and C_{T2}, and a resistance R_{T2} taken from their midpoint to earth. Very commonly C_{T1} and C_{T2} are made equal and R_{T2} is made variable so that the linearity can be adjusted. The diode D across R_{T2} serves the purpose of speeding up the recovery time of the circuit.

In both versions of the Miller circuit given, the switch Sw is shown as a normally closed mechanical switch. In practice a transistor is very often used for this purpose. In the quiescent condition this switch transistor is held bottomed, and its collector effectually presents a short circuit to the base of the Miller circuit transistor. When the switch transistor is switched off, it becomes effectually an open circuit, and we have a direct equivalent of a mechanical on-off switch.

Bootstrap Sweep Generators

Another very common non-regenerative driven sweep generator circuit is the " bootstrap " integrator. In its simplest form this takes the shape shown in Fig. 102(a). It is essentially an emitter-follower with feedback from the emitter *via* a capacitor C_F to the junction point of two resistors R_F and R_T feeding current into the transistor base, with a capacitor C_T from the base to earth. The capacitor C_T is normally short-circuited by the switch Sw. The transistor base being thus connected to earth, the transistor is cut off, and its emitter is at zero volts. If the switch Sw is now opened at time T_1, the output waveform takes the shape shown in the diagram. C_T begins to charge up through R_F and R_T. As the voltage on the base of the transistor begins to go negative, so does the emitter
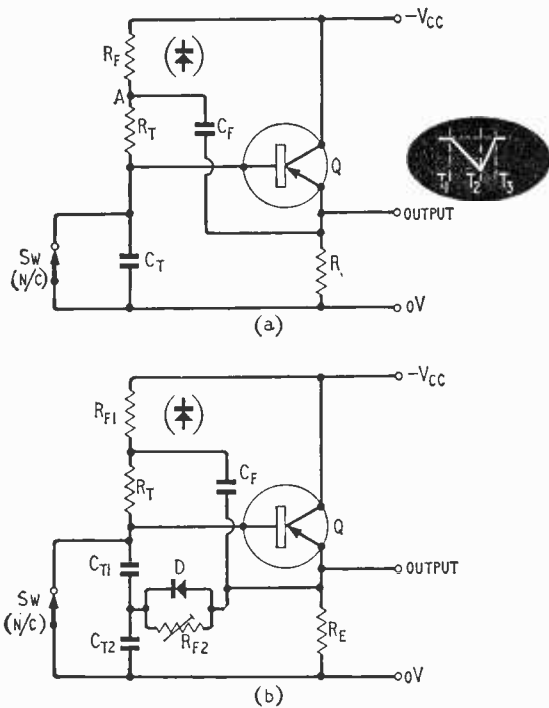
Fig. 102. Bootstrap sweep generator: (a) simple basic circuit (b) with additional linearizing network.

regenerative timebases where the action is merely triggered off by a short pulse and continues thereafter on its own. The commonest of these are the blocking oscillator and the multivibrator.

Blocking Oscillator Sweep Generators

In an earlier article we discussed in some detail the operation of the blocking oscillator circuit. One use of this circuit as a sweep generator is demonstrated in basic form in Fig. 103(a). Here the transistor base is normally d.c. short-circuited to earth *via* the transformer feedback winding, so that the transistor is cut off. The transistor collector and the output point A are both virtually at negative rail potential. Capacitor C_T is therefore charged up to $-V_{CC}$. If now a negative trigger pulse is applied at time T_1 through capacitor C_B to the transistor base, blocking action sets in and the transistor bottoms firmly for the blocking period of the oscillator. The capacitor C_T is rapidly discharged and the output voltage at point A moves rapidly to near zero volts, remaining there until time T_2, the end of the blocking pulse. The transistor now cuts itself off and the capacitor C_T begins to charge up negatively through R_T towards the rail voltage, as shown in the second part of the waveshape in Fig. 103(a),

output voltage through emitter-follower action. This emitter voltage change is transferred through the large feedback capacitor C_F to the top end of the resistor R_T, point A. As a result the voltage across the resistor R_T remains virtually constant, and provides a fixed charging current for the timing capacitor C_T. The output voltage follows the base voltage closely and we have the linear ramp shown in Fig. 102(a). At time T_2 the switch Sw is closed again. The transistor then cuts off, and the capacitor C_T discharges rapidly. This gives the flyback path from T_2 to T_3 shown in the circuit diagram illustration. The resistor R_F is often replaced by a diode as shown in brackets to remove the limitation on the recovery time which is set by the charging time constant $C_F R_F$.

