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COMMENTARY

AFTER the Salon Internationale des Composants in Paris and in anticipation of this month's Radio and Components Exhibition in London it is, perhaps, instructive to consider the future of the components industry.

The plotting of trends in statistics, unless all the variables are accounted for, is a somewhat dangerous pastime as everyone in market research is aware; but the immediate past in America is often an indication of the immediate future in Europe.

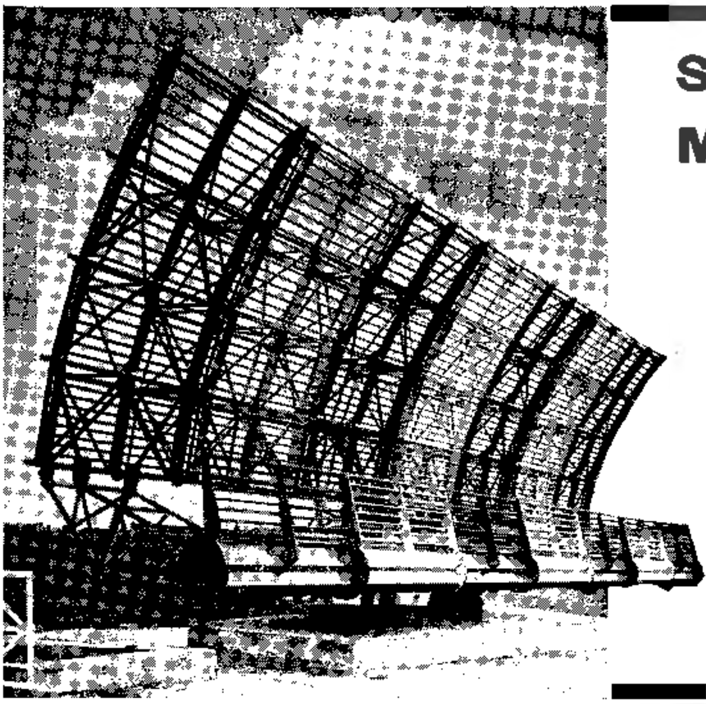
The sale of electronic valves in the U.S.A. increased by 27 per cent in 1966 compared with the previous year. Similarly, capacitor and resistor sales each jumped by 12 to 13 per cent. These three types of component have as their major competitors semiconductors and micro-electronics, both of which have shown phenomenal increases in sales during the last year. It has been natural to expect that the older electronic components would suffer a severe setback in sales but such is not the case. Indeed, intrepid visionaries say that further percentage increases are to be expected during the current year. It is quite clear, however, that both old and new devices will plough their own furrows for a long time, due partly to production economics and partly to their inherent limitations.

The long awaited entry into the domestic market of integrated circuits is not yet apparent in Europe. At the Paris components exhibition there were a very large number of passive and active circuits intended for professional as against consumer equipment. The high costs involved in producing special circuits cannot be offset without a correspondingly high guaranteed market and this has, no doubt, inhibited the electronics industry from venturing far into microcircuits for the general public. The advances made in improving the characteristics of the older established components since the war have been quite adequate to provide the wherewithal for the greater part of our electronic needs, domestic and professional. It is only where space considerations and/or the need for the utmost reliability are forced upon a designer that he normally turns to microcircuits, e.g. in space technology and military equipment, achieving ultimate satisfaction at a cost.

The imminence of colour television may change the situation so that microcircuits become commonplace. Unfortunately, the television receiver manufacturers experienced a drop in home sales of monochrome receivers in 1966 and this may very well influence long term decisions on the future of colour television design.

It has been forecast in America that France will move into second place during this year in the European electronic industry production, pushing out the U.K. by a small margin; according to this source Germany is now in the first place and is likely so to remain while the U.K. will fall into third place. The F.N.I.E. component figures for the French industry given out at the Salon in Paris show a steady rise each year in sales of electronic products generally. Components, however, have rather dragged their heels and do not show such a high rate of growth; at the same time the level of imports of components remains at a high figure, somewhat exceeding the level of exports from France. This state of affairs bears out what M. Henri Nozières said to the press conference which opened the Salon: there are two parts to the electronic industry, the professional and the public and it is evident that the former is tending to become of greater importance than the latter. At the present time a proportion of components for professional equipment has to be imported and as electronics penetrates further into general industry so the level of imports rises to satisfy this expansion.

The future scientific development of solid-state electronic components has intriguing possibilities and Prof. Werner Kleen, head of the Siemens Research Laboratory, outlined in a paper given during the Paris Exhibition the state of r. and d. from the physicist's point of view. Some of the more recent results await evaluation and, like lasers a little while ago, are in the state of solutions searching for problems to solve. There are one or two devices which he indicated may be used successfully at engineering levels. One can, for instance, use a photo-electronic device, composed of luminescent gallium arsenide and a silicon photocell as a sensor, as a very high-speed relay for telegraph communications. Oscillations on a continuous scale up to 30GHz are obtainable quite simply with Gunn effect diodes, Prof. Kleen foresaw the general introduction of varactor diodes as harmonic oscillators in radio and television receivers at all frequencies likely to be encountered. But like Nozière, Kleen saw a bright future for electronic valves, particularly those in which transit-time effects have been avoided.



SOLAR NOISE as a Means of Providing Accurate Vertical Polar Diagrams for Radars

By M. J. B. Scanlan*, B.Sc., A.R.C.S.

PART 1 General Considerations and the Measurement of Shape

This article describes how the sun and other stars can be used as signal sources to measure the size and shape of the vertical polar diagrams of radars in situ, by relatively simple means and with good accuracy. This procedure gives, for the first time, a measure of the radar itself, independent of any target and this has important consequences, both for the radar designer and for the operator. The designer can measure the vertical polar diagram objectively and without disturbance from fluctuations in the target echoing area. The operator gains by knowing what the radar diagram really is, so that variations in flight trials can safely be ascribed to the target: moreover, flight trials on one site, which now truly measure the target alone can be transferred quickly and with confidence to another. Furthermore, the operator now has an objective method of checking performance at any time and at small cost.

(Voir page 341 pour le résumé en français: Zusammenfassung in deutscher Sprache auf Seite 342)

ALTHOUGH the work to be described here is mainly on radar aerials at 600 and 1 300MHz, where the need was greatest, it is, of course, applicable to a much wider range of aerials, e.g. those used for tropospheric scatter or satellite communications. It would not be used for X-band aerials, used mainly for marine or airborne purposes, since these are normally small enough to be completely measured on an aerial range; nor perhaps for metric radars, because of the variability of the sun at these frequencies: in fact, it is noted later that the variation at 600MHz was much less than expected so that measurements at a frequency as low as 200MHz may not be out of the question.

Factors Involved in the Determination of Radar Ranges

Before proceeding to describe the solar noise method it is proposed to review briefly the factors which govern the range of a radar against a given target. The radar equation may be written:

$$\text{Maximum range} \propto \left[\frac{P_T G^2 \lambda^2 \sigma}{P_{min}}\right]^{1/4}$$

where σ is the target echoing area

G is the aerial gain

P_{min} is the minimum detectable signal

λ is the wavelength

P_T is the transmitter power

Because of their importance these will be considered in a little more detail.

Target Echoing Area. This is given first place, as being outside the radar designer's control, and as being very difficult to measure. The usual procedure in evaluating a radar is to assume a value for echoing area, and to use flight trials to measure the radar coverage, assumed not to be known otherwise. However, the echoing area is really the sum of reflections from a number of almost independent smaller areas (nose, jet intakes, wings, tail plane and fin etc.): it is not surprising that the composite area is a fluctuating quantity, or that it is impossible to say precisely what the value will be at the moment the radar pulse strikes the aircraft. It is usual to say that the echoing area of a certain aircraft is so many square metres, 'head on' or 'tail on', but these terms mean only that the aircraft track is towards or away from the radar, and do not define at all clearly its precise attitude at any moment. (Even with a precisely defined attitude, one

* The Marconi Company Limited.

would suspect a statistical fluctuation in the total echoing area because of vibration). The best that can ever be said about an echoing area is that, under carefully defined conditions, it has a certain mean value, about which it varies with known probability. To make even this limited statement with any accuracy requires that the echoing area be measured at least ten times, say, at nominally the same range, height and attitude. This condition is usually far from being fulfilled so that many flight trials are in fact attempts to measure one unknown (the radar performance) in terms of another (the echoing area).

Aerial Gain. Surveillance radar aerials are usually designed to have most gain at low angles, the gain falling off, perhaps according to a 'cosecant squared' law, at higher angles. This gain distribution, the vertical polar diagram (v.p.d.), is ideally independent of azimuth, but in practice the aerial is more or less site conscious, and the v.p.d. will vary as the aerial turns. The difficulty of measuring the v.p.d. arises from the fact that it may cover 40° or more in elevation, so that a conventional signal source cannot conveniently be positioned. Even if this can be done, of course, a measurement of shape is obtained, i.e. of relative gain, rather than of size and shape, i.e. of absolute gain. The aerial gain occurs in the radar equation for range as a square root, rather than the fourth root as do most other variables, so that accuracy is more important in the measurement of gain than of other parameters.

Minimum Detectable Signal. This is a complex of many factors. Hall[1] lists seven, one of which he defines as 'the operation and degradation factor S_o,' which is "usually added to the equation to account for the difference between theoretical and observed radar performance". However, five of the remaining factors are relatively objective and can be calculated or measured with acceptable accuracy. The remaining parameter, the noise factor, can, in the light of present knowledge, conveniently be replaced by overall noise temperature since Hall, working with poor receivers by today's standards, does not allow for the differences between aerial temperature and ambient temperature.

Wavelength. This is a precisely-known quantity and therefore presents no great problem.

Transmitter Power. This is under the control of the designer and may be accurately determined, although some care is necessary in translating mean power measurements into meaningful figures.

The Role of Flight Trials and Solar Noise Measurements

Against the background sketched in, it is now possible to put flight trials and solar noise measurements in a true perspective. At present, the radar designer may be asked to guarantee a certain 'probability of paint', say 80 per cent, against a certain target, assumed to be $10m^2$, out to a range of perhaps 100 miles, at all heights up to, say, 40 000ft. Now if, and only if, the designer knows the performance of his aerial on the proposed site and that the echoing area will in fact be $10m^2$, this is a relatively straightforward task. In practice, neither condition holds: the aerial performance is only approximately known, and the echoing area figure is a mean value, about which the true value fluctuates in an almost unknown way.

If on the other hand the aerial was known, by solar noise methods, one could begin to measure the echoing area, and to know how it fluctuates. It would no longer be necessary to over-design the radar; indeed, if enough were known of a variety of targets, and if these proved to have common features, the radar could be designed with these features in mind.

Nothing which has been said should be understood as decrying flight trials, which have an essential part to play, but rather as attempting to define and explain this essential role. Flight trials are to measure aircraft, and not to measure radars: given an adequate number of flight trials on one site to define adequately an echoing area, and provided the size and shape of a v.p.d. can be measured on any other site, then no more flight trials are needed with that aircraft at that radar frequency.

The Measurement of Shape

The gain of a surveillance radar aerial is high enough that a range of 30 or even 40dB may be of interest, corresponding to radar range variations of 30 or 100 times. To measure the variation of gain with vertical angle (the vertical polar diagram or v.p.d.) a signal source of sufficient power is needed.

The gain of a surveillance radar aerial may vary by 30 or even 40dB, over a vertical angle extending up to perhaps 40° of elevation. The corresponding radar ranges will vary from hundreds of miles, limited by the horizon, to a very few miles, when the target is almost overhead. To measure the variation of gain with vertical angle (i.e. to plot the vertical polar diagram or v.p.d.) a signal source is needed which is powerful enough to be detected at the lowest gain and which can be elevated to a sufficient angle. It should preferably be a broadband, or alternatively a very stable narrow band source and its power output must be constant for the duration of the experiment. It must also have a wide, preferably omni-directional radiation pattern, preferably emanating from a point source.

These desiderata are best met by the sun, which normally radiates noise varying only as $(\text{frequency})^2$, so that small tuning errors can be neglected, which has adequate elevation in all temperature latitudes, at least in summer, and which is an omni-directional source, unfortunately of finite size. As for the constancy of its output, there is no reason to believe that a single one of the thousands of points now plotted is in error because of fluctuations during a single sunrise or sunset: moreover, and this is more remarkable, the variation from year to year seems to to be very small, although experience now covers more than half of a sun spot cycle. Of course, if any part of a v.p.d. is suspected to be in error because of variation of the source, a repeat experiment can always be done to verify or remove the suspicion.

The outstanding question is over what range of aerial gains the sun's signal can be measured. At optical frequencies the sun can be thought of as a black body radiator with a temperature of about 6 000°K; if it were a true black body at this temperature, it would be virtually useless for the present purpose. Fortunately, however, as the wavelength increases into the microwave and u.h.f. range, the effective temperature (and size) of the sun increases, the layers of hot gas outside the photosphere becoming increasingly opaque and effective as sources of radiation. The matter is dealt with most conveniently by continuing to deal with the sun as if it were a black body, whose temperature, however, varies with wavelength. If the effective temperature is T_d at wavelength λ, then combining the Rayleigh-Jeans radiation law with the geometry of the earth's orbit around the sun, it is found that the noise flux S on the earth's surface at wavelength λ is given by:

$$S = 0{\cdot}94 \times 10^{-27} (T_d/\lambda^2) \text{ watts metres}^{-2}\,\text{Hz}^{-1} \quad \ldots . \ (1)$$

for either polarization.

T_d is generally reckoned to vary (despite what has been said above about its constancy) by a factor of four or so between sunspot minimum (1954) and maximum (1958). Its minimum value, however, can be taken to be sufficiently accurate for this purpose:

$$T_d \simeq 4 \times 10^5 \lambda \text{ degrees } (\lambda \text{ in metres}) \quad \ldots\ldots \quad (2)$$

If the aerial has a gain G in any direction, its effective area in that direction is A, given by:

$$A = G\lambda^2/4\pi$$

and the power received from the sun per unit bandwidth is SA, given by:

$$SA = \frac{0{\cdot}94 \times 10^{-27}}{4\pi} GT_d \text{ watts}$$

$$= \frac{0{\cdot}94 \times 10^{-22}}{\pi} G\lambda \text{ watts}$$

This formula is correct for the power collected by the aerial surface: all this will be delivered to the feed, provided that the beamwidth is wider than the sun (0·5°).

This power received from the sun must be detected and measured in the presence of set noise, which combines two factors, aerial noise and receiver noise. Aerial noise depends on frequency, aerial design, angle of elevation and on losses between the aerial itself and the receiver terminals: it is specified by a temperature, T_A, and will rarely be less than 100° for a horizontal looking radar aerial. The receiver noise is described by the receiver noise factor F or by the noise temperature T_R, which are related by:

$$T_R = 290\,(F - 1)$$

where the factor 290 arises because 290°K is the reference temperature for noise temperature measurements.

The constant which converts noise temperature to power is Boltzmann's constant k, so that the overall set noise power is P_N, given by:

$$P_N = k(T_R + T_A) \text{ watts Hz}^{-1}$$

$$= k(290\,(F - 1) + T_A) \text{ watts Hz}^{-1}$$

Now T_R will vary between about 1 200° (conventional mixer or t.w.t.) and 200° (parametric amplifier): taking $T_A = 100°$, a typical value, and $k = 1{\cdot}38 \times 10^{-23}$.W/°K^{-1}, it is found that R_N will almost always lie in the range:

$$4{\cdot}1 \times 10^{-21} < P_N < 1{\cdot}8 \times 10^{-20} \text{ W . Hz}^{-1}$$

Taking the lower figure as representing maximum available sensitivity, and assuming that a 1 per cent change in set noise can be detected, the sun on an aerial of gain G at wavelength λ can be detected provided that:

$$SA > 0{\cdot}01\, P_N$$

i.e. provided that:

$$\frac{0{\cdot}94 \times 10^{-22}}{\pi} G\lambda > 4{\cdot}1 \times 10^{-23}\,\lambda \text{ in metres}$$

or

$$G\lambda > 1{\cdot}3$$

Hence one has a simple table, giving the gain G_{min} required at various frequencies, to receive detectable signals from the sun; this will be valid for the quiet sun (T_d given by equation (2) using an optimum receiver, or for the sun near sun spot maximum, using a conventional receiver.

TABLE 1

Frequency (MHz)	600	1 300	3 000	5 000	10 000
G_{min}(dB) ...	4	7	11	13	16

Table 1 also allows the estimation of the range of gain which can be measured: thus the v.p.d. of a 37dB L-band aerial can be measured over a 30dB range, and −30dB side lobes will be just detectable. Experience shows, at least at the lower radar frequencies, that these figures are realistic, provided that precautions are taken to stabilize receiver gain. Thus, as long ago as 1958 (sun spot maximum), when the display system in use was comparatively insensitive (see Fig. 4), measurements at L-band, using a crystal mixer ($NF = 6{\cdot}5$dB) went down to a gain of about 10dB, while more recent experiments at 600MHz, using a parametric amplifier ($NF = 2{\cdot}5$dB) gave easily measurable signals from the radio stars, Cassiopeia A and Cygnus A and X on a 32dB aerial, the stars being about 30dB down on the sun. In both cases, the receivers used valves, and it was necessary to stabilize heater voltages by using batteries or voltage stabilizers. Further details of the star experiments will be given later, in Part 2 of this article.

There is a further point about radar receivers which deserves notice. The sun is a noise source and signals are received almost equally on both signal and image channels of a superhet receiver, provided that the r.f. stages, if any, are broadband, and cover both channels. Given a crystal mixer and a travelling wave tube (t.w.t.) with the same noise factor, the mixer will in fact be twice as sensitive to sun signals as the t.w.t. with or without image rejection filter, provided that the aerial does not squint, i.e. is non-dispersive.

This occurs because the quoted noise factor of a crystal mixer is, by convention, the radar or single channel noise factor, while the t.w.t. is quoted double channel, an image rejection filter being assumed when the t.w.t. is used in a radar. If the aerial does squint, the mixer, or t.w.t. without image rejection filter, will receive two responses for each pass of the aerial through the sun, and the mixer will be as good as a t.w.t. with filter, and twice as good as one without the filter. These difficulties do not arise with narrow band r.f. stages, such as parametric or triode amplifiers.

Experimental Methods

Being now assured that a given aerial will receive detectable solar noise, this section will discuss how the noise may be measured and used to evaluate a v.p.d. Two questions arise, that of the position of the sun (both vertically, which is vitally important for this purpose, and in azimuth, which is useful rather than vital) and that of the measurement of the received signal. It is necessary to know the received signal in relative rather than absolute terms, i.e. the ratio between signals received at various angles of elevation rather than the absolute power at any time: this eases the problem.

The position of the sun in elevation is given by a simple formula:

$$\sin h = \sin \lambda \sin \delta + \cos \lambda \cos \delta \cos \phi$$

where h is the angle of elevation, λ the latitude of the station, δ the declination of the sun and ϕ the hour angle, which is 180° when the sun is due south at the station, and varies by 15° an hour. The sun is due south at Greenwich near noon G.M.T., the difference between noon G.M.T. and the transit of the sun being given by the equation of time: it is due south at any other station four minutes earlier or later for every degree of longitude east or west of Greenwich. The equation of time and the declination of the sun at the time of the experiment must be found from an almanac (Nautical Almanac or Whitakers, for instance).

A typical calculation of the sun's elevation is given

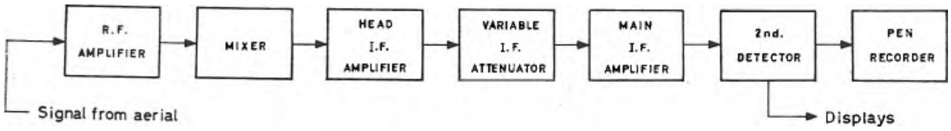


Fig. 1. Arrangement of typical radar receiver, adapted for solar noise measurements by addition of variable i.f. attenuator and pen recorder.

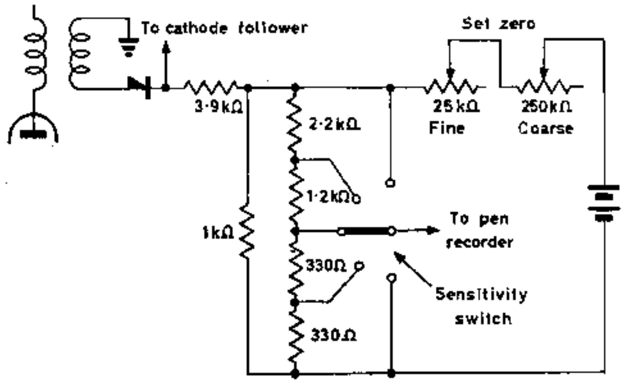


Fig. 2. Typical 2nd detector, backing off and sensitivity control circuits

in Appendix 1, in which the last two rows of the calculation give the elevation of the sun at half hourly intervals for an imaginary (as far as known!) radar station. These nine points, and as many more as may be required, form an almost linear curve of elevation against time. This curve has been corrected for refraction which is important at the small angles which are most significant for surveillance radars.

An alternative method of finding the sun's elevation can also be used[2]: this effectively does the calculations graphically, so that the elevation curve is a straight line on a specially prepared graticule.

The azimuth of the sun is not required to any accuracy, since it can almost be guessed with sufficient accuracy for the radar to find the signal. However, it is sometimes useful to know where to look with some accuracy, especially when weak signals are the first to be received in a sunset run. The azimuthal angle, A, is given by:

$$\sin A = \sin \phi \cos \delta \sec h$$

which is easily calculated once h is known.

Turning now to the detection and relative measurement of the strength of the sun signal, it is useful to calculate, using Table 1, roughly how much signal is to be expected. Thus from the table and under the conditions there mentioned, aerial gains are required of 20dB more than those listed to give a sun signal equal to the set noise. With an L-band aerial with a gain of 36dB say, the peak signal received from the sun will be (36 − 20 − 7), or 9dB up on the standing noise, and there will be a measurable range of 29dB, over most of which, however, the noise signal from the sun will be very much less than the standing set noise.

For this reason, the standing noise must be eliminated, either by using a.c. couping or, better, by using d.c. coupling, and backing off the standing current. It has been found best to use the second detector current as a measure of receiver output, so that standing current in, for example, the video cathode-follower, is eliminated. The second detector current is backed off against a variable current from a small battery, and the resultant output, normally zero, is applied to a pen recorder or a scope. As the aerial sweeps through the sun, the extra signal received will be recorded in the form of a 'horizontal' polar diagram at the appropriate angle of elevation on the scope or pen recorder, assuming, of course, that the aerial rotation is not too fast for the recording medium. If the time of the aerial sweep through the sun is recorded, one has all the basic data needed for the measurements.

There remains, however, to compare the strength of sun signals received at various times: in other words, to calibrate the system. The most convenient way of doing this is to use i.f. calibration which essentially inverts a very common method of noise factor measurement. In this method a noise source of known output is used to measure the noise factor of a receiver. If N is the noise ratio of the noise source, and α is the i.f. attenuation ratio required to restore the output to a datum when the noise source is switched on, then:

$$\text{Noise factor } F = \frac{\text{Signal/noise in}}{\text{Signal/noise out}} = \frac{N-1}{\alpha-1}$$

so that the noise factor is known from a measurement of α, N being given.

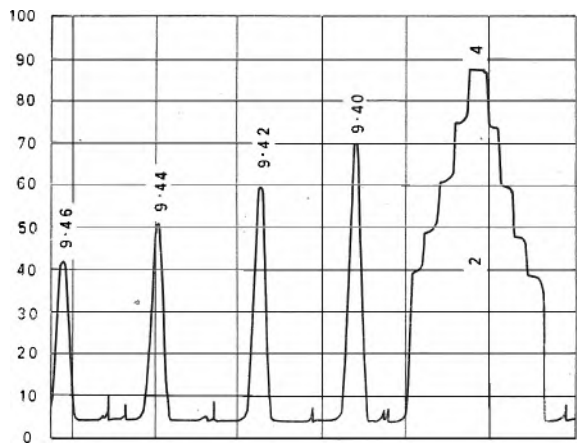


Fig. 3. Typical pen recorder trace, showing (R to L) calibration markers (one 2dB step, plus four ½dB steps) and four solar noise traces at two minute intervals

Fig. 4. An alternative method of display and measurement, using fixed coil p.p.i. display, showing two responses per aerial transit, plus calibration markers (1dB steps)

(a) Sun's elevation 12°

(b) Sun's elevation 2·6°

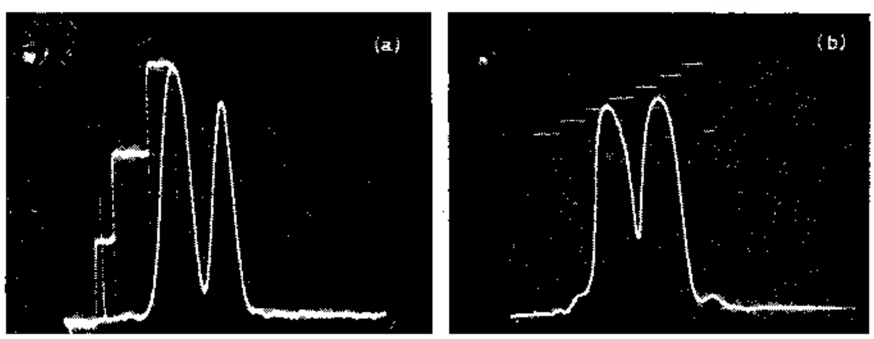


Fig. 5. Results of measurements on an L-band array, showing the v.p.d. between 0° and 36°

In the present case, the noise factor is known, or at least is assumed constant and a measure of $(\alpha - 1)$ is used as a measure of N: hence if $(\alpha - 1)$ is plotted in decibels against angle of elevation, the sun being a constant source, the desired v.p.d. is obtained. With the sun $(\alpha - 1)$ is proportional to N, rather than $N - 1$, since the sun only occupies a small part of the vertical beam, the rest receiving sky noise just as it does when the aerial is not pointing at the sun: in the noise factor measurement, on the other hand, the noise source replaces the reference source, and the signal-to-noise input is $N - 1:1$. Looking at the matter another way, if the sun's noise plus the set noise is equal to the set noise increased by the ratio α, the sun's noise is proportional to $(\alpha - 1)$.

Hence the system is calibrated by using a variable i.f. attenuator at a convenient place in the i.f. amplifier chain, preferably as early as possible to minimize non-linearities ahead of it. I.F. attenuators with $\frac{1}{2}$dB steps are readily available and quite suitable. At convenient times, if necessary after every pass of the aerial through the sun, this attenuator is adjusted to give an increased set noise nearly equal to the set noise plus sun signal. The value of α corresponding to any given sun signal can then be read off, and $(\alpha - 1)$ taken as a measure of that signal. α will, of course, be measured in decibels: it must be converted to a ratio, the one subtracted, and the result reconverted to decibels above or below datum, which is conveniently $\alpha = 3$dB, i.e. set noise = sun noise. If the receiver is sufficiently stable that only one calibration curve of i.f. attenuation against deflexion is necessary, one can conveniently tabulate $(\alpha - 1)$ in decibels, against the corresponding deflexion as follows.

Deflexion* (mm)	6	12	18	24	30	36	42	48	54
α dB	0·5	1·0	1·5	2·0	3·0	4·0	5·0	6·0	7·0
α ratio	1·12	1·26	1·41	1·58	2·0	2·5	3·16	4·0	5·0
$\alpha - 1$	0·12	0·26	0·41	0·58	1·0	1·5	2·16	3·0	4·0
$(\alpha - 1)$dB	−4·2	−5·9	−3·9	−2·3	0	1·8	3·3	4·8	6

* *Examples, not experimental figures*

From a graph plotting the calibration deflexions against $(\alpha - 1)$ in decibels the relative signal strength, which is proportional to aerial gain, can be read off for any sun signal.

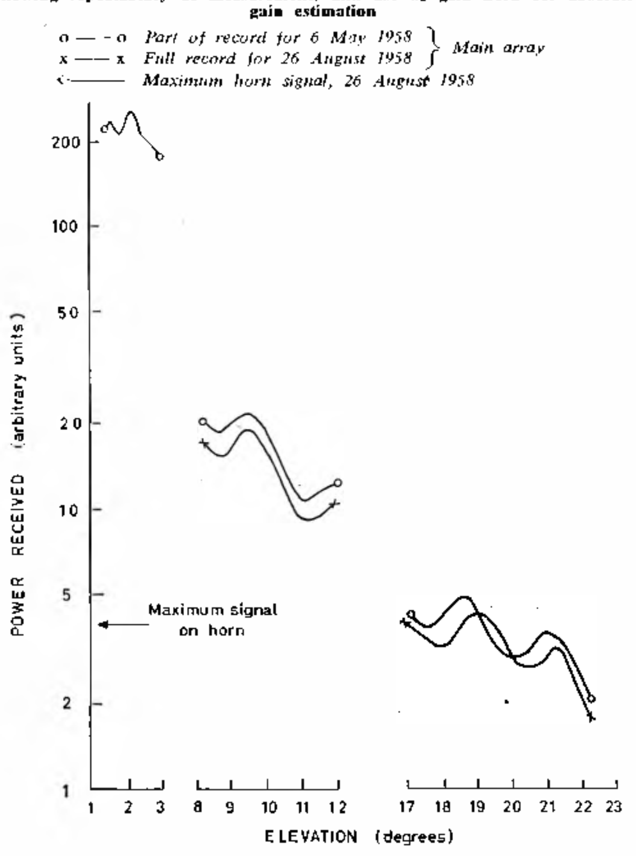
To illustrate the experimental arrangement and the calibration process further, reference should be made to the figures as follows.

Fig. 1 is a block diagram of a typical radar receiver, and shows the variable i.f. attenuator.

Fig. 2 shows a typical second detector circuit, and the backing off and range switching circuit. Range switching is necessary to cope with the largest signals, while enabling the smallest signals to be recorded at maximum sensitivity. Calibration is, of course, necessary for every range used.

Fig. 3 shows a portion of a typical pen recorder trace: on the right is a calibration, in which the receiver gain has been increased in one 2dB, followed by four 0·5dB steps, giving a total of 4dB. This is followed by four sun signal traces at two minute intervals, the times being marked opposite each signal. It will be seen that for the first of these signals $\alpha = 3·3$dB, $((\alpha - 1) = 1·14)$ while for the third $\alpha = 2·6$dB $((\alpha - 1) = 0·82)$: hence the aerial gain has changed by 114/82, or 1·43dB in 4min, which very roughly corresponds to 0·5° in elevation; a point could have been taken every 15sec if required; giving a relative gain measurement for every 0·035° or so in elevation.

Fig. 6. Results on main array on two dates, and on standard gain horn showing repeatability of measurement, and use of gain horn for absolute gain estimation



It may also be noted that the base line of the system shows no noticeable change during the record, although a change of 1mm ($\simeq$0·05dB) would be easily seen. It may be inferred that the receiver (a parametric amplifier, mixer i.f. amplifier system) is stable to about 0·01dB (0·25 per cent) over at least this period. Gain stability of this order (which is not strictly necessary for the sun) requires very careful attention to the power supplies, all of which (h.t.'s and heaters) are stabilized. Small spikes on the base line between the sun signals were caused by surges as the aerial turning gear was switched on and off.

Finally it will be noticed that the recorder response to the calibration steps is steeper than to the sun signals: it follows that the recorder is following the sun signals adequately.

Variation in detail on the above method is of course possible. Thus, in early work, a fixed coil display console was used as recorder: a slow time-base sweep was triggered just before the aerial reached the sun, and each sweep photographed, often with a calibration trace superimposed. Some results obtained in this way are shown in Fig. 4: here there are two responses, resulting from the use of a mixer as the first receiver stage together with a squinting aerial. In yet another variation, the aerial was turned at speed, the time-base and camera being triggered from azimuth marking contacts on the turning gear. R.F. calibration, in which the sun signal is compared with known noise injected via a calibrated attenuator, would be more satisfactory than i.f. calibration in that there are then no anxieties about receiver linearity: however, it complicates the equipment considerably and has rarely been used. On the other hand, it is a useful check on receiver noise to inject a known noise signal periodically and record the resultant deflexion, since i.f. calibration can correct for slow gain changes, but not for changes of overall receiver noise.

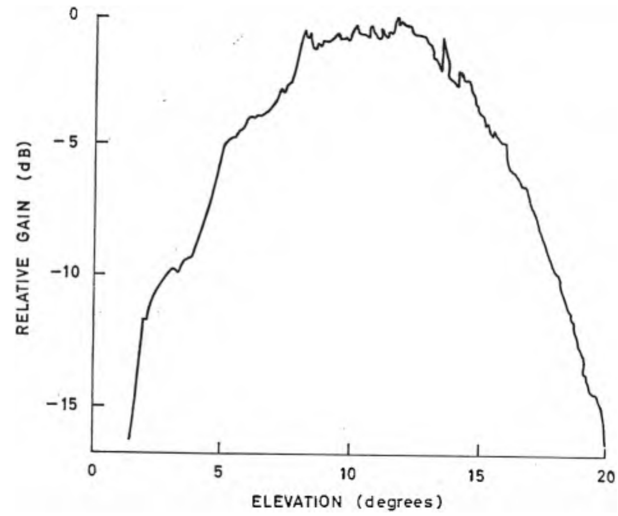


Fig. 7. Free space v.p.d. of SA.120 Aerial at 600MHz

Some Aerial Measurements Using Solar Noise

The first quantitative accurate and wide range results using solar noise were obtained in 1958 on a large L-band array in which considerable discrepancies had arisen between v.p.d. measurements on a section of the array turned on end and flight trial results. Using a crystal mixer as receiver, it proved possible to measure the v.p.d. of the array over a range of well over 30dB and to confirm that the v.p.d. measurements on a section were correct: the flight trial results gave a cover diagram, rather than a v.p.d., the difference being presumably due to the scattering pattern of the target aircraft. Some results of these first measurements are shown in Fig. 5.

Later in that same year, a horn aerial of known gain was rigged up on the same array, and sun measurements were done first on the main array, then on the horn and again on the main array, all in the one afternoon, and using the same feeder run for both aerials. By comparing the main array results with the full runs taken earlier and with the horn results, the gain of the main array was estimated. These comparisons are shown in Fig. 6, which shows that results taken months apart are highly reproducible.

Attention was then turned to radar aerials at 600MHz, which are of very great operational importance. The standard aerial, the SA120, is $52\frac{1}{2}$ft × $12\frac{1}{2}$ft overall, and is fed from a slotted waveguide linear array.

The aerial is mounted at a mean height of about 12ft,

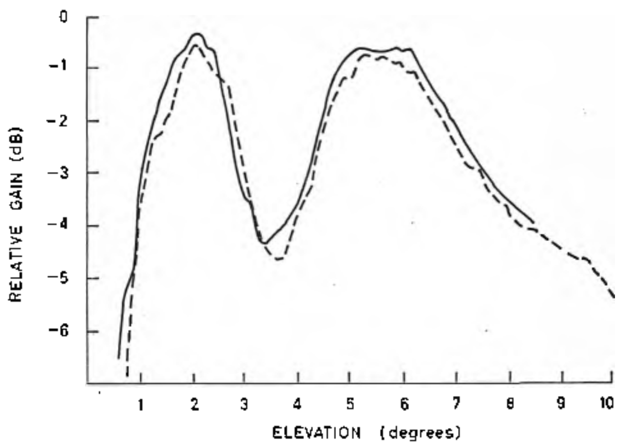


Fig. 8. V.P.D. of SA.120 Aerial at 600MHz at $4\frac{1}{4}$° tilt

Dotted line—no crop. Full line—green crop. 2ft high around aerial

and its v.p.d., which is essentially a ground reflection lobe at low angles with the free space pattern taking over above 6° or 8°, is controlled by adjustment of tilt. A tilt scale is provided, but cannot allow for the contours of the ground around the aerial. Accurate setting of the tilt angle with respect to the contours of the ground is essential if the site is to be used to best advantage, and solar noise experiments have proved ideal for checking and adjusting tilts. Very many solar noise measurements have been done on this aerial: from these are selected Fig. 7 and Fig. 8 which show respectively the free space diagram (tilt +10°), and the operational diagram (tilt $+4\frac{1}{4}$°), taken with and without a crop growing around the aerial. It will be noticed that the crop makes no significant difference to the aerial performance.

It will be noticed that on some of these diagrams, the peak of the curve shows small fluctuations (of the order 0·5dB): these are now attributed to atmosphere effects (twinkling or scintillation), which are much more marked at low angles and with sources smaller than the sun, as will be seen later, in Part 2 of this article. They are not unimportant in radar operation, since they would cause fluctuations in the radar return even from a smooth target (e.g. a sphere) at very long range: they are largely masked in normal operation by variations in the target echoing area, which are normally much larger than the ±1dB which might reasonably be ascribed to twinkling.

The examples given will perhaps suffice to show that sun measurements can give simple and repeatable measurements on the v.p.d. of a radar aerial: such results can be invaluable in adjusting tilt or focus or in commissioning a radar on a new site.

Appendix

Specimen Calculation of the Sun's Elevation

Position of radar 51°0′28″ N
1°37′33″ W

Date: 17 July, 1962.

Sun transits at Greenwich at 12h 05min 58sec.

At radar, sun transits (i.e. passes through south) 6½min later, i.e. at 12h 12min 28sec G.M.T. Declination of sun at 2100 hours (near sunset) is 21° 11″

Now $\sin h = \sin\lambda \sin\delta + \cos\lambda \cos\delta \cos\phi$

$$\log\sin\lambda = \bar{1}{\cdot}89055$$
$$\log\sin\delta = \bar{1}{\cdot}55793$$
$$\underline{\bar{1}{\cdot}44848} \quad \text{i.e. } \sin\lambda\sin\delta = 0{\cdot}28085$$

$$\log\cos\lambda = \bar{1}{\cdot}79879$$
$$\bar{1}{\cdot}96952$$
$$\bar{1}{\cdot}76831$$

Note that ϕ is chosen here at convenient intervals of 7½°. corresponding to 30min of time. The time at which ϕ is 60° is 4 hours after the sun passes through south at the station, and this gives 16h 12min 28sec in the last row, first column.

$\phi =$	60	67½	75	82½	90	97½	105	112½	120
log cos ϕ	$\bar{1}{\cdot}69897$	$\bar{1}{\cdot}58284$	$\bar{1}{\cdot}41300$	$\bar{1}{\cdot}11570$		$= -82\frac{1}{2}$	$= -75°$	$-67\frac{1}{2}°$	$-60°$
log cos λ cos δ	$\bar{1}{\cdot}76831$	$\bar{1}{\cdot}76831$	$\bar{1}{\cdot}76831$	$\bar{1}{\cdot}76831$					
log cos λ cos δ cos ϕ	$\bar{1}{\cdot}46718$	$\bar{1}{\cdot}35113$	$\bar{1}{\cdot}18131$	$\bar{2}{\cdot}88401$					
antilog	·29321	·22447	·15182	·07656	0	−·07656	−·15182	−·22447	−·29321
sin λ sin δ	·28085	·28085	·28085	·28085	·28085	·28085	·28085	·28085	·28085
sin h	·57406	·50532	·43267	·35741	·28085	·20429	·12903	·05538	−·01236
h (true)	35° 2′	30° 21′	25° 38′	20° 57′	16° 18′	11° 47′	7° 25′	3° 11′	−0° 42′
Correction	—	—	—	—	—	5′	10′	18′	52′
Apparent ht.	35° 02′	30° 21′	25° 38′	20° 57′	16° 18′	11° 50′	7° 35′	3° 29′	0° 10′
Time G.M.T.	16.12.28	16.42.28	17.12.28	17.42.48	18.12.28	18.42.28	19.12.28	19.42.28	20.12.28

References

1. Hall, W. M. Prediction of Pulse Radar Performance. *Proc. Inst. Radio Engrs.* 44, 224 (1956)
2. Shinn, D. H. A Simple Graphical Method for Finding the Elevation of the Sun to ±0·1°. Marconi Report RD/G.1602.

(To be continued)

Automation for the Tyne Tunnel

Greater safety and comfort for motorists using the new Tyne Tunnel between Howdon and Jarrow when it is opened to traffic later this year, will be assured by a telemetry monitoring and control system designed by English Electric's Industrial Control and Automation group at Kidsgrove. The telemetry and telecontrol system will aid the controller to take swift action to avoid traffic jams, to check air pollution from carbon monoxide fumes, and to maintain tunnel lighting and other services in case of emergency. The Tyne Tunnel will be the largest and longest in Britain to be equipped with an automation system to monitor and control all essential services.

English Electric is also to supply electrical distribution equipment to provide duplicate feeds from the North and South sub-stations, and two standby diesel generator sets to guarantee essential services in the event of a mains supply failure.

The telemetry system will remotely control ventilation fan speeds, pumps, the start-up of the diesel sets, and the switch-in of emergency supply circuits. Over 6000 state signals or 1000 analogue signals will be carried along a single cable to a 15ft long mimic control and instrumentation board in the Administration Building. Information will come in from the North or South ventilation stations and supply sub-stations, and from the mid-river and south portal pump houses to this central point where the outlying equipment can be controlled automatically.

The telemetry link represents a cost saving over direct wire systems, only one 10-pairs of a telephone cable being used to connect all the out-stations to the control centre. A very rapid feed back of information to the control centre is achieved by using digital techniques with a unique parallel address system and time division or time sharing principles. This allows numbers of signals transmitted in sequence to be displayed to the operator simultaneously.

English Electric's Switched Direct Wire Multiplexing Scheme designed by Telemetering Applications Department at Kidsgrove, transmits the signals for supervisory control over the ventilation fans for the tunnel, remote control of all switchgear and pumps together with constant monitoring of fan speeds, switchgear metering and alarm scanning.

The control centre in the Administration Building and the 15ft long mimic control board provides remote control of equipment in the North Ventilation Station, South Ventilation Station, Mid River Pump Station and South Portal Pump Station.

Both the North and South Stations have sub-stations housing h.v. switchgear, transformers and m.v. switchgear for 16 outgoing feeders, also fusegear boards controlled by contactors.

Back indication is provided of amperes, volts, revs per minute, power factor and kilowatt hours on 'Elite' instruments from the 43 circuits. There are 55 miniature illuminated discrepancy switches to close and trip every solenoid operated h.v. and m.v. circuit-breaker and fusegear feeder contactor in the system.

Semaphore indicators show the condition of sub-circuits and auto-manual switches allow the operator over-riding control over the automatic sequences.

Again manual over-riding control is available as proportional reference adjustment or selected level control.

The pump stations can be controlled from the mimic diagram which is a representation of the complete electrical system and has illuminated push-buttons.

The cable connecting all outstations to the central control is 1/·044 polythene insulated p.v.c. sheathed single wire armoured and p.v.c. sheathed overall of which 10-pairs are capable of carrying over 6000 'on/off' state signals or over 1000 analogue signals of 'flow', 'current', etc., by means of the time division principle.

The code generator which is situated in the Administration Building Control Room sends out discreet binary codes in sequence over 8 cores of the cable. The Outstation coded relay detectors operate when their particular code is on the line. While this code is applied, a direct connexion exists between the measurement transducer and the indicating instrument and a remote indication of the primary quantity is obtained over every pair of information cores, i.e., four analogue readings simultaneously in 2 seconds. On generation of the next code, the first coded relay will drop out and the coded relay corresponding to this code will connect the next set of readings to the line and so on.

The readings of 'current', 'voltage', 'revs per minute', and 'power factor' are transmitted in this way to 'Elite' moving-coil instruments built into the mimic diagram and scaled accordingly. The 'Elite' instruments have a pointer lock movement to hold the pointed position after every reading while the scan dwells at other points in the system.

'On/Off' indications such as 'switch closed', 'switch open', or 'motor running' are transmitted at a rate of 6 per pair of information cores i.e., 24 indications simultaneously in ½ second.

All miniature push-buttons and semaphore indicators are illuminated by this method of state indications, the lamps being red or green determined by the position of bistable relays.

A Second-order Active Filter Circuit for Tuned Amplifiers and Sinusoidal Oscillators

By E. A. Faulkner* and Viscount Downe*

This article describes a simple three-transistor active filter circuit which simulates the transfer function of an LC tuned circuit without the need for precision components; a Q stability of better than 0·1 per cent/°K, and a resonance-frequency stability of 0·02 per cent/°K are readily obtained. The same frequency stability is obtained when the filter in conjunction with a simple limiter is used as a sinusoidal oscillator.

(Voir page 341 pour le résumé en français: Zusammenfassung in deutscher Sprache auf Seite 342)

IN a recent article[1] an active filter circuit is described which employs the principle of selective positive feedback. The unsatisfactory nature of such circuits for most applications is implicit in the formula given by the authors for the Q factor:

$$Q = 1/(3-R/R_F)$$

where R and R_F are two of the resistors in the circuit. In order to obtain the Q factor of 5·0 used by the authors a value of $R/R_F = 2{\cdot}8$ is necessary; a change of this ratio by ± 1 per cent will cause a change in Q of about ± 15 per cent, and an increase of 10 per cent will cause the system to oscillate.

This excessive sensitivity to component variations is common to all active-filter circuits which employ the technique of subtracting relatively large circuit parameters to obtain small effective values, and the same fundamental objection applies to n.i.c. circuits[2,3] and their equivalents[4].

It has recently been shown[2,3] that these difficulties can be overcome by the use of gyrator circuits provided that they are correctly designed[5,6]. However, the complication of a gyrator circuit may not be justified in the frequently-occurring cases where a simple tuned circuit is required for instrumentation purposes, with a medium Q (say about 20) and a reasonable tolerance to component and temperature changes.

The circuit to be described is based on the analogue-computer simulation technique used by Sutcliffe[7]. However, the present approach to the circuit design is different from that of Sutcliffe, who used a high gain (60dB) in each of the three operational amplifiers so that the Q could be accurately determined by passive components in the system. A single silicon planar transistor is used as each operational amplifier, and the Q is allowed to be determined by the finite gain of the amplifiers; a careful choice of operating conditions ensures a low temperature coefficient and a low sensitivity to variations between individual transistors. The resulting circuit is less versatile than that of Sutcliffe, but is far simpler and has many applications in instrumentation work.

The Three-Transistor LCR Analogue Circuit

Principle of Operation

The general principle of the *LCR* analogue circuit is shown in conventional operational-amplifier form in Fig. 1.