The linearity of the sweep voltage is good in the bootstrap circuit, but it can be improved even further by using some form of compensated feedback circuit. Typical of such linearizing circuits is the arrangement shown in Fig. 102(b). Here the timing capacitor C_T is split into two parts C_{T1} and C_{T2}, and feedback is applied to the centre point via a resistor R_{F2}. The similarity to the circuit of Fig. 101(b) is immediately evident. The shunting diode D is again used to speed up recovery time. A common rule of thumb is to make C_{T1} and C_{T2} equal, and use a variable R_{F2} to adjust for best linearity.

Basic Triggered Sweep Generators

Up till now we have been considering mainly non-regenerative circuits, where the output sweep continues only as long as the input square wave continues. There is also, however, a family of
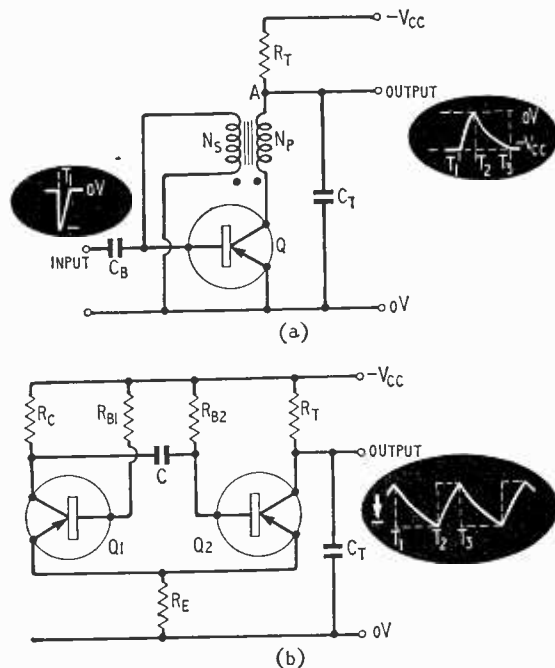

Fig. 103. Regenerative sweep generators: (a) blocking oscillator, (b) multivibrator.

eventually returning to $-V_{CC}$ at time T_3. During the interval T_2, T_3 the output voltage provides a ramp suitable for sweep generator use. The sweep is, of course, exponentially non-linear because it merely represents the charging up of the capacitor C_T through a resistor R_T from a fixed voltage $-V_{CC}$.

Various circuits can be used, however, to linearize the sweep output.

Multivibrator Regenerative Sweep Generators

By the same sort of arrangement where a timing capacitor is shunted across an output resistor, the multivibrator (also discussed in previous articles) can be used for sweep generation. One example of this is given in Fig. 103(b). Here an emitter-coupled astable multivibrator is set up with a timing capacitor C_T connected from the right-hand collector to earth. Without the capacitor C_T, the output would have the waveshape shown dotted in the diagram. The addition of the capacitor changes the output to the sawtooth waveshape shown. From T_1 to T_2 the transistor Q_2 is cut off, and C_T charges negatively through R_T with an approximate time constant $C_T R_T$. From T_2 to T_3 when the transistor Q_2 is switched hard on, the capacitor C_T discharges through R_E. As R_E is usually designed to be much lower in value than R_T, the recovery time T_3-T_2 is much shorter than the forward sweep time T_2-T_1. By suitable choice of the time constants in the multivibrator and C_T it is possible to make the recovery time much shorter than the sweep time, and give a fair approximation to a linear sweep generator. Of course, without feedback, the sweep must be quasi-exponential.