A straightforward virtual-earth analysis of this as a high-Q system (i.e., ignoring a term in $1/Q^2$) gives the following results for the resonance angular frequency ω_o, the Q factor Q_o, and the voltage gain at resonance:

$$\omega_o^2 = R_4/C_1C_2R_2R_3R_5 \quad \ldots \ldots \ldots \ldots (1)$$

$$\left.\begin{array}{l} 1/Q_o = \tan\delta_1 + \tan\delta_2 \\ \text{where } \tan\delta_1 = 1/\omega_o C_1 r_1,\ \tan\delta_2 = 1/\omega_o C_2 r_2 \end{array}\right\} \quad \ldots (2)$$

$$\text{at resonance: } v_2/v_S = jQ_oR_5/R_1 \quad \ldots \ldots \ldots \ldots (3)$$

Fig. 2 shows how the system of Fig. 1 can be realized with the use of three direct-coupled silicon planar transistors, the only major change being that the inverting amplifier VT_3 employs series feedback via the emitter resistor combination R_{EA}, R_{EB} rather than the parallel feedback shown

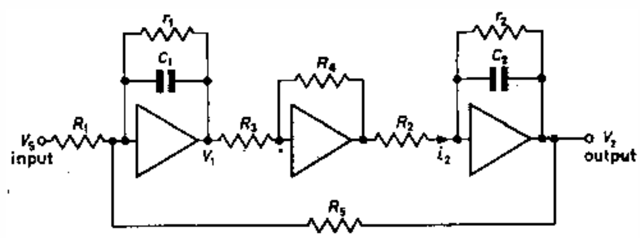


Fig. 1. LCR analogue circuit

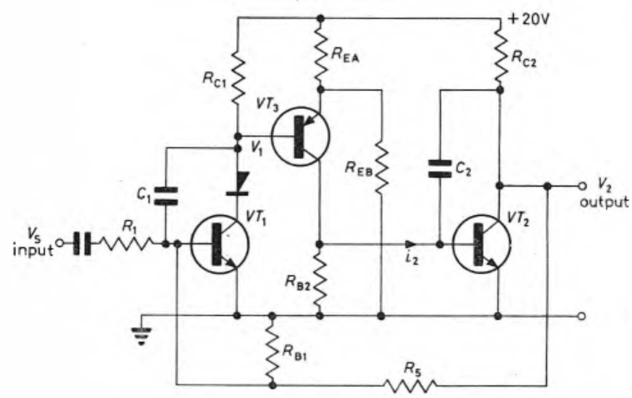


Fig. 2. Practical circuit

in Fig. 1. The function of the silicon diode in the collector lead of VT_1 is merely to prevent the circuit from going at switch-on into an unwanted operating condition with VT_1 and VT_3 saturated and VT_2 cut off.

In the usual high-feedback approximation the a.c. current i_2 flowing into the summing junction of the second integrator is related to the output voltage v_1 of the first integrator by the relation:

$$i_2 = -v_1/R_E \text{ where } R_E = R_{EA}R_{EB}/(R_{EA} + R_{EB}) \quad \ldots (4)$$

Comparing this with the corresponding relation from Fig. 1, which is:

$$i_2 = -v_1R_4/R_2R_3 \quad \ldots \ldots \ldots \ldots (5)$$

it is seen that the resonance frequency is obtained from equation (1) by replacing the quantity R_2R_3/R_4 by R_E;

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thus for the circuit of Fig. 2 one has, approximately:

$$\omega_o^2 = 1/C_1C_2R_1R_E \quad \text{.............. (6)}$$

Equation (3) for the voltage gain at resonance applies equally to Fig. 2 or to Fig. 1. On the other hand, when low-loss (e.g. silver-mica) integrating capacitors are used in the circuit of Fig. 2 the Q factor of the system is determined not primarily by the capacitor losses but by the finite gain of the operational amplifiers VT_1 and VT_2.

Calculation of Q Factor

In order to calculate the Q factor of the circuit in the low-frequency limit (that is, in the frequency range where transistor capacitances are negligible) one proceeds by calculating the phase-shift given by a single-transistor integrator in the general case. In the equivalent circuit of Fig. 3 the transistor is represented simply as an input resistance h_{ie} and a current generator $y_{fe}v_{BE}$; the signal source is represented as a voltage generator v_S in series with a resistance R_S, the collector load is a resistance R_C, and an additional resistance R_B is shown connected between the base and emitter terminals. The Thévenin equivalent of Fig. 3 is shown in Fig. 4, where the parallel combination of R_S, h_{ie}, and R_B is replaced by a single resistance R'. A straightforward analysis of this circuit gives the result:

$$v_C/v_S = R'R_C(j\omega C - y_{fe})/R_S\,[1 + j\omega C(R' + R_C + R'R_Cy_{fe})] \quad \text{............ (7)}$$

The conditions:

$$y_{fe}/\omega_oC \gg 1,\ y_{fe}R' \gg 1,\ y_{fe}R_C \gg 1 \quad \text{.......... (8)}$$

may now be introduced which will be true for any reasonable choice of circuit values. These conditions enable, assuming $\omega \simeq \omega_o$, equation (7) to be written to a good approximation in the form:

$$v_C/v_S = (j + \omega C/y_{fe} - 1/\omega CR_CR'y_{fe})/\omega CR_S \quad \text{.... (9)}$$

The second and third terms in the bracket represent the departure of the behaviour of the circuit from that of an ideal integrator, representing a loss angle δ where:

$$\tan\delta = (1/\omega CR_CR' - \omega C)/y_{fe} \quad \text{........ (10)}$$

Taking equation (10) in conjunction with equation (2), the following result is derived for the Q factor of the circuit of Fig. 2:

$$\left.\begin{aligned} 1/Q_o = {} & (1/\omega_oC_1R_{C1}R_1' - \omega_oC_1)/y_{fe1} \\ & + (1/\omega_oC_2R_{C2}R_2' - \omega_oC_2)/y_{fe2} \end{aligned}\right\} \quad \text{.... (11)}$$

where R_1' represents the parallel combination of R_1, R_{B1}, and h_{ie1}; R_2' the parallel combination of R_{B2} and h_{ie2}.

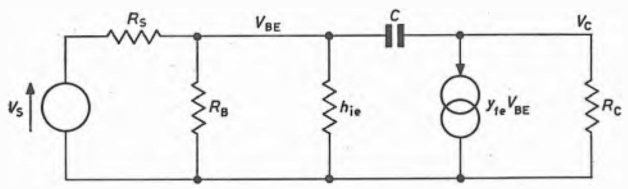

Fig. 3. Equivalent circuit of single-transistor integrator

It is clear that this formula may give a positive or a negative value for Q_o; a negative result implies that the circuit is unstable. It is part of the normal design procedure to ensure that Q_o has the required positive value, and furthermore that this value is obtained by making the magnitude of the sum of the positive terms in equation (11) much greater than that of the negative terms (at least a factor of 4) so that the operation is not critically dependent on the circuit values.

The maximum value of Q usable in practice may be estimated by assuming in equation (10) that the condition has been established that $1/\omega_oCR_CR' = 4\omega_oC$. The d.c. collector current will be taken to be 1mA and the a.c. gain to be 100 so that $y_{fe} = 38\text{m}\Omega^{-1}$ and $h_{ie} = 2{\cdot}6\text{k}\Omega$. It will further be assumed that, for the sake of temperature stability, the external base-emitter resistance R_B has been chosen to have a value of $\frac{1}{2}h_{ie}$ and that this is much less than the source resistance R_S so that the total resistance R' of the base circuit is approximately 900Ω. Finally, the d.c. supply voltage is assumed to be in the region of 20V so that one may take $R_C = 15\text{k}\Omega$. Substituting these

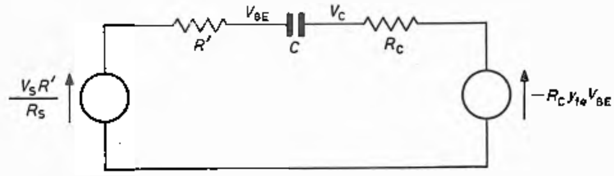

Fig. 4. Thévenin equivalent of Fig. 3

values in equation (10) it is found that $\tan\delta \simeq 0{\cdot}009$ so that, according to equation (2), the Q factor of a system incorporating two of these single-transistor integrators would be greater than 50. It should be noticed that this result does not depend on the assumption of a value of 1mA for the d.c. collector current, because a change in this value would ideally be accompanied by a proportionate scaling up of y_{fe} and scaling down of h_{ie} and R_C. It is not in fact possible to establish the optimum conditions in the circuit of Fig. 2 if VT_1 and VT_2 are run at the same d.c. collector current.

For resonance frequencies below about 100Hz the use of silver-mica capacitors may be unduly expensive; if lossy capacitors are used, additional terms must be added to equation (11) representing the loss angles of the capacitors themselves.

Frequency Limitations

When suitable circuit values have been established for a given application, the value of the resonance frequency ω_o can be changed by scaling the integrating capacitors C_1 and C_2; if the ratio C_1/C_2 is kept constant, it appears from equations (6) and (11) that the Q factor will remain unchanged. However, in deriving equation (11) no account has been taken of the transistor capacitances, and the useful frequency range of the circuit is limited at the upper end by the collector-base capacitance C_{CB} of the inverting transistor VT_3, which introduces a phase shift in such a direction as to increase the Q factor as the resonance frequency ω_o increases. Thus when ω_o has such a value that the impedance of C_{CB} is equal in magnitude to $2Q_oR_E$ (R_E is defined in equation (4)), the Q factor has double the low-frequency value Q_o given by equation (11); and when $\omega_oC_{CB} = 1/Q_oR_E$ the system becomes unstable. The effect of C_{CB} can be neutralized simply by connecting a capacitor equal to C_{CB} across R_{EA} or R_{EB}, and with this modification the circuit of Fig. 2 has been used in this laboratory at frequencies up to 470kHz. However, neutralization is not necessary in the frequency range in which this circuit has usually been used. It is interesting to notice that currently available pnp silicon planar transistors tend to have higher collector-base capacitance (about 8pF for the 2N3702) than the corresponding npn devices (about 2pF for the 2N3710) and it may be advisable to take this fact into account when deciding on the polarity of the circuit.

Temperature Coefficients

Equation (6) for the resonance frequency ω_o is not exactly correct, and to be more accurate one should rewrite it in the form:

$$\omega_0^2 = 1/C_1C_2R_5\,(R_E + R_E/h_{fe} + 1/y_{fe}) \quad \ldots\ldots \quad (12)$$

where the transistor parameters are those of VT_3. In fact this equation is still not exact because of the approximations made in writing equations (1) and (9); but the errors in equations (1) and (9) are negligible compared with the effect of the additional terms in equation (12). In a typical case each of these additional terms would have an effect of about -1 per cent on the resonance frequency, and each would have a temperature coefficient of about $-0{\cdot}6$ per cent/°K when allowance has been made for the fact that the d.c. collector current of VT_3 changes by about $-0{\cdot}3$ per cent/°K due to the variation in the base-emitter voltage of VT_2. The total resulting temperature coefficient is in the region of $+0{\cdot}01$ per cent/°K, but in practice it has not been possible to achieve a coefficient quite as low as this, presumably because of the effect of temperature on the passive components.

On the other hand, the Q factor of the circuit is seen from equation (11) to be primarily dependent on the transistor characteristics. The parameter y_{fe} has a temperature coefficient of $-0{\cdot}33$ per cent/°K at 300°K, and h_{ie} has a coefficient typically of about $+0{\cdot}9$ per cent/°K. In giving these values it has been assumed that the d.c. collector current remains constant, but in fact there is an additional (though less important) contribution to the temperature coefficient of the Q factor due to the fact that the d.c. collector current of VT_1 falls, and that of VT_2 increases, with increase of temperature.

In determining the circuit values for a given application, the external base-emitter resistors R_{B1} and R_{B2} are normally chosen to be about one-half the values of h_{ie} for the respective transistors, so that the positive temperature coefficients of R_1' and R_2' are about $+0{\cdot}3$ per cent/°K in order to cancel the negative temperature coefficient of y_{fe}. With this procedure one can normally obtain a temperature coefficient of 0·1 per cent/°K or less in the Q factor and therefore in the voltage gain at resonance, provided that capacitor losses are negligible. A further advantage of ensuring that the external base-emitter resistors are considerably less than h_{ie} is that the effect of variations in h_{ie} between individual transitors is thereby minimized.

TABLE 1

Component Values for a Typical Application

$R_{C1} = 15\text{k}\Omega$	$R_{B1} = 1{\cdot}0\text{k}\Omega$	$R_{EA} = 1{\cdot}8\text{k}\Omega$	$R_5 = 10\text{k}\Omega$
$R_{C2} = 2{\cdot}2\text{k}\Omega$	$R_{B2} = 220\Omega$	$R_{EB} = 1{\cdot}0\text{k}\Omega$	$R_1 = 10\text{k}\Omega$
C_1, C_2 = reactance 2·5kΩ at resonance frequency.			
VT_1, VT_2 = BCY43	VT_3 = 2N3702.		

Experimental Results

Performance of the Filter

As an example of the practical application of the principles discussed Table 1 gives the component values for a filter originally designed for an oscillator to give a sinusoidal output of up to 4V r.m.s.

In calculating the Q factor of this circuit from equation (11) it must be borne in mind that the transfer admittance y_{fe} of currently available low-power transistors is not accurately proportional to the d.c. collector current I_C when this exceeds about 1mA because of the effect of the base resistance, the value of which must be measured for the particular type of transistor being used. For the BCY 43 the base resistance is about 240Ω.

In a typical experiment the circuit of Fig. 2 was set up with $C_1 = C_2 = 10\,000$pF and with VT_3 having a measured a.c. gain of 110. The filter was found to have a resonance frequency of 6·1kHz and a Q factor of 15·4, the magnitude of the voltage gain at resonance being equal to the Q factor; these figures agree within experimental limits with values of 6·2kHz and 15·5 obtained from equations (12) and (11), which also show that the performance of the filter is nearly independent of the

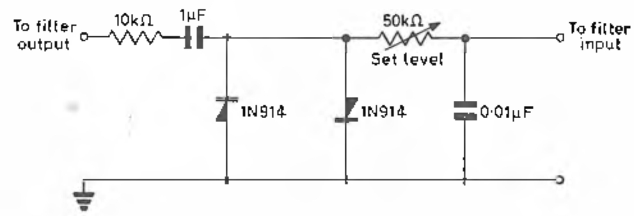

Fig. 5. Simple limiter with integrator, for use at 6KHz

a.c. gains of VT_1 and VT_3 provided that they are within the normal limits for these types (75 to 150 and > 60 respectively).

The circuit values are chosen as previously to give as low a temperature coefficient as possible, and in fact the observed temperature coefficient in Q factor and in voltage gain at resonance is substantially less than 0·1 per cent/°K. The temperature coefficient of resonance frequency is 0·02 per cent/°K, double the value calculated on the basis of equation (12); this discrepancy is taken to be due to the finite temperature coefficients of the passive components. Total distortion at an output level of 1·0V r.m.s. is less than 0·1 per cent, and the noise level less than 30μV at the output. All measurements were done with an a.c. coupled resistive load of 10kΩ connected across the output.

For certain selective-amplifier applications it has been found necessary to use a higher Q factor, and it may be calculated from equation (11) that a loaded Q factor of 28 may be expected if R_{B2} is changed from 220Ω to 470Ω, at the expense of an increase in temperature coefficient and distortion. (The distortion increase will be due to the fact that VT_3 will be running at a lower level of d.c. collector current, so that a given a.c. output represents a greater fractional swing than before.) With this modification the observed Q was 29, the temperature coefficient of Q factor and gain became $+0{\cdot}2$ per cent/°K and the total distortion at an output level of 1·0V r.m.s. increased to about 0·2 per cent.

Application to an Oscillator

In order to use the filter as a sinusoidal oscillator, it is only necessary to connect a suitable limiter between the output and the input terminals. An ideal limiter is required to provide an output having a fundamental component which is coherent in phase with the input but whose amplitude is independent of the amplitude of the input above a certain level. Although harmonics in the limiter output are not of primary importance because of the selectivity of the filter, they can lead to an undesirably high level of third harmonic in a system designed for low distortion. However, when applied to the circuit of Fig. 2 the limiter is required to give a phase lag of $\pi/2$ in the fundamental, the provision of which will enable a useful reduction of 10dB to be made in the third-harmonic content of the output.

The design of the limiter can be made as sophisticated as is necessary to provide the required degree of amplitude stability. Here the circuit will be discussed of only the simplest kind of limiter, consisting essentially of two junction diodes and a resistor, which suffers from the basic disadvantage of a relatively large temperature coefficient.

The circuit is shown in Fig. 5, where the additional capacitor is seen to be of the same size as C_1 and C_2. For a sufficiently large input swing (1V r.m.s. or more) the voltage across the diodes approximates to a square-wave, and that across the limiter output to a triangular waveform. The fundamental phase-shift is obviously never as much as $\pi/2$, but the system compensates for this by oscillating at a frequency one or two per cent higher than the centre of the filter response. When this limiter is used in conjunction with the circuit values given in Table 1, and with the amplitude of oscillation adjusted to 1·0V r.m.s. at the output, the total harmonic distortion is about 0·1 per cent with equal contributions of second and third harmonic, the temperature coefficients being $-0{\cdot}4$ per cent/°K in amplitude and $+0{\cdot}02$ per cent/°K in frequency. The amplitude temperature coefficient may be halved by operating at an output level of 3V r.m.s. at the expense of an increase in second-harmonic distortion to about 0·3 per cent. In the form described here, this oscillator has been very useful in equipment where long-term amplitude stability is not important; the frequency can be changed by changing C_1, C_2, and the limiter capacitor (keeping them still equal to one another) and the amplitude range can be extended at the lower end by the use of germanium diodes in the limiter rather than silicon diodes. Where a frequency adjustment control has been required, a variable resistor has been inserted in series or parallel with R_5 or R_E.

Acknowledgment

The authors are greatly indebted to Mr. D. W. Harding for his work in constructing and evaluating the prototypes.

REFERENCES

1. KNOTT, K. F., BOWMAN, J. S. An Active Filter for Low Frequency Noise Power Spectra Measurement. *Electronic Engng.* 38, 738 (1966).
2. ORCHARD, H. J. Inductorless Filters. *Electronics Letters.* 2, 224 (1966).
3. WOODARD, J., NEWCOMB, R. W. Sensitivity Improvement of Inductorless Filters. *Electronics Letters.* 2, 349 (1966).
4. FOSS, R. C., GREEN, B. J. Q Factor, Q Stability and Gain in Active Filters. *Electronics Letters.* 2, 99 (1966).
5. RAO, T. N., NEWCOMB, R. W. Direct-Coupled Gyrator Suitable for Integrated Circuits and Time Variation. *Electronics Letters.* 2, 250 (1966).
6. SHEAHAN, D. F., ORCHARD, H. J. Integratable Gyrator Using MOS and Bipolar Transistors. *Electronics Letters.* 2, 390 (1966).
7. SUTCLIFFE, H. Tunable Filter for Low Frequencies Using Operational Amplifiers. *Electronic Engng.* 36, 399 (1964).

The Parameters of the Current-Driven Source-Follower

By P. F. Ridler, B.E.(N.Z.), M.I.E.E., A.M.Rhod.I.E.

This article describes a solid state amplifier having unity gain, which uses a unipolar transistor and complementary bipolar current driver. Its input impedance is extremely high and output impedance low, making it suitable for the input stages of electronic instruments.

(Voir page 341 pour le résumé en français: Zusammenfassung in deutscher Sprache auf Seite 343)

FOR stages where high input impedance is necessary, the field-effect transistor is nowadays an obvious choice, but its rather low forward transfer admittance gives it an ill-defined gain in the common-drain mode unless it is used with a high load impedance. If a complementary bipolar transistor is used to supplement the transfer admittance, the combination possesses the advantages formerly typified by the 'Darlington' configuration.

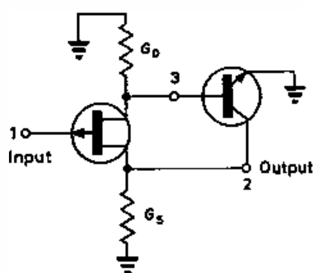

Fig. 1. Circuit of unity gain amplifier

Circuit Parameters

Considering the circuit of Fig. 1, one has a field effect transistor, with source and drain loads, feeding its drain output into the base of a complementary bipolar transistor. The output of the bipolar is fed back to the source of the f.e.t. This circuit may be subdivided into two component parts as shown in Fig. 2, and the admittance matrices written for each part. If the admittance matrices for the active devices are:

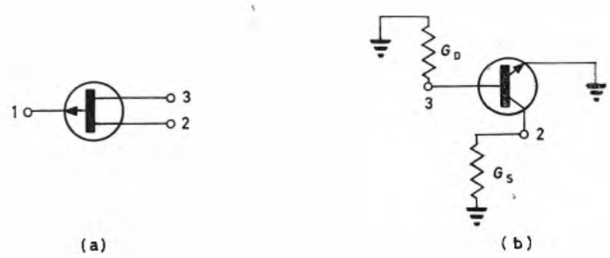

Fig. 2. Circuit of Fig. 1 subdivided for analysis

$$[Y]_{\text{F.E.T.}} = \begin{bmatrix} y_{is} & y_{rs} \\ y_{fs} & y_{os} \end{bmatrix}$$ where the y's are the common source parameters of the f.e.t.

and:

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$$[Y]_{\text{B.P.T.}} = \begin{bmatrix} y_{ie} & y_{re} \\ y_{fe} & y_{oe} \end{bmatrix}$$ where the y's are the common emitter parameters of the bipolar transistors

the matrices of the component circuits of Fig. 2 are:

$$[Y]_{2a} = \begin{bmatrix} y_{is} & -(y_{is}+y_{rs}) & y_{rs} \\ -(y_{is}+y_{fs}) & \Sigma y_s & -(y_{rs}+y_{os}) \\ y_{fs} & -(y_{fs}+y_{os}) & y_{os} \end{bmatrix} \quad \ldots (1)$$

where $\Sigma y_s = y_{is}+y_{rs}+y_{fs}+y_{os}$

$$[Y]_{2b} = \begin{bmatrix} 0 & 0 & 0 \\ 0 & y_{oe}+G_s & y_{fe} \\ 0 & y_{re} & y_{ie}+G_D \end{bmatrix} \quad \ldots\ldots\ldots (2)$$

which may be superposed to give the circuit of Fig. 1 by matrix addition, yielding:

$$[Y]_1 = \begin{bmatrix} y_{is} & -(y_{is}+y_{rs}) & y_{rs} \\ -(y_{is}+y_{fs}) & \Sigma y_s+y_{oe}+G_s & -(y_{rs}+y_{os})+y_{fe} \\ y_{fs} & -(y_{fs}+y_{os})+y_{re} & y_{os}+y_{ie}+G_D \end{bmatrix} \quad \ldots\ldots\ldots\ldots (3)$$

Making the assumptions that:

$y_{os} \ll y_{fs} \gg y_{is} \qquad y_{ie} \gg y_{os}$

$y_{fs} \gg y_{rs} \qquad y_{oe} \ll y_{fs} \gg y_{re}$

all of which are valid under normal working conditions:

$$[Y]_1 = \begin{bmatrix} y_{is} & -(y_{is}+y_{rs}) & y_{rs} \\ -y_{fs} & y_{fs}+G_s & y_{fe} \\ y_{fs} & -y_{fs} & y_{ie}+G_D \end{bmatrix} \quad \ldots\ldots (4)$$

Eliminating terminal 3 by partitioning and reducing to a 2×2 matrix, the matrix becomes:

$$[Y] = \begin{bmatrix} y_{is}-\gamma y_{rs} & -y_{is}+(\gamma-1)y_{rs} \\ -(y_{fs}+\gamma y_{fe}) & y_{fs}+\gamma y_{fe}+G_s \end{bmatrix} \quad \ldots\ldots\ldots (5)$$

where:

$$\gamma = \frac{y_{fs}}{y_{ie}+G_D}$$

Voltage Gain

The voltage gain with a load Y_L is:

$$G_V = \frac{-y_{21}}{y_{22}+Y_L} = \frac{y_{fs}+\gamma y_{fe}}{y_{fs}+\gamma y_{fe}+G_s+Y_L} \quad \ldots\ldots\ldots (6)$$

where y_{mn} is an element of $[Y]$

$$\therefore\ 1/G_V = 1 + \frac{G_s+Y_L}{y_{fs}+\gamma y_{fe}} \quad \ldots\ldots\ldots\ldots\ldots\ldots\ldots (7)$$

but $\gamma y_{fe} = \dfrac{y_{fe}y_{fs}}{y_{ie}+G_D}$ and if $G_D \ll y_{ie}$

$\simeq \beta y_{fs}$

where β is the current gain of the bipolar transistor, so that:

$$1/G_V \simeq 1 + \frac{G_s+Y_L}{\beta y_{fs}} \quad \ldots\ldots\ldots\ldots\ldots\ldots\ldots (8)$$

and:

$$G_V \simeq 1 - \frac{G_s+Y_L}{\beta y_{fs}} \quad \ldots\ldots\ldots\ldots\ldots\ldots\ldots (9)$$

if $(G_s+Y_L)/\beta y_{fs} \ll 1$.

For the typical circuit of Fig. 3 at frequencies less than 100kHz:

$$[Y]_{\text{F.E.T.}} = \begin{bmatrix} j\omega 10^{-11} & -j\omega 10^{-12} \\ 2\times 10^{-3} & 10^{-5} \end{bmatrix}$$

$$[Y]_{\text{B.P.T.}} = \begin{bmatrix} 5\times 10^{-4} & -j\omega^{-12} \\ 5\times 10^{-2} & 10^{-5} \end{bmatrix}$$

$G_s = 5\times 10^{-4}\Omega^{-1} \qquad G_D = 10^{-4}\Omega^{-1}$

which can be seen to justify the assumption made above.

Hence from equation (7):

$G_V = 0{\cdot}997 \qquad Y_L = 10^{-4}\Omega^{-1}$

$G_V = 0{\cdot}990$ for $Y_L = 10^{-3}\Omega^{-1}$

$G_V = 0{\cdot}94 \qquad Y_L = 10^{-2}\Omega^{-1}$

Input Admittance

The input admittance is:

$$Y_{IN} = \frac{y_{11}Y_L+\Delta_Y}{y_{22}+Y_L} \quad \ldots\ldots\ldots\ldots\ldots\ldots (10)$$

$$= \frac{y_{is}(G_s+Y_L) - y_{rs}[\gamma(G_s+Y_L)+y_{fs}+\gamma y_{fe}]}{y_{fs}+\gamma y_{fe}+G_s+Y_L} \quad \ldots\ldots (11)$$

where Δ_Y is the determinant of $[Y]$

$$\simeq \frac{y_{is}(G_s+Y_L) - y_{rs}(1+\beta)y_{fs}}{(1+\beta)y_{fs}} \quad \ldots\ldots\ldots\ldots\ldots (12)$$

if $y_{fe} \gg G_s + Y_L$

$$= -y_{rs} + (y_{is}/y_{fs})\frac{G_s+Y_L}{1+\beta} \quad \ldots\ldots\ldots\ldots\ldots\ldots (13)$$

Using equation (11) and the typical figures given previously, Y_{IN} becomes:

$j\omega\, 10^{-12}$ (1pF) for $Y_L = 10^{-4}\Omega^{-1}$

$j\omega\, 1{\cdot}05\times 10^{-12}$ for $Y_L = 10^{-3}\Omega^{-1}$

$j\omega\, 1{\cdot}7\times 10^{-12}$ for $Y_L = 10^{-2}\Omega^{-1}$

Output Admittance

The output admittance is:

$$Y_o = \frac{y_{22}Y_g+\Delta_Y}{y_{11}+Y_g} \quad \ldots\ldots\ldots\ldots\ldots\ldots (14)$$

where Y_g is the generator admittance

and:

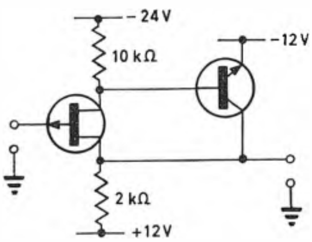

Fig. 3. Practical circuit

for $Y_g = 0 \qquad Y_o = {}_IY_o = \Delta_Y/y_{11}$ (current source)

or $Y_g = \infty \qquad Y_o = {}_VY_o = y_{22}$ (voltage source)

Y_o will therefore be between the values:

$${}_VY_o = y_{fs}+\gamma y_{fe}+G_s \quad \ldots\ldots\ldots\ldots (15)$$

$$\simeq (1+\beta)y_{fs}+G_s \quad \ldots\ldots\ldots\ldots (16)$$

and:

$${}_IY_o = \frac{y_{is}G_s - y_{rs}(y_{fs}+\gamma y_{fe}+\gamma G_s)}{y_{is}-\gamma y_{rs}} \quad \ldots\ldots (17)$$

$$= -y_{rs}\frac{y_{fs}+\gamma y_{fe}}{y_{is}-\gamma y_{rs}} + G_s$$

$$\simeq \frac{(1+\beta)y_{fs}}{-(y_{is}/y_{rs})+\gamma} + G_s \quad \ldots\ldots\ldots\ldots (18)$$

For the circuit of Fig. 3, using equation (17):

${}_VY_o = 0{\cdot}17\Omega^{-1} \qquad {}_IY_o = 1{\cdot}25\times 10^{-2}\Omega^{-1}$

The author has used the circuit described for a number of purposes, including a highly linear Bootstrap time-base for which it is ideally suited.

The gain of the amplifier may be set exactly to unity by including an impedance between the collector of the bipolar and the source of the f.e.t., and taking the output from the bipolar collector. The analysis of this circuit has been done but is rather unwieldy and is worthwhile only under exceptional circumstances.

This circuit should form a useful 'building brick' for many purposes presently filled by the 'Darlington pair'.

The Design and Manufacture of Waveguide Tchebycheff Directional Couplers

By K. E. Hancock*

In this article a simple, versatile design procedure is given for Tchebycheff array directional waveguide couplers. The design procedure is based on a single array in the broadwall of the waveguide and this is extended to stacked arrays and parallel rows of holes. Hole layouts for multi-hole couplers in waveguide sizes WG10 to WG22 are tabulated.

(Voir page 341 pour le résumé en français: Zusammenfassung in deutscher Sprache auf Seite 343)

IN modern microwave practice, particularly in the engineering and quality assurance fields, there is often a requirement for broadband directional couplers. To fully utilize the potential of such techniques as swept frequency reflectometer measurements, a coupler with full waveguide bandwidth, minimum variation of coupling with frequency, and a directivity of at least 40dB is required. The latter parameter is of particular importance as the measurement error of the reflectometer is inversely proportional to the directivity.

Directional couplers in waveguide can be constructed in many forms, perhaps the most common using the coupling through pairs of holes spaced a quarter of a guide wavelength apart in the common wall of two adjacent waveguides. By increasing the number of holes, correct positioning of the holes and by arranging the comparative coupling coefficients correctly, wide bandwidth, high directivity and a small variation of coupling with frequency may be obtained. Two types of coupling hole arrays are in common use, the Binomial, in which the ratio of coupling coefficients vary as a binomial series, and the Tchebycheff, in which the variation is in the form of a Tchebycheff polynomial. However, assuming an equal number of holes, the Tchebycheff coupler will have a wider bandwidth and less deterioration of directivity at the band edges than will the binomial coupler. This is gained at the expense of more critical design parameters, and a number of equal amplitude ripples in the directivity curve proportional to the number of elements in the array. Here a simple versatile design procedure is postulated for Tchebycheff array directional couplers. A comparison of directivity obtained with Binomial and Tchebycheff arrays with equal numbers of holes is given in Fig. 1.

The design procedure followed below is based on a basic single array in the broadwall of the waveguide. Having given the design procedure for this type of coupler, its disadvantages are shown, and more sophisticated design methods using stacked arrays and parallel rows of holes are formulated.

Design Procedure

As with any coupler the required coupling factor, directivity and bandwidth should first be decided upon.

Before going further coupling and directivity will be defined. Referring to Fig. 2.

Coupling is the ratio of the power appearing at port A to the power entering port D

that is $C = -10 \log (P_D/P_A)$ in dB .... (1)

Whereas directivity is the ratio of the power appearing at port B to the power appearing at port A, the power input again being at port D

that is $D = -10 \log (P_A/P_B)$ in dB .... (2)

In both cases it is assumed that all ports are perfectly matched.

In this type of coupler extremely high theoretical directivities can be calculated. However, factors other than the coupling array limit the directivity obtainable to about 45dB over the full waveguide bandwidth with normal batch production techniques. Even this figure can only be obtained if the direction arm is terminated so that it presents a v.s.w.r. of near unity over the required bandwith. In practice, due to the difficulty of designing very high quality small radius bends, this will limit the design to an integral termination in the directional arm for directivities greater than 35 to 40dB.

This factor must be taken into account when calculating the physical length of the component. Full details of calculating the maximum permissible v.s.w.r. of the termination for a required directivity, will be given later.

Coupling at least as close as 3dB can be obtained with this type of coupler. However, as will be shown later,

Symbols Used

C	=	Coupling in decibels
p	=	Power in watts
D	=	Directivity
λ_{g1}	=	Guide wavelength at the lower end of the required bandwidth
λ_{g2}	=	Guide wavelength at the upper end of the required bandwidth
λ_{gmid}	=	Mean guide wavelength
n	=	Number of coupling elements in basic array
D_{min}	=	Minimum theoretical directivity
ϕ	=	$\dfrac{180°}{1 + (\lambda_{g1}/\lambda_{g2})}$
λ_g	=	Guide wavelength
a	=	Broad dimension of waveguide
x	=	Axis across broad dimension of waveguide
b	=	Narrow dimension of waveguide
d	=	Diameter of hole in inches
$1 - (1{\cdot}71d/\lambda_o)^2$	=	A term giving correction for resonance phenomena
λ_o	=	Free space wavelength
t	=	Wall thickness
$32(t/d)\,[1-(1{\cdot}71d/\lambda_o)]^{\frac{1}{2}}$	=	A term giving a correction the attenuation effect on a finite wall thickness
T	=	Theoretical directivity in decibels
V	=	V.S.W.R. of integral load
r	=	Power reflection coefficient

* *Canadian Marconi Co., formerly W. H. Sanders (Electronics) Ltd.*

the closer the coupling, the larger will be the number of stacked arrays required, particularly if a small variation of coupling with frequecy is aimed for. This necessitates a physically long coupler. In the author's opinion a 3dB coupler with a variation of $\pm 0{\cdot}5$dB over the waveguide bandwidth and a minimum directivity of 40dB over the same limits, is of the maximum reasonable length for

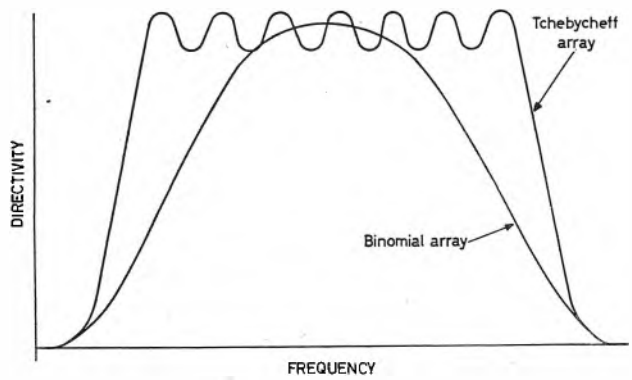

Fig. 1. Comparison of Tchebycheff and binomial arrays

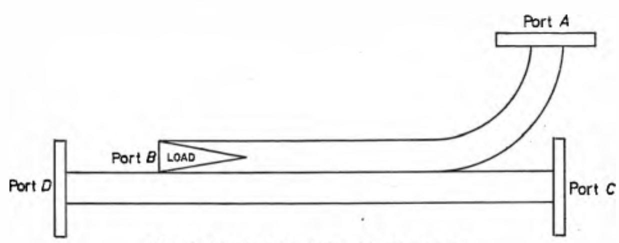

Fig. 2. Typical arrangement of coupler

TABLE 1

$n \backslash k$	0	1	2	3	4	5	6
1	2						
2		1·853					
3	4·869		3·434				
4		13·534		6·365			
5	45·294		33·442		11·795		
6		132·380		77·469		21·858	
7	445·373		355·449		172·277		40·507

bench use. If physical size is no object this performance can be considerably improved.

Calculation of Hole Spacing

As with most multi-element couplers the spacing between the counters is a quarter guide wavelength at mid-band and is given by:

$$\lambda_{gmid}/4 = \frac{\lambda_{g1}\,\lambda_{g2}}{2(\lambda_{g1} + \lambda_{g2})} \quad \ldots\ldots\ldots\ldots \quad (3)$$

Calculation of Basic Array Required

Having decided upon the required directivity and bandwidth the number of elements required in the basic array may be calculated from the following formulae[1].

$$n = 1 + \cosh^{-1}\frac{10\,(D_{min}/20)}{\cosh^{-1}(1/\cos\phi)} \ldots\ldots\ldots \quad (4)$$

Various factors such as manufacturing tolerances, attenuation in the guide comprising the coupler, and reflections from the load terminating the directional arm, will degrade the directivity below the theoretical figure.

The author has found that providing reasonable care is taken on these points, (greater details of which are given later), a figure of 3 to 5dB added to the required directivity will give an adequate theoretical minimum directivity (D_{min}) for use in the above formula. The nearest integer above the figure found in formula (4) should be taken.

Calculation of Coupling Coefficients

Having determined the number of holes required in the basic array to achieve the required directivity, the next step is to calculate the coupling coefficients such that their ratios are in the form of a Tchebycheff expansion. This is most conveniently achieved by the method of Graves, which although apparently unpublished has been used by other authors[2].

The factor x_0 is the first determined where:

$$x_0 = 1/\cos\phi \frac{1}{\cos\dfrac{180°}{1 + (\lambda_{g1}/\lambda_{g2})}} \quad \ldots\ldots\ldots\ldots \quad (5)$$

The coupling coefficients are then related to the Tchebycheff response as follows.

To clarify the explanation the figures for a 7-hole array in WG16 (λ_{g1} = 6·089cm, λ_{g2} = 2·849cm) are given. In this case $x_0 = 1{\cdot}853$.

To avoid confusion the minus signs in front of the coupling coefficients will be omitted in all coupling coefficient calculations. Referring to Table 1.

(a) In the first column, first row, insert the number 2. In the second column, second row, insert the value of x_0 (1·853).

(b) Along the diagonal formed by the number 2 and the value of x_0 insert the value of $(x_0)^k$, e.g. $(1{\cdot}853)^2$ is placed in the third column, third row, $(1{\cdot}853)^3$ is placed in the fourth column, fourth row, etc.

(c) For any other square in the first column, multiply the element above and to the right by $2x_0$ and then subtract the element in the second row above the entry to be found; e.g. the value for the element in the first column, third row, is found by multiplying (1·853) (3·7065) = 6·869, then 6·869 − 2 = 4·869.

(d) In any other square, add the element appearing one row above and to the left of the entry to be found to the element appearing one row above and to the right of the entry to be found. Multiply this sum by x_0 and then subtract the element in the second row above the entry to be found; e.g. the value for the element in the second column, fourth row is found by adding 4·869 + 3·434 = 8·303 and then multiplying (8·303) (1·853) = 15·387 and subtracting 1·853 from 15·387 = 13·534.

The figures in the horizontal rows obtained in Table 1 are proportional to the coupling coefficients for the required array. To normalize these coefficients the figures in the required row (row 7 in the example) are divided through by the smallest number, giving:

A	*B*	*C*	*D*	*C*	*B*	*A*
1·000	4·253	8·775	10·995	8·775	4·253	1·000

The relative coupling of each element in decibels is now derived by summing the coefficients and dividing the sum by each coefficient. Twenty times the log of this figure will give the required coupling in decibels of each hole.

In the example:

$$
\begin{aligned}
2A &= 2{\cdot}000 \\
2B &= 8{\cdot}506 \\
2C &= 17{\cdot}550 \\
1D &= 10{\cdot}995 \\
\hline
&\ 39{\cdot}051
\end{aligned}
$$

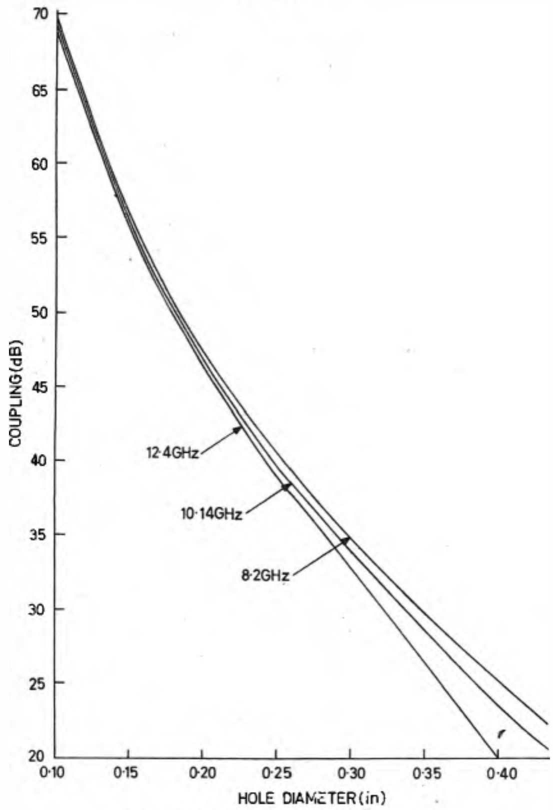

Fig. 3. Hole diameter against coupling

$$\text{Holes } A = 20 \log \frac{39{\cdot}051}{1} \text{ dB} = 31{\cdot}832\text{dB}$$

$$\text{Holes } B = 20 \log \frac{39{\cdot}051}{4{\cdot}253} \text{ dB} = 19{\cdot}259\text{dB}$$

$$\text{Holes } C = 20 \log \frac{39{\cdot}051}{8{\cdot}775} \text{ dB} = 13{\cdot}968\text{dB}$$

$$\text{Holes } D = 20 \log \frac{39{\cdot}051}{10{\cdot}051} \text{ dB} = 11{\cdot}776\text{dB}$$

Having obtained the relative coupling of each hole, the required attenuation of the coupler is added to each figure to give the final required coupling of each hole.

In the example for a 10dB couple, Holes $A = 41{\cdot}832$, Holes $B = 29{\cdot}259$dB, Holes $C = 23{\cdot}968$dB and Holes $D = 21{\cdot}776$dB.

Calculation of Hole Diameter

Hole diameters of coupling arrays in general are based on formulae postulated by Bethe[3] for small holes. It can be shown, however, that for coupling holes in the common broad wall of two parallel waveguides, the coupling through holes located at the centre line varies with λ_g, while if located at the waveguide wall the coupling varies with $1/\lambda_g$. By plotting coupling against x/a it is found that for a value of $x/a = 0{\cdot}203$ the coupling is flat. If this value is used in Bethe's formula, the equation for the coupling of a hole located at this point across the broad wall between two parallel waveguides becomes:

$$c = 20 \log \left[12a^2b/\pi d^3 \left(1 - (1{\cdot}71d/\lambda_0)^2\right)\right] + 32{\cdot}0\,(t/d)\left[1 - (1{\cdot}71d^2/\lambda_0)\right]^{\frac{1}{2}} \quad \ldots\ldots\ldots\ (6)$$

A graph of this equation showing coupling against hole diameter for the relevant waveguide size and wall thickness provides a convenient method of design. A contracted version is given in Fig. 3 for WG16 and a wall thickness of 0·025in. This yields hole diameters as follows:

$A = 0{\cdot}234$in $B = 0{\cdot}343$in $C = 0{\cdot}397$in $D = 0{\cdot}421$in

This is for a single array of seven holes positioned $0{\cdot}203a$ from one wall.

Disadvantages of a Single Array

The single array described above has two serious disadvantages. It will be seen that particularly where close coupling is required the hole sizes become quite large and tend to overlap. If the hole spacing is calculated for the example it will be seen that both the C and D holes would overlap. Also the coupling hole formula is based on small hole theory and the author has found empirically that if the hole diameter exceeds about $0{\cdot}187\lambda_g$ the error introduced by a finite hole size ceases to be negligible.

The second disadvantage is apparent from Fig. 3 that as the hole size decreases, so does the variation of coupling with frequency. Thus to design a component with close coupling and/or a small spread of coupling with frequency a more sophisticated hole layout is required.

Methods of Obtaining Close Coupling and Low Frequency Spread

The disadvantages of a single array may be overcome by either or both of two methods:

A number of arrays may be superimposed to give the same coupling with a larger number of small holes, while the high directivity and the Tchebycheff form are maintained. This may be done as follows:

Returning again to the example given above and considering the normalized coupling coefficients of the basic array:

A	B	C	D	C	B	A
1·000	4·253	8·775	10·995	8·775	4·253	1·000

Let this array be superimposed, say six times, in the following manner:

ABCDCBA
ABCDCBA
ABCDCBA
ABCDCBA
ABCDCBA
ABCDCBA

Summing the elements of these arrays gives:

A, *B*, *C*, *AD*, *BC*, *BC*, *ADA*, *BC*, *BC*, *ADA*, *BC*, *BC*, *ADA*, *BC*, *BC*, *ADA*, *BC*, *BC*, *AD*, *C*, *B*, *A*

Following the procedure given earlier these coefficients are now summed, the sum being divided by the individual coefficient and the result converted to decibels to which the required attenuation of the coupler is added.

The second method of reducing the required hole diameter is to simply parallel two rows of holes, each being $0{\cdot}203a$ from opposite sides of the guide. This will increase by 6dB the total power coupled through the guide, the hole sizes remaining the same.

In the example combining these two methods will give the figures given in Table 2.

It will be seen that this will give substantial reduction in the required hole size and therefore a reduction of coupling spread, at the cost of a larger number of holes. In the example the number of holes required will be 44.

Minimizing the Length of the Array

Using both the methods given above and returning to the example, the following results will be obtained.

It will be seen that the largest holes are now only about $0{\cdot}156\lambda_{g\,mid}$ in diameter and the coupling spread over the waveguide band will be about $\pm 0{\cdot}5$dB. Referring to the summation of the arrays above it will be noted that each stack will increase the sum of the coefficients by $2(B+C)+2A+D$. By working backwards therefore the largest holes may be made of the order of $0{\cdot}17\lambda_{g\,mid}$ in diameter and the number of stacks required to give this figure calculated. This will normally give the shortest stack that will provide the required performance. In certain cases an increase in the number of elements in the basic array will slightly decrease the number of stacks required thus further reducing the length of the final array.

TABLE 2

HOLE	BASIC COUPLING (dB)	ATTENUATION OF REQUIRED COUPLER (dB)	ADDITION FOR TWO ROWS IN PARALLEL (dB)	TOTAL (dB)	HOLE DIAMETER (in.)
$A = 20\text{Log}\dfrac{234{\cdot}4}{1} =$	47·384	10	6	63·384	0·120
$B = 20\text{Log}\dfrac{234{\cdot}4}{4{\cdot}253} =$	34·826	10	6	50·826	0·1765
$C = 20\text{Log}\dfrac{234{\cdot}4}{8{\cdot}775} =$	28·534	10	6	44·534	0·215
$A+D = 20\text{Log}\dfrac{234{\cdot}4}{11{\cdot}995} =$	25·918	10	6	41·918	0·235
$2A+D = 20\text{Log}\dfrac{234{\cdot}4}{12{\cdot}995} =$	25·120	10	6	41·120	0·2405
$B+C = 20\text{Log}\dfrac{234{\cdot}4}{13{\cdot}028} =$	25·102	10	6	41·102	0·2400

Empirical Modification to Coupling Formula for Very Large Numbers of Holes

It has been found in practice that if the number of holes used is greater than about fifty, the expression for attenuation effect of a finite wall thickness does not fully compensate for this error. However for up to at least 70 holes, an empirical correction of 0·5dB to the final coupling figure for each hole is found to fully compensate for this effect.

Modification of Arrays to Optimize a Particular Parameter

It will readily be appreciated that the methods detailed above can be used to give couplers in which any parameter such as coupling spread, directivity and bandwidth may be optimized.