Sweep Generator Linearity Considerations

We have skated rather lightly over linearity considerations up to this point, but as they are of great importance in practical applications, we must take a closer look at the requirements in this respect. In the case of a general purpose c.r.t. sweep circuit, one important requirement of the sweep is that the sweep speed, i.e. the rate of change of output signal with time, be constant. A useful definition of deviation from linearity in this case is the ratio of the difference in slope at the beginning and the end of the sweep to the initial value. This is known as the slope (or sweep-speed) error and is usually denoted by the symbol ϵ_s.

In connection with other timing applications, however, a more important index of linearity is the maximum difference between the actual sweep signal and the linear sweep which passes through the beginning and the end points of the actual sweep as shown in Fig. 104. This index is called the displacement error, usually denoted ϵ_d, and defined as the ratio of the maximum deviation of the actual from the linear sweep to the final sweep amplitude. It is usually expressed as a percentage rather than as a fraction.

Another non-linearity index sometimes met with is the transmission error, ϵ_T. This arises when a sweep voltage is transmitted through an RC high-pass network and the output falls away from the linear sweep towards the end. The transmission error is defined as the ratio of the output "droop" to the input signal measured at the end of the sweep.

Provided the deviation from linearity due to the phenomena mentioned is small, so that the actual sweep voltage may be approximated by the sum of a linear and a quadratic term in T, it can be shown from the above definitions that to a first approximation $\epsilon_d = \dfrac{\epsilon_s}{8} = \dfrac{\epsilon_T}{4}$. In the remainder of this

discussion, therefore, we will confine our considerations to the displacement error, ϵ_d, to simplify matters.

To put some values to the errors discussed above, so you can see what sort of linearity to expect, it can be taken as a rule of thumb that for a general-purpose oscilloscope a displacement error of about 1.0-1.25% is tolerable. Now for the straight exponential charging-up of a capacitor C through a resistor R from voltage V_s, it can be shown that the displacement error is $V_{max}/(8V_s)$ where V_{max} is the voltage to which the capacitor is allowed to charge by the end of the sweep. If the sweep time is substantially less than the CR time constant, the displacement error can then be shown to be approximately $T/(8RC)$. Hence, if the sweep is to be reasonably linear the time constant RC must be large compared with the sweep time, T. In general-purpose oscilloscopes, the circuits are usually arranged so that V_{max} is about a tenth of V_s. This gives a displacement error of about 1.25% (as noted earlier).
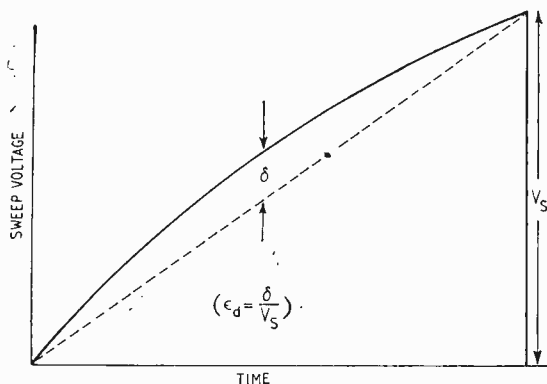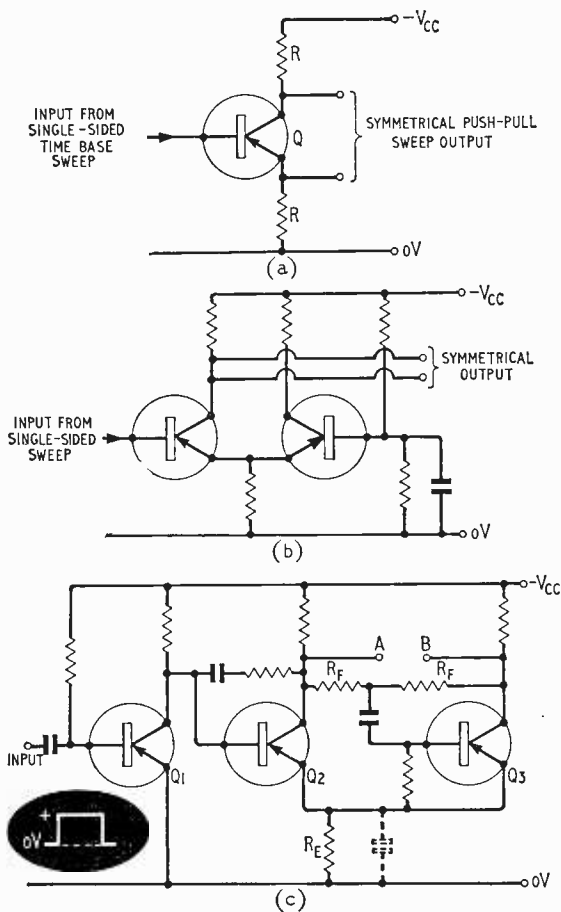
In many timing applications, however, a much