The Effect on Directivity of Reflections from the Load Arm

Any reflections occurring at the load arm, that is part B in Fig. 2, will have a deleterious effect on the directivity. To understand this and to calculate its effect, consider again the coupler shown in Fig. 2. The wanted output at port A will be $10^{c/10}$ the input at port D (from equation (1)), port A being isolated from port C by the sum of the coupling and directivity. Considering the effect on directivity of reflection from port B referring to an unwanted input at port C, port B being perfectly matched.

As previously defined

$$C = -10\log(P_c/P_B)$$

$$D = -10\log(P_B/P_A)$$

$$\therefore P_B = P_c 10^{c/10}$$

$P_A = P_c 10^{(c+T)/10}$ if P_B is perfectly matched. If P_B is not perfectly matched the power reflected from

$$\text{port } B = P_c\,(10^{c/10})r \quad \ldots\ldots\ldots\ldots\ldots\ldots \quad (7)$$

and the true power at

$$\text{port } B = P_c 10^{c/10} - P_c 10^{c/10} r \quad \ldots\ldots\ldots \quad (8)$$

Assuming a constant mismatch with frequency and ignoring power lost by recoupling back to the main guide, the total power at P_A will vary between

$$P_c 10^{(c+T)/10} + P_c 10^{c/10} r \text{ and } P_c 10^{(c+10)/10} \;\; P_c 10^{(c/10)} r$$

depending on the electrical distance between P_B and P_A. Substituting equations (7) and (8) in (2) and simplifying one obtains a directivity between the limits

$$-10\log\frac{1-r}{10^{T/10}\pm r} \quad \ldots\ldots\ldots\ldots\ldots\ldots \quad (9)$$

In the example using equation (4) transposed to give

$$T = -\ 20\log\ (\cosh\ (n-1)\ \cosh^{-1}\ (1/\cos\phi)\text{dB}$$

$$T = -47{\cdot}894\text{dB}$$

Assuming a load v.s.w.r. of 1·02 : 1

$$r = \left(\frac{1{\cdot}02-1}{1{\cdot}02+1}\right)^2 = {\cdot}0000984 = 10^{-4.0068}$$

giving a minimum effective directivity of $-39{\cdot}407$dB while as the power reflected from the load is greater than that reaching port A due to the finite theoretical directivity, the effective directivity will reach an infinite peak and return to a finite value in the opposite phase when the reflected power at port *A* is in anti-phase with the direct power.

It will readily be appreciated therefore that any mismatch of the load and waveguide will have a profound effect on the directivity of the coupler, and hence the accuracy of any measurement circuit of which it is a part.

Manufacture of Multihole Couplers

Manufacture of the couplers may conveniently be divided into four parts as follows: the breakdown into piece parts; the fabrication of the holes; the assembly of the piece parts: and the tolerancing of the various dimensions.

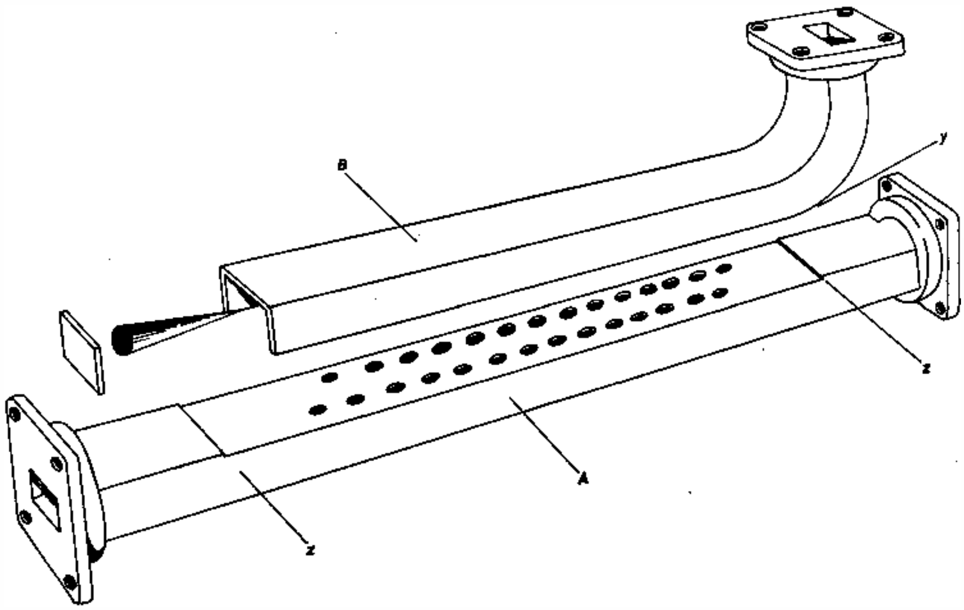

Fig. 4. Exploded view of coupler

The Breakdown into Piece Parts

It has been found that the best results are obtained when the hole arrays are fabricated directly in the upper wall of the lower waveguide. An alternative is to manufacture a separate hole plate and mill off one wall from both main waveguides. The disadvantage of this method is the difficulty of keeping the hole plate flat and free from wrinkles, and the awkwardness of fabrication. An exploded diagram showing one method which has been found to be most satisfactory is shown as Fig. 4.

The holes are fabricated in waveguide *A* which is then milled down to the required wall thickness between points *z-z*

The lower wall of waveguide *B* is removed completely. Great care should be taken during this operation to preserve a 'feather edge' at the bend (point *y*) in order to prevent a 'step' in the assembled coupler. The bend is put into the guide before machining, and if there is any doubt as to the wideband performance of the bend, its v.s.w.r. may be measured at this stage.

Any bowing of the waveguides after machining should be straightened out with mandrels. It has been found that stress releaving before machining has little or no effect.

Fabrication of the Holes

As the positioning and diameters of the holes must be held to fairly tight tolerances, precision methods of machining must be used. Either jig-boring or spark machining is normally employed. Jig-boring is most useful for 'one-offs' or small runs, but has the disadvantages of being difficult to deburr without distorting or chamfering the holes, and of creating stresses which may cause distortion of the guide. Spark machining, while being more expensive in the first instance, requiring a precision electrode jig, enables all the holes to be machined in one operation, giving good repeatability, raises no burrs, and gives rise to very little stress.

Assembly of the Piece Parts

This has been found to be the most critical operation in the manufacture of the coupler. The main pieces of waveguide are soldered together. It has been found, however, that the long solder joints, while appearing to be perfect to the most careful external inspection both inside and outside of the guide, are very prone to have internal gaps, often several inches long which appear to interrupt the wall currents, giving rise to high v.s.w.r.'s and subsequent serious degrading of the performance. This may be overcome entirely by the use of a flux which permits a capillary action to take place, and eliminates gaps, fillets and blobs.

Tolerancing of Dimensions

In this type of coupler it is more than ordinarily important to maintain waveguide dimensions to the normal tolerances. However, the main tolerancing problem is of course the hole arrays. The tolerances given below have been found by trial and error, and are adequate, however, they should not be regarded as optimum. It has been found that for waveguide sizes between 10 and 18 ±·001in tolelrances on hole diameters, hole centre spacing, and longitudinal positioning will be required. For waveguide 22 ±·005in for these parameters will be required.

As mentioned in the text the positioning of the hole centres from the waveguide wall will critically effect the coupling spread. To keep this parameter to within ±0·5dB over the waveguide bandwidth the run of the complete array from nominal parallel should be not greater than ±·005in in waveguide sizes 10 and 12, ±·002in in waveguide sizes 14, 15 and 16, and ±·001in in waveguide size 22.

Final Note

It should be borne in mind that the components used in the test circuit for high directivity couplers should be of comparative performance as the coupler. This applies in particular to the load used to terminate the main waveguide. Any degrading of performance of this component or its poor alignment with the coupler will have an effect comparable with a poor integral load, and may cause an apparent severe decrease of directivity.

Appendix

Hole Layouts for Multi-hole Couplers

Note: All holes are arranged in two parallel rows, staggered by one hole, the centres of each row of holes being on a line 0·203a from the inside wall of the guide. In each row the holes run alphabetically to letter *D*, (or *E* in the case of some couplers) all the *D* holes are placed adjacent then from *D* back to *A*, so that the row is symmetrical, starting with the small holes, expanding to the largest, then contracting back to the smallest. All dimensions are in inches.

References

1. Levy, R. Guide to the Practical Application of Tchebycheff Functions to the Design of Microwave Components. *Proc. Instn. Elect. Engrs. 106C*, 193 (1959).
2. Cohn, S. B. Optimum Design of Stepped Transmission Line Transformers. *I.R.E. Trans. Microwave Theory & Tech. MTT3*, 16 (Apr. 1955).
3. Bethe, H. S. Theory of Diffraction by Small Holes. *Physical Rev. 66*, 163-182, 2nd ser. (1 and 15 October 1944).

WG10

	HOLE	SIZE	NUMBER
3dB	*A*	·423	4
	B	·645	4
	C	·785	4
	D	·861	50
6dB	*A*	·418	4
	B	·617	4
	C	·747	4
	D	·828	38
10dB	*A*	·388	4
	B	·568	4
	C	·690	4
	D	·758	32
20dB	*A*	·382	4
	B	·557	4
	C	·680	4
	D	·740	4
	E	·759	4

Hole spacing 1·1975
Wall thickness ·080

WG12

	HOLE	SIZE	NUMBER
3dB	*A*	·294	4
	B	·431	4
	C	·525	4
	D	·576	50
6dB	*A*	·281	4
	B	·412	4
	C	·509	4
	D	·556	38
10dB	*A*	·262	4
	B	·382	4
	C	·465	4
	D	·511	32
20dB	*A*	·258	4
	B	·376	4
	C	·460	4
	D	·498	4
	E	·508	4

Hole spacing ·8073
Wall thickness ·064

WG14

	HOLE	SIZE	NUMBER
3dB	*A*	·188	4
	B	·2865	4
	C	·368	4
	D	·404	50
6dB	*A*	·180	4
	B	·281	4
	C	·352	4
	D	·388	38
10dB	*A*	·166	4
	B	·258	4
	C	·323	4
	D	·356	32
20dB	*A*	·164	4
	B	·254	4
	C	·318	4
	D	·341	4
	E	·352	4

Hole spacing ·537
Wall thickness ·030

WG15

	HOLE	SIZE	NUMBER
3dB	*A*	·1547	4
	B	·239	4
	C	·298	4
	D	·325	4
	E	·3305	46
6dB	*A*	·1460	4
	B	·2295	4
	C	·2875	4
	D	·3180	38
10dB	*A*	·1368	4
	B	·2105	4
	C	·2635	4
	D	·2915	32
20dB	*A*	·135	4
	B	·210	4
	C	·260	4
	D	·279	4
	E	·288	4

Hole spacing ·447
Wall thickness ·025

WG16

	HOLE	SIZE	NUMBER
3dB	*A*	·1355	4
	B	·201	4
	C	·245	4
	D	·2725	50
6dB	*A*	·1300	4
	B	·1925	4
	C	·2340	4
	D	·2610	38
10dB	*A*	·1200	4
	B	·1763	4
	C	·2145	4
	D	·2360	32
20dB	*A*	·1180	4
	B	·1745	4
	C	·2130	4
	D	·2320	4
	E	·2360	4

Hole spacing ·383
Wall thickness ·025

WG18

	HOLE	SIZE	NUMBER
3dB	*A*	·0960	4
	B	·1438	4
	C	·1762	4
	D	·1890	50
6dB	*A*	·0924	4
	B	·1380	4
	C	·1700	4
	D	·1856	38
10dB	*A*	·0852	4
	B	·1275	4
	C	·1564	4
	D	·1716	32
20dB	*A*	·08420	4
	B	·12570	4
	C	·1540	4
	D	·1670	4
	E	·1707	4

Hole spacing ·2533
Wall thickness ·020

WG22

	HOLE	SIZE	NUMBER
3dB	*A*	·0519	4
	B	·0741	4
	C	·0880	4
	D	·0950	50
6dB	*A*	·0513	4
	B	·0715	4
	C	·0857	4
	D	·0923	38
10dB	*A*	·0484	4
	B	·0685	4
	C	·0798	4
	D	·0858	32
20dB	*A*	·0480	4
	B	·0666	4
	C	·0794	4
	D	·0837	4
	E	·0861	4

Hole spacing ·1178
Wall thickness ·020

A Rapidly Reversible 700kHz Digital Decade Counter

By J. F. Brandwijk* and P. H. Sydenham*, B.E.

In precise digital instrumentation fast counters capable of being rapidly reversed are required. A solid state decade counter is described that can count to $0{\cdot}7 \times 10^6$p/s and be reversed in 4μsec. Counting is in the 'excess-3' binary code using direct coupled feedback to obtain a decimal count. All circuits, including decoding and display lampdrives, are given along with performance figures.

(Voir page 341 pour le résumé en français: Zusammenfassung in deutscher Sprache auf Seite 343)

IN digital instrumentation and control it is necessary to have counters that can count in either direction in order to record the resultant value of excursions made by the parameter being monitored. As the precision is improved more counts per unit excursion are produced increasing the pulse repetition rates generated and decreasing the available time for the counter to reverse direction without loss of counts or significant digits.

The requirements needed in the digital control application measuring shaft position, were that the counter should count to about 10^6p/s and be able to reverse satisfactorily in about 5μsec. To enable the counter stages to be easily coupled to a digital computer, the counting code used in the decade stages should be some continuous code, it is necessary to introduce feedback which forces the four bistable counter stages to pass through only ten stable transition states instead of the naturally occurring sixteen. In the 'excess-3' code, see Table 1, decimal 0 is chosen as decimal 3 in the 1248 code. Feedback is used to return the four binary stages to the desired state when ten counts have been registered. The bistables are designated A, B, C, D with A the least significant digit. When the counter reaches 0011 the next pulse will give state 1011, which is detected by a feedback gate returning the bistables to 1100 thus recounting from zero. A decimal carry pulse is derived from the 9 to 0 transition to increment the next decade stage.

To reverse the counting direction the input to each

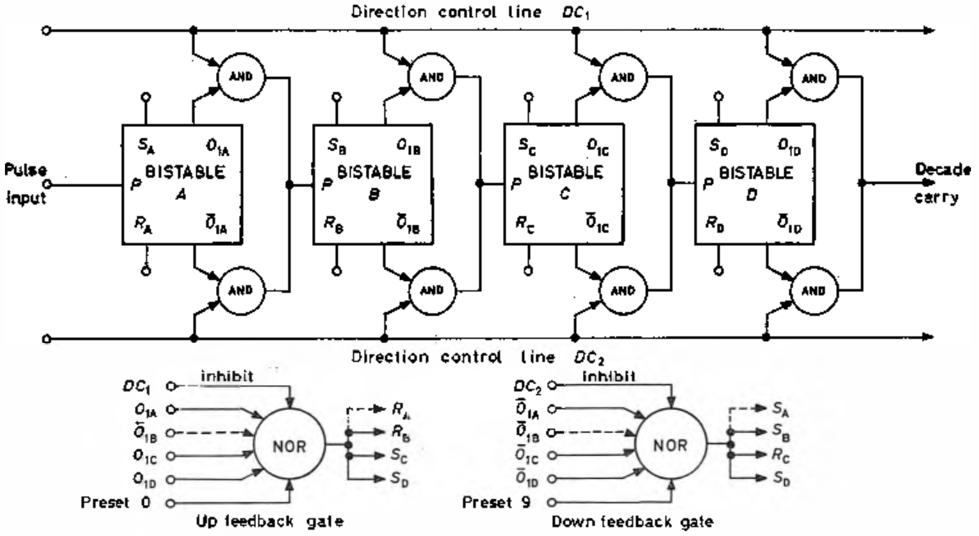

Fig. 1. Logical arrangement of a bi-directional, decade counter stage using excess —3 logic

group of the 1248 binary code which occurs naturally from cascading binary stages.

Search of recent literature showed that no reversible counter with all these specifications had been described. The subsequent design was modelled in principle on that of Tassinari[1], who reported a bi-directional decade counter using an 'excess-3' binary code with direct coupled feedback. (Whereas this article adequately reported the principles, the performance of the circuits given was not quoted.)

Logic Principles

In order to count with a base of 10 using two state storage devices connected to count in the 1248 binary stage is obtained from the complement output (instead of the true side) of the preceding bistable. Direction control gates are therefore needed to select the output used to trigger the following bistable. When counting down, another feedback gate is needed to sense the 1100 to 0100 transition, returning the four bistables to 0011.

Combining the four bistables with the direction-control and feedback gates leads to the logic diagram representing the basic decade shown in Fig. 1. Inspection of Table 1 shows that in the decimal 0-9 group of values of the 'excess-3' code only three inputs and three feedback paths are essential for each feedback gate, the fourth condition (dotted in Fig. 1) being redundant. There are, however, advantages in using the redundancies as mentioned below.

Merits of the 'excess-3' systems are that the only feed-

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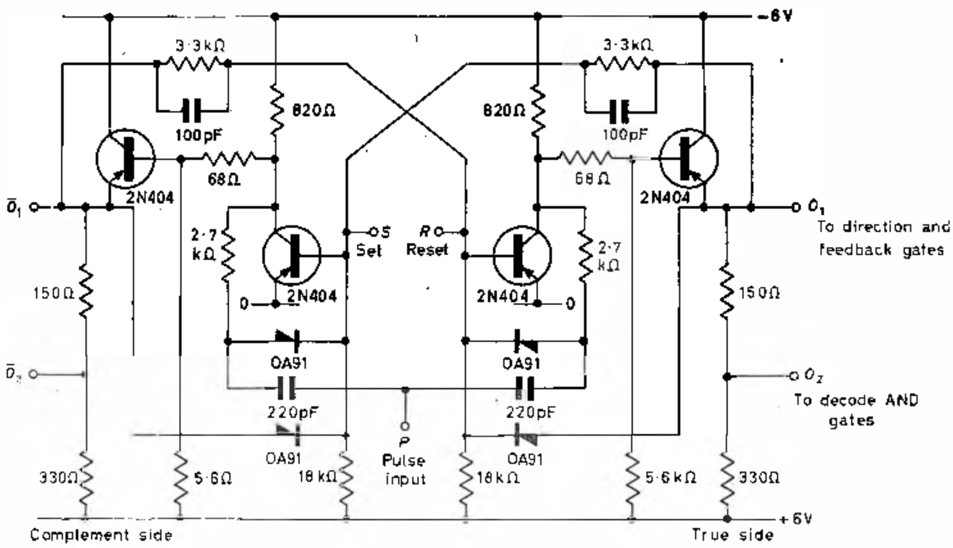

Fig. 2. Bistable used in the decade counter

back operation occurs when a decade carry pulse is transferring to the next stage, the counting operation is in an undisturbed 1248 binary code sequence and that the ten states are in 9's complement form. Four carries, however, occur in the 4 to 5 ('excess-3' code) transition limiting the maximum counting rates to less than that of the individual bistables. This objection may be overcome by using unit distance codes in which all transitions only require a single carry. (The outputs are not then in the required 1248 code sequence.) An example of the unit distance code counter has been reported by Farrel and Moore[2].

In the reversing system outlined above the reversing action must be maintained in a direction for sufficient time to allow decade carries to ripple through all stages when say 9...9 to 0...0 occurs. A reversal too soon will cause significant digits to be incorrectly stored. As will be described below, the minimum reversal time with this type of directional control can be reduced by a special technique to a time comparable with that needed to reverse one decade stage. (This time is limited by the duration of the return from 9 to 0 or 0 to 9 or by the duration of the carries needed in the 4 to 5 transition whichever is the longer. In the counter described these times were 300nsec and 500nsec, respectively.)

The counting state of the bistables can be decided with three input AND gates to yield output signals directly indicating the decimal count. A fourth output is redundant and can be omitted. The logic required for decoding the 'excess-3' binary code into decimal is given in Table 1.

The Bistables

A circuit diagram of the bistables used in the counter is given in Fig. 2. (This was based on a design already available within the department). As emitter-follower stages are usually needed to buffer the bistable outputs it is advantageous to include them internally, thus obtaining minimum switching times while retaining low output impedance. The bistable has been designed to use a minimum number of supply voltages; non-saturated performance being achieved without the use of a collector clamping bus. Driving a load of 30mA, a rise time of 100nsec and a fall time of 150nsec were measured using the 10 to 90 per cent change criterion. The maximum switching rate was about 2MHz being limited by the inherent time delays of the transistor used. Input pulse requirements were determined to be 80 to 500nsec (5V amplitude) and 3·2 to 8V (250nsec duration).

Two outputs are available from each side of the bistable.

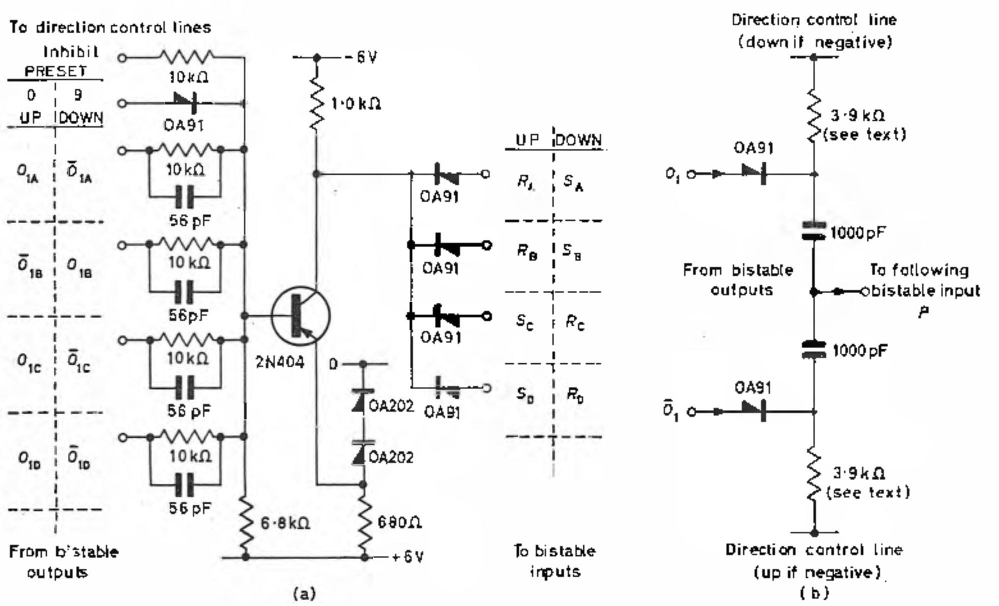

Fig. 3. (a) UP and DOWN NOR feedback gates

(b) Direction control interstage coupling gates

Output O_1 (and $\overline{O}_1$ from the complement side), used to operate the feedback gates and to generate carry pulses, switches from $+0{\cdot}5$V to $-4{\cdot}2$V. The emitter-follower resistor is tapped to obtain a swing of $\pm 2{\cdot}0$V needed to correctly switch the transistor driving the display lamp. In the logic given above a bistable output of $+0{\cdot}5$V (transistor ON) represents a '1' and output of $-4{\cdot}2$V (transistor OFF) represents a '0' in the true outputs.

Feedback Gates

The UP feedback gate must detect when a logical output of 1011 occurs (see Table 1) and return the *A*, *B*, *C* and *D* bistables to 1100, respectively. The DOWN gate must sense and recondition the stages in the minimum time. These functions are performed with NOR gates using resistor-transistor logic driving diode feedback paths. The circuits for both gates are identical, differing only in the connexions, as shown in Fig. 3(a).

Diode feedback is used to obtain maximum isolation and minimum response times. Forward biased silicon diodes are used to ensure that the feedback diodes are reversed-biased when the gate transistor is held ON by any of its inputs. The redundant states have been included to improve the response. Although the feedback gates are apparently independent, it was found that interference occurred in practice. A fifth input resistor was added to inhibit the action of the unwanted NOR gate when counting in the chosen direction.

Direction Control Gates

As already mentioned, the counting direction is controlled by using either the true or complement output to produce the carry pulse for the following bistable. An inter-bistable coupling network is shown in Fig. 3(b). The 3·9kΩ resistor is used to forward bias one diode when the direction control bus is made negative, the other bus being earthed to reverse-bias the unwanted diode path. Large value capacitors form an OR gate coupling the two outputs into a common input channel. At the higher pulse rates a minimum of -6V is needed to ensure that the diode forward resistance is sufficiently low so that it does not significantly attenuate the trigger pulse. To further improve the performance the 3·9kΩ resistor should be replaced with 1·8kΩ for the gates coupling the *A* and *B* bistables.

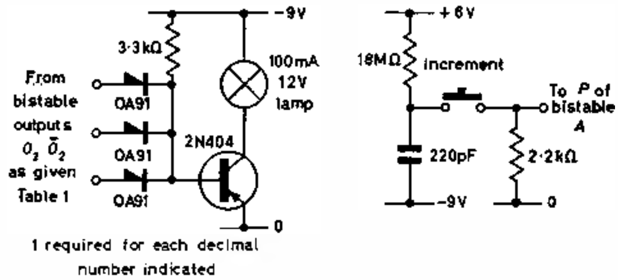

Fig. 4. (a) Decode and display lamp driver
(b) Increment pulse circuit incorporating time delay

With this type of direction control there is the possibility of generating incorrect triggering pulses when the control lines are reversed. Although other designs have exhibited this effect[3] a decade could be reliably reversed in 500nsec using a bistable to switch the direction-control lines.

The direction-control lines are also used to inhibit the unwanted feedback gates via the inhibit resistor provided (see Fig. 3(a)).

Decoding and Lamp Display

In order to obtain visual indication of the decimal state of each decade it is necessary to interpret the *A*, *B*, *C*, and *D* bistable outputs which are in 'excess-3' binary code. The most economical type of visual display is to use a separate lamp for each decimal state, a photographically produced graticule being used to physically form the required number. While this type of display is not ideal in its presentation it does not need high voltage d.c. supplies as do display tubes and, as only one lamp is illuminated in each decade at any time, the power supply requirements are minimized.

An extra transistor is required to drive the lamp of each decimal number from the available signals. The transistor, as shown in Fig. 4(a), is controlled by a three input diode AND gate operated from the bistable O_2, $\overline{O}_2$ outputs, connected directly as indicated in Table 1. To ensure that the transistor is maintained in a fully OFF state the inputs are derived from the tapped outputs of the bistables as seen in Fig. 2. A -9V bus is used to operate the 100mA, 12V lamps (Philips 8003D) at a reduced voltage.

TABLE 1
Logic states for reversible decades counting in 'Excess-3' code

BINARY CODE* 1248 WEIGHTS BISTABLES (TRUE SIDE)	DECIMAL VALUES CORRESPONDING TO 1248 CODE	DECIMAL VALUES CORRESPONDING TO EXCESS-3CODE	DECODING AND LOGIC CONVERTING EXCESS-3 TO DECIMAL EQUIVALENT
A B C D (L.S.D) (M.S.D)			
0 0 0 0	0	Not used	
1 0 0 0	1	Not used	
0 1 0 0	2	DOWN DETECTION STATE	
1 1 0 0	3	0	$\overline{A}.C.D$
0 0 1 0	4	1	$A.B.D$
1 0 1 0	5	2	$\overline{A}.B.D$
0 1 1 0	6	3	$A.\overline{B}.\overline{C}$
1 1 1 0	7	4	$\overline{A}.\overline{B}.\overline{C}$
0 0 0 1	8	5	$A.B.C$
1 0 0 1	9	6	$\overline{A}.B.C$
0 1 0 1	10	7	$A.\overline{B}.\overline{D}$
1 1 0 1	11	8	$\overline{A}.\overline{B}.\overline{D}$
0 0 1 1	12	9	$A.\overline{C}.\overline{D}$
1 0 1 1	13	UP DETECTION STATE	
0 1 1 1	14	Not used	

* Transistor ON represents a '1'.

Coupling the Decades

In order to form a multi-digit decade counter, identical decade stages, constructed as detailed by the circuits and connexions given above, are coupled in cascade. In unidirectional counters the pulse repetition rate decreases with the significance of the position allowing slower speed decades to be used. In reversible systems using the above outlined direction control method, however, all stages must be equally fast to minimize the time needed to propagate the decimal carries.

To reverse all decades simultaneously, a common direction control bistable was used to switch all decade stages, its circuit being identical with that given in Fig. 2 but in which $-9V$ is used instead of the $-6V$ bus to obtain an adequate negative level for the direction control lines. If more than six decades are to be reversed the bistable output would need buffering.

With the single bistable control the counter must be held in each direction for sufficient time to ripple through the decimal carries, a time approximately equal to the feedback gate cycle time (500nsec) multiplied by the number of stages used. In the six decade counter constructed, this time was measured at 3·8μsec. To obtain a faster reversing time, each decade stage can be reversed by a separate reversing bistable controlled by the reversing bistable of the preceding decade. A time delay, just less than the 500nsec needed, should be interposed between successive direction control bistables. With such an arrangement the reversing time for a six decade counter should be reducible to about 1μsec.

The direction control gates at the output of the decade must be arranged to obtain positive pulses at the 0 to 9 and 9 to 0 transitions in order to advance the following stage at the correct count.

Two pulse inputs are used, one channel for UP pulses the other for DOWN pulses. These are used to set or reset the reverse control bistable into the appropriate state. Both pulse channels are also combined into a common count input with an OR gate.

Facilities

The conventional method to preset the counter stages to 0 is to drive all bistables into the correct state with preset resistors from a negative bus. When many stages are to be driven in this manner the impedance of the many parallel paths becomes too low and diodes are needed instead to provide isolation. A simpler solution used in this counter was to actuate the UP feedback gate by driving its transistor OFF with a single diode as shown in Fig. 3(b). This allows preset to 0 and also to 9 by using the DOWN gate instead.

Preset to other than 0 or 9 values would need complex switching to set the four bistables into the required 'excess-3' binary code state. A compromise solution has been incorporated whereby each decade may be individually incremented one count at a time with its own push-button. To ensure that contact bounce does not cause multiple triggering a 4msec time-constant circuit has been used as shown in Fig. 4(b).

Anti-coincident circuits, such as those mentioned by Tassinari[3], have not been included, being unnecessary in the digital shaft position application for which the counter was primarily designed.

REFERENCES

1. TASSINARI, A. F. A Bi-directional Decade Counter with Anti-coincident Network and Digital Display. *Electronic Engng. 36*, 73 (1964).
2. FARRELL, R. L., MOORE, G. E. An Automatic Co-ordinate Measuring Equipment Using Moiré Fringes. *Electronic Engng. 37*, 644 (1965).
3. SCOLLAR, I. An Economical Reversible Transistor Decade Counter. *Electronic Engng. 33*, 597 (1961).

A Small British Satellite Launcher

An order covering the construction and supply of three Black Arrow 3-stage satellite launchers has been received by Westland Aircraft from the Ministry of Technology, Space Department, and involves a programme which had its inception in 1964.

Black Arrow has been designed as a technological satellite launcher and is a direct development from the successful Black Knight high altitude experimental rocket. Its primary purpose is to develop and space test new satellite components which will contribute to the communications programmes, and in addition, be used as a tool for advanced space research of all types.

In addition to developing British Space Technology in all types of satellite equipment, it will also improve British Industry's capability to compete in the European Space Markets. The Black Arrow Launcher may well be suitable as a launch vehicle for some of the future Esro series of satellites, and may eventually be considered in conjunction with the large Eldo launch vehicle. The programme is in two distinct phases. In the development phase three vehicles are being launched. The first of these, which is programmed for mid-1968, will provide information on the performance of both the first and second stages, together with the separation of a dummy third stage. The remaining two vehicles will be full orbital attempts with satellites containing instrumentation aimed primarily at giving information on the performance of the complete launch vehicle, and particularly the behaviour of the third stage and of the satellite in orbit.

One prototype has already been built and is being used to determine the resonance characteristics of the complete vehicle. These tests will be completed in time for the whole vehicle to be displayed at the Paris Air Show. The first complete flight vehicle is now in an advanced stage of construction and will commence tests at the High Down Rocket Test Site at The Needles on the Isle of Wight, in the middle of next year. The design of the vehicle which will make the first orbital attempt has passed the 75 per cent completion point.

When the first phase involving the three development vehicles has been completed, Black Arrow will be used at the rate of approximately one firing per year to put into orbit a series of so-called utilization satellites, the first of which is aimed at developing and space testing the fundamental equipment that will be needed in all future experiments.

Although Black Arrow is really just the launcher vehicle, this name is now being used to indicate the whole programme involving both the development of the launcher vehicle and the development of the satellites. It is now a Ministry of Technology programme in which the Royal Aircraft Establishment, Farnborough, hold the research, development and design authority and co-ordinate the work of the main contractors. Westland Aircraft is the contractor responsible for the design, manufacture and testing of the first and second stages. Bristol Siddeley Engines provide the liquid propellant Gamma engines, which are a direct development of the Black Knight engines, for both the first and second stages. The solid propellant third stage is being developed jointly by the Rocket Propulsion Establishment at Westcott and Bristol Aerojet.

The final assembly and launching of the complete 3-stage vehicles will be carried out at the Rocket Establishment, Woomera, Australia.

Simplified Information Theory and Data Transmission

(Part 2)

By A. E. Cawkell*, C.Eng., M.I.E.R.E., Sen.M.I.E.E.E.

(Voir page 275 pour le résumé en français: Zusammenfassung in deutscher Sprache auf Seite 276)

Timing; Noise; Synchronous and Asynchronous Systems

Suppose, in a two state system, that one state is transmitted continuously for 80msec to indicate 10 1's. Unless the receiver knows the time allocated for a single 1 it cannot tell that the 80msec interval is made up of 10 unit time intervals. If a clock is switched on and off by the signal transitions, and generates 8msec timing spikes while it is on, 10 spikes, corresponding to the 10 1's will be generated. Synchronous timing information of this kind must be present at the receiver in all complete data transmission systems, to provide advance information about the bit period.

With 'synchronous' systems, the timing information may be provided in various ways. One method is to use the transitions between 0's and 1's to establish the unit time interval. The transitions are used to control the phase of a local oscillator which produces a waveform to strobe the demodulated signals; alternatively the transitions may be used to generate a timing waveform directly which is correctly phase shifted to sample the centre of the binary waveform elements.

In an alternative arrangement, the 'start-stop' system, special timing signals—for instance signals having a unique level—are interspersed with the information signals, and these are used to carry timing information. In this case there need not be a known fixed interval between groups of signals as the special signal indicates that a group is about to be transmitted. However as these special signals are relatively infrequent, as opposed to the frequent timing signals in the synchronous system, the start-stop system timing is more likely to be confounded by noise.

If it is desired to have a versatile data link which may be used to transmit signals at any rate dictated by the capacity of an available channel, an asynchronous link is very suitable. The timing information must then be supplied by associated equipment, according to the modulation rate selected, in order to make the system complete.

Timing is of importance in another respect. It was shown in Fig. 1(b), that only at the instant of peak of a signal waveform is the waveform free of interfering components from adjacent waveforms in time, if signalling is being carried out at the Nyquist rate. It may be advantageous to arrange for the receiver to 'look' at the waveform at this instant. At this instant also, the receiver has the best chance of distinguishing the intended signal from the noise. In Fig. 8, the maximum amplitude of a signal state is at t_1 and t_2. At these instants noise peaks of opposite phase to signal peaks may have an amplitude of up to 0·5 of the peak-to-peak signal value before a 1 will be altered to a 0 (or vice versa) by the noise. In Fig. 8(b) with correct timing a noise peak has just not caused an error, while in Fig. 8(c) with incorrect timing the same noise peak has made a 0 look like a 1 to the receiver.

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In a practical system with nearly perfect timing, the number of errors will be very much greater than the 'arbitrarily small number' when information is sent at the rate C for various reasons, including the unrealizability of 'ideal coding'. Error rates achieved in a well implemented practical system with Gaussian noise are shown in Fig. 9. Other forms of noise are likely to be present which will further increase the number of errors.

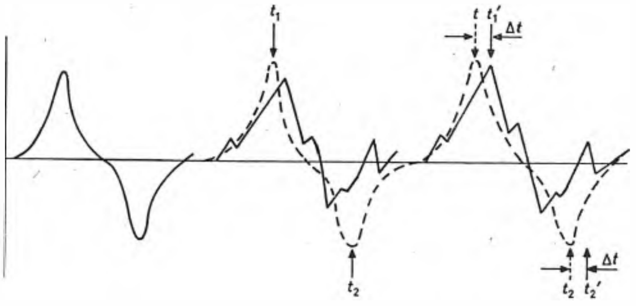


Fig. 8. Error caused by noise (a) transmitted signal, (b) Received signal with noise; receiver 'looks' at instant t_1, (c) the same signal with the same noise, but receiver 'looks' at instant t_1' and interprets a '0' as a '1'

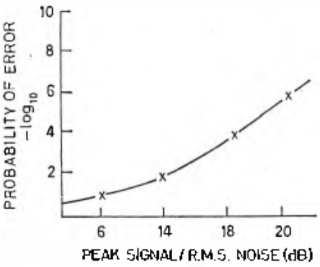


Fig. 9. Binary signalling error probability

Modulation

Discrete or analogue signals may be transmitted along a channel without further processing if the channel is suitable. For instance, 'mark-space' signals may be directly impressed on a pair of wires connecting a transmitter and receiver spaced a short distance apart, and at low signalling rates the signals will not be much affected by the channel. A communication system of this kind is called a baseband system.

Typical baseband signals (before shaping) are shown in Fig. 10. For direct current applied to the channel and a random signal sequence, the d.c. component of the on-off signal represents a power equal to the average power of the polar signal. The total average power in the on-off signal is the sum of the d.c. and polar components which is twice the average power of the polar signal. The polar signal, therefore, provides a 3dB signal/noise advantage, as for the same error rate 3dB less signal power is required.

Very often a channel is unsuitable for conveying signals in this form; for instance telephone lines will not normally pass signal frequencies below about 250Hz or above 3 500Hz. Frequency or time multiplexing may be required in order to send a number of messages simultaneously along the same channel. In such cases the electrical signals must be processed into a suitable form for transmission, using some form of modulation.

Modulation principles and methods are described in many textbooks and require rather lengthy treatment. Only facets relevant to data transmission will be described here.

An alternating sinusoidal current imposed upon a communication channel may be described by:

$$I = A \sin (2\pi f t + \phi) \quad \ldots\ldots\ldots\ldots (13)$$

where A = peak amplitude

f = frequency

t = time

ϕ = phase

Information may be conveyed by variation of any of the parameters A, f, or ϕ.

The waveforms generated in such variations are shown in Fig. 11. Symbols must be recovered at the receiver after demodulation in the face of channel attenuation, distortion, and noise. The modulation method chosen will depend upon the equipment complexity, the types of channel impairment, and the channel capacity and desired error rate.

The considerations about information rate, inter-symbol interference, channel capacity, and waveform sampling already discussed apply to modulated carrier waves, although this may not be easy to show. For instance modulation may be considered like baseband signalling if after the processing of the baseband spectrum into upper and lower sidebands it is assumed to be detected in such a manner that each sideband is restored to its original position in the baseband. Phase modulated signals may be regarded as suppressed carrier a.m. signals. Direct application of the above considerations to f.m. is more difficult, although some approximations have been given[6].

For comparison purposes it is convenient to consider the performance of modulated carrier systems in terms of two signal states, when each state carries one bit of information. The term 'baud' is often used in this connexion. The baud is the unit of modulation rate, which may or may not be related to information content. For binary signalling with an information content of 1 bit per modulated signal state, the bit/sec and baud/sec rates are the same. If there are n bits per state as in multi-level signalling ($n \geqslant 2$) the speed in bauds is still the number of states per second without regard to n.

Table 1 shows the signal-to-noise ratio necessary for n bits/sec per Hz signalling with various modulation systems and 10^{-4} error rate, where n is 1 bit per signal state for two state signalling if feasible with that system, or 1 bit per two signal states (that is per Hz) if 1 bit per state is not feasible, as indicated. The figures given apply to perfectly implemented systems with optimum signal shaping and noise filtering.

P.M. and f.m. systems are usually more effective in discriminating against noise than a.m. systems. The choice requires a consideration of the kind of channel impair-

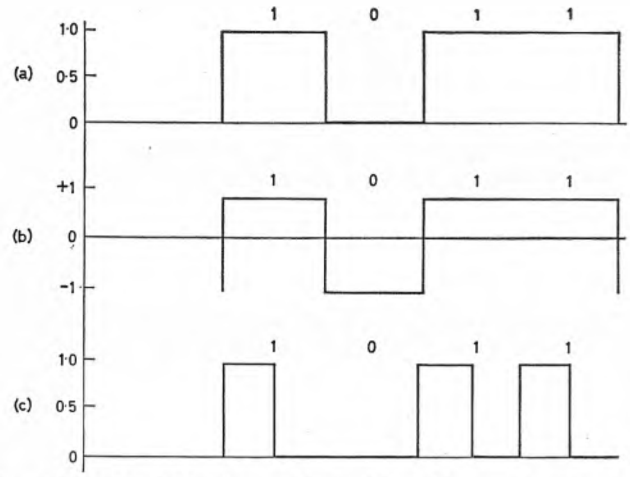


Fig. 10. Baseband signalling waveforms (a) on-off, non-return to zero (N.R.Z.), (b) polar, non-return to zero, (c) on-off, return to zero

Fig. 11. Modulation waveforms

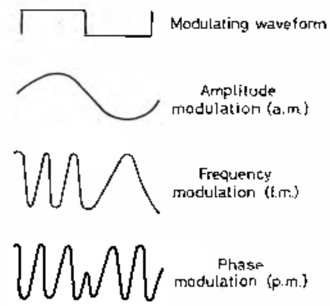


TABLE 1

Comparison of Modulation Systems[9]

TYPE OF SYSTEM	SPEED, BITS PER SECOND, PER HZ OF BANDWIDTH	SIGNAL-TO-NOISE RATIO FOR 10^{-4} BIT ERROR RATE: AVERAGE SIGNAL TO AVERAGE NOISE (GAUSSIAN) POWER	SIGNAL-TO-NOISE RATIO FOR 10^{-4} BIT ERROR RATE: AVERAGE SIGNAL TO AVERAGE NOISE (GAUSSIAN) POWER, CORRECTED FOR 2BIT/SEC/PER HZ. FIXED SPEED BASIS
Shannon..	2	5	5
On-off baseband ..	2	14·4	14·4
Polar baseband ..	2	11·4	11·4
A.M. coherent (synchronous) detection	1	11·4	14·4
A.M. suppressed carrier coherent detection ..	1	8·4	11·4
P.M. coherent detection	1	8·4	11·4
P.M. differential detection	1	9·3	12·3
F.M.	1	11·7	14·3

NOTE—2 : 1 in bits per second per cycle bandwidth is equivalent to 3dB S/N. For example, for comparing f.m. with polar baseband on a fixed speed basis, 3dB must be added to f.m. S/N.

ments, the method of detection and synchronization to be used, and equipment complexity and cost, according to which arrangement is likely to provide the highest efficiency in particular circumstances. It has been shown that for certain kinds of delay distortion f.m. is to be preferred while a.m. is better if other kinds of delay distortion are present[7]. A.M. is most affected by channel level variations, but is least affected by small changes of carrier frequency caused in the channel. F.M. or p.m. with differential phase detection, are least affected by phase jumps.

In order that the interfaces of communication equipment may be compatible, international recommendations have been made by the C.C.I.T.T., and systems are available which comply from various sources. For telephone line signalling up to 1 200 bauds, frequency modulation has been adopted, not without some controversy[8]. Further recommendations are awaited for signalling at high speeds.

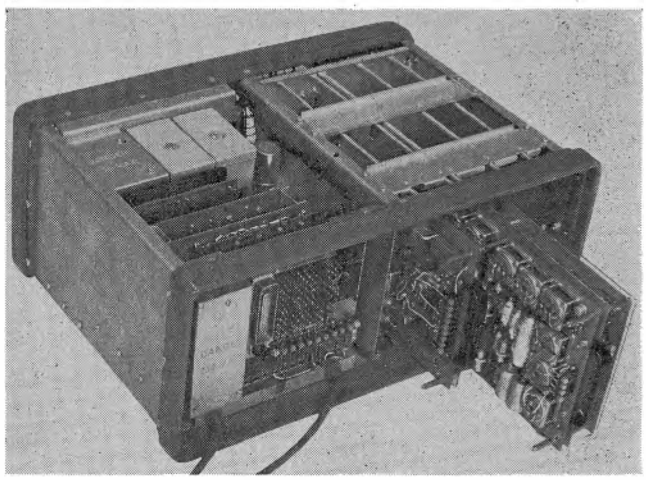


Fig. 12. Post office 1A modem (photograph by permission of the P.O.E.E. Journal)

General Considerations in Telephone Line Transmission

In order to describe data transmission systems in any detail in the space available it will be necessary to restrict the discussion to those using telephone lines as the communication channel. However, telephone lines have the widest distribution and probably the greatest potential; a much greater use of them for data transmission will be seen in the near future.

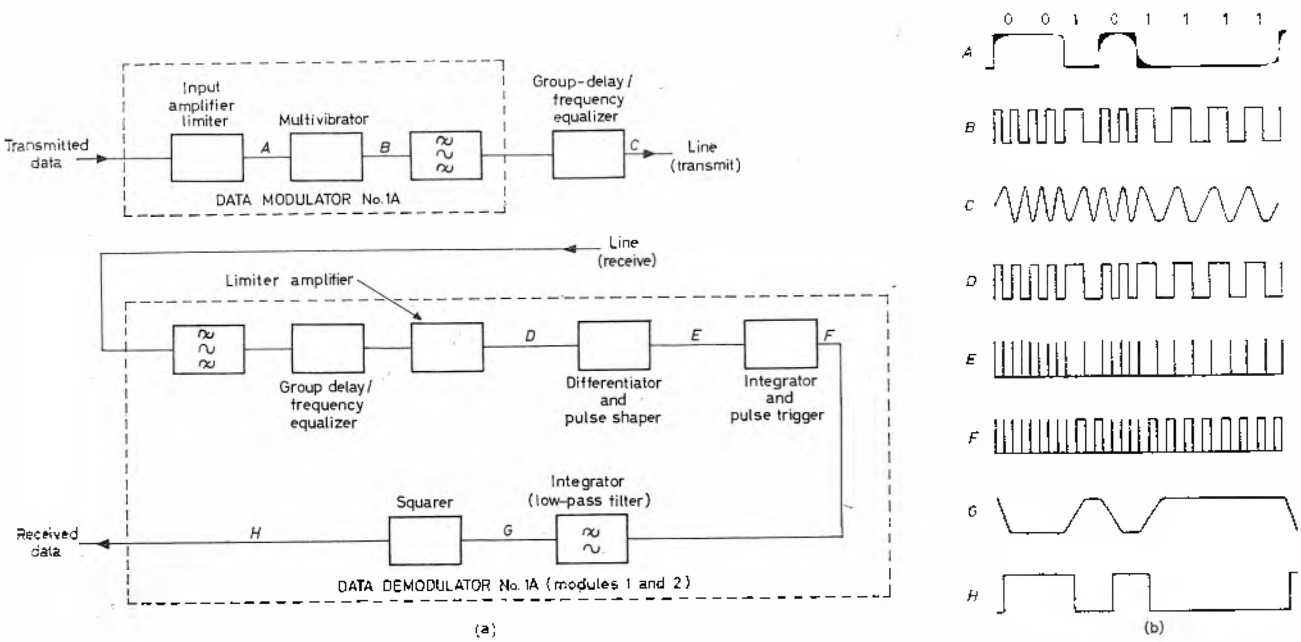
A telephone line suitable for speech of good intelligibility and fair quality should have a bandwidth of about 300 to 3000Hz, within which band most of the energy in speech is concentrated. Assuming a signal-to-noise ratio of 30dB, the channel capacity (from equation (7)) would be $2700 \log_2 (1\text{-}1000) = 27\,000$ bits/sec However the most readily available lines—those in the switched public network—may have, in extreme cases, an attenuation of 50dB at 2000Hz, and 30dB at 800Hz. The transmission delay in the pass-band is not constant, and impulsive noise is present. There may also be crosstalk, line interruptions, and an effect which causes frequency excursions of the signal spectrum. These impairments, together with reasons already discussed for the non-realization of the theoretical channel capacity, limit the actual channel capacity to about 1200 bits/sec. on switched lines, and to 600 bits/sec, in extreme cases.