Fig. 104. Displacement errors, ϵ_d, in linear sweep.

higher precision that this is required, and it is usually then necessary to go to the Miller integrator or bootstrap circuits. As a rough approximation, it can be taken that with a Miller integrator the linearity is improved by a factor of A, where A is the voltage gain of the Miller amplifier. Similarly in the case of the bootstrap, the linearity is improved by a factor of $1/(1-A')$, where A' is the voltage gain of the emitter follower. As A' is slightly less than unity, $1/(1-A')$ is quite large and will be of the same order of magnitude as A in the Miller integrator case. This illustrates clearly why in the main Miller and bootstrap integrators are used in high-precision sweep generators.

Apart from the basic circuit arrangement, various compensating networks are added in practice to improve linearity even further. Two of these have been illustrated earlier in Fig. 101(b) and 102(b).
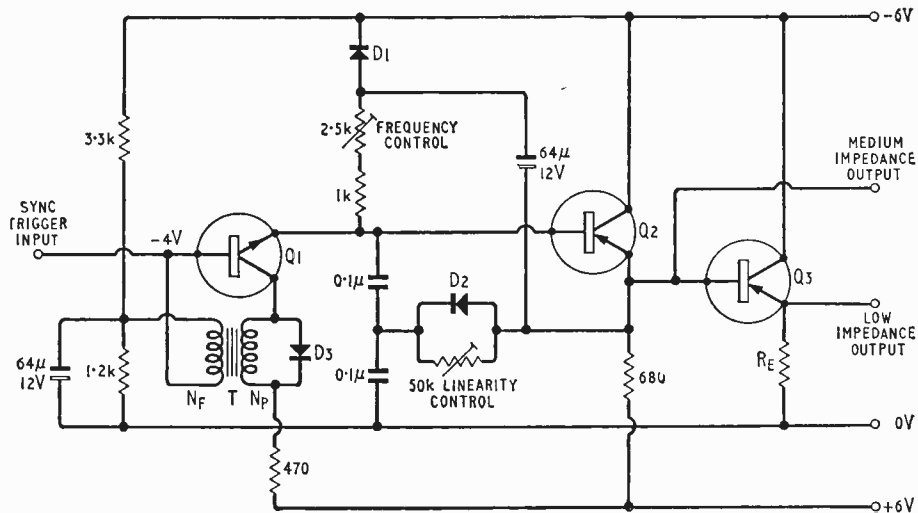
Symmetrical Timebases

So far we have been concerned mainly with single-ended timebases. Often there is a requirement for a symmetrical push-pull timebase. The simplest way to achieve this is to feed the output of a single-ended timebase into a phase splitter. In Fig. 105(a) this is done by a conventional single-transistor phase-splitter. A more refined splitter is the long-

tailed pair circuit shown in Fig. 105(b). Finally a separate phase-inverting transistor may be inserted. A typical sort of circuit with this last arrangement is given in Fig. 105(c). Here Q1 is the " switch " transistor for the Miller integrator transistor Q2. The output from Q2 collector is fed via R_F to the base of the inverter transistor Q3, and an equal resistor R_F is used for feedback from the collector to the base of Q3. Thus Q3 acts as an operational phase inverter and the output at the Q3 collector is positive-going when the Q2 collector is negative-going. The symmetrical antiphase voltages at points A and B can be used to drive the horizontal plates of a cathode ray tube. The emitter resistor R_E common to Q2 and Q3 does not introduce degeneration into the Q2, Q3 circuit because the current through it remains constant, since the sum of the currents through Q2 and Q3 does not change. Sometimes a decoupling capacitor is used (as shown dotted) to eliminate switching transients.

Free-running Timebases

The timebases discussed so far have been mostly " driven " by a train of input pulses. In practice timebases are often required to be free running. An illustration of such a circuit operating at about 10kc/s is shown in Fig. 106. In this the stage Incorporating the n-p-n transistor Q1 is an astable blocking oscillator, which supplies the drive to the bootstrap integrator circuit Q2. The output from Q2 can be taken off directly at its emitter as a medium-impedance output or, if a higher load current is required, can be taken off at the output of the second buffer emitter follower Q3. From either output a voltage sweep is provided. Although the free running frequency of the blocking oscillator is about 10kc/s, its frequency can be varied over a range by means of the 2.5kΩ variable resistor shown as " frequency control." Also it is possible to feed a train of synchronising input pulses into the base of Q1 to synchronize the timebase accurately to an external control frequency, provided the control frequency is slightly higher than the preset free-running frequency of the blocking oscillator. The variable 50kΩ resistor is provided to enable the linearity of the output sweep voltage to be optimized.

Above: Fig. 105. Symmetrical sweep generators: (a) phase splitter inversion of single-ended sweep (b) long-tailed pair balanced inversion, (c) separate phase-inversion, transistor.

Below: Fig. 106. Practical example of complete free-running voltage sweep generator.

Current Timebases

Up to this point we have tacitly been considering voltage rather than current sweeps. In electromagnetic deflection systems, however, there is a specific requirement for a current rather than a voltage sweep. How a current sweep is achieved depends on whether the output transistor sees an inductive or a resitive load.

In television field timebases, which run at relatively slow speed (e.g., 50c/s), the impedance of the field coils is for all practical purposes resistive. In this case, the linear current sweep is effected by using a transistor as a linear power amplifier as shown in Fig. 107(a). A linear sweep (sawtooth) voltage is
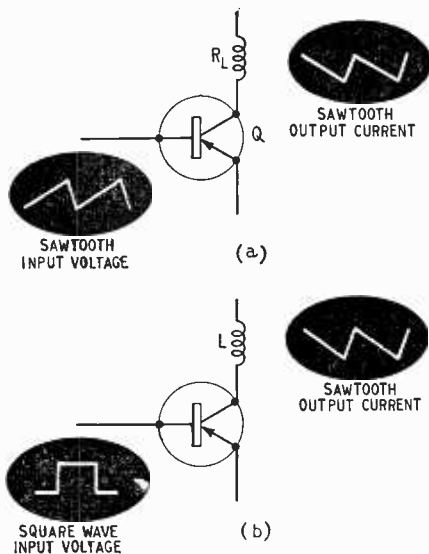
R_L

SAWTOOTH
OUTPUT CURRENT

Q

SAWTOOTH
INPUT VOLTAGE   (a)

L

SAWTOOTH
OUTPUT CURRENT

SQUARE WAVE
INPUT VOLTAGE   (b)

Fig. 107. Current timebases: (a) slow speed (load predominantly resistive, (b) high speed (load inductive).

fed to the transistor base, and produces a sawtooth collector current in the deflection coil R_L. The inductance of R_L is so low compared with the load resistance at the slow sweep speed that the usual $L \dfrac{di}{dt}$ equation does not apply.