As mentioned in the previous section, frequency modulation has been recommended for international data transmission by the C.C.I.T.T., for speeds up to 1200 bits/sec. The recommendation, V23 1964, includes the alternatives of synchronous or asynchronous transmission, and an optional return path for signalling at up to 75 bits/sec. Frequencies, power levels, and other matters relating to the interface, supervisory signals, and so on, are also specified.

British Post Office Datel Equipment

The Post Office can provide Datel modem No. 1 in various versions for data transmission/reception, at an annual rental of £120 to £150. Switched line call charges are the same as STD voice call charges, and the signalling rate is up to 600 or 1200 bits/sec, depending on conditions. Private line tariff S charges vary from £168 p.a., for a 10 mile radius, up to £2200 for a 200 mile radius; up to 1200 bits/sec is normally obtained on these lines.

Fig. 13. Post office 1A modem; block diagram and waveform (by permission of the P.O.E.E. Journal)



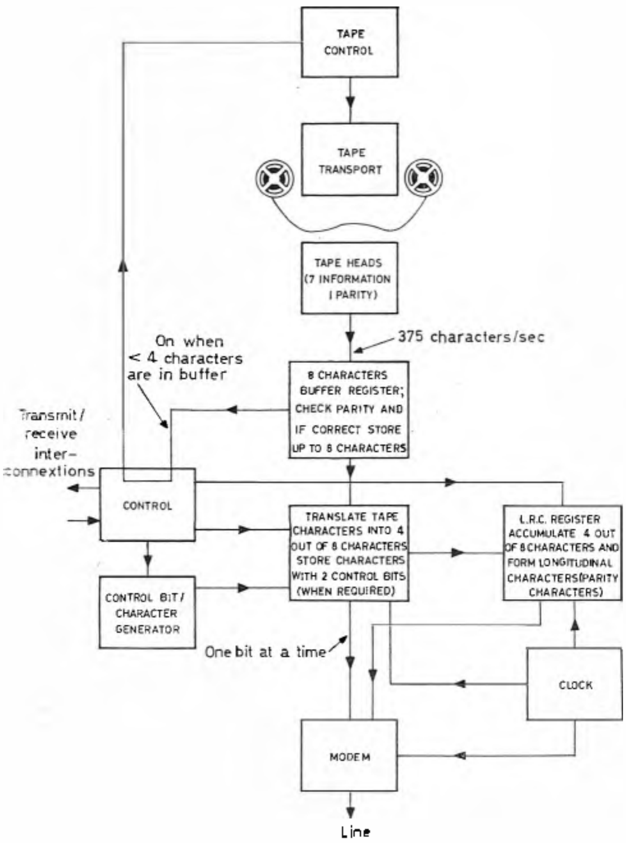


Fig. 14. IBM 7702 transmitter

International calls are being introduced, and the charge to the United States is 28s. per minute (switched line).

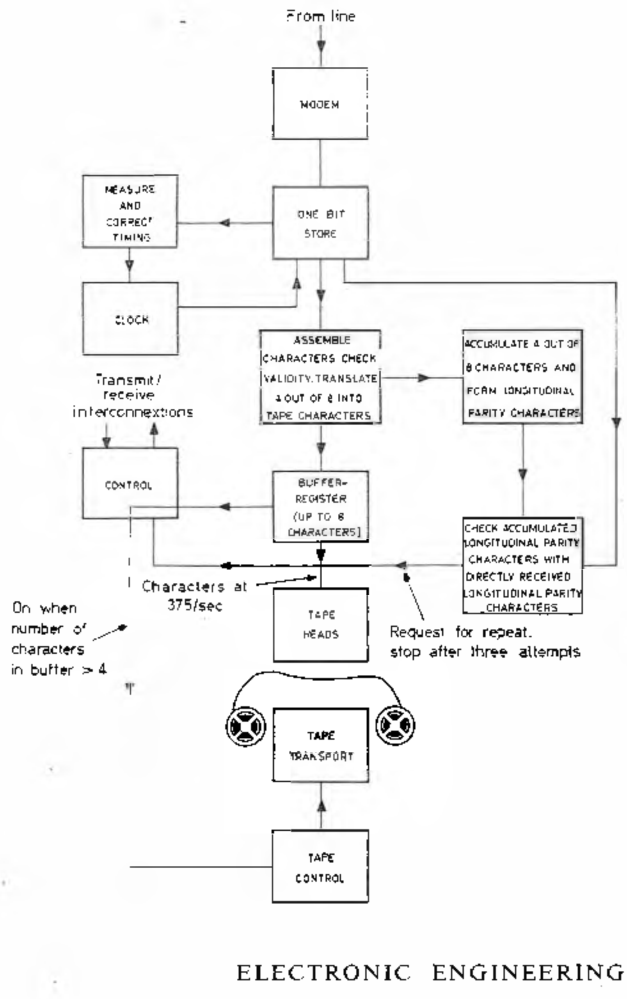
A photograph of the modem is shown in Fig. 12, and a block diagram in Fig. 13.

It will be remembered that for signalling at the Nyquist rate, and for good discrimination against noise, various criteria have to be met in respect of filter performance, timing, and shape of waveforms. In the practical case of the 1A modem operation, matters relating to bandwidth restrictions, equipment versatility, and the requirement for working into channels whose characteristics may be widely different from one installation to another, largely override these considerations. For instance, the transmitting filter must attenuate sharply below 900Hz because trunk line signalling is carried on below this frequency. The receiving filter must be designed to minimize crosstalk from the return channel, as well as to reduce other forms of noise. Filter characteristics dictated by these requirements introduces severe low frequency group delay distortion which makes equalization necessary. Very little freedom is left for other considerations such as the rather sophisticated requirements of implementing optimum waveforms for f.m. signalling[20].

The transmitter in modem No. 1A[11] embodies a voltage controlled multivibrator. A voltage between −3 and −25 causes a 1300Hz carrier to be generated corresponding to a '1', and a +3 to +25 voltage causes 1700Hz to be generated for signalling at up to 600 bits/sec, or 2100Hz for signalling at up to 1200 bits/sec, corresponding to a '0'. A method of shifting frequency by the alteration of control bias, as opposed to the switching between two oscillators, or between elements such as capacitors in a single oscillator, ensures phase coherence and a gradual transition between states. The system is asynchronous, and a 'zero crossing detector' is used at the receiver. The baseband signal is derived from crossings (transitions) of the carrier which occur at the rate $2f$, which is 2600 per second for the duration of a '1' waveform, and 4200 per second (at up to the 1200 bits/sec, rate) for the duration of a '0' waveform. The receiver needs no prior knowledge of the bit signalling element duration or of the character code, so the modem may be fed with serial data from a variety of types of communication equipment. This equipment must supply timing information to make the system complete. It will be noted from Fig. 13 that the carrier is squared and differentiated for the purpose of triggering a circuit whose output is a series of 'long' pulses when the repetition rate is 2600 per second, and a series of 'short' pulses when it is 4200 per second. The low-pass filter which follows cuts off at a frequency above that of the major components of the baseband waveform, but below the lower sideband of the signalling waveform, and a baseband signal of suitable d.c. levels is recovered.

With this type of circuit working at these frequencies and bit rates, operation at 1200 bit/sec, will cause a 1300Hz carrier to be transmitted having a period of 770μsec, that is 385μsec between transitions. The modulating waveform has a duration of 830μsec, and as the carrier is not synchronized with the modulation, either two or three transitions may be generated during the bit waveform. Jitter is likely to be present because the triggering impulse may arrive early or late depending on the signal/noise relationship at the triggering instant. In the two transition case this may give rise to errors, as the opportunity for the receiver to decide which symbol is intended is at its smallest. The introduction of timing in-

Fig. 15. IBM 7702 receiver



formation by the associated communication equipment may bring about an improvement; for instance, if the bit waveform is sampled at dead centre, the margin of error against noise is greatest although the improvement in the two transition case is likely to be small.

Error tests have been reported[12] using real typical channels and 1A modems. In 31 tests of 31 different telephone line connexions, the error rate was less than 1 in 10^4 in 63 per cent of the tests, at the 1200 bit/sec rate. 90 per cent of 100 bit blocks in error had 3 or less errors per block during tests at the 1000 bit/sec rate.

IBM 7702 Magnetic Tape Transmission Terminal and 3977 Modem

These equipments, when used together, are a good example of a current data communication system, and exemplify a number of the points made in the theoretical section of this article. They are, of course, one of the many forms of computer communications—a fast growing field using a telephone line as the communication channel.

Error detection is provided by the initial parity check of the tape code, and by the use of the two co-ordinate 4-out-of-8 code (fixed weight p out of q code) for error detection at the receiver. This is a combination of two of the methods described previously. 64 out of the available 4-out-of-8 characters are allocated to the symbols encoded on the tape, while the remaining 6 are allocated meanings for control purposes such as 'start', 'end of record', 'acknowledge', 'idle', and so on. The 'idle' signal is sent at the beginning of a transmission and during gaps between records in order to establish and maintain synchronism at the receiver. Additionally each message has an odd-even count at the receiver, so that it is not possible to 'loose' a message (for instance by the incorrect intervention of a human operator).

At the receiver, a number of mirror image operations take place as will be evident from the figure, together with certain special operations. A character error, or a longitudinal parity character comparison which is incorrect, initiates a 'repeat' signal from receiver to transmitter at the end of the record. Both tapes are then backstepped, and the record is repeated. If after three re-transmissions the record is still received incorrectly, the alarm is given at the transmitter.

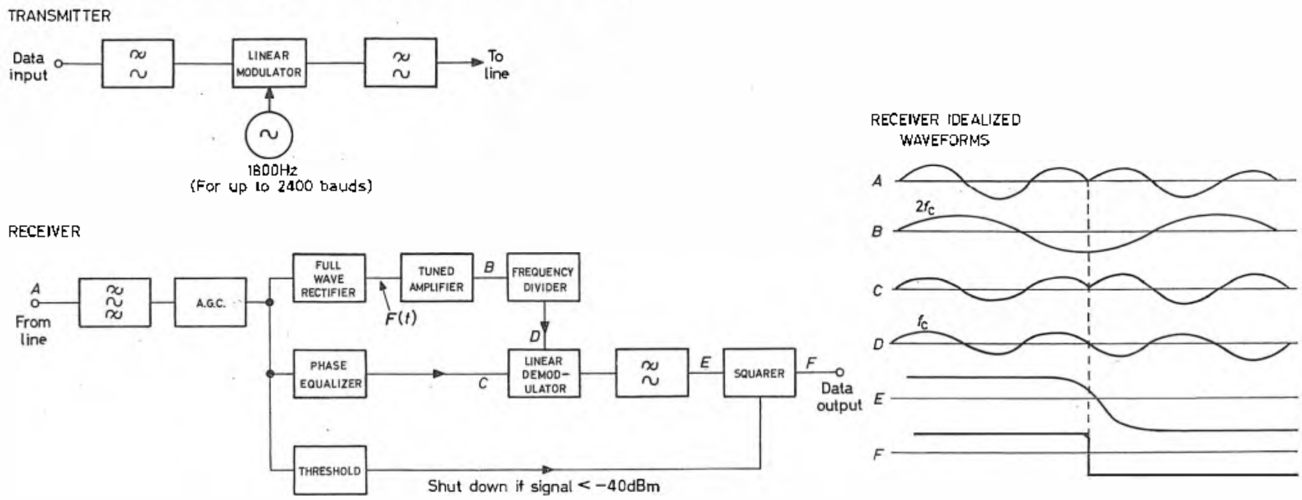


Fig. 16. IBM Modem type 3977

Figs. 14 and 15 show the transmit and receive functions diagrammatically for the 7702. A block diagram of the modem is shown in Fig. 16.

For transmission of coded information derived from information stored on tape, the 7702 and 3977 may work at up to 2400 bits/sec, if line conditions permit.

7702 operations are supervised by a control unit which commands the tape transport to deliver sequences of 7 information and 1 parity bits in parallel (= 1 character) to the read heads at the rate of 375 characters/second. The operations, which may be followed through on the block diagram, are as follows. Check tape data code character (binary or b.c.d.) for validity by reference to the parity bit; regulate the supply of characters so that when the transmission rate is slower than the tape delivery rate, more than 4 characters in the 8 character buffer-store causes the tape transport to be switched off until the number stored is less than four; translate data code into 4-out-of-8 code; accumulate 4-out-of-8 characters to form longitudinal parity character; shift characters serially bit by bit to the modem for transmission, and shift accumulated longitudinal parity character from the *LRC* register to the modem at the end of a record.

Timing is under the control of a crystal controlled clock which synchronizes all operations, and from which timing information is conveyed to the remote receiver.

The clocks consist of circuits which count down from a crystal oscillator. At the receiver each bit is mid-point sampled by reference to transitions in the received signal, and to receiver clock pulses. Any timing shift necessary to maintain mid-point sampling causes a command to be passed to a variable counter in the clock to advance or retard; this is a continuous process. The receiver clock stability is adequate to hold the timing for 10sec without receipt of a transition from the transmitter. The problem of channel introduced phase shift will be discussed later. The above method is inherently jitter-free, and long noise bursts will not upset synchronization.

The net transmission rate depends on channel characteristics, length of records (normally between 300 and 3000 characters), number of re-transmissions required, and delays in receiver turn-around with half-duplex working. Either duplex (four wire) or half-duplex (two wire) working may be used. With a complete system comprising transmitting equipment, receiving equipment, modems, and channel, it is extremely unlikely that an undetected error will reach the receiver tape when the system is working normally.

The 3977 modem (Fig. 17) is designed for the serial transmission of binary information using phase modulation. Each transition of the modulating waveform causes a phase reversal of the carrier. This mode of operation overcomes malfunctioning caused by sudden and variable phase-shift of the carrier, as can occur in the telephone line. It is unlikely that the line could contribute a phase-shift of such magnitude and rapidity that it could be confused with the modulation 180° phase-shift.

It is of interest to consider the two principle methods of phase detection[13]. In a phase demodulator it is necessary to compare the phase modulated data signal with a reference sine wave which must be in phase with one or other of the phase states of the signal. This may be done in a differentially-coherent detector, in which the signal waveform is applied directly to the demodulator, while the reference sine wave is the signal waveform applied to the demodulator, after being delayed by a time equivalent to the duration of one binary element in a delay line[14]. Thus each binary signal element is compared with its predecessor, with which it will be more nearly in phase than out of phase, or vice versa, and from this information a decision may be made that a phase reversal has or has not occurred. One disadvantage is that the reference sine wave, being also the signal, may be noisy or distorted; additionally there is a degree of inflexibility, which may give rise to a degradation of performance, arising from the relationship of the delay, the carrier, and the modulation rate; for instance, if the carrier is not synchronized to the modulating waveform (as is desirable if the modem is to be suitable for use with various types of communication equipment), the margin of error against noise is reduced.

The advantages are that a wrong decision due to channel phase-shift is unlikely, because for this to happen there would have to be a phase-shift of greater than 90° between elements—an unlikely occurrence; a further advantage with this method is that correct phasing is re-established almost immediately following upon a break in transmission.

In the 3977 modem the demodulation is carried out using the alternative method—phase coherent detection. Theoretically this form of detection provides a 3dB margin of improvement against noise compared with the differentially coherent detector, because the reference sine wave is almost free from noise and distortion; moreover the disadvantages associated with a delay line and with restriction of application because of the desirability of synchronization between carrier and modulation, are absent. Certain disadvantages associated with phase coherent detection have been largely overcome in the 3977, as will be described.

The output waveform $F(t)$ from a full wave rectifier contains terms including a component at twice the carrier frequency; an amplifier is tuned to this component, from which the reference sine wave is obtained via a divide by two circuit.

The baseband component, present in the demodulator output waveform, is passed through a low-pass filter to remove higher frequency components. The waveform is then squared to provide a binary waveform closely resembling the original.

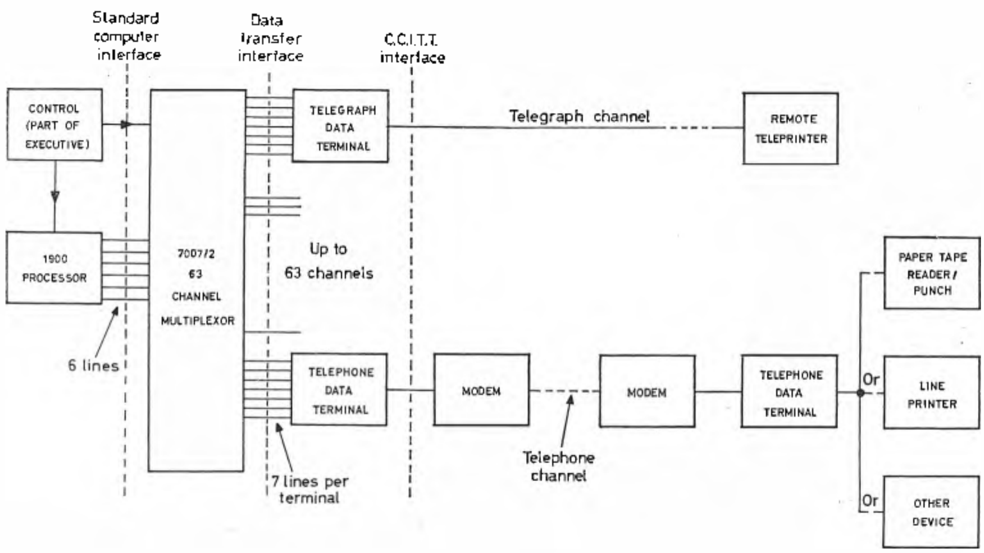


Fig. 17. ICT 1900 computer communication system

Flywheel effect in the tuned amplifier will maintain an output of the $2f_0$ component in the event of the component being momentarily absent at the input to the amplifier because of some adverse combination of signal elements.

Provision is made for phase equalization adjustment, but this usually only requires attention when signalling is being carried out at 2400 baud. Provision is also made for alteration of the carrier frequency according to the modulation rate: for up to 1200 baud signalling the carrier is set to 1500Hz, while for signalling at up to 2400 baud, the carrier frequency is normally set to 1800Hz; at this rate a bandwidth of from 300 to 3100Hz is required, so a private line must be used

A disadvantage of the method is that as the modulation rate approaches or exceeds the carrier frequency (the carrier is not synchronized with the modulating waveform) inter-modulation products are generated depending on the phase relationship between carrier and modulation. Careful filtering is required to remove these products from the transmitted frequency spectrum.

A further disadvantage lies in the fact that this method of deriving a pure reference carrier via a narrow band amplifier causes the response to rapid phase changes to be slower through this path than the response through the information path; the phasing between the reference sine wave, and the information signal at the modulator, may be unsuitable for correct operation in the event of a rapid phase shift being introduced in the channel.

It will be appreciated that this design requires a number of compromises, but consistently good results are achieved in practice, inclusive of 2 400 baud signalling through private lines.

I.C.T. Multiplexed Telephone/Telegraph Computer Communications

A large computer is an extremely useful device, but unless it can be put to work for most of the time, its high cost cannot be justified. However the computer may be made available to many users by means of multi-access equipment arranged in such a manner that each user apparently has exclusive and almost instantaneous communication with it; moreover users may be remotely situated and need possess only an appropriate peripheral unit connected to the computer by a telephone line along which information flows to and fro.

The information rate per user is limited by the line to, say, 2 400 bits/sec, but this is adequate for most purposes.

I.C.T. series 1900 computers may be used in this manner with the aid of a 7007/2 multiplexer and suitable data terminals and modems. Up to 63 remote users may be connected to the central processor, and the multi-access arrangements, providing apparently exclusive use, are automatically controlled by the executive supervisory system. The general arrangement is shown in Fig. 17.

To receive, serial information consisting of one character comprised of seven information and one parity bits is accumulated in a data terminal (telephone line channel). The multiplexer scanner, acting in the manner of a stepping switch, constantly scans the data terminals. When reaching a terminal in a 'ready to transfer' condition, the scan stops, the channel number is sent to the computer, and the core store address is determined. The character is then transferred, the address is incremented, and the scan carries on to the next data terminal. Conversion to I.C.T. six bit computer code together with the insertion of shift characters (upper case, lower case, etc.) may be carried out optionally at this stage, and the total time spent per channel during this operation plus time spent in scan-and-transfer, is about 70μsec. If all 63 data terminals were 'ready' and allowing a small margin, the total time taken for one complete scan would be about 5msec. All channels may therefore safely operate at 200 characters/sec, that is 1 600 bits/sec, and the rate at which information is then being fed into the computer is over 100 kilobits/sec. If however any channel is not ready when scanned, the scan time per channel is only 8μsec, and in a practical installation mixed telegraph and telephone channels may be used, rather than telephone channels only. Under these conditions the only speed limitation of telephone line data is likely to be the channel itself, so telephone channels may be working at, say, 2 400 bits/sec, while the telegraph channels are working at 50 baud.

To transmit, similar but oppositely directed events occur, and at any instant the computer will be receiving on some channels, and transmitting on others.

A message received from any channel is first transferred into a store having a capacity of 128 characters. Messages will normally be shorter than this and will end with a longitudinal parity character. If the message is found to be incorrect a request for repeat may be transmitted, otherwise it will be made available for processing. In certain circumstances the uncorrected message may be processed without correction, as desired. Messages longer than 128 characters may be handled by using an overflow area.

Special signals are used for control and indication purposes (for instance 'start of message,' or 'synchronizing group'), and each telephone data terminal and modem is individually clocked; the modems may be of any type which conform to the C.C.I.T.T. recommendations, such as Dataset 1A.

This description applies to only one version of the 1900 computer communication equipment, and various alternative equipments are available.

Future Developments

Improvements in data transmission systems can be expected in the near future, with computer communication equipment in the forefront, and becoming widely used.

As already mentioned, C.C.I.T.T. recommendations are expected shortly with regard to the method of modulation and other matters to be adopted for signalling at speeds at above 2 400 bits/sec. It seems likely that some form of phase modulation will be recommended, as although there is a p.m.-f.m. controversy about signalling at slower rates, there is some agreement about the adoption of p.m. for faster signalling. In addition to the phase reversal system already discussed, 4, 8, or more phase-shifts may be used, and the system becomes effectively a 'multi-level' system, as more than one bit of information per signal state may be conveyed. In quaternary phase modulation, for instance, the information rate is double that of phase reversal modulation; the coding may be $+45° = 11$, $-45° = 10$, $+135° = 01$, $-135° = 00$. It may be shown that as the number of phase-shifts is increased, the margin against noise decreases, as might be expected, but when comparing the noise discrimination for p.m. against a.m. as the number of states is doubled and re-doubled, the former shows a decided advantage. This advantage diminishes for a number of states exceeding 8.

Another system which has received some attention, and which may be further developed and used is the so-called 'duobinary' signalling system. Intuitively it can be visualized that if a system is adjusted for signalling with raised-cosine waveforms spaced by one Nyquist interval, signalling may also be carried on using half amplitude waveforms spaced at one half of a Nyquist interval, without inter-symbol interference. This waveform passes through amplitude values of 0 and 0·5 spaced at half an interval, as compared with the values 0 and 1, spaced at one interval for the full amplitude waveform. For the same power this gives rise to the possibility of three level signalling, namely, 0, 0·5, and 1 at double the Nyquist rate, with a suitably devised code; a form of bandwidth compression has been achieved. The theoretical error rate for random polar three level baseband signalling with an 18dB average signal power to average noise power ratio is 1 in 10^6. Although some reduction in the margin against noise might be expected for a system resembling a three level one, the same error rate with a similar signal-to-noise ratio has been claimed for duobinary signalling at double the signalling rate[15]. This is surprising, as such a result might be expected only if the signal-to-noise ratio was increased by at least 3dB.

It seems likely that greater attention will be paid to p.c.m. signalling in wideband lines in the future. These lines may consist of specially laid cables, or the bandwidth may be obtained by a further grouping of cables from the existing arrangements of 12 (48kHz), and supergroups and mastergroups, to still larger groupings, or to the use of repeaters and equalization at more frequent intervals, making possible rates of tens of megabits or more[16-18].

An interesting development known as 'adaptive equalization' has been described recently[19]; this could bring about considerable improvements, at least in cases where the cost of extra equipment can be justified. In the first experiments, speeds of 9 600 bits/sec were obtained in private lines by the manual adjustment of filters derived from an adjustment algorithm. Later an automatic system was used in which control signals were derived from the information signals, no special test signals being required. These signals were used to adjust a filter network in order to move the instant of sampling into the 'eye opening'. (An eye opening may be observed on an eye diagram which is a pattern of superimposed signals displayed on an oscilloscope synchronized with the bit rate.) As channel conditions deteriorated the system operated to re-adjust parameters for optimum performance. Error rate improvements of up to three orders of magnitude are claimed.

Finally mention must be made of the possibility of sending television pictures along telephone lines[20,21]; pictures can convey a great deal of information in a very acceptable form, and possible applications include the transmission of pictures of bank statements, drawings, photographs, tables of stock exchange prices or race results, to name but a few. Considerable work has been carried out in this field, and the indications are that systems already working could be made technically and commercially feasible if the necessary investment were made for further development and 'productionizing'. Possibly a range of techniques such as picture data reduction, bandwidth compression, use of special monitors, and special private lines would need to be combined to form the right compromises for the desired results.

Corrections

Several errors occurred in Part 1 of this article, they are: In the caption to Fig. 1(b), $t' = 1/280$ should read $t' = 1/2f_o$; in equation (11) the square-root sign should be deleted; in the caption to Fig. 6, loding should read coding; on p. 217 the 8th line below Fig. 7 should read, $q!/p!\ (q-p)! \ldots$; in reference 4 the volume number should be 29.

REFERENCES

6. Turin, G. L. Error Probabilities for Binary Symmetric Ideal Reception Through Slow Fading and Noise. *Proc. Inst. Radio Engrs. 46*, 1603 (1958).
7. Sunde, E. D. Pulse Transmission by A.M., F.M. and P.M. in the Presence of Phase Distortion. *Bell Syst. Tech. J. 40*, 353 (1961).
8. Clark, A. P. Considerations in the Choice of the Optimum Data Transmission System for use over Telephone Lines. *J. Brit. Instn. Radio Engrs. 23*, 331 (1962).
9. After Bennett and Davey (see Bibliog.).
10. Sunde, E. D. Ideal Binary Pulse Transmission by A.M. and F.M. *Bell Syst. Tech. J. 38*, 1357 (1959).
11. Allery, G. D., Smith, N. G. A Standard Modem for Data Transmission over the Telephone Network at 600/1 200 Bits Per Second. *P.O. Elect. Engng. J.* (July, 1966).
12. Allery, G. D., Roberts, L. W. Contribution Number 6. I.E.E. Colloquium on Data and Telegraph Transmission, 11th October, 1965.
13. Clark, A. P. *Loc. Cit.*
14. Hopner, E Phase Reversal Data Transmission System for Switched and Private Telephone Lines. *I.B.M. J. R. & D. 5*, 93 (April, 1961).
15. Lender, R. Faster Digital Communications with Duobinary Techniques. *Electronics. 36*, 61 (22 March, 1963).
16. Stevens, A. D. A P.C.M. System for Junction Telephone Cables. *Point to Point Telecommunications.* (Part 1) *10*, 2 (Feb., 1966). (Part 2) *10*, 3 (June, 1966).
17. James, R. T. High Speed Information Channels. *I.E.E.E. Spectrum*, 3, 4, 79 (1966).
18. Franklin, R. H., Law, H. B. Trends in Digital Communication by Wire. *I.E.E.E. Spectrum*. *3*, 11, 52 (1966).
19. Lucky, R. W. Techniques for Adaptive Equalization of Digital Communication Systems. *Bell Syst. Tech. J. 45*, 255 (1966).
20. Cherry, C., Kubba, M. H. An Experimental Study of the Possible Bandwidth Compression of Visual Image Signals. *Proc. Inst. Elect Electronic Engrs. 51*, 1507 (1963).
21. Cawkell, A. E. Cathode Ray Storage Devices; Television by Telephone Line. *Electronic Engng. 36*, 142 (1964).

London Educational Television

Two television studios are now complete and being used for the training of teachers in programme production techniques at the Inner London Education Authority's new television centre, expected to be the largest educational closed-circuit television system in Britain. It is being established at Laycock School in Islington, and will provide programmes for over 1300 schools and colleges in London by 1970.

The cameras and studio equipment in the first two studios have been supplied by The Marconi Company. These studios, now in full operation, are being used primarily for training teachers and making experimental programmes. The first transmissions will be made in September 1968 to 300 schools in North London. By then the work of adapting the building as a television centre will be completed with two additional studios.

The studio camera is the type V322B which is very simple to use and provides a high quality picture. For the ILEA scheme, the cameras have been provided with remote manual control as well as automatic control facilities, increasing the flexibility of operation in order to allow the equipment to be used when engineers are not available. In the larger of the first two studios, three V322Bs are employed, and two in the other studio, although either area can operate with the full complement of five cameras.

The V322B is one of a series of cameras, produced by The Marconi Company, which are self-contained, versatile units designed specifically for educational television, though they are equally suitable for many types of small studio installation. The basic camera is the V322A which is usually operated with a fixed lens; the V322B has an integrated viewfinder situated on top of this camera to provide the cameraman with a high-brightness 7in picture.

The V322B is equipped with a 4 lens turret or alternatively with a wide variety of zoom lenses. In the turret version the camera accepts four C type or broadcast-mount vidicon lenses with focal lengths of ½ to 4in or 1 to 6in. For the ILEA Television Centre one of the cameras has a 10:1 zoom lens with a continuously variable angle of acceptance of 50° to 5°.

The basic camera comprises a vidicon tube, scanning yoke and two printed boards carrying the video and scanning circuits including the power supplies. All this is contained in a compact rectangular metal case 15¼in by 6½in by 4¾in. The V322B version contains basically the same camera although power supplies for the whole unit, including the viewfinder, are housed in the viewfinder section of the equipment.

Most of the Marconi equipment in the control rooms, such as vision and audio mixing equipment, television monitors and control room/studio intercom systems, has been specially designed for educational studio applications.

In the Islington scheme, facilities are available to mix vision signals from up to six cameras and sound signals from up to 12 sources such as microphones, telecine machines and disk reproducers. A Marconi telecine bench is installed in the larger control room, providing for the televising of films, slide material and captions. Local and remote control facilities are provided in order that the equipment on the bench can be operated from either studio control desk.

Impulse Testing of Dynamic Systems

Using Cross-Correlation Techniques

By F. Jones*, Dip.El., M.I.E.E., A.F.R.Ae.S., A.M.I.E.R.E.

The essential principle of the cross-correlation method of measuring a system's impulse response is that if white noise is applied to a linear system the cross-correlation of the system input and output gives the system's impulse response.

The use of pseudo-random binary sequences, whose auto-correlation function approximates that of white noise, has eased the practical instrumentation difficulties previously encountered in delay, multiplication and integration.

A practical cross-correlator, the Sperry 'Inspectron', is described and the significance of the measured impulse response of second and higher order systems is related to step function and frequency response methods of testing.

(Voir page 341 pour le résumé en français: Zusammenfassung in deutscher Sprache auf Seite 343)

Statement of the Problem

Briefly it may be said that in system testing the general problem is to deduce from measured system data certain information concerning the dynamic characteristics of that system. What information is desired will depend on the depth of analysis required from the test. During the development of a system it is often necessary to check design concepts by detailed and comprehensive measurements in order to highlight the effect of design parameters on the overall system performance. There are many situations, however, such as pre-flight checks on aircraft and pre-launch tests on missiles, in which it is necessary only to determine that one or more system parameters are out of tolerance: the actual value of each parameter is not of interest. Basic to all forms of testing is the requirement that the system will produce a specified function of time $y(t)$ as its output when excited by a prescribed input function of time, $x(t)$. This often has to be determined by the interpretation of system behaviour in the real or complex frequency domains. Nevertheless, the correct time response is the design aim and information concerning this response should be the fundamental objective of any measurements made on dynamic systems.

What is chosen as the correct relationship between system output and input depends on the operational requirement and obviously varies from system to system but any specified relationship should ideally be related to this requirement. In complex systems, formulating test criteria often involves simultaneous solution of the 'system identification' equations defining the input-output characteristics of the system. Furthermore, for out-of-tolerance rectification a diagnostic procedure must be evolved which relates input/output anomalies with individual system parameters and hence allows the identification of faulty components or corrective adjustments to be made. This ideal test philosophy is often impractical when test instrumentation is considered and test procedures may have to be tailored to suit the available test hardware. Consequently, although a system's time response is desired the main test method is often one employing frequency response techniques.

The Significance of the Impulse Response

It can be appreciated then that knowledge of the system input/output relationship is fundamental to the testing problem, although in practical feedback systems parameter sensitivity is often deliberately reduced in the closed-loop configuration and this can detract from the amount of diagnostic information available from the input/output measurements. As two functions are concerned, the input $x(t)$ and the output $y(t)$, and since the system presumably has one function which relates the output to the input for any given set of conditions the first step is to obtain this describing function. From Laplace transform theory a

Fig. 1. Convolution and its Laplace equivalent

function $h(t)$ is known that relates the input and output of a system or network. It is called the impulse response of the system and its Laplace transform $H(S)$ is generally called the system transfer function. Unfortunately if one is working in the time domain the input and output are related through a finite integral known as the convolution integral. This means that in order to get the impulse response $h(t)$ it is necessary to solve an integral equation, viz.,

$$y(t) = \int_0^\infty h(t) \,.\, x(\tau - t)\, dt \quad .... \quad (1)$$

Convolution in the time domain and its Laplace equivalent are shown in Fig. 1. The convolution integral appearing in equation (1) provides a direct method of obtaining the output signal from a knowledge of the input signal and the impulse response of the system. In practical terms, convolution involves the operations of time delaying a folded input function, multiplication and integration.

If, instead of an arbitrary input function $x(t)$ which complicates the convolution process, a constant amplitude impulse of short time duration is used and this is swept (i.e. controlled delay) across the significant time period of the systems response then the system output y(t) will be proportional to the systems impulse response $h(t)$. The practical advantage of impulse testing is that it represents a disturbance applied to a system at some time, after which the system is allowed to settle back into its original state rather than to some new constant state, as is the case of a step function, or a steadily changing state as with a velocity or ramp function. Unfortunately an input impulse is not a very practical test signal and therefore sinusoidal, step and ramp functions have of necessity been used in the past for the testing of dynamic systems. Satura-

* *Sperry Gyroscope Co. Ltd.*

tion difficulties and correct distribution of spectral power in order to make the system responsive have limited the usefulness and application of impulse testing prior to the advent of correlation techniques employing digital methods. The use of random and pseudo-random binary sequences as input functions has eased the practical instrumentation difficulties previously encountered in delay, multiplication and integration.

Using the Impulse Response

When the impulse response of a linear system is known it is possible to find its response to any other driving function. Information concerning system parameters, their deviations and their effect on the system steady-state performance and transient behaviour can all be deduced from the measured impulse response of the system operating in its closed-loop configuration. As engineers are more familiar with frequency response techniques, root locus techniques and step function testing it is important to consider their inter-relationship with impulse testing. A step function response, for example, is the integral of the impulse response while the system response time can be related to the system bandwidth in the frequency domain. In the simple systems described in the following paragraphs this inter-relationship is noted where important.

Simple First-Order System

If an impulse $\delta(t)$ is applied as the input to the simple differentiating network shown in Fig. 2(a) the output time response is given by:

$$y(t) = \delta(t) - 1/CR \exp(-t/CR) \quad \ldots\ldots\ldots \quad (2)$$

A graph of this function is shown in Fig. 2(b). If this output waveform is integrated with respect to time then the unit step response, shown in Fig. 2(c), is obtained. If the input $x(t)$ had been a sinusoidal function then the asymptotic gain and phase characteristics as the input frequency (ω) is varied would be as illustrated in Fig. 2(d). Thus

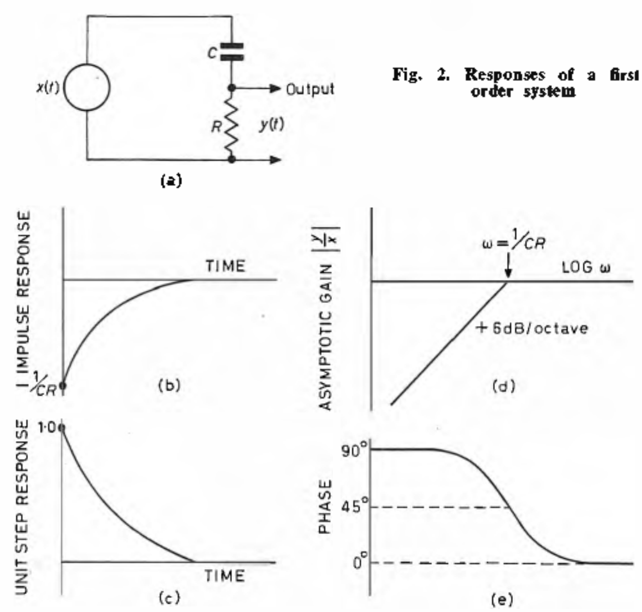


Fig. 2. Responses of a first order system

differentiation is achieved by working within the bandwidth of the network, i.e. below $\omega = 1/CR$ rad/sec, on a rising gain characteristic of +6dB/octave associated with a phase advance. The process of differentiation would cease for frequencies higher than the bandwidth of the network. As the time response bears an inverse relationship to the bandwidth i.e., decreasing the response time is equivalent to increasing the bandwidth, any restrictions of bandwidth immediately impose restrictions on the action of the network to certain time functions. This is generally true for physical systems. In this example the exponential component in the output response is the error introduced by the finite bandwidth of the network and reduction of this error can only be achieved at the expense of an increase in bandwidth.

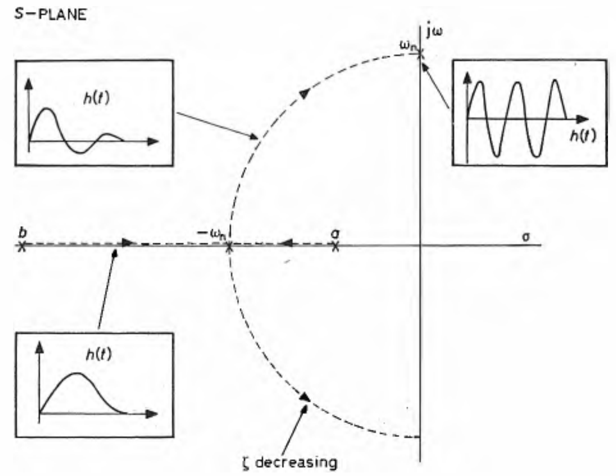


Fig. 3. Root locus of $\frac{1}{S^2/\omega_n^2 + 2\zeta S/\omega_n + 1}$ as ζ varies between 0° and ∞

Second-Order Systems

A second order system has a basic differential equation of motion in which the highest derivative of interest is the second. The dynamic characteristics of the system will be determined by the roots of this second-order characteristic equation. Many basic problems in system analysis can be understood by studying second-order systems and the lessons learned subsequently applied to higher order systems. Due to the simplicity of second-order systems it is possible to present performance characteristics as simple non-dimensional curves, thus allowing performance criteria to be interpreted directly in the form of a system transfer function or impulse response.

The impulse of a system whose transfer function can be approximated by:

$$H(S) = \frac{1}{(S^2/\omega_n^2) + (2\zeta/\omega_n)S + 1} \quad \ldots\ldots\ldots \quad (3)$$

Where ζ = system damping ratio and ω_n = natural undamped pulsatance is given by:

$\zeta > 1$ *(overdamped)*

$$h(t) = \omega_n/\sqrt{(\zeta^2 - 1)} \exp(-\zeta\omega_n t) \sinh \omega_n \sqrt{(\zeta^2 - 1)}\, t \quad \ldots \quad (4)$$

$\zeta = 1$ *(critically damped)*

$$h(t) = \omega_n^2\, t \exp(-\omega_n t) \quad \ldots\ldots\ldots\ldots \quad (5)$$

$\zeta < 1$ *(underdamped)*

$$h(t) = \omega_n/\sqrt{(1 - \zeta^2)} \exp(-\zeta\omega_n t) \sin \omega_n \sqrt{(1 - \zeta^2)}\, t \quad \ldots \quad (6)$$

As stated earlier, the dynamic characteristics of the system are determined by the position of the roots of its transfer function in the s-plane. Thus the path of the roots (root locus) as the damping ratio is varied indicates the change in relative stability of the system and thus the character of the impulse response. Fig. 3 shows the path of the roots as ζ is varied from infinity through to zero and illustrates the general change in character of the impulse response, $h(t)$.

When ζ is very large, i.e. the system is overdamped, there are two real roots, one close to the origin at $s = -a$ and the other well displaced from the origin at $s = -b$. The closer the root is to the imaginary axis the more it will influence the impulse response, thus the contribution of the root at $s = -b$ is an exponential which dies out relatively quickly while the root at $s = -a$ effectively determines the form of the impulse response. Reducing ζ causes the real roots to approach one another until at a value of $\zeta = 1$ (critical damping) they coalesce to give a double root at a point $s = -\omega_n$. For $\zeta < 1$ (underdamped system) the roots leave the real axis to become complex conjugate as ζ tends towards zero and they move on arcs of circles centre the origin, radius ω_n. This results in the impulse response becoming oscillatory at a damped natural pulsatance $\omega_d = \omega_n \sqrt{(1-\zeta^2)}$. In the limit when $\zeta = 0$ the roots are purely imaginary and the impulse response is one of constant amplitude oscillations at a natural pulsatance of ω_n rad/sec. Fig. 4, 5 and 6 show non-dimensionalized curves of the impulse response, unit step response and frequency response for a zero displacement error, second order system.

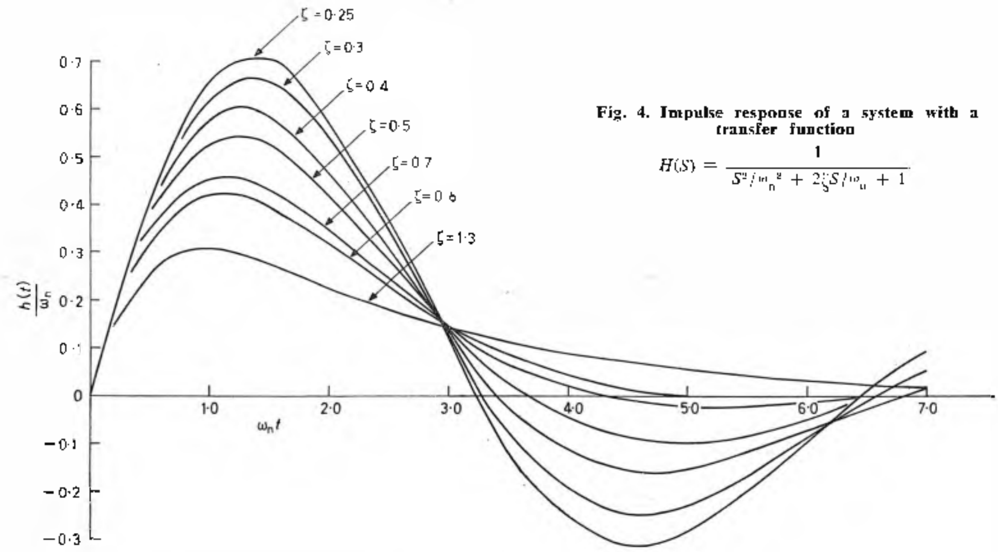


Fig. 4. Impulse response of a system with a transfer function

$$H(S) = \frac{1}{S^2/\omega_n^2 + 2\zeta S/\omega_n + 1}$$

It is often required that some criterion of how stable a system is, should be available in a convenient form. This really means having information about the damping ratio (ζ) of the system. A common criterion for this control of relative stability has been the percentage overshoot in response to a unit step-function input. If step function testing is already specified for a system check-out procedure, the position and magnitude of the peak overshoot and the rate of change of the step response can easily be obtained from the measured derivative function i.e. the impulse response.

With the advent of practical instrumentation capable of measuring the impulse response directly, the control of relative stability can be obtained in three ways by measurement of (a) the logarithmic decrement, (b) the impulse response area ratio and (c) the difference between positive and negative areas of the impulse response. These three criteria are illustrated in Fig. 7.

(a) *Logarithmic Decrement.* Defined as the natural logarithm of the ratio of amplitude of the impulse response measured one period apart. Referring to Fig. 7 this can be stated, for second-order systems, as:

$$\text{Log Decrement} = \ln f(t_1)/f(t_2) = 2\pi\zeta/\sqrt{(1-\zeta^2)} \quad \ldots (7)$$

where $t_2 - t_1 = 2T = 2\pi/\omega_n\sqrt{(1-\zeta^2)}$. Thus, both the damping ratio (ζ) and the undamped natural pulsatance (ω_n) can be determined from these formulae.

(b), (c) *Area Criteria.* Suitable criteria to control relative stability can be constructed from the positive and negative areas (A^+, A^-) of the impulse response. For a second-order system the positive area is given by,

$$A^+ = \frac{1}{1 - \exp\left[\dfrac{-\pi\zeta}{\sqrt{(1-\zeta^2)}}\right]}$$

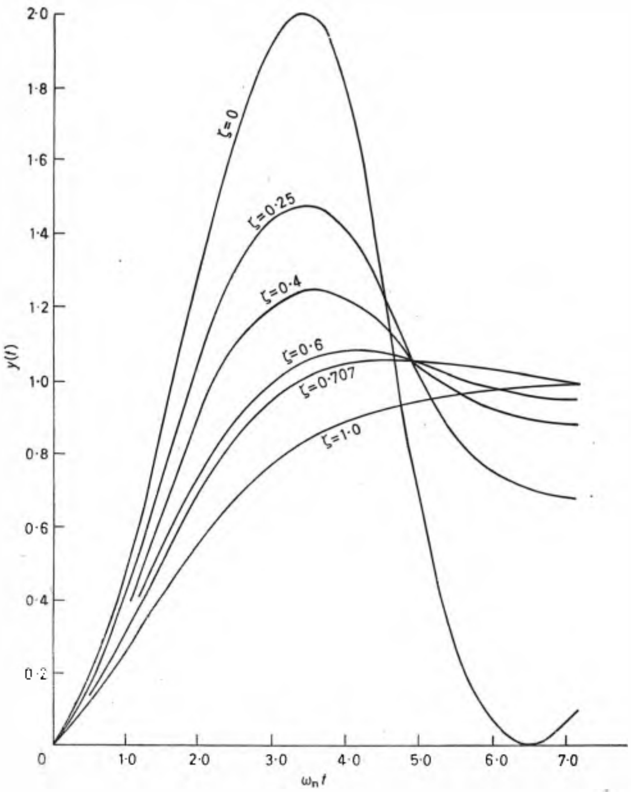


Fig. 5. Unit step response of a second order system

Similarly

$$A^- = \frac{\exp\left[\dfrac{-\pi\zeta}{\sqrt{(1-\zeta^2)}}\right]}{1 - \exp\left[\dfrac{-\pi\zeta}{\sqrt{(1-\zeta^2)}}\right]}$$

Thus,

$$\text{Impulse Response Area Ratio} = \exp\left[\frac{\pi\zeta}{\sqrt{(1-\zeta^2)}}\right] \quad \ldots\ldots\ldots\ldots (8)$$

The criterion using the area difference would take the form:

$$(A^-) - (KA^-) = \frac{1 - K \exp\left[\frac{-\pi\zeta}{\sqrt{(1-\zeta^2)}}\right]}{1 - \exp\left[\frac{-\pi\zeta}{\sqrt{(1-\zeta^2)}}\right]} \quad \ldots\ldots\ldots \quad (9)$$

The value of K could be chosen to make the criterion value equal to zero when the desired degree of damping is obtained. Any deviations from this design value would then be in the form of positive or negative signals depending upon whether the stability had increased or decreased. This could be particularly useful in adaptive systems or when a qualitative assessment of the effects of non-linearities on system stability is required.