In line timebases, on the other hand (operating at greater than 10kc/s) the deflection coils are largely inductive at this high line frequency. To get a linear output current sweep it is necessary to feed a square wave to the transistor input as shown in Fig. 107(b). In effect the transistor operates as a switch to transfer the supply voltage across the deflection coil for the sweep duration. The equation $V = L \dfrac{di}{dt}$ than gives a quasi-linear rising coil sweep current. For a fuller discussion of current timebases, readers should consult "Transistor Television Receivers" by T. D. Towers (Iliffe Books Ltd.).

Conclusion

In this article we have had a look at the types of sweep generators that have been worked out with conventional transistors, but it should not be overlooked that over recent years new generations of semiconductor devices have grown up which in some cases have proved peculiarly suited to sweep generators. V.h.f. silicon epitaxial transistors enable timebases with very fast speeds to be devised, while field effect and metal-oxide-silicon type transistors with their exceptional low leakage and high input resistance make possible timebases with very long periods. Other devices which have been widely used in timebase circuits are unijunctions, silicon controlled rectifiers and GTO's (gate turn-off units). In general, however, it will be found that these more refined circuits tend merely to be adaptations of the basic " switch," " Miller " or " bootstrap " circuits described earlier.

It would not be fitting to close this final article of the series without placing on record my appreciation of the help I have had from discussions with many engineers on the topics covered. It would be diffcult to give all the credits due, but I should particularly like to express my gratitude to some of my colleagues at Newmarket Transistors, in particular Messrs. T. J. H. Dilley, R. A. L. Warrington, K. L. Hall and S. Mitra.

Commercial Literature

The 1964/65 edition of the " **Mazda Valves and Picture Tubes** " pocket data booklet has recently been published and is available, free of charge, from the publicity department of Thorn-AEI Radio Valves and Tubes Ltd., 155 Charing Cross Road, London, W.C.2. It runs to 160 pages and gives abridged information on 259 current and obsolescent valves and c.r. tubes. A complete list of obsolete Mazda types and their applications is also included for reference. The equivalents list now contains over 1,200 fully cross-indexed types.
15WW 301 for further details

Mazda have also recently publicized their pocket data booklet " **Mazda Semiconductors.**" This covers devices used in the entertainment field and does not include industrial semiconductors. Features of the booklet include a " Comparables List ", which can be used as a guide for substitution and a "Device Identification List", which enables the user to determine the name of manufacturer from a semiconductor's type number. More than 220 devices are listed in this section. Copies of the booklet are also available from the publicity department of Thorn-A.E.I. Radio Valves and Tubes Ltd.
15WW 302 for further details

Avo Valve Manual.—The 16th edition of the Avo Valve Data Manual, price 35s, covers some 6,000 different types of valves, including a number now obsolete. A number of cross-reference tables are included covering Service types to Commercial types and vice-versa, the American Army VT Series to British Service types, the British Post Office VT Series to British Service types, and B.V.A. Utility equivalents. Copies are available from 92-96 Vauxhall Bridge Road, London, S.W.1.

Details of the extended range of Belclere transformer design kits are given in a leaflet available from the Belclere Company, 385-387 Cowley Road, Oxford. This brings the total number of different kits to 50.
15WW 303 for further details

Correction.—The address of the United Kingdom agents of the German Elac organization, the High Fidelity Centre, is 61 West Street, Dorking, Surrey, and not as given in last month's issue. A leaflet describing a unit comprising a transcription unit (mentioned last month) and a stereo transistor amplifier with a 16-watt per channel rating is now available. A technical specification is given.
15WW 304 for further details

NEWS FROM INDUSTRY

d-mac limited (*sic*) is the new name of Dobbie McInnes (Electronics) Ltd., of 55 Kelvin Avenue, Glasgow, which was formed in 1952 by the parent company Dobbie McInnes Ltd. to concentrate its electronics interest. In 1958 it was decided to increase the capital of the company with money from outside Dobbie McInnes to enable the company to become independent. Dr. A. R. Boyle, the technical director, then became managing director.

Submarine Cables Ltd. have received a £500,000 contract from the British Post Office for a repeatered submarine cable, providing 480 telephone circuits, to be installed between Bournemouth and Jersey in 1967. The contract also includes special transistor terminal-station equipment made by A.E.I., one of Submarine Cables parent companies —the other is B.I.C.C.

The **M-O Valve Company** has received an order from the British Post Office for a number of high-power travelling-wave tubes, which are to be used at the Goonhilly Down satellite ground station. Their output is rated at 10 kW at a frequency of 6301 Mc/s.

The **Canadian Westinghouse Company** has been awarded a contract for some $500,000 worth of mobile electronics equipment to test and calibrate the airborne fire control systems in the Royal Canadian Air Force CF-104 type of aircraft while on the ground.

Winston Electronics Ltd., who recently became the exclusive manufacturing licensees of International Electronics Research Corporation, of California, U.S.A., have created a new division to make and market the I.E.R.C. range of heat dissipating and mounting devices for valves and semiconductors. Known as the I.E.R.C. Division, it will operate from Govett Avenue, Shepperton, Middx. (Tel.: Walton-on-Thames 26321.)

EMI to Market Olivetti Control System.—The Italian Olivetti continuous-path machine tool control system is to be marketed and maintained by EMI Electronics Ltd. in the United Kingdom and a number of other countries on the Continent and in the Commonwealth. EMI Electronics will also provide a computer programming service for the recorded magnetic tapes used with the Olivetti system of numerically controlled machine tools.

Monitoring Receivers:—An order for twenty-two additional BRT 400 communications receivers for the B.B.C.'s receiving station at Crowsley Park, near Reading, has been placed with G.E.C. (Electronics) Ltd. Already more than twenty of these receivers, which cover 150 to 385 kc/s and 0·51 to 30·0 Mc/s, are being used at this station. Nearly 3.000 of these receivers. which were first introduced in 1947, are now in use throughout the world.
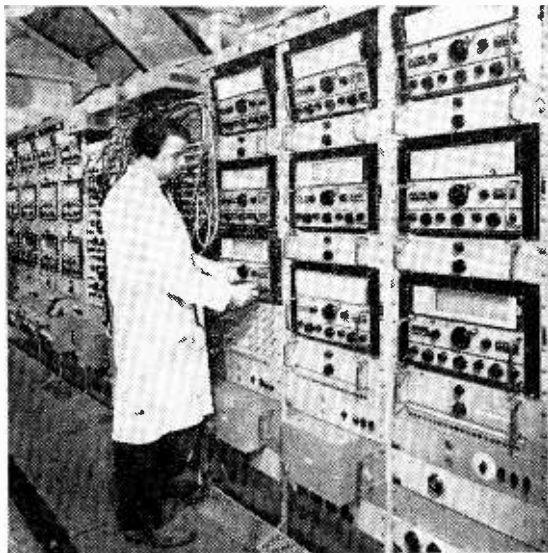
Newton Brothers (Derby) Ltd., Alfreton Road, Derby (Tel.: Derby 47676) have been appointed U.K. agents for a number of overseas manufacturers of electrical and electronic equipment. Included are **Gebr. Ruhstrat,** of Western Germany, producers of variable transformers, fixed and variable resistors and potentiometers, and **Nordqvist & Berg,** of Sweden, manufacturers of transistor stabilized power supplies.

Livingston Laboratories Ltd. have been appointed sole United Kingdom agents for the sale of " Flexwell " waveguides, high-frequency cables and associated accessories manufactured by Hackethal-Draht-und Kabel-Kerke Aktiengesellschaft, of Hanover, West Germany.

German Outlet for Ferranti.—The automatic systems division of Ferranti Ltd., of Wythenshawe, Manchester, has entered into a marketing agreement with the recently formed German company Gesellschaft fur Rechnergesteuerte Anlagen mbH (G.R.A.), of Wuppertal. Under the agreement, the company will market Ferranti automatic control systems in the Federal Republic of Germany.

Denham & Morley Ltd., of Denmore House, 173-175 Cleveland Street, London, W.1, have been appointed U.K. agents for **Electronics Industry,** of Yugoslavia. Goods made by E.I. include radio components and a range of domestic receivers and radiograms.