Higher Order Systems

Most practical systems have a characteristic equation which is higher than second order. This results from shaping networks, transducers etc., which are an inevitable part of the main system configuration. It is convenient in system analysis to assume a principal quadratic factor as the significant term which characterizes the impulse response. Practical measurements will confirm whether the assumption is valid for the system 'hardware' in use. Fig. 8 shows the impulse responses for a unity numerator, cubic denominator system when the coefficients are varied. The destabilizing tendency caused by the introduction of a numerator coefficient in the cubic system is illustrated in Fig. 9. The impulse responses for two cascaded quadratic lags of fixed damping ratios but varying bandwidths are plotted in Fig. 10.

Non-linear Systems

All physical systems are non-linear in some degree. Any general deterioration with time of the components of the system due to wear, ageing or fatigue will cause time varying system parameters. Amplifier saturation, transport lags, stiction and backlash in mechanical components will all contribute to produce systems which cannot be analysed as autonomous systems and to which the principle of superposition does not apply. In a constant parameter linear system, for example, the form of the time response is independent of the size of the input or initial conditions; stability is a property only of the system. In a non-linear system the nature of the time response, and therefore the dynamic characteristics of the system generally, is dependent on the input or initial conditions. The effects of saturation, stiction, backlash and transport lag on the system's impulse response are shown in Fig. 11. The destabilizing tendency of these non-linearities can be clearly seen. From a practical viewpoint it is important to realize that if a non-linear element is within a closed-loop system, its position within that loop is important and the input conditions at that element must be considered. For example, if an error amplifier in a closed-loop system saturates it is of little use to keep the system input constant unless the system output is also monitored and the actuating signal (system input—system output) is determined. This effect is particularly important where frequency response techniques are being used and the error signal increases as the system response falls off with frequency.

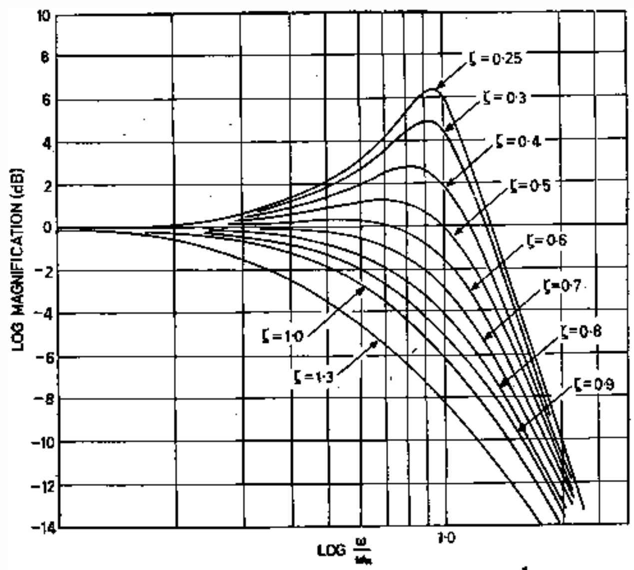


Fig. 6. Frequency Response of the Function $\frac{1}{1 - (\omega/\omega_n)^2 + 2j\zeta(\omega/\omega_n)}$

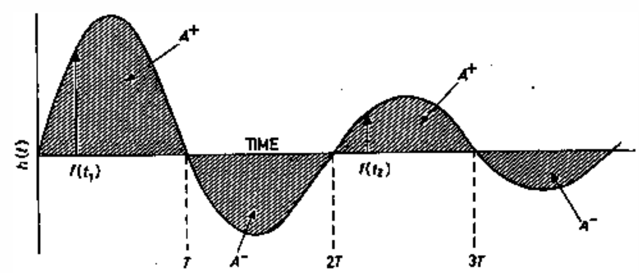


Fig. 7. Impulse response criteria

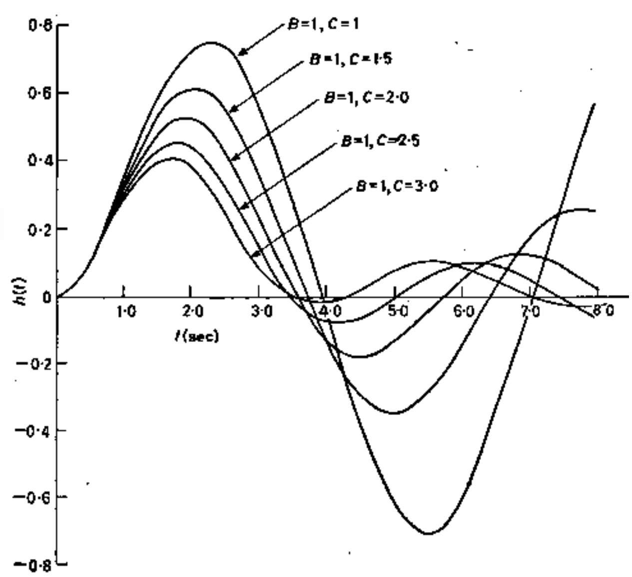


Fig. 8. Impulse response of a system with a transfer function
$$H(S) = \frac{1}{1 + CS + BS^2 + S^3}$$

Use of Cross-Correlation in Impulse Testing

Principle of the Method

The two main methods of determining empirically the dynamic characteristics of a system use either transient or sinusoidal test inputs. Step function inputs tend to cause saturation and are not usually realistic inputs while steady-state frequency response methods, to be really meaningful, require measurements to be made at several discrete frequencies which can be a lengthy process. Both step and sinusoidal inputs must be well above the system noise level and this fact usually precludes 'on line' testing. Small test perturbations can only be used if the resulting measurements reduce the effects of system noise. Cross-correlation techniques have been shown to produce significant improvements in the empirical measurement of dynamic systems. The essential principle of the cross-correlation method of measuring system impulse response is that if white noise is applied to a linear system the cross-correlation of system input and output gives the systems impulse response. If the input is $x(t)$ and the output is $y(t)$ then the system impulse response at time $t = \tau$ is given by:

$$\phi_{xy}(\tau) = \int_{0}^{\infty} h(t)\, \phi_{xx}(\tau - t)\,.\,dt \quad \ldots\ldots\ldots \quad (10)$$

where,

ϕ_{xy} = input/output cross-correlation function.
ϕ_{xx} = input auto-correlation function.
$h(t)$ = impulse response of the system.

If this is compared with the convolution integral (equation (1)) it can be seen to be of the same form if ϕ_{xy} is identified as the output and ϕ_{xx} as the input. Thus, if a test signal is chosen whose auto-correlation function is an impulse, measurement of the input/output cross-correlation function will give the impulse response of the system. White noise, whose power density spectrum is constant at a value $2\pi K$, has an auto-correlation function which is an impulse of magnitude K i.e., $\phi_{xx}(\tau) = K\delta(\tau)$. If this is applied as an input to the system an output $\phi_{xy}(\tau) = K\,.\,h(\tau)$ is obtained. Hence, it is seen that by measuring the cross-correlation function between the system input and output, when the input is white noise, the value of the systems impulse function at the instant $t = \tau$ is obtained. By varying τ the entire impulse response can be plotted.

Basic Components of a Cross-Correlator

The essential components of a basic cross-correlator can be deduced from the mathematical description of the cross-correlation function, $\phi_{xy}(\tau)$.

$$\phi_{xy}(\tau) = \lim_{T \to \infty} (1/2T) \int_{-T}^{+T} x(t)\,.\,y(t+\tau)\,.\,dt \quad \ldots\ldots \quad (11)$$

Here the input is multiplied with a time advanced version of the output; the same effect can be achieved by keeping the output undisplaced and then time delaying the input. In addition to this delay, ideal multiplication and integration are required together with a noise source. These essential basic components of a cross-correlator are shown in Fig. 12. There are certain practical difficulties which make a cross-correlation in this basic form inadequate. White noise, characterized by a flat power spectrum of infinite bandwidth, is unrealistic and therefore band limited noise has to be used. With conventional physical noise sources it is difficult to generate stable low frequency outputs even with feedback control. The integrator averagers have to be stable over long integration periods even though the system dynamics may require short measurement periods. Stable and accurate multiplication is also a problem. These considerations led to the use of a binary noise generator which gives an output, known as a maximal length pseudo-random binary sequence, whose auto-correlation function approximates an impulse of known amplitude which is repeated at pre-determined intervals. Because of the repetitive nature of the auto-correlation function integration is only required over the repetition period and the binary nature of the signal makes multiplication a simple switching operation, while delays can be readily generated through logical combinations of the binary sequences.

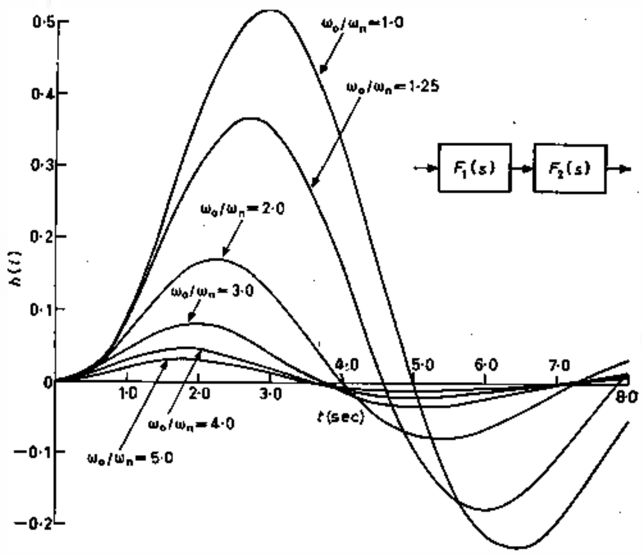


Fig. 10. Convolution of cascaded quadratic lag functions F_1, F_2 for various bandwidth ratios

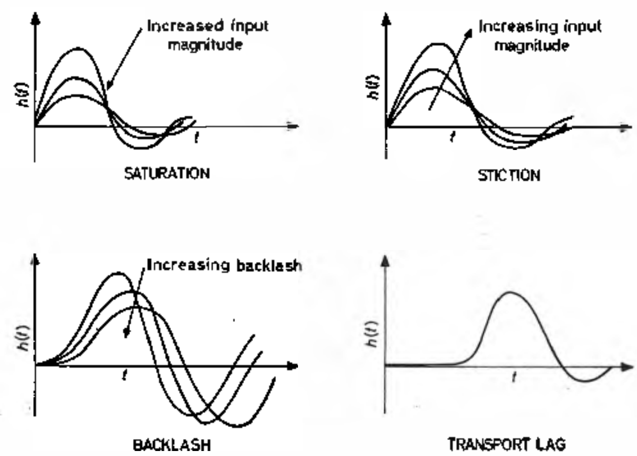


Fig. 11. Effect of non-linearities on impulse response

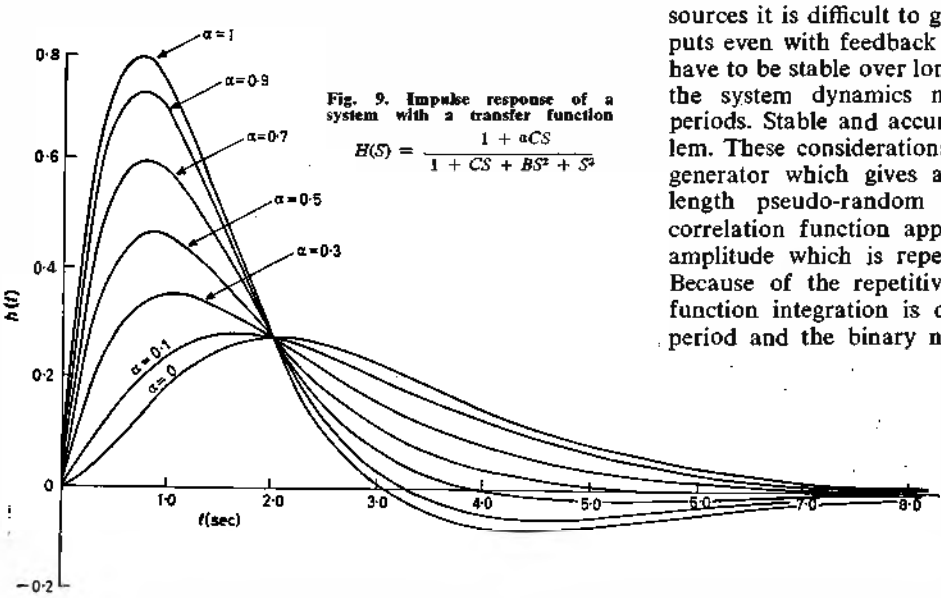


Fig. 9. Impulse response of a system with a transfer function $H(S) = \dfrac{1 + \alpha CS}{1 + CS + BS^2 + S^3}$

Practical Considerations

To obtain the correct impulse response of the system under test it is necessary to ensure that the repetition period between impulses is much greater than the length of the significant part of the impulse response and that the impulse width is narrow compared to this length in each period. Inevitably, with practical test configurations, the impulse response of transducers, networks, etc. which are cascaded with the main system element must be considered. For example if two stages are in series the resulting impulse response of them together is given by the time domain convolution of the individual impulse responses. If the impulse response of one of the stages is narrow compared with that of the other, the response obtained is essentially that of the stage with the significantly longer response. This is no different from frequency domain considerations where, if the bandwidth of one stage of a two-stage cascaded system is large compared with the other its effect will be negligible on the overall frequency response. This can be verified by inspection of Fig. 10 which illustrates the effects of convolution of two quadratic lags.

An Automatic Servo Tester

The Sperry 'Inspectron' automatic servo tester is an instrument which uses correlation techniques to enable an unskilled operator to rapidly check out a wide range of servo systems. The test is done entirely automatically, the system's impulse response being checked at pre-selected delays against tolerance limits set by a programmable plug-in unit. The operational status of the system is given as an unequivocal GO, NO-GO indication on a meter. The testers application is not confined to servo systems and any active or passive system having bandwidths between 0·025Hz to 190Hz can be tested to an accuracy of 1 per cent. A manual facility allows the operator to plot the system's cross-correlation function, i.e., impulse response, at a variety of time delays which are controlled from the front panel. The arrangement of the instrument as used in the automatic mode is given in Fig. 13.

The test signal, a pseudo-random binary sequence, is generated by a 10-stage shift register with feedback from stages 10 and 3 to the input via a modulo-2 sum circuit. This gives a maximal length sequence of 1 023 bits; the individual bit periods being chosen from three frequency settings (1 600Hz, 200Hz, 25Hz) derived from the master clock. The latter comprises a temperature compensated alloy tuning fork, a drive coil and a pick-up coil on permanent magnets and a silicon transistor oscillator circuit. The whole is assembled on a common base which in turn is resiliently mounted within a hermetically sealed can. The output frequency is 1 600Hz and the accuracy is $\pm 0{\cdot}1$ per cent over a temperature range of -5°C to $+55^\circ$C.

A unique characteristic feature of a maximal length *m* sequence is its 'shift and add' property. That is, if an *m* sequence is added (modulo 2) to one of its delayed version on a bit by bit basis, still another version is obtained with a different delay. Delayed versions of the sequence can readily be obtained by tapping off the output of selected stages of the shift register and logical combination of these in a modulo-2 summer will produce a resultant *m* sequence which is delayed by an integral number of clock periods; the actual delay being a function of the register stages initially selected. In the 'Inspectron' up to four stages can be selected which results in a total number of 441 possible delays. Extra logic circuits would be required to obtain all the 1 022 delays covering the sequence length. In the automatic mode 11 pre-selected delays are used as test

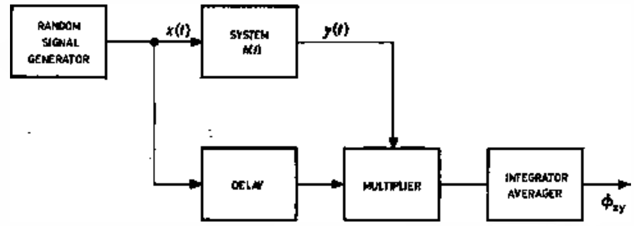


Fig. 12 Basic components of a cross-correlator

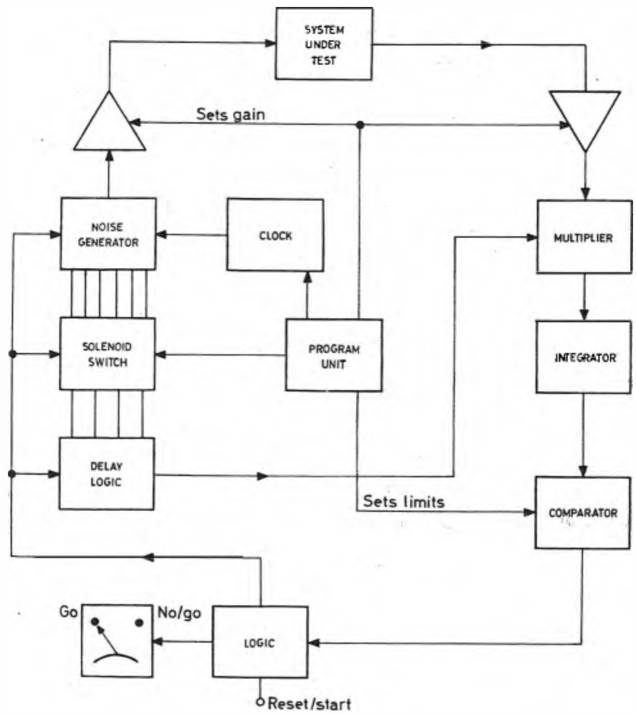


Fig. 13. Sperry automatic servo tester

points; an exponential delay distribution ensures that fast rise times and steady-state response are adequately covered.

The use of binary test signals means that multiplication can be a simple switching operation. Thus the multiplier consists of a reversing amplifier and a switch. The latter is actuated by the polarity of the reference binary noise and feeds the system output or an inverted system output to the integraor, dependent upon whether the reference noise is positive or negative.

The integrator output is then fed to a comparator which is used for the GO/NO-GO assessment. This comprises a computing amplifier in which the integrated output voltage is subtracted from a pre-set voltage. The former represents the measured magnitude of the system's impulse response at a finite time, the latter represents the desired value of the impulse response. The amplifier gain effectively sets the tolerance limits about this desired value, both these being automatically set by the programme unit. The amplifier output is then used as the input to a non-linear circuit which gives a GO/NO-GO indication depending upon whether the measured value of the impulse response is within or has exceeded the tolerance limits.

Acknowledgments

The author wishes to thank the Directors of the Sperry Gyroscope Co. Ltd for their permission to publish this article.

A Note on A.C. Compensating Networks

By D. R. Wilson*, M.Sc., Ph.D., Mem.I.E.E.E.

A survey of Whiteley's demodulating compensating network is presented which illustrates that the network can be of use as an a.c. servo compensating circuit, which has four particular modes of operation: (1) as a lead network, (2) as a lag network, (3) as a lag-lead network, (4) as a lead-lag network.

(Voir page 342 pour le résumé en français: Zusammenfassung in deutscher Sprache auf Seite 343)

A RECENT article[1] drew attention to a method of synthezing a.c. compensating networks. Particularly, lead-lag and lag-lead configurations must have at least a bi-quartic transfer function, thus the resulting networks produced by conventional synthesis techniques are always extremely complex. This note draws attention to the demodulating compensating network[2] which may be used as a lead, lag, lead-lag or lag-lead, network with the same basic element.

Lead Network

Fig. 1 illustrates the network components where the switches may be replaced by any suitable demodulating means. Under the conditions of an ideal switching demodulator i.e., the switch action may be represented by:

$$M(t) = u(t) - 2u\,(t-a) + 2u\,(t-2a) - \ \ldots\ldots \quad (1)$$

such that:

$$[M(t)]^2 = 1 \quad \ldots\ldots \quad (2)$$

the transfer function may be written[3]:

$$Y(s) = \frac{S^2 + \dfrac{(2-8/\pi^2)}{T}\,S + \omega_s{}^2 + \dfrac{(1-8/\pi^2)}{T^2}}{(S + 1/T)^2 + \omega_s{}^2} \quad \ldots\ldots \quad (3)$$

and hence the frequency response deduced.

The notch in the characteristic occurs at $\omega = \omega_n$ say, where:

$$\omega_n{}^2 = \omega_s{}^2 - 1/T^2 \quad \ldots\ldots \quad (4)$$

ω_s being the reference switching frequency of the demodulator.

In the case of a non-ideal switching demodulator i.e., a demodulator with a dead space in the characteristic defined by:

$$M_1(t) = u(t-b) - u(t-(a-b)) - u(t-(a-b)) + u(t-(2a-b)) - \ldots\ldots \quad (5)$$

See Fig. (2), the transfer function may be expressed[4]:

$$Y(s) = \frac{\left(1-\dfrac{2b}{a}\right)s^2 + \dfrac{2}{T'}\left(1-\dfrac{2b}{a}-\dfrac{4}{\pi^2}\cos^2\pi\dfrac{b}{a}\right)s + \left(\dfrac{1}{(T')^2}+\omega_s{}^2\right)\left(1-\dfrac{2b}{a}\right) - \left(\dfrac{1}{T'}\right)^2\dfrac{8}{\pi^2}\cos^2\pi\dfrac{b}{a}}{\left(s+\dfrac{1}{T'}\right)^2 + \omega_s{}^2} \quad \ldots\ldots \quad (6)$$

$$\text{where } T' = \frac{T}{\left(1-\dfrac{2b}{a}\right)}$$

and hence again the frequency response predicted.

Lag Network

Fig. 3 represents the lag configuration where the operator $M(t)$ represents the demodulator. For the case of ideal switching the transfer function may be written[5]:

$$Y(s) = \frac{8/\pi^2\,(1+sT)}{(1+sT)^2 + (\omega_s T)^2} \quad \ldots\ldots \quad (7)$$

* *A.E.I. Electronics Apparatus Division.*

From which it can be seen that the frequency response, Fig. 4, is only acceptable as an a.c. phase lag compensator when $(\omega_s T) \gg 1$. The frequency response and hence the compensating action may be predicted in any particular case by the selection of the time-constant.

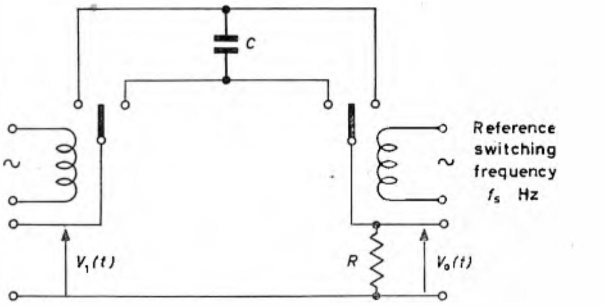


Fig. 1. Whiteley's demodulating compensating network in its phase advance mode

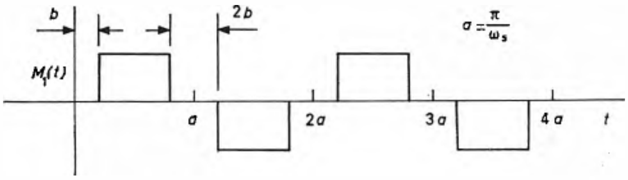


Fig. 2. Switching wave with non-conducting time b

Lag-Lead

Fig. 5 shows the network in its lag-lead configuration. In the most general form the network transfer function has been shown[6,7] to be:

$$\frac{V_0(s)}{V_1(s)} = Y(s)$$

where:

$$Y(s) = R_j \sum_{m,\text{odd}}^{\infty} \frac{8}{(m\pi)^2}\,G(s + jm\omega_s) \quad \ldots\ldots \quad (8)$$

where $G(s)$ is defined by the relation:

$$\frac{X(s)}{V(s)} = G(s) \quad \ldots\ldots \quad (9)$$

and:

$$X(s) = \mathcal{L}\,\frac{V_0(t)}{M(t)} \quad \ldots\ldots \quad (10)$$

$$V(s) = \mathcal{L}\, M(t)\, V_1(t) \quad \ldots\ldots\ldots\ldots\ldots \quad (11)$$

In this particular case it can be shown[5] that for $R_1 \neq 0$:

$$G(s) = \frac{s^2 + s\,(R_1/L) + 1/LC}{s^2 + (R_1/L) + 1/CR)s + (1 + R_1/R)\,(1/LC)} \quad \ldots.(12)$$

and in the case of negligible inductor resistance, i.e., $R_1 = 0$, then:

$$G(s) = \frac{S^2 + 1/LC}{S^2 + (1/CR)S + 1/LC} \quad \ldots\ldots\ldots \quad (13)$$

from which the overall transfer function of the device may be computed in any particular case by substitution of equation (12) or (13) in equation (8).

Lead-Lag

Fig. 6 shows the network configuration for the network acting in a lead-lag mode of operation. Once again the network may be analysed by the method indicated in reference (8) where it is shown that for $R_1 \neq 0$ the transfer

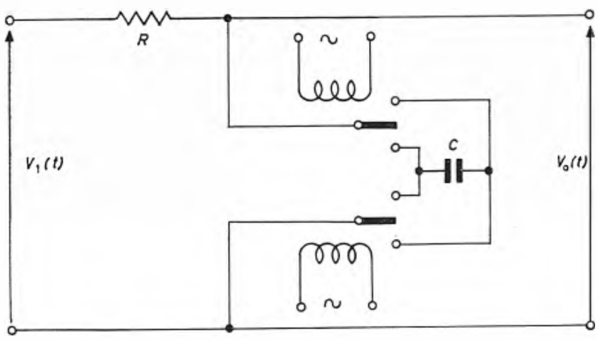


Fig. 3. Whiteley's demodulating compensating network in its phase lag mode

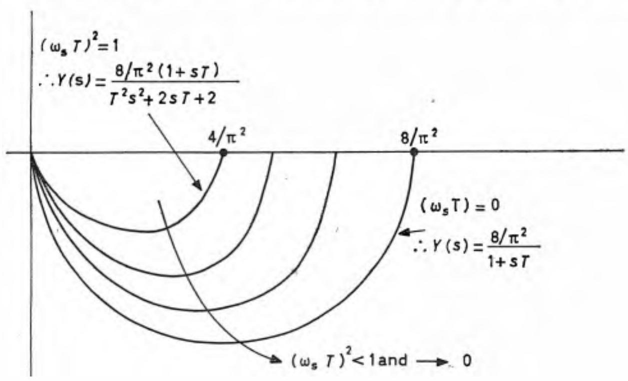


Fig. 4. Frequency response for the condition $1 \geq (\omega_s T)^2 \geq 0$

	Symbols	
$M(t)$	=	Square wave switching function
$M'(t)$	=	Square wave switching function with non-conducting time b
$Y(s)$	=	Network transfer function
$\omega_n = 2\pi f_n$, f_n		is the network notch frequency
$\omega_s = 2\pi f_s$, f_s		is the reference demodulator switching frequency
$T = RC$	=	time-constant of the simple lead or lag network
$a = \pi/\omega_s$	=	half period of the reference switching frequency
$T' = \dfrac{T}{\left(1 - \dfrac{2b}{a}\right)}$	=	an apparent network time-constant
R_j	=	real part off with respect to j

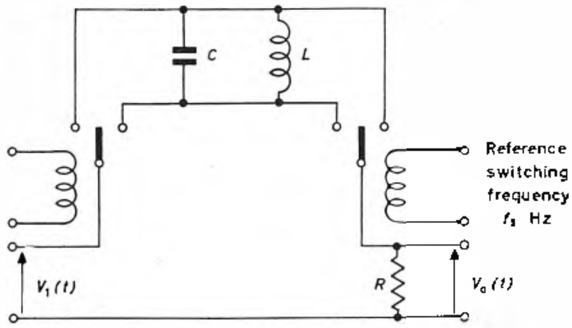


Fig. 5. A.C. lag-lead network
Ideal inductance shown i.e. $R_1 = 0$

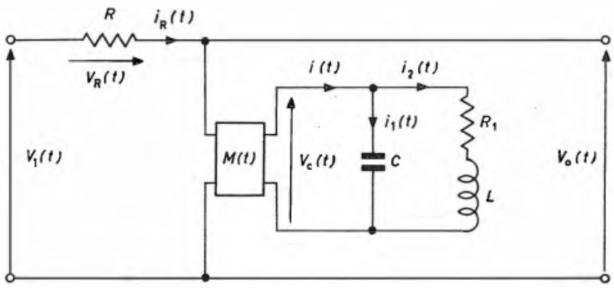


Fig. 6. Circuit model of the lead-lag network with a non-ideal inductor
i.e. $R_1 \neq 0$

function may be written:

$$G(s) = \frac{1/(T_1T_2)(1 + sT_1)}{s^2 + (1/T_1 + 1/T_2)\,s + \dfrac{(1 + R/R_1)}{T_1T_2}} \quad \ldots\ldots \quad (14)$$

where $T_1 = L/R_1$, $T_2 = RC$ and in the case of $R_1 = 0$ it follows:

$$G(s) = \frac{(1/T_2)\,s}{s^2 + s/T_2 + 1/LC} \quad \ldots\ldots \quad (15)$$

by substitution of equation (14) or (15) in equation (8) any particular frequency response may be predicted.

Conclusion

A survey of the demodulating compensating network has been presented in which four modes of operation have been drawn out, thus enabling compensation of an a.c. servo to be achieved by the same basic network lead, lag, lead-lag, or lag-lead.

Acknowledgment

The author thanks the Directors of A.E.I. Electronics Apparatus Division for permission to publish and Dr. A. L. Whiteley for useful discussions.

REFERENCES

1. Advani, J. G., Neelkantan, M. W. Frequency Transformations for A.C. Networks. *Electronic Engng. 38*, 525 (1966).
2. Patent No. 620 028. Dec. 1964.
3. Wilson, D. R. Transfer Function of an Active A.C. Phase-Advance Network. *Electronic Letters. 1*, 162 (1965).
4. Wilson, D. R. *Inst. Elect. Electronic Engrs. P.G.A.C. 1*, No. 6 (1966).
5. Wilson, D. R. A Note on Whiteley's Network *Proc. Instn. Elect. Engrs.* (Jan. 1967).
6. Wilson, D. R., Walker, F. Design of Switched Networks for A.C Servo Compensation. (In two parts). *Control. 10*, 299 and 370 (1966).
7. Wilson, D. R. I.F.A.C. Congress, 1966.
8. Wilson, D. R. Lead-Lag and Lag-Lead A.C. Compensating Networks. To be published.

Further references not directly referred to in the text.

9. Murphy, G. J., Egan, J. F. *Nat. Electronics Conf. 15*, 1056 (1959).
10. Murphy, G. J., Egan, J. F. The Analysis of Demodulating Compensating Networks. *I.R.E. Trans. Automatic Control. AC-4*, 71 (1959).
11. Bohn, E. V. A Simple Method for the Analysis of Demodulator Compensating Networks. *I.R.E. Trans. Circuit Theory (U.S.A.). CT-8*, 306 (1961).
12. Bohn, E. V. Demodulator Lead Networks. *I.R.E. Trans. Circuit Theory. CT-7*, 56 (1960).
13. Knox, A., Murphy, G. J. A Method for the Design of a Phase Lag Demodulating Compensator for use in Carrier Control Systems. *I.E.E.E. Trans. Applic. Industr. (U.S.A.). 73*, 252 (1964).
14. Weiss, G. Synchronous Networks. *I.R.E. Trans. Automatic Control (U.S.A.). AC-7*, 45 (1962).
15. Nagrath, I. J., Pratap, R I.F.A.C. Congress, London, 1966.

Design of Low Level D.C. Chopper Amplifiers

By S. K. Bhola*, M.Sc.(Hons) R. H. Murphy*, B.Sc.(Eng)

Choppers have been widely used in d.c. amplifiers to reduce the effective drift. However, the overall performance of a d.c. amplifier with chopper signal modulation, cannot be better than that of the chopper itself. A number of chopper devices have been developed for low level detection, but each has one or more of the problems of drift, noise, operating frequency, life, cost or stability. Recent advances in semiconductor technology have produced semiconductor choppers which surpass most other choppers in many respects. This article discusses the application of semiconductor choppers to low level d.c. signal amplification and points out the basic limitations of the chopper technique.

(Voir page 342 pour le résumé en français: Zusammenfassung in deutscher Sprache auf Seite 343)

IN low level chopper circuits for high quality d.c. amplifiers, a balanced transistor pair is used as the input modulator. Fig. 1 shows a balanced transistor chopper along with its equivalent circuit in 'on' and 'off' conditions. In the 'on' condition the transistor is characterized by two parameters, i.e. offset voltage $V_{EC(off)}$ and saturation resistance $r_{EC(on)}$.

For a single transistor, it can be shown[1] that:

$$V_{EC(off)} = (kT/q) \ln (1 + (1/h_{FE})) + I_B r_C \quad \ldots\ldots (1)$$

and

$$r_{EC(on)} = (kT/qI_B) \frac{h_{FE} + h_{FE(I)}}{h_{FE(I)}[1 + h_{FE}]} + r_C + r_E \quad \ldots\ldots\ldots\ldots (2)$$

where k = Boltzmann constant = $1{\cdot}38 \times 10^{-23}$ joules/°K
T = Absolute temperature
q = Electronic charge = $1{\cdot}59 \times 10^{-19}$ coulomb
I_B = Base drive current
r_C = Collector bulk resistance
r_E = Emitter bulk resistance
h_{FE} = D.C. gain of transistor in normal connexion
$h_{FE(I)}$ = D.C. gain of transistor in inverted connexion.

The 'off' condition is defined by an offset current, which, for a single transistor is given by:

$$I_{EBO} = I_{EO} + V_{BE}/r_{EO} \quad \ldots\ldots\ldots\ldots (3)$$

where I_{EO} is the saturation current which flows through the transistor under no bias conditions. V_{BE} is the reverse bias across diode junction.

The second term represents a leakage current proportional to voltage which may be accounted for by the ohmic leakage resistance across the diode junction. In equation (3) r_{EO} is the emitter diode leakage resistance.

It is evident from equations (1) and (3) that if tight control is maintained in manufacture and units are matched for various parameters, $V_{EC(off)}$ and I_{EBO} can be reduced to very small value. Any temperature effects on these parameters also likewise cancel. The practical limit to these residual errors is the closeness of matching and constancy of matching over the temperature range and varying drive and signal currents. As is seen the $r_{EC(on)}$ terms add up for the balanced configuration.

In addition to these static or d.c. errors, there is a dynamic error which is produced by the coupling of the base drive waveform to the output. This appears in the form of spikes of different magnitude and duration for positive and negative parts of the driving waveform. The main reason for these spikes is the output capacitance although the magnitude and duration are strongly circuit dependent. In short the magnitude depends upon the magnitude, frequency and rise time of the driving waveform and transistor capacitance; the duration depends upon load and source resistance, load capacitance and amplifier bandwidth[3].

The factors which must be considered for the design of low level chopper amplifiers are:

(1) *Bandwidth*

The bandwidth of the amplifier should be rationally computed for the particular applications involved, because bandwidth dictates the chopping frequencies. It is generally found that the chopping frequency should be about 50 times the highest frequencies to be amplified so that the attenuation introduced at the output filter for band-pass is negligible. Now, the effects of transients becomes more and more severe as the chopping frequency increases and so does the error due to this effect when signals are of quite low level. Where good accuracy and linearity are required, this becomes a serious limitation. In that case, it is advisable to use the chopper channel only for d.c. and very low frequency signals and a separate channel for a.c. signals and then combine the two at the output stage. This arrangement is popularly known as the parallel-T configuration. For low level d.c. signals, a bandwidth of about 1 to 3Hz is quite adequate for most applications.

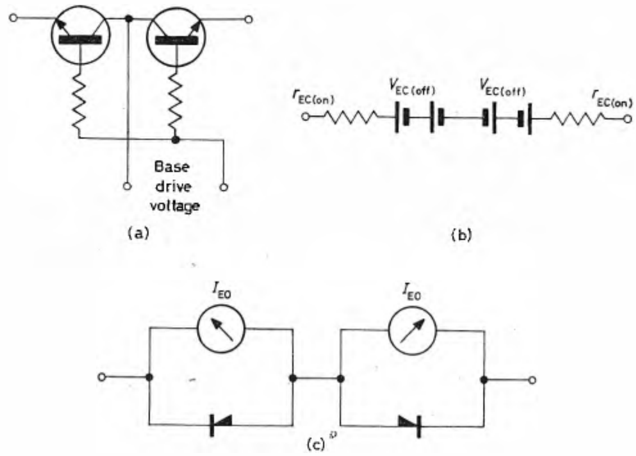


Fig. 1. Balanced transistor chopper and equivalent circuits

(2) *Noise*

The overall noise in the circuit can be considered to be of two parts. One is due to the passive components between the signal and the a.c. amplifier and the second is due to the a.c. amplifier itself. Before the a.c. amplifier, both the bias and source resistance have thermal noise characteristics which are dependent upon temperature and

* *Transitron Electronic Ltd.*

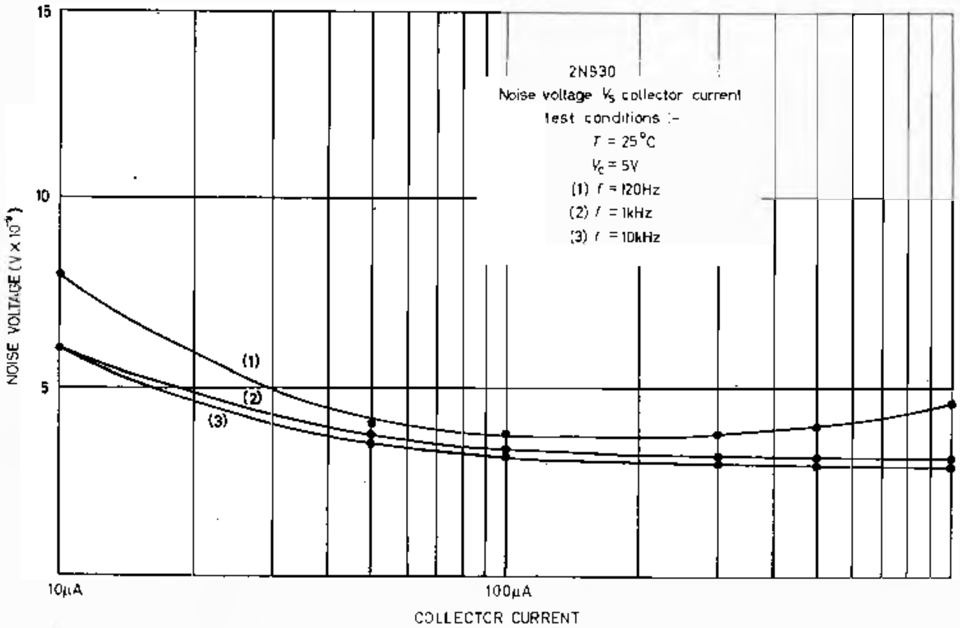


Fig. 2. Typical variation of noise voltage with collector current for the 2N930

crease. A compromise is made such that the transistor is biased for best offset and lowest temperature coefficient and then the signal current is reduced so that the error due to r_{CE} is reduced. Then the effective offset V_O will be:

$$V_O = V_{EC(off)} \pm I_s r_{EC(on)}$$

where

I_s is the signal current.

At the demodulator, it is more important to decrease $r_{EC(on)}$ because if it is large, it will introduce non-linearity, and the gain for positive and negative signals will be slightly different. This is because $r_{EC(on)}$ changes with load current, so if $r_{EC(on)}$ is small its change will also be made small.

bandwidth. In actual components, the noise voltage can be more than the theoretical value, depending upon the quality of components used.

The noise introduced by the a.c. amplifier depends mainly on the transistor used and the input circuit. Low noise transistors having high gain at low collector currents are selected. The variation of noise voltage against collector current for the 2N930 is shown in Fig. 2. Fig. 3 indicates the dependence of noise voltage on source resistance which clearly shows the advantage of low source resistance. Transistor noise also depends upon the operating frequency. It remains constant in the range of 1 to 10kHz. At lower frequencies, it varies inversely with frequency while at higher frequencies the f^2 law is followed.

(3) *Chopper Drive Voltage*

The chopping frequency should be about 50 times the bandwidth of the d.c. amplifier. In cases where low bandwidths are involved, a chopping frequency from 100Hz to 2kHz should be used to reduce the errors due to transients and noise due to the a.c. amplifier.

The magnitude of drive depends upon input signal voltage and current. In low level applications, only turn-on drive is required. The drive voltage should be high enough so that the transistor has a low offset voltage, low temperature coefficient of offset voltage and low dynamic resistance. In dual transistors, the tracking also depends upon base drive. Generally offset voltage will increase with drive base current I_B and dynamic resistance will de-

(4) *Output Stage*

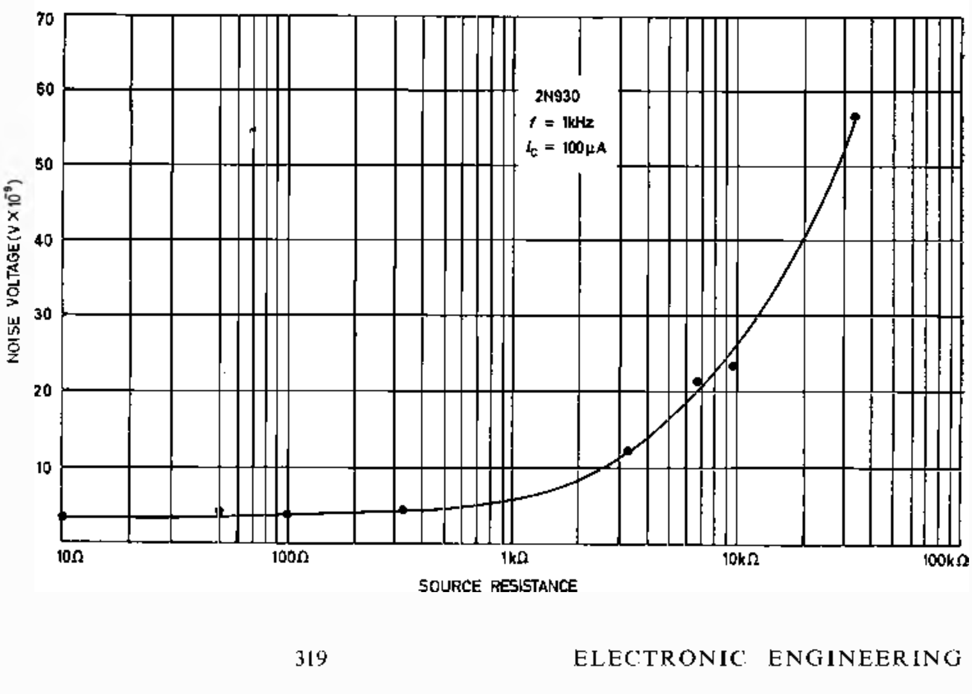
When a low level amplified signal delivers a load current of a few milliamperes, the drift in the output stage becomes significant unless high d.c. gain is achieved. If a single transistor is used as a cathode-follower at the output stage, its V_{BE} varies at the rate of $-2{\cdot}4\text{mV}/°\text{C}$ for silicon transistors. The d.c. gain must be 10 000 so that it introduces a drift of only $0{\cdot}25\mu\text{V}/°\text{C}$.

This problem is made less critical if the arrangement as shown in Fig. 4 is adopted at the output. The balanced configuration tends to compensate the changes due to temperature. It can be easily derived from semiconductor theory that the temperature coefficient of the base emitter voltage of a transistor is given by[3]:

$$\delta V_{BE}/\delta T = (k/T)\,[\ln I_c - 3\ln AT] + E_g - (3k/q) \quad \ldots\ldots\ldots\ldots\ldots \text{(4)}$$

where:

Fig. 3. Typical variation of noise voltage with source resistance for the 2N930



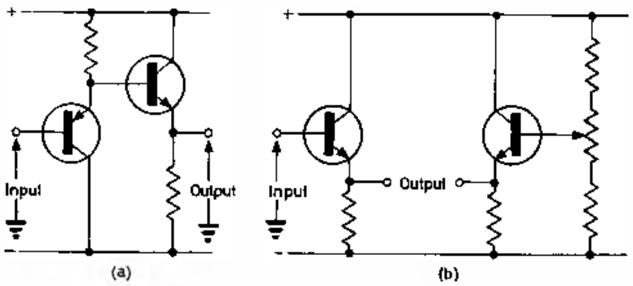


Fig. 4. Output stages (*a*) *complementary*, (*b*) *differential*

A = Constant for the semiconductor device depending upon the material, doping, etc.

E_g = Energy gap between the valence and conduction bands in the semiconductor material.

T, V_{BE}, I_c, k, q and I_c have their usual meaning.

From equation (4) it is quite easy to show that the residual drift of a differential pair such as in Fig. 4(b) is given by:

$$\delta V_{BE(12)}/\delta T = (k/q)\ln(I_{c1}/I_{c2})$$

By adjusting the currents in the two transistors, the output can have positive, negative or zero temperature coefficients. Knowing the temperature coefficient of offset voltage; the output stage can be adjusted for zero temperature coefficient at the output.

The same compensation technique can be applied with the arrangement of Fig. 4(a) but due to differences in A and E_g of the two transistors, a complete cancellation becomes very difficult. In practice the output stage will have a drift of 10 to 50μV/°C if V_{BE}'s are matched to within 5 per cent and good thermal equilibrium is maintained. The practical limitations of non-uniformity of temperature and non-tracking of V_{BE} sets a lower limit.

Fig. 4(b) also gives an easy means of adjusting output to zero. It should be noted that here the output impedance of the stage is equal to the sum of those of single output transistors.

Where large offsets are not encountered, zero can be obtained by varying the collector current in either of the transistors in Fig. 4(a). This is attractive in that it offers a much lower output impedance, without any sacrifice of performance.

Fig. 5 shows a practical circuit of a low level d.c. chopper amplifier.

Input Circuit

In order to investigate the limitations of this technique a 3N120 dual-emitter chopper was employed at the input stage. It has a maximum offset voltage of 10μV for base currents at 0·3mA to 3mA with temperature coefficient of offset voltage as low as 0·2μV/°C for the same base current range over 0 to 100°C. At a base current of 1mA, it has a dynamic resistance of 20Ω.