The Gas Purification and Chemical Company announce a pre-tax profit for the year ended 30th June of £661,765. This represents an increase of £208,051 on the previous year's result. Taxation provisions this year totalled £351,729 compared with £226,007 in 1962/63. Subsidiaries of this company include A.B. Metal Products, B & R Relays, Greencoat Industries, Grundig (Great Britain) and Wolsey Electronics.

British Electric Resistance Company pre-tax profit for the year ended 31st July totalled £84,863 compared with £55,494 in the previous year. Tax this year took £43,800 (£29,600) leaving a net profit of £41,063 (£25,894). Tax over-provision of £1,472 (£642) is to be credited to the 1963/64 result.

The **C.G.S. Resistance Company** have recently completed an extension to their existing factory at Marsh Lane, Lymington, Hants. The new building houses a considerably enlarged winding section, fitted with up-to-date resistance winding and testing equipment.

Associated American Winding Machinery Inc., of New York, have been appointed the main distributors of **Avo instruments** for the whole of the United States.

Claude Lyons Ltd. have been appointed exclusive U.K. distributors for **Spectra-Physics Inc.,** of Mountain View, California, U.S.A., who manufacture a range of helium-neon gas lasers.

Industrial Type Drawer Slides.— A comprehensive range of drawer slides manufactured by the Grant Pulley and Hardware Corporation, of New York, is now being handled through its recently set-up British subsidiary Grant Slides Ltd., 16 Frederick Street, Brighton, Sussex. Many of these slides have been specifically designed for the electronics industry and are now available through the U.K. distributors Technical Components (Weybridge) Ltd., of Farnham, Surrey.

New London headquarters for **Standard Telephones & Cables Ltd.** are being built in the Strand opposite St. Clement Danes church. The ten-storey building, with a floor area of 47,500 sq ft, will be completed next spring and is to be known as " S.T.C. House."

International Computers and Tabulators' head office has been transferred to I.C.T. House (formerly Hotpoint House), Putney, London, S.W.15. (Tel.: PUTney 7272.) The former I.C.T. House in Putney has been renamed Bridge House and will continue to be the company's sales headquarters. (Tel.: RENown 3322.) The premises at 17 Old Park Lane has been vacated.

Ferranti Ltd. has moved its London office from Kern House, Kingsway, to Millbank Tower, Millbank S.W.1. (Tel.: VICtoria 6611.)

The Closed Circuit Television Development Laboratories of the **Marconi Company** are now at The Lane Works, Waterhouse Lane, Chelmsford, Essex. (Tel.: Chelmsford 3281.)

Aircraft-Marine Products (G.B.) Ltd. moved from Saffron House, London, E.C.1, to new and larger offices at Terminal House, Stanmore, Middx., during September. All administration, sales, data processing, production planning and control, research and development are centralized at Terminal House. (Tel.: STOnegrove 6341.)

A joint sales department is announced by **Pamphonic Reproducers Ltd. and W. Brian Savage Ltd.** at Heath Works, Baldcck Road, Royston, Herts. (Tel.: Royston 2412.)
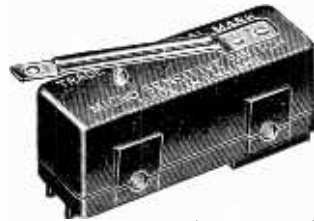
Plantex Electronics Ltd., manufacturers and contractors to the radio and electronics industry, have moved from Bowes Park to 11 Mayes Road, Wood Green, London, N.22. (Tel.: MULberry 1091.)

George Kent Ltd. has moved its registered office and London sales office from 199-201 High Holborn to Kingsway House, 103 Kingsway, London, W.C.2. During September an office was opened at 14-17 George Road, Edgbaston, B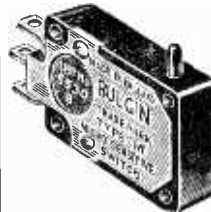irmingham, 15. (Tel.: Edgbaston 5385.)