As is evident from Fig. 1, if the chopper is to be used in single-ended operation, the base drive must be isolated from ground and should be supplied through a transformer. Now if the transformer winding capacitance is not very small, it will differentiate the drive voltage and very large transients will appear at the output. Also, the circuit has to be very carefully shielded against stray capacitance.

If on the other hand, the chopper is used as a differential unit, the drive can be earthed at one end. Also here, since the signal is being fed only to one half of the balanced chopper, the effective dynamic resistance is half that of the dual chopper. At the same time this circuit has the advan-

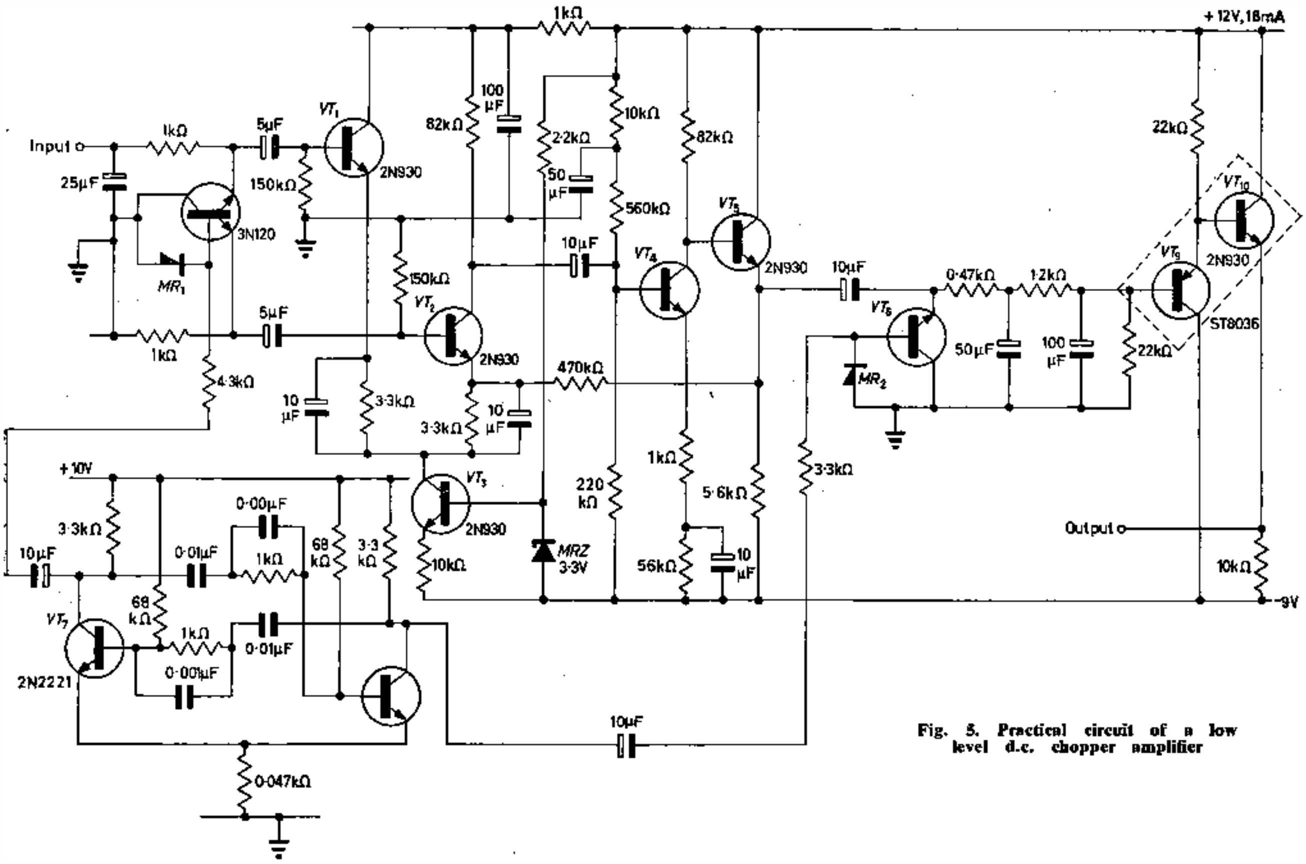


Fig. 5. Practical circuit of a low level d.c. chopper amplifier

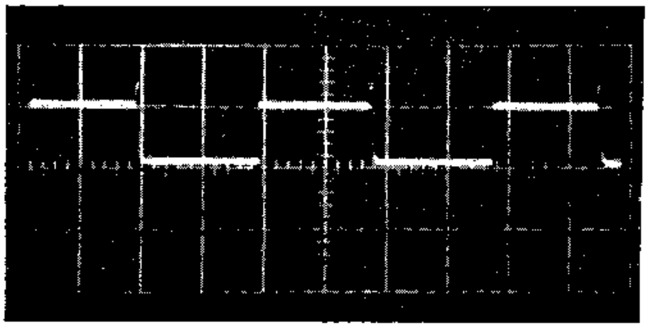


Fig. 6. Waveform at the input of the a.c. amplifier for a d.c. signal of 1mV

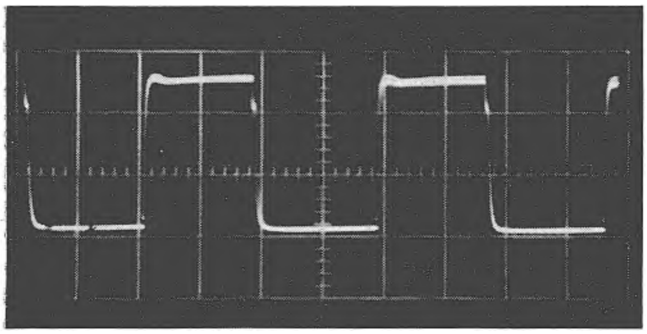


Fig. 7. Waveform at the output of the a.c. amplifier for a d.c. signal of 1mV

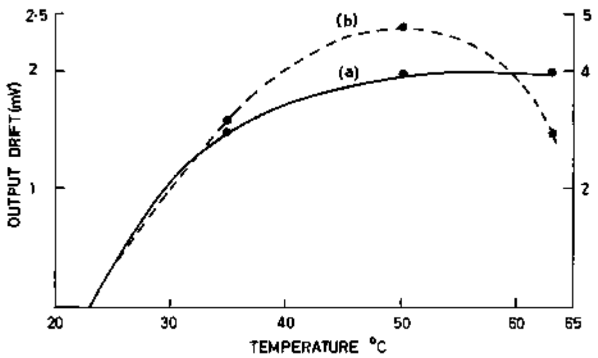


Fig. 8. Temperature drift compensation
(a) drift with output stage, (b) drift without output stage

tage of cancellation of static and dynamic errors inherent in the transistor switch.

The drive voltage is an 8V peak-to-peak, 1kHz square wave supplied by a conventional multivibrator. The chopper operates at a base current of 1mA to give the lowest offset and lowest temperature coefficient of offset voltage. Diode MR_1 provides a low impedance path for the drive voltage when the transistor is 'off', thus reducing the magnitude and duration of the turn-off transients.

A current limiting resistor of 1kΩ is added to each section of the chopper. Any variation in source resistance will alter the magnitude of the signal fed to the a.c. amplifier and so change the effective d.c. gain of the system. The signal developed across the a.c. amplifier is:

$$[V_s \pm V_{EC(off)} \pm r_d I_s]\left(\frac{R_L}{R_L + R_s}\right)$$

The upper limit of R_s is the minimum signal voltage. The signal current flowing through the transistor chopper in the 'off' condition must be more than the offset current which also flows through R_s. For the 3N120, the offset current at 25°C is 0·5nA, and considering the minimum signal current to be 20 times more, the minimum signal voltage is restricted to 10μV. The lower limit of R_s is set by the conditions that I_s should be small enough to ensure that the chopper is always in saturation. Apart from the above considerations, the value of R_s is dictated by the noise introduced by the a.c. amplifier as discussed earlier. For optimum performance R_s is chosen to be 1kΩ.

Figs. 6 and 7 show the waveform at the input and output of the a.c. amplifier for a d.c. signal of 1mV.

A.C. Amplifier

The input stage consists of a differential pre-amplifier being fed by a constant current generator VT_3. The base-emitter variation of transistor VT_3 is compensated by the temperature coefficient of zener diode MR_3. The input resistance of the amplifier is reduced to about 100kΩ by the 150kΩ input shunt resistors. The output from the collector is a.c. coupled to the next stage which feeds a d.c. coupled cathode-follower. The open loop gain of the amplifier is 4000 but can be adjusted to any desired value by the feedback resistance R_{fb} which connects the emitter of VT_5 to the emitter VT_1.

Demodulator

Transistor VT_6 acts as a demodulator to restore the amplified a.c. signal to d.c. The waveform is supplied by the other collector of the same multivibrator.

In VT_5 the base current is about 1·5mA at which the offset voltage is about 1·5mV with a dynamic resistance of 4Ω. This ensures that the transistor is well in saturation at all values of signal current and also that the error introduced by changes in dynamic resistance is small. The temperature coefficient of offset is not important. Diode MR_2 clamps the base to ground to reduce transients.

Performance of Chopper Amplifier

The output stage as shown in Fig. 3(a) is used here. The ST8036 operates at an emitter current of 600μA while the 2N930 has a 1mA emitter current. Under these conditions, both transistors have equal base-emitter voltage drops. In practice the currents were adjusted such that this transistor pair produced a net offset of about −5mV. This procedure gives a negative temperature coefficient of differential base-emitter voltage which tends to compensate the temperature coefficient of offset voltage of the chopper, which, in this case is positive.

The amplifier was immersed in a liquid bath and its drift with temperature was measured. Figs. 8(a) and (b) give the variation of output when measured with and without the output stage. It is clear that the output stage produces a compensating effect and effectively reduces the equivalent input drift.

Theoretically, a complete compensation is possible, but in practice, it is limited by the thermal equilibrium and thermal tracking errors of the output unit, as discussed earlier.

The amplifier exhibits a noise of less than 0·5μV as referred to the input. This figure can be improved if R_s is reduced and more attention is paid to the detailed design of the a.c. amplifier. With R_s equal to 470Ω, the peak-to-peak noise was found to be less than 0·2μV.

The output impedance of the amplifier with a 2·2kΩ load was found to be 30Ω. This amplifier can be used for 100μV to 1mV signals with 1 per cent accuracy. The recommended source impedance is less than 1kΩ.

Conclusion

It is possible to design a d.c. amplifier using differential techniques without a chopper[3], to give a drift of less than $0{\cdot}5\mu V/°C$, but this requires accurate adjustment for each unit and performance is guaranteed over a limited range of temperature. In addition, the amplifier has to be placed in an oil bath for good thermal equilibrium.

On the other hand, a chopper d.c. amplifier does not need any compensation, the performance is nearly uniform over a greater temperature range, and the system does not need an oil bath. Cost-wise, chopper operated d.c. amplifiers are less expensive though they are larger in size.

Present techniques in semiconductor technology offer such drift-free devices that the limiting factor in the performance of d.c. amplifiers is the noise generated in the transistors. Further improvement in this respect will necessarily give a further improvement in the performance.

Acknowledgement

The authors record their gratitude to the Directors of Transitron Electronic Ltd for permission to publish this article.

REFERENCES

1. EBERS, J. J., MOLL, J. L. Large Signal Behaviour of Junction Transistors. *Proc. Inst. Radio Engrs. 42*, 1761 (1954).
2. EKISS, J. A., HALLIGAN, J. W. The Application Characterization and performance of the SPAT as a Transistor Chopper. *Instrum. Practice 17*, 387 (1963).
3. BHOLA, S. K., MURPHY, R. H. Low Drift Design of Differential D.C. Amplifiers. *Electronic Engng. 38*, 659 (1966)

Electronics in Aviation

Bristol Siddeley Engines Ltd has recently completed a new Electronics Laboratory at the Patchway headquarters of the Company's Aero Division near Bristol. The Laboratory is divided into separate Development and Research sections.

Development

Providing a service to all departments within the company, the main activities of the development department conveniently divide into two main categories. The first, covering instrumentation, data recording and processing, includes equipment for measurement of noise, speed, fuel flow, vibration and temperature. Analogue circuits, accounting for the bulk of instrumentation both in bench and flight test, are giving place to digital methods which improve accuracy and simplify analysis. Several major digital systems are in use and intensive development is proceeding.

The second category of development, relating to engine controls and ancillary equipment, embraces systems engineering, advanced control theory, simulation computer programming, transducer and actuator design and assessment, and both analogue and digital hardware and circuit design.

Both of these categories read across to the research department where fundamental investigation is initially carried out.

The Olympus 593 programme for Concorde accounts for a large proportion of current activity. One of the most important is the actual analogue engine control system.

Analogue Engine Control

In this work, Ultra Electronics as suppliers of the engine control system co-operated very closely with Bristol Siddeley Engines in the basic servo design.

For utmost realism, the laboratory test-bench arrangement for developing the analogue control system for the Olympus 593 incorporates the actual analogue control amplifier, together with fuel flow and primary nozzle area actuators that form part of the flight control system.

Responding to an engine speed command from a representative pilot's throttle lever, the control amplifier varies the actuator settings to give the corresponding flight performance parameters. Angular positional settings on each actuator are converted by potentiometer to an input signal which is fed to an 'electrical engine' simulating the 593 servo loop performance.

This linked method of actuator control closely relates the speeds of both low and high-pressure spools to give minimum fuel consumption at any selected load condition from flight idle to maximum output. Speed of the low-pressure compressor is controlled by that of the high-speed spool to give a direct relationship which avoids the surge boundaries of the low and high-pressure sections.

Two natural loops occur on the 593. Fuel flow is proportional to turbine jet pipe temperature and hence speed of high-pressure spool. Therefore high-pressure spool speed and fuel flow provide one loop. The second loop involves low-pressure spool speed which is governed by altering the primary nozzle area.

A further design feature clearly demonstrating the flexibility of the twin-spool Olympus 593 layout is its ability to absorb servo loop changes as development of the engine and intake proceed. Servo loop performance covers high-speed spool positioning governing, low-speed schedule and governing via the final nozzle acceleration, deceleration temperature control, safety system, and other features such as starting and relight.

14 Channel Data Recorder

For the joint Bristol Siddeley/SNECMA Olympus 593 test programme, Bristol Siddeley engineers have built four 14 channel magnetic tape analogue data recorders which continuously monitor 12 essential engine parameters one speed and one speech and timing circuit.

Incorporating record replay and on-line visual monitoring features, this equipment may be adapted for any engine test programme by simply changing the fully transistorized amplifying circuits to introduce other systems of recording. These may well be frequency modulated grid, radio slip ring or strain gauge techniques.

Digital Recording Equipment

The increase in the number of parameters to be measured coupled with the higher accuracy required has led to the installation of digital recording equipment in the engine test plant. Several systems are installed or under development for both steady state and transient recording of digital data.

The whole of the engine test plant data will be handled automatically into a computer facility. Thus by the application of digital techniques, the information from several hundred parameters on each of several test beds can be reduced in a few hours to give complete engineering performance records.

Digital data recording equipment (d.d.r.) has already been commissioned for the Vulcan flying test bed aircraft which is currently evaluating the entire subsonic performance envelope that the Olympus 593 will encounter in Concorde.

Radio Telemetry Slip Ring

For twin-spool engines, a valuable extension of conventional slip-ring recording techniques which are limited to measurement of the low-pressure (outer shaft) signals, is the radio telemetry slip ring. By this method, printed circuit electronic transmitters buried in an annulus on the rotating compressor disk and thus subjected to high temperature and extremely high '*g*' loadings, transmit frequency impulses across a narrow gap from the very heart of the high-pressure rotating assembly. Already this system has been successfully applied to the Pegasus vectored thrust engine and to both low-pressure and high-pressure shafts of the Olympus 593 which is subjected to high inlet temperatures at supersonic Mach 2·2 cruise condition.

Research

Although analogue techniques account for the bulk of instrumentation used in bench and flight testing, increasing emphasis is placed on digital methods for improved accuracy and simplification of analysis. Thus it is work on such systems either for data recorders or for on-line control of a jet engine in flight, that accounts for the major activity of the Research Department.

Coupling Dekatron Selectors with Low-Cost Transistors

By J. F. Young*, C.G.I.A., M.I.E.E., A.M.I.E.R.E.

A transistor form of cascode circuit, in which the output pulse voltage amplitude can be nearly twice the voltage rating of the transistors, is used to couple Dekatron selector stages operating at speeds up to 20kHz. Low-cost high-voltage transistors are used to obtain pulse output voltage swings of up to 200V, so making the arrangement of interest economically.

(Voir page 342 pour le résumé en français: Zusammenfassung in deutscher Sprache auf Seite 343)

When the Dekatron cold-cathode counting tube was first introduced[1], methods of coupling using both cold-cathode tubes and hot-cathode vacuum valves were described These methods have been widely used. However, the cold-cathode tube coupling has the disadvantage of limited speed of operation, while the hot-cathode valves have the disadvantages associated with heated cathodes. Consequently, there have been various attempts during the past decade to use transistors to couple Dekatron stages. In some cases the Dekatrons have been avoided completely in counters, and transistor decade counters have been used. Sometimes these have been operated at high enough voltages to operate cold-cathode indicator tubes[2]. However, although the price of Dekatron tubes has not fallen as much as might perhaps have been expected, they are still economically attractive for use in counters. Unfortunately, the interstage coupling arrangements can still add greatly to the cost of the equipment. In such circumstances, it is therefore of interest that low-cost transistors capable of operating at reasonably high voltages have recently been introduced.

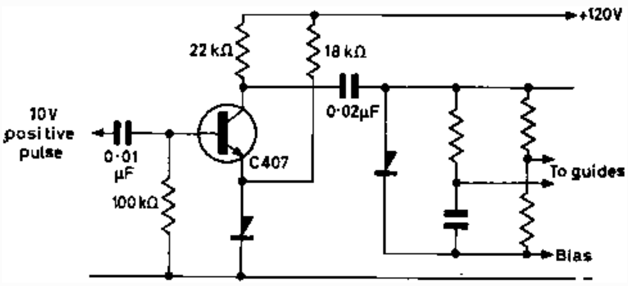


Fig. 1. Single-transistor coupling circuit

Of necessity, early methods of coupling Dekatron stages using transistors incorporated low-voltage-rating transistors driving voltage step-up transformers to increase the amplitude of the pulses to a suitable level for application to Dekatron guides. The circuits used[3,4] were based on blocking-oscillator techniques. A notable feature was the ingeniously simple way in which two diodes were included in the guide circuits to convert the bidirectional pulse from the blocking oscillator transformer into two separate but overlapping unidirectional pulses.

When higher-voltage transistors became available, attempts were made to dispense with the coupling pulse-transformer. Johanson[6] used two transistors per stage to couple Dekatron selectors. An interesting feature of the published circuits is that some of the high-voltage transistors have the base connected only to a capacitor with no normal form of d.c. return path at all. A more controlled design using the same types of transistors was described by Fraser[7]. For inter-stage coupling of Dekatron counter tubes he used a single high-voltage transistor with a 105V collector supply, connected in a circuit reminiscent of some of those used with early cold-cathode tube coupling arrangements. In the input stage he used two transistors, one per guide electrode, with a certain amount of positive feedback from the collector of the second to the base of the first. This helped to reduce the length of the trailing edge of the first transfer pulse, and it decreased the rise time of the second transfer pulse. The circuits are stated to be very satisfactory for use with medium-speed counters.

More recently, very low-cost npn transistors rated at 120V have become available. Such devices would appear to be ideally suited to Dekatron coupling applications, though they are actually intended to drive cold-cathode numerical indicator tubes. The first circuit investigated is

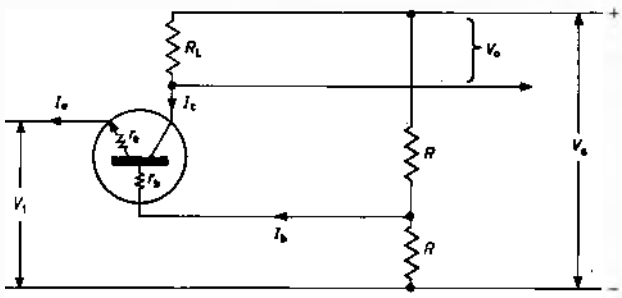


Fig. 2. Voltage-doubling common-base stage

shown in Fig. 1. The transistor is normally biased to the non-conducting condition by the forward drop of the diode in the emitter circuit, since the base is connected to the negative supply line. However, a positive input pulse can overcome the bias and cause forward base current to flow. When this happens, the transistor conducts and a negative pulse, of amplitude nearly equal to the supply voltage, appears at the collector. This pulse is coupled to the conventional guide circuit via a capacitor, a d.c. restoring diode being included. While this arrangement was reasonably satisfactory for use with GS10D Dekatron selectors, the amplitude of the output pulse was not really adequate to give reliable operation of selectors over a wide range of possible repetition frequencies. However, it would probably be adequate for medium-speed counter tubes. Unfortunately, the size of output pulse is limited by the voltage rating of the transistor, and it could not be increased without risk of unreliability.

The cost of these transistors is so low compared with that of the Dekatron selectors that the possibility of using two transistors in each coupling circuit need not be dismissed on economic grounds. However, the two-transistor

* *University of Aston.*

circuits mentioned above do not give an increased output pulse amplitude. A method of obtaining an increased output amplitude from a transistor of given voltage rating is shown in Fig. 2. The transistor is in common-base connexion, the input going to the emitter while the base is biased from the two resistors R across the supply V_S. If the transistor is cut-off, no voltage appears across the load R_L. As the emitter is made negative to the base, the transistor starts to conduct and the output voltage V_O increases. When the output voltage has increased to rather over one-half of the supply voltage, then the collector-to-base voltage of the transistor becomes zero and the transistor is in the bottomed condition.

If the emitter is now made even more negative, the transistor remains bottomed and the output voltage continues to increase as the emitter and collector move negatively together. When the input voltage V_1 is zero, the output voltage almost equals the supply voltage V_S. Now in this arrangement, the collector-to-base voltage of the transistor has never exceeded $V_S/2$, and the required swing of the input voltage V_1 is also $V_S/2$, yet the output voltage can swing over the full range of the supply voltage V_S. Consequently, a transistor with a collector-voltage rating of 120V should be capable of providing a maximum output voltage swing of nearly 240V, provided that no current gain is required in this stage. Note that there is no phase inversion in this stage.

A practical arrangement incorporating this circuit is shown in Fig. 3. Here, transistor VT_1 provides the input voltage for transistor VT_2. The forward drop of the diode MR_1 provides a bias voltage for VT_1, so that VT_1 is normally cut-off. A positive input voltage pulse to the base of VT_1 causes it to conduct, so providing emitter current for VT_2. This arrangement has been successfully used to couple together Dekatron selectors, the input coming from a Dekatron cathode and the output going to a conventional guide-feed circuit.

Analysis of the circuit of Fig. 2, assuming that

$$I_c = \alpha I_e + I_{co}$$

shows that the output voltage V_o is given by

$$V_o = I_c R_L = \frac{\alpha V_1 + \alpha V_s/2 + I_{co}(r_c + r_b + R/2)}{r_e + (r_b + R/2)(1-\alpha)} R_L$$

where α is the emitter-collector current gain,

I_{co} is the temperature-dependent part of the collector current,

r_e and r_b are the internal emitter and base resistances of the transistors.

The equation shows that in the linear region the voltage gain approaches

$$V_o/V_1 \rightarrow \frac{2\alpha}{1-\alpha} \cdot (R_L/R)$$

and this can be high. Once this transistor is bottomed, its voltage gain falls to unity. Thus there is a large change of gain as the output voltage swings. However, by making

$$R/R_L = \frac{\alpha}{1-\alpha}$$

then the linear voltage gain is 2 and the circuit need never operate in the bottomed condition. The change of gain with operating point would then be avoided.

In a pulse application involving a large voltage swing, the change of gain would probably not matter anyway. However, there is a further possible problem with this arrangement. The variation of output voltage with the temperature-dependent current I_{co} approaches

$$V_o/I_{co} \rightarrow \frac{R_L}{1-\alpha}$$

so that the circuit is potentially very temperature dependent. This is particularly undesirable in high-voltage work[3]. In addition, the effect of supply-voltage variation approaches

$$V_o/V_s \rightarrow \frac{\alpha}{1-\alpha} \cdot (R_L/R)$$

The points mentioned are unlikely to be too important in many applications, especially if silicon transistors are

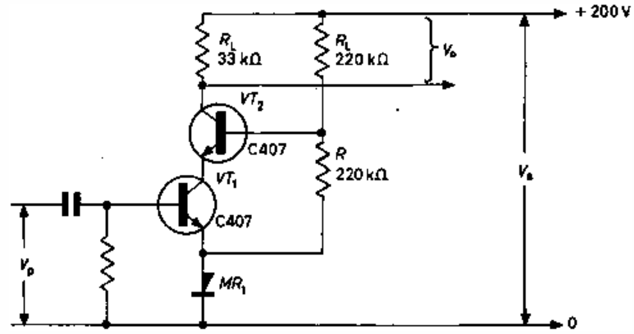


Fig. 3. Cascode arrangement without negative feedback

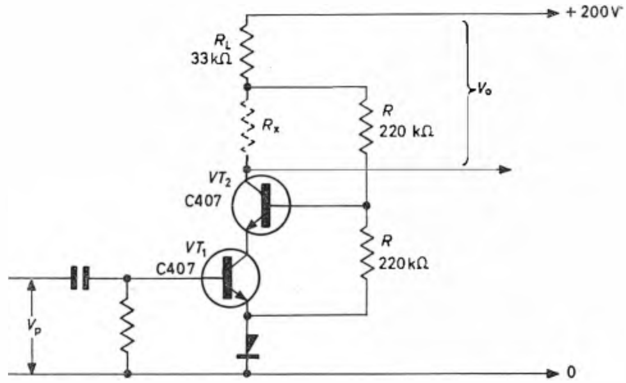


Fig. 4. Cascode arrangement with negative feedback

used. However, the introduction of collector-base feedback in the second transistor circuit[9,10] can anyway give a considerable improvement. The circuit then becomes that shown in Fig. 4. Analysis of this arrangement shows that

$$V_o = \frac{V_1 R R_L(1+\alpha) + V_s R_L(R+r_e+r_b(1-\alpha)) + I_{co} R R_L(R+r_e+2r_b(1-\alpha))}{(R_L+2R)(R+r_e+r_b(1-\alpha)) - (1+\alpha)R^2}$$

Thus in this case, the dependence on temperature approaches

$$V_o/I_{co} \rightarrow \frac{R_L}{R_L/R + (1-\alpha)}$$

As might be expected, this can be less than was obtained with the circuit of Fig. 3.

Also in Fig. 4, the effect of supply voltage variations approaches

$$V_o/V_s \rightarrow \frac{R_L}{R_L + R(1-\alpha)}$$

This again indicates the improvement expected due to the negative feedback. However, the linear voltage gain

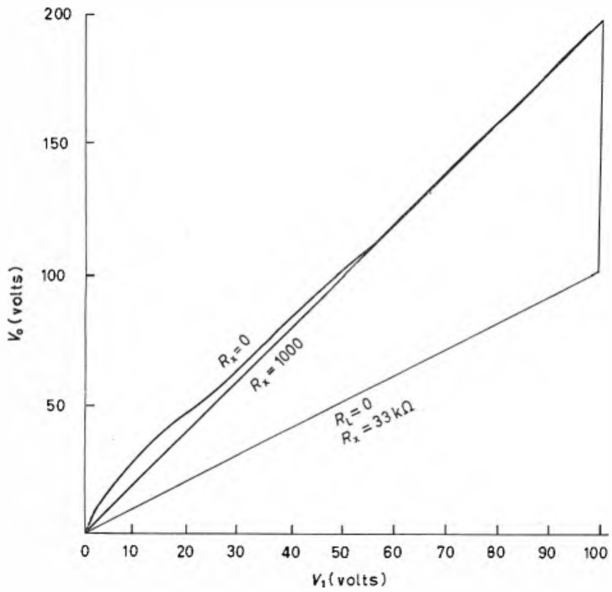


Fig. 5. Output voltage-input voltage characteristics

approaches

$$V_o/V_1 \to \frac{1+\alpha}{1+(1-\alpha)\,R/R_L}$$

In order to make this approach a value of 2, it would be necessary to make the value of the ratio R/R_L approach zero. Fortunately, in practice a voltage gain in the second stage of less than two is acceptable. An output voltage-input voltage curve for the second stage is given in Fig. 5. In a pulse application the slight variation of slope is unimportant, but it can be avoided if required by the addition of a resistor of about 1kΩ at R_x. The overall voltage gain of the two stages is about 800 and the current gain is about 75, while an output voltage swing of about 200V is obtainable.

REFERENCES

1. BACON, R. C., POLLARD, J. R. The Dekatron, A new Cold Cathode Counting Tube. *Electronic Engng. 22,* 173 (1950).
2. YOUNG, J. F. Counters and Shift Registers Using Transistor High Voltage Static Switching Units. *Electronic Engng. 37,* 161 (1965).
3. WARMAN, J. B., BIBB, D. M. Transistor Circuits for use with Cold-Cathode Gas-Filled Multi-Cathode Counter Valves. *Electronic Engng. 30,* 136 (1958).
4. CHAPLIN, G. B. B., WILLIAMSON, R. Dekatrons and Electro-Mechanical Registers Operated by Transistors *Proc. Instn. Elect. Engrs. 105B,* 231 (1958).
5. KLEIN, A. G., PRYOR, A. W. The Application of Transistors to Nucleonic Instrumentation. *Proc. Inst. Radio Engrs. Aust. 20,* 388 (1959).
6. JOHANSON, E. W. Glow-Tube Programmer Controls Neutron Spectrometer Experiments. *Electronics, 34,* 65 (12 May, 1961).
7. FRASER, H. J. Transistorized Dekatron Driving Circuits. *Electronic Engng. 34,* 40 (1962).
8. YOUNG, J. F. The Use of Transistors in Industrial Timer Circuits. *Electronic Engng. 35,* 366 (1963).
9. GRINICH, V. H. An Eighty-Volt-Output Transistor Video Amplifier. *Trans. Inst. Radio Engrs. CT3,* 61 (1956).
10. SALAMAN, R. G. Receiver Video Transistor Amplifiers. *Trans. Inst. Radio Engrs. BTR4,* 68 (1958).

Scanning Electrical Information Channels

With long-running machinery left unattended for any length of time, it is often desirable to provide an automatic cut-out so that in the event of, say, a bearing overheating or oil pressure rising to an undesirably high level, the machinery is de-energized before any damage results.

Proprietary pre-set trip devices are available for this purpose and these usually also provide a continuous measurement of the particular parameter under surveillance. If several information channels however are to be monitored and each has a different trip level, then it is usually necessary to provide a separate pre-set trip device for each channel since most pre-set trip devices can only be adjusted to one specific trip level. This means that organizations undertaking this kind of multi-channel monitoring on some scale have to stock relatively expensive trip device in some quantities.

An idea put forward by the National Coal Board's Central Engineering Establishment, near Burton-on-Trent, overcomes this problem and provides a scanning system which while incorporating a single pre-set trip device only, can nevertheless monitor and trip several electrical information channels having different trip levels.

The proposal includes a switch circuit which automatically connects each channel in turn to the pre-set trip device. In the first instance a channel is directly connected to the pre-set trip device by the switch circuit so that an actual measurement of the voltage signal level is obtained, then the channel is switched-in via a further variable voltage input which is set by adjusting a variable resistance to bias the signal in such a way that its true trip level is shifted to a nominal trip level corresponding to the trip setting of the pre-set trip device.

Each channel in turn is scanned in this manner and by setting the trip on the trip device to a nominal value and adjusting the variable resistances to give biasing voltages matching the difference between the true trip levels of each channel voltage signal and nominal trip setting a trip condition in each channel can be detected. A system of coloured lamps is used to clearly indicate whether the pre-set trip device is indicating at any given instance a true measure of the signal voltage in a particular channel or whether it is searching for a trip condition in that channel.

Upon a trip condition in a particular channel being detected, the pre-set trip device trips and de-energizes the machinery.

Simultaneously, the switch circuit ceases further switching operations and maintains the connexion between the tripped channel and the trip device so that the lamp appropriate to that channel remains illuminated.

Prototype equipment made by the Central Engineering Establishment has been applied with some success to monitor bearing temperatures on machinery undergoing long-running tests in the Establishment's test laboratories.

SHORT NEWS ITEMS

The Fifth International Aerospace Instrumentation Symposium will be held at the College of Aeronautics, Cranfield, Bedford, on 25 to 28 March 1968. The symposium is sponsored by the College of Aeronautics and the Instrument Society of America Aerospace Industry Division. Papers are invited for presentation at this symposium and authors are requested to submit a 300 to 500 word summary for consideration by the sponsors not later than 1 June this year.

Appropriate subjects in the applied science and technology of instrumentation of aerospace vehicles may be chosen; however, papers concerned with:

Instrumentation for Helicopters and/or VTOL Aircraft;

Development in Transducing Techniques;

On Line Data Processing for Aerospace Flight Testing;

Analysis of Non-Stationary Data;

will be particularly welcome.

Authors are asked to submit their summaries to:

Symposium Organizer, Department of Flight, The College of Aeronautics, Cranfield, Bedford,

or to:

E. K. Merewether, Chairman, International Affairs Committee, ISA Aerospace Industry Division, 20801 Sherman Way, Canoga Park, California, U.S.A.

A Conference organized by the Institution of Electronic and Radio Engineers with the collaboration of the Institution of Production Engineers and the Institution of Electrical Engineers, and entitled "Integration of Design and Production in the Electronics Industry" is to be held at the University of Nottinghame on 10 to 13 July this year.

Further information, including registration and accommodation details, may be obtained from the Institution of Electronic and Radio Engineers, 8-9 Bedford Square, London, W.C.1.

A two-day study course is to be held at the Fulmer Research Institute, Stoke Poges, Buckinghamshire, on 16 to 17 May. The course is sponsored by Associated Electrical Industries Ltd, Leicester, and the Fulmer Research Institute and will include an introduction to spark source mass spectrometry and demonstrations using the Fulmer Research Institute's AEI MS702 double focusing mass spectrometer.

This course is designed to show the latest techniques for analysing and measuring trace elements in natural products, dental samples, semiconductor materials and metals, among other uses.

The Society of Instrument Technology (S.I.T.) is to hold a meeting on 1 May at Manchester Literary and Philosophical Society, 36 George Street, Manchester 1, under the auspices of the Manchester Section.

The paper to be read is: 'Simulation and Computer Control of Road Traffic' by M. G. Hartley and D. H. Green.

Further details are available from R. Maley, National Industrial Fuel Efficiency Service, 89 Oxford Street, Manchester 1.

Another S.I.T. meeting is to be held on 2 May at Cleveland Scientific Technical Institute, Corporation Road, Middlesbrough, under the auspices of the Tees-Side Section.

The paper to be read is: 'Lasers and their Application to Measurement' by Professor O. S. Heavens, joint meeting with the Institute of Physics and Physical Society.

Further details are available from R. W. Dean, 23 Thornfield Road, Middlesbrough, Yorks.

The Council of the Institution of Electrical Engineers has approved the acquisition and joint use of an ICT Model 1901 computer with the Computer Services Centre (London) Ltd through the medium of a joint company Faraday Computers Ltd.

The computer will handle programs devised by each partner. Those of the IEE will deal with such office matters as subscription collection, addressing its 58 000 members and the like, and will also play a vital part in a program known as INSPEC, which is concerned with providing a substantially mechanized information service in physics, electrotechnology and control, for which a groundwork is provided by the IEE's long experience in this field as the publisher, since 1898, of 'Science Abstracts', and latterly of the current awareness of publications which now supplement the original work.

The computer with its full range of ancillary equipment has now been installed in the IEE Centre at Savoy Place.

The Acoustics Group of the Institute of Physics and the Physical Society is to give greater attention to physical acoustics and ultrasonics together with the use of their techniques in the fundamental studies of gases, liquids and solids. The Group is also interested in the development of atmospheric, medical and underwater acoustics, the non-destructive testing of materials, and noise, especially those aspects which are of interest to physicists. The Group also stimulates the study of acoustics at all educational levels.

A number of symposia and lectures are arranged each year, either independently or jointly with other groups and societies. In addition, conferences are organized and visits are arranged to industrial or other establishments of interest to members.

Details of membership can be obtained from The Registrar, The Institute of Physics and The Physical Society, 47 Belgrave Square, London, S.W.1.

BBC engineers have now solved the problem of the electronic conversion of monochrome and colour television pictures exchanged between America and Europe.

In 1963 the BBC developed the first all-electronic television standards convertor, giving high-quality conversion of pictures between systems using different numbers of lines, but having the same number of pictures per second. This type of convertor is capable of operating unattended and is widely used to change pictures between the two main systems used in Western Europe; the 405-line system used by BBC-1 and ITV and the 625-line one used by BBC-2 and the rest of Europe. The optical convertor has, however, been the only type which could be used in the exchange of programmes between Europe and America, because of the difference in the numbers of pictures per second used in the two continents.

It is expected that equipment using the new principle will be available in time for the opening of the colour service on BBC-2 later this year, allowing exchanges to be made between Europe and America of colour pictures. A firm commitment has also been accepted to provide pictures from the Olympic Games in Mexico in 1968, converted from the American colour pictures, for transmission throughout Europe by the Eurovision Distribution Network of the European Broadcasting Union.

The National Research Development Corporation has negotiated an agreement with Welwyn Electric Ltd of Bedlington, near Newcastle, on the automated low-cost production of microcircuits.

Production rates up to 200 000 micro-assemblies a week are expected at a cost which will enable Britain to compete on favourable terms in world markets. It is the first venture of its kind to be launched in the North East development area.

The micro-assemblies to be produced are thick-film and thin-film products. These consist of a glass-like substrate of alumina ceramic on which resistors and capacitors can be chemically deposited, printed or etched.

These are the passive components. Active elements, diodes and transistors are made in the same factory and attached to the films to make hybrid assemblies.

Welwyn have spent six years in developing the new technique. Ultimately their research, and the new automated production line, will cost around three-quarters of a million pounds. NRDC has agreed to make a substantial contribution to this cost.

The production methods chosen by Welwyn are chemical deposition and fired-on glazes instead of vacuum deposition. The Company has worked with all three methods and finds that the chemical method is to be preferred to the vacuum method, is more effective, cheaper, and can be rigorously controlled for accuracy. Glazes are even cheaper, though not quite so accurate, and very much more suited to automation techniques.

A separate organization has been set up at Bedlington for the manufacture of chip, or flake, planar silicon transistors and diodes which can be soldered directly into the thick-film circuits. These have already been successfully produced in pre-production runs of 1 000 a week, but by the end of this year output will have gone up to 25 000 a week. These semiconductors have been under development for four years.

Elliott-Automation Ltd has supplied its modified 920B type computers to the British Army for artillery fire control in the forward area.

The Field Artillery Computer Equipment (FACE) will introduce automation into command posts and survey sections and will improve the effectiveness of artillery fire. It is based on a mobile version of the general-purpose Elliott 920B computer and can be programmed for use with many types of ordnance in the battery command posts of close-support, general-support and missile regiments. In the survey role, FACE will be used for calculating the survey data required to fix and orient guns and target acquisition devices.

FACE will be able to accept the more frequent meteorological data which the Army will obtain from the complementary Artillery Meteorological System (AMETS) now being developed. AMETS will also be based on the Elliott 920B computer.

The initial orders for 76 FACE systems, which will cover the first batch of equipments only, will bring to more than 150 the number of Elliott 920 Series computers to have been ordered for use by Britain's three armed services.

The Association Internationale PRO ELECTRON has recently been formed and based at 10 avenue Hamoir, Brussels, with the following objectives:

(a) To study and define systems of nomenclature for electronic devices falling within the categories so far created, viz: valves and tubes; semiconductors, including integrated circuits.

(b) To issue and register type numbers, in accordance with the systems defined, at the request of manufacturers, whether they are members of the Association or not. PRO ELECTRON will utilize the systems already evolved and perform the service previously given by l'Association PRO ELECTRON, Luxembourg, which has now ceased activity.

(c) To promote the use of these systems of nomenclature.

(d) To promote scientific research capable of contributing to the development of improved systems of nomenclature of electronic devices.

The Association is a non-profit-making body. Its members are manufacturers of electronic tubes and/or semiconductors in European countries which are members of O.E.E.C. The constitution of the Association was published in the 'Annexe au Moniteur Belge' of 26 January 1967.

GEC Road Signals Ltd, Wembley, has now completed a £20 000 Ministry of Transport contract for the installation of signals at 45 new X-way pedestrian crossings in West London, Reading and Lincoln. Twenty-eight of the installations are in West London, six in Reading and eleven in Lincoln. Those in West London will form part of the area's computerized traffic control system.

The X-ways, announced by the Ministry of Transport on 22 February, supersede the 'Panda' system and will operate experimentally in a number of centres in England and Scotland.

The Autoflex X-ray signal is a fixed-time controller incorporating an electronic timer which automatically controls the sequence in which the signals presented to oncoming traffic and pedestrians are displayed.

A Board of Trade probe into the air traffic control system at Heathrow Airport, London, is to continue this summer using equipment supplied by Hawker Siddeley Dynamics' Industrial Automation Division.

A record of aircraft movements, at various points on the airport, will be made in a form suitable for computer analysis, providing a complete picture of the traffic pattern. This analysis is hoped to show ways of improving the flow of traffic.

The contract with Hawker Siddeley Dynamics is for a new data acquisition system, consisting of a tape-recorder and playback equipment. Details of aircraft movement will be coded on to the tape through a push-button electronic coder. This information will be converted, on the playback equipment, to punched tape which will be fed into the computer. Additional information can be recorded vocally and coded on to the punched tape by hand.

Included in the recent list of new Fellows of the Royal Society are Dr. F. E. Jones, Managing Director, Mullard Ltd, London, for his contributions to radar and infra-red technology, and for his outstanding technological leadership in an advanced industry, and J. M. Ziman, Professor of Theoretical Physics at the H. H. Wills Physics Laboratory in the University of Bristol, for his theoretical contributions to solid state physics, especially the study of transport phenomena and for his work on the electronic properties of liquid metals.

Saab, the Swedish aircraft, car and electronics manufacturer, has received an order valued at 19 million francs for the development of the telecommunication systems for the TD 1 and TD 2 satellites to be launched by the European Space Research Organization.

Saab is a member of the MESH group of companies which also includes Engins Matra (France), Entwicklungsring Nord (Germany), and Hawker Siddeley Dynamics (Great Britain) with TRW Systems, USA, as technical consultants. The group members are at present engaged in the ESRO II satellite, the

ELDO-A satellite and in national space programmes.

The telecommunication system, with associated test equipment, to be developed by Saab includes on the transmitting side units for collecting and coding measurement data from the satellites' scientific instrumentation and systems and the storage of these data on tape recorders. On command from the ground, the stored data are transmitted to the ground stations by means of a telemetry link.

For the receiving and execution of commands from the ground, the telecommunication system includes a command receiver, a decoder and switching logics. A considerable number of the units in the system are duplicated to ensure maximum reliability. Extensive test equipment will be fabricated to facilitate testing and trimming of the system.

Among sub-contractors to be used by Saab for this project are SRA and L. M. Ericsson for telemetry transmitter and command receiver and ASEA for parts of the test equipment.

AEG-Telefunken have erected for Radiodifussion Télévision Belge (RTB) an h.f. antenna system near Wavre, south-east of Brussels, which is intended for broadcast coverage of the former Belgian Congo and Rwanda Burundi/Central Africa. The system comprises four directional radiators (wideband dipole curtains).

By remote control the main radiation direction of each antenna may be slewed electrically by $\pm 5°$ and $\pm 10°$. The antenna lengths extend from 30m (for the 17/14/11m bands) up to 92m (for the 49/41/31m bands).

For the suspension of the curtain arrays the height of two existing towers had to be increased to 55m and 80m, and two new towers 36m and 67m high had to be erected.

A five-year technical agreement for the exchange of specialized knowledge and joint co-operation in the field of automation and electronic equipment and components between the Plessey Co. Ltd and the State Committee of the U.S.S.R. Council of Ministers for Science and Technology has been signed in Moscow.

As a result of this agreement an extensive 'package deal' in automation and electronic systems, equipment and components will be submitted to the State Committee.

This 'package deal', which is expected to lead to substantial business for Plessey and will, therefore, contribute considerably to the British trade balance, has the full approval of the British Government.

Avo Ltd (MI Group) has published the third edition of its Transistor Data Manual.

This international reference book gives in-line data for more than 8 000 transistors including those of Russian manufacture. It provides a rapid and convenient guide for use not only with Avo instruments but also for wider application by laboratory and service engineers.

In addition to in-line data for every transistor, a comprehensive list of transistor equivalents is included with commercial equivalents of Service transistors.

Copies of the manual are available from the Spares Department, Avo Ltd, Avocet House, Dover, Kent, price £2 5s. (postage paid in the United Kingdom).

The General Post Office has ordered from Recognition Equipment Ltd, Maidenhead, for the National Giro—the new current account banking service due to come into operation in October next year—electronic equipment that can read Giro documents and capture the information on them at the rate of 2 400 characters per second, then sort them so that they can be returned to customers each 24 hours if their account moves. A million or more transactions are expected to be handled at the centre each day when the system is in full swing.

The equipment, valued at £825 000, includes two Retina Computing Readers which achieve a character resolution in reading approaching that of the human eye. Each machine has 256 elements of information compared with the 400 that the human eye achieves in character recognition—the missing elements, however, are concerned with the interpretation of colour which is unnecessary to the machine.

The efficiency of the new retina reader is due to the three-dimensional approach which bases recognition on height, width and grey shades of character giving it a multi-font reading capacity so that it can read a variety of typing.

Each reader will have a programme controller—this is a high-speed, low-cost, general purpose computer for real time control of character recognition and paper handling plus on-line processing. Through the use of stored programmes, the programmed controller directs the reader continuously. It acquires the needed data while the document is being processed.

Other equipment in the order includes three bar-code reader sorter transports each with programme controllers and auxiliary equipment such as tape units and high speed printers.

The Ministry of Technology has contributed £7 000 for the purchase of equipment at the Advisory Centre for Low Cost Automation which has recently been opened at the Paisley College of Technology, Scotland.

The centre is the first to be set up in the United Kingdom as part of the Ministry of Technology's 'Approaching Automation' campaign to encourage industry, and especially the smaller firms, to make greater use of low-cost automation techniques.

One of the aims of the campaign is to show how automatic control can be introduced step by step, simply and cheaply, to increase output, improve the quality of the product and make more efficient use of skilled labour. The emphasis is on inexpensive techniques and devices which will pay for themselves in months rather than years and can be introduced without major reorganization or disruption of production.

The Paisley Centre is equipped with a wide range of electronic, hydraulic and pneumatic equipment which can be combined in various ways to simulate and solve industrial control problems. Specialist staff in the College will help industrialists to identify the methods and equipment best suited to their processes. The Centre will also run short training courses in low-cost automation techniques at frequent intervals.

Keyswitch Relays Ltd has modified an existing punched data card system and adapted it, for what is believed to be the first time, as a simpler, faster, more efficient aid to component selection.

The new aid, which is now freely available to users and prospective users of relays, is known as the KMRF (Keyswitch Magic Relay Finder) and consists of over twenty illustrated punched cards each containing full specification, delivery and price details of a basic Keyswitch relay type.

The cards, supplied in a protective wallet, are edge coded to cover 13 parameters, with which are associated a total of 87 different values. The parameters chosen are: coil voltage, coil minimum operating power, contact rating, number of contacts, type of contacts, working voltage, type approvals, size, mechanical life, special applications, connexions, timings and accessories. The required value of one parameter after another is specified by pushing a plastic needle through the appropriate holes. After each pass a gentle shake releases all cards which do not conform to specification, so that the remaining cards are still in alignment for the next selection.

Keyswitch estimate that the typical user will get down to one or a very few remaining cards in well under one minute. The cards can be re-used stacked in any order, provided the clipped corners are aligned. Part of the company's service will include the provision of Relay Finder cards for all subsequent new products.

NEW BOOKS

Theory and Applications of Active Devices

By H. J. Reich, J. G. Skalnik and H. L. Krauss. 739 pp. Med. 8vo. van Nostrand. 1966. Price £5 2s.

IN the last few years a very large number of textbooks have appeared dealing with the theory and application of active electronic devices. Many of these books have only dealt with transistors but a substantial number have attempted an integrated treatment of both thermionic valves and of transistors. Unfortunately most texts of the latter type have given a fairly good treatment of valve aspects but have restricted the transistor work to the rather more theoretical aspects. Often this approach has led to such books giving a complete discussion of the various types of transistor equivalent circuit with very little of the way in which these relate to actual transistor usage. The present book undertakes an integrated approach but does not suffer from the shortcomings of many of its predecessors. Practical results are given to illustrate the text and each chapter concludes with a set of problems. These problems cover the full range from pure theoretical analysis to design studies where engineering judgment is required in order to assign values to circuit components.

The first two chapters deal with the physics of electron devices and cover both electron emission and the nature of carrier flow in semiconductors. Chapter 3 discusses the pn junction diode followed by a brief chapter on the thermionic vacuum diode. The basic material is then completed by a chapter on electrical conduction in gases. Chapter 6 covers transistor operation and includes discussion of the field-effect transistor and four-layer diodes and Chapter 7 deals with multielectrode valves. The book then proceeds to consider the various ways in which active devices are employed in actual circuits. The topics include power rectification and control, untuned and tuned amplifiers, feedback amplifiers, regenerative circuits, oscillators and modulation. The transistor circuits use npn transistors so that transistors are drawn with the collector upwards, hence circuit diagrams for transistor and for valve circuits having the same function are easily compared. This is particularly useful with regenerative circuits as the waveforms have the same polarities for both thermionic valves and for transistors.

Any textbook must represent a compromise between depth of treatment and width of coverage; in the present volume the balance is nicely judged and the book is very well suited to a general course in electronics at undergraduate level.

V. H. Attree

Electronic Automatic Control Devices

By A. A. Bulgakov. 549 pp. Demy 8vo. Pergamon Press. 1966. Price £6

THIS book, which is a translation of a Russian textbook published in 1958, deals with electronic devices for automatic control.

The first chapter is introductory, and defines the concept of control, gives a short mathematical treatment of feedback, classifies the type of control device, and presents the principles of automatic control. The next three chapters deal with small signal aspects of the control loop, namely, high-gain voltage amplifiers, switching circuits for changing the loop configuration, and error sensing circuits. Chapter 5 is concerned with the ways in which grid control is achieved in thyratrons, mercury-arc rectifiers, and ignitrons.

In chapter 6 the characteristics of thyratron power amplifiers are examined both in rectifier and inverse-parallel connexions. In chapter 7, thyratron drive of d.c. motors is discussed, and in chapter 8, this is extended to the overall automatic control of a system. In the next chapter the problems involved in thyratron drive of induction motors are considered, and in chapter 10, the control of d.c. drives with rotating amplifiers is mentioned.

Chapter 12 is devoted to multi-motor drives where, for example, in maintaining constant speed or tension of materials being wound on to reels, the speed regulation of a number of motors must be matched. Chapter 13 considers programmed machine-tool control, and the last chapter deals with multi-loop automatic control systems.

This book, which is intended for engineers working in the field of automation, is well written and easy to understand. Some of the translation of technical terms is a little strange, for example 'indicating circuits' for error sensing circuits, 'counter-parallel' connexions and 'rotary' amplifiers, but this does not often obscure the meaning. The great defect of this book is that it is out of date. Between the publication of this translation and the original version, great changes have taken place in the devices used in this field. Thermionic devices are described in applications where, to a large extent, they have been superseded by thyristors and transistors.

The editor has recognized this fault, and his introduction consists mainly of a guide to semiconductor devices, but this is not a real help as it is too elementary. This book is not recommended unless one is working specifically with thermionic valves, or is prepared for the chore of adapting the results given to transistors and thyristors.

G. D. Bergman

Untersuchungen über ebene Beugungsprobleme Elektromagnetischer Wellen für Rechtwinkligkeilförmige Gebiete (Plane Difraction Problems of Electro-magnetic Waves in Right-angled, Wedge-shaped Areas) Untersuchungen an Dielektrischen Stielstrahlern über den Einfluss der Strahlungskopplung auf deren Fusspunktsimpedanz (The Influence of the Radiation Coupling on the Base Impedance of Dielectric Rod Radiators)

By N. Latz and J. Ehrhardt. Research Report of the Land Rheinland-Westfalen No. 1710. 66 pp. Royal 8vo. Westdeutscher Verlage, Cologne. 1966. Paperback, price DM.47,60

THE first paper is a contribution to the theory of the radiation fields of dielectric aerials. N. Latz comes to the conclusion that the plane diffraction of electromagnetic waves may be dealt with uniformly by the method of two-dimensional Laplace transformations. The holomorphy of the transformed into half-plane pairs and the zero conditions resulting from it are used to obtain a representation of the transformed field function. Special cases are considered and the derivation of the integral equation system is given.

In the second paper J. Ehrhardt investigates the coupling of the radiation cross-section of groups of dielectric rods. It does not only affect the radiation characteristic but also the base impedance of the individual rods, thus causing undesirable mismatching. The experimental results are used to compile a comprehensive atlas of radiation coupling. The author comes to the conclusion that the base impedance is relatively insensitive to couplings of surface waves as long as no elements capable of resonating are present in the cross-section conducting energy, and as long as the v.s.w.r. is near 1. Even rods close by do not couple much under these conditions. The impedance reacts very sharply, however, against e.g. a $\lambda/2$ dipole arranged in the neighbourhood of the rod.

E. R. FRIEDLAENDER

Microwave Valves

By C. H. Dix and W. H. Aldous. 275 pp. Demy 8vo. Iliffe Books Ltd. 1966. Price 55s.

THIS book is one of many on microwave devices but with an important difference in that it is intended as an introduction to microwave valves providing, to use the authors' words, "a readily absorbed account of the basic physical processes and operation of microwave valves." One may ask if the microwave valve has any future with the rapid advances in microwave solid state devices. The authors themselves considered this problem and the evidence supports their contention that microwave valves will be important devices for many years to come.

The book deals with all the main types of microwave valves, klystrons, travelling wave tubes, backward-wave oscillators, magnetrons and related cross-field devices and includes electron beam parametric amplifiers.

The first four chapters establish the foundations dealing with the motion of electrons in electric and magnetic fields, the properties of r.f. circuits and their utilization in conventional valves. Chapter 5 then deals with the operation of such valves at high frequencies and shows how limitations to their performance comes about.

Chapter 6 introduces the principles of velocity modulation and the space charge wave concept on electron beams. The remainder of the book, with the exception of one chapter on electron beams, details how these principles are utilized in the various microwave valves.

Chapter 7 deals briefly with klystrons.

Chapters 8 and 9 are concerned with travelling wave tubes both low and high power and includes the backward-wave oscillator.

Chapter 10 is concerned with the design of electron guns and the various methods of maintaining the beam by the use of permanent magnets. Electrostatic focusing is also described with its application to a backward-wave oscillator.

Chapters 10 and 11 are concerned with crossed-field devices including the magnetron and the amplitron.

The last chapter is devoted to noise and the design principles associated with its reduction in beam type devices.

The only criticism is the lack of sufficient references for further study. Very few mistakes have been detected but an obvious one is on page 189 where the energy products of Alcomax II and Ticonal G should be around 5×10^6.

The book succeeds in its purpose and goes a long way to remove some of the mystique associated with microwave valves and can be thoroughly recommended to all who wish to improve their knowledge of the subject.

D. E. LAMBERT.

Dictionary of Electro-Technology, German-English

By E. Höhn. 705 pp. Chapman & Hall Ltd. 1966. Price 6 gns.

THE ever-increasing specialization in science and technology brings with it a continuous demand for specialized dictionaries, and the present volume is intended to meet this demand in the field of electro-technology.

On thumbing through this volume one of the first things one notices is the surprisingly large number of entries referring to commercial, financial and legal terms, which in a dictionary of this type seems to be somewhat out of place. Inclusion of electronics terms instead might have been more appropriate in a volume stated to deal not only with heavy-current engineering but also with control, automation and nuclear energy.

A technical dictionary must be useful and concise. Terms like "usufractuary" or "O-electron" will hardly be of much use to engineers (even if the latter is translated, as "O-Elektron"). "Diode" is the same word in both languages and should not be translated as "rectifier". The one and only proper translation for rectifier is "Gleichrichter" even though a diode *may* operate as a rectifier. A similar confusion refers to "Verbindung" which can mean either a chemical compound or a physical connexion; it cannot be used for a "junction" in semiconductor devices, the appropriate German term being "Übergang". Similarly "halbleitende Verbindung" is a "semiconducting compound", while "intermetallic compound" (given as the corresponding translation) is only one of various types of semiconducting compounds.

According to the foreword a number of novel features were incorporated in this dictionary; the reviewer would, however, not agree that these are necessarily advantageous. Thus, listing the diphthongs ä, ö, ü under ae, oe, ue is against convention and will irritate many users, without bringing any obvious advantage. (It is unfortunate that a mistake has crept into the illustrative example in the corresponding explanatory notes, both in the English and the German version stating exactly the opposite of what is intended.) Another feature, i.e. to enter the past tense of verbs as independent items (e.g. "*ge*steuert" as well as "steuern") is certainly novel but rather extravagant. Taken to its logical conclusion, this might lead to doubling the number of entries referring to all German verbs! The reason given for this procedure, "to help the translator without a full command of the German language", is hardly acceptable. Another moot point in German dictionaries is always, how far to go in listing word combinations, and again insufficient consideration seems to have been given to their selection. Finally there can be little justification for giving automatically all the slightly different English and American spellings, unless different *words* are used. Surely any English reader would understand "color, analog, analyze" etc just as any American will understand "colour, analogue, analyse".

In conclusion: Although the issue of a new dictionary covering recent developments in a rapidly expanding technological field will always be welcomed, it would appear that in this case better use of available space might have been made. By considerable pruning and a more judicious selection of entries it might be possible to produce a *two-way* dictionary, perhaps even covering electronic terms, without any increase in the overall number of entries and in the cost of the book.

E. BILLIG

Analysis and Synthesis of Tunnel Diode Circuits

By S. O. Scanlan. 274 pp. Med. 8vo. John Wiley & Sons Ltd. 1966. 63s.

THE content of this book is less wide than its title indicates. In fact it does not cover digital circuits at all, but limits itself to tuned amplifiers, both narrow and broad band, and a few non-linear circuits of closely related types such as frequency convertors and amplitude demodulators. A concise first chapter summarizes the physical principles underlying tunnel diode operation, and although without material on device fabrication gives an adequate basis for what follows. The second chapter develops an equivalent circuit for the tunnel diode and discusses the dependence of the circuit parameters on working point. A discussion of the effects of temperature would have been welcome here.

Once the equivalent circuit has been derived the approach to the subject becomes very much that of the network analyst, and a thorough treatment of the properties of all the usual amplifier configurations is given, with some attention to synthesis. This part of the book is commendably clear and readable: only elementary mathematical techniques are used. Anyone concerned with problems of the kind would undoubtedly find it a valuable working reference. Finally a section on mixers and demodulators gives a useful review, highlighting some results of particular significance.

What the book does, it does well. Only a few errors of a trivial nature were noted. Criticism must relate more to what the book omits. Tunnel diodes are not being widely used at the present time, and it is arguable that, with the recent improvements in high frequency transistors and the introduction of new microwave solid state signal sources, they have now been totally eclipsed. This book, however, ignores the possibility and makes no serious effort to disclose the area (if any there be) where the advantages of tunnel diodes make them the preferred active device. The decision not to relate tunnel diode circuits to those using other active devices also hampers the exposition. Many properties of tunnel diode circuits have exact analogues elsewhere: to quote a simple example, the loss of gain stability in a transmission amplifier as the magnitude of the gain is increased is precisely what is observed in any amplifier with positive feedback. Connexions of this kind are not brought out.

Having made these criticisms, however, it is only fair to repeat that Mr. Scanlan has written an entirely satisfactory book within his self-imposed limitations. If the future of the tunnel diode appears very limited then at least this volume is a creditable obituary.

W. GOSLING

Beiträge zur Analyse des dynamischen Verhaltens von Regelkreisen (Papers relating to the analysis of the dynamic behaviour of control circuits)

By O. Schäfer, J. Janzig, H. Schwarz and K. W. Plessmann. Research Report No. 1645 of the Land Nordrhein-Westfalen. 141 pp. Royal 8vo. Westdeutscher Verlag, Cologne. 1966. Paperback, Price DM 91.30

The authors, members of the Institute for Control Engineering at the Technical University Aachen, deal in three papers with systems which go beyond the elementary theory in respect of their structure and their reaction to interference. Schäfer and Janzig report on a study of the effect of stochastic interference on linear control circuits. This is followed by proposals for the elimination of coupling effects in multiple control circuits by Schäfer and Schwarz. The concluding paper is by Schäfer and Plessmann and covers the control of systems with all-pass characteristics.

Fundamentals of Digital Machine Computing

By Guenther Hintze. 225 pp. Med. 8vo. Springer-Verlag, Berlin. 1966. DM 25.60

This book deals with basic circuits, logical organization and programming concepts common to all large modern computers.

Lithium-Drifted Germanium Detectors—Proceedings of Panel, Vienna, 6-10 June, 1966

International Atomic Energy Agency, Vienna. 225 pp. Med. 8vo. Her Majesty's Stationery Office. 1966. Price 35s. 4d.

This book contains the fourteen papers presented at the conference held in Vienna in June 1966 and organized by the International Atomic Energy Agency.

The subjects dealt with include coaxial lithium—drifted germanium diodes for gamma ray spectroscopy, current problems with germanium for gamma-ray detectors and the spectroscopy for mesonic atoms.

Membranes—Technology and Economics

By R. N. Pickles. 197 pp. Noyes Development Corporation, New Jersey. 1967. $1967.

Reports on Progress in Physics Parts I and II

Edited by A. C. Stickland. 756 pp. Crown 4to. The Institute of Physics and the Physical Society. 1966. Price £5 5s. each

Parts I and II of this volume contain 15 papers on various aspects ranging from Plasma Oscillations to Spin Waves in Ferromagnets.

Proceedings of IFIP Congress, 1965 New York City, May 24-29, Vol. 2

Edited by W. A. Kalenich. 647 pp. Demy 4to. MacMillan & Co. Ltd. 1967. Price 95s.

The main contents of this second volume consist of the summaries of the symposium sessions and the reports of panel discussions which took place at the IFIP Congress held in New York during May 1965 and organized by the International Federation for Information Processing.

Electronic Testing

By L. L. Farkas. 304 pp. Med. 8vo. McGraw-Hill Pub. Co. Ltd. Price 96s. 1966.

Three types of electronic testing are covered in this book namely (1) Receiving and transmitting equipment; (2) Special equipment and (3) Computers. The first section deals with the testing of receiving and transmitting equipment as applied to telemetry, ranging, tracking and safety systems, and to the detection and suppression of radio interference.

In the second section special devices are dealt with including sensors, gyros, inertial guidance platforms and flight control systems.

The third section covers the testing of analogue and digital computers and their internal circuits.

The Optical Properties of Solids

Edited by J. Tauc. 434 pp. Crown. George Newnes Ltd. 1967. Price 176s.

This book contains the proceedings of the "Enrico Fermi" International School of Physics' Course 34 held at Lake Como, Italy, in 1965, at which seventeen papers were presented.

Mathematics for Radio and Electronics Technicians

By Dr. I. F. Bergtold. 304 pp. Demy 8vo. George Newnes Ltd. 1967. Price 50s.

Starting with the simplest equations and basic methods of calculation, this book deals with such subjects as involution, curves, the slide rule, trigonometrical functions, founier and other series, differentiation and integration, polar co-ordinates and the Gaussion plane.

Electrical & Electronic Trader's Year Book 1967

476 pp. Demy 8vo. Iliffe Press Ltd. 1967. Price 35s.

Extensive revision has been carried out in all sections of the thirty-seventh edition of this yearbook.

A new feature gives the addresses and telephone numbers of manufacturers' own service depots and of their service agents throughout the U.K.

RCA Receiving Tube Manual

608 pp. Demy 8vo. RCA of America. 1966. Price $1.25

Partial Differential Equations

By I. G. Petrovskii. Iliffe Books Ltd. 1967. Price 50s.

This book is based on a course of lectures given by the author at the Moscow State University, and deals primarily with linear second-order equations with ine unknown function, which includes many important cases arising in physical problems such as the Cauchy problem for analytical equations with non-analytical initial conditions.

It is suitable as a source of reference for mathematicians, physicists and engineers.

A Simple Approach to Electronic Computers

By E. H. W. Hersee. 261 pp. Crown 8vo. Gordon & Breach. Price $7.50. 1967.

The additional material published in this second edition is largely confined to three chapters. Chapter 4 has been enlarged to include descriptions of computer storage systems. Chapter 7 is new and deals with simplifications and limitations of digital computing and Chapter 11, also new, introduces a special application of digital techniques not covered in the original edition.

Digest of Literature on Dielectrics, Vol. 29

457 pp. 1966. National Academy of Sciences—National Research Council. Price $20.00

This volume has been prepared by the Committee on Digest of Literature of the Conference on Electrical Insulation and Dielectric Phenomena Division of Engineering of the National Research Council, Washington D.C.

Electronics & Nucleonics Dictionary

By J. Markus. 743 pp. Med. 8vo. McGraw-Hill Publishing Ltd. 1966. Price 132s.

The Third edition of this book now contains the preferred meaning, spelling, hyphenation, abbreviation and usage of some 16 300 terms—an increase of 25 per cent over the previous edition—supported by over 1500 illustrations.

These terms ase to be found in industrial, medical, military and space electronics, television, radio, radar and nuclear engineering.

Eddy Currents

By J. Lammeraner & M. Stafl. 233 pp. Demy 8vo. Iliffe Books Ltd. 1966. Price 37s. 6d.

Originally published in Czechoslovakia in 1964 and now translated into English by G. A. Toombs, this book introduces the reader to the elements of the subject and then proceeds to the more complex problems occurring in the design of modern electrical machines. Full mathematical treatment is given throughout.

History of the House of Siemens Vol. I & II. 1914-45

By G. Siemens. 326 pp. Med. 8vo. Siemens Aktiengesellschaft. 1957

Originally published in Germany in 1957 this book has now been translated into English by A. F. Rodger and gives a full account of the history of this firm over a century of existence. Volume I covers the history from 1847 to 1914, and Volume II from 1914 to 1945.

NEW EQUIPMENT

A description, compiled from information supplied by the manufacturers, of new components, accessories and test instruments.

ISOLATED POWER SUPPLY

Bell and Howell Ltd, Consolidated Electrodynamics Division, 14 Commercial Road, Woking, Surrey

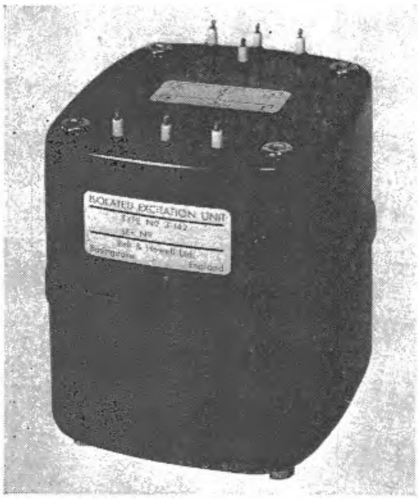
(Illustrated below)

The type 3-142-2 isolated mains power supply newly introduced by Consolidated Electrodynamics Division, Bell & Howell Ltd, is designed to provide a stable 10V output to instrumentation systems with complete freedom from problems of earth loops and common mode voltages.

The maximum output current of 320mA, at 10V allows the parallel excitation of ten 350Ω strain gauge transducers. The circuit design permits a maximum primary-to-secondary leakage current of 0·6μA peak to be specified.

Output voltage change is less than 0·2 per cent for mains voltage changes from 200 to 260V, or 100 to 130V; output resistance is less than 0·25Ω. Operational temperature range 0 to 50°C.

An alternative version—the type 3-142-1, is available for output currents in the range 0 to 32mA.



For more information circle **No. 291**

VEHICLE DETECTOR

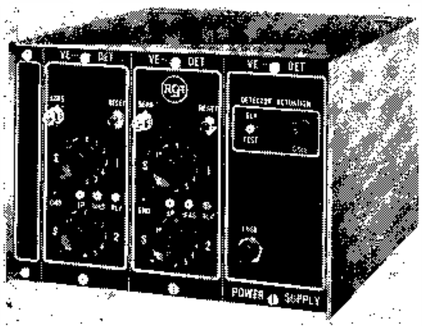
RCA (Great Britain) Ltd, Lincoln Way, Windmill Road, Sunbury-on-Thames, Middlesex

(Illustrated above right)

A compact, all-transistor system for the detection of stationary or moving vehicles is being manufactured and marketed by RCA Great Britain Ltd.

The RCA Ve-det is a product of earlier development work carried out by the David Sarnoff Research Laboratories, Princetown, New Jersey, on automated road systems. Vehicles ranging in size from bicycles to articulated lorries, buses or even taxiing aircraft can be reliably detected.

The RCA Ve-det has been selected by the Ministry of Transport for use in the



West London Traffic Experiment and the Ministry has also granted approval for Ve-det installation on selected sites, on British roads, in conjunction with approved traffic signals. Other potential applications include traffic counting, monitoring of parking facilities and incorporated in a traffic light system, control and direction of haulage vehicles in dock areas or railway termini.

Installation is less expensive than traditional methods: a wire loop, which can cover several traffic lanes, is laid in a slot, 1 to 2in deep $\frac{1}{4}$ to $\frac{3}{8}$in wide, cut in the road. The wire loop is connected to the RCA Ve-det, which can be located up to 750ft away, and mounted in a control box or in a pavement pillar.

Another version of the RCA Ve-det has been designed for the detection of rolling stock in railway systems.

The method of detection is based on the phase characteristic of a parallel resonant circuit, comprising the inductance of the loop and the tuning capacitor of the Ve-det. Tuning is carried out at the time of installation and takes less than one minute. By driving the tuned circuit with an r.f. current constant in amplitude and phase, the voltage across the *LC* combination reflects the change in inductance of the loop as a vehicle enters it. The detector then uses this phase change to de-energize a relay in the sensor unit.

No maintenance is normally required for the road loop since this is unaffected by weather conditions or mechanical damage by heavy traffic and can be left unattended for indefinite periods. Normal road repairs can be made without removing the loop. It is a practical proposition to lay loops at strategic points when relaying road surfaces, permitting rapid connexion of the RCA Ve-det to the loop at a later date for traffic counting, speed testing or the permanent installation of light controls.

For more information circle **No. 292**

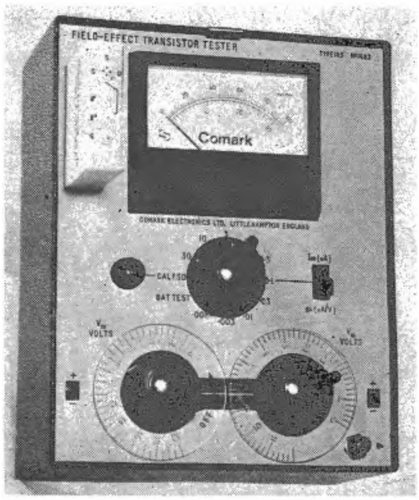
ELECTRONIC ENGINEERING will occupy stand number 453 at the BRITISH ELECTRONIC COMPONENTS SHOW to be held at Olympia, London from 23 to 26 May and visitors will be welcome

F.E.T. TESTER

Comark Electronics Ltd, Gloucester Road, Littlehampton, Sussex

(Illustrated below)

The field-effect transistor tester manufactured by Comark Electronics Ltd, is a battery powered portable instrument for measuring the static and dynamic



characteristics of f.e.t.'s and m.o.s.t.'s.

It will measure: drain current, pinch-off voltage and mutual conductance at various settings of drain and gate voltage or current. The drain and gate breakdown voltage may also be checked.

The polarities of the drain and gate supplies are separately reversible so that both enhancement and depletion f.e.t. and m.o.s.t. of -n channel and -p channel may be tested.

A simple and self-explanatory front panel layout facilitates operation of the instrument. It may be used for component evaluation, quality control checks, field service, goods inwards inspection, component selection and many other applications demanding the determination of f.e.t. and m.o.s.t. parameters.

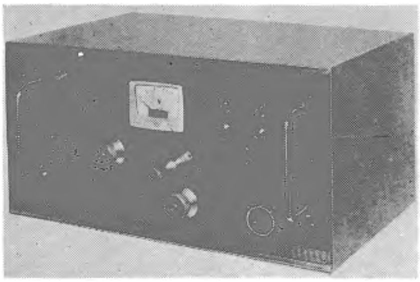
Overload protection has been built-in to reduce the risk of damage to the tester and tested component through misuse.

The drain voltage is Zener stabilized and may be adjusted to any voltage between 0 and ±20V, with an accuracy of ±5 per cent. Drain currents from 1μA (f.s.d.) to 30mA may be measured to an accuracy of ±3 per cent in ten ranges. The gate voltage is variable from 0 to ±11V and may be set to an accuracy of ±3 per cent.

The sum of the drain and gate potentials may be connected across the device to measure drain-gate breakdown.

Mutual conductance is measured using an internally generated 500Hz square wave. Eight ranges permit measurements from 10μA/V(f.s.d.) to 30mA/V, with an accuracy to ±5 per cent.

For more information circle **No. 293**



MICROWAVE NOISE MEASURING EQUIPMENT

James Scott (Electronic Engineering) Ltd, Carntyne Industrial Estate, Glasgow, E.2

(Illustrated above)

James Scott (Electronic Engineering) Ltd, has announced a new addition to its range of microwave noise measuring equipment—the single channel microwave series, type 60.

Recent developments in the field of solid state sources and general purpose klystrons have brought with them the need for an assessment of the purity of their c.w. outputs, since this factor is of importance in some of their applications. The purity of the c.w. outputs will be affected by the random noise originating in the sources and from imperfections in the power supplies serving them, causing amplitude and frequency modulation of the outputs.

This equipment, type 60, will measure amplitude modulation of a carrier expressed in terms of signal-to-noise ratio, and frequency modulation of a carrier expressed in terms of the r.m.s. frequency deviation produced in a 70Hz bandwidth of noise (alternative bandwidths available are 30Hz, 100Hz, 500Hz. It will also measure a.m. signal-to-noise ratios of the order of 126dB to 130dB and, in the absence of overriding a.m. noise, f.m., r.m.s. deviations down to values of the order of ½Hz in a 70Hz bandwidth.

It is to be noted that this equipment is particularly suitable for the evaluation of solid state sources having f.m. noise and spurious modulations of a high order (e.g. tens of Hz r.m.s. deviation).

The equipment is currently available in X-band and J-band versions denoted type 60X and 60J covering the frequency ranges 8·2 to 12·4GHz and 12·4GHz to 18·0GHz respectively.

For more information circle **No. 294**

INVERTOR

Industrial Instruments Ltd, Stanley Road, Bromley, Kent

(Illustrated above)

Industrial Instruments Ltd now offer 1 250VA static invertors, which have been designed for arduous industrial or military applications.

This 'Transipack' inverts a 24V d.c. input and produces a simulated mains output of 240V a.c. (sinusoidal) at 50Hz. It is of rugged construction and has undergone rigorous trials at Service units during its development period.

Features of the electrical specification are the incorporation of protection against short-circuit and reversal of input polarity, frequency stability of better than plus or minus one per cent, a regulated output and the ability to accept a wide input voltage swing. The power circuit is so designed as to enable satisfactory starting of motors of poor power factor.

For more information circle **No. 295**

U.H.F. SWITCHED ATTENUATOR

Marconi Instruments Ltd, St. Albans, Hertfordshire

(Illustrated below)

The u.h.f. switched attenuator type TF2163 is intended for use over the

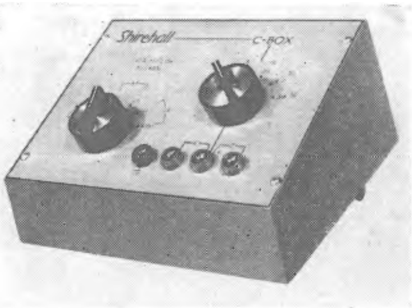
frequency range d.c. to 1GHz. It has a characteristic impedance of 50Ω and an attenuation range of 142dB in 1dB steps. This facility—142dB up to 1GHz—is believed to be unique.

The attenuator consists of two castings, in a module case, which have been carefully dimensioned to maintain the characteristic impedance. This has resulted in a v.s.w.r. better than 1·1 : 1 up to 200MHz, 1·25 up to 500MHz and 1·5 up to 1GHz.

Accuracy and freedom from standing waves has been achieved by the use of a novel design in which special resistive pads are selected by cam-operated microswitches. Careful attention to internal screening prevents errors due to leakage fields at high attentuation.

A programmable version of this attenuator, type TF 2163/1M2, is also being developed. It provides remote selection, controlled by parallel b.c.d. input. In this model, the microswitches are driven by electromagnetic actuators.

For more information circle **No. 296**



CAPACITANCE BOX

Shirehall Electronics Ltd, Station Yard, Borough Green, Sevenoaks, Kent

(Illustrated above)

The Shirehall 'C-Box' provides quick selection of capacitance, in values ranging from 100pF to 5 000μF. With only two controls, one for the significant figure and one for the tens multiple, it is very easy to use and read the value selected.

From 100pF to 6 800pF, the internal capacitors are polyester film and foil construction while from 0·01μF to 0·68μF they are metallized film. Both types have a tolerance of 5 per cent and are rated at 400V d.c. From 1μF to 5 000μF, they are electrolytic with the usual tolerance of −20 per cent to +50 per cent and rated at 15V d.c.

The C-Box is intended as a general purpose aid for design and development engineers and should find wide application in the selection of optimum values for filters, timing circuits, feedback networks, servo loops, etc. The electrolytic capacitors extend the range of application to decoupling and power-supply requirements.

The C-Box is fitted with an 'earthing' terminal, connected to the metal case for screening purposes, and supplies complete with flexible leads terminated in versatile plug/socket connectors.

For more information circle **No. 297**

BLOOD FLOWMETERS

Distributed by: S.E. Laboratories (Engineering) Ltd, Astronaut House, Feltham, Middlesex

(Illustrated below)

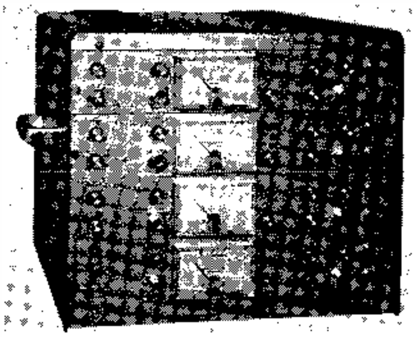
A range of electromagnetic blood flowmeters and electronic units manufactured by Biotronex Laboratory Inc. are now marketed in the U.K. by S.E. Laboratories (Engineering) Ltd.

Both slotted flow transducers and cannulating types are available in many sizes. They contain a small iron cored magnet which is excited from the main amplifier unit, and small electrodes which are mounted in the wall of the transducer.

The BL410 multi-channel console contains a power supply and filtering for operating up to four plug-in flow amplifiers. It also has two oscillators (one at 400Hz, the other 1 000Hz), which provide a sine wave drive signal for the flowmeter elements. From one to four plug-in flow amplifier units can be fitted with either 400Hz or 1 000Hz as the carrier centre frequency. This prevents transducer interference where the measurement sites are in close proximity.

Outputs are available with switched low-pass filters covering mean flow, as well as offering steps of 20Hz response, up to the maximum of 100Hz.

The flow range of the amplifiers is from less than 2ml/min to well over 20l/min according to the lumen diameter of the flowmeter.



For more information circle **No. 298**

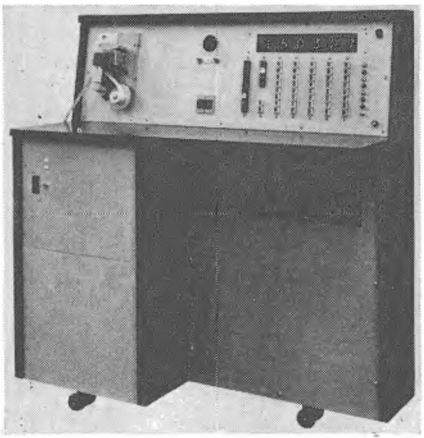
GENERAL PURPOSE COMPUTER

Digico Ltd, 10 Albion Road, St. Albans, Hertfordshire

(Illustrated above right)

The DIGIAC is a British low cost general-purpose digital computer using a magnetic drum store memory. Specially developed for scientific, engineering, and educational applications and suitable for general computation, it is particularly valuable as a laboratory instrument in controlling experimental systems or pilot plants and for data collection, reduction and analysis. The computer may be operated 'on-line' to the user's own equipment, through interfaces easily devised by the user and plugged into the DIGIAC.

The instruction set of the DIGIAC is sufficiently logical and comprehensive for it to be simple to program directly in octal machine code. A machine oriented assembler code further simplifies programming, especially where used in conjunction with an input-output



typewriter. A library of tested programs, mathematical sub-routines and test routines is constantly expanding, and is available to all users.

The DIGIAC is housed in a small console complete with desk-top for books and papers, yet the unit is easy to wheel from one laboratory to another. The computer is ready to operate as soon as it is plugged in, no installation or air-conditioning being necessary.

Some of the more important features are as follows:

All arithmetic is performed on 16 bit two's complement numbers. The circuits are based on silicon planar transistors and metal oxide resistors, to produce a system with high noise immunity that operates correctly in a wide range of environments. Exceptionally low cost has been achieved through the use of serial arithmetic and by the incorporation of a magnetic drum memory. Use of advanced logical techniques has made it possible to minimise the drawbacks of such inherently slow devices.

Any one of the four main registers can be selected for display on large numerical indicators. If the computer is stopped, the contents of the selected register can be changed by touching the push-buttons.

Both the register display and setting-in buttons are in octal, which is far more convenient than working in pure binary. The buttons on the control panel enable the operator to check through a program one instruction at a time, or to obey the instruction currently in the I register. These facilities are of great value during program development.

An instruction may directly access any one work in the standard 4 096 word store, thus avoiding the complications of indirect accessing.

This register contributes greatly to efficiency and simplicity of operation, especially when handling large data blocks.

Most of the instructions not requiring an address may be grouped so that two or more instructions are specified in a single program word. In this way, about 200 useful instructions are available.

Input/output instructions do not need to wait for the completion of the mechanical cycle of the input/output device. Times are thereby reduced, permitting the time-sharing of slow peripherals.

For more information circle **No. 299**

MICROWAVE SWITCH

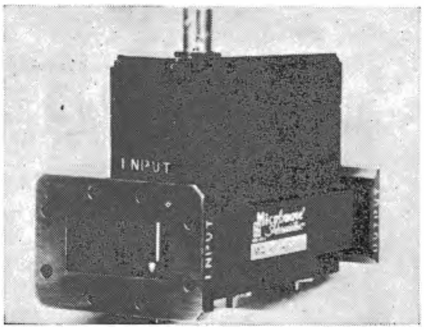
Microwave Associates Ltd, Cradock Road, Luton, Bedfordshire

(Illustrated below)

Microwave Associates Ltd has developed a totally new diode switch with a high performance and reliability rating for use in microwave communications systems in the 6·5 to 7·2GHz frequency range. It has a low bias power of about 0·25W, a very high isolation of 90dB, insertion loss of 1·0dB and a v.s.w.r. of 1·15.

The MA8 319-1C3 switch will be used primarily for hot-standby operation of which two r.f. transmitters, tuned to the same frequency, are switched into a single transmit antenna. When one equipment fails, the MA8319-1C3 automatically switches operation to the standby transmitter.

Its capacity to reliably switch up to 7W of c.w. microwave power in 500nsec provides this solid state switch with a considerable performance advantage over previous electromechanical switches, since its inherently high switching speed removes the possibility of switch failure in r.f. circuits during maintenance switching or conditions of path fading.



For more information circle **No. 300**

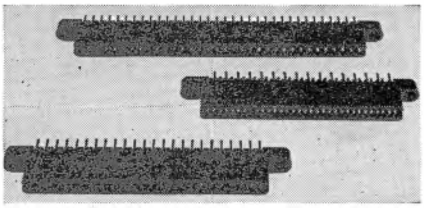
PRINTED CIRCUIT EDGE CONNECTORS

Belling and Lee Ltd, Great Cambridge Road, Enfield, Middlesex

(Illustrated below)

Three new printed circuit edge connectors have been added by Belling & Lee Ltd to their slim-line range. These are L1740 a 24 pole connector, and the L1741, a 34 pole, both on 0·15in centres and the L1486 32 pole connector on 0·1in centres.

These connectors are moulded in a black nylon filled phenolic resin and are only 0·312in wide. The domed contacts are of beryllium copper, heat treated to



ensure consistent and minimum wear with low contact resistance. The contacts, either gold or silver plated, are removeable. The contact posts are tubular in construction to facilitate dip soldering on to a printed circuit mother board.

Current rating is 3A per contact with printed wiring and 4A per contact with free wiring. Breakdown voltage between contacts is greater than 3kV d.c. The contact resistance is about 2·5mΩ.

For more information circle **No. 301**

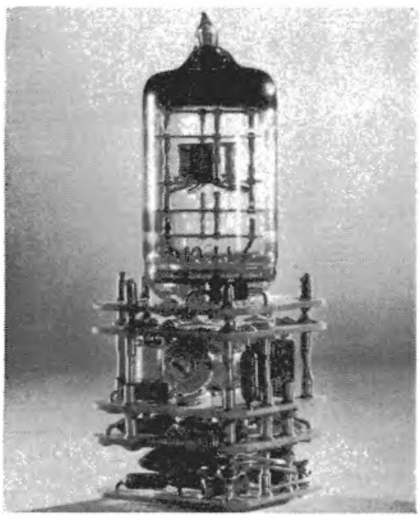
MINIATURE STABILIZED OSCILLATORS

The Marconi Co. Ltd, Chelmsford, Essex

(Illustrated below)

A new temperature stabilization technique, employing a microelectronic circuit, is embodied in a new range of miniature master oscillators announced by Marconi's Specialized Components Division. The oscillators, which have a short-term stability of 1 part in 10^8, have applications in airborne equipment and portable man-pack receivers employing the most advanced methods of radio communication.

The first model, type F3180, is designed to operate in a very wide temperature range, $-55°C$ to $+90°C$, and can be supplied pre-set to any frequency between 10 and 15MHz. In order to ensure that the power consumption of the oscillator is low, the whole crystal



assembly has been designed so that a minimum of heating is required to keep the temperature stable. It is achieved by using a tiny quartz crystal, about 3/16in in diameter, cold-welded into a transistor can, together with a very sensitive microcircuit acting as a heating element. The transistor can is in turn contained in an evacuated glass envelope to insulate the circuit from the temperature at which the equipment operates.

The microelectronic circuit is used purely for stabilizing the temperature of the crystal. If the surrounding atmosphere becomes colder, the circuit conducts more and heats up, compensating for the change outside.

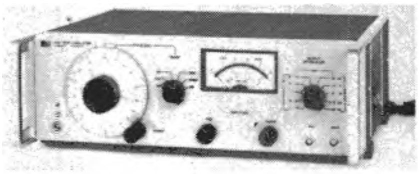
For more information circle **No. 302**

TEST OSCILLATOR

Hewlett Packard Ltd, 224 Bath Road, Slough, Buckinghamshire

(Illustrated below)

Covering the spectrum from 10Hz to 10MHz in six continuously-variable ranges, the new improved solid-state Hewlett-Packard model 651B test oscillator has outstanding amplitude stability. A dual amplitude control (coarse and fine) makes amplitude adjustments easier and more precise.



Short-term frequency stability of the new instrument, in normal laboratory environments, is typically $\pm 0·02$ per cent/22 hours, even at the least stable frequencies. Similarly, typical amplitude stability is $\pm 0·1$ per cent/17 hours.

Model 651B is housed in a Hewlett-Packard rack-convertible module, measuring 5¼in in height. It is 16¾in wide, 13¼in deep, and weighs 17lb.

The frequency response is flat within ± 2 per cent from 100Hz to 1MHz. The dial is accurate within ± 2 per cent, from 100Hz to 1MHz. Available output is 200mW into 50Ω, 16mW into 600Ω, or 6·32V open-circuit. Either output is controlled by a 90dB attenuator in 10dB steps (± 1 per cent). The model 651B with 75Ω output impedance is available on special order.

Model 651B was designed for fast, accurate measurements of television amplifiers, audio amplifiers, filter networks, tuned circuits, and telephone and telegraphic carrier equipment.

Two options are available. With option 01, the 651B's output monitor top scale is calibrated in dBm/600Ω, bottom scale calibrated in volts. With option 02, the 651B output is available in 75Ω and 600Ω.

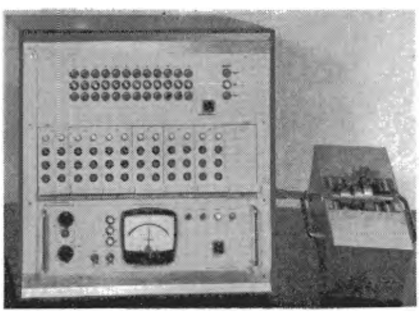
For more information circle **No. 303**

AUTOMATIC GAUGING EQUIPMENT

The Rank Organization, Taylor Hobson Division, Phoenix Works, Great West Road, Brentford, Middlesex

(Illustrated below)

New high speed electronic automatic gauging equipment giving a checking rate of 40 dimensions a second, is announced by The Rank Organization's Taylor Hobson Division.



It is called the Mitronic Multigauge and has been developed to meet a demand for a rapid and accurate system of inspection for use by unskilled labour.

The system can be made to customer requirements and will check, at one setting, any number of dimensions on a component in an inspection fixture. Results are given by light signals, meter readings and by external output. This latter facility can also be used for machine or sorting control, or to operate print-out or statistical display facilities for quality control.

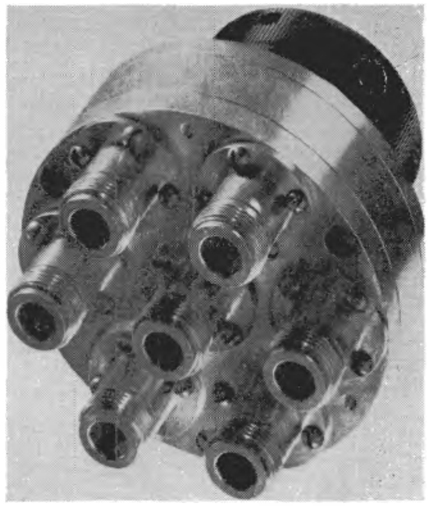
Settings are made from a standard workpiece of known dimensions and tolerances can be checked and, if necessary, adjusted at any time.

Side acting, axial and miniature axial gauging heads can be used singly or in combination to measure length, internal and external diameters, angle of tapers, parallelism and concentricity.

In operation, a master comparator circuit is sequentially switched to inspect each dimension as measured by one or more of the gauge heads. If any dimension is outside the set tolerance this is displayed.

Any three full scale tolerance ranges can be provided between $\pm 0·0003$ and $\pm 0·020$in or metric equivalents. The appropriate one for each dimension is selected on the programme pin board.

For more information circle **No. 304**



MULTI-PORT COAXIAL SWITCH

Decca Radar Ltd, Decca House, Albert Embankment, London, S.E.1

(Illustrated above)

Decca Radar Ltd has added a multiport coaxial switch to its range of microwave switches. It covers the frequency range d.c. to 12·4GHz and can be used for switching one input into any one of up to six outputs. Type N connectors are standard.

High isolation, generally better than 60dB, low insertion loss and good v.s.w.r. characteristics are features of this switch which is suitable for uses in development, test and system engineering, particularly where broadband performance is required.

The switch is compact, robust and

designed to meet military specifications; and has a long life expectancy. This is achieved by minimizing friction and incorporating automatic adjustment to ensure constant contact pressure over long periods of operation.

Both manual and automatic models are available. The automatic model uses a demountable solenoid actuator to give a fast operating time. The standard supply is 28V d.c. and no holding current is required.

For more information circle **No. 305**

DIGITAL COUNTERS

Distributed by: Britec Ltd, 17 Charing Cross Road, London, W.C.2

(Illustrated below)

'Elesta' offers a comprehensive range of electronic counters with compatible accessories and modules, enabling the construction of complete digital systems to count, measure, control and time.

Especially designed for industrial applications, these counters are uniform in format and have standardized input/output characteristics; they are fully transistorized instruments of modern industrial construction using silicon semiconductors. Their maximum counting

frequency is 100kHz, and they incorporate neon numerical tubes giving an in-line digital display. They have a pulse-forming input stage with gate facilities, and have manual, external or automatic reset. They are suitable for an ambient temperature range of 0° to +55°C and for operation from 220 to 240V a.c. supplies. The standard types are:

Type 'CP' Batch Counters: Compact single-preselecting counters, with 3- or 4-decades, in a metal housing, for bench top or for panel mounting, measuring only 7¼in wide × 3½in high × 10in deep. They have a solid state output for relays or additional transistor logic. The scope of these instruments can be greatly extended with additional 'Elesta' modular units. Typical applications are: Batch counting on automatic packing machines, cutting to length, blending of liquids, machine control, winding and positioning.

Type 'CPT' Universal Preselecting Counters: These instruments, with 4- or 5-decades, are available with single or with double preselection in a housing 13in × 3½in × 10in. They offer the same facilities as types 'CP' and in addition can be supplied with a print-out outlet. The double preselection version gives a pre-batch signal for starting or braking control. Both types 'CP' and 'CPT' can be supplied in a housing 4in wider to incorporate a quartz standard with decade outputs from 10Hz to 100kHz for accurate timing, or with a 4- or 5-decade unit for additional preselection.

Type 'CM' Measuring Counters: 4- or 5-decade interval counters with a crystal-controlled time base for revolutions per minute, frequency and slip measurement. A 2- or 3-decade switch permits incremental adjustment of the 10μsec to 100sec gating time so that the counter reads directly in the units required. The input gate may be controlled externally making possible measurement of ratio of two frequencies, slip measurement and determination of extension. The standard housing is 13in × 3½in × 10in, but a 17in wide case is available with preselection facilities, or for 19in rack mounting.

Accessories include photo-diode and photo-electric sensors, inductive proximity detectors, cable head amplifiers for distances up to 300ft between sensor and counter, plug-in printed-circuit modules consisting of a start-stop flip-flop to control gate, a contact-bounce suppressor, a 1 000Hz oscillator, a monostable output for double-preselection, and as 4in modular units: a 10Hz to 100kHz quartz standard, 4- or 5-decade auxiliary preselection unit, 4-channel co-incidence input stage, and a 5-decade electromagnetic counter as a batch totalizer or to extend the count.

For more information circle **No. 306**

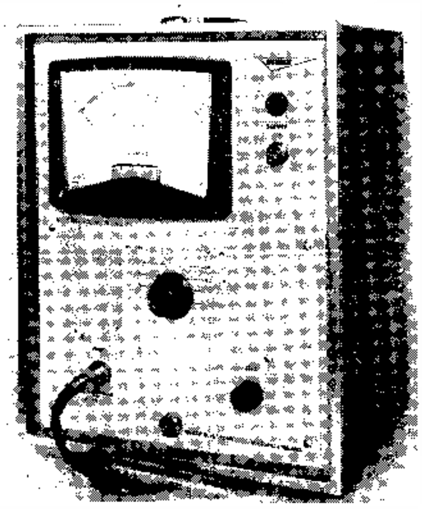
V.H.F. MILLIVOLTMETER

Dymar Electronics Ltd, Rembrandt House, Whippendell Road, Watford, Hertfordshire

(Illustrated below)

The type 711 v.h.f. millivoltmeter extends the a.c. measuring capability of the Dymar System into the v.h.f. and u.h.f. frequencies. It will measure with ease and accuracy voltages as low as 300μV over a frequency range of 50kHz to 850MHz.

It features a lightweight and compact r.f. probe which houses two high speed rectifying diodes in a full-wave circuit. The resulting low level d.c. voltage is amplified in a low noise chopper stabilized d.c.-amplifier. The rectifying characteristic of these diodes is such that below approximately 30mV they follow a square law thus giving true r.m.s. response. Special linearizing circuits are used on each range which make the meter reading linear between ⅓ f.s.d. and full scale.



A full range of accessories can be supplied which extend the versatility of the instrument. These include a T-connector and 50Ω load for measurements on coaxial circuits, a probe to B.N.C. adaptor, and also a high impedance capacitive divider for measuring voltages up to 300V.

For more information circle **No. 307**

VARIABLE CAPACITORS

Wingrove and Rogers Ltd, Polar Division, 75 Uxbridge Road, Ealing, London, W.5

(Illustrated below)

Specially designed for printed circuit mounting is a new piston-type trimmer introduced by Wingrove & Rogers Ltd, makers of the range of 'Polar' variable capacitors.

The new trimmer, type S60-01, has an overall diameter of 0·25in with 0·2in fixing centres and is variable from 2pF to 25pF by means of a screwdriver slot in the rotor. Overall height at minimum capacitance is 0·928in and at maximum 0·678in. It has a polystrene dielectric, but for operation in extremes of tem-

perature, a p.t.f.e. dielectric can be supplied.

Another new type is the CG80-03, a miniature, three-section, ganged capacitor designed for v.h.f. receivers. Each section has a capacitance of from 2·5pF to 17pF and the vanes are spaced at 0·12in. An internal gear and pinion drive gives a 3 : 1 reduction, and includes a robust end stop. High-grade ceramic posts insulate the stator from the frame, and the capacitor is 1·5in long, 0·65in wide and 0·85in high overall.

For more information circle **No. 308**

RING CURRENT TRANSFORMERS

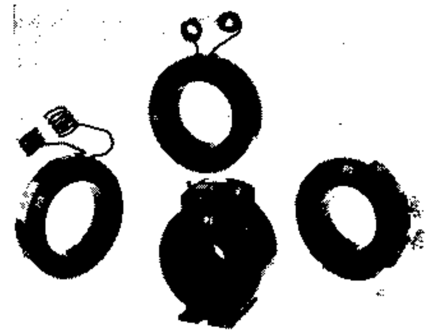
Atkins, Robertson & Whiteford Ltd, Industrial Estate, Thornliebank, Glasgow

(Illustrated on page 337)

Ring current transformers to the requirements of BSS.3938(1965) are available from Atkins, Robertson & Whiteford Ltd. To ensure quick delivery and competitive prices standard ranges have been designed covering the majority of metering applications.

Ratios range from 50/5 to 4 000/5 and internal diameters from 1¼in to 4½in. Standard burdens are 2·5, 5, 10 and 15VA. Alternative mounting arrangements are available and secondary terminations can be either wire ends or terminal lugs. Transformers with 1A secondaries can also be supplied.

For exposed situations or for applications where adverse atmospheric conditions may be encountered users may select from a range of current transformers cast in epoxy resin. In addition



to the protection provided by the epoxy resin covering the appearance of the transformers is greatly enhanced.

Standards of accuracy are strictly controlled by the use of precision test equipment and certificates of compliance with BSS.3938 are supplied with every order.

Enquiries are invited for Protective Current Transformers or for other types not included in the standard ranges.

For more information circle **No. 309**

MULTI-METER

Avo Ltd, Avocet House, Dover, Kent

(Illustrated below)

An improved version of its panclimatic Avometer is announced by Avo (MI Group).

The new model, which has been granted Ministry of Defence approval (to specification DEF-155), is designated Avo multirange test set No. 1. It will replace the previous models 8SX, 9SX and 40SX which for some years have been extensively used in the Armed Services.

The new version, like its predecessors, is specially designed for use in conditions of extreme temperature or high humidity.

The multirange test set No. 1 retains design features of earlier Avometers, including mechanical cut-out, movement reverse button and interlocked range switching. The 5in scale has markings for both voltage and current measurements, with a separate scale for resistance measurements.

Additional overload protection is provided by a fuse connected in series with the ohms range. Components are selected to ensure stability of calibration.

The instrument is supplied with an ever-ready carrying case, a pair of rubber-covered test leads, two spring slips and a pair of test prods.

For more information circle **No. 310**

BATCH COUNTER

Distributed by: M.L. Industrial Products Ltd, 292 Leigh Road, Trading Estate, Slough, Buckinghamshire

(Illustrated below)

The CB 23 is a preset batch counter designed and manufactured by Thorn Electronics Ltd, and marketed in the U.K. by ML Industrial Products Ltd.

The CB 23 is designed to withstand the heavy wear and tear of industrial usage, and is housed in a rugged steel case which is sealed to prevent the ingress of dust or moisture. The circuit is completely transistorized and employs printed circuit boards.

Batch counts of up to 9 999 can be made by presetting the selector switches. The count is made from the selected figure down to zero, and this makes it possible to arrange a pre-batch warning pulse where required. The normal count rate is up to 10kHz, but under special

circumstances speeds of up to 50kHz can be achieved.

The output is via a relay with double-pole changeover contacts rated at 450VA. Re-set is automatic, with manual override, and there is no loss of count during the re-set operation.

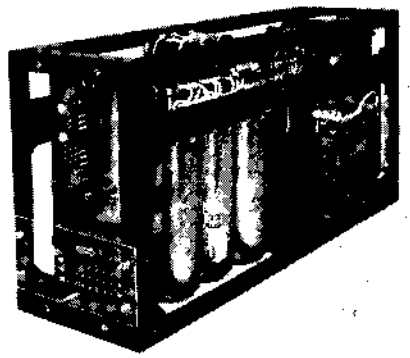
For more information circle **No. 311**

D.C. POWER SUPPLY

International Electronics Ltd, 132-135 Sloane Street, London, S.W.1

(Illustrated above right)

The DS.369 is an extension of the 'I.E.' range of modular d.c. power supply units. This all silicon, high efficiency unit gives an output of 10V at 10A suitable for use with integrated circuits. Designed to operate up to 65°C ambient with wide mains input tolerance, the unit continues to operate within specification for 20msec after the interruption of the input supply. Type approved components, run within their RCS ratings, used where possible and highly reliable computer grade electrolytic capacitors give a high m.t.b.f. The unit measures 14in × 6¾in × 4¾in.



For more information circle **No. 312**

PIRANI-PENNING GAUGE

Edwards High Vacuum Ltd, Manor Royal, Crawley, Sussex

(Illustrated below)

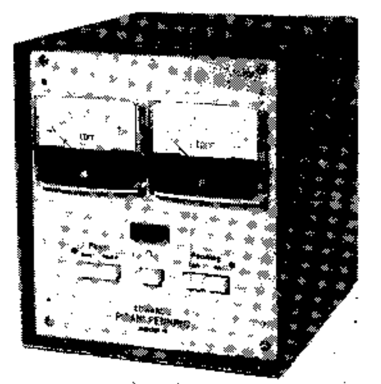
Edwards' new model 4 Pirani-Penning gauge combines a Penning gauge with a Pirani gauge to provide in one unit a versatile, wide-range instrument suitable for general purpose vacuum applications. The overall range of the instrument is 3 torr to 10^{-6} torr made up as follows:

Pirani: 3 to 10^{-3} torr
Penning: 10^{-2} to 10^{-5} torr—range HIGH
2×10^{-5} to 10^{-6} torr—range LOW

One model 6 Penning and two series 6B Pirani gauge heads can be accommodated covering the full vacuum measurement requirements of the normal high vacuum pumping system. The Penning head would normally be sited on the vacuum chamber, with one Pirani head on the rough pumping line and the other Pirani head on the backing line. However, a wide variety of alternative arrangements can be covered with this basic head array.

Range changing and gauge head selection are provided by push buttons on the front panel. Voltage stabilized transistor circuits eliminate the need for manual voltage adjustment and pressure is indicated on two clear-front meters calibrated directly in torr.

An important feature for process control applications is the provision of output terminals which can accommodate two recorders and two Edwards VS9-1 relay units on the Pirani section and one Edwards VS9-2 relay unit on the Penning section all of which can work



concurrently but independently of each other.

For more information circle **No. 313**

STABILIZED POWER SUPPLIES

Kingshill Electronic Products Ltd. Torrens Works, Torrens Street, London, E.C.1

(Illustrated above)

The 'Compact' series of laboratory stabilized power supplies by Kingshill consists of five all-silicon units measuring only 7¾in × 6½in × 6½in overall, with maximum ratings of 18V 1A, 36V 1A, 50V ½A, 100V ¼A and 150V 150mA.

All units are similar in performance and operation, having coarse and fine voltage controls giving continuous adjustment from zero to maximum. Protection against short-circuit and overloads is by means of an adjustable current limiter which is fitted with an output shorting push button for easy setting, independent of output voltage. A 3in high accuracy volt-ammeter with one current range and two voltage ranges is provided, selection of voltage/current is by means of a slider switch, the voltmeter reading being doubled at the touch of a push-button.

Standard output is two terminal; 4 terminal output for remote sensing and constant current operation is also available if required. Brief specification. Stabilization ratio >1 000, regulation 0·1 per cent, ripple and noise <1mV, mains tolerance ±10 per cent, maximum ambient temperature 45°C.

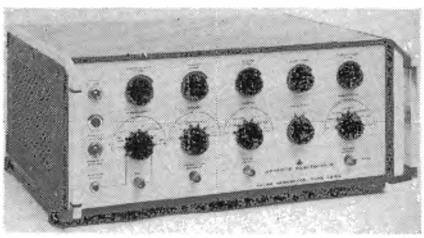
For more information circle **No. 314**

PULSE GENERATOR

Advance Electronics Ltd. Roebuck Road, Hainault, Ilford, Essex

(Illustrated below)

Advance Electronics Ltd announce a new high power pulse generator—PG55—which gives a 50V output into 50Ω and is fully protected against short-circuit and open-circuit output conditions. The output rise time is variable between 12nsec and 100nsec and width and delay are variable between 100nsec



and 100msec. The PG55 can be externally triggered and the output can be gated by an external negative pulse. Up to 1A can be delivered into low impedance loads and a double pulse output is provided at frequencies up to 500kHz.

The repetition rate is continuously variable from 1Hz to 5MHz in decade steps with a 10 : 1 fine control. A push-button is provided for single-shot control.

For more information circle **No. 315**

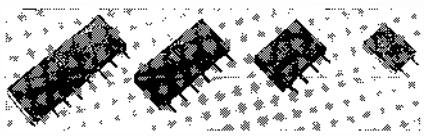
MINIATURE RESISTOR MODULES

A.B. Metal Products Ltd. North West House, 119-127 Marylebone Road, London, N.W.1

(Illustrated below)

These compact resistor modules are produced by screening formulations of conductive and resistive materials on to an alumina substrate. After firing, the resistive and conductive elements become an integral part of the substrate and so resistant to vibration and shock. The high overload capacity afforded by the excellent heat sink qualities of the substrate, prevents temperature high spots.

The modules are available as 2, 4, 6 or 8 pin designs with single resistors or networks or wide ohmic range. Standard tolerance ±5 per cent (±1 per cent available): power dissipation is $25W/in^2$ at 70°C ambient temperature, derated to no load as 125°C; temperature coefficient ±300 in 10^6/°C (standard), lower coefficients available—can be matched to ±25 in 10^6/°C on individual substrates; high stability <½ per cent change as a result of humidity or thermal shock. Standardized dimensions permit magazine loading and automatic insertion; 0·125in pin spacing.



For more information circle **No. 316**

CARRIER FREQUENCY AMPLIFIER

Vibro-Meter S.A., British Office: Haletop Cevio Centre, Wythenshawe, Manchester, 22

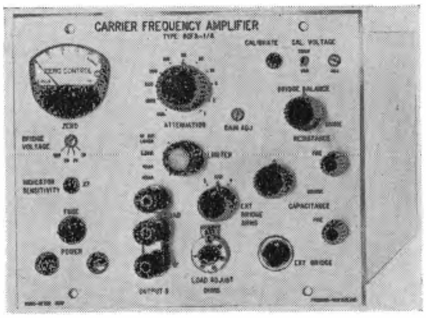
(Illustrated above right)

Vibro-Meter SA, the Swiss instrument company, has introduced a new version of its 8kHz carrier frequency amplifier system. The new instrument type number 8-CFA-1/A, is fully transistorized and is constructed from module units for ease of maintenance.

It is designed for use with half or full bridge strain gauge circuits, resistive and inductive transducers. Vibro-meter transducers are available for the measurement of pressure, load, displacement, acceleration, force and torque.

The amplifier has a frequency response between 0 to 1 800Hz, the bridge voltage is adjustable between 1 and 10V, the zero stability is 0·1 per cent f.s.d. per day and a similar figure is quoted for gain stability. The instrument has a maximum gain of 8 000. Output is 40mA into 100Ω.

Although calibration of the complete measurement system is usually made by static calibration of the transducer, the 8-CFA-1/A amplifier includes a built-in calibration facility by which the amplifier gain can be re-set or checked. The new version of the 8kHz carrier frequency amplifier can be operated either from the 240V 50Hz mains supply or from a 24V d.c. battery. The instrument



is particularly suitable therefore for use in motor vehicles, for example, road tests on torque transmission systems etc.

A notable feature is the output circuit; the output load can be adjusted and the output current can be limited to 2·5, 10 or 40mA to match the requirements of galvanometer recorders.

For more information circle **No. 317**

DELAY LINE MODULE

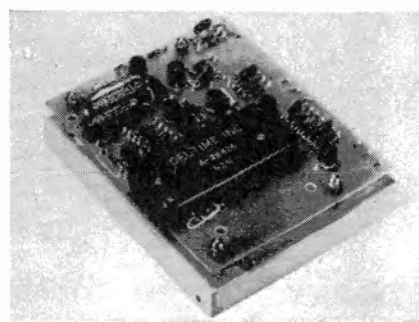
Sealectro Ltd. Walton Road, Farlington, Portsmouth, Hampshire

(Illustrated below)

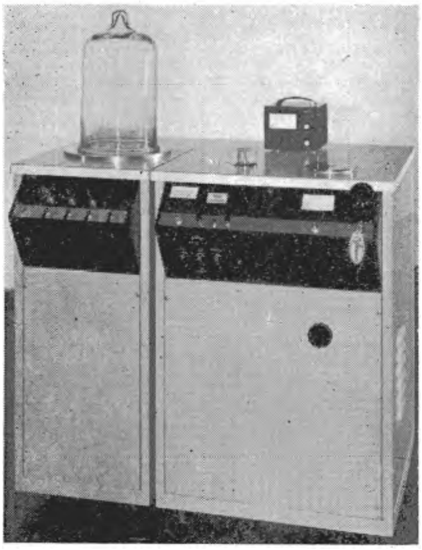
For magnetostrictive delay line applications, a unity gain, wide pulse width circuit module is available from Sealectro.

When combined with a magnetostrictive delay line, the RZN-1 solid state module provides delays up to 5 000μsec of random width pulses at repetition frequencies to 1MHz in RZ mode and 2MHz in NRZ mode. The output pulse width of the module is equal to the input pulse width within ±0·08μsec; no limit is designated on the maximum pulse width.

The RZN-1 solid state module measures 5 5/16in × 4⅜in × ⅞in. Complementary delayed output logic levels are −7·6V and the input logic level is −6·6V. The rise time is 0·075μsec and the fall time, 0·05μsec. The RZN-1 is designed to operate in temperature ranges from 0°C to 55°C. The input power requirements are 12V d.c. For delays up to 1 000μsec model 190A delay line is recommended; for delays to 2 000μsec, model 192A; for delays to 5 000μsec, model 214A.



For more information circle **No. 318**



LEAK DETECTOR

Veeco Instruments Ltd, 21 Aston Road, Waterlooville, Portsmouth, Hampshire

(Illustrated above)

A new helium mass spectrometer leak detector, the MS9/VBC has been announced by Veeco Instruments Ltd. The instrument extends the application of the firm's standard line of mass spectrometer leak detectors to provide a facility for leak testing, backfilling, and sealing devices such as relays, switches, gyros, and aircraft instruments.

Devices may be evacuated, backfilled with controllable amounts of gas and leak tested. Furthermore, the completely sealed and backfilled unit may be leak tested so that the problem of checking for leaks at the final seal or 'pinch-off' is overcome.

The helium tracer gas is detected by a specially tuned 60° magnetic sector mass spectrometer and leak rates of less than 1×10^{-10}cm^3/sec may be determined.

The instrument features automatic operation complete with safety interlocks which fully protect it against misuse. A knowledge of high vacuum technique is not necessary to effect a quick and accurate leak test of any device.

For more information circle **No. 319**

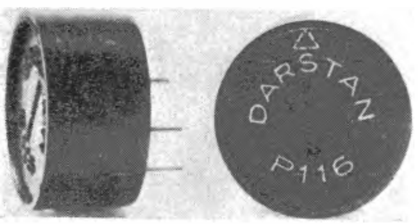
WIREWOUND POTENTIOMETER

Darstan Ltd, 263 Church Road, Thundersley, Essex

(Illustrated below)

The P116 is a wirewound component which is complementary to the P109, and is 8·2mm high and overall diameter of 16mm. Electrically similar to the P109 but with snap type lid which eliminates the aluminium can.

Redesigned wiper, and terminations



together with fully sealed base offer greater robustness, and the component is suitable where either vibration is evident or a 'set and forget' requirement is needed. The leads (4·5mm × 0·82mm diameter) are on 0·1in module for printed circuits.

Available in resistance range 5Ω to 20kΩ. Centre threaded and plain stud versions are also available and shortly Darstan will announce a spindle version.

For more information circle **No. 320**

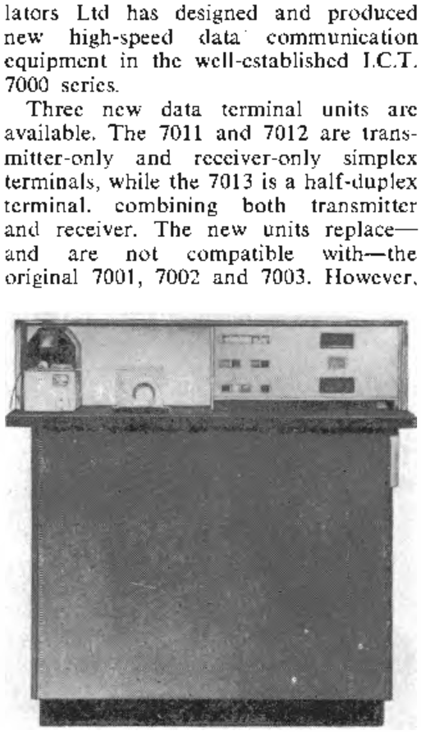
DATALINK EQUIPMENT

International Computer and Tabulators Ltd, ICT House, Putney, London, S.W.15

(Illustrated below)

International Computers and Tabulators Ltd has designed and produced new high-speed data communication equipment in the well-established I.C.T. 7000 series.

Three new data terminal units are available. The 7011 and 7012 are transmitter-only and receiver-only simplex terminals, while the 7013 is a half-duplex terminal, combining both transmitter and receiver. The new units replace—and are not compatible with—the original 7001, 7002 and 7003. However,



the new units are compatible in that a data transmission network can be built up of any mix of the three terminals. Moreover, they can readily be connected on-line to a 1900 series computer, thus allowing direct data communication with remote installations.

Using existing switched or private telephone lines, I.C.T. 7000 datalink equipment is designed to transmit data very accurately at high speed. Information punched in 5, 6, 7, or 8-track paper tape at the transmitting end is automatically reproduced in paper tape at the receiving end. Transmission speeds of 600 or 1 200 bits/sec can be selected by a two-position switch, giving data throughput of 60 or 110 characters/sec.

Data are transmitted in 80-character blocks and, in the event of a transmission error, punching is inhibited and the block of data concerned is automatically re-transmitted. As a result, the output tape at the receiving end is always error-free.

For more information circle **No. 321**

ELECTROLUMINESCENT NUMERICAL INDICATORS

Thorn Special Products Ltd, Great Cambridge Road, Enfield, Middlesex

Designed for use singly or in banks as a modern digital read-out, the electrolumisescent numerical indicator by Thorn Special Products will reproduce green numerals from 0 to 9 and several letters of the alphabet. Only 9/32in thick, with a height of 2⅜in and wdith of 1 13/16in, the indicator has a current consumption of 2·5mA at a nominal supply of 115V a.c. 400Hz.

The use of electroluminescent segments directly behind a transparent face eliminates numeral distortion problems encountered in some conventional indicators and allows the numerals to be read over a viewing angle of 160°. The proven reliability of the electroluminescent lamp ensures trouble free operation over many thousands of hours.

The segmented design of the indicator enables the letters A, C, E, F, H, J, L, P, U to be illuminated. In addition the letters G, I, O and S may be lit by using the figures 6, 1, 0 and 5 but the possibilities of confusion should be considered before use.

The indicator is housed in either a clear or green polycarbonate case and weighs approximately one ounce. Eight solder pins provide wiring connexions to the segments.

For more information circle **No. 322**

Meetings THIS Month

THE INSTITUTION OF ELECTRICAL ENGINEERS

All evening meetings will be held at Savoy Place, London W.C.2, commencing at 5.30 p.m. (tea at 5 p.m.), unless otherwise stated.

Date: 1 May.
Discussion: Electronic Aspects of Chemico-Physical Instrumentation.
(Further details to be announced in IEE News).

Date: 1 May.
Colloquium: Applications of Computers to Field Analysis.
(Time to be announced in IEE News).

Date: 2 May.
Lecture: Mathematics in Control. (Second Annual Lecture of the Control and Automation Division).
By: M. J. Lighthill.

Date: 2-4 May.
Conference: Integrated Circuits. (International Conference).
Held at: Eastbourne.

Date: 3 May.
Visit: Technical Visit to Bowaters U.K. Pulp and Paper Mills Ltd, Ellesmere Port, Cheshire.
(Further details to be announced in IEE News).

Date: 4 May.
Lecture: Application of Transistor Techniques to Relays and Protection for Power Systems.
By: F. L. Hamilton, M. Legg and J. B. Patrickson.

Date: 4 May.
Discussion: An Automatic Stroboscope for Measuring Machine Losses.
Opened by: J. Greig.

Date: 8 May.
Discussion: Are Aircraft Electrics too Complicated?
Opened By: H. Zeffert.
Held at: The Royal Aeronautical Society, 4 Hamilton Place, London W.1.
(At 6 p.m., tea at 5.30 p.m.).

Date: 9 May.
Lecture: High-Frequency Errors in Manual Tracking.
By: G. W. Lange and P. H. Hammond.
(Joint meeting with the Automatic Control Group of the I.Mech.E.).

Date: 10 May.
Lecture: Results of Multiple Carrier Experiments with a t.w.t. Amplifier Operating near to Saturation.
By: R. J. Westcott.

Date: 10 May.
Discussion: The Education of Computer Engineers.
(Joint meeting with the I.E.R.E. Computer Group).
Held at: The London School of Hygiene and Tropical Medicine, Keppel Street, London W.C.1.
Time: 6 p.m. (Tea at 5.30 p.m.).

Date: 11 May.
Lecture: Canals as Cable Routes.
By: E. P. C. Watson, E. J. Brooks and C. H. Gosling.

Date: 11 May.
Discussion: The Influence of Changes in School Curricula on the Intake of Engineering Courses.
(Further details to be announced in IEE News).

Date: 15 May.
Discussion: Development of Electrical Accessories in the U.K. and a Review of Continental Wiring Practice.
Opened by: E. J. Sutton and another speaker.
Lecture: Factors Affecting the Design of Electrical Accessories.
By: E. J. Sutton.
Paper: Electrical Installation Practice in Europe.
By the late D. A. Picken.

Date: 15 May.
Discussion: Developments in Transformer Ratio Arm Bridges.
Opened by: A. H. Silcocks.

Date: 16 May.
Discussion: E.H.D.
Opened by: W. D. Allen.

Date: 17 May.
Discussion: Recent Developments in Medical Thermography.
Opened by: K. Lloyd Williams, D. Shawe and M. Gross.
(Joint meeting with the I.E.R.E. Medical and Biological Group).
Held at: The I.E.R.E., 9 Bedford Square, London, W.C.1.
Time: 6 p.m. (tea at 5.30 p.m.).

Date: 17 May.
Lecture: Developments in Power-System Control.
By: P. F. Gunning and H. E. Pulsford.

Date: 18 May.
Meeting: Annual General Meeting.
Followed at about 6.30 p.m. by a technical film show lasting approximately one hour.

Date: 19 May.
Colloquium: Advances in Measurements brought about by recently introduced Semiconductor Devices.
Time: 9.30 a.m.
(All wishing to attend must apply to the Secretary for a ticket. No fee for IEE members: non-members £1 5s. 0d.).

Date: 22-24 May.
Conference: Frequency Generation and Control for Radio Systems.

Date: 24 May.
Visit: Technical Visit to British Railways Research Dept., Derby.
(Further details to be announced in IEE News).

Date: 24 May.
Lecture: The Postal Service and the Electronics Engineer.
By: J. Piggot and T. Pilling.
Time: 6 p.m. (tea at 5.30 p.m.).

Date: 25 May.
Colloquium: Computer Solution of Waveguide Problems.
Time: 2.30 p.m.

INSTITUTION OF ELECTRONIC AND RADIO ENGINEERS

Date: 10 May.
Time: 6 p.m.
Held at: London School of Hygiene and Tropical Medicine, Keppel Street, London, W.C.1.
Discussion: The Education of Computer Engineers.
(Joint meeting with I.E.R.E.-I.E.E. Computer Groups).

Date: 11 May.
Time: 6 p.m.
Held at: Institution of Electronic and Radio Engineers 8-9 Bedford Square, London, W.C.1.
Lecture: The Scientific Information Problem: New Methods for Retrospective and Current Awareness in the Face of a Constantly Expanding Literature.
By: A. E. Cawkell.

Date: 12 May.
Time: 6 p.m.
Held at: The Edward Lewis Lecture Theatre, Middlesex Hospital Medical School, Cleveland Street, London, W.1.
Symposium: Electronic Engineering Training for the Royal Electrical and Mechanical Engineers.
By: H. G. Frost and members of the Staff of R.E.M.E., Arborfield.

Date: 17 May.
Time: 6 p.m.
Held at: Institution of Electronic and Radio Engineers, 8-9 Bedford Square, London, W.C.1.
Symposium: Recent Developments in Medical Thermography.
By: K. Lloyd Williams, D. Shaw and M. Gross.
(Joint with I.E.R.E.-I.E.E. Medical and Biological Electronics Group).

Date: 24 May.
Time: 5.30 p.m.
Held at: Institution of Electronic and Radio Engineers, 8-9 Bedford Square, London, W.C.1.
Symposium: Television Network Switching at the Post Office Tower.
By: R. W. Fenton, H. F. Lloyd, H. Mirzwinski, R. G. Moore, E. D. Probert and J. G. Thomasson.

South Western Section

Date: 2 May.
Time: 7 p.m.
Held at: Plymouth College of Technology.
Meeting: Microelectronics.
Opened by: S. S. Forte.
(Joint with the Western Centre of the I.E.E. Electronics Section).

Thames Valley Section

Date: 23 May.
Time: 7.30 p.m.
Held at: J. J. Thomson Physical Laboratory, University of Reading.
Lecture: Astronomical Instrumentation.
By: P. B. Fellgett.

THE INSTITUTION OF POST OFFICE ELECTRICAL ENGINEERS
(Formal Meetings)

Date: 16 May.
Time: 5 p.m.
Held at: The Institution of Electrical Engineers, Savoy Place, London, W.C.2.
Meeting: Communication Systems—Research and Development in the Post Office.
Opened by: W. J. Bray.
(This meeting will be preceded by the Annual General Meeting of the Institution, commencing at 5.00 p.m.).

THE ROYAL TELEVISION SOCIETY

Date: 19 May.
Annual Dinner and Dance.
Held at: The Dorchester Hotel, London.
(Admission by ticket only).

Publications Received

PAL COLOUR TV is a new publication by Mullard Ltd.

Briefly, the book describes a complete hybrid dual-standard PAL colour receiver, as well as giving a general introduction to the subject.

Chapters are devoted to the vision and sound i.f. amplifiers; the luminance a.g.c. and synchronizing circuits; colour decoding circuits; a shunt stabilized line timebase circuit; a field timebase circuit; convergence and raster corrections circuits; coils and circuits for automatic degaussing; and power supply circuits.

Items of circuity features are: 405-line monochrome—625-line colour; hybrid luminance amplifier; mean beam-current limiting; true signal level a.g.c.; hybrid colour processing circuits; 25in 90 degree scanning; dual-standard convergence; and automatic chrominance control.

The price of the book is 12s 6d. Radio and television dealers may obtain it through wholesalers. Alternatively, it may be ordered (post free to the trade) from Home Trade Sales Division, Mullard Ltd, Mullard House, Torrington Place, London, W.C.1. Cash should be sent with orders.

THE TRANSFORMER RATIO-ARM BRIDGE is the subject of Wayne Kerr Monograph No. 1. This monograph describes in detail the theory of the Transformer Ratio-Arm Bridge, including the extension of the range to cover very low impedances using a T-network. The text is general, without reference to any commercial instruments, and the principles given apply equally at low audio frequencies and vhf.

This type of circuit, known also as 'in situ' bridge has recently received much wider attention because of its suitability for the measurement of small-signal parameters of semiconductors and the ease with which extremely fine discrimination can be achieved by means of external standards for 'backing-off'.

Enquiries for this monograph should be addressed to the Wayne Kerr Co., Ltd, Sycamore Grove, New Malden, Surrey.

ENGLISH ELECTRIC VALVES 1967 provides a guide to EEV products correct at the time of going to press. In addition to abridged data on the range of valves and accessories manufactured at Chelmsford and Lincoln an equivalents index complete with CV numbers. Requests for this brochure should be made, on company headed notepaper to, English Electric Valve Co. Ltd, Chelmsford, Essex.

ELECTRONIC COMPONENTS is a new short form catalogue illustrating the range of components manufactured by A.B. Metal Products Ltd including 'Smart and Brown Connectors' and 'Wolsey Electronics'. Copies are available on request from, A.B. Metal Products Ltd, Abercynon, Glamorgan.

THE HEAT TREATMENT SERVICE AT TIPTON is a new brochure which sums up briefly the facilities afforded by the Wild Barfield Heat-Treatment Division at Tipton. The service now available to the large number of manufacturers and users of tools and dies situated within a reasonable distance is directed to curb costs in two ways—by a reduction in finishing costs of tools and by an increase in tool life resulting from scientifically controlled heat-treatment.

The equipment installed and dealt with in the brochure represents the latest advances in plant for commercial heat-treatment operations, and is backed by a well-equipped metallurgical laboratory, the services of which are also available on a contract basis. Copies of this brochure are available from Wild Barfield Ltd, Elecform Works, Otterspool Way, Watford By-Pass, Watford, Hertfordshire.

MODERN SOLDERS is a comprehensive publication which includes over 40 illustrations and 15 reference tables.

Apart from specific information regarding Ersin and Arax Multicore Soldiers, Liquid Fluxes, Solder Preforms and printed Circuit Materials, it contains much useful technical data on solders, soldering processes and fluxes of interest to electronic and chemical executives.

Résumés des Principaux Articles

Le bruit solaire comme moyen d'assurer des diagrammes polaires verticaux de précision aux radars par M. J. B. Scanlan

Résumé de l'article aux pages 280 à 286

Cet article traite de la manière d'utiliser le soleil et d'autres étoiles comme sources de signaux pour mesurer, de façon assez simple et avec précision, les dimensions et la forme des diagrammes polaires verticaux de radars en service. Le procédé en question donne, pour la première fois jusqu'à ce jour, la mesure du radar lui-même, independamment de toute cible. Ce sont là, bien entendu, des indications précieuses tant pour l'opérateur que pour le réalisateur d'appareils de radar. Ce dernier peut, en effet, mesurer le diagramme polaire vertical objectivement et sans risque de perturbations dues à des fluctuations dans la zone d'échos de la cible. L'opérateur, quant à lui, gagne à connaître le diagramme réel du radar car il peut ainsi attribuer à la cible, en toute certitude, les variations dans les essais de vol. De plus, les essais de vol en un lieu donné, qui maintenant mesurent effectivement la cible proprement dite, peuvent être transférés rapidement et avec confiance à un autre lieu. Enfin, l'opérateur dispose aujourd'hui d'une méthode objective pour mesurer la performance à n'importe quel moment et à peu de frais.

Un circuit à filtre actif secondaire pour amplificateurs accordés et oscillateurs sinusoidaux
par E. A. Faulkner et le Vicomte Downe

Résumé de l'article aux pages 287 à 290

Il s'agit d'un simple circuit à filtre actif à trois transistors qui simule la fonction de transfert d'un circuit accordé capacitif-inductif en se passant de composants de précision. On obtient facilement une stabilité du facteur de surtension supérieure à 0.1%/°K ainsi qu'une stabilité de la fréquence de résonance de 0.02%/°K. La même stabilité de fréquence s'obtient lorsque le filtre, en liaison avec un simple limiteur, est utilisé comme oscillateur sinusoïdal.

Les paramètres du palpeur de source à excitateur de courant par P. F. Ridler

Résumé de l'article aux pages 290 à 291

Cet article décrit un amplificateur constitué d'éléments solides à gain unitaire utilisant un transistor unipolaire et un mécanisme d'excitation de courant bipolaire complémentaire. Son impédance d'entrée est extrêmement élevée et son impédance de sortie est basse, ce qui le rend approprié pour les étages d'entrée d'instruments électroniques.

L'étude et la réalisation des coupleurs de guide d'ondes directionnels Tchebycheff par K. E. Hancock

Résumé de l'article aux pages 292 à 297

Cet article indique une méthode de réalisation simple et pratique pour les coupleurs de guides d'ondes directionnels à réseau Tchebycheff. Cette méthode est basée sur l'emploi d'un seul réseau dans le guide d'ondes. Ce procédé peut être étendu à des réseaux entassés et à des rangées parallèles d'ouvertures. La disposition des ouvertures des coupleurs à plusieurs ouvertures pour guides d'ondes des formats WG10 à WG22 est indiquée.

Un compteur à décades numérique de 700Kc/s à inversion rapide par J. F. Brandwijk et P. H. Sydenham

Résume de l'article aux pages 298 à 301

Des compteurs rapides et pouvant être facilement inversés sont exigés dans l'instrumentation numérique de précision. Cet article décrit un compteur à décades constitué d'éléments solides qui peut compter jusqu'à 0,710⁶i.p.s. et, inversé, en 4μs. Le comptage s'effectue dans le code binaire "plus 3" en utilisant la réaction à couplage direct pour obtenir un comptage décimal. Tous les circuits, y compris les mécanismes d'entraînement du décodage et de la lampe d'affichage, sont décrits, ainsi que les chiffres de performance.

Le contrôle des impulsions des systèmes dynamiques par inter-corrélation par F. Jones

Résumé de l'article aux pages 310 à 315

Le principe essentiel de la méthode dite d'inter-corrélation, telle qu'elle est utilisée pour la réponse des impulsions d'un système donné, est que si l'on applique un son blanc à un système linéaire, l'inter-corrélation de l'entrée et de la sortie du système indique la réponse de ses impulsions.

L'emploi de séquences binaires pseudo-aléatoires, dont la fonction d'auto-corrélation est analogue à celle du son blanc, a rendu moins onéreuses les difficultés d'ordre pratique rencontrées jusqu'ici dans l'instrumentation, en ce qui concerne le retard, la multiplication et l'intégration.

L'auteur décrit un inter-corrélateur pratique, l'INSPECTRON' de la société SPERRY, et rattache la valeur de la réponse d'impulsion mesurée dans des systèmes secondaires et supérieurs aux méthodes de contrôle de la réponse de fréquence et de la fonction pas-à-pas.

Note sur les réseaux de compensation de la tension alternative par D. R. Wilson

Résumé de l'article aux pages 316 à 317

Cette note constitue une étude du réseau de compensation de la démodulation conçu par Whiteley. Elle montre que ce réseau peut être utilisé comme circuit de compensation asservi à tension alternative comportant quatre modes particuliers de fonctionnement, à savoir: (1) comme réseau d'établissement, (2) comme réseau à retardement, (3) comme réseau d'établissement-retardement, (4) comme réseau à retardement-établissement.

La réalisation d'amplificateurs de tension continue réduite à relais modulateur par S. K. Bhola et R. H. Murphy

Résumé de l'article aux pages 318 à 322

Les relais modulateurs sont d'usage courant dans les amplificateurs de tension continue pour réduire la dérive effective. Le rendement général d'un amplificateur de tension continue à modulation du signal par relais modulateur ne peut, cependant, être supérieur à celui du relais lui-même. Un certain nombre de relais de ce genre ont été rélaisés pour le repérage à niveau réduit mais aucun d'eux n'est exempt des problèmes de dérive, de bruit, de fréquence de fonctionnement, de durée d'utilisation, de coûtet de stabilité. Les progrès récents dans le domaine de la technologie des semi-conducteurs ont permis de mettre au point des relais modulateurs à semi-conducteurs d'une nette supériorité par rapport à la plupart des autres relais. Cet article examine l'utilisation des relais modulateurs à semi-conducteurs pour l'amplification de signaux à tension continue à niveau réduit et indique les limites fondamentales de la méthode de transformation du courant par dispositif interrupteur ou relais modulateur.

Sélecteurs à Dékatron de couplage avec transistors à bas prix par J. F. Young

Résumé de l'article aux pages 323 à 325

Afin d'accoupler des étages sélecteurs à Dékatron fonctionnant à des vitesses maximales de 20KHz, on a utilisé un circuit en cascade à transistor dont l'amplitude de tension des impulsions de sortie peut être deux fois celle de l'amplitude de tension nominale des transistors. Des transistors à haute tension et à bas prix sont utilisés pour obtenir des oscillations de tension d'impulsion de sortie allant jusqu'à 200V ce qui rend le dispositif d'un grand intérêt économique.

Zusammenfassung der wichtigsten Beiträge

Sonnenrauschen als Hilfsmittel zur Ermittlung genauer Vertikalpolardiagramme für Radars von N. J. B. Scanlan

Zusammenfassung des Beitrages auf Seite 280-284

Der Beitrag beschreibt, wie die Sonne und andere Sterne als Signalquellen ausgenutzt werden können, die es gestatten, mit verhältnismässig einfachen Mitteln und grosser Genauigkeit Grösse und Form der Vertikalpolardiagramme von Radars im eingebauten Zustand zu messen. Nach diesem Verfahren erhält man zum ersten Mal unabhängig von jedem Ziel ein Mass für das Radar selbst, was sowohl für den Radar-Entwurfsingenieur wie auch das Bedienungspersonal von grosser Bedeutung ist. Der Entwurfsingenieur kann das Vertikalpolardiagramm objektiv und ohne Störung durch Schwankungen im Zielechogebiet messen. Die Kenntnis des wahren Vertikalpolardiagramms kommt dem Bedienungspersonal zustatten, da Schwankungen in der Flugerprobung nunmehr mit Sicherheit dem Ziel zugeschrieben werden können. Ausserdem kann man Ergebnisse von Flugerprobungen über einem Gelände, die nun wirklich nur das Ziel betreffen, schnell und ohne Bedenken auf ein anderes übertragen. Darüber hinaus steht dem Personal jederzeit und mit geringem Kostenaufwand ein objektives Verfahren zur Überprüfung der Leistungskennwerte zur Verfügung.

Schaltung eines aktiven Filters zweiter Ordnung für abgestimmte Verstärker und Sinusoszillatoren
von E. A. Faulkner und Viscount Downe

Zusammenfassung des Beitrages auf Seite 287-290

Der Beitrag beschreibt eine mit drei Transistoren bestückte aktive Filterschaltung, die ohne Präzisionsbauelemente die Übertragungsfunktion einer LC-abgestimmten Schaltung simuliert, wobei ohne Schwierigkeiten eine Q-Konstanz von besser als 0,1 Prozent/°K und eine Resonanz-Frequenzkonstanz von 0,02 Prozent/°K erreicht werden. Dieselbe Frequenzkonstanz wird erreicht, wenn das Filter zusammen mit einem einfachen Begrenzer als Sinusoszillator betrieben wird.

Die Parameter eines stromgesteuerten Sourcefolgers von P. F. Ridler

Zusammenfassung des Beitrages auf Seite 290-291

Dieser Beitrag beschreibt einen Festkörperverstärker mit Verstärkungsfaktor 1, in dem ein unipolarer Transistor und komplementärer bipolarer Stromtreiber Anwendung finden. Da seine Eingangsimpedanz äusserst hoch und die Ausgangsimpedanz niedrig ist, eignet er sich als Eingangsstufe elektronischer Instrumente.

Entwurf und Fertigung von Tschebyscheff-Hohlleiter-Richtkopplern von K. E. Hancock

Zusammenfassung des Beitrages auf Seite 292-297

In diesem Beitrag wird ein einfaches, vielseitiges Entwurfsverfahren für Tschebyscheff-Strahler in Hohlleiter-Richtkopplern gegeben. Das Entwurfsverfahren beruht auf einem Einzelstrahler in der Breitseite eines Hohlleiters und wird weiter auf gestockte Strahler und Parallelreihen von Löchern ausgedehnt. Die Lochanordnung für Koppler mit Vielfachlöchern in Hohlleitergrössen WG10 und WG22 sind tabelliert.

Ein schnell umkehrbarer 700-kHz-Digitaldekadenzähler von J. F. Brandwijk und P. H. Sydenham

Zusammenfassung des Beitrages auf Seite 298-301

Für digitale Präzisionsinstrumentierung sind sehr schnell umkehrbare Schnellzähler erforderlich. Ein beschriebener Festkörper-Dekadenzähler kann bis zu 0,7 × 10⁶ Impulse je Sekunde zählen und in 4 µs umgekehrt werden. Es wird nach dem Drei-Überschuss-Binärcode mit direktgekoppelter Rückführung gezählt, um ein Dezimalresultat zu erreichen. Alle Schaltungen, einschliesslich Entschlüsselung und Treiber der Anzeigelampen, werden zusammen mit den erzielten Leistungswerten gegeben.

Impulstesten dynamischer Systeme nach dem Kreuzkorrelationsverfahren von F. Jones

Zusammenfassung des Beitrages auf Seite 310-315

Wenn weisses Rauschen an ein lineares System gelegt wird, gibt die Kreuzkorrelation des Systemeingangs und -ausgangs die Impulscharakteristik des Systems.

Die Anwendung von binären Pseudostreufolgen, deren Selbstkorrelationsfunktion annähernd weissem Rauschen entspricht, hat die bisher auftretenden praktischen Instrumentierungsschwierigkeiten in Verzögerung, Multiplikation und Integration vermindert.

Ein praktischer Kreuzkorrelator—das "Inspectron"—wird beschrieben und die Bedeutung der gemessenen Impulscharakteristik von Systemen zweiter und höherer Ordnung mit der nach dem Sprungfunktions- oder Frequenzgangverfehren ermittelten verglichen.

Bemerkungen über wechselstromkompensierende Netzwerke von D. R. Wilson

Zusammenfassung des Beitrages auf Seite 316-317

In einer Besprechung des demodulation-kompensierenden Netzwerkes von Whitley wird gezeigt, dass das Netzwerk als kompensierende Schaltung eines Wechselstromreglers Anwendung finden kann und dann vier getrennte Betriebsarten hat: (1) als Voreilnetzwerk, (2) als Verzögerungsnetzwerk, (3) als Voreil-Verzögerungsnetzwerk und (4) als Verzögerungs-Voreilnetzwerk.

Entwurf eines Chopper-Gleichstromvorverstärkers von S. K. Bhola und P. H. Murphy

Zusammenfassung des Beitrages auf Seite 318-322

Chopper finden weitgehend zur Minderung der Effektivdrift in Gleichstromverstärkern Anwendung. Die Gesamtleistungskennwerte eines Gleichstromverstärkers mit Choppermodulation des Signals können jedoch nicht besser sein als die des Choppers. Für den Nachweis niedriger Pegel wurden eine Anzahl von Chopper-Bausteinen entwickelt, die jedoch alle Nachteile in bezug auf einen oder mehrere der folgenden Parameter haben: Drift, Geräusch, Betriebsfrequenz, Lebensdauer, Kosten oder Stabilität. Jüngste Fortschritte in der Halbleitertechnologie haben zum Bau von Halbleiter-Choppern geführt, die den meisten anderen Choppern in vielen Beziehungen überlegen sind. In diesem Beitrag wird die Anwendung von Halbleiter-Choppern für die Vorverstärkung von Gleichstromsignalen besprochen und auf die grundsätzlichen Grenzen der Chopper-Technik hingewiesen.

Das Koppeln von Dekatrons mit preiswerten Transistoren von J. F. Young

Zusammenfassung des Beitrages auf Seite 323-325

Eine transistorierte Ausführung der Kaskodenschaltung, in der die Spannungsamplitude des Ausgangsimpulses fast doppelt so gross sein kann wie die zulässige Spannung der Transistoren, wird für die Kopplung von Dekatron-Stufen angewendet, die mit Geschwindigkeiten bis zu 20 kHz arbeiten. Man bedient sich preiswerter Hochspannungstransistoren, um Impuls-Ausgangsspannungshübe von bis zu 200 V zu erzielen, wodurch die Anordnung wirtschaftlich interessant wird.